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Analytical Analysis and Small-Signal Modeling of a QR Power Converter

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Course Agenda

- What is a QR Converter?
- Deriving Operating Conditions
- The Over Power Problem
- Three Operating Modes
- Small-Signal Analysis – QR
- Small-Signal Analysis – VCO
- Compensating with TL431
- Control Loop Design Example
- Conclusion

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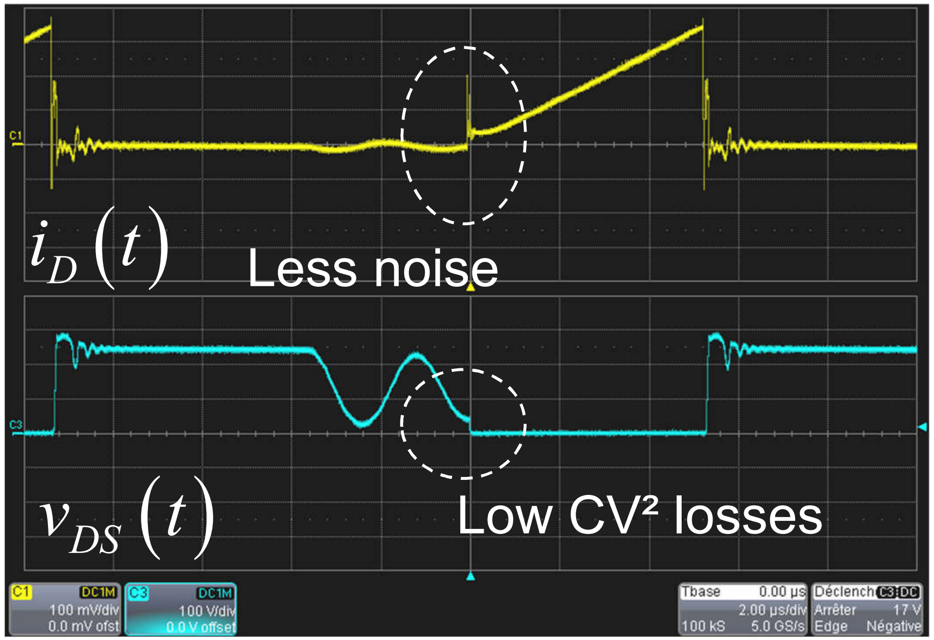
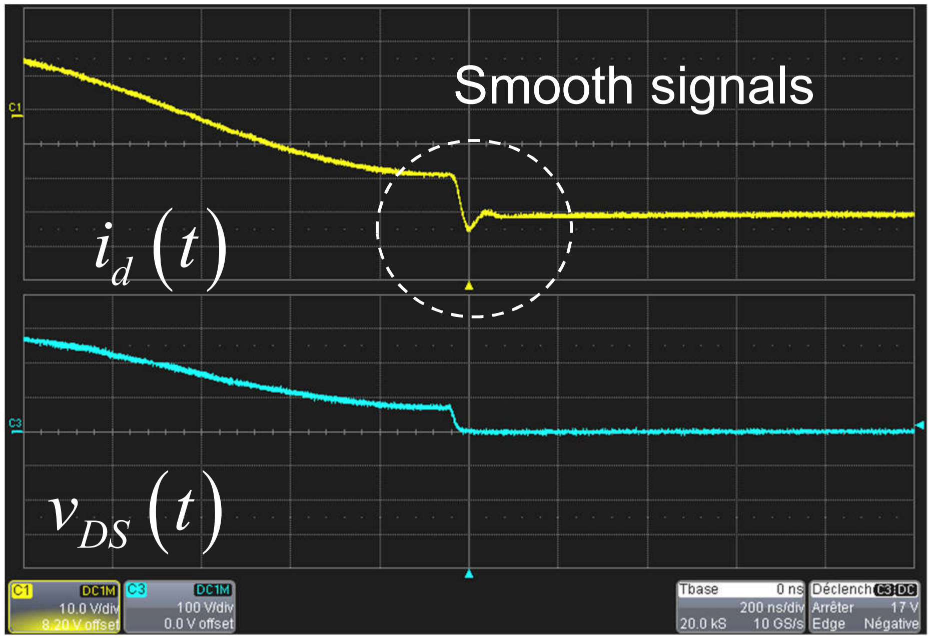
Why QR Operation?

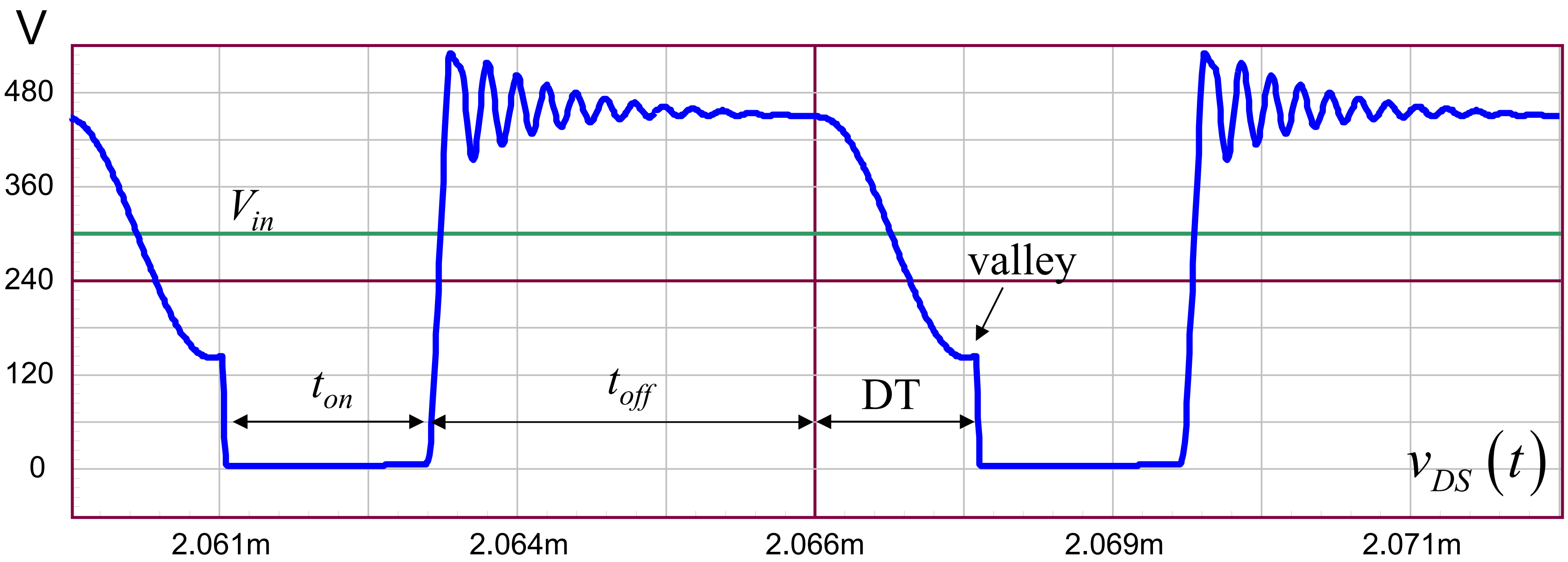
- ❑ More converters are using variable-frequency operation
- ❑ This is known as Quasi-Square Wave Resonant mode: QR
- ➢ Valley switching ensures extremely low capacitive losses
- ➢ DCM operation saves losses on the secondary diode
- ➢ Easier synchronous rectification
- ➢ The Right Half-Plane Zero is pushed to high frequencies





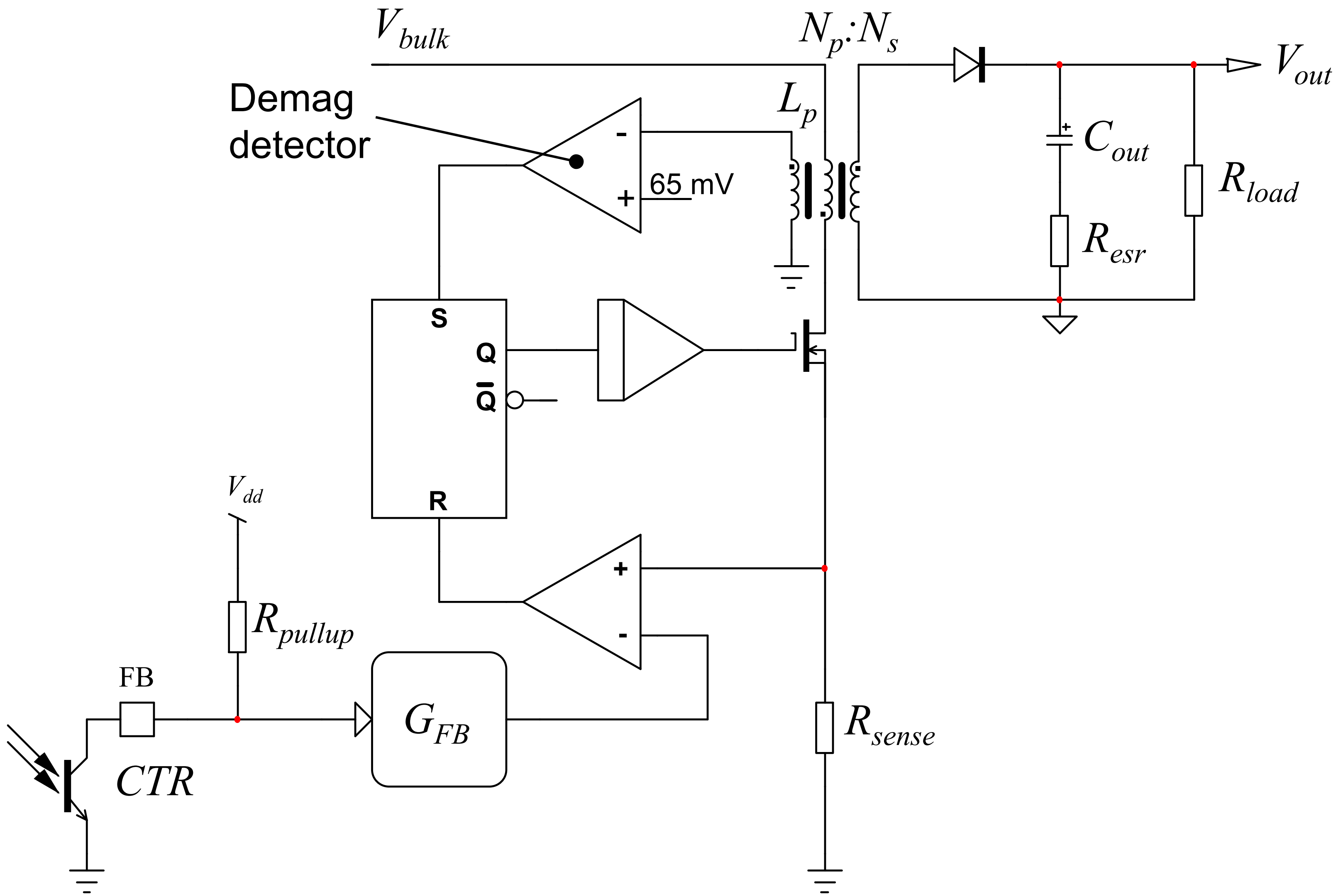
What is the Principle of Operation?

- ❑ The drain-source signal is made of peaks and valleys
- ❑ A valley presence means:
  - ➢ The drain is at a minimum level, capacitors are naturally discharged
  - ➢ The converter is operating in the discontinuous conduction mode



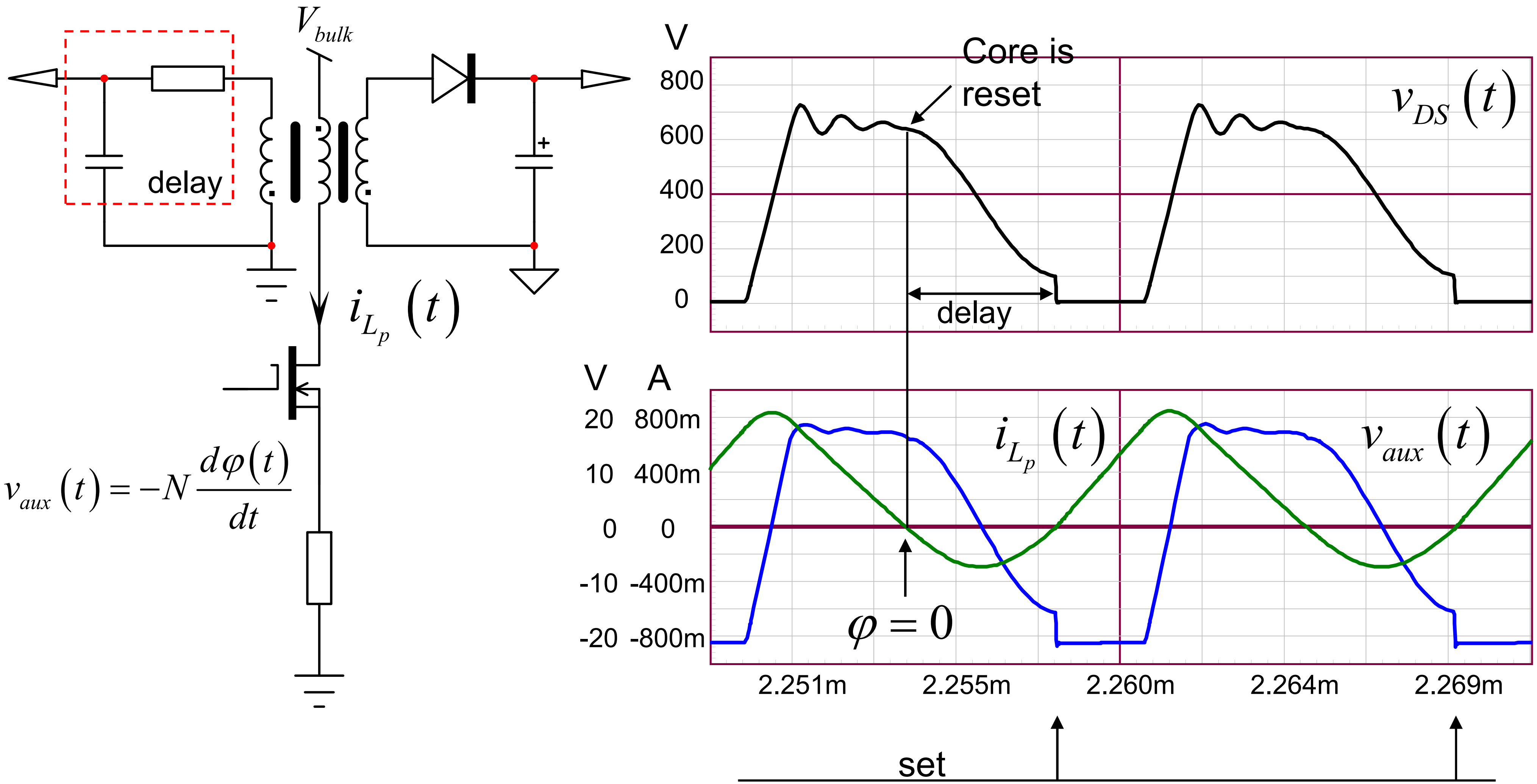
A QR Circuit Does not Need a Clock

- The system is a self-oscillating current-mode converter



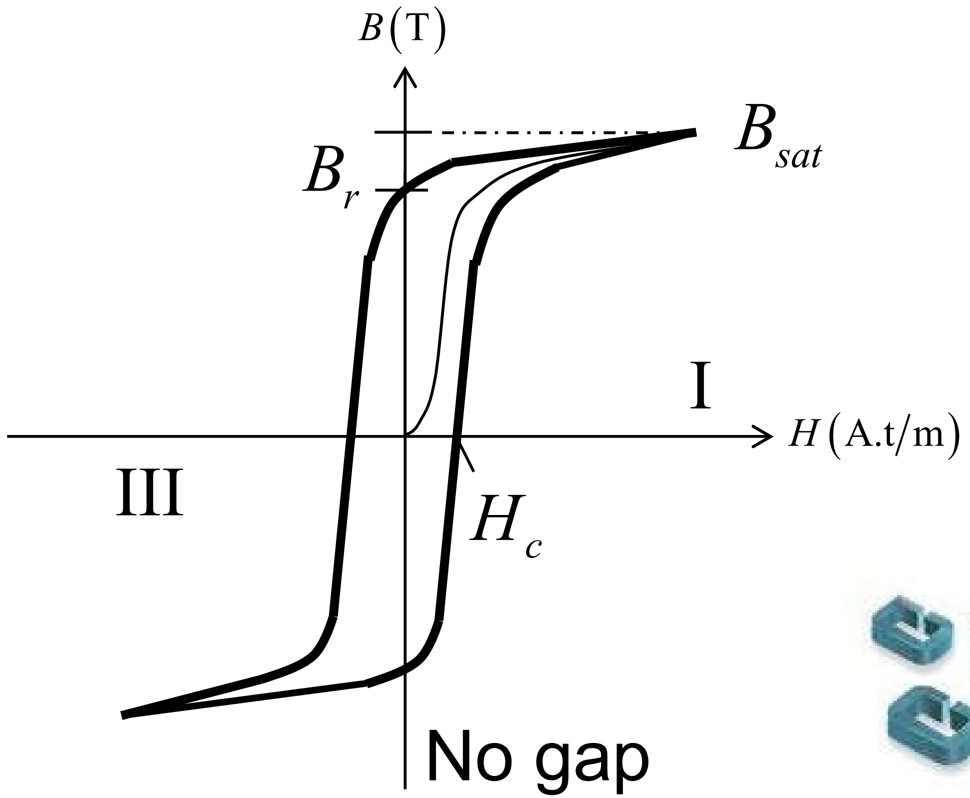
A Winding is Used to Detect Core Reset

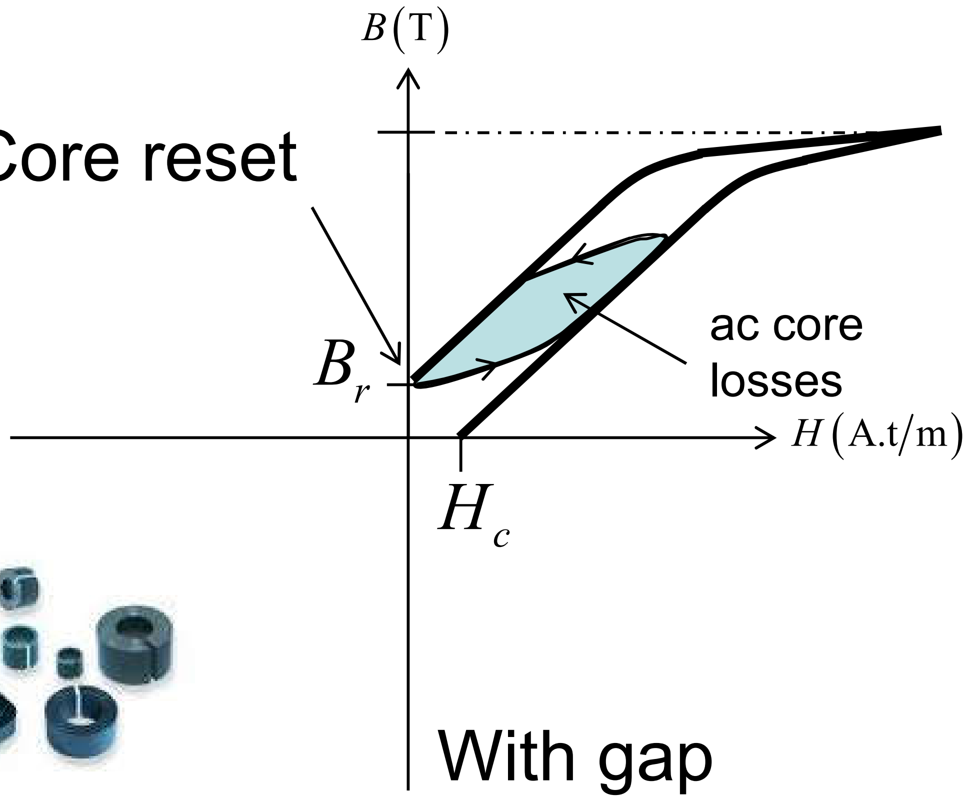
- ❑ When the flux returns to zero, the aux. voltage drops
- ❑ Discontinuous Mode is always maintained



What Does Core Reset Mean?

- The core flux density returns to zero in a switching cycle

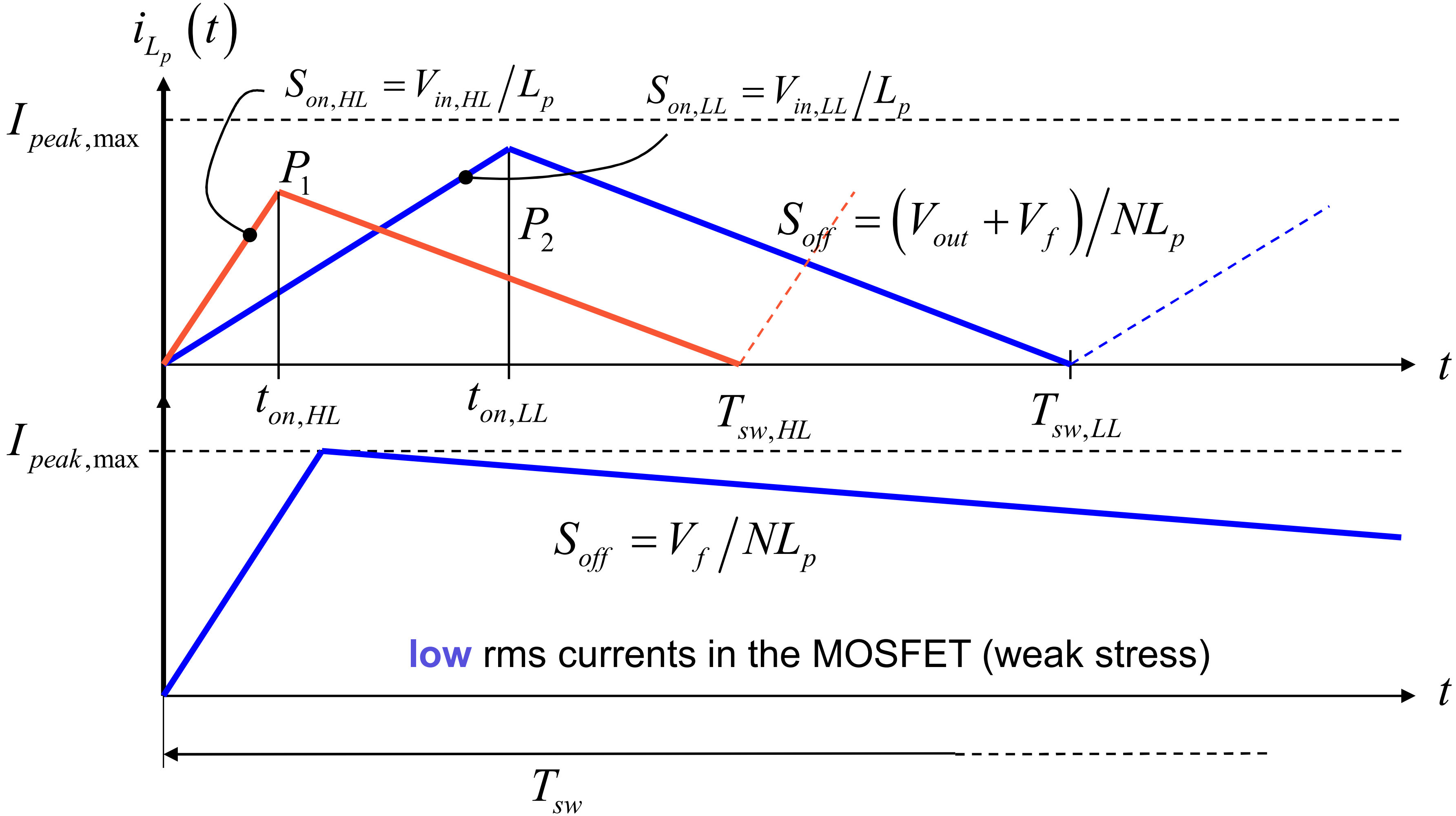




- There is always a remanent flux in the transformer core

Frequency Changes over Line and Load

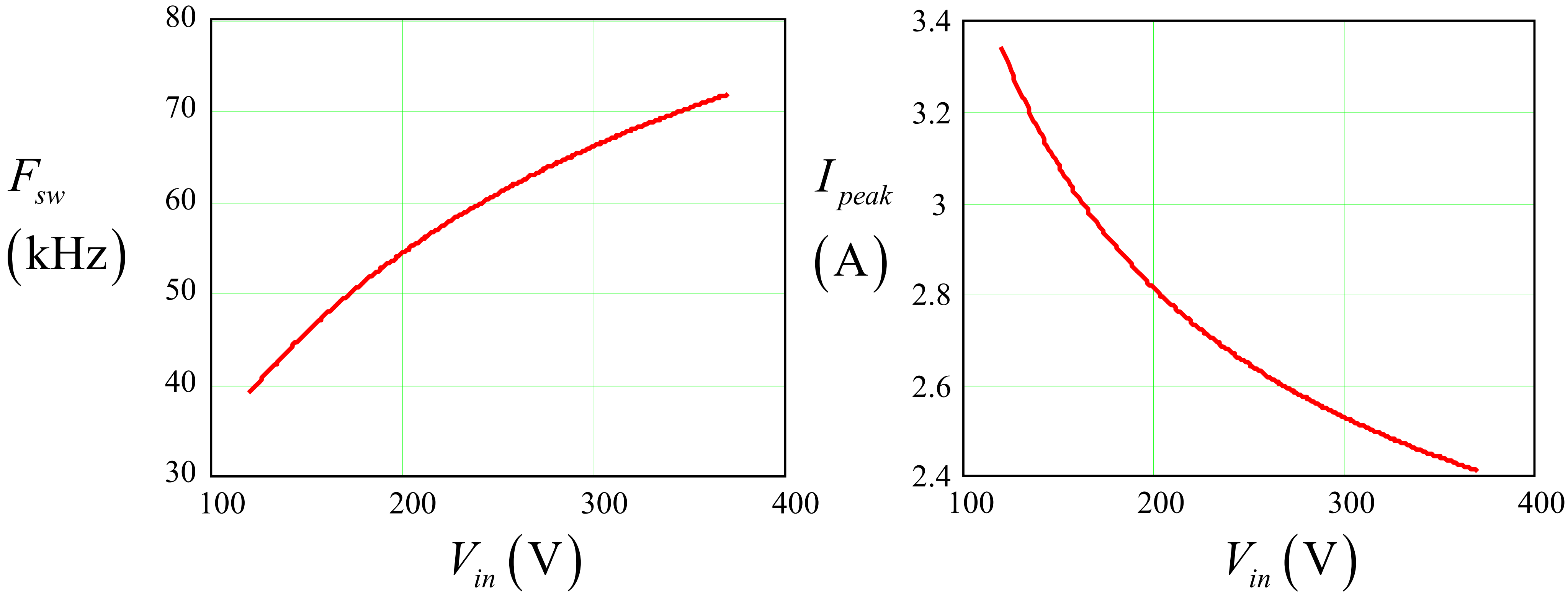
- As the peak current and the on-slope vary, T_{sw} changes



- Excellent behavior in short circuit conditions!

The Excursion Can be Quite Large

- In heavy load low-line conditions, F_{sw} decreases
- In light-load and high-line operations, F_{sw} can go very high



- EMI and switching losses are at stake as F_{sw} goes up
- Standby power obviously suffers from this condition

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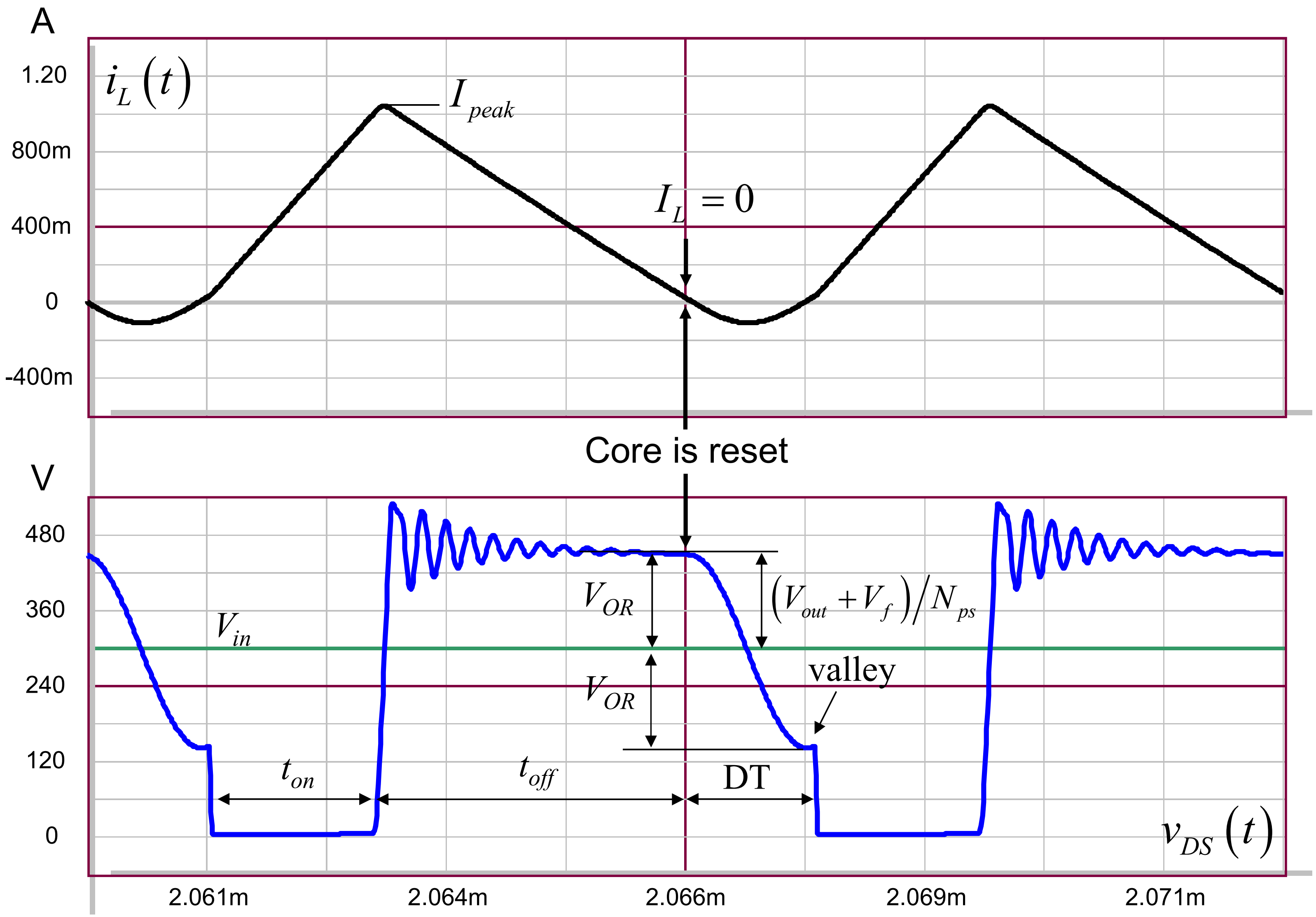
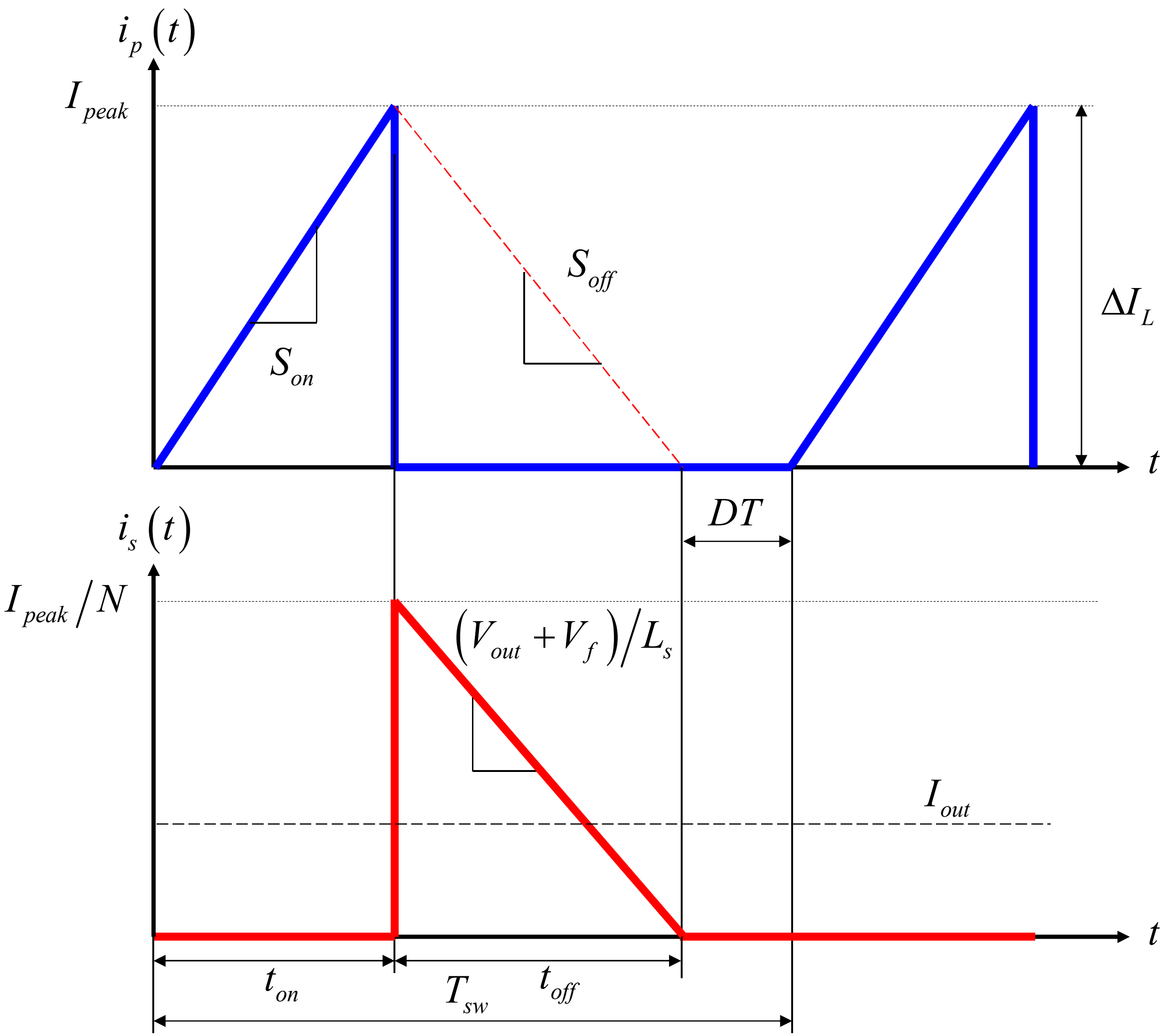
QR Converter Typical Waveforms

- The waveforms at the switch opening are smooth



A Variable Frequency Converter

- S_{on} and S_{off} are constant parameters at steady state

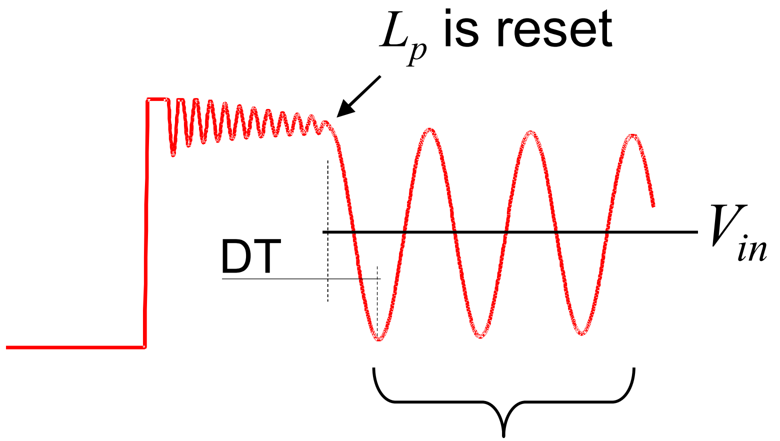


A Variable Frequency Converter

- Frequency depends on the *on* and *off* slopes

$$S_{on} = \frac{V_{in}}{L_p} \qquad S_{off} = -\frac{V_{out} + V_f}{NL_p}$$

- …but also on the valley-switching deadtime



$$f_{OSC} = \frac{1}{2\pi\sqrt{\left(L_p + l_{leak}\right)C_{lump}}}$$

1[st] valley

$$\mathrm{DT} = \pi\sqrt{\left(L_p + l_{leak}\right)C_{lump}}$$

n[th] valley

$$\mathrm{DT} \approx \left(2n-1\right)\pi\sqrt{L_p C_{lump}}$$

What is the Switching Period?

- The total period depends on t_{on} and t_{off} plus DT

$$T_{sw} = t_{on} + t_{off} + \mathrm{DT}$$

- From slope definitions, we have:

$$I_{L,peak} = S_{on} t_{on} \qquad I_{L,peak} = \frac{V_{in}}{L_p} t_{on} \qquad \longrightarrow \qquad t_{on} = \frac{L_p I_{L,peak}}{V_{in}}$$

$$I_{L,peak} = S_{off} t_{off} \qquad I_{L,peak} = \frac{V_{out} + V_f}{N L_p} t_{off} \qquad \longrightarrow \qquad t_{off} = I_{L,peak} \frac{N L_p}{V_{out} + V_f}$$

- A flyback operated in the Discontinuous Mode delivers

$$P_{out} = \frac{1}{2} F_{sw} L_p I_{L,peak}{}^2 \eta \qquad \longrightarrow \qquad I_{L,peak} = \sqrt{\frac{2 T_{sw} P_{out}}{L_p \eta}}$$

What is the Switching Period?

- Update t_{on} and t_{off} definitions with $I_{L,peak}$

$$t_{on} = \frac{L_p}{V_{in}}\sqrt{\frac{2T_{sw}P_{out}}{L_p\eta}} \qquad t_{off} = \sqrt{\frac{2T_{sw}P_{out}}{L_p\eta}}\frac{NL_p}{V_{out}+V_f}$$

- Include the dead-time contribution

$$T_{sw} = \sqrt{\frac{2T_{sw}P_{out}}{L_p\eta}}\left(\frac{L_p}{V_{in}}+\frac{NL_p}{V_{out}+V_f}\right)+\mathrm{DT}$$

- Rearrange and rewrite

$$-T_{sw} + \sqrt{T_{sw}}\sqrt{\frac{2P_{out}}{L_p\eta}}\left(\frac{L_p}{V_{in}}+\frac{NL_p}{V_{out}+V_f}\right)+\mathrm{DT}=0$$

$$x = \sqrt{T_{sw}}$$

$$B = \sqrt{\frac{2P_{out}}{L_p\eta}}\left(\frac{L_p}{V_{in}}+\frac{NL_p}{V_{out}+V_f}\right)$$

The Switching Frequency Definition

- Solve the equation to extract the final definition

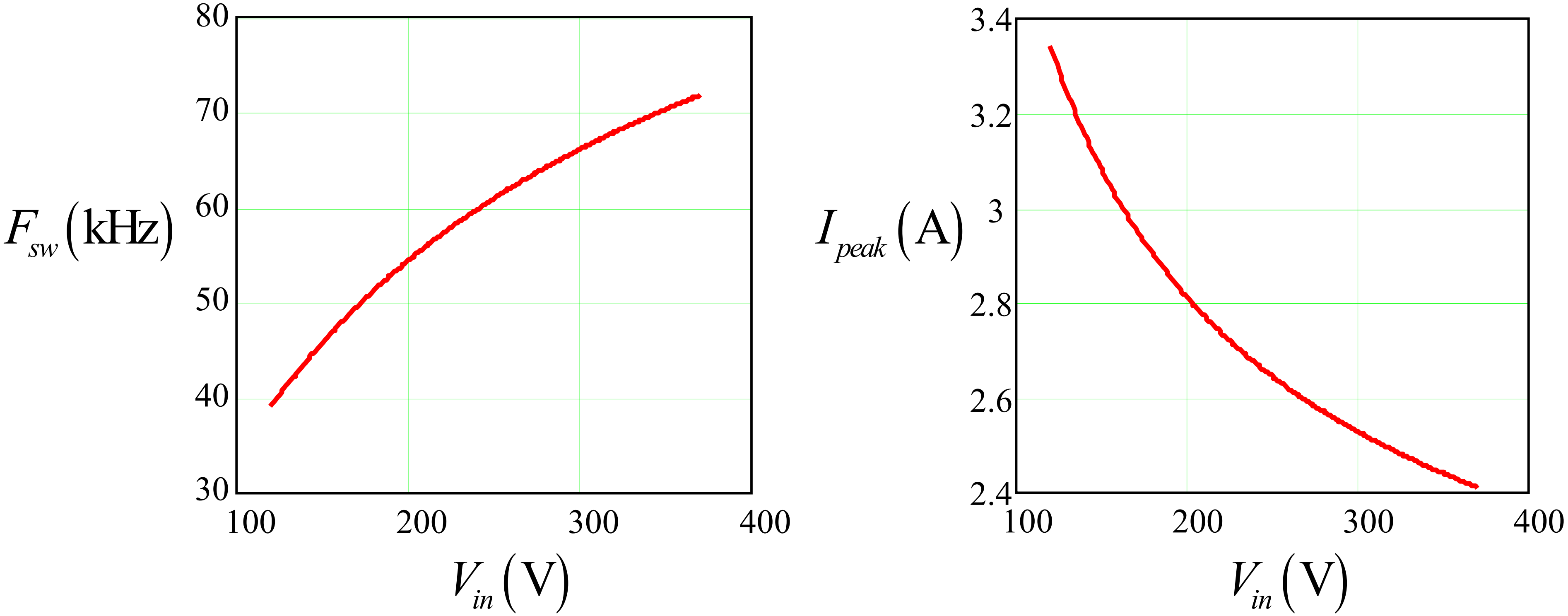
$$-x^2 + Bx + DT = 0 \longrightarrow x = \frac{\left(B + \sqrt{B^2 + 4DT}\right)^2}{4}$$

- Replace and trust Mathcad®!

$$F_{sw} = \frac{4}{\left(\sqrt{4\text{DT} + \frac{2L_p P_{out}\left(V_f + V_{out} + NV_{in}\right)^2}{\eta V_{in}{}^2\left(V_{out} + V_f\right)^2}} + \frac{\sqrt{2}L_p\left(V_f + V_{out} + NV_{in}\right)\sqrt{\frac{P_{out}}{\eta L_p}}}{V_{in}\left(V_{out} + V_f\right)}\right)^2}$$

Frequency and Peak Current Change

- Frequency increases in light-load conditions



- Peak current demand is larger at low line than high line…
- More power capability at high line brings OPP problems!

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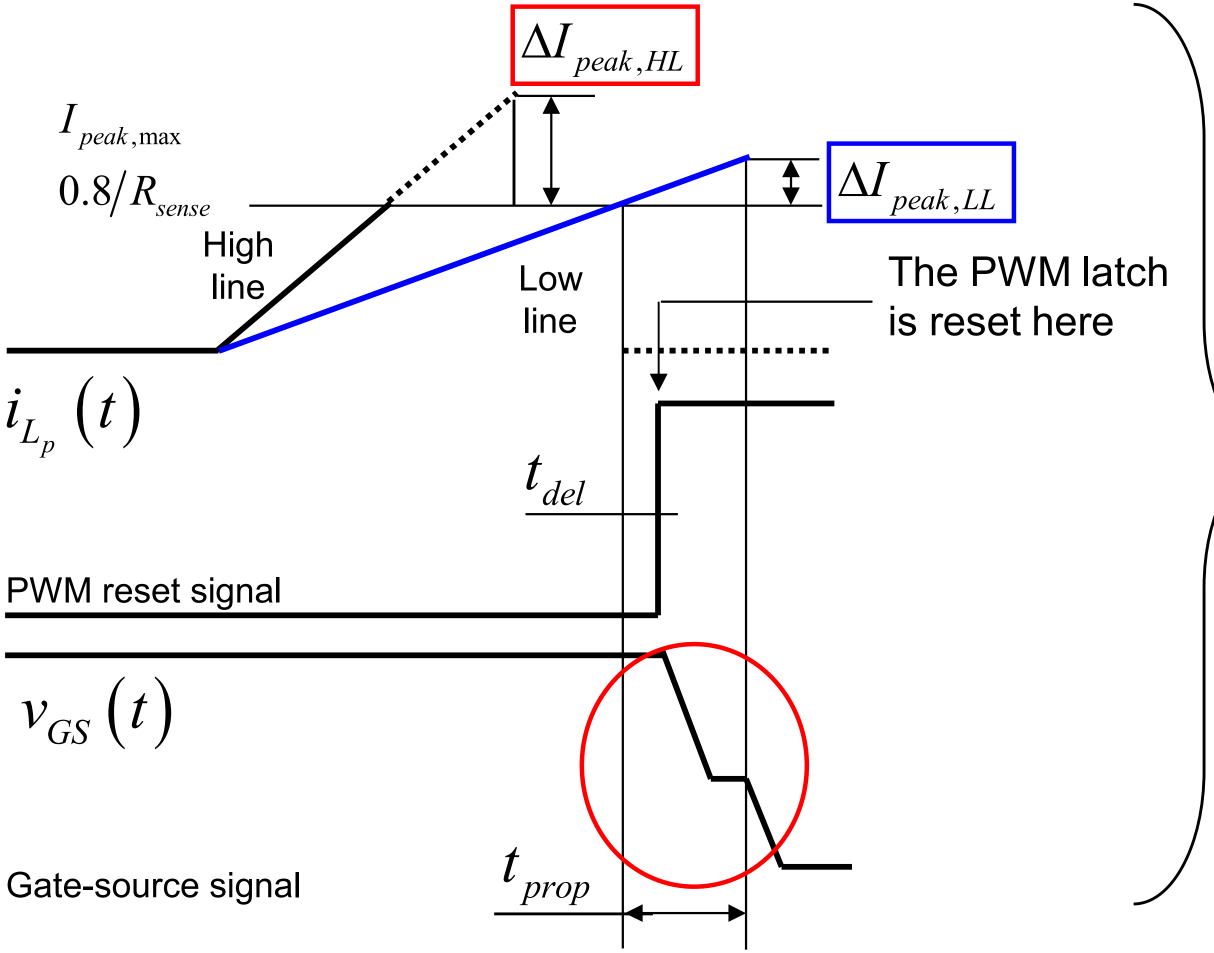
Over Power in a DCM Converter

- What maximum peak current at different input voltages?

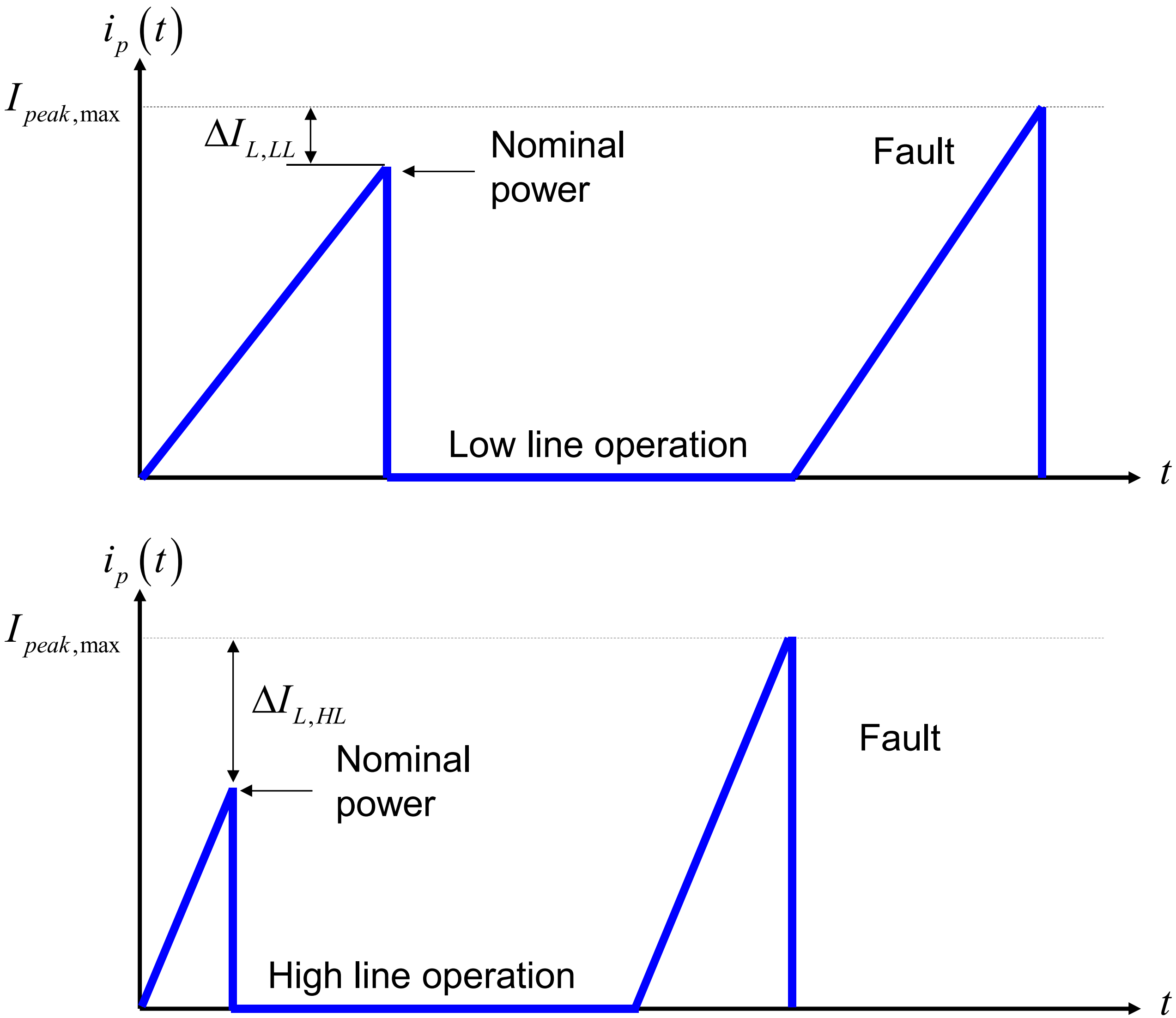


$$I_{peak} = I_{peak,\max} + \frac{V_{in,\max}}{L_p} t_{prop}$$

Watch saturation!

Power Runaway is Worse in QR

- ❑ As frequency is high, nominal peak at low line is small



Low line: small headroom before protection trips

High line: large Headroom, current can grow further

Quantifying Power Excursion

- Let us assume the following components values

$$R_{sense} = 0.2\,\Omega \longrightarrow I_{p,\max} = 0.8/0.2 = 4\,\text{A}$$

$$L_p = 350\,\mu\text{H} \qquad N = 0.25 \qquad t_{prop} = 350\,\text{ns} \qquad \text{DT} = 831\,\text{ns}$$

- The operating frequency with these elements is

$$F_{sw} = \frac{1}{I_{L,\max} L_p \left(\frac{1}{V_{in}} + \frac{N}{V_{out} + V_f} \right) + \text{DT}}$$

$$F_{sw} = 32\,\text{kHz low line}$$

$$F_{sw} = 40.7\,\text{kHz high line}$$

- Applying the DCM output power formula, we calculate

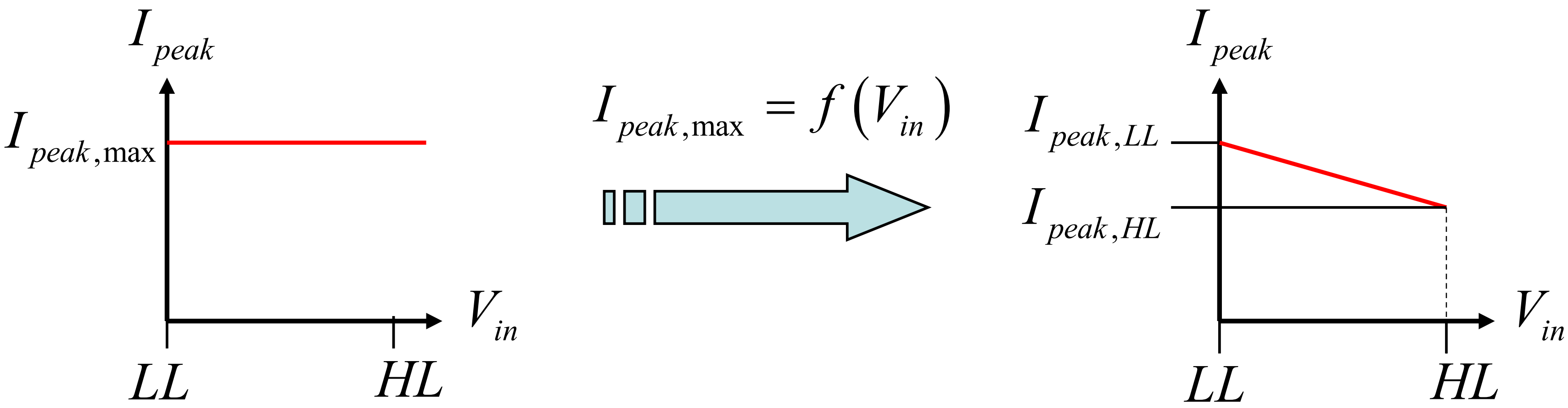
$$P_{out} = \frac{1}{2} F_{sw} L_p I_{L,peak}{}^2 \eta \Rightarrow$$

$$P_{out} = 80.6\,\text{W} \quad \text{low line}$$

$$P_{out} = 121\,\text{W} \quad \text{high line} \longrightarrow$$

Harnessing the Maximum Output Power

- ❑ There are easy options to pursue:
- ➢ we reduce the maximum peak current at high line



- ❑ How to calculate the compensated hi-line current?
- ❖ Equate lo-line power with hi-line power and solve for I_{peak}

$$P_{out,\max,LL} = \frac{1}{2} L_p I_{peak,\max,HL}{}^2 F_{sw} \eta_{HL}$$

Solve for I_{peak}

With a QR Converter Frequency Varies

- ❑ Fix the power level you can accept at high line: 80 W
- ❑ Calculate the switching frequency for 80 W high line

$$F_{sw,HL}=\frac{4}{\left(\sqrt{4\times 831\text{n}+\frac{2\times 350u\times 80\times(0.5+19+0.25\times 370)^2}{0.89\times 370^2(19+0.5)^2}}+\frac{\sqrt{2}\times 350u\times(0.5+19+0.25\times 370)\sqrt{\frac{80}{0.89\times 350u}}}{370\times(19+0.5)}\right)^2}=59.2\,\text{kHz}$$

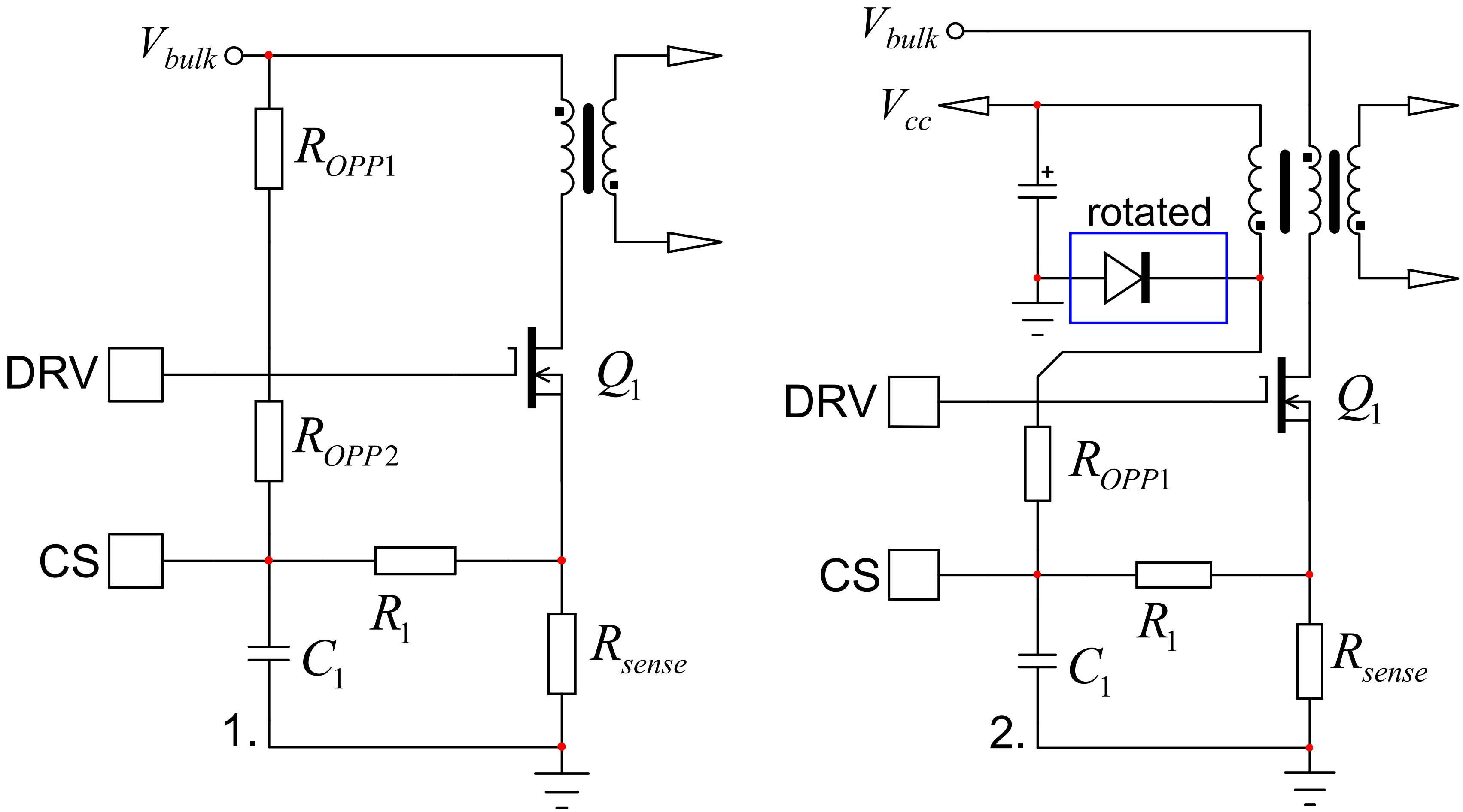
- ❑ Calculate the corresponding peak current at this frequency

$$I_{L,\max,CMP}=\sqrt{\frac{2P_{out}}{\eta L_p F_{sw}}}=\sqrt{\frac{2\times 80}{0.89\times 350u\times 59.2k}}=2.95\,\text{A}$$

- ❑ Question, how to reduce the peak setpoint versus line?

Offsetting the Current Sense

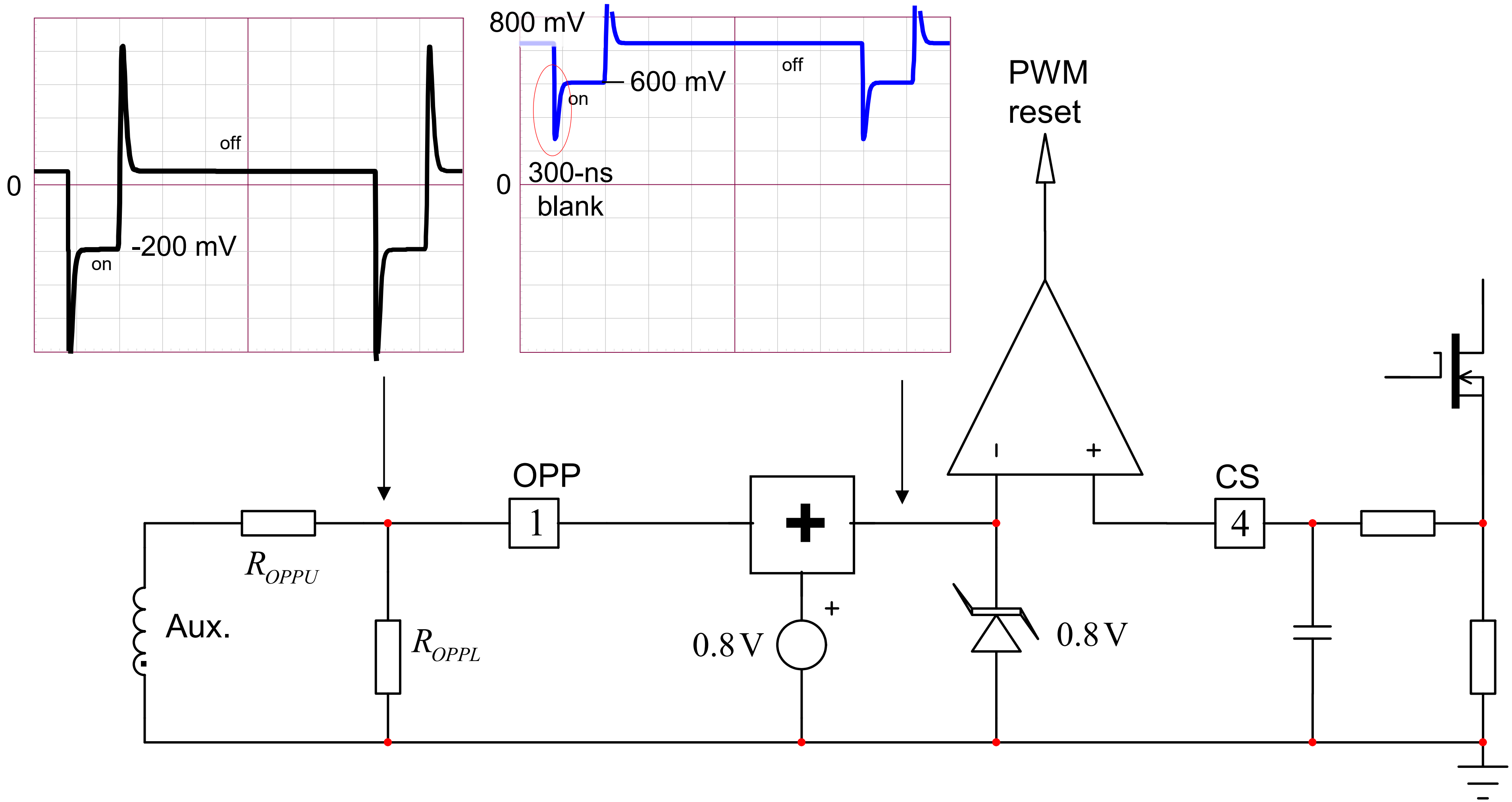
- ❑ This offset must be proportional to the input voltage



- ❑ Offset degrades light-load operation
- ❑ Radiated EMI can be affected in solution 2

ON Semiconductor Proprietary OPP Scheme

- ❑ This is a non-dissipative OPP circuitry



- ❑ The aux. swings to $-V_{in}$ and reduces setpoint → OPP

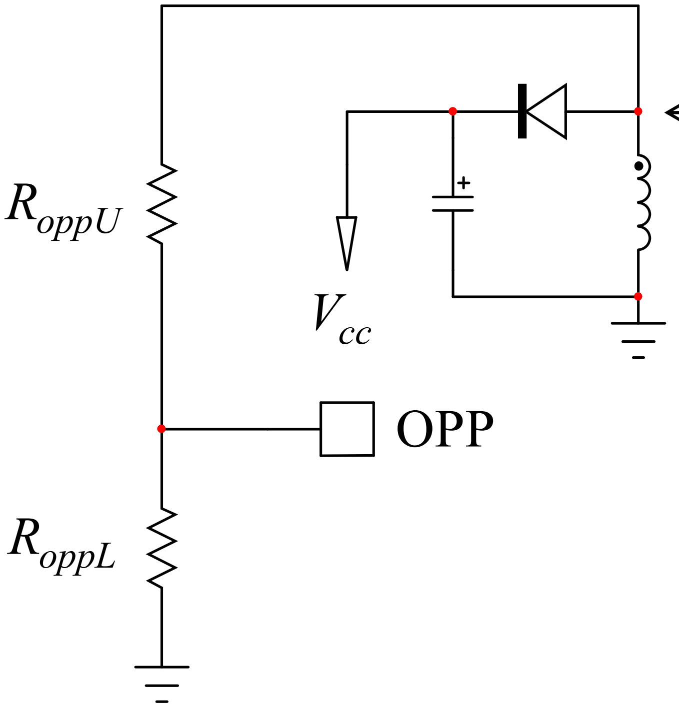
Peak Current Reduction is Easy Then

- We need 2.95 A from previous calculation

$$V_{sense,\max,CMP} = R_{sense}\left(I_{L,\max,CMP} - \frac{V_{in,HL}}{L_p} t_{prop}\right) = 0.2 \times \left(2.95 - \frac{370}{350u} \times 350n\right) = 516\,\text{mV}$$

Accounts for prop. delay

- The negative contribution must simply be 284 mV



Swings to -60 V during on-time hi line

1. Fix R_{oppL} to 1 kΩ
2. Calculate current: $I_{R_{oppL}} = \frac{284m}{1k} = 284\,\mu\text{A}$
3. Calculate R_{oppU}: $R_{oppU} = \frac{60 - 284m}{284u} = 210\,\text{k}\Omega$
4. You are done!

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A Controller With 3 Operating Modes

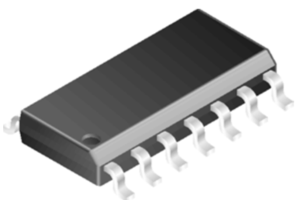
- Modern controllers need to cope with specifications on

✓ acoustic noise: vibrations in audible range must be avoided
✓ efficiency: full-load efficiency but also light-load efficiency
✓ standby power: in no-load conditions, absorbed power is close to 0 W

- Need to find solutions to fulfill all these requirements

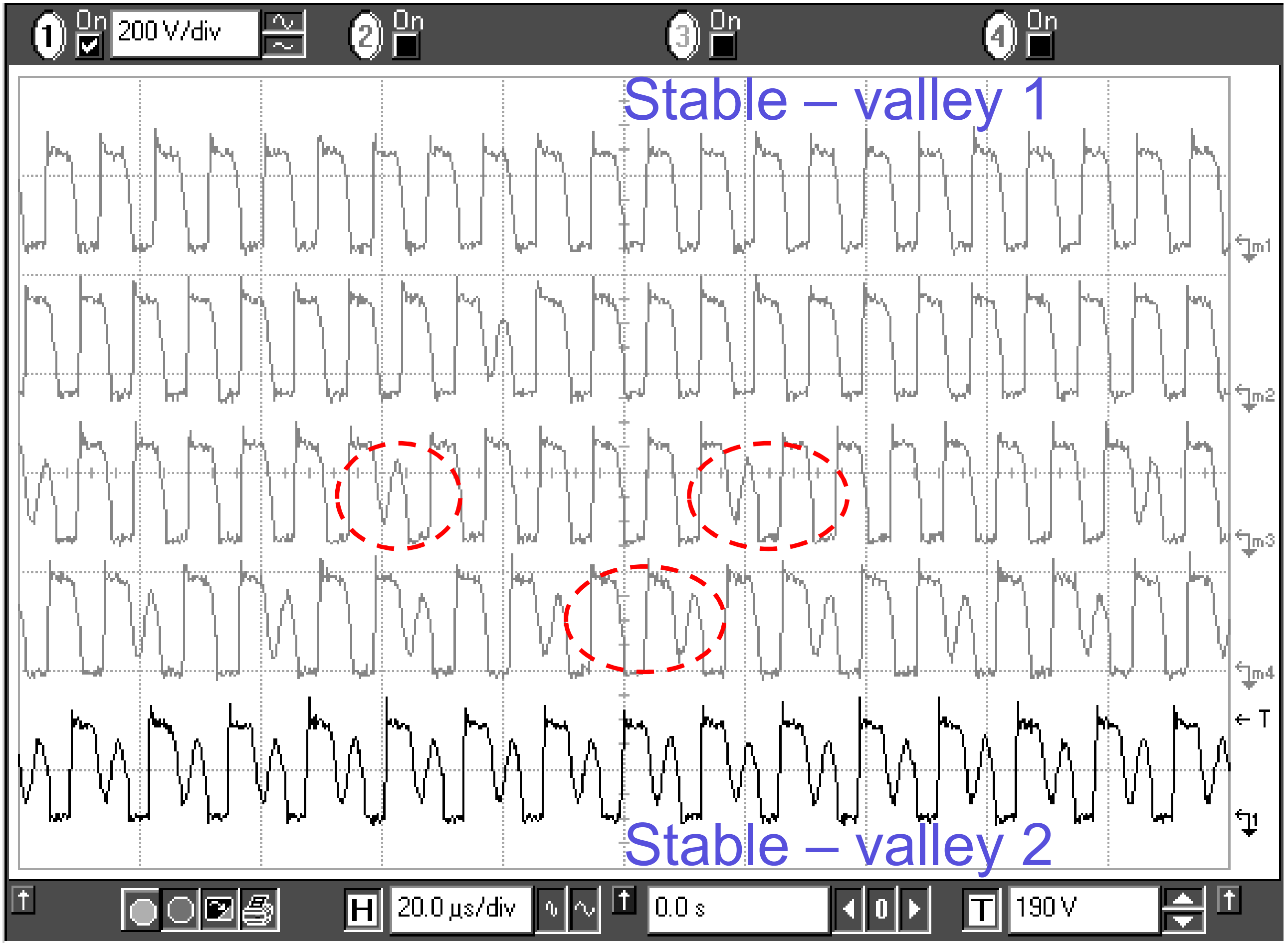
1. Valley lockout elegantly solves this issue by going down the valleys
2. Voltage-Controlled Oscillator folds the frequency back in light load
3. Skip cycle helps reduce consumption in no-load standby

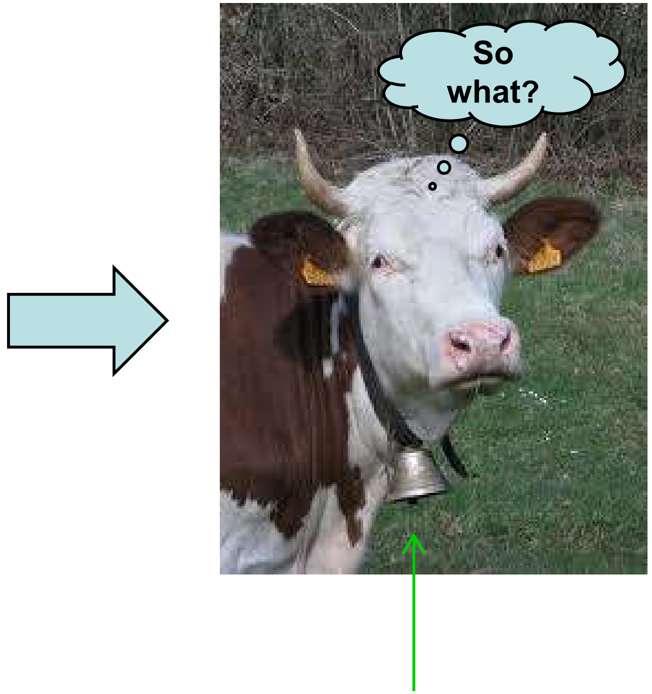
→ NCP1336/39 operate this way



Valley Jumping and Lockout

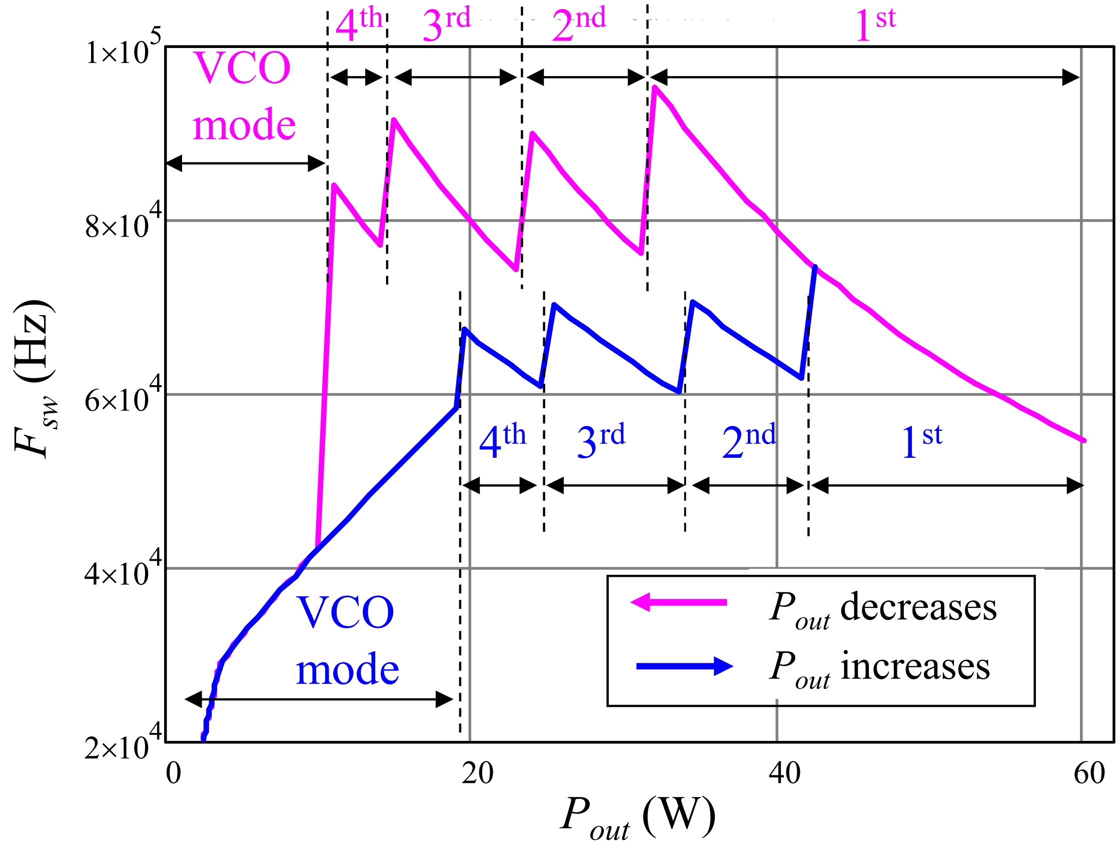
❑ Jumps between valleys generate instability and potential noise





New Controllers Lock in the Valleys

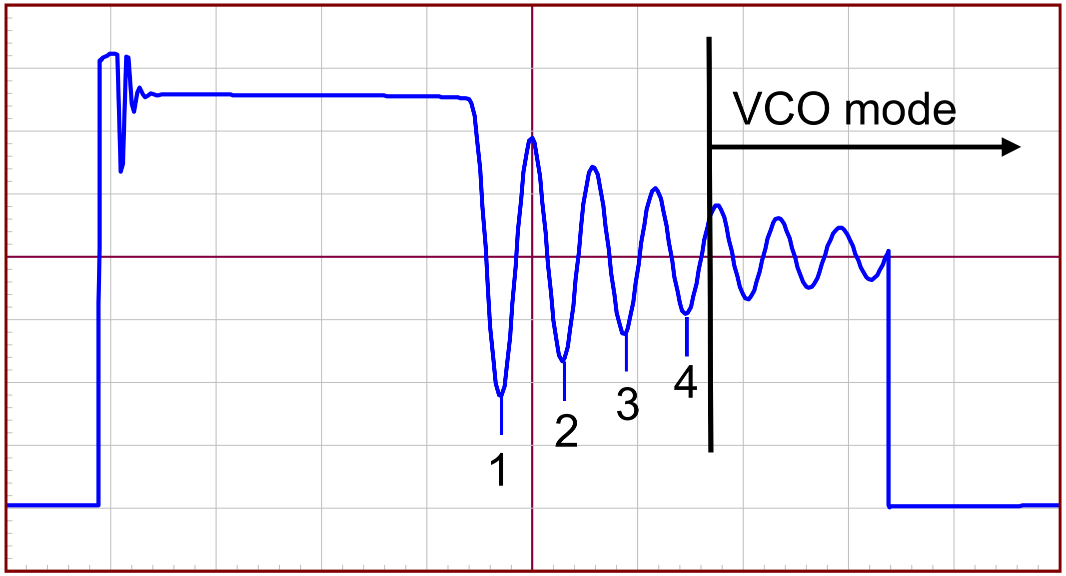
- To prevent the noise, the NCP1336/39 lock the valley
- The current is allowed to move within a certain limit
- When it exceeds this limit, the controller selects a new valley
- As the load gets lighter, a VCO takes over and reduce F_{sw}



NCP1336/1339

Three Operating Modes

1. In Quasi-Resonant mode, controlled variable is I_{peak}
- ❖ Switching frequency changes with valley position
- ❖ Transfer function can be affected by different operating points



2. In VCO mode, the controlled variable is F_{sw}, I_{peak} is frozen

3. In skip cycle, frequency and peak current are frozen: hysteretic

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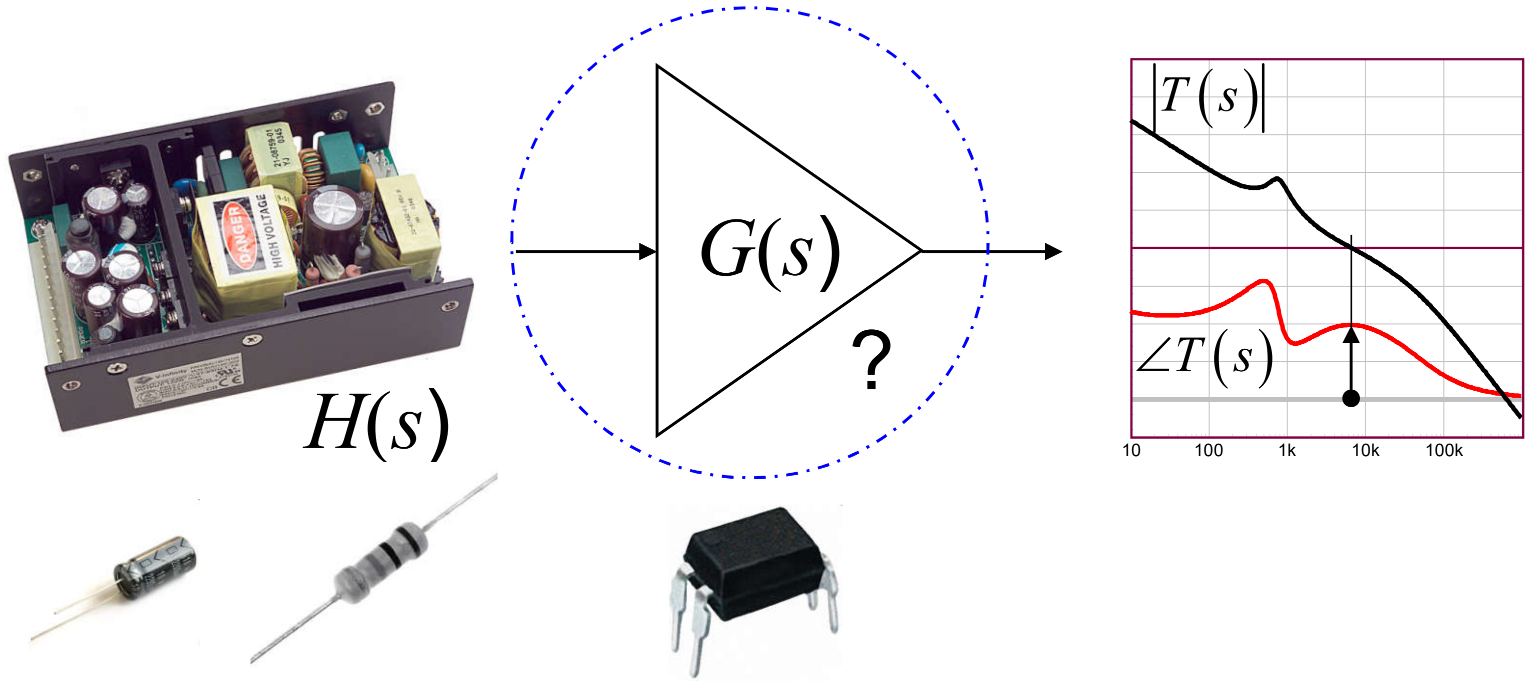
Small-Signal Analysis

- ❑ Loop instability is a common issue in production
- ❑ Due to time pressure, designers often use trial and error
  - ➢ no indication on design margins
  - ➢ offenders are ignored, robustness is at stake



❖ Understand and counteract their variations when building $G(s)$

There are Two Options

❑ Analytical analysis of the power stage:

✓ best to see where the offenders are hidden (ESR, opto pole etc.)

✓ equations are complex but litterature abounds

❖ transfer function are for DCM or CCM

❖ difficult to predict transient response

❑ SPICE models:

✓ easy-to-implement averaged models

✓ can work in ac or transient mode

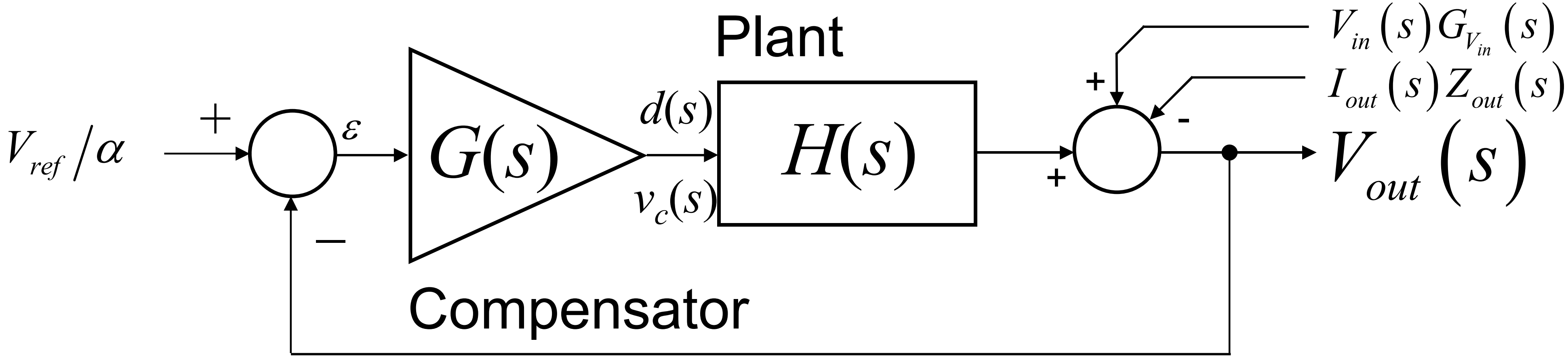
✓ easily transition between CCM and DCM

❖ do not explictly disclose the position of poles and zeros

A measurement on the bench is mandatory, whatever you choose!

Analytical Analysis

- ❑ You must first characterize the "plant" transfer function
  - ➢ what are your power stage ac characteristics?



$$H(s) = \frac{V_{out}(s)}{v_c(s)}$$

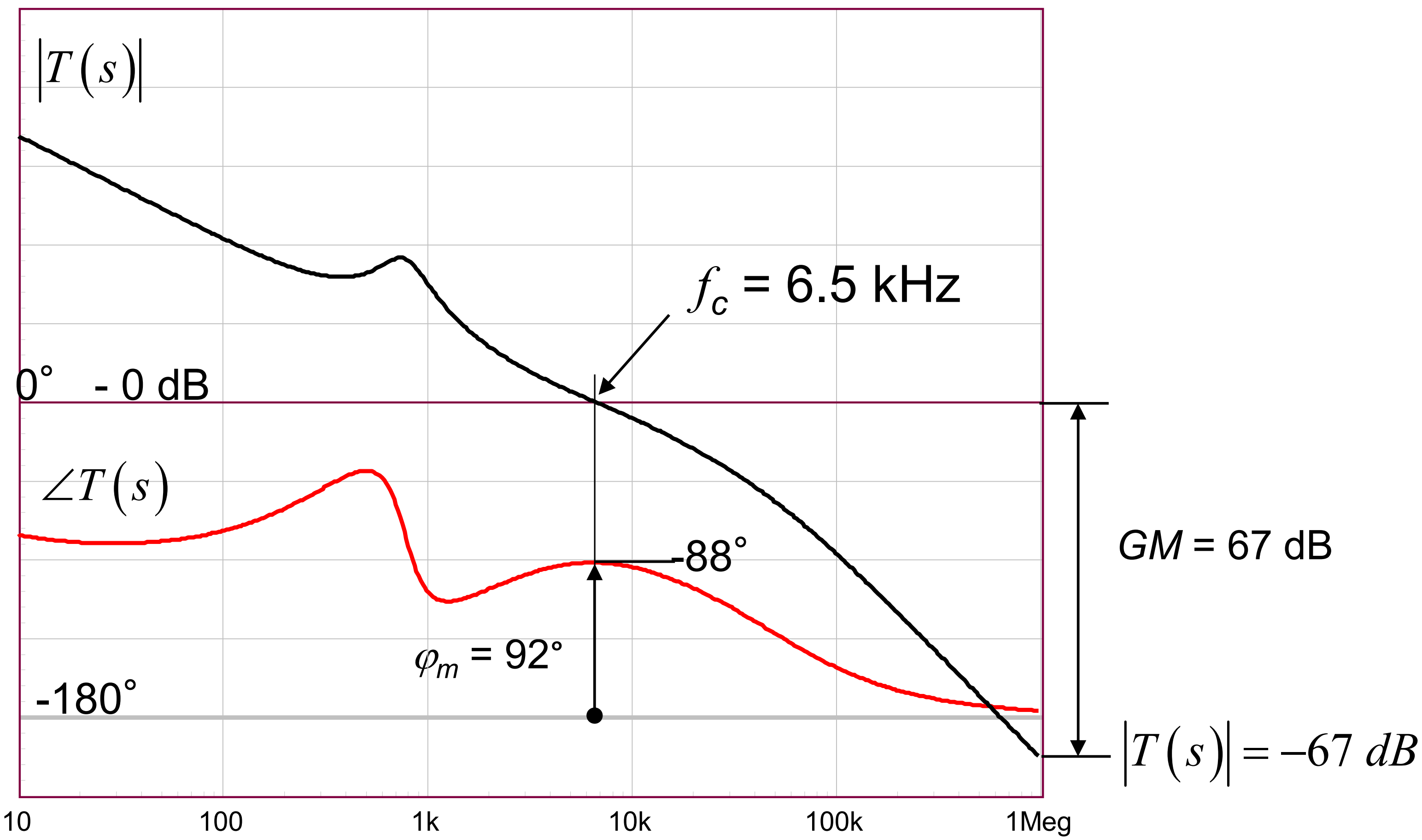
Current-mode control

$$H(s) = \frac{V_{out}(s)}{d(s)}$$

Voltage-mode control

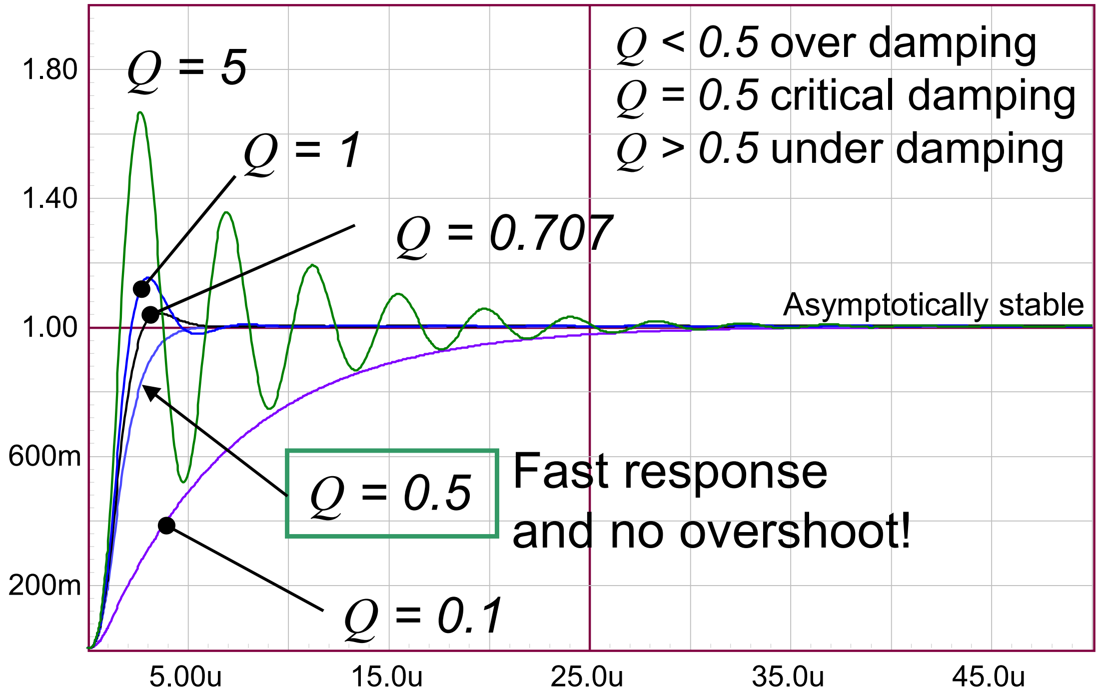
How do we Stabilize a Converter?

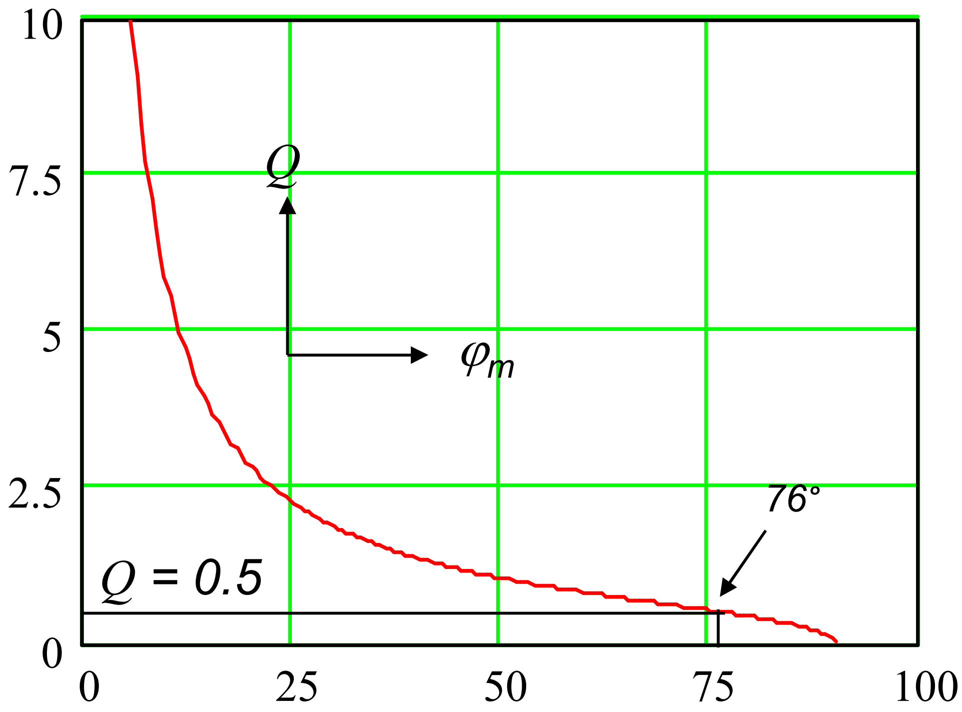
- ❑ We need a high gain at dc for a low static error
- ❑ We want a sufficiently high crossover frequency for response speed
- ➢ Shape the compensator G(s) to build phase and gain margins!



How much phase margin to chose?

- 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!

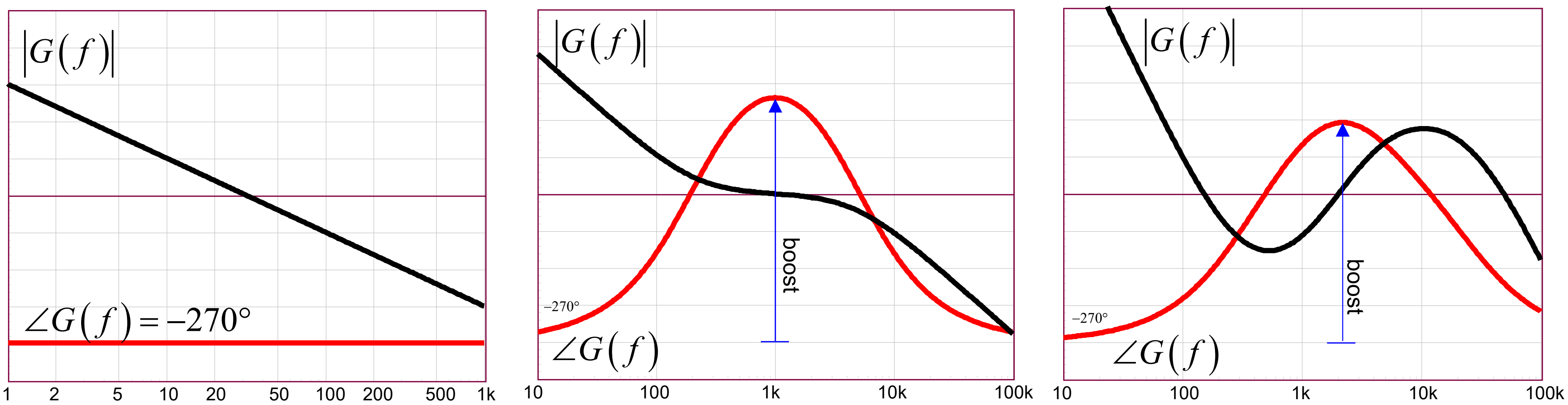




- phase margin depends on the needed response: fast, no overshoot…
- good practice is to shoot for 60° and make sure φ_m always > 45°

What Compensator Types do we Need?

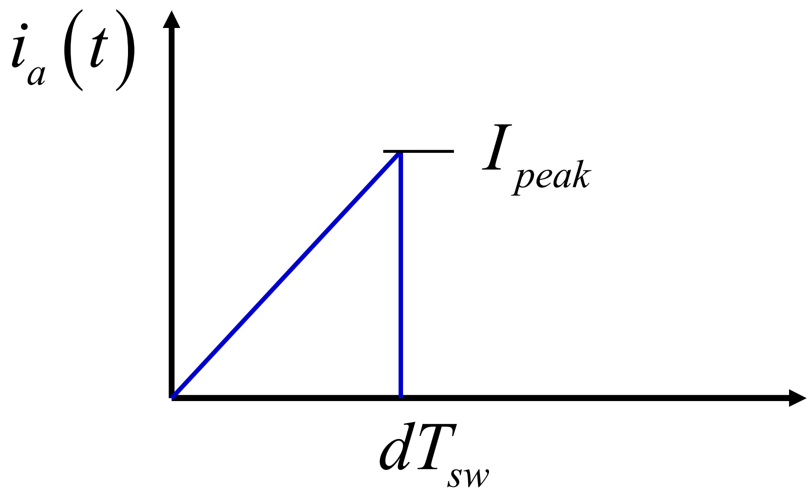
- ❑ There are basically 3 compensator types:
- ➢ type 1, 1 pole at the origin, no phase boost
- ➢ type 2, 1 pole at the origin, 1 zero, 1 pole. Phase boost up to 90°
- ➢ type 3, 1 pole at the origin, 1 zero pair, 1 pole pair. Boost up to 180°

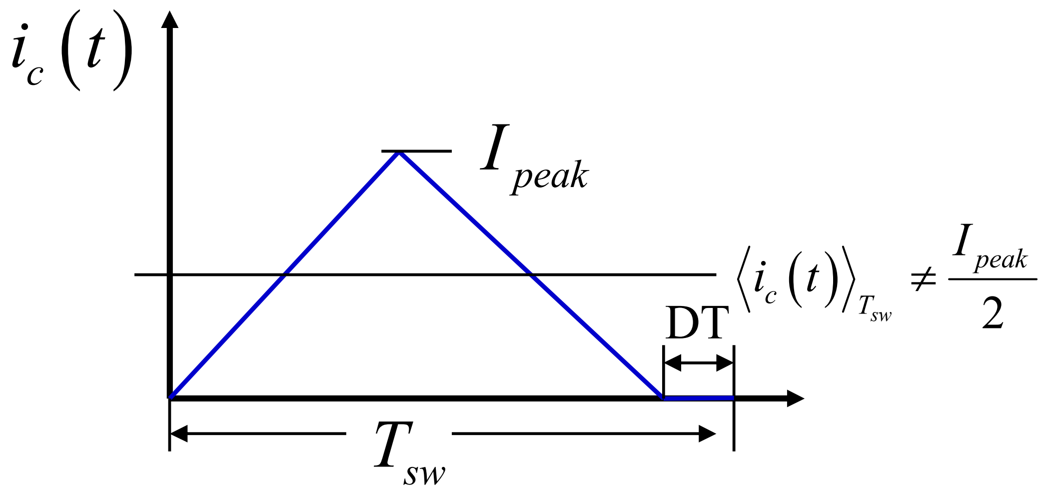


Type 1 Type 2 Type 3

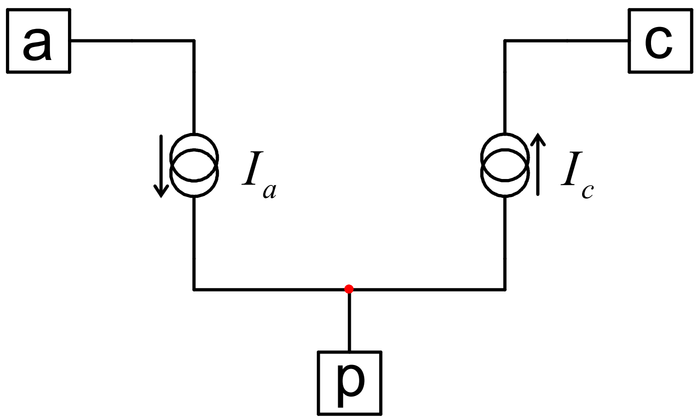
Variable-Frequency Current-Mode

- ❑ Observing the waveforms helps us to derive an average model





- ❑ It leads to a large-signal model



Equations Account for the DeadTime

- Former equations no longer work and need to include DT

$$d_1 = \frac{V_c}{R_i}\frac{L}{V_{ac}T_{sw}}$$

V_{in} ↑

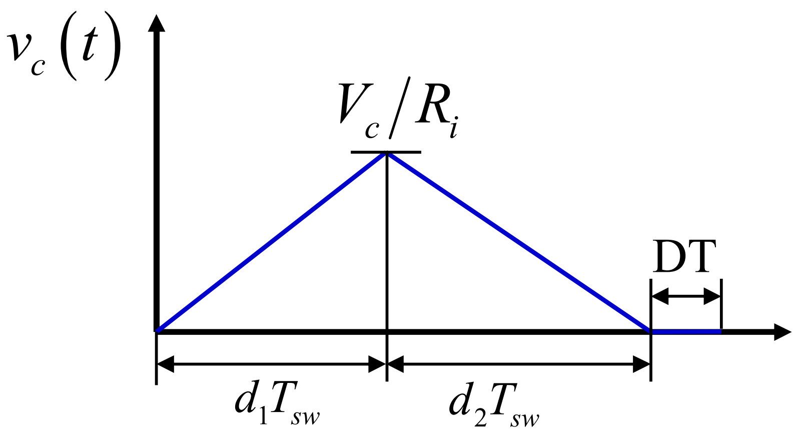
$$d_2 = \frac{V_c}{R_i}\frac{L}{V_{cp}T_{sw}}$$

V_{out}/N ↑

$$t_{on} = \frac{V_c}{R_i}\frac{L}{V_{ac}} \qquad t_{off} = \frac{V_c}{R_i}\frac{L}{V_{cp}}$$

$$\Rightarrow \quad T_{sw} = \frac{V_c L}{R_i}\left(\frac{1}{V_{ac}} + \frac{1}{V_{cp}}\right) + \mathrm{DT}$$

- The average current in terminal c now depends on d_1 and d_2



$$I_c = \frac{V_c}{R_i}\left(\frac{d_1 + d_2}{2}\right) = \left(\frac{V_c}{R_i}\right)^2 \frac{L\left(V_{ac} + V_{cp}\right)}{2T_{sw}V_{ac}V_{cp}}$$

Define Terminals a and c Currents

- Substitute d_1 and d_2 in I_c definition and rearrange

$$I_c = \frac{V_c}{2R_i} \frac{V_{ac} + V_{cp}}{V_{ac} + V_{cp} + \frac{\mathrm{DT} \cdot R_i V_{ac} V_{cp}}{L V_c}}$$

- Same method for I_a, express I_{peak}, substitute, rearrange

$$I_a = \frac{I_{peak}}{2} d_1 = I_c \frac{d_1}{d_1 + d_2} = I_a = I_c \frac{V_{cp}}{V_{ac} + V_{cp}}$$

$$\Rightarrow \quad I_a = \frac{V_c}{2R_i \left(\frac{V_{ac}}{V_{cp}} + 1 + \mathrm{DT} \frac{R_i}{L} \frac{V_{ac}}{V_c} \right)}$$

Large-Signal to Small-Signal

- We have large-signal equations that we must linearize

$$I_c = f\left(V_c, V_{ac}, V_{cp}\right) \qquad I_a = f\left(V_c, V_{ac}, V_{cp}\right)$$

- Apply partial differentiation to I_a and I_c

$$\hat{i}_c = \frac{\partial I_c\left(V_c, V_{ac}, V_{cp}\right)}{\partial V_c}\hat{v}_c + \frac{\partial I_c\left(V_c, V_{ac}, V_{cp}\right)}{\partial V_{ac}}\hat{v}_{ac} + \frac{\partial I_c\left(V_c, V_{ac}, V_{cp}\right)}{\partial V_{cp}}\hat{v}_{cp}$$

$$\hat{i}_c = k_1\hat{v}_c + k_2\hat{v}_{ac} + k_3\hat{v}_{cp}$$

$$\hat{i}_a = \frac{\partial I_a\left(V_c, V_{ac}, V_{cp}\right)}{\partial V_c}\hat{v}_c + \frac{\partial I_a\left(V_c, V_{ac}, V_{cp}\right)}{\partial V_{ac}}\hat{v}_{ac} + \frac{\partial I_c\left(V_c, V_{ac}, V_{cp}\right)}{\partial V_{cp}}\hat{v}_{cp}$$

$$\hat{i}_a = k_4\hat{v}_c + k_5\hat{v}_{ac} + k_6\hat{v}_{cp}$$

Small-Signal Expressions

- Calculate and collect all individual coefficients

$$k_1 = \frac{L_p V_c \left(V_{ac} + V_{cp}\right)\left(L_p V_c V_{ac} + L_p V_c V_{cp} + 2 \cdot \mathrm{DT} \cdot R_i V_{ac} V_{cp}\right)}{2R_i \left(L_p V_c V_{ac} + L_p V_c V_{cp} + \mathrm{DT} \cdot R_i V_{ac} V_{cp}\right)^2} \qquad k_2 = -\frac{\mathrm{DT} \cdot L_p V_c^{\,2} V_{cp}^{\,2}}{2\left(L_p V_c V_{ac} + L_p V_c V_{cp} + \mathrm{DT} \cdot R_i V_{ac} V_{cp}\right)^2}$$

$$k_3 = -\frac{\mathrm{DT} \cdot L_p V_c^{\,2} V_{ac}^{\,2}}{2\left(L_p V_c V_{ac} + L_p V_c V_{cp} + \mathrm{DT} \cdot R_i V_{ac} V_{cp}\right)^2}$$

Terms for I_c

$$k_4 = \frac{L_p V_c V_{cp} \left(L_p V_c V_{ac} + L_p V_c V_{cp} + 2 \cdot \mathrm{DT} \cdot R_i V_{ac} V_{cp}\right)}{2R_i \left(L_p V_c V_{ac} + L_p V_c V_{cp} + \mathrm{DT} \cdot R_i V_{ac} V_{cp}\right)^2} \qquad k_5 = -\frac{V_c \left(\dfrac{1}{V_{cp}} + \dfrac{\mathrm{DT} \cdot R_i}{L_p V_c}\right)}{2R_i \left(\dfrac{V_{ac}}{V_{cp}} + \dfrac{\mathrm{DT} \cdot R_i V_{ac}}{L_p V_c} + 1\right)^2}$$

$$k_6 = \frac{V_c V_{ac}}{2R_i V_{cp}^{\,2} \left(\dfrac{V_{ac}}{V_{cp}} + \dfrac{\mathrm{DT} \cdot R_i V_{ac}}{L_p V_c} + 1\right)^2}$$

Terms for I_a

A Small-Signal PWM Switch Model in QR

- ❑ Associate current sources together to form the small-signal model

parameters

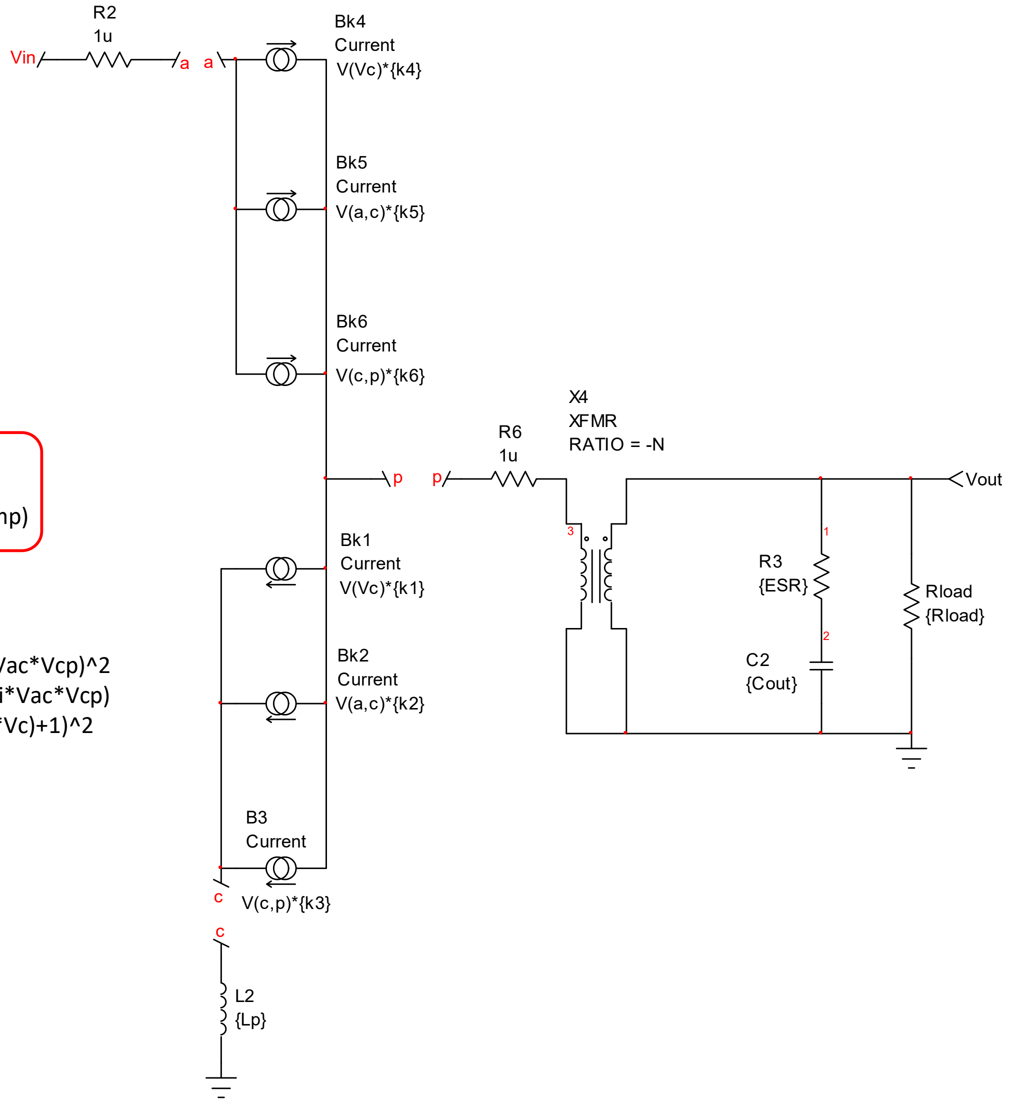
Vout=12
Vin=100
Pout=70
Lp=450u
Ri=0.25
N=1/7.5
Cout=1500u
ESR=50m
Rload=Vout^2/Pout
Vc=947m

Deadtime calculations

Clump=200p
nv=6
DT=(2*nv-1)*3.14159*sqrt(Lp*Clump)

Vac=Vin
Vcp=Vout/N

A=(Lp*Vc*Vac+Lp*Vc*Vcp+DT*Ri*Vac*Vcp)^2
B=(Lp*Vc*Vac+Lp*Vc*Vcp+2*DT*Ri*Vac*Vcp)
C=2*Ri*((Vac/Vcp)+DT*Ri*Vac/(Lp*Vc)+1)^2
k1=Lp*Vc*(Vac+Vcp)*B/(2*Ri*A)
k2=-DT*Lp*Vc^2*Vcp^2/(2*A)
k3=-DT*Lp*Vc^2*Vac^2/(2*A)
k4=Lp*Vc*Vcp*B/(2*Ri*A)
k5=-Vc*((1/Vcp)+DT*Ri/(Lp*Vc))/C
k6=Vc*Vac/(Vcp^2*C)



Build a SPICE Model

- Modify the original QR model to account for valley switching

```
.subckt PWMQR a c p vc ton fsw params : L=1.2m Ri=0.5 DT=1
*
* This subckt is a current-mode BCM model, version 1
*
.subckt limit d dc params: clampH=0.99 clampL=16m
*
Gd 0 dcx d 0 100u
Rdc dcx 0 10k
V1 clpn 0 {clampL}
V2 clpp 0 {clampH}
D1 clpn dcx dclamp
D2 dcx clpp dclamp
Bdc dc 0 V=V(dcx)
.model dclamp d n=0.01 rs=100m
.ENDS
*
Btsw tsw 0 V= ( (V(vc)*{L}/{Ri}) * ( 1/v(a,cx)  + 1/v(cx,p) )+{DT}) *1Meg
Bdc dcx 0 V=V(vc)*{L}/({Ri}*V(a,cx)*(V(tsw)/1Meg))
Xdc dcx dc limit params: clampH=0.99 clampL=7m
Blap a p I=I(VM)*V(dc)/(V(dc)+V(d2))
Bd2 d2 0 V=V(vc)*{L}/({Ri}*V(c,p)*(V(tsw)/1Meg))
Blpc p cx I=V(vc)/{Ri}
Blmju cx p I=(v(cx,p)/{L})*V(d2)*(V(tsw)/1Meg)*(1-(V(dc)+V(d2))/2)
Bton ton 0 V = V(dc)*v(tsw)
Bfsw fsw 0 V = (1/((V(tsw)/1Meg)))/1k
Rdum1 vc 0 1Meg
VM cx c
*
.ENDS
```

DT changes in relationship to valley selection

Run a Simple Test Circuit

- ❑ Before going through the analysis, validate the model!

parameters

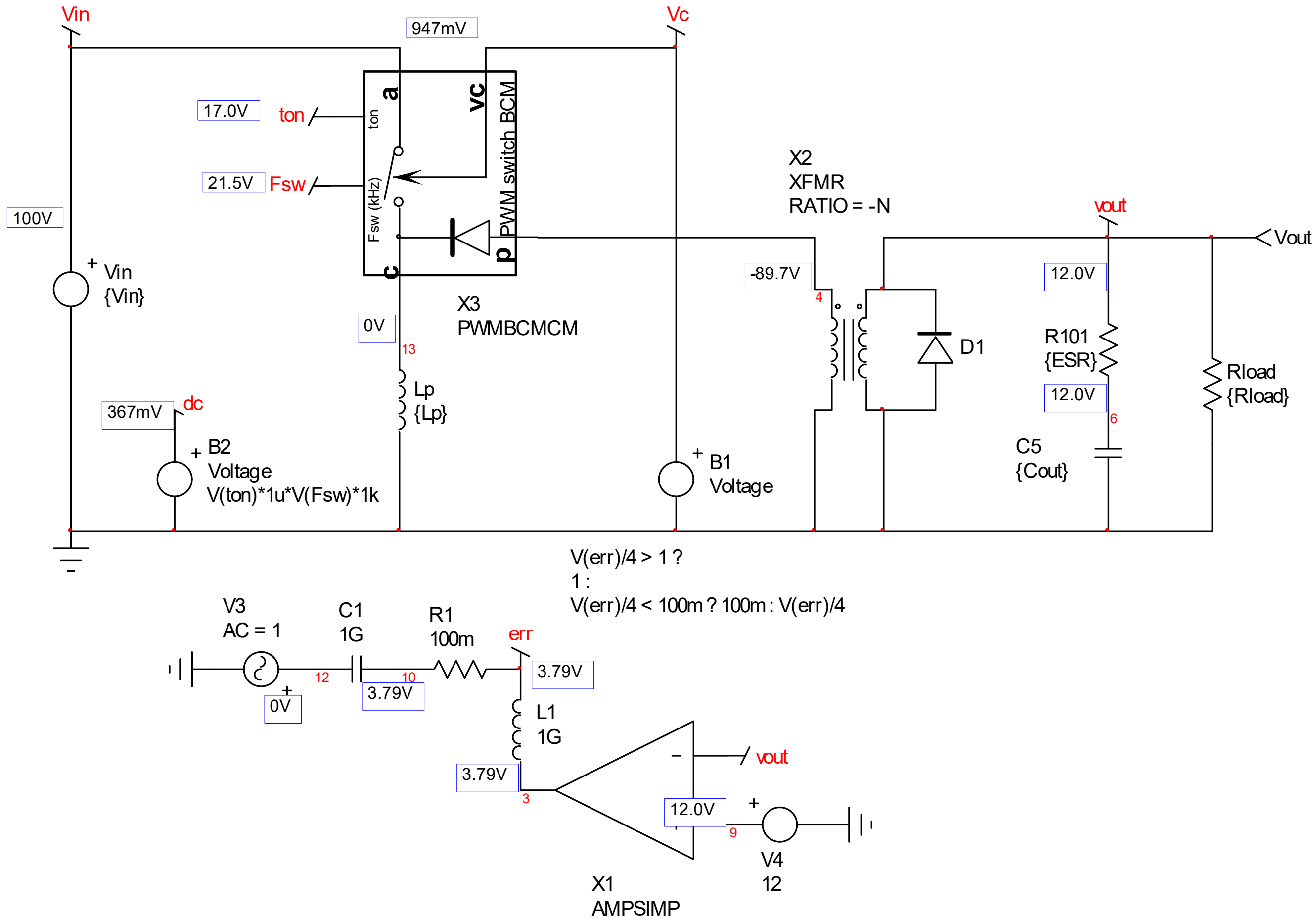
QR Model with valley switching

Vout=12
Vin=100
Pout=70
Lp=450u
Ri=0.25
N=1/7.5
Cout=1500u
ESR=50m
Rload=Vout^2/Pout
Vc=947m

Clump=200p
nv=6
DT=(2*nv-1)*3.14159*sqrt(Lp*Clump)

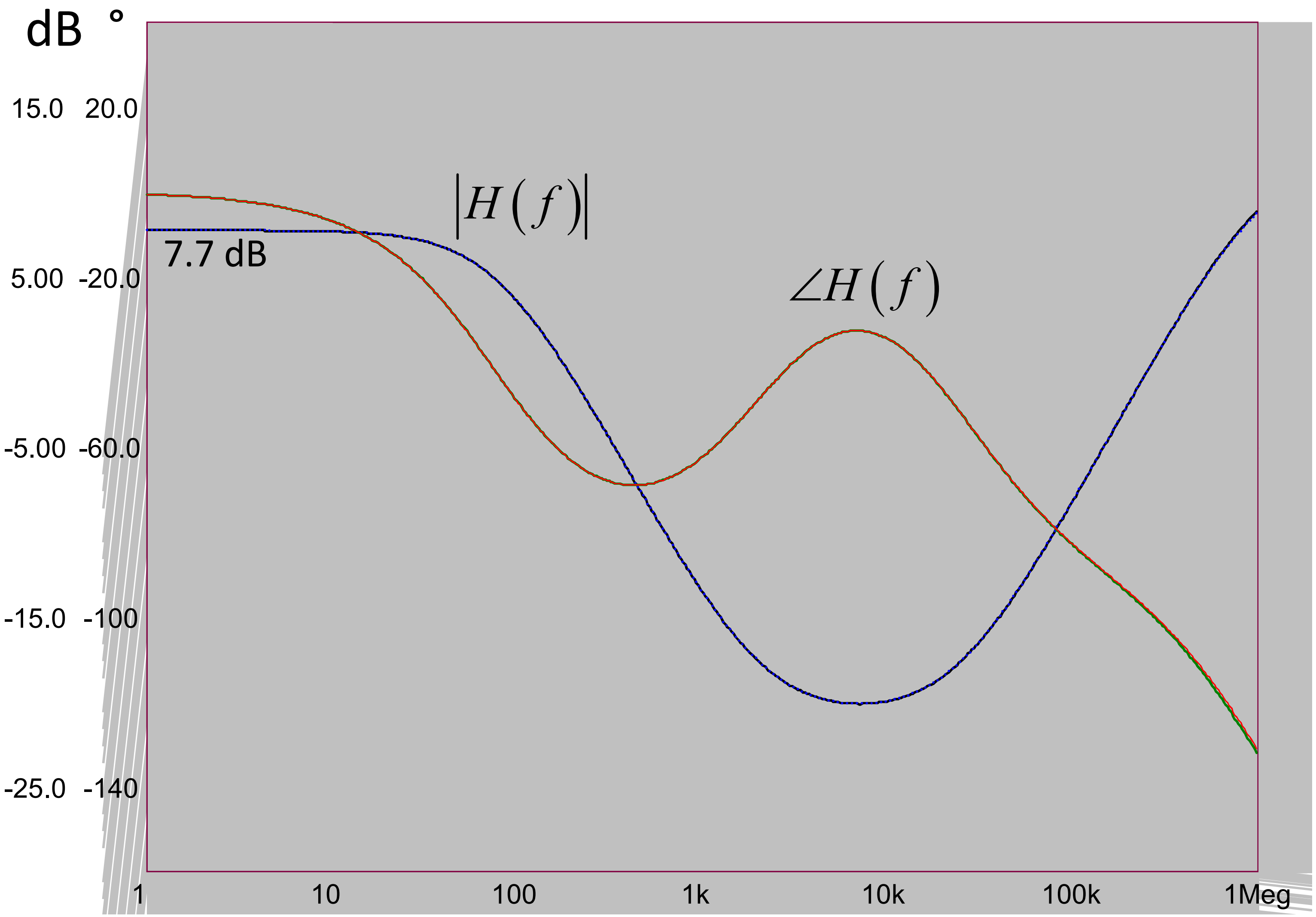
Vac=Vin
Vcp=Vout/N

A=(Lp*Vc*Vac+Lp*Vc*Vcp+DT*Ri*Vac*Vcp)^2
B=(Lp*Vc*Vac+Lp*Vc*Vcp+2*DT*Ri*Vac*Vcp)
C=2*Ri*((Vac/Vcp)+DT*Ri*Vac/(Lp*Vc)+1)^2
k1=Lp*Vc*(Vac+Vcp)*B/(2*Ri*A)
k2=-DT*Lp*Vc^2*Vcp^2/(2*A)
k3=-DT*Lp*Vc^2*Vac^2/(2*A)
k4=Lp*Vc*Vcp*B/(2*Ri*A)
k5=-Vc*((1/Vcp)+DT*Ri/(Lp*Vc))/C
k6=Vc*Vac/(Vcp^2*C)



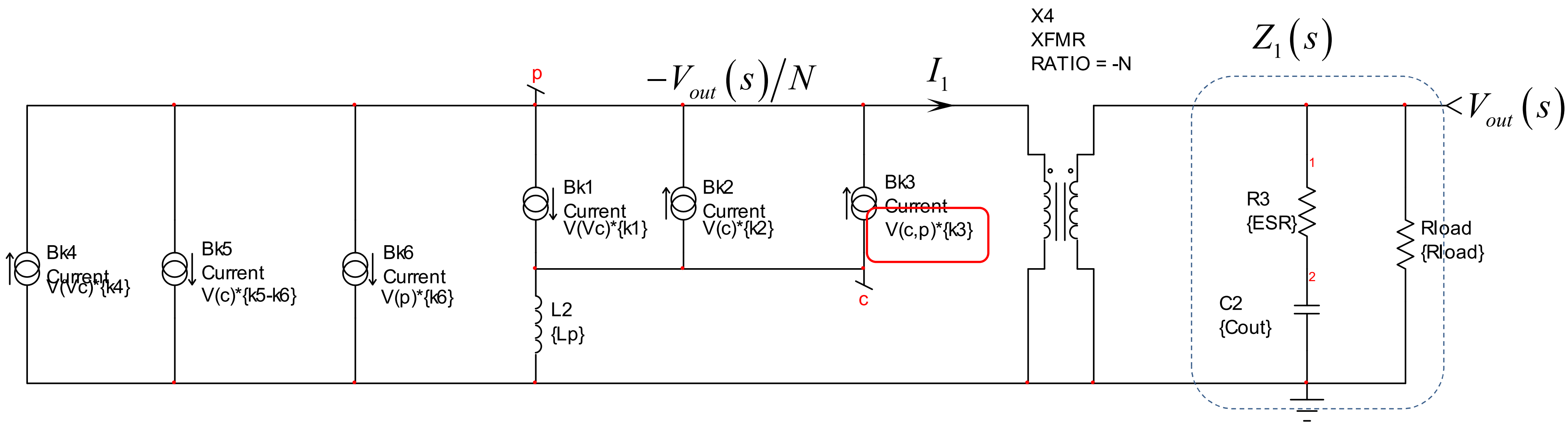
Compare Analytic to Simulated Curves

- ❑ Curves have to be identical before proceeding



Simplify, Simplify and Simplify!

- The key to solving transfer functions is to simplify networks
- Here, the input voltage is not modulated, hence $V_{(a)}(s) = 0$



$$V_{(c)}(s) = I_c(s) s L_p B_{k_3} = V(c,p) k_3 = V_{(c)} k_3 - V_{(p)} k_3 = k_3 \left(V_{(c)} + \frac{V_{out}}{N} \right)$$

$$V_{(c)}(s) = I_c(s) s L_p$$

$$I_c(s) = V_c(s) k_1 - V_{(c)} k_2 + \left(V_{(c)} + \frac{V_{out}(s)}{N} \right) k_3$$

Solving the Transfer Function

- Substitute $V_{(c)}$ into I_c and solve for I_c

$$I_c(s) = \frac{V_{out}(s)k_3 + NV_c(s)k_1}{N\left(1 + sL_p k_2 - sL_p k_3\right)}$$

- The current I_1 entering the transformer is given by:

$$I_1(s) = V_c(s)k_4 - I_c(s)sL_p(k_5 - k_6) + \frac{V_{out}}{N}k_6 - I_c(s)$$

- A nice equation comes out:

$$I_1(s) = -\frac{V_{out}(s)(k_3 - k_6) + NV_c(s)(k_1 - k_{4)}) + sL_p V_{out}(s)(k_3 k_5 - k_2 k_6) + sV_c(s)NL_p(k_5 k_1 - k_2 k_4 - k_1 k_6 + k_3 k_4)}{N\left[1 + sL_p(k_2 - k_3)\right]}$$

- This current enters a complex impedance $Z_1(s)$

$$Z_1(s) = \frac{\left(\frac{r_c}{N^2} + \frac{1}{sC_{out}N^2}\right)\frac{R_{load}}{N^2}}{\left(\frac{r_c}{N^2} + \frac{1}{sC_{out}N^2}\right) + \frac{R_{load}}{N^2}} = \frac{R_{load}}{N^2}\frac{1 + sr_C C_{out}}{1 + sC_{out}\left(R_{load} + r_C\right)}$$

Rearranging the Raw Expression

- Collecting $V_{out}(s)$ and $V_c(s)$, we have our transfer function

$$H(s)=\frac{V_{out}(s)}{V_c(s)}=\frac{NR_{load}}{\mathrm{Div}\cdot\left[N^2+R_{load}\left(k_6-k_3\right)\right]}\frac{\left(1+sr_C C_{out}\right)\left(k_1-k_4+sL_p\left(k_1k_5-k_2k_4-k_1k_6+k_3k_4\right)\right)}{1+a_1s+a_2s^2}$$

In which we have

$$a_1=\frac{C_{out}N^2R_{load}+C_{out}N^2r_C+L_pN^2k_2-L_pN^2k_3-C_{out}R_{load}k_3r_C+C_{out}R_{load}k_6r_C-L_pR_{load}k_2k_6-L_pR_{load}k_3k_5}{N^2+R_{load}\left(k_6-k_3\right)}$$

$$a_2=\frac{\mathrm{Div}\cdot C_{out}L_p\left(N^2R_{load}k_3-N^2R_{load}k_2-N^2k_2r_C+N^2k_3r_C+R_{load}r_Ck_2k_6+R_{load}r_Ck_3k_5\right)}{\mathrm{Div}\cdot\left[R_{load}\left(k_3-k_6\right)-N^2\right]}$$

- Rearranging the formula, we have

$$H(s)=\frac{NR_{load}\left(k_1-k_4\right)}{\mathrm{Div}\cdot\left[N^2+R_{load}\left(k_6-k_3\right)\right]}\frac{\left(1+sr_C C_{out}\right)\left(1+sL_p\frac{k_1k_5-k_2k_4-k_1k_6+k_3k_4}{k_1-k_4}\right)}{1+a_1s+a_2s^2}$$

The Low-Q Approximation

- In this expression, we have a quality factor Q: $Q = \frac{\sqrt{a_2}}{a_1}$
- The resonant angular frequency Q is: $\omega_0 = \frac{1}{\sqrt{a_2}}$

With a low Q
2nd order form simplifies

$$1+\frac{s}{\omega_o Q}+\left(\frac{s}{\omega_o}\right)^2 \approx \left(1+\frac{s}{\omega_{p_1}}\right)\left(1+\frac{s}{\omega_{p_2}}\right)$$

$$\longrightarrow \omega_{p_1} = \omega_0 Q$$

$$\longrightarrow \omega_{p_2} = \frac{\omega_0}{Q}$$

$$\omega_{p_1} = \frac{N^2 + R_{load}\left(k_6 - k_3\right)}{C_{out}N^2\left(R_{load}+r_C\right)+L_pN^2\left(k_2-k_3\right)-C_{out}R_{load}r_C\left(k_6-k_3\right)-L_pR_{load}\left(k_2k_6+k_3k_5\right)}$$

$$L_p \text{ and } r_C \ll 1 \longrightarrow \quad \omega_{p_1} \approx \frac{1}{C_{out}\dfrac{R_{load}+r_C}{1+\dfrac{R_{load}\left(k_6-k_3\right)}{N^2}}}$$

The Complete Expression

- The second pole is at high frequencies and can be neglected
- There is a RHP Zero defined as:

$$\omega_{z_2} = \frac{k_1 - k_4}{L_p\left(k_1 k_5 - k_2 k_4 - k_1 k_6 + k_3 k_4\right)}$$

- The dc gain depends on the load and other coefficients

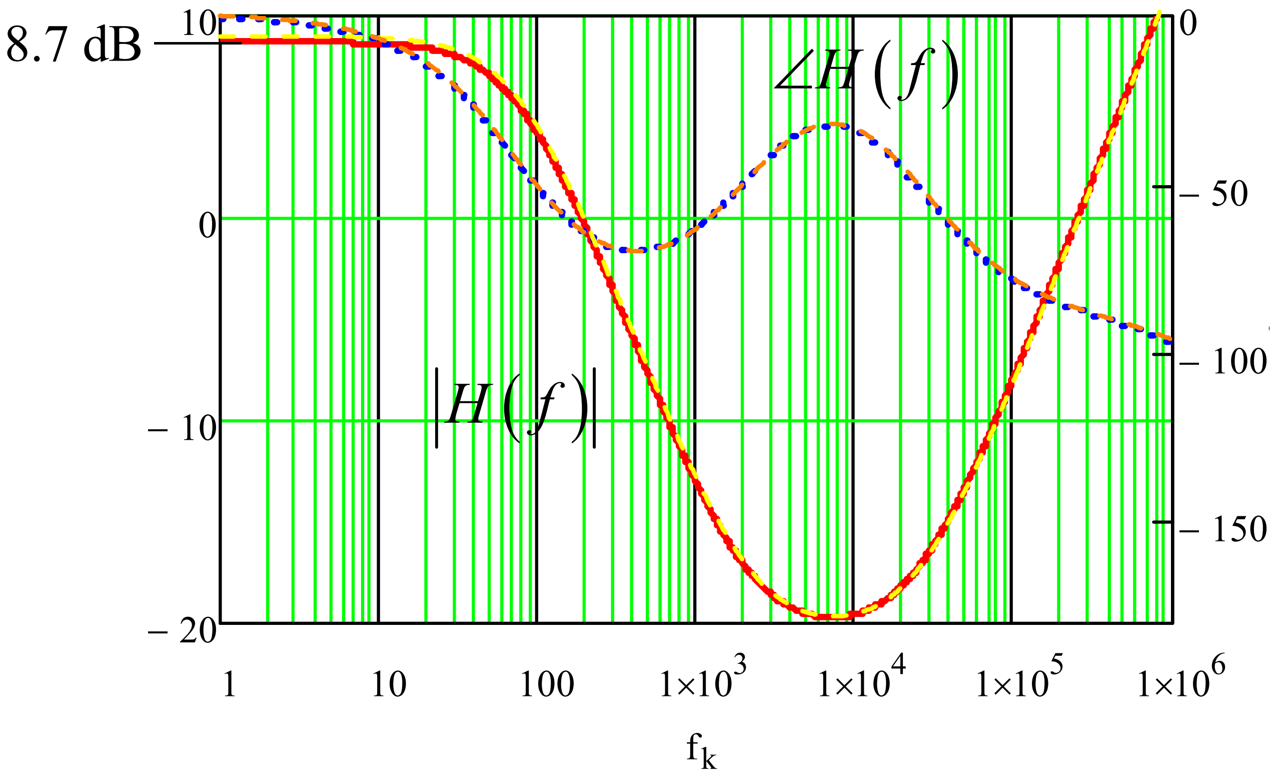
$$H_0 = \frac{N R_{load}\left(k_1 - k_4\right)}{\mathrm{Div}\cdot\left[N^2 + R_{load}\left(k_6 - k_3\right)\right]}$$

- Finally, the complete transfer function looks like

$$H(s) \approx H_0 \frac{\left(1+\frac{s}{\omega_{z_1}}\right)\left(1+\frac{s}{\omega_{z_2}}\right)}{1+\frac{s}{\omega_{p_1}}}$$

Simulation versus Equations

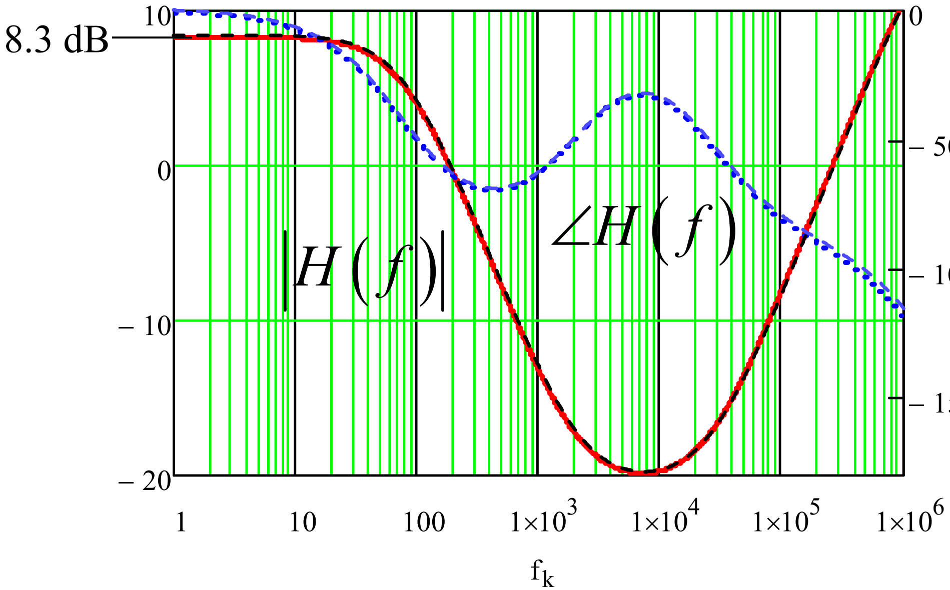
- It is important to test matching between model and equations



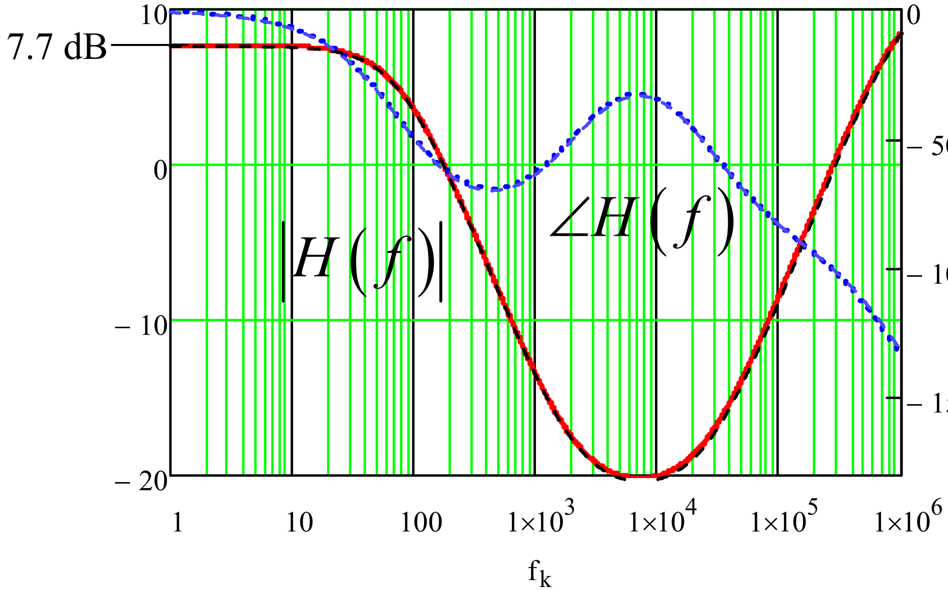
1st valley operation

Simulation and Calculations

❑ The dc gain slightly changes as valley number is modified



3rd valley operation

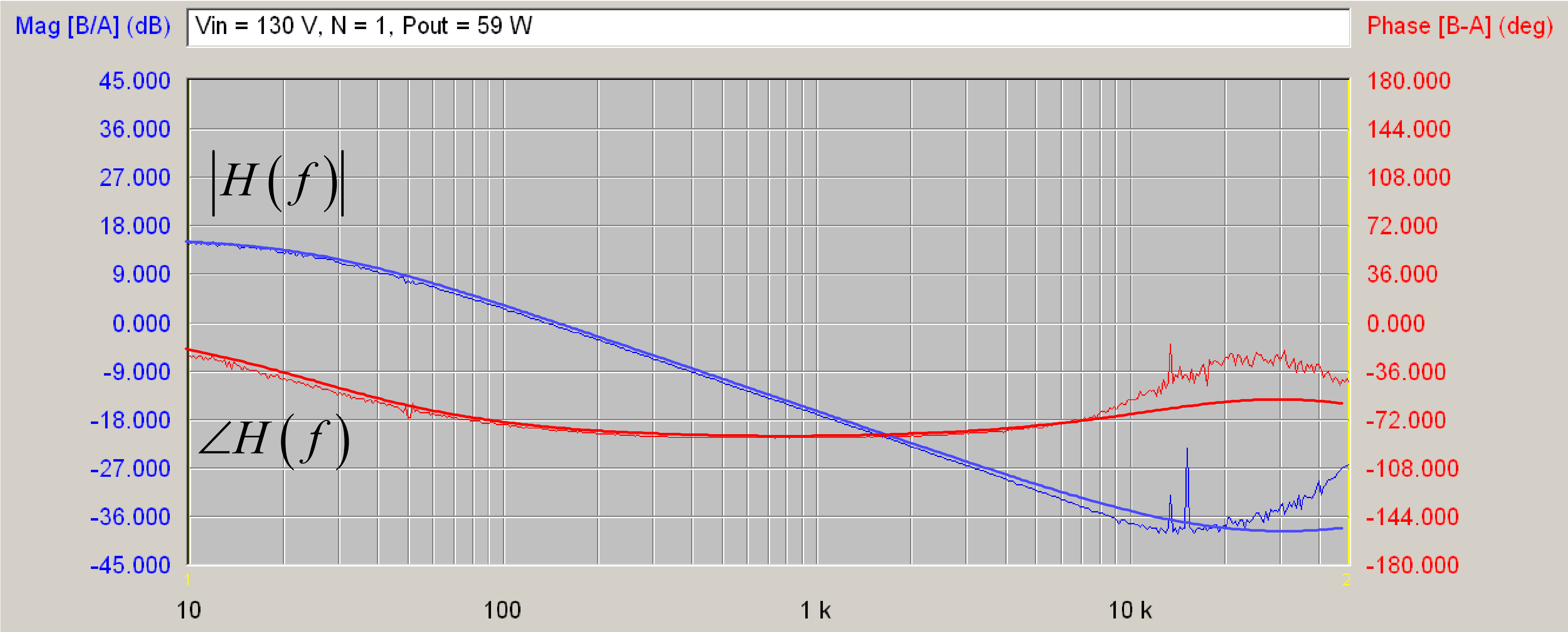


6th valley operation

❑ The dc gain slightly changes as valley number is modified

Laboratory Tests (1)

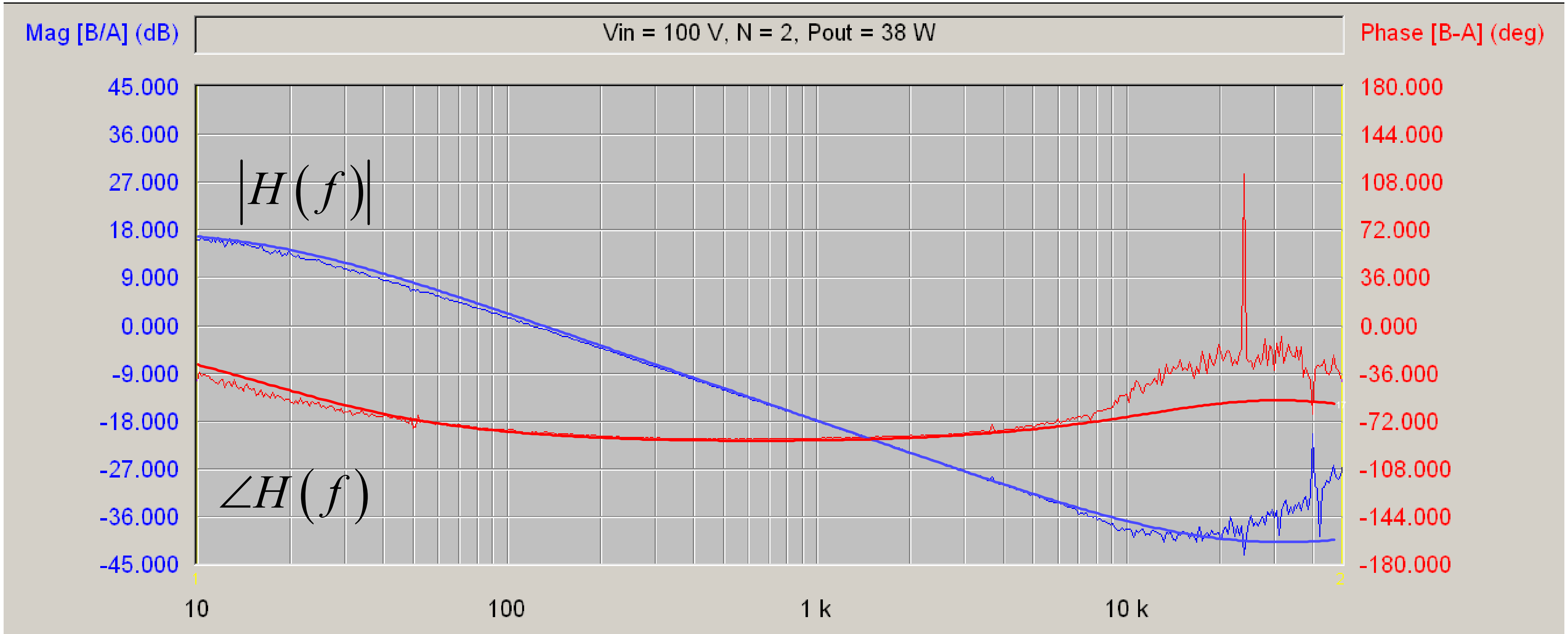
❑ The low to middle freq. agreement is good, deviation beyond



V_{in} = 130 V, valley = 1

Laboratory Tests (2)

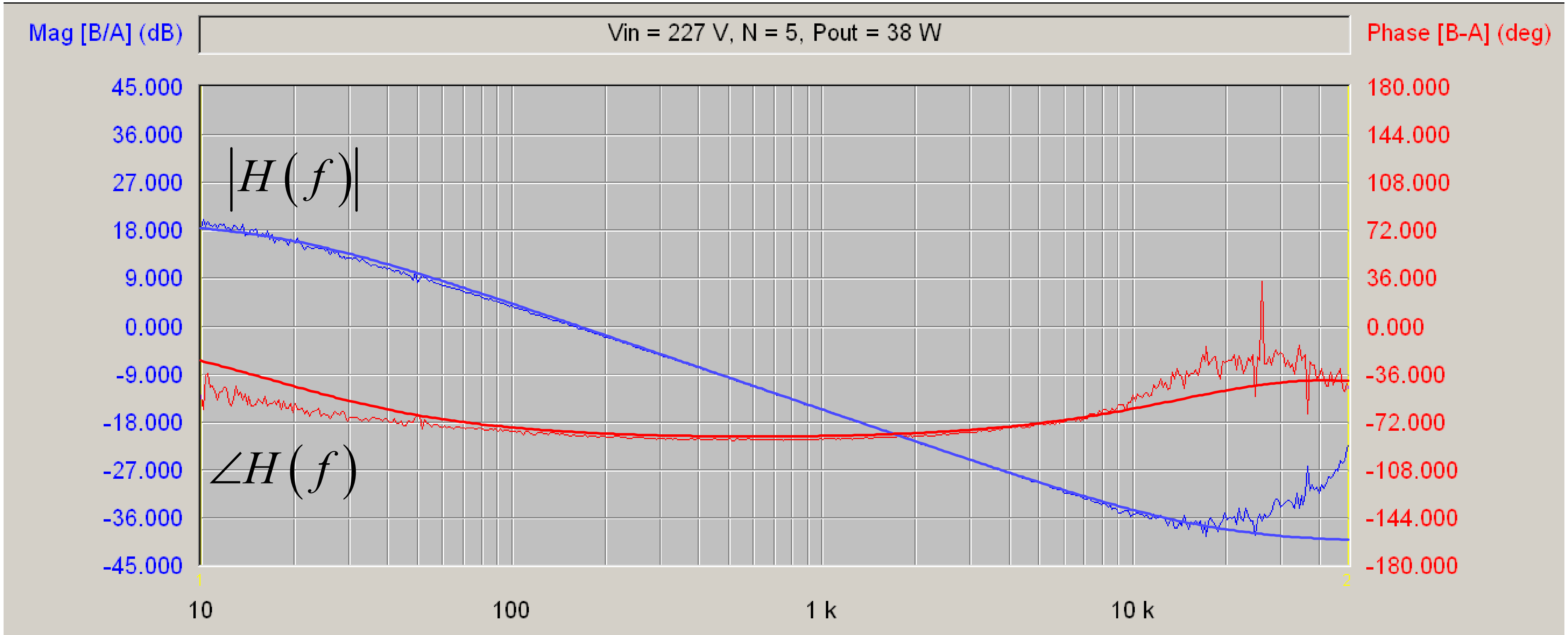
- The output power is changed and valley number 2 is engaged



V_{in} = 100 V, valley = 2

Laboratory Tests (3)

- The intput voltage is increased and valley changes to 5



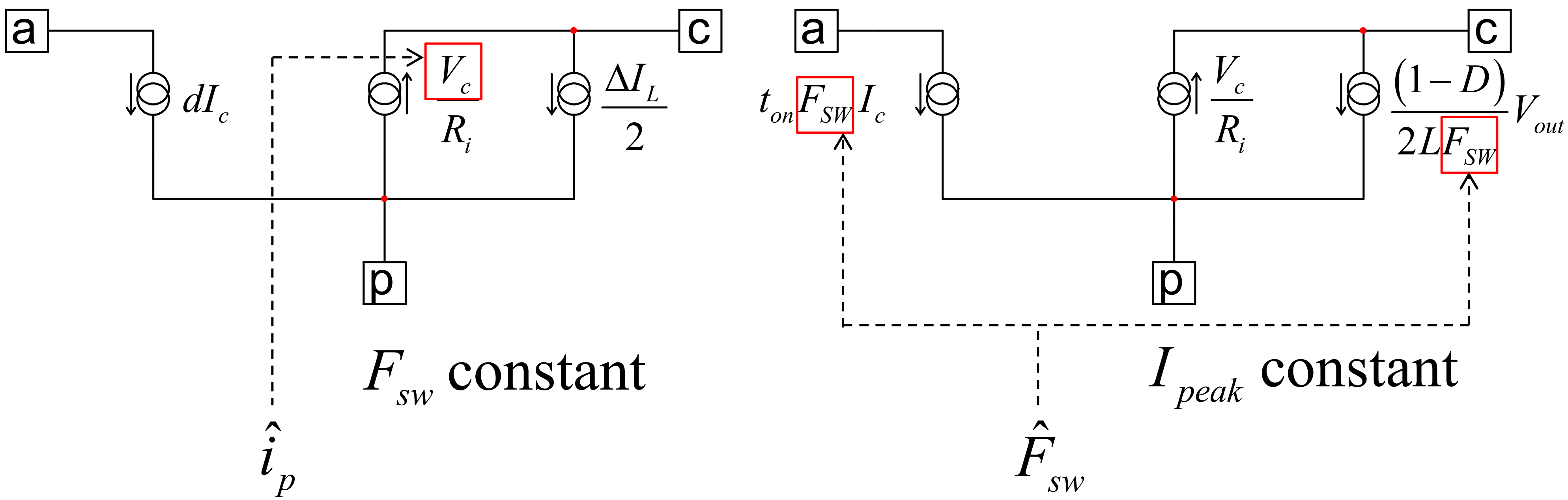
V_{in} = 227 V, valley = 5

Course Agenda

- ❑ What is a QR Converter?
- ❑ Deriving Operating Conditions
- ❑ The Over Power Problem
- ❑ Three Operating Modes
- ❑ Small-Signal Analysis – QR
- ❑ **Small-Signal Analysis – VCO**
- ❑ Compensating with TL431
- ❑ Control Loop Design Example
- ❑ Conclusion

Where do We Start From?

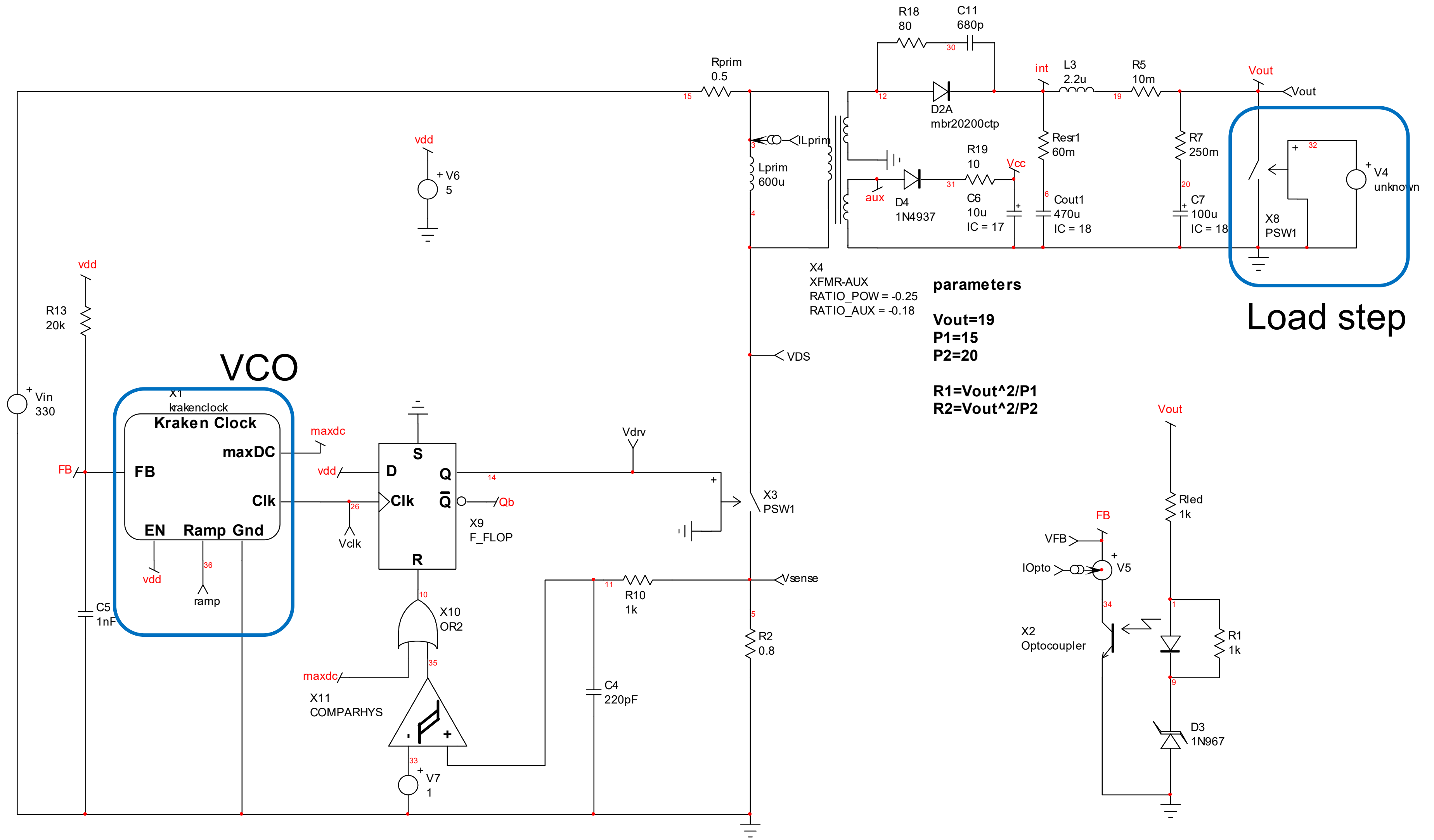
- We have a DCM peak-current mode control large-signal model
- The PWM switch controls the peak current at a fixed F_{sw}
- Why not fixing the peak current and controlling F_{sw}?



- The peak current is frozen and the frequency is controlled

A Transient Load Step Response First

- We can check the cycle-by-cycle response to a load step

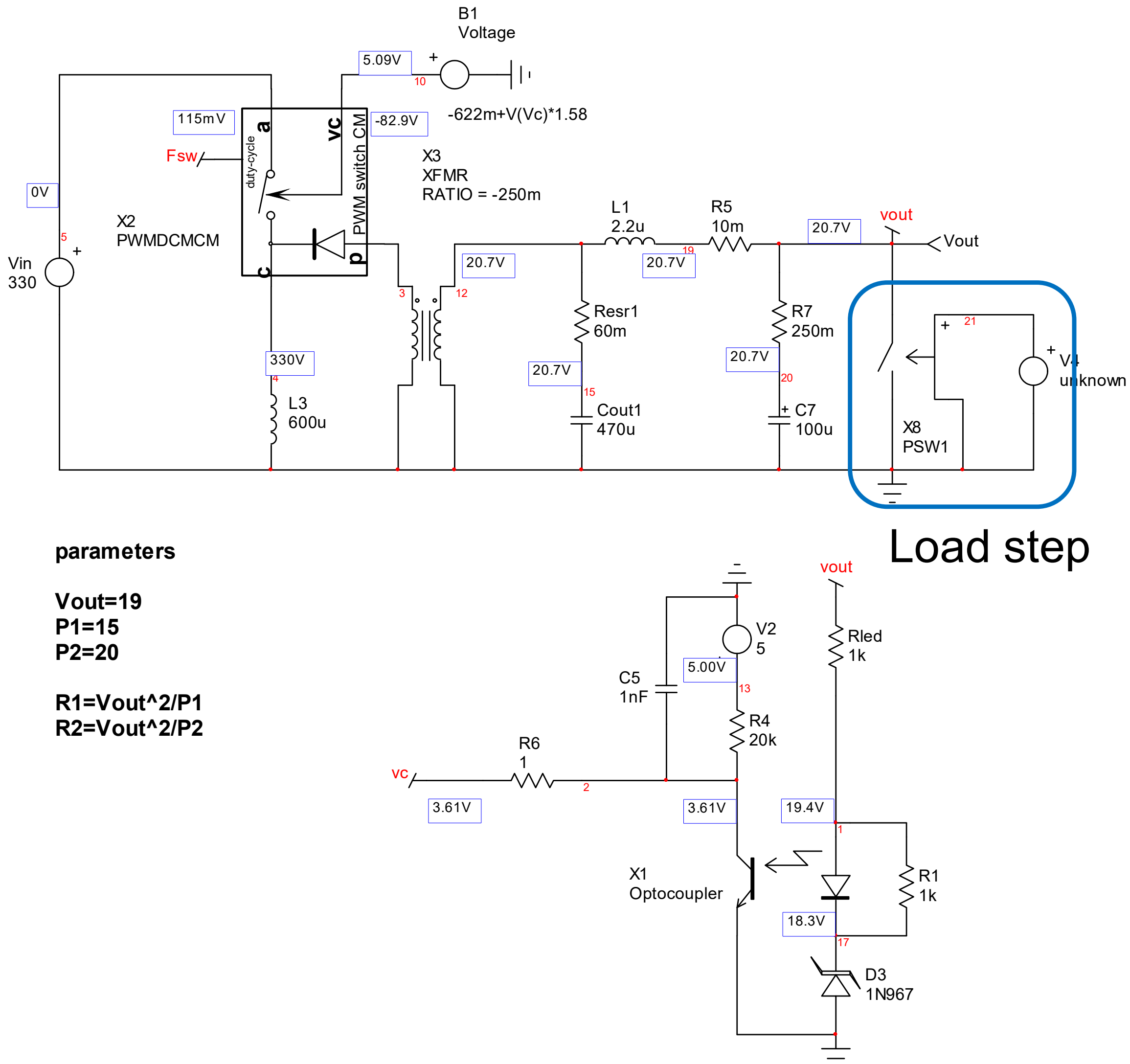


Cycle by cycle model

F_{sw} = 51 kHz

A Transient Load Step Response First

- And compare it to that of a modified PWM switch model



Averaged model

```
*
.param {Vc}=1
*
Bdc dcx 0 V =
{Vc}*V(Fsw)*10k/({Se}+(abs(v(a,c))*{Ri})/{L}+1u)
Xdc dcx dc limit params: clampH=0.99
clampL=7m
BVcp 6 p V=(V(dc)/(V(dc)+V(d2)))*V(a,p)
Blap a p I=(V(dc)/(V(dc)+V(d2)))*I(VM)
Bd2 d2X 0 V=(2*I(VM)*{L}-
v(a,c)*V(dc)^2*1/(V(Fsw)*10k)) / (
+v(a,c)*V(dc)*1/(V(Fsw)*10k)+1u )
Xd2 d2X dc d2 limit2
Rdum1 dc 0 1Meg
Rdum2 vc 0 1Meg
RS 7 c 1u
VM 6 7
*
.ENDS
```

Current frozen to $1/R_{sense}$

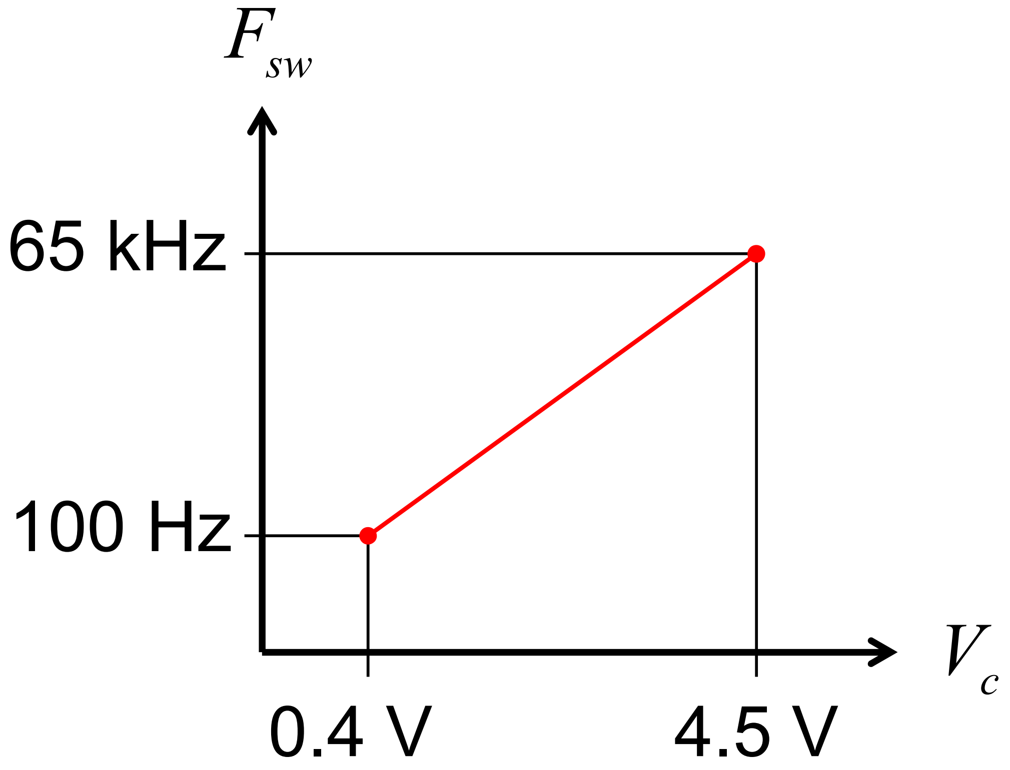
1 V = 10 kHz

Modified netlist of the PWM switch

F_{sw} = 50.9 kHz

You Need to Adjust the VCO Modulator

- ❑ The control voltage to the switching frequency is the VCO gain



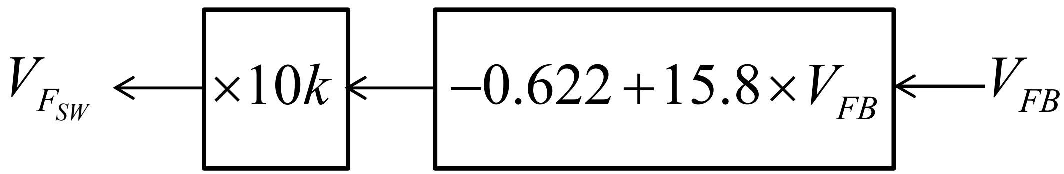
$$y = ax + b$$

$$y = \frac{\Delta F_{sw}}{\Delta V_c} = \frac{64.9k}{4.1} = 15.8\,\text{kHz/V}$$

$$100 = ax + b = 0.4 \times 15.8k + b$$

1/10000 Scaling factor

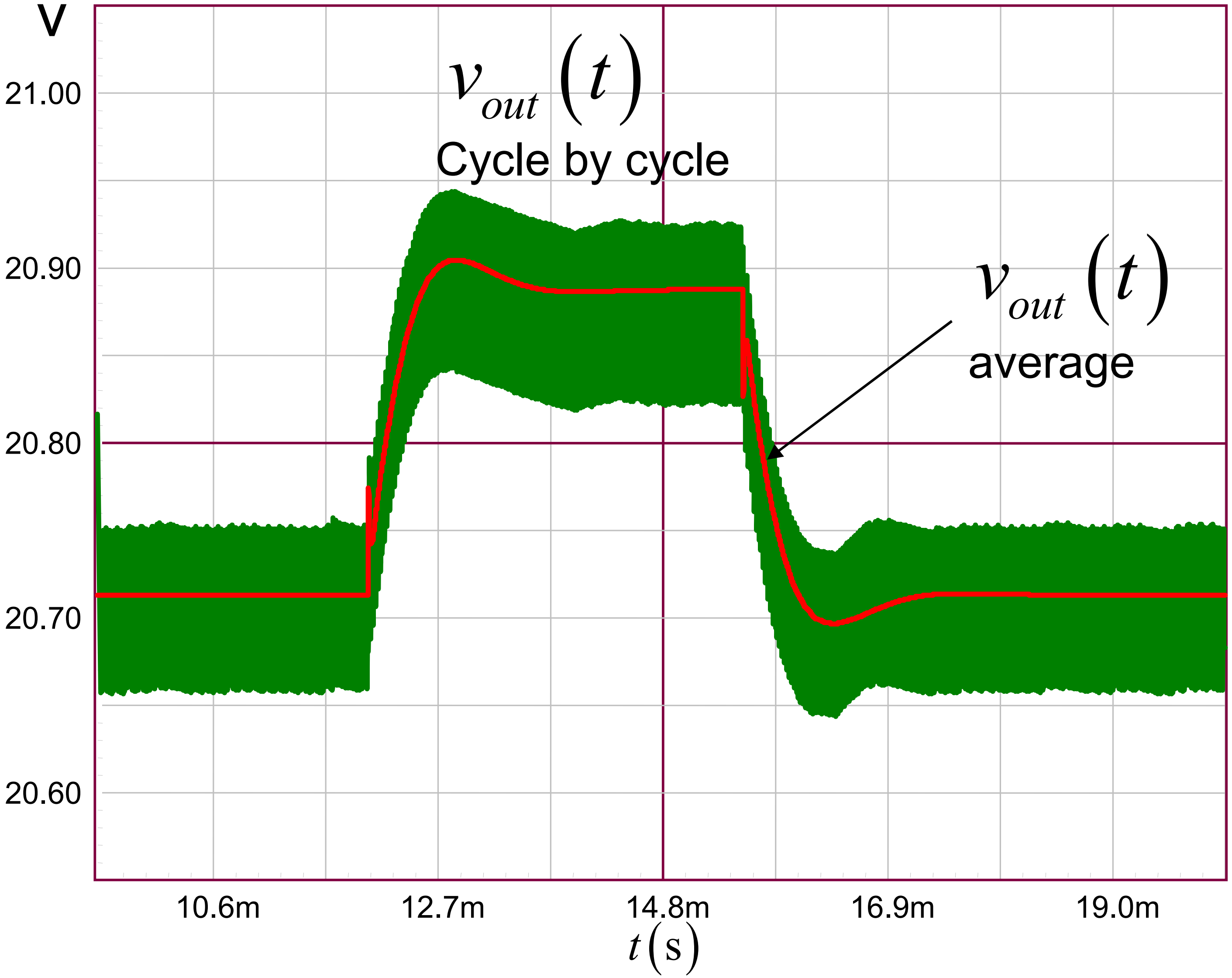
$$b = 0.10 - 0.4 \times 15.8 = -622\,\text{mV}$$



VCO gain G_{VCO} is $15.8\,\text{kHz/V}$

Compare the Transient Responses

- Responses are identical, the averaged version looks correct



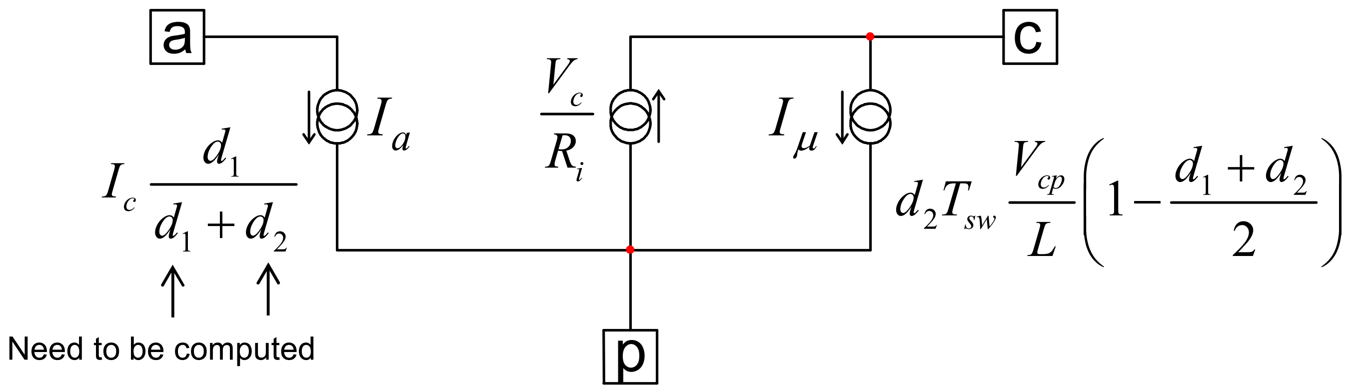
Keys of Small-Signal Analysis

❑ Rather than going full speed into small-signal analysis:

➢ break down the system into smaller parts

➢ run simplifications whenever you can

➢ go step by step and verify the answer always fits the original



Start from here, DCM model

"Switch-mode Power Supplies: SPICE Simulations and Practical Designs", C. Basso, McGraw-Hill 2008, page 161

Simplifying and Compacting the Model

- ❑ Compact the DCM model to suppress variables calculation

$$I_a = I_c \frac{d_1}{d_1 + d_2} \qquad I_\mu = \frac{d_2 V_{cp}}{F_{sw} L_p}\left(1 - \frac{d_1 + d_2}{2}\right)$$

$$d_1 = \frac{L_p V_c F_{sw}}{V_{ac} R_i} \qquad d_2 = \frac{2 I_c F_{sw} L_p}{d_1 V_{ac}} - d_1$$

substitute

$$I_a = I_c \frac{d_1}{d_1 + d_2}$$

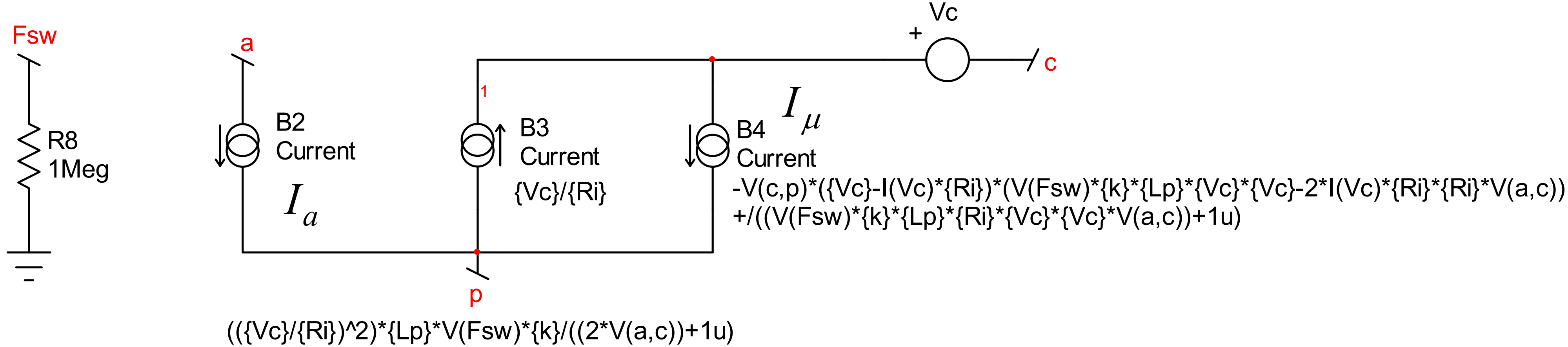
"Let the craziness begin"

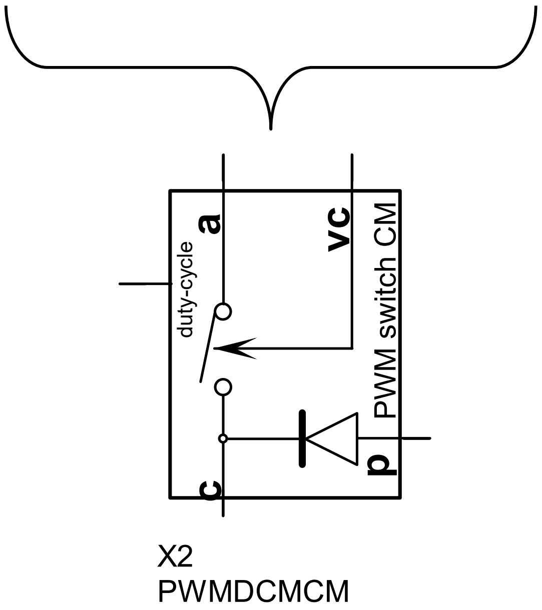
$$I_a = \frac{F_{sw} L_p V_c^{\,2}}{2 R_i^{\,2} V_{ac}}$$

$$I_\mu = -\frac{V_{cp}\left(V_c - I_c R_i\right)\left(F_{sw} L_p V_c^{\,2} - 2 I_c R_i^{\,2} V_{ac}\right)}{F_{sw} L_p R_i V_c^{\,2} V_{ac}}$$

The New Model Looks Simpler

- ❑ The equations no longer include a computed variable

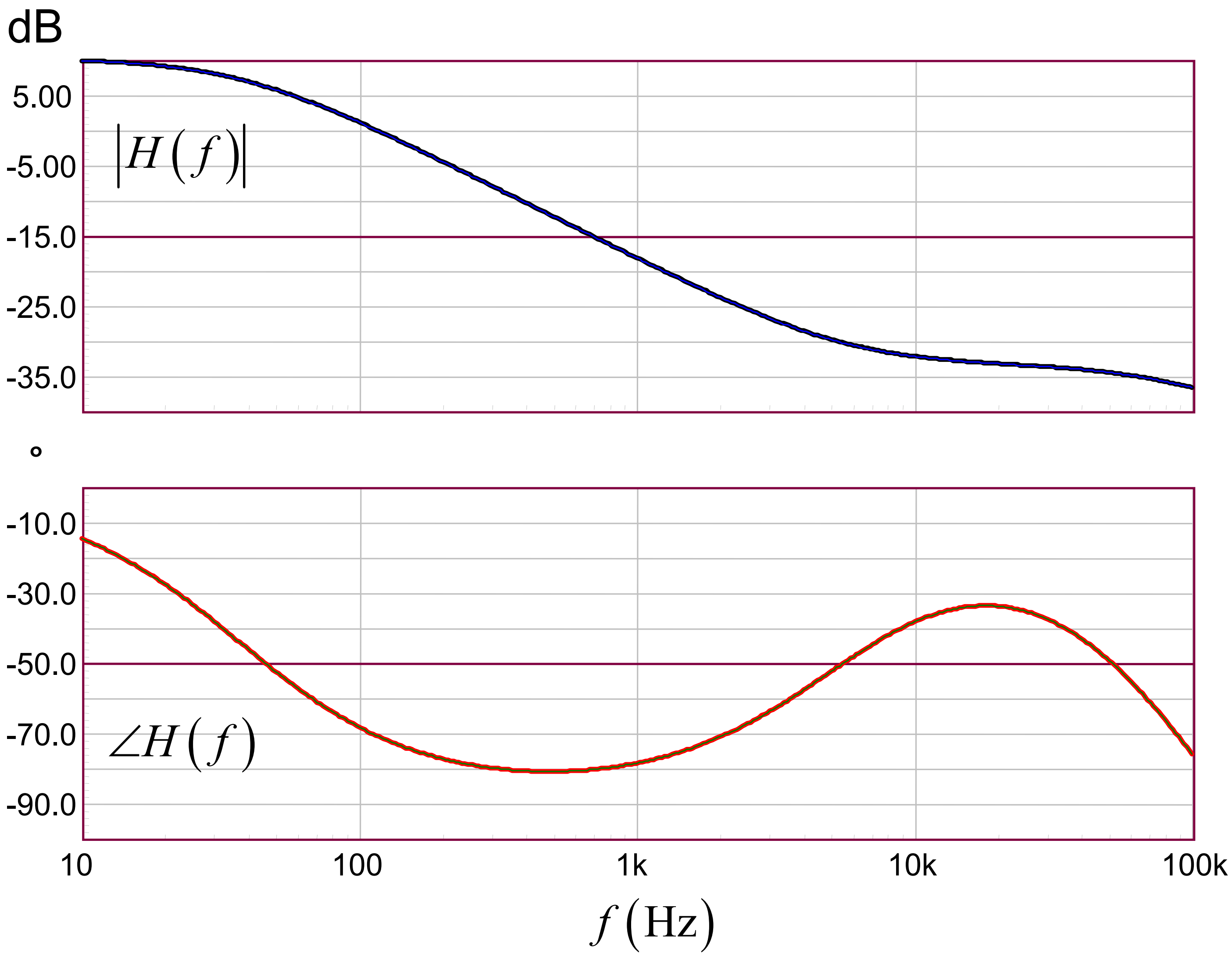




Check against the complete model

Ac Responses are Similar

- ❑ The curves perfectly superimpose, 1st step is ok



Second Step, Linearize the Sources

- ❑ To apply Laplace equations, we need linear elements
- ❑ Linearization can be done in different ways:

✓ perturb all equations with a small quantity (the "hat" notation)
✓ re-arrange the terms and collect dc and ac contributors
❖ can be tedious to re-arrange, you neglect cross products

$$V_1 = R_1 I_1 + D V_3$$

$$V_1 + \hat{v}_1 = R_1\left(I_1 + \hat{i}_1\right) + \left(\hat{d} + D\right)\left(V_3 + \hat{v}_3\right)$$

$$V_1 = R_1 I_1 + D V_3$$ dc equation (bias point)

$$\hat{v}_1 = R_1 \hat{i}_1 + \hat{d} V_3 + \underbrace{\boxed{\hat{d}\hat{v}_3}}_{\approx 0} + D\hat{v}_3$$ ac equation

Second Step, Linearize the Sources

- ❑ A second option is to calculate partial derivative coefficients

✓ the process can be automated by Mathcad®
✓ you only have ac terms, no sort needed

$$V_1(I_1, D, V_3) = R_1 I_1 + D V_3$$

$$dV_1 = \left(\frac{\partial V_1(I_1, D, V)}{\partial I_1}\right)\Bigg|_{D,V_3} dI_1 + \frac{\partial V_1(I_1, D, V)}{\partial D}\Bigg|_{I_1,V_3} dD + \frac{\partial V(I_1, D, V)_1}{\partial V_3}\Bigg|_{D,I_1} dV_3$$

$$\hat{v}_1 = \frac{\partial V_1(I_1, D, V)}{\partial I_1}\Bigg|_{D,V_3} \hat{i}_1 + \frac{\partial V_1(I_1, D, V)}{\partial D}\Bigg|_{I_1,V_3} \hat{d} + \frac{\partial V_1(I_1, D, V)}{\partial V_3}\Bigg|_{D,I_1} \hat{v}_3$$

$$\hat{v}_1 = R_1 \hat{i}_1 + D\hat{v}_3 + \hat{d} V_3$$

ac equation, no cross products

Second Step, Linearize the Sources

❑ Now, identify the variables in each source

$$I_a = \frac{F_{sw} L_p V_c^{\,2}}{2 R_i^{\,2} V_{ac}} \longrightarrow \begin{matrix} F_{sw} \\ V_{ac} \end{matrix}$$

$$I_\mu = -\frac{V_{cp}\left(V_c - I_c R_i\right)\left(F_{sw} L_p V_c^{\,2} - 2 I_c R_i^{\,2} V_{ac}\right)}{F_{sw} L_p R_i V_c^{\,2} V_{ac}} \longrightarrow \begin{matrix} V_{cp} \\ V_{ac} \\ F_{sw} \\ I_c \end{matrix}$$

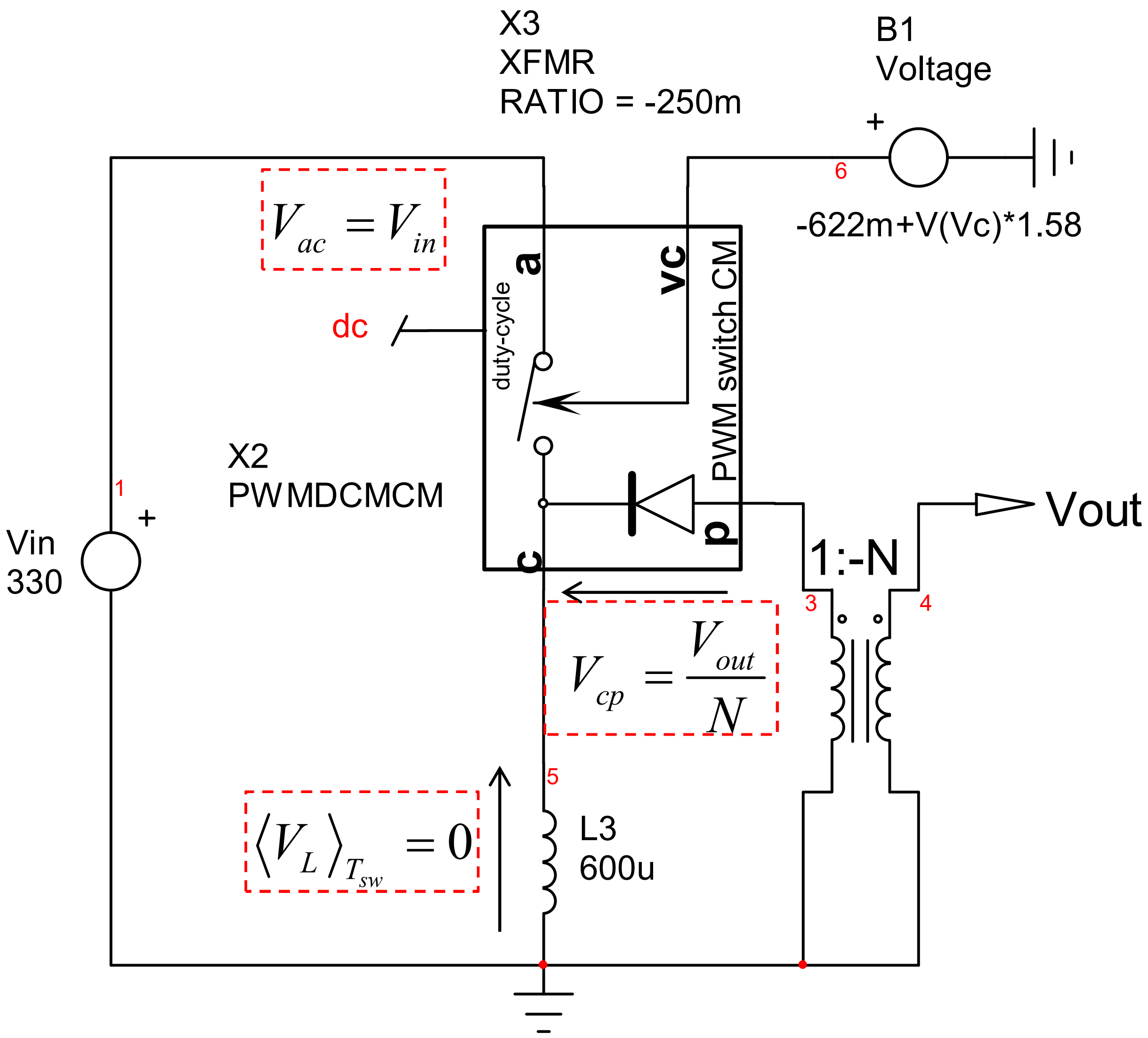
❑ 6 variables imply six partial derivatives, 6 coefficients

❑ You must identify these static variables first

➢ Look at the PWM switch configuration

Identify the Variables in the Schematic

- the average voltage across L_3 is 0: point c is grounded.



Sources Derivation

- We can now individually derive all these sources

$$I_a = \frac{F_{sw} L_p V_c^2}{2 R_i^2 V_{ac}}$$

$$\hat{i}_a = \left(\frac{\partial I_a}{\partial F_{sw}}\right)_{V_{ac}} \hat{F}_{sw} + \left(\frac{\partial I_a}{\partial V_{ac}}\right)_{F_{sw}} \hat{v}_{ac}$$

$$k_1 = \frac{L_p V_c^2}{2 R_i^2 V_{ac}} \qquad k_2 = -\frac{F_{sw} L_p V_c^2}{2 R_i^2 V_{ac}^2}$$

$$I_\mu = -\frac{V_{cp}\left(V_c - I_c R_i\right)\left(F_{sw} L_p V_c^2 - 2 I_c R_i^2 V_{ac}\right)}{F_{sw} L_p R_i V_c^2 V_{ac}}$$

$$\hat{i}_\mu = \underbrace{\left(\frac{\partial I_\mu}{\partial V_{cp}}\right)_{I_c, F_{sw}, V_{ac}}}_{k_3} \hat{v}_{cp} + \underbrace{\left(\frac{\partial I_\mu}{\partial I_c}\right)_{F_{sw}, V_{cp}, V_{ac}}}_{k_4} \hat{i}_c + \underbrace{\left(\frac{\partial I_\mu}{\partial F_{sw}}\right)_{I_c, V_{cp}, V_{ac}}}_{k_5} \hat{F}_{sw} + \underbrace{\left(\frac{\partial I_\mu}{\partial V_{ac}}\right)_{V_{cp}, I_c, F_{sw}}}_{k_6} \hat{v}_{ac}$$

Sources Derivation

- ❑ Yes, Mathcad® or an equivalent software is of great help…

$$k_3 = -\frac{\left(V_c - I_c R_i\right)\left(F_{sw} L_p V_c^2 - 2 I_c R_i^2 V_{ac}\right)}{F_{sw} L_p R_i V_c^2 V_{ac}}$$

$$k_4 = \frac{V_{cp}\left(F_{sw} L_p V_c^2 - 2 I_c R_i^2 V_{ac}\right)}{F_{sw} L_p V_c^2 V_{ac}} + \frac{2 R_i V_{cp}\left(V_c - I_c R_i\right)}{F_{sw} L_p V_c^2}$$

$$k_5 = \frac{V_{cp}\left(V_c - I_c R_i\right)\left(F_{sw} L_p V_c^2 - 2 I_c R_i^2 V_{ac}\right)}{F_{sw}^2 L_p R_i V_c^2 V_{ac}} - \frac{V_{cp}\left(V_c - I_c R_i\right)}{F_{sw} R_i V_{ac}}$$

$$k_6 = \frac{V_{cp}\left(V_c - I_c R_i\right)\left(F_{sw} L_p V_c^2 - 2 I_c R_i^2 V_{ac}\right)}{F_{sw} L_p R_i V_c^2 V_{ac}^2} + \frac{2 I_c R_i V_{cp}\left(V_c - I_c R_i\right)}{F_{sw} L_p V_c^2 V_{ac}}$$

Evaluate all These Coefficients

- Select a converter at a certain operating point

$$V_{out} = 21.1\text{ V}$$
$$R_{load} = 18\ \Omega$$
$$R_i = 0.8\ \Omega$$
$$V_c = 1\text{V}$$
$$V_{ac} = 330\text{ V}$$
$$r_C = 0.06\,\Omega$$
$$C_{out} = 470\,\mu\text{F}$$
$$N_1 = 0.25$$
$$V_{cp} = \frac{V_{out}}{N_1} = 84.4\text{ V}$$
$$L_p = 600\,\mu\text{H}$$
$$k_F = 10000$$

$$F_{sw} = \frac{2I_{out}R_i^2V_{out}}{L_pV_c^2} = 52.8\,\text{kHz}$$

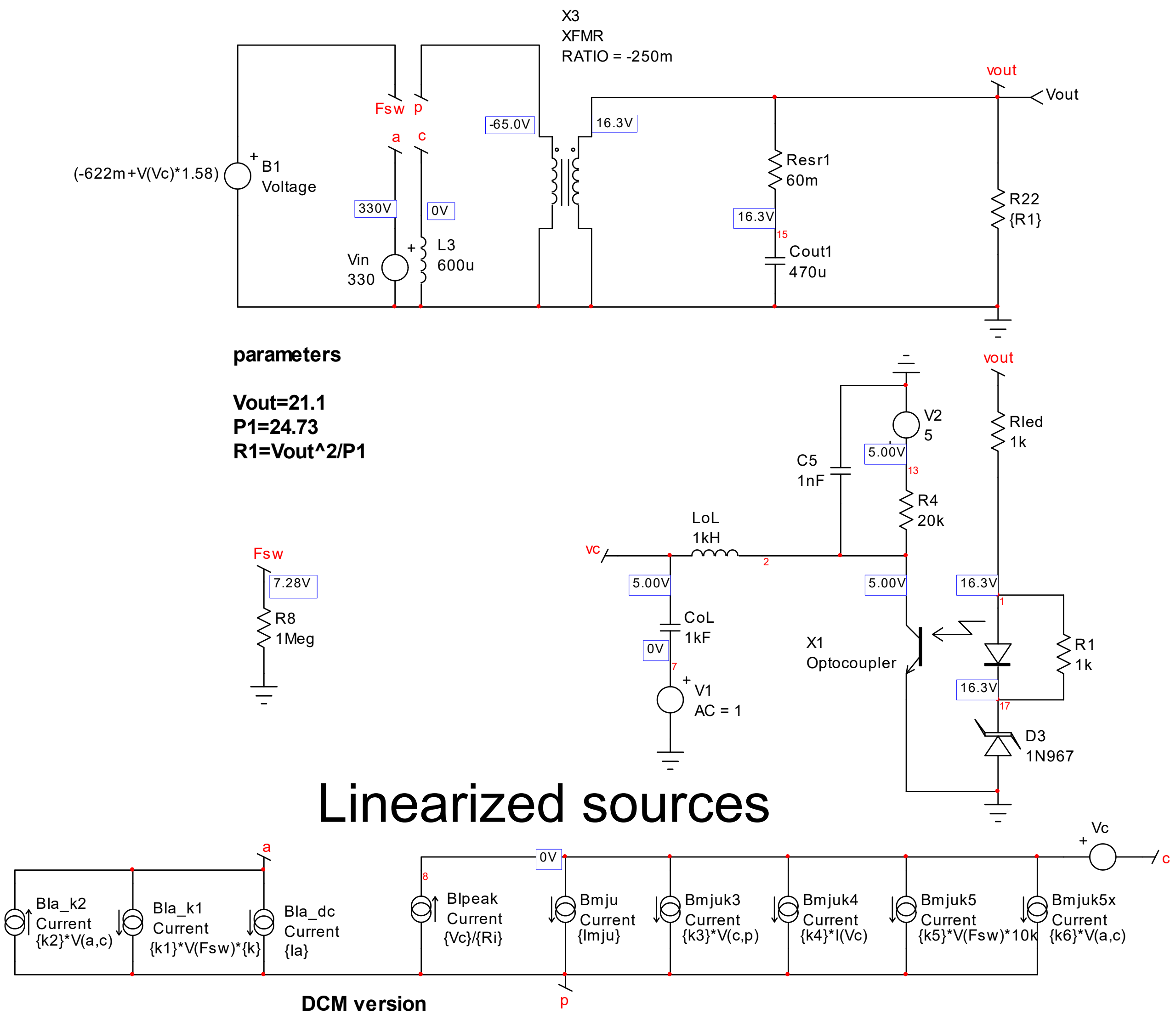
$$I_c = \frac{F_{sw}L_pV_c^2\left(V_{ac}+V_{cp}\right)}{2R_i^2V_{ac}V_{cp}} = 0.368\,\text{A}$$

$$I_a = \frac{F_{sw}L_pV_c^2}{2R_i^2V_{ac}} = 0.075\,\text{A}$$

$$I_\mu = 0.882\,\text{A}$$

Test the Coefficient Values with the Sources

- ❑ Capture a new schematic with the linearized sources



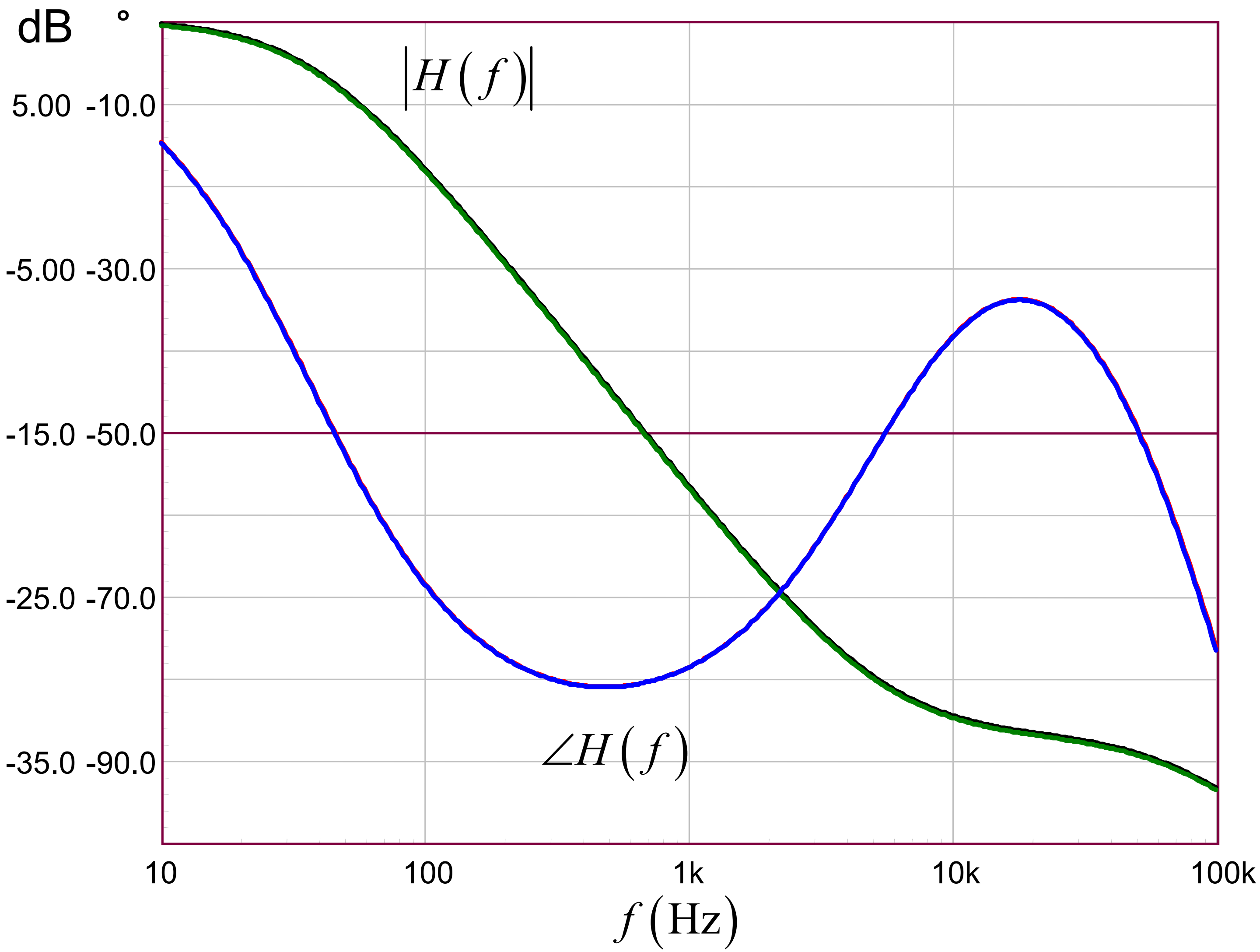
parameters

Vc=1
Lp=600u
Ri=0.8
Lp=600u
k=10k
N=250m
Fsw=52.766k
Iout=P1/Vout
Vac=330
Vcp=Vout/N
Ia=Fsw*Lp*Vc^2/(2*Ri^2*Vac)
k1=(Lp*Vc^2)/(2*Ri^2*Vac)
k2=(Fsw*Lp*Vc^2)/(2*Ri^2*Vac^2)
AA=Fsw*Lp*Vc^2*Vac+Fsw*Lp*Vc^2*Vcp
BB=2*Ri^2*Vac*Vcp
Ic=AA/BB
A=-Vcp*(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
B=Fsw*Lp*Ri*Vc^2*Vac
Imju=A/B
AAA=-(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
BBB=Fsw*Lp*Ri*Vc^2*Vac
k3=AAA/BBB
k4A=Vcp*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
k4B=Fsw*Lp*Vc^2*Vac
k4C=2*Ri*Vcp*(Vc-Ic*Ri)/(Fsw*Lp*Vc^2)
k4=(k4A/k4B)+k4C
k5A=Vcp*(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
k5B=Fsw^2*Lp*Ri*Vc^2*Vac
k5C=Vcp*(Vc-Ic*Ri)/(Fsw*Ri*Vac)
k5=(k5A/k5B)-k5C
k6A=Vcp*(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
k6B=Fsw*Lp*Ri*Vc^2*Vac^2
k6C=2*Ic*Ri*Vcp*(Vc-Ic*Ri)/(Fsw*Lp*Vc^2*Vac)
k6=(k6A/k6B)+k6C

Mathcad® coefficients

Responses with Previous Models are Similar

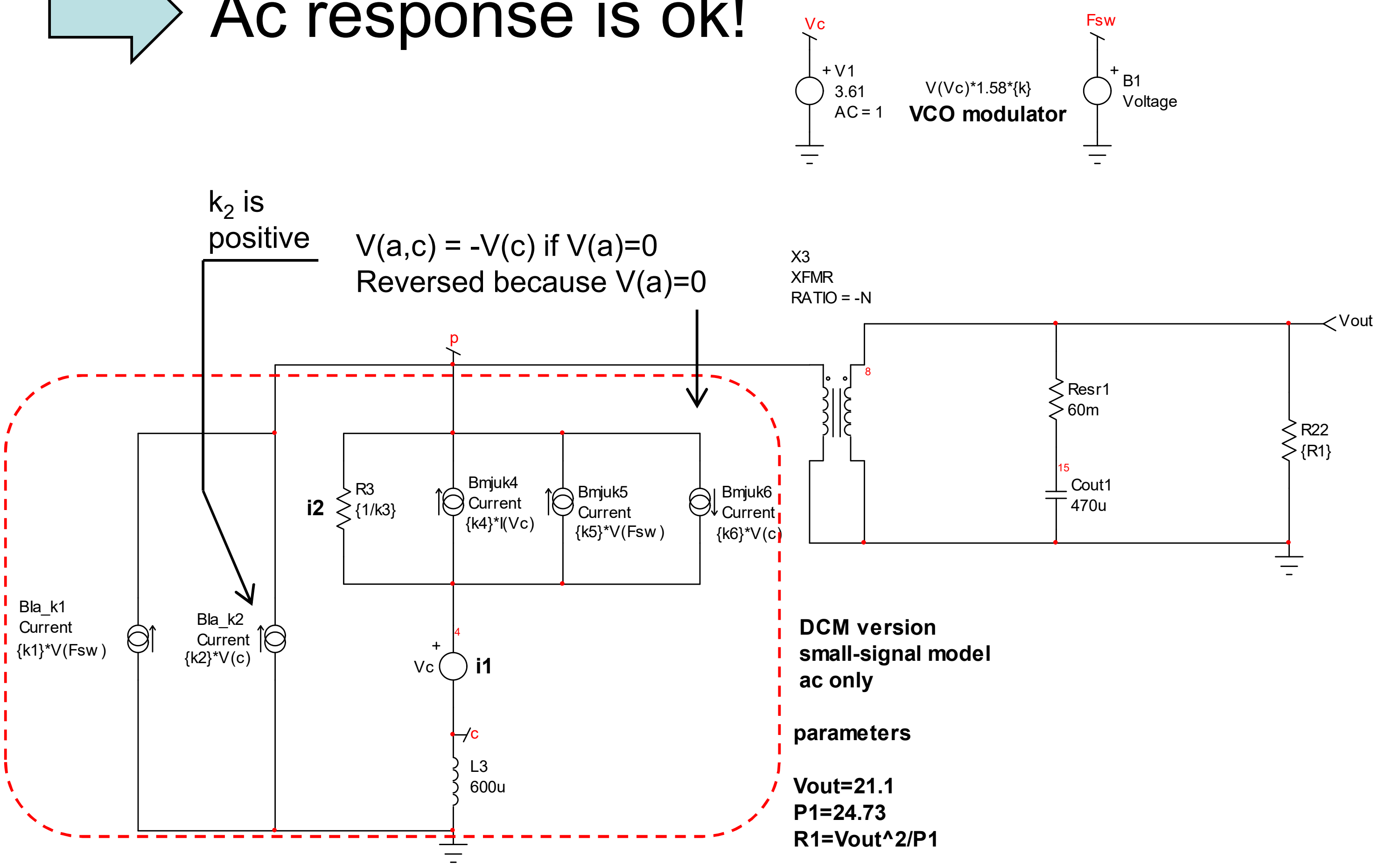
- The curves perfectly superimpose, 2nd step is ok



Combine and Arrange the Sources

- Now, re-arrange the sources in a more convenient way

Ac response is ok!



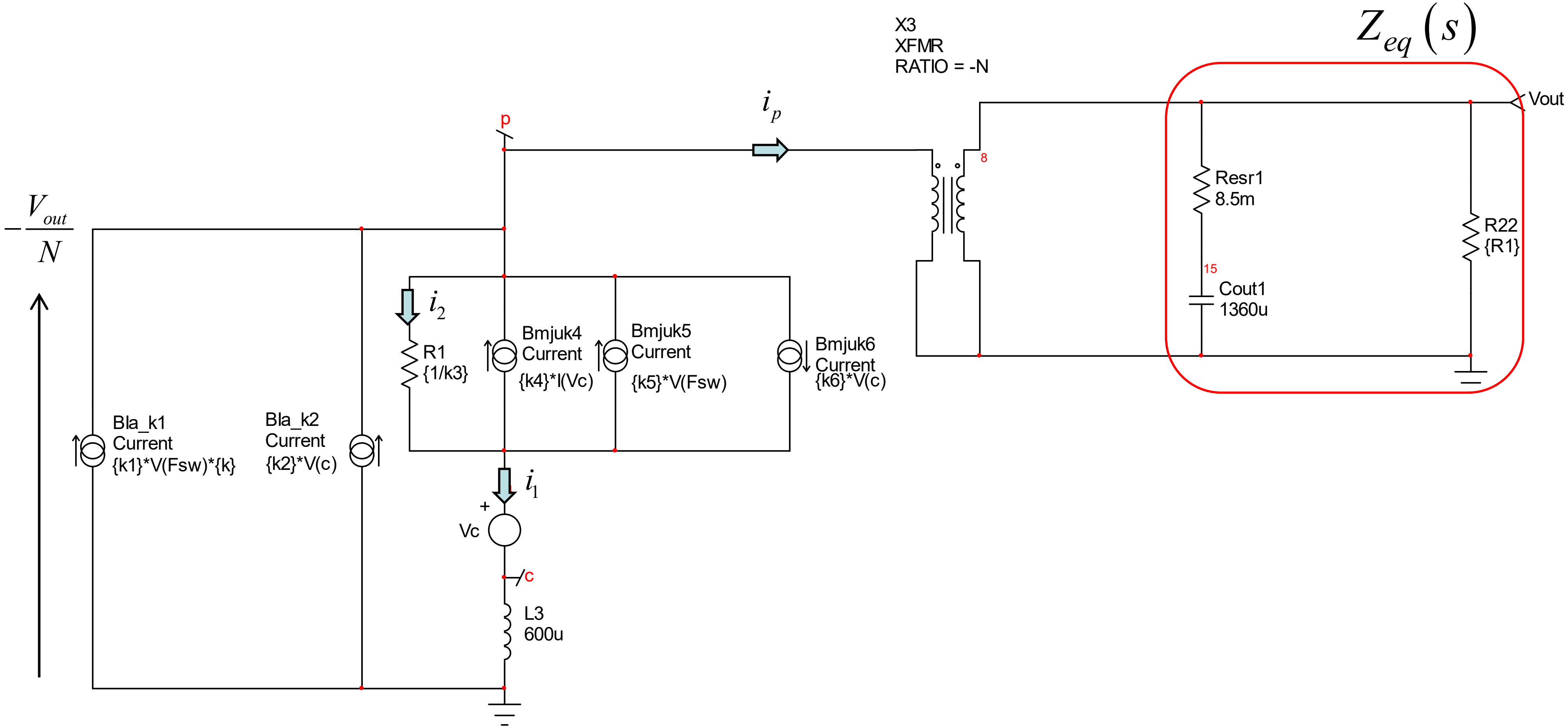
Linearized PWM switch

parameters

Vc=1
Lp=600u
Ri=0.8
Lp=600u
k=10k
N=250m
Fsw=52.766k
Iout=P1/Vout
Vac=330
Vcp=Vout/N
Ia=Fsw*Lp*Vc^2/(2*Ri^2*Vac)
k1=(Lp*Vc^2)/(2*Ri^2*Vac)
k2=(Fsw*Lp*Vc^2)/(2*Ri^2*Vac^2)
AA=Fsw*Lp*Vc^2*Vac+Fsw*Lp*Vc^2*Vcp
BB=2*Ri^2*Vac*Vcp
Ic=AA/BB
A=-Vcp*(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
B=Fsw*Lp*Ri*Vc^2*Vac
Imju=A/B
AAA=-(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
BBB=Fsw*Lp*Ri*Vc^2*Vac
k3=AAA/BBB
Rk3=1/k3
k4A=Vcp*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
k4B=Fsw*Lp*Vc^2*Vac
k4C=2*Ri*Vcp*(Vc-Ic*Ri)/(Fsw*Lp*Vc^2)
k4=(k4A/k4B)+k4C
k5A=Vcp*(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
k5B=Fsw^2*Lp*Ri*Vc^2*Vac
k5C=Vcp*(Vc-Ic*Ri)/(Fsw*Ri*Vac)
k5=(k5A/k5B)-k5C
k6A=Vcp*(Vc-Ic*Ri)*(Fsw*Lp*Vc^2-2*Ic*Ri^2*Vac)
k6B=Fsw*Lp*Ri*Vc^2*Vac^2
k6C=2*Ic*Ri*Vcp*(Vc-Ic*Ri)/(Fsw*Lp*Vc^2*Vac)
k6=(k6A/k6B)+k6C

Go for Mesh and Node Analysis

- Express the current and voltage in the primary side



Mesh and Node Analysis

- KCL: the sum of currents arriving at a node equals the sum of currents leaving the node:

$$k_4 i_1 + k_5 V(F_{sw}) + i_1 = i_2 + s k_6 L_p i_1 \qquad i_2 = i_1 + i_1 k_4 + k_5 V(F_{sw}) - s L_p i_1 k_6$$

$$i_1 = \frac{-\frac{V_{out}}{N_1} - R_1 i_2}{s L_p}$$

$$i_1(s) = -\frac{V_{out}(s) + N_1 R_1 i_1(s) + N_1 R_1 i_1(s) k_4 + N_1 R_1 k_5 V(F_{sw}) - s L_p R_1 i_1(s) k_6 N_1}{L_p N_1 s}$$

- solve for i_1:

$$i_1(s) = -\frac{V_{out}(s) + N_1 R_1 k_5 V(F_{sw})}{N_1 R_1 (1 + k_4) + s L_p N_1 (1 - k_6 R_1)}$$

Mesh and Node Analysis

- Apply similar technique to get the primary current:

$$i_p(s) = k_1 V(F_{sw}) + k_2 i_1(s) sL_p - i_1(s)$$

$$i_p(s) = \frac{V_{out}(s) - L_p V_{out}(s) k_2 s + N_1 R_1 V(F_{sw})(k_1 + k_5 + k_1 k_4) + sL_p N_1 V(F_{sw})(k_1 - R_1 k_1 k_6 - R_1 k_2 k_5)}{N_1 (R_1 + R_1 k_4 + sL_p - sL_p R_1 k_6)}$$

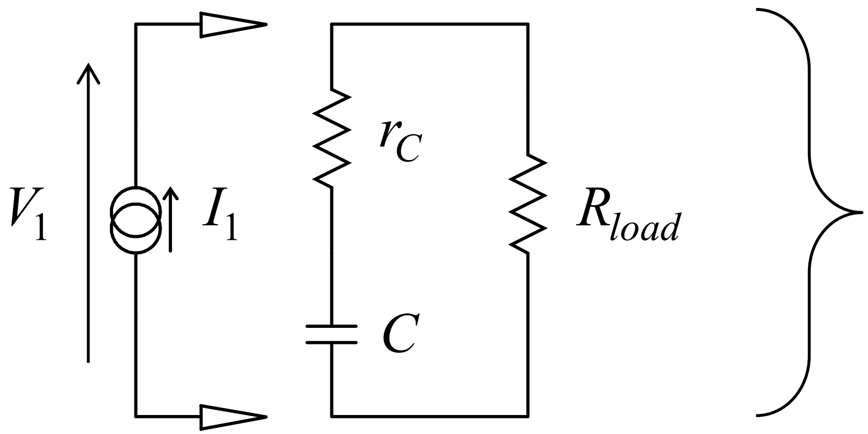
$$i_{out} = -\frac{i_p}{N_1} \qquad V_{out} = i_{out} Z_{eq}$$

$$V_{out}(s) = \frac{V_{out}(s) - L_p V_{out}(s) k_2 s + N_1 R_1 V(F_{sw})(k_1 + k_5 + k_1 k_4) + sL_p N_1 V(F_{sw})(k_1 - R_1 k_1 k_6 - R_1 k_2 k_5)}{N_1^{\,2} (R_1 + R_1 k_4 + sL_p - sL_p R_1 k_6)} Z_{eq}(s)$$

$$\frac{V_{out}(s)}{V(F_{sw})} = \frac{(N_1 R_1 (k_1 + k_5 + k_1 k_4) + sL_p N_1 (k_1 - R_1 k_1 k_6 - R_1 k_2 k_5))}{N_1^{\,2} (sR_1 L_p k_6 - sL_p - R_1 k_4 - R_1) - Z_{eq}(s) + sL_p k_2 Z_{eq}(s)} Z_{eq}(s)$$

Fast Analytical Techniques

- Fast analytical techniques unveil Z_{eq} in a second!



1st-order system

$$Z_{eq}(s) = Z_0 \frac{1 + s/\omega_z}{1 + s/\omega_p}$$

1. In dc, open the capacitor: $Z_0 = R_{load}$

2. What prevents the excitation I_1 from reaching the output V_1?

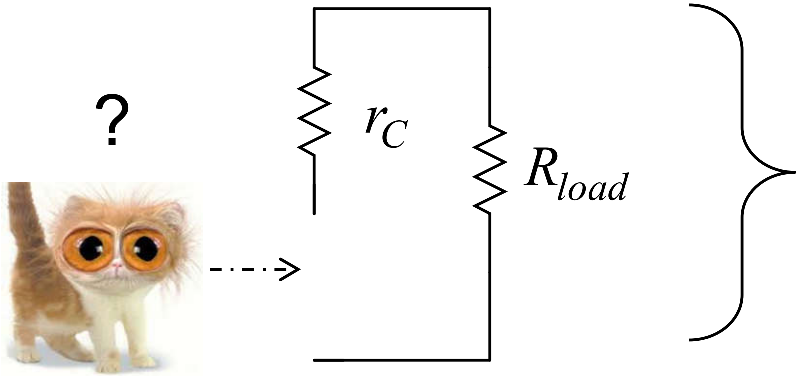
➢ A short-circuit between r_C and C:

$$r_C + \frac{1}{sC} = \frac{s r_C C + 1}{sC} = 0 \quad \Rightarrow \quad \omega_z = \frac{1}{r_C C}$$

Fast Analytical Techniques

❑ Get the time constant by putting the excitation to zero:

➢ open the current source and look at the cap. driving R



$$\tau = \left(r_C + R_{load}\right)C$$

❑ The equivalent impedance is therefore:

$$Z_{eq}\left(s\right) = R_{load}\frac{1 + sr_C C}{1 + sC\left(r_C + R_{load}\right)}$$

You cannot beat equation-solving by inspection!

Almost There…

- Develop the expression with $Z_{eq}(s)$, cry and re-arrange:

$$H(s) = H_0 \frac{N(s)}{D(s)} \qquad H_0 = -G_{VCO} \frac{N_1 R_{load} R_1 (k_1 + k_5 + k_1 k_4)}{R_{load} + N_1^2 R_1 (1 + k_4)} = \frac{V_{out}}{2F_{sw}} G_{VCO}$$

$$N(s) = \left(1 + sL_p \left(\frac{k_1 - R_1 k_1 k_6 - R_1 k_2 k_5}{R_1 (k_1 + k_5 + k_1 k_4)}\right)\right)(1 + sr_C C) \qquad D(s) = 1 + as + bs^2$$

$$a = C_{out}\left(R_{load} + r_C - \frac{R_{load}^2}{R_{load} + N_1^2 R_1 (1 + k_4)}\right) + L_p \left(\frac{N_1^2 - R_{load} k_2 - N_1^2 R_1 k_6}{R_{load} + N_1^2 R_1 (1 + k_4)}\right)$$

$$b = L_p N_1^2 C_{out} \left(\frac{R_{load} + r_C - \frac{R_{load}}{N_1^2} k_2 r_C - R_1 R_{load} k_6 - R_1 k_6 r_C}{R_{load} + N_1^2 R_1 (1 + k_4)}\right)$$

A Few More Minutes, Keep the Faith…

- Put the denominator under a second-order form and identify

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

$$f_0 = \frac{1}{2\pi\sqrt{L_p C_{out}}}\sqrt{\frac{\frac{R_{load}}{N_1^{\,2}} + R_1\left(1+k_4\right)}{R_{load} + r_C - \frac{R_{load}}{N_1^{\,2}}k_2 r_C - R_1 k_6\left(R_{load} + r_C\right)}} = 1.75\,\text{kHz}$$

$$Q = \frac{1}{\omega_0\left(C_{out}\left(R_{load} + r_C - \frac{R_{load}^{\;2}}{R_{load} + N_1^{\,2}R_1\left(1+k_4\right)}\right) + L_p\left(\frac{N_1^{\,2} - R_{load}k_2 - N_1^{\,2}R_1 k_6}{R_{load} + N_1^{\,2}R_1\left(1+k_4\right)}\right)\right)} = 0.021$$

A Few More Minutes, Keep the Faith…

- Extract the zeros:

$$f_{z_1} = \frac{1}{2\pi L_p \left(\frac{k_1 - R_1 (k_1 k_6 - k_2 k_5)}{R_1 (k_1 + k_5 + k_1 k_4)} \right)} = -487\,\text{kHz} \qquad f_{z_2} = \frac{1}{2\pi r_C C_{out}} = 5.6\,\text{kHz}$$

↑ RHPZ

- Extract the low-frequency poles:

$$f_{p_1} = \frac{1}{2\pi C_{out} \left(R_{load} + r_C - \frac{R_{load}^2}{R_{load} + N_1^2 R_1 (1 + k_4)} \right)} = \frac{1}{2\pi C_{out} \left(R_{load} + r_C - \frac{R_{load}}{2} \right)} \approx \frac{1}{\pi R_{load} C_{out}}$$

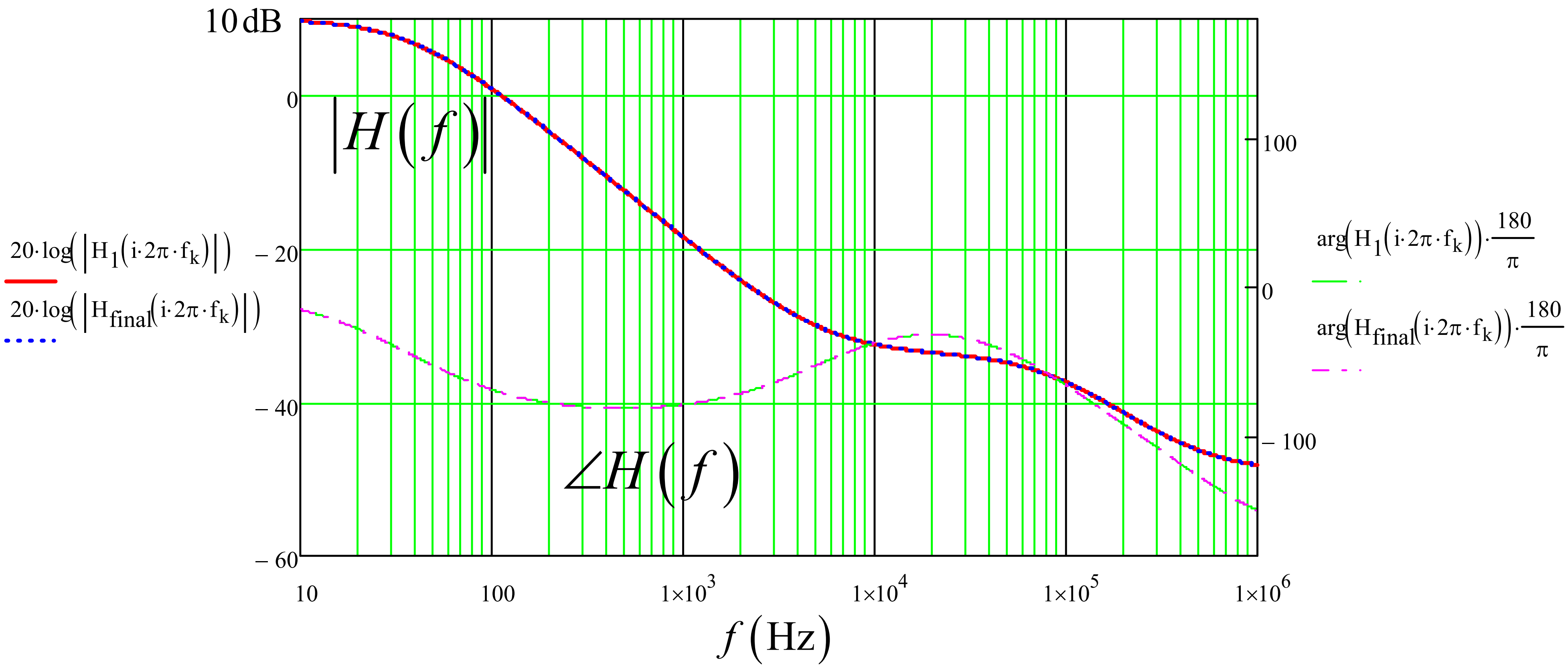
$\downarrow$ $r_C \ll R_{load}$

$$f_{p_1} = \frac{1}{\pi C_{out} (R_{load} + r_C)} = 37.73\,\text{Hz}$$

$$f_{p_2} = \frac{1}{2\pi L_p \left(\frac{N_1^2 - R_{load} k_2 - N_1^2 R_1 k_6}{R_{load} + N_1^2 R_1 (1 + k_4)} \right)} = 152\,\text{kHz}$$

Final Lap, Compare the ac Plots

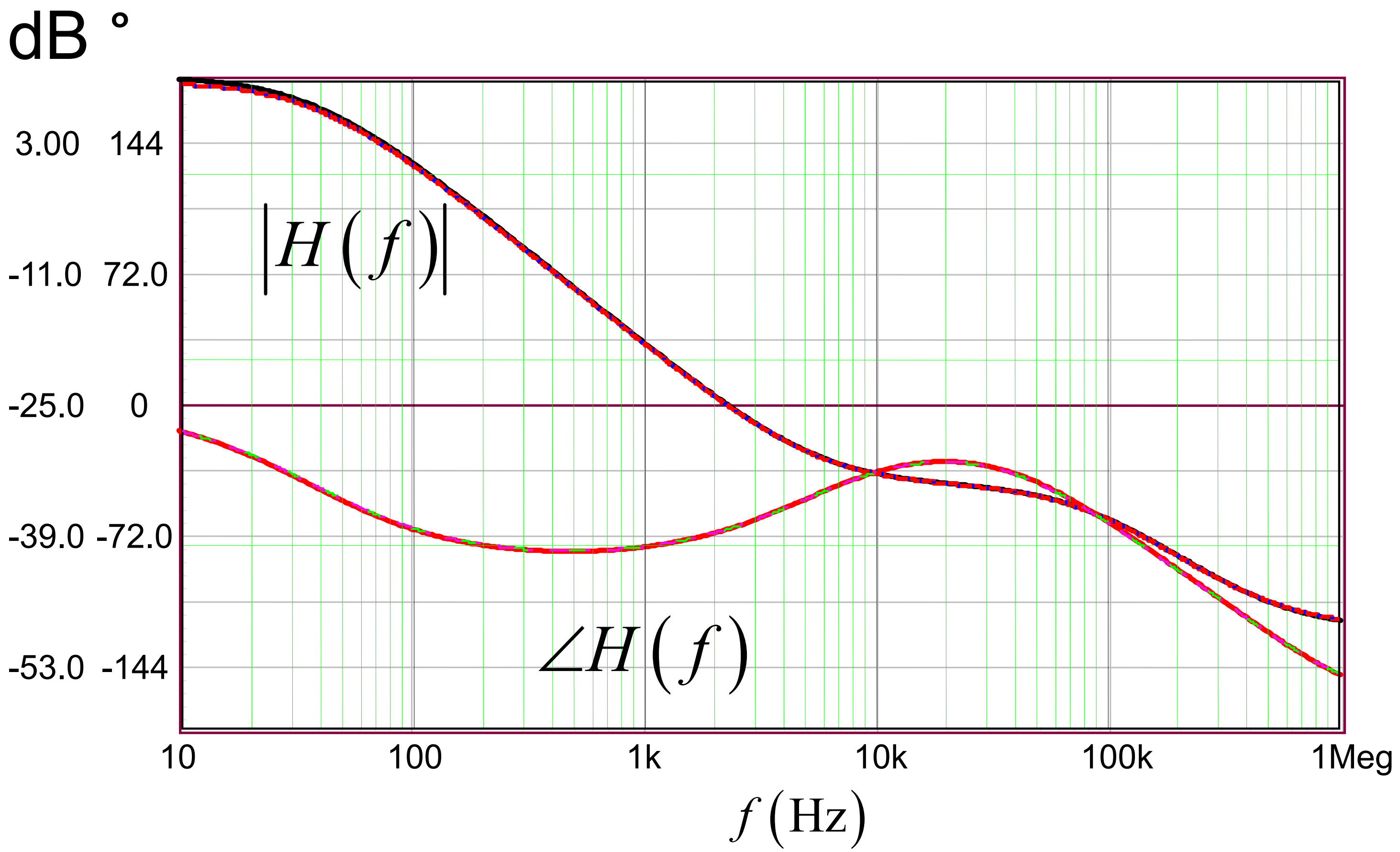
- Compare the original equation and its re-arranged form:



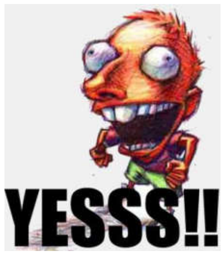
- It confirms the derivation is correct!

The Final Test: SPICE vs Mathcad

- ❑ If all is well, the curves must perfectly superimpose

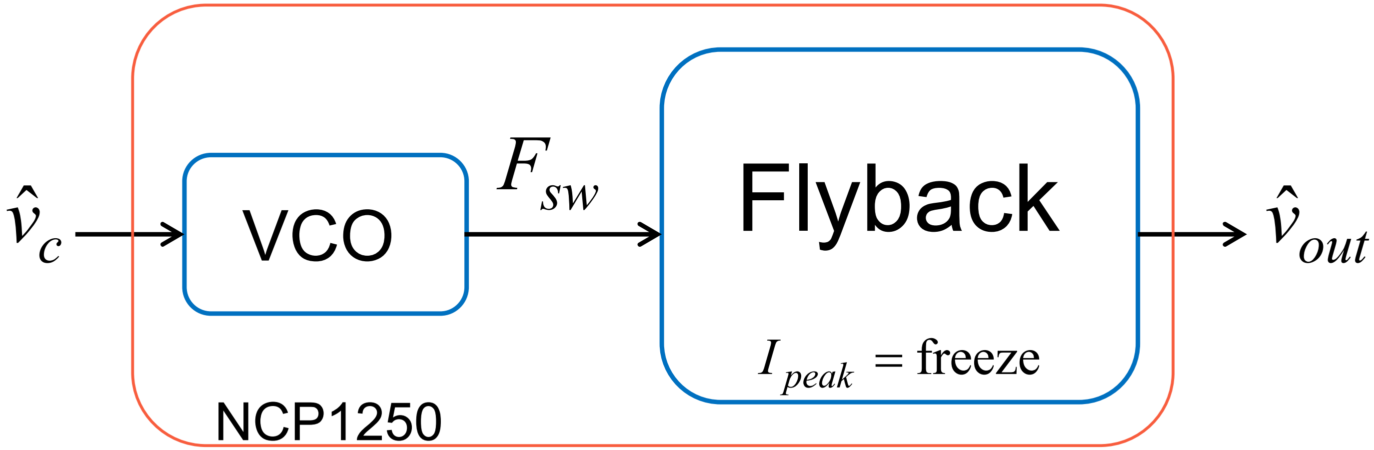


- ➢ If not, there is a hidden mistake: chase it (good luck)



The Final Transfer Function

- The current-mode flyback converter is operated in DCM
- The peak current is fixed and F_{sw} is controlled
- What is the simplified transfer function?

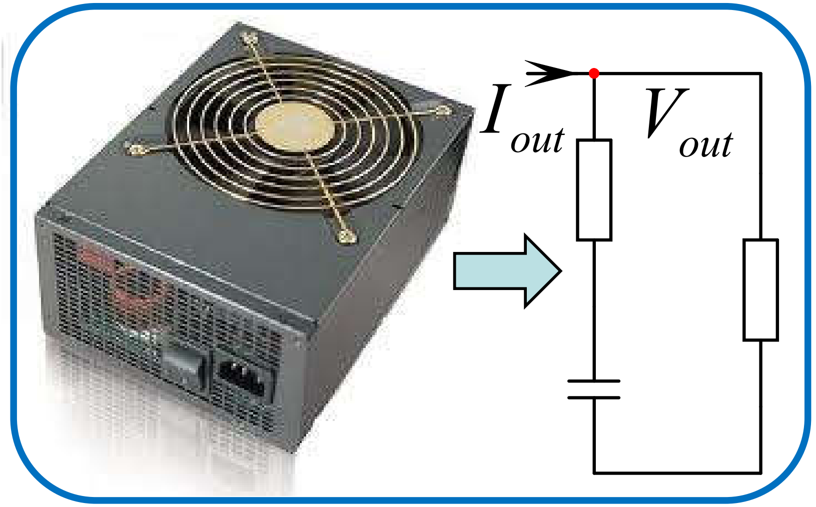


$$H(s) = \frac{\hat{v}_{out}}{\hat{v}_c} = H_0 \frac{1 + s/\omega_z}{1 + s/\omega_p} = \frac{V_{out}}{2F_{sw}} G_{VCO} \frac{1 + s r_C C_{out}}{1 + s \frac{R_{load}}{2} C_{out}}$$

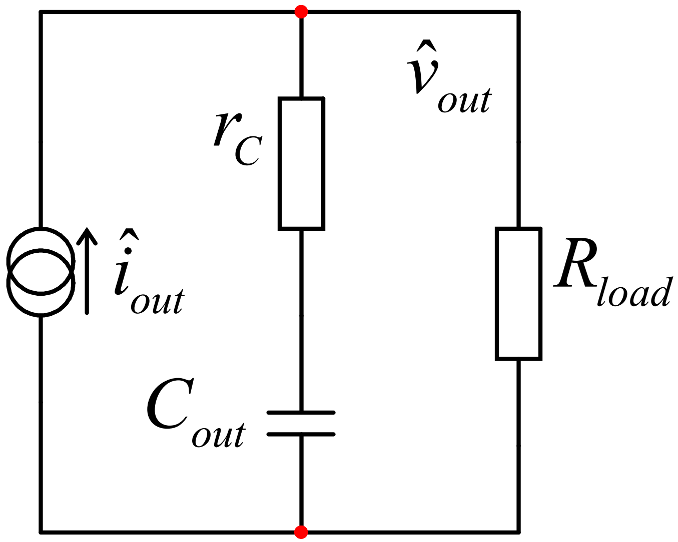
A Faster Way?

- If you are in a hurry and looking for a faster way:

  - Formulate the output current expression
  - Differentiate the expression to its variables
  - Draw an equivalent schematic and solve the equations



A DCM/CCM
current-mode converter



Simplified linearized model

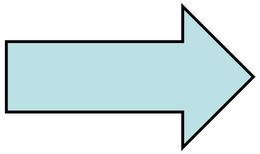
The Founding Equation is Well Known

- A flyback converter running in DCM obeys:

$$P_{out} = I_{out}V_{out} = \frac{1}{2}L_p I_{peak}{}^2 F_{sw} \rightarrow I_{peak} = \frac{V_c}{R_i} \rightarrow I_{out}V_{out} = \frac{1}{2}L_p\left(\frac{V_c}{R_i}\right)^2 F_{sw}$$

$$I_{out} = \frac{L_p\left(\frac{V_c}{R_i}\right)^2 \boxed{F_{sw}}}{2\boxed{V_{out}}}$$

Two variables



$$\hat{i}_{out} = \left(\frac{\partial I_{out}(F_{sw}, V_{out})}{\partial V_{out}}\right)_{F_{sw}} \hat{v}_{out} + \left(\frac{\partial I_{out}(F_{sw}, V_{out})}{\partial F_{sw}}\right)_{V_{out}} \hat{F}_{sw}$$

First Expression Comes Easily

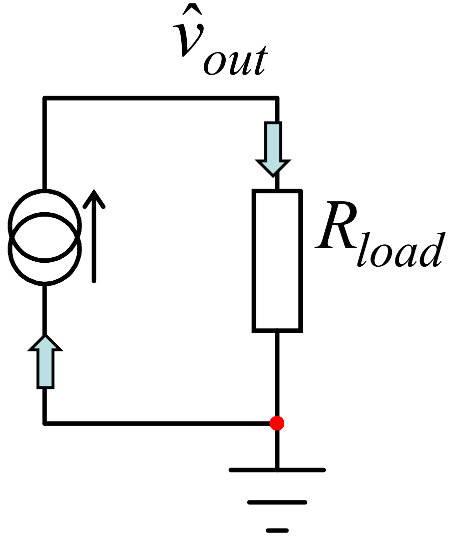
- Rework the individual coefficients to a simpler form

$$\left(\frac{\partial I_{out}\left(F_{sw},V_{out}\right)}{\partial V_{out}}\right)_{F_{sw}}\hat{v}_{out}=-\frac{F_{sw}L_p}{2V_{out}^{2}}\left(\frac{V_c}{R_i}\right)^{2}\hat{v}_{out}\;\Rightarrow\; g_m\cdot\hat{v}_{out}=\frac{1}{R}\hat{v}_{out}$$

$$g_m \rightarrow \left[\Omega^{-1}\right]$$

$$-\frac{F_{sw}L_p}{2V_{out}^{2}}\left(\frac{V_c}{R_i}\right)^{2}=-\frac{\frac{1}{2}I_{peak}^{2}F_{sw}L_p}{V_{out}^{2}}$$

$$-\frac{P_{out}}{V_{out}^{2}}=-\frac{V_{out}I_{out}}{V_{out}V_{out}}=-\frac{1}{R_{load}}$$

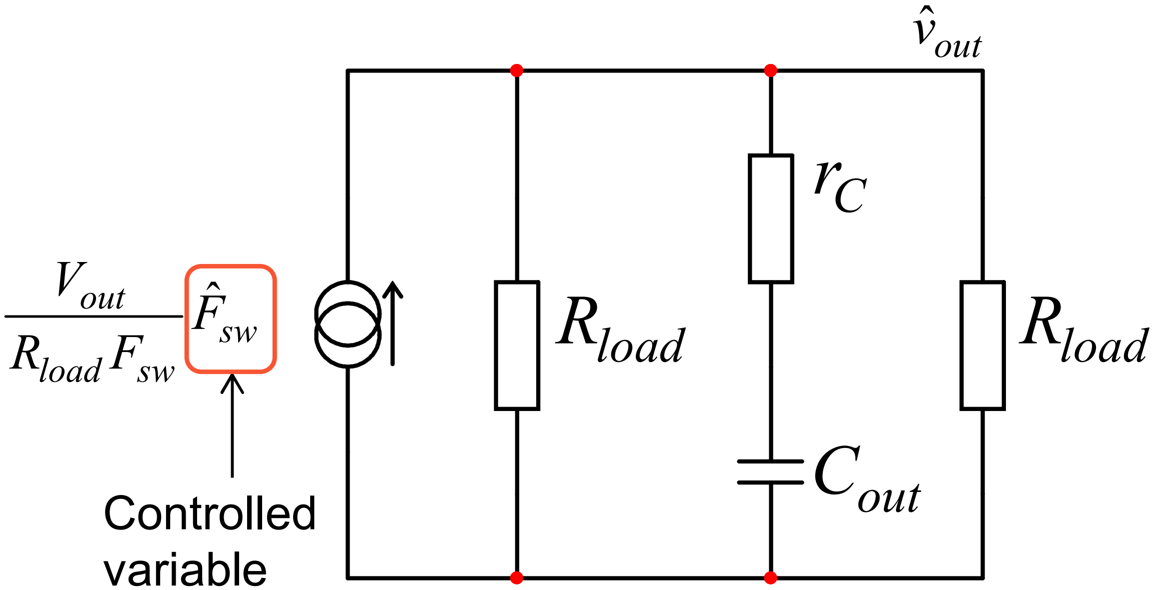


Second Expression is Also Simple

- Identify the power in the equation, simplify…

$$\left(\frac{\partial I_{out}\left(F_{sw},V_{out}\right)}{\partial F_{sw}}\right)_{V_{out}} \hat{F}_{sw} = \frac{L_p}{2V_{out}}\left(\frac{V_c}{R_i}\right)^2 \hat{F}_{sw}$$

$$\frac{L_p}{2V_{out}}\left(\frac{V_c}{R_i}\right)^2 \hat{F}_{sw} = \frac{P_{out}}{F_{sw}V_{out}}\hat{F}_{sw} = \frac{V_{out}}{R_{load}}\frac{V_{out}}{F_{sw}V_{out}}\hat{F}_{sw} = \frac{V_{out}}{R_{load}F_{sw}}\hat{F}_{sw}$$



Can't beat such a simple schematic!

The Transfer Function is Immediate

- The transfer function is now straightforward

$$\hat{v}_{out}(s) = \hat{F}_{sw}(s)\frac{V_{out}}{R_{load}F_{sw}} R_{load} \parallel Z_{eq}(s) \longrightarrow \approx \frac{R_{load}}{2}\frac{1+sr_C C}{1+sC\left(\frac{R_{load}}{2}\right)}$$

$$\hat{v}_{out}(s) = \hat{v}_c(s) k_F G_{VCO}\frac{V_{out}}{2F_{sw}}\frac{1+sr_C C}{1+sC\left(\frac{R_{load}}{2}\right)}$$

- Put it under the normalized form

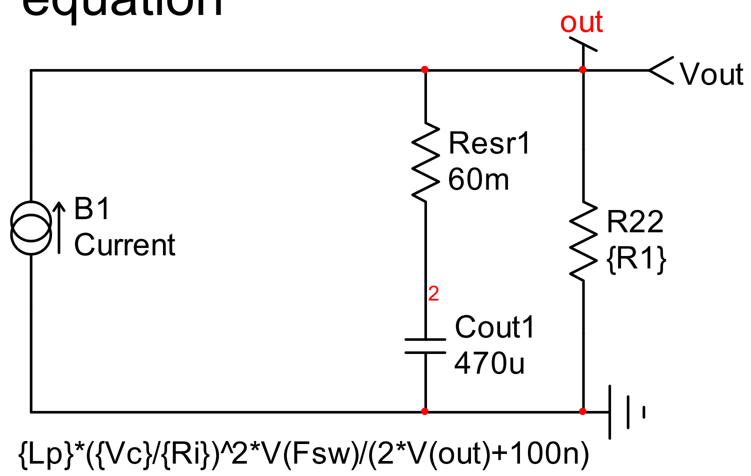
$$H(s) = \frac{\hat{v}_{out}}{\hat{v}_c} = H_0\frac{1+s/\omega_z}{1+s/\omega_p} = \frac{V_{out}}{2F_{sw}} k_F G_{VCO}\frac{1+sr_C C_{out}}{1+s\frac{R_{load}}{2}C_{out}}$$

- No RHPZ and no high-frequency pole prediction
- Good for low-frequency analysis only, but good enough for us!

Simulation Shows the Differences

- SPICE only manipulates linear equations

Large-signal equation

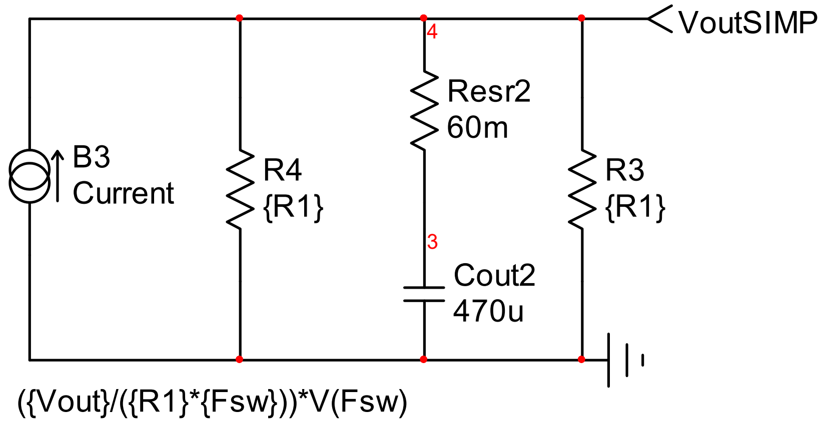


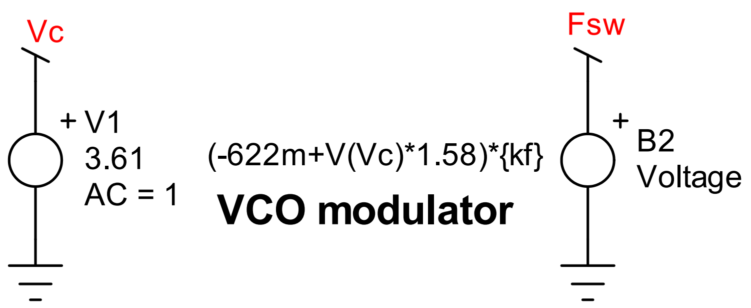
parameters

Vc=1
Lp=600u
Ri=0.8
Lp=600u
kf=10k

Vout=21.1
P1=24.73
R1=Vout^2/P1
Fsw=50.8k

Small-signal equation



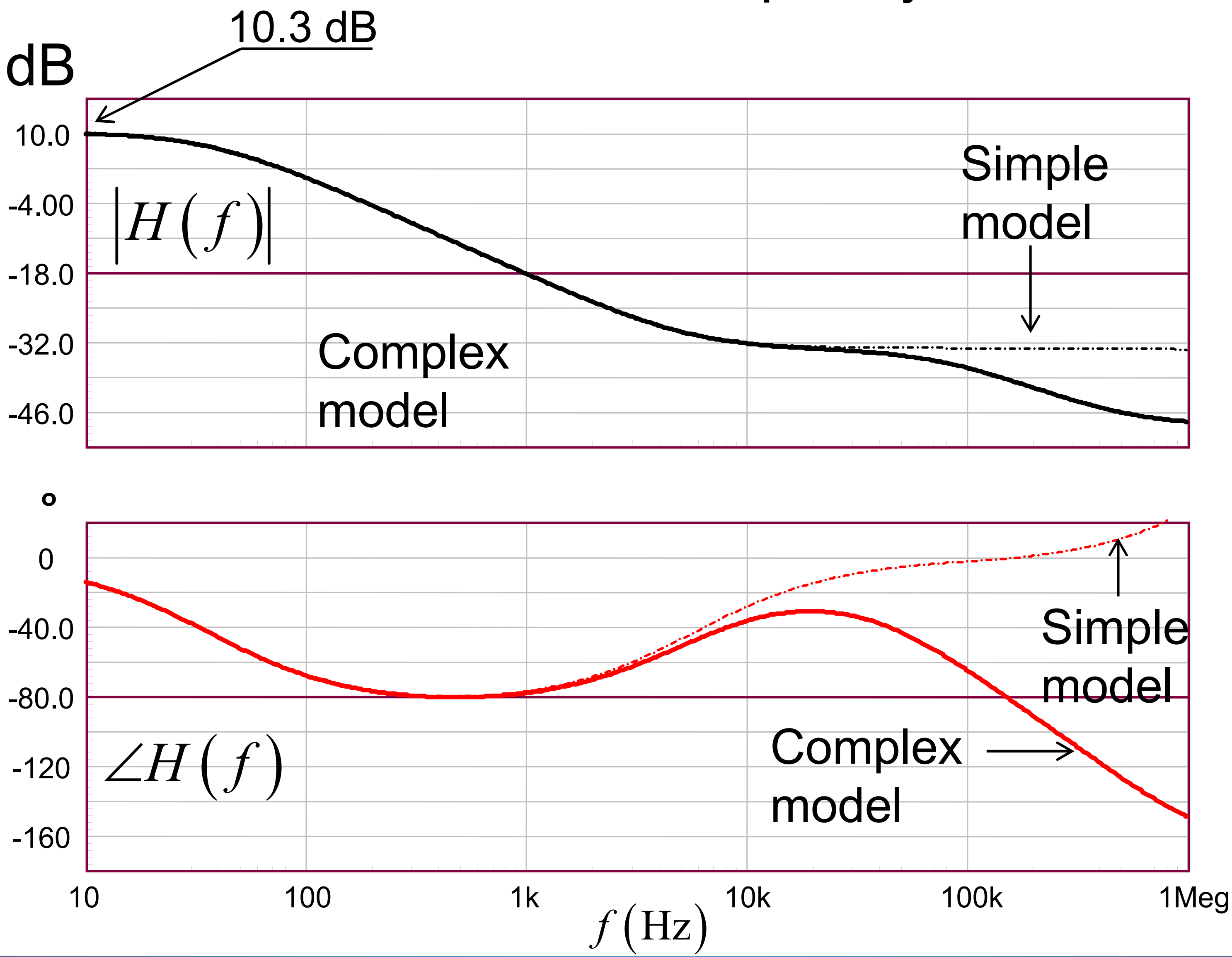


VCO modulator

$$20\log_{10}\left|H_0\right| = 20\log_{10}\left(\frac{V_{out}}{2F_{sw}}k_F G_{VCO}\right) = 20\log_{10}\left(\frac{21.1}{2\times 50.8k}10k\cdot 1.58\right) = 10.3\,\text{dB}$$

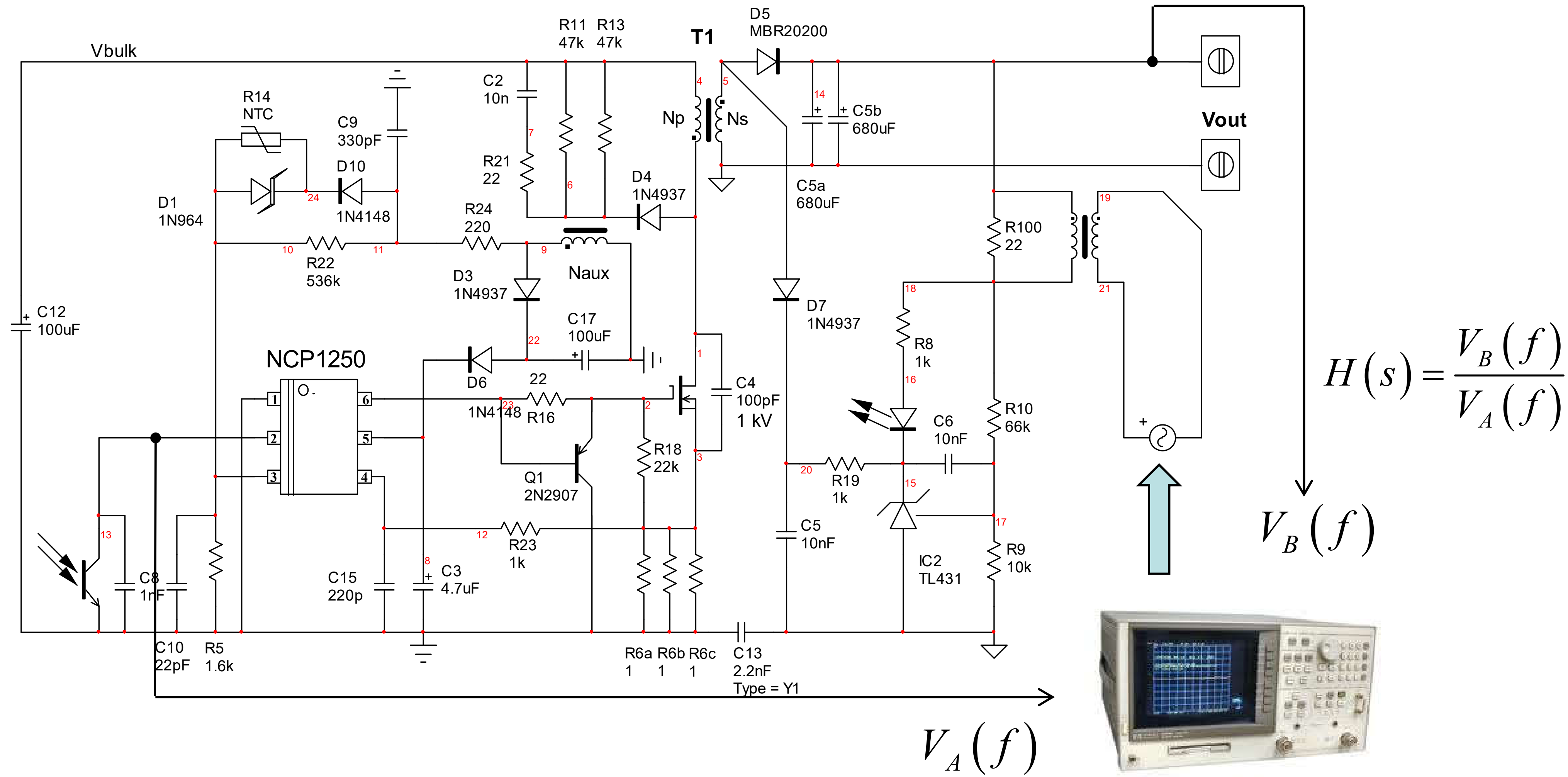
Simulation Shows the Differences

- ❑ Ac results are identical at low frequency



Test Fixture

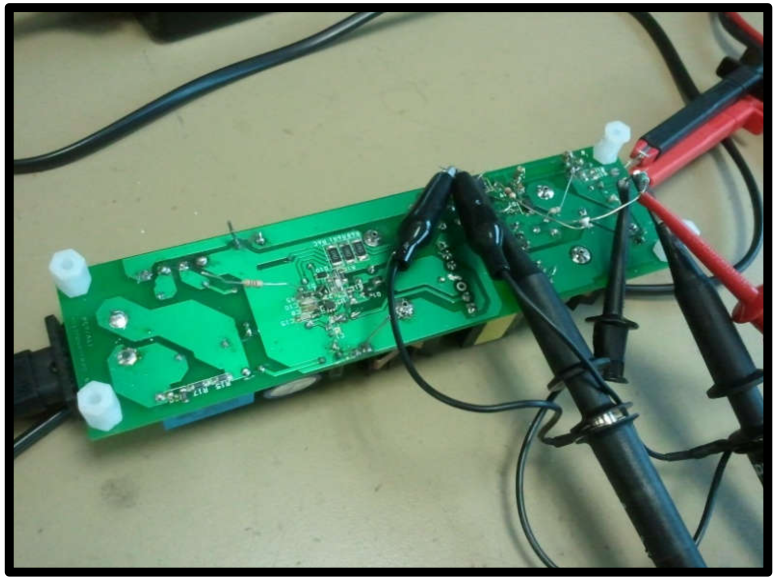
- ❑ Plant frequency response with a NCP1250
- ❑ Load is reduced to force DCM where I_{peak} is frozen



Connections

- ❑ Connect probe grounds to a quiet point: opto emitter is ok
- ❑ Use an isolated current-limited dc-supply
- ❑ Short primary and secondary grounds

100 V dc

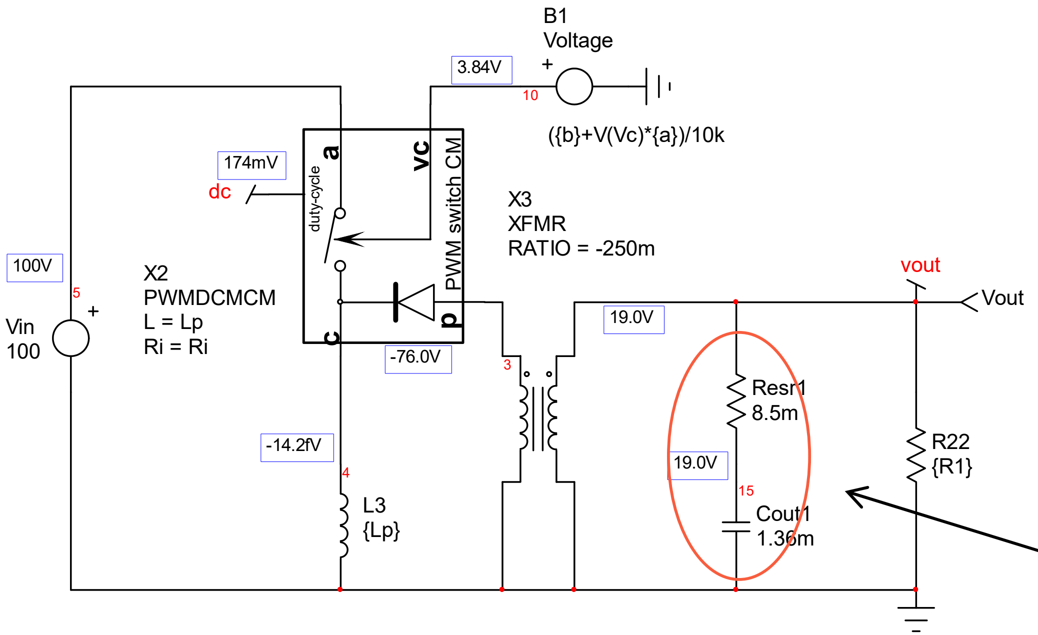


Power supply under test

$H(f)$

Injection transformer

Run the SPICE Simulation



- Check operating points:

$$F_{sw} = 38.6\,\text{kHz}(\text{bench})\ 38.4\,\text{kHz}(\text{SPICE})$$

$$V_{FB} = 0.71\,\text{V}(\text{bench})\,0.79\,\text{V}(\text{SPICE})$$

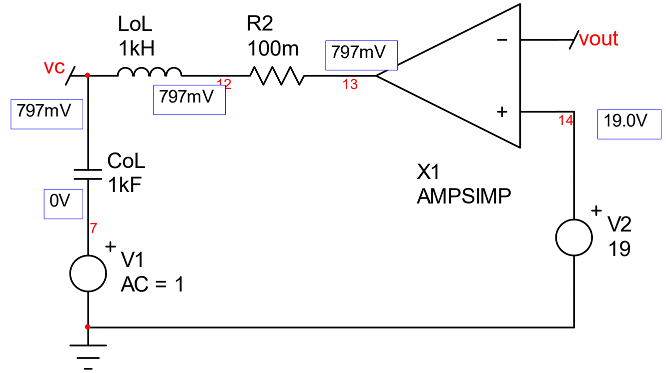
Main contributor to errors is the capacitor ESR

parameters

Vout=19
P1=6.6
R1=Vout^2/P1

Lp=600u
Ri=0.33

Fmax=65k
Fmin=26k
Vfold=1.5
Vmin=0.35
a2=(Fmax-Fmin)/(Vfold-Vmin)
a=27.75k
b=Fmin-Vmin*a

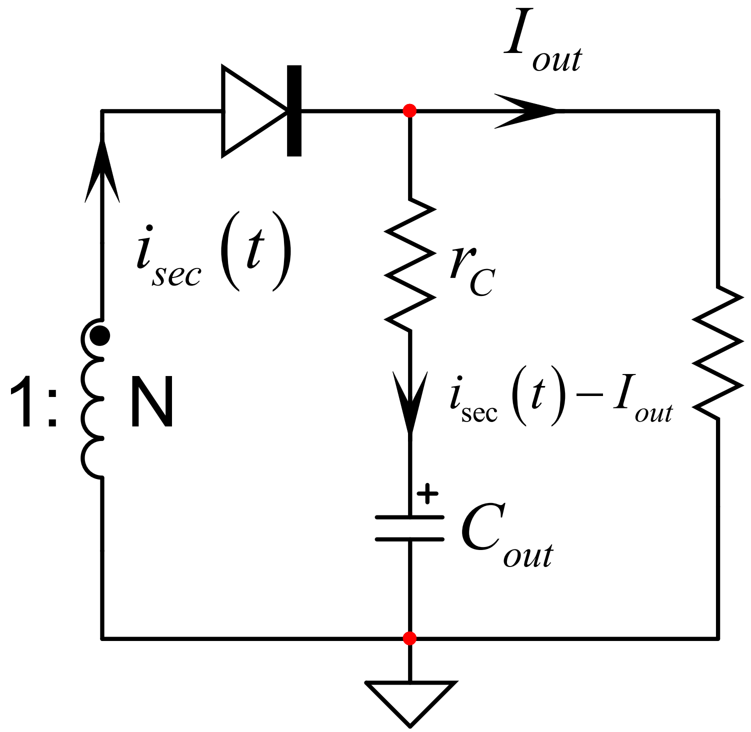


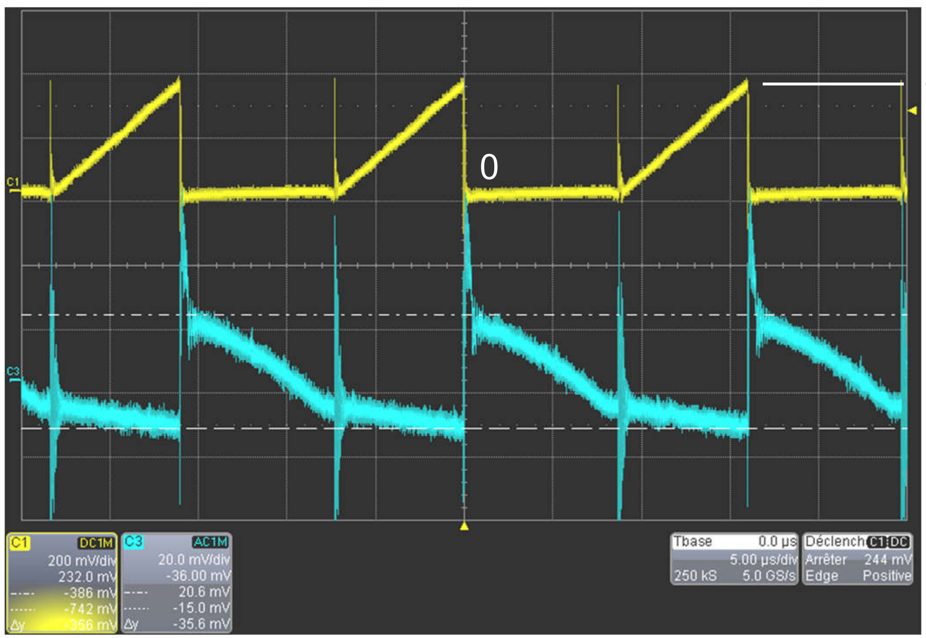
VCO slope measured to 27.8 kHz/V

ZL series Rubycon

Approximate ESR Extraction

- A triangular output shows that ESR is the main contributor





$I_{p,peak} = 1\,\text{A}$

$v_{out}(t)$

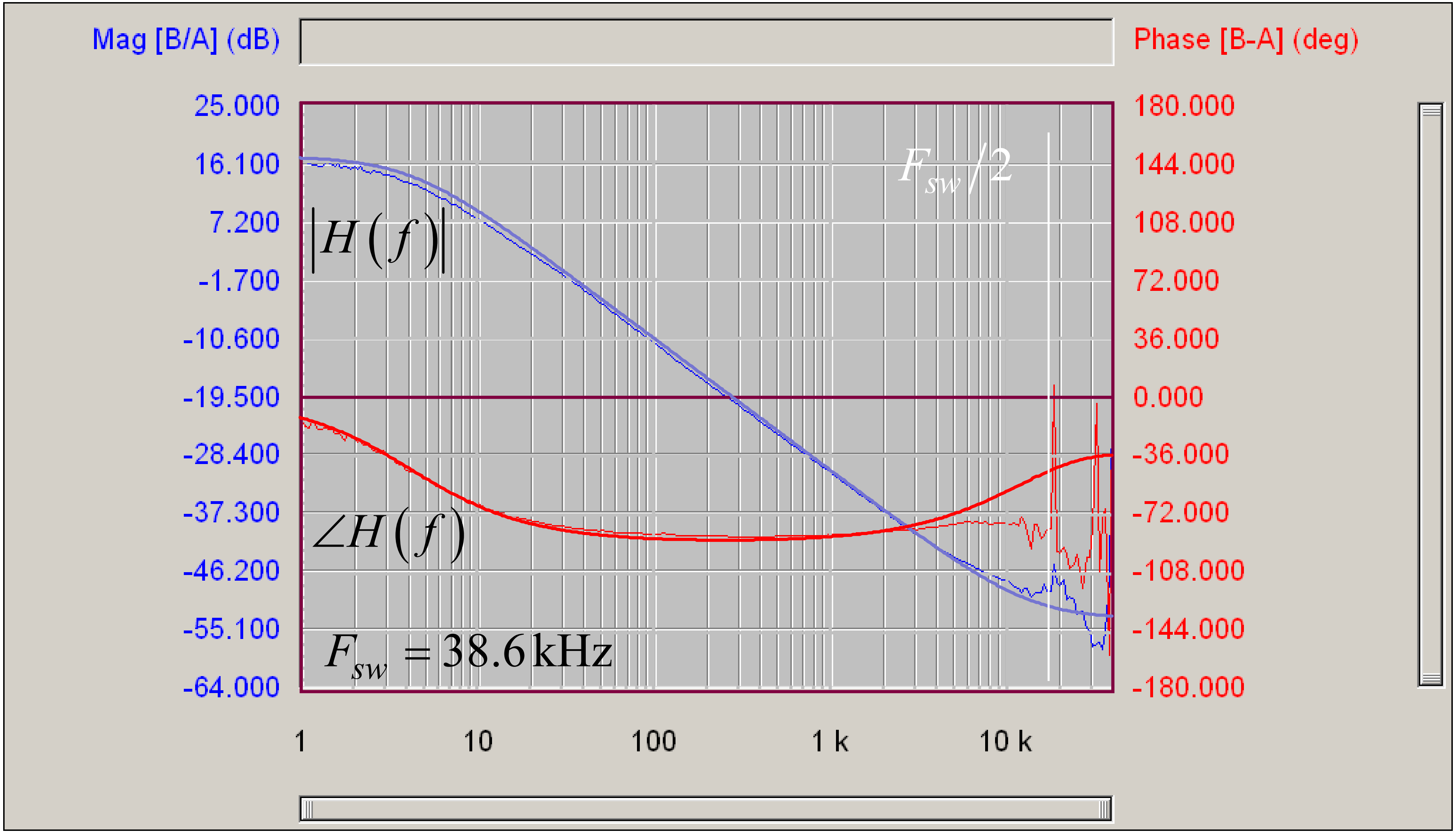
$$I_{sec,peak} = \frac{I_{p,peak}}{N} = \frac{1}{0.25} = 4\,\text{A}$$

$$V_{out,peak} \approx I_{sec,peak} r_C$$

$$r_C \approx \frac{V_{out,pp}}{I_{sec,peak}} = \frac{35m}{4} = 8.75\,\text{m}\Omega$$

DS says 11.5 mΩ

Plant Frequency Response

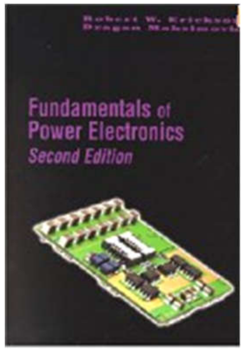


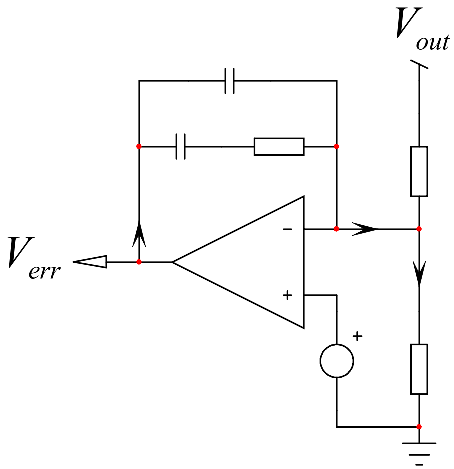
Agenda

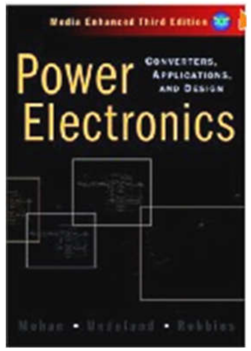
- ❑ What is a QR Converter?
- ❑ Deriving Operating Conditions
- ❑ The Over Power Problem
- ❑ Three Operating Modes
- ❑ Small-Signal Analysis – QR
- ❑ Small-Signal Analysis – VCO
- ❑ **Compensating with TL431**
- ❑ Control Loop Design Example
- ❑ Conclusion

How is regulation performed?

- Text books only describe op amps in compensators…



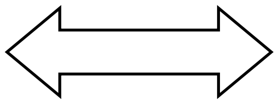


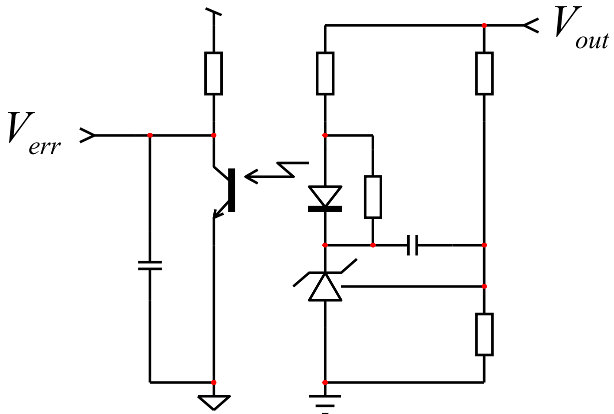


- The market reality is different: the TL431 rules!

TL431

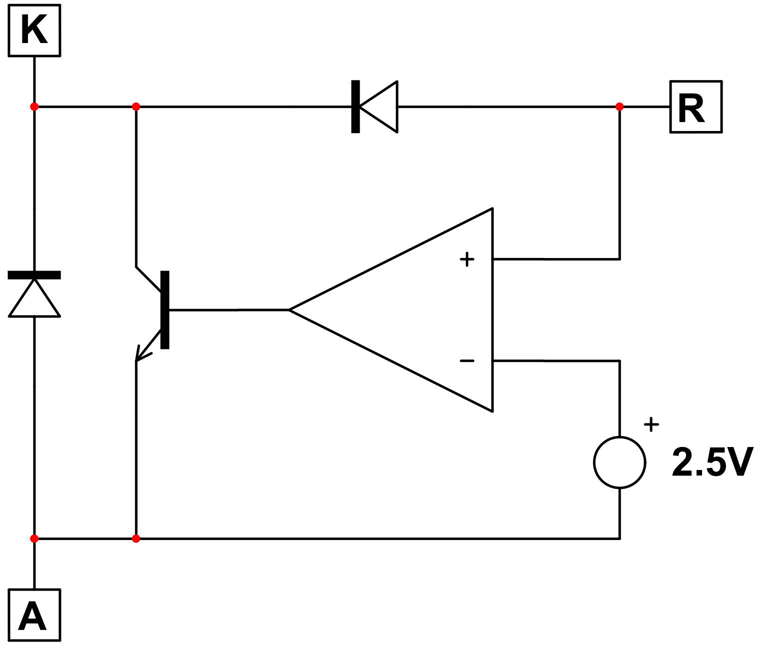
optocoupler

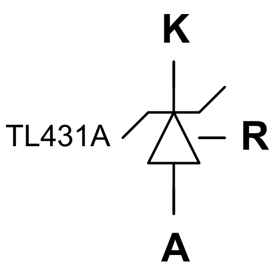


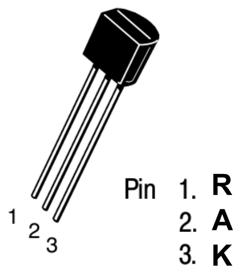


The TL431 Programmable Zener

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



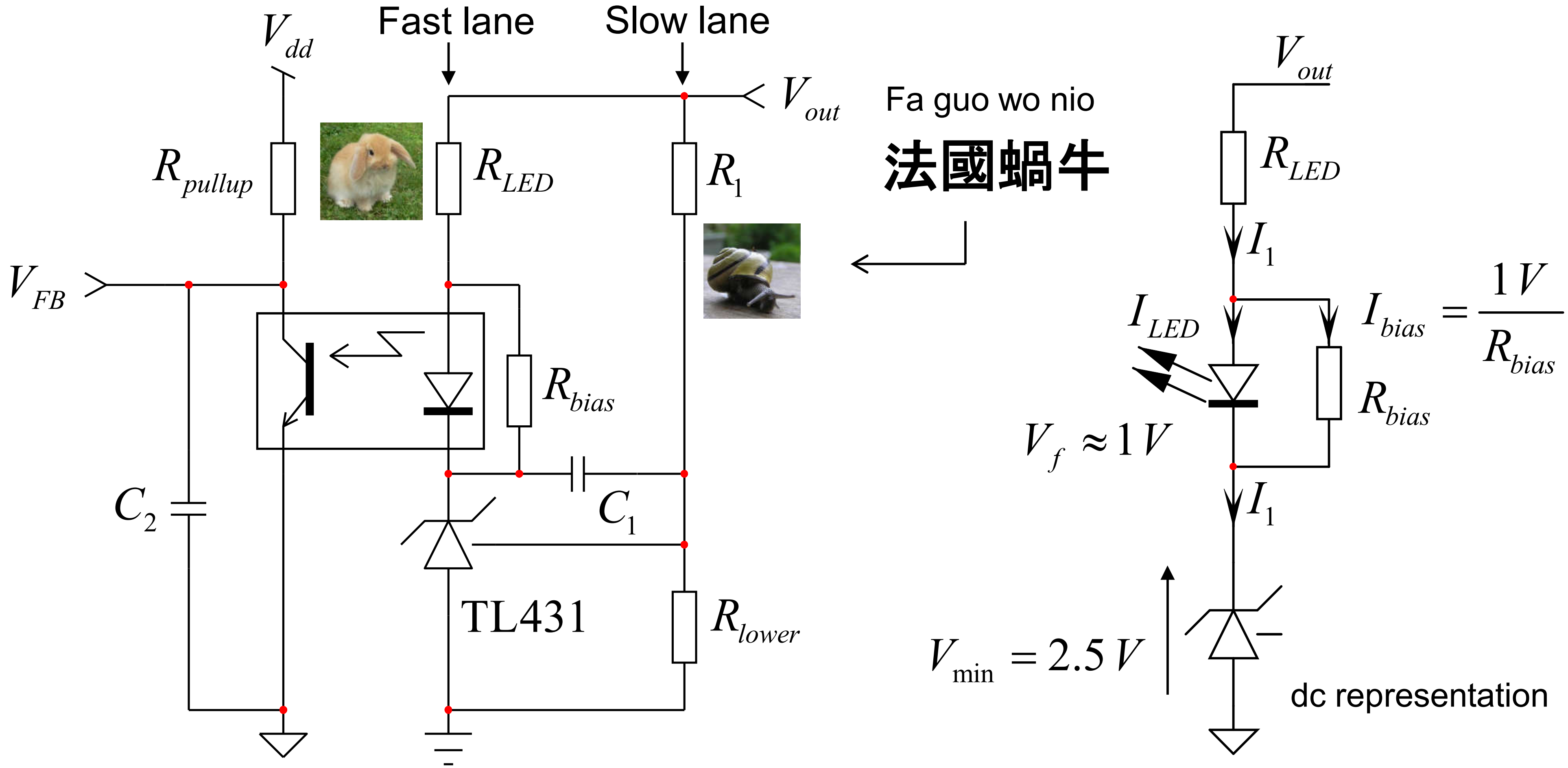




- The TL431 is a shunt regulator

A Rabbit and a (French) Snail…

- The TL431 lends itself very well to optocoupler control



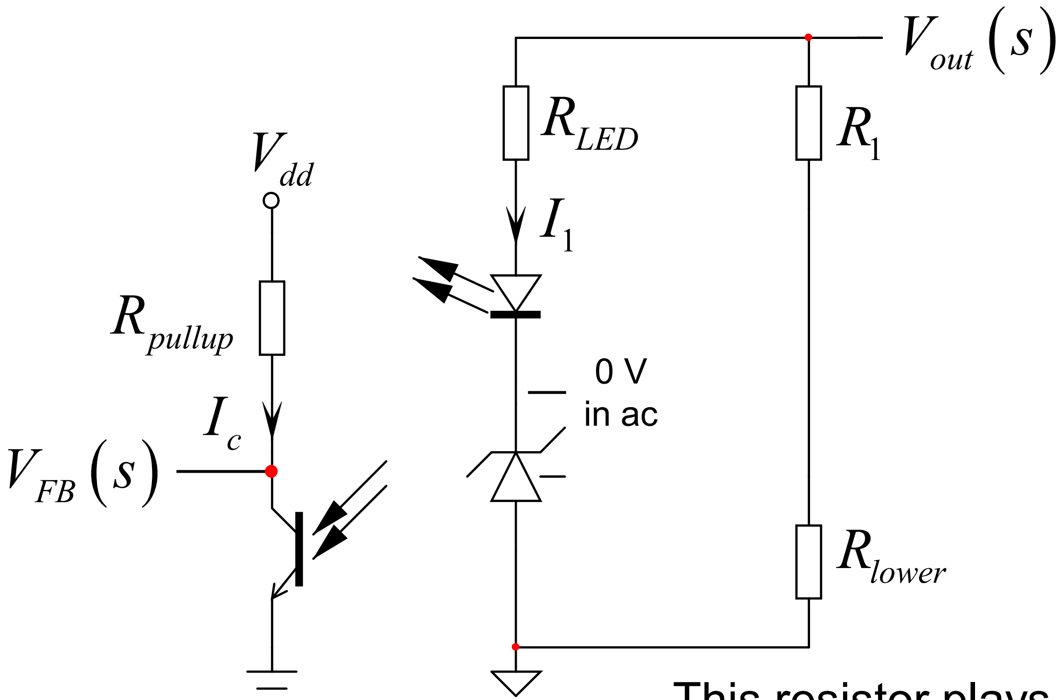
- R_{LED} must leave enough headroom over the TL431: upper limit!

Understanding the Fast Lane Drawback

- This LED resistor is a design limiting factor in low output voltages:

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

- When the capacitor C_1 is a short-circuit, R_{LED} fixes the fast lane gain



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

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

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

This resistor plays a role in dc too!

The Static Gain Limit

- Let us assume the following design:

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

$$R_{LED,\max} \leq \frac{5-1-2.5}{4.8-0.3+1m\times0.3\times20k}\times20\text{k}\times0.3$$

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

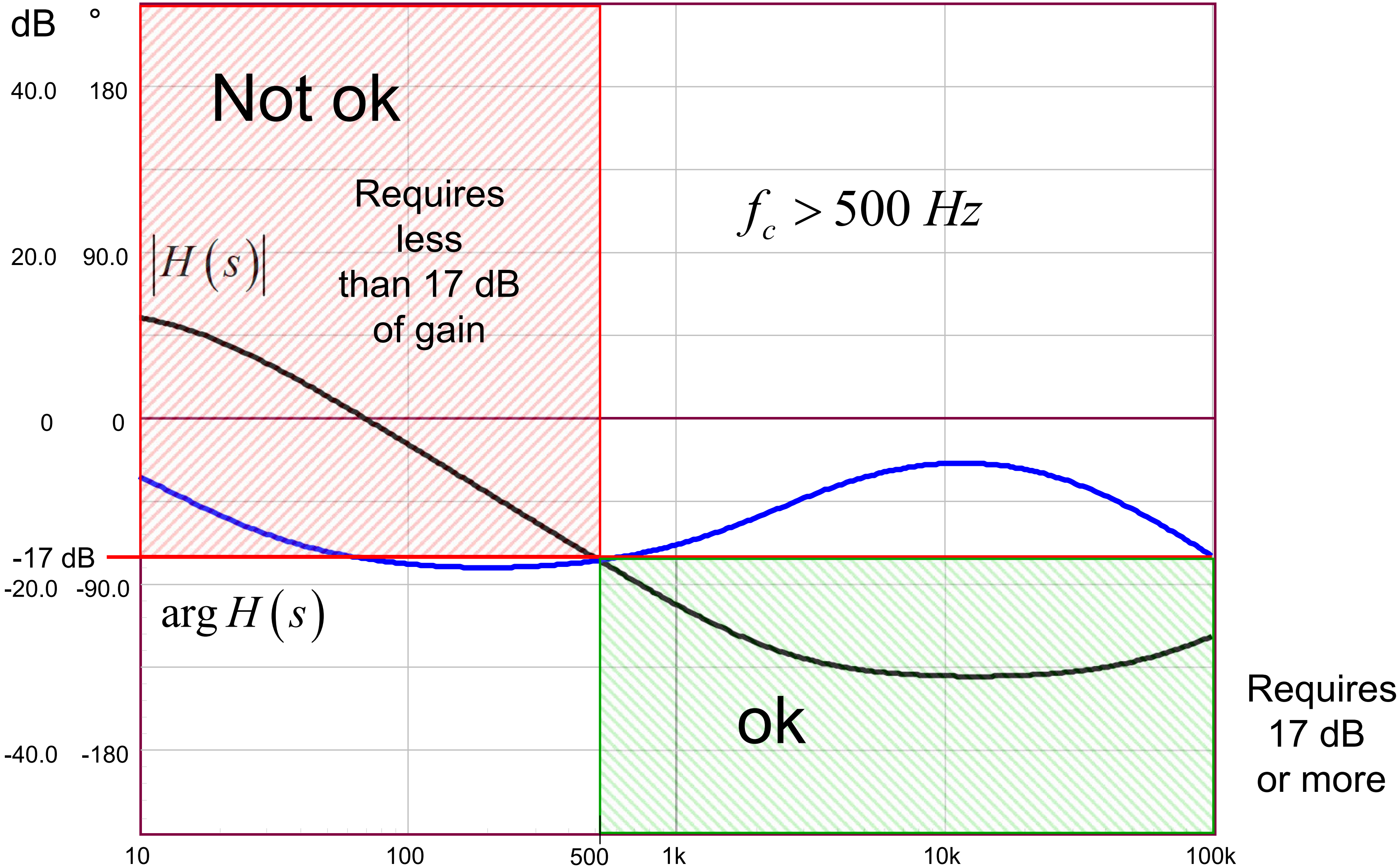
$$G_0 > \text{CTR}\frac{R_{pullup}}{R_{LED}} > 0.3\frac{20}{0.857} > 7 \text{ or} \approx 17\text{ dB}$$

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

⇨ You cannot “amplify” by less than 17 dB

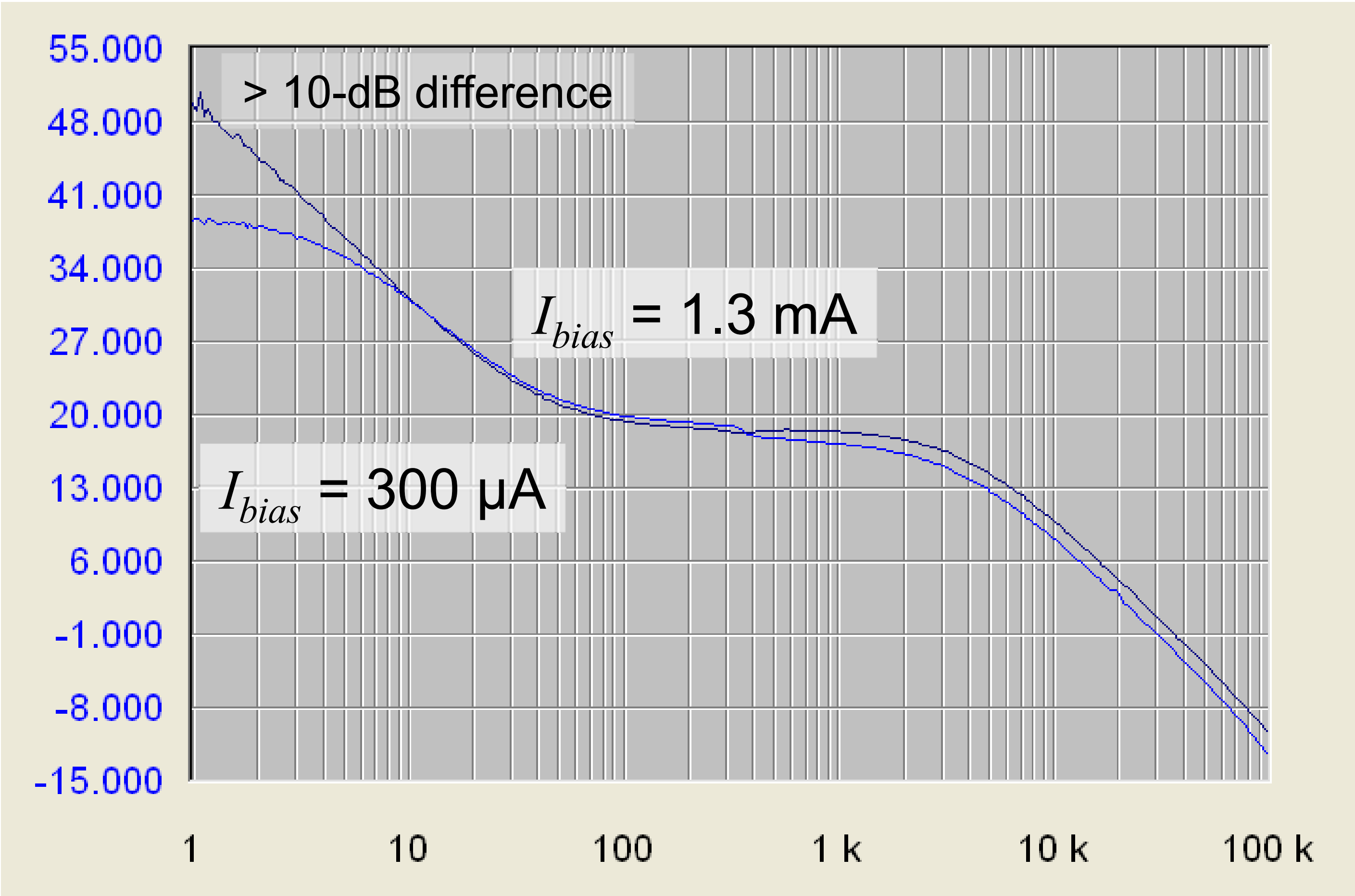
Forbidden Compensation Areas

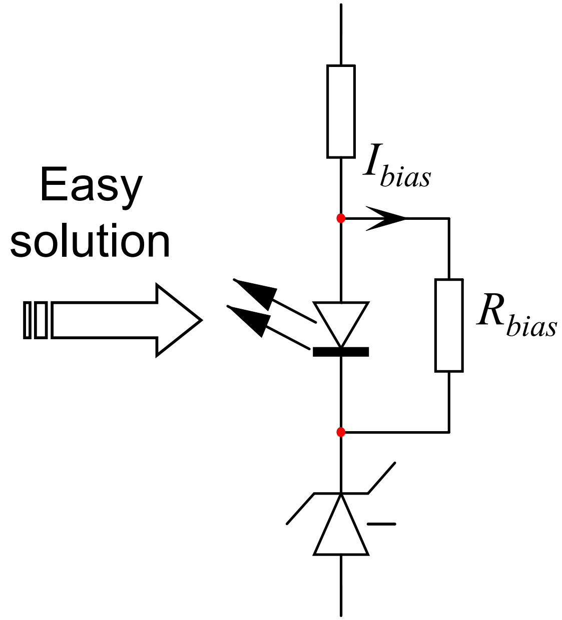
- You must identify the areas where compensation is possible



Injecting Bias Current

- Make sure enough current always biases the TL431
- If not, its open-loop suffers – a 10-dB difference can be observed!

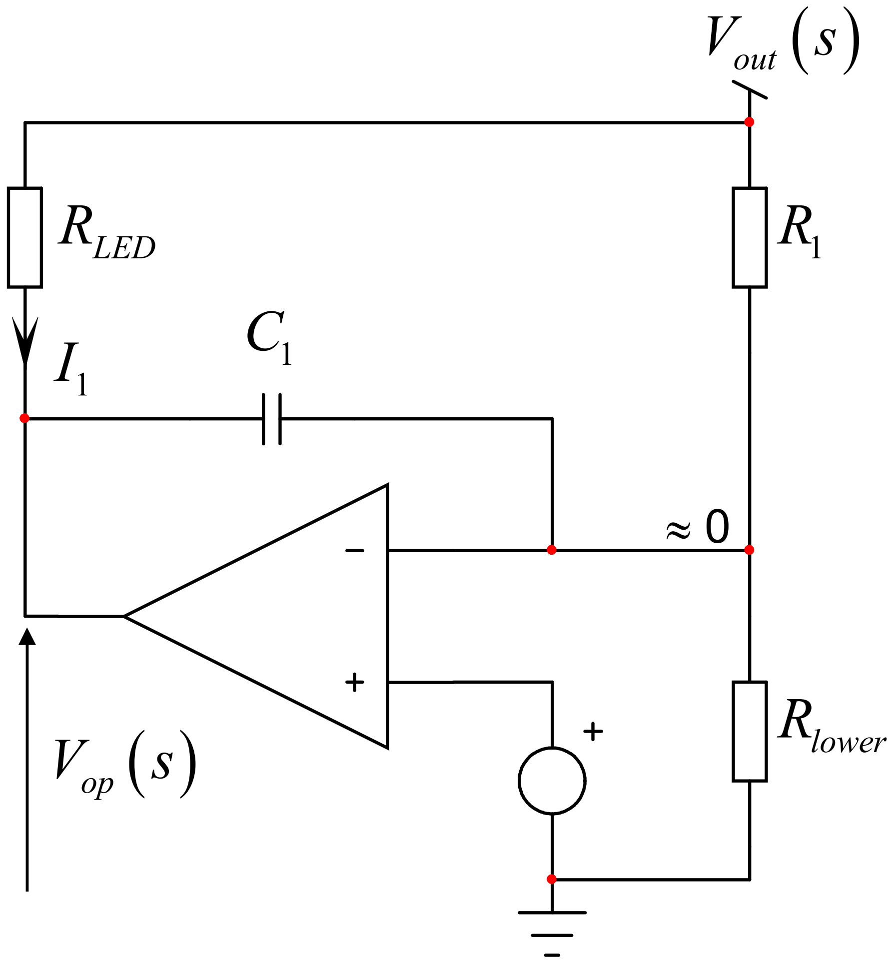




$$R_{bias} = \frac{1}{1m} = 1\,k\Omega$$

Small-Signal Analysis

- The TL431 is an open-collector op amp with a reference voltage
- Neglecting the LED dynamic resistance, we have:



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

$$V_{op}(s) = -V_{out}(s)\frac{\frac{1}{sC_1}}{R_{upper}} = -V_{out}(s)\frac{1}{sR_{upper}C_1}$$

$$I_1(s) = V_{out}(s)\frac{1}{R_{LED}}\left[1 + \frac{1}{sR_{upper}C_1}\right]$$

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

$$\frac{V_{FB}(s)}{V_{out}(s)} = -\frac{R_{pullup}\mathrm{CTR}}{R_{LED}}\left[\frac{1 + sR_{upper}C_1}{sR_{upper}C_1}\right]$$

Creating a High-Frequency Pole

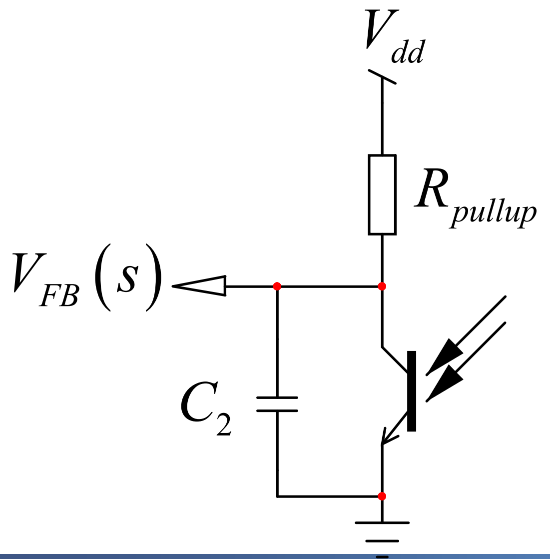
- ❑ In the previous equation we have:

✓ a static gain $G_0 = \mathrm{CTR}\dfrac{R_{pullup}}{R_{LED}}$

✓ a 0-dB origin pole frequency $\omega_{po} = \dfrac{1}{C_1 R_{upper}}$

✓ a zero $\omega_{z_1} = \dfrac{1}{R_{upper} C_1}$

- ❑ We are missing a pole for the type 2!



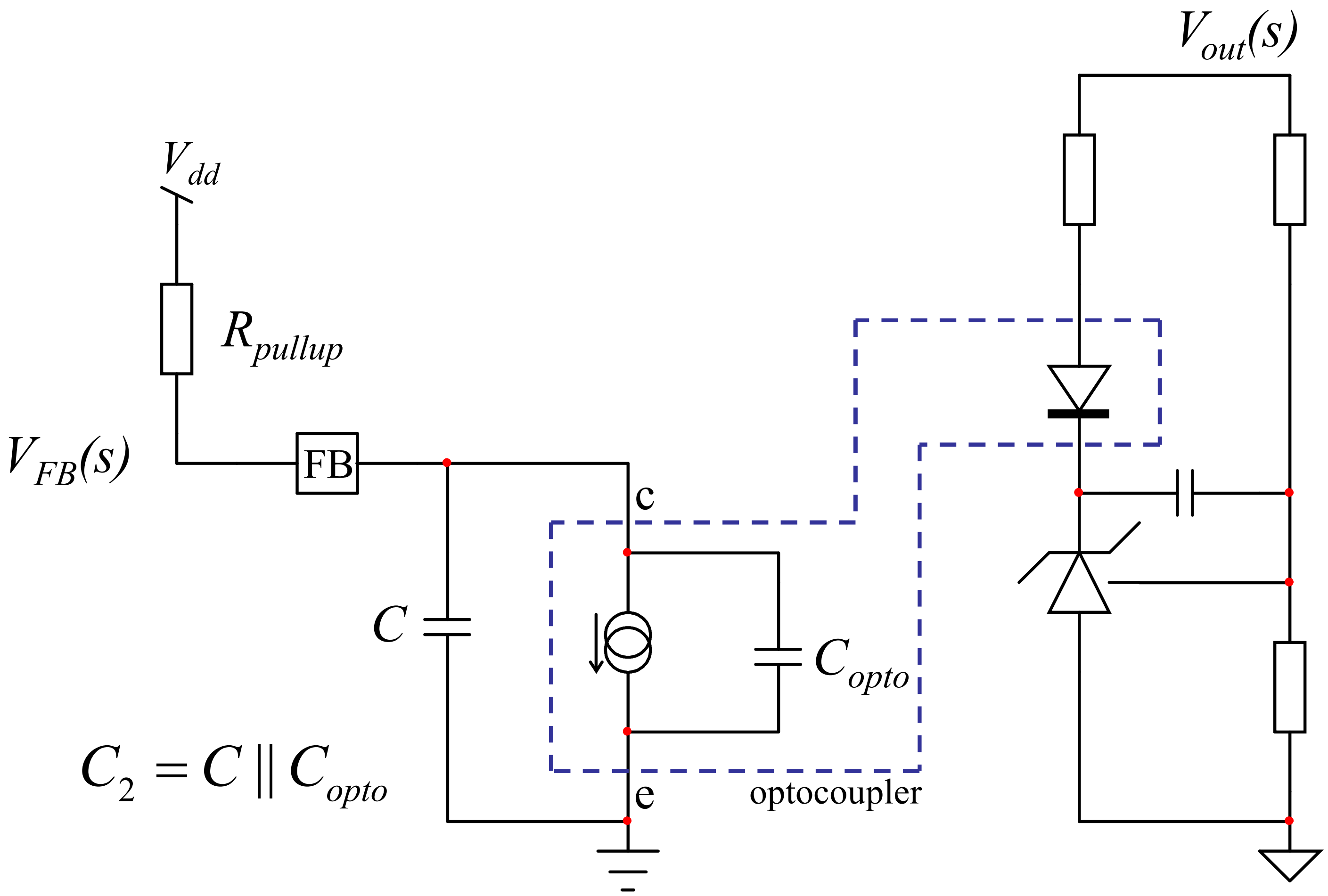
Add a cap. from collector to ground

Type 2 transfer function

$$\frac{V_{FB}(s)}{V_{out}(s)} = -\frac{R_{pullup}\mathrm{CTR}}{R_{LED}}\left[\frac{1+sR_{upper}C_1}{sR_{upper}C_1\left(1+sR_{pullup}C_2\right)}\right]$$

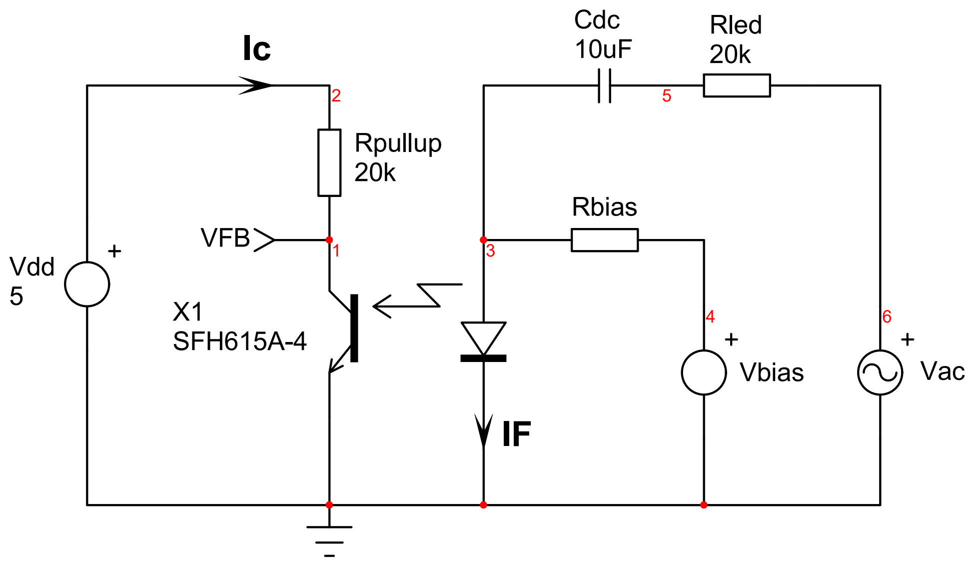
Understanding the Optocoupler Pole

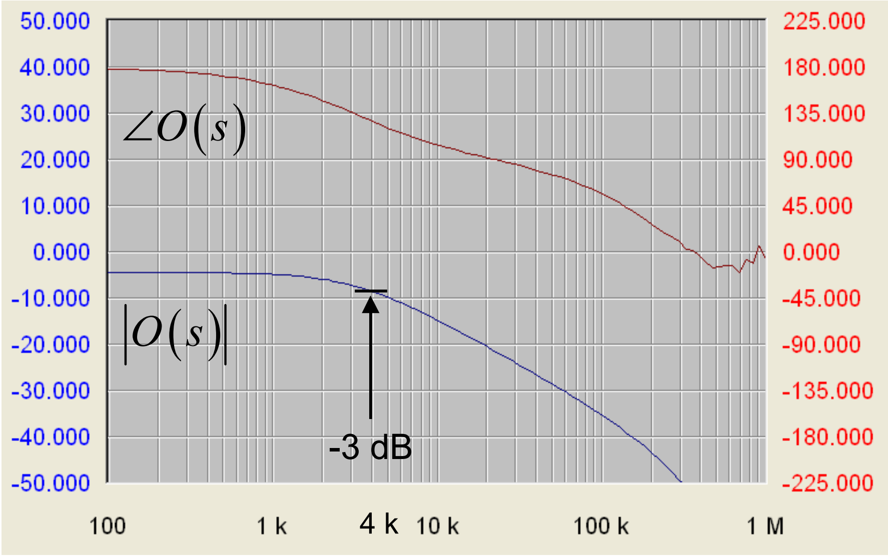
- ❑ The optocoupler also features a parasitic capacitor
- ➢ it comes in parallel with C_2 and must be accounted for



Extracting the Pole

- ❑ The optocoupler must be characterized to know where its pole is





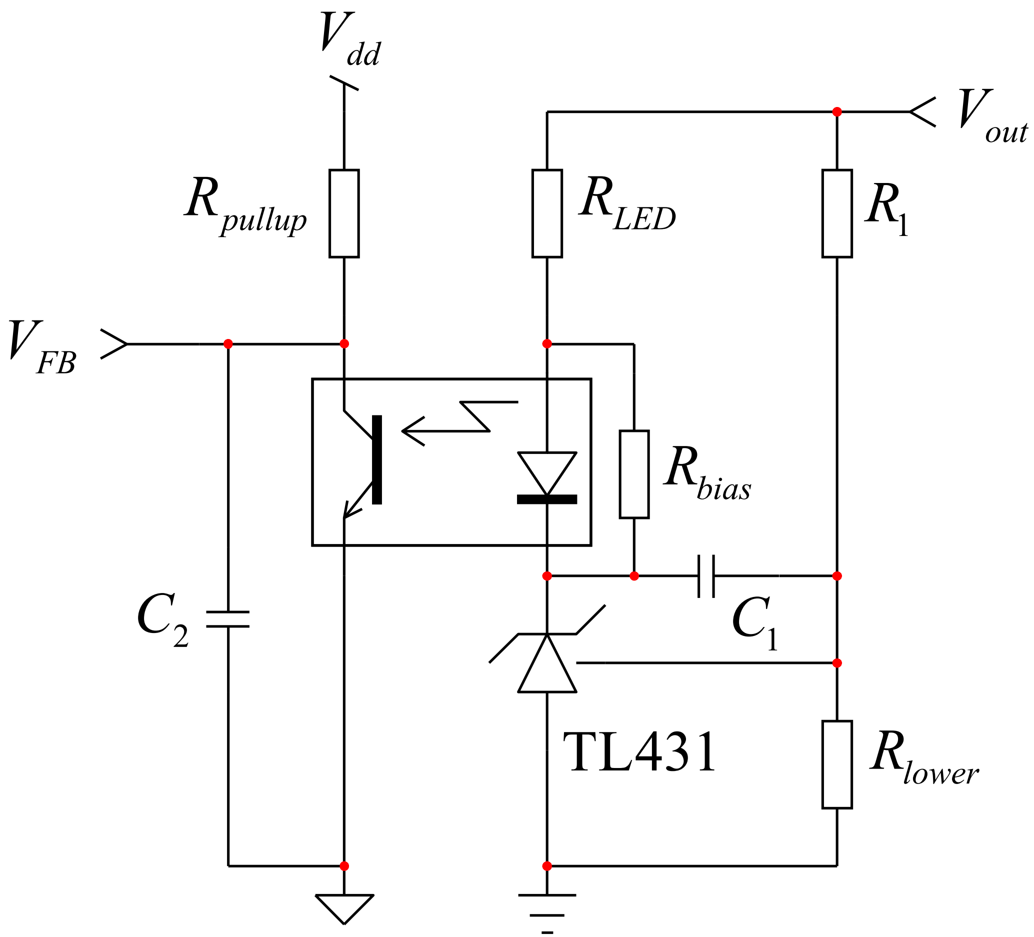
- ❑ Adjust V_{bias} to have V_{FB} at 2-3 V to be in linear region, then ac sweep
- ❑ The pole in this example is found at 4 kHz

$$C_{opto} = \frac{1}{2\pi R_{pullup} f_{pole}} = \frac{1}{6.28 \times 20k \times 4k} \approx 2\ \text{nF}$$

⇨ Another design constraint!

Most Popular Compensator is Type 2

- Our first equation was already a type 2 definition, we are all set!



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

$$\omega_{z_1} = \frac{1}{R_1 C_1}$$

$$\omega_{p_1} = \frac{1}{R_{pullup} C_2}$$

- Just make sure the optocoupler contribution is involved…

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- ❑ **Control Loop Design Example**
- ❑ Conclusion

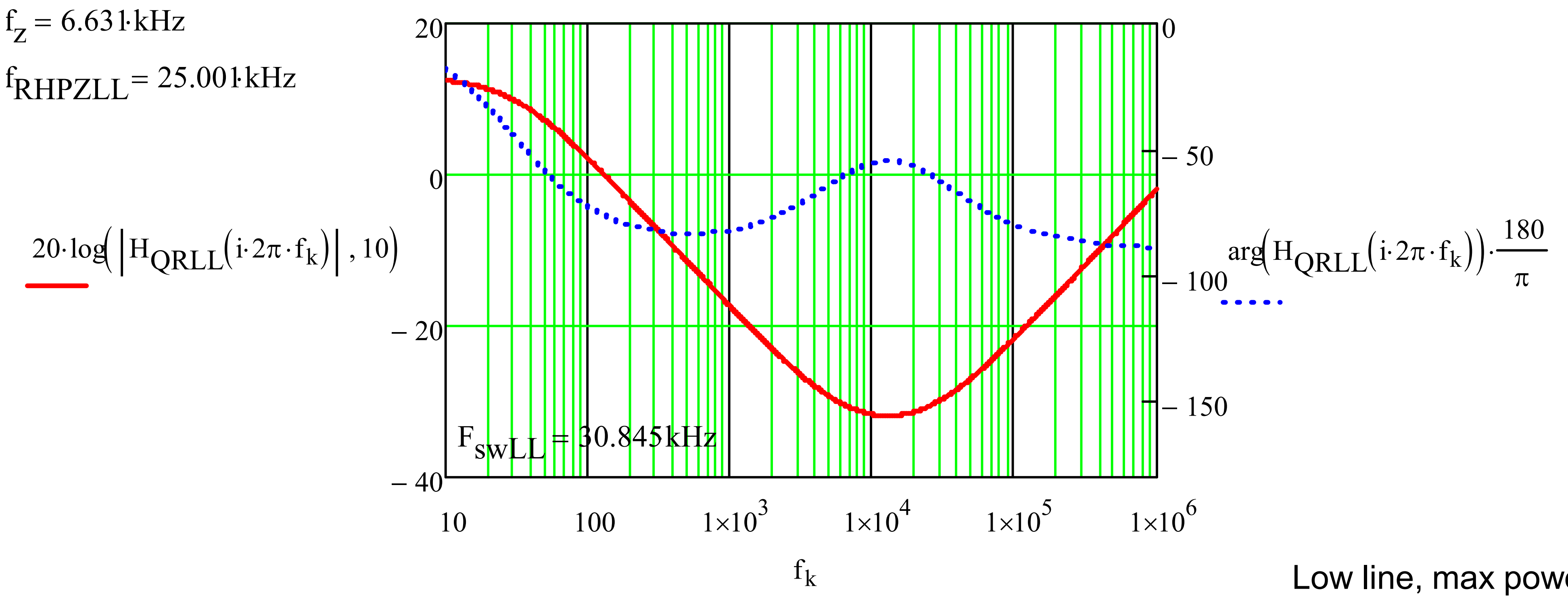
Stabilizing a QR Converter

- Assume a 60-W QR converter with the following values

$$V_{in,\min} = 120\,\text{V} \;\; N = 0.25 \;\; R_{sense} = 0.3\,\Omega \;\; C_{out} = 1.2\,\text{mF}$$

$$V_{out} = 19\,\text{V} \;\; L_p = 600\,\mu\text{H} \;\; R_{esr} = 20\,\text{m}\Omega \;\; P_{out} = 60\,\text{W}$$

- Plot the ac transfer function, select crossover (1 kHz)



Low line, max power

Tailor the Compensator

- Extract attenuation and phase lag at 1 kHz from the graph

$$\left|H\left(1\text{kHz}\right)\right| = -17.4\,\text{dB} \qquad \arg H\left(1\text{kHz}\right) = -82°$$

- Shoot for a phase margin of 70°
- Lift the gain curve by 17.4 dB at 1 kHz, provide a boost of

$$\text{Boost} = \text{PM} - \text{argH}\left(f_c\right) - 90 = 70 + 82 - 90 = 62°$$

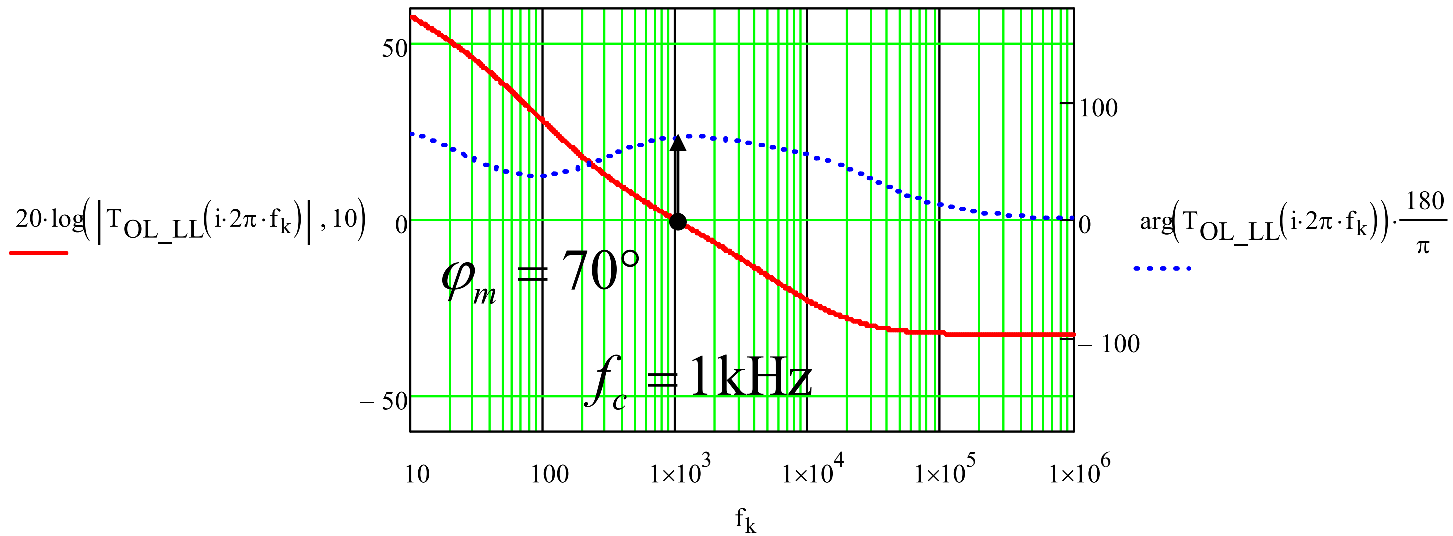
- Position the pole and the zero to boost the phase by 62° at 1 kHz

$$k = \tan\left(\frac{\text{boost}}{2} + 45°\right) = 4 \longrightarrow f_z = \frac{f_c}{k} = 249\,\text{Hz} \quad f_p = k \cdot f_c = 4\,\text{kHz}$$

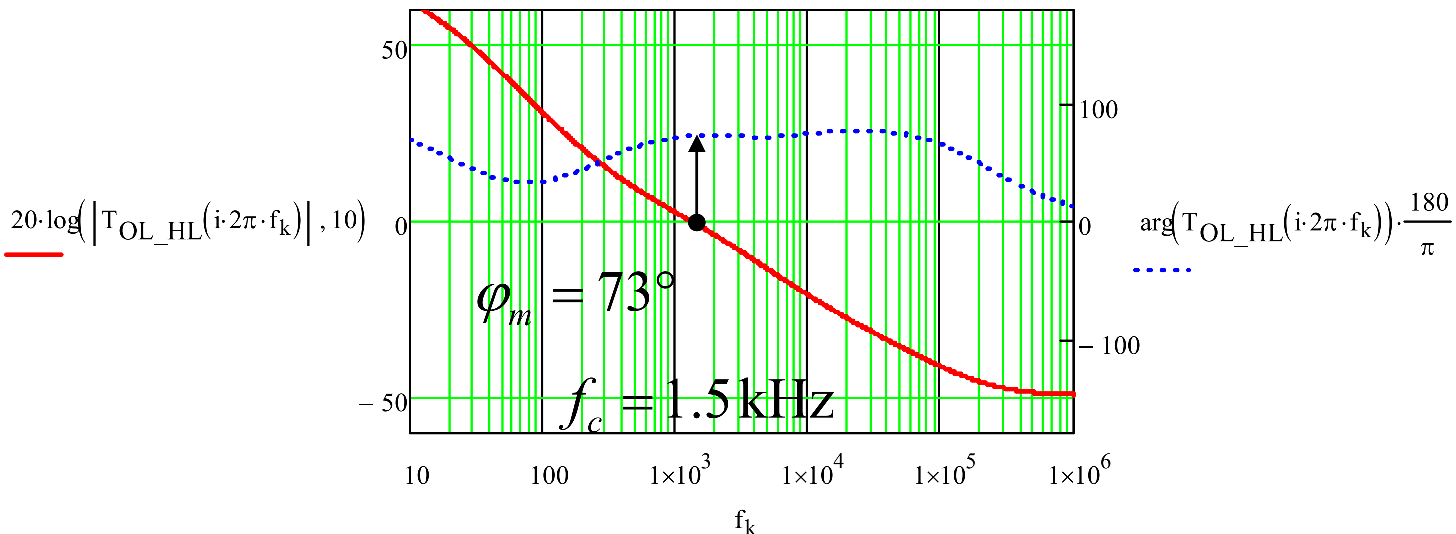
- Apply this compensation parameters to the plant transfer function

Check Final Bode Plot

- Plot the loop gain and check margins $T(s) = H(s)G(s)$



Low line



High line

Deriving Component Values for the Type 2

- You need to provide a 17.4-dB gain at 1 kHz with a 62° boost

$$f_p = 4\text{ kHz} \qquad f_z = 249\text{ Hz} \qquad G_0 = \text{CTR}\frac{R_{pullup}}{R_{LED}} = 10^{17.4/20} = 7.4$$

- With a 250-µA bridge current, the divider resistor is made of:

$$R_{lower} = 2.5/250u = 10\text{ k}\Omega \qquad R_1 = (19 - 2.5)/250u = 66\text{ k}\Omega$$

- The pole and zero respectively depend on R_{pullup} and R_1:

$$R_{pullup} \rightarrow 20\text{ k}\Omega$$

$$C_2 = 1/2\pi f_p R_{pullup} \approx 2\text{ nF} \qquad C_1 = 1/2\pi f_z R_1 \approx 10\text{ nF}$$

- The LED resistor depends on the needed mid-band gain:

$$R_{LED} = \frac{R_{pullup}\text{CTR}}{G_0} = 810\ \Omega \xrightarrow{\text{ok}} R_{LED,\max} \le 8.85\text{ k}\Omega$$

- The optocoupler is at a 4-kHz frequency. Its capacitance is 2 nF
- As it comes in // with C_2, add a 100-pF cap. between FB and GND

Test the Computed Values with SPICE

❑ This fixture gives the type 2 response with a TL431

parameters

Vout=19
Rupper=(Vout-2.5)/250u
fc=1k
pm=70
Gfc=-17.4
pfc=-82

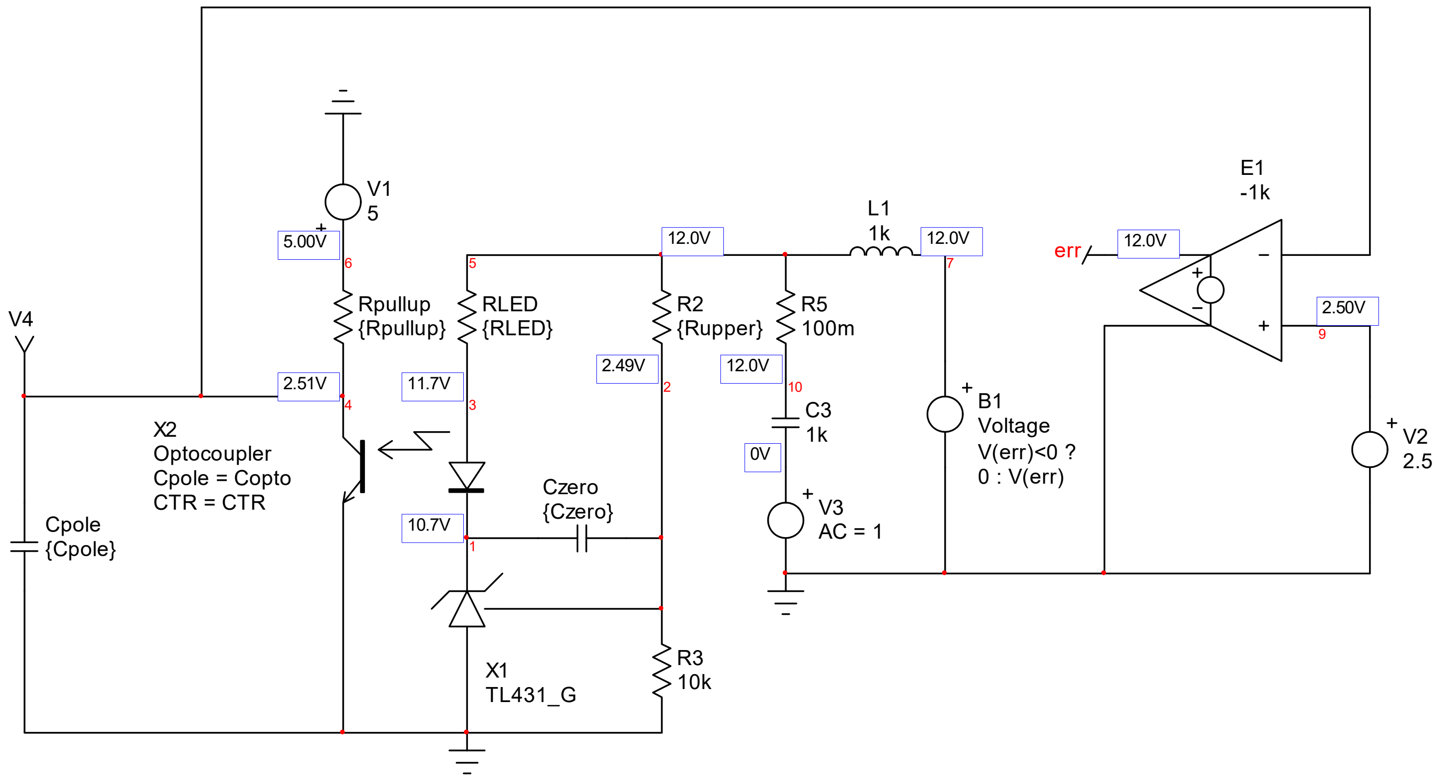
G=10^(-Gfc/20)
boost=pm-(pfc)-90
pi=3.14159
K=tan((boost/2+45)*pi/180)
C2=1/(2*pi*fc*G*k*Rupper)
C1=C2*(K^2-1)
R2=k/(2*pi*fc*C1)

Fzero=fc/k
Fpole=k*fc

Rpullup=20k
RLED=CTR*Rpullup/G
Czero=1/(2*pi*Fzero*Rupper)
Cpole1=1/(2*pi*Fpole*Rpullup)

Cpole=Cpole1-Copto

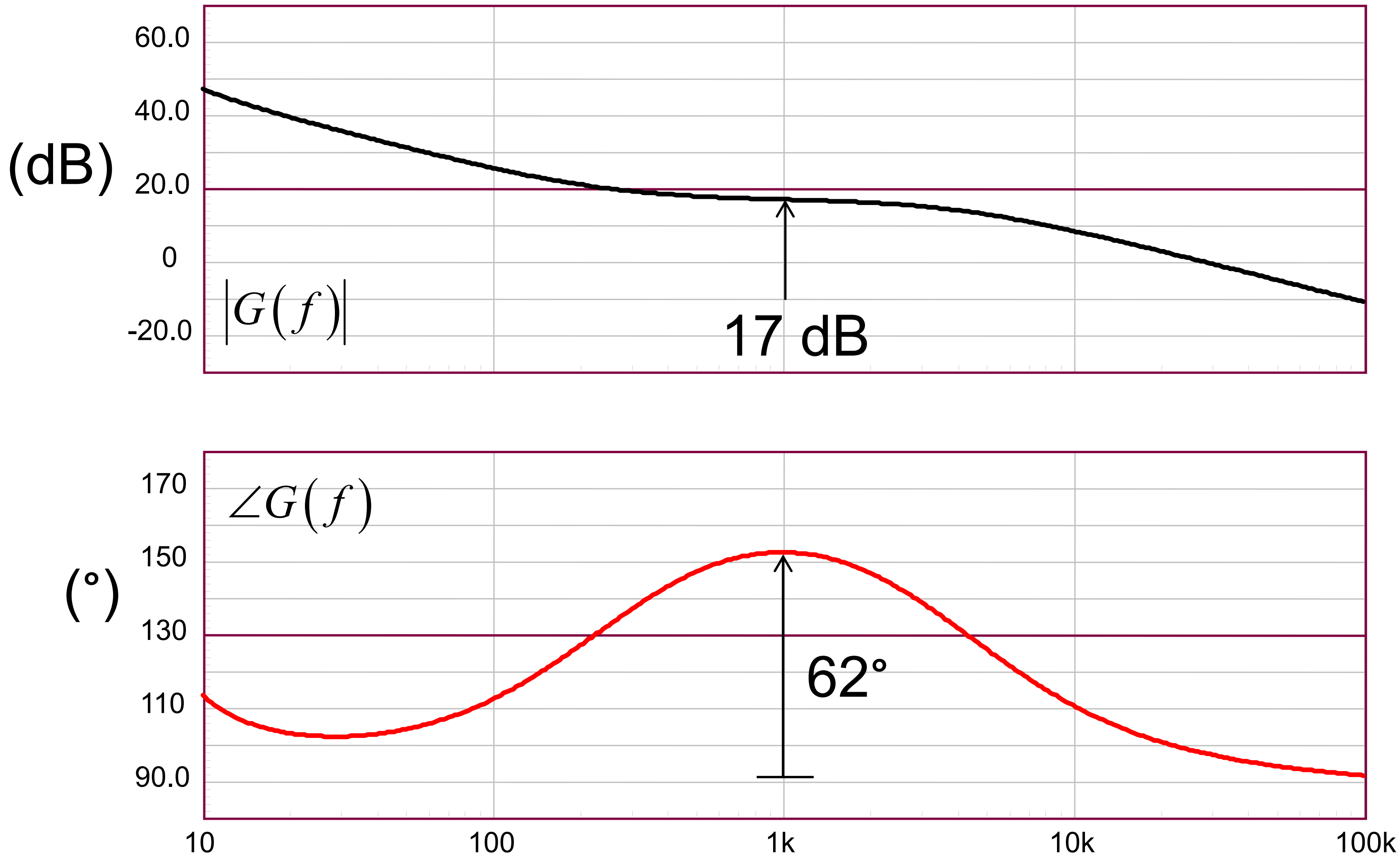
Fopto=4k
Copto=1/(2*pi*Fopto*Rpullup)
CTR = 0.3



Automated type 2 compensator using a TL431

TL431 type 2 Design Example

- The 1-dB gain difference is linked to R_d and the bias current



ON Semiconductor Offer in QR

Product	Data Sheet	Compliance	Status	Description	Topology	Control Mode	f_{sw} Typ (kHz) Variable	Stand-by Mode	UVLO (V)	Short Circuit Protection	Latch	Soft Start	V_{CC} Max (V)	Drive Cap. (mA)	Package Type
NCP1336		Pb-free Halide free	Active NEW	Quasi-Resonant Current Mode Controller w/ HV Start-up	Flyback	Current Mode	Variable	Yes	9 - 15	Yes	Yes	Yes	28	500 / 800	SOIC-14 NB, Less Pin 13
NCL30051 Design it		Pb-free Halide free	Active	PFC and Resonant Half Bridge Combo Controller for LED Lighting		Voltage Mode	Variable				No				SOIC-16
NCP1308		Pb-free Halide free	Active	Current Mode Controller for Free Running Quasi-Resonant Operation	Flyback	Current Mode	Variable	Yes	Yes	Yes	Yes	Yes	16	500 / 500	SOIC-8
NCP1337		Pb-free Halide free	Active	Quasi-Resonant Current Mode Controller Featuring Overpower Compensation	Flyback	Current Mode	Variable	Yes	Yes	Yes	Yes	Yes	20	500 / 500	PDIP-8 SOIC-8
NCP1338		Pb-free Halide free	Active	PWM Current Mode Controller for Free Running Quasi Resonant Operation	Flyback	Current Mode	Variable	Yes	Yes	Yes	Yes	Yes	20	500 / 500	SOIC-8
NCP1351		Pb-free Halide free	Active	Variable Off-Time Current Mode Controller Featuring Short Circuit Protection	Flyback	Current Mode	Variable	No	Yes	Yes	Yes	No	28	400 / 400	PDIP-8 SOIC-8
NCP1377		Pb-free Halide free	Active	Current-Mode Controller for Free Running Quasi-Resonant Operation	Flyback	Current Mode	Variable	Yes	Yes	No	Yes	Yes	16	500 / 500	PDIP-8 SOIC-8
NCP1379		Pb-free Halide free	Active	Quasi-Resonant Current Mode Controller	Flyback	Current Mode	Variable	Yes	9	Yes	Yes	Yes	28	500 / 800	SOIC-8
NCP1380		Pb-free Halide free	Active	Quasi-Resonant Current Mode Controller	Flyback	Current Mode	Variable	Yes	9	Yes	No Yes	Yes	28	500 / 800	SOIC-8

New Book on Loop Control

Conclusion on QR Converters

❑ Quasi Square-Wave Resonant Converters feature:

- minimal or zero turn-on losses, Zero Voltage Switching, ZVS
- limited amount of radiated EMI
- natural frequency dithering at low line
- safe operation in short circuit
- easy synchronous rectification implementation
- 1st order system in small signal, easy to compensate

❑ Despite these bullets, QR converters suffer from:

- variable operating frequency, line and load dependency
- discontinuous operation leads to high conduction losses
- large output power delivery discrepancy between line levels
- higher losses in transformer core compared to CCM