ON Semiconductor®

Dc-dc converters feedback and control

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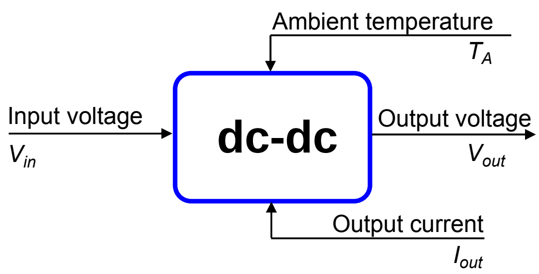
Course agenda

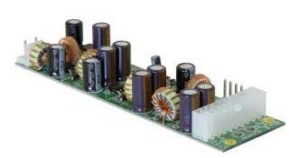
- ❑ Feedback generalities
- ❑ Building an oscillator
- ❑ Poles and zeros
- ❑ Phase margin and quality coefficient
- ❑ Undershoot and crossover frequency
- ❑ Compensating the converter
- ❑ Current-mode converters
- ❑ Automated pole-zero placement
- ❑ Manual pole-zero placement
- ❑ Compensating with a TL431
- ❑ Watch the optocoupler!
- ❑ Input filter
- ❑ A real case example
- ❑ Conclusion

What do we expect from a dc-dc?

- A stable output voltage, whatever loading, input, temperature and aging conditions.
- A fast reaction to a incoming perturbation such as a load transient or an input voltage change.
- A quick settling time when starting-up or recovering from a transient state.

→ A stable and noiseless dc source we can trust!

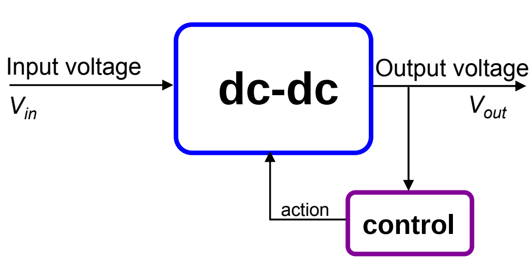


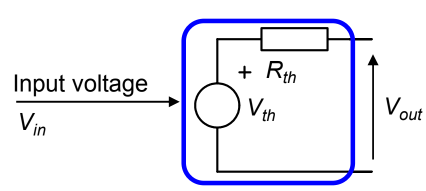


What is feedback?

- A target is assigned to one or several state-variables, e.g. V_{out} = 12 V.
- A dedicated circuit monitors V_{out} deviations.
- If V_{out} deviates from its target, an error is created and fed-back to the power stage for action.
- The action is a change in the control variable: duty ratio (VM), peak current (CM) or switching frequency.

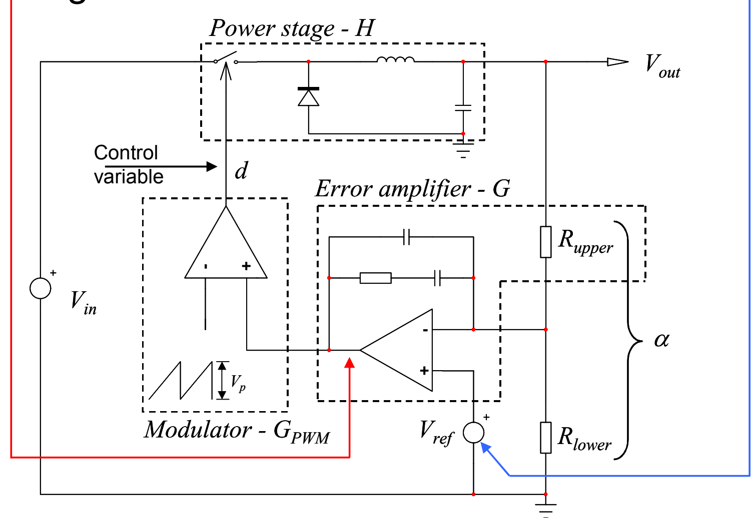
→ Compensating for the converter shortcomings!





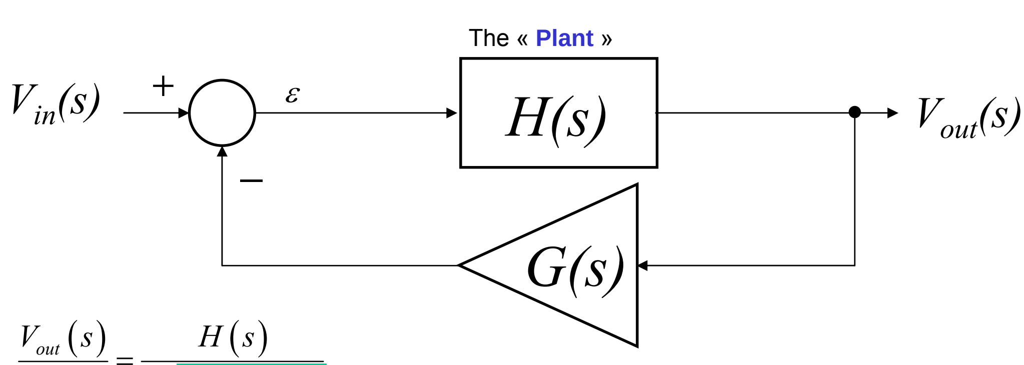
The feedback implementation

- V_{out} is permanently compared to a reference voltage V_{ref}.
- The reference voltage V_{ref} is precise and stable over temperature.
- The error, $\varepsilon = V_{ref} - \alpha V_{out}$, is amplified and sent to the control input.
- The power stage reacts to reduce ε as much as it can.



Positive or negative feedback?

- Do we want to build an oscillator?



$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{H(s)}{1 + \boxed{H(s)G(s)}}$$ → Open-loop gain *T(s)*

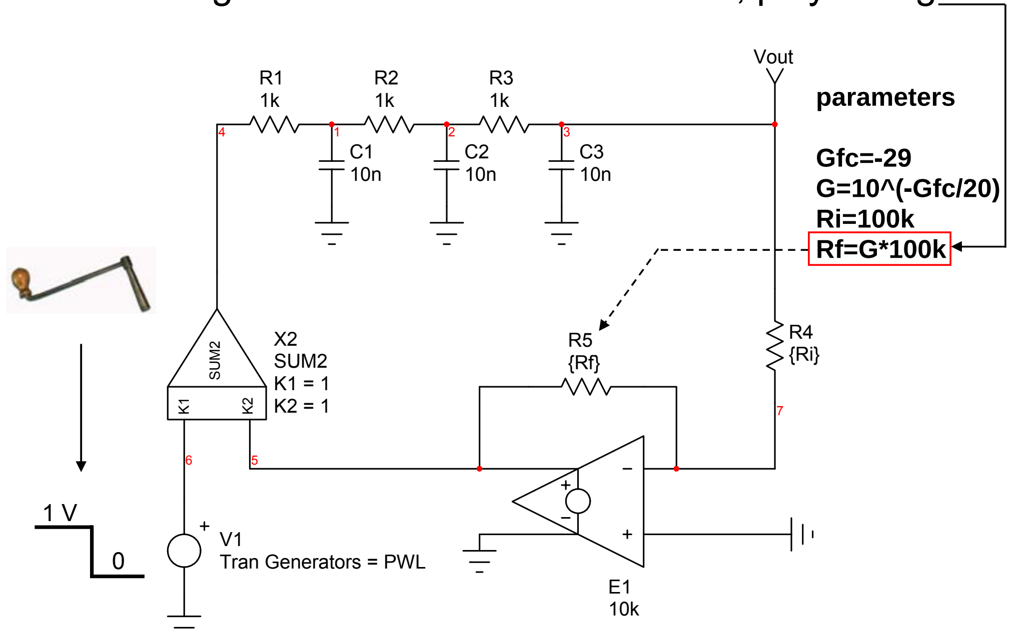
$$V_{out}(s) = \lim_{V_{in}(s)\to 0}\left[\frac{H(s)}{1+G(s)H(s)} V_{in}(s)\right]$$

To sustain self-oscillations, as $V_{in}(s)$ goes to zero, quotient must go infinite

$$1 + G(s)H(s) = 0 \longrightarrow \begin{cases} |G(s)H(s)| = 1 \\ \angle G(s)H(s) = -180° \end{cases} \longrightarrow \underset{-1,\, j0}{\overset{Nyquist}{+}}$$

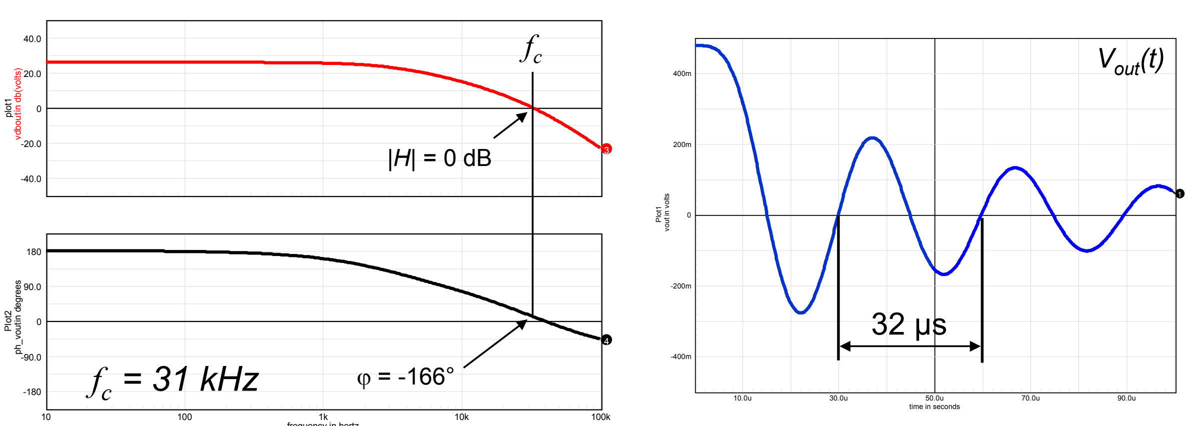
A constant gain with a 180° rotation

- ❑ Starting the oscillator with a « crank », play with gain



A simple oscillator

- ❑ Case n°1, at 0-dB gain crossover, phase φ is below -180°
- ❑ Oscillations are damped, system is asymptotically stable



- ❑ The oscillation frequency is the crossover frequency f_c
- ❑ The poles have an imaginary and a *negative* real part (LHPP)

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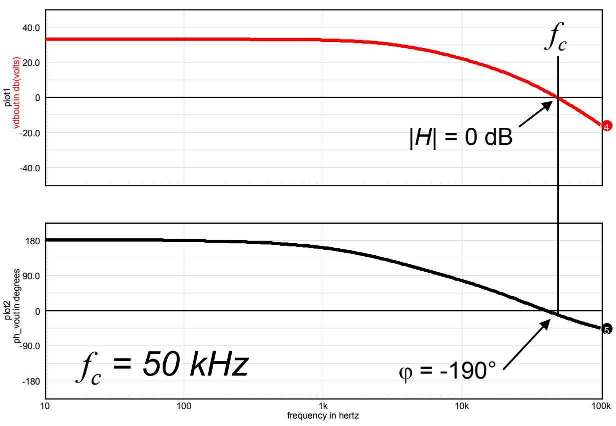
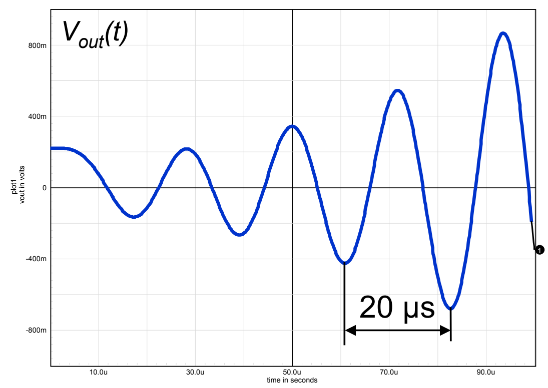
A simple oscillator

- ❑ Case n°2, at 0-dB gain crossover, φ is beyond -180°
- ❑ Oscillations are not damped, system diverges





- ❑ The poles have an imaginary and a *positive* real part (RHPP)

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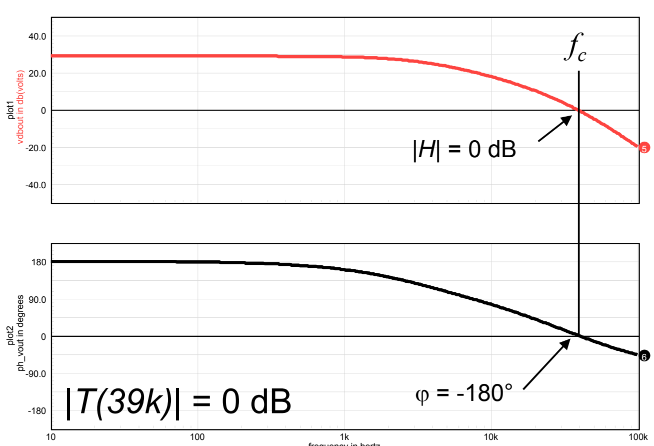
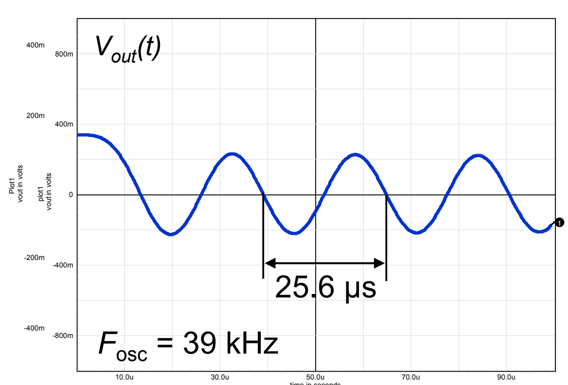
A simple oscillator

- ❑ Case n°3, at 0-dB gain crossover, φ is exactly -180°
- ❑ Oscillations are sustained, we have an oscillator!

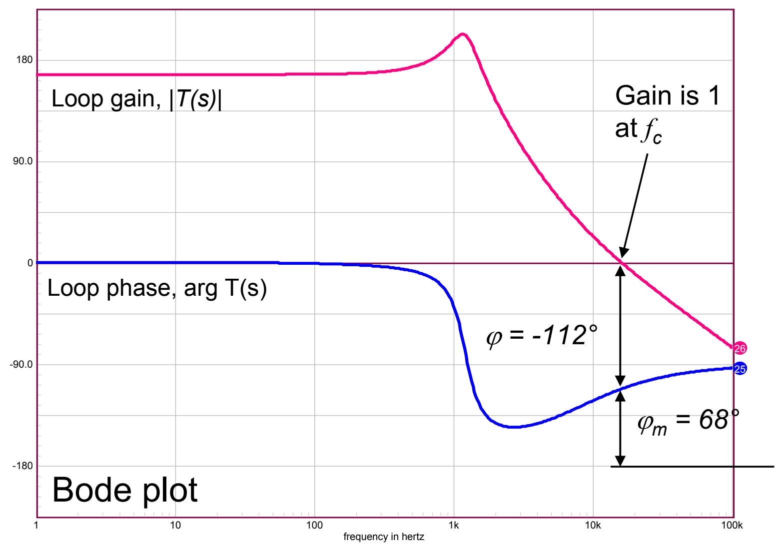




- ❑ The poles are pure imaginary: no damping

Conditions for steady-state stability

- ❑ We do not want to create an oscillator!
- ❑ Conditions for non-permanent oscillations are:
- ➢ total phase rotation less than -360° at the crossover point



Total phase delay at f_c:

-112° *H(s)* power stage
-180° *G(s)* opamp

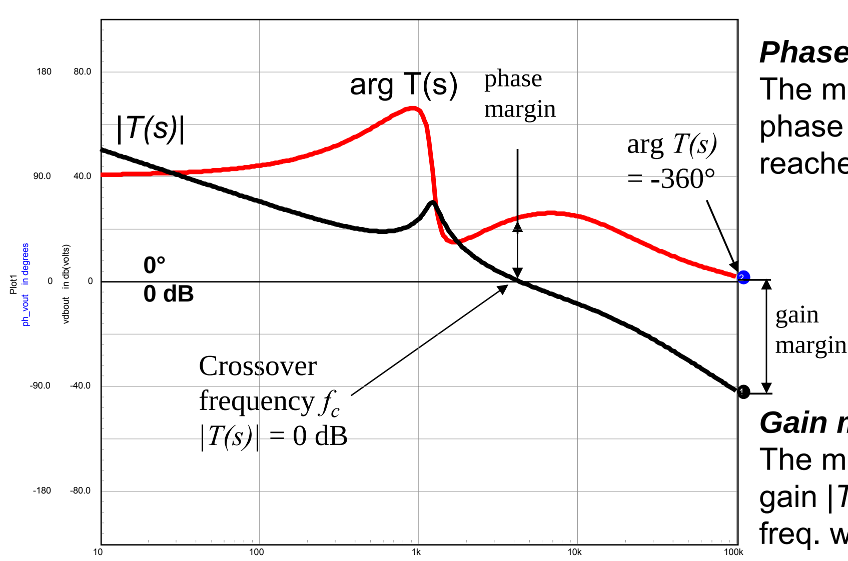
total = $\boxed{-292°}$

Stable!

H.W. Bode – 1905-1982

The need for phase margin

- ❑ we need phase margin when |*T(s)*| = 0 dB
- ❑ we need gain margin when arg *T(s)* = -360°



Phase margin:
The margin before the loop phase rotation arg *T(s)* reaches -360° at |*T(s)*| = 0 dB

Gain margin:
The margin before the loop gain |*T(s)*| reaches 0 dB at a freq. where arg *T(s)* = -360°

Poles, zeros and s-plane

- A plant loop gain is defined by:

$$H(s)=\frac{N(s)}{D(s)}\begin{array}{l}\longrightarrow \text{numerator}\\ \longrightarrow \text{denominator}\end{array}$$

- solving for *N(s)* = 0, the roots are called the **zeros**
- solving for *D(s)* = 0, the roots are called the **poles**

$$H(s)=\frac{(s+5k)(s+30k)}{s+1k}$$

Numerator roots

$$\boxed{\begin{array}{l} s_{z_1}=-5k \\ s_{z_2}=-30k \end{array}}$$

$$\boxed{s_{p_1}=-1k}$$

Denominator root

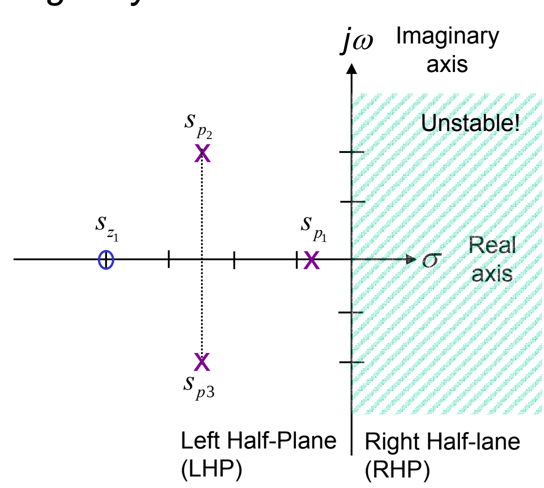
$$\longrightarrow \quad f_{z_1}=\frac{5k}{2\pi}=796\ Hz \qquad f_{z_2}=\frac{30k}{2\pi}=4.77\ kHz \qquad f_{p_1}=\frac{1k}{2\pi}=159\ Hz$$

Poles, zeros and s-plane

- The roots can either be *real* or *imaginary*:

$$H(s)=\frac{s+4}{(s+0.8)\left[\underbrace{(s+2.5)^2+4}_{\pm 2j}\right]}$$

$$s_{z_1}=-4$$
$$s_{p_1}=-0.8$$
$$\left.\begin{array}{l} s_{p_2}=-2.5+2j \\ s_{p_3}=-2.5-2j \end{array}\right\}\ \text{Conjugated roots}$$



- We can place these roots in the imaginary plane
  - ➢ root-locus analysis in the *s*-plane
  - ➢ their positions in the *s*-plane affect the time domain response

Poles, zeros and s-plane

- ❑ How the poles do influence the time domain response of the plant?
- ➢ assume an input-step response is wanted:
- ✓ multiply *H(s)* by 1/s
- ✓ take the inverse Laplace transform
- ✓ plot the response

$$H(s)=\frac{2}{(s+1)(s+2)} \xrightarrow{\times\frac{1}{s}} H(s)\frac{1}{s}=\frac{1}{s}\frac{2}{(s+1)(s+2)}$$

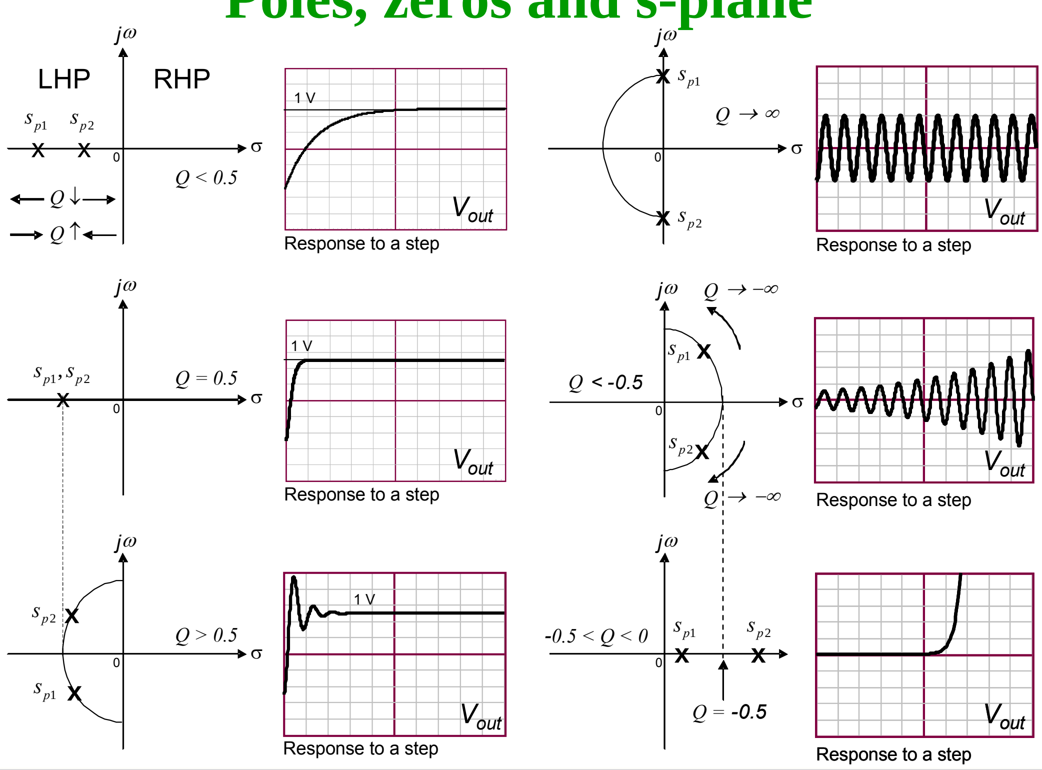
$$s_{p_1}=-1 \qquad s_{p_2}=-2$$

Take inverse Laplace transform

$$\mathcal{L}^{-1}\{H(s)\frac{1}{s}\}=0.5-e^{\boxed{-1}t}+0.5e^{\boxed{-2}t}$$

- ❑ The roots are the exponentials exponents: -1 and -2
- ❑ If the denominator roots are negative, the signal is decaying
- ❑ If the denominator roots are positive, the response diverges

Poles, zeros and s-plane



Poles, zeros and s-plane

- the pole magnitude at the cutoff frequency is -3 dB
- Its asymptotic phase at $f = \infty$ is -90°
- The pole "lags" the phase

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{1}{1+sRC} = \frac{1}{1+s/\omega_0} \longrightarrow \left|\frac{V_{out}(\omega_0)}{V_{in}(\omega_0)}\right| = \frac{1}{\sqrt{1+[\omega_0/\omega_0]^2}} = \boxed{\frac{1}{\sqrt{2}}}$$

$$\arg\frac{V_{out}(\infty)}{V_{in}(\infty)} = \arg(1) - \arg\left(1+\frac{\infty}{s_{p_1}}\right) = -\arctan(\infty) = -\frac{\pi}{2} = -90°$$ At $f = \infty$

- the zero magnitude at the cuttoff frequency is +3 dB
- Its asymptotic phase at $f = \infty$ is +90°
- The zero "boosts" the phase

$$\frac{V_{out}(s)}{V_{in}(s)} = 1+\frac{s}{\omega_0} \longrightarrow \left|\frac{V_{out}(\omega_0)}{V_{in}(\omega_0)}\right| = \sqrt{1+[\omega_0/\omega_0]^2} = \boxed{\sqrt{2}}$$

$$\arg\frac{V_{out}(\infty)}{V_{in}(\infty)} = \arg\left(1+\frac{\infty}{s_{p_1}}\right) = \arctan(\infty) = +\frac{\pi}{2} = 90°$$ At $f = \infty$

Poles, zeros and s-plane

- Poles and zeros can sometimes appear "at the origin"

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{\frac{s}{\omega_0}}{D(s)}$$ Zero for s = 0: zero at the origin $\longrightarrow$ As f increases the gain increases with a +1 slope (+20 dB/decade)

$$\arg\frac{V_{out}(s)}{V_{in}(s)} = \arg\left(\frac{\frac{s}{s_{zo}}}{0}\right) = \arctan(\infty) = +\frac{\pi}{2}$$ For $f > f_{zo}$

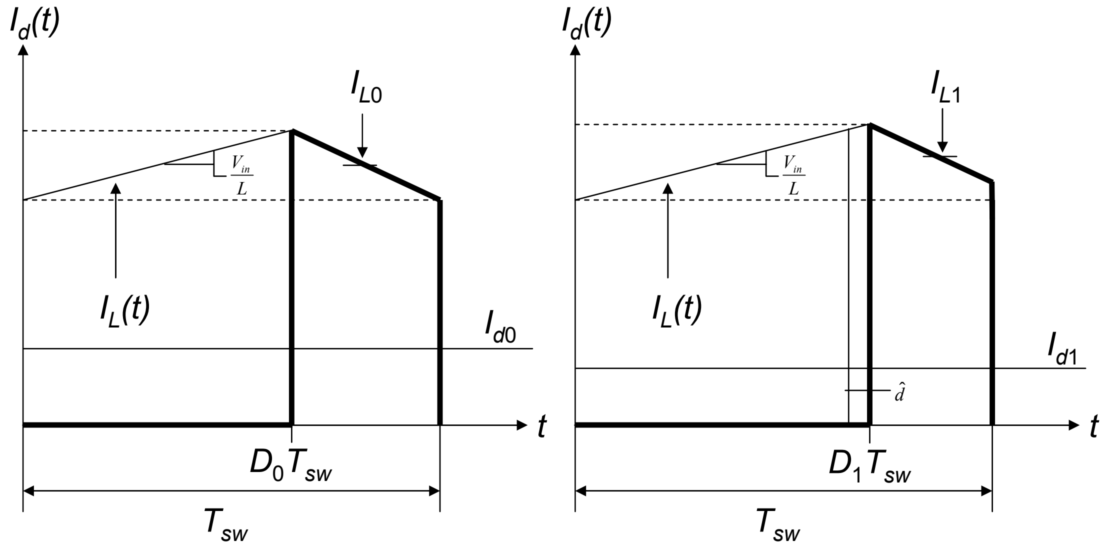
$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{N(s)}{\frac{s}{\omega_0}}$$ Pole for s = 0: pole at the origin $\longrightarrow$ As f increases the gain decreases with a -1 slope (-20 dB/decade)

$$\arg\frac{V_{out}(s)}{V_{in}(s)} = \arg(1) - \arg\left(\frac{\frac{s}{s_{po}}}{0}\right) = -\arctan(\infty) = -\frac{\pi}{2}$$ For $f > f_{po}$

A pole at the origin introduces a fixed phase rotation of -90°

The Right Half-Plane Zero

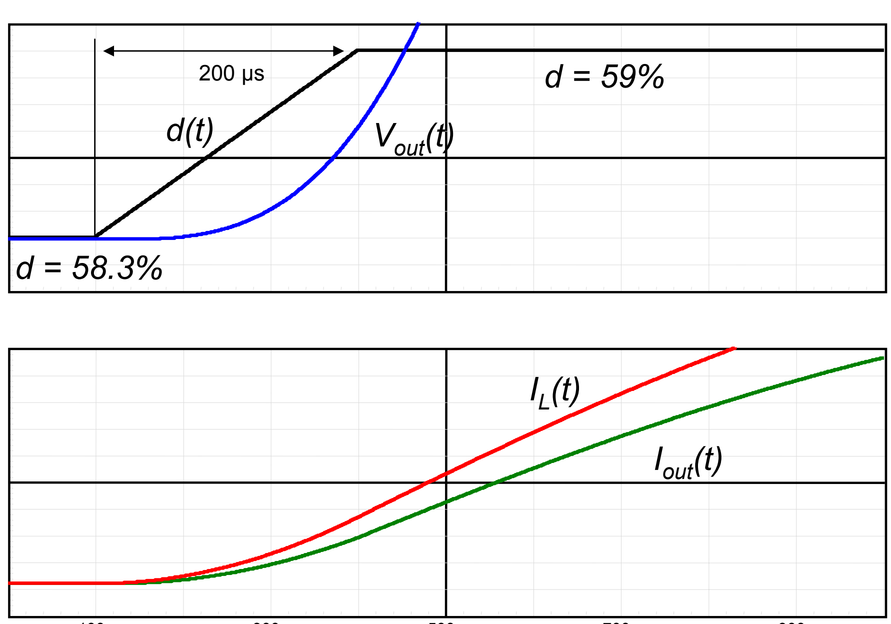
- In a CCM boost, I_{out} is delivered during the off time: $I_{out} = I_d = I_L(1-D)$



- If D brutally increases, D' reduces and I_{out} drops!
- What matters is the inductor current slew-rate $\longrightarrow \frac{d\langle V_L\rangle(t)}{dt}$

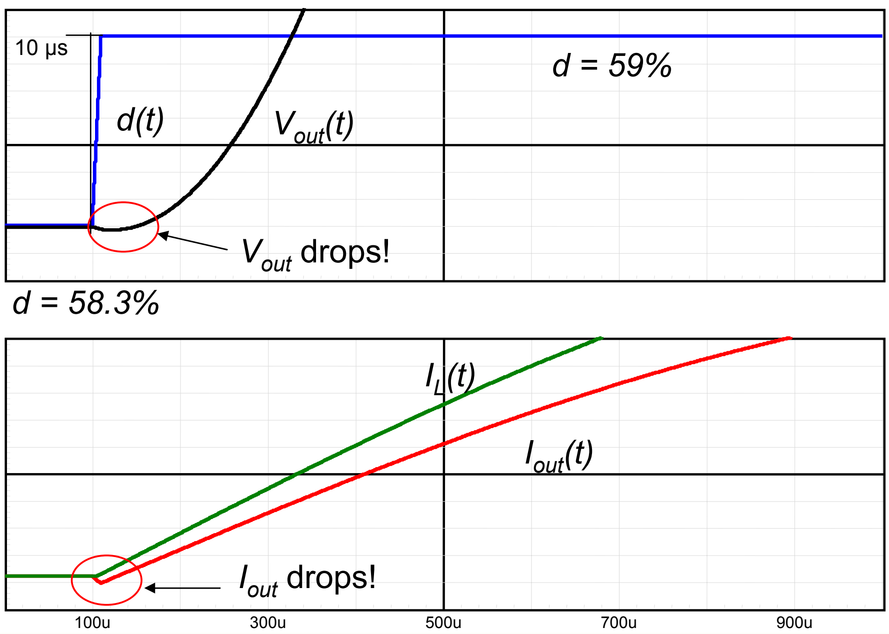
The Right Half-Plane Zero

- If $I_L(t)$ can rapidly change, I_{out} increases when D goes up



The Right Half-Plane Zero

- If $I_L(t)$ is limited because of a big L, I_{out} drops when D increases



The Right Half-Plane Zero

- Small-signal equations can help us to formalize it

$$I_{out} = I_L(1-D)$$

$$\hat{i}_{out} = \left(\frac{\partial I_{out}}{\partial I_L}\hat{i}_L\right)_D + \left(\frac{\partial I_{out}}{\partial D}\hat{d}\right)_{I_L} = \hat{i}_L(1-D) - \hat{d}I_L$$

The negative sign indicates a positive root!

$$\frac{\hat{i}_{out}(s)}{\hat{d}(s)} = \frac{V_{out}D'}{sL}\left(1-\frac{sL}{D'^2 R_{load}}\right) = \frac{\left(1\boxed{-}\frac{s}{\omega_{z_2}}\right)}{\frac{s}{\omega_0}} \qquad \omega_0 = \frac{V_{out}D'}{L} \qquad \omega_{z_2} = \frac{R_{load}D'^2}{L}$$

Voltage mode

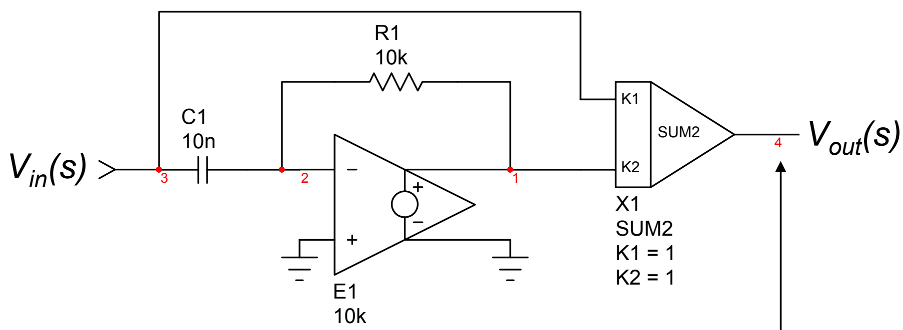
$$\frac{\hat{i}_{out}(s)}{\hat{v}_c(s)} = \frac{D'}{R_{sense}} - \frac{sL}{D'R_{load}R_{sense}} = G_0\left(1-\frac{s}{\omega_{z_2}}\right) \qquad \omega_{z_2} = \frac{R_{load}D'^2}{L}$$

Current mode

- Voltage mode or current mode, the RHPZ remains the same

The Right Half-Plane Zero

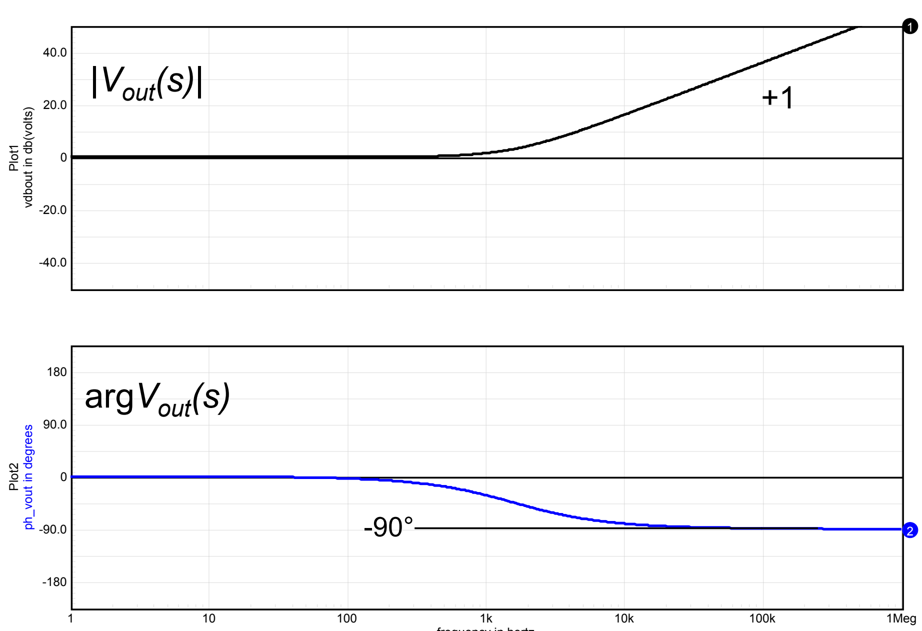
- ❑ To limit the effects of the RHPZ, limit the duty ratio slew-rate
- ❑ Chose a cross over frequency equal to 20-30% of RHPZ position
- ▪ A simple RHPZ can be easily simulated:



$$V(s) = V_{in}(s) - V_{in}(s)\frac{R_1}{\frac{1}{sC_1}} = V_{in}(s)\left(1 - \frac{s}{\omega_0}\right)$$

The Right-Half-Plane-Zero

- ❑ With a RHPZ we have a boost in gain but a lag in phase!



LHPZ

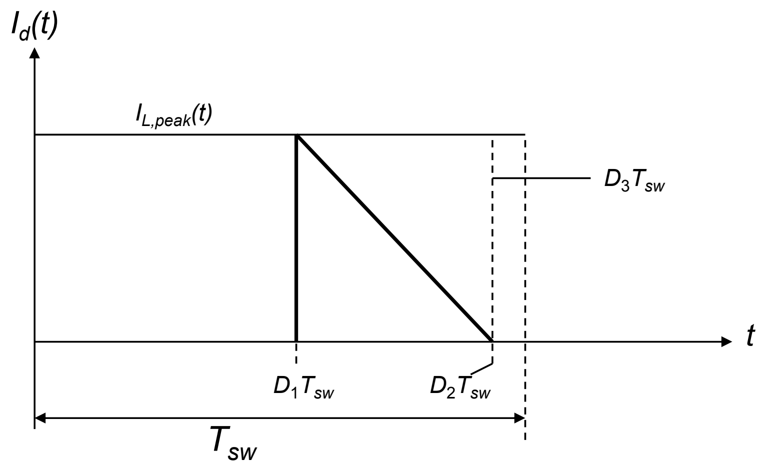
$$G(s) = 1 + \frac{s}{\omega_0}$$

RHPZ

$$G(s) = 1 \boxed{-} \frac{s}{\omega_0}$$

The Right-Half-Plane-Zero

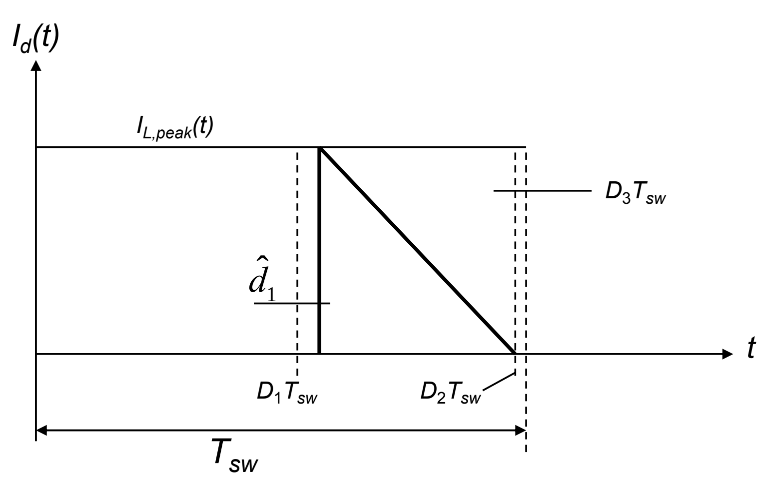
- A RHPZ also exists in DCM boost, buck-boost converters…



- When D_1 increases, $[D_1,D_2]$ stays constant but D_3 shrinks

The Right-Half-Plane-Zero

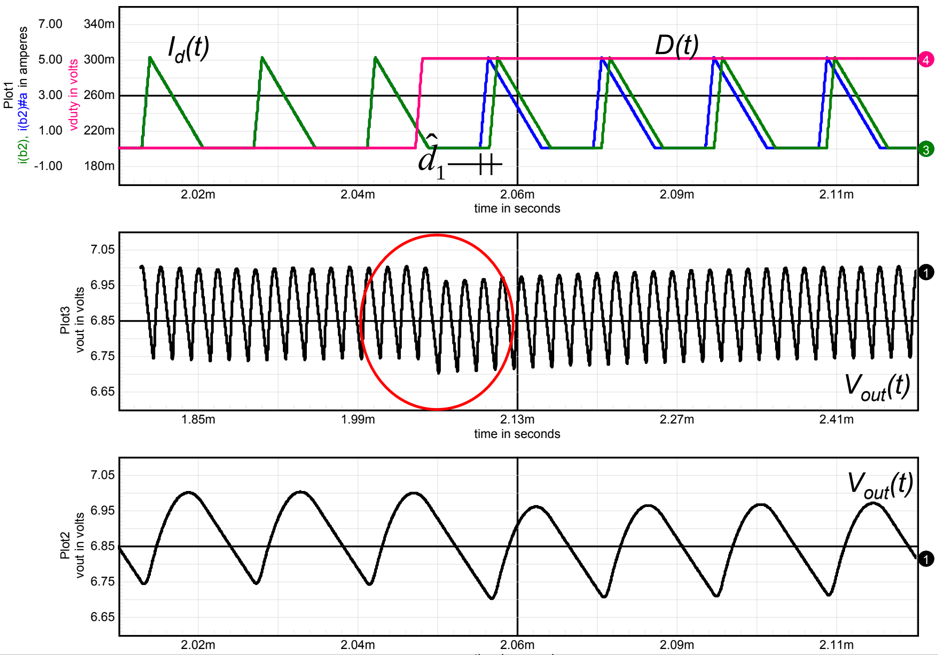
- The triangle is simply shifted to the right by $\hat{d}_1$



- The refueling time of the capacitor is delayed and a drop occurs

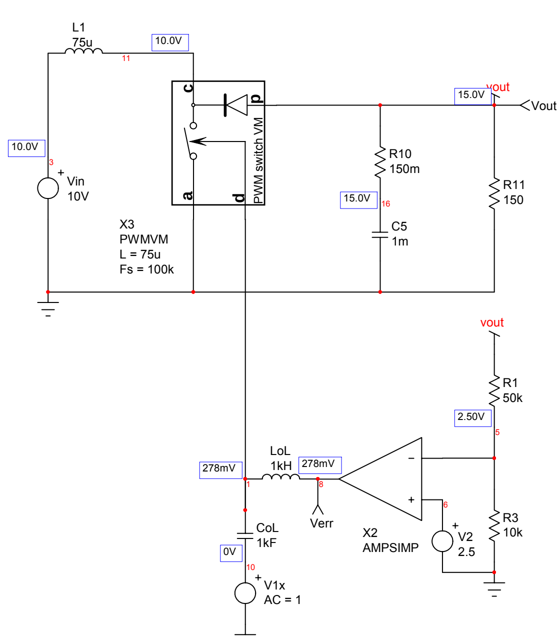
The Right-Half-Plane-Zero

- If *D* increases, the diode current is delayed by $\hat{d}_1$



The Right-Half-Plane-Zero

- Averaged models can predict the DCM RHPZ



$$\frac{\hat{v}_{out}(s)}{\hat{d}(s)} = H_d \frac{\left(1+s/s_{z_1}\right)\left(1-s/s_{z2}\right)}{\left(1+s/s_{p_1}\right)\left(1+s/s_{p2}\right)}$$

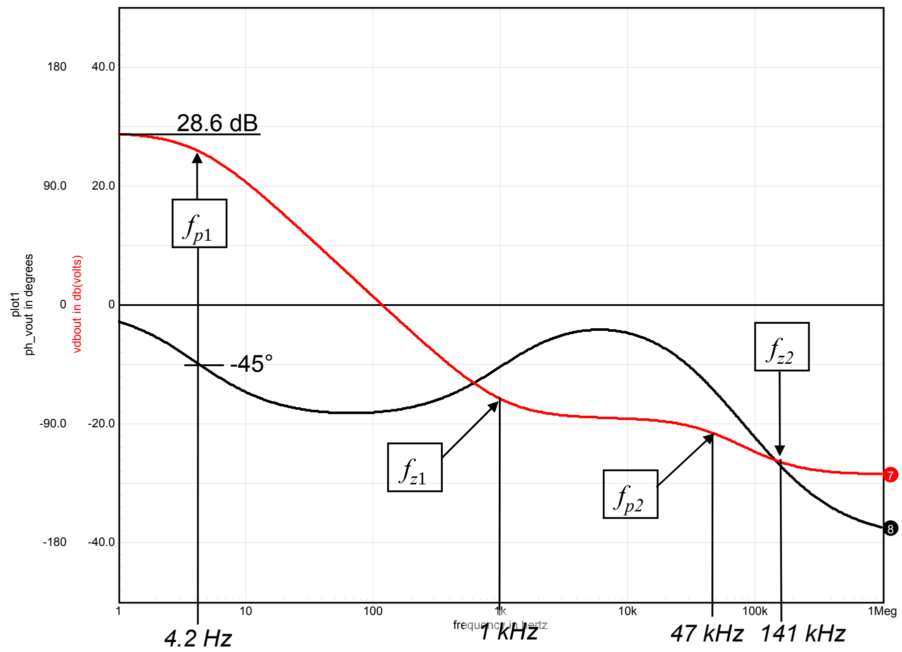
$$s_{z_1} = \frac{1}{C_{out}R_{ESR}} \qquad \boxed{s_{z_2} = \frac{R_{load}}{M^2 L}}$$

$$s_{p_1} = \frac{2M-1}{M-1}\frac{1}{C_{out}R_{ESR}} \qquad \boxed{s_{p_2} = 2F_{sw}\left(\frac{1-1/M}{D}\right)^2}$$

$$H_d = \frac{2V_{out}}{D}\frac{M-1}{2M-1}$$

The Right-Half-Plane-Zero

❑ Averaged models can predict the DCM RHPZ



$$H_d = 28.75\ dB$$

$$f_{z_1} = 1.06\ kHz$$

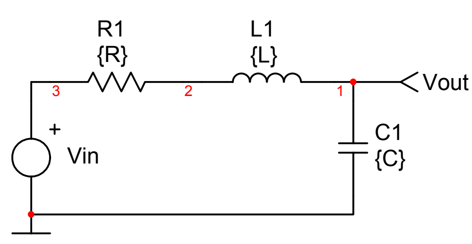
$$f_{z_2} = 141\ kHz$$

$$f_{p_1} = 4.2\ Hz$$

$$f_{p_2} = 47.1\ kHz$$

How much margin? The *RLC* filter

❑ let us study an *RLC* low-pass filter, a 2nd order system



$$T(s) = \frac{1}{LCs^2 + RCs + 1}$$

$$T(s) = \frac{1}{\frac{s^2}{\omega_r^2} + 2\zeta\frac{s}{\omega_r} + 1} = \frac{1}{\frac{s^2}{\omega_r^2} + \frac{s}{\omega_r Q} + 1}$$

$$\omega_r = \frac{1}{\sqrt{LC}}$$

$$\boxed{\zeta} = R\sqrt{\frac{C}{4L}} \qquad Q = \frac{1}{2\zeta}$$

zeta

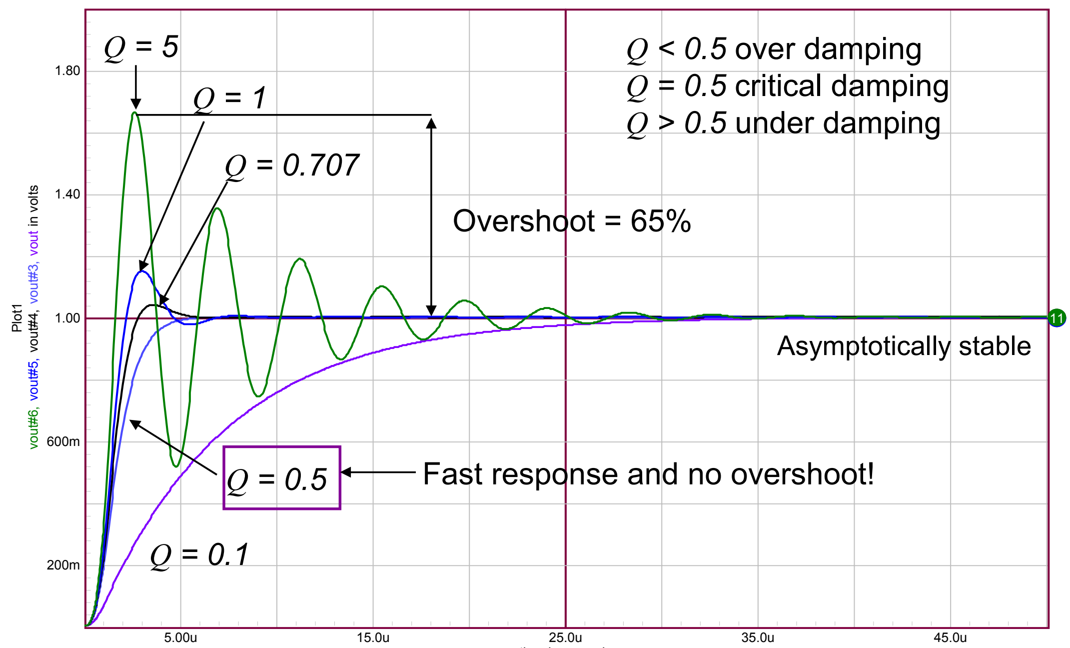
parameters

f0=235k
L=10u
C=1/(4*3.14159^2*f0^2*L)
w0=({L}*{C})^-0.5
Q=10
R=1/((({C}/(4*{L}))^0.5)*2*{Q})

Change Q and run the simulation

The *RLC* response to an input step

- changing Q affects the transient response



The *RLC* response to an input step

- Q affects the poles position

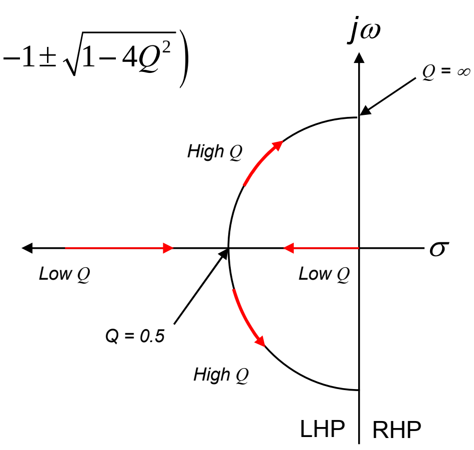
$$T(s) = \frac{1}{\frac{s^2}{\omega_r^2} + \frac{s}{\omega_r Q} + 1} = \frac{N(s)}{D(s)}$$

$\longrightarrow$ Solve $N(s)$ = 0 to obtain zeros

$\longrightarrow$ Solve $D(s)$ = 0 to obtain poles

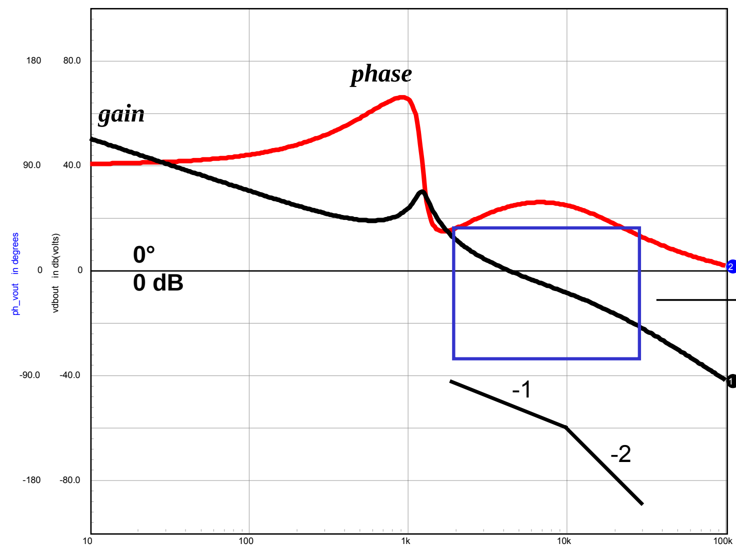
$$\frac{s^2}{\omega_r^2} + \frac{s}{\omega_r Q} + 1 = 0 \longrightarrow \text{2 roots, } s_1 \text{ and } s_2 = \frac{\omega_r}{2Q}\left(-1 \pm \sqrt{1-4Q^2}\right)$$

Q < 0.5, two real negatives roots
Q = 0.5, two real coincident negative roots
Q > 0.5, two complex roots
Q = ∞, two imaginary conjugate roots



Where is the analogy with *T(s)*?

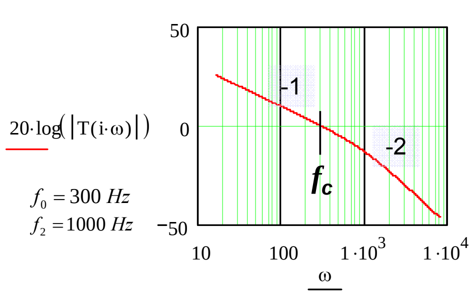
❑ in the vicinity of the crossover point, *T(s)* combines:
- one pole at the origin, ω_0
- one high frequency pole, ω_2

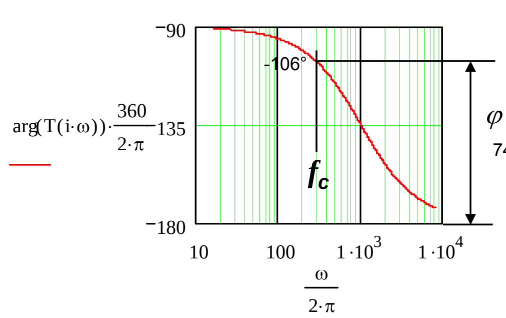


$$T(s) = \frac{1}{\left(\frac{s}{\omega_0}\right)\left(1+\frac{s}{\omega_2}\right)}$$

Closed-loop gain study

❑ Linking the open-loop phase margin to the closed-loop Q





$$T(s) = \frac{1}{\left(\frac{s}{\omega_0}\right)\left(1+\frac{s}{\omega_2}\right)} \xrightarrow{\text{Closed loop}} \frac{T(s)}{1+T(s)} = \frac{1}{\frac{s^2}{\omega_0\omega_2}+\frac{s}{\omega_0}+1}$$

$$\frac{1}{\frac{s^2}{\omega_0\omega_2}+\frac{s}{\omega_0}+1} = \frac{1}{\frac{s^2}{\omega_r^2}+\frac{s}{\omega_r Q}+1} \xrightarrow{\text{solve}} Q = \sqrt{\frac{\omega_0}{\omega_2}} \qquad \omega_r = \sqrt{\omega_0\omega_2}$$

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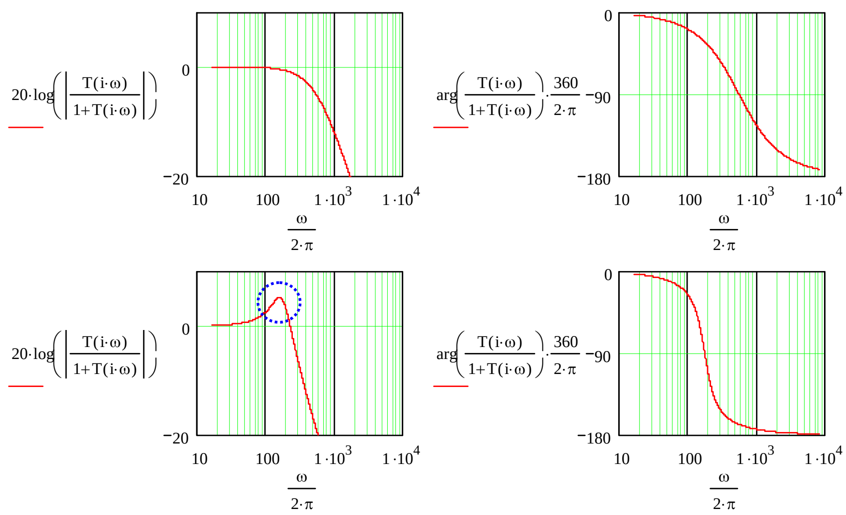
Closed-loop gain study

- if we plot the closed-loop expression:
  - $\omega_2 > \omega_0$, low Q, no peaking
  - $\omega_2 < \omega_0$, high Q, peaking

$$\frac{T(s)}{1+T(s)} = \frac{1}{\frac{s^2}{\omega_r^2}+\frac{s}{\omega_r Q}+1}$$



$f_0 = 300\ Hz$
$f_2 = 1000\ Hz$
$Q = 0.55$
$\boxed{\varphi_m} = 74°$

Open-loop phase

$f_0 = 300\ Hz$
$f_2 = 100\ Hz$
$\boxed{Q} = 1.7$
$\varphi_m = 32°$

Closed-loop Q

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Linking φ_m and Q

- an open-loop phase margin leads to a closed-loop quality coeff. Q
- we have seen that Q affects the transient response (*RLC* filter)
- let us link the phase margin to the quality coefficient:

1. calculate the crossover frequency for which $|T(s)| = 1$

$$\left|\frac{1}{\left(\frac{j\omega_c}{\omega_0}\right)\left(1+\frac{j\omega_c}{\omega_2}\right)}\right| = 1 \longrightarrow \omega_c = \frac{\omega_2\sqrt{\left(\sqrt{1+4Q^4}-1\right)}}{\sqrt{2}}$$

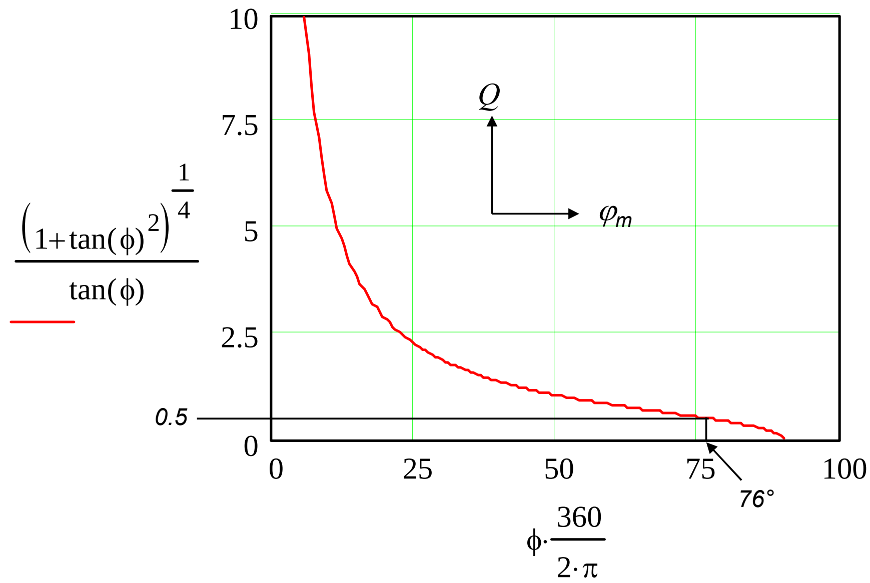
2. substitute ω_c into *T(s)*, calculate its argument (phase margin)

arg *T(s)*@ f_c = $\arctan\left(\sqrt{\frac{2}{\sqrt{\left(1+4Q^4\right)}-1}}\right)$

3. extract the quality coefficient Q: $Q = \frac{\sqrt[4]{1+\tan(\varphi_m)^2}}{\tan(\varphi_m)} = \frac{\sqrt{\cos(\varphi_m)}}{\sin(\varphi_m)}$

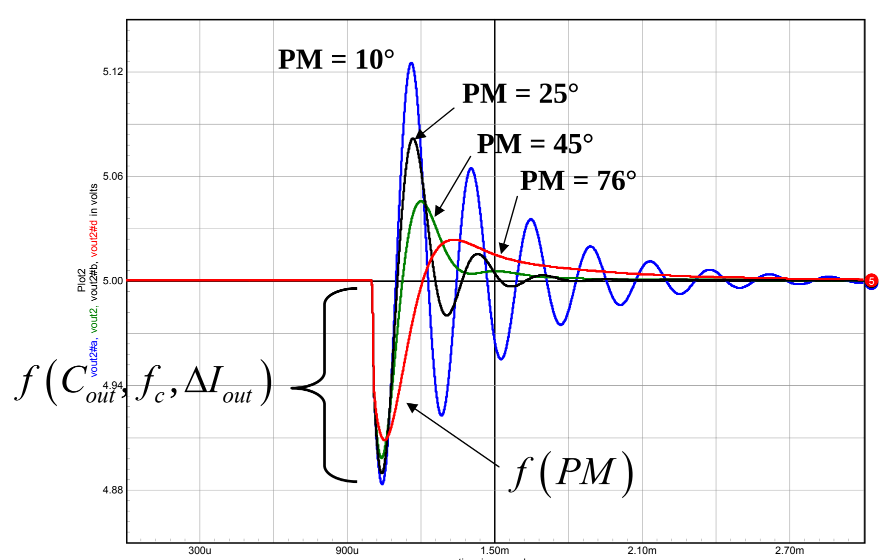
We can now plot Q versus φ_m

- ❑ a Q factor of 0.5 (critical response) implies a φ_m of 76°
- ❑ a 45° φ_m corresponds to a Q of 1.2: oscillatory response!



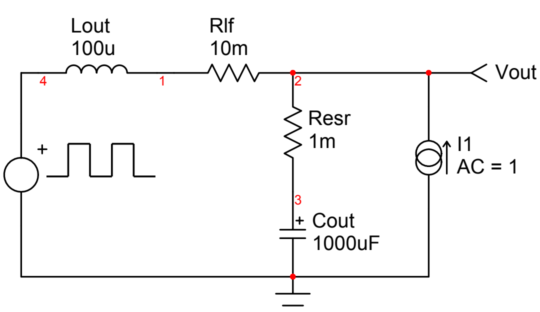
Summary on the design criteria

- ❑ compensate the open-loop gain for a phase margin of 70°
- ❑ make sure the open-loop gain margin is better than 15 dB
- ❑ do not accept a phase margin lower than 45° in worst case



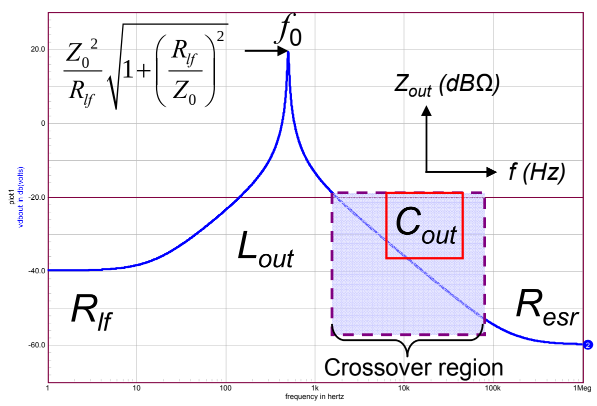
Dc-dc output impedance

- A dc-dc conv. combines an inductor and a capacitor
- As f is swept, different elements dominate Z_{out}



A buck equivalent circuit

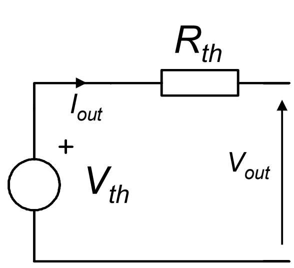
$$Z_{out} = \left(sL_{out} + r_{Lf}\right) \| \left(R_{esr} + \frac{1}{sC_{out}}\right)$$



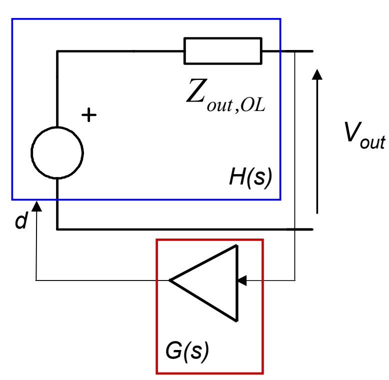
To avoid stability issues,
$f_c >> f_0$

Closing the loop…

- Any circuit can be represented by its Thévenin model
- At high frequency, C_{out} impedance dominates
- Once in closed-loop, Z_{out} goes down as $T(s)$ is high



$$\boxed{Z_{out,OL} = \frac{1}{2\pi C_{out} f}}$$

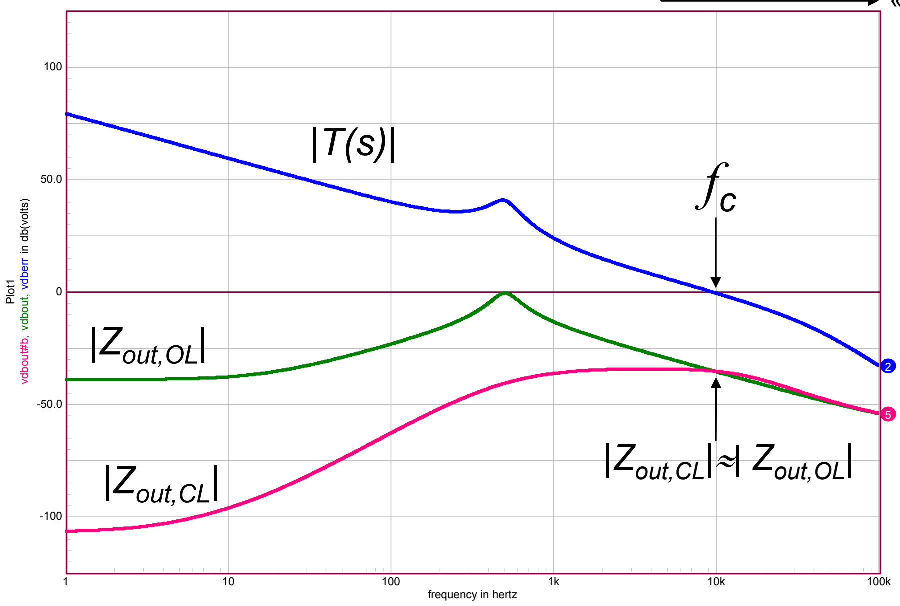


$$\left|Z_{out,CL}\right| = \left|Z_{out,OL}\right|\left|\frac{1}{1+T(s)}\right|$$

Closing the loop…

- At the crossover frequency $Z_{out,CL} \boxed{\approx} Z_{out,OL}$

Let's assess « almost » :-)



Calculating the output impedance

- closed-loop output impedance is dominated by C_{out}

$$\left|Z_{out,CL}\right| \approx \frac{1}{2\pi f_c C_{out}} \left|\frac{1}{1+T(s)}\right|$$

- we have calculated the crossover value, $|T(s)| = 1$

$$\omega_c = \frac{\omega_2 \sqrt{\left(\sqrt{1+4Q^4}-1\right)}}{\sqrt{2}}$$

- substitue into 1/ (1+$T(s)$) and extract module

$$\left|\frac{1}{1+T(s)}\right| = \frac{1}{\sqrt{\left(\dfrac{-\omega_0}{\omega_2\left(1+\dfrac{\omega^2}{\omega_2^{\,2}}\right)}+1\right)^2 + \dfrac{\omega_0^{\,2}}{\omega^2\left(1+\dfrac{\omega^2}{\omega_2^{\,2}}\right)^2}}}$$

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Calculating the output impedance

- ❑ Introduce the quality factor coefficient

$$Q = \sqrt{\frac{\omega_0}{\omega_2}} \qquad \left|\frac{1}{1+T(s)}\right| = \frac{1}{2Q\sqrt{\frac{2Q^2+1-\sqrt{1+4Q^4}}{\left(1+\sqrt{1+4Q^4}\right)\left(\sqrt{1+4Q^4}-1\right)\omega_2{}^2}}}$$

- ❑ Now replace Q by its definition $Q = \frac{\sqrt{\cos(\varphi_m)}}{\sin(\varphi_m)}$

$$\left|\frac{1}{1+T(s)}\right| = \frac{1}{\sqrt{\frac{1}{\cos(\varphi_m)}\left(2\cos(\varphi_m)+\left[\frac{1+\cos^2(\varphi_m)}{\sin^2(\varphi_m)}\right]\left[\cos^2(\varphi_m)-1\right]+1-\cos^2(\varphi_m)\right)}}$$

- ❑ Simplify $\longrightarrow$ $\left|\frac{1}{1+T(s)}\right| = \frac{1}{\sqrt{2-2\cos(\varphi_m)}}$

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An example with a buck

- ❑ Let's assume an output capacitor of 1 mF
- ❑ The spec states a 80-mV undershoot for a 2-A step
- ❑ How to select the crossover frequency?

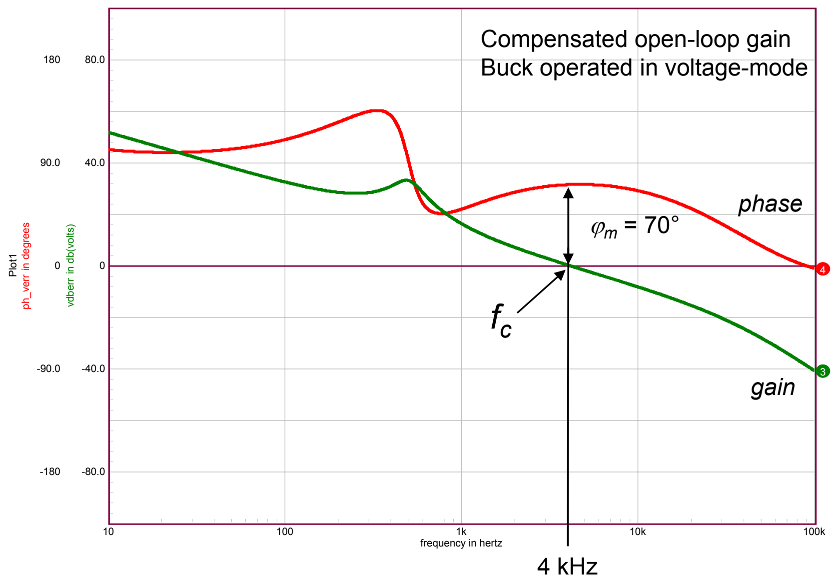
$$\Delta V_{out} \approx \frac{\Delta I_{out}}{2\pi f_c C_{out}} \longrightarrow f_c \approx \frac{\Delta I_{out}}{\Delta V_{out} C_{out} 2\pi}$$

$$f_c \approx \frac{2}{80m \times 1m \times 2\pi} = \boxed{4\ kHz} \qquad Z_{C_{out}} @ 4\ kHz = \frac{1}{2\pi \times 4k \times 1m} = 40\ m\Omega$$

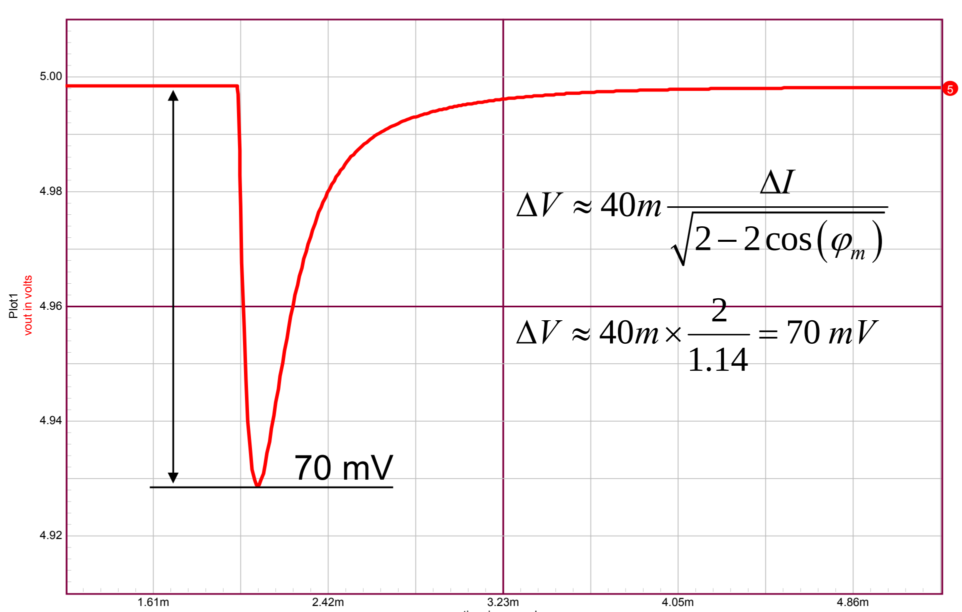
➡ Select a 1000-µF capacitor featuring less than 40-mΩ ESR

Setting the right crossover frequency

- ❑ Compensate the converter for a 4-kHz f_c

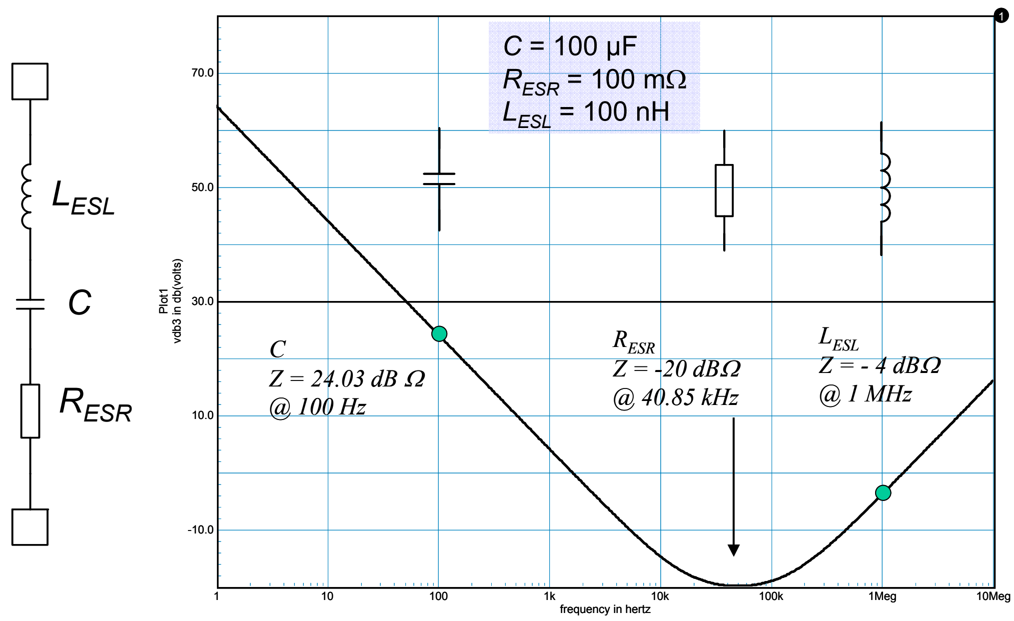


Measure the obtained undershoot



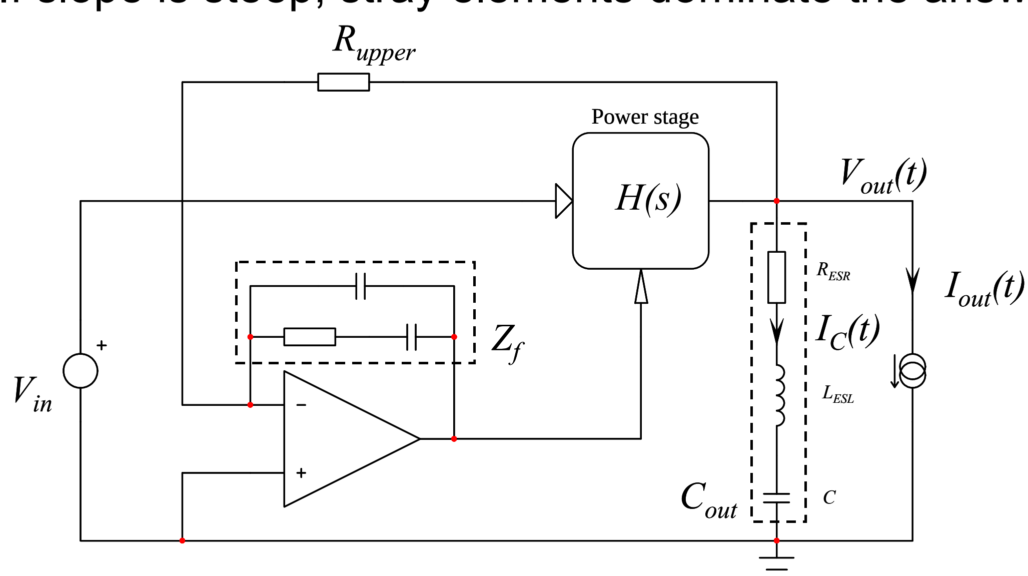
Is my capacitor a real capacitor?

- A capacitor is made of parasitic elements



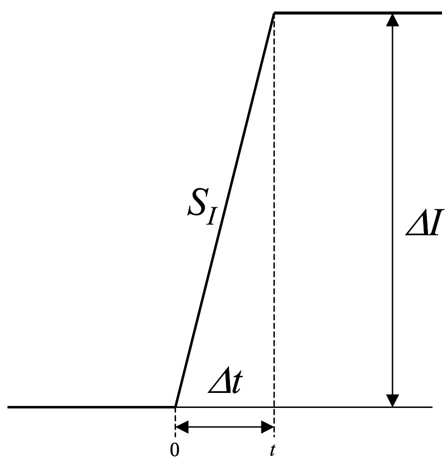
How these elements affect the undershoot?

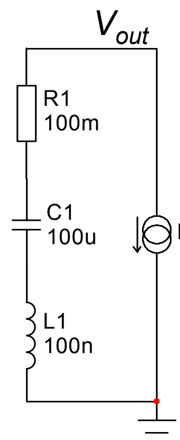
- The output current slope affects the undershoot
- If slope is steep, stray elements dominate the answer



How these elements affect the undershoot?

- ❑ Because of bandwidth limits, R_{ESR} and L_{ESL} play alone



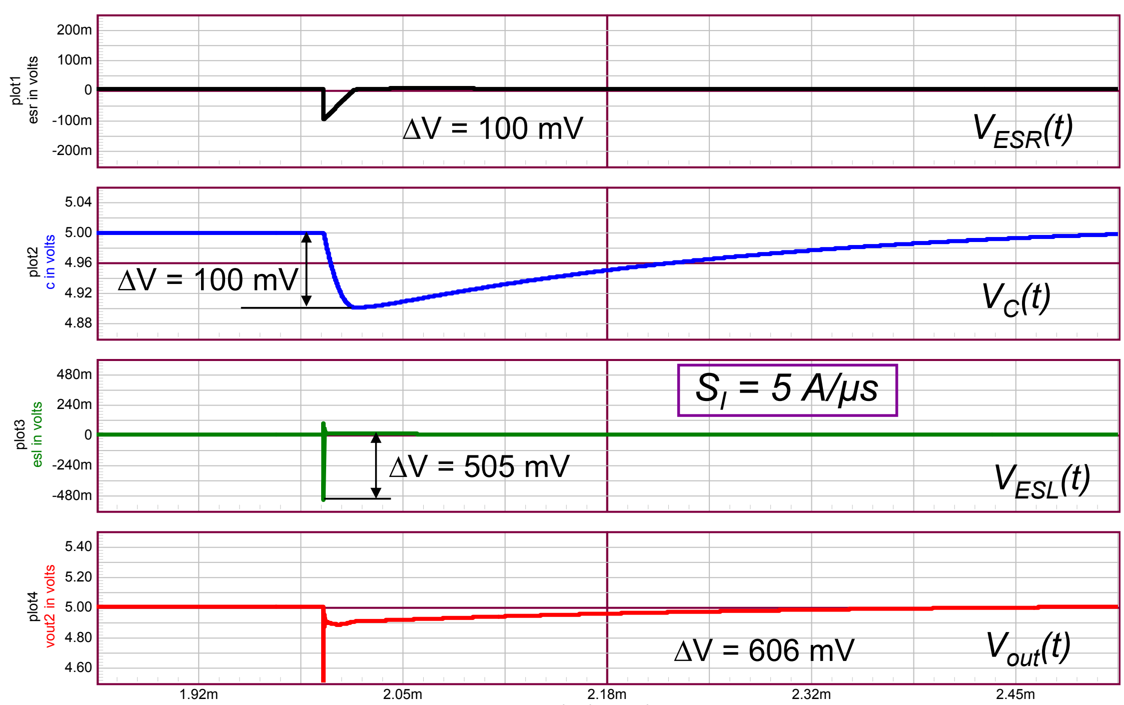


$$I_{out}(t) = \Delta I \frac{t}{\Delta t} = S_I t$$

$$\Delta V_{out} = R_{ESR} S_I t + L_{ESL} S_I + \frac{1}{C}\int_0^t S_I t \cdot dt$$

$$\Delta V_{out} = R_{ESR} S_I t + L_{ESL} S_I + \frac{S_I t^2}{2C}$$

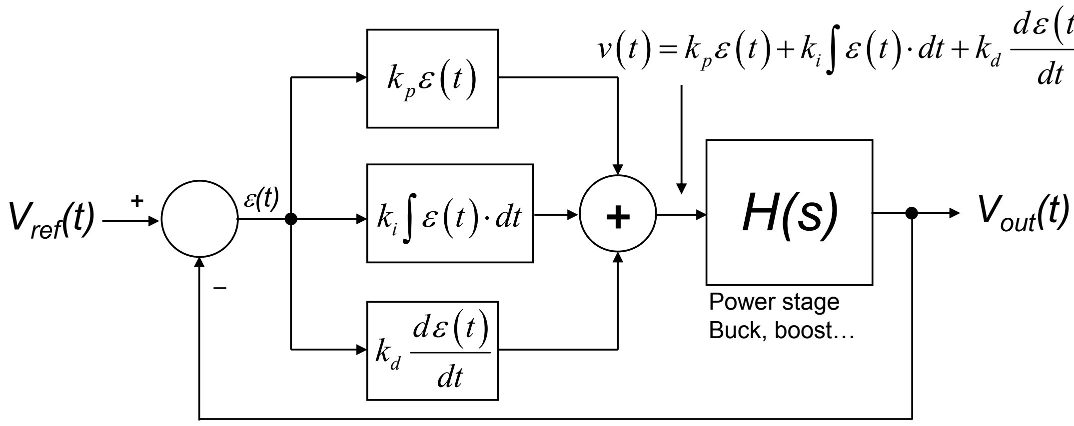
The capacitor contribution is small…



Compensating the converter

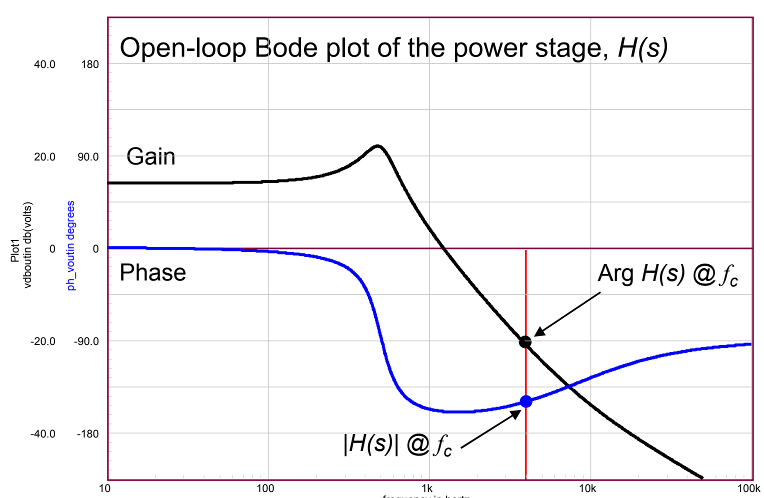
- ❑ Fix the current error with a **proportional** term (P)
- ▪ The proportional gain gives fast reaction time but also overshoot
- ❑ Fix the long-term static error with an **integral** term (I)
- ▪ The integral term cancels the static error but slows down the response
- ❑ Fix the immediate error by observing the slope with a **derivative** term (D)
- ▪ The derivative term decreases overshoot but slows down the response

$$v(t) = k_p\varepsilon(t) + k_i\int\varepsilon(t)\cdot dt + k_d\frac{d\varepsilon(t)}{dt}$$



How do we stabilize the converter?

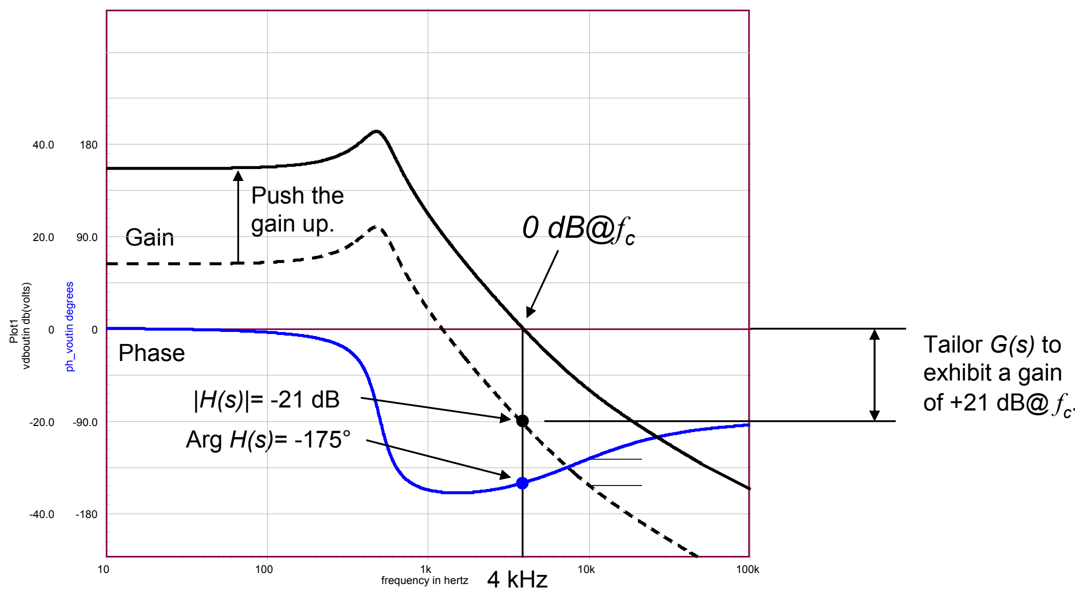
1. Select the crossover frequency f_c (assume 4 kHz)
2. Provide a high dc gain for a low static error and good input rejection
3. Shoot for a 70° phase margin at f_c
4. Evaluate the needed phase boost at f_c to meet (3)
5. Shape the *G(s)* path to comply with 1, 2 and 3



$$A_{sc,CL}(s) = \frac{A_{sc,OL}(s)}{1+T(s)}$$

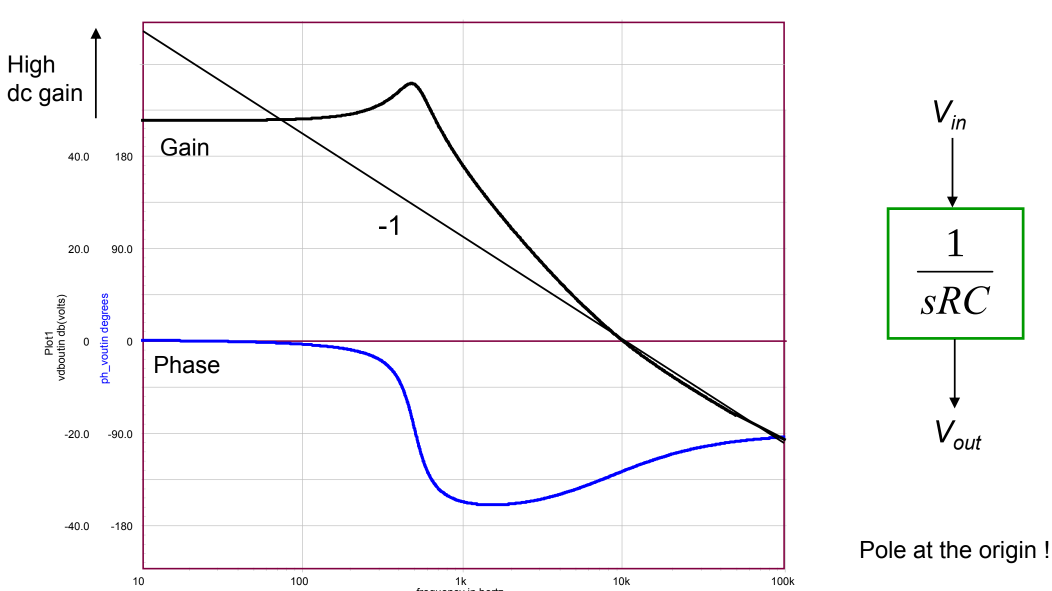
First, provide mid-band gain at crossover

1. Adjust *G(s)* to boost the gain by +21 dB at crossover



Second, provide high gain in dc

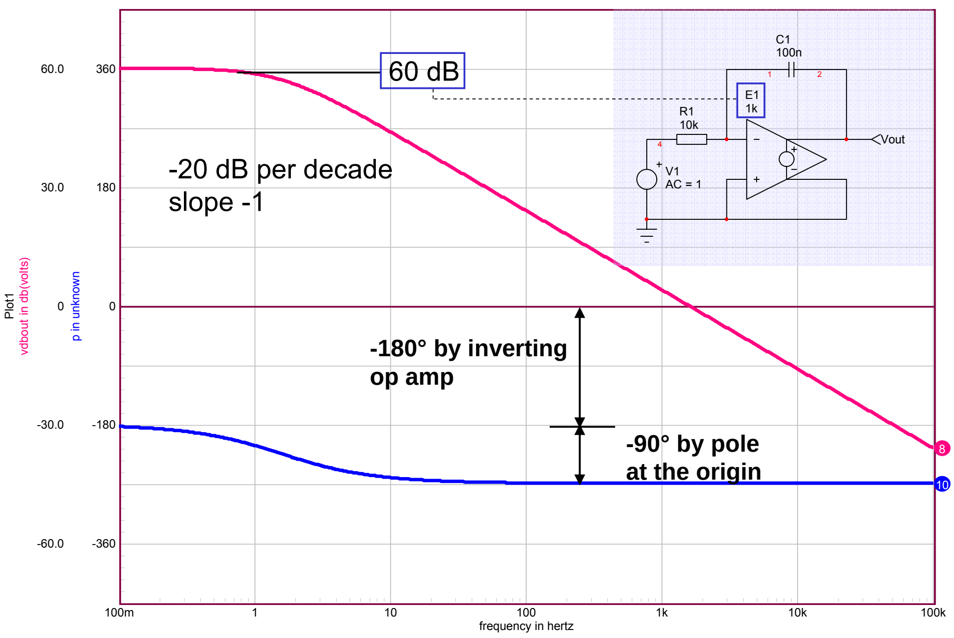
2. High dc gain lowers static error and brings good input rejection



Pole at the origin !

Second, provide high gain in dc

2. An integrator provides a high dc gain but rotates by -270°



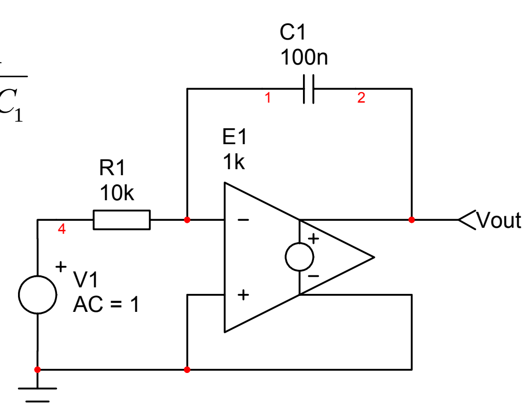
Second, provide high gain in dc

2. An integrator provides a high dc gain but rotates by -270°

$$G(s)=\frac{V_{out}(s)}{V_{in}(s)}=\frac{1}{sR_1C_1}=\frac{1}{\frac{s}{\omega_{po}}}\qquad \omega_{po}=\frac{1}{R_1C_1}$$

$$\arg G(j\omega)=-\arg\left(\frac{j\omega}{\omega_{po}}\right)$$

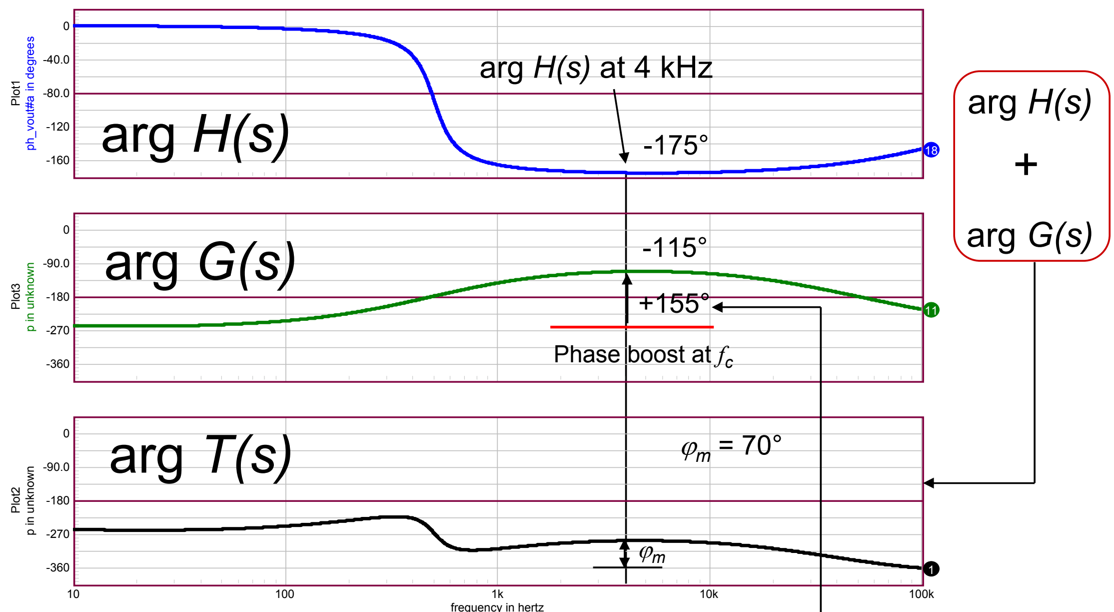
$$\lim_{s\to\infty}\arg G(s)=\lim_{\omega\to\infty}-\arctan\left(\frac{\frac{\omega}{\omega_{po}}}{0}\right)=-\frac{\pi}{2}$$



→ Total phase lag brought by an origin pole is -3π/2 or -270°

All compensators (1, 2 or 3) feature an origin pole

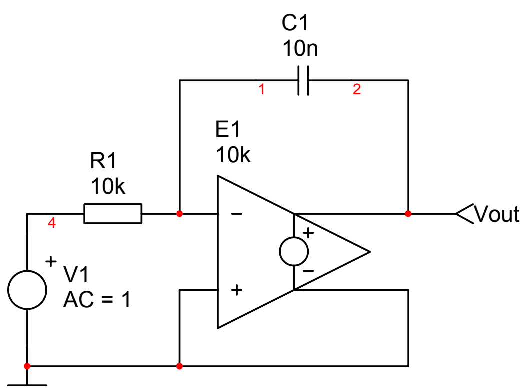
Third, evaluate the phase boost at f_c



$$\arg H(f_c) - 270° + BOOST - \varphi_m = -360°$$
$$BOOST = \varphi_m - \arg H(f_c) - 90° = 70° + 175 - 90 = \boxed{155°}$$

How do we boost the phase at f_c?

- ❑ The type 1 configuration
- ❑ No phase boost, pure integral term
- ❑ Permanent phase lag of -270°
- ❑ Ok if arg$H(f_c)$ < -45° for a φ_m of 45°



$$G(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{1}{sR_1C_1} = \frac{1}{\frac{s}{\omega_{po}}}$$

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

$$G = \frac{f_{po}}{f_c}$$

Select f_{po} to have G at f_c

1 pole at the origin

Type 1

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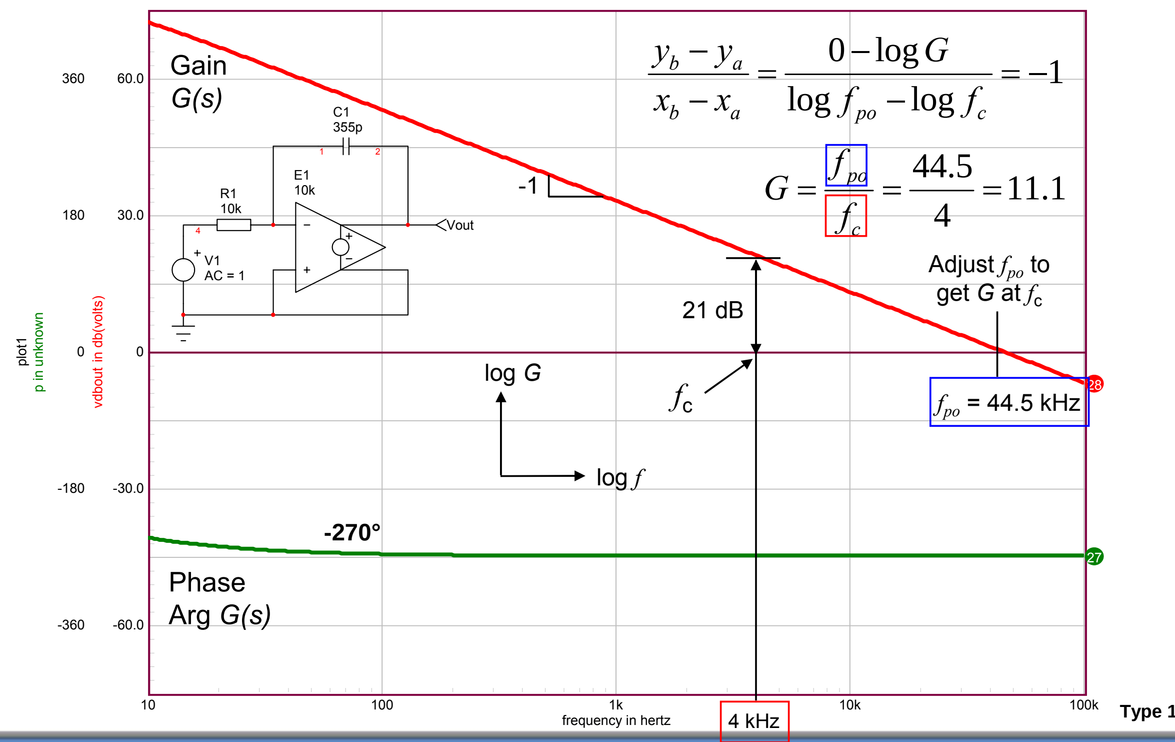
How do we boost the phase at f_c?



Type 1

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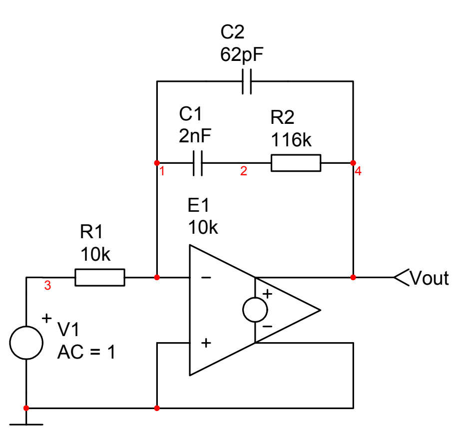
How do we boost the phase at f_c?

- ❑ The type 2 configuration
- ❑ Phase boost up to 90°
- ❑ Ok if arg$H(f_c)$ < -90°



$$G(s) = -\frac{1 + sR_2C_1}{sR_1(C_1 + C_2)\left(1 + sR_2\left[\frac{C_1C_2}{C_1 + C_2}\right]\right)}$$

If $C_2 << C_1$

$$\omega_{po} = \frac{1}{R_1C_1} \quad \omega_{p1} = \frac{1}{R_2C_2} \quad \omega_{z1} = \frac{1}{R_2C_1}$$

1 pole at the origin
1 zero
1 pole

Type 2

Pole/zero placement and boost at f_c

- Phase boost appears between the zero and the pole

$$G(j\omega)=\frac{\left(1+j\frac{\omega}{\omega_{z1}}\right)}{\left(1+j\frac{\omega}{\omega_{p1}}\right)}\qquad \arg G(j\omega)=boost=\arg\frac{\left(1+j\frac{\omega}{\omega_{z1}}\right)}{\left(1+j\frac{\omega}{\omega_{p1}}\right)}\qquad \begin{array}{l}G=a+jb\\ \arg G=\arctan\left(\frac{b}{a}\right)\end{array}$$

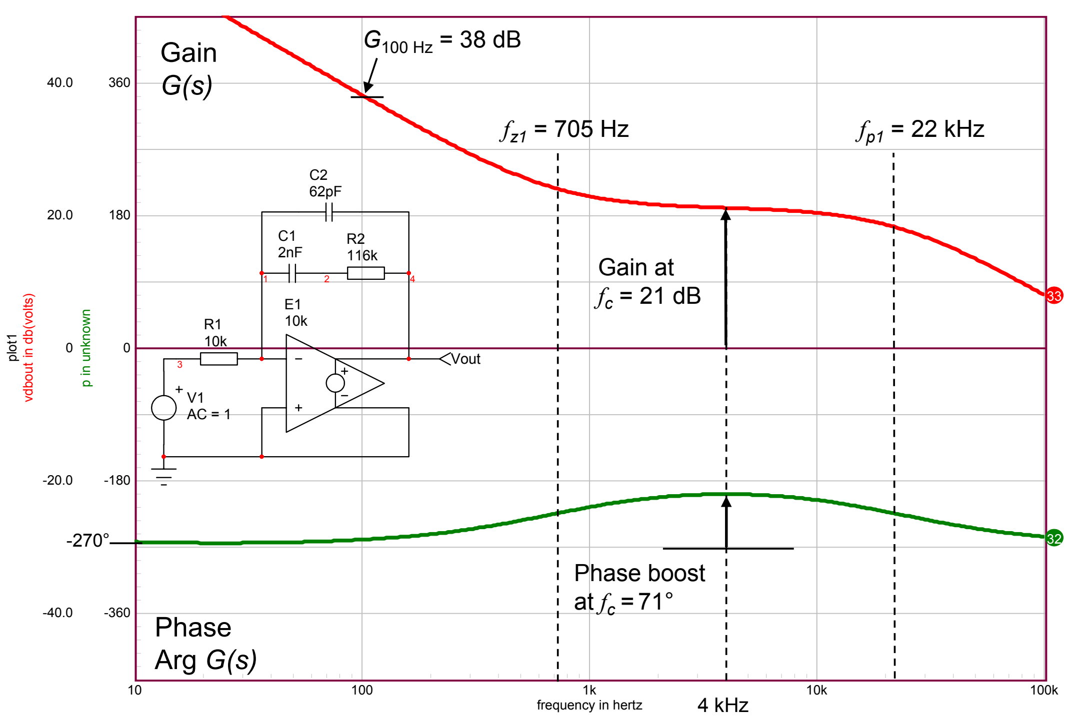
$$\arg G(f)=\arctan\left(\frac{f}{f_{z1}}\right)-\arctan\left(\frac{f}{f_{p1}}\right)\xrightarrow{\text{Peaks to a max at}} f=\sqrt{f_{z1}f_{p1}}$$

Assume 1 zero placed at 705 Hz, 1 pole at 22 kHz and a 4-kHz crossover:

$$\arg G(4\,kHz)=\arctan\left(\frac{4k}{705}\right)-\arctan\left(\frac{4k}{22k}\right)=80-10.3\approx 70°$$

Type 2

How do we boost the phase at f_c?

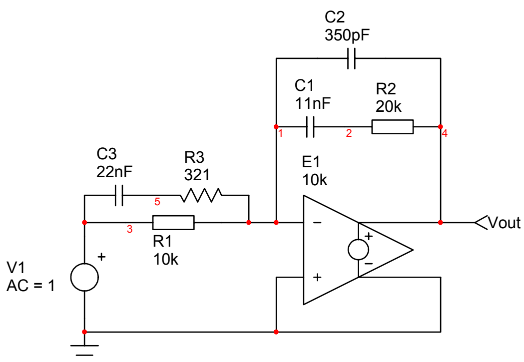


Type 2

How do we boost the phase at f_c?

- The type 3 configuration
- Phase boost up to 180°
- Ok if arg$H(f_c)$ < -180°

$$G(s)=-\frac{sR_2C_1+1}{sR_1(C_1+C_2)\left(1+sR_2\frac{C_1C_2}{C_1+C_2}\right)}\frac{sC_3(R_1+R_3)+1}{(sR_3C_3+1)}$$



If $C_2 \ll C_1$ and $R_3 \ll R_1$

$$\omega_{z1}=\frac{1}{R_2C_1}\quad \omega_{z2}=\frac{1}{R_1C_3}\quad \omega_{po}=\frac{1}{R_1C_1}$$

$$\omega_{p1}=\frac{1}{R_3C_3}\quad \omega_{p2}=\frac{1}{R_2C_2}$$

1 pole at the origin
2 zeros
2 poles

Type 3

Pole/zero placement and boost at f_c

- Phase boost appears between the zero and the pole

$$G(j\omega)=\frac{\left(1+j\frac{\omega}{\omega_{z1}}\right)\left(1+j\frac{\omega}{\omega_{z2}}\right)}{\left(1+j\frac{\omega}{\omega_{p1}}\right)\left(1+j\frac{\omega}{\omega_{p2}}\right)}\qquad \arg G(j\omega)=boost=\arg\frac{\left(1+j\frac{\omega}{\omega_{z1}}\right)\left(1+j\frac{\omega}{\omega_{z2}}\right)}{\left(1+j\frac{\omega}{\omega_{p1}}\right)\left(1+j\frac{\omega}{\omega_{p2}}\right)}$$

$$\arg G(f)=\arctan\left(\frac{f}{f_{z1}}\right)+\arctan\left(\frac{f}{f_{z2}}\right)-\arctan\left(\frac{f}{f_{p1}}\right)-\arctan\left(\frac{f}{f_{p2}}\right)$$

Assume 2 zeros placed at 500 Hz, 2 poles at 50 kHz and a 4-kHz crossover:

$$\arg G(4\,kHz)=2\arctan\left(\frac{4k}{500}\right)-2\arctan\left(\frac{4k}{50k}\right)=166-4.6\approx 161°$$

Type 3

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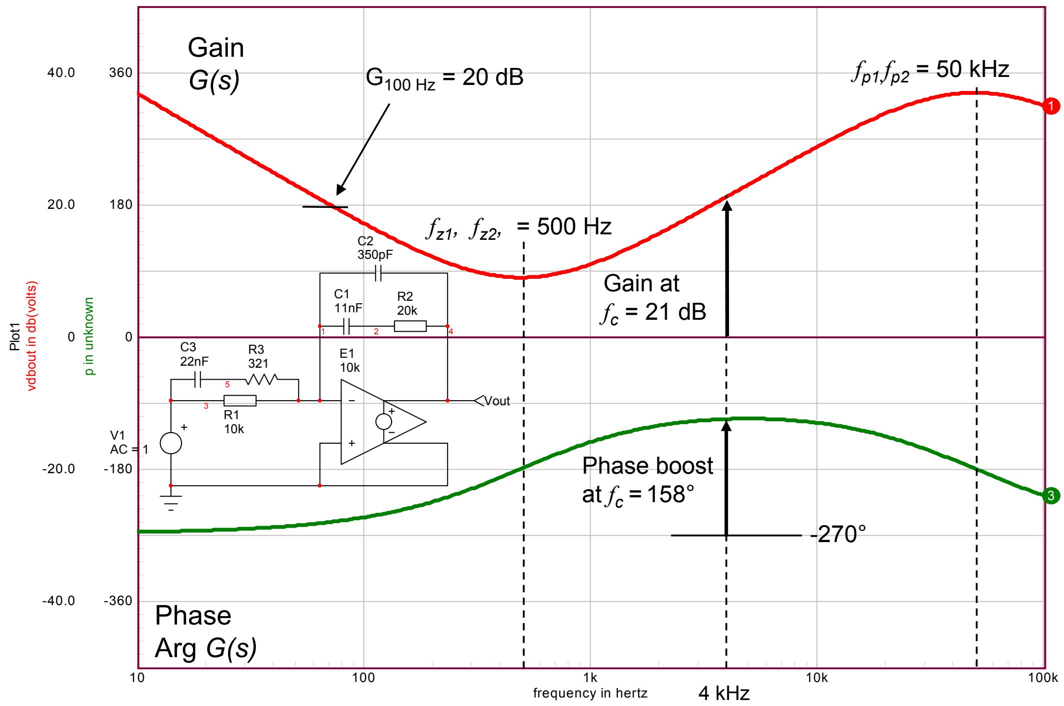
How do we boost the phase at f_c?



Type 3

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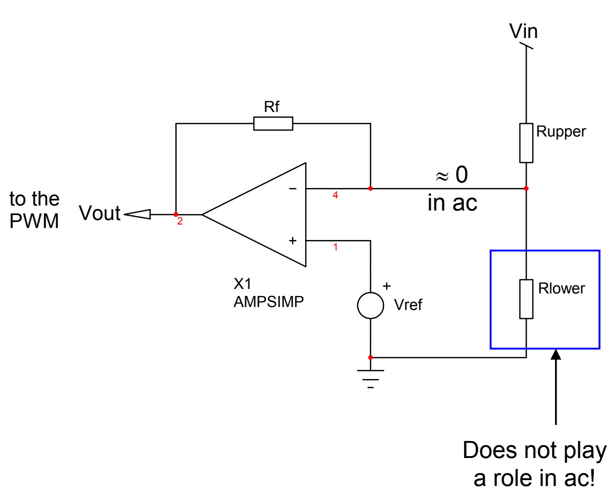
Why R_{lower} never appears in the ac functions?

- Because of the virtual ground, R_{lower} disappears in ac!



dc equation

$$V_{out} = V_{ref}\left(\frac{R_f}{R_{upper} \parallel R_{lower}} + 1\right) - \frac{R_f}{R_{upper}} V_{in}$$

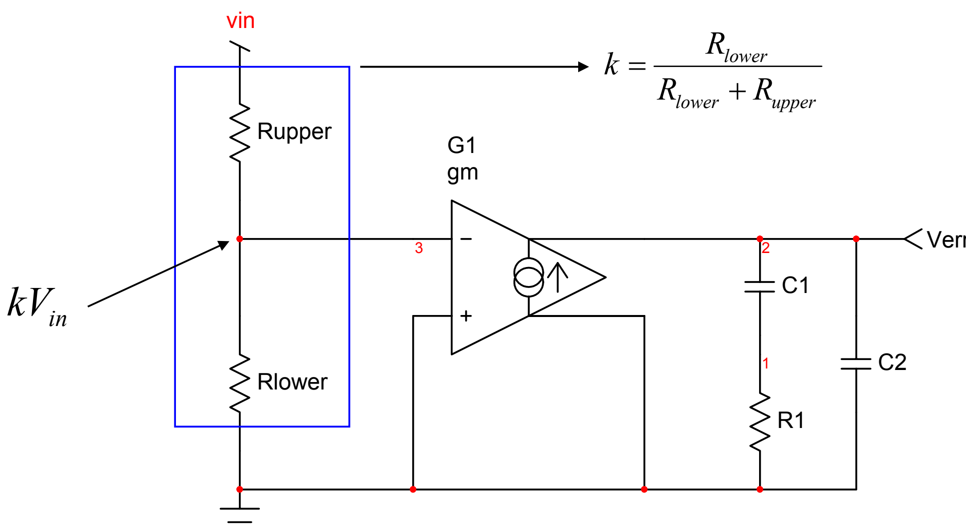
ac equation

$$\frac{V_{out}(s)}{V_{in}(s)} = -\frac{R_f}{R_{upper}}$$

As long as a virtual ground exists, the division ratio plays no role!

In absence of virtual ground, R_{lower} comes back

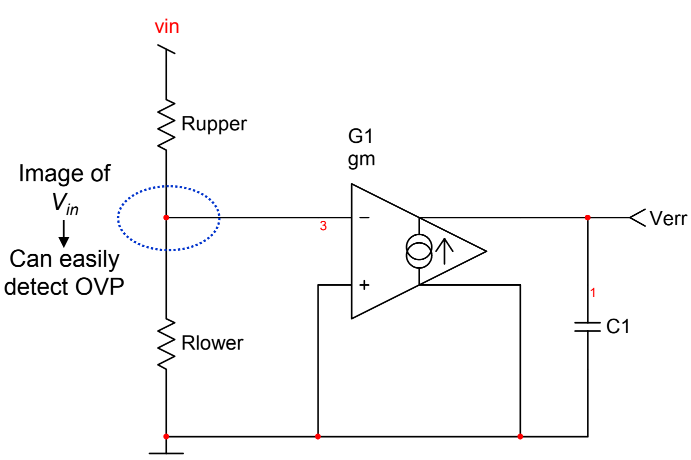
- ❑ In an OTA-based circuit, there is no virtual ground
- ❑ R_{lower} now affects $G(s)$



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

Boosting the phase at f_c with OTAs

- ❑ The type 1 configuration with an OTA
- ❑ OTA types are popular in PFC circuits
- ❑ No virtual grounds, the divider enters the picture!



$$G(s) = \frac{1}{s\dfrac{R_{lower} + R_{upper}}{gmR_{lower}}C_1}$$

$$\omega_{po} = \frac{1}{C_1\dfrac{R_{lower} + R_{upper}}{gmR_{lower}}}$$

$$G = \frac{f_{po}}{f_c}$$

Select f_{po} to have G at f_c

1 pole at the origin

Type 1 - OTA

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Boosting the phase at f_c with OTAs

- The calculation is simple:

parameters

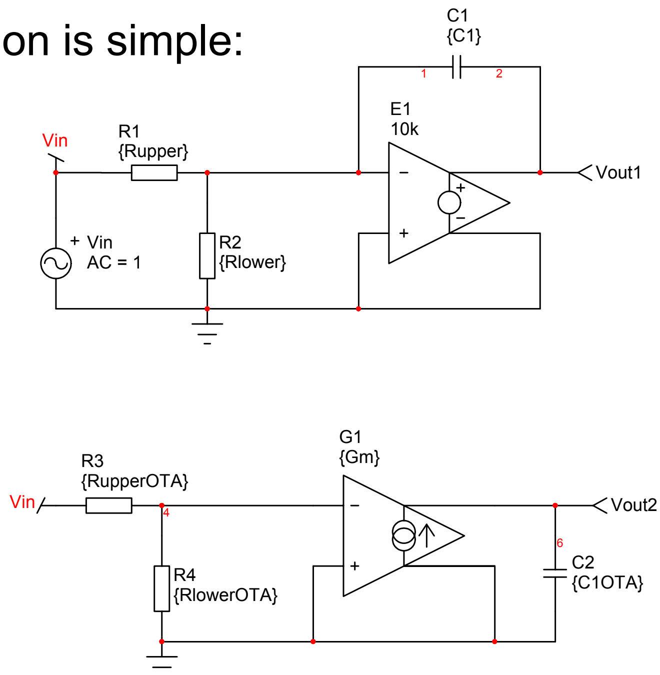
Vout=385V
Ib=250u
Vref=2.5

Rupper=(Vout-Vref)/Ib
Rlower=2.5/Ib

fc=20
Gfc=46

gm=200u
fpo=G*fc
RlowerOTA=Vref/Ib
RupperOTA=(Vout-Vref)/Ib
Req=(Rlower+Rupper)/(gm*Rlower)
C1OTA=1/(2*pi*fpo*Req)

G=10^(-Gfc/20)
pi=3.14159
C1=1/(2*pi*fc*G*Rupper)



Type 1 - OTA

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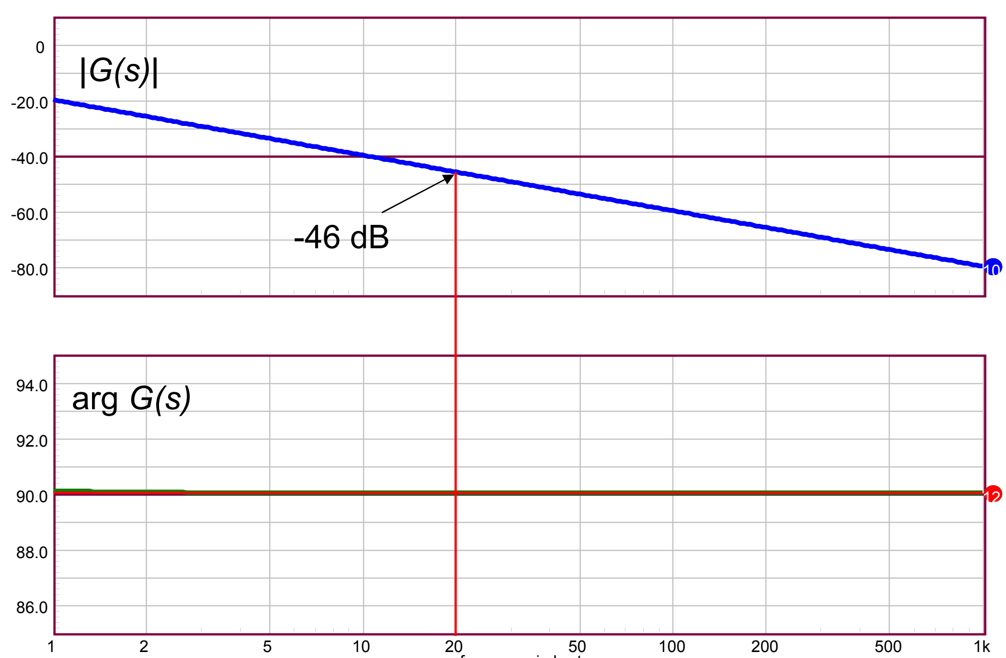
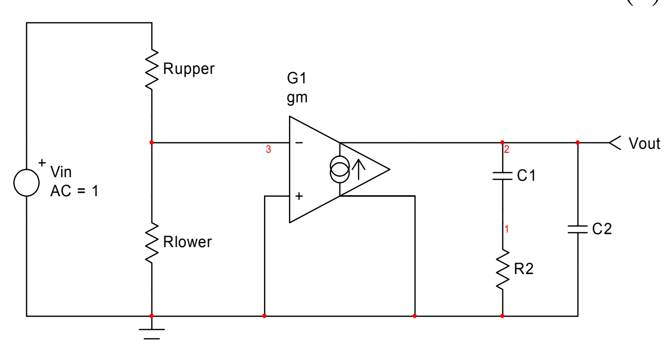
Boosting the phase at f_c with OTAs



Type 1 - OTA

Boosting the phase at f_c with OTAs

- The type 2 configuration with an OTA
- Type 2 improves transient response in BCM PFCs
- No virtual grounds, the divider enters the picture!



Mid-band gain

$$G(s) \approx \boxed{\frac{R_{lower}\, gmR_2}{R_{lower} + R_{upper}}} \frac{1 + sR_2C_1}{sR_2C_1\left(1 + sR_2C_2\right)}$$

If $C_2 << C_1$

$$\omega_{po} = \omega_z = \frac{1}{R_2C_1} \quad \omega_p = \frac{1}{R_2C_2}$$

1 pole at the origin
1 zero
1 pole

Type 2 - OTA

Boosting the phase at f_c with OTAs

parameters

Rupper=3.6Meg
Rlower=23k
gm=200u
fc=10
Gfc=-15

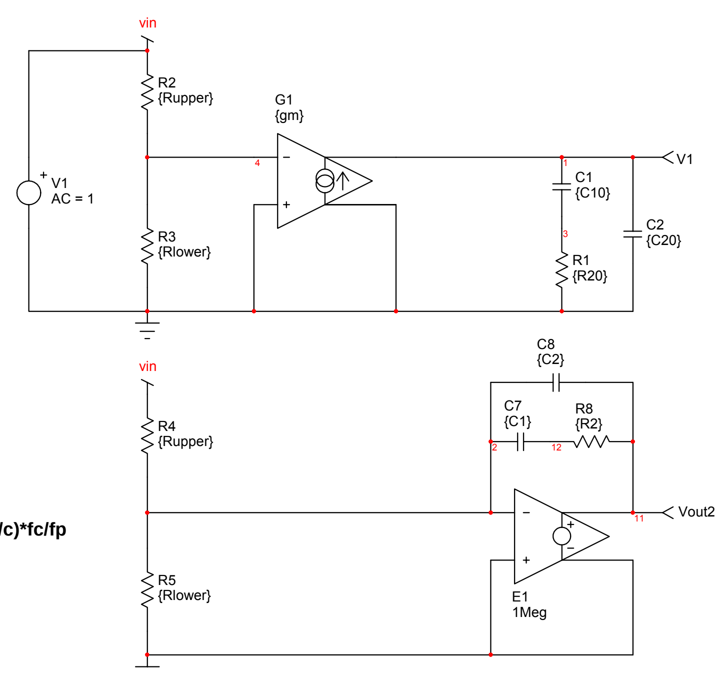
G=10^(-Gfc/20)
pi=3.14159

fp=100
fz=1

a=sqrt((fc^2+fp^2)*(fc^2+fz^2))
c=(fc^2+fz^2)

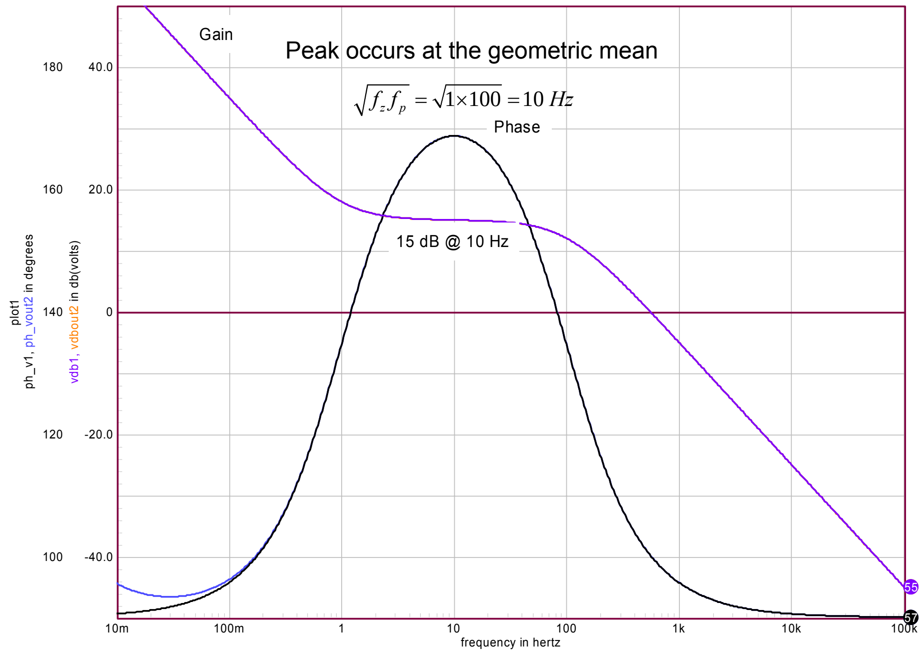
R2=(a/c)*Rupper*fc*G/fp
C2=1/(2*pi*fp*R2)
C1=1/(2*pi*fz*R2)

R20=G*(Rupper+Rlower)/(Rlower*gm)*(a/c)*fc/fp
C10=1/(2*pi*fz*R20)
C20=C10/(2*pi*fp*C10*R20-1)



Type 2 - OTA

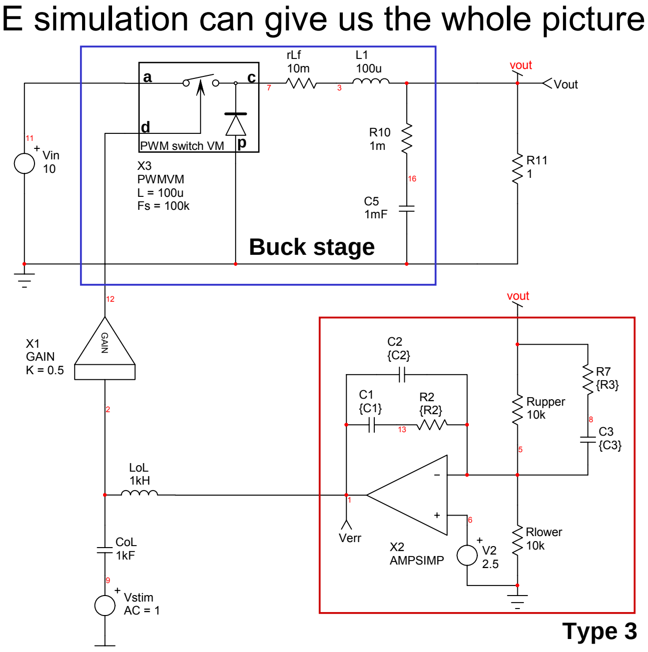
Boosting the phase at f_c with OTAs



Type 2 - OTA

Finally, we test the open-loop gain

5. Given the necessary boost of 155°, we select a type-3 amplifier
6. A SPICE simulation can give us the whole picture!



1 pole at the origin
2 zeros at 500 Hz
2 poles at 50 kHz

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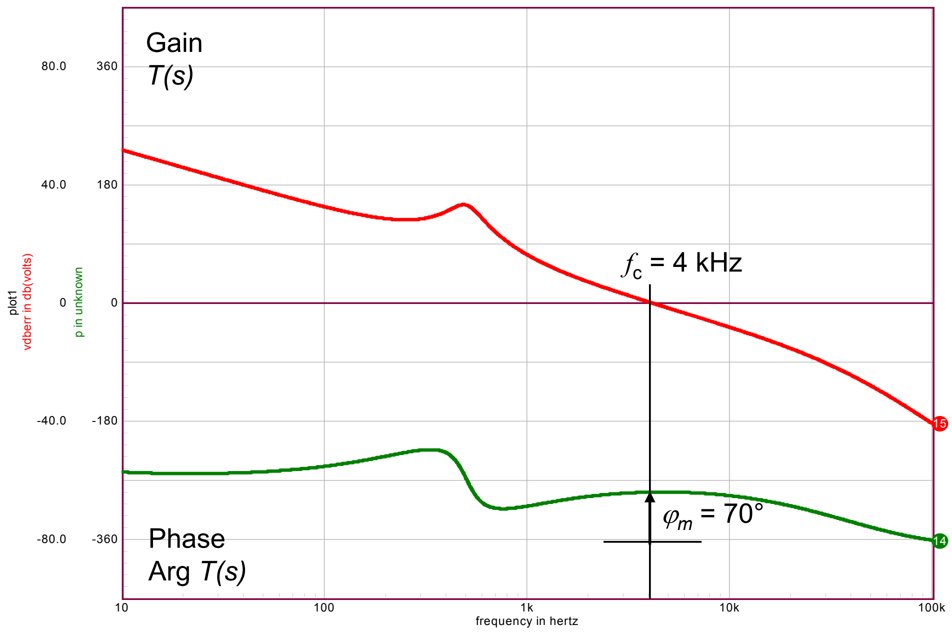
Finally, we test the open-loop gain

An ac simulation gives us the open-loop Bode plot



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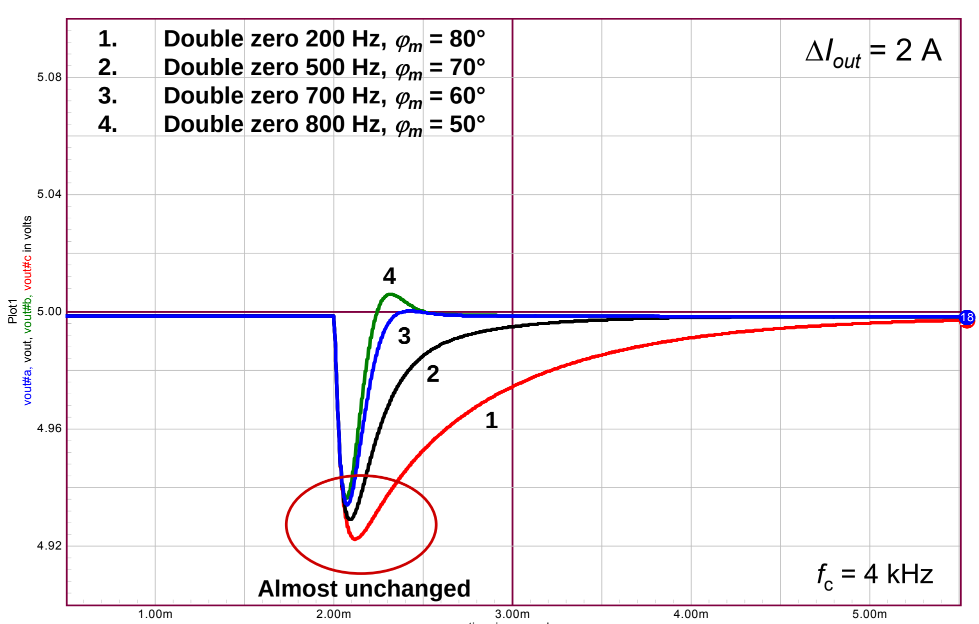
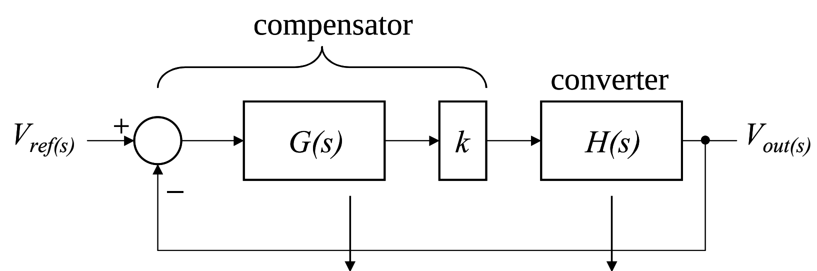
What transient response?

- ❑ The crossover frequency is constant, but zeros position is changed



Zeros become poles??

- ❑ *T(s)* is shaped by introducing zeros/poles in *G(s)*
- ❑ In closed-loop conditions, the zeros of *G(s)* become poles…



$$\boxed{G(s)=\frac{N_G(s)}{D_G(s)}} \qquad \boxed{H(s)=\frac{N_H(s)}{D_H(s)}}$$

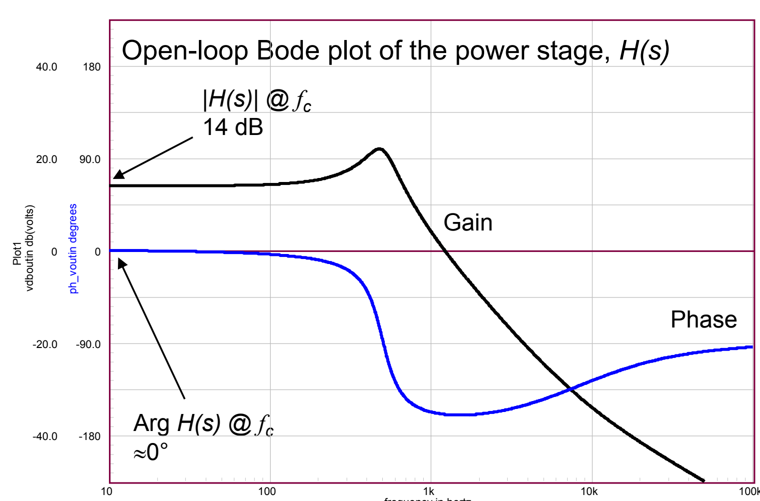
$$\frac{V_{out}(s)}{V_{ref}(s)}=\frac{kG(s)H(s)}{1+kG(s)H(s)} \longrightarrow \frac{V_{out}(s)}{V_{ref}(s)}=\frac{k\dfrac{N_G(s)}{D_G(s)}\dfrac{N_H(s)}{D_H(s)}}{1+k\dfrac{N_G(s)}{D_G(s)}\dfrac{N_H(s)}{D_H(s)}}=\frac{kN_G(s)N_H(s)}{D_G(s)D_H(s)+k\boxed{N_G(s)}N_H(s)}$$

For *k* >> 1
zeros of *G(s)* appear
in denominator as poles

→ Pushing the zeros towards low frequency slows down the response!

If we roll-off the BW in the low frequencies?

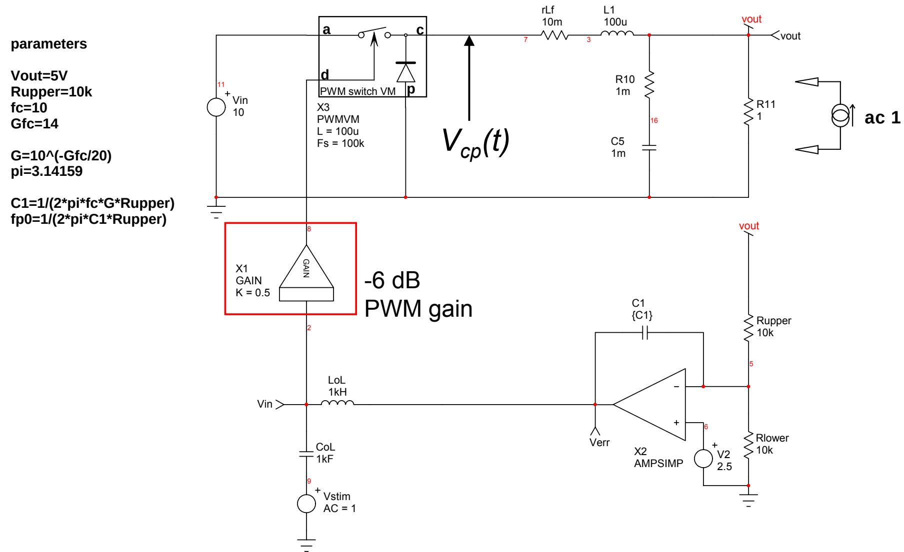
- ❑ Could we avoid the resonance with a 10 Hz crossover point?



- ❑ We have almost no phase rotation at 10 Hz, a type 1 could do??

If we roll-off the BW in the low frequencies?

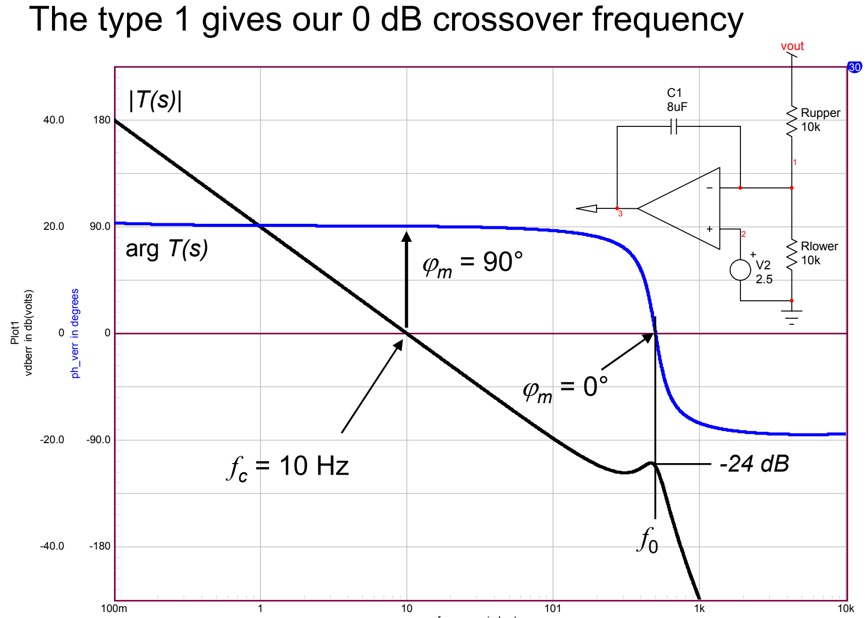
- A simple SPICE ac simulation gives us the open-loop gain



- An ac current source sweeps the output impedance

The open-loop gain looks good…

- The type 1 gives our 0 dB crossover frequency



- We have plenty of phase margin, a slow system, ok…

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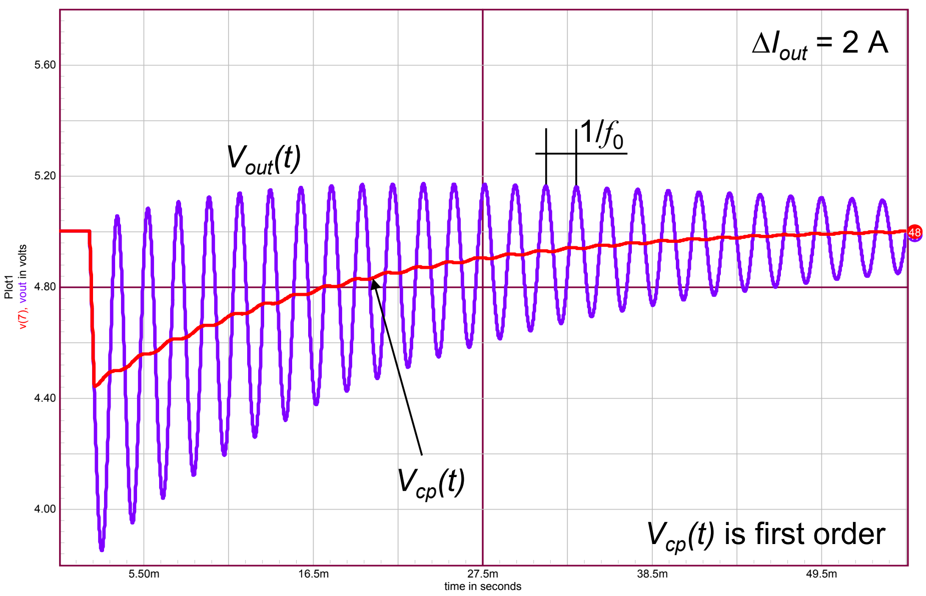
Oh no, it is ringing!

- ❑ The load step reveals a ringing ac output



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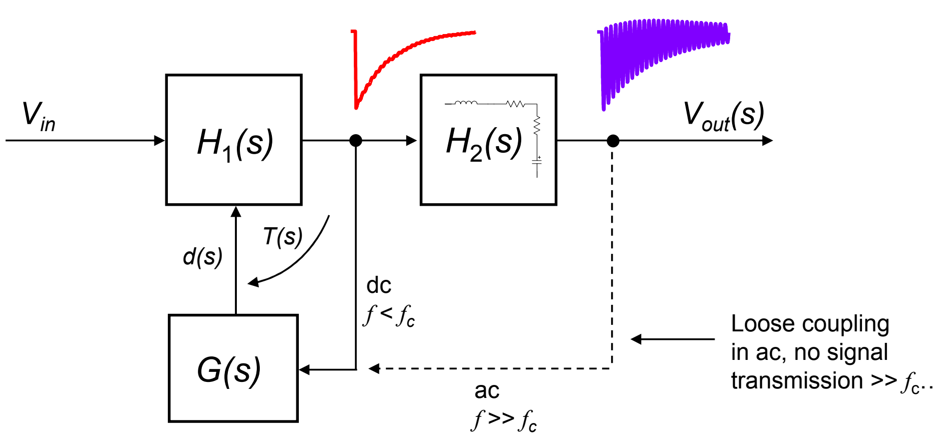
The *RLC* network rings alone…

- ❑ H_1 is stable per Bode analysis, but H_2 is out of the loop…



- ❑ The dc is fed back via the loop but not the ac…
- ❑ Oscillations are NOT due to the loop!

"Fast Analytical Techniques for Electrical and Electronic Circuits"
Vatché Vorpérian, Cambridge Press, 2002

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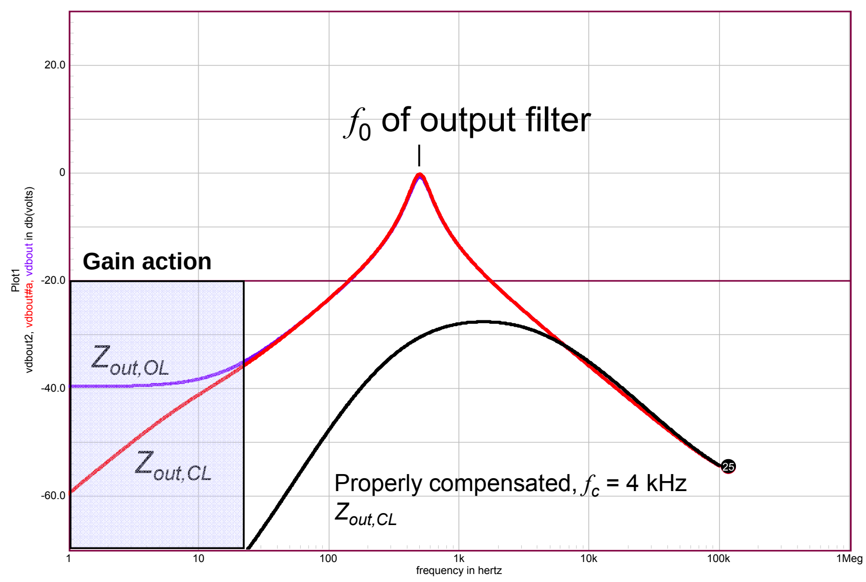
There is no gain to compensate the peaking!

- No gain when the resonance occurs: the *RLC* network runs open loop



- The system cannot reduce the Q at the resonant frequency

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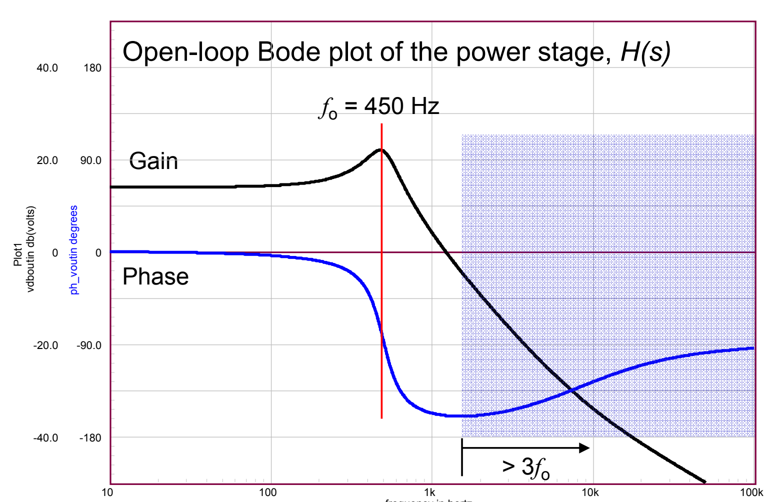
The crossover must be above f_0

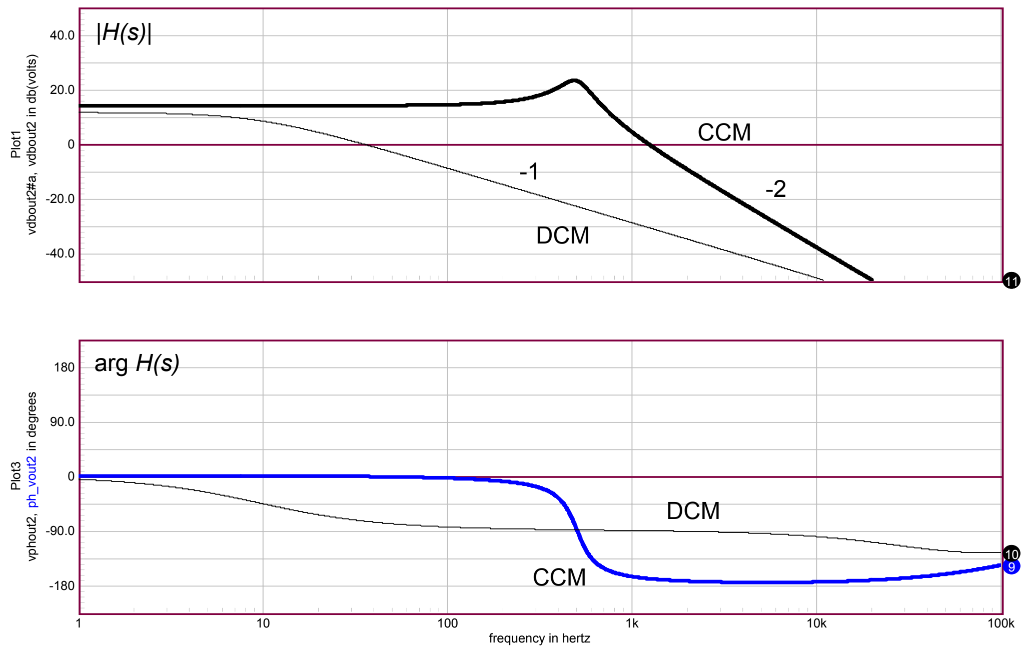
- There still must be gain at the resonance to damp the filter
- f_c should be far from f_0 to reduce phase stress at resonance



Place f_c at least three times above resonance!

CCM to DCM, the transfer function changes

- In DCM, the buck voltage-mode becomes a first order system



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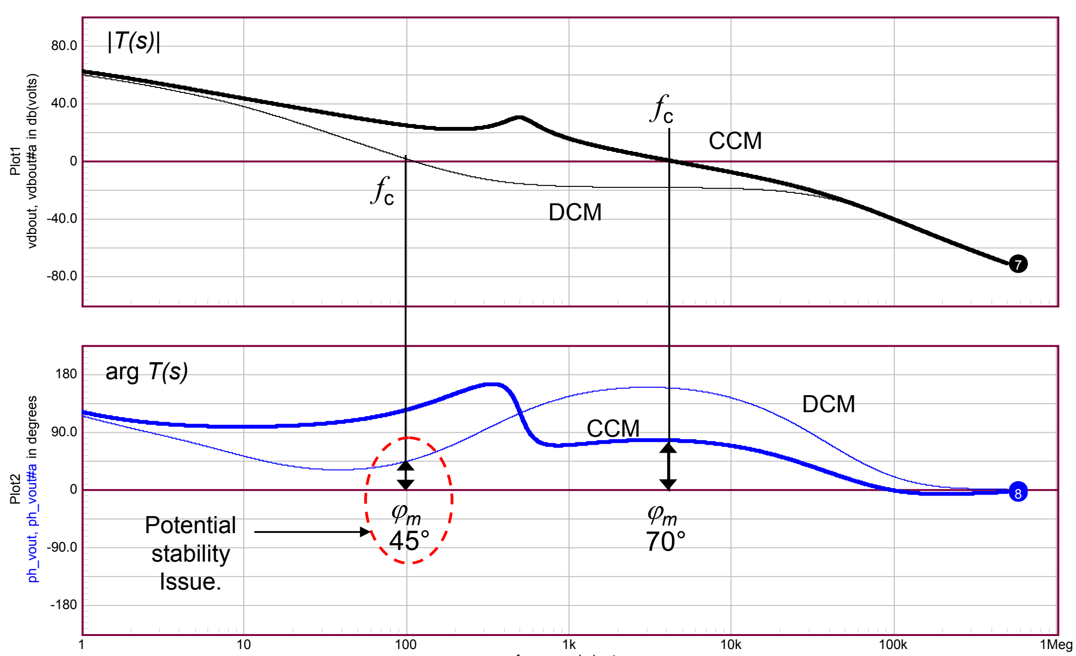
CCM to DCM, the transfer function changes

- In DCM, the compensated system becomes slower than in CCM



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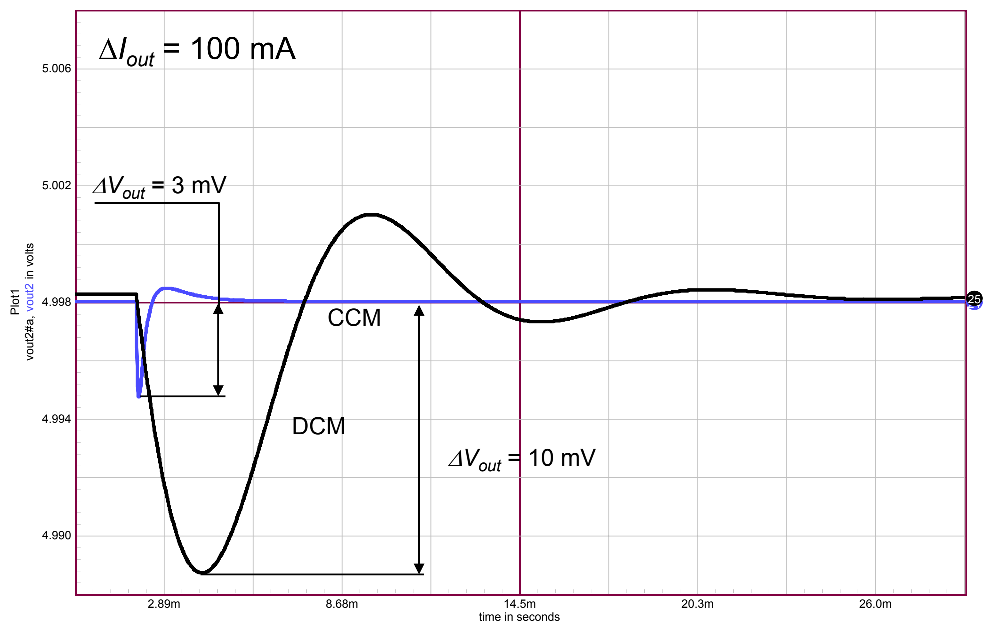
The DCM response is slower than in CCM

❑ A 100-Hz crossover in DCM versus a 4-kHz crossover in CCM



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General methods for compensation

❑ For a buck in CCM voltage-mode:

- Place a double zero at the LC_{out} resonance
- If the ESR zero f_z is below f_c, put a pole $f_{p1}=f_z$
- If the ESR zero f_z is above f_c, put a pole $f_{p1}=F_{sw}/2$
- Put a second pole f_{p2} at $f_{p2}=F_{sw}/2$

$$\frac{V_{out}(s)}{V_{err}(s)}=\frac{V_{in}}{\boxed{V_{peak}}}K_c\frac{\left(1+\frac{s}{\omega_{z1}}\right)}{1+\frac{s}{Q\omega_0}+\left(\frac{s}{\omega_0}\right)^2}$$

This zero is brought by the output capacitor ESR

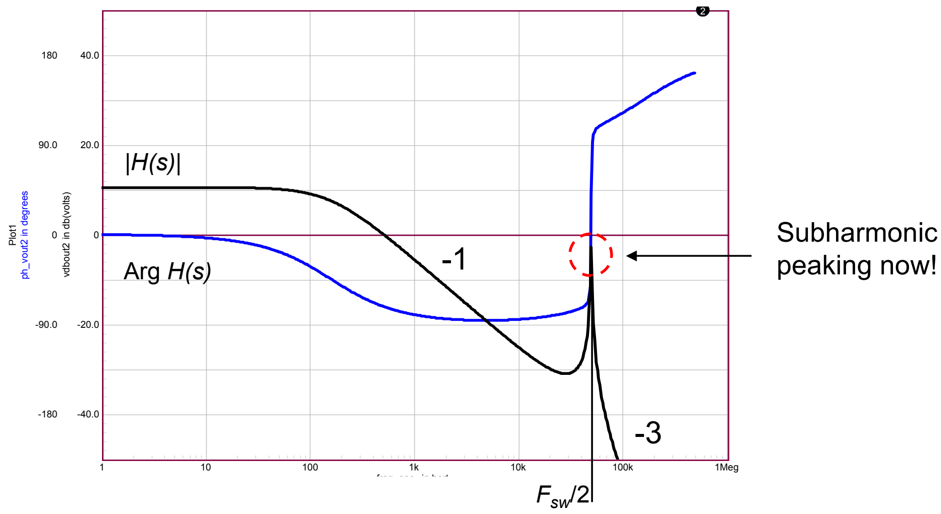
PWM gain

$$K_c=\frac{R_{load}}{r_{Lf}+R_{load}}\qquad \omega_{z1}=\frac{1}{r_{Cf}C}\qquad \omega_0=\frac{1}{\sqrt{LC\frac{R+r_{Cf}}{R+r_{Lf}}}}\qquad Q=\frac{1}{\frac{Z_o}{r_{Lf}+R_{load}}+\frac{r_{Cf}+r_{Lf}\parallel R_{load}}{Z_o}}$$

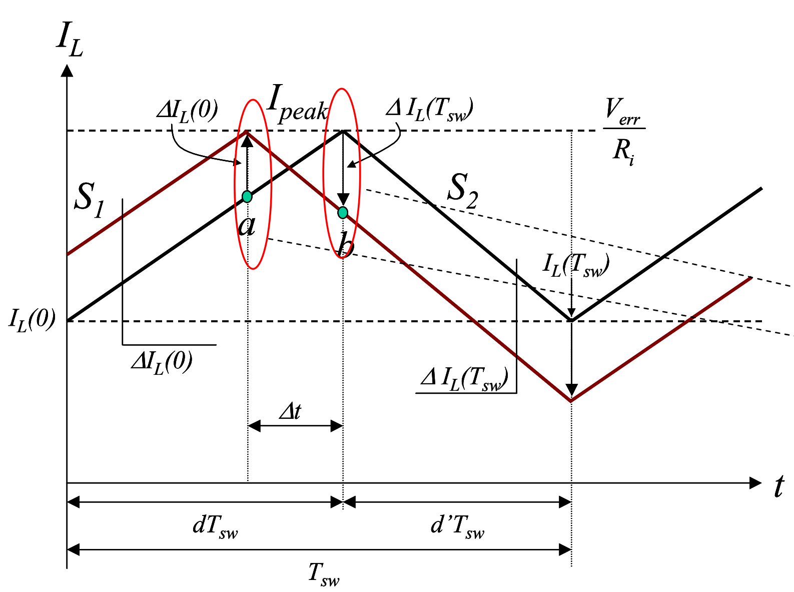
Type 3

The buck in current-mode

- ❑ A CCM current-mode converter acts as a third order system
- ❑ It looks like a first order system at $f_c << F_{sw}/2$
- ➢ No *LC* peaking anymore!
- ➢ But a subharmonic peaking at $F_{sw}/2$ now appears!



CCM operation, current instabilities



$$I_{peak} = a + S_1 \Delta t$$

$$b = I_{peak} - S_2 \Delta t$$

Solving Δt

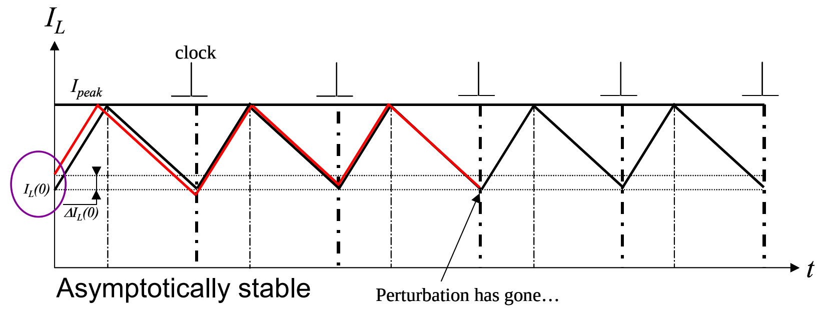
$$\frac{I_{peak} - a}{S_1} = \frac{I_{peak} - b}{S_2}$$

$$\frac{\boxed{\Delta I_L(0)}}{S_1} = \frac{\boxed{\Delta I_L(T_{sw})}}{S_2}$$

$$\frac{S_2}{S_1} = \frac{d}{d'}$$

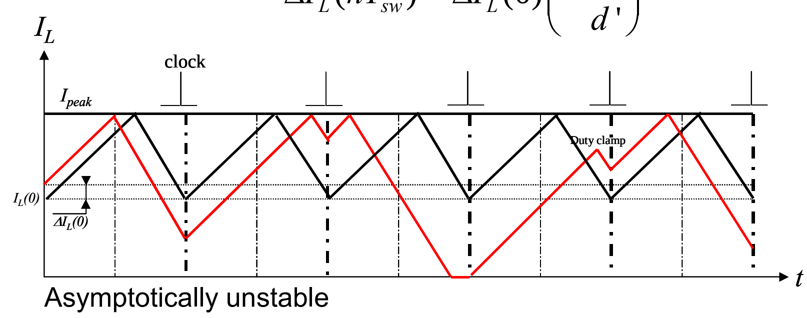
$$\Delta I_L(nT_{sw}) = \Delta I_L(0)\left(-\frac{d}{d'}\right)^n$$

CCM operation, current instabilities



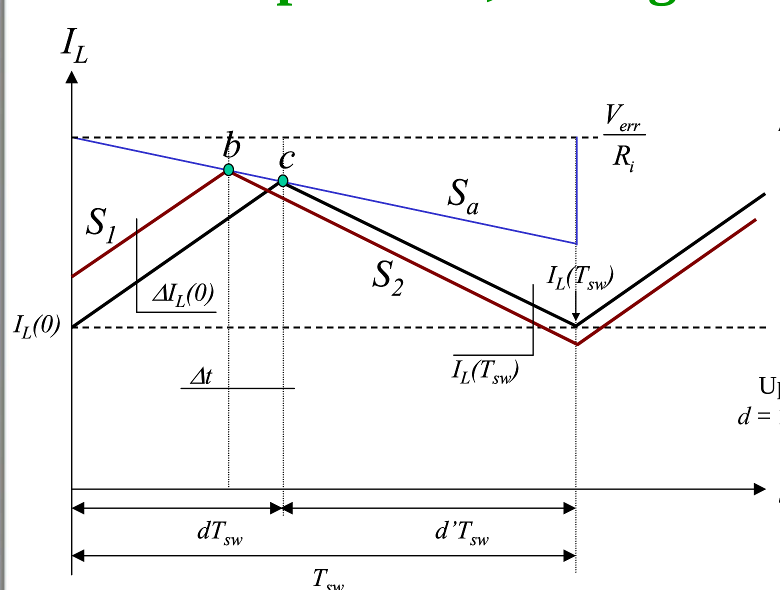
Duty-cycle < 50%

$$\Delta I_L(nT_{sw}) = \Delta I_L(0)\left(-\frac{d}{d'}\right)^n$$



Duty-cycle > 50%

CCM operation, curing current instabilities



$$\Delta I_L(nT_{sw}) = \Delta I_L(0)\left[-\frac{1-\frac{S_a}{S_2}}{\frac{d'}{d}+\frac{S_a}{S_2}}\right]^n = \Delta I_L(0)(-a)^n$$

Must stay below 1

Up to d = 100%

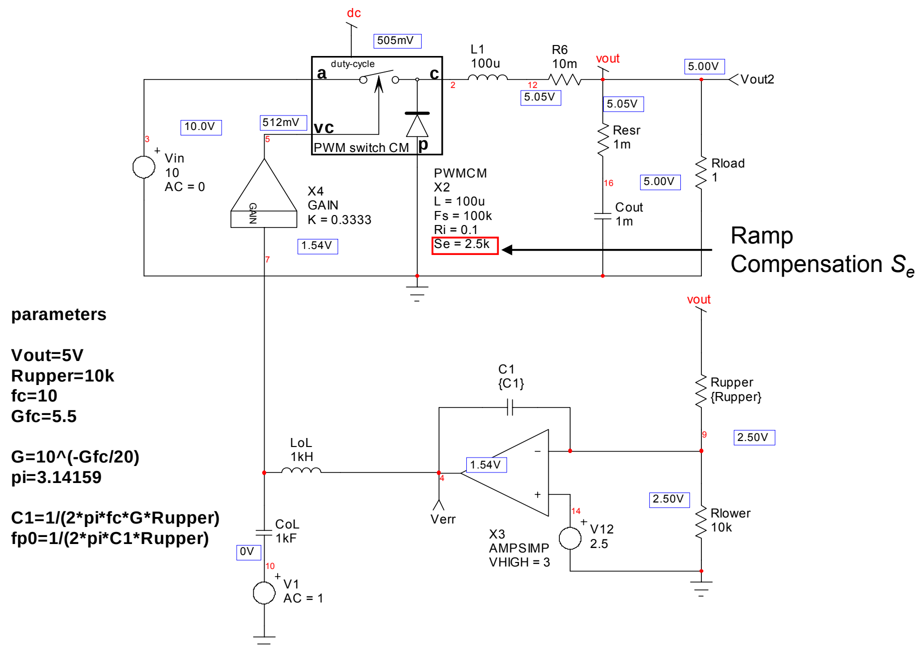
$$\left|\frac{1-\frac{S_a}{S_2}}{0+\frac{S_a}{S_2}}\right| < 1$$

Inject ramp compensation

$$\boxed{S_a > 50\% S_2}$$

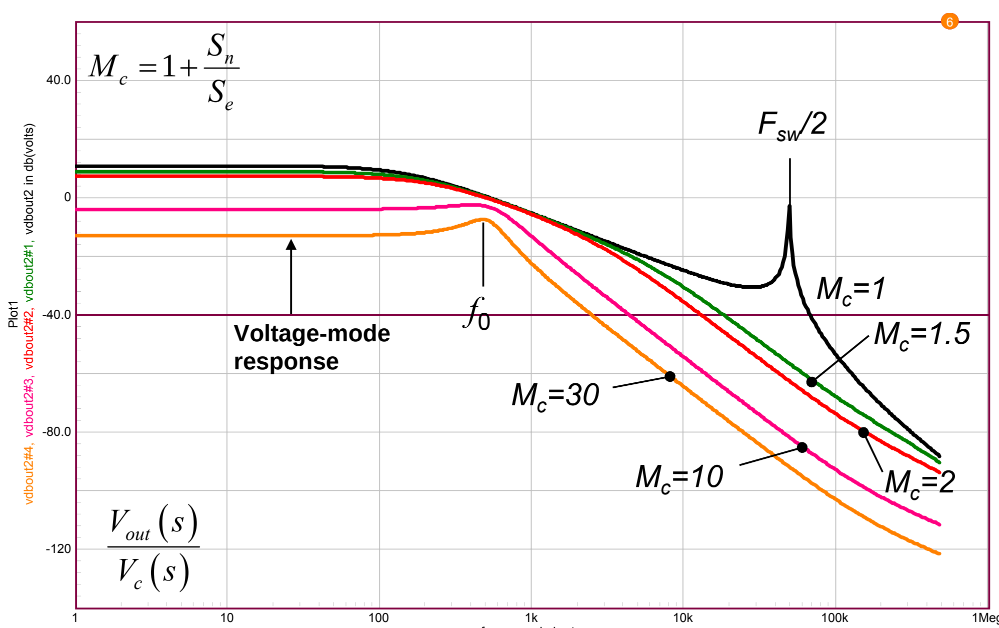
A buck CCM in current-mode

- ❑ A SPICE model can predict subharmonic instabilities

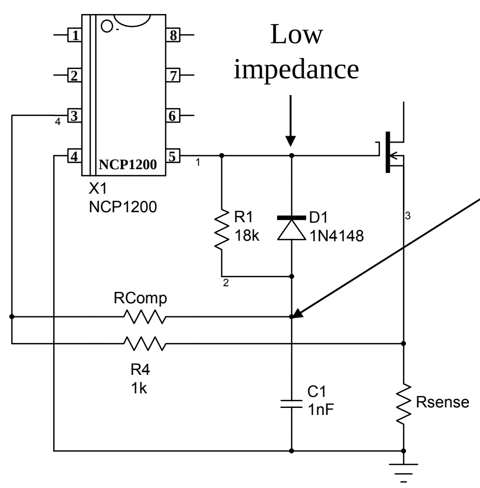


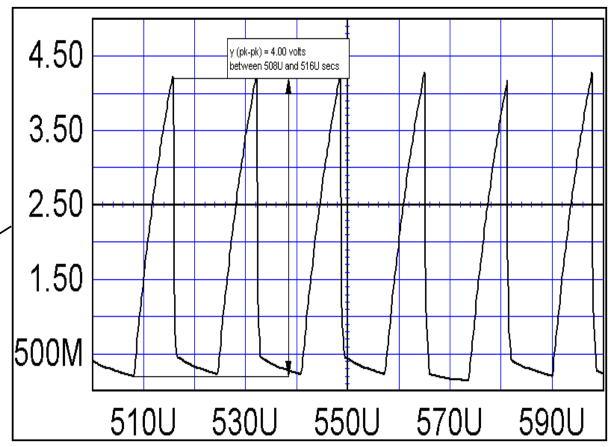
Injecting ramp damps the double pole

- ❑ Too much ramp turns the converter into voltage-mode!



The right way to inject ramp compensation



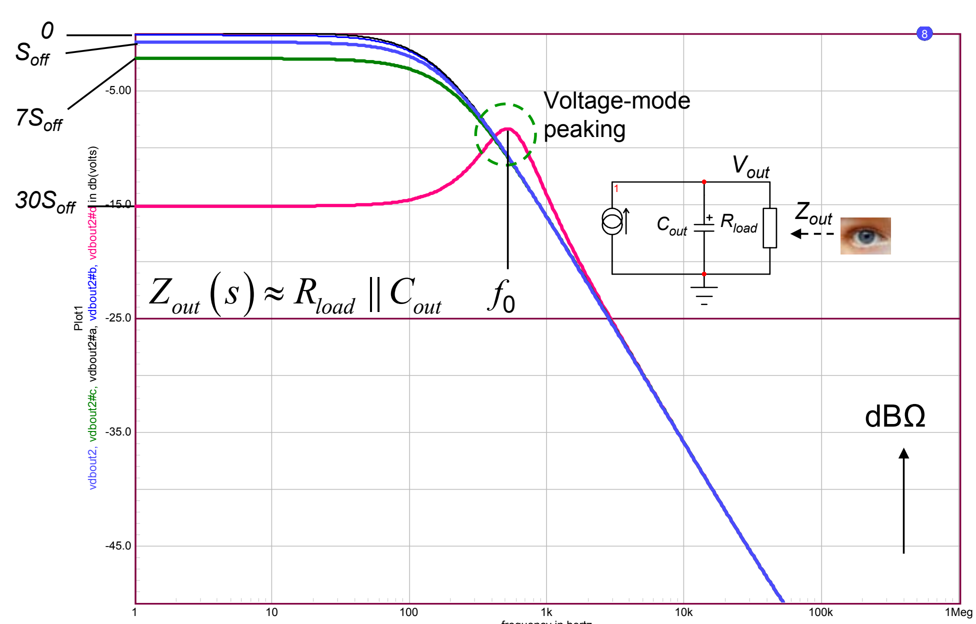


$$R_{comp} = 1k \cdot \frac{S}{S' \cdot M}$$

S = generated ramp
S' = reflected sensed ramp
M = amount of ramp, 0.5-0.75

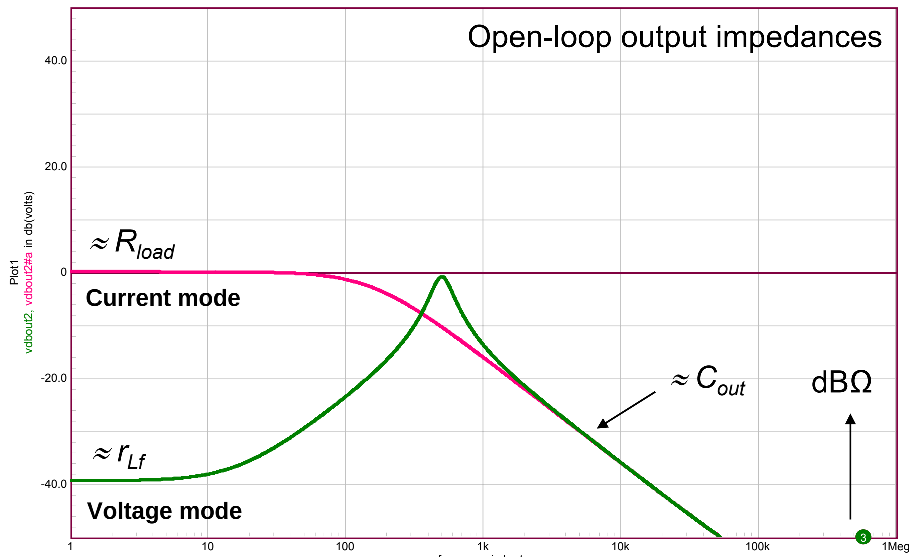
The open-loop impedance of a CM converter

- ❑ The current-mode output impedance depends on the ramp level



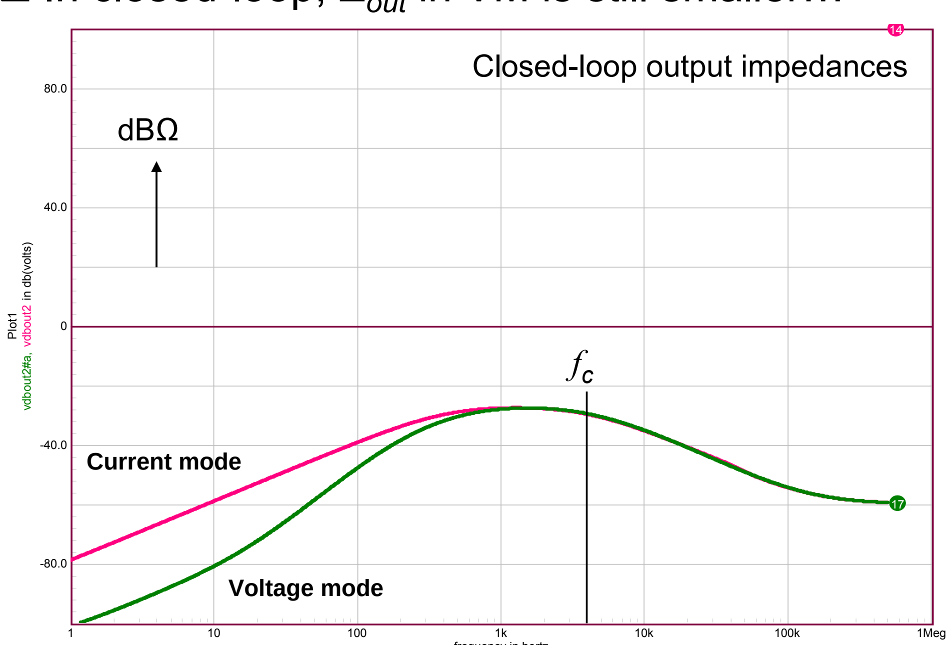
The closed-loop output impedance of CM and VM

- Z_{out} CM is naturally larger than Z_{out} VM



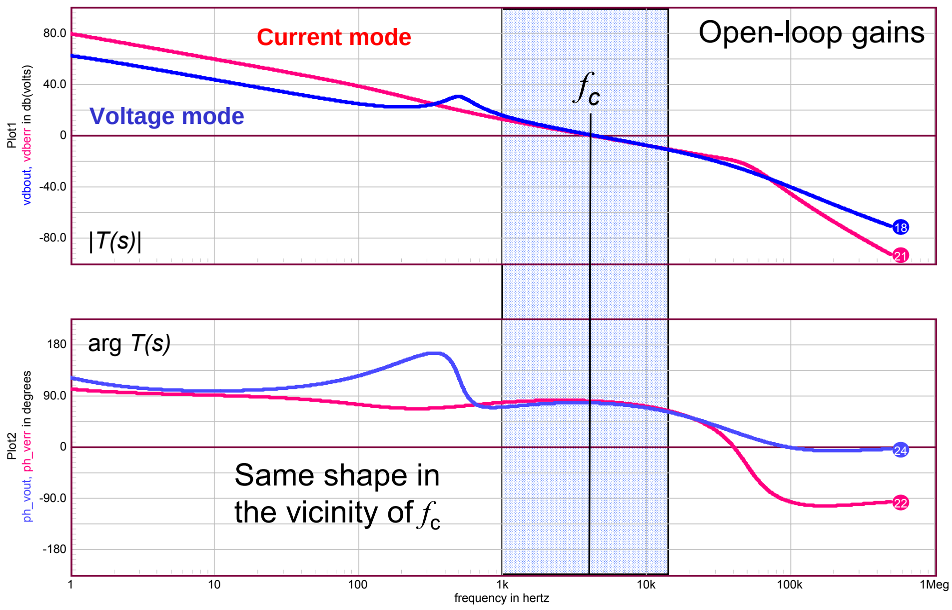
The closed-loop output impedance of CM and VM

- In closed-loop, Z_{out} in VM is still smaller…



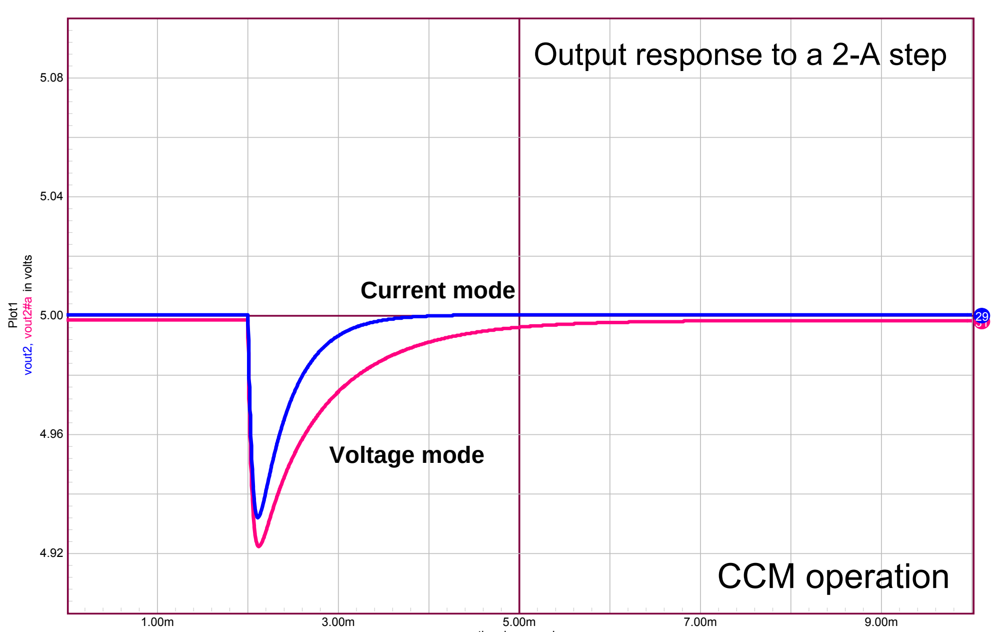
The closed-loop output impedance of CM and VM

❑ The dc gain in VM is smaller because of zeros



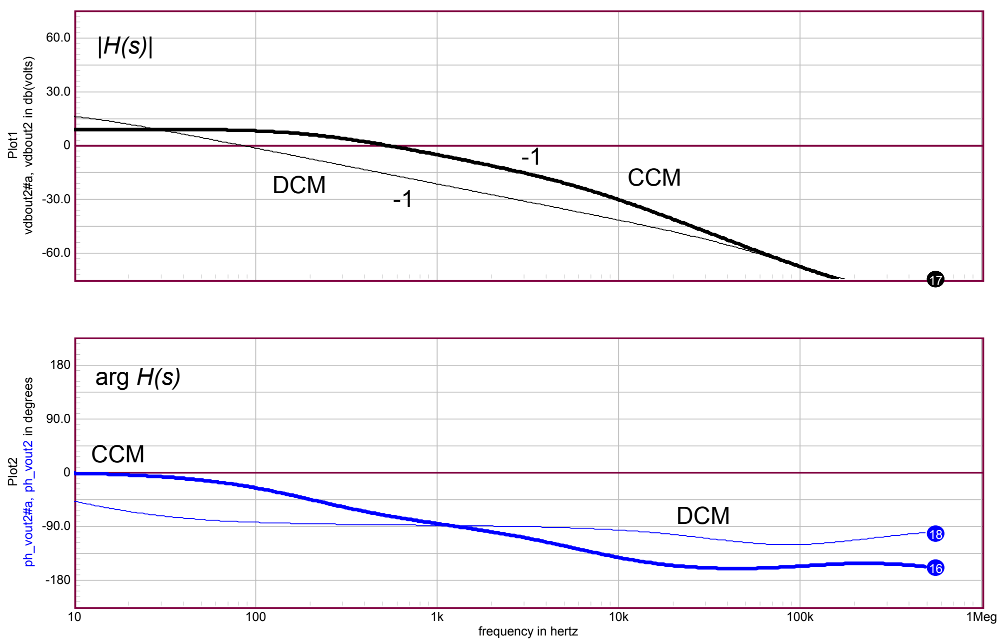
With similar crossover frequency…

❑ The voltage mode is slightly slower than current-mode



Transition from CCM to DCM in current-mode

- ❑ Current-mode remains a 1st order system in both cases



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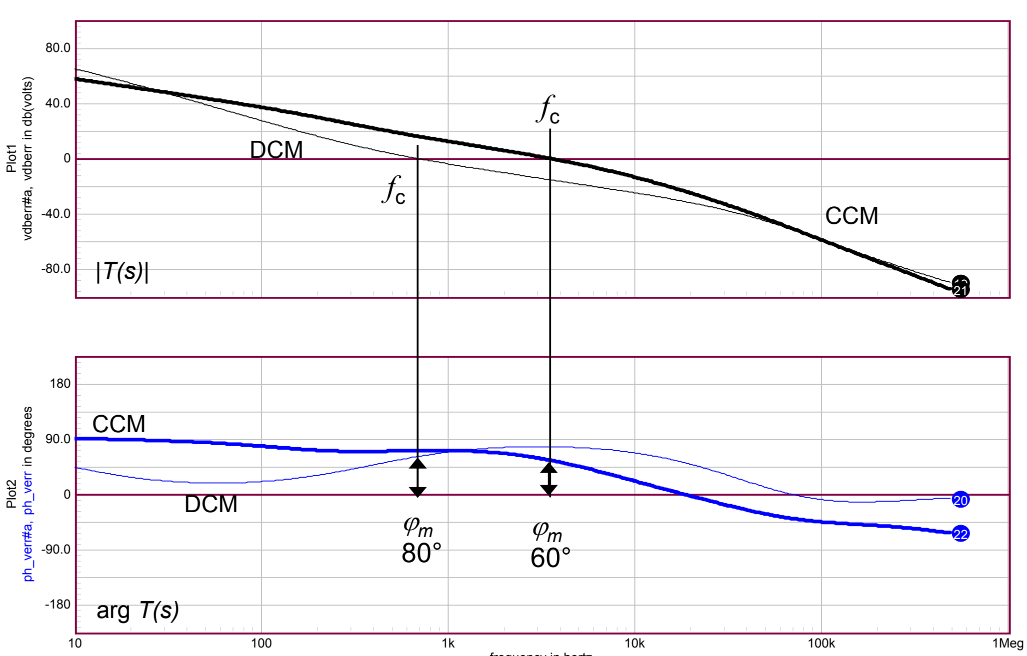
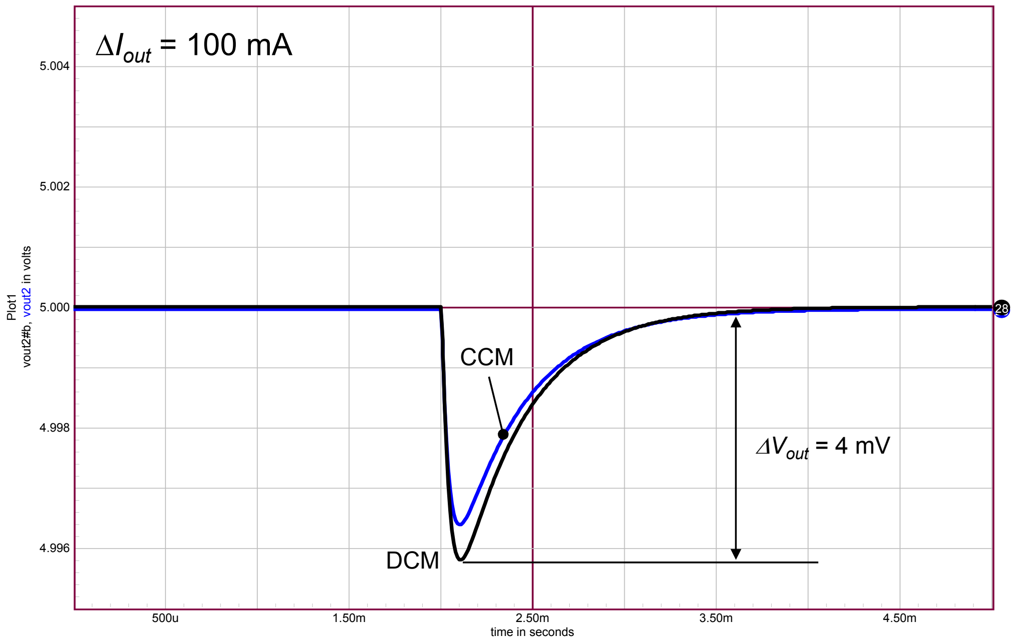
Transition from CCM to DCM in current-mode

- ❑ Once compensated, f_c still changes but φ_m is still ok



The DCM response is not too much affected

❑ The step response is almost unchanged…



The k factor in an automated type 2

❑ Poles and zeros boost the phase at the crossover freq.

❑ How to place poles/zeros for a precise boost at f_c?

✓ Use the k factor technique introduced by Dean Venable

From a 1 zero/ 1 pole compensation circuit, we have:

$$\arg\left(T\left(f_c\right)\right) = boost = \arg\left(\frac{1+\dfrac{f_c}{f_{z0}}}{1+\dfrac{f_c}{f_{p0}}}\right) = \arctan\left(\frac{f_c}{f_{z0}}\right) - \arctan\left(\frac{f_c}{f_{p0}}\right)$$

If we place one pole at kf_c and one zero at f_c/k, we have:

$$boost = \arctan\left(k\right) - \boxed{\arctan\left(\frac{1}{k}\right)}$$

$$\arctan\left(x\right) + \boxed{\arctan\left(\frac{1}{x}\right)} = 90° \longrightarrow k = \tan\left(\frac{boost}{2} + 45\right)$$

The k factor in an automated type 2

- Suppose we have the following specs:
  - f_c = 1 kHz, |*H(1kHz)*|= –10 dB
  - Arg *H(1kHz)* = -100°
  - φ_m = 70° $BOOST = \varphi_m - \arg H(f_c) - 90° = 70° + 100 - 90 = 80°$
  - For a 80-° phase boost, select a type 2
- Calculate *k*

$$k = \tan\left(\frac{80}{2} + 45\right) = 11.4$$

- Place a zero at 1k/11.4 = 90 Hz
- Place a pole at 1k×11.4 = 11.4 kHz
- Adjust mid-gain to reach 10 dB@1 kHz

Type 2

The k factor in an automated type 2

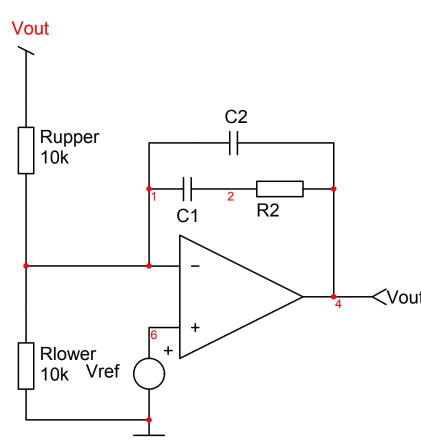
- Calculate the elements for a type 2 amplifier

$$C_2 = \frac{1}{2\pi f_c G k R_{upper}} = \frac{1}{6.28 \times 1k \times 3.16 \times 11.4 \times 10k} = 442\ pF$$

$$C_1 = C_2\left(k^2 - 1\right) = 442p \times \left(11.4^2 - 1\right) = 57\ nF$$

$$R_2 = \frac{k}{2\pi f_c C_1} = \frac{11.4}{6.28 \times 1k \times 57n} = 31.8\ k\Omega$$

$$G = 10^{\frac{Gf_c}{20}} = 10^{10/20} = 3.16$$



D. Venable, "The k-factor: a new mathematical tool for stability analysis and synthesis", proceedings of Powercon 10, 1983, pp. 1-12

Type 2

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The k factor in an automated type 2

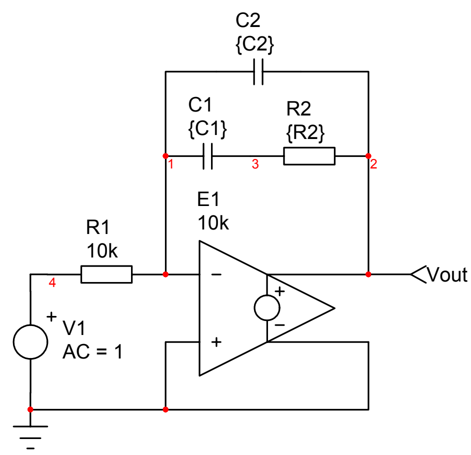
❑ Macros can be used to automate the calculation

parameters

Rupper=10k
fc=1k
pm=70
Gfc=-10
pfc=-100

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

fp=1/(2*pi*R2*C2)
fz=1/(2*pi*R2*C1)



results

```
**************************
*  DEFAULT PARAMETERS
*
**************************

***** mainckt

RUPPER = 1.000e+004
FC = 1.000e+003
PM = 7.000e+001
GFC = -1.000e+001
PFC = -1.000e+002
G = 3.162e+000
BOOST = 8.000e+001
PI = 3.142e+000
K = 1.143e+001
C2 = 4.403e-010
C1 = 5.709e-008
R2 = 3.187e+004
FP = 1.134e+004
FZ = 8.749e+001
```

Type 2

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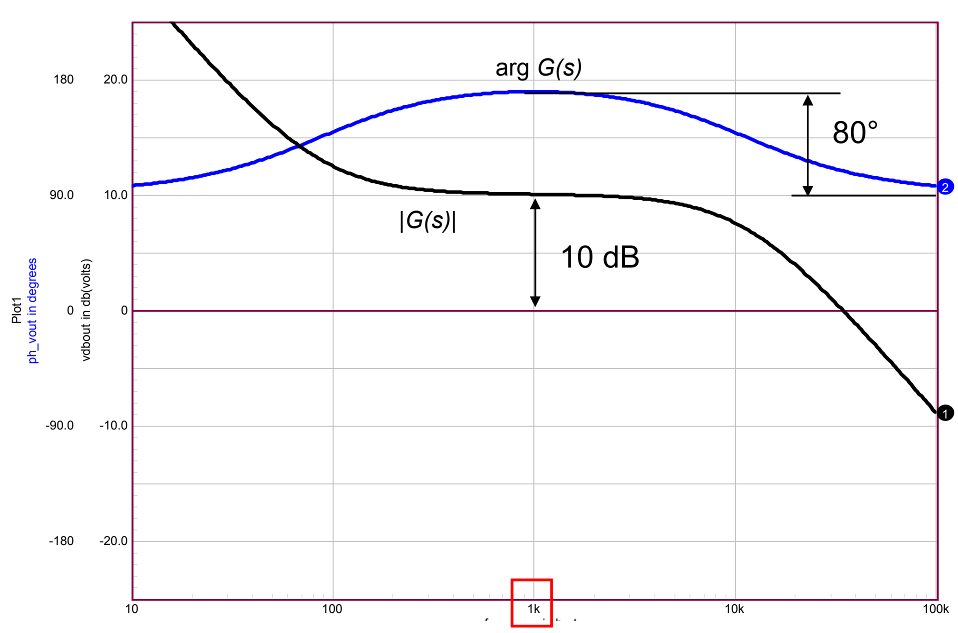
The k factor in an automated type 2

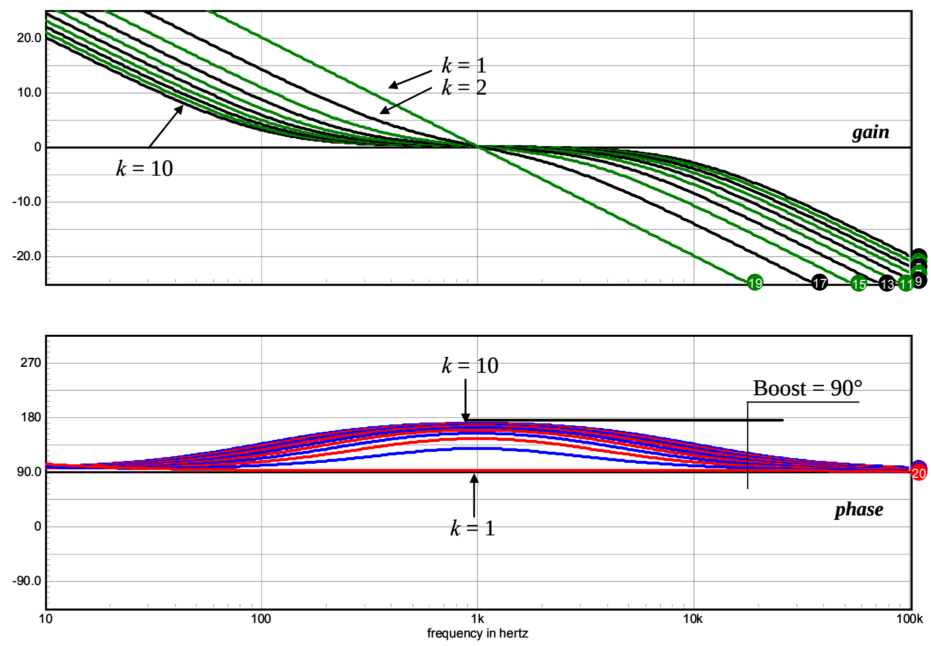
❑ The numbers exactly match the predictions



Type 2

The k factor in an automated type 2

- ❑ Adjusting k modulates the resulting phase boost



Type 2

The k factor in an automated type 3

- ❑ 2 poles and 2 zeros give a higher boost at crossover
- ❑ These 2 poles and 2 zeros are coincident
- ❑ How to place them for a precise boost at f_c?

From a 1 zero pair/ 1 pole pair compensation circuit, we have:

$$\arg\left(T\left(f_c\right)\right) = boost = \arg\left(\frac{\left(1+\dfrac{f_c}{f_{z0}}\right)^2}{\left(1+\dfrac{f_c}{f_{p0}}\right)^2}\right) = 2\arctan\left(\frac{f_c}{f_{z0}}\right) - 2\arctan\left(\frac{f_c}{f_{p0}}\right)$$

If we place two poles at $\sqrt{k}\,f_c$ and two zeros at $f_c/\sqrt{k}$,we have:

$$boost = 2\left[\arctan\left(\sqrt{k}\right) - \arctan\left(\frac{1}{\sqrt{k}}\right)\right]$$

$$boost = 2\left[\arctan\left(\sqrt{k}\right) + \arctan\sqrt{(k)} - 90\right] = 4\arctan\sqrt{(k)} - 180 \longrightarrow k = \left[\tan\left(\frac{boost}{4} + 45\right)\right]^2$$

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The k factor in an automated type 3

- Suppose we have the following specs:
  - f_c = 1 kHz, $|H(1kHz)|$= –15 dB
  - Arg $H(1kHz)$ = -140°
  - φ_m = 70° $BOOST = \varphi_m - \arg H\left(f_c\right) - 90° = 70° + 140 - 90 = 120°$
  - For a 120-° phase boost, select a type 3
- Calculate k

$$k = \left[\tan\left(\frac{120}{4} + 45\right)\right]^2 = 13.9$$

- Place a zero at 1k/3.7 = 268 Hz
- Place a pole at 1k×3.7 = 3.7 kHz
- Adjust mid-gain to reach 15 dB@1 kHz

Type 3

ON Semiconductor® Selection. Service. Support.

The k factor in an automated type 3

- Calculate the elements for a type 3 amplifier

$$C_2 = \frac{1}{2\pi f_c G R_1} = \frac{1}{6.28 \times 1k \times 5.6 \times 10k} = 2.85\ nF$$

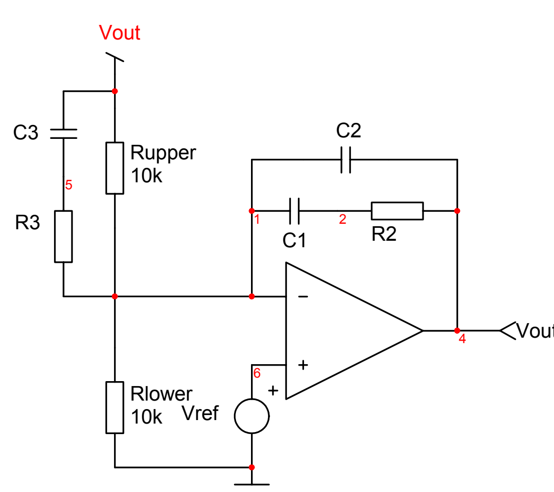
$$C_1 = C_2\left(k-1\right) = 2.85n \times \left(13.9-1\right) = 36.7\ nF$$

$$R_2 = \frac{\sqrt{k}}{2\pi f_c C_1} = \frac{\sqrt{13.9}}{6.28 \times 1k \times 36.7n} = 16.2\ k\Omega$$

$$R_3 = \frac{R_1}{k-1} = \frac{10k}{13.9-1} = 775\ \Omega$$

$$C_3 = \frac{1}{2\pi f_c \sqrt{k} R_3} = \frac{1}{6.28 \times 1k \times \sqrt{13.9} \times 775} = 55\ nF$$

$$G = 10^{\frac{G_{f_c}}{20}} = 10^{15/20} = 5.6$$



Type 3

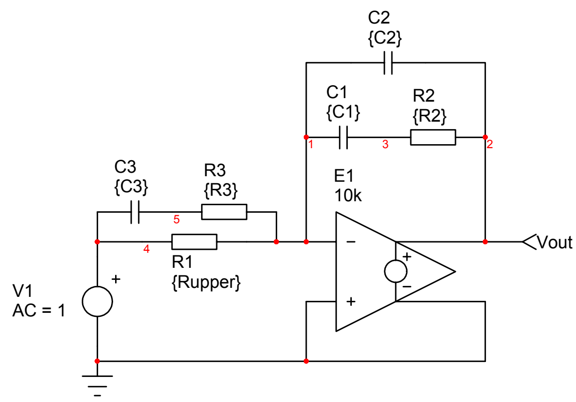
The k factor in an automated type 3

❑Macros can be used to automate the calculation

parameters
Rupper=10k
fc=1k
pm=70
Gfc=-15
ps=-140

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)



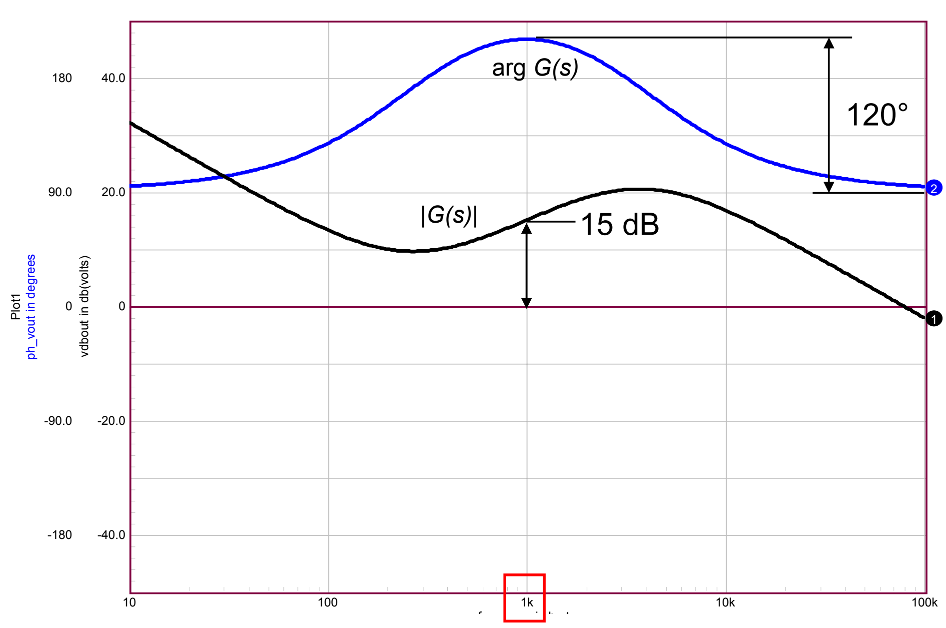
results

```
*************************
*   DEFAULT  PARAMETERS
*
*************************
***** mainckt

RUPPER = 1.000e+004
FC = 1.000e+003
PM = 7.000e+001
GFC = -1.500e+001
PS = -1.400e+002
G = 5.623e+000
BOOST = 1.200e+002
PI = 3.142e+000
K = 1.393e+001
C2 = 2.830e-009
C1 = 3.659e-008
R2 = 1.623e+004
R3 = 7.735e+002
C3 = 5.513e-008
FP1 = 3.464e+003
FP2 = 3.732e+003
FZ1 = 2.680e+002
FZ2 = 2.887e+002
```


The k factor in an automated type 3

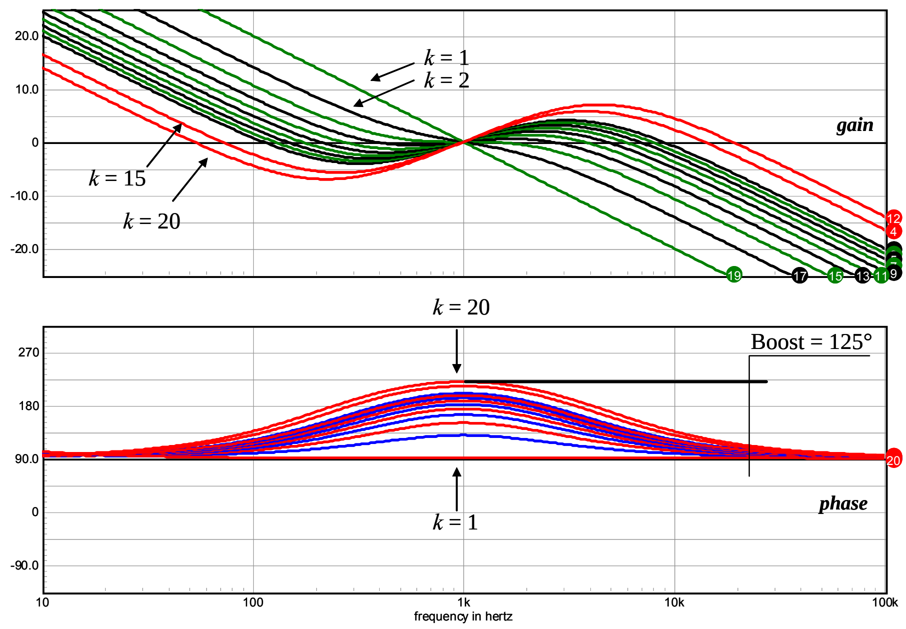
❑ The numbers exactly match the predictions



Type 3

The k factor in an automated type 3

- ❑ Adjusting k modulates the resulting phase boost



Type 3

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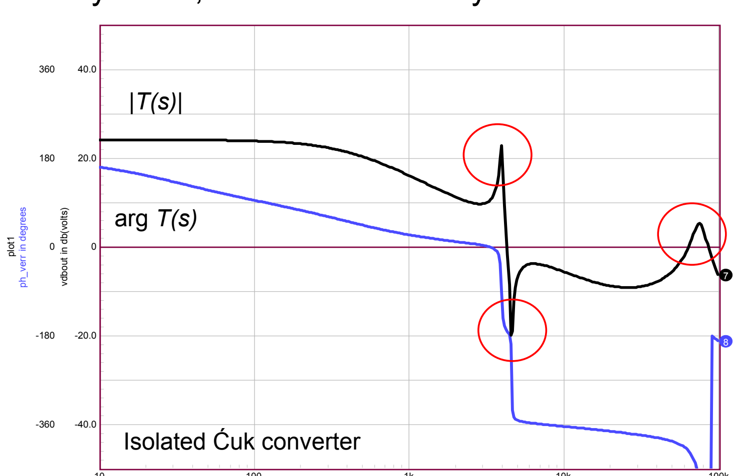
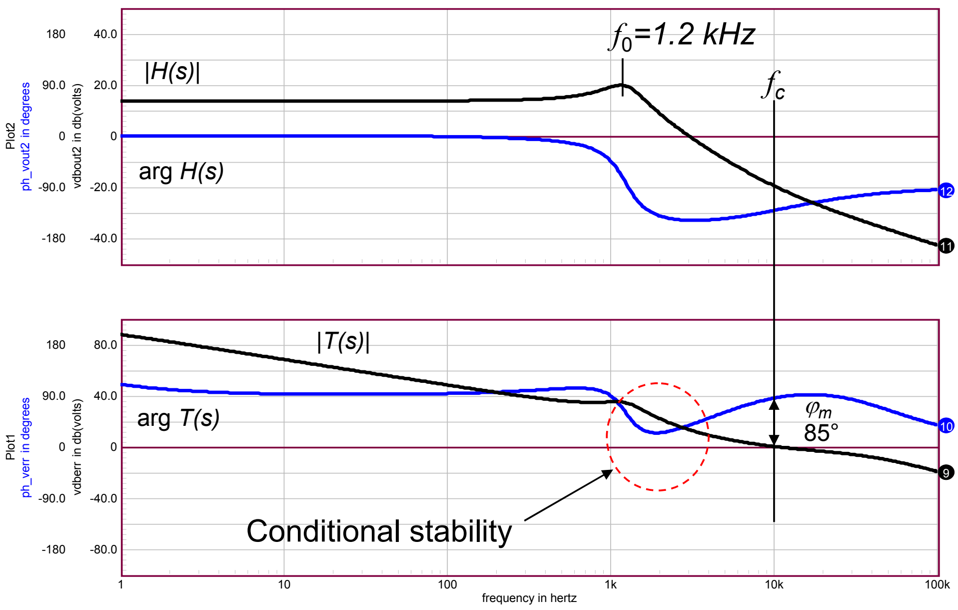
Is the k factor the panacea?

- ❑ The tool is a quick and easy means to stabilize the converter
- ❑ However, it solely focuses on the crossover frequency
- ▪ What is happening before and beyond f_c?
- ➢ The k factor technique is blind to these effects.
- ➢ In resonant systems, conditional stability can occur.



The k factor can lead to conditional stability

- ❑ In CCM voltage-mode, conditional stability occurs



k factor gives

$f_{p1,2}$= 40 kHz

$f_{z1,2}$= 2.3 kHz

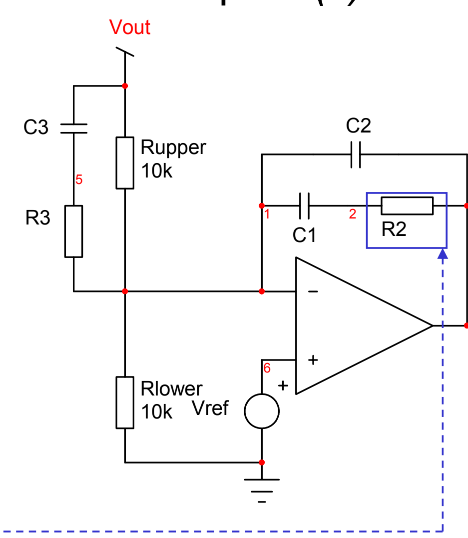
Manual pole/zero placement is a solution

- ❑ Avoid coincident poles and zeros brought by k factor
- ❑ Place them wherever you wish to shape *G(s)*

$$G(s) \approx \frac{R_2}{R_3} \frac{\left(\frac{1}{sR_1C_3}+1\right)\left(\frac{1}{sR_2C_1}+1\right)}{(1+sR_2C_2)\left(\frac{1}{sR_3C_3}+1\right)}$$

$$G(f) \approx \left| \frac{R_2}{R_3} \frac{\left(1+\frac{s_{z1}}{s}\right)\left(1+\frac{s_{z2}}{s}\right)}{\left(1+\frac{s}{s_{p1}}\right)\left(1+\frac{s_{p2}}{s}\right)} \right|$$

$$R_2 = \sqrt{\frac{\left(f_{p1}^2+f_c^2\right)\left(f_{p2}^2+f_c^2\right)}{\left(f_{z1}^2+f_c^2\right)\left(f_{z2}^2+f_c^2\right)}} \frac{Gf_cR_3}{f_{p1}}$$



$$f_{z1} = \frac{1}{2\pi R_1C_3} \qquad f_{z2} = \frac{1}{2\pi R_2C_1}$$

$$f_{p1} = \frac{1}{2\pi R_2C_2} \qquad f_{p2} = \frac{1}{2\pi R_3C_3}$$

Type 3

Manual pole/zero placement is a solution

parameters

Rupper=10k
Rlower=Rupper

fc=10k
Gfc=-20

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

fz1=1k
fz2=1k
fp1=26k
fp2=26k

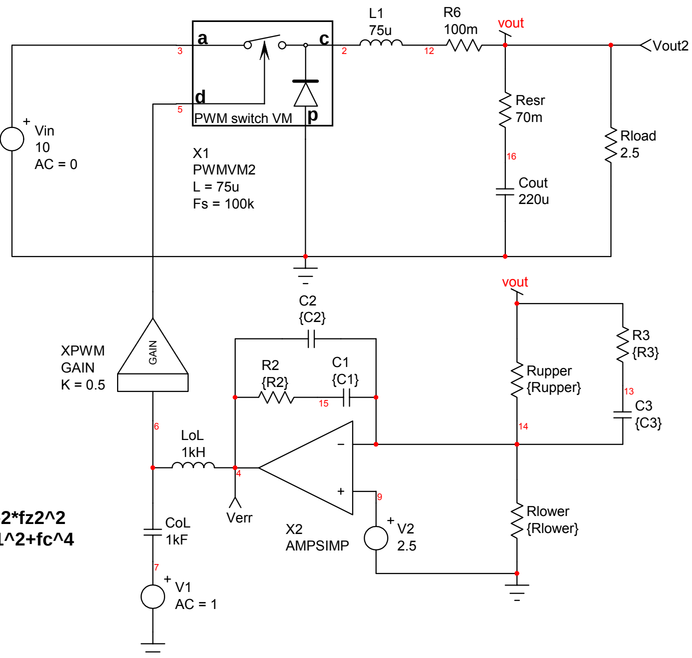
C3=1/(2*pi*fz1*Rupper)
R3 =1/(2*pi*fp2*C3)

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

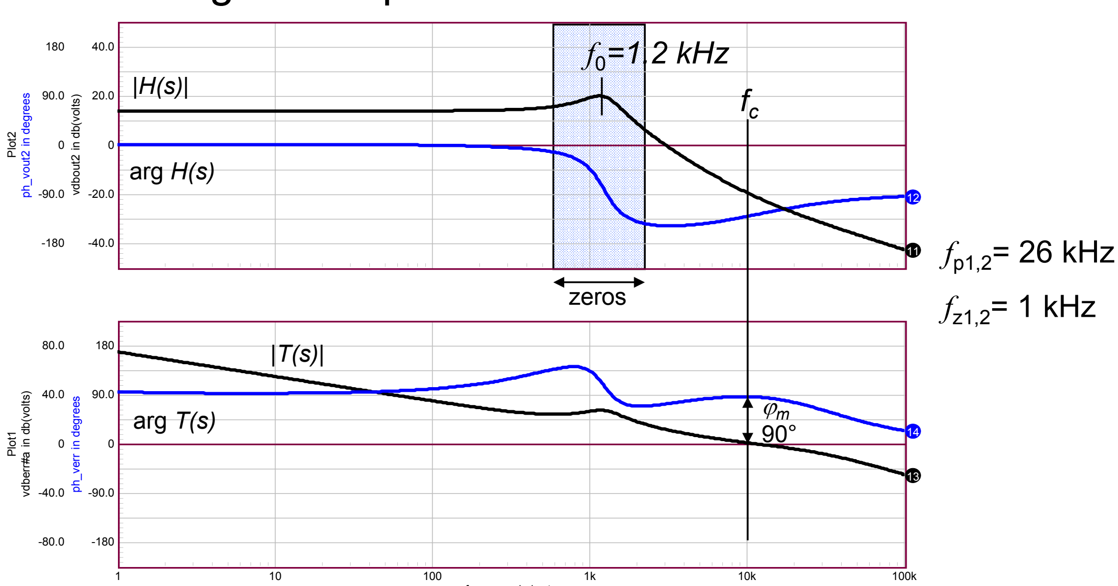
fz1x=1/(2*pi*C1*R2)
fz2x=1/(2*pi*C3*(Rupper+R3))
fp1x=1/(2*pi*(C1*C2/(C1+C2))*R2)
fp2x=1/(2*pi*C3*R3)



- ❑ Automated calculations help iterations!

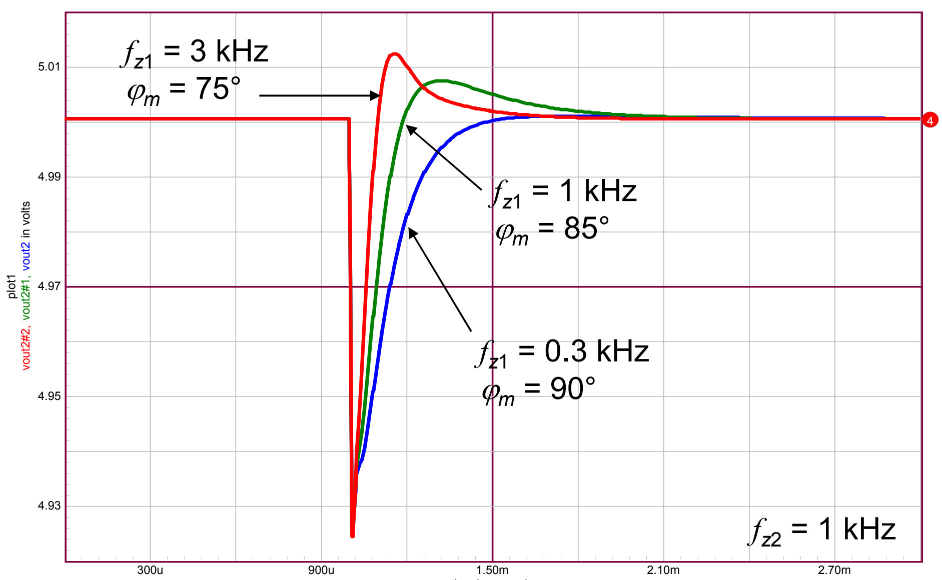
Manual pole/zero placement is a solution

- ❑ Conditional stability is gone
- ❑ Fine tuning is now possible



The zeros position affects the response

- ❑ Transient response changes as one zero position moves

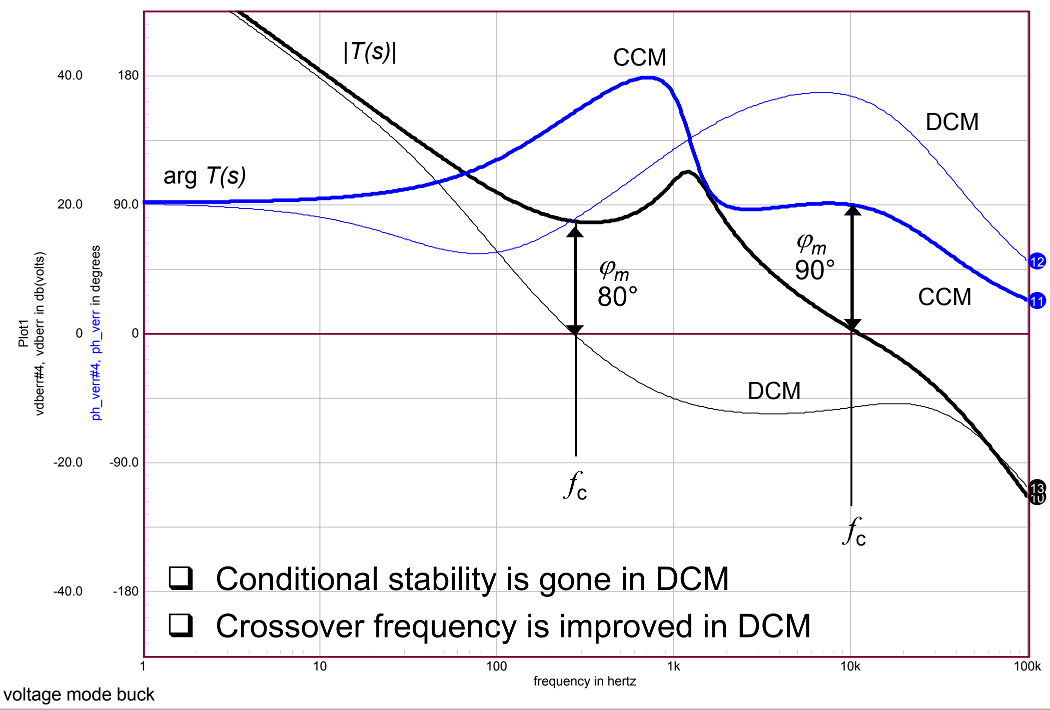


The zeros position affects the response

- ❑ Splitting the zeros can fix DCM stability issue in buck VM

	Frequency	Overshoot	Settling time	Phase margin
f_{z1}	↗	↗	faster	↘
	↘	↘	slower	↗
f_{z2}	↗	↗	faster	↘

By decreasing f_{z1}, DCM stability is improved

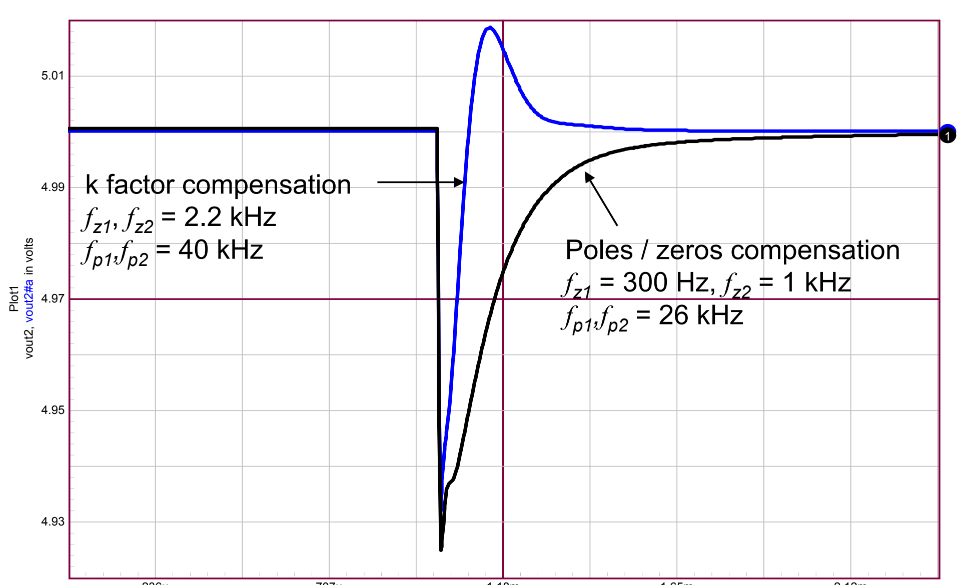


- ❑ Conditional stability is gone in DCM
- ❑ Crossover frequency is improved in DCM

CCM/DCM voltage mode buck

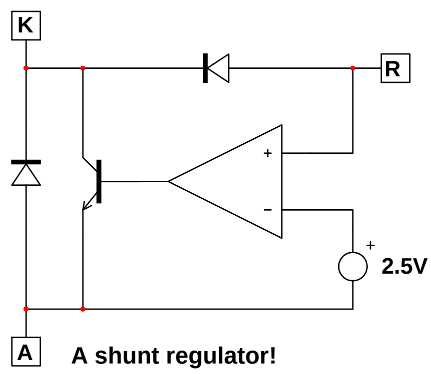
Improving DCM stability slows down the buck

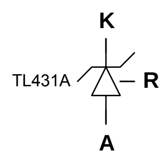
- ❑ f_{z1} moving low, it hampers the response time…

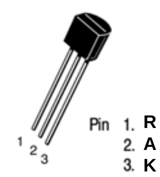


Type 2 and 3 with a TL431

- ❑ Litterature examples use op amps to close the loop.
- ❑ Reality differs as the TL431 is widely implemented.
- ❑ How to adapt type 2 and 3 to a TL431 circuit?
- ❑ Question: who hides behind the TL431 anyway?

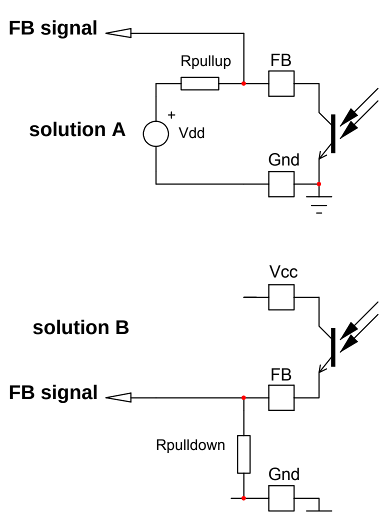


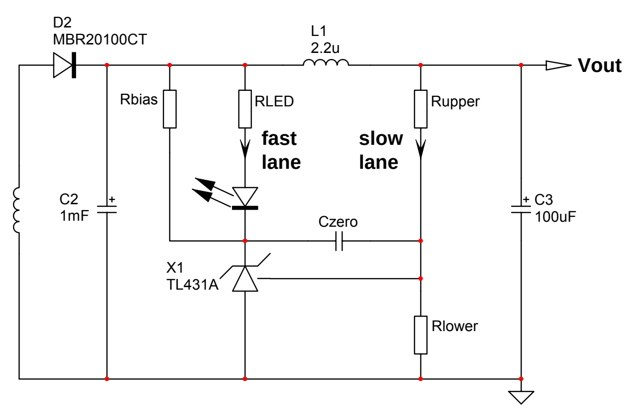




Feedback with a TL431

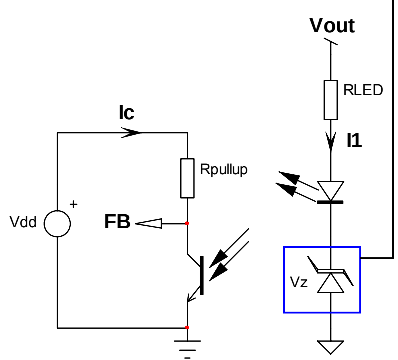
- ❑ A TL431 implements a two-loop configuration





Feedback with a TL431

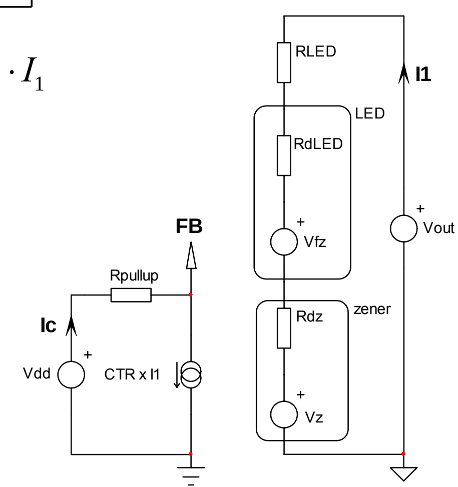
- When C_{zero} is a short circuit, the slow lane is off
- The TL431 turns into a static Zener diode



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

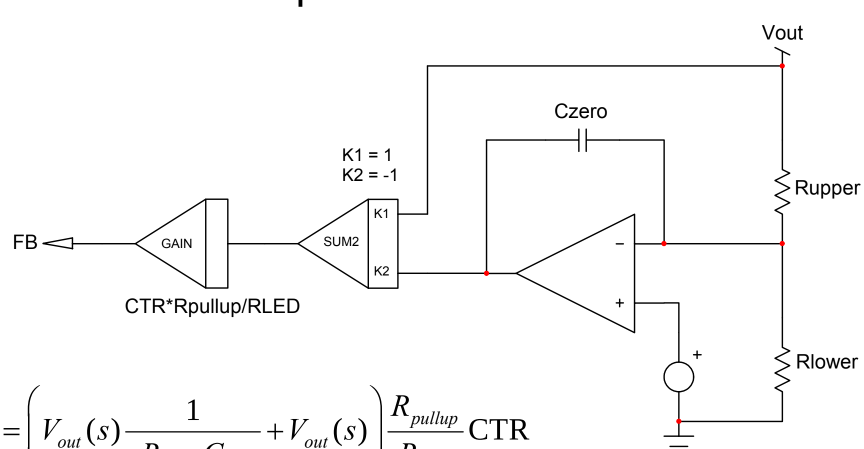
$$I_1 = \frac{V_{out}(s)}{R_{LED}}$$

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



The TL431 and its equivalent schematic

- The fast lane presence adds a zero in the equation



$$V_{FB}(s) = \left(V_{out}(s)\frac{1}{sR_{upper}C_{zero}} + V_{out}(s)\right)\frac{R_{pullup}}{R_{LED}}\text{CTR}$$

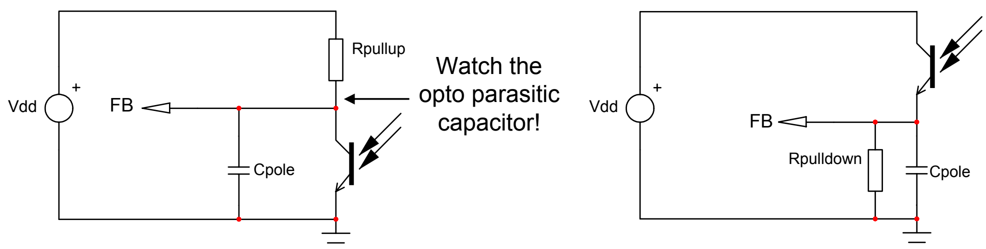
$$\frac{V_{FB}(s)}{V_{out}(s)} = -\left(\frac{1}{sR_{upper}C_{zero}} + 1\right)\frac{R_{pullup}}{R_{LED}}\text{CTR} = -\left(\frac{sR_{upper}C_{zero}+1}{sR_{upper}C_{zero}}\right)\frac{R_{pullup}}{R_{LED}}\text{CTR}$$

No high frequency pole?

Type 2

Adding a pole for a type 2 circuit

❑ The pole is a simple capacitor on the collector



Or on the emitter

$$G(s)=\frac{V_{FB}(s)}{V_{out}(s)}=-\left(\frac{sR_{upper}C_{zero}+1}{sR_{upper}C_{zero}}\right)\left(\frac{1}{1+sR_{pullup}C_{pole}}\right)\frac{R_{pullup}}{R_{LED}}\mathrm{CTR}$$

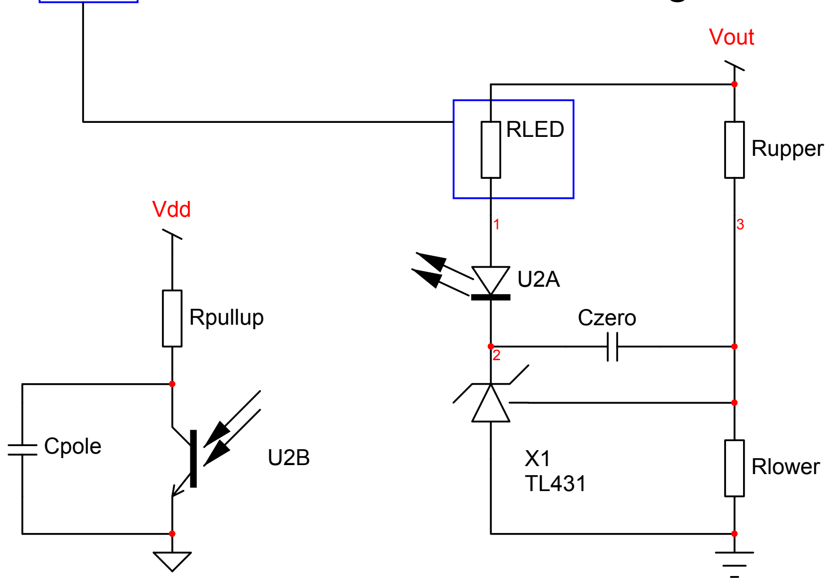
$$f_{po}=\frac{1}{2\pi R_{upper}C_{zero}} \qquad f_z=\frac{1}{2\pi R_{upper}C_{zero}} \qquad G=\frac{R_{pullup}}{R_{LED}}\mathrm{CTR} \qquad f_p=\frac{1}{2\pi R_{pullup}C_{pole}}$$

Pole at the origin | Low frequency zero | Mid-band gain | High frequency pole

Type 2

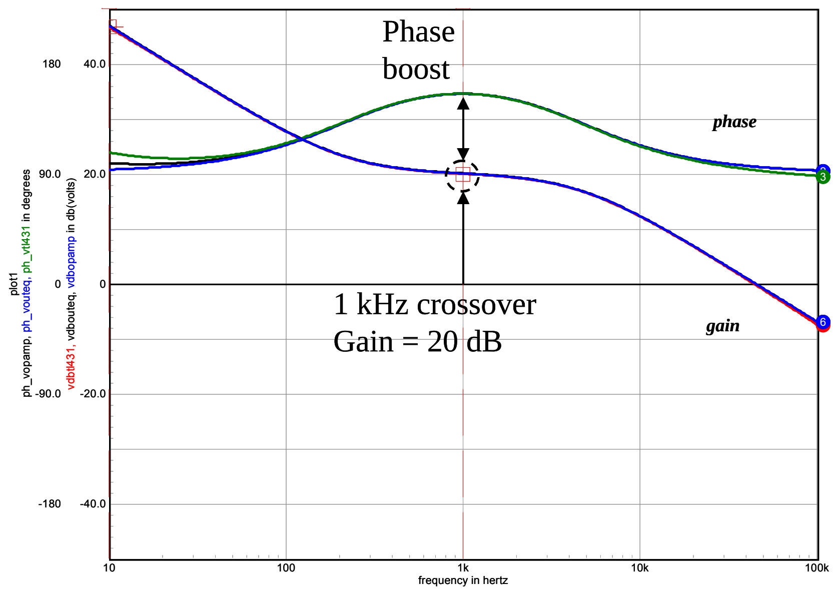
The type 2 final implementation

❑ The LED resistor fixes the mid-band gain



Comparing a type 2 op amp with a TL431

❑ The ac plot shows no significant differences



Type 2

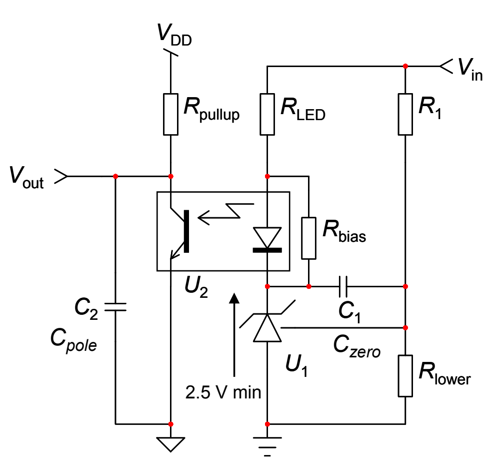
The TL431 type 2 is not a panacea!

❑ Because R_{LED} must also ensures sufficient bias current:

➢ There is an upper resistor limit

➢ The minimum mid-band gain you can obtain is clamped!

➢ Worst case is low V_{out}, e.g. 5 V



$$R_{LED,\max} \le \frac{V_{out} - V_f - V_{TL431,\min}}{V_{dd} - V_{CE,sat}} R_{pullup} \text{CTR}_{\min}$$

$$R_{LED,\max} \le \frac{5-1-2.5}{5-0.3} \times 0.3 \times 20k \le 1.9\ k\Omega$$

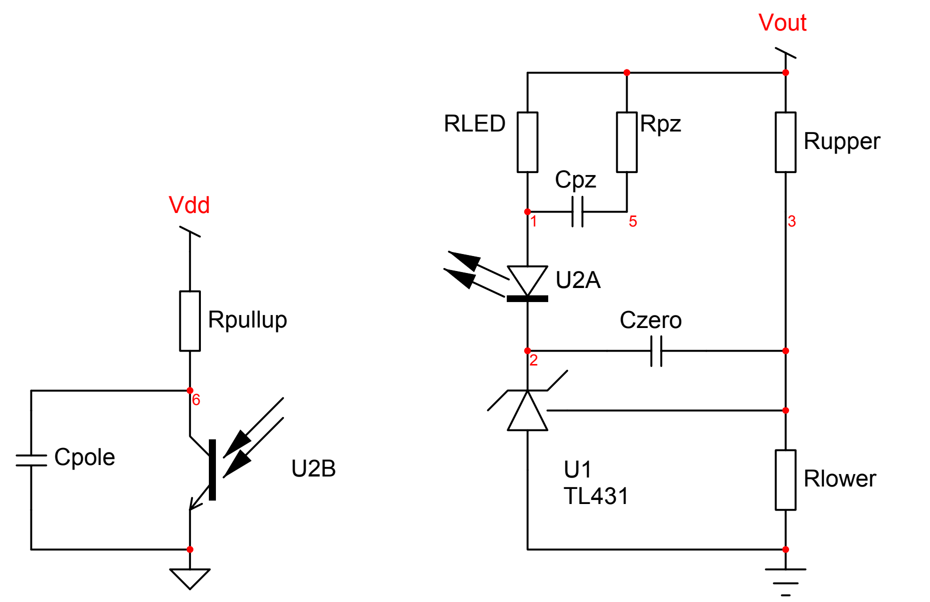
$$G_0 \ge \frac{R_{pullup}}{R_{LED,\max}} \text{CTR}_{\min}$$

Cannot go below this!

$$G_0 \ge \frac{V_{dd} - V_{CE,sat}}{V_{out} - V_f - V_{TL431,\min}} \ge 3.13 \text{ or} \approx \boxed{10\ dB}$$

A type 3 is also possible

- ❑ The type 3 is less flexible given the R_{LED} role



Type 3

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If arg$H(s)$ @ f_c is less than 45°…

- ❑ We can use a true type 1 without gain limits
- ➢ No phase boost!

$$R_{LED,\max} \le \frac{5-1-2.5}{5-0.3}\times 0.3\times 20k \le 1.9\ k\Omega \xrightarrow{\text{20\% margin}} R_{LED,\max} = 0.8\times 1.9\ k\Omega = 1.5\ k\Omega$$

coincident

$$G(s) = -\left(\frac{\boxed{sR_{upper}C_{zero}+1}}{s\dfrac{R_{upper}R_{LED}}{R_{pullup}\mathrm{CTR}}C_{zero}}\right)\left(\frac{1}{\boxed{1+sR_{pullup}C_{pole}}}\right)$$

Origin pole $\longrightarrow$ $\omega_{po} = \dfrac{1}{s\dfrac{R_{upper}R_{LED}}{R_{pullup}\mathrm{CTR}}C_{zero}}$

$$f_{po} = f_c G \longrightarrow C_{pole} = \frac{\mathrm{CTR}}{2\pi f_{po}R_{LED}} \qquad C_{zero} = \frac{R_{pullup}}{R_{upper}}C_{pole}$$

Selection. Service. Support.

The type 3 with a TL431

- R_{LED} affects the gain and the zero position

$$G(s)=\frac{V_{FB}(s)}{V_{out}(s)}=-\left(\frac{sR_{upper}C_{zero1}+1}{sR_{upper}C_{zero1}}\right)\left(\frac{1}{1+sR_{pullup}C_{pole2}}\right)\frac{\left[sC_{pz}\left(R_{LED}+R_{pz}\right)+1\right]}{\left(sR_{pz}C_{pz}+1\right)}\frac{R_{pullup}}{R_{LED}}\mathrm{CTR}$$

$$f_{po}=\frac{1}{2\pi R_{upper}C_{zero1}}$$

$$f_{z1}=\frac{1}{2\pi R_{upper}C_{zero1}}\qquad f_{z2}=\frac{1}{2\pi\left(\boxed{R_{LED}}+R_{pz}\right)C_{pz}}$$

$$G=\frac{R_{pullup}}{\boxed{R_{LED}}}\mathrm{CTR}$$

$$f_{p1}=\frac{1}{2\pi R_{pz}C_{pz}}\qquad f_{p2}=\frac{1}{2\pi R_{pullup}C_{pole2}}$$

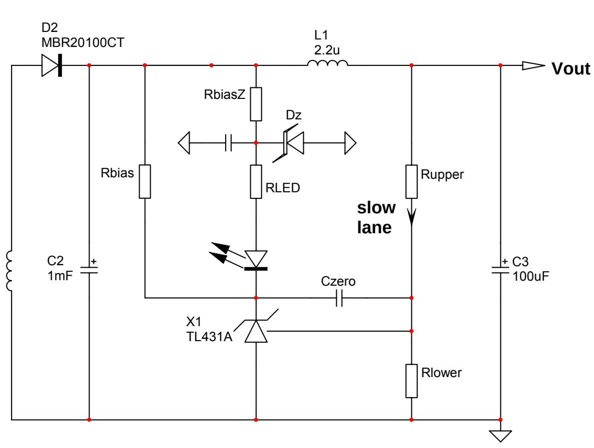
Pole at the origin — Two zeros — Mid-band gain — Two poles

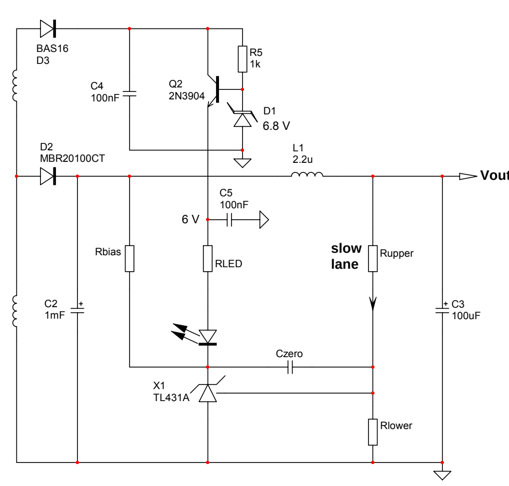
Type 3

Selection. Service. Support.

The problem is the fast lane…

- The difference with the TL431 comes from the fast lane
- Can we get rid of it?

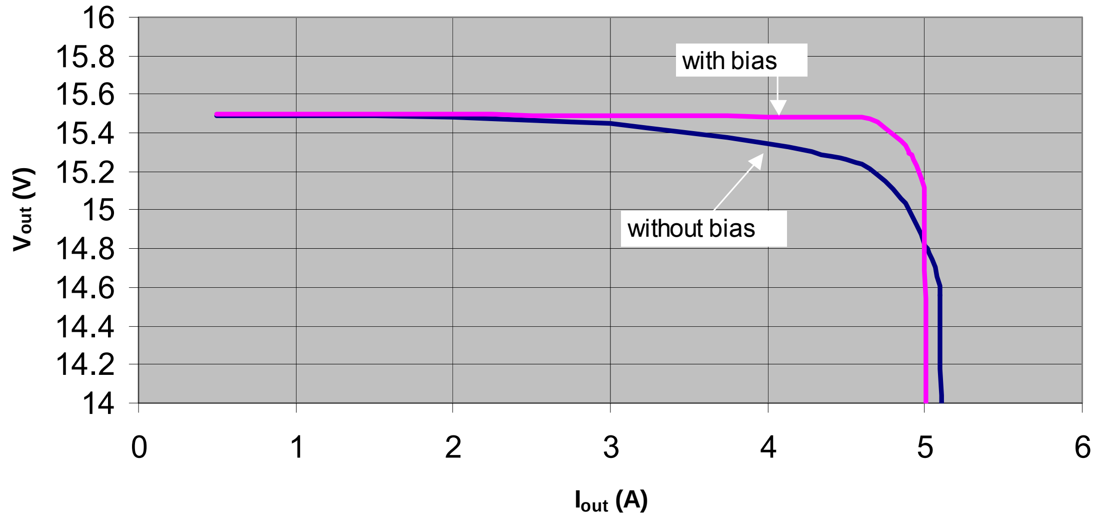




- Rather costly solutions…

Watch for the TL431 bias!

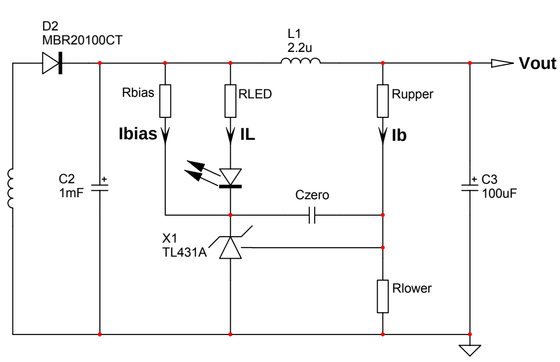
- ❑ A TL431 requires 1 mA minimum to operate within specs
- ❑ A TLV431, 100 µA only…



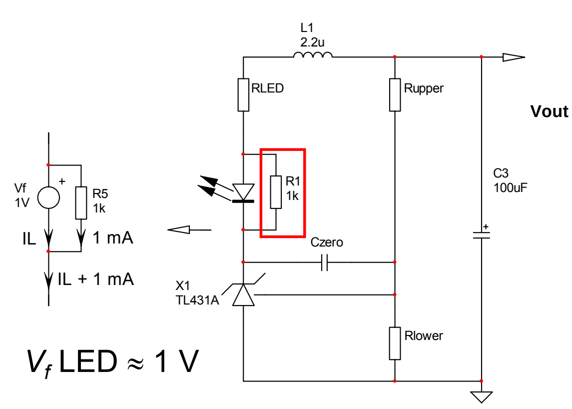
- ❑ Changes in Z_{out} implies a change in $T(0)$

How to provide more bias?

- ❑ Connect a resistor to V_{out}
- ❑ Use the LED to form a constant current source



Solution ***a***



Solution ***b***

- ❑ R_1 on the right picture steals current away from the LED

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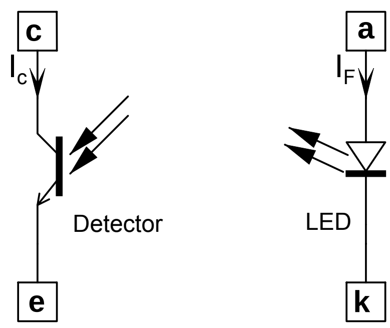
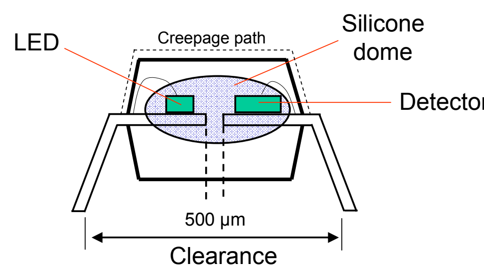
The optocoupler is the traitor here!

- ❑ 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

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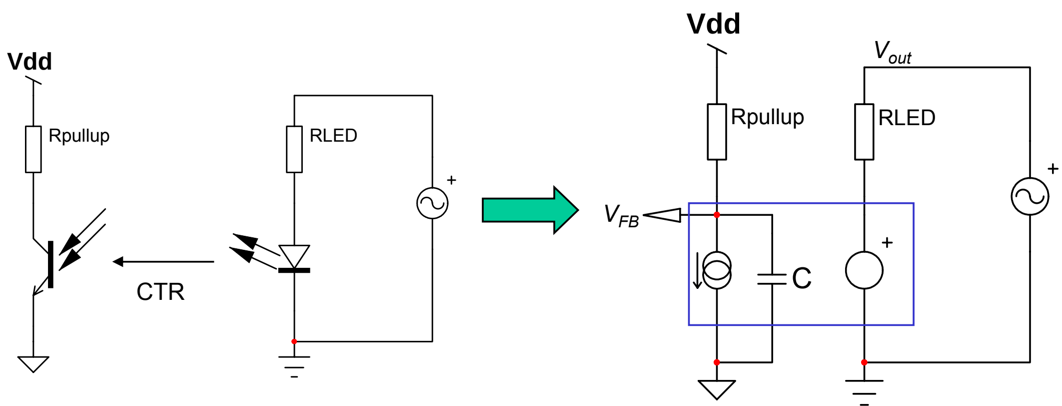
The internal pole should be known

- ❑ The photons are collected by a collector-base area.
- ❑ This area offers a large 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 above 5 times f_c, its effect is negligible
If f_p is close to f_c, phase margin degradation

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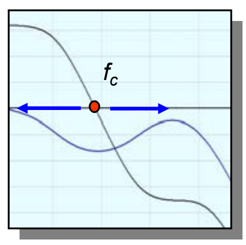
Assess the CTR variations

- ❑ 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**

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 +2 dB



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

SFH-615

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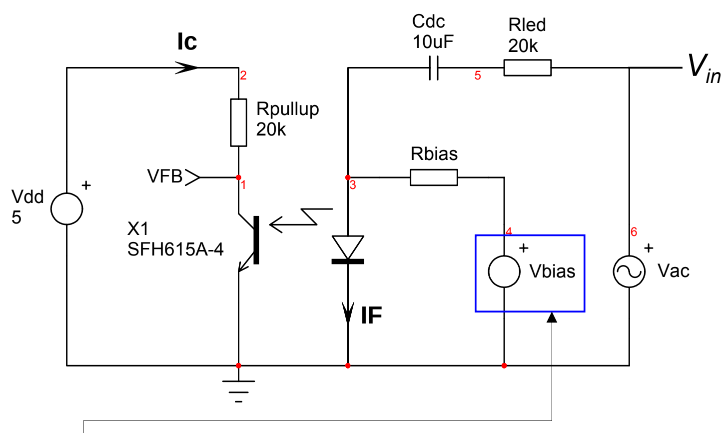
Selecting the right optocoupler

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

Understand the optocoupler impact on your design!

Find the pole position on your optocoupler

❑ Once your optocoupler is selected, characterize it:

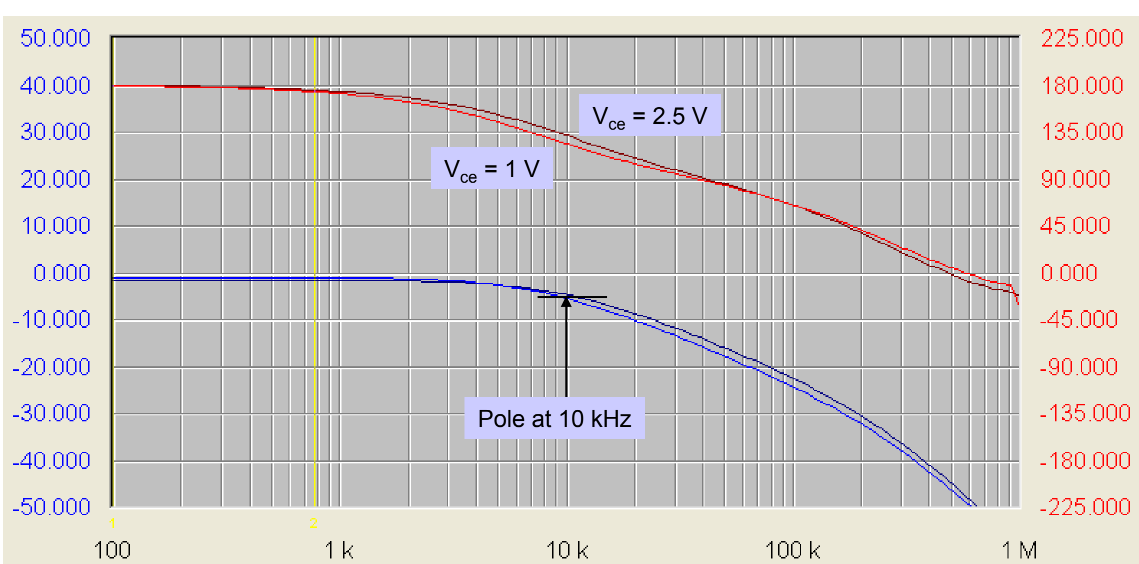


❑ Adjust V_{bias} to V_{FB} = 2.5 V (or V_{dd}/2) then ac sweep

❑ Observe V_{FB} with a scope or a network analyzer

Select the bias point as in your circuit

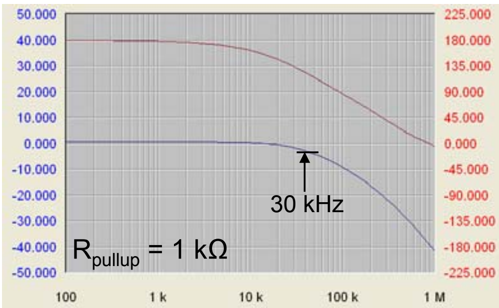
❑ Identify the pole position. Here it is located at 10 kHz

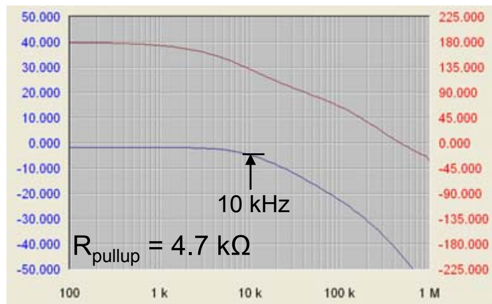


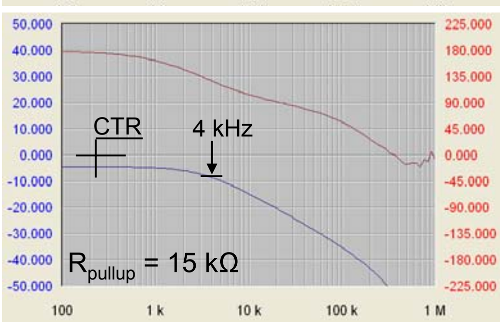
R_{pullup} = 4.7 kΩ SFH615A-2

Changing the pullup affects the pole position

- ❑ A low pullup resistor offers better bandwidth!







- ❑ Changing the bias point affects the CTR

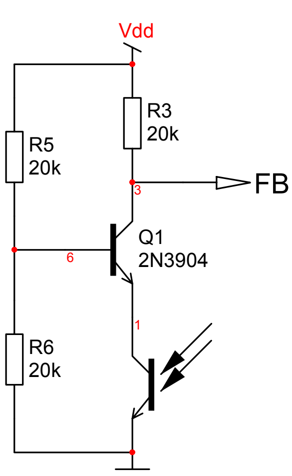
$$\frac{V_{FB}(s)}{V_{out}(s)} = -\frac{R_{pullup}}{R_{LED}}\mathrm{CTR}$$

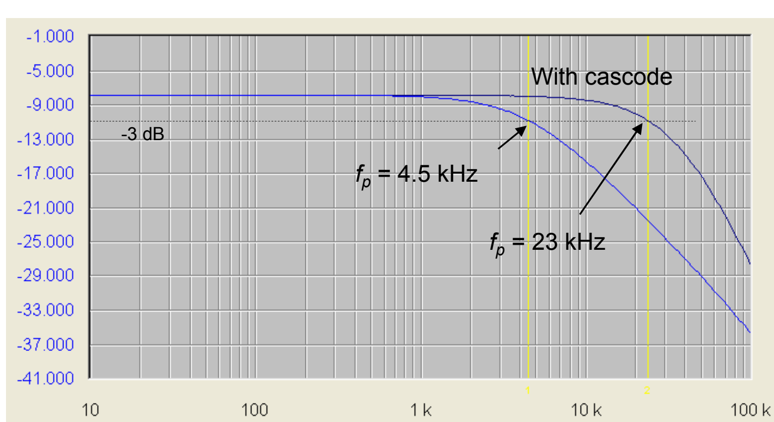
- ❑ If $R_{pullup} = R_{LED}$, then $|G_0| = 0$ dB…?

SFH615A-2

To push the pole farther, use a cascode

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

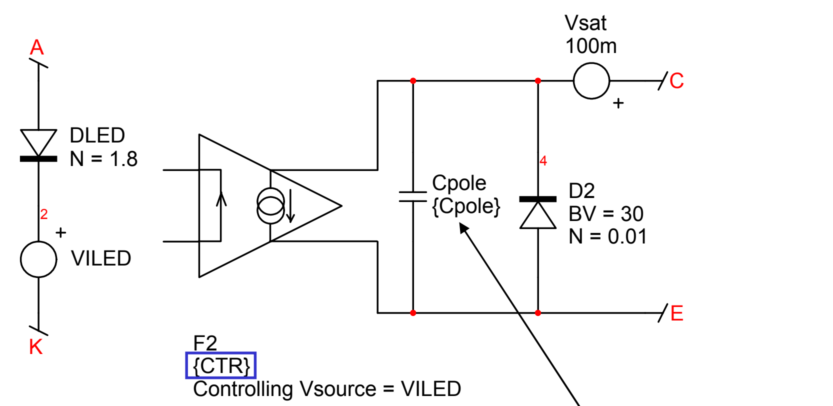




SFH615A-2

A simple optocoupler model

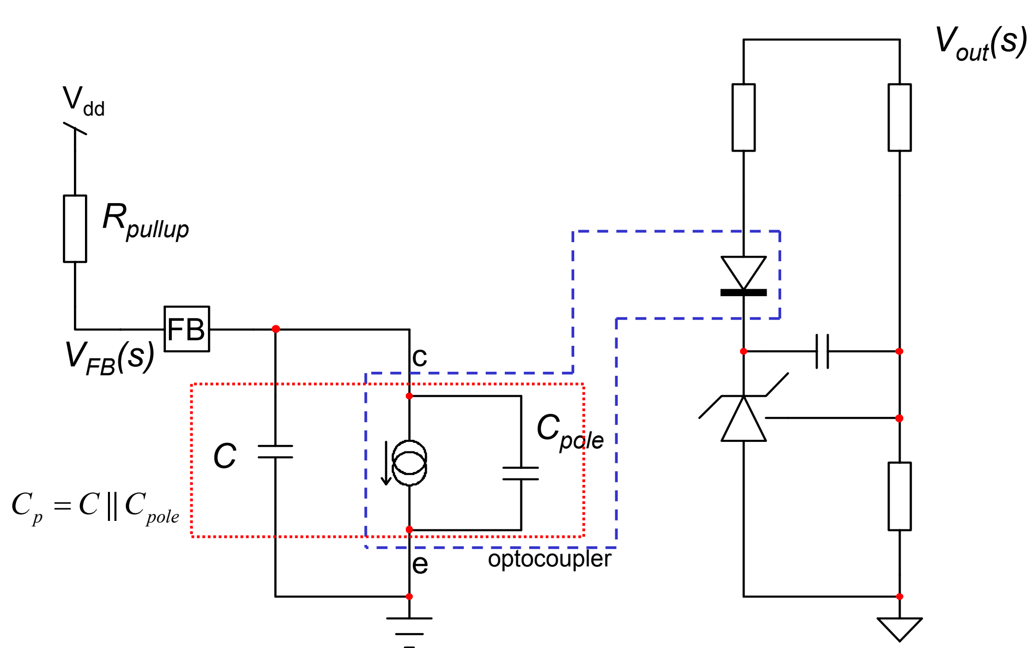
- ❑ Read the pole position and the CTR value
- ❑ Adjust the internal capacitor value to fix the pole



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

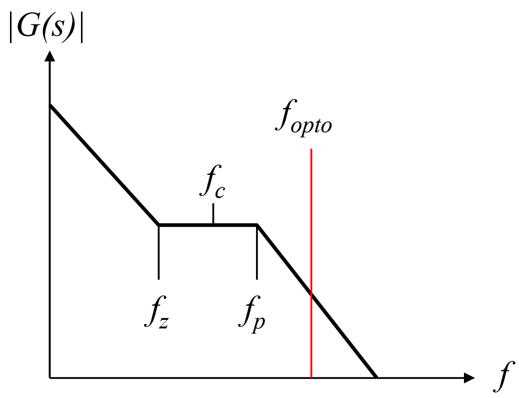
With the TL431, the situation is different

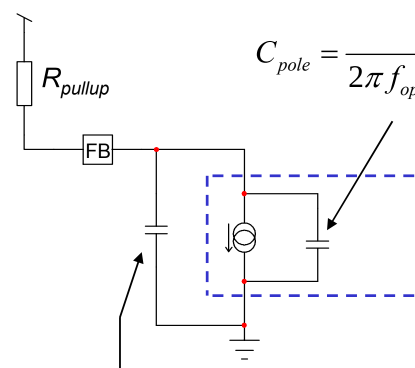
- ❑ In the TL431 application, the pole is inherently present



With the TL431, the situation is different

- ❑ First case: the optocoupler pole is beyond *G(s)* pole
- ➢ Calculate C_p to combine with C_{pole}



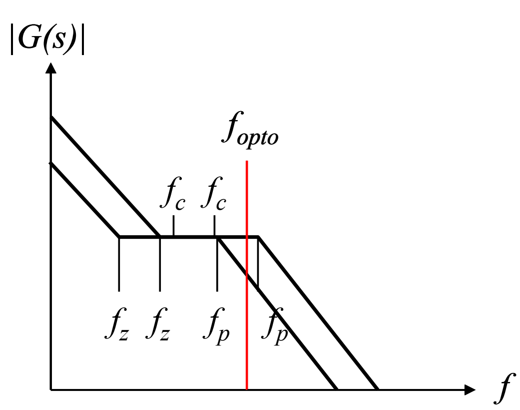


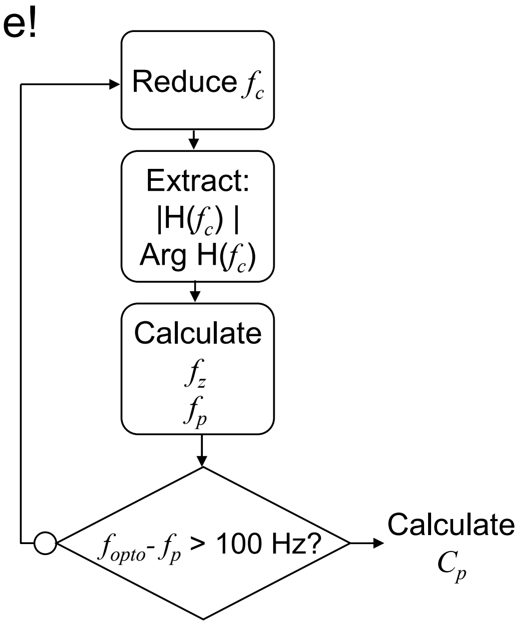
$$C_{pole} = \frac{1}{2\pi f_{opto} R_{pullup}}$$

From *G(s)* calculations → *C* → $\boxed{C_p = C - C_{pole}}$

With the TL431, the situation is different

- ❑ Second case: the optocoupler pole is below *G(s)* pole
- ➢ The optocoupler sets *G(s)* pole!





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Design example: a DCM flyback converter

- ❑ We want to stabilize a 20-W DCM adapter
- ❑ V_{in} = 85 to 265 V rms, V_{out} = 12 V/1.7 A
- ❑ F_{sw} = 65 kHz, R_{pullup} = 20 kΩ
- ❑ Optocoupler is SFH-615A, pole is at 6 kHz
- ❑ Cross over target is 1 kHz
- ❑ Selected controller: NCP1216

1. Obtain a power stage open-loop Bode plot, *H(s)*
2. Look for gain and phase values at cross over
3. Compensate gain and build phase at cross over, *G(s)*
4. Run a loop gain analysis to check for margins, *T(s)*
5. Test transient responses in various conditions

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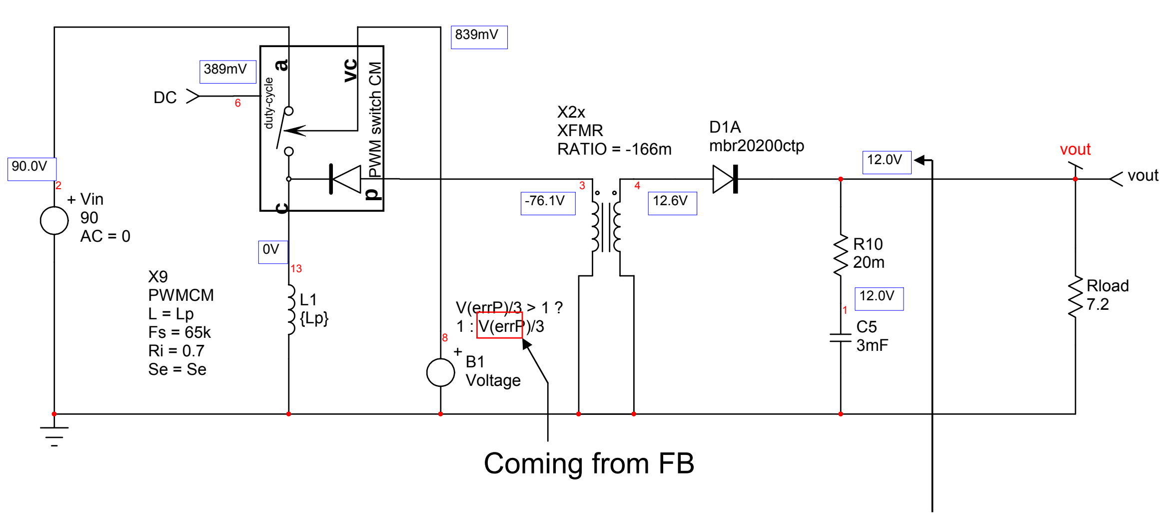
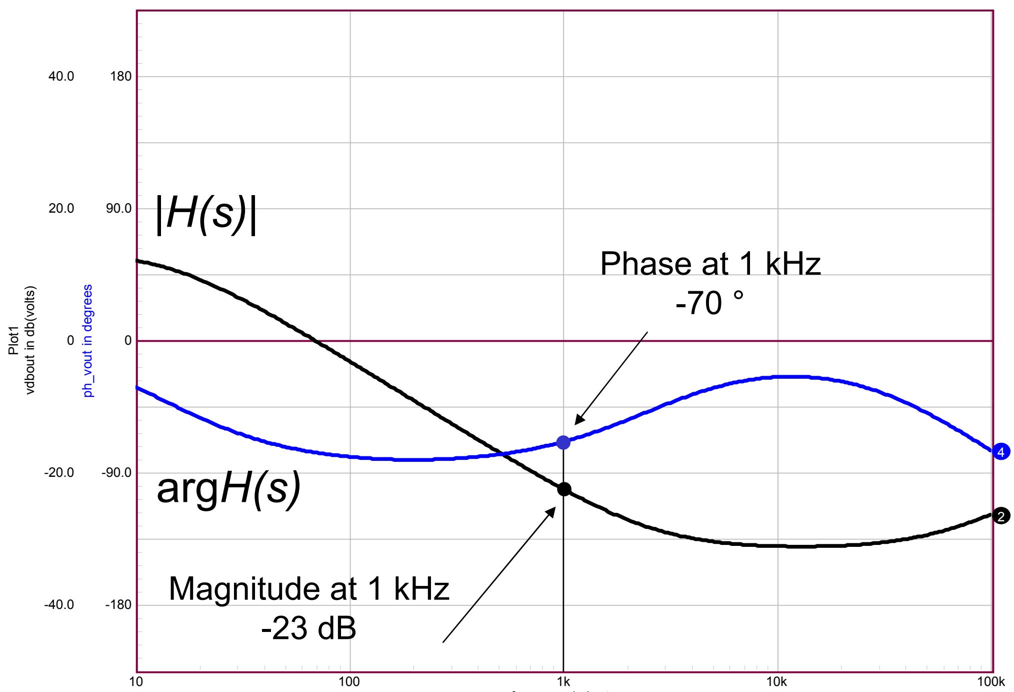
Stabilizing a DCM flyback converter

- ❑ Capture a SPICE schematic with an averaged model



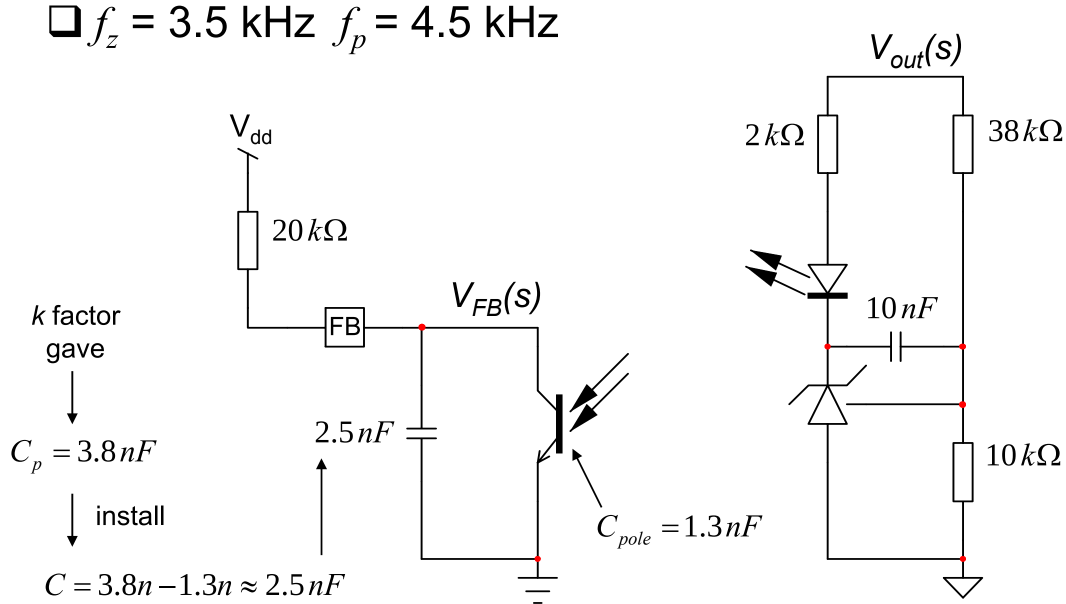
- ❑ Look for the bias points values: V_{out} = 12 V, ok

Stabilizing a DCM flyback converter

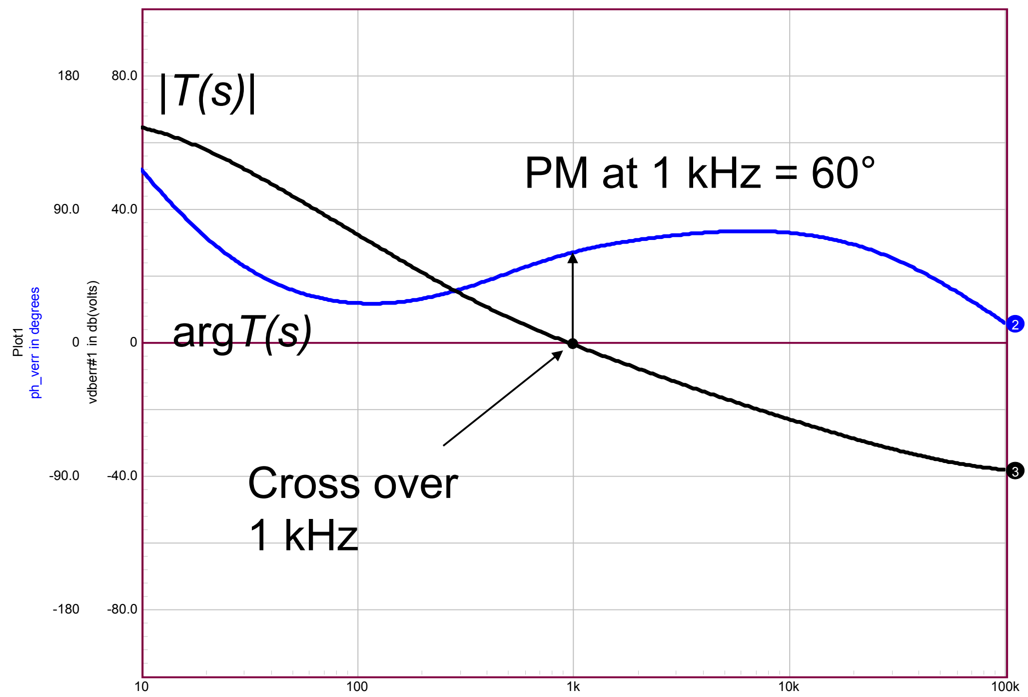


Stabilizing a DCM flyback converter

- ❑ Apply k factor or other method, get f_z and f_p
- ❑ f_z = 3.5 kHz f_p = 4.5 kHz



Stabilizing a DCM flyback converter



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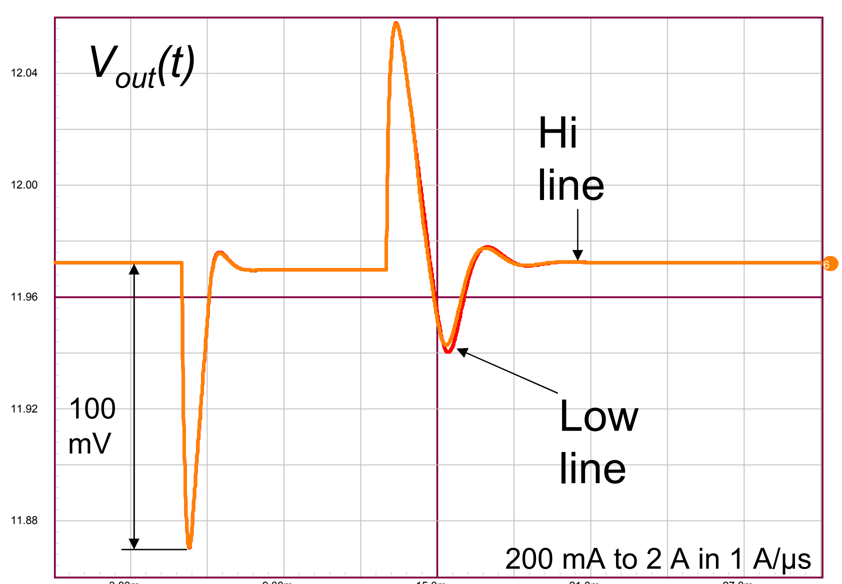
Stabilizing a DCM flyback converter

- ❑ Sweep ESR values and check margins again



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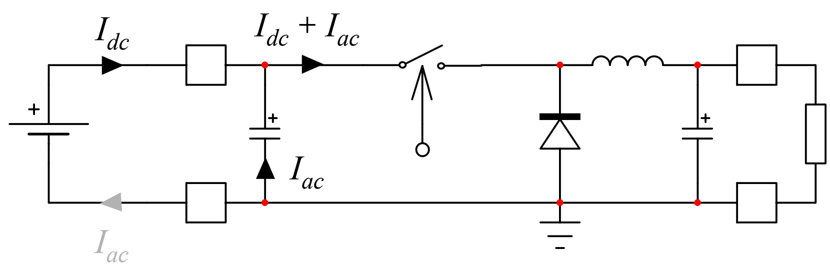
The need for an input filter

- ❑ The dc-dc converter is supplied from a dc source
- ❑ A capacitor locally decouples the line. In theory:
- ➢ The capacitor supplies the ac current
- ➢ The source only sees a dc current



- ❑ In reality, the capacitor is not perfect (limited cap., ESR and ESL)
- ❑ Some ac current manages to enter the source
- ➢ Source pollution, adjacent converters disturbance, radiated noise

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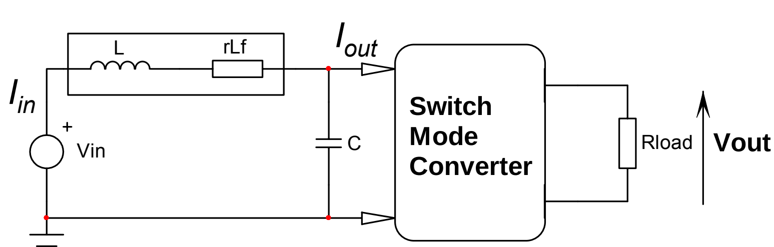
The need for an input filter

- ❑ A front-end filter has to be installed
- ❑ The ac current will only flow in *C* as *L* opposes its circulation
- ❑ The remaining ac current in the source must pass the specs!



The need for an input filter

- Our dc-dc ensures P_{out} is constant regardless of V_{in}
  - If V_{in} increases, I_{in} reduces to keep $V_{in} \cdot I_{in}$ constant
  - If V_{in} decreases, I_{in} increases to keep $V_{in} \cdot I_{in}$ constant

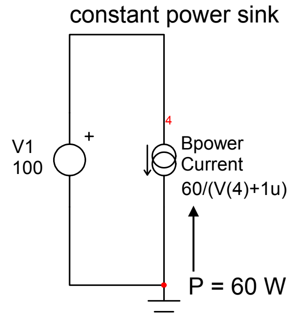
η = 100%

$$R_{in} = \frac{V_{in}}{I_{in}} \qquad V_{in} = \frac{P_{in}}{I_{in}} \qquad P_{in} = R_{load} I_{out}^2$$

Neg. sign!

$$\frac{dV_{in}}{dI_{in}} = \frac{d}{dI_{in}} \frac{R_{load} I_{out}^2}{I_{in}} = \boxed{-} R_{load} \left(\frac{I_{out}}{I_{in}} \right)^2$$

SPICE simulation
constant power sink

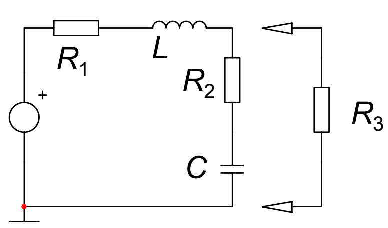


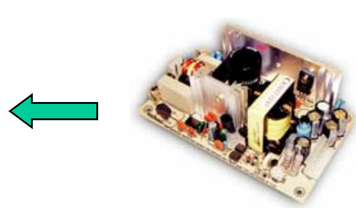
.TF V(4) V1 ; transfer function analysis
***** SMALL SIGNAL DC TRANSFER FUNCTION
output_impedance_at_V(4) 0.000000e+000
v1#Input_impedance -1.66667e+002
Transfer_function 1.000000e+000

Neg. sign!

The need for an input filter

- Ohmic paths in the *RLC* filter are losses that damp the network





- If these losses are compensated, the damping factor is zero

$$T(s) = \frac{1}{\frac{s^2}{\omega_0^2} + \frac{s}{\omega_0 Q} + 1} \qquad Q = \frac{1}{2\boxed{\zeta}}$$

If ohmic losses are gone, the damping factor is zero.

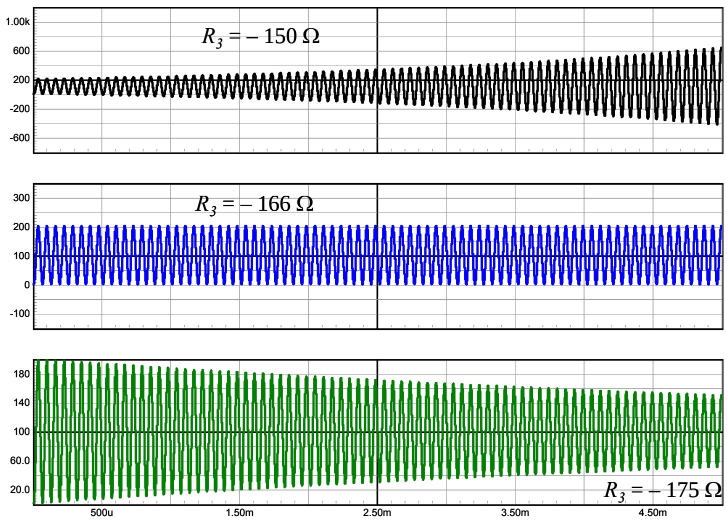
- In a lossy *RLC* filter, the damping factor is:

$$\zeta = \frac{\boxed{L + C(R_2R_3 + R_1R_3 + R_2R_1)}}{2(R_1 + R_3)}\omega_0 \xrightarrow{=0} R_3 = -\frac{R_1R_2C + L}{C(R_1 + R_2)}$$

A tunnel-based oscillator?

- Loading an *RLC* filter with a negative impedance:
  - You build a negative resistance oscillator!

$$R_3 = -\frac{R_1 R_2 C + L}{C\left(R_1 + R_2\right)} = -\frac{100m \times 500m \times 1u + 100u}{1u \times \left(600m\right)} = -166\ \Omega$$



Diverging oscillations
$\zeta < 0$

Steady-state oscillations
$\zeta = 0$

Damped oscillations
$\zeta > 0$

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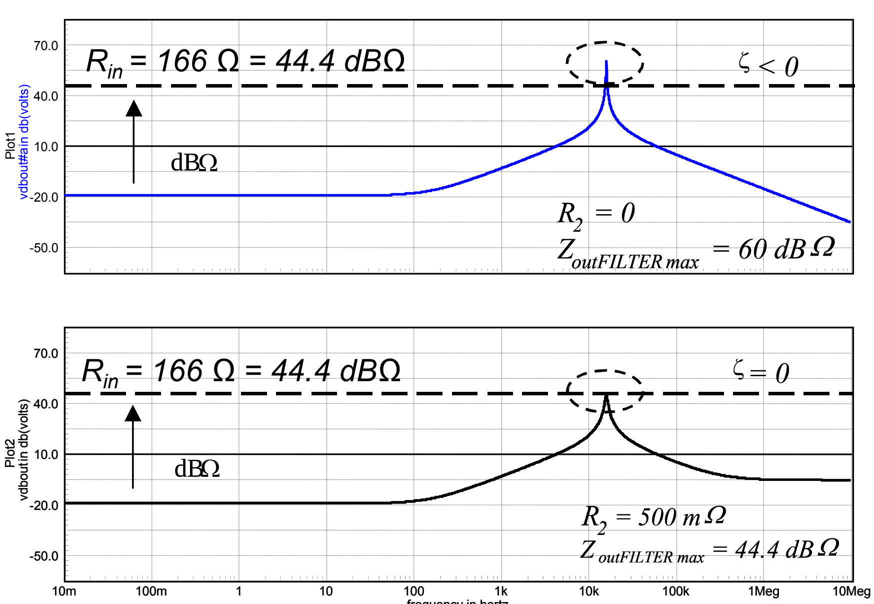
Watch the output/input impedances

- A possibility to look for oscillations is to ac sweep Z_{out} of *RLC*



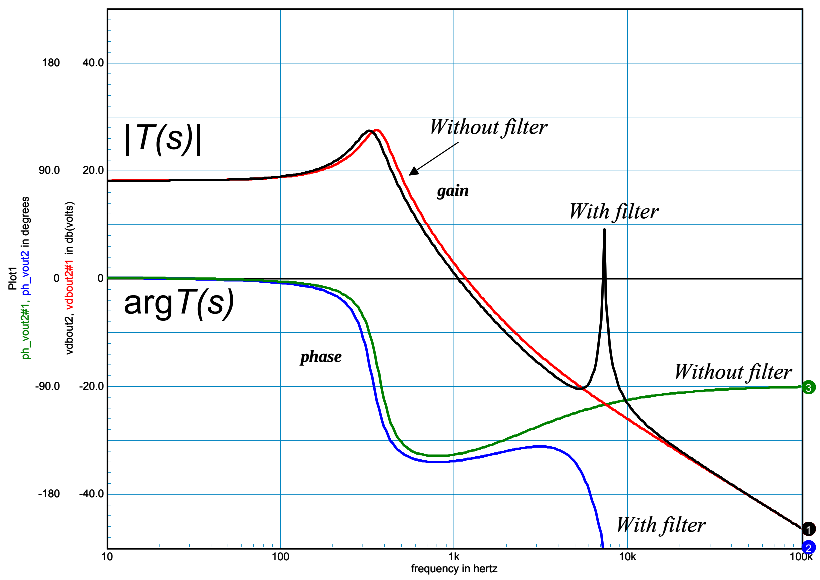
$$\left\| Z_{out} \right\|_{\max} = \frac{Z_0^{\ 2}}{R_1}\sqrt{1 + \left(\frac{R_1}{Z_0}\right)^2}$$

To be safe

$$\left\| Z_{out} \right\| << \left\| Z_{in} \right\|$$

The filter effect also appears on the Bode plot

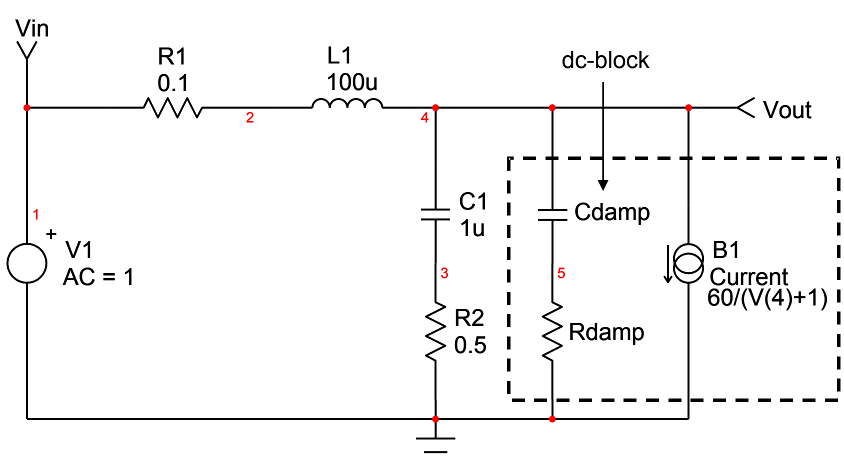
- ❑ The filter peaking brings instabilities…



Damp that *RLC* filter!

How to damp the *RLC* network

- ❑ Damping can be obtained through different arrangements:
- ➢ a resistor in parallel with C_1
- ➢ a resistor in parallel with L_1
- ➢ Numerous other possible combinations!



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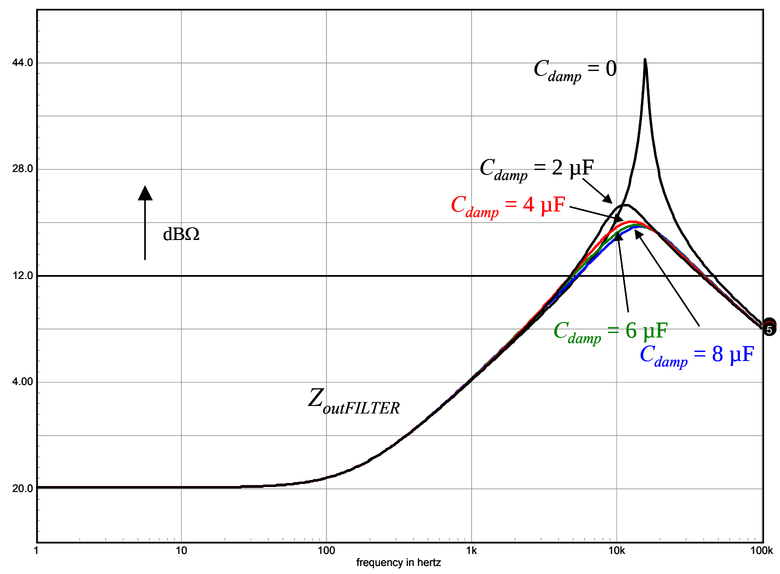
Buck, a design example

- ❑ The damping resistor is calculated to 9.5 Ω in this example
- ❑ The dc-block capacitor is swept from 2 to 10 times C_1



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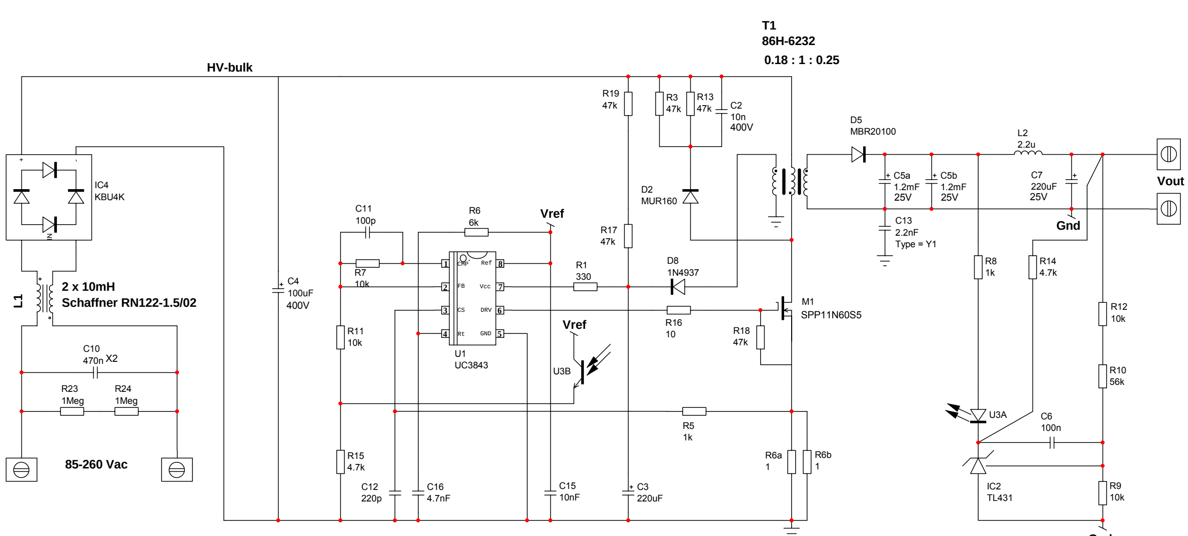
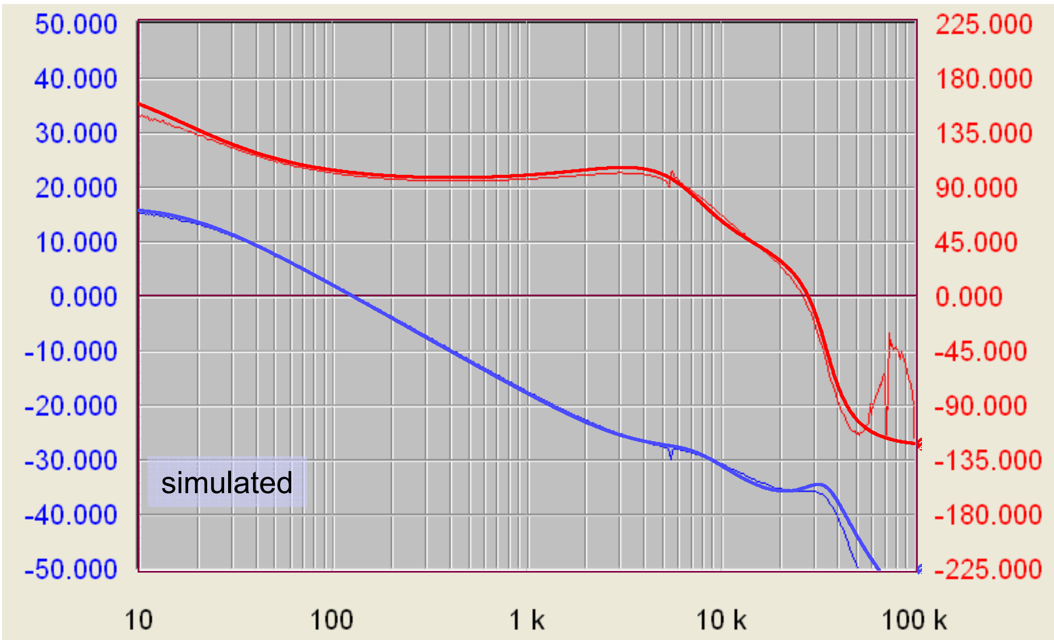
Stabilizing a UC3843 converter

- ❑ A 19 V/3 A converter is built around an UC3843

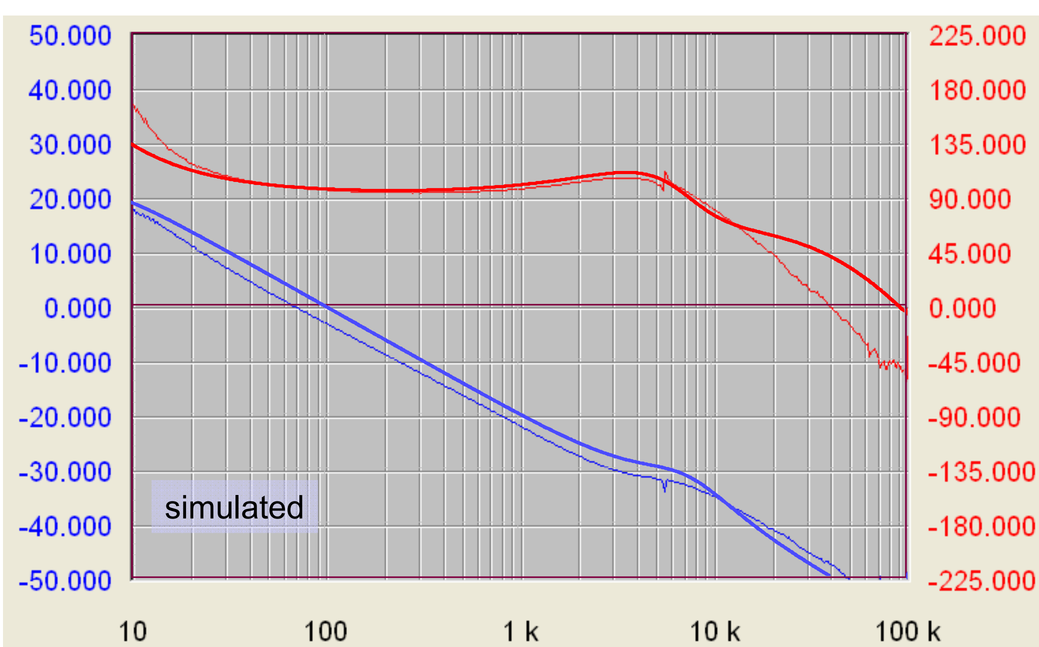


Change the operating conditions easily



CCM operation, R_{load} = 6.3 Ω

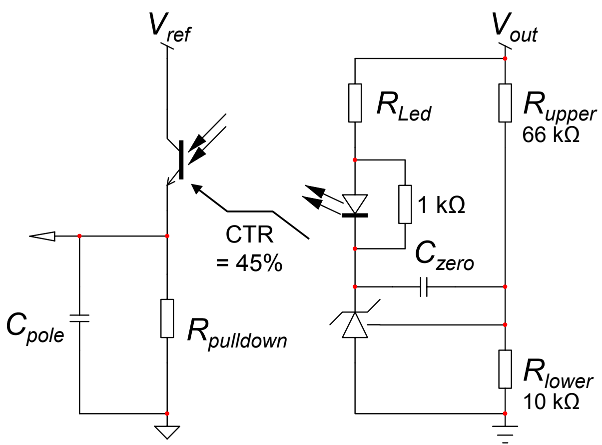
Reduce the load to enter in DCM



DCM operation, R_{load} = 20 Ω

From the open-loop Bode plot, compensate

❑ The TL431 is tailored to pass a 1-kHz bandwidth



Calculate mid-band gain: +18 dB

$$R_{LED} = \frac{R_{pullup}\text{CTR}}{10^{\frac{18}{20}}} = \frac{4.7k \times 0.45}{7.94} = 266\,\Omega$$

We place a zero at 300 Hz:

$$C_{zero} = \frac{1}{2\pi f_{zero} R_{upper}} = \frac{1}{6.28 \times 300 \times 66k} = 8\,nF$$

We place a pole at 3.3 kHz:

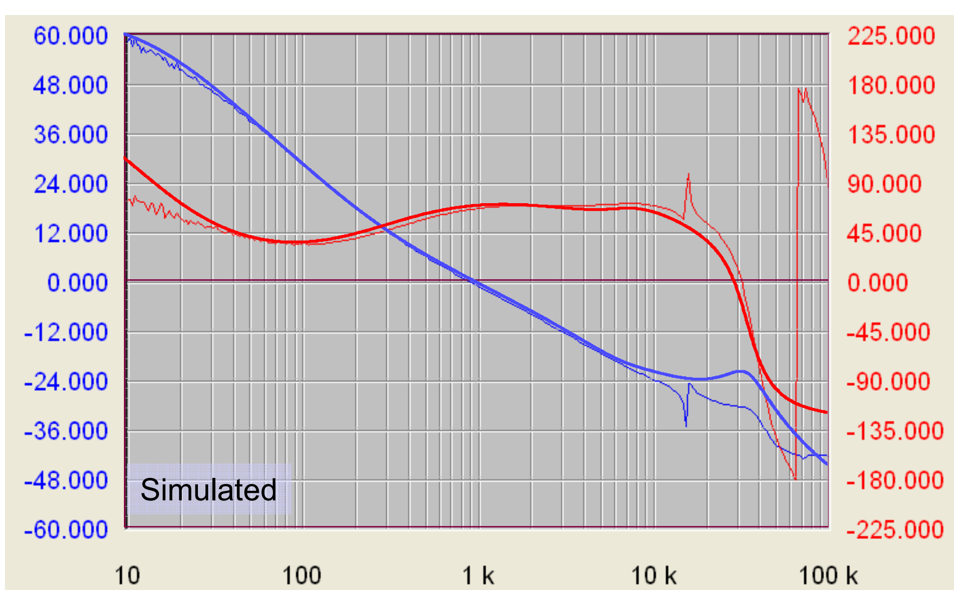
$$C_{pole} = \frac{1}{2\pi f_{pole} R_{pulldown}} = \frac{1}{6.28 \times 3.3k \times 4.7k} = 10\,nF$$

k factor method

"Switch-Mode Power Supplies: SPICE Simulations and Practical Designs", McGraw-Hill

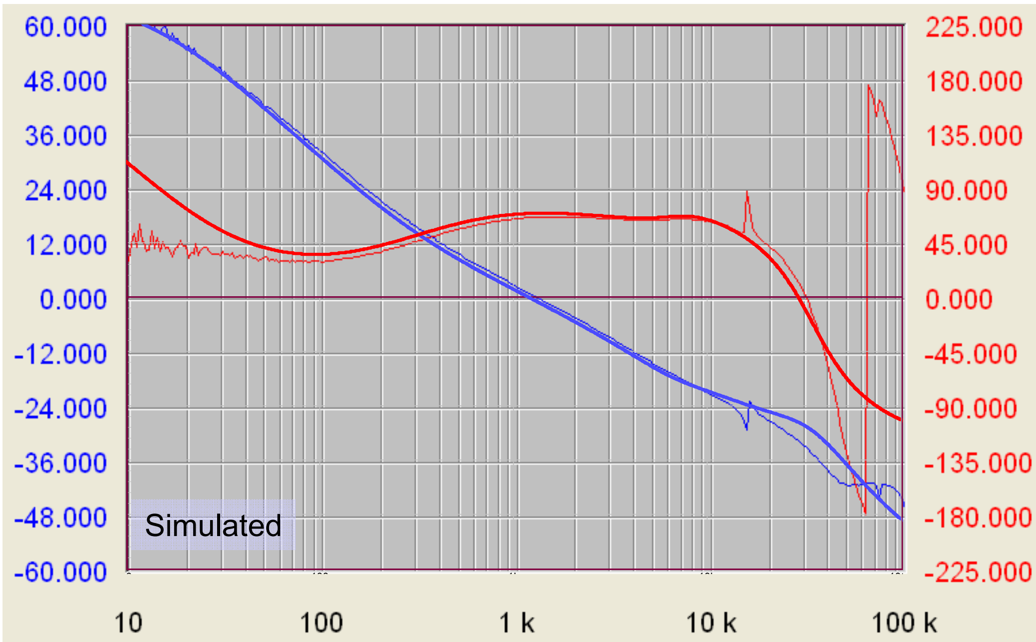
Verify in the lab. the open-loop gain

❑ Sweep extreme voltages and loads as well!



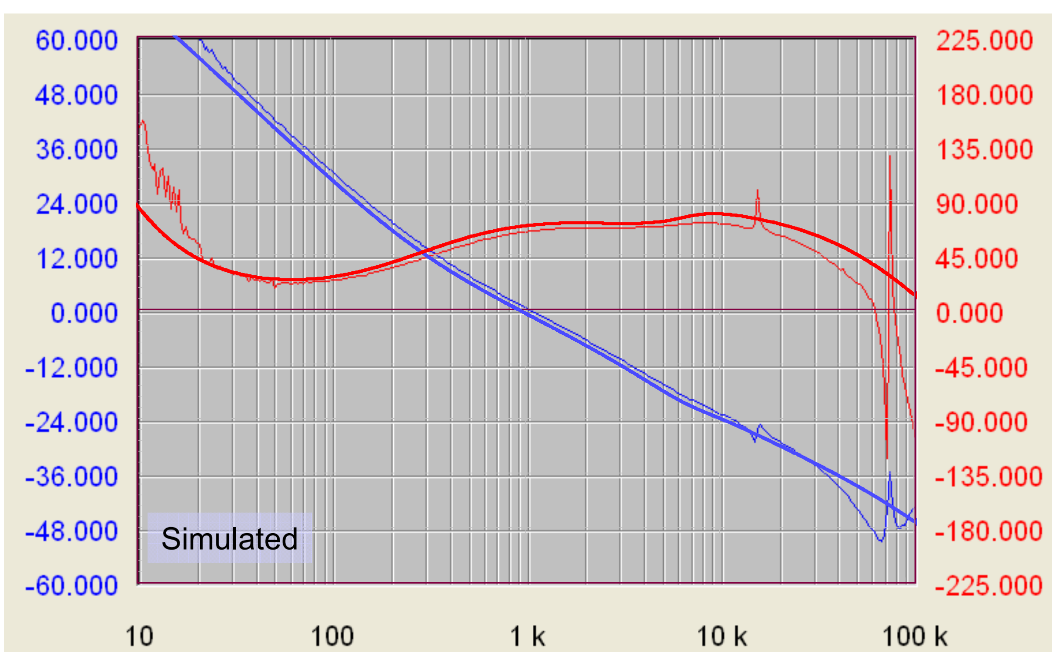
CCM operation, R_{load} = 6.3 Ω, V_{in} = 150 Vdc

Verify in the lab. the open-loop gain



CCM operation, R_{load} = 6.3 Ω, V_{in} = 330 Vdc

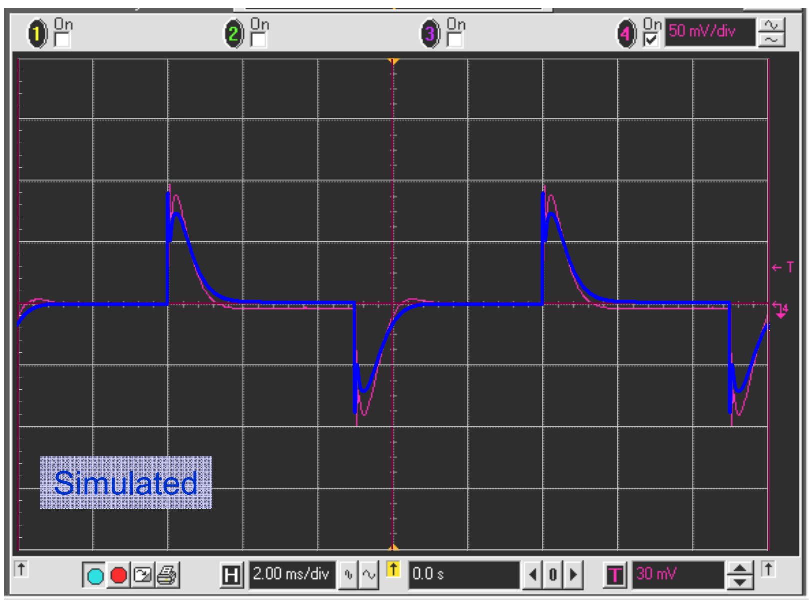
Verify in the lab. the open-loop gain



DCM operation, R_{load} = 20 Ω, V_{in} = 330 Vdc

As a final test, step load the output

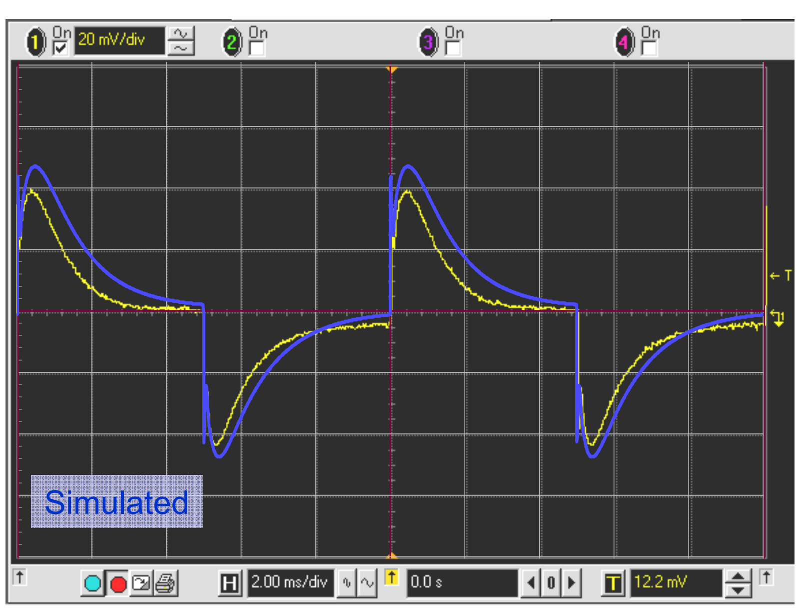
- ❑ Good agreement between curves!



V_{in} = 150 V
CCM
2 to 3 A
1 A/µs

As a final test, step load the output

- ❑ DCM operation at high line is also stable



V_{in} = 330 V
DCM
0.5 to 1 A
1 A/µs

Conclusion

- ❑ We now understand the origins of phase margin needs
- ❑ The crossover frequency value is analytically derived
- ❑ Current-mode technique simplifies the compensation
- ❑ Operating mode transition is not a problem for CM
- ❑ Type 2 and type 3 are also available with OTAs and TL431
- ❑ The optocoupler's pole can degrade the phase margin
- ❑ Do NOT forget the influence of the EMI filter
- ❑ SPICE eases the design with multi-output converters
- ❑ A real-case example confirmed the validity of the approach!