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Practical Implementation of Loop Control in Power Converters

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IEEE Senior Member

Course Agenda

- Control System Basics
- Operational Amplifier and Low-Frequency Pole
- Gain-Bandwidth Impact on Phase Boost
- Op Amp Slew Rate Effects in Loop Control
- Start-Up Sequence and Auxiliary Supply
- Characterizing the Optocoupler Pole
- Dealing with the Fast Lane
- Going Around the TL431 Fast Lane

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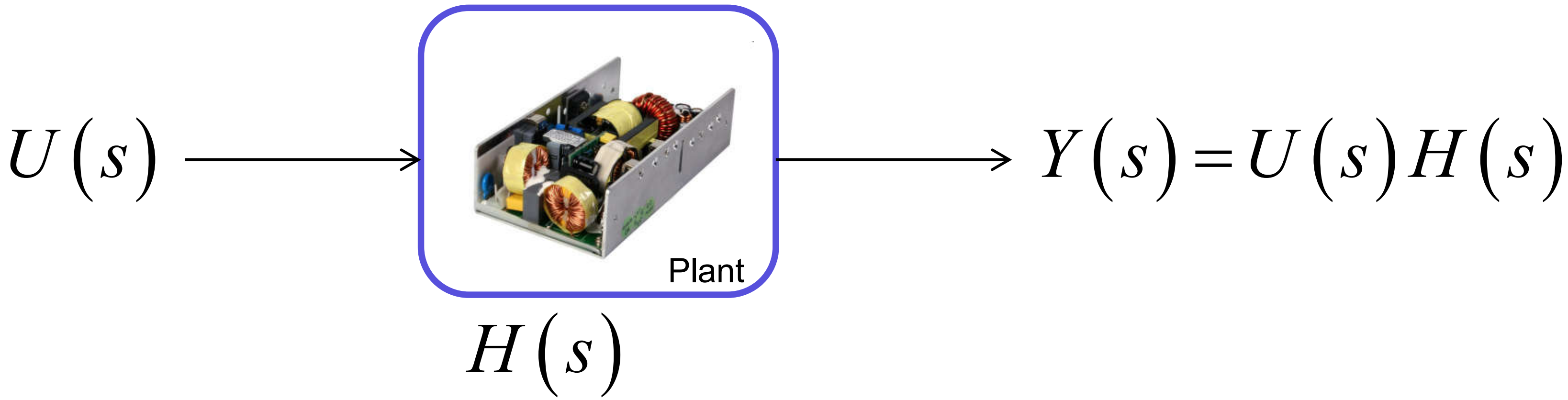
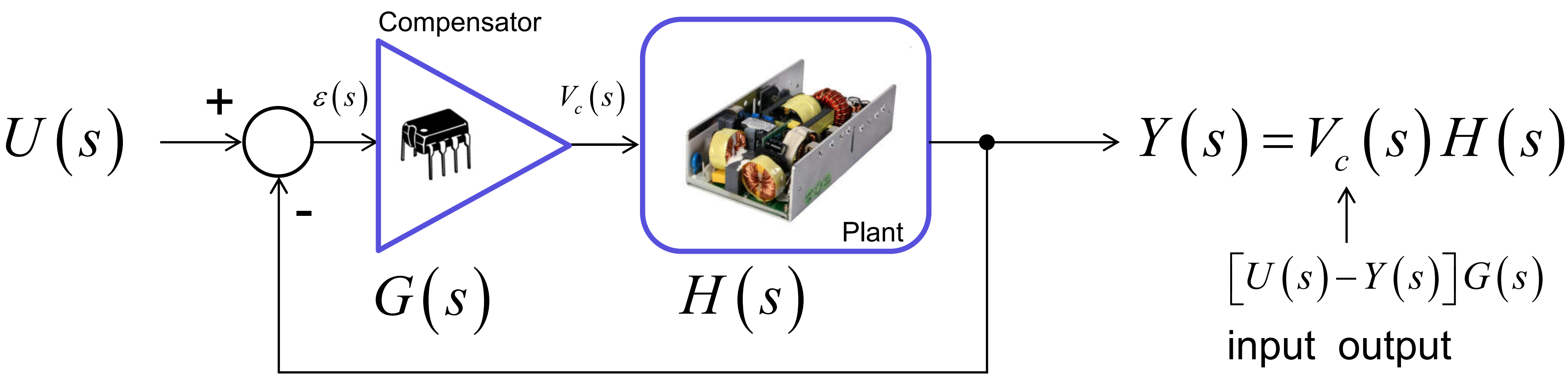
What is a Control System?

- An *open-loop* system links the output to the control variable



- A *control system* observes the output to induce correction



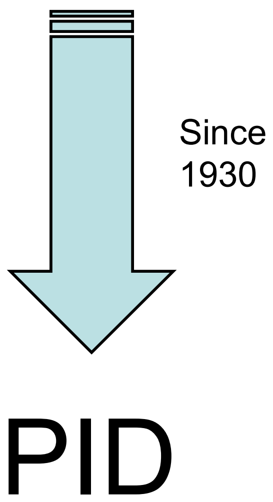
Performance of a Control System

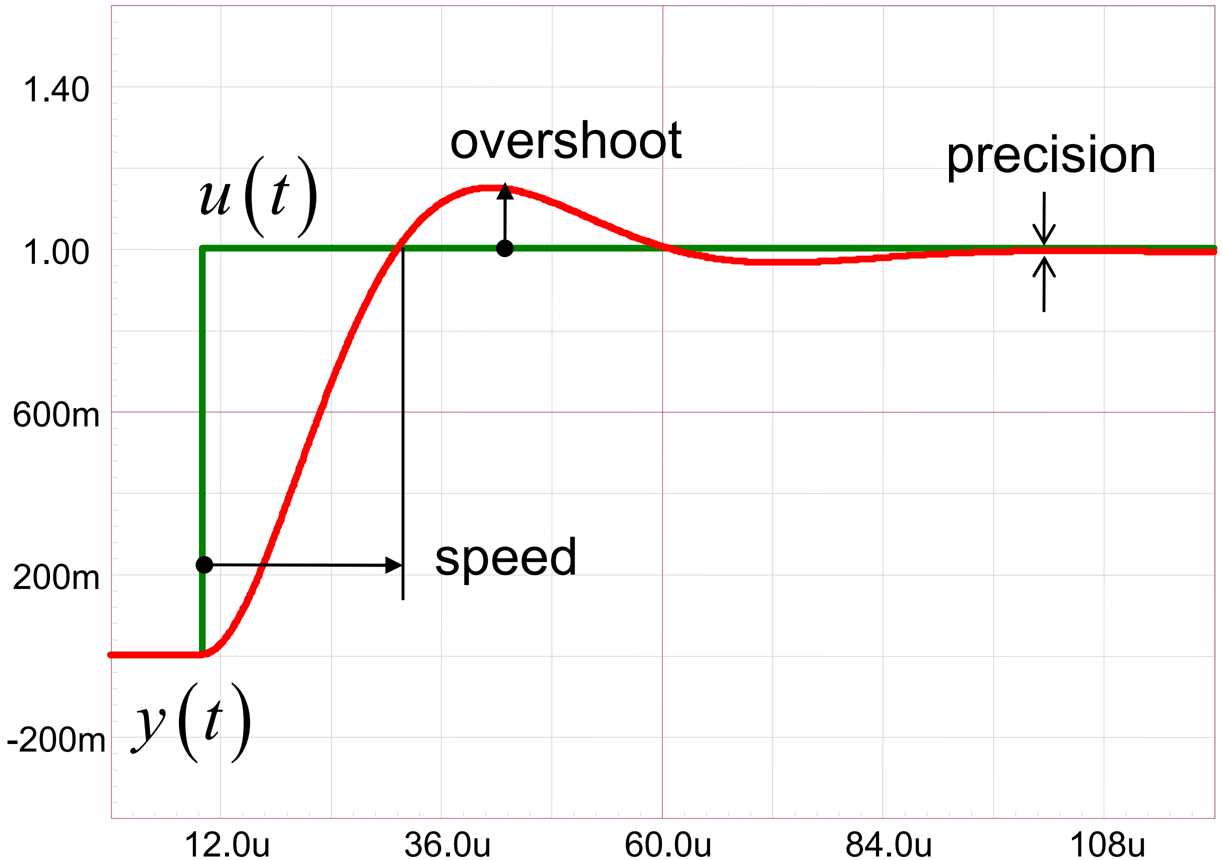
❑ We want the following operating characteristics:

✓ Speed

✓ Precision

✓ Robustness

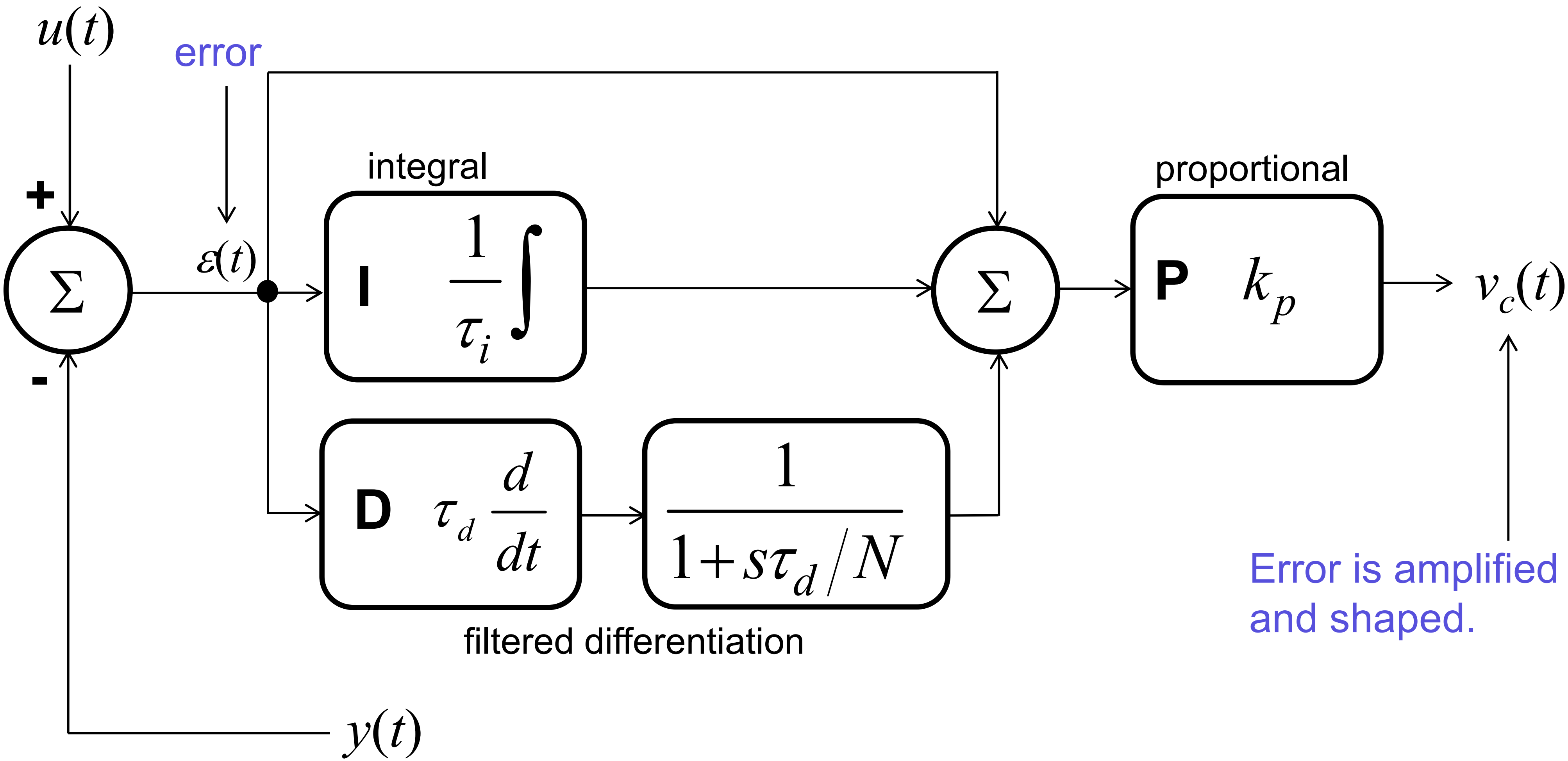




PID = Proportional Integral Derivative

Combine Three Terms for a PID

- This is the filtered standard form of the PID

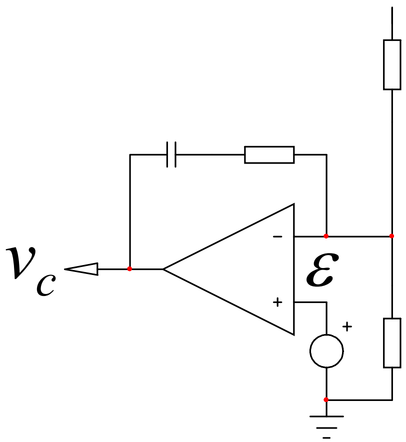


How do You Implement the PID?

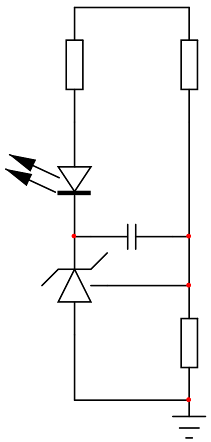
- The compensator is the place where you apply corrections

The compensator: G

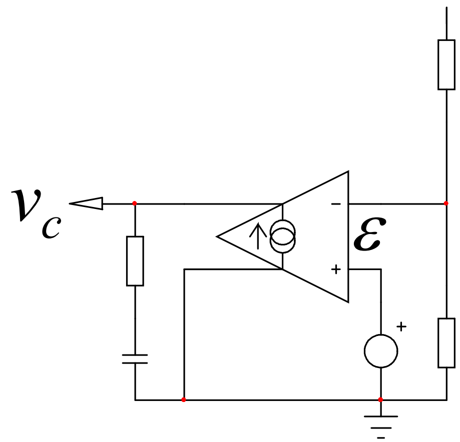
- The compensator is built with an error amplifier:



Op amp



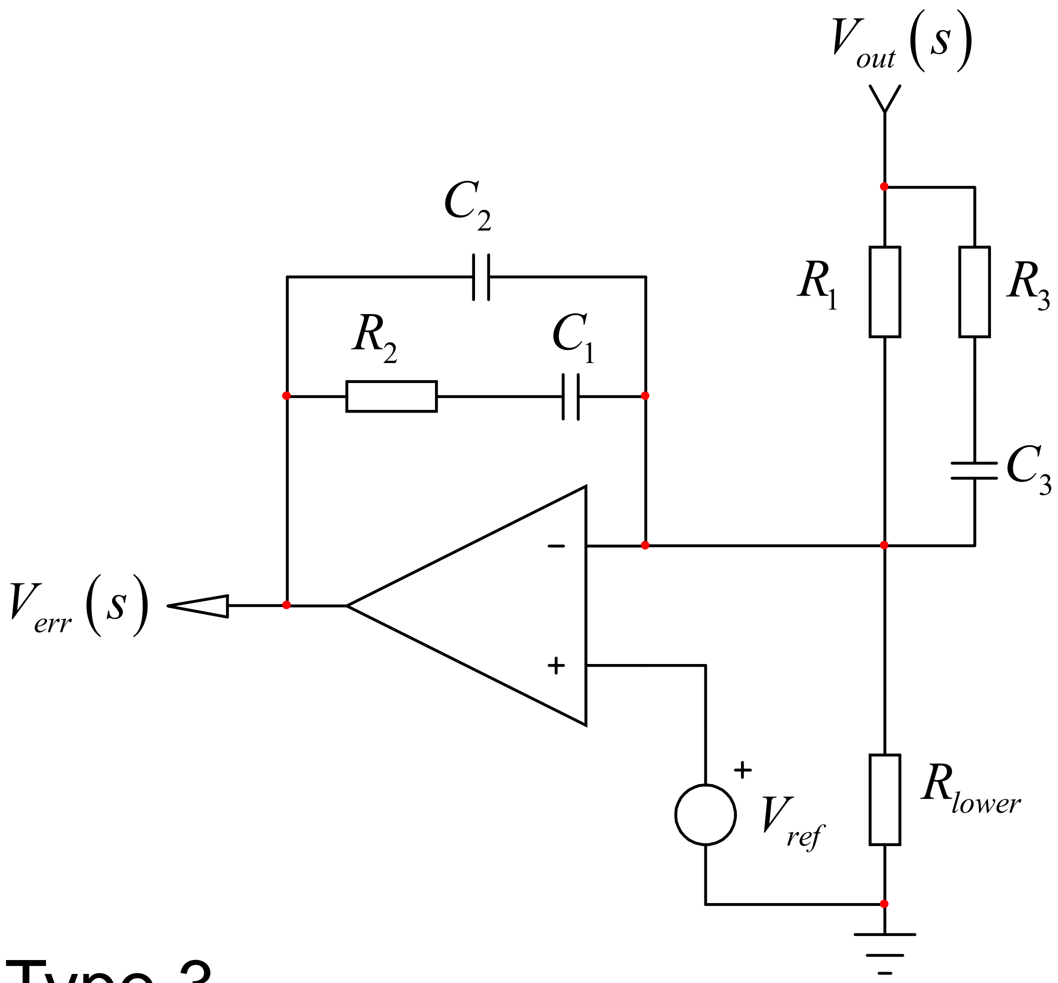
TL431



OTA

Build the PID with an Op Amp

- A type 3 compensator is a PID with an extra pole



Type 3

$$G(s) = -\frac{1 + s\left(\frac{1}{s_{z_1}} + \frac{1}{s_{z_2}}\right) + s^2\left(\frac{1}{s_{z_1}s_{z_2}}\right)}{\frac{s}{s_{po}}\left(1+\frac{s}{s_{p_1}}\right)\boxed{\left(1+\frac{s}{s_{p_2}}\right)}}$$

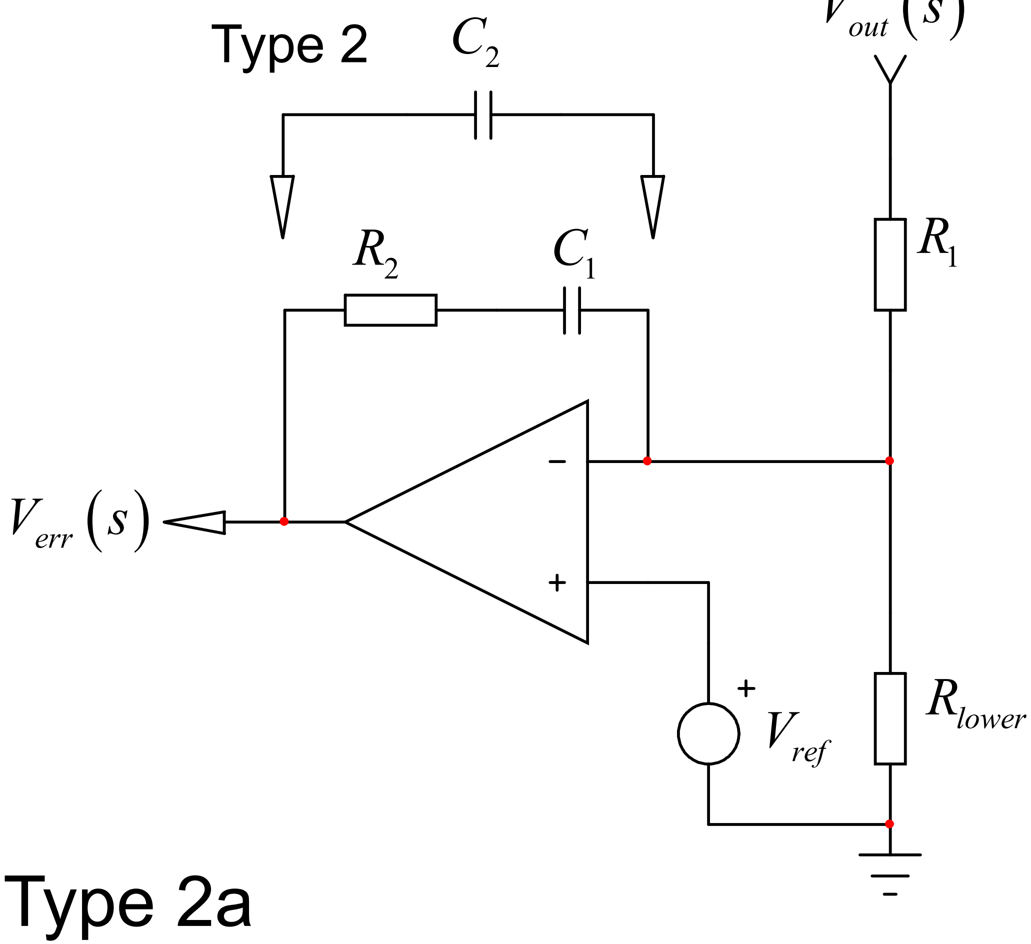
PID to type 3

Added pole

$$G(s) = -\frac{\left(1+\frac{s}{s_{z_1}}\right)\left(1+\frac{s}{s_{z_2}}\right)}{\frac{s}{s_{po}}\left(1+\frac{s}{s_{p_1}}\right)\boxed{\left(1+\frac{s}{s_{p_2}}\right)}}$$

A PI Compensator is Also Useful

- A PI version does not include a differentiation term



$$G(s) = -k_p\left(\frac{1+s\tau_i}{s\tau_i}\right)$$

PI to type 2a

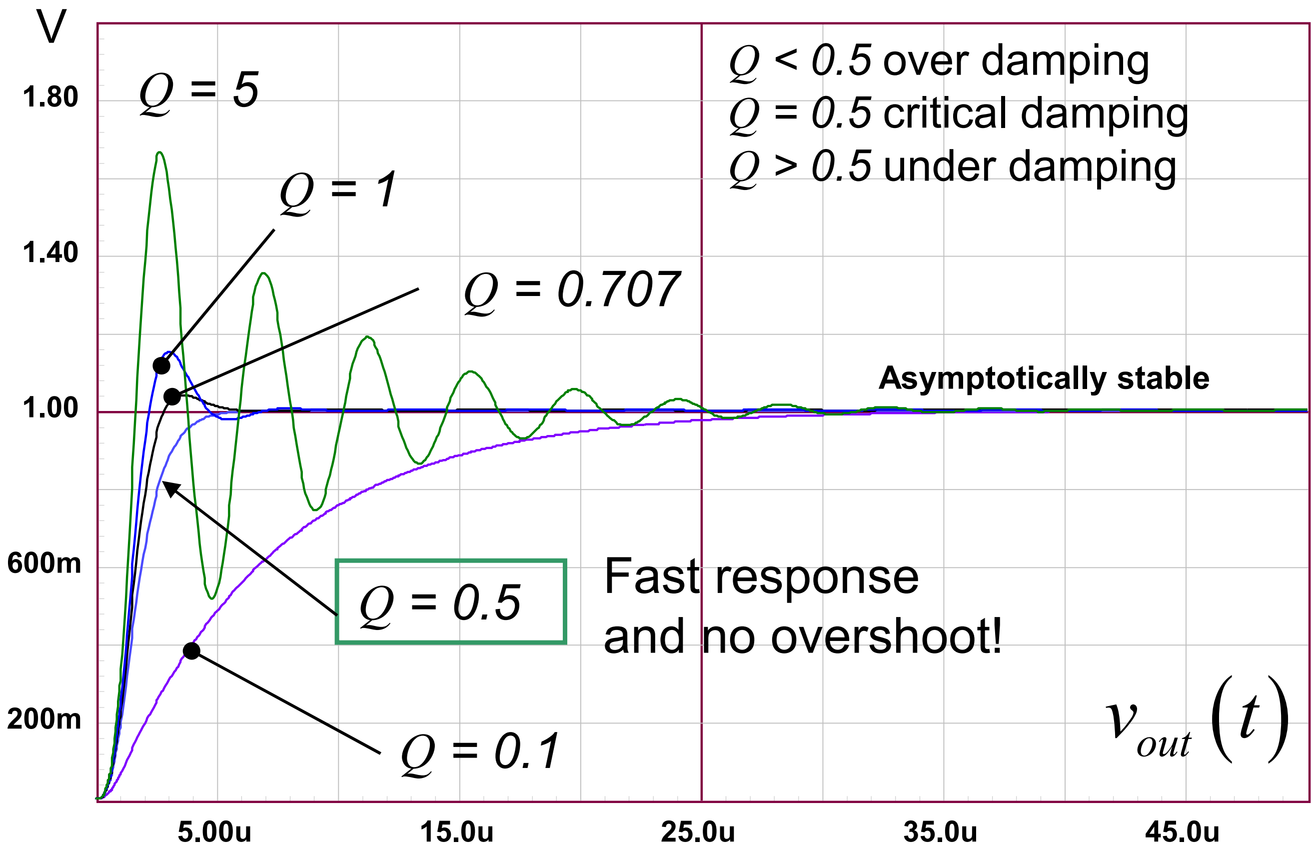
$$G(s) = -\frac{1+s/s_{z_1}}{s/s_{po}}$$

2a to type 2

$$G(s) = -\frac{1+s/s_{z_1}}{s/s_{po}\left(1+s/s_{p_1}\right)}$$

Compensation Strategy

1. Force crossover at the selected frequency f_c
2. Build phase margin at crossover
3. Ensure robustness with gain margin

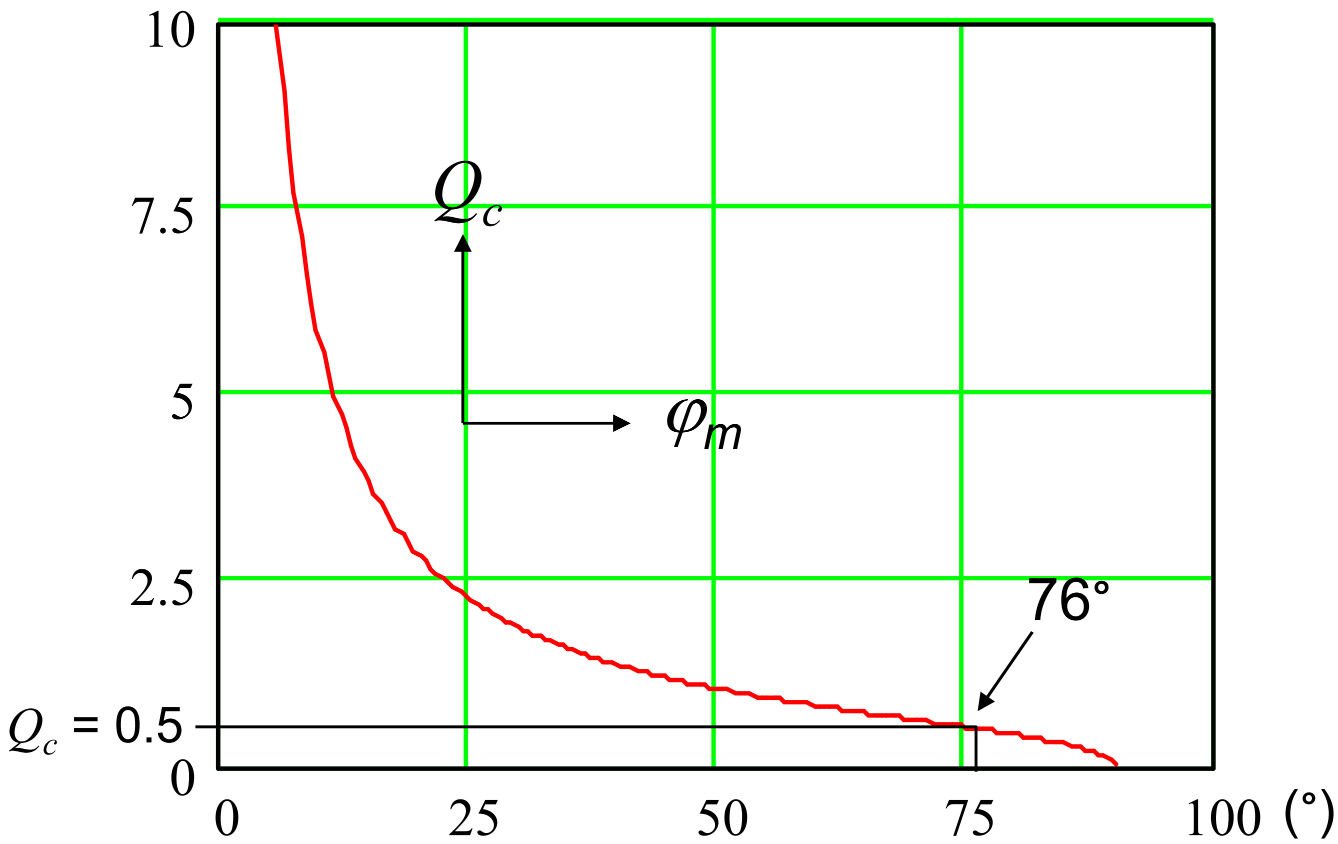


Impact transient response

$$H(s)=\frac{1}{1+\frac{s}{Q\omega_0}+\left(\frac{s}{\omega_0}\right)^2}$$

Phase Margin Affects the Response

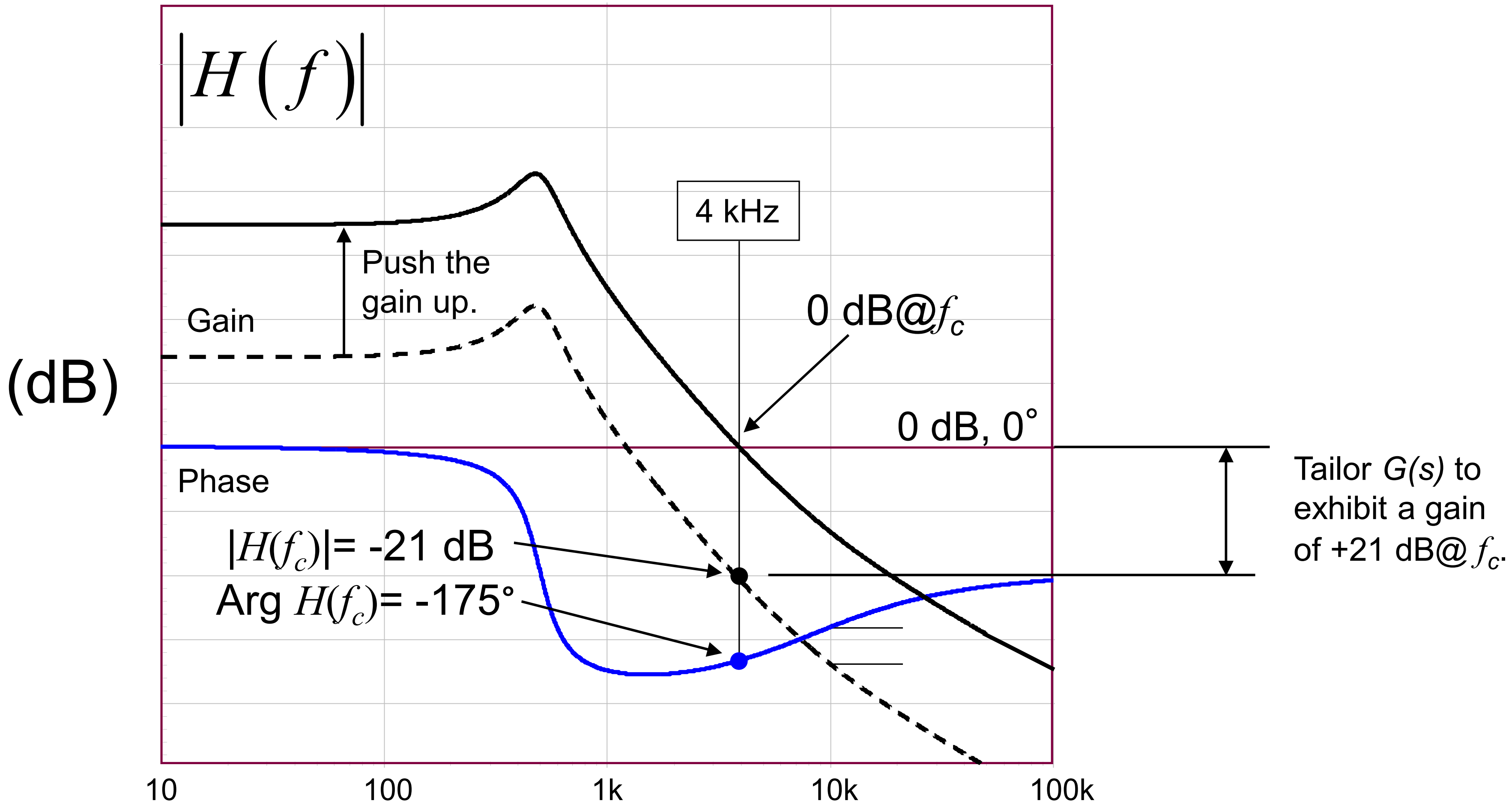
- ❑ **Open-loop** phase margin affects **closed-loop** response



- ❑ Select the phase margin based on the response you want

Imposing Crossover Frequency

- ❑ Compensate power stage gain excess or attenuation at f_c



- ❑ Tailor the compensator to provide attenuation at crossover

Combine Poles and Zeros to Boost Phase

- An transfer function where a pole and a zero are combined looks like

$$G(s)=\frac{\left(1+\frac{s}{s_{z_1}}\right)}{\left(1+\frac{s}{s_{p_1}}\right)}=\frac{N(s)}{D(s)}$$

- The argument of a quotient is: $\arg N-\arg D$

$$\arg G(f)=\arctan\left(\frac{f}{f_{z_1}}\right)-\arctan\left(\frac{f}{f_{p_1}}\right)$$

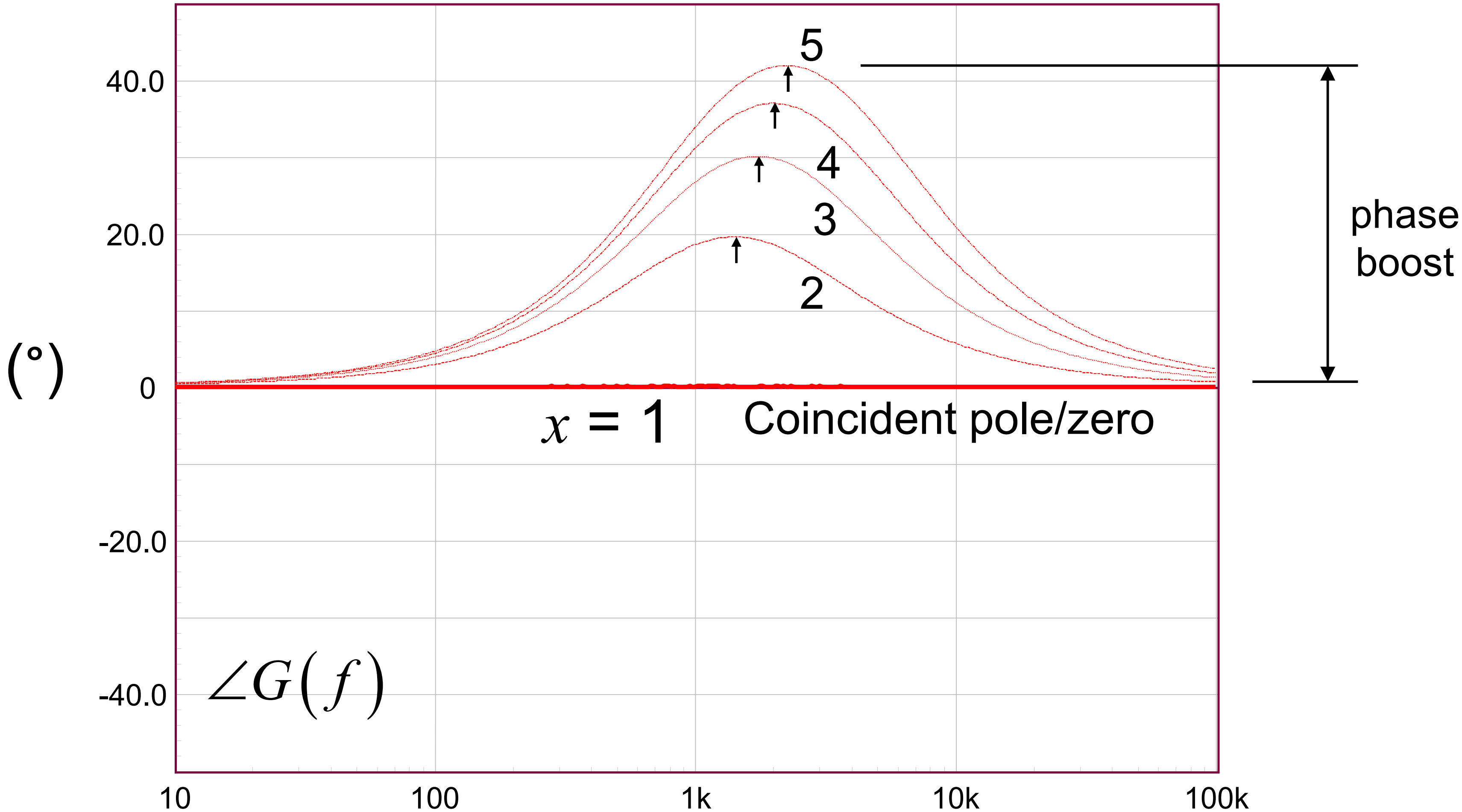
- Where does the phase peak (the boost) occur?

$$\frac{d\left(\arctan\left(\frac{f}{f_{z_1}}\right)-\arctan\left(\frac{f}{f_{p_1}}\right)\right)}{df}=0$$

Max boost occurs at: $\Rightarrow$ $\boxed{f=\sqrt{f_{z_1}f_{p_1}}}$

Split Poles and Zeros to Boost Phase

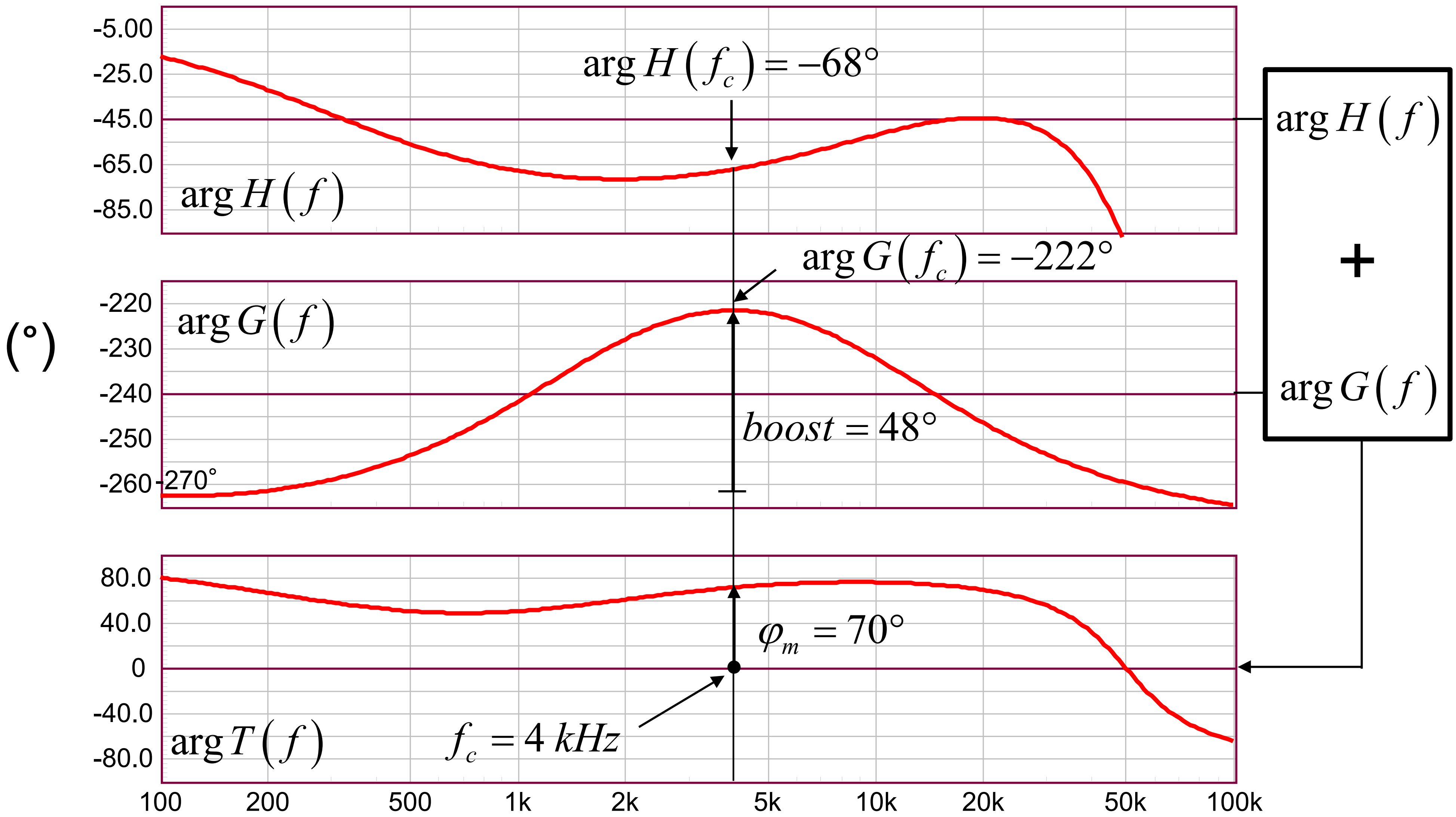
- ❑ A pole and zero are linked by $f_p = x \cdot f_z$



- ❑ Phase boost grows as pole(s) and zero(s) spread appart

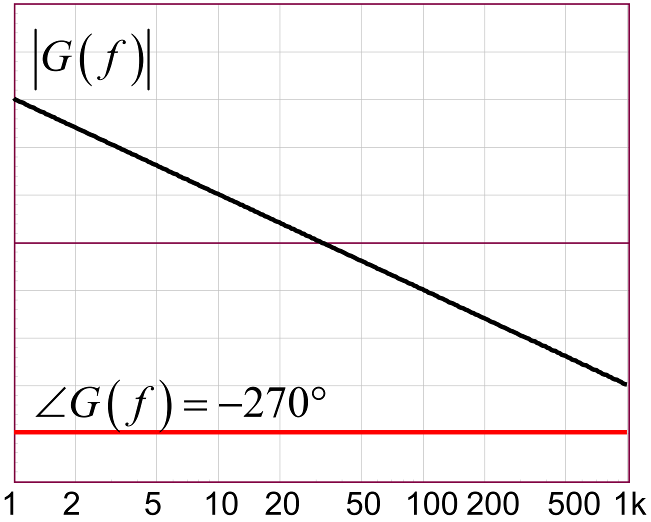
Boost the Phase to Create Margin at f_c

- ❑ The pole has been placed at 10.4 kHz and the zero at 1.5 kHz

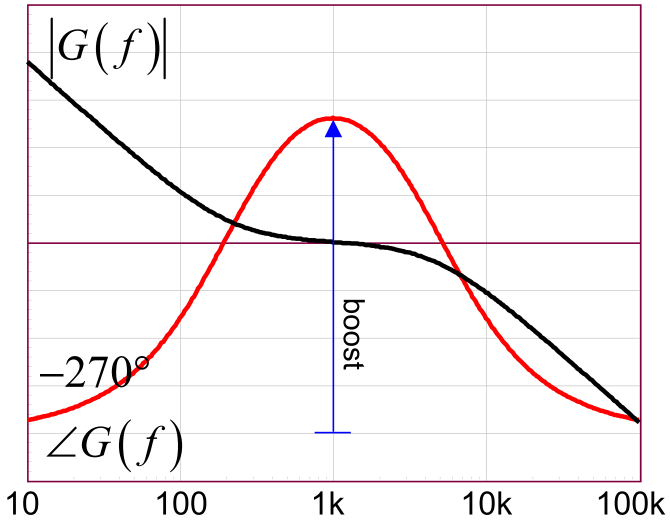


Three Compensator Types Exist

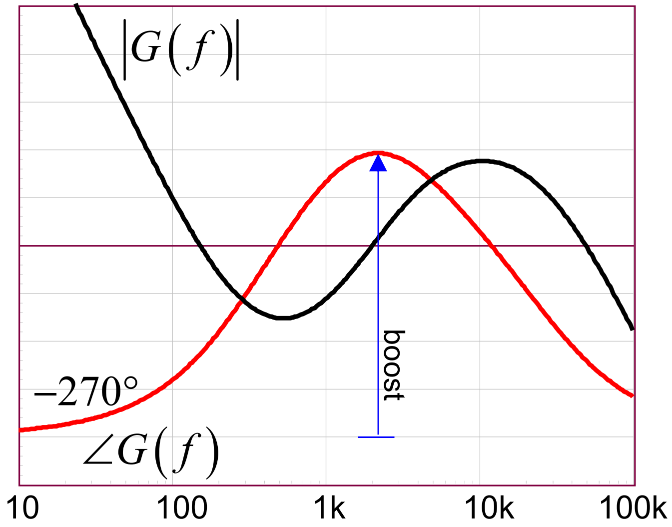
- ❑ Boost the phase from 0 to 180° with 3 different filters



Type 1 – no boost



Type 2 – up to 90°



Type 3 – up to 180°

$$G_1(s) = -\frac{1}{\dfrac{s}{\omega_{po}}}$$

$$G_2(s) = -G_0 \frac{1 + \dfrac{s_{z_1}}{s}}{1 + \dfrac{s}{s_{p_1}}}$$

$$G_3(s) = -G_0 \frac{\left(1 + \dfrac{s_{z_1}}{s}\right)\left(1 + \dfrac{s}{s_{z_2}}\right)}{\left(1 + \dfrac{s}{s_{p_1}}\right)\left(1 + \dfrac{s}{s_{p_2}}\right)}$$

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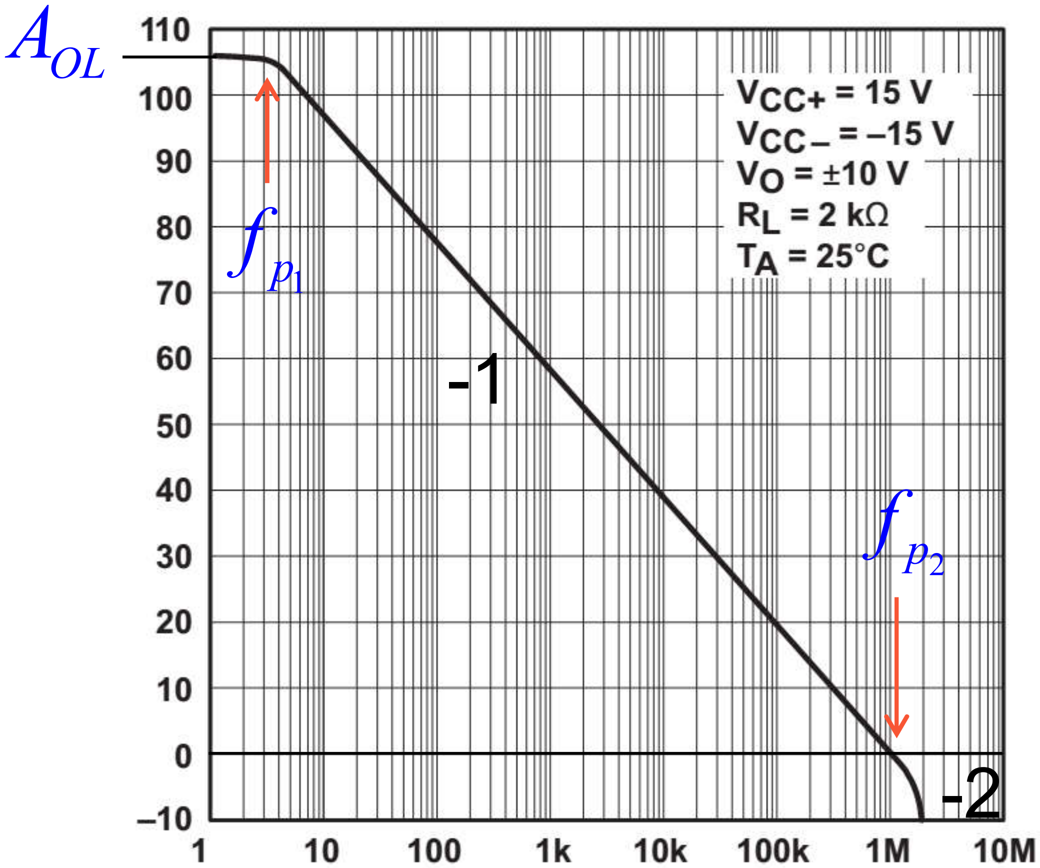
The Operational Amplifier

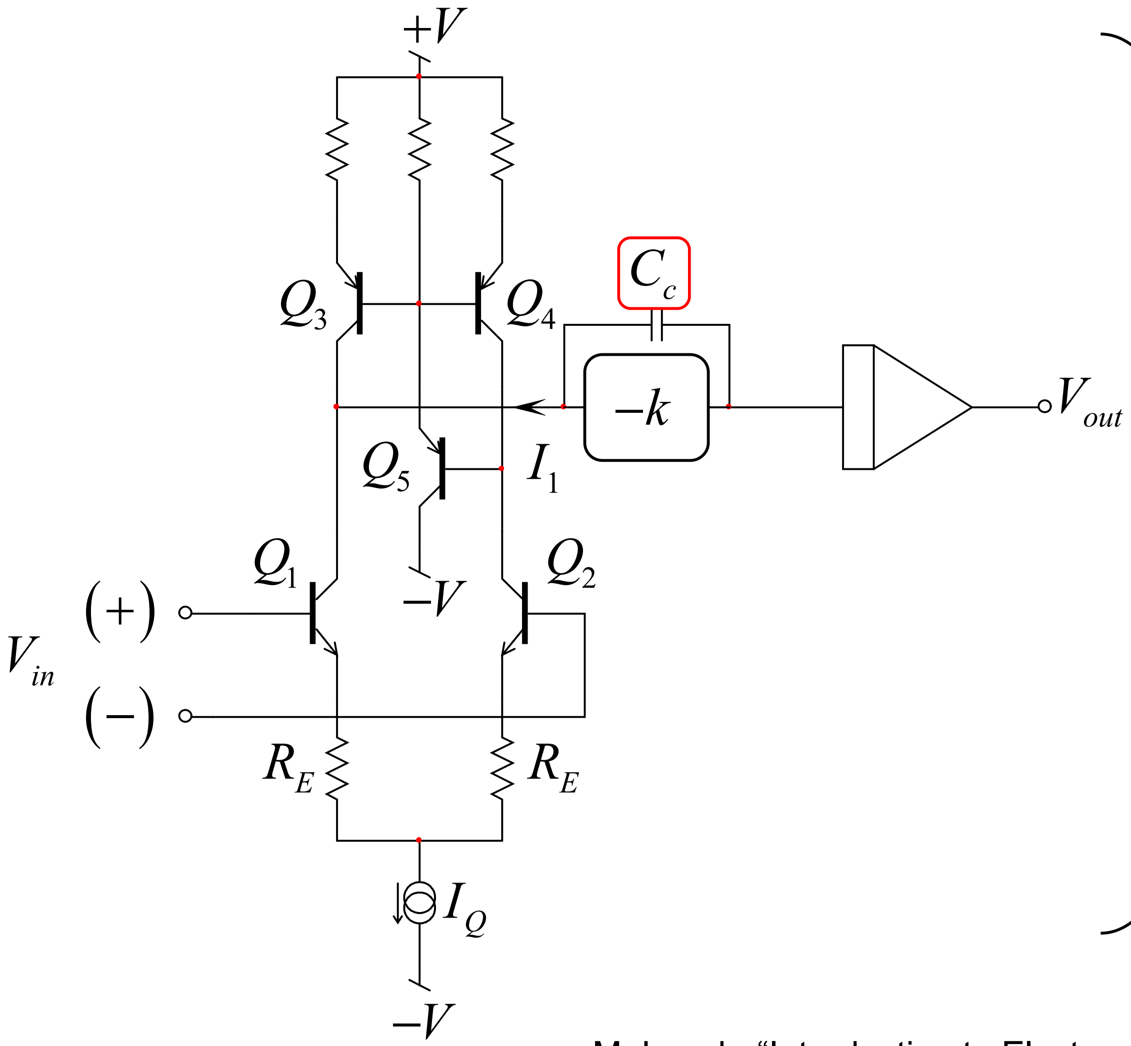
- For high bandwidth systems, compensators use op amps
- Op amps are affected by an open-loop gain and two poles



- How do these variables affect compensator performance?

Op Amp Simplified Internals

- ❑ Capacitor C_c is key as compensating element



$$G(s) = \frac{A_{OL}}{1 + \frac{s}{\omega_{p_1}}}$$

$$A_{OL} = \frac{k \cdot R_{eq}}{r_\pi + R_E}$$

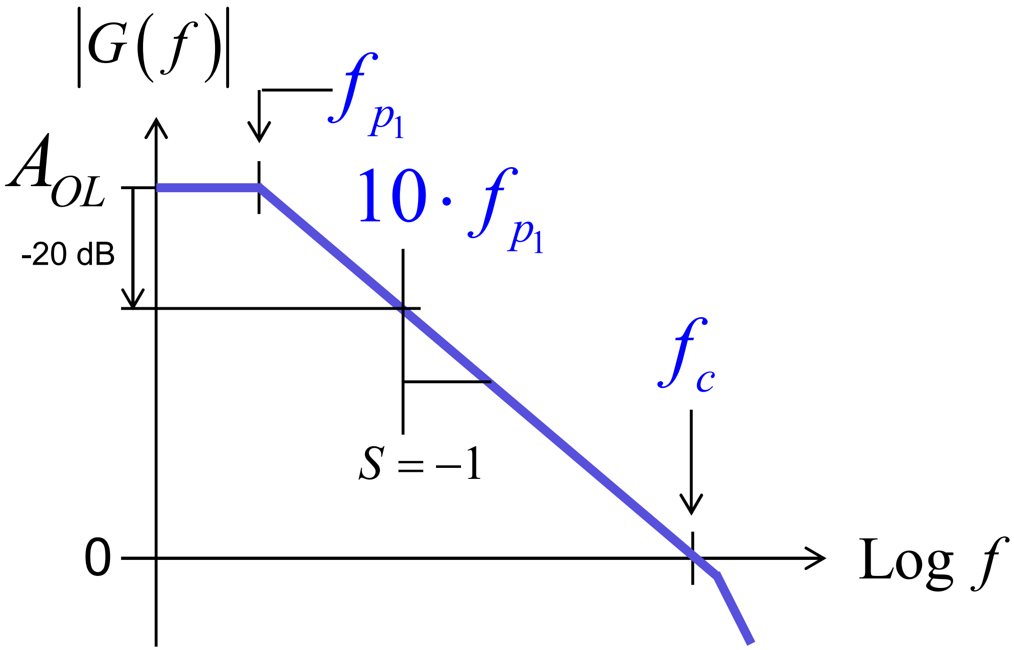
$$\omega_{p_1} = \frac{1}{(1+k)\boxed{C_c}R_{eq}}$$

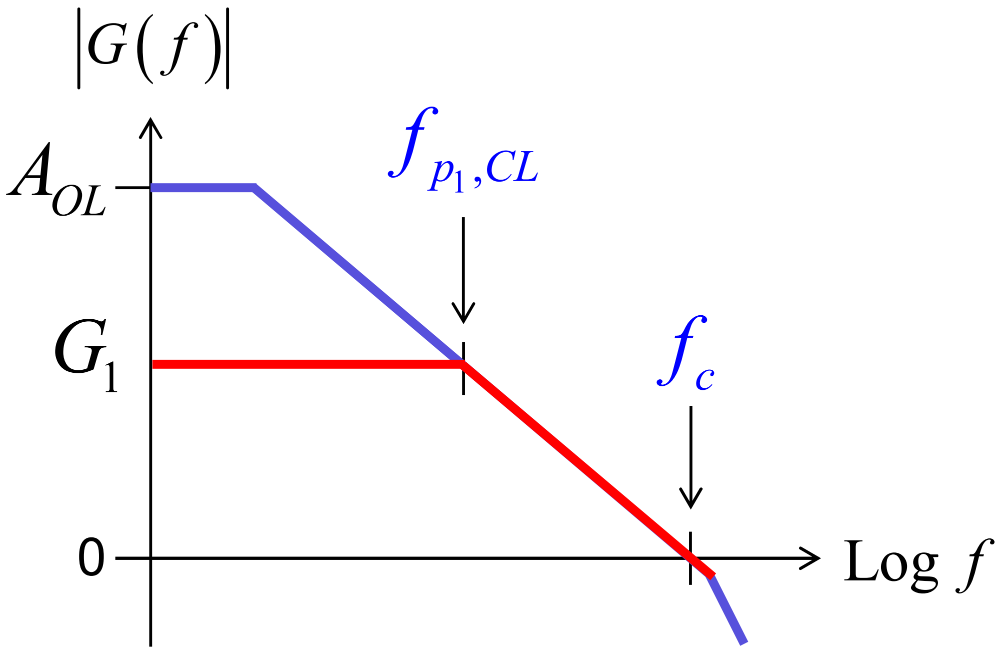
Miller effect

M. Leach, "Introduction to Electroacoustics and Audio Amplifier Design", Kendall/Hunt, 2003

Op Amp Gain-Bandwidth Product

- The 0-dB crossover point is at unity-gain frequency or GBW





$$S = \frac{\Delta y}{\Delta x} = \frac{-20\text{ dB}}{\log\left(\frac{10 f_{p_1}}{f_{p_1}}\right)} = -20\,\text{dB/dec}$$

$$f_c = A_{OL} \times f_{p_1} \longrightarrow \text{GBW}$$

$$f_{p_1,CL} = \frac{f_c}{G_1}$$

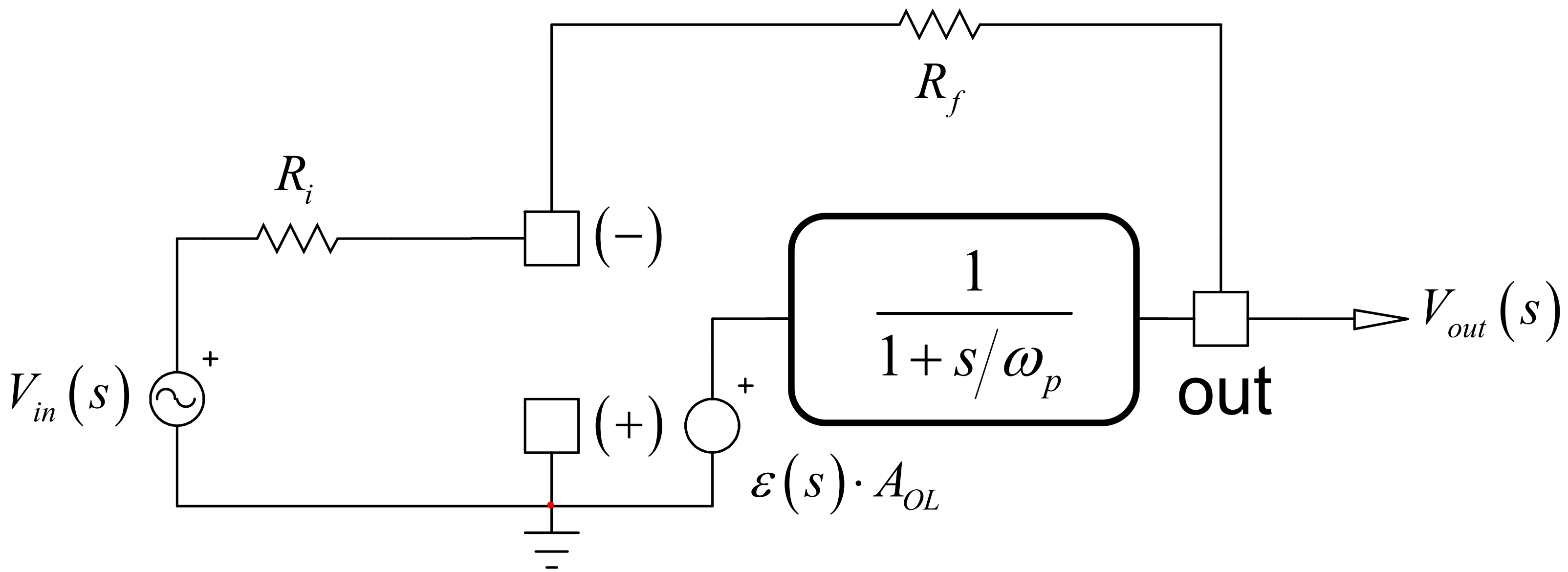
Accounting for the Open-Loop Gain

❑ The open-loop gain can vary quite a bit, see µA741 below

A_{VD}	Large-signal differential voltage amplification	$R_L \geq 2k\Omega$, $V_O = \pm 10$ V	25°C	20	200	50	200	V/mV
			Full range	15		25		

❖ How does it impact our compensator transfer function?

❑ Let's take a simple inverting case, a proportional gain



C. Basso, "Designing Control Loops for Linear and Switching Power Supplies", Artech House, 2012

The Low-Frequency Pole Moves

- The transfer function follows the form

$$G(s) = -\frac{R_f}{R_i + \dfrac{R_f + R_i}{A_{OL}}} \frac{1}{1 + s\dfrac{R_i + R_f}{\omega_p\left(R_f + R_i + A_{OL}R_i\right)}} = -G_0\frac{1}{1+\dfrac{s}{\omega_{eq}}}$$

- The equivalent pole expression can be re-arranged

$$\omega_{eq} = \omega_p\frac{R_f + R_i + A_{OL}R_i}{R_f + R_i} \longrightarrow \frac{R_f}{R_i} \approx G \longrightarrow R_f \approx R_i G$$

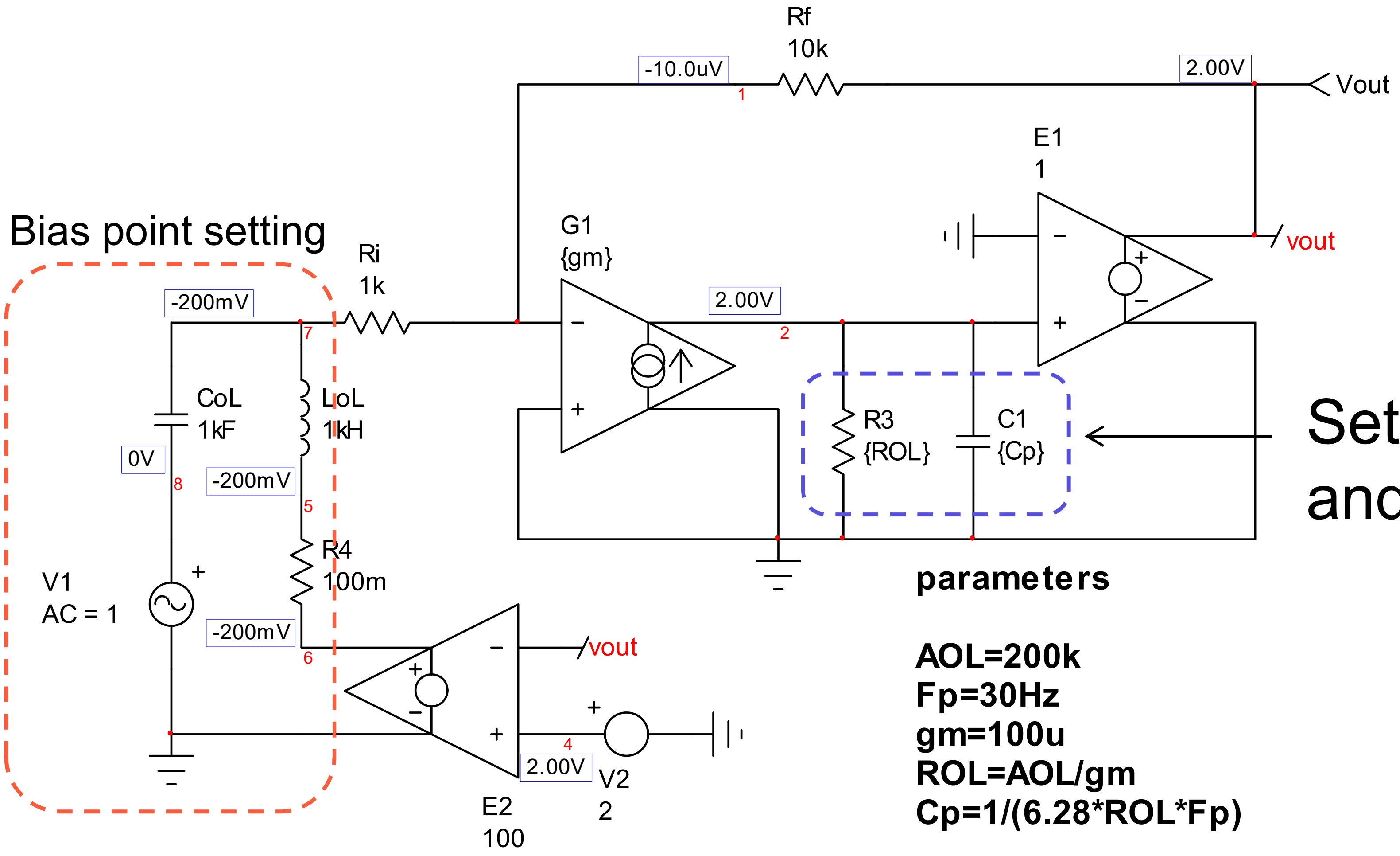
$$\omega_{eq} = \omega_p\frac{R_i G + R_i + A_{OL}R_i}{R_i G + R_i} = \omega_p\frac{1 + G + A_{OL}}{1 + G}$$

Factor $1 + G$ and simplify

$$\omega_{eq} = \omega_p\left(1 + \frac{A_{OL}}{1+G}\right) \approx \frac{\omega_p A_{OL}}{1+G} \Rightarrow \omega_{eq} \approx \frac{GBW}{1+G}$$

What is the Impact on Transfer Function?

❑ You need a wide-bandwidth voltage buffer – A_{OL} = 200k



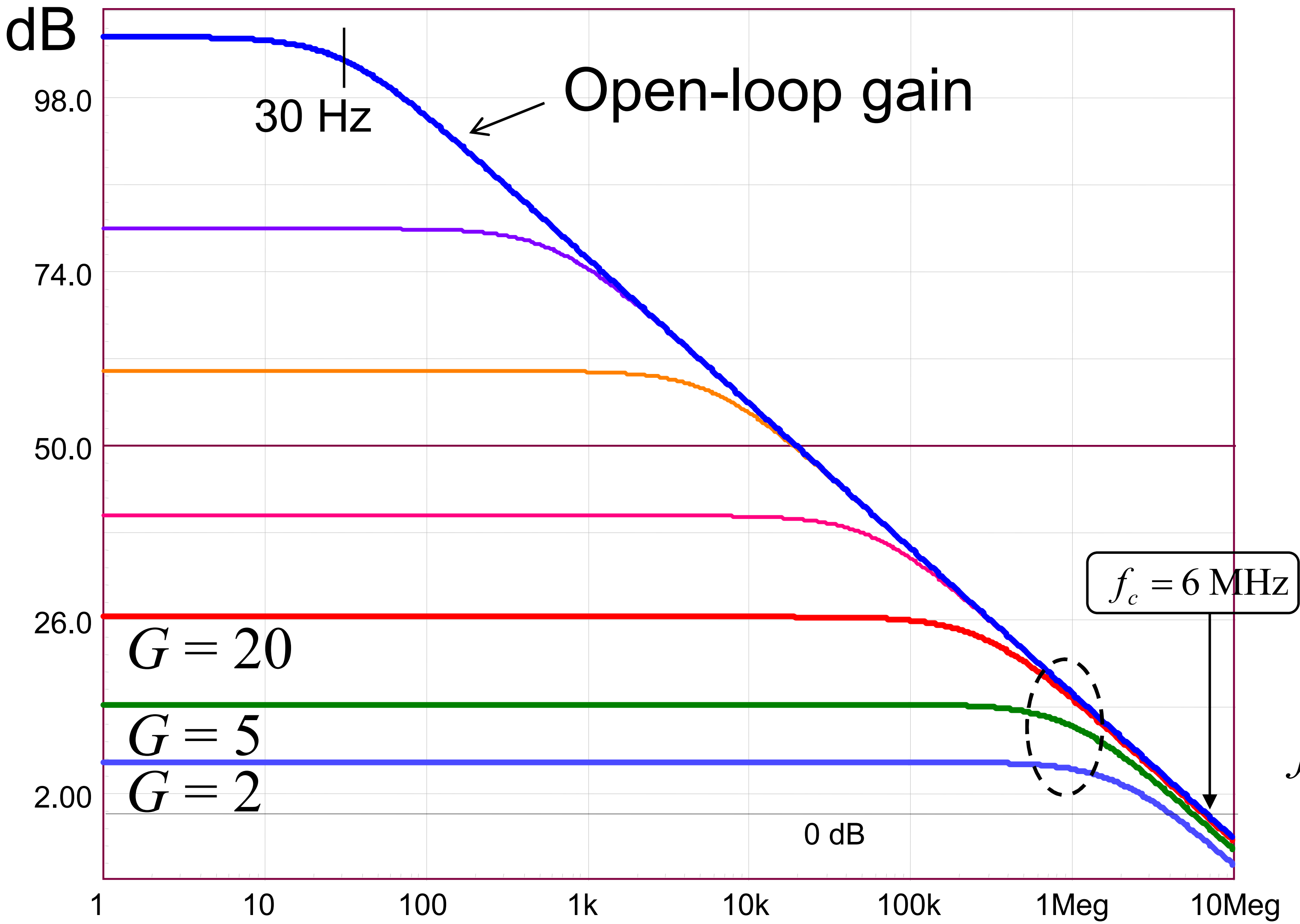
Sets A_{OL} and ω_p

parameters

AOL=200k
Fp=30Hz
gm=100u
ROL=AOL/gm
Cp=1/(6.28*ROL*Fp)

Closed-Loop Bandwidth vs Gain

- You need a wide-bandwidth voltage buffer – A_{OL} = 200k



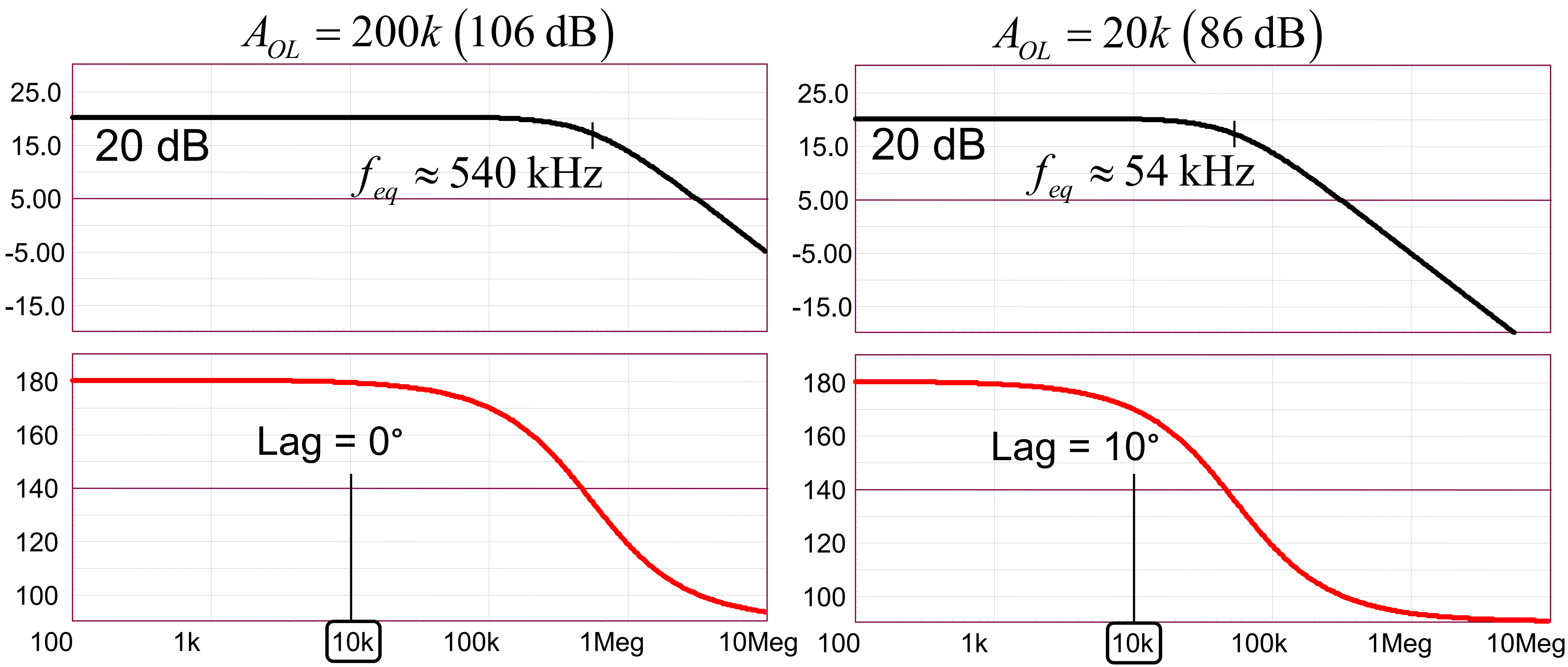
$$f_{p_1,CL} \approx \frac{f_c}{G_1}$$

$$f_{p_1,CL} = \frac{f_c}{G_1+1}$$

$\text{Log} f$

Low-Frequency Pole Moves

- Too wide an open-loop gain dispersion brings additional lag



- A 20-dB difference in the open-loop gain divides BW by 10

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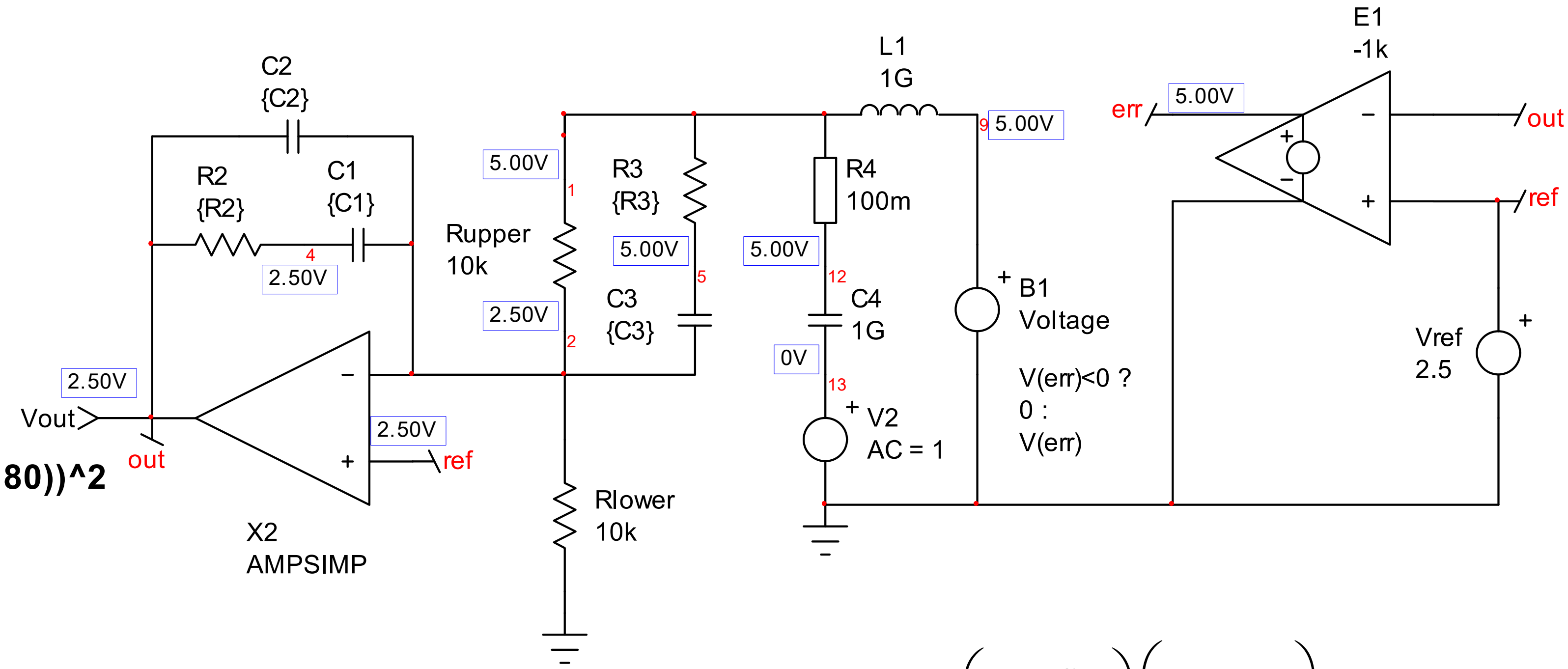
What Gain-Bandwidth is Needed?

- Assume a type III design for 130° boost and 25 dB at 200 kHz

parameters
Rupper=10k
fc=200k
pm=70
Gfc=-25
ps=-150

G=10^(-Gfc/20)
boost=pm-(ps)-90
pi=3.14159
K=(tan((boost/4+45)*pi/180))^2
C2=1/(2*pi*fc*G*Rupper)
C1=C2*(K-1)
R2=sqrt(k)/(2*pi*fc*C1)
R3=Rupper/(k-1)
C3=1/(2*pi*fc*sqrt(k)*R3)

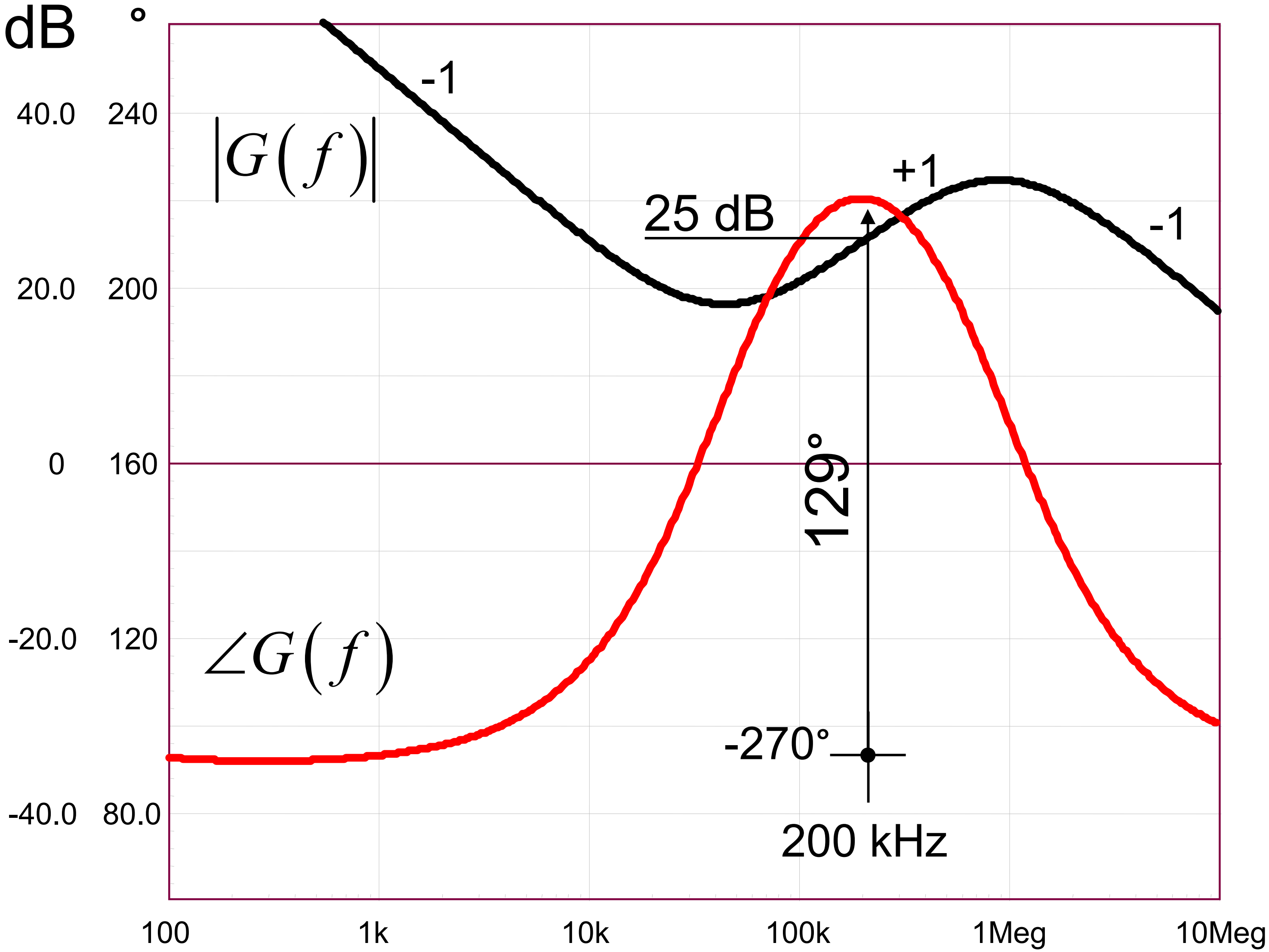
fp1=1/(2*pi*R2*C2)
fp2=1/(2*pi*R3*C3)
fz1=1/(2*pi*R2*C1)
fz2=1/(2*pi*Rupper*C3)



$$G(s) = -G_0 \frac{\left(1+\frac{s_{z_1}}{s}\right)\left(1+\frac{s}{s_{z_2}}\right)}{\left(1+\frac{s}{s_{p_1}}\right)\left(1+\frac{s}{s_{p_2}}\right)}$$

Type III Response with a Perfect Op Amp

- Excellent ac response at 200 kHz with a perfect op amp



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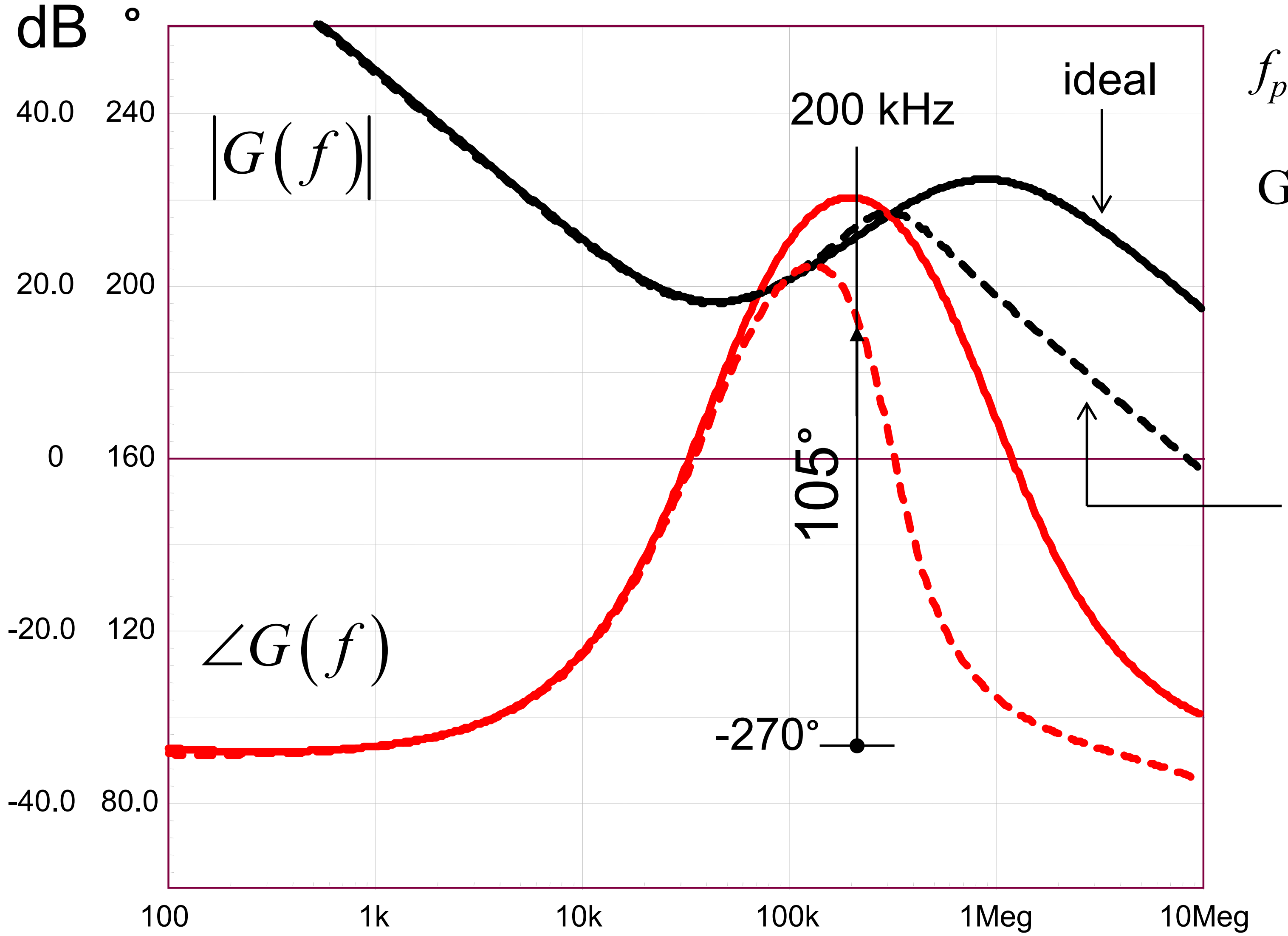
True Op Amp Degrades Phase Boost

- ❑ Phase boost is affected by a 10-MHz gain bandwidth product



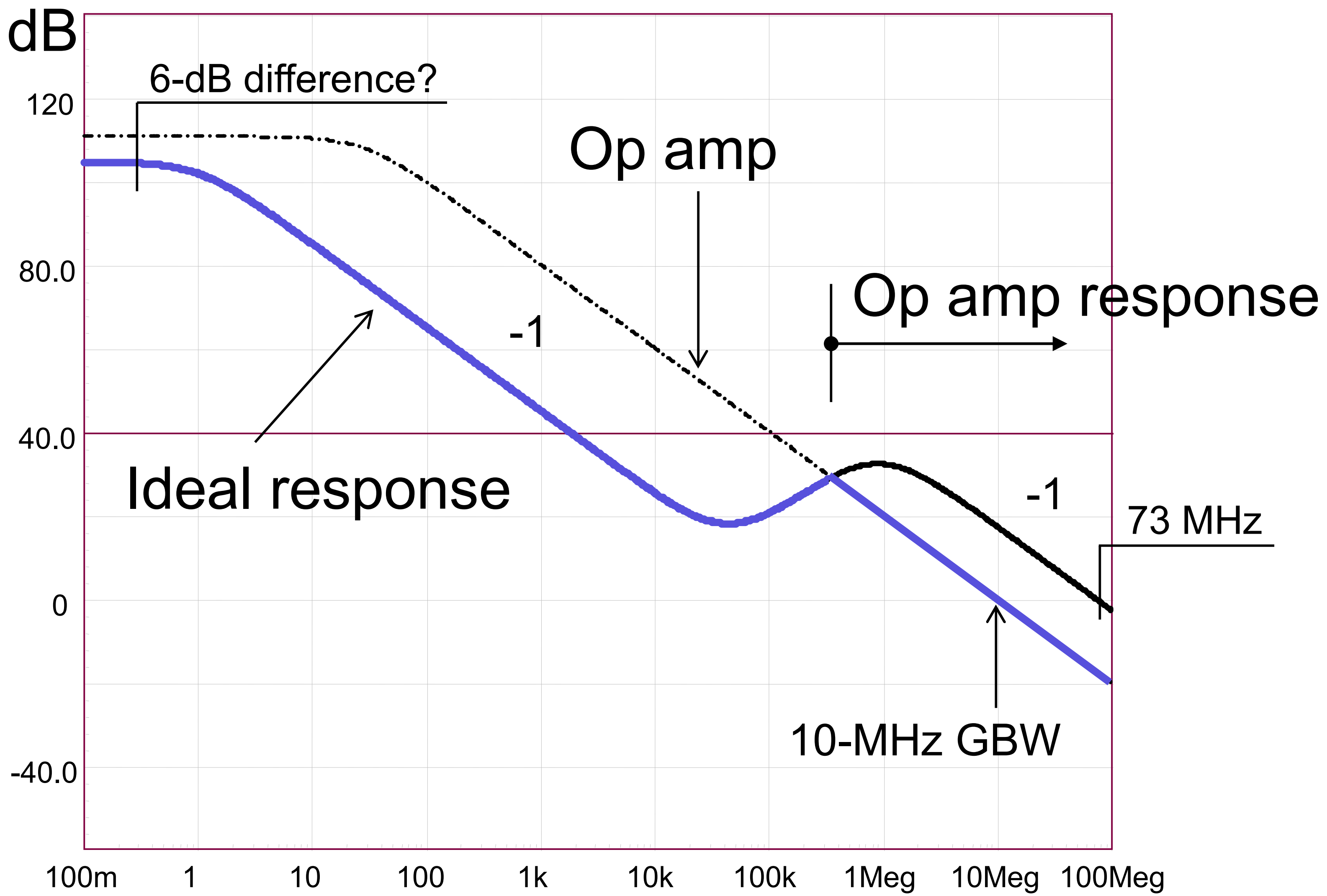
f_p = 30 Hz

GBW=10 MHz

Loss of 24°

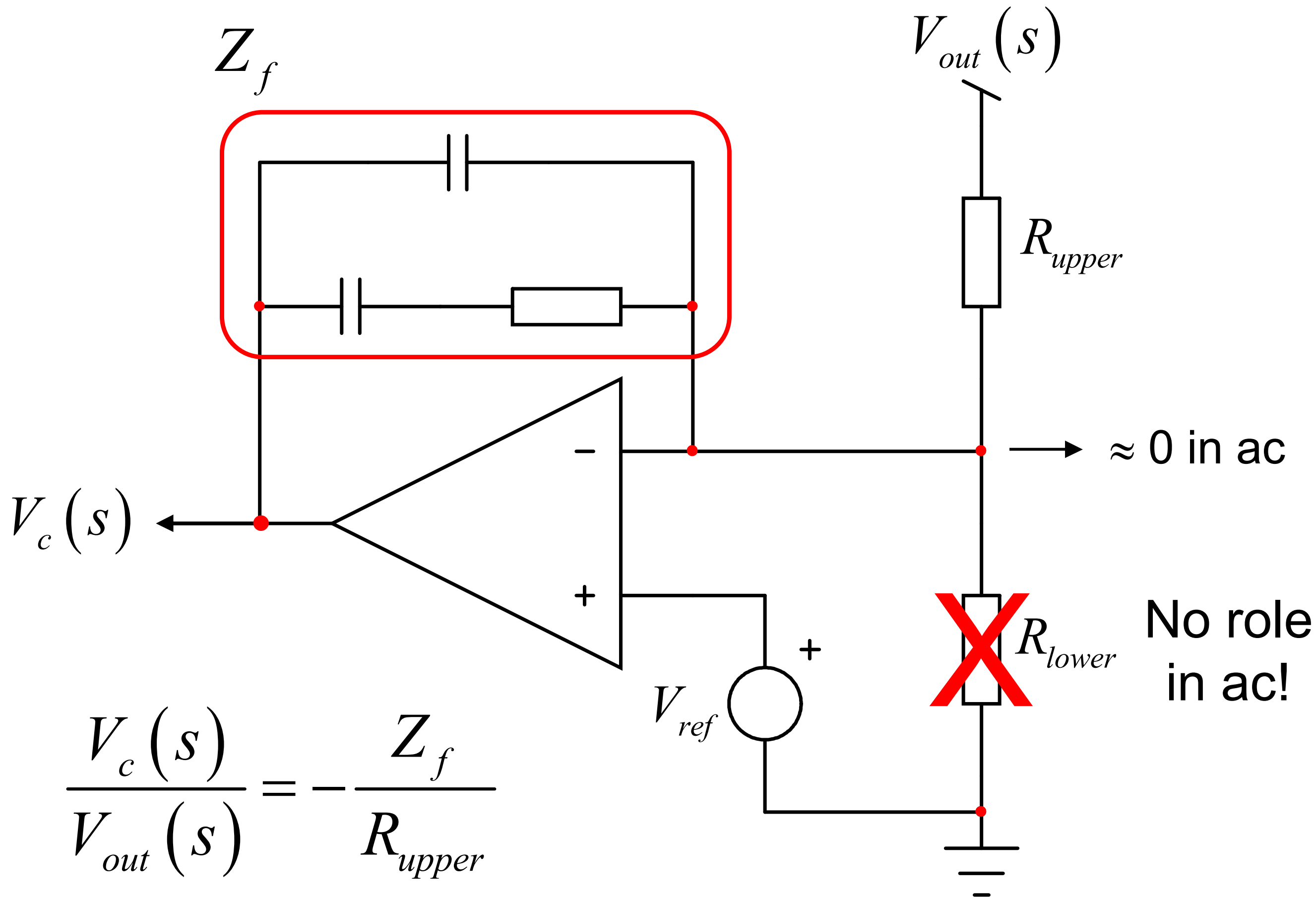
Superimpose the Op Amp Ac Response

- The selected op amp GBW leads response at high frequency



Why is there a Change in the Dc Gain?

- ❑ Because of local feedback, the op amp builds a virtual ground

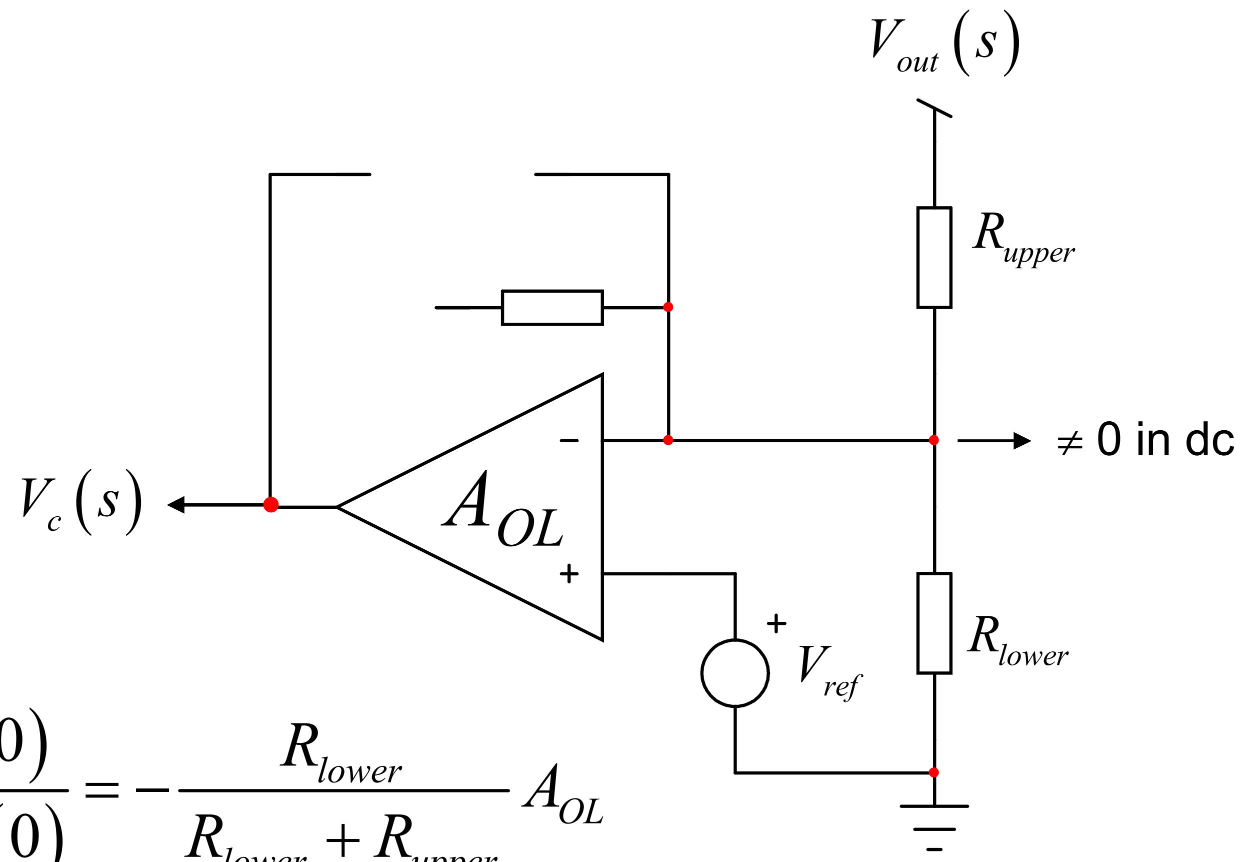


$$\frac{V_c(s)}{V_{out}(s)} = -\frac{Z_f}{R_{upper}}$$

- ❑ The virtual ground excludes the lower resistor R_{lower}

The Divider Ratio is Back in Dc

- In a type 1, 2 or 3, the local feedback is lost for $s = 0$

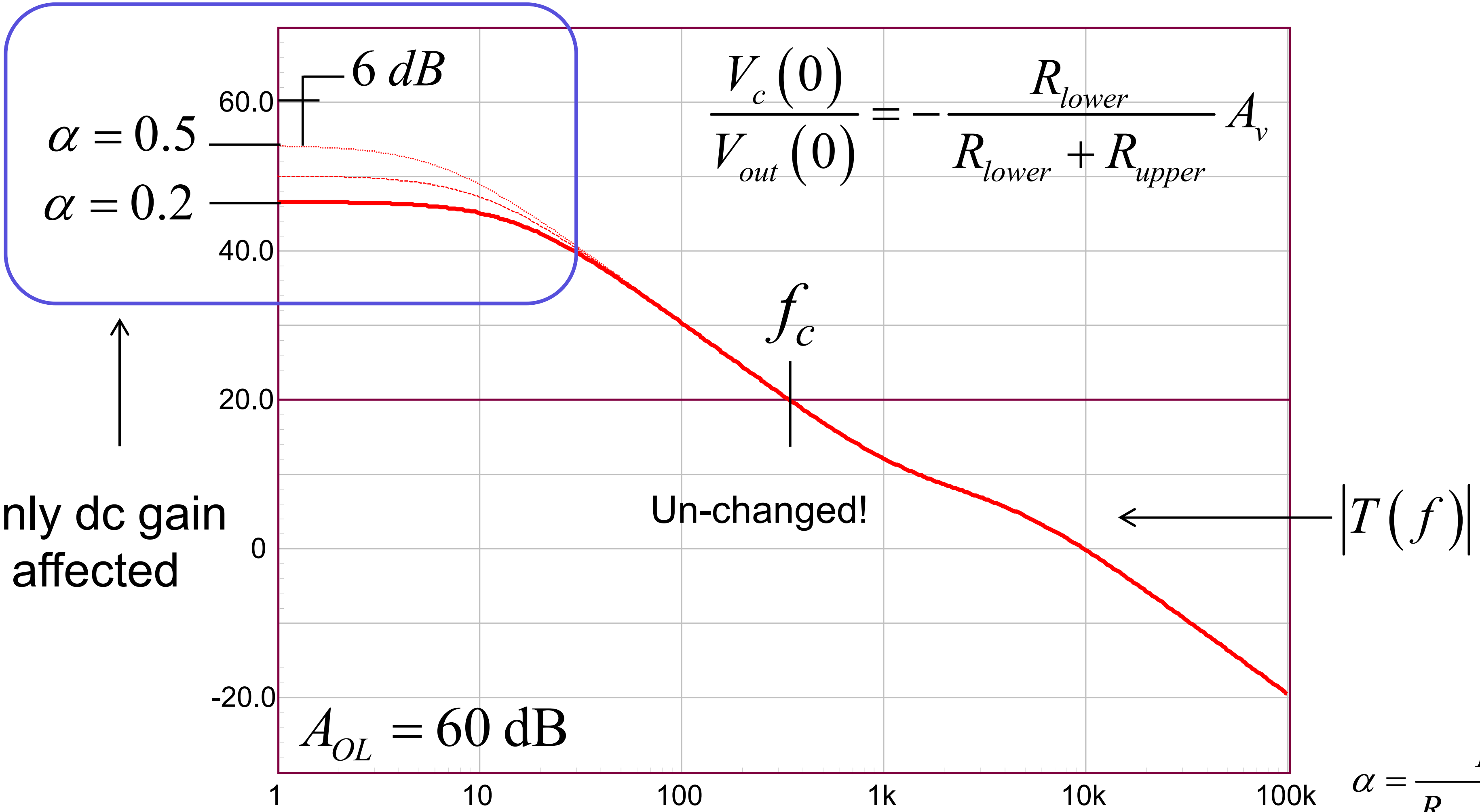


$$\frac{V_c(0)}{V_{out}(0)} = -\frac{R_{lower}}{R_{lower} + R_{upper}} A_{OL}$$

- The 0-Hz gain is indeed changed but not f_c!

The Division Ratio Affects the Dc Gain Alone

- The division ratio is changed to adjust V_{out}: f_c is unchanged

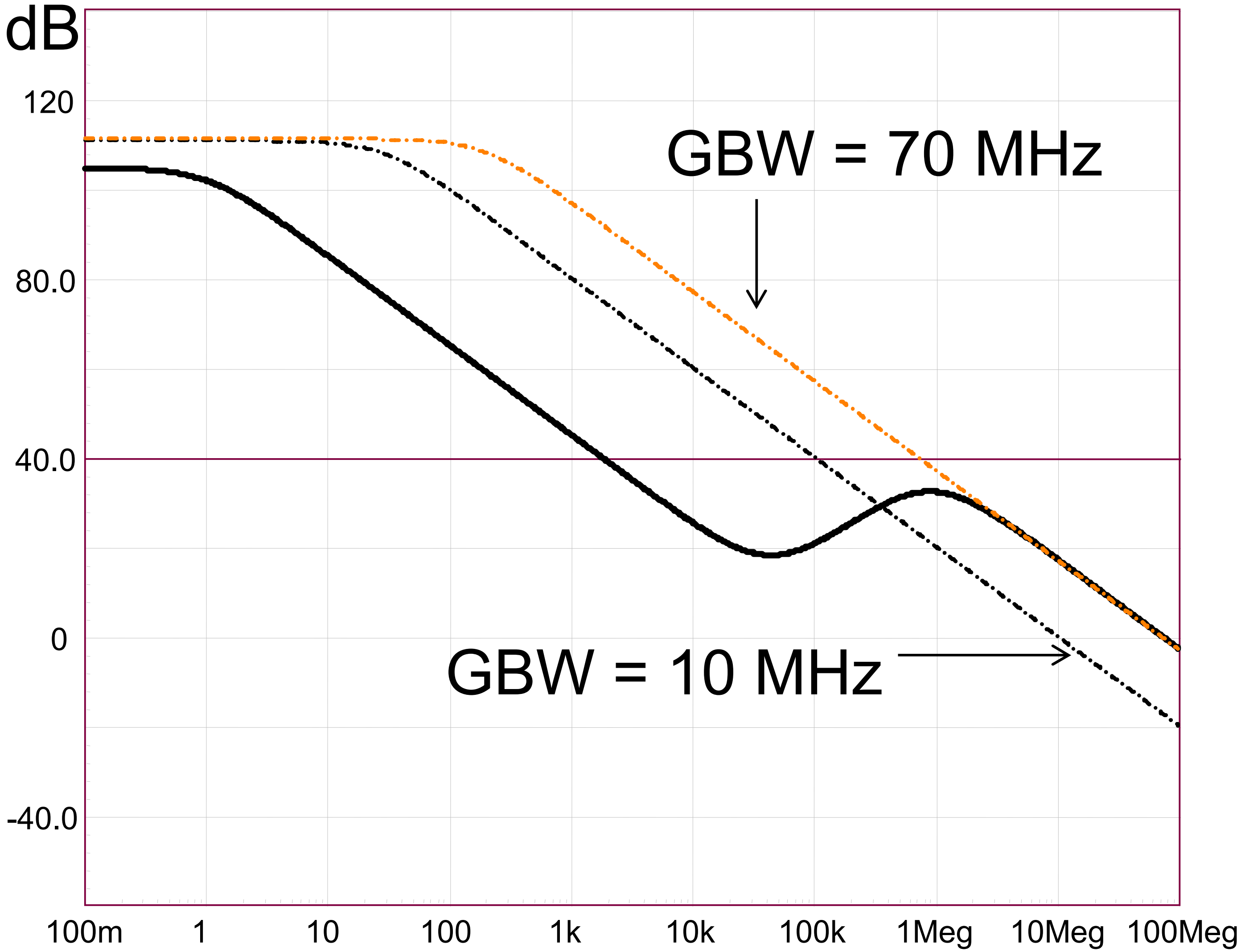


Only dc gain is affected

$$\alpha = \frac{R_{lower}}{R_{lower}+R_{upper}}$$

Select Higher GBW Op Amplifiers

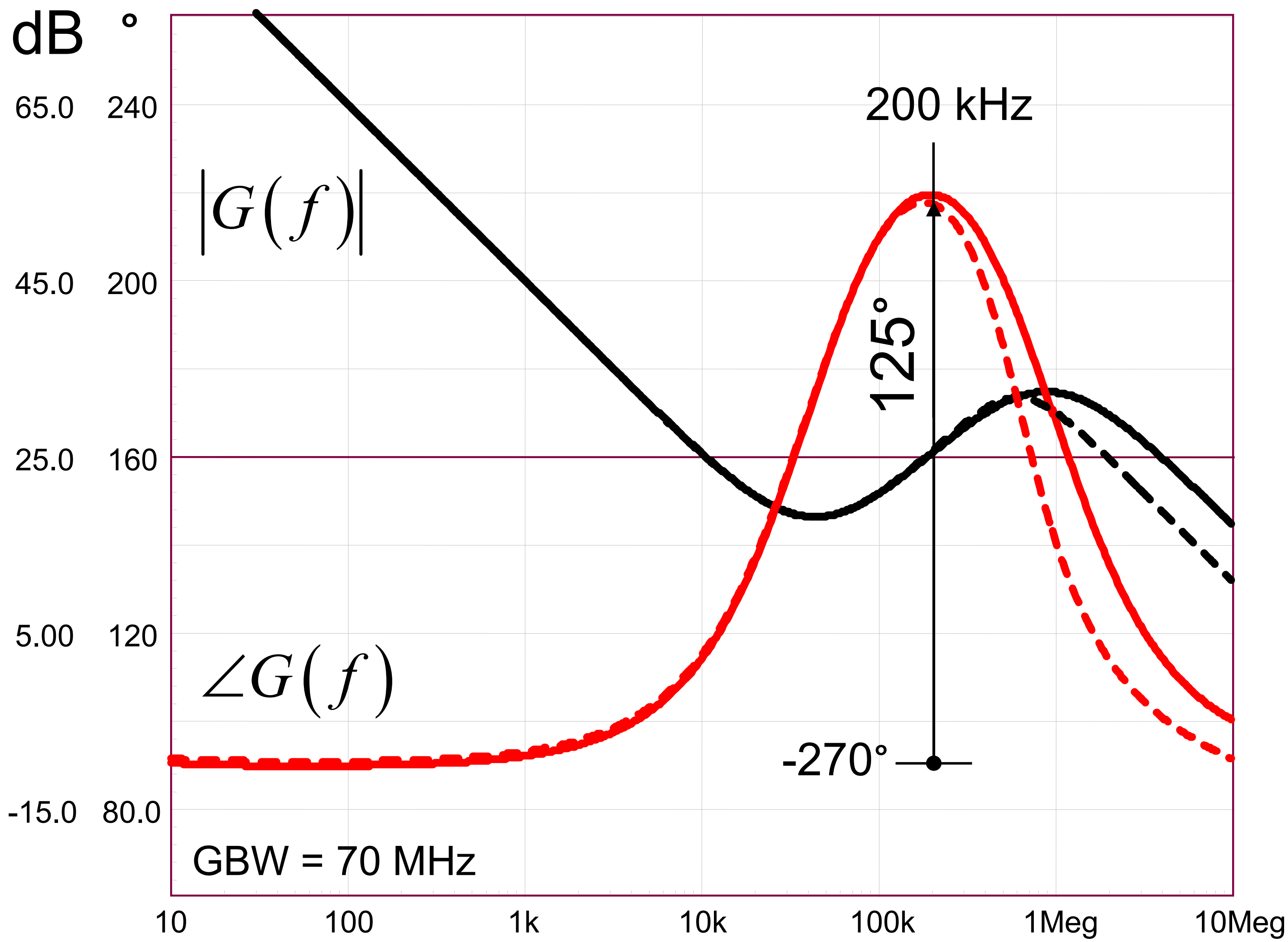
- Choose an op amp whose GBW is the type 3 0-dB f_c



T. Hegarty, "Error Amp. Limitations in High Performance Regulator Applications", AN-1997, Texas Inst.

Final Phase Distortion is Low

- With the right GBW, impact on the final phase boost is low



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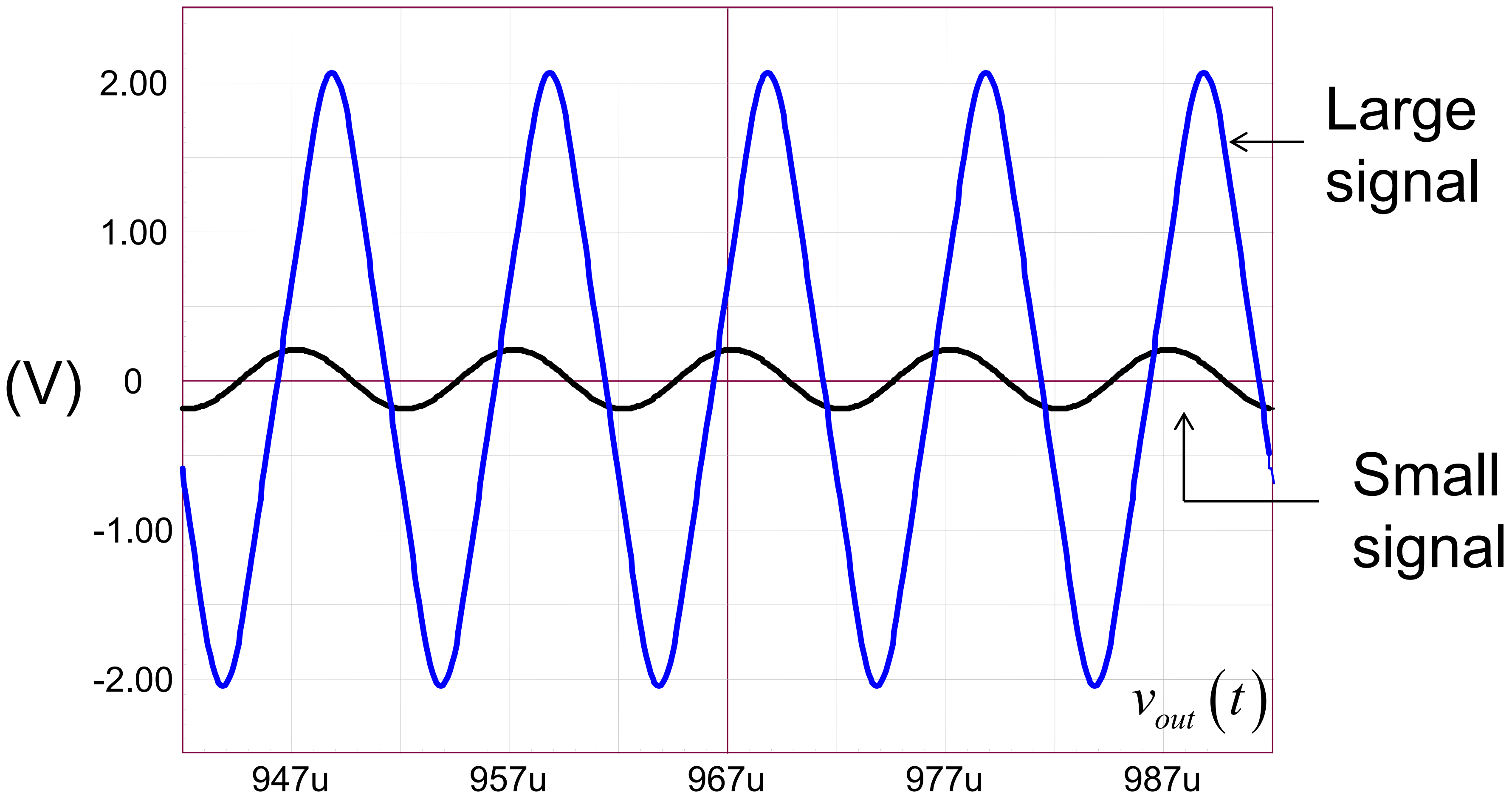
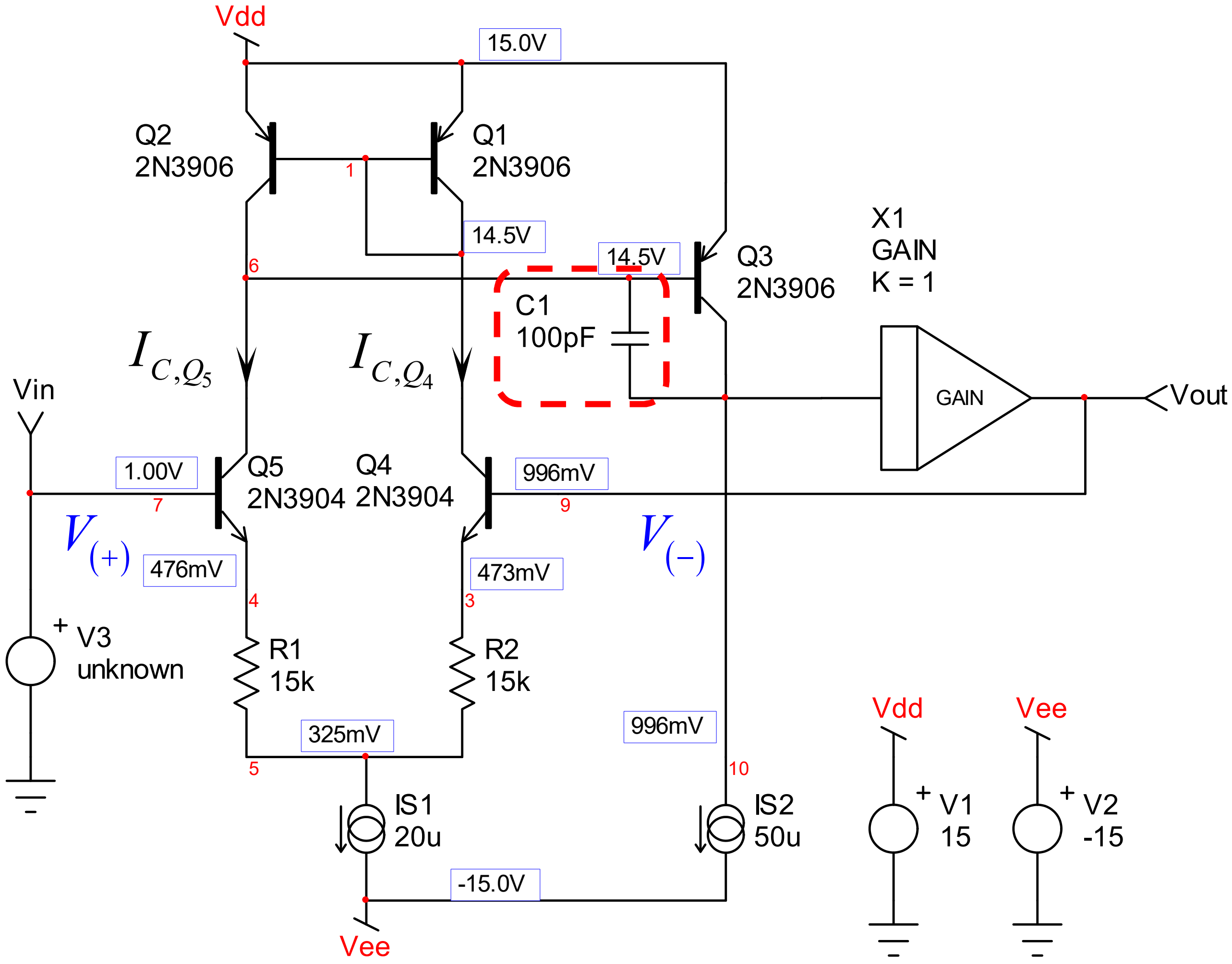
Slew Rate in an Op Amp

- ❑ Small-signal analysis implies small output dynamics
- ❑ Step load response often involves large-signal behavior



Why Does the Op Amp Slew?

- SR limit occurs if differential input stage equilibrium is lost

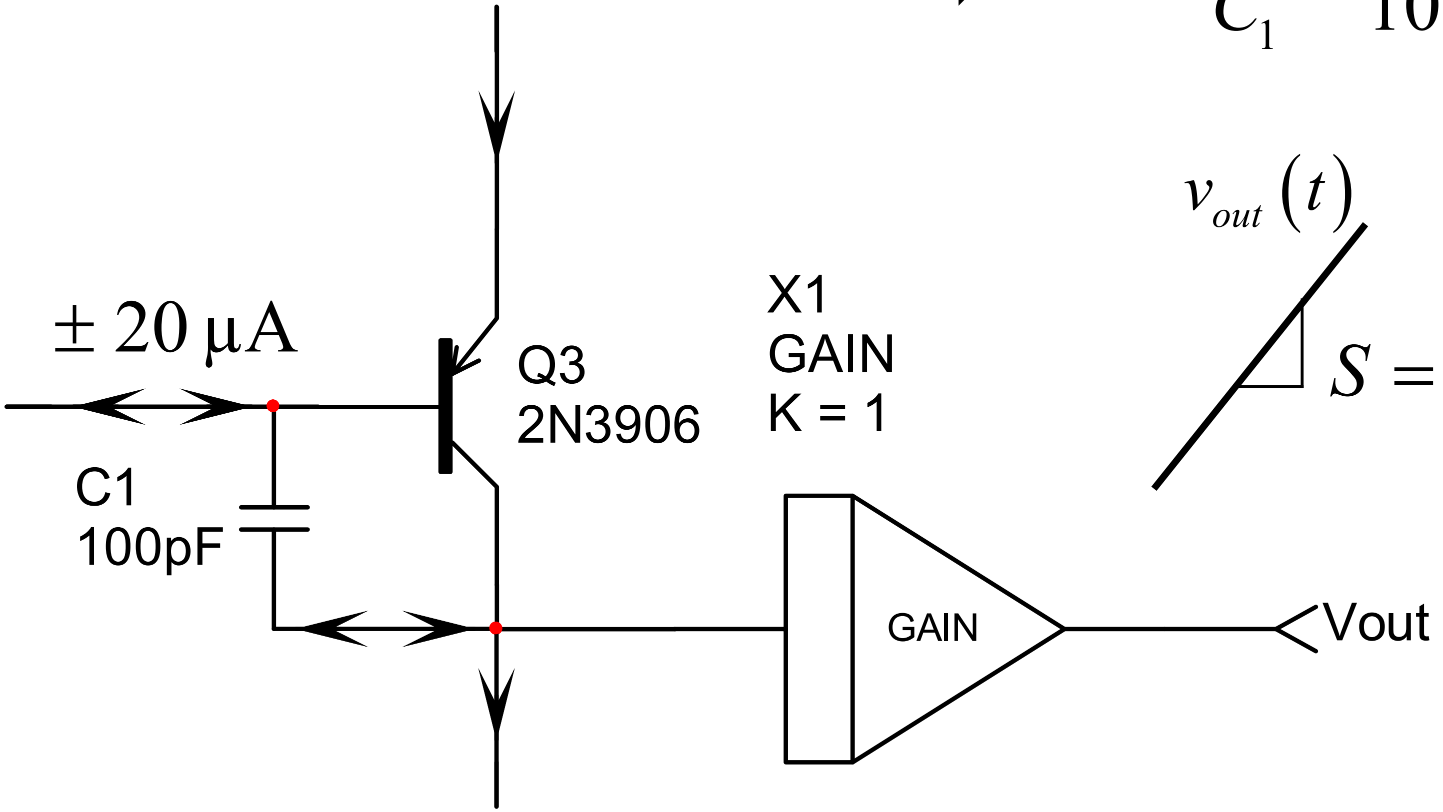


$$I_{C,Q_5} \neq I_{C,Q_4}$$

http://www.edn.com/electronics-blogs/the-signal/4415482/Slew-Rate-the-op-amp-speed-limit

Capacitor C_1 Limits the Output Slew Rate

- The Miller capacitor is charged/discharged by 20-µA
- Neglecting base currents ⇒ $SR = \frac{I_{S_1}}{C_1} = \frac{20u}{100p} = 200\ \mathrm{mV/\mu s}$

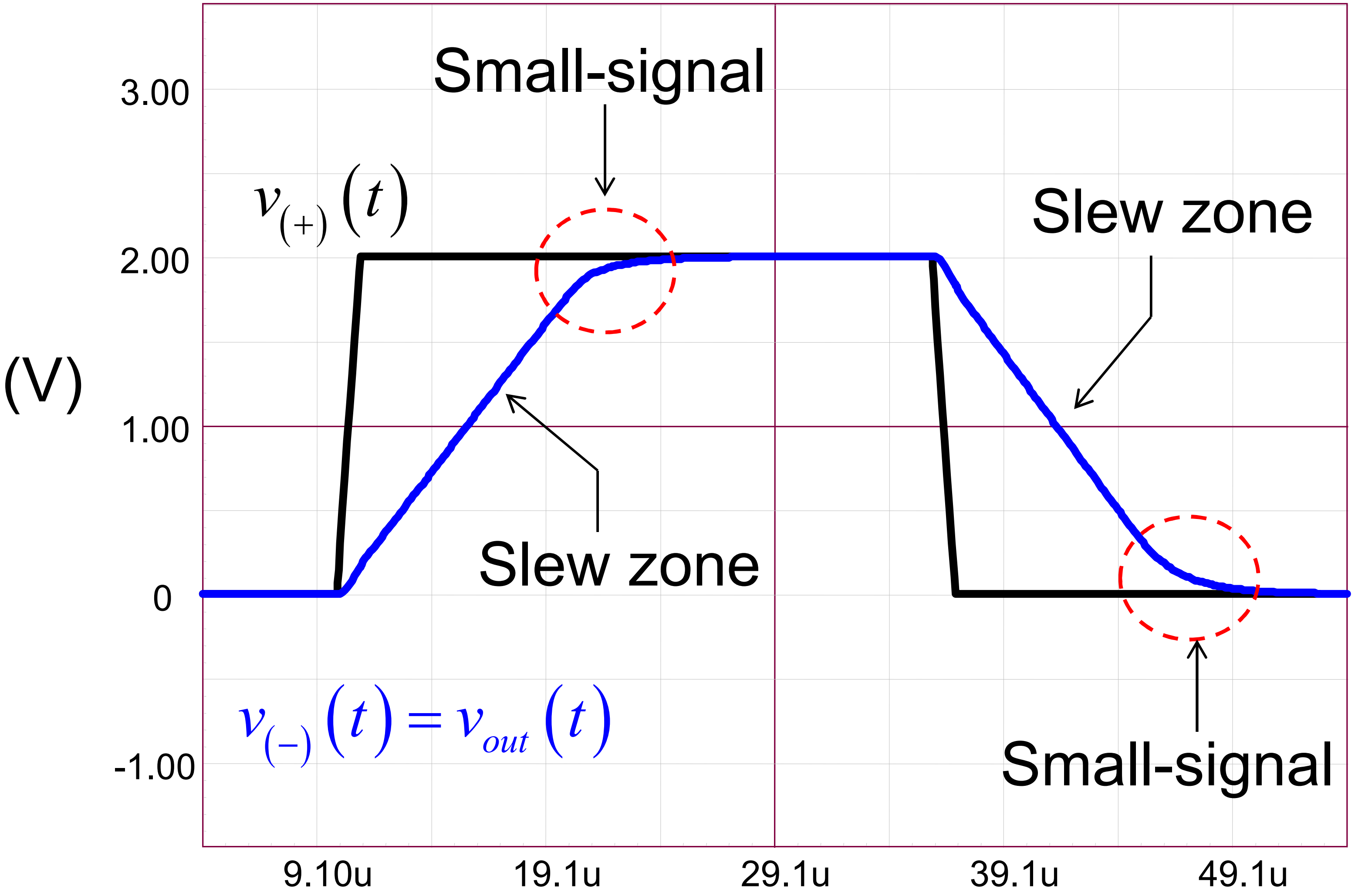


$$S = \frac{dv_{out}(t)}{dt}\ [\mathrm{V/s}]$$

- The slew rate is the maximum rate of change you can get

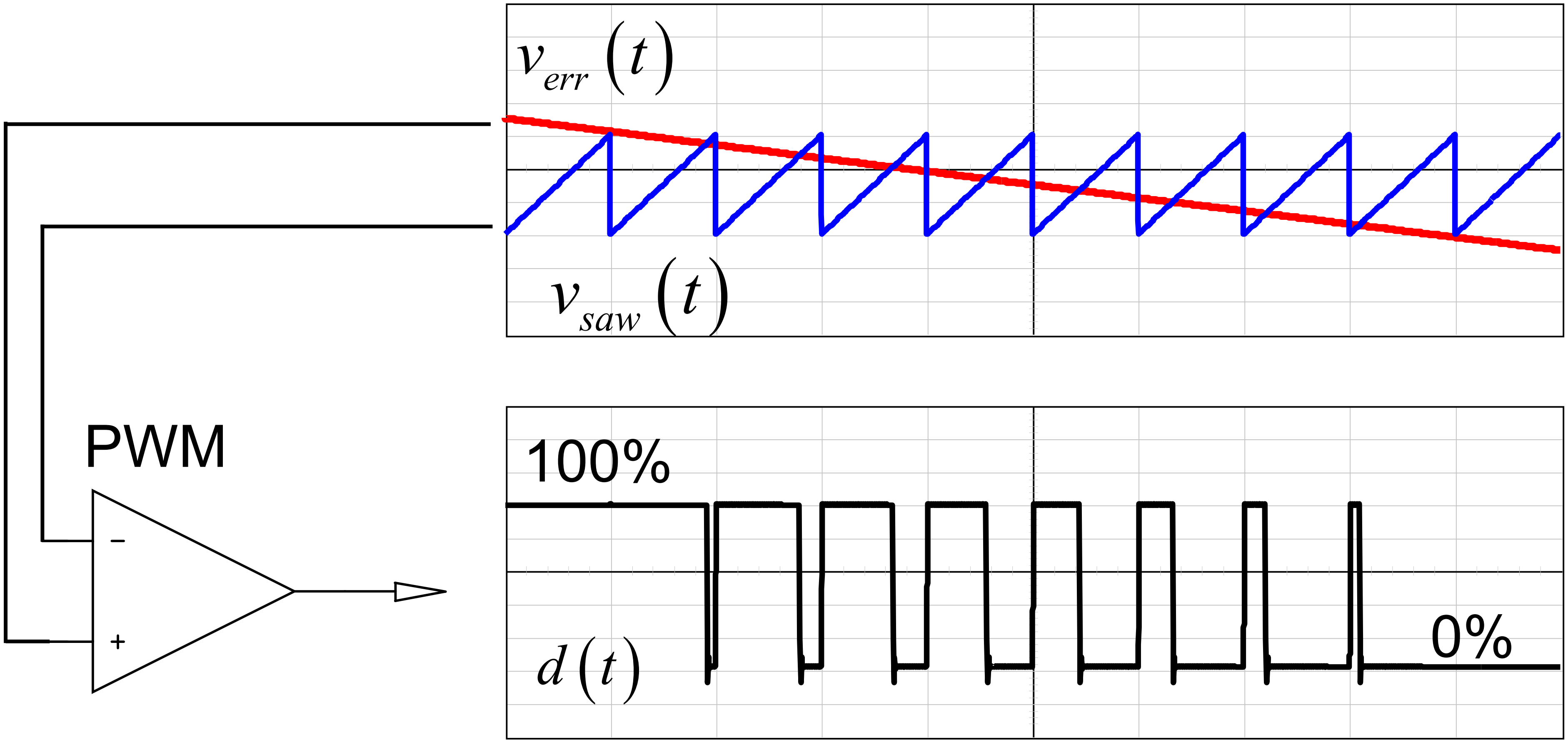
Miller Capacitor Charging/Discharging

- The op amp output cannot keep up the change: slewing



Duty Ratio Modulator

- The duty ratio can change from 0 to 100%



- What if the op amp slews while the loop asks for a change?

Slew-Rate Impact on a Buck Regulator

- ❑ This converter switches at 1 MHz and features a 200-kHz f_c

parameters

Rupper=38k
fc=200k
Gfc=-30

G=10^(-Gfc/20)
pi=3.14159

fz1=10k
fz2=5k
fp1=50k
fp2=500k

C3=1/(2*pi*fz1*Rupper)
R3 =1/(2*pi*fp2*C3)

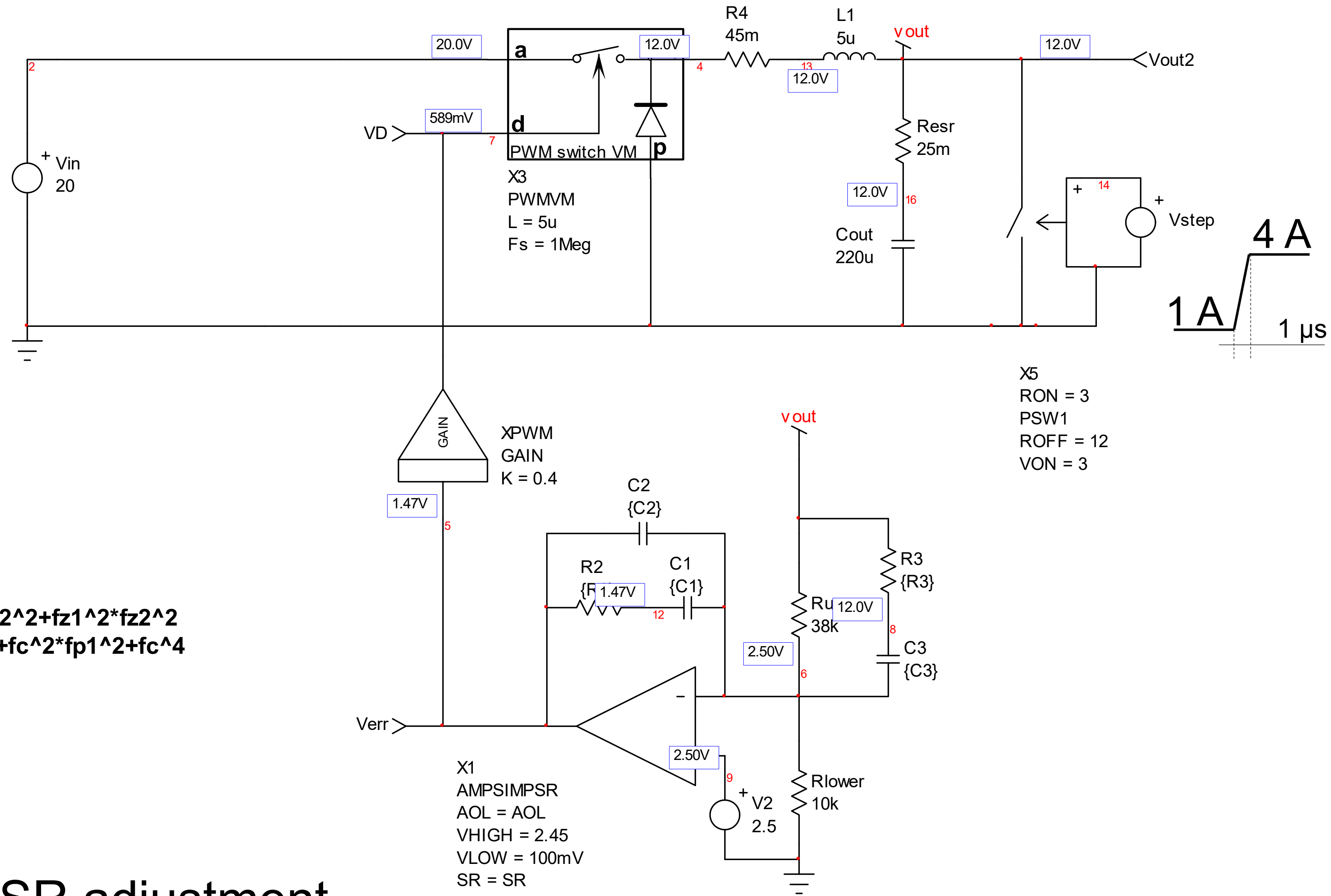
C1=1/(2*pi*fz2*R2)
C2=1/(2*pi*(fp1)*R2)

a=fc^4+fc^2*fz1^2+fc^2*fz2^2+fz1^2*fz2^2
c=fp2^2*fp1^2+fc^2*fp2^2+fc^2*fp1^2+fc^4

R2=sqrt(c/a)*G*fc*R3/fp1

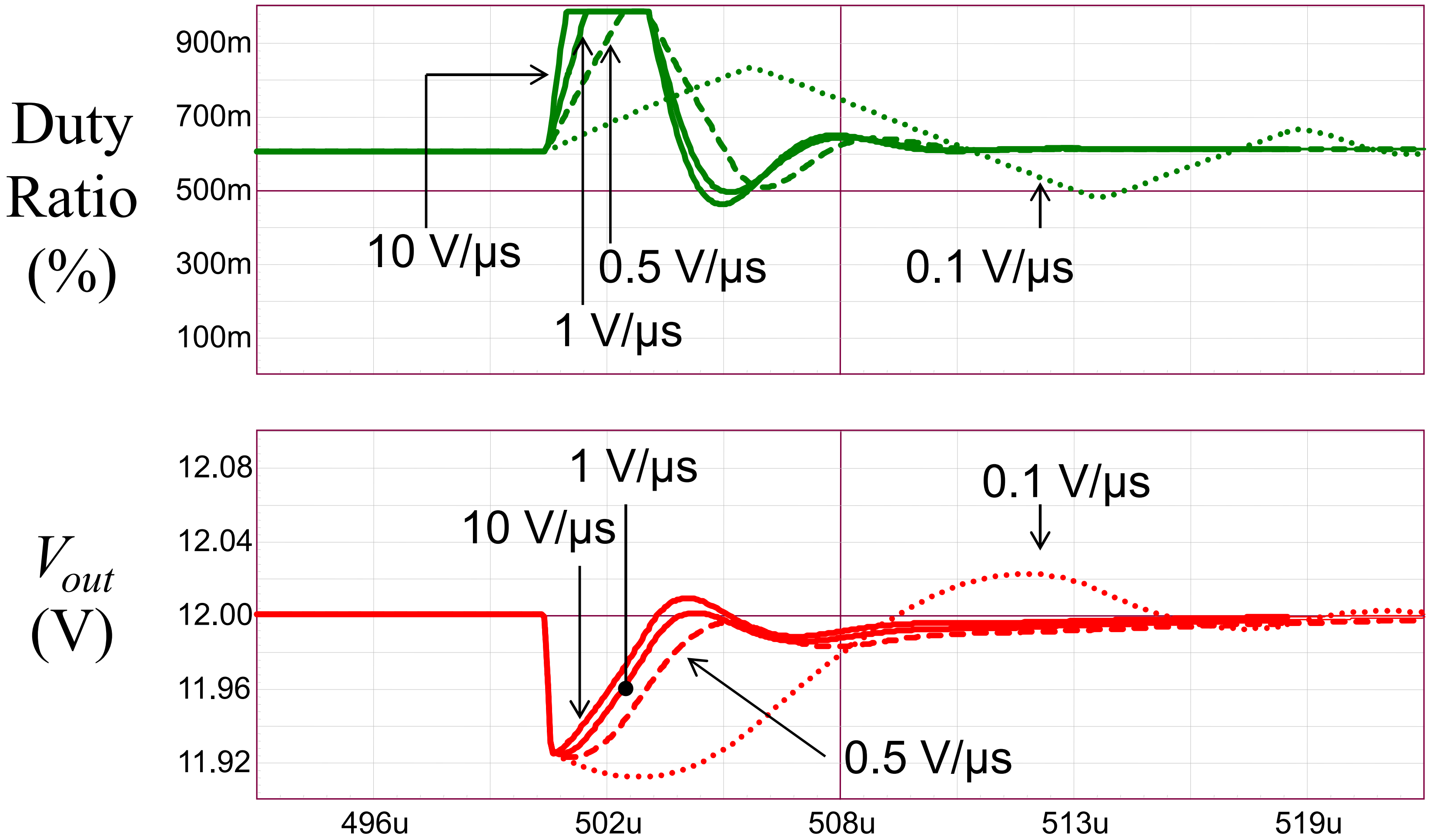
GBW=10Meg
LFP=30
AOL=20*LOG(GBW/LFP)
SR=0.1

SR adjustment



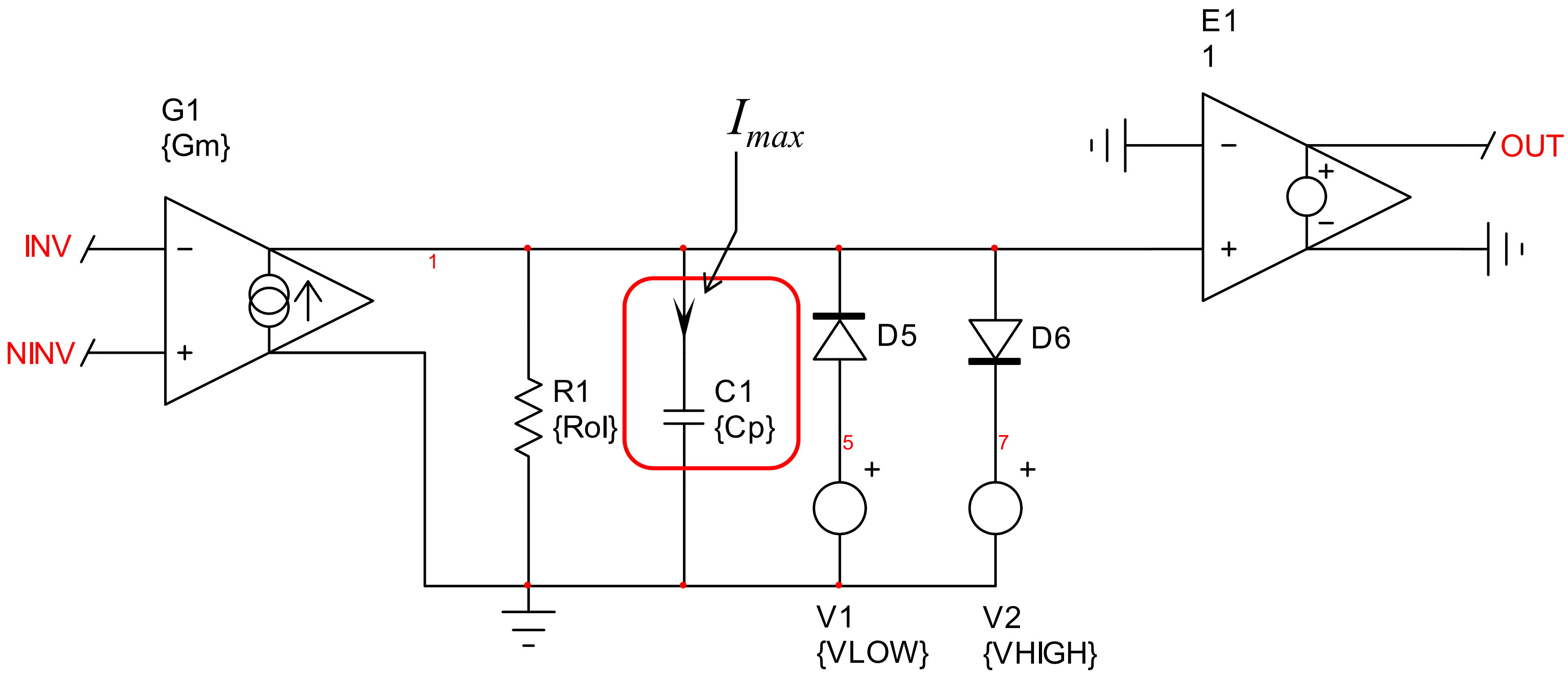
Too Low a Slew-Rate Brings Oscillations

- Changing the op amp slew-rate affects transient response



Controlling the Slew-Rate in SPICE

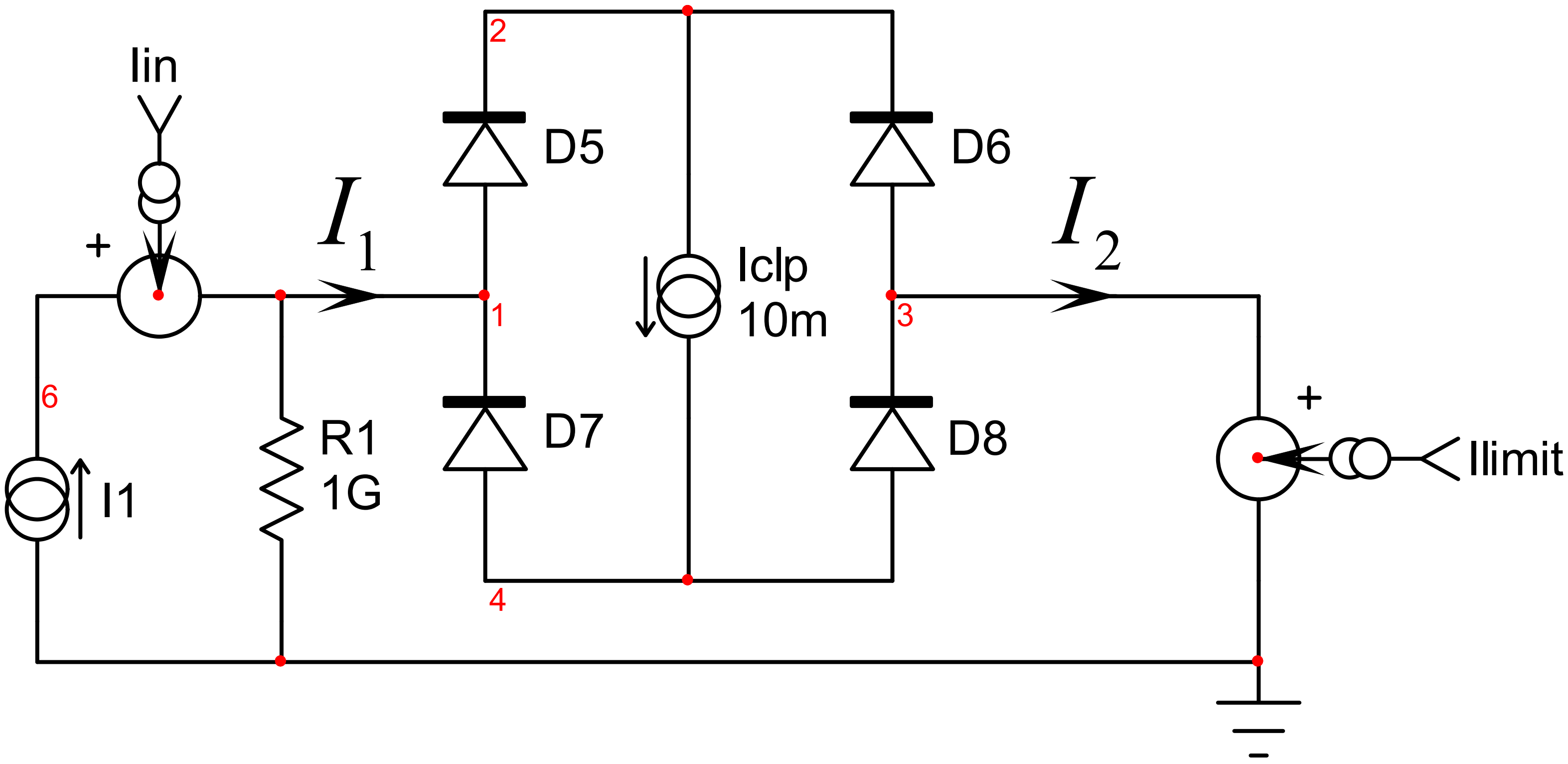
- What defines slew-rate in an op amp model?



- The current charging capacitor C_1 limits the slew-rate
- How do we clamp the current through C_1?

A Diode Bridge as a Current Clamp

❑ A diode bridge is used as a current clamp

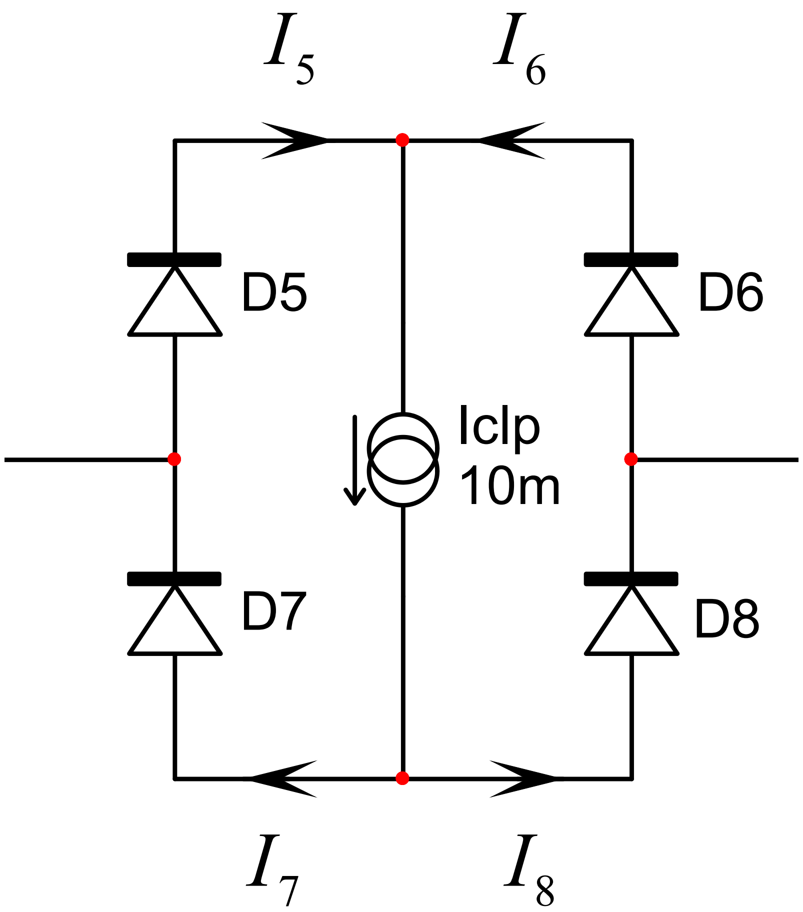


❑ When current source I_1 is less than 10 mA, $I_2 = I_1$

❑ How does the current split between diodes?

Current Source Equally Splits

- The trick is to assume all diodes are always biased
- All diodes are of similar characteristics, same V_f and r_d

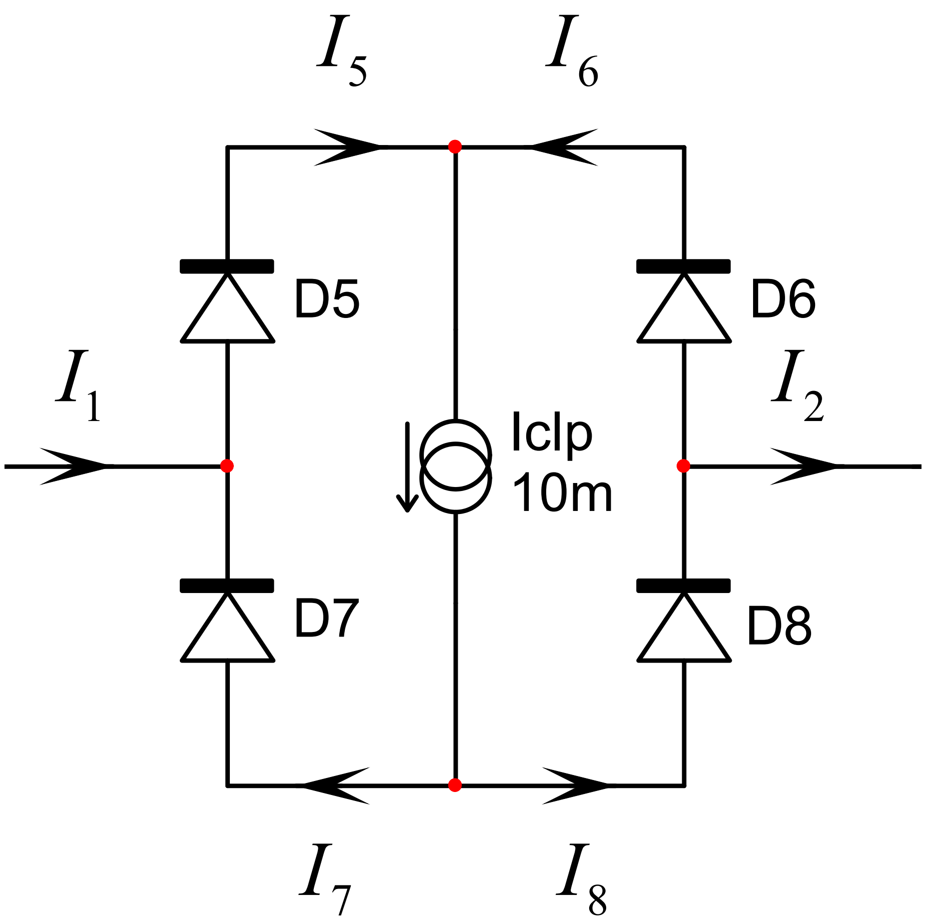


$$I_5 = I_6 = \frac{I_{clp}}{2} = 5\text{ mA}$$

$$I_7 = I_8 = \frac{I_{clp}}{2} = 5\text{ mA}$$

Splitting and Adding Currents

- Write KCL considering conducting diodes



$$I_5 + I_6 = I_{clp}$$

$$I_7 + I_8 = I_{clp}$$

$$I_7 + I_1 = I_5 \qquad I_2 + I_6 = I_8$$

subtract

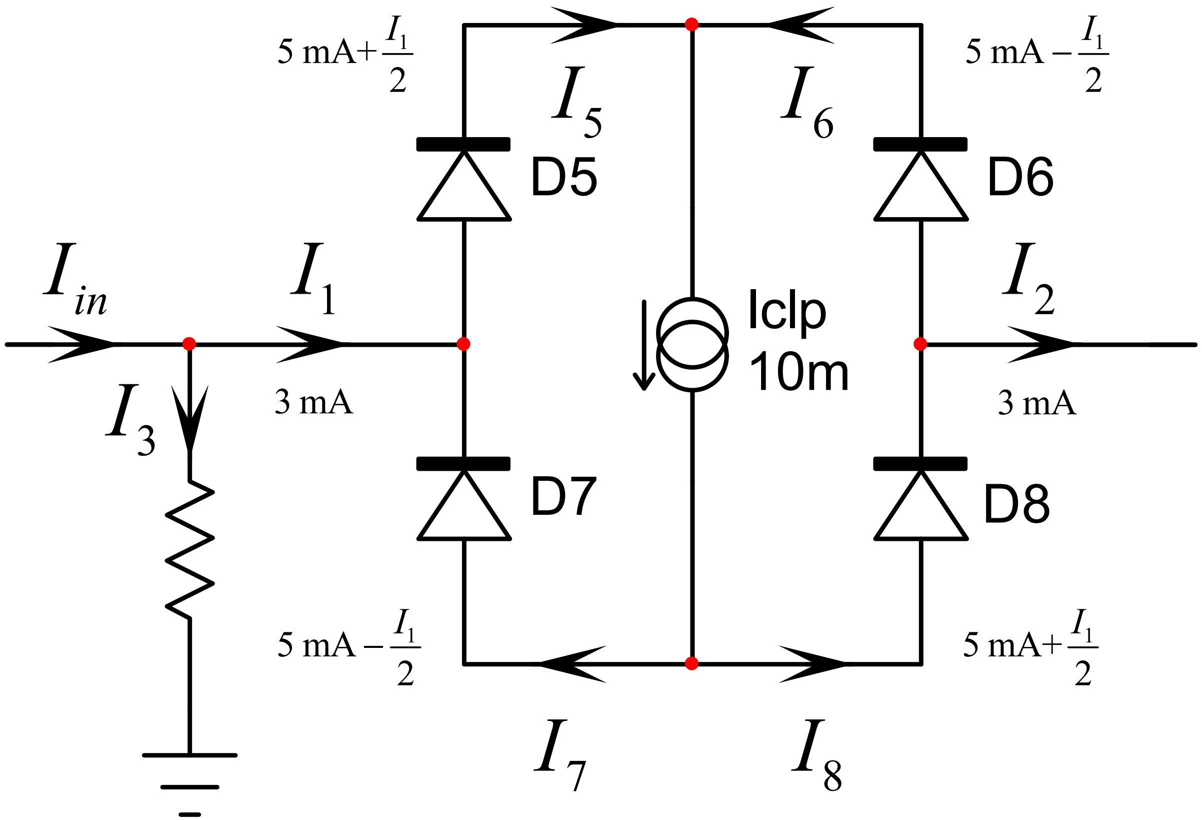
$$I_7 + I_1 - I_2 - I_6 = I_5 - I_8$$

$$I_1 - I_2 = \underbrace{\boxed{I_5 + I_6}}_{I_{clp}} - \underbrace{\boxed{\left(I_8 + I_7\right)}}_{I_{clp}} = 0$$

$$\boxed{I_1 = I_2}$$

Input Current Goes to the Output

- ❑ Diodes are always forward-biased and imply $I_1 = I_2$



$$I_1 = 3\ \text{mA}$$

$$\Downarrow$$

10 mA: $I_5 = 6.5\ \text{mA}$, $I_7 = 3.5\ \text{mA}$

10 mA: $I_8 = 6.5\ \text{mA}$, $I_6 = 3.5\ \text{mA}$

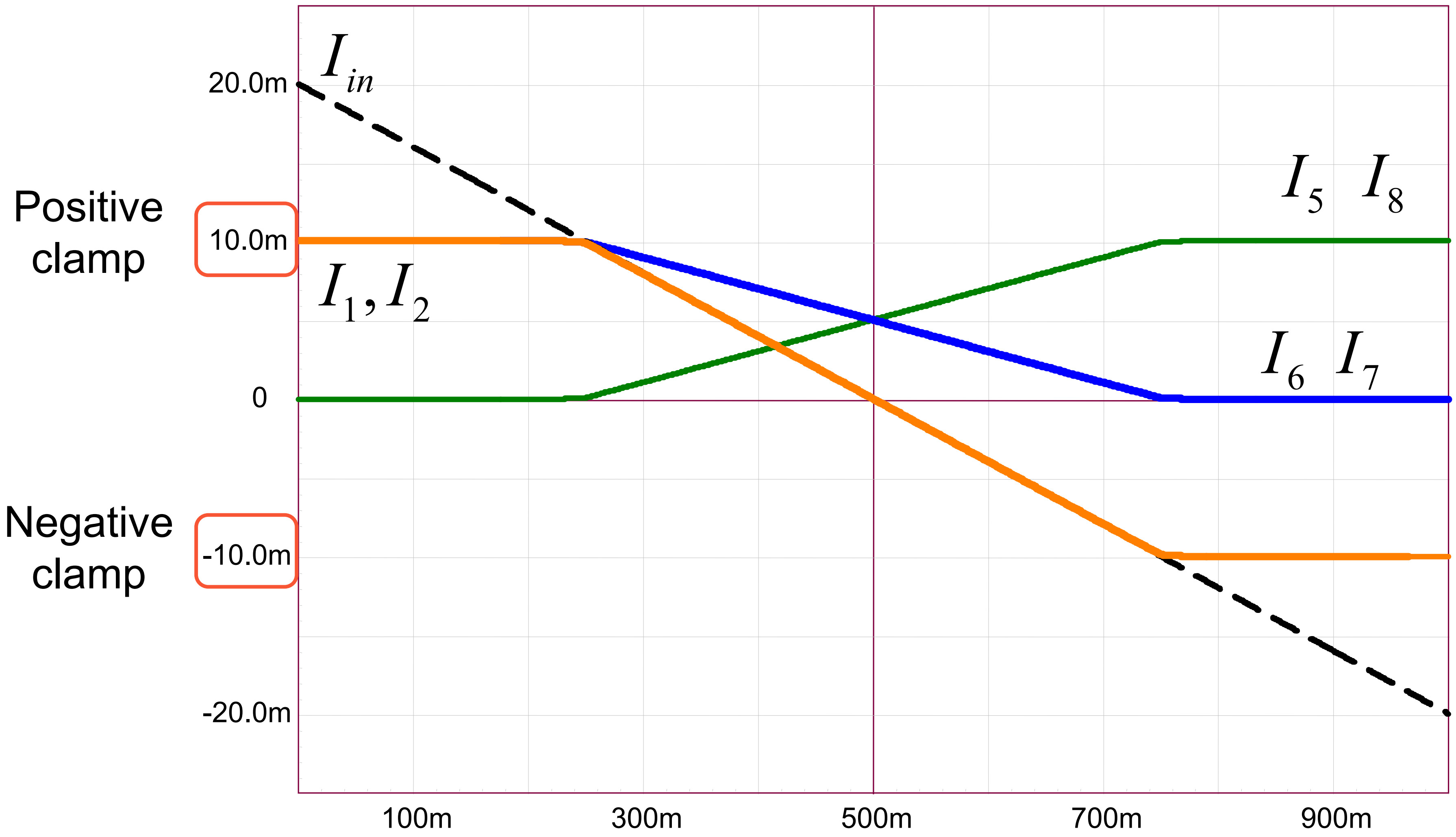
$$\Downarrow$$

$$I_2 = 3\ \text{mA}$$

- ❑ When diodes block, I_1 and I_2 are clamped to 10 mA
- ❑ I_3 is the excess current in clamped state

Clamp Circuit at Work

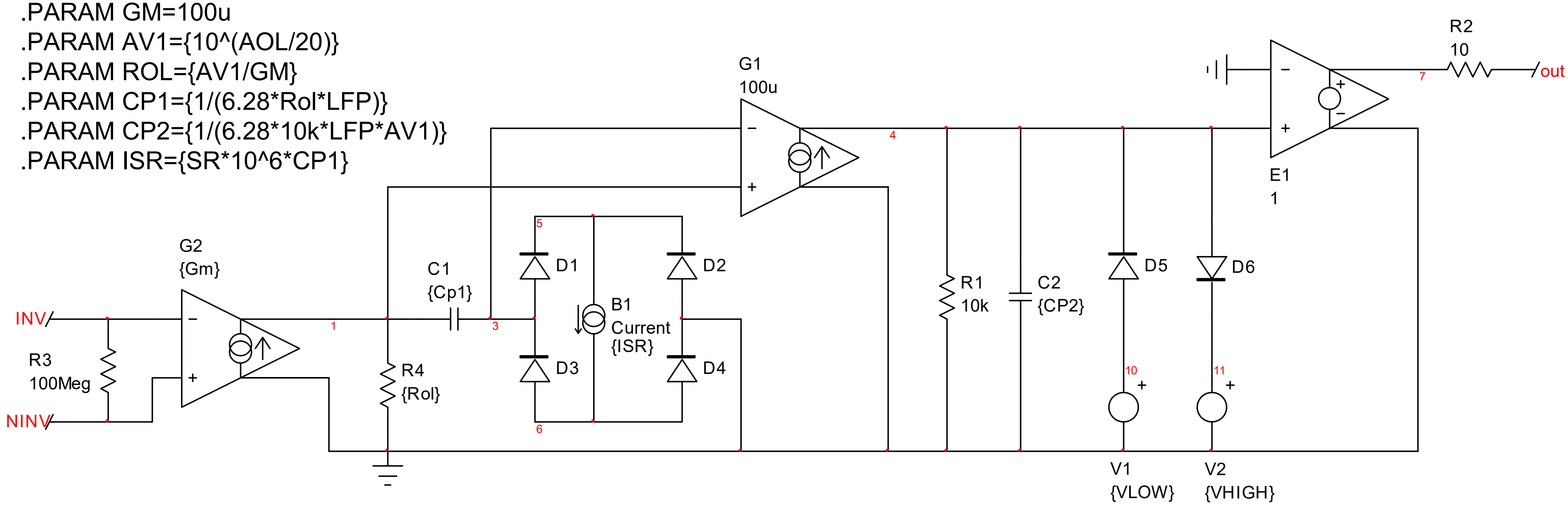
- ❑ Output current is clamped to 10 mA regardless of polarity



Slow-Down Low Frequency Pole Charging

- ❑ Create SR limit by clamping the capacitor charging current

```
.PARAM GM=100u
.PARAM AV1={10^(AOL/20)}
.PARAM ROL={AV1/GM}
.PARAM CP1={1/(6.28*Rol*LFP)}
.PARAM CP2={1/(6.28*10k*LFP*AV1)}
.PARAM ISR={SR*10^6*CP1}
```



- ❑ Complete op amp macro model featuring SR and 2nd pole
- ❑ The slope (slew rate) is I_{SR}/C_{p1}

$$\Rightarrow \quad I_{SR} = C_{p1}\mathrm{SR}\times 10^{-6}$$

V/µs

When the Op Amp Slews

- ❑ The slewing op amp distorts the input signal

  ⇨ Check expected slope is always smaller than max SR

- ❖ What is the highest slope of a sinusoidal signal?

$$S(t)=\frac{d}{dt}V_p\sin(\omega t)=V_p\cdot\omega\cdot\cos(\omega t)$$

$$\omega t = 0$$

$$\frac{d}{dt}S(t)=0$$

$$\Downarrow$$

$$-V_p\omega^2\sin(\omega t)=0$$

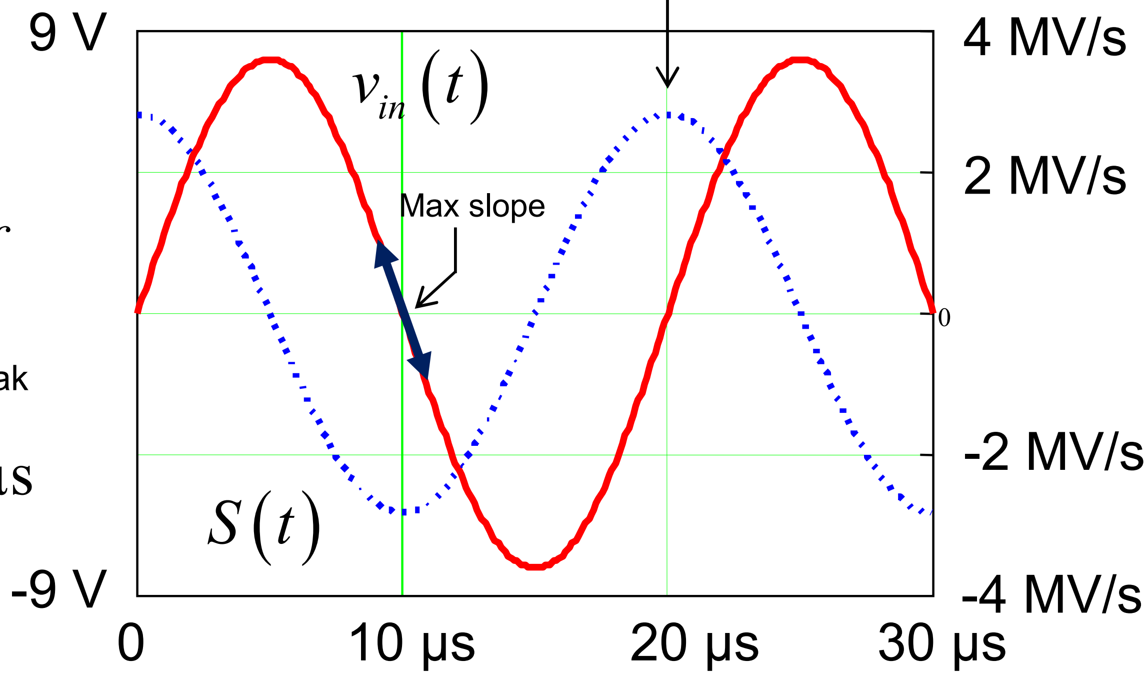
$$\Downarrow$$

$$\omega t = 0$$

$$SR = V_p 2\pi f$$

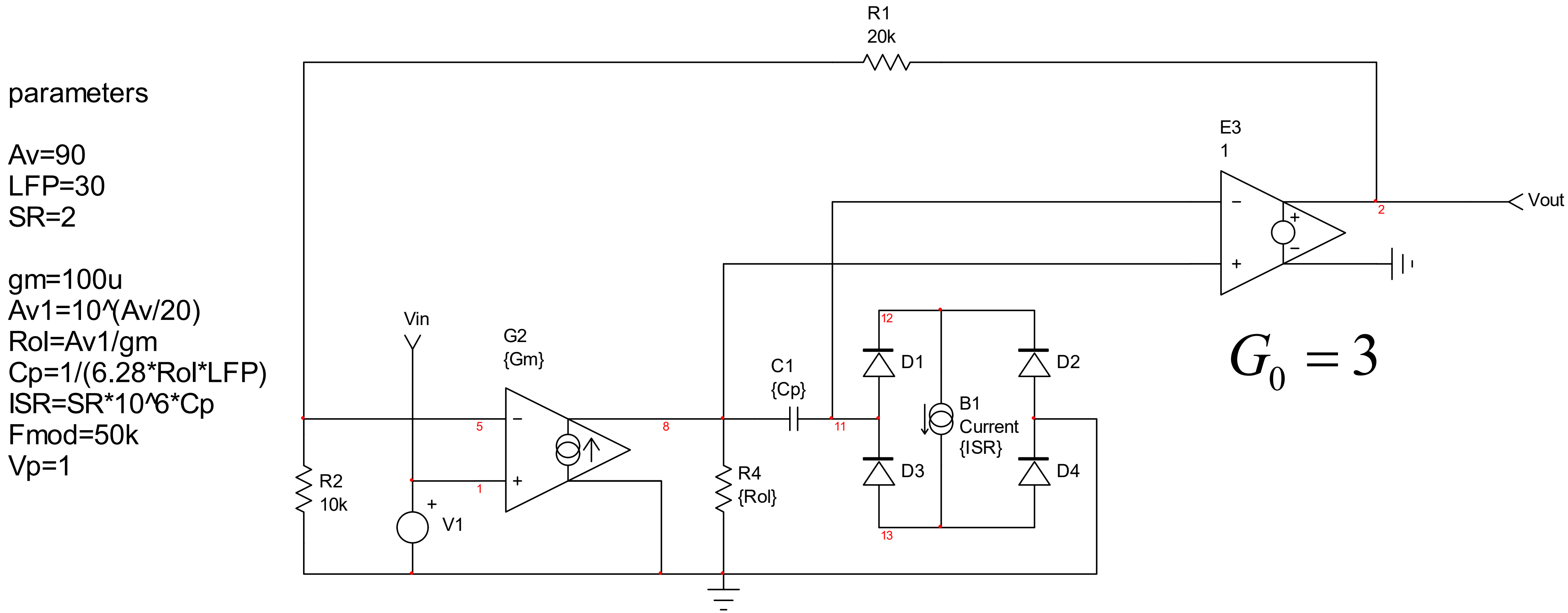
$$\Downarrow$$ 50 kHz, 9 V peak

$$SR = 2.8\ \mathrm{V/\mu s}$$



Testing the Model in Transient

- A 50-kHz 1-V_{peak} waveshape drives an SR-limited op amp



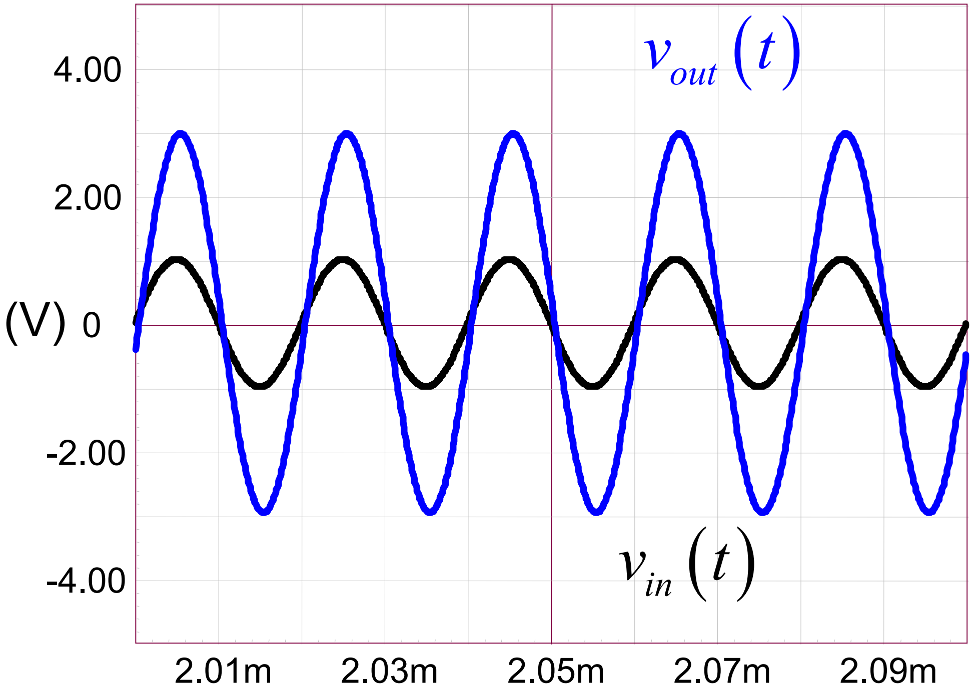
- The signal max SR is: $SR(V_{in}) = V_p 2\pi f = 1 \times 6.28 \times 50k \approx 0.3\ \text{V}/\mu\text{s}$

$G_0 = 3$

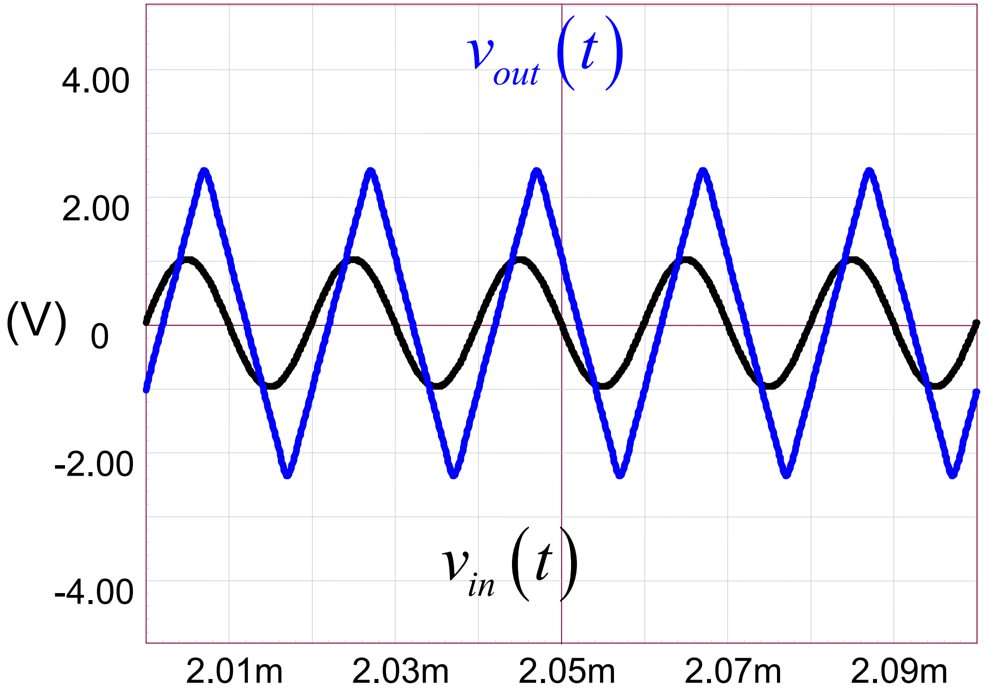
$$\Rightarrow SR(V_{out}) = V_p 2\pi f = 3 \times 6.28 \times 50k \approx \boxed{1\ \text{V}/\mu\text{s}}$$

Slewing in Action

- ❑ The slew-rate is adjusted while the output is monitored



SR = 2 V/µs

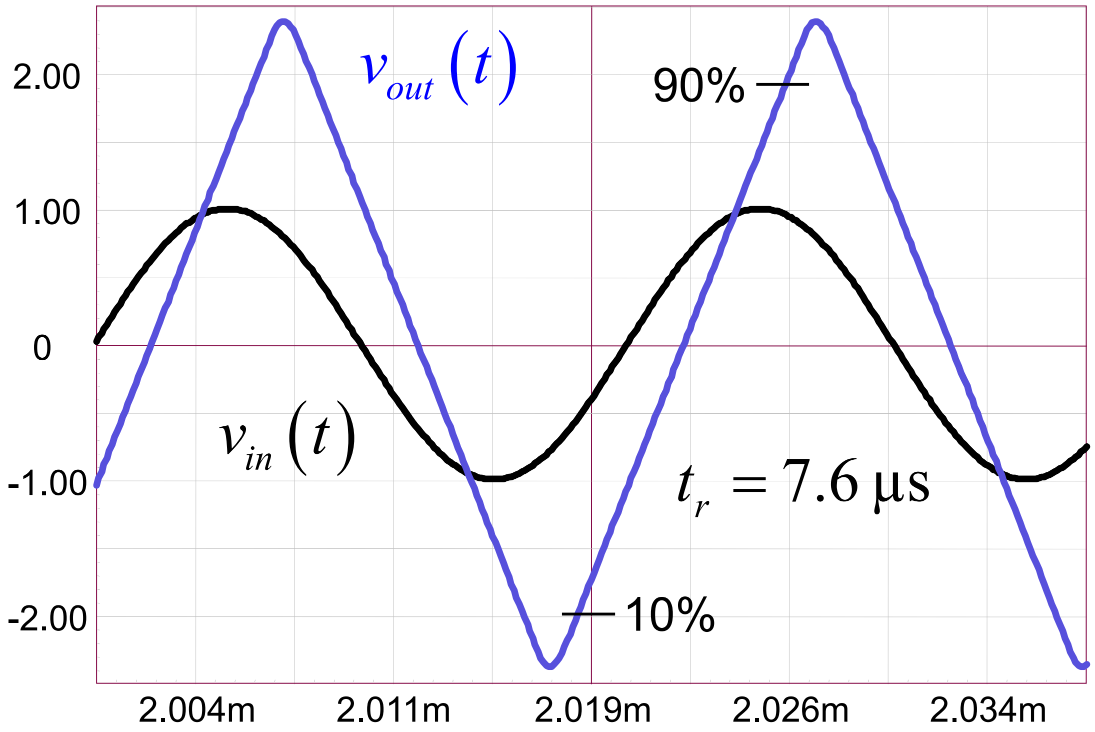


SR = 0.5 V/µs

- ❑ If the op amp slews, it cannot track the input signal

Adding a Low-Frequency Pole

❑ A slewing op amp behaves like adding a low-frequency pole



For a 1st-order system:

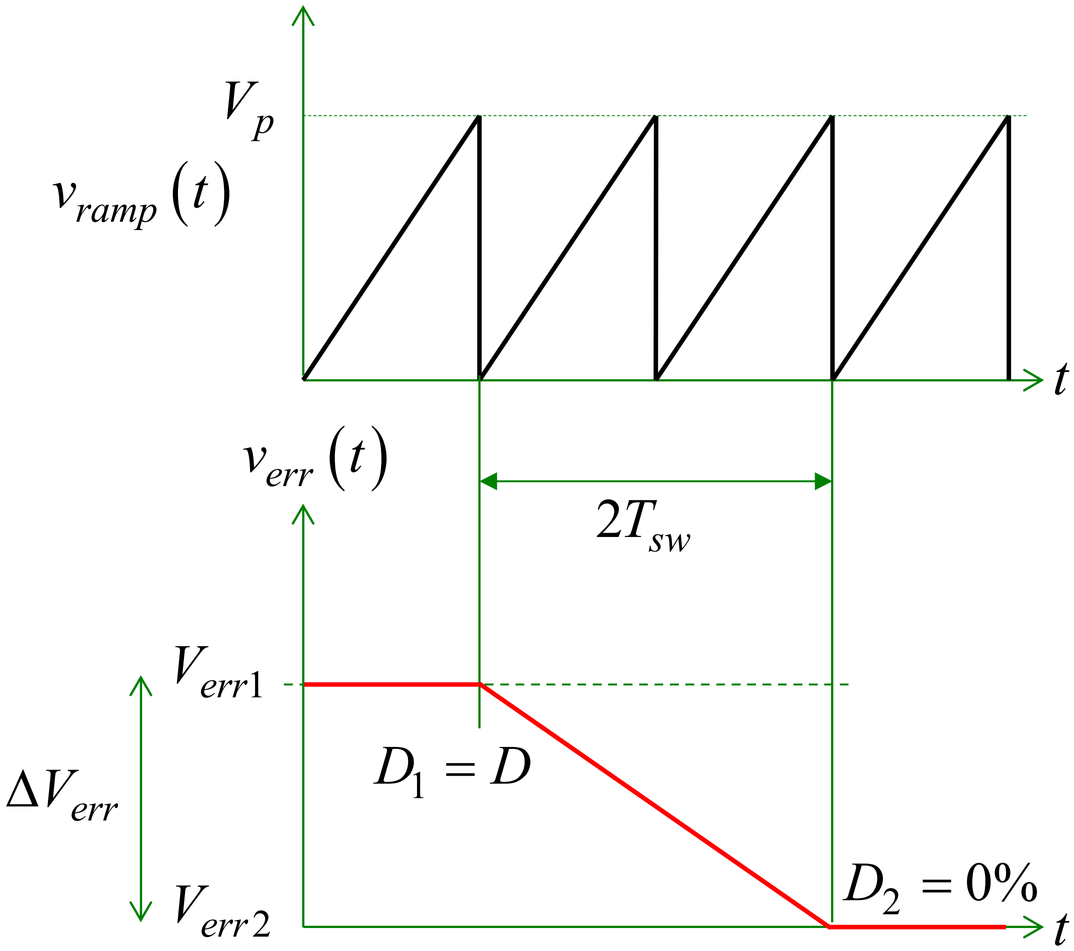
$$f_p \approx \frac{0.35}{t_r} = \frac{0.35}{7.6u} = 46\ \text{kHz}$$

❑ A new pole is added to the system

⇨ Phase margin degradation, oscillatory response

Slew Rate and Pulse Width Modulation (1)

- 1st case, the load is released and duty ratio drops



$$SR_1 = \frac{\Delta V_{err}}{2T_{sw}} = \frac{V_{err1} - V_{err2}}{2T_{sw}}$$

$$D = \frac{V_{err}}{V_p}$$

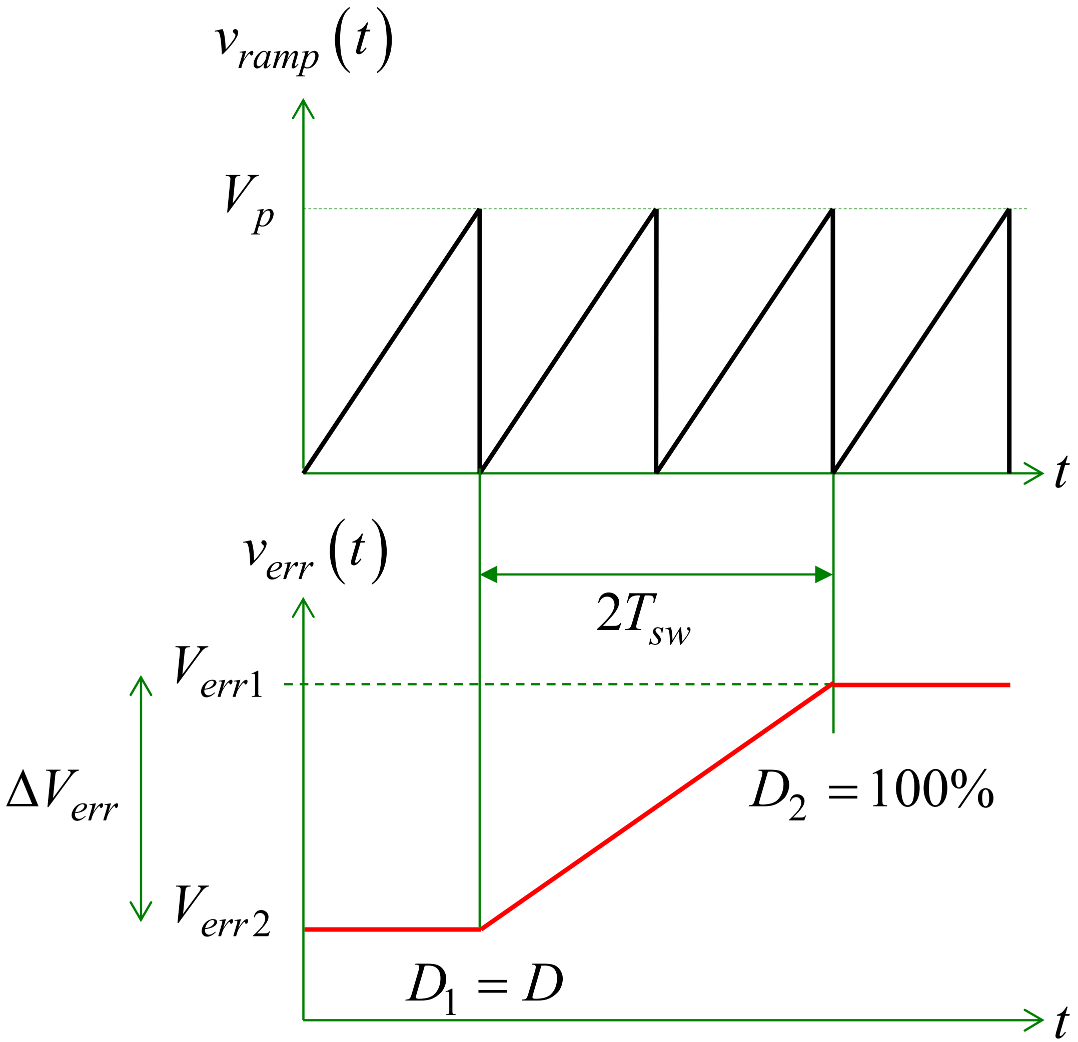
buck

$$SR_1 = \frac{DV_p - 0}{2T_{sw}} = \boxed{\frac{V_p V_{out}}{2V_{in,\min}} F_{sw}}$$

M. Jovanović et al. “Design Considerations for Low-Voltage On-Board Dc-Dc Modules”, IEEE Transactions on Power Electronics, Vol. 11, no 2, March 1996

Slew Rate and Pulse Width Modulation (2)

- 2nd case, the load is connected and duty ratio jumps high



$$SR_2 = \frac{\Delta V_{err}}{2T_{sw}} = \frac{V_{err1} - V_{err2}}{2T_{sw}}$$

$$D = \frac{V_{err}}{V_p}$$

buck

$$SR_2 = \frac{V_p - DV_p}{2T_{sw}} = \frac{V_p(1-D)}{2T_{sw}}$$

$$\boxed{SR_2 = \frac{V_p}{2}\left(1 - \frac{V_{out}}{V_{in,\max}}\right)F_{sw}}$$

Slew Rate Selection Example

- ❑ Take the highest result and select the op amp SR

$$V_{in} = 20\text{ V}\ V_{out} = 12\text{ V}\ F_{sw} = 1\text{ MHz}\ V_p = 2.5\text{ V}$$

$$SR_1 = \frac{V_p V_{out}}{2V_{in}} F_{sw} = \frac{2.5 \times 12}{2 \times 20} 10^6 = 750\,\text{kV/s} = 0.75\,\text{V/µs}$$

$$SR_2 = \frac{V_p}{2}\left(1 - \frac{V_{out}}{V_{in}}\right) F_{sw} = \frac{2.5}{2}\left(1 - \frac{12}{20}\right) 10^6 = 500\,\text{kV/s} = 0.5\,\text{V/µs}$$

Select an op amp with a SR at least equal to 1 V/µs

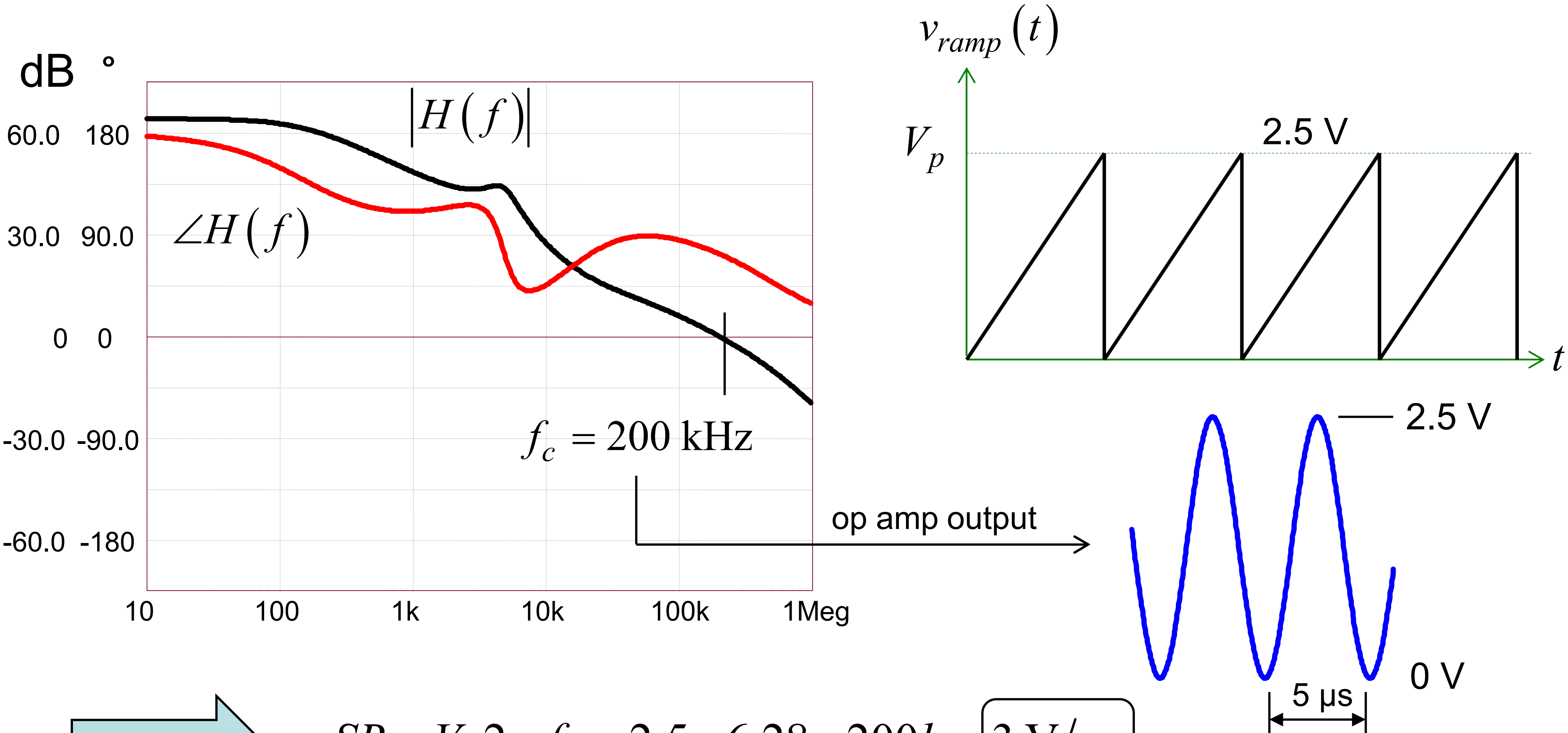
TLV271

Slew Rate at Unity Gain	SR	$V_{O(pp)} = V_{DD}/2$, $R_L = 10\ k\Omega$, $C_L = 50\ pF$	$V_{DD} = 2.7\ V$	1.35	2.1		V/µS
		$T_A = -40°C$ to $+105°C$		1			
		$V_{O(pp)} = V_{DD}/2$, $R_L = 10\ k\Omega$, $C_L = 50\ pF$	$V_{DD} = 5\ V$	1.45	2.3		
		$T_A = -40°C$ to $+105°C$		1.2			
		$V_{O(pp)} = V_{DD}/2$, $R_L = 10\ k\Omega$, $C_L = 50\ pF$	$V_{DD} = \pm 5\ V$	1.8	2.6		
		$T_A = -40°C$ to $+105°C$		1.3			

- ❑ Watch for SR variations depending on op amp configuration!

Look at the Open-Loop Bode Plot

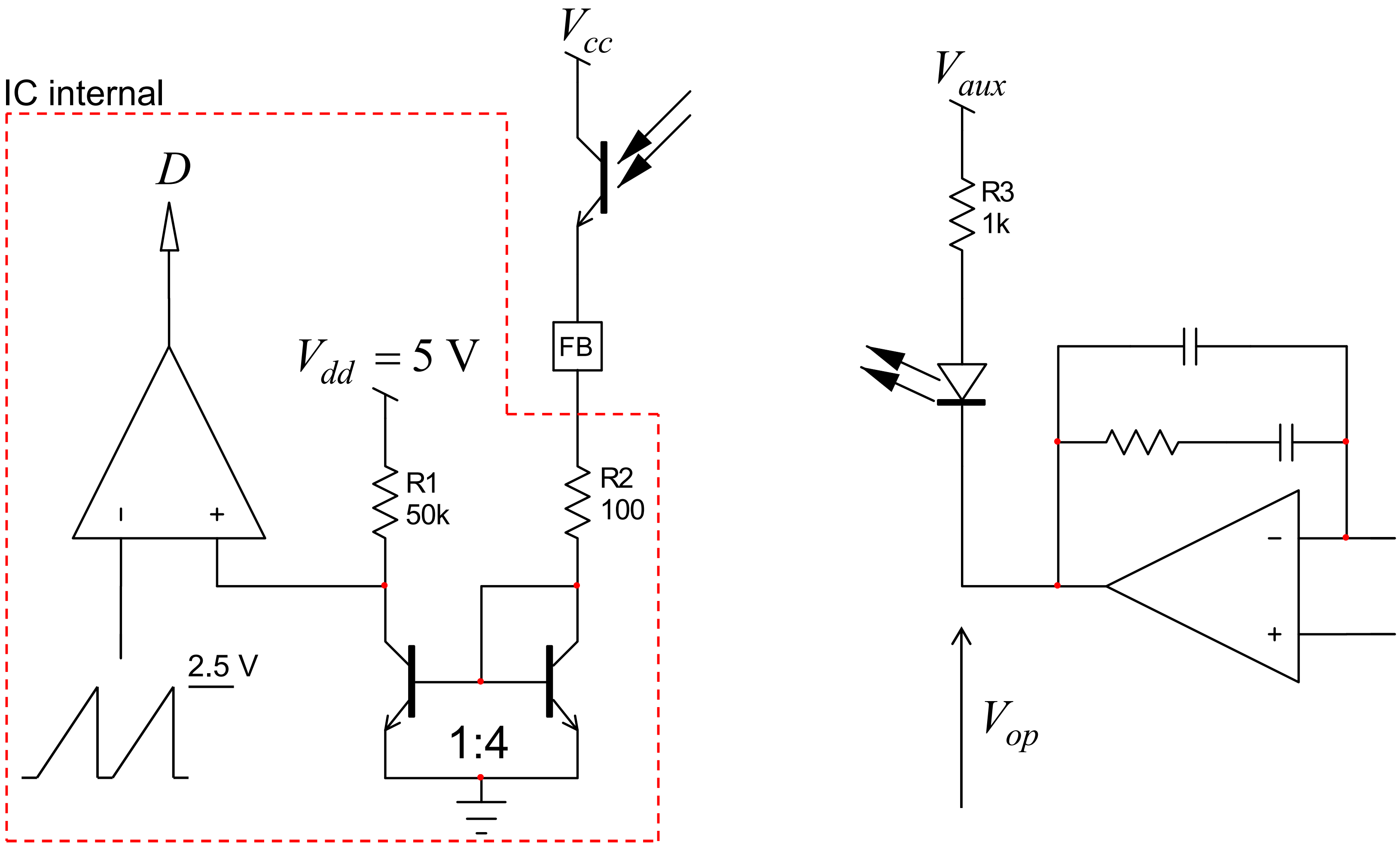
- ❑ The system exhibits gain up to 200 kHz



$$SR = V_p 2\pi f_c = 2.5 \times 6.28 \times 200k \approx \boxed{3\ \text{V/µs}}$$

Slew Rate Depends on Architecture

- ❑ For opto-isolated dc-dc, op amp dynamics is reduced



- ❑ What is the op slew rate to sweep the 2.5-V ramp?

NCP1565 – Active Clamp Controller

A Lower SR for the Opto-Isolated Version

- The 2.5-V ramp must be swept from 0 to 2.5 V
- An external resistor adjusts the maximum duty ratio

$$I_{R_1} = \frac{V_{dd} - V_{\mathrm{CE,sat}}}{50k} \approx 100\ \mu\mathrm{A} \xrightarrow[1:4]{\text{mirror}} I_{FB} \approx 400\ \mu\mathrm{A}$$

- Considering a 60% CTR worst case, the LED current is

$$I_{LED} \approx 670\ \mu\mathrm{A}$$

- Across the $1\text{-k}\Omega$ resistor, it induces a voltage swing of

$$V_{op,ac} = 670u \times 1k = 670\ \mathrm{mV}$$

- If we adopt 0.67 V, then for a 200-kHz crossover

$$SR = V_p 2\pi f_c = 0.67 \times 6.28 \times 200k \approx 0.84\ \mathrm{V/\mu s}$$

⇒ Select an op amp with a 1-V/µs slew rate

Course Agenda

- ❑ Control System Basics
- ❑ Operational Amplifier and Low-Frequency Pole
- ❑ Gain-Bandwidth Impact on Phase Boost
- ❑ Op Amp Slew Rate Effects in Loop Control
- ❑ **Start-Up Sequence and Auxiliary Supply**
- ❑ Characterizing the Optocoupler Pole
- ❑ Dealing with the Fast Lane
- ❑ Going Around the TL431 Fast Lane

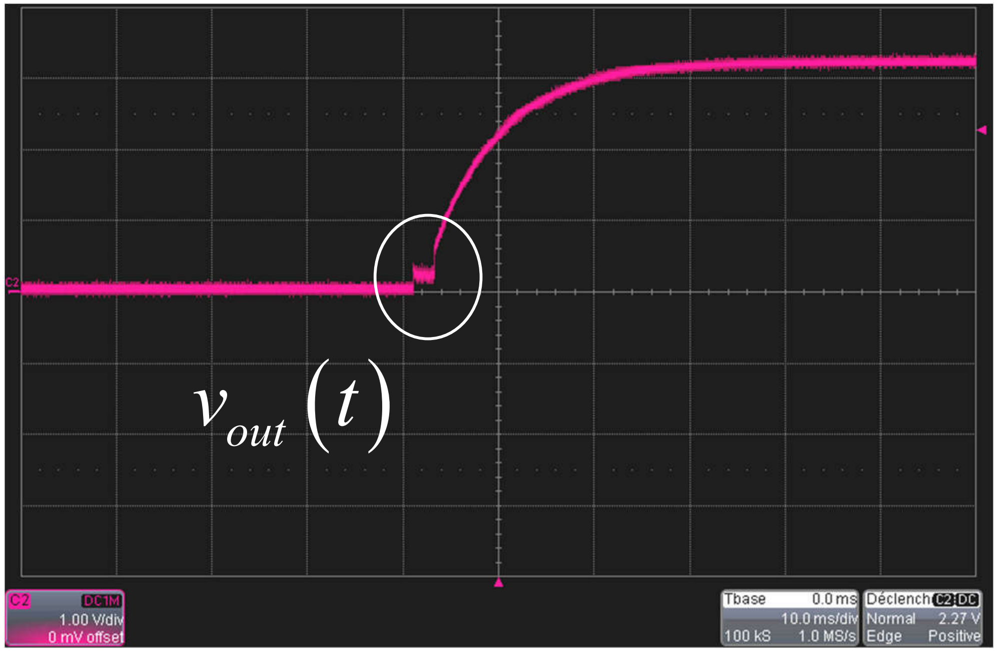
Start Up Characteristics for Dc-Dc Bricks

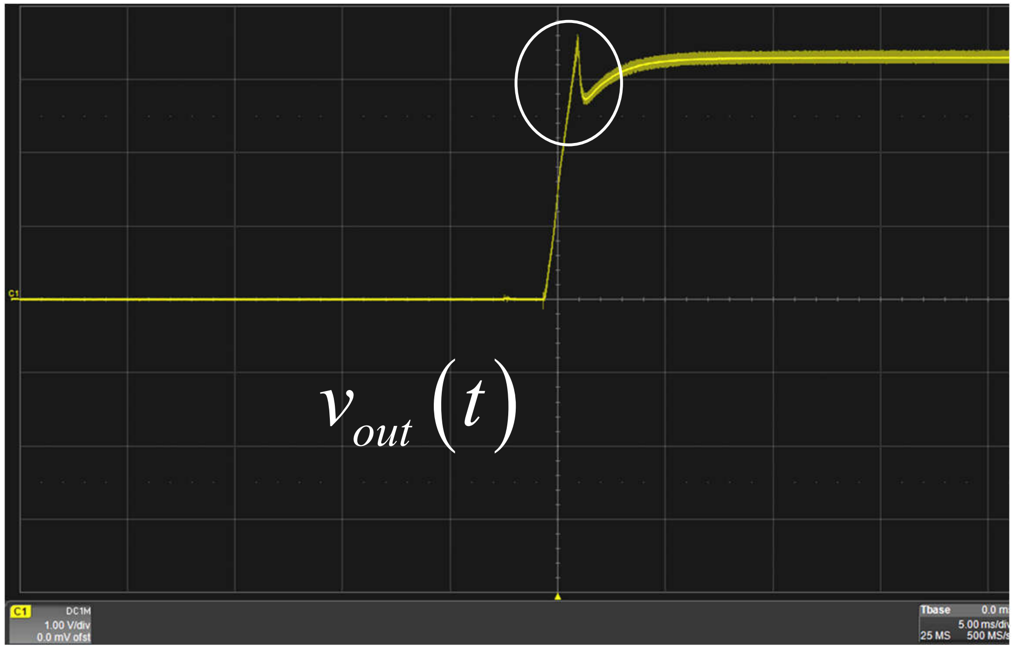
❑ Monotonicity of V_{out} at start-up is important:

❖ no hiccup or glitch while V_{out} ramps up

❖ no negative slope at any time during start up

❖ no overshoot of any kind, smooth V_{out} landing





❑ How do you prevent these from happening?

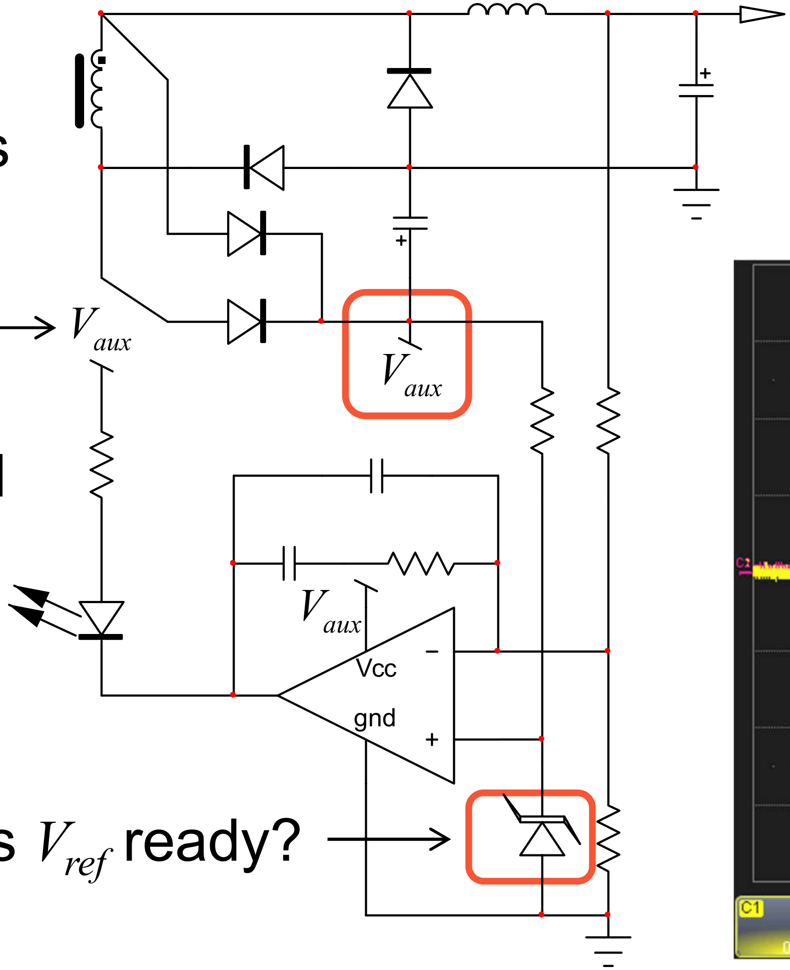
ON Semiconductor®

Controlling V_{out} at Start Up

- In a power supply, the loop at start-up is blind to V_{out}
  - Output control must occur at the very first switching cycles

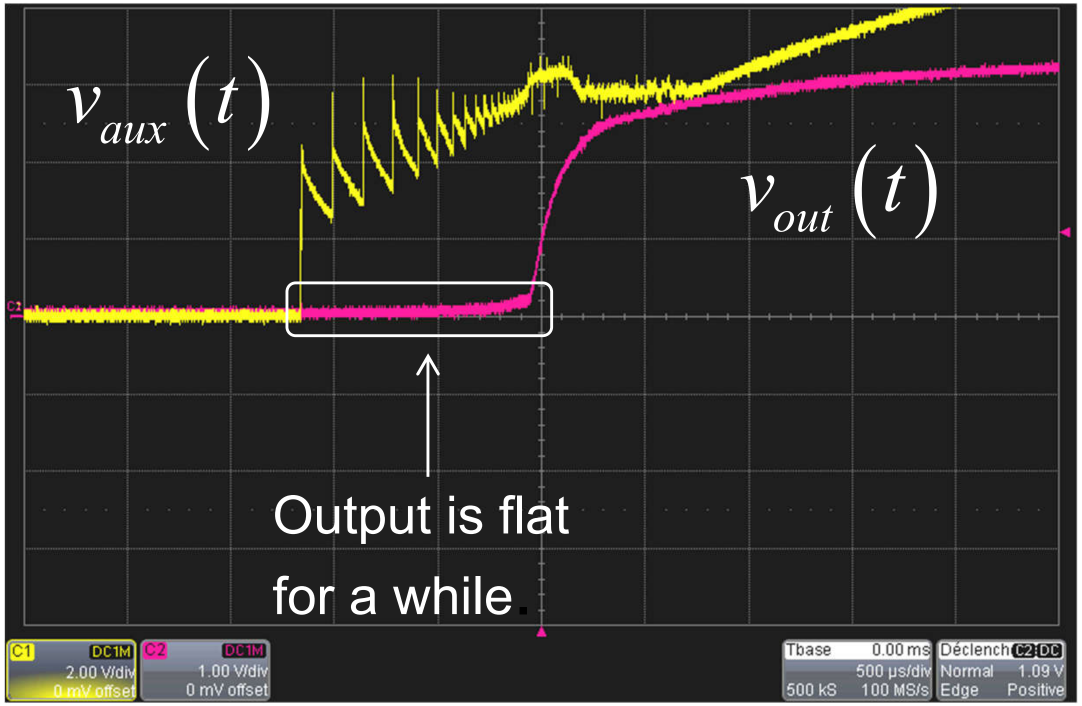
How does V_{aux} rise?

No V_{aux} no control



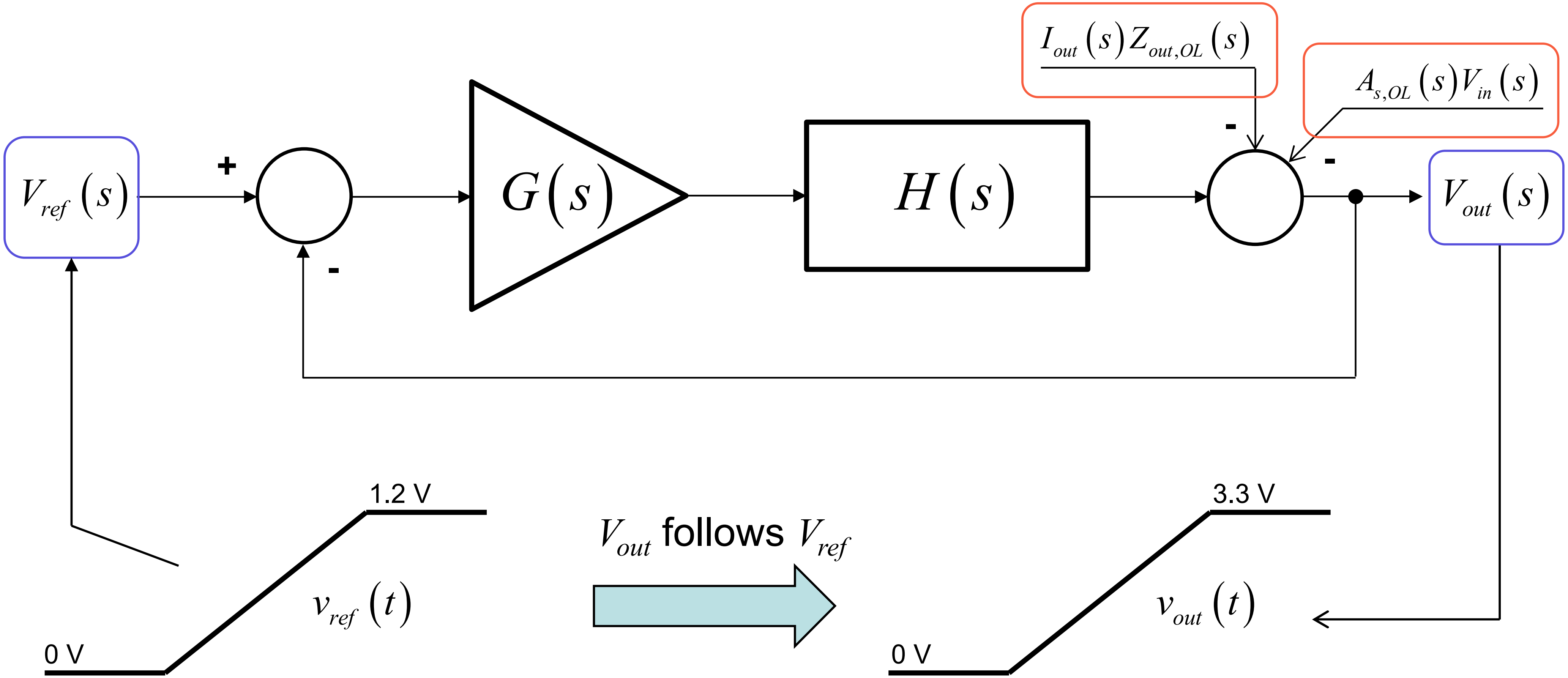
Is V_{ref} ready?

Typical ugly start up



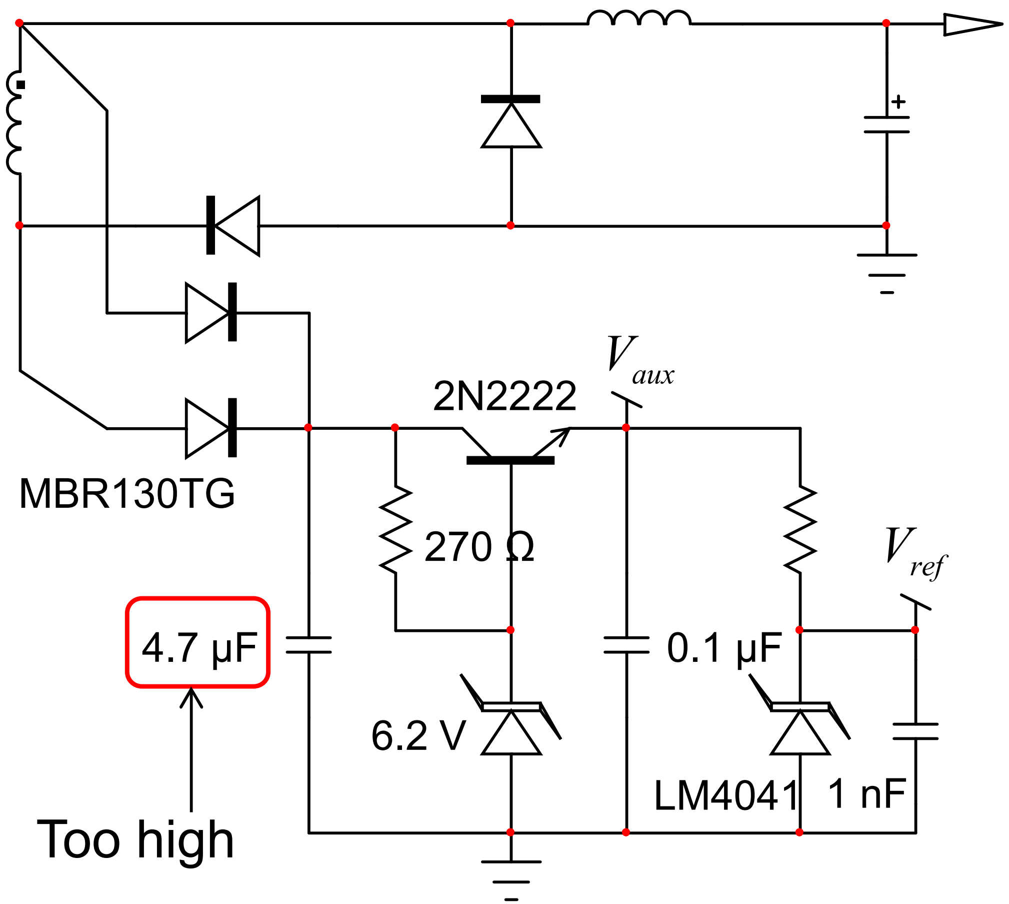
A Two-Loop System

- At power up, V_{ref} takes over and V_{out} follows it
- When target is reached, loop observes V_{out} only

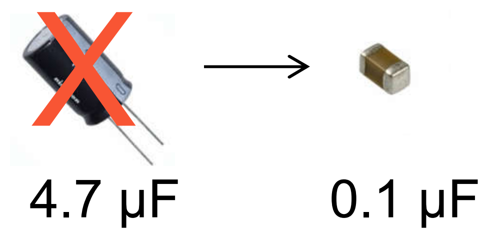


Immediate Auxiliary Voltage Availability

- ❑ The key is to have V_{aux} showing up immediately
- ❖ V_{ref} and the op amp are biased
- ❖ Control current exists to take the lead

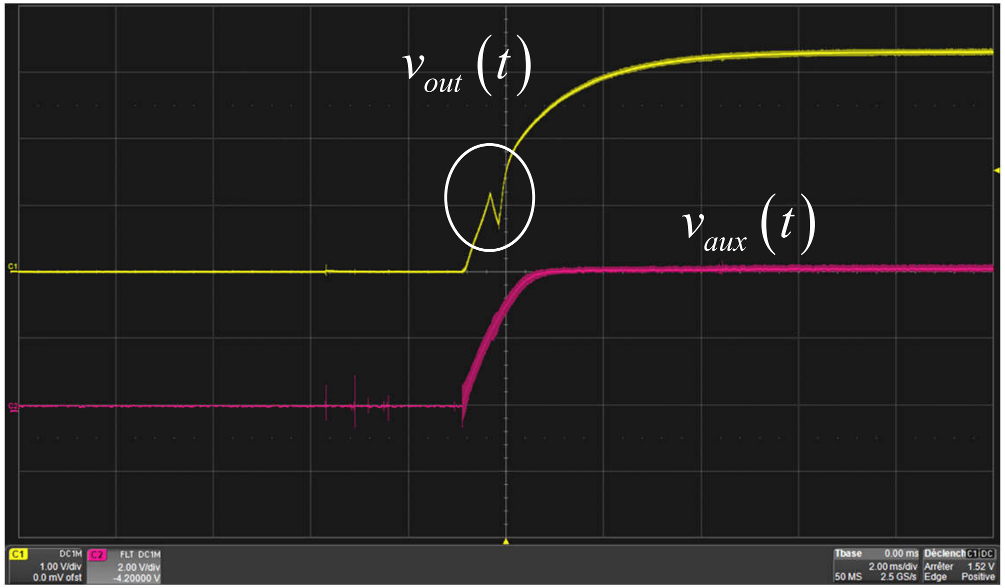


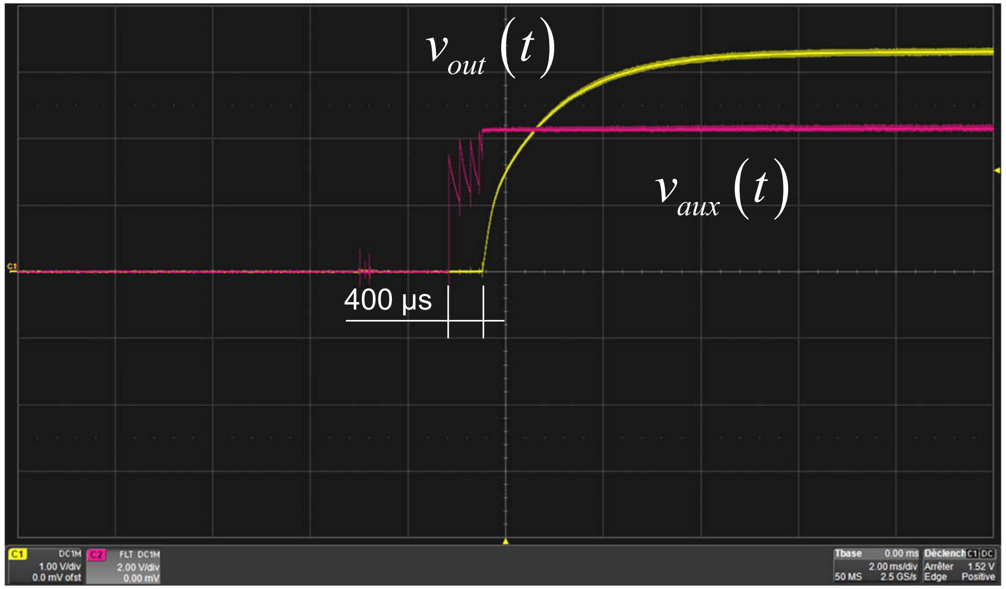
If decoupling capacitors are too big, V_{aux} slugs!



Optimum Capacitor Value

- The auxiliary voltage has to come up immediately





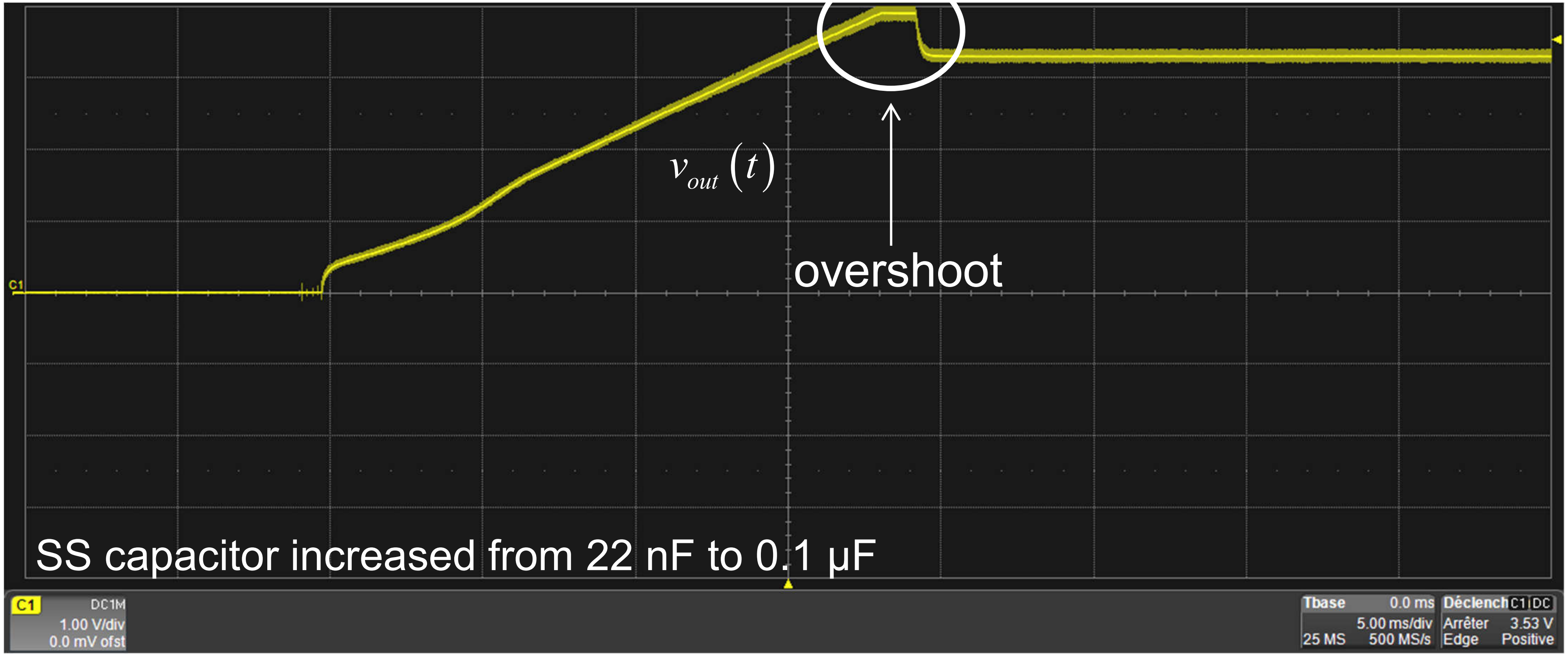
V_{aux} is late, discontinuity
C = 4.7 µF

V_{aux} is immediate, no discontinuity
C = 0.1 µF

- The secondary side must take the lead: V_{out} follows V_{ref}

Too Much Soft Start Brings Overshoot

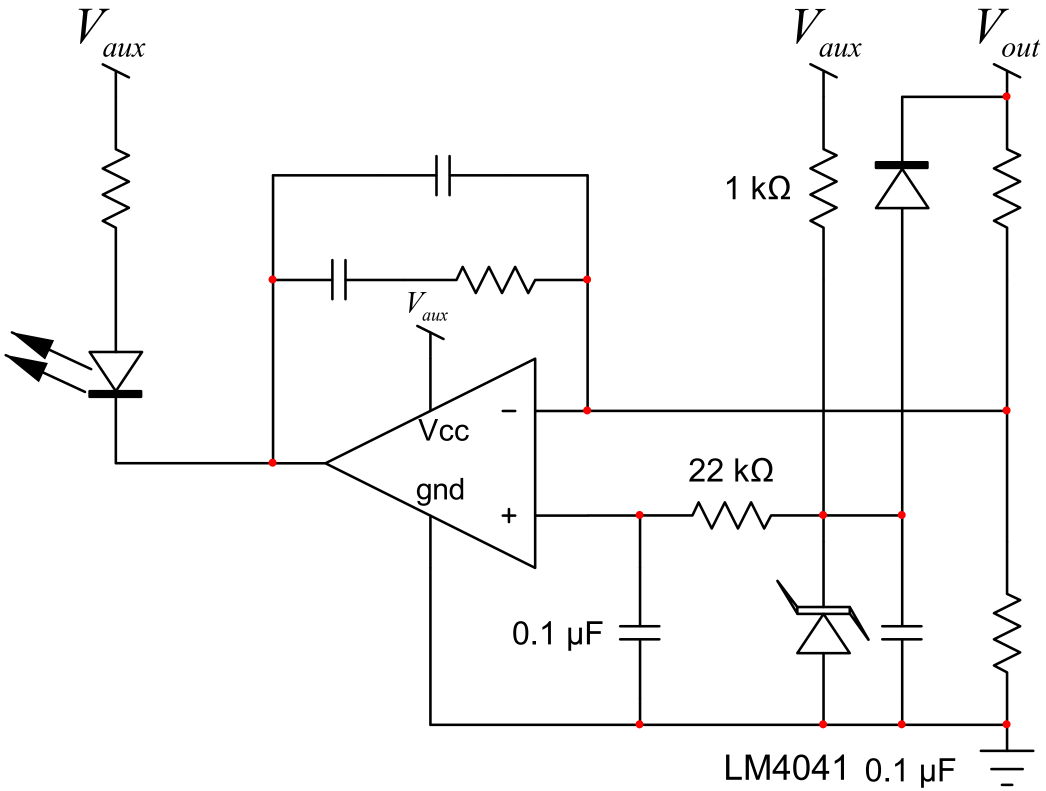
- ❑ Primary-side duty ratio soft start must not be oversized

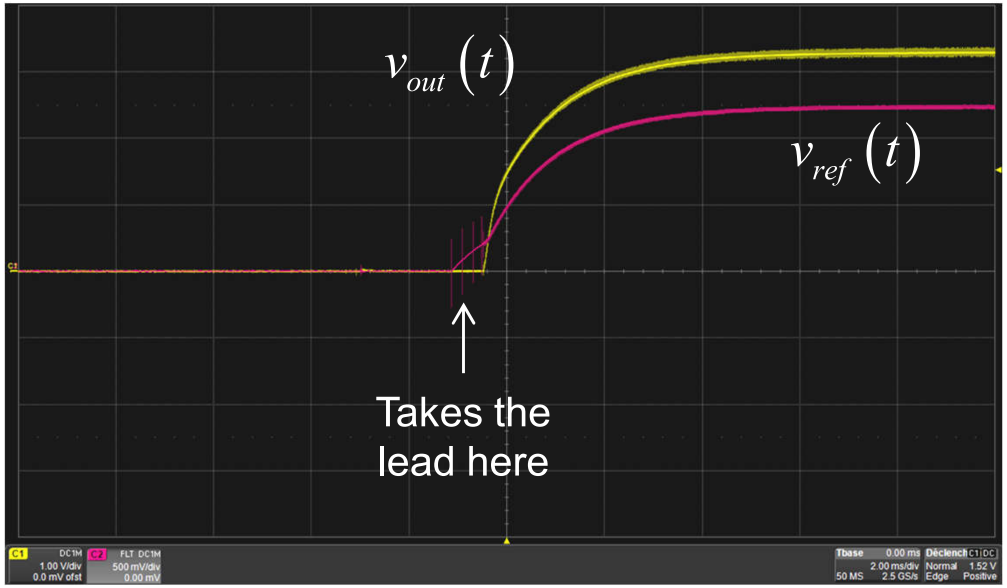


- ❑ Compensation capacitors impose op amp recovery time

Soft-Started V_{ref} is Key to Smooth Start-Up

- ❑ Monotonic output is obtained when V_{ref} is soft-started

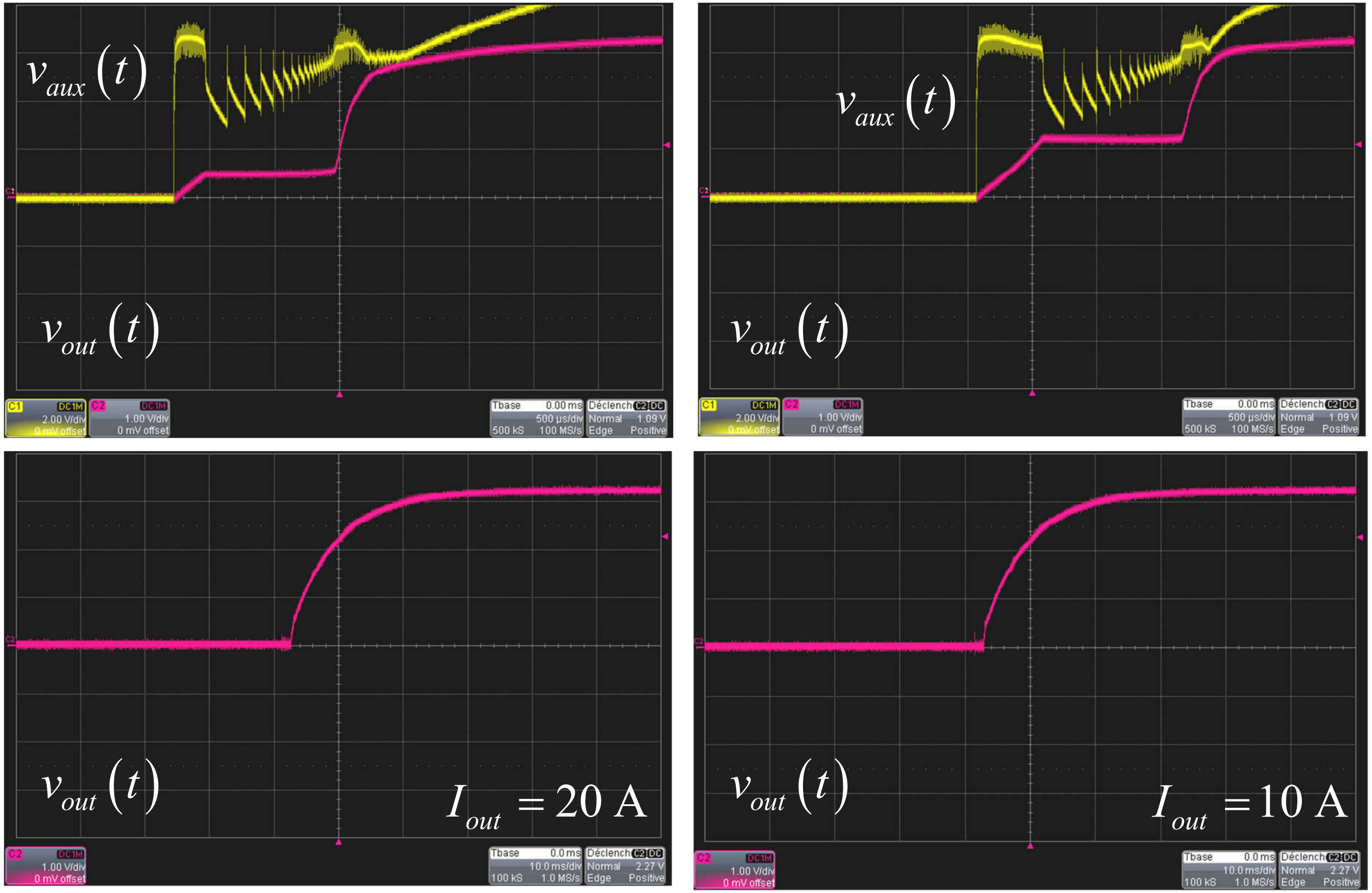




- ❑ The diode is important to discharge the SS cap. at power off

Clean Start-Up Sequences

- ❑ Apply these rules to a prototype and see immediate changes!



Raw results

After optimization

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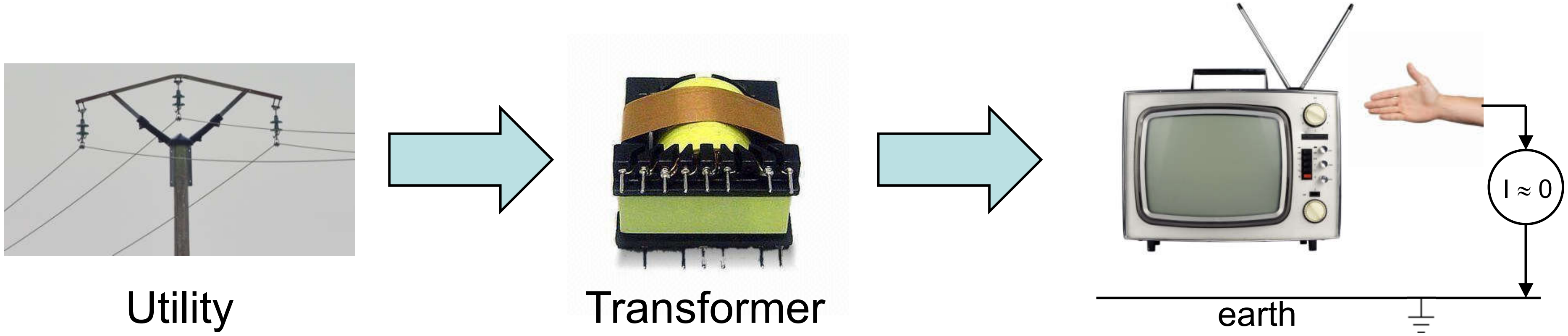
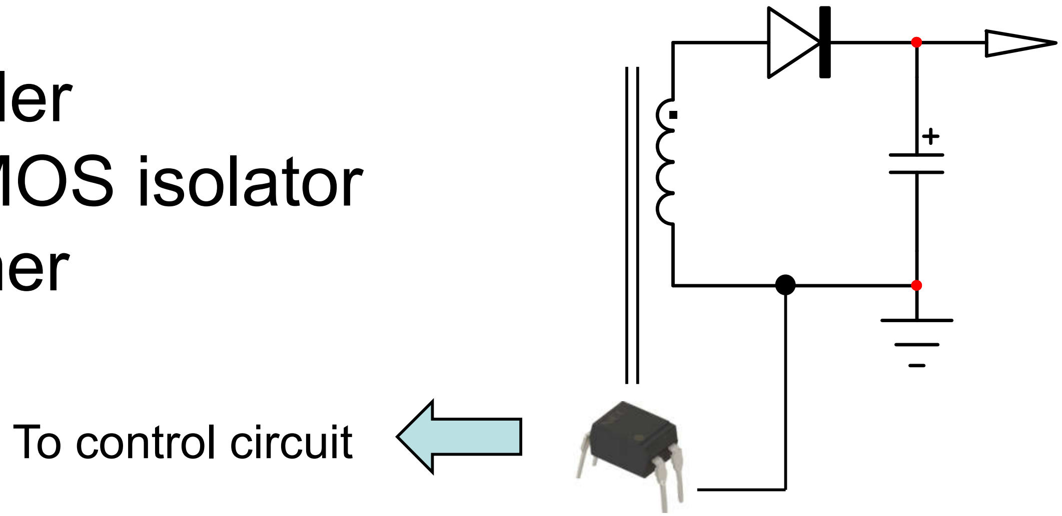
Isolated Switching Converters

- ❑ The transformer ensures isolation from source to load



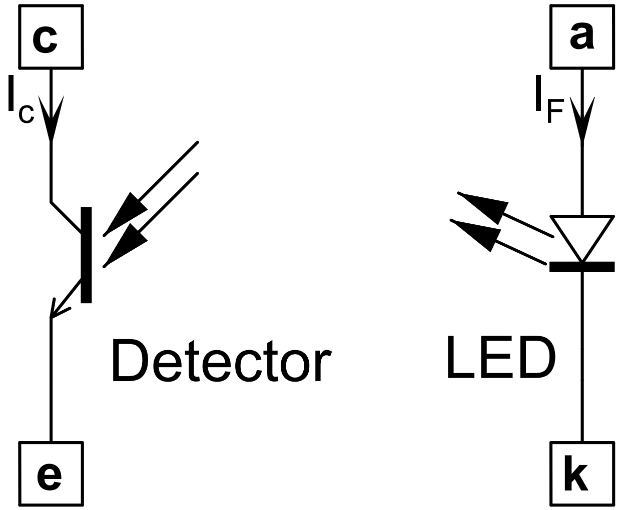
- ❑ In closed-loop systems, the information must be fed back

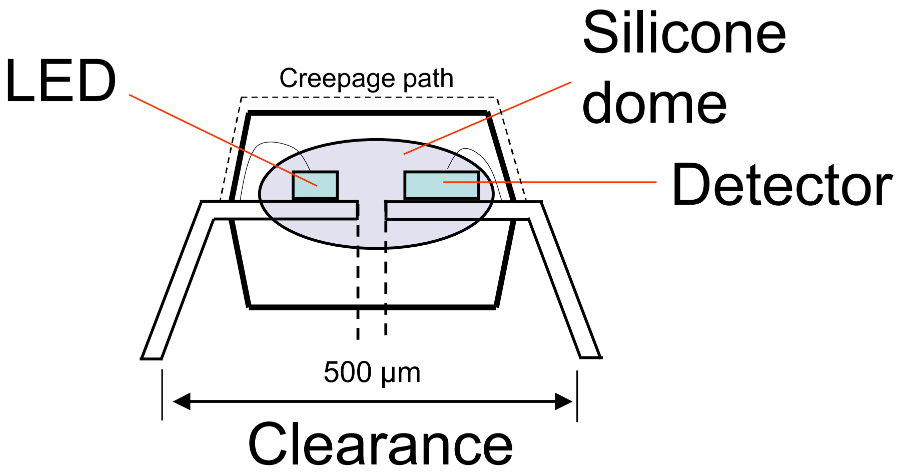
  - ❖ Optocoupler
  - ❖ Digital CMOS isolator
  - ❖ Transformer



Emitting and Receiving Light

- ❑ You need galvanic isolation between the prim. and the sec.
- ❑ An optocoupler transmits light only, no electrical link



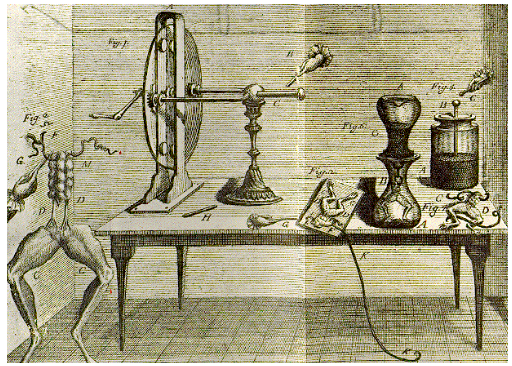


$$\mathrm{CTR} = \frac{I_c}{I_F} \times 100$$

Current Transfer Ratio

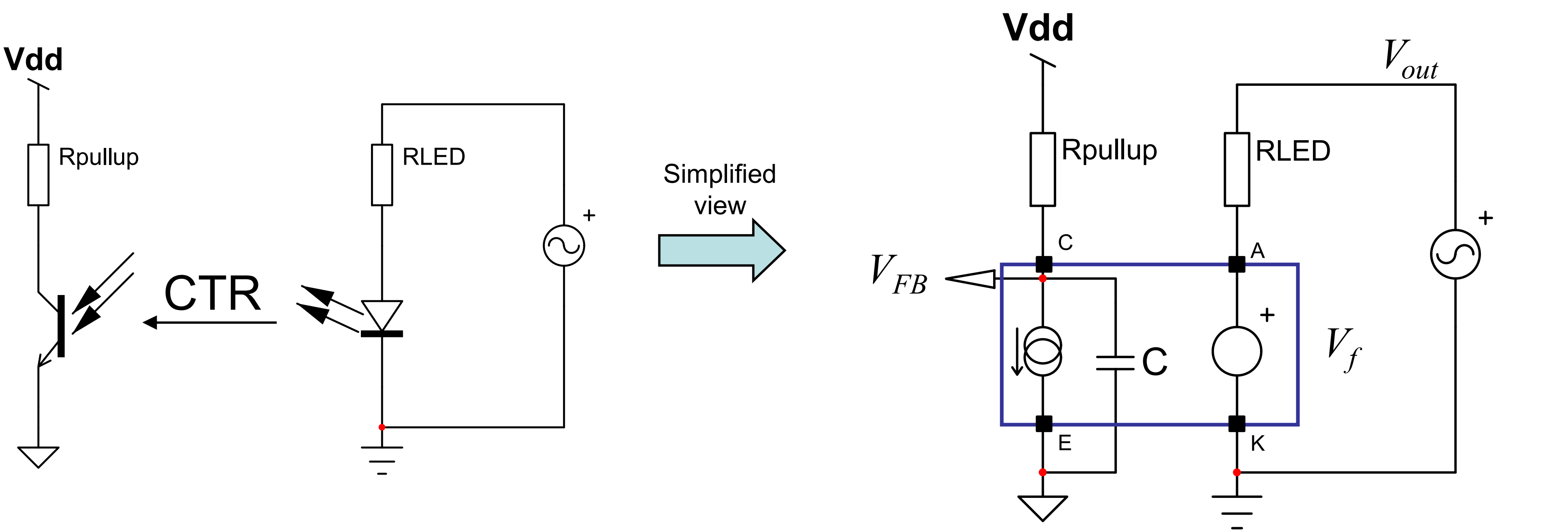
Luigi Galvani, 1737-1798
Italian physician and physicist

French specimen →



The Optocoupler Corrupts the Return Path

- Photons are collected by a collector-base area
- This area brings a large CE parasitic capacitance



$$\frac{V_{FB}(s)}{V_{out}(s)} = -\frac{R_{pullup}\mathrm{CTR}}{R_{LED}} \frac{1}{1+sR_{pullup}C}$$

If f_p is well above f_c, its effect is negligible

If f_p is close to f_c, phase margin degradation

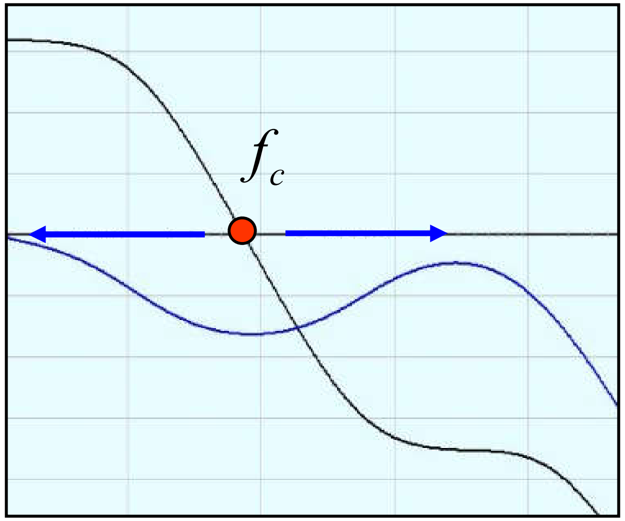
Assess the CTR variations

- ❑ CTR changes with the operating current
- ❑ Try to select collector bias currents around 2-5 mA

Current Transfer Ratio (I_C/I_F at V_{CE}=5.0 V) **and Collector-emitter Leakage Current** SFH-615

Parameter	-1	-2	-3	-4	-12	-23	-34	-13	-24	-14	Unit
I_C/I_F (I_F=10 mA)	40–80	63–125	100–200	160–320	40–125	63–200	100–320	40–200	63–320	40–320	%
I_C/I_F (I_F=1.0 mA)	30(>13)	45(>22)	70(>34)	90(>56)	30(>13)	45(>22)	70(>34)	30(>13)	45(>22)	30(>13)	
Collector-Emitter Leakage Current, I_{CEO}, V_{CE}=10 V	2.0(≤50)	2.0(≤50)	5.0(≤100)	5.0(≤100)	2.0(≤50)	5.0(≤100)	5.0(≤100)	5.0(≤100)	5.0(≤100)	5.0(≤100)	nA

CTR between 0.63 and 1.25
Normalized to 1 (0 dB)
0.63 gives –4 dB
1.25 gives +1.9 dB



Watch out for crossover frequency changes and phase margin at CTR extremes!

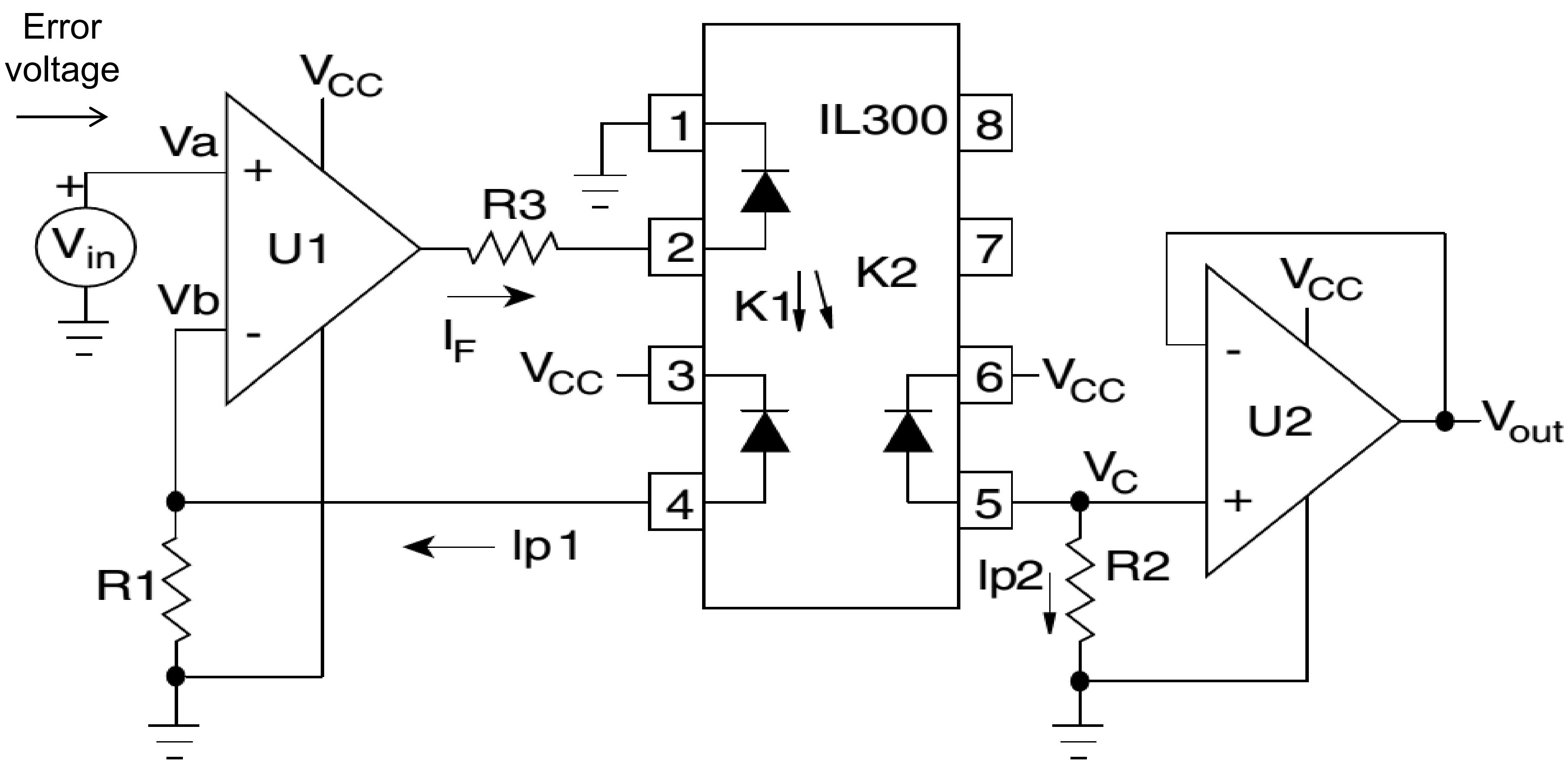
Selecting the Right Optocoupler

- ❑ High temperature shortens the optocoupler lifetime
- ❑ Low LED currents:
  - ▪ expand lifetime
  - o CTR dispersion increases
- ❑ Low CTR optocouplers have low internal capacitance
  - ▪ Select them for high-bandwidth applications
- ❑ Higher drive currents improve speed but:
  - o Shorten lifetime
  - o Degrade standby power

⇨ Understand the optocoupler impact on your design!

LED Photons Emission Suffers with Age

- ❑ For long-life applications, compensate for LED degradation

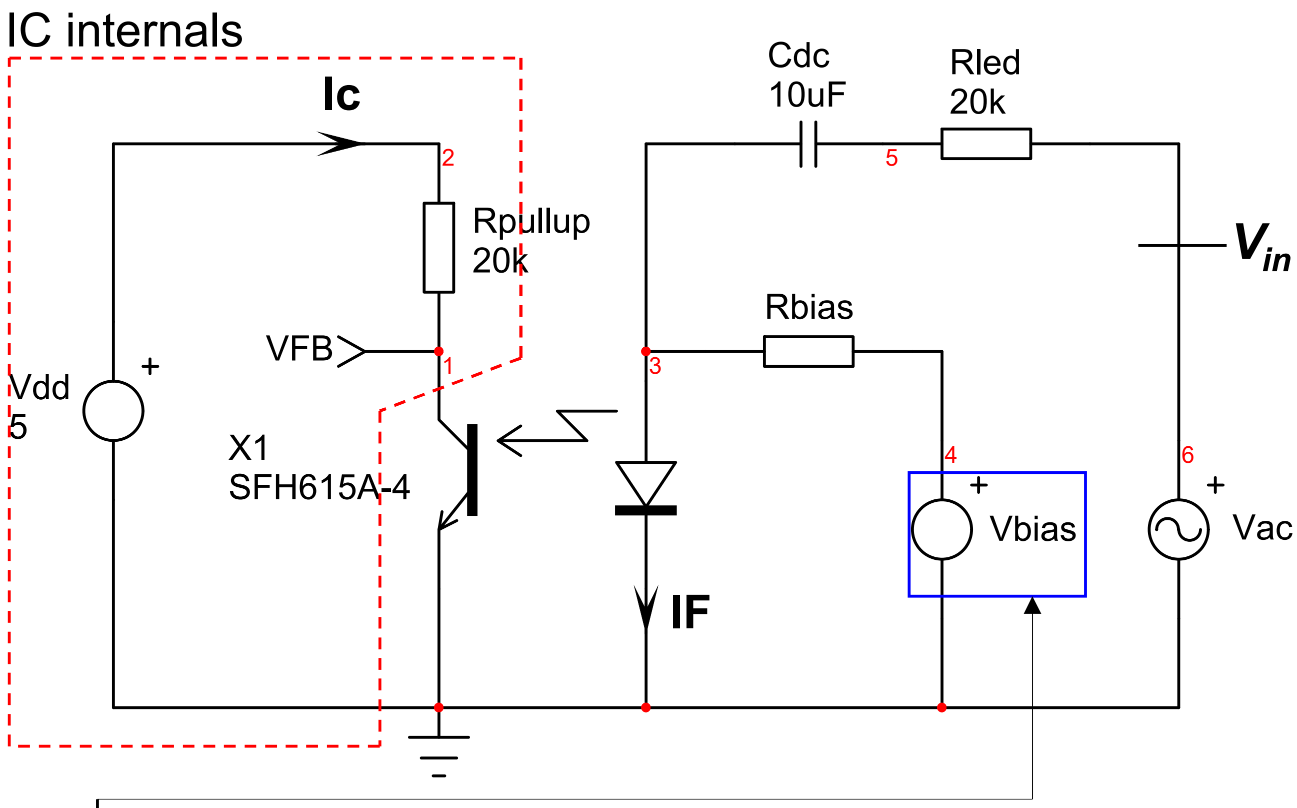


- ❑ IL300 senses the LED light and adjusts driving current

http://www.vishay.com/docs/83622/il300.pdf

Where is the Pole?

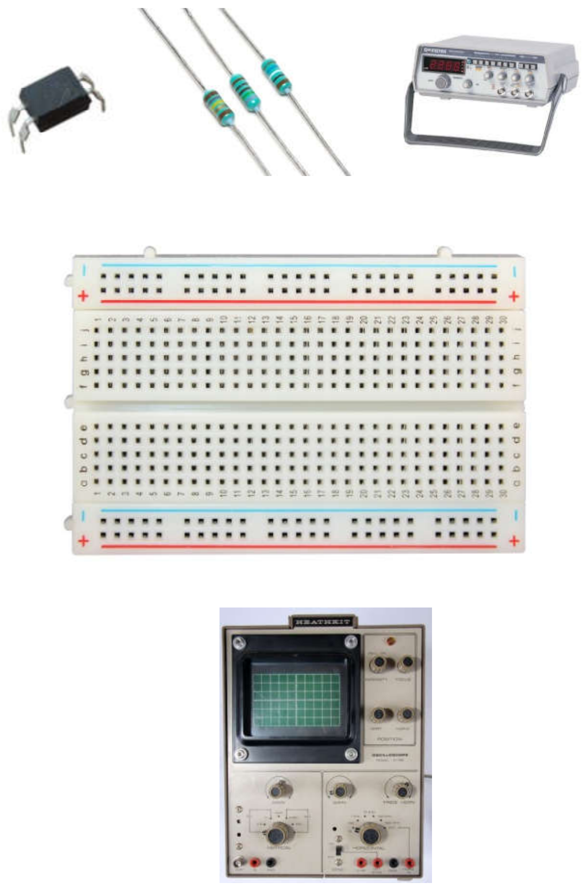
- ❑ Once your optocoupler is selected, characterize it

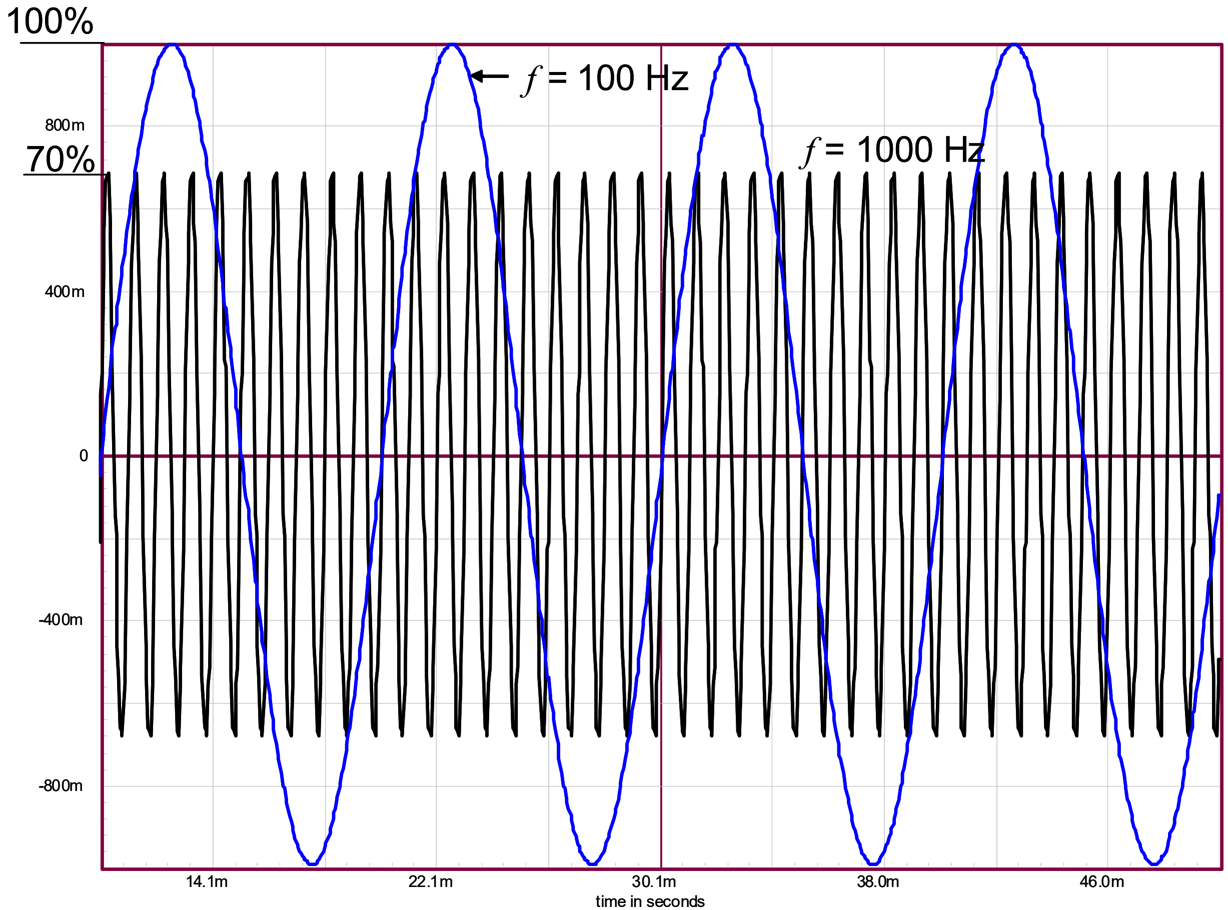


- ❑ Adjust $\boxed{V_{bias}}$ to V_{FB} = 2.5 V (or $V_{dd}/2$) then ac sweep
- ❑ Observe V_{FB} with a scope or a network analyzer

No Need for a Network Analyzer

- ❑ Pole extraction can also be done with an oscilloscope

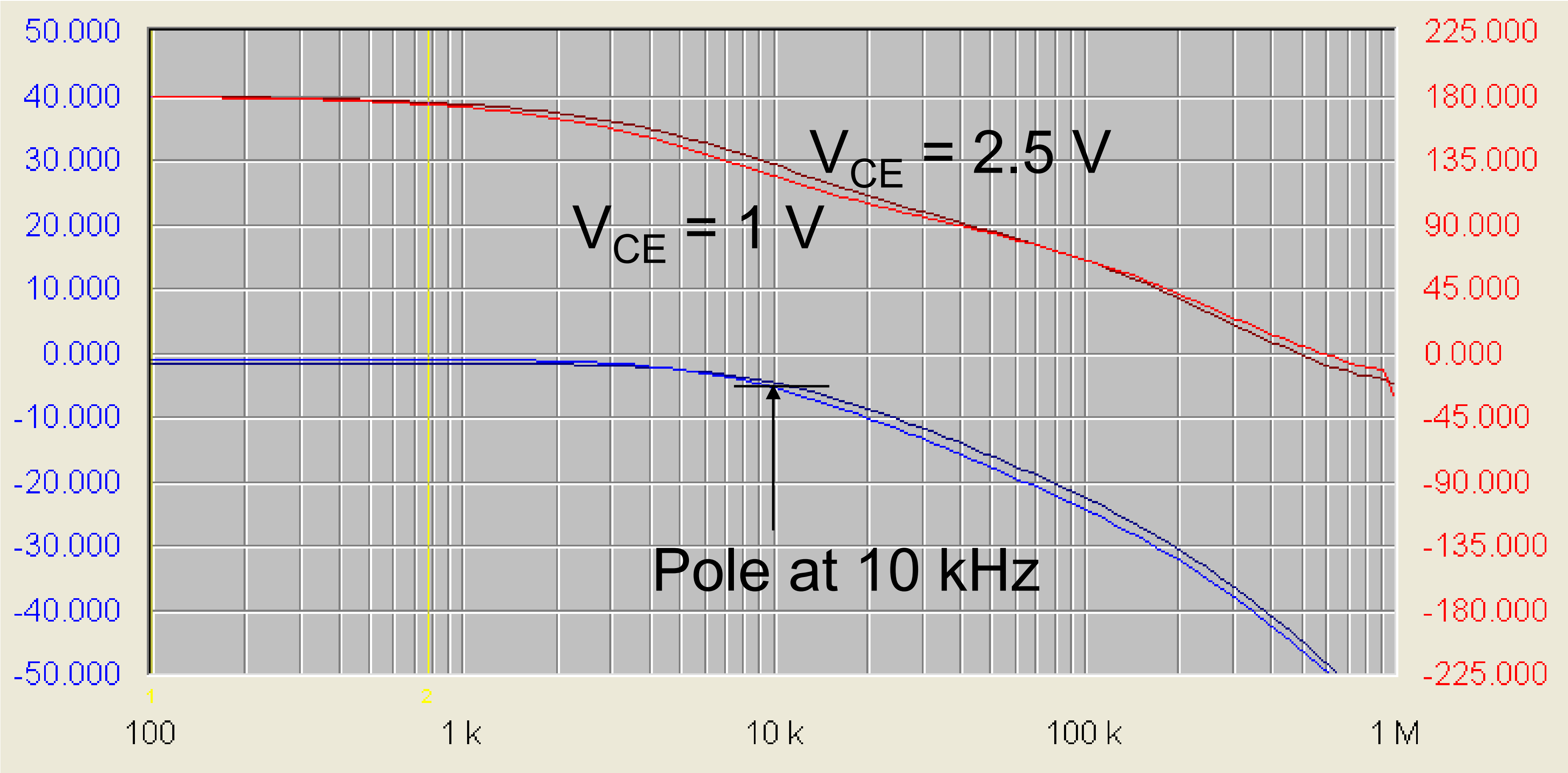




- ❑ 10 division is the low-frequency reference, 7 divisions is f_p

Adapt the Bias Point to Your Circuit

- The dc bias does not change the pole position

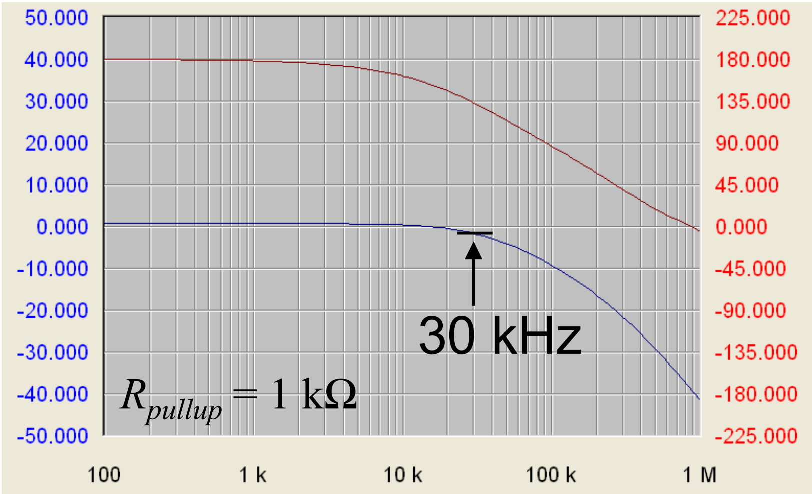


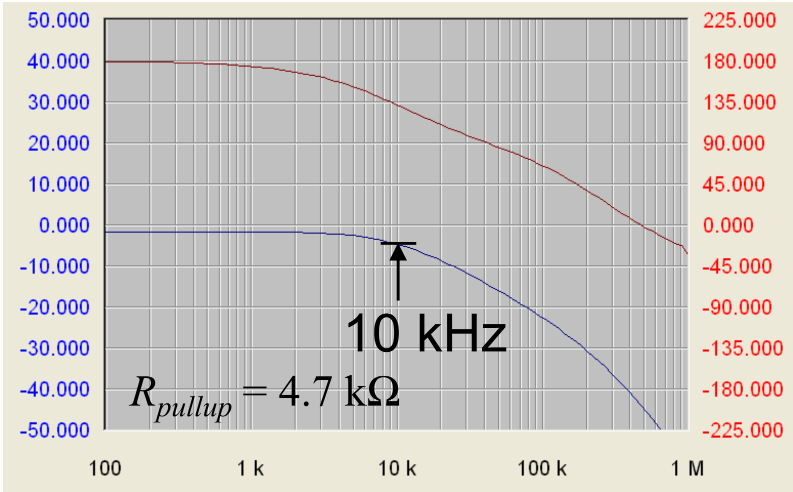
R_{pullup} = 4.7 kΩ

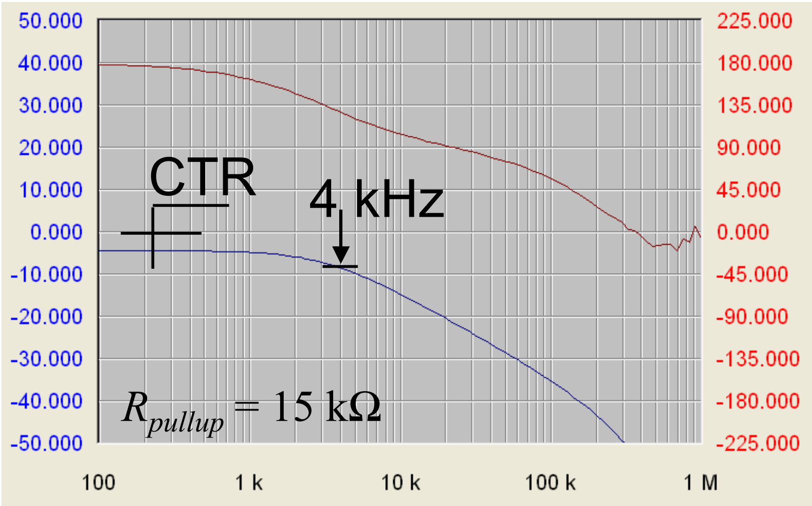
SFH615A-2

The Pull-Up Resistor Affects the Pole

- ❑ A low value pull-up resistor offers better bandwidth!







- ❑ Changing the bias point affects the CTR

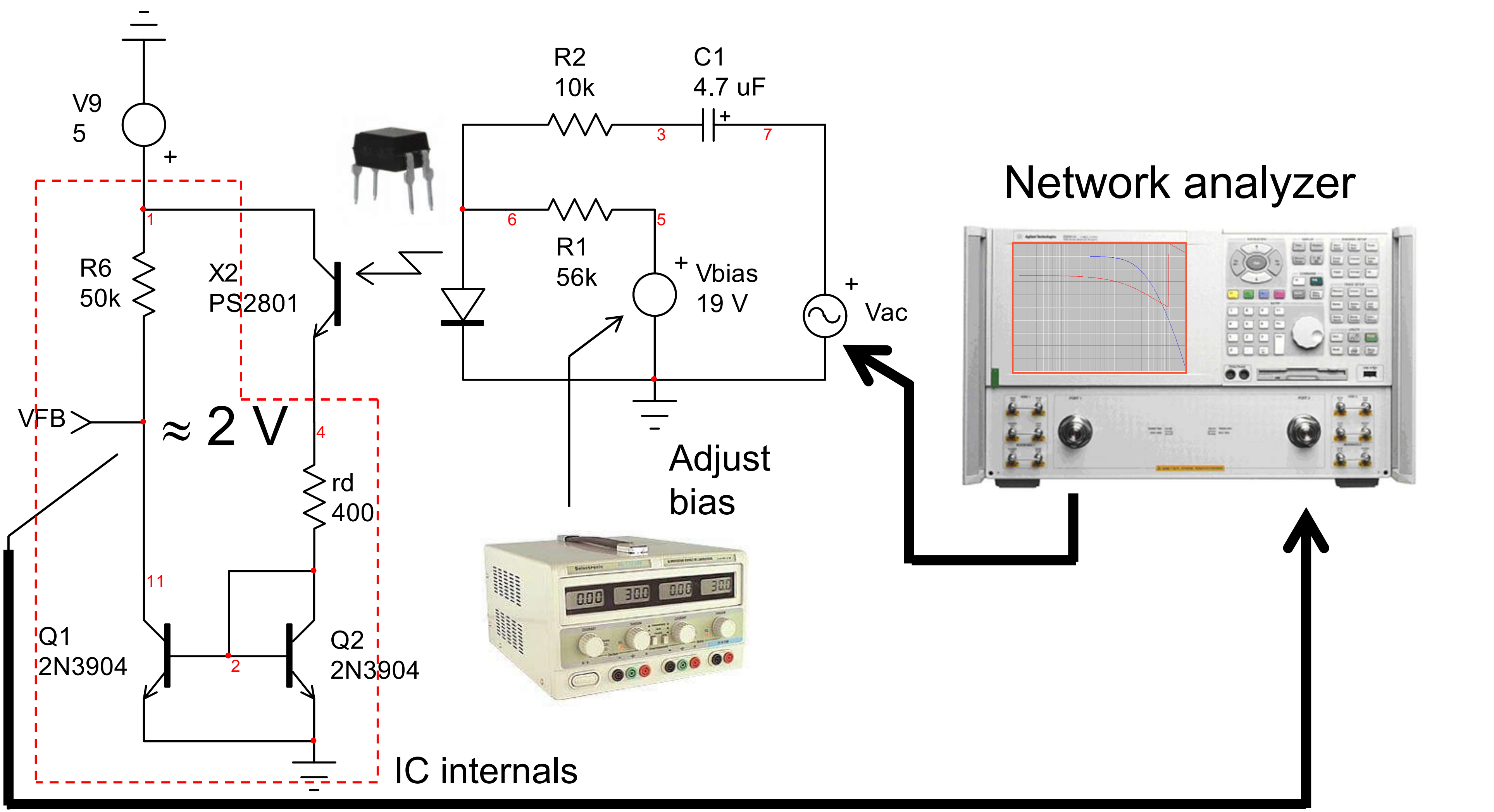
$$\frac{V_{FB}(s)}{V_{out}(s)} = -\frac{R_{pullup}}{R_{LED}}\mathrm{CTR}$$

- ❑ Low resistor values degrade standby

SFH615A-2

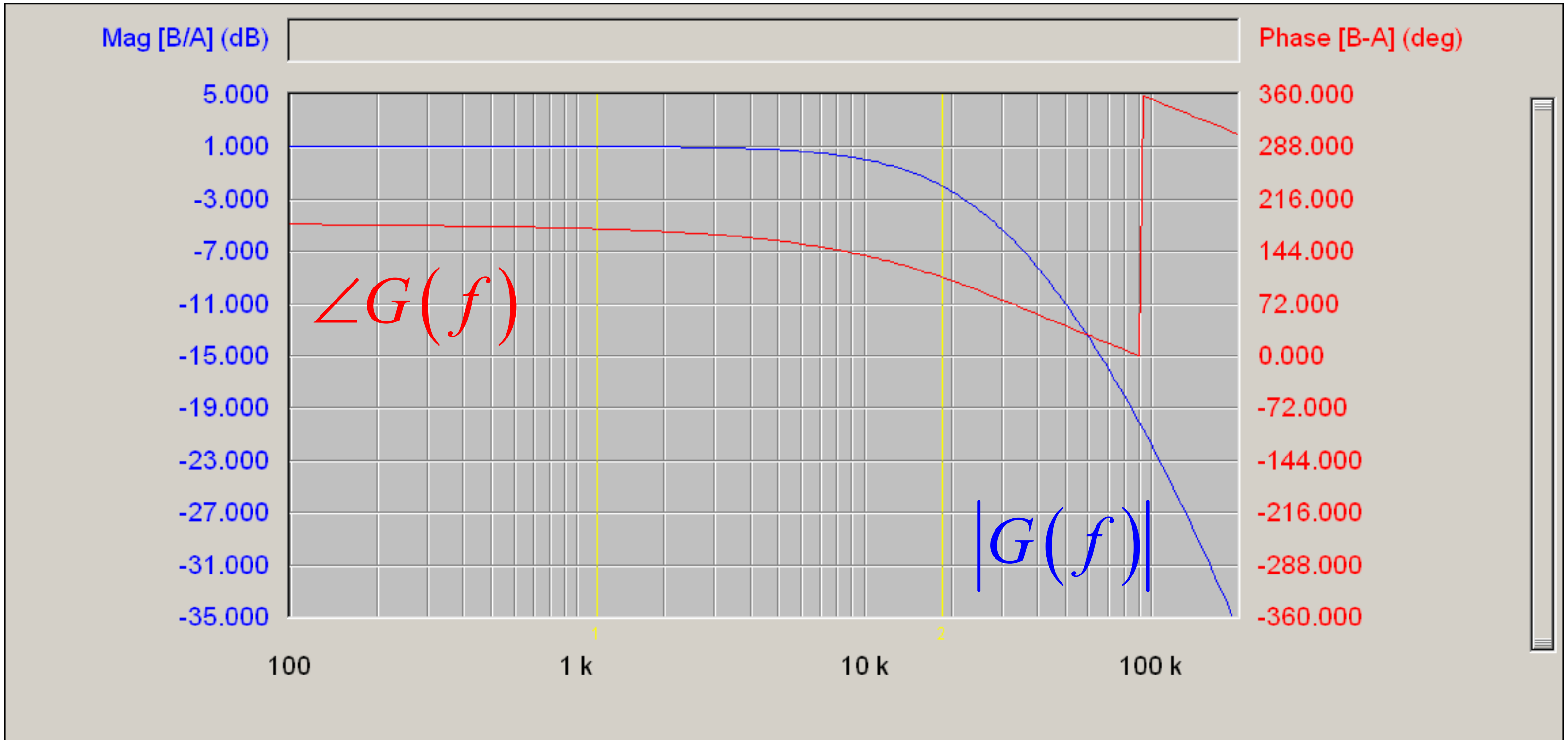
What is the Application Circuit?

❑ You need to recreate the optocoupler biasing conditions



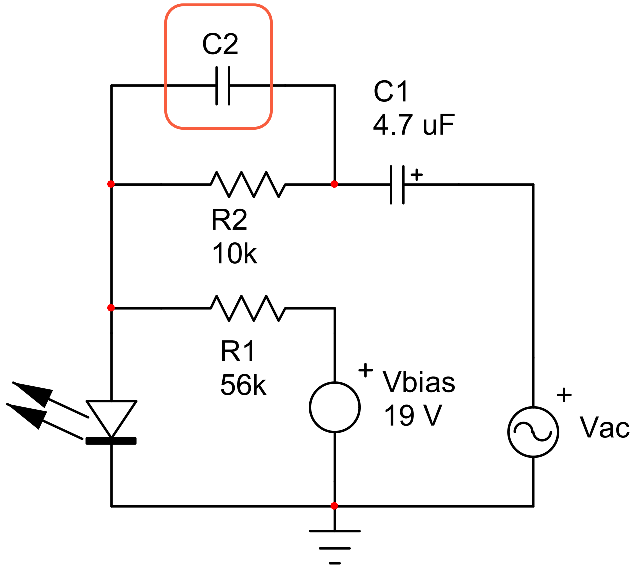
Pole Will Stress the Phase at Crossover

❑ The pole is located at 17.5 kHz, R_d is 400 Ω



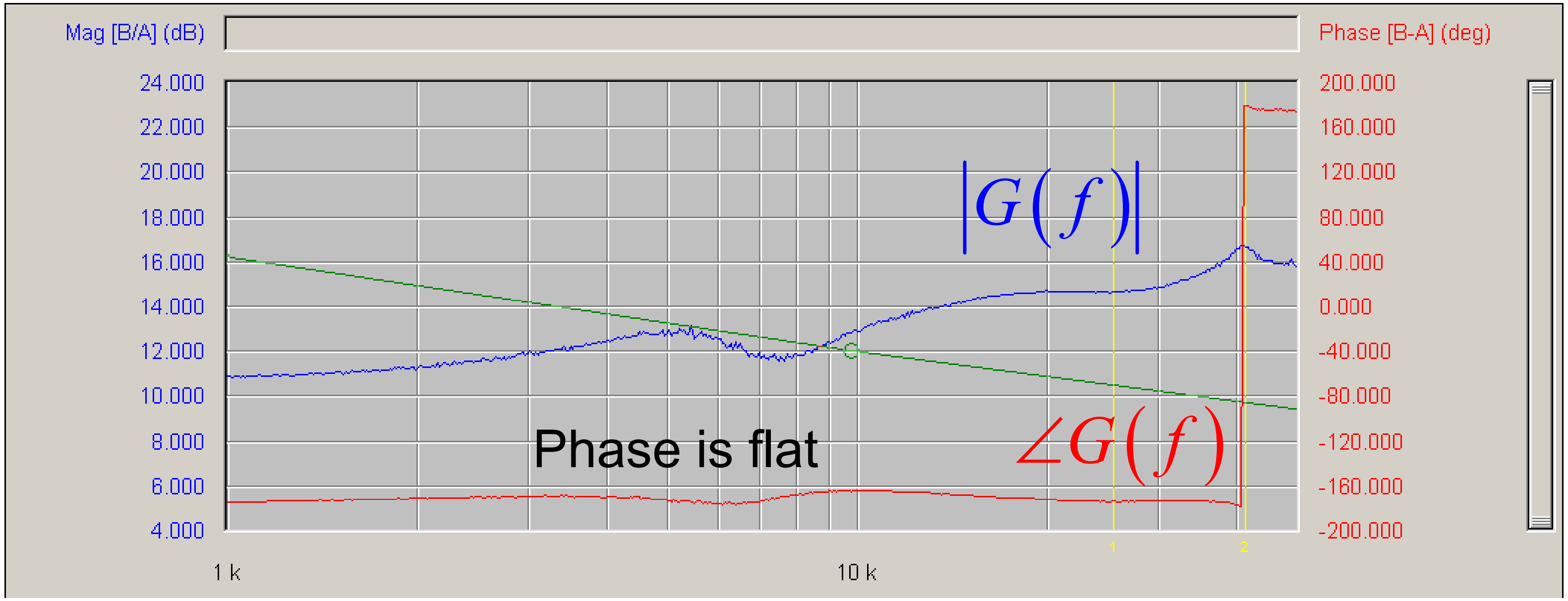
❖ Phase lag can potentially bother compensation

Compensating the Optocoupler Poke



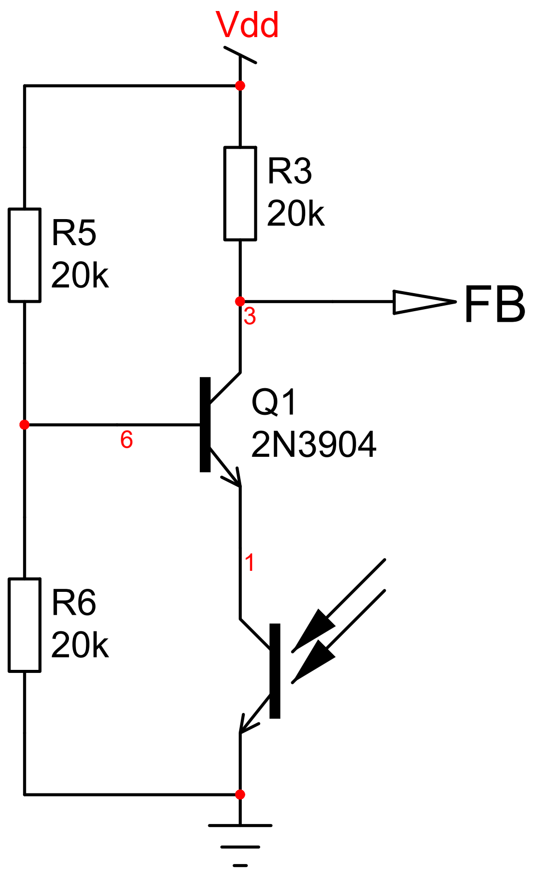
- A zero compensates the pole

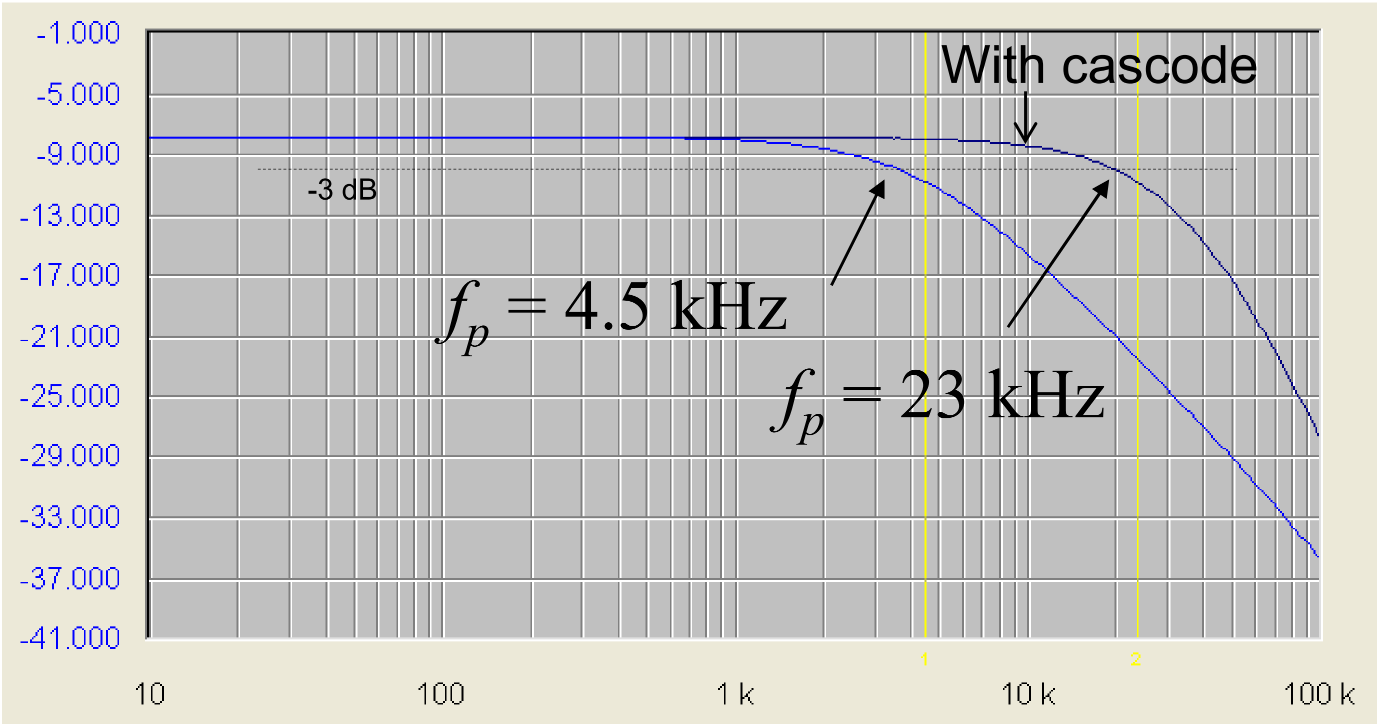
$$C_2 = \frac{1}{2\pi f_{pole} R_2} \approx 870 \text{ pF}$$



Push the Pole Farther with a Cascode

- The cascode fixes the optocoupler collector potential
- It neutralizes the Miller capacitance in the optocoupler

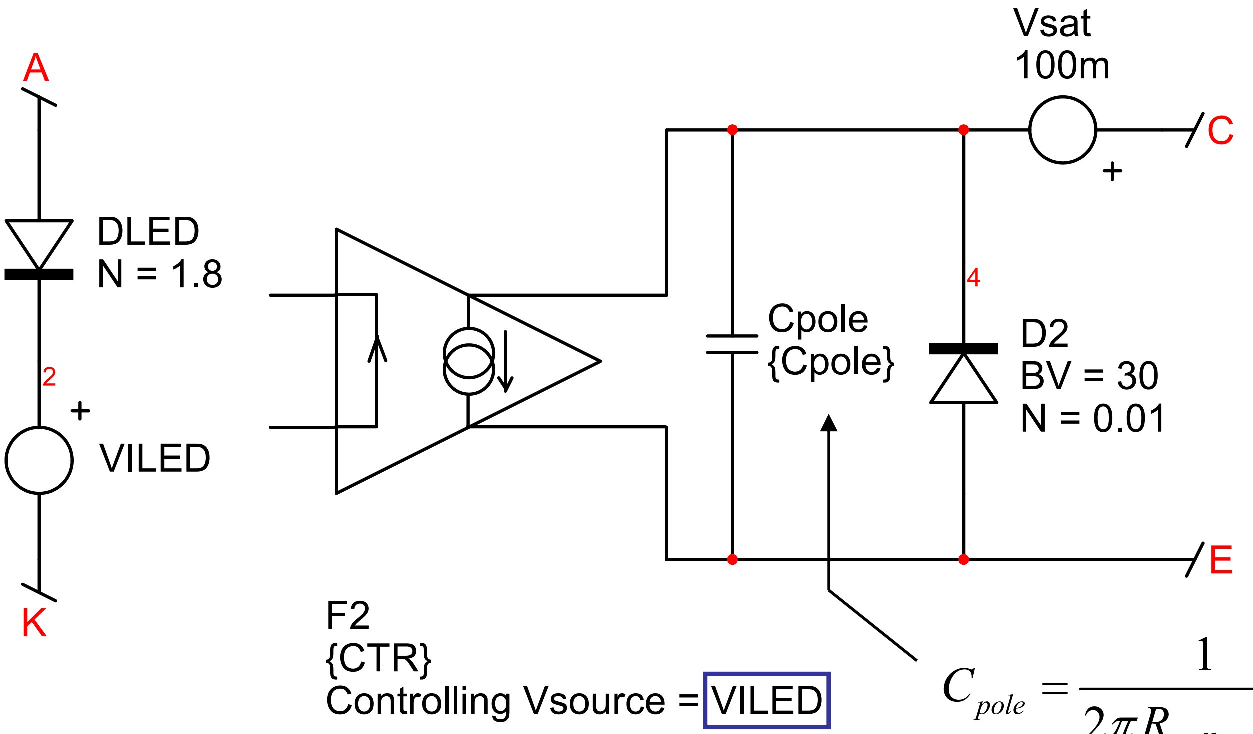




SFH615A-2

A Simple Optocoupler Model

- ❑ Read the pole position and the CTR value
- ❑ Adjust the internal capacitor value to fix the pole



$$C_{pole} = \frac{1}{2\pi R_{pullup} f_{opto}}$$

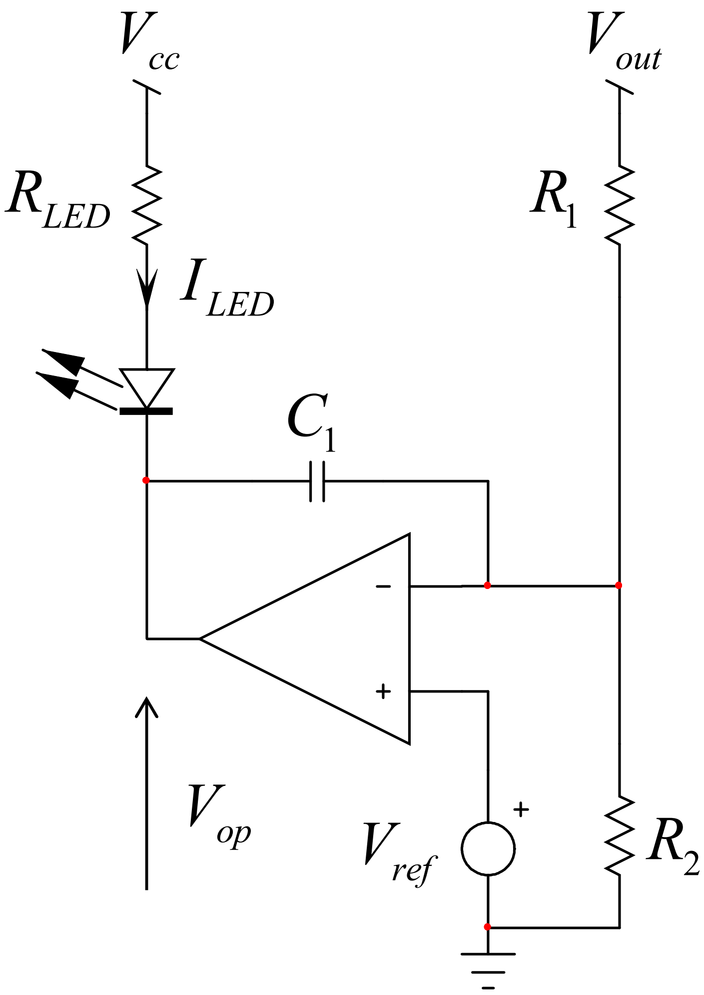
- ❑ Ok for low-frequency exploration, no 2nd-order effects

Course Agenda

- ❑ Control System Basics
- ❑ Operational Amplifier and Low-Frequency Pole
- ❑ Gain-Bandwidth Impact on Phase Boost
- ❑ Op Amp Slew Rate Effects in Loop Control
- ❑ Start-Up Sequence and Auxiliary Supply
- ❑ Characterizing the Optocoupler Pole
- ❑ **Dealing with the Fast Lane**
- ❑ Going Around the TL431 Fast Lane

A Single Current Path

- There is a single ac path which defines the LED current: V_{out}



$$I_{LED}(s) = \frac{\cancel{V_{cc}} - V_{op}(s)}{R_{LED}} = -\frac{V_{op}(s)}{R_{LED}}$$

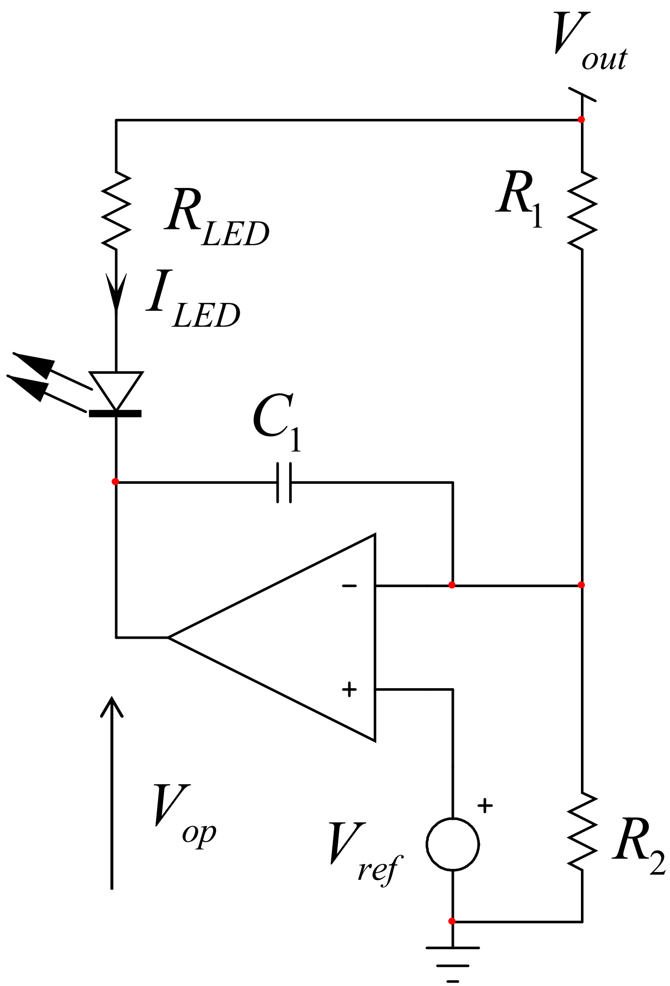
$$V_{op}(s) = -\frac{\frac{1}{sC_1}}{R_1} = -\frac{1}{sR_1C_1} = -\frac{1}{\frac{s}{\omega_{po}}}$$

$$\Rightarrow \quad I_{LED}(s) = \frac{1}{R_{LED}} \frac{1}{\frac{s}{\omega_{po}}}$$

Integrator, single pole

A Second Path Creates the Fast Lane

- The optocoupler current now depends on V_{out} via two paths



$$I_{LED}(s) = \frac{V_{out}(s) - V_{op}(s)}{R_{LED}}$$

$$V_{op}(s) = -\frac{\frac{1}{sC_1}}{R_1} = -\frac{1}{sR_1C_1}$$

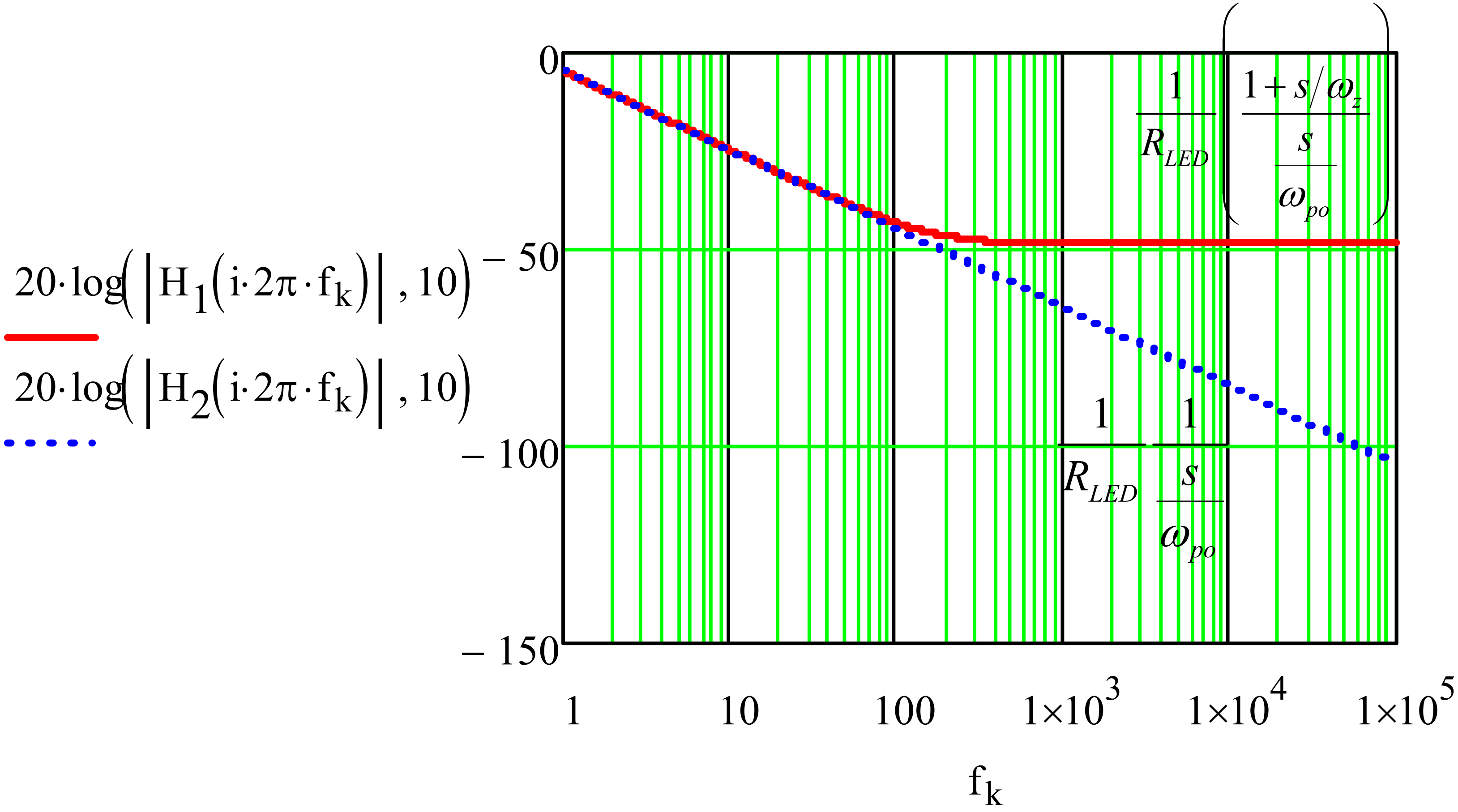
$$I_{LED}(s) = \frac{1}{R_{LED}}\left(1 + \frac{1}{sR_1C_1}\right)$$

$$= \frac{1}{R_{LED}}\left(\frac{1+sR_1C_1}{sR_1C_1}\right) = \frac{1}{R_{LED}}\left(\frac{1+\boxed{s/\omega_z}}{\frac{s}{\omega_{po}}}\right)$$

Integrator plus a zero

The Fast Lane Prevents Gain Roll-off

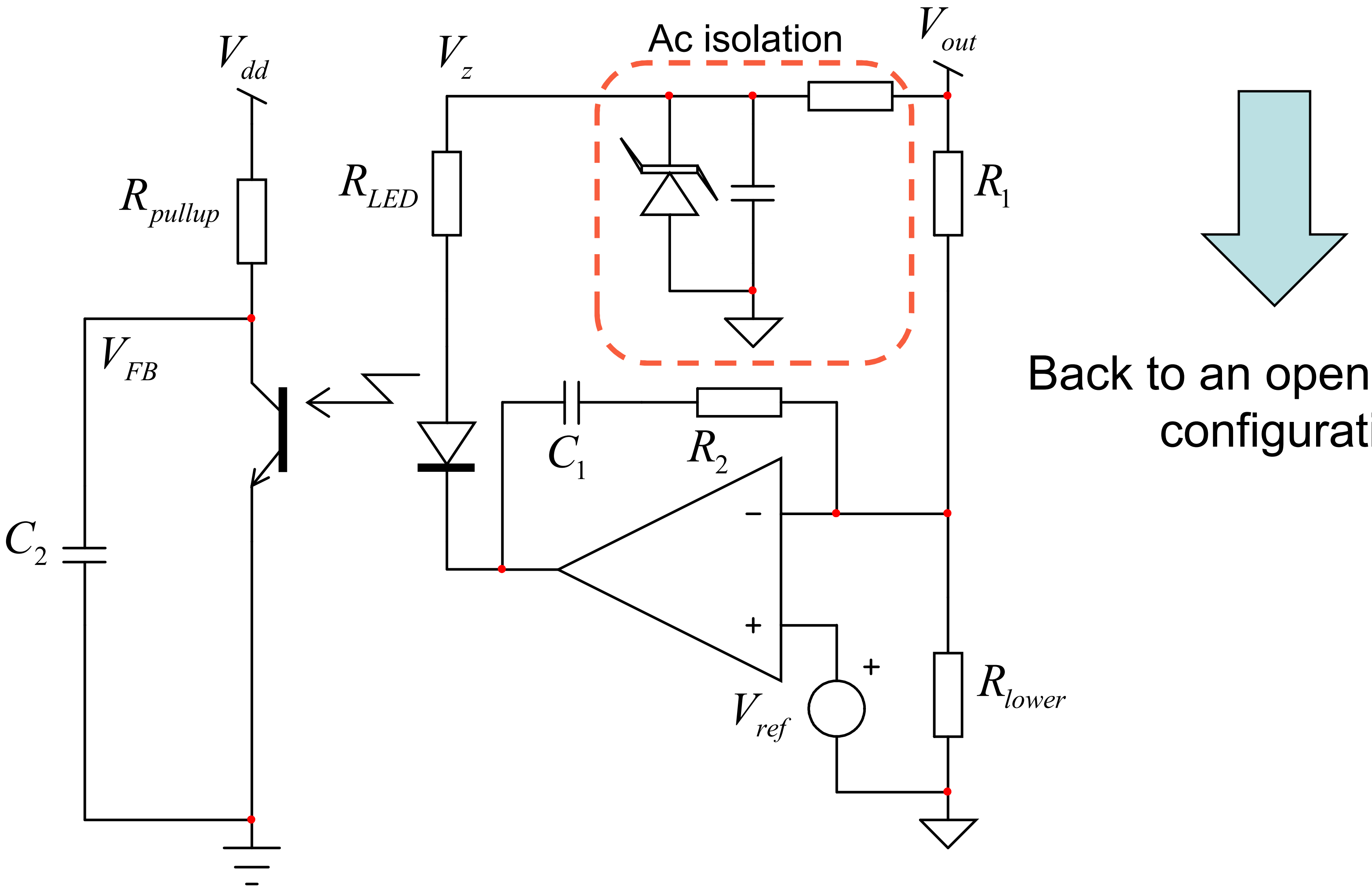
- With the fast lane presence, you cannot roll off the gain



- The zero imposes a fixed gain as s approaches infinity

Disconnect the Fast Lane!

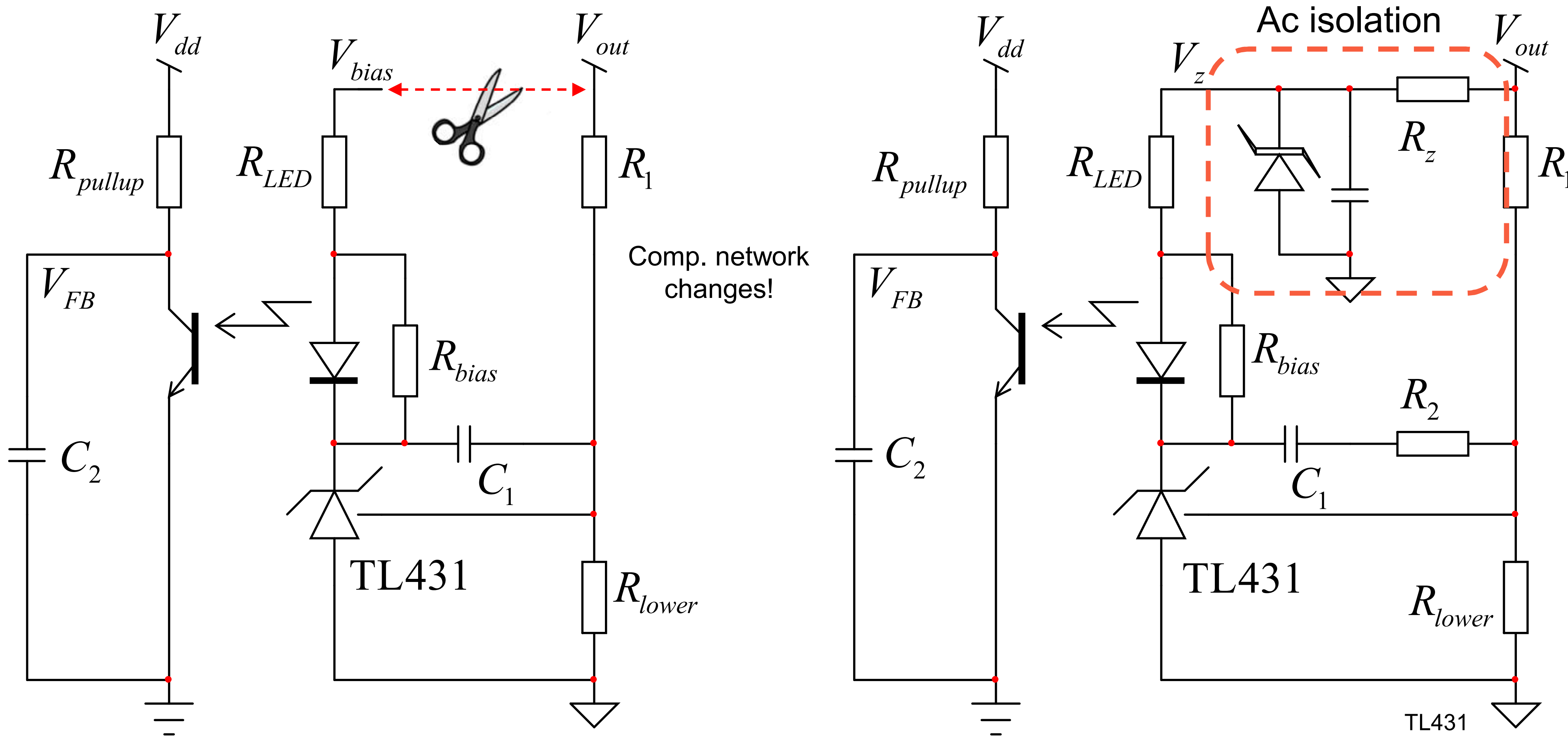
- One solution is to get rid of the fast lane via ac isolation



Back to an open-collector configuration

Same Issue with a TL431

- ❑ The TL431 exhibits exactly the same fast lane problem



- ❑ The solution is to also ac-isolate the fast lane from V_{out}

ON Semiconductor®

What is the Impact on Compensator?

- ❑ A poor ac-isolation will corrupt the compensator response

parameters

Rpullup=1k
Fopto=15k
Copto=1/(2*pi*Rpullup*Fopto)
CTR=0.8
G=10^(-Gfc/20)

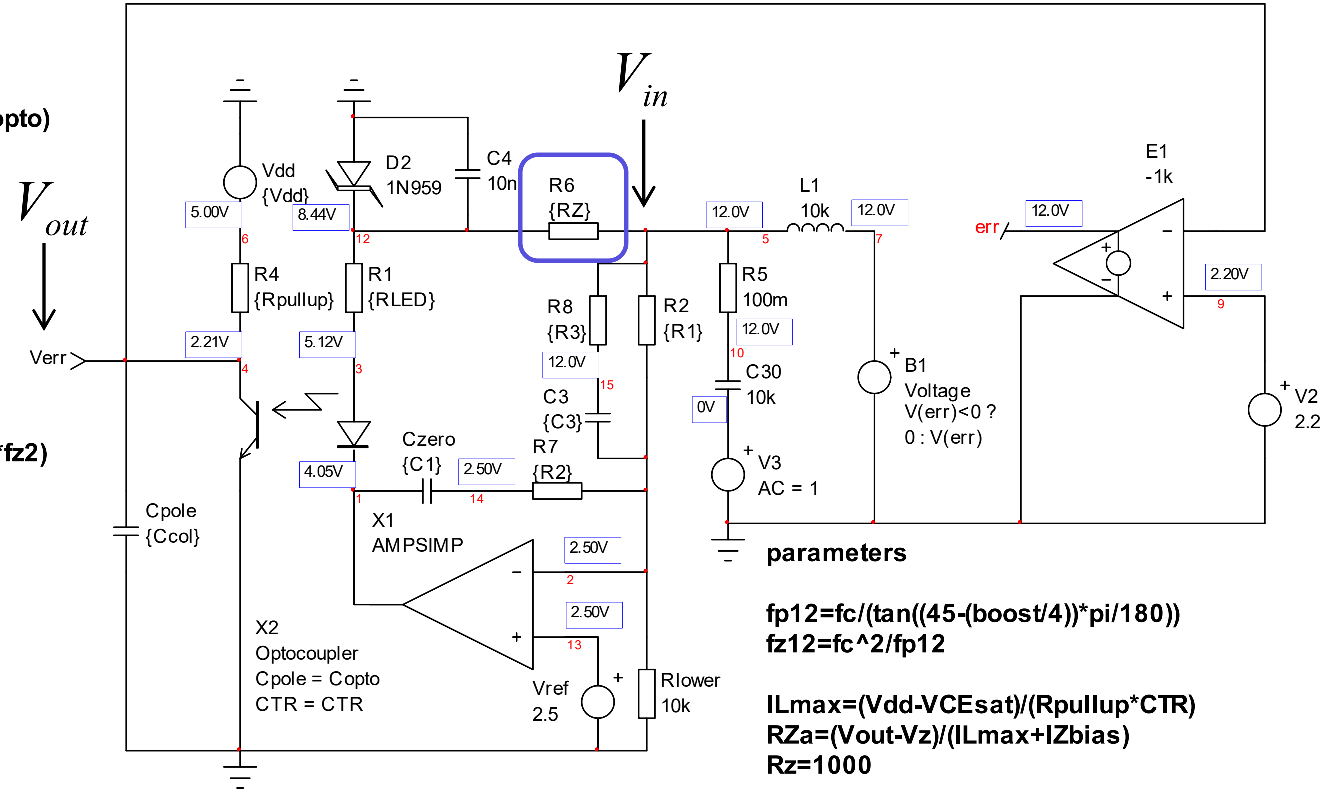
RLED=Rmax*0.8
RLED1=910

G1=Rpullup*CTR/RLED
G2=G/G1

C1=1/(2*pi*fz1*R2)
C3=(fp2-fz2)/(2*pi*R1*fp2*fz2)
R3=R1*fz2/(fp2-fz2)

a=sqrt((fc^2/fp1^2)+1)
b=sqrt((fc^2/fp2^2)+1)
c=sqrt((fz1^2/fc^2)+1)
d=sqrt((fc^2/fz2^2)+1)

R2=(a*b/(c*d))*R1*G2
C2=1/(2*pi*fp1*Rpullup)
Ccol=C2-Copto



parameters

fp12=fc/(tan((45-(boost/4))*pi/180))
fz12=fc^2/fp12

ILmax=(Vdd-VCEsat)/(Rpullup*CTR)
RZa=(Vout-Vz)/(ILmax+IZbias)
Rz=1000

parameters

Vout=12V
R1=38k
fc=1.8k
Gfc=10
boost=150
G=10^(-Gfc/20)
pi=3.14159

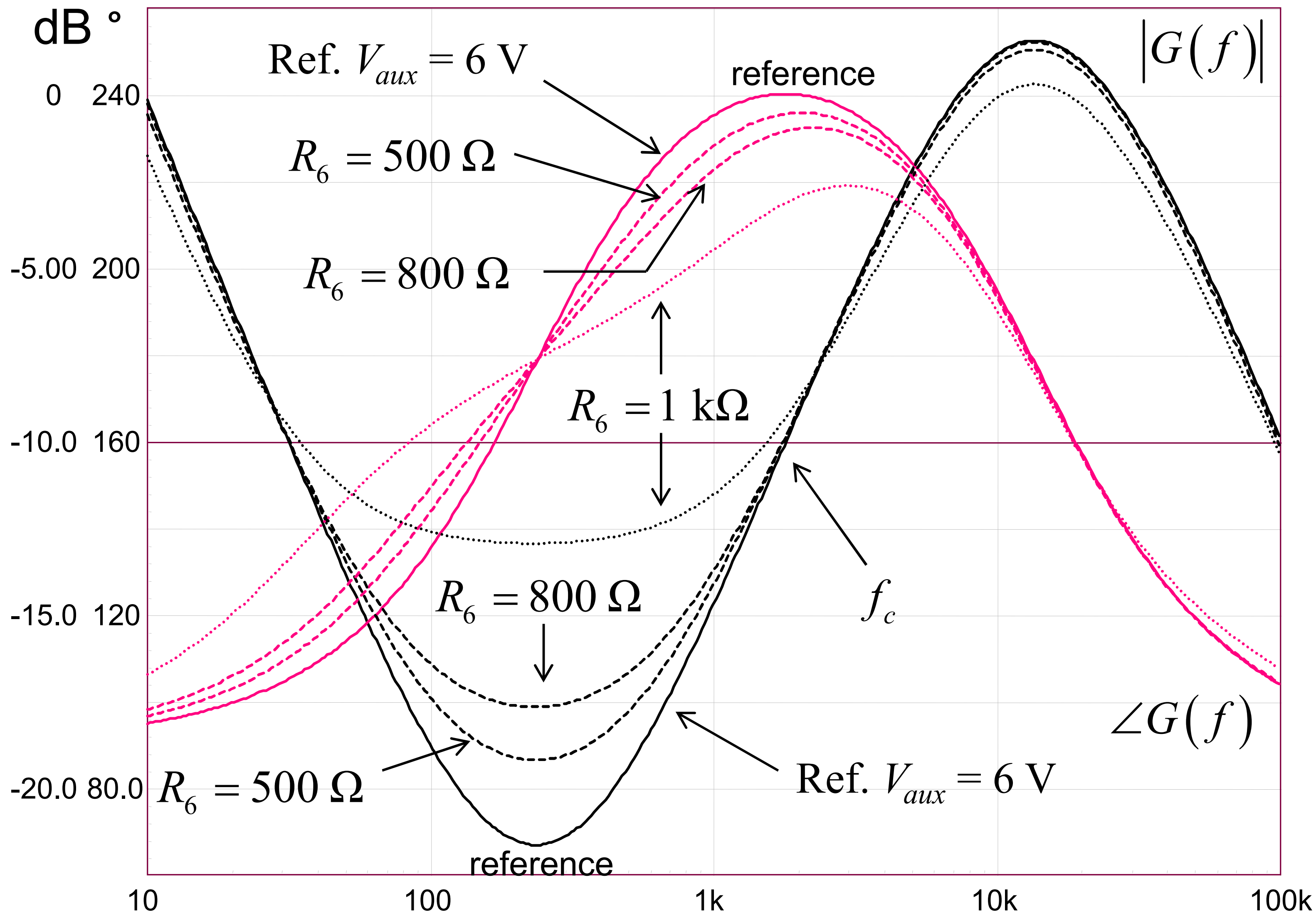
fz1=fz12
fz2=fz12
fp1=fp12
fp2=fp12

VOL=0.2
VCEsat=0.3
Vdd=5
Vz=8.2
Vf=1
IZbias=1m
AA=Vz-Vf-VOL
BB=Vdd-VCEsat
Rmax=(AA/BB)*Rpullup*CTR

- ❑ The Zener diode bias point is changed by adjusting R_6

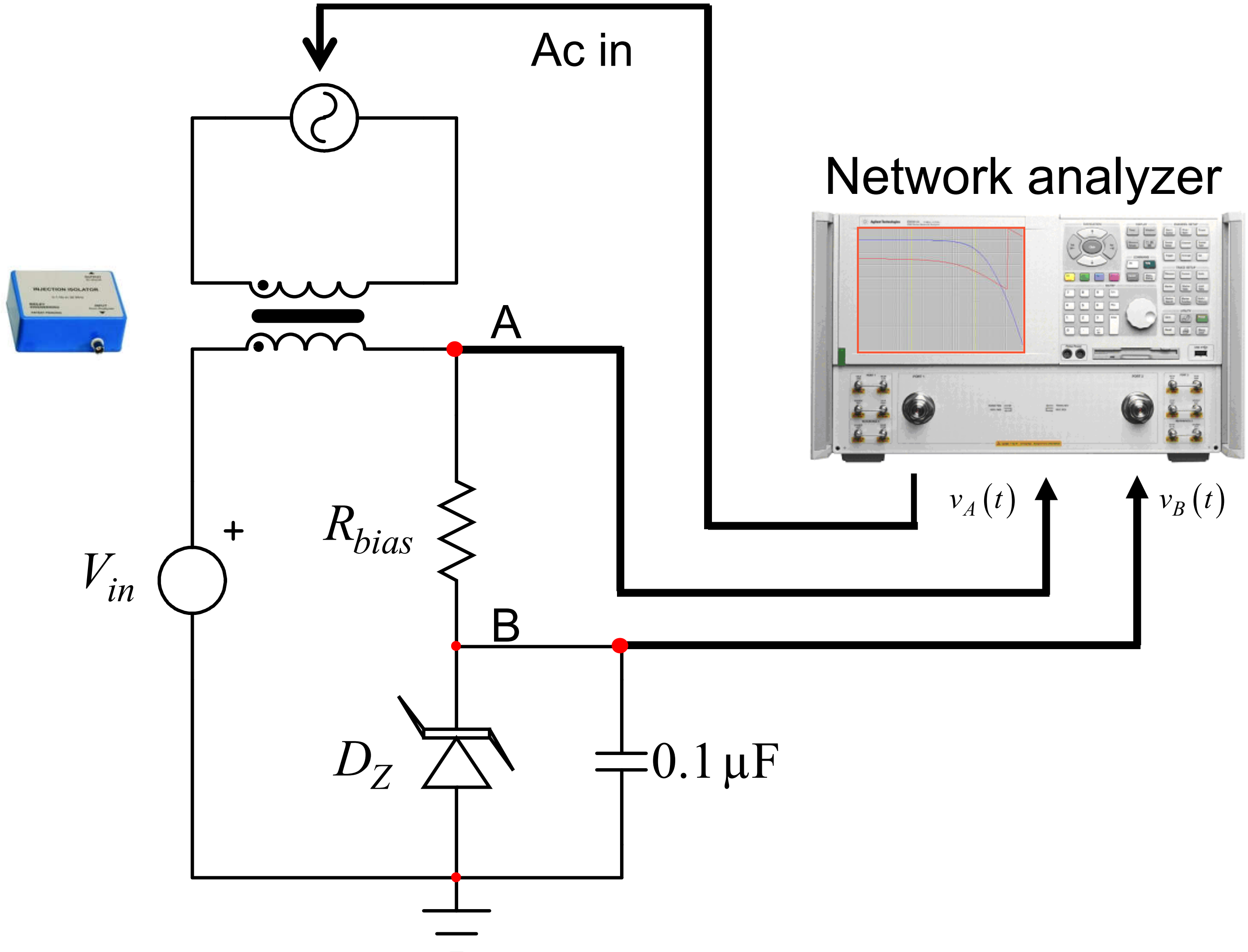
Poor Isolation Corrupts Phase Boost

- As dynamic resistance changes, ac response degrades



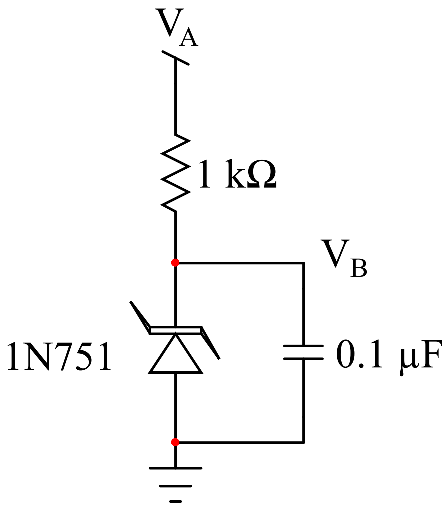
Practical Experimentation Around the Zener

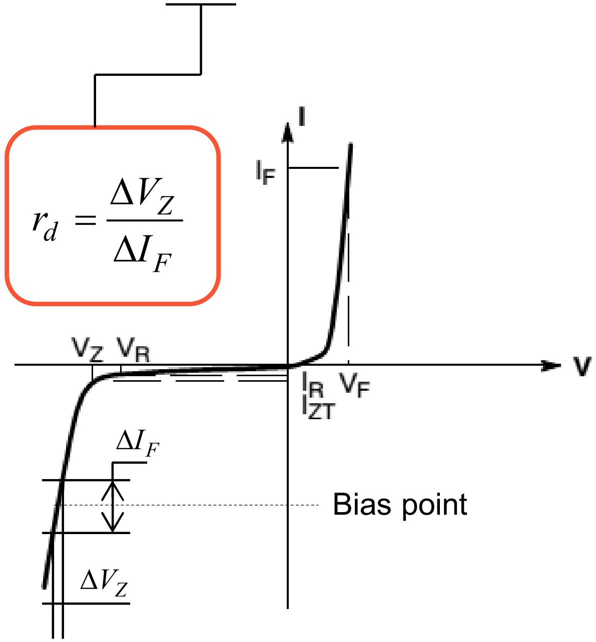
- ❑ How much isolation does a biased Zener diode provide?

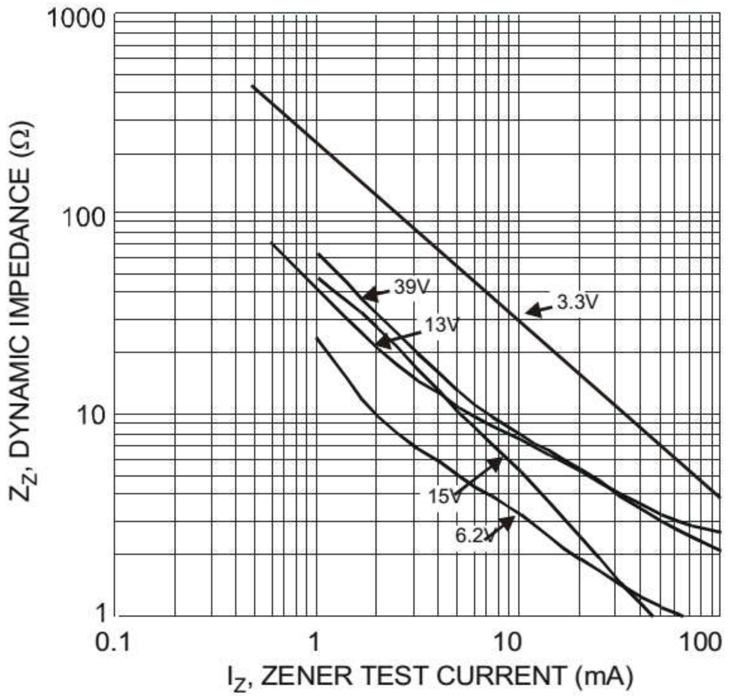


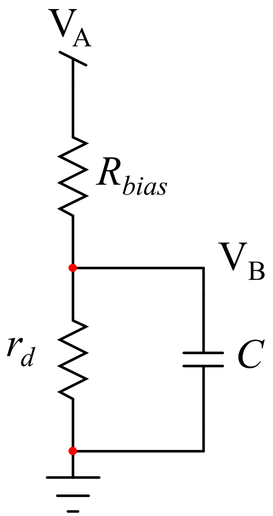
Check Zener Isolation Capability

- ❑ Dynamic resistance r_d changes as bias conditions move





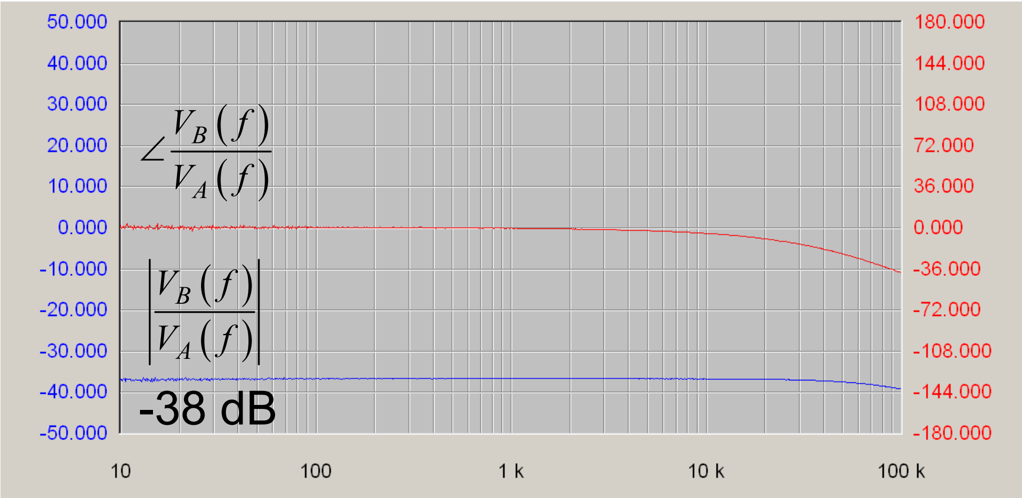


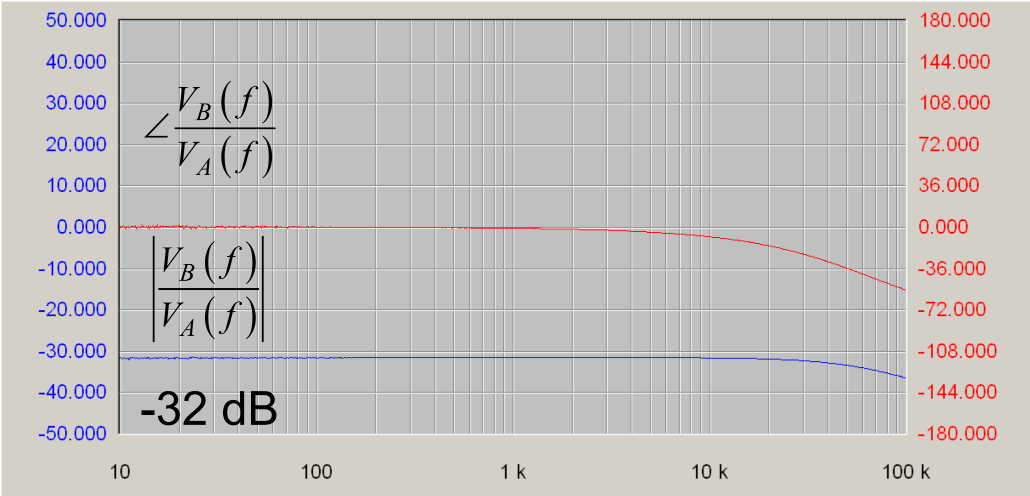


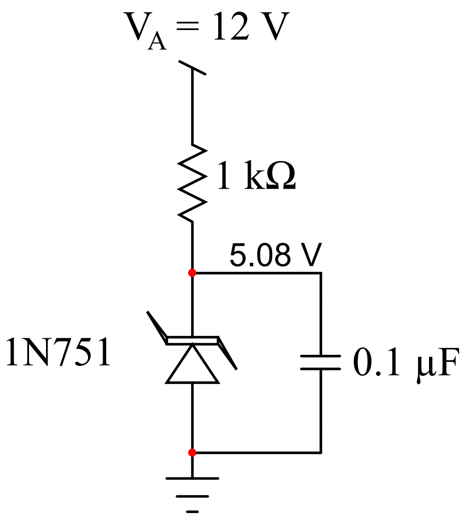
$$H(s) = \frac{V_B(s)}{V_A(s)} = \frac{r_d}{r_d + R_{bias}} \frac{1}{1 + s\left(r_d \parallel R_{bias}\right)C}$$

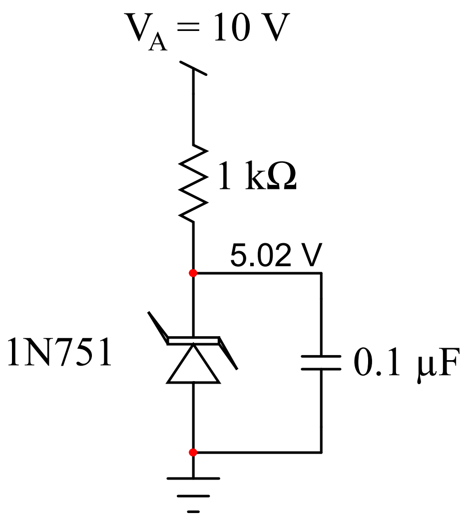
The Dc Bias Affects the Resistor Drop

- Attenuation degrades as dc bias reduces



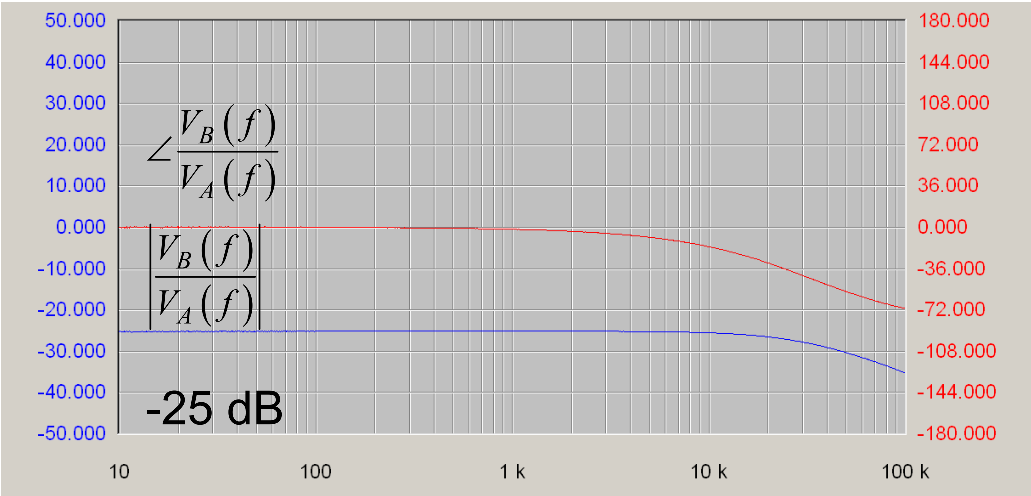


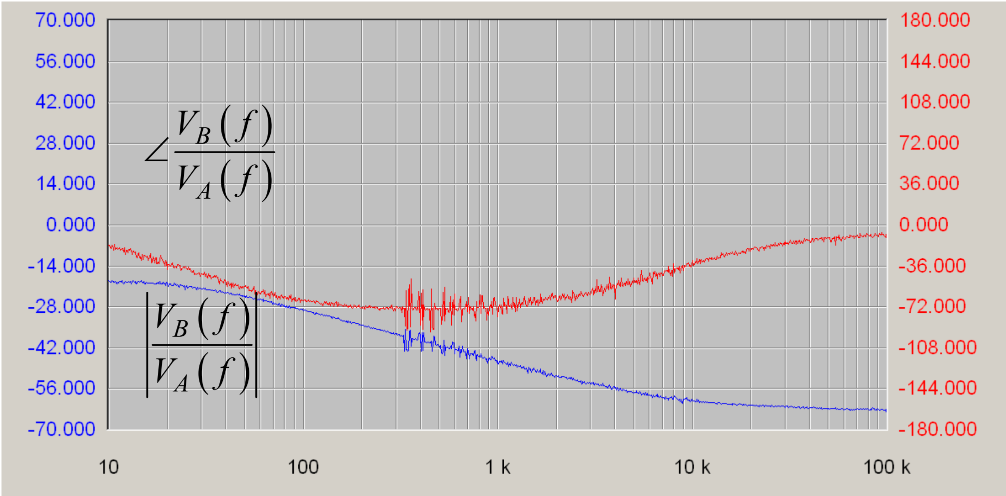


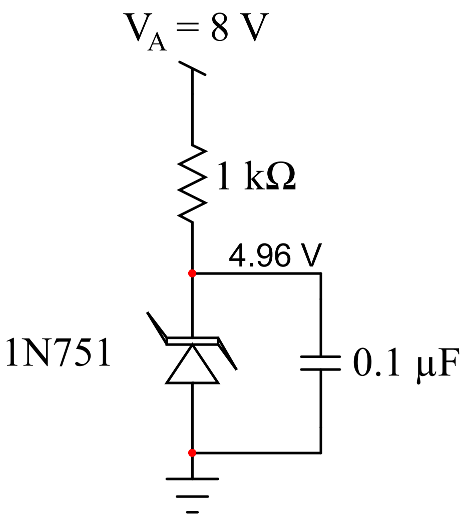


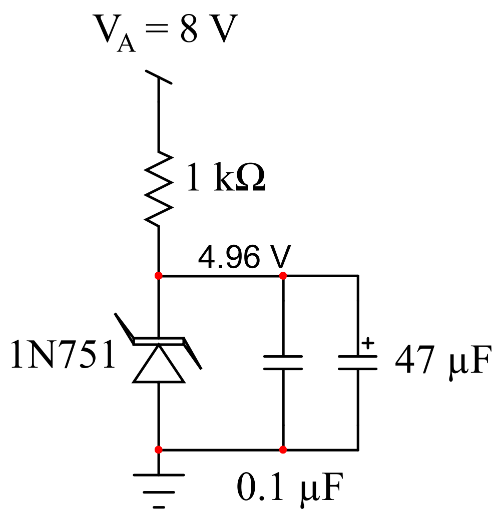
Decoupling Helps at High Frequency

- A larger capacitor helps but not enough at low frequency



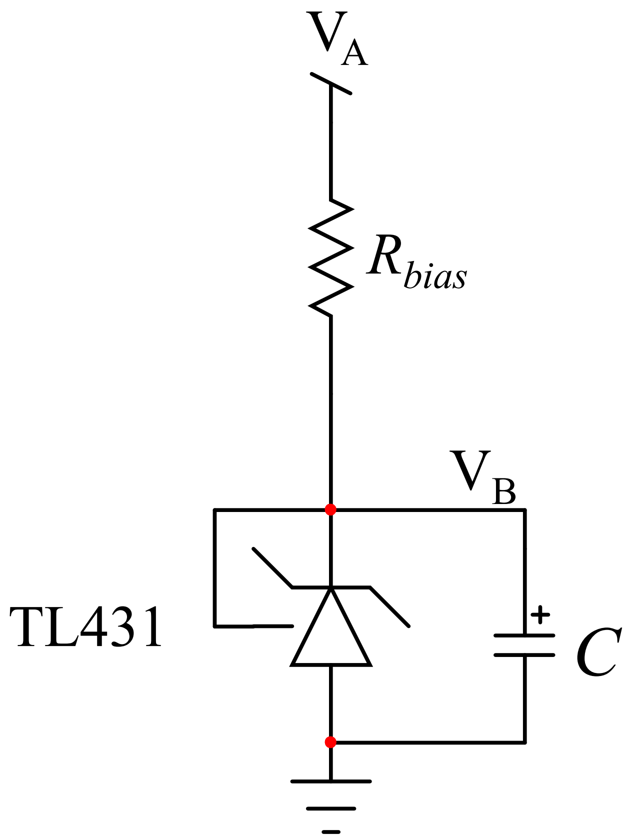


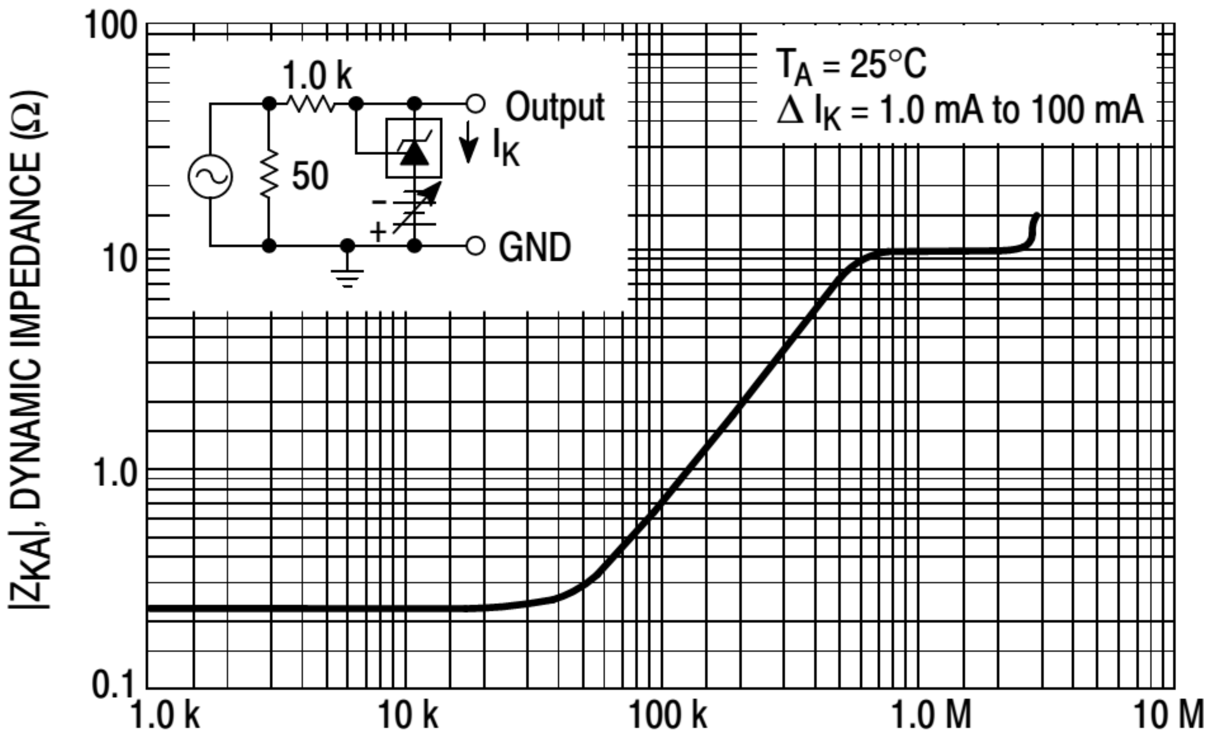




Calling an Active Zener Diode

❑ Why not using a programmable Zener diode?

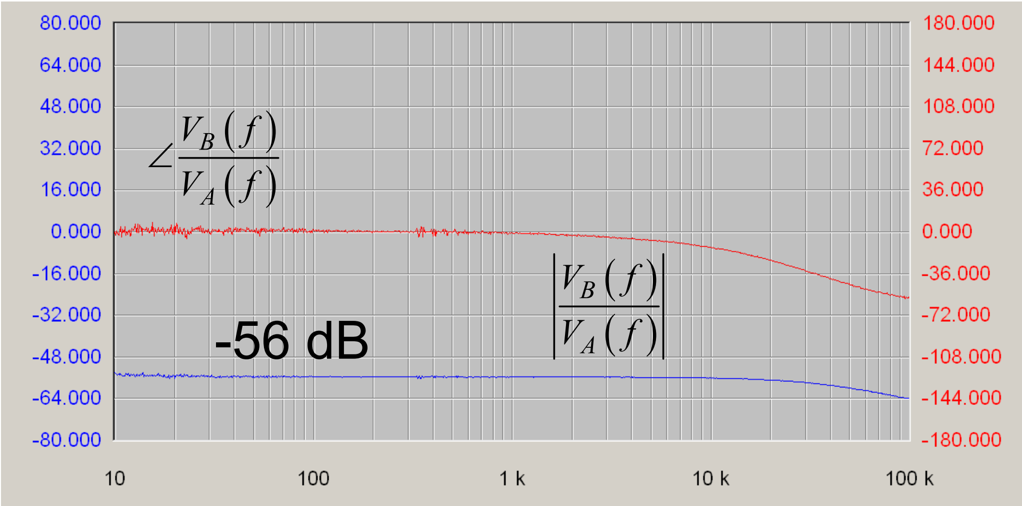


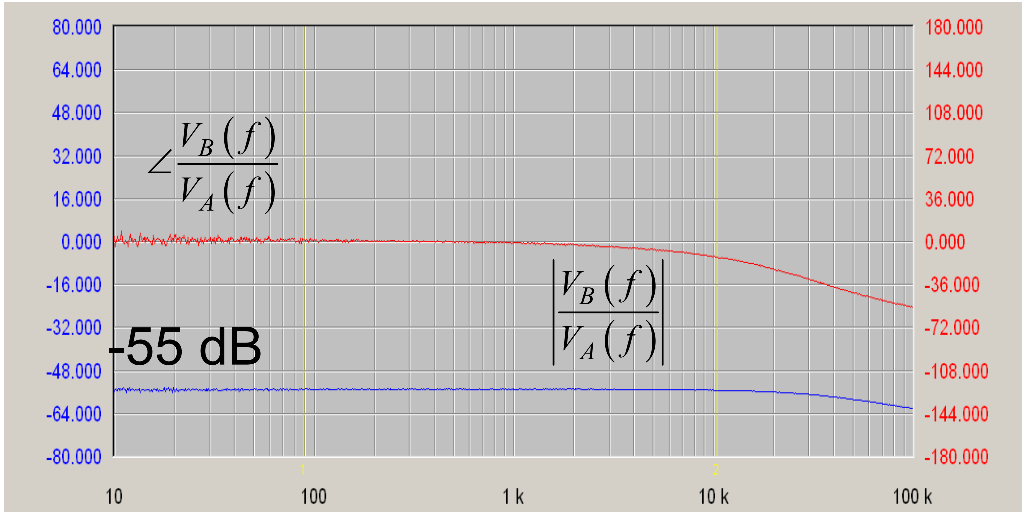


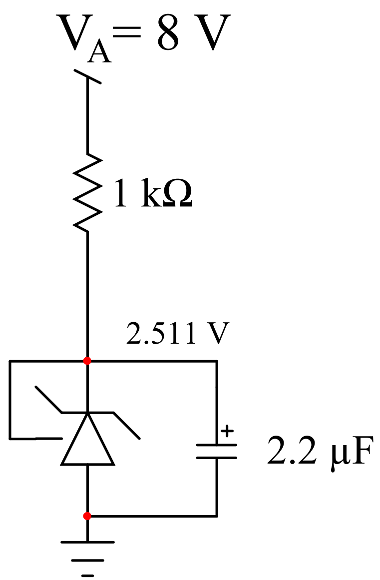
❑ The TL431 exhibits an excellent dynamic impedance

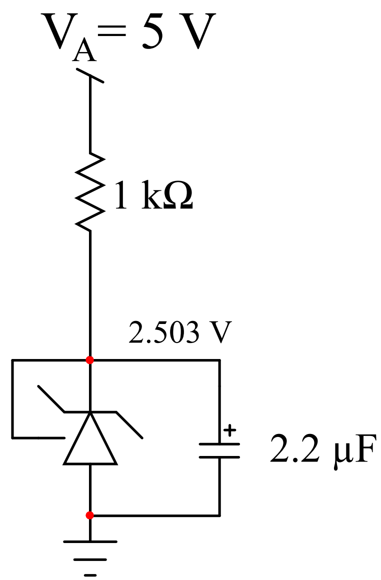
Dc-Bias Changes with the Active Zener

- The change in dc bias is not that dramatic with TL431



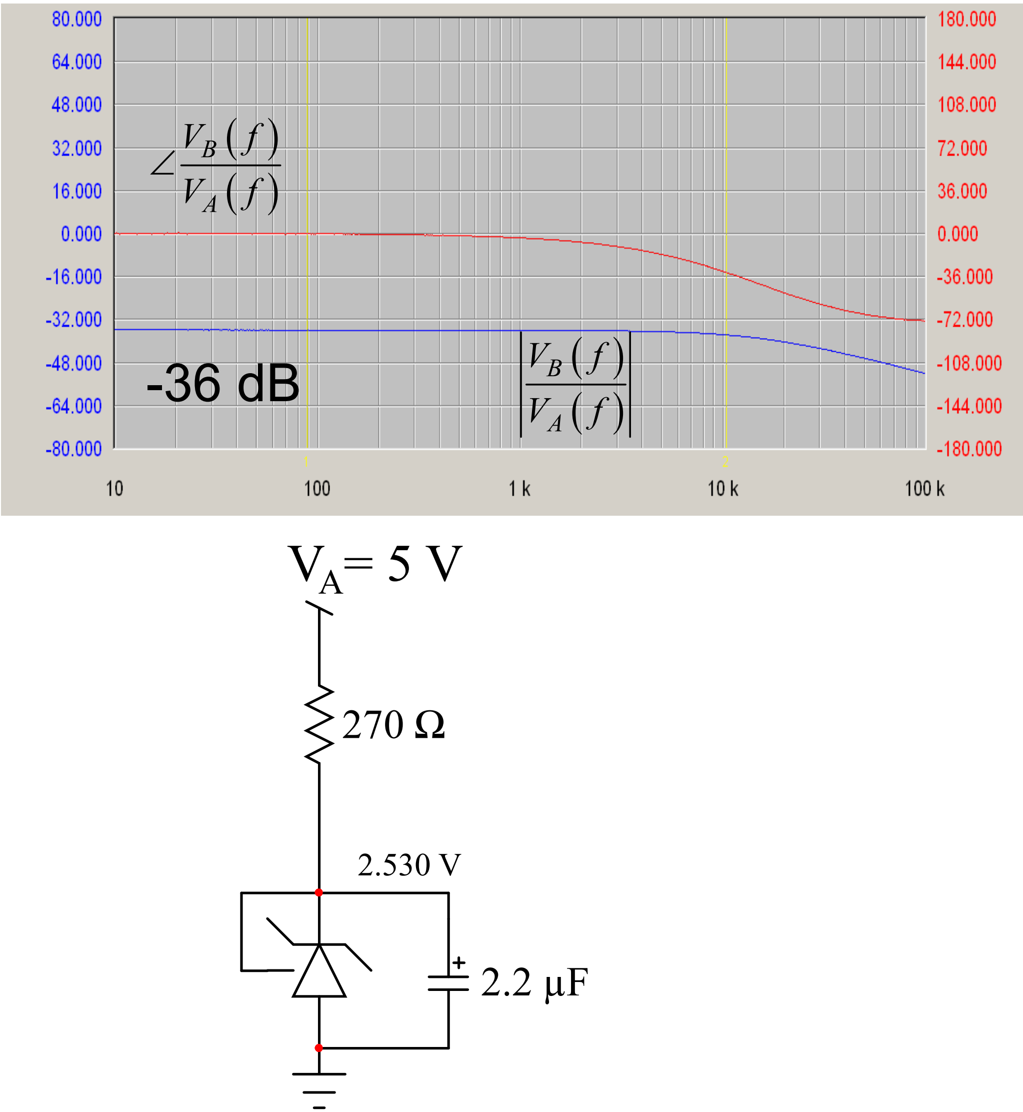


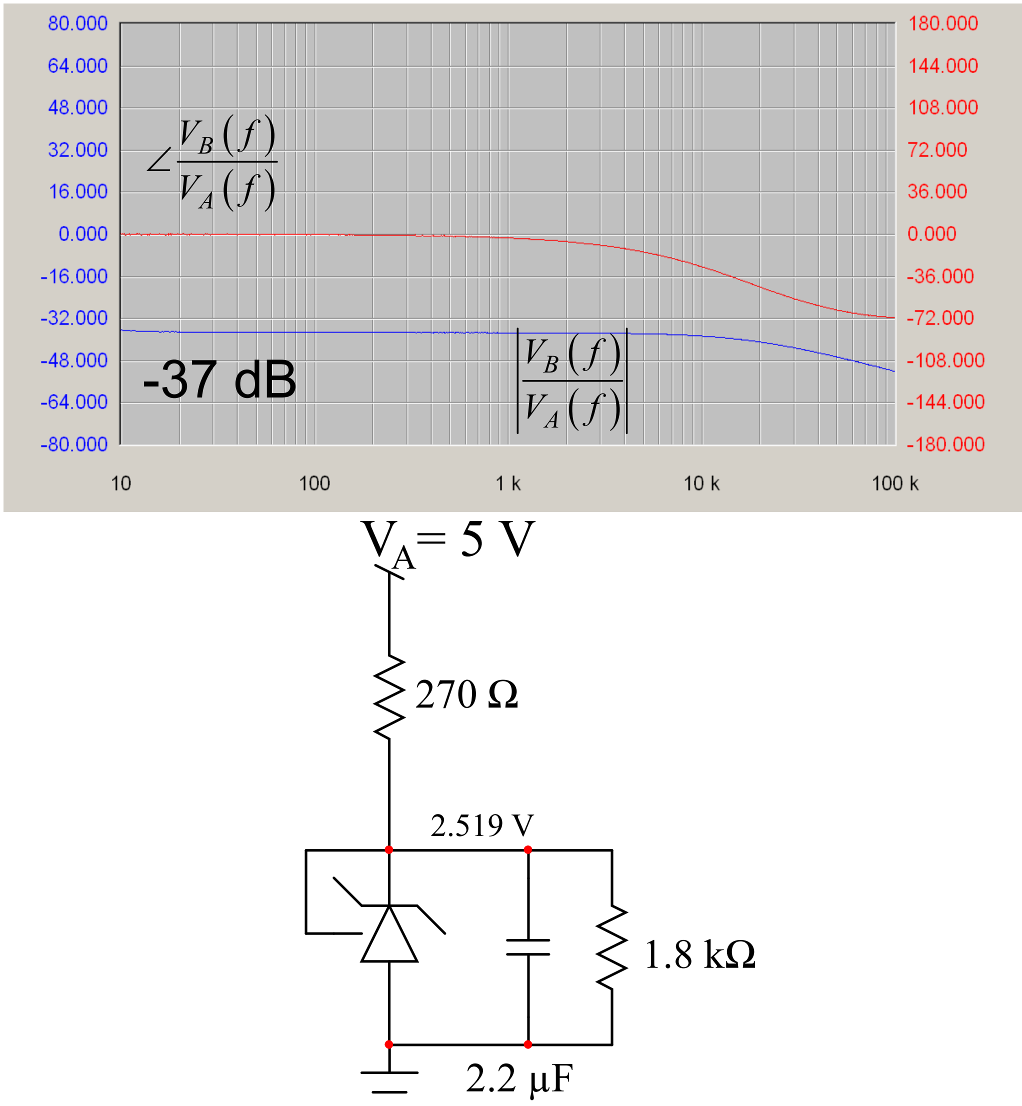




Watch Out for the Bias Resistor Selection

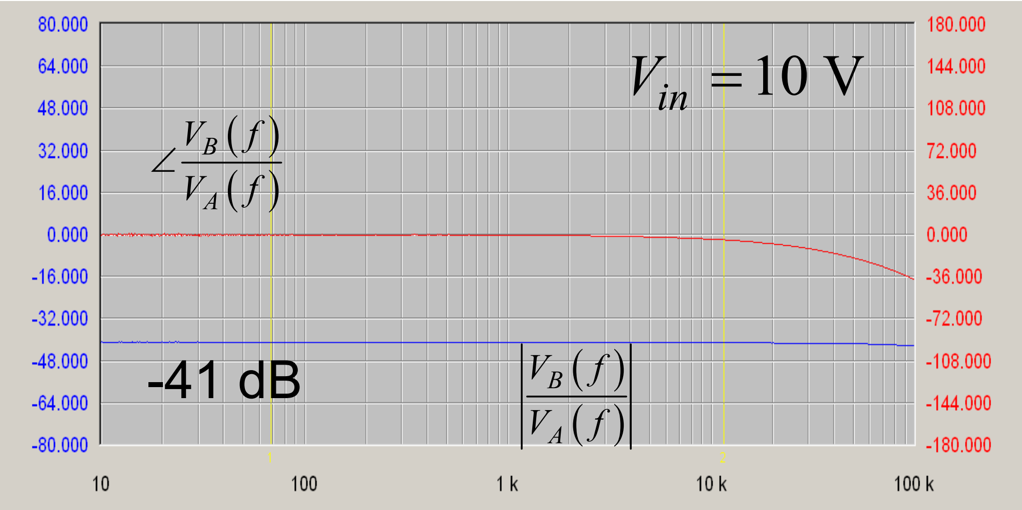
- Reducing the dropping resistor degrades rejection ratio

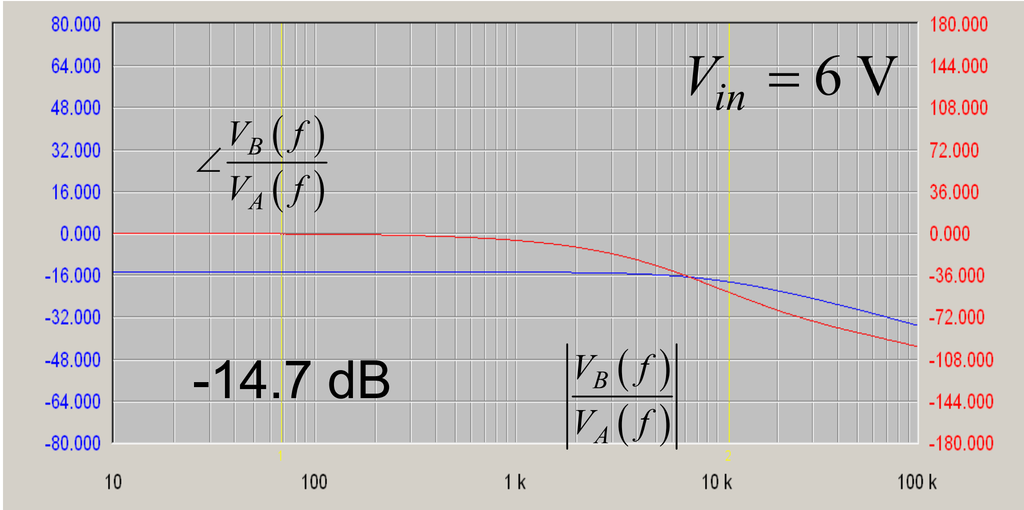


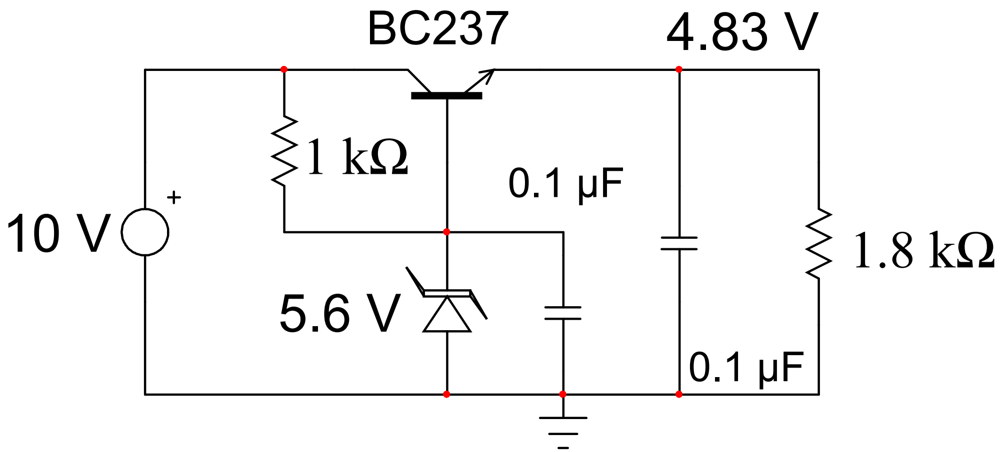


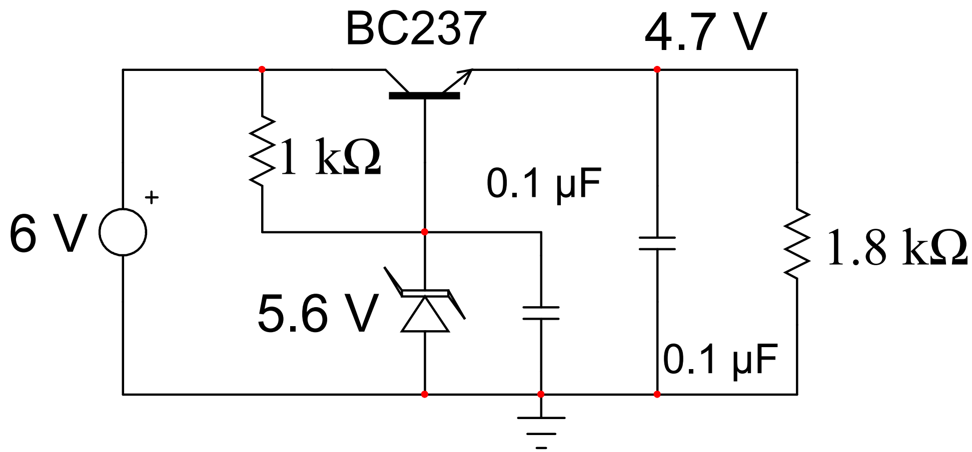
A Bipolar Also Brings Isolation

- ❑ A series transistor is slightly better but sensitive to V_{in}



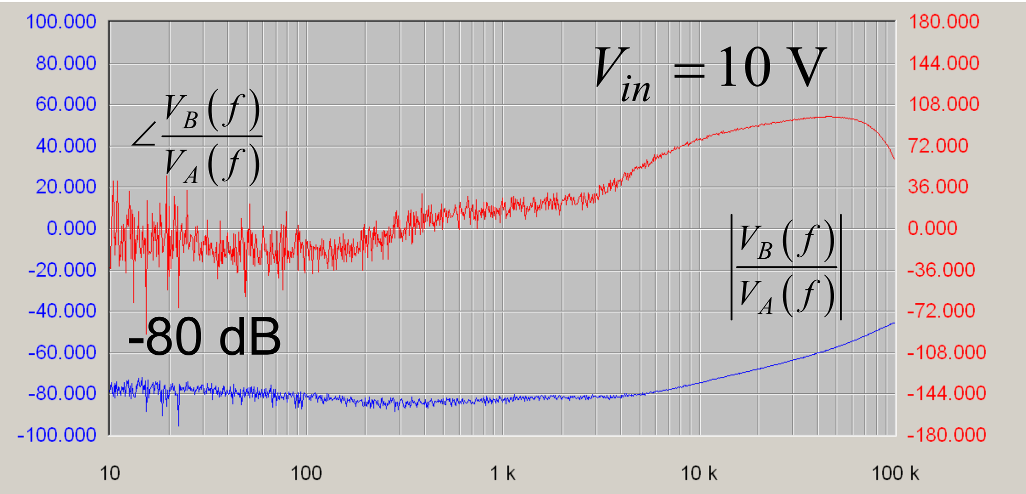


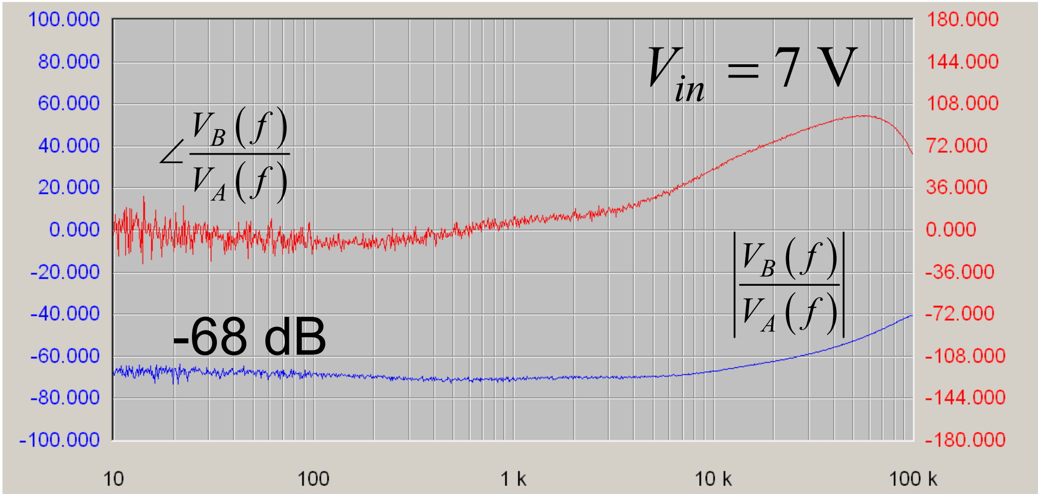


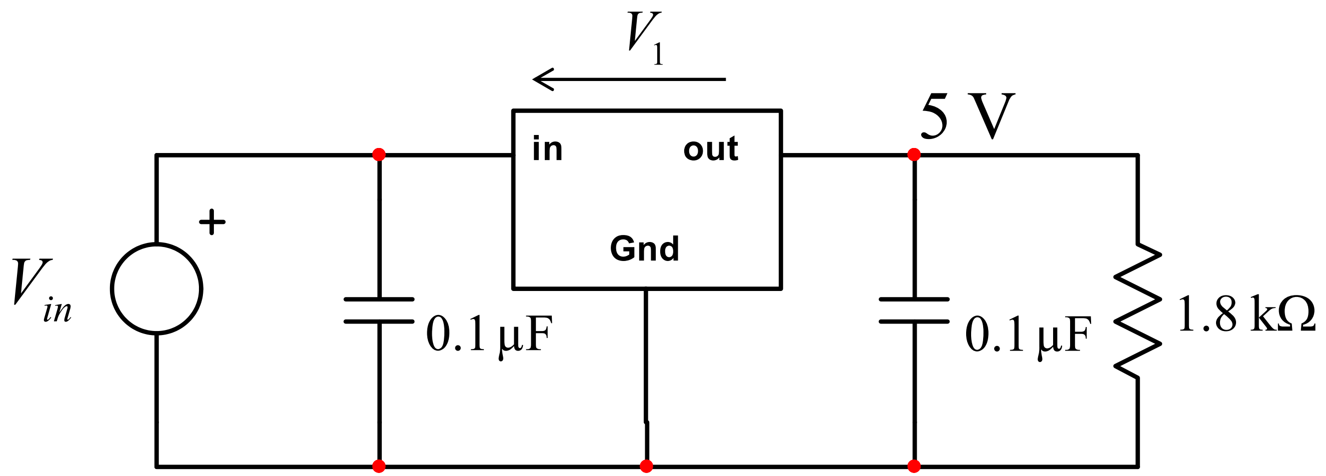


Regulators Can Also Do the Job

- Good rejection from a 78L05. Expect poor start-up behavior





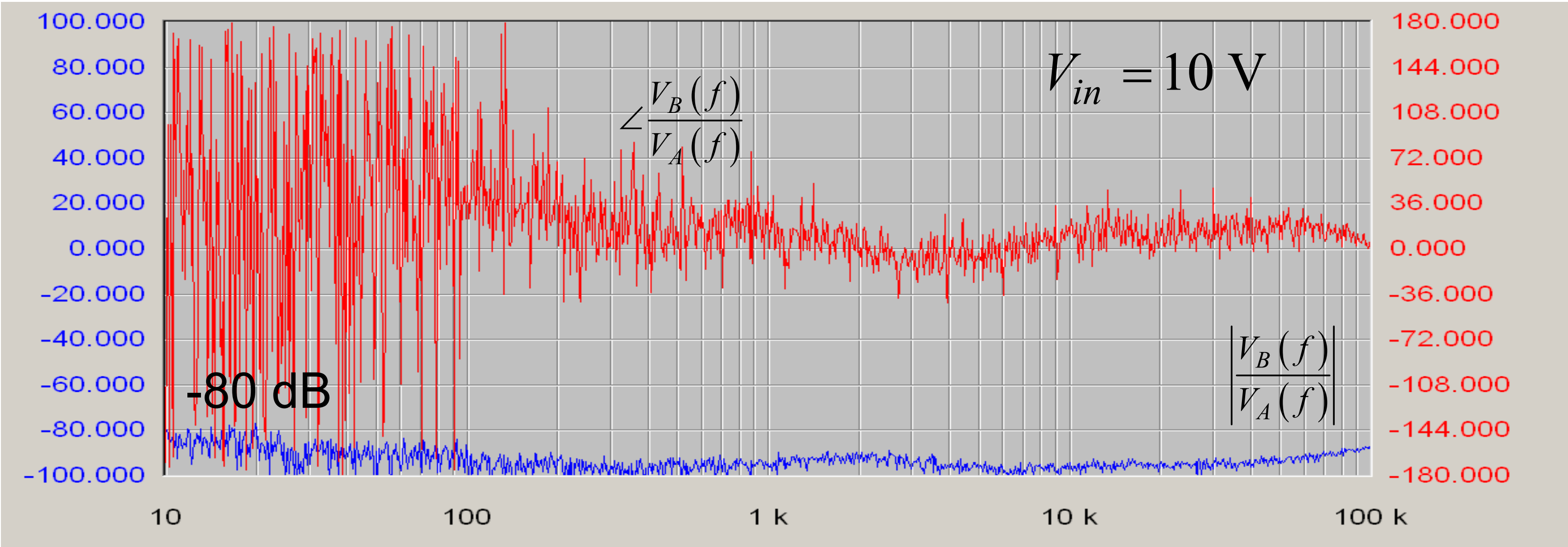


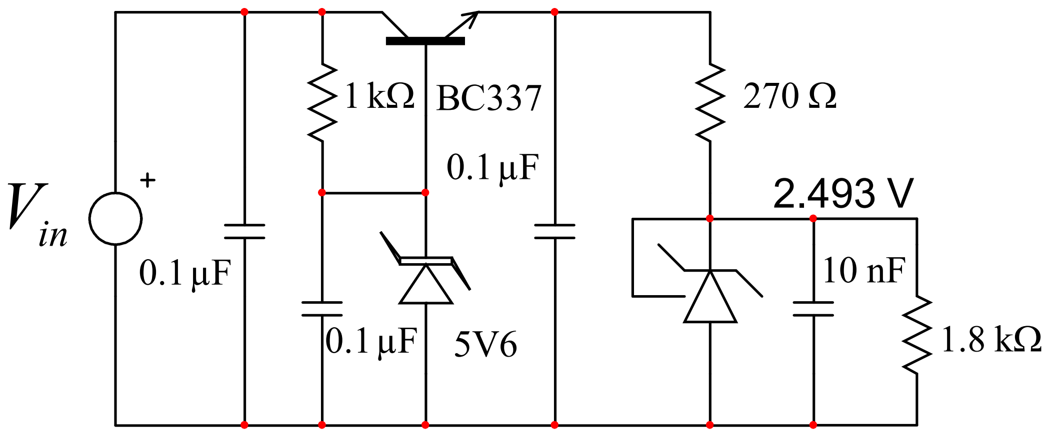
MC78L05

- Rejection depends on regulator dropout voltage V_1

Best Combination

- The strongest rejection is obtained by cascading stages

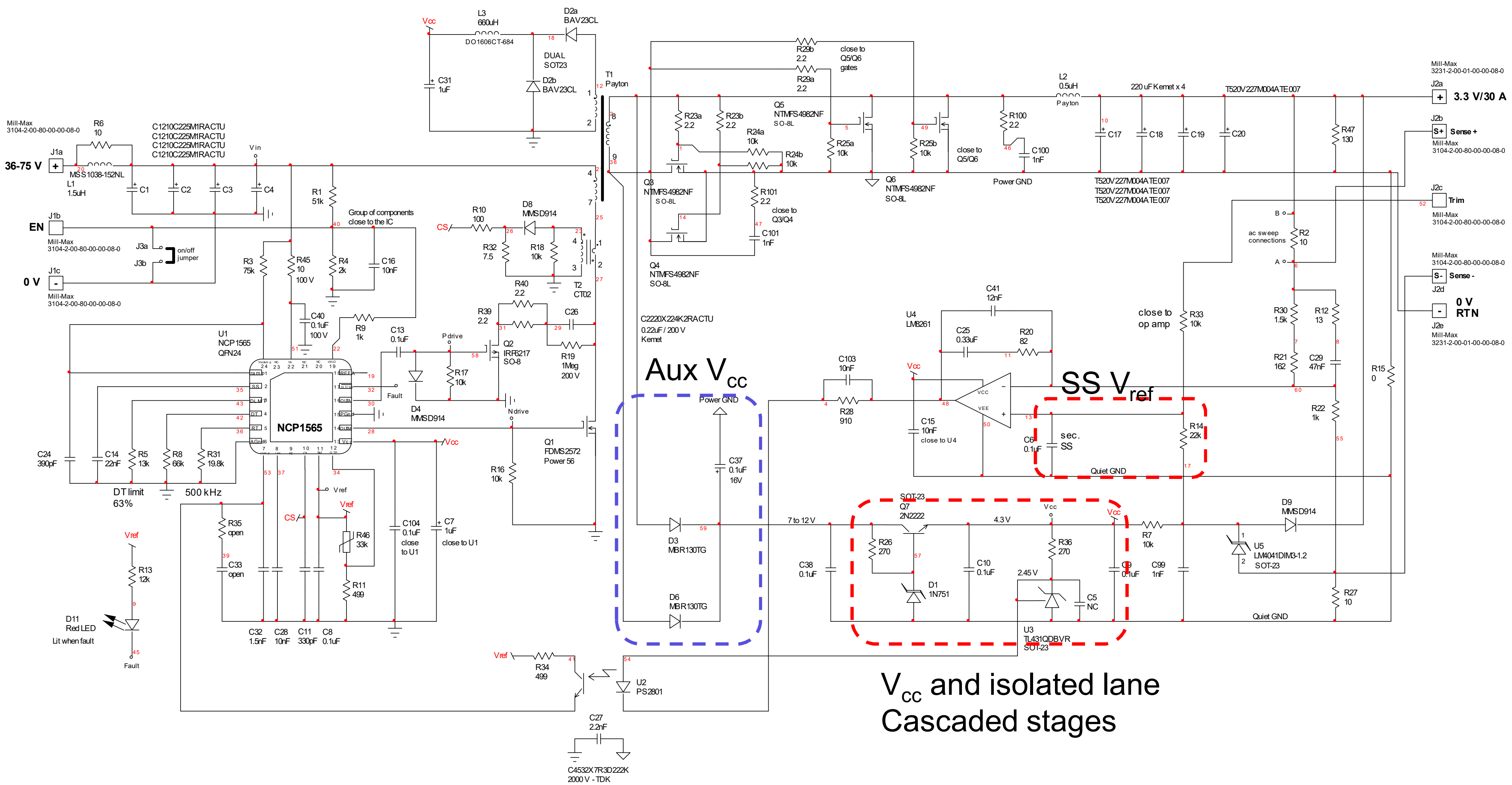




Low V_{out} is ok with op amp and LED connection

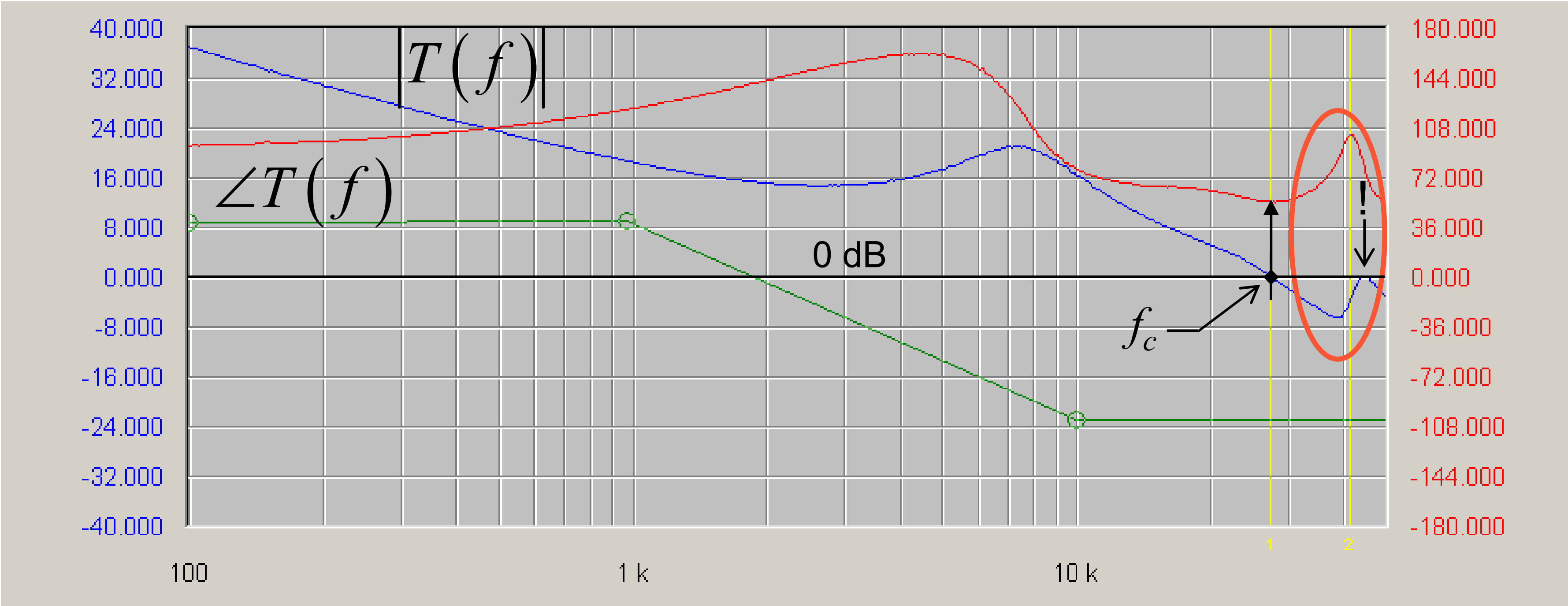
Practical Implementation

❑ Stabilizing a 3.3-V/20-A Active Clamp Forward converter



Check the Compensated Loop Response

- Open-loop Bode plot shows good phase margin at 27-kHz f_c



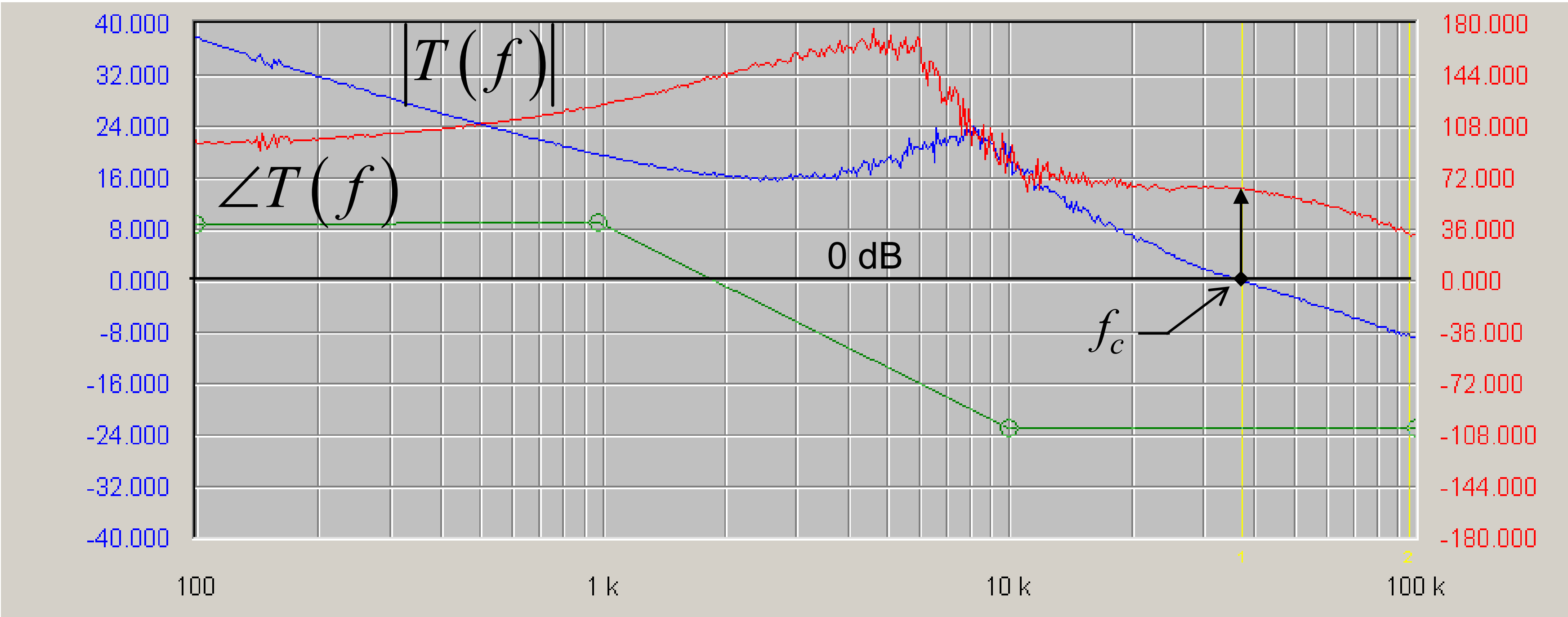
V_{in} = 36 V, 110-mΩ constant-resistance load

- Resonance in the upper spectrum is linked to the EMI filter

Apply damping

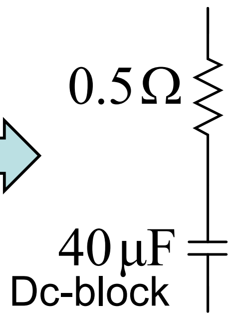
Damping Filter Helps Removing the Glitch

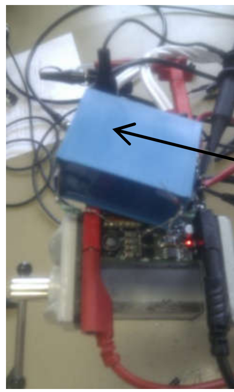
- Damping removes glitches and improves f_c to 35 kHz



V_{in} = 36 V, 110-mΩ constant-resistance load

$$R_{damp} := R_{in} \cdot \frac{L_1 + C_{tot} \cdot r_C \cdot r_L - \frac{r_L}{\omega_0}}{-2 \cdot R_{in} \cdot C_{tot} \cdot r_L + \frac{R_{in}}{\omega_0} + L_1 + C_{tot} \cdot r_L \cdot r_C - \frac{r_C}{\omega_0}} = 0.422\Omega$$

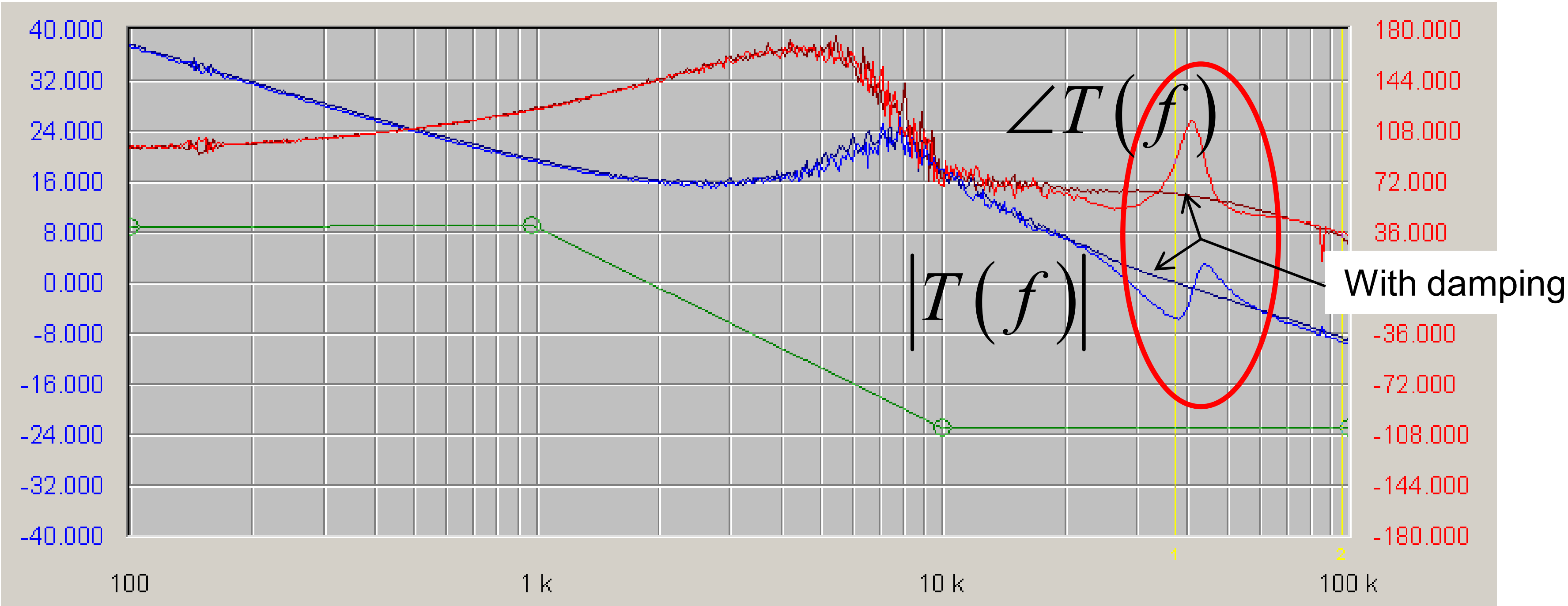




40 µF (couldn't find bigger!)

Comparing Filter Effects on Loop Gains

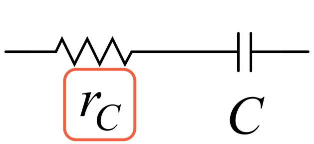
❑ EMI filter resonance degrades open-loop response



❑ Damping means increased losses per cycles

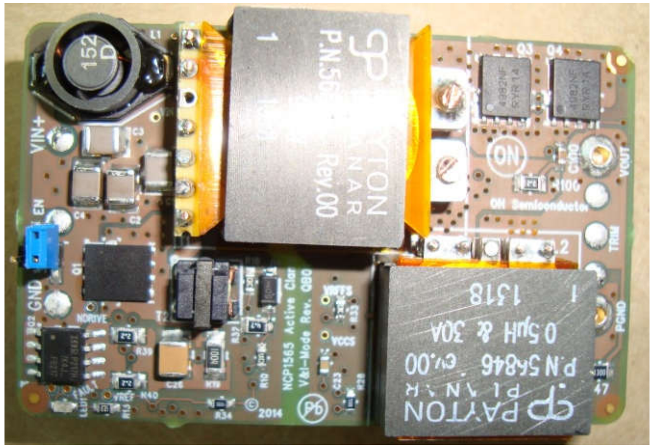
❑ An electrolytic cap. features an ESR

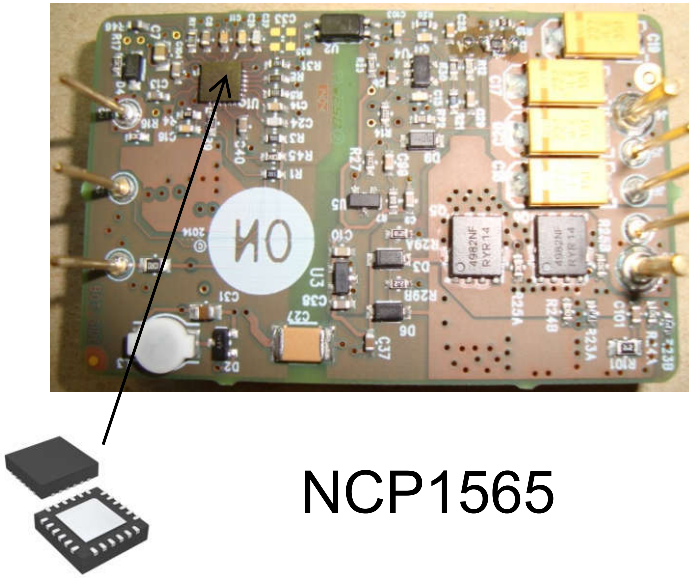
Naturally lossy cap.



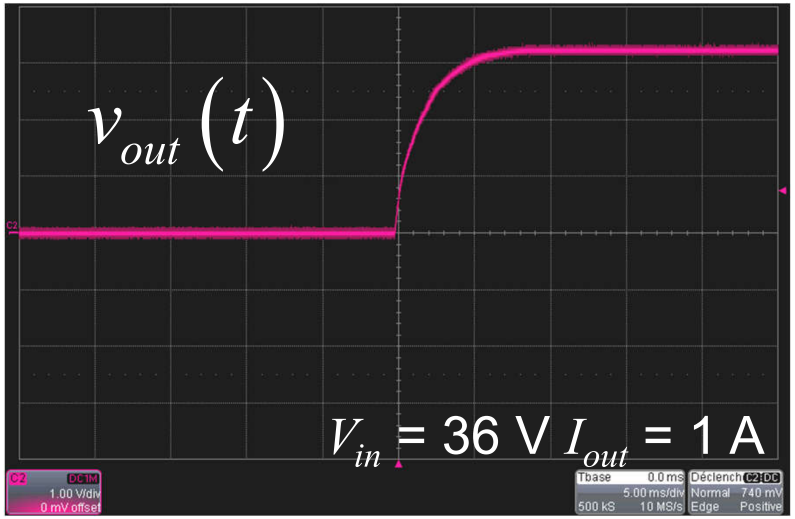
Run Transient Tests

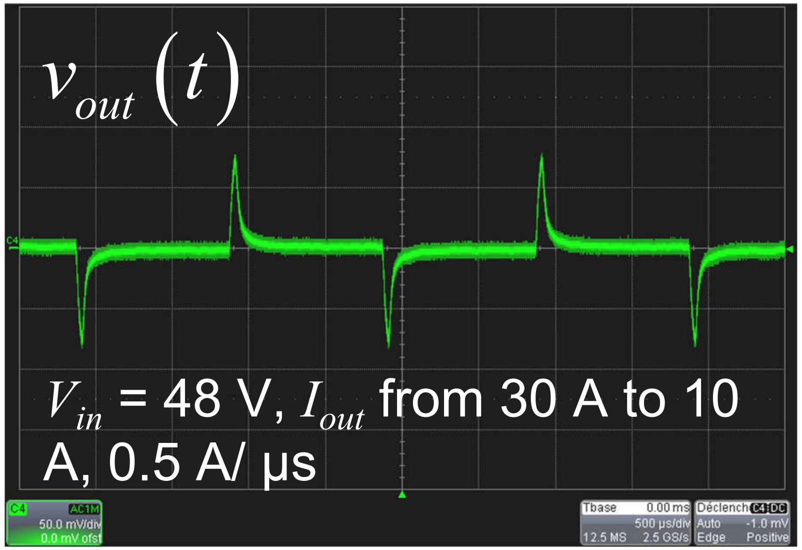
- Final test, run start-up sequences and transient steps





NCP1565



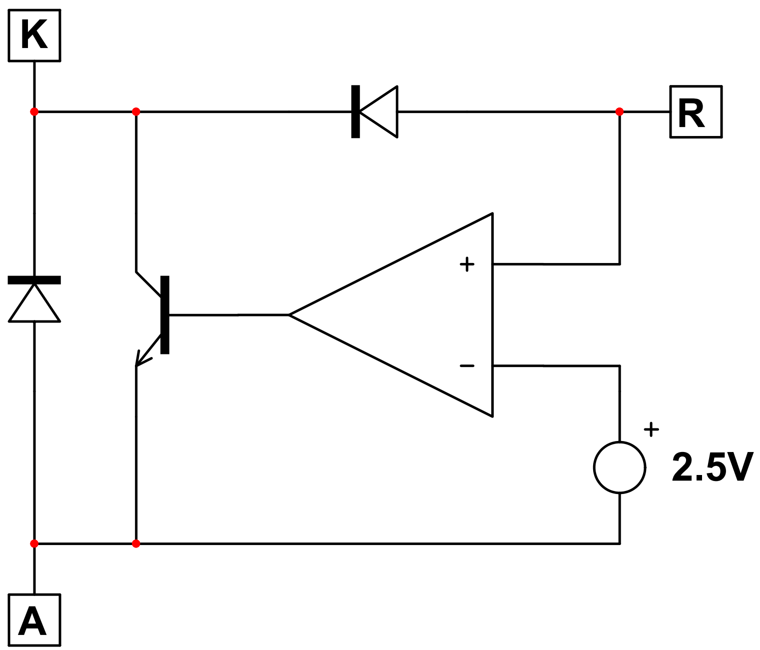


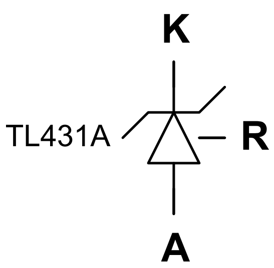
Course Agenda

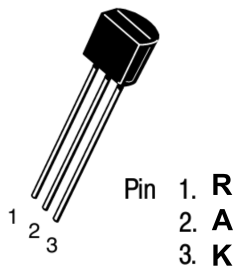
- ❑ Control System Basics
- ❑ Operational Amplifier and Low-Frequency Pole
- ❑ Gain-Bandwidth Impact on Phase Boost
- ❑ Op Amp Slew Rate Effects in Loop Control
- ❑ Start-Up Sequence and Auxiliary Supply
- ❑ Characterizing the Optocoupler Pole
- ❑ Dealing with the Fast Lane
- ❑ **Going Around the TL431 Fast Lane**

The TL431 Programmable Zener

- ❑ The TL431 is the most popular choice in nowadays designs
- ❑ It associates an open-collector op amp and a reference voltage
- ❑ The internal circuitry is self-supplied from the cathode current
- ❑ When the R node exceeds 2.5 V, it sinks current from its cathode



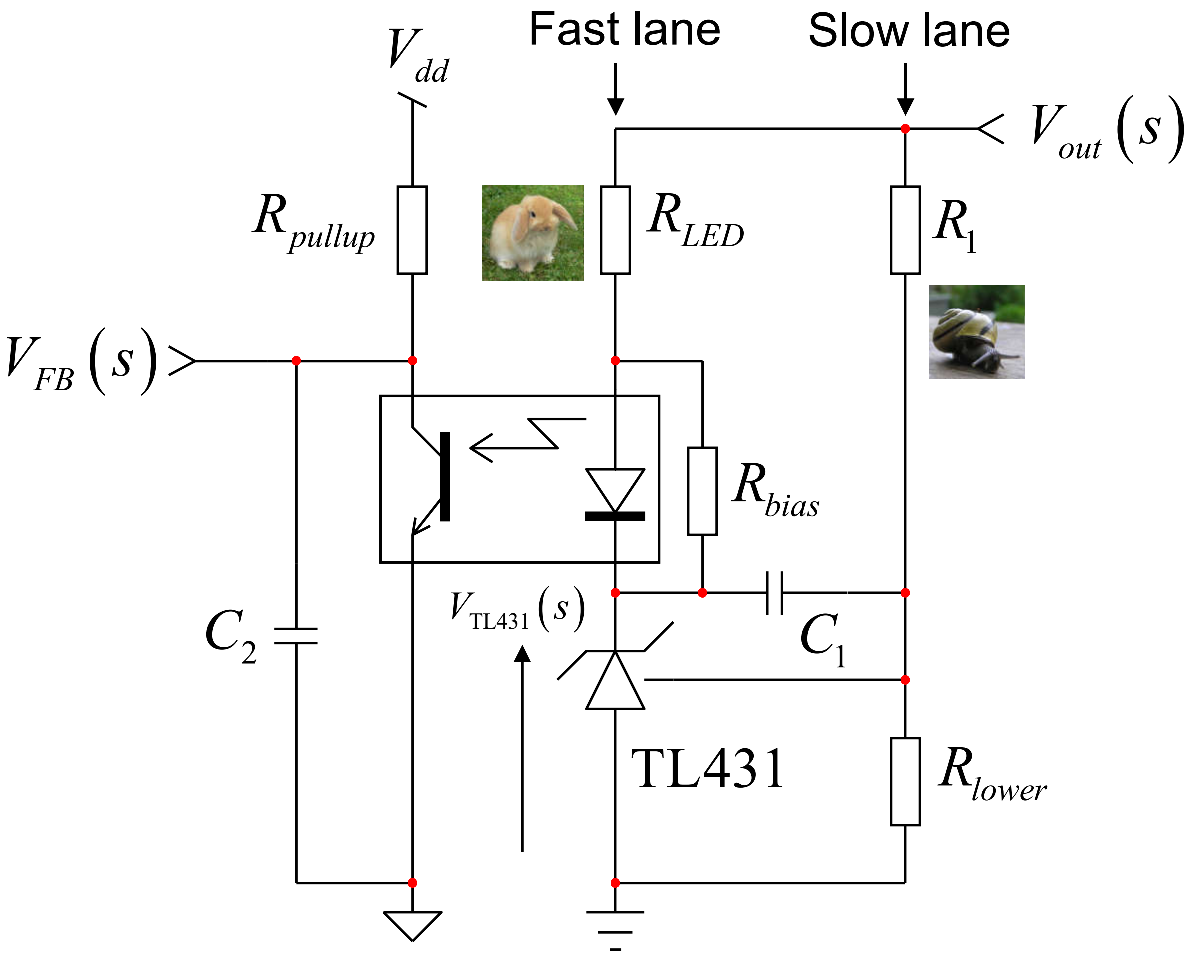




- ❑ The TL431 is a shunt regulator

Two Lanes to Drive the LED

- The TL431 lends itself very well to optocoupler control



$$I_{LED}(s) = \frac{V_{out}(s) - V_{TL431}(s)}{R_{LED}}$$

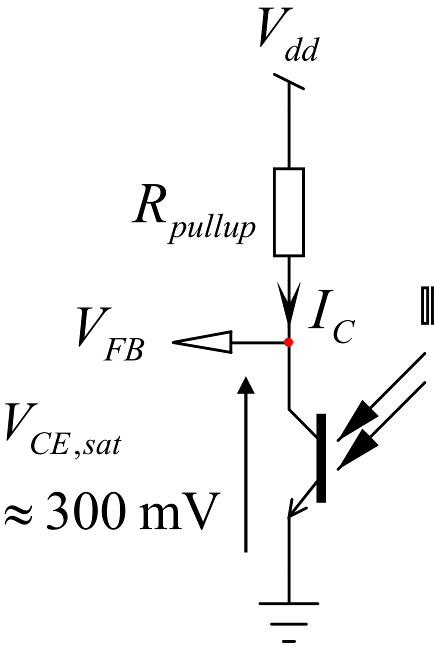
$$I_{LED}(s) = \boxed{\frac{V_{out}(s)}{R_{LED}}} - \boxed{\frac{V_{TL431}(s)}{R_{LED}}}$$

Fast lane Slow lane

- R_{LED} connected to V_{out} offers a direct path to the LED: fast lane!

Bias Conditions Set an Upper Limit

- R_{LED} cannot exceed a certain value because of bias limits
- V_{FB} must swing between $V_{CE,sat}$ and V_{dd}

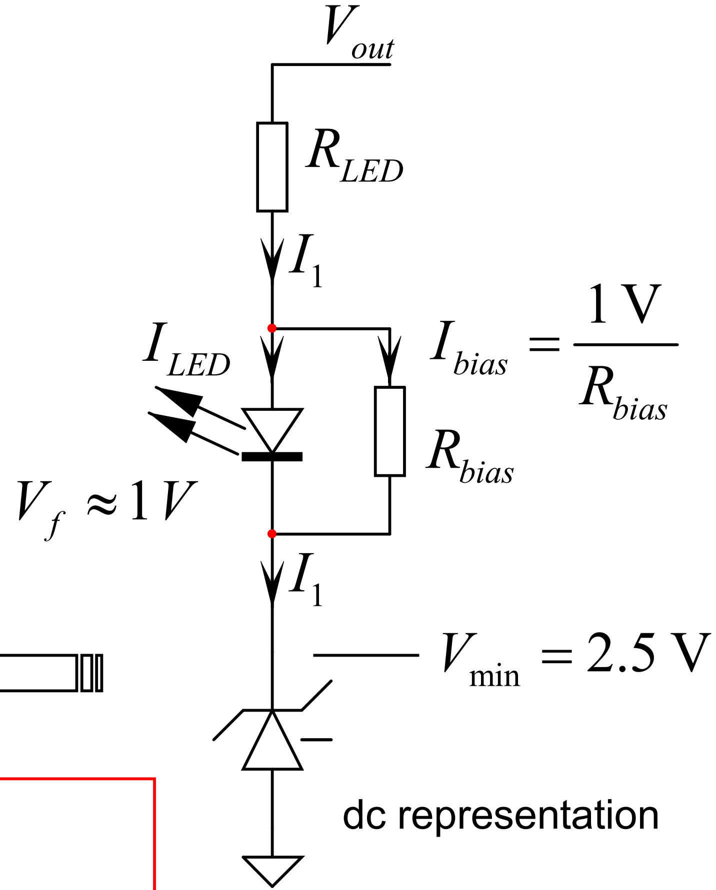


$$I_{C,\max} = \frac{V_{cc} - V_{CE,sat}}{R_{pullup}}$$

$$I_{1,\max} = \frac{V_{dd} - V_{CE,sat}}{R_{pullup}\,\mathrm{CTR}_{\min}} + I_{bias}$$

$$\Updownarrow$$

$$I_{1,\max} = \frac{V_{out} - V_f - V_{\mathrm{TL431,min}}}{R_{LED}}$$

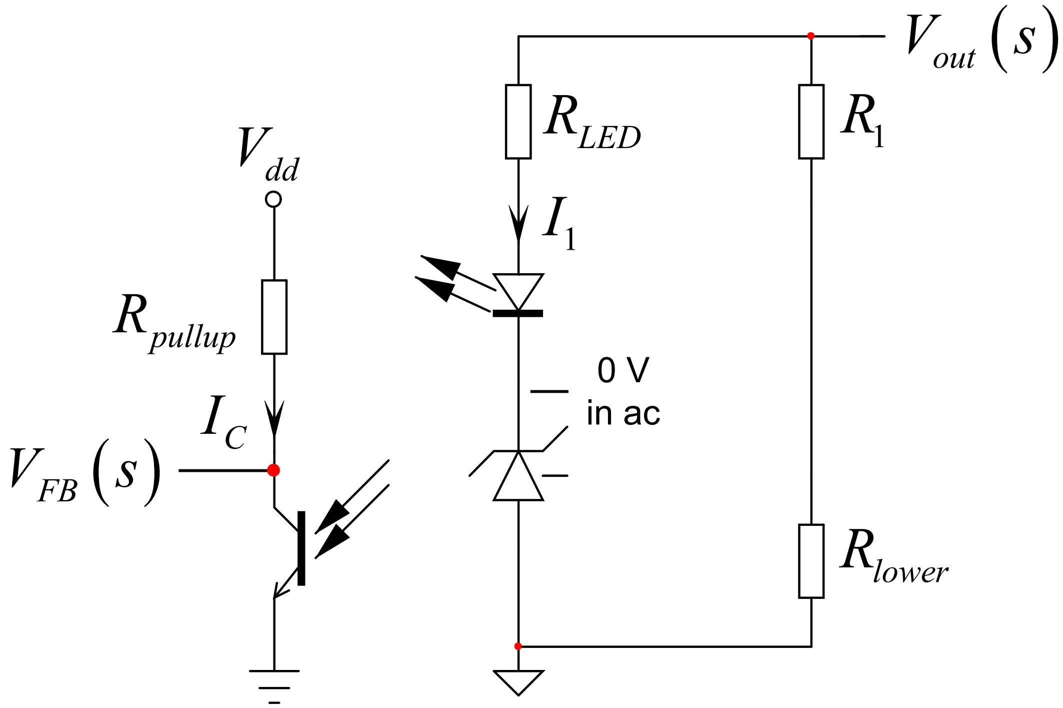


$$\boxed{R_{LED,\max} = \frac{V_{out} - V_f - V_{TL431,\min}}{V_{dd} - V_{CE,sat} + I_{bias}\,\mathrm{CTR}_{\min}\,R_{pullup}}\,R_{pullup}\,\mathrm{CTR}_{\min}}$$

ON Semiconductor®

The Gain is Affected by Bias Limits

- At high frequencies, the TL431 ac output is zero, C_1 is a short-circuit
- R_{LED} alone fixes the fast-lane gain



$$V_{FB}(s) = -\mathrm{CTR} \cdot R_{pullup} \cdot I_1$$

$$I_1 = \frac{V_{out}(s)}{R_{LED}}$$

$$\frac{V_{FB}(s)}{V_{out}(s)} = -\mathrm{CTR}\frac{R_{pullup}}{R_{LED}}$$

- R_{LED} must also leave headroom to bias the TL431: upper limit!

A Quick Design Example

- Assume the following design:

$$V_{out} = 5\text{ V}$$
$$V_f = 1\text{ V}$$
$$V_{TL431,\min} = 2.5\text{ V}$$
$$V_{dd} = 4.8\text{ V}$$
$$V_{CE,sat} = 300\text{ mV}$$
$$I_{bias} = 1\text{ mA}$$
$$\text{CTR}_{\min} = 0.3$$
$$R_{pullup} = 20\text{ k}\Omega$$

$$R_{LED,\max} \le \frac{5-1-2.5}{4.8-0.3+1m\times 0.3\times 20k}\times 20\text{k}\times 0.3$$

$$\Downarrow$$

$$R_{LED,\max} \le 857\ \Omega$$

$$\Downarrow$$

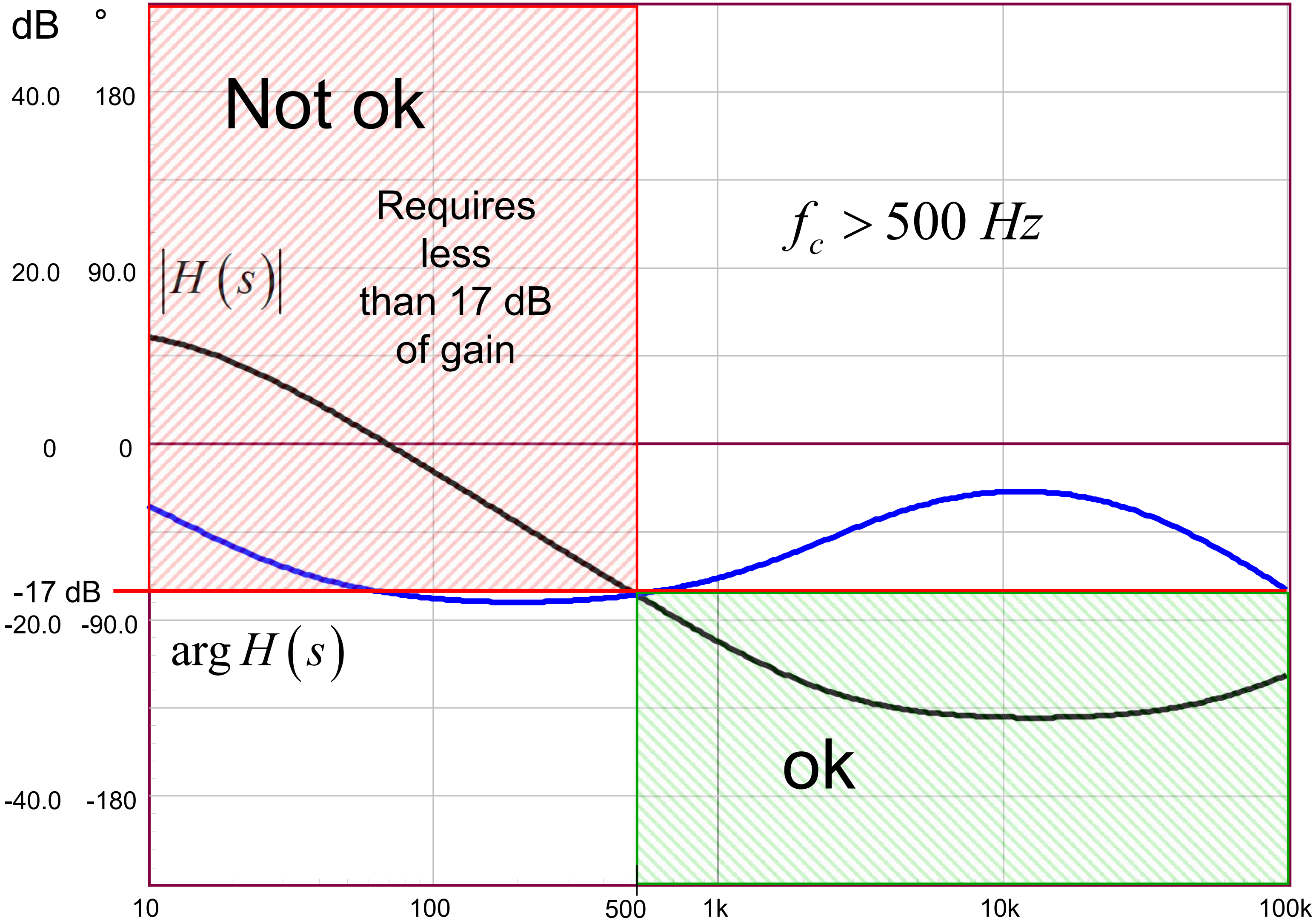
$$G_0 > \text{CTR}\frac{R_{pullup}}{R_{LED}} > 0.3\frac{20}{0.857} > 7 \text{ or } \approx 17\text{ dB}$$

- In designs where R_{LED} fixes the gain, G_0 cannot be below 17 dB

⇨ You cannot "amplify" by less than 17 dB

Crossover Frequency Choices are Limited

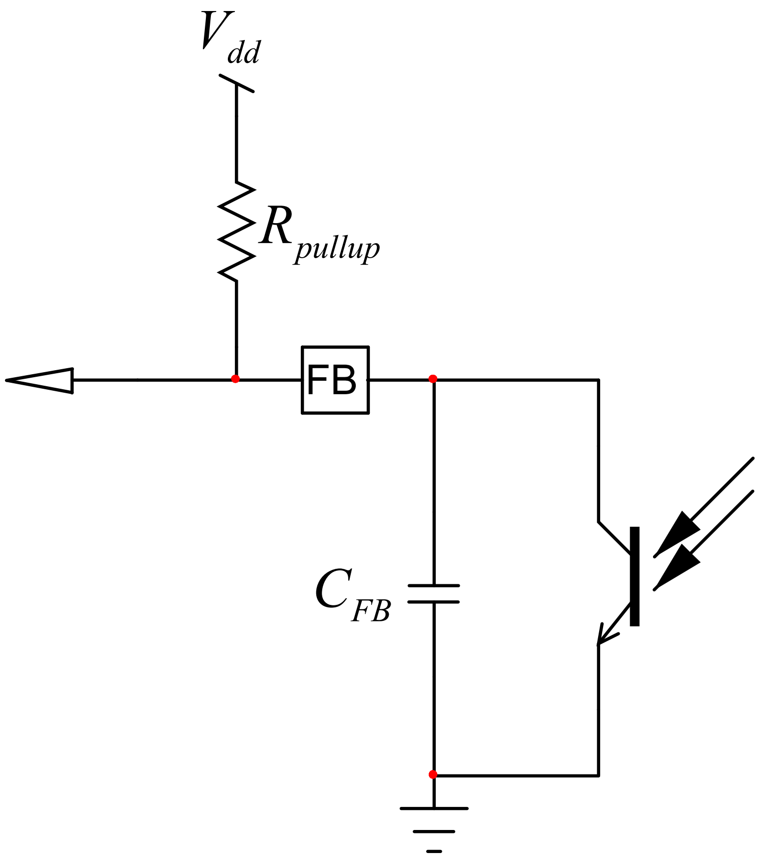
- You must identify areas where crossover is possible

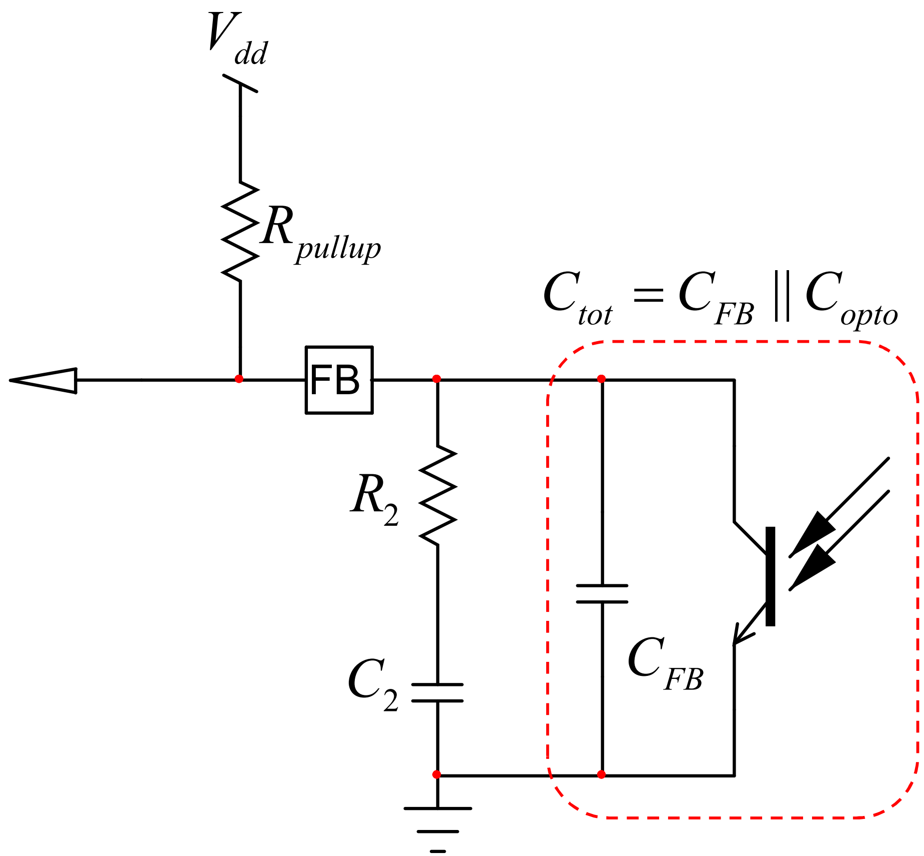


Requires 17 dB or more

Going Around the Bias Limit

- The pull-up resistor is internal and cannot be altered
- One way to modify the mid-band gain is by adding an RC circuit



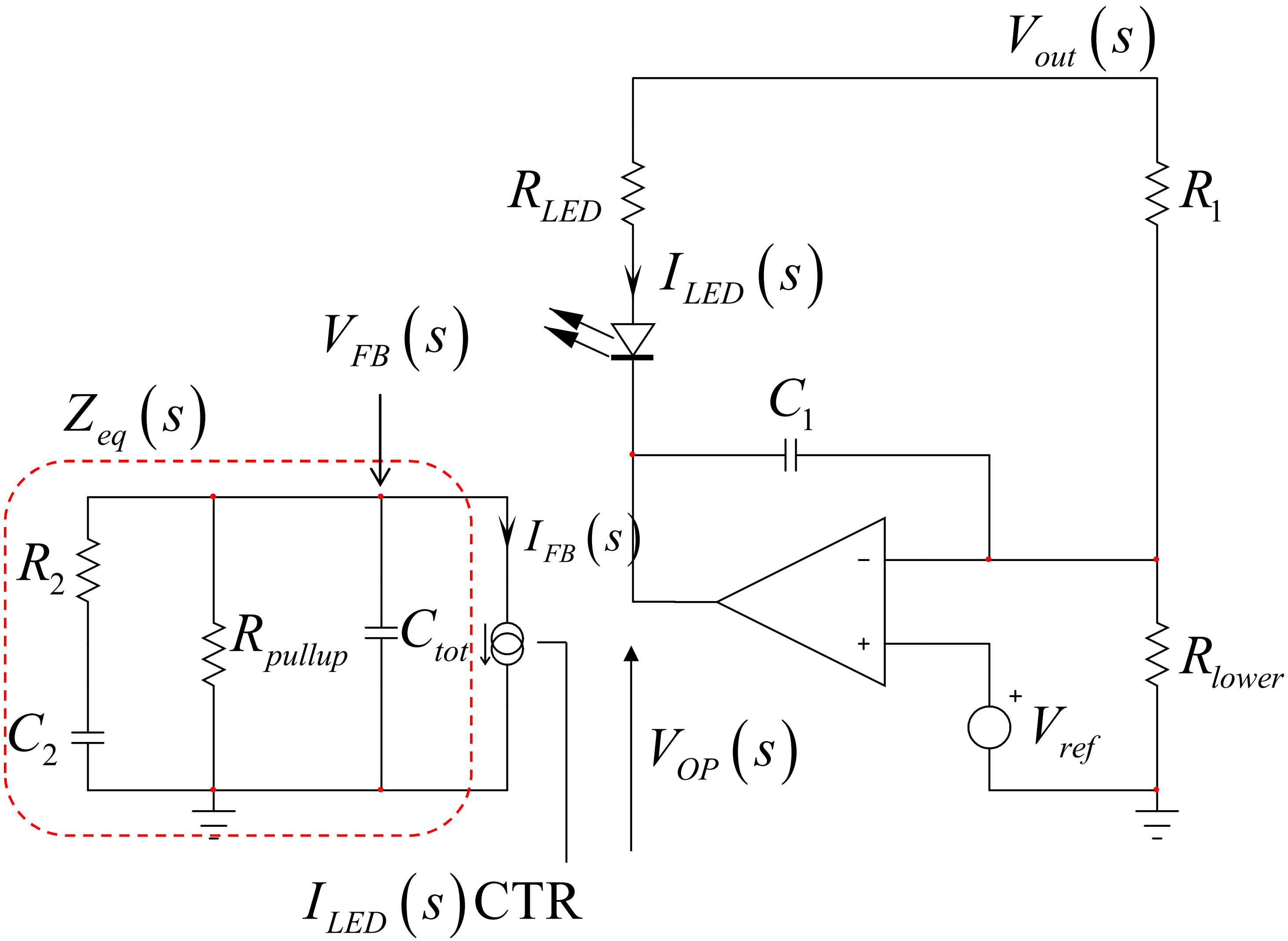


- The capacitance at the FB pin involves the optocoupler parasitics

C. Richardson, http://www.powerguru.org/a-simple-low-cost-technique-for-compensating-isolated-converters/

An Equivalent Small-Signal Model

- The new small-signal circuit includes the extra R_2C_2 circuit



Define the LED Current

- The LED current can be defined by

$$I_{LED}(s)=\frac{V_{out}(s)-\boxed{V_{op}(s)}}{R_{LED}} \qquad V_{op}(s)=-\frac{V_{out}(s)}{sR_1C_1}$$

- Rearrange and factor

$$I_{LED}(s)=\frac{V_{out}(s)}{R_{LED}}\left(1+\frac{1}{sR_1C_1}\right)=\frac{V_{out}(s)}{R_{LED}}\left(\frac{1+sR_1C_1}{sR_1C_1}\right)=\frac{V_{out}(s)}{R_{LED}}\frac{1+\frac{s}{s_{z_1}}}{\frac{s}{s_{po}}}$$

- Rearrange and factor to unveil $V_{FB}(s)$

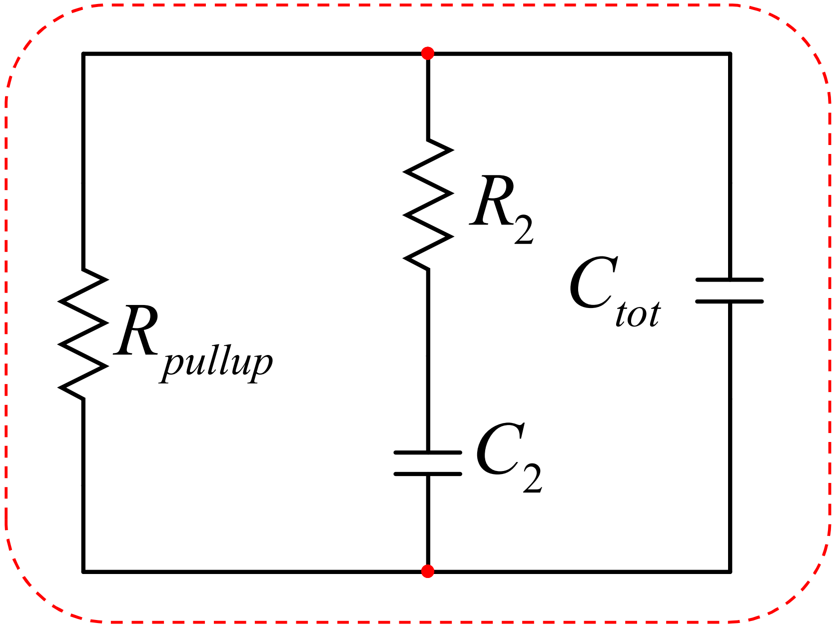
$$I_{FB}(s)=I_{LED}(s)\mathrm{CTR}$$

$$V_{FB}(s)=-I_{LED}(s)Z_{eq}(s)\mathrm{CTR}$$

$$\omega_{po}=\omega_{z_1}=\frac{1}{R_1C_1}$$

A 2nd-Order Impedance

- The equivalent impedance Z_{eq} is a second-order network
- It's time to call the FACTs!



$Z_{eq}(s)$

You apply brute-force analysis…

$$Z_{eq}(s) = \left(R_{pullup} \,||\, \frac{1}{sC_{tot}}\right) || \left(R_2 + \frac{1}{sC_2}\right)$$

or use FACTs for a low-entropy result

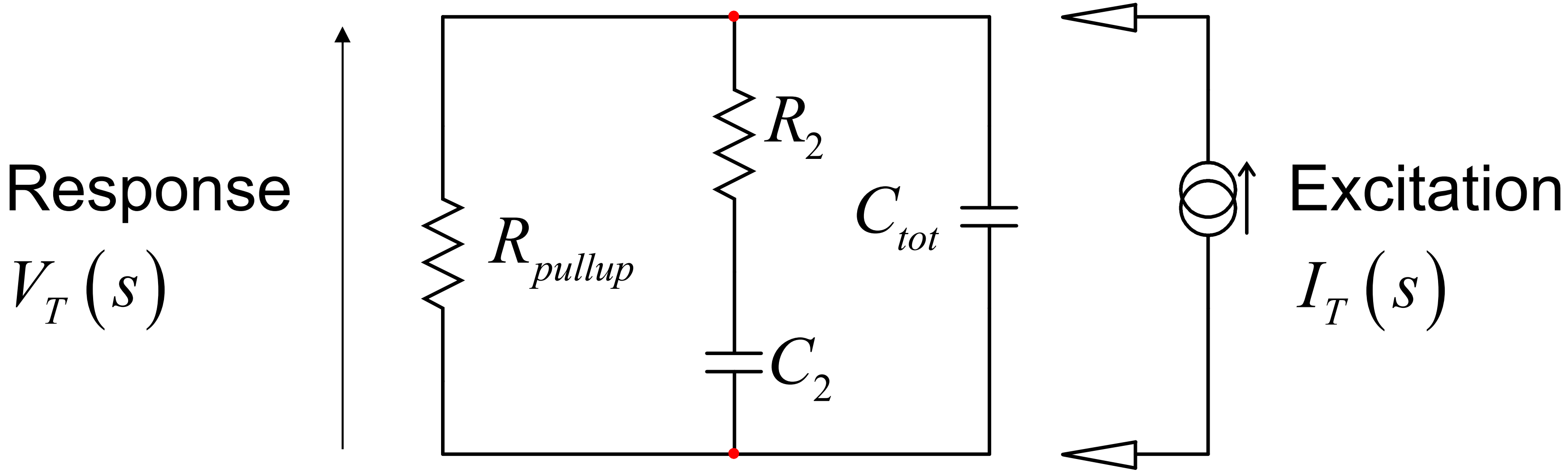
$$Z_{eq}(s) = R_0 \frac{1 + \frac{s}{s_{z_1}}}{1 + a_1 s + a_2 s^2}$$

- There are two storage elements, this is a 2nd–order system

V. Vorpérian, "Fast Analytical Techniques for Electrical and Electronic Circuits", Cambridge Press 2004

Identify Stimulus and Response

- The circuit is driven by a current source, the excitation
- The voltage is the response

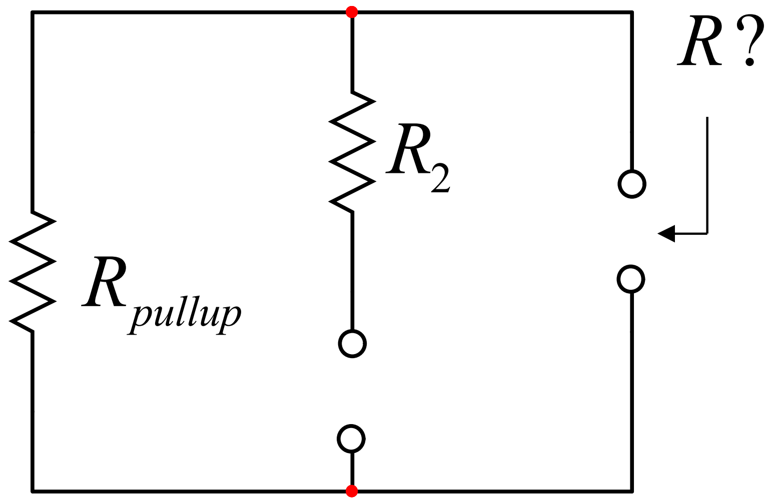


- The impedance is the ratio of V_T over I_T

$$Z_{eq}(s) = \frac{V_T(s)}{I_T(s)} \begin{matrix} \longleftarrow \text{response} \\ \longleftarrow \text{stimulus} \end{matrix}$$

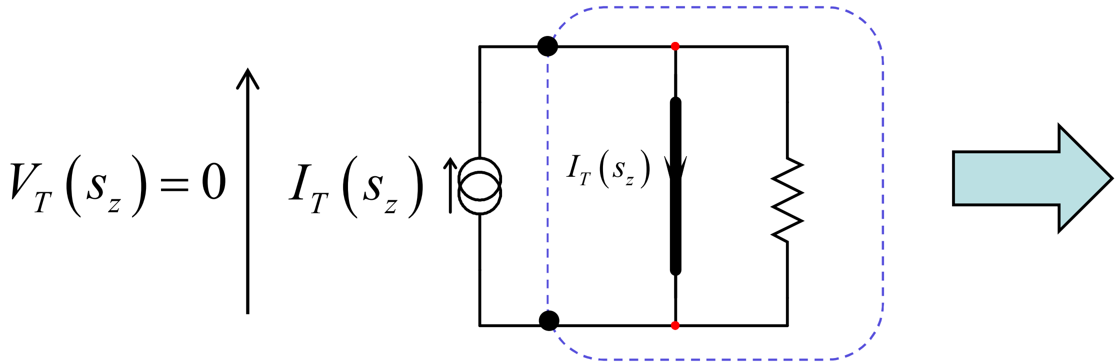
Start by the Circuit Examined in Dc

- For $s = 0$, open the capacitors and define R_0



$$R_0 = R_{pullup}$$

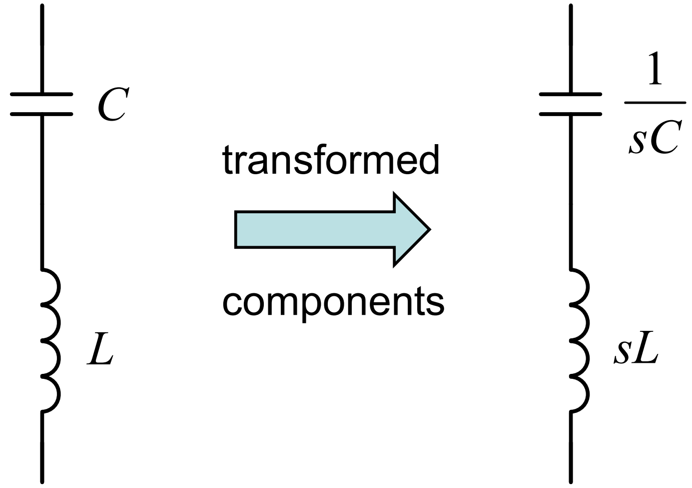
- Examine circuit conditions that could lead to $V_{out} = 0$ at $s = s_z$

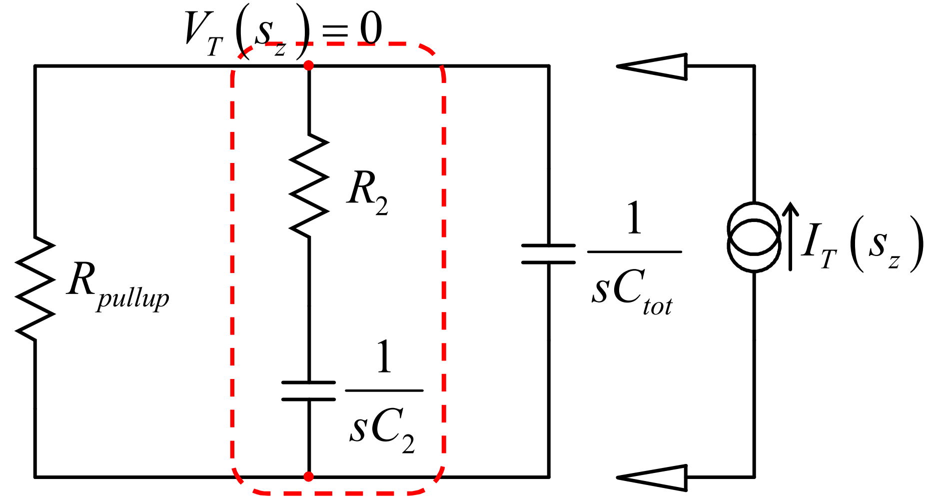


A transformed short could null the response

Examine the Transformed Circuit

- A transformed circuit associates C and L impedances





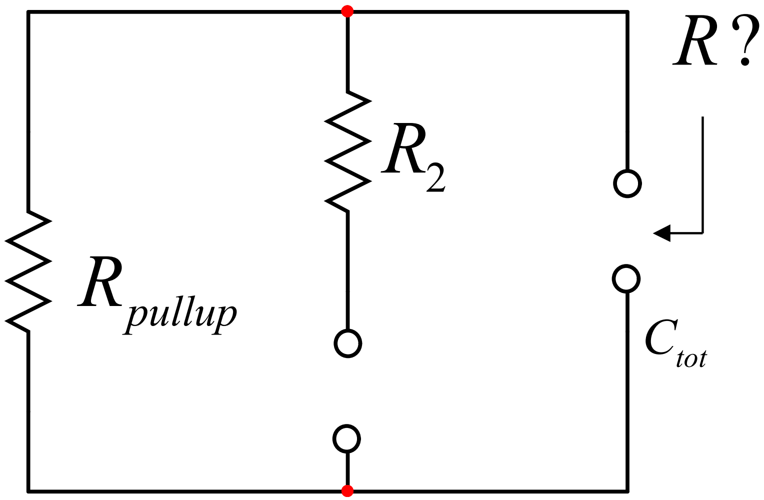
- A transformed short circuit brings the response to a null

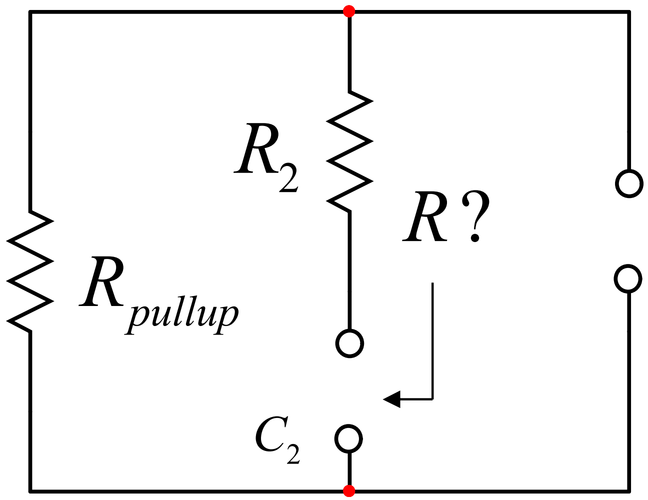
$$R_2+\frac{1}{sC_2}=\frac{1+sR_2C_2}{sC_2}=0 \longrightarrow 1+sR_2C_2=0 \longrightarrow s_{z_2}=-\frac{1}{R_2C_2}$$

$$\Rightarrow \omega_{z_2}=\frac{1}{R_2C_2}$$

Calculate Time Constants

- A transformed circuit associates C and L impedances





$$\tau_2 = R_{pullup}C_{tot}$$

$$\tau_1 = \left(R_{pullup} + R_2\right)C_2$$

- The denominator is unitless

$$D(s) = 1 + a_1 s + a_2 s^2 \left\} \begin{array}{l} a_1\left[\mathrm{s}\right] \\ a_2\left[\mathrm{s}^2\right] \end{array} \right.$$

$$a_1 = \tau_1 + \tau_2$$

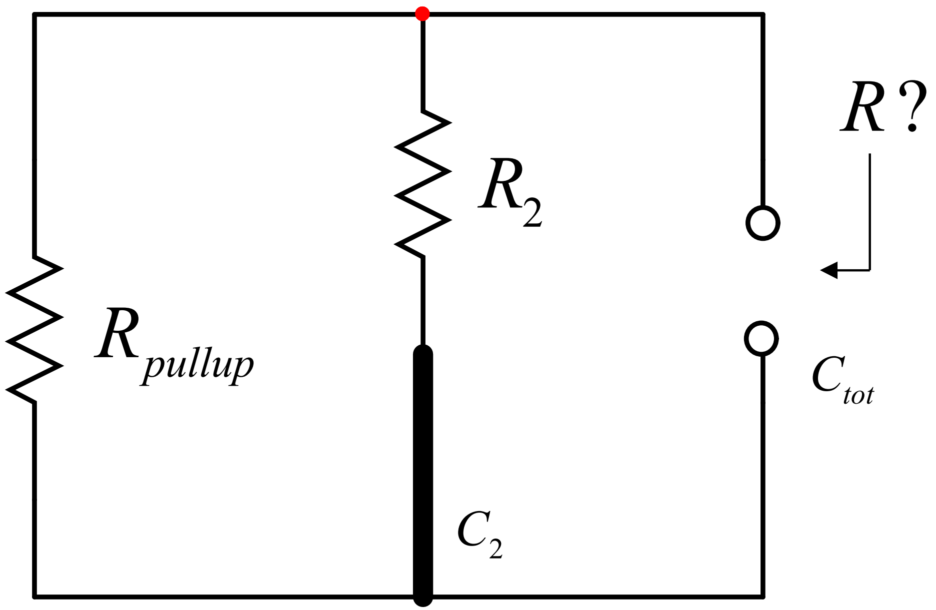
$$a_2 = \tau_1\tau_2^1 = \tau_2\tau_1^2$$

Calculate Higher Order Time Constants

- Set one of the capacitor in its high-frequency state

$\tau_2^1 \longrightarrow$ Put storage element involved in τ_1 in HF

Look at driving resistance of element involved in τ_2



$$\tau_2^1 = C_{tot}\left(R_2 \,||\, R_{pullup}\right)$$

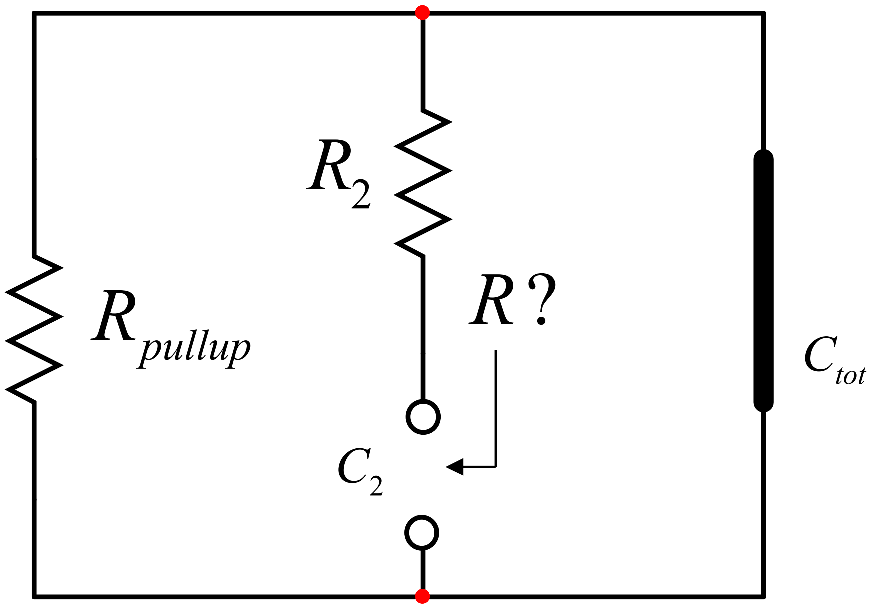
$$a_2 = \tau_1\tau_2^1 = \left(R_{pullup} + R_2\right)C_2\left(R_2 \,||\, R_{pullup}\right)C_{tot}$$

Calculate Higher Order Time Constants

- Check if the second combination brings a simpler result

$\tau_1^2 \longrightarrow$ Put storage element involved in τ_2 in HF

Look at driving resistance of element involved in τ_1



$$\tau_1^2 = R_2 C_2$$

$$a_2 = \tau_2 \tau_1^2 = R_{pullup} C_{tot} R_2 C_2$$

Much simpler definition for a_2

Rewrite Denominator with Cascaded Poles

- Identify variables in the second-order polynomial

$$D(s) = 1 + \frac{s}{\omega_0 Q} + \left(\frac{s}{\omega_0}\right)^2$$

$$D(s) = 1 + a_1 s + a_2 s^2$$

$$a_1 = \frac{1}{\omega_0 Q} \quad a_2 = \frac{1}{\omega_0^{\ 2}}$$

$$\omega_0 = \frac{1}{\sqrt{a_2}} \quad Q = \frac{\sqrt{a_2}}{a_1}$$

- You can apply the low-Q approximation to $D(s)$

$$D(s) = 1 + a_1 s + a_2 s^2 \approx \left(1 + \frac{s}{s_{p_1}}\right)\left(1 + \frac{s}{s_{p_2}}\right)$$

$$\omega_{p_1} = Q\omega_0 = \frac{1}{a_1}$$

$$\omega_{p_2} = \frac{\omega_0}{Q} = \frac{a_1}{a_2}$$

Express Poles and Zeros

- C_{tot} is a few nF while C_2 will be around the µF: $C_{tot} << C_2$

$$\omega_{p_1} = \frac{1}{R_{pullup}C_{tot} + \left(R_{pullup} + R_2\right)C_2} \qquad f_{p_1} = \frac{1}{2\pi\left[R_{pullup}C_{tot} + \left(R_{pullup} + R_2\right)C_2\right]}$$

$$f_{p_1} \approx \frac{1}{2\pi\left(R_{pullup} + R_2\right)C_2} \longrightarrow$$ Low-frequency pole

- The second pole involves the optocoupler capacitor C_{tot}

$$\omega_{p_2} = \frac{R_{pullup}C_{tot} + \left(R_{pullup} + R_2\right)C_2}{R_{pullup}C_{tot}R_2C_2} = \frac{1}{R_2C_2} + \frac{1}{C_{tot}\left(R_2 \parallel R_{pullup}\right)}$$

$$\longrightarrow \quad f_{p_2} = \frac{1}{2\pi}\left[\frac{1}{R_2C_2} + \frac{1}{C_{tot}\left(R_2 \parallel R_{pullup}\right)}\right] \qquad f_{p_2} \approx \frac{1}{2\pi C_{tot}\left(R_2 \parallel R_{pullup}\right)}$$

Assembling Poles and Zeros

- Assemble gain, poles and zeros to form the transfer function

$$V_{FB}(s) = -I_{LED}(s)\,Z_{eq}(s)\,\mathrm{CTR}$$

$$V_{FB}(s) = -\frac{V_{out}(s)\,\mathrm{CTR}}{R_{LED}} R_0 \frac{1+\frac{s}{s_{z_1}}}{\frac{s}{s_{po}}} \frac{1+\frac{s}{s_{z_2}}}{\left(1+\frac{s}{s_{p_1}}\right)\left(1+\frac{s}{s_{p_2}}\right)}$$

$$\Rightarrow \quad G(s) = -\frac{\mathrm{CTR}}{R_{LED}} R_{pullup} \frac{(1+sR_1C_1)(1+sR_2C_2)}{sR_1C_1\left[1+\left(R_{pullup}+R_2\right)C_2\right]\left[1+sC_{tot}\left(R_2 \| R_{pullup}\right)\right]}$$

- Factor sR_1C_1 in the numerator

$$G(s) = -\frac{\mathrm{CTR}}{R_{LED}} R_{pullup} \frac{\cancel{sR_1C_1}}{\cancel{sR_1C_1}} \frac{\left(1+\frac{1}{sR_1C_1}\right)(1+sR_2C_2)}{\left[1+\left(R_{pullup}+R_2\right)C_2\right]\left[1+sC_{tot}\left(R_2 \| R_{pullup}\right)\right]}$$

Simplify Final Transfer Function

❑ Rearrange the low-frequency pole and zero

$$G(s) = -\frac{\text{CTR}}{R_{LED}} R_{pullup} \frac{1+\frac{1}{sR_1C_1}}{1+sC_{tot}\left(R_2 \parallel R_{pullup}\right)} \frac{1+\boxed{sR_2C_2}}{1+\boxed{s\left(R_{pullup}+R_2\right)C_2}}$$

Factor

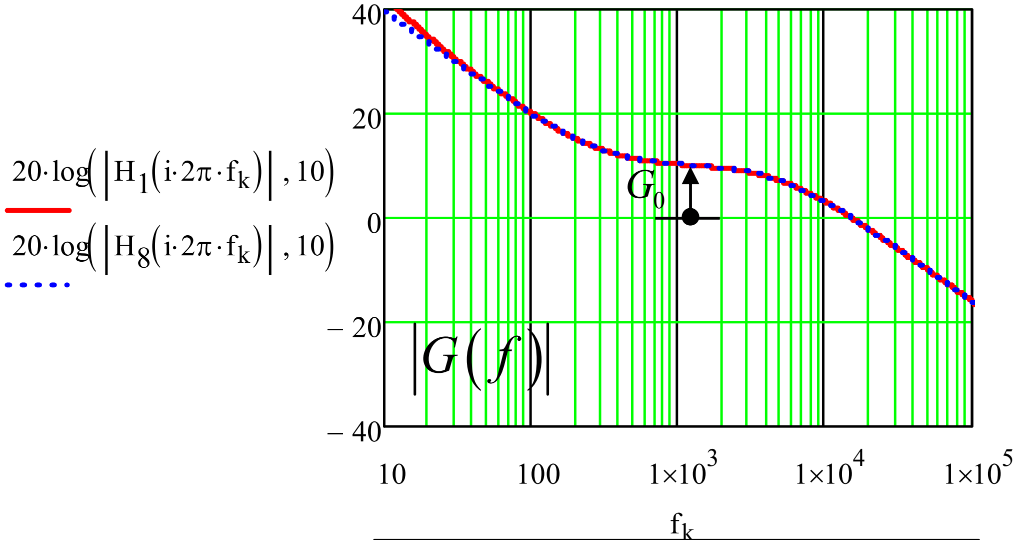
$$G(s) = -\frac{\text{CTR}}{R_{LED}} R_{pullup} \frac{\cancel{s}R_2\cancel{C_2}}{\cancel{s}\left(R_{pullup}+R_2\right)\cancel{C_2}} \frac{1+\frac{1}{sR_1C_1}}{1+sC_{tot}\left(R_2 \parallel R_{pullup}\right)} \frac{1+\frac{1}{sR_2C_2}}{1+\frac{1}{s\left(R_{pullup}+R_2\right)C_2}}$$

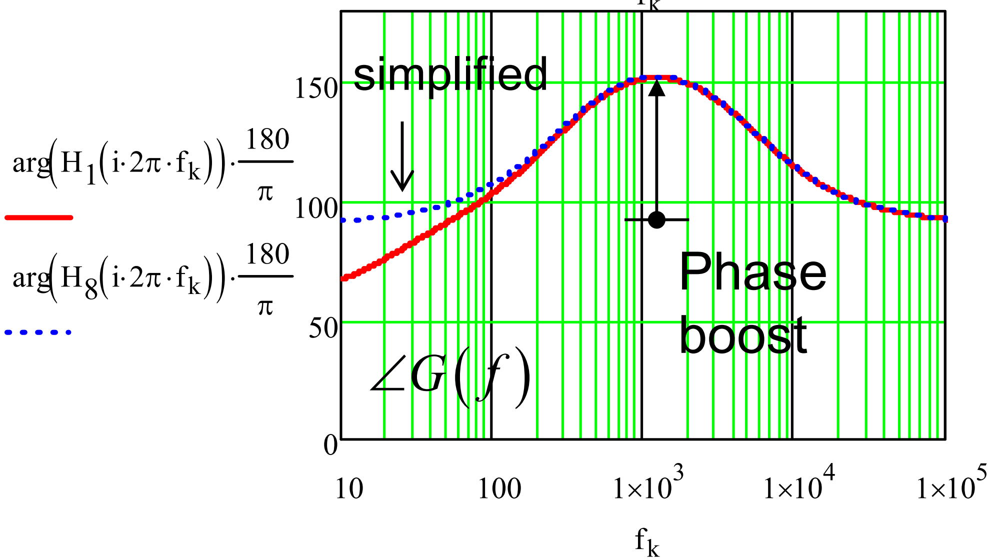
$$\frac{1+\frac{1}{sR_2C_2}}{1+\frac{1}{s\left(R_{pullup}+R_2\right)C_2}} \approx 1 \quad \Rightarrow \quad G(s) = -\boxed{\frac{R_{pullup} \parallel R_2}{R_{LED}}\text{CTR}} \frac{1+\frac{1}{sR_1C_1}}{1+sC_{tot}\left(R_2 \parallel R_{pullup}\right)}$$

New mid-band gain

Validating the Transfer Function

- Compare both expressions with Mathcad®





Full formula

$$G(s)=-\frac{\mathrm{CTR}}{R_{LED}}R_{pullup}\frac{(1+sR_1C_1)(1+sR_2C_2)}{sR_1C_1\left[1+(R_{pullup}+R_2)C_2\right]\left[1+sC_{tot}(R_2\,\|\,R_{pullup})\right]}$$

Simplified formula

$$G(s)=-\frac{R_{pullup}\,\|\,R_2}{R_{LED}}\mathrm{CTR}\frac{1+\frac{1}{sR_1C_1}}{1+sC_{tot}(R_2\,\|\,R_{pullup})}$$

SPICE Helps Validate Calculations

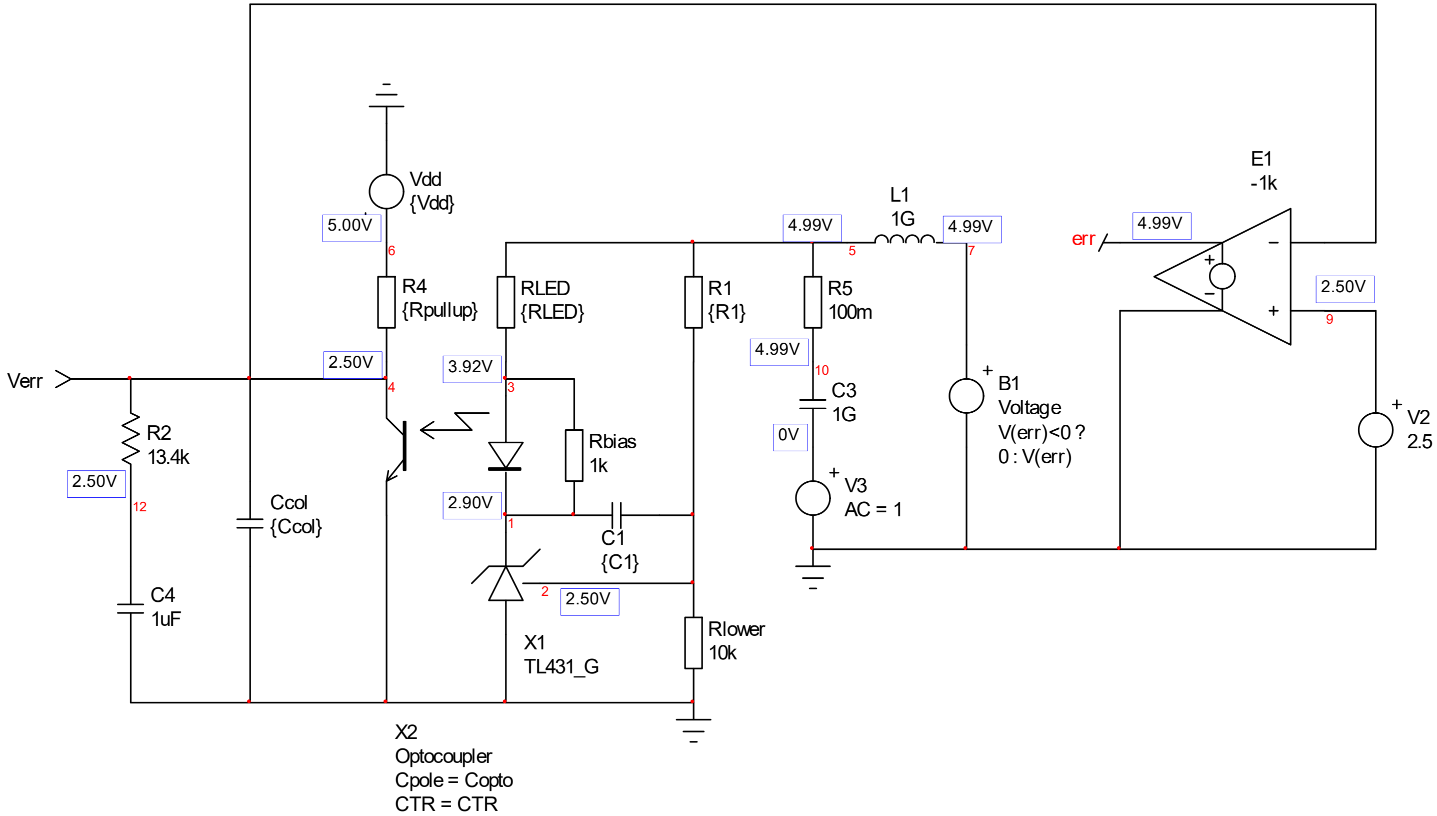
❑ A SPICE simulation gives the response in a second

parameters

Vout=5
R1=(Vout-2.5)/250u
fc=1k
Gfc=-10

Vref=2.5
VCEsat=300m
Vf=1
Ibias=1m
Vdd=5
G=10^(-Gfc/20)
pi=3.14159
Fzero=300
Fpole=2k
A=Vout-Vf-Vref
B=Vdd-VCEsat+Ibias*CTR*Rpullup
Rmax=(A/B)*Rpullup*CTR
Rpullup=20k
RLEDa=CTR*Rpullup/G
RLED=750
C1=1/(2*pi*Fzero*R1)
C2=1/(2*pi*Fpole*Rpullup)
Ccol=C2-Copto
Fopto=6k
Copto=1/(2*pi*Fopto*Rpullup)
CTR = 0.3

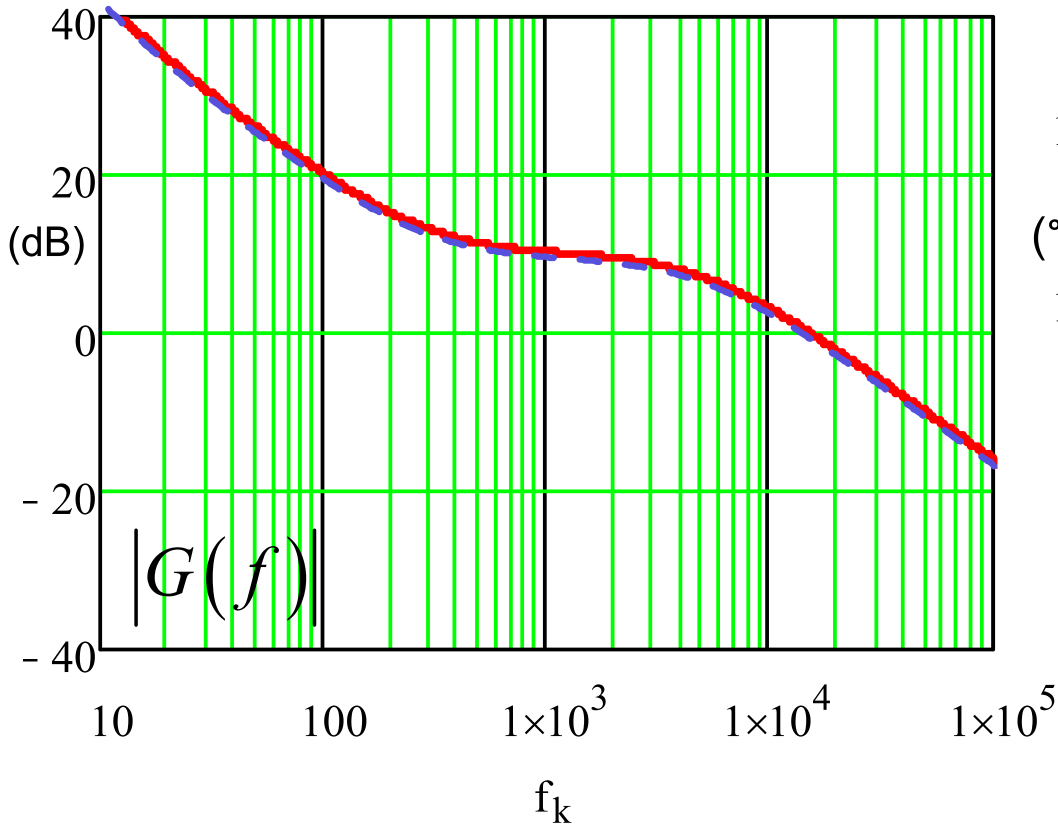
Rx=G*Rpullup*Rled/(Rpullup*CTR-G*Rled)

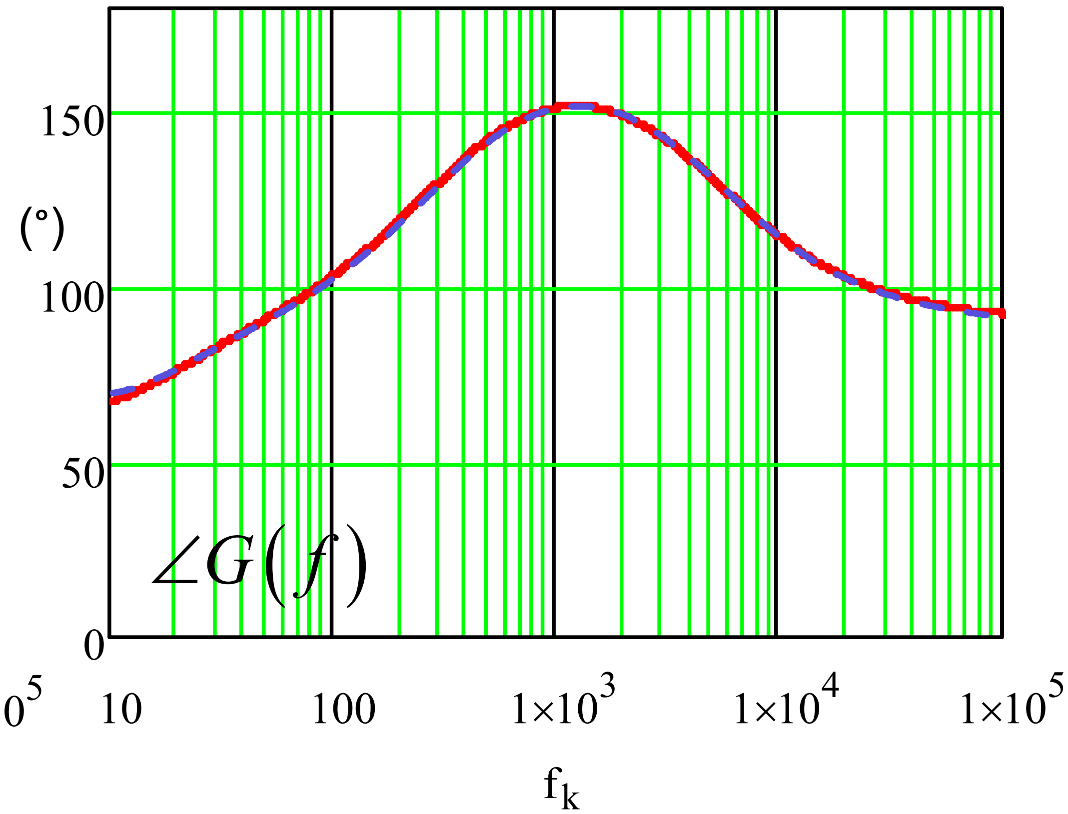


5-V design

Comparing SPICE and Mathcad®

- ❑ The simulation matches analytical results

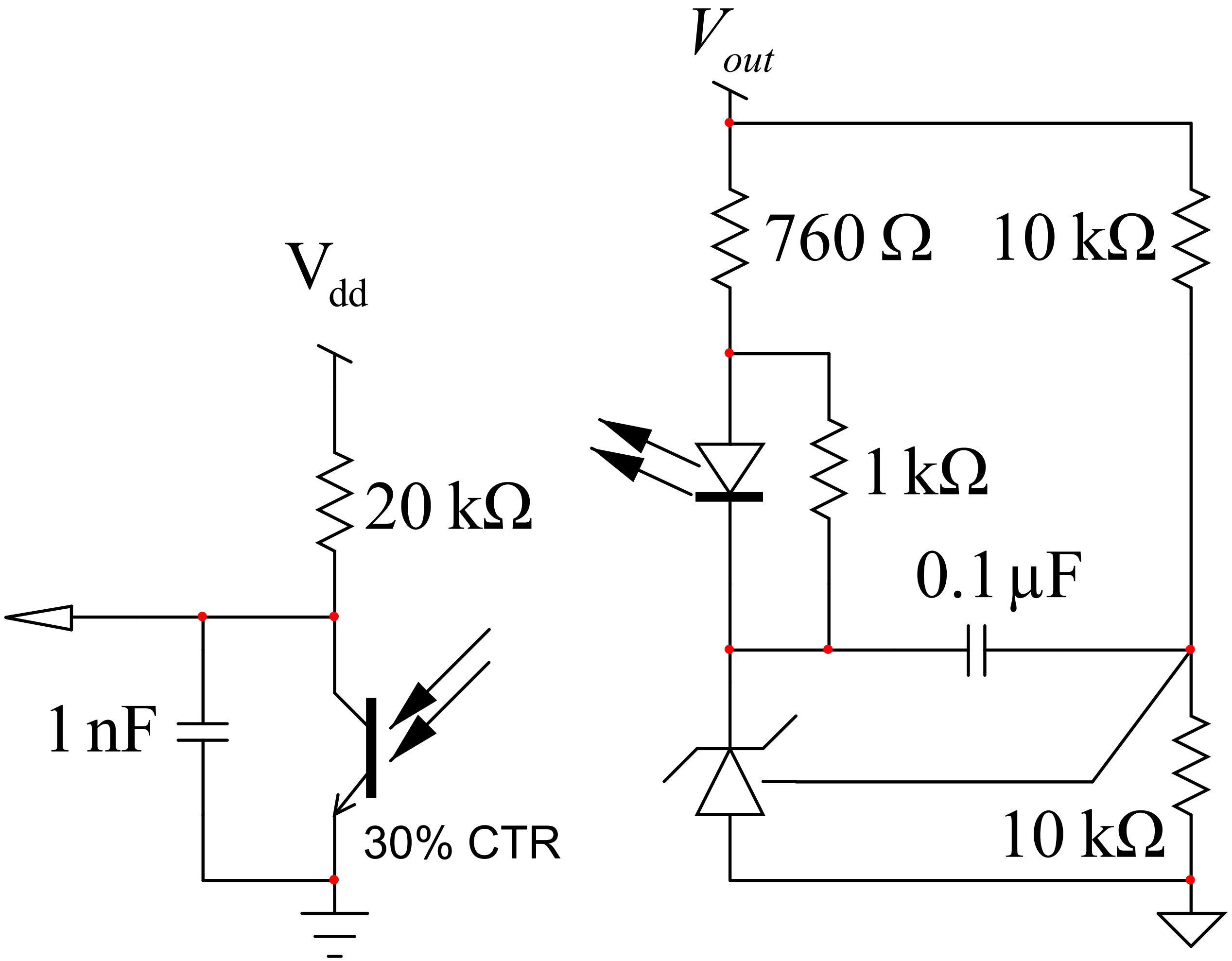




- ❑ Let's exercise these formulas in a practical example

A Typical Design Case

- ❑ Assume a 5-V power supply, crossover requires a 10-dB gain
- ❑ R_{LED} max is computed to 840 Ω



Minimum gain is

$$G_0 = \mathrm{CTR}\frac{R_{pullup}}{R_{LED}}$$

$$= 0.3 \times \frac{20}{0.840}$$

$$= 7.2 \text{ or } \approx \boxed{17\text{ dB}}$$

Reducing the Mid-Band Gain

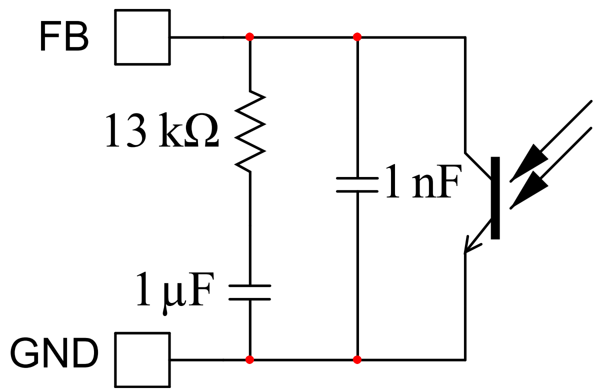
- The LED resistor is selected with margins to satisfy bias
- The extra resistor is computed based on the required gain

$$G_0 = \frac{R_{pullup} \parallel R_2}{R_{LED}} \mathrm{CTR} \longrightarrow R_2 = \frac{G_0 R_{LED} R_{pullup}}{\mathrm{CTR} \cdot R_{pullup} - G_0 R_{LED}}$$

- A 10-dB gain means G_0 equals 3.16

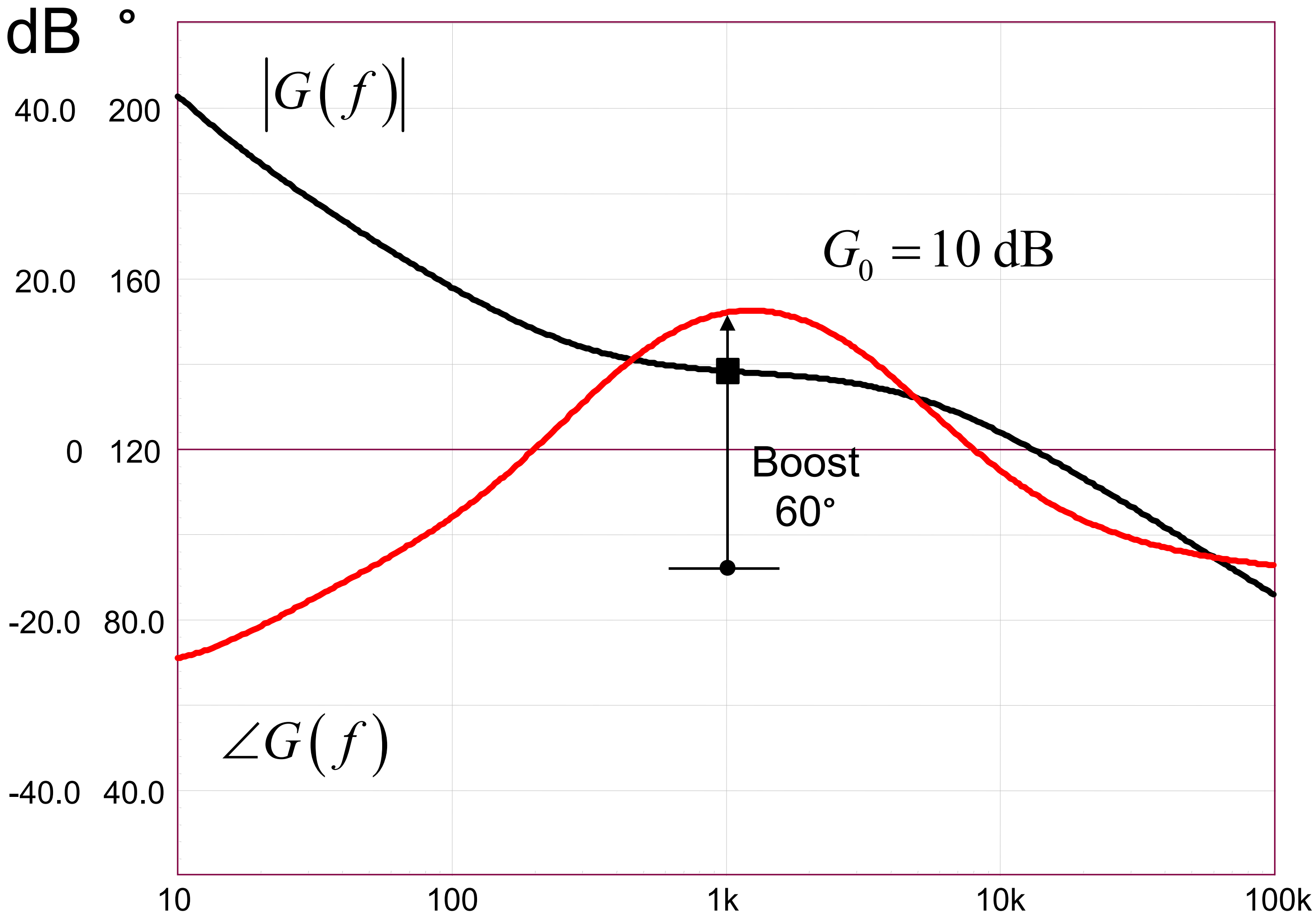
$$R_2 = \frac{3.16 \times 760 \times 20k}{0.3 \cdot 20k - 3.16 \times 760} = 13.3\ \mathrm{k\Omega}$$

- Select a high-value capacitor for C_2, 1 µF



SPICE Simulation of Compensator

- Simulation results confirm calculations



Conclusion

- It is important to consider non-ideal components in studies
- Models show the impact of op amps imperfections
- Simulation helps assess contribution of parasitics
- Optocoupler is key in isolated designs and must be characterized
- The fast lane affects optocoupler-based return paths
- Dc-dc bricks control require soft-start from the secondary side
- Analysis, simulation and bench are recipes for success!