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 $\varepsilon = V_{ref} - \alpha V_{in}$ is amplified and sent to the control input.
- The power stage reacts to reduce $\varepsilon$ as much as it can.

Positive or negative feedback?

- Do we want to build an oscillator?

\[
\frac{V_{out}(s)}{V_{in}(s)} = \frac{H(s)}{1 + H(s)G(s)}
\]

Open-loop gain $T(s)$

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

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

\[
1 + G(s)H(s) = 0 \quad \implies \quad \{ G(s)H(s) \} = 1 \quad \implies \quad \angle G(s)H(s) = -180^\circ
\]

\[
\angle G(s)H(s) = -180^\circ \quad \implies \quad \gamma_{\text{quas}} = -1, 0
\]
A constant gain with a 180° rotation

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

Parameters
- Gfc = -29
- G = 10^(-Gfc/20)
- Rf = G * 100k

A simple oscillator

- Case n°1, at 0-dB gain crossover, phase $\phi$ 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)
A simple oscillator

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

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

A simple oscillator

- Case n°3, at 0-dB gain crossover, $\phi$ 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 < -360° at the crossover point

**Conditions for non-permanent oscillations are:**

- Total phase rotation < -360° at the crossover point

**Total phase delay at \( f_c \):**
- \(-112°\) H(s) power stage
- \(-180°\) G(s) opamp
- \(\text{total} = -292°\)

**Stable!**

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:
  \[ H(s) = \frac{N(s)}{D(s)} \]

- 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:
  \[ s_n = -5k \]
  \[ s_n = -30k \]
  Denominator root:
  \[ s_p = -1k \]

  \[ f_n = \frac{5k}{2\pi} = 796 \text{ Hz} \]
  \[ f_n = \frac{30k}{2\pi} = 4.77 \text{ kHz} \]
  \[ f_n = \frac{1k}{2\pi} = 159 \text{ Hz} \]

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

\[ H(s) = \frac{s + 4}{(s + 0.8)[(s + 2.5)^2 + 4]} \]

  - \( s_n = -4 \)
  - \( s_n = -0.8 \)
  - \( s_p = -2.5 + 2j \)
  - \( s_p = -2.5 - 2j \)

- 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

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

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

Take inverse Laplace transform

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

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

LHP RHP

$Q < 0.5$

Response to a step

$Q > 0.5$

Response to a step

$Q = 0.5$

Response to a step

$Q < -0.5$

Response to a step

$Q > -0.5$

Response to a step

$-0.5 < Q < 0$

Response to a step

$Q = -0.5$

Response to a step
Poles, zeros and s-plane

- The pole magnitude at the cutoff frequency is \(-3\, \text{dB}\).
- The pole "lags" the phase.

\[
\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = \frac{1}{1 + sRC} \quad \Rightarrow \quad \frac{V_{\text{out}}(\omega_b)}{V_{\text{in}}(\omega_b)} = \frac{1}{\sqrt{1 + [\omega_b/\omega_c]^2}} = \frac{1}{\sqrt{2}}
\]

\[
\arg\left(\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)}\right) = \arg(1) - \arg\left(1 + \frac{s}{\omega_b}\right) = -\arctan(\infty) = -\frac{\pi}{2} = -90° \quad \text{At } f = \infty
\]

- The zero magnitude at the cutoff frequency is \(+3\, \text{dB}\).
- The zero "boosts" the phase.

\[
\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = 1 + \frac{s}{\omega_b} \quad \Rightarrow \quad \frac{V_{\text{out}}(\omega_b)}{V_{\text{in}}(\omega_b)} = \sqrt{1 + [\omega_b/\omega_c]^2} = \sqrt{2}
\]

\[
\arg\left(\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)}\right) = \arg(1 + \frac{s}{\omega_b}) = \arctan(\infty) = \frac{\pi}{2} = 90° \quad \text{At } f = \infty
\]

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

\[
\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = \frac{s}{\omega_b} \quad \text{Zero for } s = 0: \text{ zero at the origin} \quad \Rightarrow \quad \text{As } f \text{ increases the gain increases with a +1 slope (+20 dB/decade)}
\]

\[
\arg\left(\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)}\right) = \arg\left(\frac{s}{\omega_b} \right) = \arctan(\infty) = \frac{\pi}{2} \quad \text{For } f > f_{zo}
\]

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

\[
\arg\left(\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)}\right) = \arg\left(1 - \frac{s}{\omega_b} \right) = -\arctan(\infty) = -\frac{\pi}{2} \quad \text{For } f > f_{po}
\]

A pole at the origin introduces a fixed phase rotation of \(-90°\).
The Right Half-Plane Zero

- In a CCM boost, $I_{\text{out}}$ is delivered during the off time: $I_{\text{out}} = I_d = I_1 (1 - D)$

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

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

- $d = 58.3\%$
- $d = 59%$
The Right Half-Plane Zero

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

Small-signal equations can help us to formalize it:

$$
\frac{\partial I_{out}}{\partial L} + \frac{\partial I_{out}}{\partial D} I_L = I_L (1 - D) - \hat{d} I_L
$$

The negative sign indicates a positive root!

$$
\hat{i}_{out}(s) = \frac{\partial I_{out}}{\partial L} \frac{sL}{s}\left(1 - \frac{sL}{D^2 R_{load}}\right) = \frac{\partial I_{out}}{\partial D} \frac{\partial \hat{d}}{\partial D} = \frac{\partial \hat{d}}{\partial D} I_L
$$

$$
\omega_h = \frac{V_{out} D'}{L}, \quad \omega_z = \frac{R_{load} D'^2}{L}
$$

- 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:

\[ V(s) = V_n(s) - V_n(s) \frac{R}{sC} = V_n(s) \left(1 - \frac{s}{\omega_0}\right) \]

![Diagram of RHPZ circuit](image)

The Right-Half-Plane-Zero

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

| \( |V_{out}(s)| \) |
| --- |
| \( \frac{G(s) = 1 + \frac{s}{\omega_0}}{s} \) |
| \( \frac{G(s) = 1}{s} \) |

<table>
<thead>
<tr>
<th>( \text{arg}V_{out}(s) )</th>
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<tr>
<td>( -90^\circ )</td>
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Plot 1: \( v_{db_{out}} \text{ in } \text{db(volts)} \)

Plot 2: \( |V_{out}(s)| \text{ vs frequency in hertz} \)

- LHPZ
- RHPZ
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 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 \( \hat{d}_1 \).

\[ \hat{d}_1 \]

\[ D(t) \]

\[ V_{out}(t) \]

Averaged models can predict the DCM RHPZ

\[ \dot{v}_{out}(s) = H_d \left( \frac{1+s/s_1}{1+s/s_2} \right) \left( \frac{1-s/s_2}{1+s/s_1} \right) \]

\[ s_{1_1} = \frac{1}{C_{out}R_{ESR}} \quad s_{2_1} = \frac{R_{load}}{M^2L} \]

\[ s_{p_1} = \frac{2}{M-1} \left( \frac{1}{C_{out}R_{ESR}} \right) \]

\[ s_{p_2} = 2F_{sw} \left( \frac{1-1/M}{D} \right) \]

\[ H_d = \frac{2V_{in}}{D} \left( \frac{M-1}{2M-1} \right) \]
The Right-Half-Plane-Zero

- Averaged models can predict the DCM RHPZ

- \( H_j = 28.75 \text{ dB} \)
- \( f_{\text{h}} = 1.06 \text{ kHz} \)
- \( f_{\text{r}} = 141 \text{ kHz} \)
- \( f_{\text{f}} = 4.2 \text{ Hz} \)
- \( f_{\text{p}} = 47.1 \text{ kHz} \)

How much margin? The RLC filter

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

\[
T(s) = \frac{1}{LCs^2 + RCS + 1}
\]

\[
T(s) = \frac{1}{s^2 + \frac{2\zeta}{\omega_n} s + \frac{1}{\omega_n^2}} = \frac{1}{s^2 + \frac{s}{\omega_n} + \frac{1}{\omega_n^2 + \frac{1}{\omega_n Q} + 1}}
\]

\[
\omega_n = \frac{1}{\sqrt{LC}}
\]

\[
\zeta = R \sqrt{\frac{C}{4L}} \quad Q = \frac{1}{2\zeta}
\]

parameters:
- \( f_0 = 235k \)
- \( L = 10u \)
- \( C = 1/(4\times3.14159^2 \times f_0^2 \times L) \)
- \( w_0 = ((L)^*(C))^*\times0.5 \)
- \( Q = 10 \)
- \( R = 1/(((C)/(4*(L)))^*\times0.5)^*2*(Q)) \)

Change \( Q \) and run the simulation
The RLC response to an input step

- changing $Q$ affects the transient response

$$Q = 0.1$$ 
$$Q = 0.5$$ 
$$Q = 5$$

$Q < 0.5$ over damping  
$Q = 0.5$ critical damping  
$Q > 0.5$ under damping

Overshoot $= 65\%$

Fast response and no overshoot!

$Q < 0.5$, two real negatives roots  
$Q = 0.5$, two real coincident negative roots  
$Q > 0.5$, two complex roots  
$Q = \infty$, two imaginary conjugate roots

The RLC response to an input step

- $Q$ affects the poles position

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

Solve $N(s) = 0$ to obtain zeros  
Solve $D(s) = 0$ to obtain poles

$$s^2 + \frac{s}{\omega Q} + 1 = 0$$

2 roots, $s_1$ and $s_2 = \frac{\omega}{2Q} \left(-1 \pm \sqrt{1 - 4Q^2}\right)$
Where is the analogy with $T(s)$?

- In the vicinity of the crossover point, $T(s)$ combines:
  - one pole at the origin, $\omega_0$
  - one high frequency pole, $\omega_2$

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

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Closed-loop gain study

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

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

Closed loop:

$$\frac{T(s)}{1 + T(s)} = \frac{1}{s^2 + \frac{\omega_0^2}{\omega_2^2} + \frac{s}{\omega_2} + 1}$$

Solve for $Q$:

$$Q = \sqrt{\frac{\omega_0^2}{\omega_2^2} + \frac{1}{\omega_0^2 \omega_2^2}}$$
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

\[
T(s) = \frac{1}{1 + T(s)} = \frac{1}{s^2 + \frac{s}{\omega_0} + 1}
\]

\( f_1 = 300 \text{ Hz} \)
\( f_2 = 1000 \text{ Hz} \)
\( Q = 0.55 \)
\( \phi_m = 74^\circ \)

\[
\phi_m = 0^\circ \\
\phi_m = 32^\circ \\
\phi_m = 74^\circ \\
\phi_m = 133^\circ
\]

Linking \( \phi_m \) 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 \(|T(s)| = 1\)

\[
\left(\frac{\omega_1}{\omega_0} \right) \frac{1}{1 + j \frac{\omega_1}{\omega_0}} = 1 \quad \Rightarrow \quad \omega_1 = \frac{\omega_0 \sqrt{1 + 4Q^2} - 1}{2}
\]

2. substitute \( \omega_c \) into \( T(s) \), calculate its argument (phase margin)

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

3. extract the quality coefficient \( Q \):

\[
Q = \frac{\sqrt{1 + \tan^2(\phi_m)}}{\tan(\phi_m)} = \frac{\cos(\phi_m)}{\sin(\phi_m)}
\]
We can now plot $Q$ versus $\varphi_m$

- A $Q$ factor of 0.5 (critical response) implies a $\varphi_m$ of 76°
- A 45° $\varphi_m$ 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 $Z_{out}$

\[ Z_{out} = \left( sL_{out} + r_{lf} \right) \parallel \left( R_{esr} + \frac{1}{sC_{out}} \right) \]

A buck equivalent circuit

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

\[ Z_{out,OL} = \frac{1}{2\pi C_{out} f} \]

\[ |Z_{out,OL}| = \left| Z_{out,OL} \right| \frac{1}{1 + T(s)} \]
Closing the loop…

- At the crossover frequency \( Z_{out,CL} \approx Z_{out,OL} \)

Let's assess « almost » :-)

Calculating the output impedance

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

\[
|Z_{out,CL}| \approx \frac{1}{2\pi f_{c} C_{out}} \left| \frac{1}{1 + T(s)} \right|
\]

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

\[
\omega_c = -\frac{\omega_s \sqrt{\left( 1 + 4Q^2 \right) - 1}}{\sqrt{2}}
\]

- substitute into \( \frac{1}{1 + T(s)} \) and extract module

\[
\frac{1}{1 + T(s)} = \frac{1}{\sqrt{\left( \frac{-\omega_s}{\omega_c} + 1 \right)^2 + \frac{\omega_s^2}{\omega_c^2 \left( 1 + \frac{\omega_s^2}{\omega_c^2} \right)}}}
\]
Calculating the output impedance

Introduce the quality factor coefficient

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

Now replace \( Q \) by its definition

\[
Q = \frac{\cos(\varphi_m)}{\sin(\varphi_m)}
\]

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

Simplify

\[
\frac{1}{1+T(s)} = \frac{1}{\sqrt{2} - 2 \cos(\varphi_m)}
\]

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?

\[
\Delta V_{out} \approx \frac{\Delta I_{out}}{2\pi f_c C_{out}} \quad f_c \approx \frac{\Delta I_{out}}{\Delta V_{out} C_{out}} \frac{2\pi}{2\pi}
\]

\[
f_c \approx \frac{2}{80 \times 1 \times 2\pi} = 4 \text{ kHz}
\]

\[
Z_{out} \text{ @ } 4 \text{ kHz} = \frac{1}{2\pi \times 4 \times 1 \times 1} = 40 \text{ m\Omega}
\]

Select a 1000-µF capacitor featuring less than 40-mΩ ESR
Setting the right crossover frequency

- Compensate the converter for a 4-kHz $f_c$

![Compensated open-loop gain](image1)

- Buck operated in voltage-mode

- $\phi_m = 70^\circ$

Measure the obtained undershoot

\[ \Delta V \approx 40m \cdot \frac{\Delta I}{\sqrt{2 - 2 \cos(\phi_m)}} \]

\[ \Delta V \approx 40m \cdot \frac{2}{1.14} = 70 \text{ mV} \]

- 70 mV
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.

$$I_{out}(t) = \Delta I \frac{t}{\Delta t} = S_1 t$$

$$\Delta V_{out} = R_{ESR} S_1 t + L_{ESL} S_1 t + \frac{1}{C_0} \int_0^t S_1 t \cdot dt$$

Because of bandwidth limits, $R_{ESR}$ and $L_{ESL}$ play alone.

The capacitor contribution is small...

- $S_1 = 5 \, A/\mu s$

$$\Delta V = 505 \, mV$$

$$\Delta V = 100 \, mV$$

$$\Delta V = 606 \, mV$$

$V_{ESR}(t)$

$V_{ESL}(t)$

$V_{out}(t)$
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

\[
V(t) = k_p e(t) + k_i \int e(t) \, dt + k_d \frac{d e(t)}{d t}
\]

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

\[
A_{ac,cl}(s) = \frac{A_{ac,ol}(s)}{1 + T(s)}
\]
First, provide mid-band gain at crossover

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

![Graph showing gain and phase response with 0 dB@ $f_c$ and -21 dB@ $f_c$.]

Second, provide high gain in dc

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

![Graph showing gain and phase response with pole at the origin labeled $\frac{1}{sRC}$ and $V_{in}$ and $V_{out}$ connections.]
Second, provide high gain in dc

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

\[ G(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{1}{sRC} = \frac{1}{s} \omega_{po} = \frac{1}{R_iC_i} \]

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

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

Total phase lag brought by an origin pole is \(-3\pi/2\) or -270°

All compensators (1, 2 or 3) feature an origin pole
Third, evaluate the phase boost at $f_c$

\[ \arg H(s) \]
\[ \arg G(s) \]
\[ \arg T(s) \]

\[ \arg H(f_c) - 270^\circ + \text{BOOST} - \phi_m = -360^\circ \]
\[ \text{BOOST} = \phi_m - \arg H(f_c) - 90^\circ = 70^\circ + 175 - 90 = 155^\circ \]

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 $\arg H(f_c) < -45^\circ$ for a $\phi_m$ of 45°

\[ G(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{1}{sRC} = \frac{1}{s} \]
\[ \omega_{po} = \frac{1}{RC} \]
\[ G = \frac{f_{po}}{f_c} \quad \text{Select } f_{po} \text{ to have } G \text{ at } f_c \]
How do we boost the phase at $f_c$?

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

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

If $C_2 \ll C_1$

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

1 pole at the origin
1 zero
1 pole
Pole/zero placement and boost at $f_c$

- Phase boost appears between the zero and the pole

$$G(j\omega) = \left(\frac{1 + j\frac{\omega}{\omega_z}}{1 + j\frac{\omega}{\omega_p}}\right)$$

$$\arg G(j\omega) = \arg\left(\frac{1 + j\frac{\omega}{\omega_z}}{1 + j\frac{\omega}{\omega_p}}\right)$$

$$\arg G(f) = \arctan\left(\frac{f}{\omega_z}\right) - \arctan\left(\frac{f}{\omega_p}\right)$$

Peaks to a max at

$$f = \sqrt{\frac{\omega_z}{\omega_p}}$$

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

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

---

How do we boost the phase at $f_c$?

- Phase boost at $f_c = 71^\circ$

![Diagram showing phase and gain at different frequencies]

- Gain at $f_c = 21\text{ dB}$

- $f_z = 705\text{ Hz}$

- $f_p = 22\text{ kHz}$

- $G_{500\text{ Hz}} = 38\text{ dB}$

- Phase boost at $f_c = 71^\circ$
How do we boost the phase at $f_c$?

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

$$G(s) = \frac{sR_1C_1 + 1}{sR_1(C_1 + C_2) \left(1 + sR_2 \frac{C_2}{C_1 + C_2}\right)} \frac{sC_1(R_1 + R_f) + 1}{(sR_1C_1 + 1)}$$

If $C_2 << C_1$ and $R_3 << R_1$

- $\alpha_{p_1} = \frac{1}{R_1C_1}$
- $\alpha_{p_2} = \frac{1}{R_3C_2}$

1 pole at the origin
2 zeros
2 poles

Pole/zero placement and boost at $f_c$

- Phase boost appears between the zero and the pole

$$G(j\omega) = \left[\frac{1 + j\frac{\omega}{\omega_{z_1}}}{1 + j\frac{\omega}{\omega_{p_1}}}\right] \left[\frac{1 + j\frac{\omega}{\omega_{z_2}}}{1 + j\frac{\omega}{\omega_{p_2}}}\right] \quad \arg G(j\omega) = \text{boost} = \arg \left[\frac{1 + j\frac{\omega}{\omega_{p_1}}}{1 + j\frac{\omega}{\omega_{p_2}}}\right]$$

$$\arg G(f) = \arctan\left(\frac{f}{f_{z_1}}\right) + \arctan\left(\frac{f}{f_{z_2}}\right) - \arctan\left(\frac{f}{f_{p_1}}\right) - \arctan\left(\frac{f}{f_{p_2}}\right)$$

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

$$\arg G(4\,\text{kHz}) = 2 \arctan\left(\frac{4k}{500}\right) - 2 \arctan\left(\frac{4k}{50k}\right) = 166 - 4.6 \approx 161°$$
How do we boost the phase at $f_c$?

![Graph showing gain and phase boost at $f_c$]

**Gain** $G(s)$
- $G_{500\text{ Hz}} = 20 \text{ dB}$
- $f_p, f_{p0} = 500 \text{ Hz}$

**Gain at** $f_o = 21 \text{ dB}$

**Phase boost**
- at $f_c = 158^\circ$
- $-270^\circ$

**Equations**
- $G(s)$

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

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

**dc equation**

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

**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!

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

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

$$G = \frac{f_{po}}{f_c} \quad \text{Select} \ f_{po} \ \text{to have} \ G \ \text{at} \ f_c$$
Boosting the phase at $f_c$ with OTAs

The calculation is simple:

Parameters:
- $V_{out} = 385V$
- $I_b = 250\mu A$
- $V_{ref} = 2.5$
- $R_{upper} = \frac{(V_{out} - V_{ref})}{I_b}$
- $R_{lower} = \frac{2.5}{I_b}$
- $f_c = 20$
- $G_{fc} = 46$
- $g_m = 200u$
- $f_{po} = G_{fc} f_c$
- $R_{lowerOTA} = \frac{V_{ref}}{I_b}$
- $R_{upperOTA} = \frac{(V_{out} - V_{ref})}{I_b}$
- $R_{eq} = \frac{(R_{lower} + R_{upper})}{g_m R_{lower}}$
- $C_{1OTA} = \frac{1}{2\pi f_{po} R_{eq}}$
- $G = 10^{-G_{fc}/20}$
- $\pi = 3.14159$
- $C_1 = \frac{1}{2\pi f_c G R_{upper}}$

Boosting the phase at $f_c$ with OTAs

Type 1 - OTA

<table>
<thead>
<tr>
<th>$\arg G(s)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>90.0</td>
</tr>
<tr>
<td>89.0</td>
</tr>
<tr>
<td>88.0</td>
</tr>
<tr>
<td>87.0</td>
</tr>
<tr>
<td>86.0</td>
</tr>
</tbody>
</table>

$\frac{G(s)}{\arg G(s)}$
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!

\[
G(s) \approx \frac{R_{low} \cdot g_m \cdot R_2}{R_{low} + R_{upper}} \cdot \frac{1 + s \cdot R \cdot C_1}{s \cdot R \cdot C_1 \cdot (1 + s \cdot R \cdot C_2)}
\]

If $C_2 << C_1$

\[
\omega_1 = \frac{1}{R \cdot C_1}, \quad \omega_p = \frac{1}{R \cdot C_2}
\]

1 pole at the origin
1 zero
1 pole

**Type 2 - OTA**

---

**Type 2 - OTA**

---
Boosting the phase at $f_c$ with OTAs

\[ \sqrt{f_H f_p} = \sqrt{10 \times 100} = 10 \, \text{Hz} \]

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!

Type 3

1 pole at the origin
2 zeros at 500 Hz
2 poles at 50 kHz
Finally, we test the open-loop gain

An ac simulation gives us the open-loop Bode plot

![Open-loop Bode plot](image)

- Gain $T(s)$
- Phase $\phi_m = 70^\circ$
- $f_c = 4$ kHz

What transient response?

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

1. Double zero 200 Hz, $\phi_m = 80^\circ$
2. Double zero 500 Hz, $\phi_m = 70^\circ$
3. Double zero 700 Hz, $\phi_m = 60^\circ$
4. Double zero 800 Hz, $\phi_m = 50^\circ$

$\Delta I_{out} = 2$ A

$I_{out}$ vs. time in seconds

$f_c = 4$ kHz

Almost unchanged
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…

$V_{ref}(s) \rightarrow + \quad G(s) \quad k \quad H(s) \rightarrow V_{out}(s)$

$G(s) = \frac{N_c(s)}{D_c(s)}$  
$H(s) = \frac{N_p(s)}{D_p(s)}$

$V_{out}(s) = \frac{kN_c(s)N_p(s)}{D_c(s)D_p(s)} + \frac{kN_c(s)N_p(s)}{D_c(s)D_p(s)}$

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?

Open-loop Eode plot of the power stage, $H(s)$

- 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...
Oh no, it is ringing!

- The load step reveals a ringing ac output

![Graph showing ringing output](image)

\[ \Delta I_{\text{out}} = 2 \, \text{A} \]

\[ V_{\text{out}}(t) \]

\[ V_{\text{cp}}(t) \]

\[ V_{\text{cp}}(t) \text{ is first order} \]

The **RLC** network rings alone…

- \( H_1 \) is stable per Bode analysis, but \( H_2 \) is out of the loop…

![Block diagram](image)

\[ V_{\text{in}} \rightarrow H_1(s) \rightarrow H_2(s) \rightarrow V_{\text{out}}(s) \]

- 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
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

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

\[
|H(s)|
\]

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

\[
arg H(s)
\]
The DCM response is slower than in CCM

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

![Graph showing the response comparison between DCM and CCM](image)

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 $f_{p1}=Fsw/2$
  - Put a second pole $f_{p2}$ at $f_{p2}=Fsw/2$

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

This zero is brought by the output capacitor ESR

| 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!

<table>
<thead>
<tr>
<th>freq enc in hertz</th>
<th>-40.0</th>
<th>-20.0</th>
<th>0</th>
<th>20.0</th>
<th>40.0</th>
</tr>
</thead>
<tbody>
<tr>
<td>vdbout2 in volts</td>
<td></td>
<td></td>
<td>0</td>
<td></td>
<td></td>
</tr>
<tr>
<td>ph_vout2 in degrees</td>
<td>-180</td>
<td>-90.0</td>
<td>0</td>
<td>90.0</td>
<td>180</td>
</tr>
</tbody>
</table>

Subharmonic peaking now!

CCM operation, current instabilities

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

$\Delta I_{peak} = b = I_{peak} - S_1\Delta t$

Solving $\Delta t$

$I_{peak} - a = I_{peak} - b$

$S_1 = a$

$S_2 = b$

$S_2 = \frac{d}{d_1}$

$\Delta I_L(nT_{sw}) = \Delta I_L(0) \left( -\frac{d}{d_1} \right)^n$
CCM operation, current instabilities

Duty-cycle < 50%

\[ \Delta I_L(nT_{sw}) = \Delta I_L(0) \left(-\frac{d}{d^*}\right)^n \]

Asymptotically stable

Perturbation has gone.

Duty-cycle > 50%

Asymptotically unstable

Inject ramp compensation

\[ \frac{S_a}{S_2} > 50\%S_2 \]

Must stay below 1

\[ \Delta I_L(nT_{sw}) = \Delta I_L(0) \left[ 1 - \frac{S_a}{S_2} \right]^{d/d_a} \]

Up to \( d = 100\% \)
### A buck CCM in current-mode

- **A SPICE model can predict subharmonic instabilities**

![Diagram of a buck CCM in current-mode](image)

**Parameters**
- $V_{out}=5V$
- $R_{upper}=10k$
- $f_c=10$
- $G_{fc}=5.5$
- $G=10^{-(G_{fc}/20)}$
- $\pi=3.14159$
- $C_1=1/(2\pi f_c G R_{upper})$
- $f_{po}=1/(2\pi C_1 R_{upper})$

**Biasing**
- $R_{load}=1$
- $V_{in}=10$
- $AC=0$

**Injecting ramp damps the double pole**

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

![Graph of the voltage-mode response](image)
The right way to inject ramp compensation

\[ R_{\text{comp}} = 1k \cdot \frac{S}{S' M} \]

- \( S \) = 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

\[ Z_{\text{out}}(s) \approx R_{\text{load}} \parallel C_{\text{out}} \frac{f_0}{s} \]

\( v_{\text{out}} \) = \( V_{\text{out}} \)

- \( V_{\text{out}} \) = output voltage
- \( Z_{\text{out}} \) = output impedance
- \( f_0 \) = frequency of the output signal
- \( s \) = complex frequency
- \( dB\Omega \) = decibels per ohm

- Frequency in Hz
- dBΩ
- Voltage-mode peaking

\[ Z_{\text{out}}(s) \text{ in } \Omega \]
The closed-loop output impedance of CM and VM

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

![Graph showing open-loop output impedances with $R_{load}$, $C_{out}$, and $f_{L}$ for current mode and $f_{HF}$ for voltage mode.]

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

![Graph showing closed-loop output impedances with $f_{c}$.]
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

Output response to a 2-A step
Transition from CCM to DCM in current-mode

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

<table>
<thead>
<tr>
<th>H(s)</th>
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<tbody>
<tr>
<td>CCM</td>
</tr>
<tr>
<td>DCM</td>
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<th>T(s)</th>
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<td>CCM</td>
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<td>DCM</td>
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<tbody>
<tr>
<td>CCM</td>
<td>DCM</td>
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</table>

- Once compensated, $f_c$ still changes but $\phi_m$ is still ok

<table>
<thead>
<tr>
<th></th>
<th>T(s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>CCM</td>
<td>DCM</td>
</tr>
</tbody>
</table>

$|H(s)|$, $\arg H(s)$, $|T(s)|$, $\arg T(s)$
The DCM response is not too much affected

- The step response is almost unchanged...

![Graph showing DCM response](image)

- $\Delta I_{out} = 100 \text{ mA}$
- $\Delta V_{out} = 4 \text{ mV}$

- For $v_{out2}$, $v_{out2}$ in volts

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:

$$\arg(T(f_c)) = \text{boost} = \arg\left(\frac{1 + \frac{f_c}{f_{z_0}}}{1 + \frac{f_c}{f_{p_0}}}\right) = \arctan\left(\frac{f_c}{f_{z_0}}\right) - \arctan\left(\frac{f_c}{f_{p_0}}\right)$$

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

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

$$\arctan(x) + \arctan\left(\frac{1}{x}\right) = 90° \quad \Rightarrow \quad k = \tan\left(\frac{\text{boost}}{2} + 45\right)$$
The k factor in an automated type 2

- Suppose we have the following specs:
  - $f_c = 1$ kHz, $|H(1kHz)| = -10$ dB
  - $\text{Arg } H(1kHz) = -100^\circ$
  - $\varphi_m = 70^\circ$  
    $\text{BOOST} = \varphi_m - \text{arg } H(f_c) - 90^\circ = 70^\circ + 100^\circ - 90^\circ = 80^\circ$
  - 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 $1k \times 11.4 = 11.4$ kHz
- Adjust mid-gain to reach 10 dB@1 kHz

The k factor in an automated type 2

- Calculate the elements for a type 2 amplifier

  $$C_1 = \frac{1}{2\pi f_c G k R_{\text{upper}}} = \frac{1}{6.28 \times 1k \times 3.16 \times 11.4 \times 10k} = 442 \ pF$$

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

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

  $$G = \frac{G f_c}{10} = \frac{10^{10}/20}{10} = 3.16$$
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)
R2=k/(2*pi*fc*C1)

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

- Results

The numbers exactly match the predictions
The k factor in an automated type 2

- Adjusting k modulates the resulting phase boost

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:

\[
\arg(T(f_c)) = \text{boost} = \arg \left( \frac{1 + \frac{f}{f_p}}{1 + \frac{f}{f_z}} \right) = 2 \arctan \left( \frac{f}{f_p} \right) - 2 \arctan \left( \frac{f}{f_z} \right)
\]

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

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

\[
\text{boost} = 2 \left[ \arctan \left( \sqrt{k} \right) + \arctan \left( \sqrt{k} - 90 \right) \right] = 4 \arctan \left( \sqrt{k} \right) - 180
\]

\[
k = \left[ \tan \left( \frac{\text{boost}}{4} + 45 \right) \right]^2
\]
The k factor in an automated type 3

Suppose we have the following specs:

- \( f_c = 1 \text{ kHz} \), \(|H(1kHz)| = -15 \text{ dB}\)
- \( \text{Arg } H(1kHz) = -140^\circ \)
- \( \varphi_m = 70^\circ \) \( \text{BOOST} = \varphi_m - \text{arg } H(f_c) - 90^\circ = 70^\circ + 140 - 90 = 120^\circ \)
- 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 \( 1k/3.7 = 268 \text{ Hz} \)
Place a pole at \( 1k \times 3.7 = 3.7 \text{ kHz} \)
Adjust mid-gain to reach 15 dB@1 kHz

The k factor in an automated type 3

Calculate the elements for a type 3 amplifier

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

\[
C_1 = C_2 (k - 1) = 2.85n \times (13.9 - 1) = 36.7 \text{ nF}
\]

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

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

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

\[
G = 10^{\frac{20}{30}} = 10^{\frac{5}{20}} = 5.6
\]
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*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)
fx1=1/(2*pi*R2*C1)
fx2=1/(2*pi*Rupper*C3)
```

*** mainckt
RUPPER = 1.000e+004
FC = 1.000e+003
PM = 7.000e+001
GFC = -1.500e+001
PS = -1.400e+002
G = 5.823e+000
BOOST = -1.200e+002
PI = 3.142e+000
K = 1.393e+01
C2 = 2.336e+006
C1 = 5.055e+006
R2 = 1.623e+004
R3 = 7.734e+002
C3 = 5.134e+000
R1 = 1.000e+004
FX1 = 3.732e+003
FX2 = 3.668e+002
FX3 = 2.877e+002

```
results
```
**The k factor in an automated type 3**

- Adjusting k modulates the resulting phase boost

![Graph showing phase and gain for different k values.]

**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

![Conditional stability diagram]

- 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 $G(s)$

$$G(s) = \frac{R_2}{R_1} \left( \frac{1}{sR_1C_3} + 1 \right) \left( \frac{1}{sR_2C_2} + 1 \right)$$

$$G(f) = \left| \frac{R_2}{R_1} \right| \begin{pmatrix} \frac{1}{s} + \frac{1}{s_p} \\ \frac{1}{s} + \frac{1}{s_{p1}} \end{pmatrix}$$

$$R_2 = \frac{\sqrt{(f_{p1}^2 + f_c^2)(f_{p2}^2 + f_c^2)}}{GfR_1} \frac{R_1}{f_{p1}}$$

Type 3
Manual pole/zero placement is a solution

- Conditional stability is gone
- Fine tuning is now possible

- **Automated calculations help iterations!**

<table>
<thead>
<tr>
<th>Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rupper = 10k</td>
</tr>
<tr>
<td>Rlower = Rupper</td>
</tr>
<tr>
<td>G = 10^(-Gfc/20)</td>
</tr>
<tr>
<td>( \pi = 3.14159 )</td>
</tr>
<tr>
<td>( f_z1 = 1k )</td>
</tr>
<tr>
<td>( f_z2 = 1k )</td>
</tr>
<tr>
<td>( f_p1 = 26k )</td>
</tr>
<tr>
<td>( C_3 = \frac{1}{2\pi f_z1 Rupper} )</td>
</tr>
<tr>
<td>( R_3 = \frac{1}{2\pi f_z1 R_2 C_3} )</td>
</tr>
<tr>
<td>( a = f_z1^4 + f_z1^2 f_z2^2 + f_z2^4 + f_z1^2 f_z2^2 )</td>
</tr>
<tr>
<td>( c = f_p2^2 f_p1^2 + f_z1^2 f_p1^2 + f_z1^2 f_p2^2 )</td>
</tr>
<tr>
<td>( R_2 = \sqrt{ac}/0.5 f_z1 R_3 f_p1 )</td>
</tr>
<tr>
<td>( f_z1 x = \frac{1}{2\pi C_1 R_2} )</td>
</tr>
<tr>
<td>( f_z2 x = \frac{1}{2\pi C_3 (Rupper + R_3)} )</td>
</tr>
<tr>
<td>( f_p1 x = \frac{1}{2\pi (C_1 C_2 / (C_1 + C_2)) R_2} )</td>
</tr>
<tr>
<td>( f_p2 x = \frac{1}{2\pi C_3 R_3} )</td>
</tr>
<tr>
<td>( X_1 = 1 )</td>
</tr>
<tr>
<td>( X_2 = 2.5 )</td>
</tr>
<tr>
<td>AC = 1000 V</td>
</tr>
<tr>
<td>AC = 0.10 V</td>
</tr>
<tr>
<td>AC = 0 V</td>
</tr>
</tbody>
</table>

- **|H(s)|**
- **|T(s)|**

- **f_0 = 1.2 kHz**
- **f_p1,2 = 26 kHz**
- **f_z1,2 = 1 kHz**
The zeros position affects the response

- Transient response changes as one zero position moves

\[ f_{z1} = 3 \text{ kHz} \]
\[ \phi_m = 75^\circ \]
\[ f_{z1} = 1 \text{ kHz} \]
\[ \phi_m = 85^\circ \]
\[ f_{z1} = 0.3 \text{ kHz} \]
\[ \phi_m = 90^\circ \]
\[ f_{z2} = 1 \text{ kHz} \]

The zeros position affects the response

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

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Overshoot</th>
<th>Settling time</th>
<th>Phase margin</th>
</tr>
</thead>
<tbody>
<tr>
<td>( f_{z1} )</td>
<td>( f_{z2} )</td>
<td>faster</td>
<td>slower</td>
</tr>
</tbody>
</table>
By decreasing $f_{z1}$, DCM stability is improved

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

Poles / zeros compensation
- $f_{z1} = 300$ Hz, $f_{z2} = 1$ kHz
- $f_{p1}, f_{p2} = 26$ kHz

Improving DCM stability slows down the buck

- $f_{z1}$ moving low, it hampers the response time...

K factor compensation
- $f_{z1}, f_{z2} = 2.2$ kHz
- $f_{p1}, f_{p2} = 46$ kHz
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 $C_{\text{zero}}$ is a short circuit, the slow lane is off
- The TL431 turns into a static **Zener diode**

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

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

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

---

The TL431 and its equivalent schematic

- The fast lane presence adds a zero in the equation

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

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

No high frequency pole? Type 2
Adding a pole for a type 2 circuit

- The pole is a simple capacitor on the collector.

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

- \( f_p = \frac{1}{2\pi R_{upper}C_{zero}} \)
- \( f_z = \frac{1}{2\pi R_{upper}C_{zero}} \)
- \( G = \frac{R_{pullup}}{R_{LED}} \)
- \( f_p = \frac{1}{2\pi R_{pullup}C_{pole}} \)

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

The TL431 type 2 is not a panacea!

- Because $R_{LED}$ must also ensure 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

\[ \frac{V_{out} - V_{I} - V_{T,431,\text{min}}}{V_{dd} - V_{CE,\text{sat}}} \leq R_{pullup} \cdot CTR_{\text{min}} \]

\[ R_{LED,\text{max}} \leq \frac{5 - 1 - 2.5}{5 - 0.3} \times 0.3 \times 20k \leq 1.9 \Omega \]

\[ G_{b} \geq \frac{R_{pullup}}{R_{LED,\text{max}}} \]

\[ G_{b} \geq \frac{V_{dd} - V_{CE,\text{sat}}}{V_{out} - V_{f} - V_{T,431,\text{min}}} \geq 3.13 \text{ or } \approx 10 \text{ dB} \]
A type 3 is also possible

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

If $\arg H(s)@f_c$ is less than 45°…

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

$$R_{LED,max} \leq \frac{5 - 1 - 2.5}{5 - 0.3} \times 0.3 \times 20k \leq 1.9 \, k\Omega \quad \text{20\% margin}$$

$$R_{LED,max} = 0.8 \times 1.9 \, k\Omega = 1.5 \, k\Omega$$

$$G(s) = \left[ \frac{sR_{upper}C_{zero} + 1}{sR_{upper}R_{LED}C_{zero}} \right] \left[ \frac{1}{1 + sR_{pullup}C_{pole}} \right]$$

$$f_{po} = f_c \cdot G \quad \rightarrow \quad C_{pole} = \frac{CTR}{2\pi f_p R_{LED}} \quad C_{zero} = \frac{R_{pullup}}{R_{upper}}C_{pole}$$

$$\omega_{po} = \frac{1}{sR_{upper}R_{LED}C_{zero}}$$
The type 3 with a TL431

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

$$G(s) = \frac{V_{fb}(s)}{V_{out}(s)} = \left(\frac{sR_{upper}C_{zero} + 1}{sR_{upper}C_{zero}}\right) \left(\frac{1}{1 + sR_{pullup}C_{pole2}}\right) \left[\frac{sC_{pc}(R_{LED} + R_{pc}) + 1}{R_{pullup}CTR}\right]$$

$$f_{pz} = \frac{1}{2\pi R_{upper}C_{zero}} \quad f_{p1} = \frac{1}{2\pi R_{pullup}C_{pole2}}$$

$$f_{p2} = \frac{1}{2\pi R_{LED}CTR} \quad f_{p3} = \frac{1}{2\pi R_{LED}C_{pole2}}$$

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

Type 3

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
The optocoupler is the traitor here!

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

\[
CTR = \frac{I_c}{I_F} \times 100
\]

Current Transfer Ratio

Luigi Galvani, 1737-1798
Italian physician and physicist

The internal pole should be known

- The photons are collected by a collector-base area.
- This area offers a large parasitic capacitance.

\[
\frac{V_{FB}(s)}{V_{out}(s)} = \frac{\frac{R_{pullup} \cdot CTR}{R_{LED}} \cdot \frac{1}{1 + sR_{pullup}C}}{V_{dd}}
\]

If \( f_p \) is above 5 times \( f_c \), 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

normalized to 1 (0 dB)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>-4</th>
<th>-3</th>
<th>-2</th>
<th>-13</th>
<th>-24</th>
<th>-14</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$I_{C}$/% $I_{B}$=10 mA</td>
<td>40-80</td>
<td>40-125</td>
<td>63-200</td>
<td>63-200</td>
<td>100-320</td>
<td>100-320</td>
<td>%</td>
</tr>
<tr>
<td>% $I_{C}$/% $I_{B}$=1.0 mA</td>
<td>30&gt;13</td>
<td>30&gt;13</td>
<td>30&gt;13</td>
<td>30&gt;13</td>
<td>45&gt;12</td>
<td>45&gt;12</td>
<td></td>
</tr>
<tr>
<td>Collector Emitter Leakage Current, $I_{CBO}$, $V_{CE}$=10 V</td>
<td>2.0&lt;500</td>
<td>2.0&lt;500</td>
<td>5.0&lt;1000</td>
<td>5.0&lt;1000</td>
<td>5.0&lt;1000</td>
<td>5.0&lt;1000</td>
<td>nA</td>
</tr>
</tbody>
</table>

CTR between 0.63 and 1.25

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
  - CTR dispersion increases
- Low CTR optocouplers have low internal capacitance
- Higher drive currents improve speed but:
  - Shorten lifetime
  - Degrade standby power

Understand the optocoupler impact on your design!
Find the pole position on your optocoupler

- Once your optocoupler is selected, characterize it:

  ![Diagram of optocoupler circuit]

  - \(V_{bias}\) 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

  ![Graph showing pole at 10 kHz]

  - \(R_{pullup} = 4.7\) kΩ  SFH615A-2

\(V_{ce} = 1\) V
\(V_{ce} = 2.5\) V

\(R_{bias}\)
Changing the pullup affects the pole position

- A low pullup resistor offers better bandwidth!

![Graphs showing pullup resistor effects](image)

- Changing the bias point affects the CTR
  
  $$\frac{V_{fb}(s)}{V_{out}(s)} = -\frac{R_{pullup}}{R_{LED}} \cdot CTR$$

- If $R_{pullup} = R_{LED}$, then $|G_{ol}| = 0 \text{ dB}$...

To push the pole farther, use a cascode

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

![Cascoding circuit](image)
A simple optocoupler model

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

\[
C_{\text{pole}} = \frac{1}{2\pi R_{\text{pullup}} f_{\text{opto}}}
\]

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 $C_{pole}$

  \[ \text{From } G(s) \text{ calculations} \rightarrow C \rightarrow C_p = C - C_{pole} \]

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

\[ f_{z} \quad f_{p} \quad f_{zoom} \quad f_{o} \quad f_{c} \quad f_{p} \]

\[ |G(s)| \]

\[ R_{pullup} \]

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

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

Apply k factor or other method, get $f_z$ and $f_p$

- $f_z = 3.5$ kHz
- $f_p = 4.5$ kHz

$C_p = 3.8 \text{nF}$

install

$C = 3.8n - 1.3n \approx 2.5nF$
Stabilizing a DCM flyback converter

- Sweep ESR values and check margins again

- PM at 1 kHz = 60°
- Cross over 1 kHz
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

\[ I_{dc} + I_{ac} \]

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

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

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

Neg. sign!

\[
\frac{dV_{in}}{dl_{in}} = \frac{d}{dl_{in}} R_{load} I_{out}^2 = -R_{load} \left(\frac{I_{out}}{I_{in}}\right)^2
\]

SPICE simulation
constant power sink

\[\text{Transfer function analysis}\]

\[
\text{output_impedance at } V(4) = 0.000000e+000
\]

\[
v1\#input_impedance = -1.66667e+002
\]

\[
\text{Transfer function} = 1.000000e+000
\]

P = 60 W

The need for an input filter

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

\[\begin{align*}
R_1 & \quad L & \quad R_2 \\
\downarrow & \quad C & \quad R_3
\end{align*}\]

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

\[
T(s) = \frac{1}{s^2 + \frac{s}{\omega_0} + 1} \quad Q = \frac{1}{Q}
\]

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

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

\[
\zeta = \frac{L + C \left( R, R_1, R_2, R_3, R_4 \right)}{2 \left( R_1 + R_2 \right)} \quad \omega_0 = 0 \quad R_3 = \frac{R, R, C + L}{C \left( R_1 + R_2 \right)}
\]
A tunnel-based oscillator?

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

\[ R_t = \frac{R \cdot R_s + L}{C \cdot (R_s + R_t)} = \frac{100 \text{m} \cdot 500 \text{m} \cdot 100 \mu \text{F}}{1 \text{kHz} \cdot (600 \text{m})} = -166 \Omega \]

- Diverging oscillations \( \zeta < 0 \)
- Steady-state oscillations \( \zeta = 0 \)
- Damped oscillations \( \zeta > 0 \)

Watch the output/input impedances

- A possibility to look for oscillations is to ac sweep \( Z_{\text{out}} \) of RLC

\[ Z_{\text{out}} = R_{\text{out}} = \begin{cases} 166 \Omega & 44.4 \text{ dB} \Omega \\ 0 \Omega & \zeta < 0 \\ \frac{Z_0^2}{R} \sqrt{1 + \left( \frac{R}{Z_0} \right)^2} \text{ dB} \Omega \end{cases} \]

To be safe

\[ \left| Z_{\text{out}} \right| \ll \left| Z_{\text{in}} \right| \]

\[ R_{\text{in}} = 166 \Omega \text{ at } 44.4 \text{ dB} \Omega \]

\[ R_{\text{in}} = 500 \text{ m} \Omega \text{ at } 44.4 \text{ dB} \Omega \]
The filter effect also appears on the Bode plot

- The filter peaking brings instabilities…

![Bode plot with filter effect](image)

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!

![RLC network diagram](image)
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 $C_1$

![Graph showing damping resistor calculation](image)

Stabilizing a UC3843 converter

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

![Circuit diagram of UC3843 converter](image)
Change the operating conditions easily

CCM operation, $R_{\text{load}} = 6.3 \, \Omega$

Reduce the load to enter in DCM

DCM operation, $R_{\text{load}} = 20 \, \Omega$
From the open-loop Bode plot, compensate

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

Calculate mid-band gain: +18 dB

\[ R_{LED} = \frac{R_{pulse} \cdot CTR}{10^{\pi}} = \frac{4.7k \times 0.45}{7.94} = 266 \Omega \]

We place a zero at 300 Hz:

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

We place a pole at 3.3 kHz:

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


Verify in the lab. the open-loop gain

- Sweep extreme voltages and loads as well!

Simulated operation, \( R_{load} = 6.3 \Omega \), \( V_{in} = 150 \text{ Vdc} \)
Verify in the lab. the open-loop gain

CCM operation, $R_{\text{load}} = 6.3 \, \Omega$, $V_{\text{in}} = 330 \, V_{\text{dc}}$

DCM operation, $R_{\text{load}} = 20 \, \Omega$, $V_{\text{in}} = 330 \, V_{\text{dc}}$
As a final test, step load the output

- Good agreement between curves!

Vin = 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

Vin = 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!

Merci !
Thank you!
Xiè-xiè!