FUTURE ELECTRONICS

An Introduction to the Dual Active Bridge Converter or DAB

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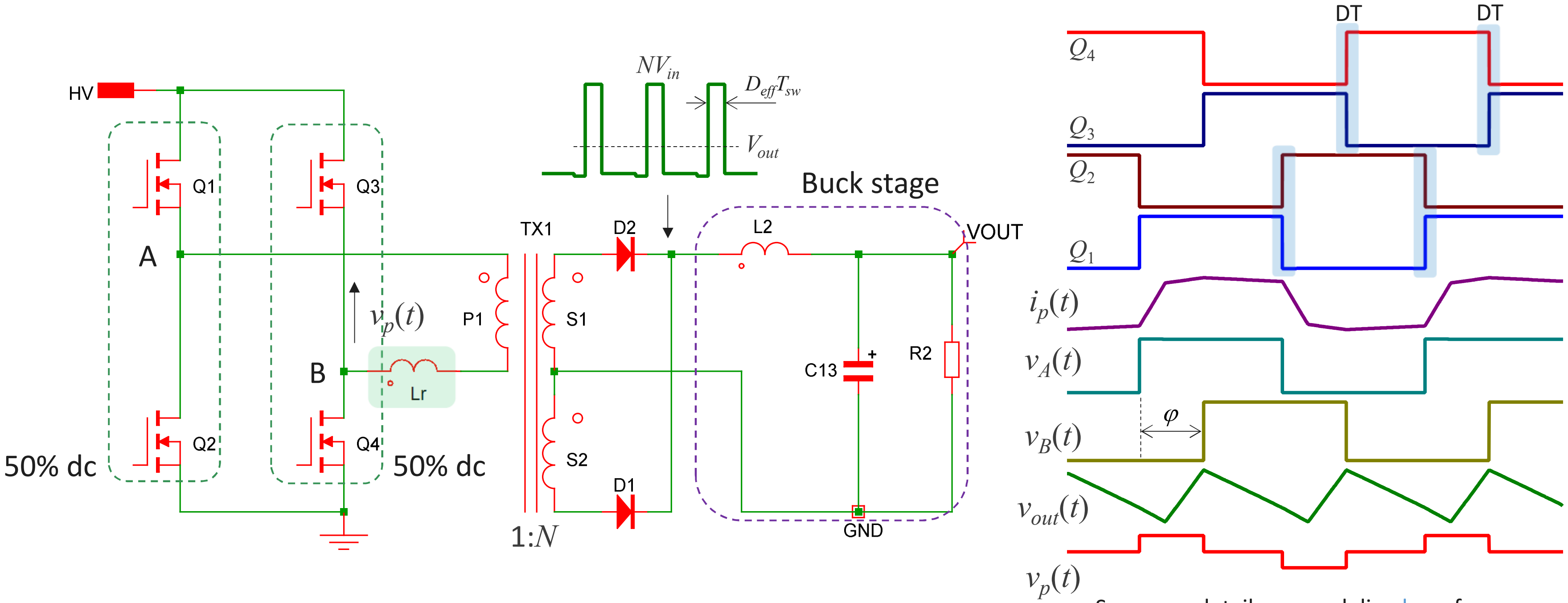
Agenda

- Phase-Shift Control
- Determining Inductor Currents
- Controlling the Power Flow
- Current Stress in the Converter
- Transformer Operating Point
- Switching Losses
- Small-Signal Response

Phase-Shift Modulation

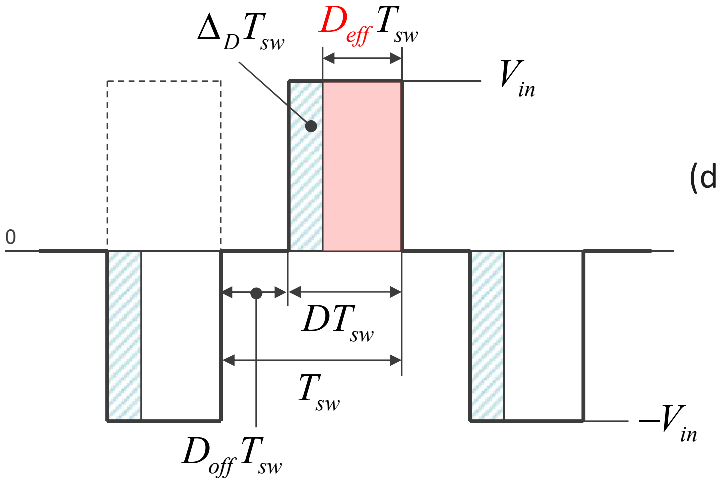
- A classical full-bridge converter is a buck-derived topology
- The dc transfer characteristic depends on the effective duty ratio D_{eff}
- ZVS can be obtained in the primary side but may be a problem in light load



See more details on modeling here from my webpage

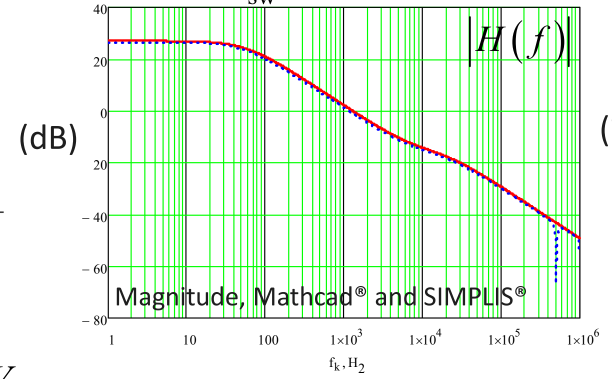
Small-Signal Response of the PSFB

- The control-to-output transfer function of the PSFB shows no resonating peak
  - ➢ Inherent current feedback effectively damps the system: d_{eff} reduces if I_{out} increases

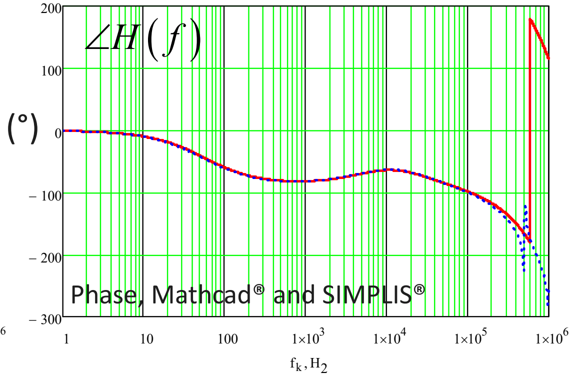


$$\Delta D = \frac{n\left[2I_{out} - T_{sw}\frac{V_{out}}{L}(1-D)\right]}{\frac{V_{in}}{l_{leak}}T_{sw}}$$

$$\Delta t := \Delta D \cdot T_{sw} = 434.87248 \cdot ns$$



Magnitude, Mathcad® and SIMPLIS®



Phase, Mathcad® and SIMPLIS®

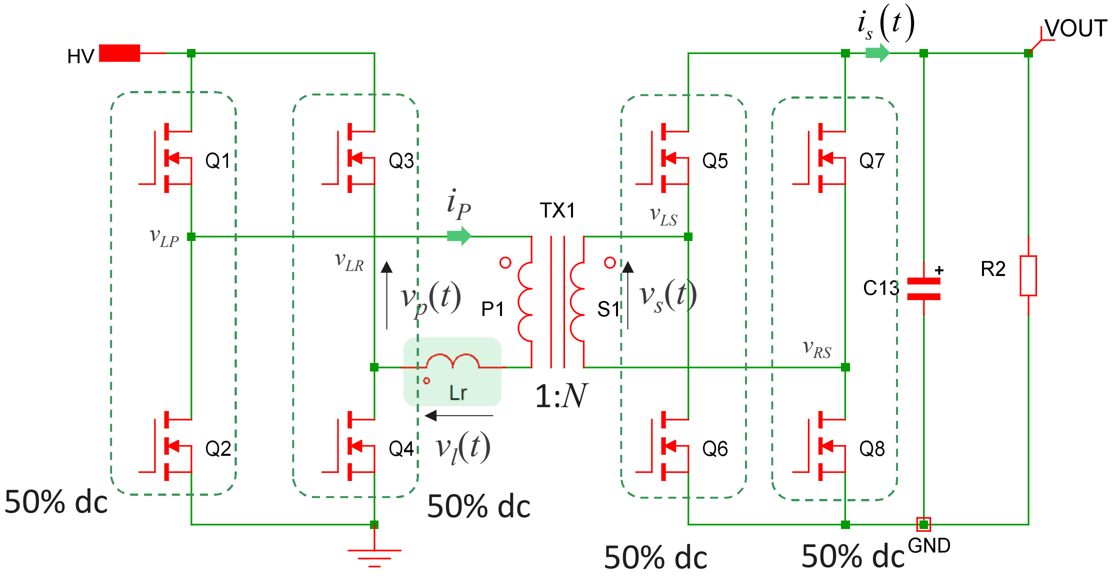
$$H_2(s) := \frac{N_1 \cdot R_L \cdot V_{in}}{R_L + R_d} \cdot \frac{1 + s \cdot r_C \cdot C_1}{1 + s \cdot \left(\frac{L_1 + C_1 \cdot R_L \cdot R_d}{R_L + R_d}\right) + s^2 \cdot \frac{C_1 \cdot L_1 \cdot R_L}{R_L + R_d}} \cdot e^{-s \cdot \Delta t}$$

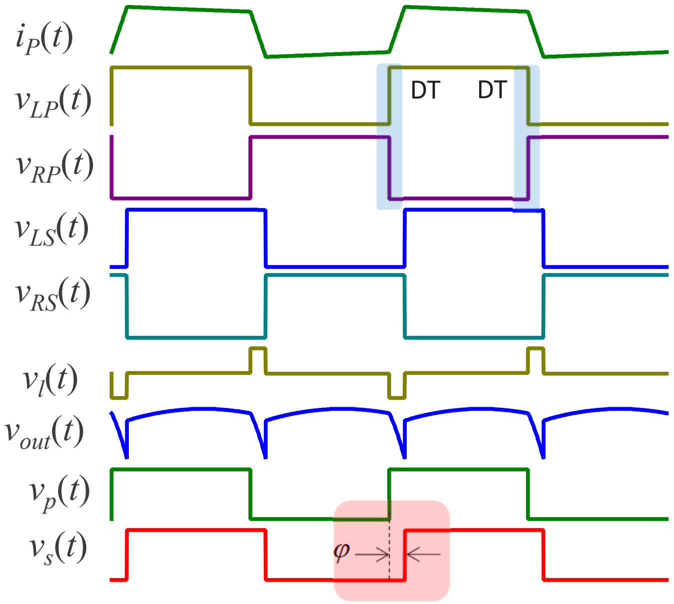
$$H(s) = H_0 \frac{1 + \frac{s}{\omega_z}}{1 + \frac{s}{\omega_0 Q} + \left(\frac{s}{\omega_0}\right)^2} e^{-s\Delta t} \qquad Q = \frac{\sqrt{b_2}}{b_1} \qquad \omega_0 = \frac{1}{\sqrt{b_2}}$$

Controlled Rectification

- The dual active bridge or DAB implies a second controlled bridge for rectification
- A single series inductor located in the primary side sizes the power flow
- Power transfer takes place from the leading bridge to the lagging bridge

✓ Depending on the angle polarity, power can flow in either direction





R.W. De Doncker et al., *A Three-Phase Soft-Switched High Power Density Dc-Dc Converter for High-Power Applications*, IEEE Transactions on Industry Applications, 1991

Power Transfer between Sinusoidal Sources

- A simplified view to the converter implies two sinusoidal sources and an inductor
- The sources operate at the same frequency and the phase between them is adjusted

✓ By changing the phase shift between sources, power flow is adjusted

```
.param Vp1=100
.param Vp2=150
.param phi=45
.param Fsw=100k
```

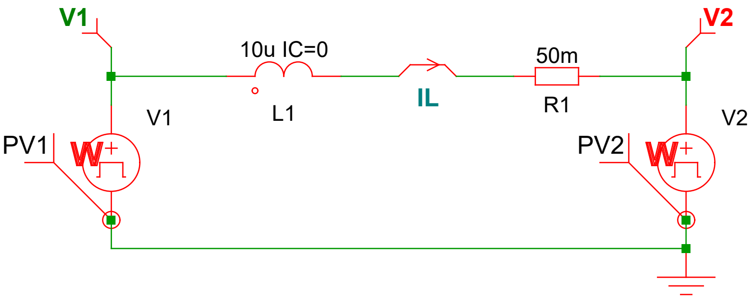
$t := 0, 100ns .. 20\mu s$ $\qquad V_1 := 100V$

$F := 100kHz \qquad L := 10\mu H$ $\qquad V_2 := 150V$

$\omega := 2\cdot\pi\cdot F \qquad \varphi := 45°$ $\qquad r := 0.05\Omega$

$v_1(t) := V_1\cdot\sin(\omega\cdot t)$

$v_2(t) := V_2\cdot\sin(\omega\cdot t - \varphi)$



Instantaneous power

$$p(t) = \overline{\left[\frac{V_1\cdot e^{i\cdot\omega\cdot t} - V_2\cdot e^{i\cdot(\omega\cdot t+\varphi)}}{i\cdot L\cdot\omega}\right]}\cdot\left(V_1\cdot e^{i\cdot\omega\cdot t}\right)$$

Instantaneous current

$$P = 2\cdot F_{sw}\cdot\int_0^{\frac{1}{2\cdot F_{sw}}}\left(\frac{\overline{v_L(t)}}{Z_L}\right)\cdot v_1(t)\,dt$$

Complex power is computed as P = UI*

$$\boxed{\frac{V_1\cdot V_2\cdot\sin(\varphi)}{\omega\cdot L} = 1.688kW}$$

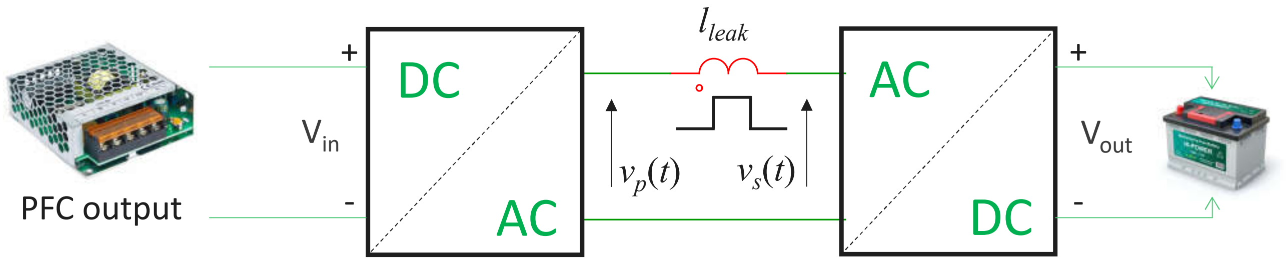
Processed power

$$\mathrm{Re}\left(\frac{V_1\cdot V_2\cdot\sin(\varphi) - V_1\cdot V_2\cdot\cos(\varphi)\cdot i + V_1^{2}\cdot i}{L\cdot\omega}\right) = 1.688kW$$

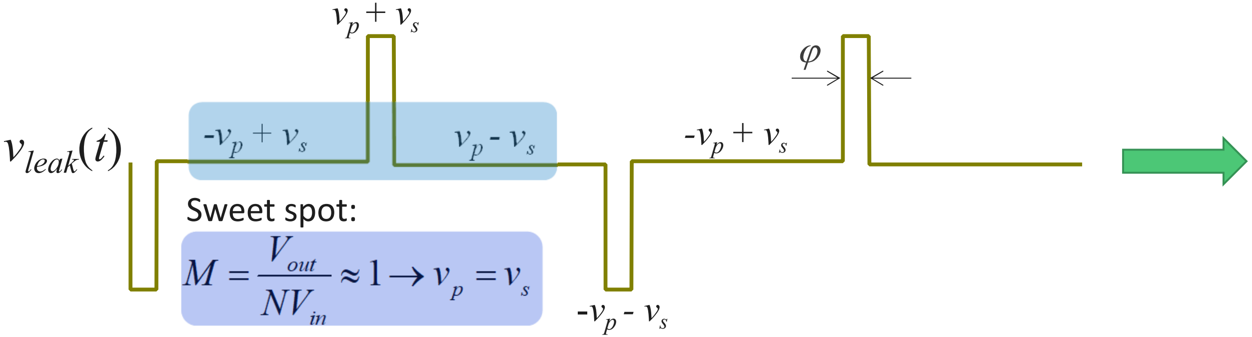
- Changing the sign of the phase reverses the power flow

Power Transfer between Square-Wave Sources

- Power is adjusted by controlling the voltage applied across the leakage inductance
- A transformer brings isolation and provides boost or input voltage reduction



✓Adjusting the phase shift between square waves controls the inductor current



Sweet spot:

$$M = \frac{V_{out}}{NV_{in}} \approx 1 \rightarrow v_p = v_s$$

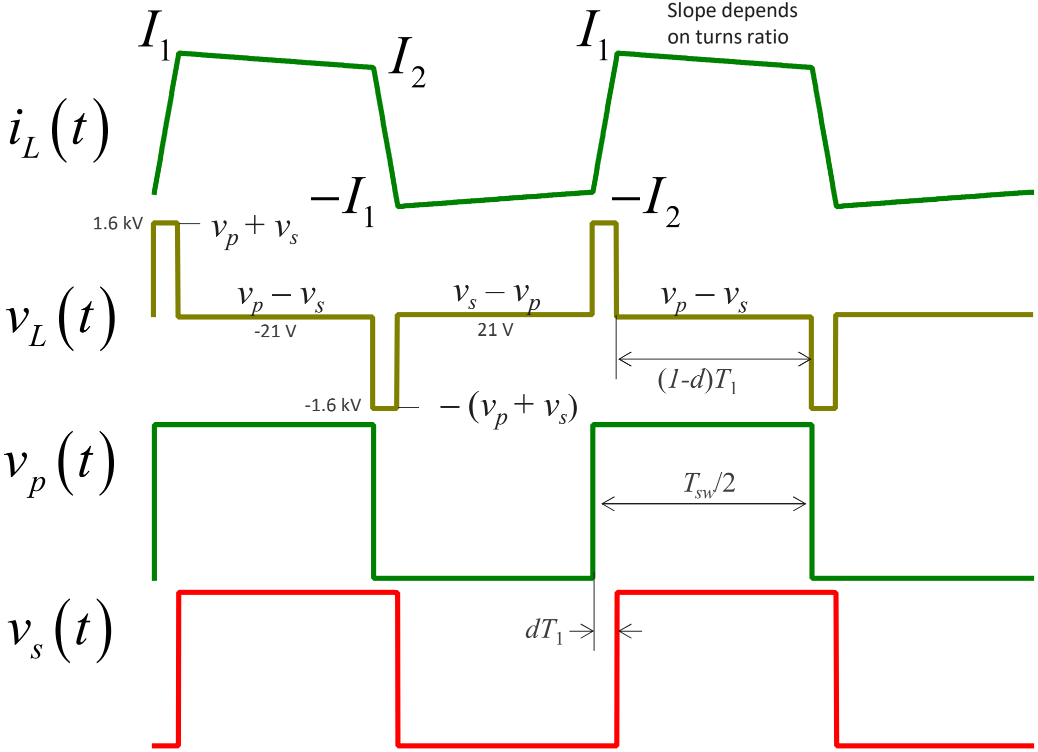
$$\frac{di_L(t)}{dt} = \frac{v_p(t) - v_s(t)}{l_{leak}}$$

Agenda

- Phase-Shift Control
- **Determining Inductor Currents**
- Controlling the Power Flow
- Current Stress in the Converter
- Transformer Operating Point
- Switching Losses
- Small-Signal Response

Determining Inductor Currents

- Power is adjusted by controlling the voltage applied across the leakage inductance
- A transformer brings isolation and scale for boosting or stepping down the voltage


Slope depends on turns ratio

$$I_1 + I_2 = \frac{V_{in} + V_{out}/N}{l_{leak}} dT_1$$

$$I_1 - I_2 = \frac{V_{out}/N - V_{in}}{l_{leak}} (1-d)T_1$$

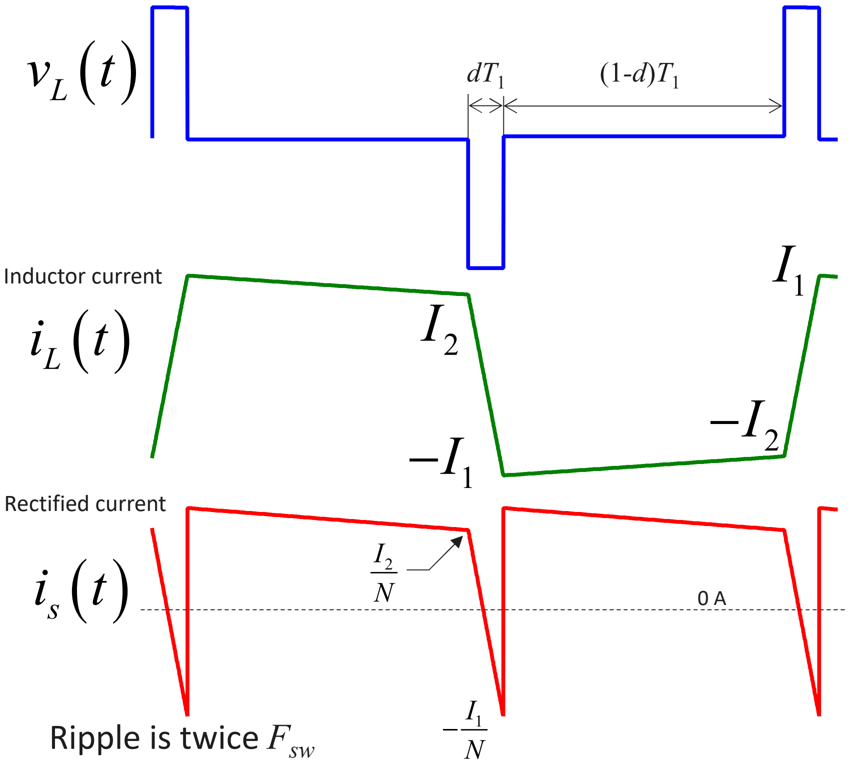
$$I_1 = \frac{T_1\left(V_{out} - NV_{in} + 2N \cdot d \cdot V_{in}\right)}{2Nl_{leak}}$$

$$I_2 = \frac{T_1\left(2 \cdot d \cdot V_{out} + NV_{in} - V_{out}\right)}{2Nl_{leak}}$$

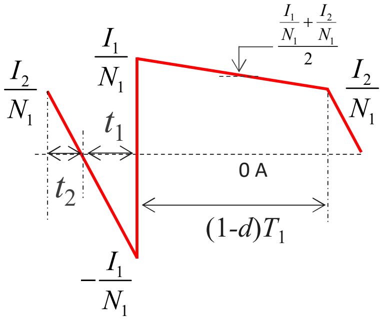
$T_1 = 1/\left(2F_{sw}\right)$

Calculating the Output Power

- The primary inductive current is scaled by the turns ratio and flows in the secondary
- Determining its average rectified value leads to the delivered power



Ripple is twice F_{sw}



$$t_1 = \frac{I_1 N l_{leak}}{V_{out} + N V_{in}}$$

$$t_2 = \frac{I_2 N l_{leak}}{V_{out} + N V_{in}}$$

$$t_1 + t_2 = dT_1$$

$$\langle i_s(t) \rangle_{T_{sw}} = \frac{\left(\frac{I_2}{N}\frac{t_2}{2} - \frac{I_1}{N}\frac{t_1}{2}\right) + \frac{\frac{I_2}{N} + \frac{I_1}{N}}{2}(1-d)T_1}{T_1}$$

Insert I_1 and I_2 from previous slide, rearrange

$$P_{out} = I_s V_{out} = \frac{d(1-d)T_{sw} V_{in} V_{out}}{2 N l_{leak}}$$

$T_1 = 1/(2F_{sw})$

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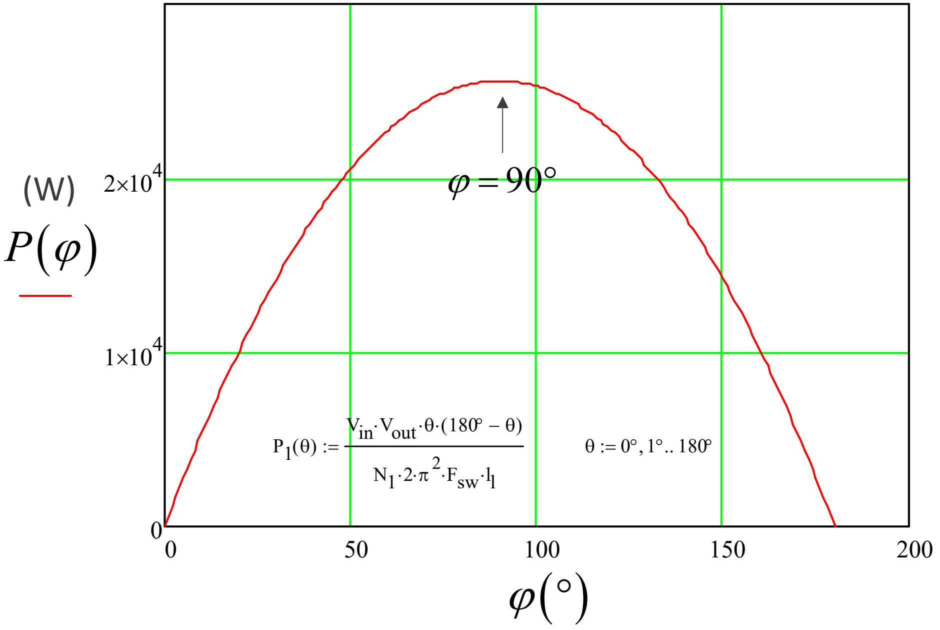
Controlling the Power Flow

- Varying the duty ratio or the phase-shift between bridges adjusts the power flow
- The transmitted power is maximal when the angle reaches 90° or π/2



✓ The sign of the angle determines the power transfer direction

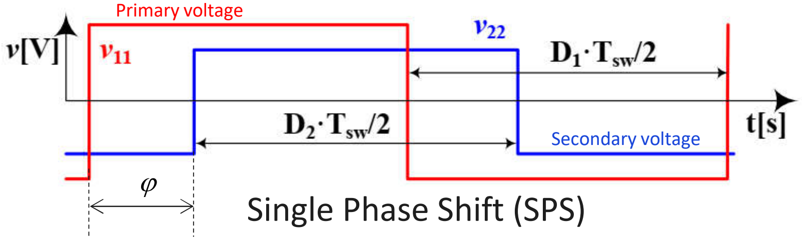
✓ The controlling angle must be bounded to keep the same control law slope:

$$D_1 \in [0;50\%] \quad D_2 \in [0;50\%] \quad \varphi \in \left[-\frac{\pi}{2};\frac{\pi}{2}\right]$$

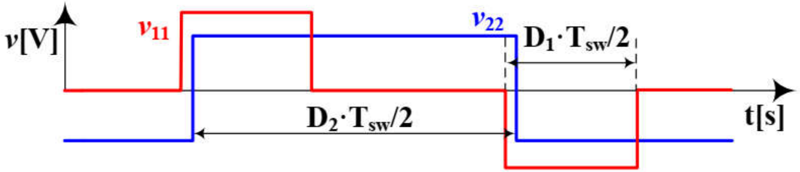
$$P_{out} = \frac{V_{in}V_{out}\cdot\varphi\cdot(180° - \varphi)}{2\pi^2 N F_{sw} l_{leak}}$$

Different Types of Modulations

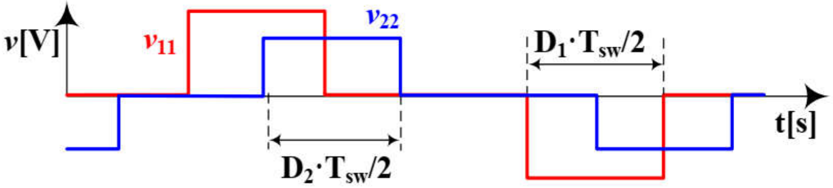
- Different strategies exist to modulate the power flow with different control variables
- ✓ Single phase shift implies 50% duty ratio for both bridges and phase modulation
- ✓ Dual phase shift with variable *equal* duty ratios and phase shift adjustment
- ✓ Triple phase shift where both duty ratios are modulated with phase shift control



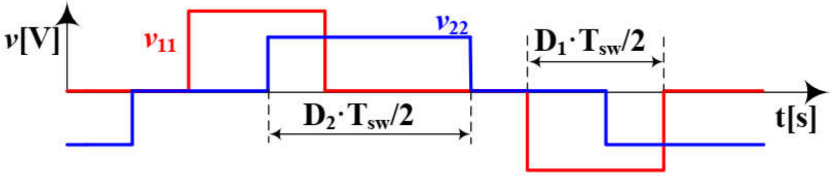
Single Phase Shift (SPS)



Extended Phase Shift (EPS)



Dual Phase Shift (DPS)

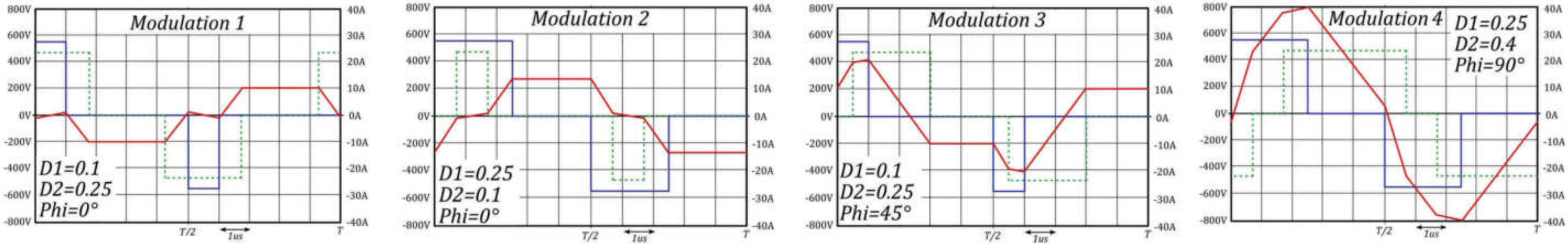


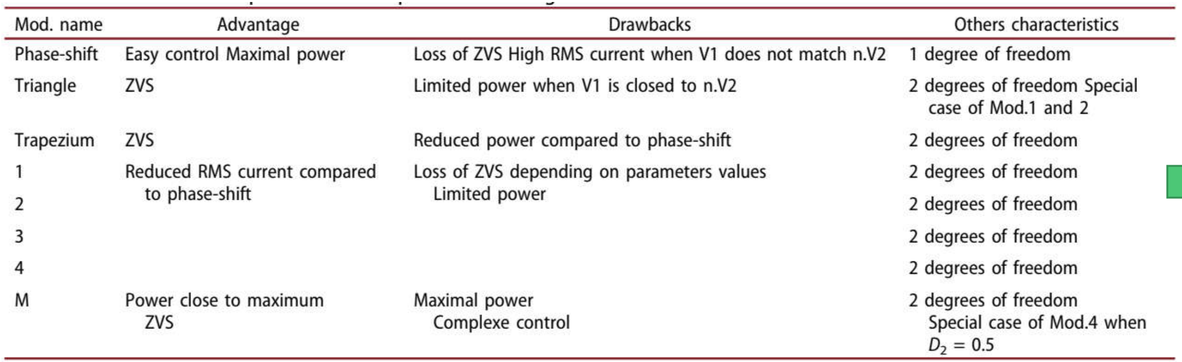
Triple Phase Shift (TPS)

C. Calderon et al., *General Analysis of Switching Modes in a Dual Active Bridge with Triple Phase Shift Modulation*, MDPI, September 2018

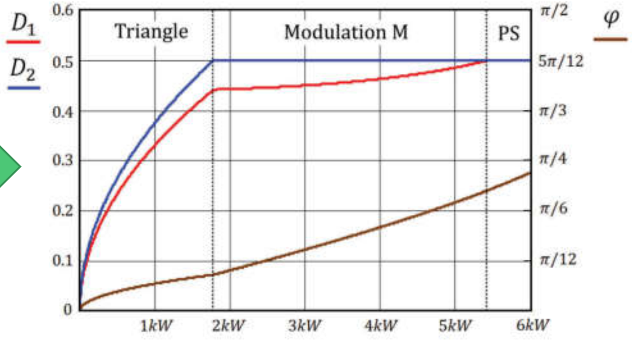
Minimizing Rms Currents

- Various modulation strategies can be explored to minimize the circulating rms currents
- The aim is to reduce conduction losses and reactive power while ensuring ZVS





Mod. name	Advantage	Drawbacks	Others characteristics
Phase-shift	Easy control Maximal power	Loss of ZVS High RMS current when V1 does not match n.V2	1 degree of freedom
Triangle	ZVS	Limited power when V1 is closed to n.V2	2 degrees of freedom Special case of Mod.1 and 2
Trapezium	ZVS	Reduced power compared to phase-shift	2 degrees of freedom
1	Reduced RMS current compared to phase-shift	Loss of ZVS depending on parameters values Limited power	2 degrees of freedom
2			2 degrees of freedom
3			2 degrees of freedom
4			2 degrees of freedom
M	Power close to maximum ZVS	Maximal power Complexe control	2 degrees of freedom Special case of Mod.4 when $D_2 = 0.5$



M. Blanc, JP Ferrieux and al., *Optimization of a Dc-Dc Dual Active Bridge Converter for Aircraft Application*, PCIM Europe 2017, Intelligent Motion and Renewable Energy and Energy Management

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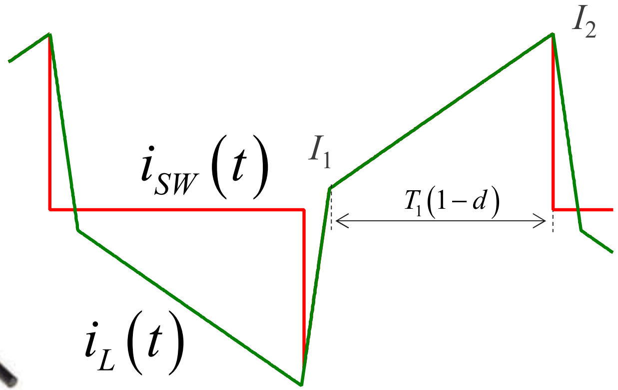
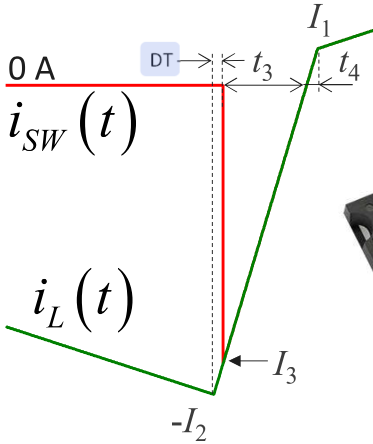
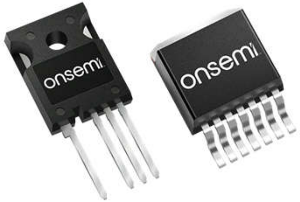
Rms Current in Primary Switches

- The deadtime plays a role in the circulating current and must be accounted for



Zoom in





- Determine the switch current I_3 after deadtime:

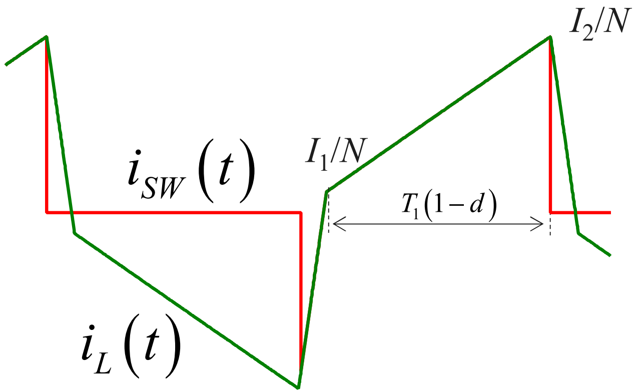
$$I_3 = -\left(I_2 - \frac{V_{in} + \frac{V_{out}}{N}}{l_{leak}} \cdot \mathrm{DT}\right) \longrightarrow t_3 = \frac{N I_3 l_{leak}}{V_{out} + N V_{in}}$$

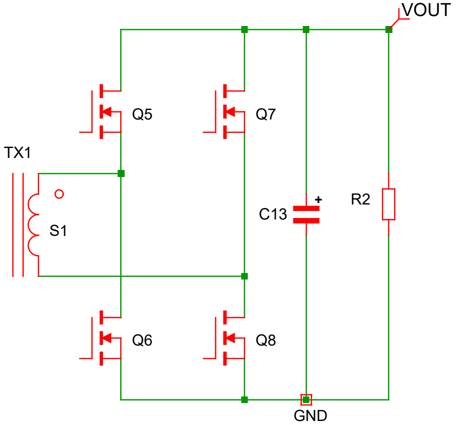
- Determine the time from 0 A to I_1: $t_4 = \frac{N I_1 l_{leak}}{V_{out} + N V_{in}}$
- Sanity check to verify $dT_1 = \mathrm{DT} + t_3 + t_4$
- Conduction time $t_5 = T_1(1-d)$

$$I_{SW,\mathrm{rms}} = \sqrt{\frac{1}{3T_{sw}}\left(I_3^2 t_3 + I_1^2 t_4\right) + \frac{1}{3}\left(I_1^2 + I_2^2 + I_1 I_2\right)\frac{t_5}{T_{sw}}}$$

Rms Current in the Secondary Switches

- The secondary-side bridge operates as a synchronous rectifier
- It is important to size the amount of rms current which circulates in these switches



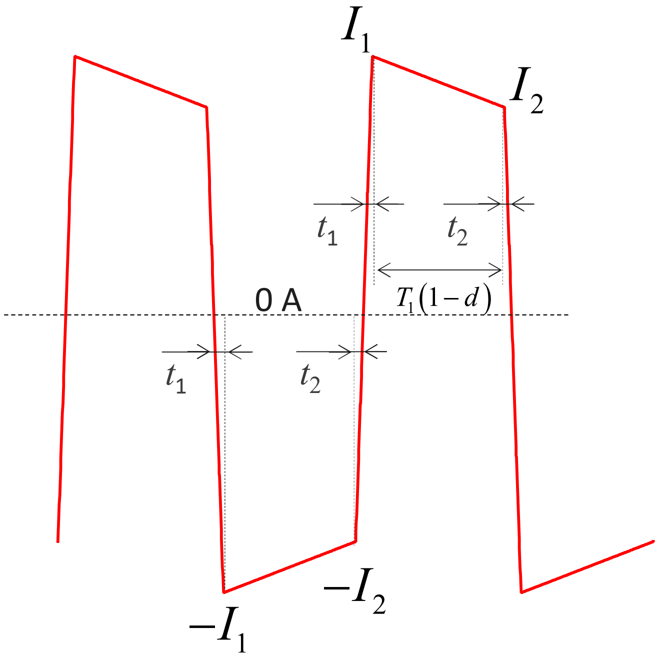


- The secondary-side currents are those of the primary switches scaled by the turns ratio

$$I_{SW,\mathrm{rms}} = \sqrt{\frac{1}{3T_{sw}}\left[\left(\frac{I_3}{N}\right)^2 t_3 + \left(\frac{I_1}{N}\right)^2 t_4\right] + \frac{1}{3}\left[\left(\frac{I_1}{N}\right)^2 + \left(\frac{I_2}{N}\right)^2 + \frac{I_1 I_2}{N^2}\right]\frac{t_5}{T_{sw}}}$$

Rms Current in the Transformer

- The transformer and the primary-side inductor see the entire primary current
- It is important to size this rms current as it affects copper losses



$$I_{L,\mathrm{rms}} = \sqrt{\frac{1}{3T_1}\left(I_1{}^2 t_1 + I_2{}^2 t_2\right) + \frac{1}{3}\left(I_1{}^2 + I_2{}^2 + I_1 I_2\right)(1-d)}$$

$$I_{L,\mathrm{rms}} = \sqrt{\frac{I_1{}^2}{3} + \frac{I_2{}^2}{3} + I_1 I_2 \frac{1-2\dfrac{\varphi}{180^\circ}}{3}}$$

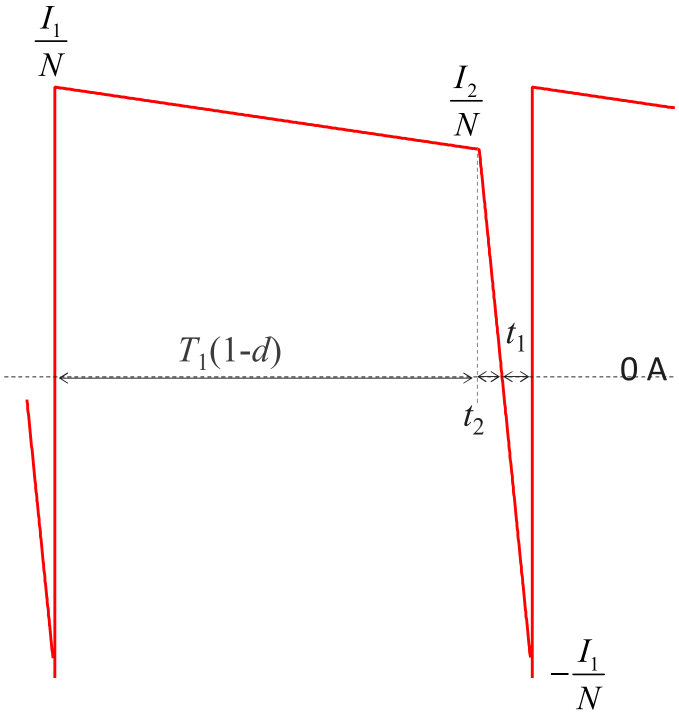
From TI application note

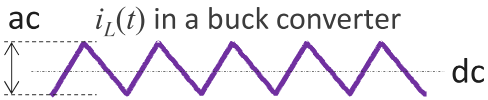
Design Guide: TIDA-010054 - Bidirectional, Dual Active Bridge Reference Design for Level 3 Electric Vehicle Charging Stations

Rms Current in the Output Capacitor

- It is important to determine the rms current flowing in the output capacitor

✓ Operating life-time of the capacitor depends on the correct type selection





$$I_{L,rms} = \sqrt{I_{ac}^{\ 2} + I_{dc}^{\ 2}}$$

Flows in the cap.

Flows in the load.

$$\left\langle i_s(t)\right\rangle_{T_{sw}} = \frac{\left(\frac{I_2}{N}\frac{t_2}{2} - \frac{I_1}{N}\frac{t_1}{2}\right) + \frac{\frac{I_2}{N}+\frac{I_1}{N}}{2}(1-d)T_1}{T_1}$$

Dc current

Capacitor rms current

$$I_{s,\text{rms}} = \sqrt{\frac{1}{3T_1}\left[\left(\frac{I_1}{N}\right)^2 t_2 + \left(\frac{I_2}{N}\right)^2 t_1\right] + \frac{1}{3}\left[\left(\frac{I_1}{N}\right)^2 + \left(\frac{I_2}{N}\right)^2 + \frac{I_1 I_2}{N^2}\right](1-d)}$$

Checking Computed and Simulated Data - 1

- SIMPLIS can deliver the operating point in a few seconds
- It is important to check computed and simulated data before validating calculated values

1st design:

$V_{in} = 600\ \mathrm{V}$

$V_{out} = 500\ \mathrm{V}$

$l_{leak} = 32\ \mu\mathrm{H}$

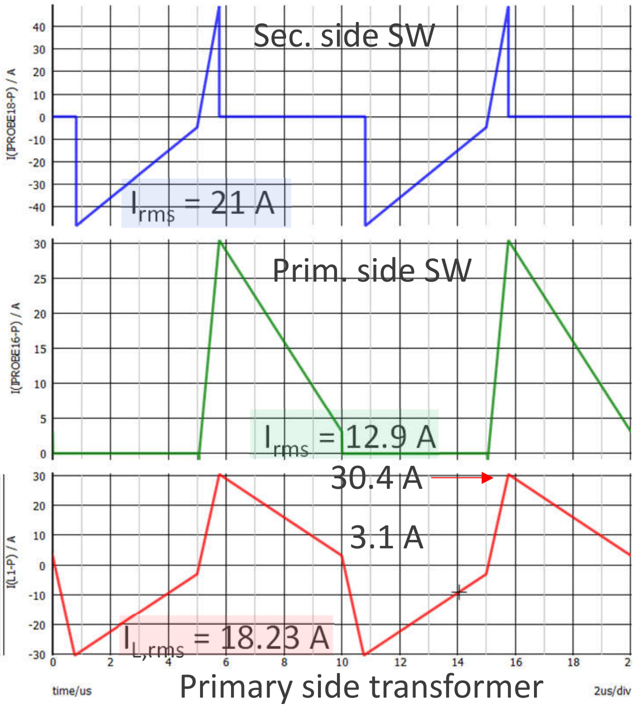
Turns ratio $N = 625\mathrm{m}$

Mathcad®

$$I_1 := \frac{T_1 \cdot (V_{out} - N_1 \cdot V_{in} + 2 \cdot N_1 \cdot V_{in} \cdot d)}{2 \cdot N_1 \cdot l_l} = 30.275\mathrm{A}$$

$$I_2 := \frac{T_1 \cdot (N_1 \cdot V_{in} - V_{out} + 2 \cdot V_{out} \cdot d)}{2 \cdot N_1 \cdot l_l} = 3.402\mathrm{A}$$

$$P_{out} := \frac{(1-d) \cdot d \cdot T_{sw} \cdot V_{in} \cdot V_{out}}{2 \cdot l_l \cdot N_1} = 9.79\,\mathrm{kW}$$



Primary side transformer

Inductor rms current

$$I_L := \sqrt{\frac{1}{3} \cdot I_2^2 \cdot \frac{t_2}{T_1} + \frac{1}{3} \cdot I_1^2 \cdot \frac{t_1}{T_1} + \frac{1}{3} \cdot \left[\left(I_1^2 + I_2^2 + I_1 \cdot I_2\right) \cdot (1-d)\right]} = 18.253\mathrm{A}$$

Sec. switches rms current

$$I_{SW} := \sqrt{\frac{1}{3} \cdot \left(\frac{I_3}{N_1}\right)^2 \cdot \frac{t_3}{T_{sw}} + \frac{1}{3} \cdot \left(\frac{I_1}{N_1}\right)^2 \cdot \frac{t_4}{T_{sw}} + \frac{1}{3} \cdot \left[\left(\frac{I_1}{N_1}\right)^2 + \left(\frac{I_2}{N_1}\right)^2 + \frac{I_1 \cdot I_2}{N_1^2}\right] \cdot \frac{t_5}{T_{sw}}} = 20.649\mathrm{A}$$

Prim. switches rms current

$$I_{SW} := \sqrt{\frac{1}{3} \cdot I_3^2 \cdot \frac{t_3}{T_{sw}} + \frac{1}{3} \cdot I_1^2 \cdot \frac{t_4}{T_{sw}} + \frac{1}{3} \cdot \left(I_1^2 + I_2^2 + I_1 \cdot I_2\right) \cdot \frac{t_5}{T_{sw}}} = 12.906\mathrm{A}$$

Output capacitor rms current

$$I_C := \sqrt{I_{outR}^2 - I_{out}^2} = 21.74\mathrm{A}$$

22 A rms

Check operating temperature in worst case

Checking Computed and Simulated Data - 2

- In this second operating point, the output voltage is down to 308 V

✓ The duty ratio is reduced to 8.6% for that purpose

2nd design:

$V_{in} = 600\ \text{V}$

$V_{out} = 308\ \text{V}$

$l_{leak} = 32\ \mu\text{H}$

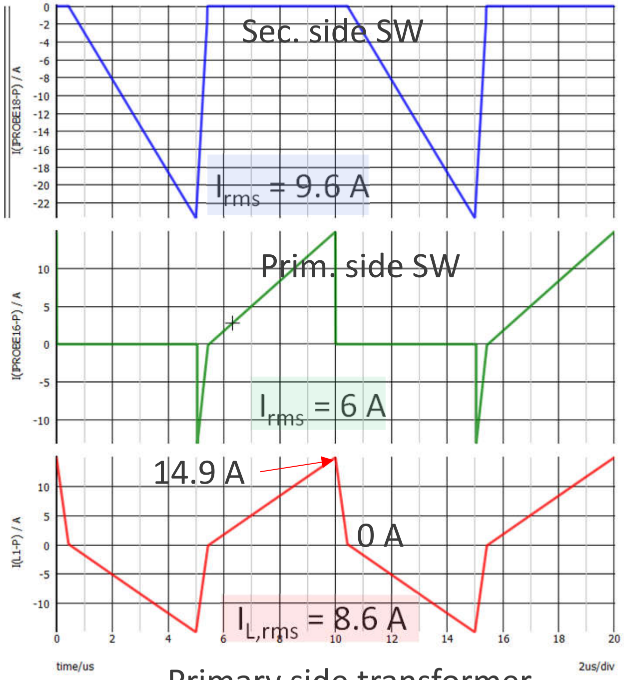
Turns ratio $N = 625\text{m}$

Mathcad®

$$I_1 := \frac{T_1 \cdot (V_{out} - N_1 \cdot V_{in} + 2 \cdot N_1 \cdot V_{in} \cdot d)}{2 \cdot N_1 \cdot l_l} = -0.301\text{A}$$

$$I_2 := \frac{T_1 \cdot (N_1 \cdot V_{in} - V_{out} + 2 \cdot V_{out} \cdot d)}{2 \cdot N_1 \cdot l_l} = 15.006\text{A}$$

$$P_{out} := \frac{(1-d) \cdot d \cdot T_{sw} \cdot V_{in} \cdot V_{out}}{2 \cdot l_l \cdot N_1} = 3.636\text{kW}$$



Primary side transformer

Inductor rms current

$$I_P := \sqrt{\frac{I_1^2}{3} + \frac{I_2^2}{3} + I_1 \cdot I_2 \cdot \frac{1 - 2 \cdot \frac{\theta_1}{180°}}{3}} = 8.593\text{A}$$

Sec. switches rms current

$$I_{SW} := \sqrt{\frac{1}{3} \cdot \left(\frac{I_3}{N_1}\right)^2 \cdot \frac{t_3}{T_{sw}} + \frac{1}{3} \cdot \left(\frac{I_1}{N_1}\right)^2 \cdot \frac{t_4}{T_{sw}} + \frac{1}{3} \cdot \left[\left(\frac{I_1}{N_1}\right)^2 + \left(\frac{I_2}{N_1}\right)^2 + \frac{I_1 \cdot I_2}{N_1^2}\right] \cdot \frac{t_5}{T_{sw}}} = 9.589\text{A}$$

Prim. switches rms current

$$I_{SW} := \sqrt{\frac{1}{3} \cdot I_3^2 \cdot \frac{t_3}{T_{sw}} + \frac{1}{3} \cdot I_1^2 \cdot \frac{t_4}{T_{sw}} + \frac{1}{3} \cdot (I_1^2 + I_2^2 + I_1 \cdot I_2) \cdot \frac{t_5}{T_{sw}}} = 5.993\text{A}$$

Output capacitor rms current

$$I_C := \sqrt{I_{outR}^2 - I_{out}^2} = 7.048\text{A}$$

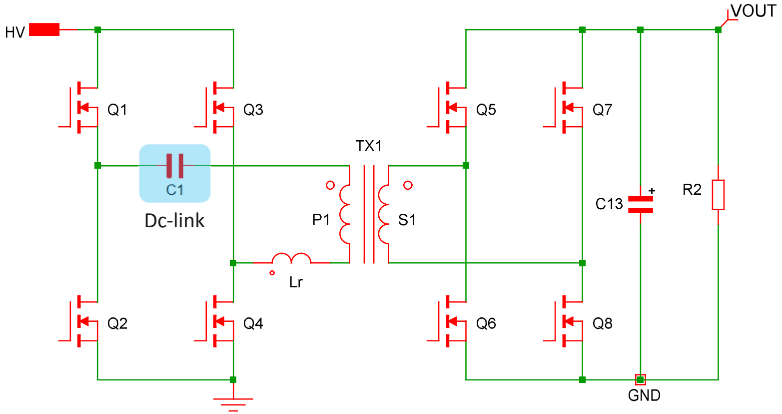
6.82 A rms

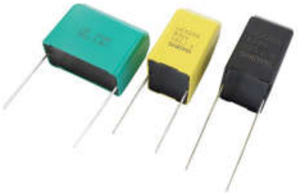
Agenda

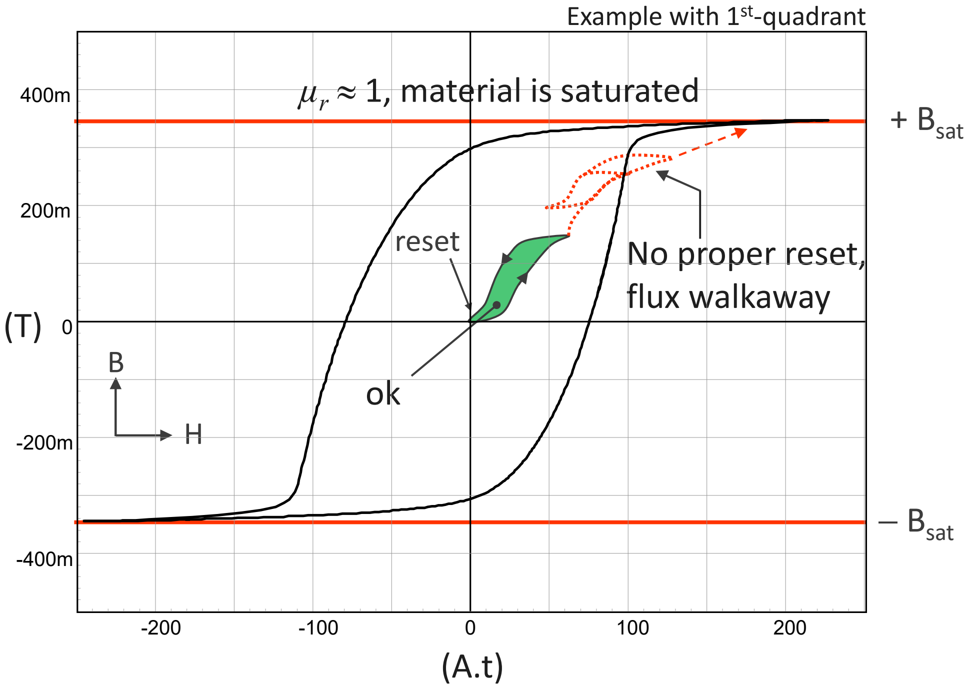
- Phase-Shift Control
- Determining Inductor Currents
- Controlling the Power Flow
- Current Stress in the Converter
- Transformer Operating Point
- Switching Losses
- Small-Signal Response

The Need to Demagnetize

- Core losses are linked to the magnetizing current and associated ΔB in the material
- Dc bias shifts the operating point and can cause saturation with catastrophic failure
- A dc-block capacitor is a solution, but it sees a large rms current and hampers cost







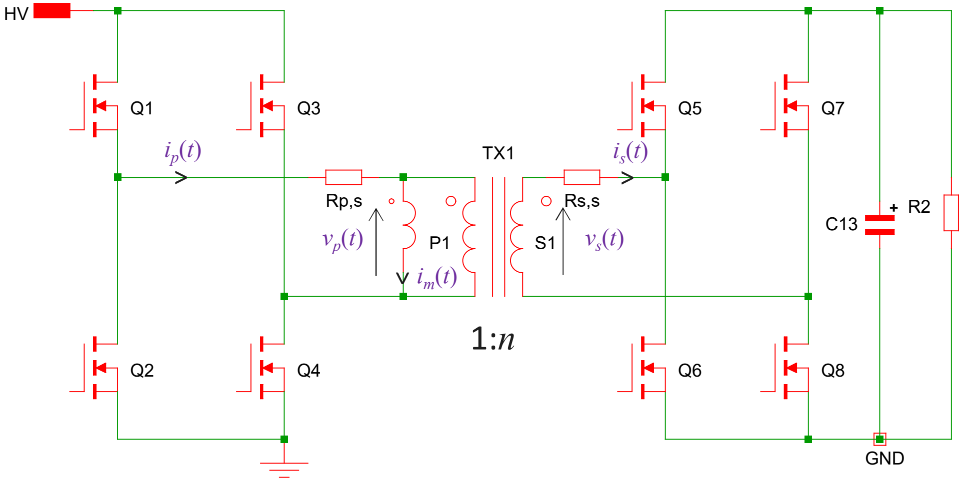
The Origin of the Drift

- Small variations in the control variables can affect the transformer volt-seconds:

✓ Semiconductor forward voltage drops move with temperature

✓ Insufficient pulse width modulator resolution can cause severe mismatches

✓ Gate driving signal delays affect the effective on- and off-time durations



- The primary current i_p is made of the magnetizing current i_m plus the reflected output current:

$$i_m(t) = i_p(t) - n \cdot i_s(t)$$

- It is possible to lump all ohmic drops into equivalent resistances in series with the primary and secondary:

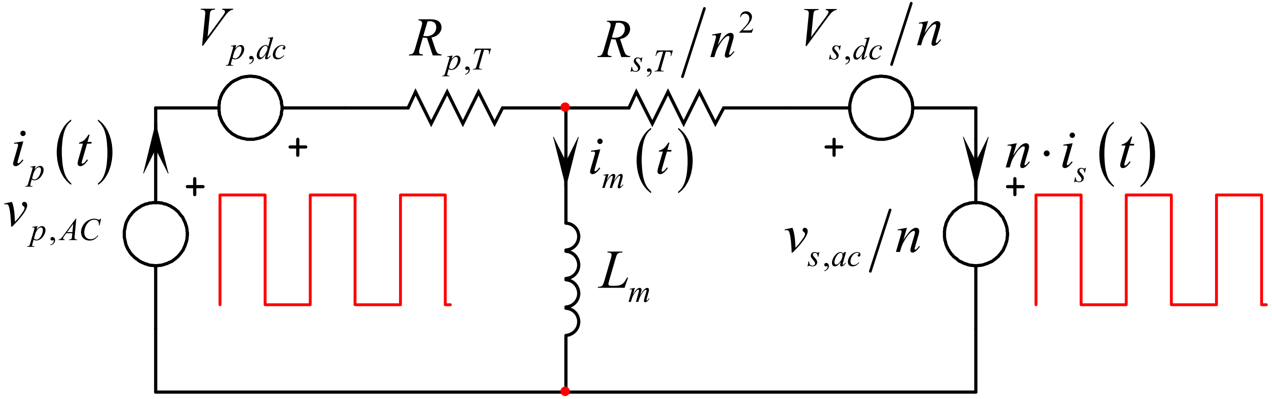
$$I_{m,dc} = I_{p,dc} - n \cdot I_{s,dc} = \frac{V_{p,dc}}{R_{p,T}} - n\frac{V_{s,dc}}{R_{s,T}}$$

G. Ortiz et al., *Flux Balancing of Isolation Transformers and Application of "The Magnetic Ear" for Closed-Loop Volt–Second Compensation*, *IEEE Transactions*, vol. 29, no. 8, Aug. 2014

Modeling the Drift

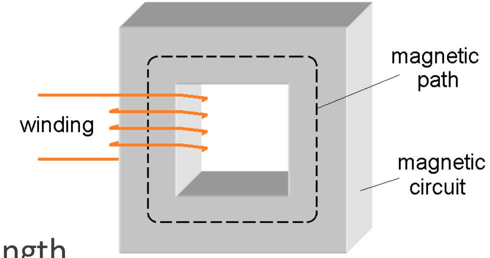
- The dc magnetic flux density depends on the winding and the core
- It is limited by the equivalent series resistances $R_{p,T}$ and $R_{s,T}$

✓ These resistances are extremely low for the best possible efficiency



$$H = \frac{NI}{l_m} \qquad B = \mu_0 \mu_r H$$

$$B = \frac{N_p \cdot I_{m,dc}}{l_m} \mu_0 \mu_r$$



$$B_{dc} = \left(\frac{V_{p,dc}}{R_{p,T}} - n \frac{V_{s,dc}}{R_{s,T}} \right) \cdot \frac{N_p}{l_m} \cdot \mu_0 \mu_r$$

l_m → mean magnetic path length

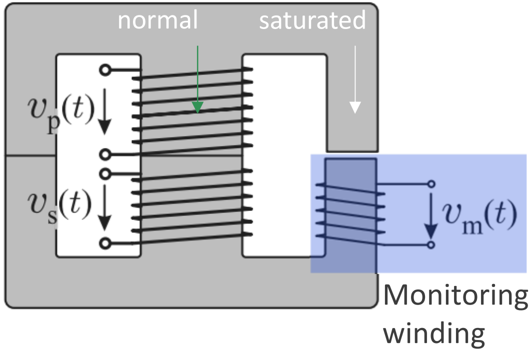
Assessing Transformer Flux Imbalance

- The drift can be significant in a high-power converter:

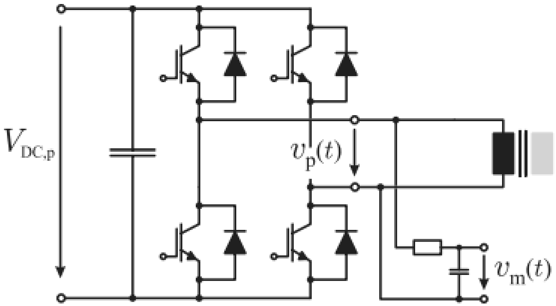
$$R_{p,T} = 1.7\,\mathrm{m\Omega} \quad F_{sw} = 20\,\mathrm{kHz} \quad P_{out} = 166\,\mathrm{kW} \quad F_{sw} = 20\,\mathrm{kHz} \quad \mu_r \approx 1950$$

- Material is N87 with a μ_r of 1950 and the error in on- off-primary duration is 0.0125%
- This is a 2.5-ns timing error which leads to dc flux density component of B_{dc} = 50 mT
- Risks of saturation, nonlinear magnetizing current and higher core losses under dc bias

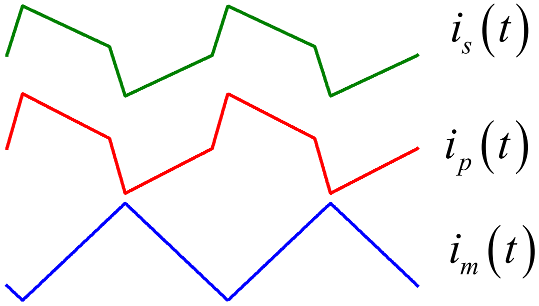
Saturation detection by parallel flux path



Dynamic flux measurement



Magnetizing current reconstruction



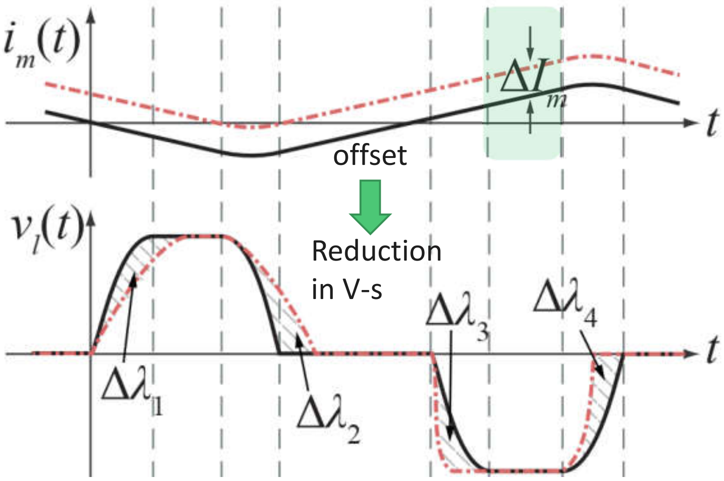
$$i_m(t) = i_p(t) - n \cdot i_s(t)$$

J. Muhlethaler et al., *Core Losses Under the Dc Bias Condition Based on Steinmetz Parameters*, *IEEE Transactions on Power Electronics*, vol. 27, no. 2, 2012

Fighting Transformer Flux Imbalance

- A series capacitor is possible with all the associated penalties (passive balancing)
- A gap is introduced to increase the saturation current, but it does not fix the problem
- Zero-Voltage Switching can partially compensate mismatch in V-s

Offset on mag. current

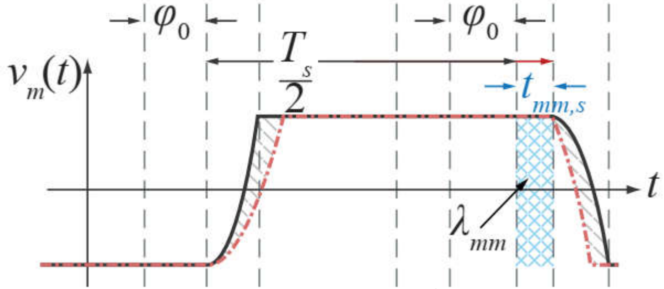


offset

Reduction in V-s

Add time mismatch

- ✓ The current offset speeds up the resonant transition and reduces flux linkage
- ✓ The time mismatch t_{mm} increases the V-s but the resulting dc offset speeds-up transitions which offers a counter measure



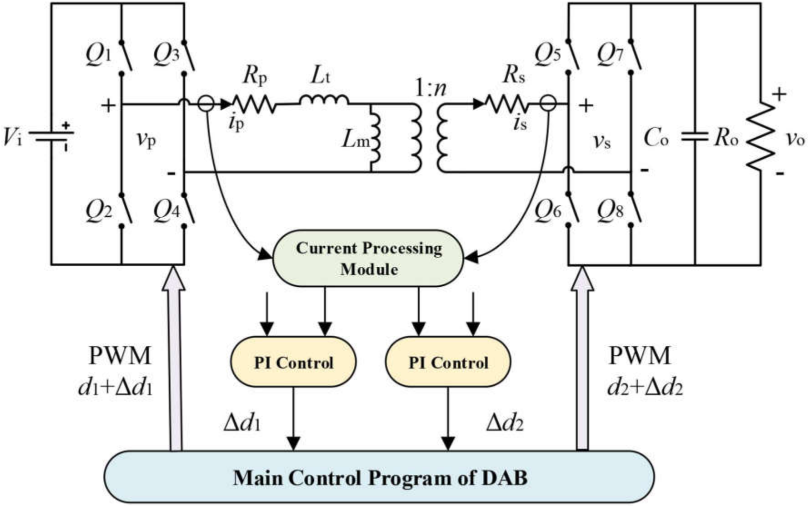
Primary-side voltage

$$\Delta I = \frac{2V_g t_{mm,\varphi}}{2\sum_{i=1}^{2}\frac{\Delta\lambda_i}{\Delta I}}$$

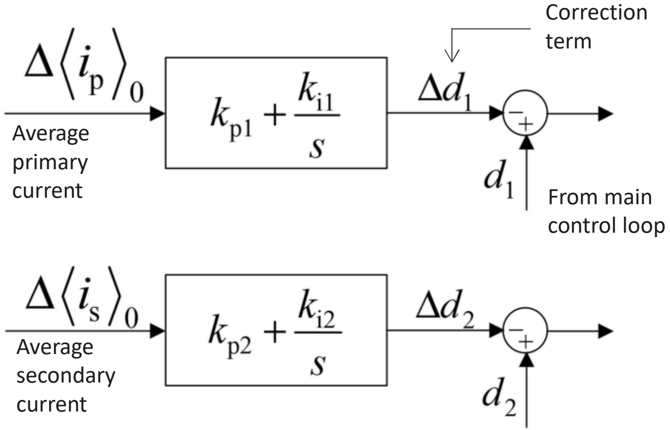
D. Costinett, D. Seltzer, D. Maksimovic and R. Zane, *Inherent volt-second balancing of magnetic devices in zero-voltage-switched power converters*, APEC *2013*

Active Control

- Active flux control requires the monitoring of the primary- and secondary-side currents
- ✓ Two loops are implemented with two different processing paths
- ✓ One loop keeps a zeroed magnetizing current, a second zeroes the average prim. current
- ➢ Having i_m and i_p zeroed on average implies a zeroed average sec. side current



- ➢ A small PWM Δd modulation is added to the setpoint and keeps the dc offset to zero as a goal



Z. Wang, J. Chai, X. Sun, *Method to control flux balancing of high-frequency transformers in dual active bridge dc–dc converters*, IET Journals, 2018

Y. Panov, M. M. Jovanović and B. T. Irving, *Novel transformer-flux-balancing control of dual-active-bridge bidirectional converters*, *APEC* 2015

Agenda

- Phase-Shift Control
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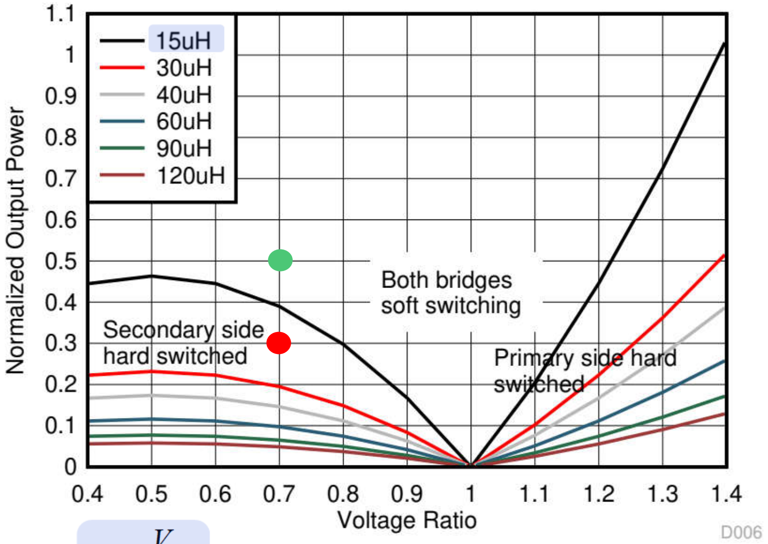
Effects of the Leakage Inductance

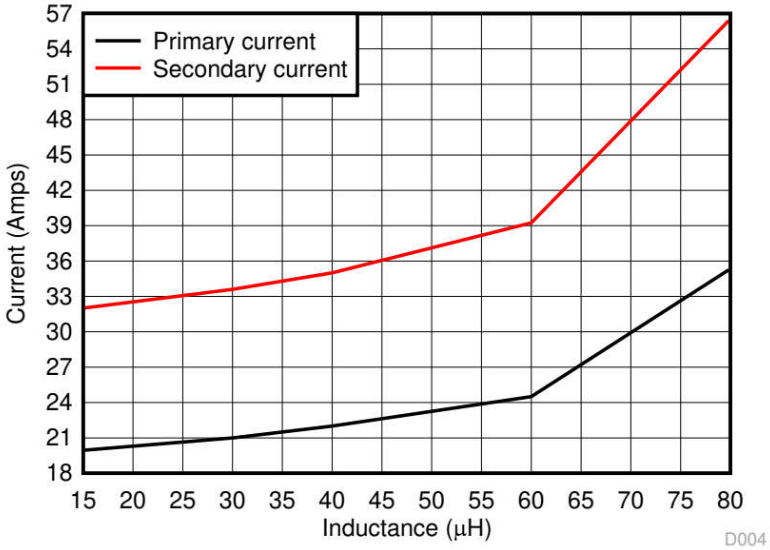
- The sweet spot for operating in soft switching is when M is unity
- The leakage inductance also determines the regions where soft switching is kept

✓ Having inductive current still circulating in light load ensures wide ZVS operation

❖ Too high an inductive current increases rms content and conduction losses



$$M = \frac{V_{out}}{NV_{in}}$$

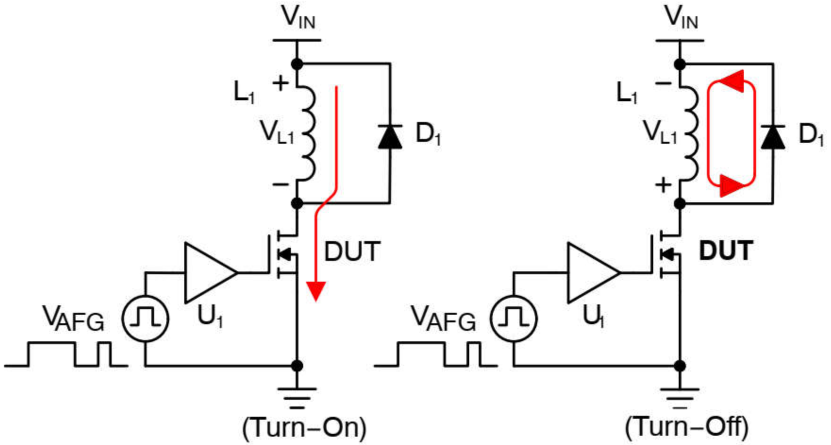


Design Guide: TIDA-010054 - Bidirectional, Dual Active Bridge Reference Design for Level 3 Electric Vehicle Charging Stations

Double-Pulse Test

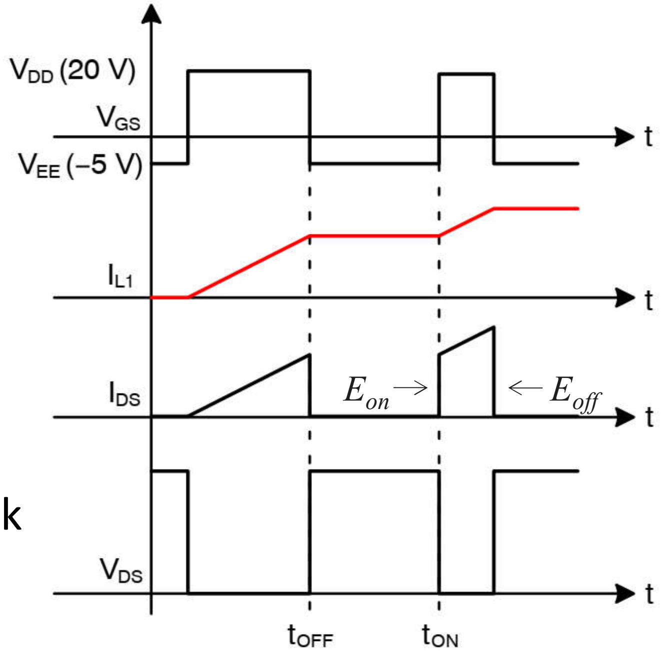
- Transistors manufacturers provide curves graphing turn-on and -off losses
- These losses are evaluated using the double-pulse test

✓ Scaling to the operating parameters is necessary to evaluate the dissipated power



(Turn–On)

(Turn–Off)

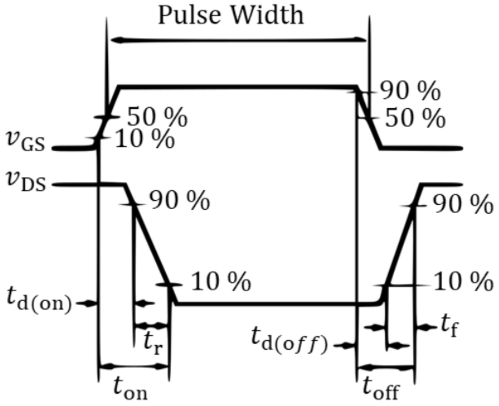


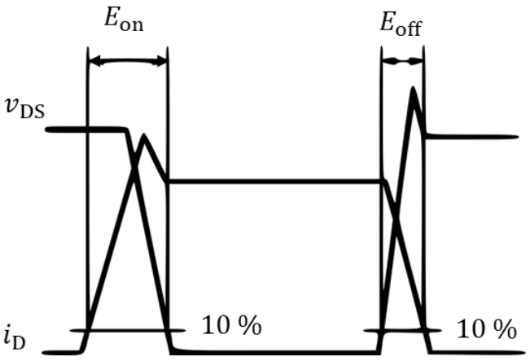
- A first pulse energizes the inductor at a given peak
- A second pulse ensures CCM operation

SiC MOSFETs: Gate Drive Optimization, TND6237-D, onsemi application note

Evaluating Turn-On and Off Losses

- Transition losses are evaluated according to the IEC 60747-8 standard (MOSFETs)





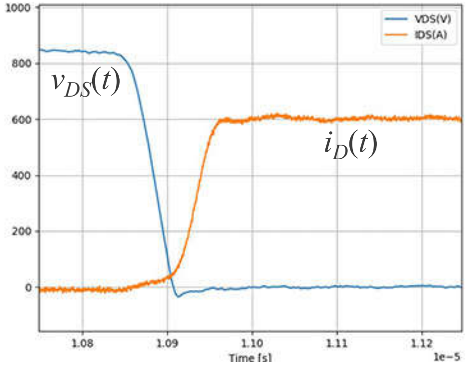
$$E_{on} = \int_{t_1}^{t_2} i_D(t) \cdot v_{DS}(t) \cdot dt$$

$$E_{off} = \int_{t_3}^{t_4} i_D(t) \cdot v_{DS}(t) \cdot dt$$

Typical values found in a data-sheet

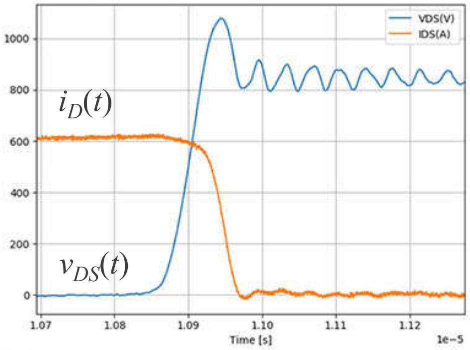
E_{ON}	Turn-On Switching Energy (Body Diode FWD)		246		µJ
E_{OFF}	Turn Off Switching Energy (Body Diode FWD)		99		

V_{DS} = 600 V, V_{GS} = -4 V/15 V, I_D = 35 A,
$R_{G(ext)}$ = 2.5Ω, L= 59 µH, T_J = 150°C



Turn-on losses

Wolfspeed 300-kW Inverter



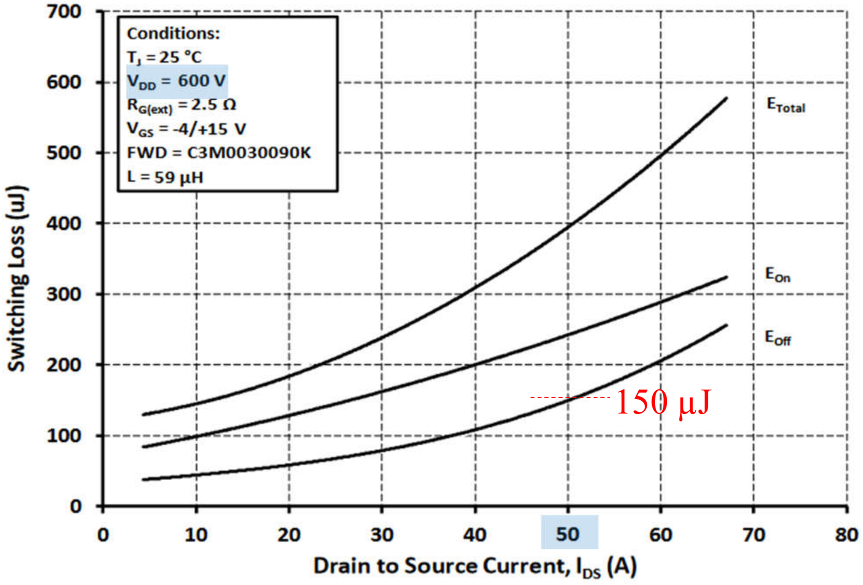
Turn-off losses

J. Schweickhardt et al., *Tips & Tricks on Double-Pulse testing*, Application Note Rohde & Schwarz

Scaling Losses in the SiC Transistors

- Transistors manufacturers provide curves plotting turn-on and -off losses
- These losses are evaluated using the double-pulse test in *ideal* conditions

✓ Scaling to the operating parameters is necessary to evaluate the dissipated power



C3M0030090K

Primary-side losses

$$P_{off,\mathrm{prim}} = F_{sw} E_{off} \cdot k_2 \qquad k_2 = \frac{I_{\mathrm{prim}} V_{in}}{V_{DS} I_D}$$

Secondary-side losses

$$P_{off,\mathrm{sec}} = F_{sw} E_{off} \cdot k_1 \qquad k_1 = \frac{I_{\mathrm{sec}} V_{out}}{V_{DS} I_D}$$

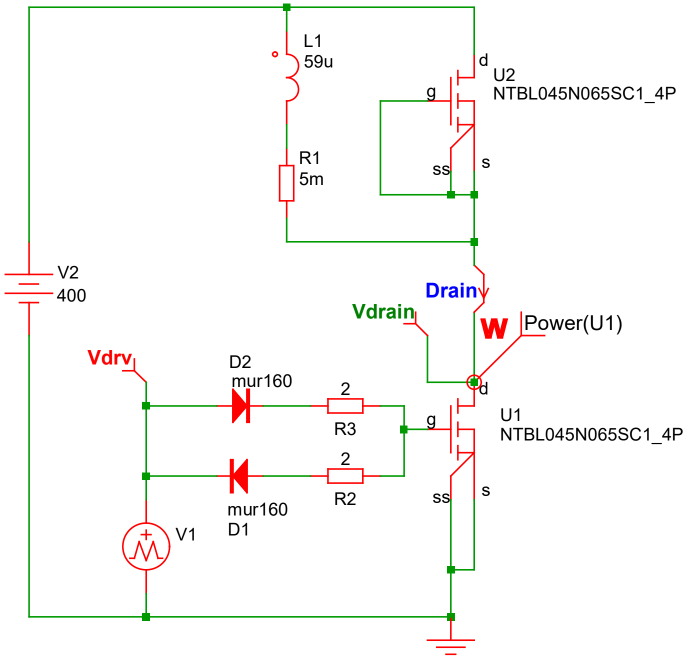
$$F_{sw} = 100\ \mathrm{kHz} \quad V_{out} = 500\ \mathrm{V}$$

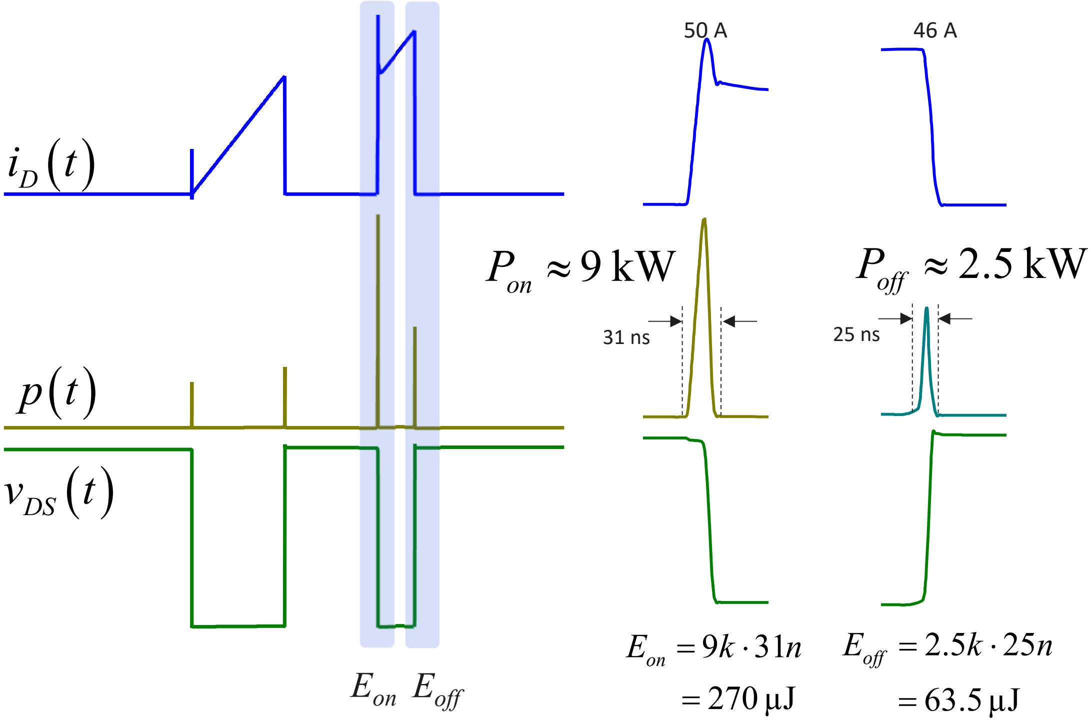
$$E_{off} = 150\ \mu\mathrm{J} \quad I_{\mathrm{sec}} = 35\ \mathrm{A}$$

➡ $$P_{off,\mathrm{sec}} = 100k \times 150u \times \frac{35 \times 500}{50 \times 600} = 10.5\ \mathrm{W}$$

SIMetrix® Double Pulse Simulation

- *onsemi* provides scalable SPICE models that can be imported into a simulator
- It is possible to measure switching losses in test conditions reflecting the application





$P_{on} \approx 9\,\mathrm{kW}$

$P_{off} \approx 2.5\,\mathrm{kW}$

$$E_{on} = 9k \cdot 31n = 270\,\mu\mathrm{J}$$

$$E_{off} = 2.5k \cdot 25n = 63.5\,\mu\mathrm{J}$$

Agenda

- Phase-Shift Control
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Small-Signal Model of the DAB – Output Sources

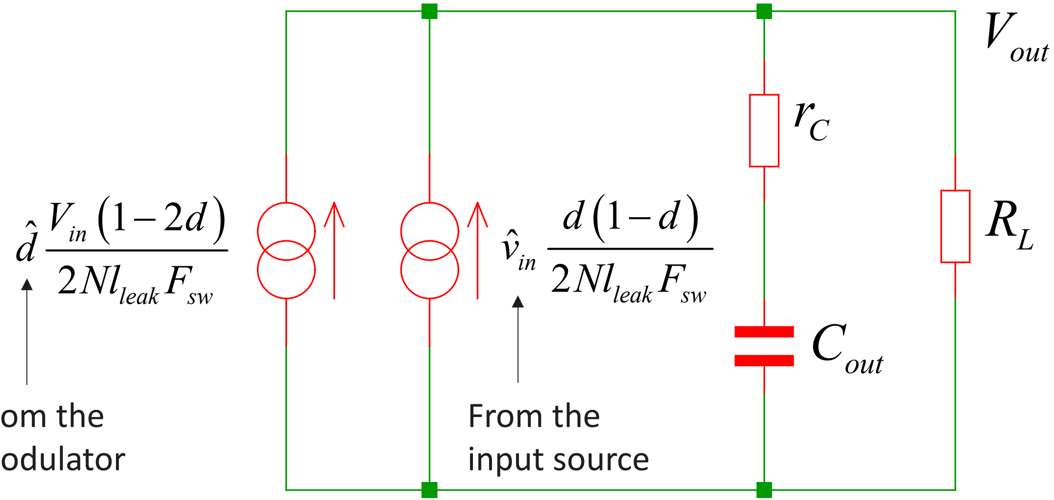
- A simple 1st-order model can easily be determined via an averaging process
  - Express the output power and differentiate with respect to control variables d and V_{in}

Large-signal equation

$$I_{out} = \frac{P_{out}}{V_{out}} = \frac{(1-d)\,dT_{sw}V_{in}}{2l_{leak}N}$$

$$\Downarrow$$

$$\hat{i}_{out} = \frac{\partial I_{out}(d)}{\partial d}\hat{d} + \frac{\partial I_{out}(V_{in})}{\partial V_{in}}\hat{v}_{in}$$



$$\hat{d}\,\frac{V_{in}(1-2d)}{2Nl_{leak}F_{sw}}$$

$$\hat{v}_{in}\,\frac{d(1-d)}{2Nl_{leak}F_{sw}}$$

From the modulator

From the input source

- You obtain the control-to-output and input-to-output transfer functions

Y. Guan et al., An Active Damping Strategy for Input Impedance of Bidirectional Dual Active Bridge DC-DC Converter: Modelling, Shaping, Design, and Experiment, IEEE Transactions on Industrial Electronics, Vol. 68, Issue 2, Feb. 2021

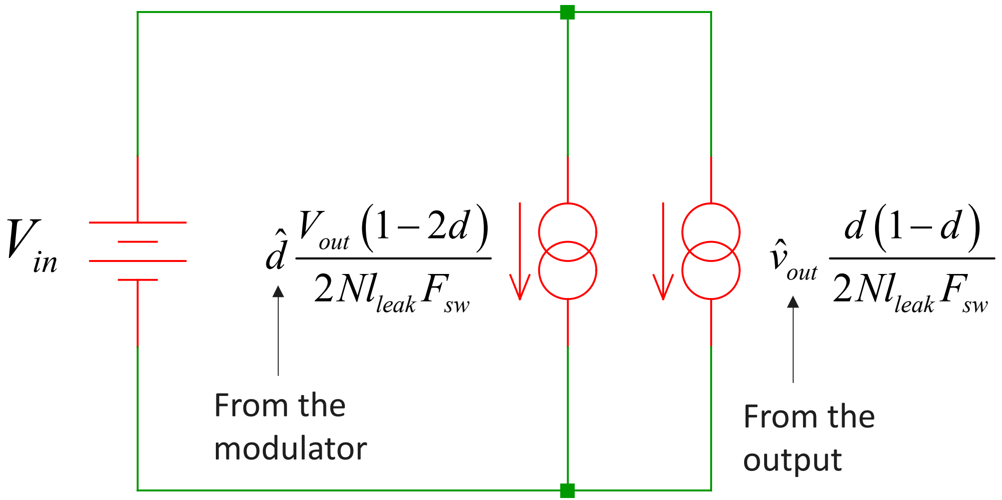
Small-Signal Model of the DAB – Input Sources

- Proceed with a similar process for the input side
- Express the input power and differentiate with respect to d and V_{out}

Large-signal equation (η = 100%)

$$I_{in} = \frac{P_{in}}{V_{in}} = \frac{(1-d)\,d T_{sw} V_{out}}{2 l_{leak} N}$$

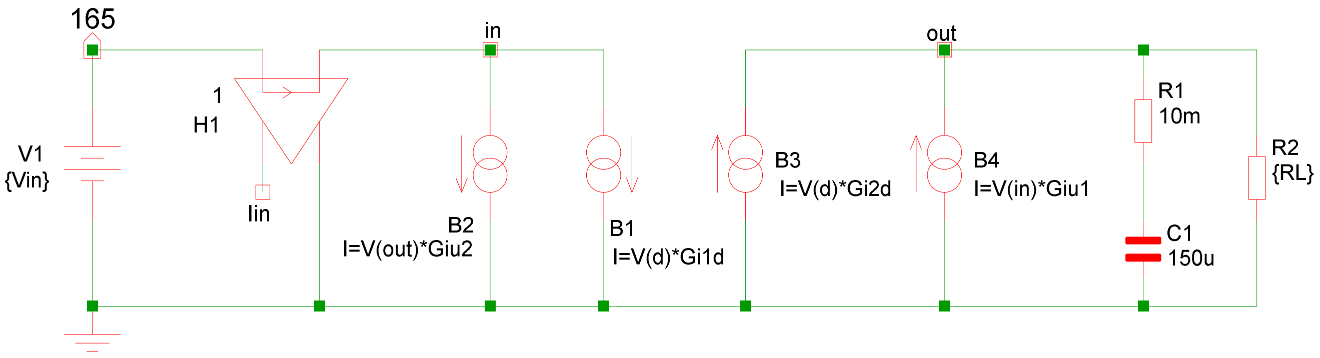
$$\hat{i}_{in} = \frac{\partial I_{in}(d)}{\partial d}\hat{d} + \frac{\partial I_{in}(V_{out})}{\partial V_{out}}\hat{v}_{out}$$

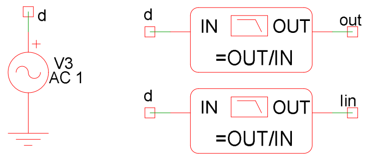


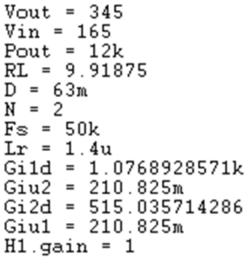
- It is easy to derive the small-signal open-loop input impedance of the DAB

Checking Responses with SIMetrix

- A SPICE model can be assembled with these four current sources
- It becomes easy to plot all needed transfer functions







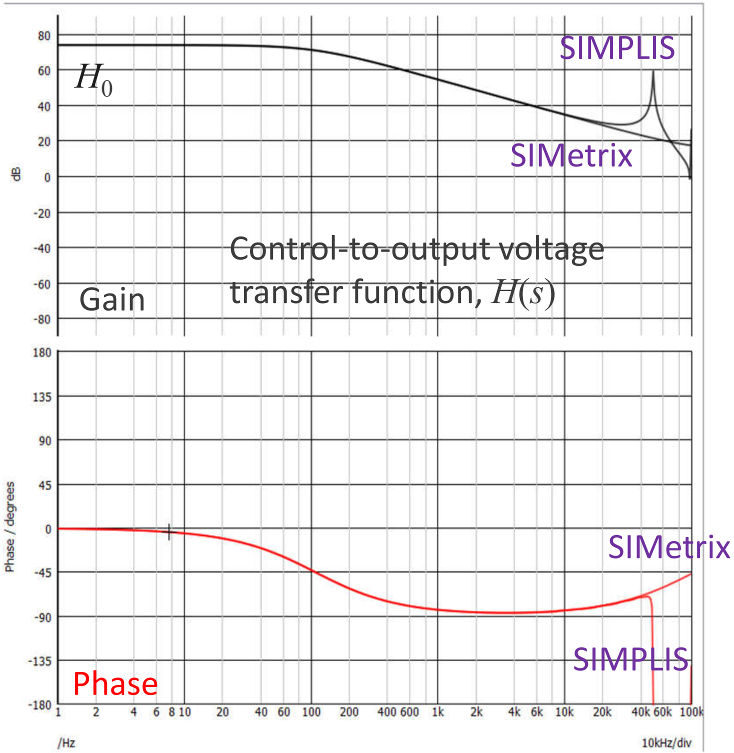
```
Vout = 345
Vin = 165
Pout = 12k
RL = 9.91875
D = 63m
N = 2
Fs = 50k
Lr = 1.4u
Gi1d = 1.0768928571k
Giu2 = 210.825m
Gi2d = 515.035714286
Giu1 = 210.825m
H1.gain = 1
```

```
*
.PARAM Vout=345
.PARAM Vin=165
.PARAM Pout=12k
.PARAM RL={Vout^2/Pout}
*
.PARAM D=0.063
*
* DAB Components values
*
.PARAM N=2
.PARAM Fs=50k
.PARAM Lr=1.4u
*
.PARAM Gi1d={Vout*(1-2*D)/(2*N*Fs*Lr)}
.PARAM Giu2={D*(1-D)/(2*N*Fs*Lr)}
.PARAM Gi2d={Vin*(1-2*D)/(2*N*Fs*Lr)}
.PARAM Giu1={D*(1-D)/(2*N*Fs*Lr)}
*
{'*'}
{'*'}
{'*'} Gi1d = {Gi1d}
{'*'} Giu2 = {Giu2}
{'*'} Gi2d = {Gi2d}
{'*'} Giu1 = {Giu1}
{'*'}
{'*'}
*
```

SIMetrix and SIMPLIS Ac Responses

- The two simulators deliver identical responses for these transfer functions:

✓ Control-to-output voltage and control-to-input current



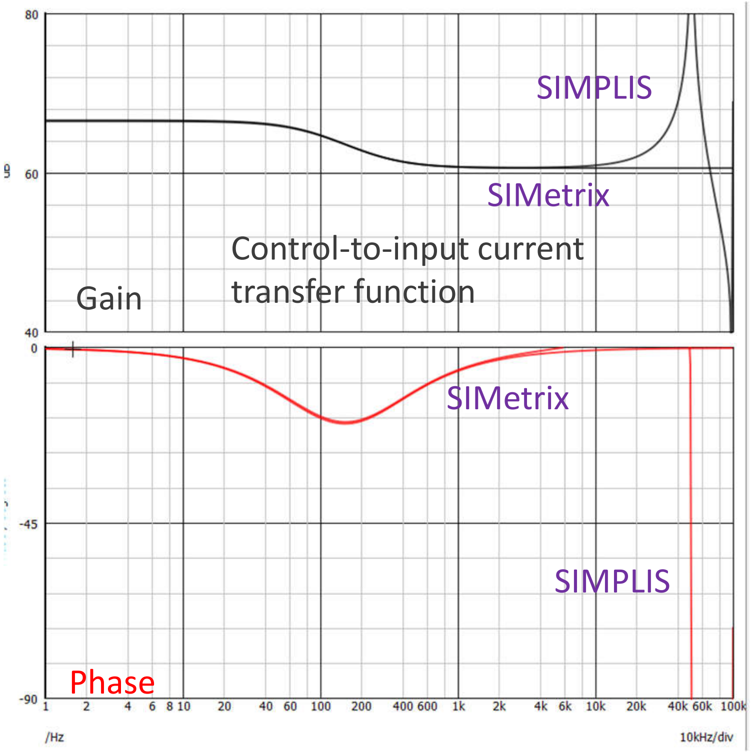
Control-to-output voltage transfer function, $H(s)$

$$H(s) = H_0 \frac{1 + \dfrac{s}{\omega_z}}{1 + \dfrac{s}{\omega_p}}$$

$$H_0 = \frac{V_{out}}{V_p} \frac{(1-2D)}{(1-D)D}$$

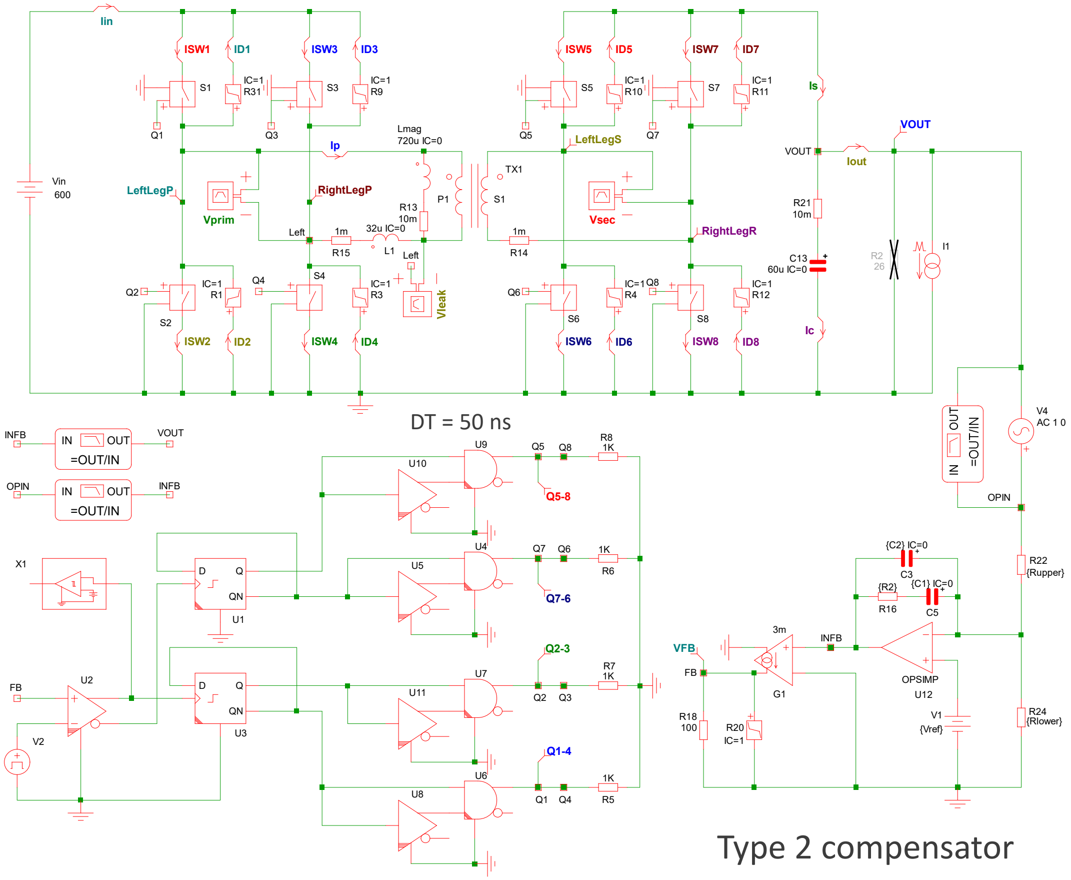
$$\omega_z = \frac{1}{r_C C_{out}}$$

$$\omega_p = \frac{1}{C_{out} R_L}$$



Control-to-input current transfer function

Compensation of the Dual Active Bridge



DT = 50 ns

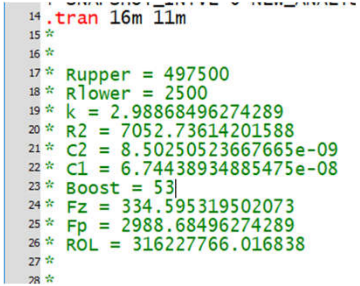
Type 2 compensator

Extracted from plant TF

```
.VAR Gfc=38 * magnitude at crossover *
.VAR PS=-83 * phase lag at crossover *
*
* Enter Design Goals Information Here *
*
.VAR fc=1k * targeted crossover *
.VAR PM=60 * choose phase margin at crossover
*
*
* Enter the Values for Vout and Bridge Bias
Current *
*
.VAR Vout=500
.VAR Ibias=1m
.VAR Vref=2.5
.VAR Rlower=Vref/Ibias
.VAR Rupper=(Vout-Vref)/Ibias
*
* Do not edit the below lines *
.VAR boost=PM-PS-90
.VAR G=10^(-Gfc/20)
.VAR k=tan((boost/2+45)*pi/180)
.VAR fp=fc*k
.VAR fz=fc/k
.VAR C2=1/(2*pi*fc*G*k*Rupper)
.VAR C1=C2*(k^2-1)
.VAR R2=k/(C1*2*pi*fc)
*
* Choose op amp characteristics *
*
.GLOBALVAR AOL=90 * open-loop gain in dB *
.GLOBALVAR POLE=30 * low-frequency pole *
.GLOBALVAR VHIGH=5 * upper output level *
.GLOBALVAR VLOW=100m * lower output level
*
*
* Do not edit these lines *
.GLOBALVAR gm=100u
.GLOBALVAR GAIN={10^(AOL/20)}
.GLOBALVAR COL={1/(6.28*(GAIN/100u)*POLE)}
.GLOBALVAR ROL={GAIN/100u}
```

```
*
{ '*' }
{ '*' }
{ '*' } Rupper = {Rupper}
{ '*' } Rlower = {Rlower}
{ '*' } k = {k}
{ '*' } R2 = {R2}
{ '*' } C2 = {C2}
{ '*' } C1 = {C1}
{ '*' } Boost = {boost}
{ '*' } Fz = {Fz}
{ '*' } Fp = {Fp}
{ '*' } ROL = {ROL}
{ '*' }
{ '*' }
*
```

Automatically-calculated values

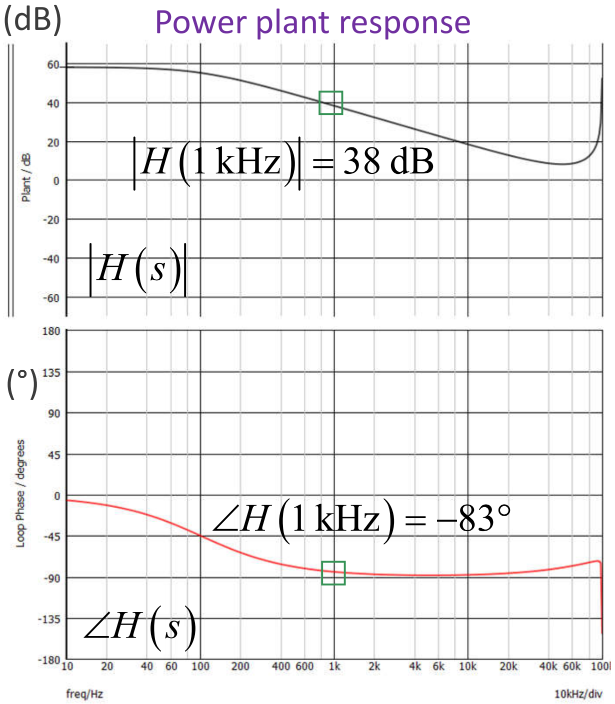


```
.tran 16m 11m
*
*
*  Rupper = 497500
*  Rlower = 2500
*  k = 2.98868496274289
*  R2 = 7052.73614201588
*  C2 = 8.50250523667665e-09
*  C1 = 6.74438934885475e-08
*  Boost = 53
*  Fz = 334.595319502073
*  Fp = 2988.68496274289
*  ROL = 316227766.016838
*
*
```

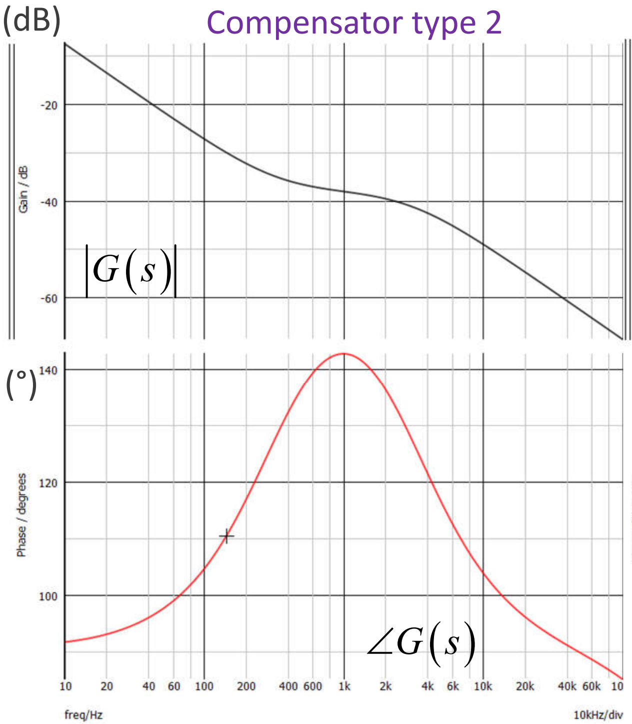
Control-to-Output Transfer Function

- Read the ac response magnitude and phase graphs and extract data at crossover
- Place a pole and a zero to boost the phase at crossover and meet phase margin goals

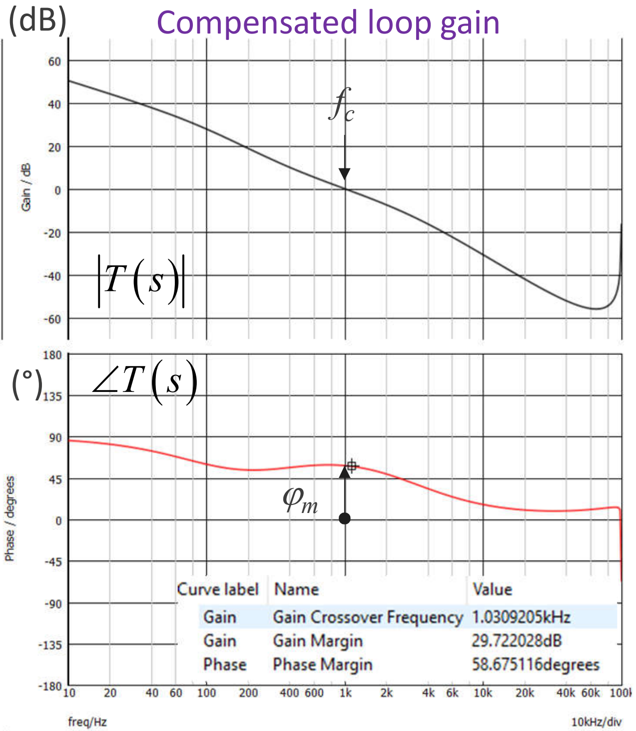
Power plant response



Compensator type 2



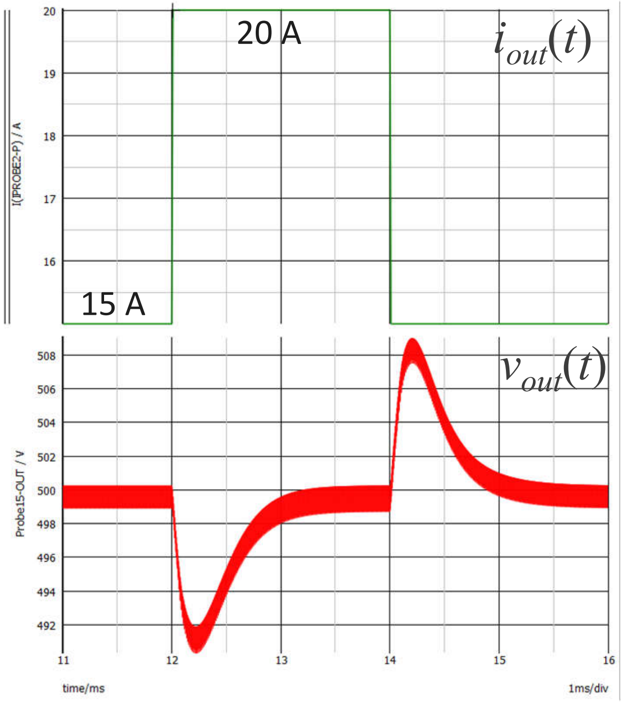
Compensated loop gain

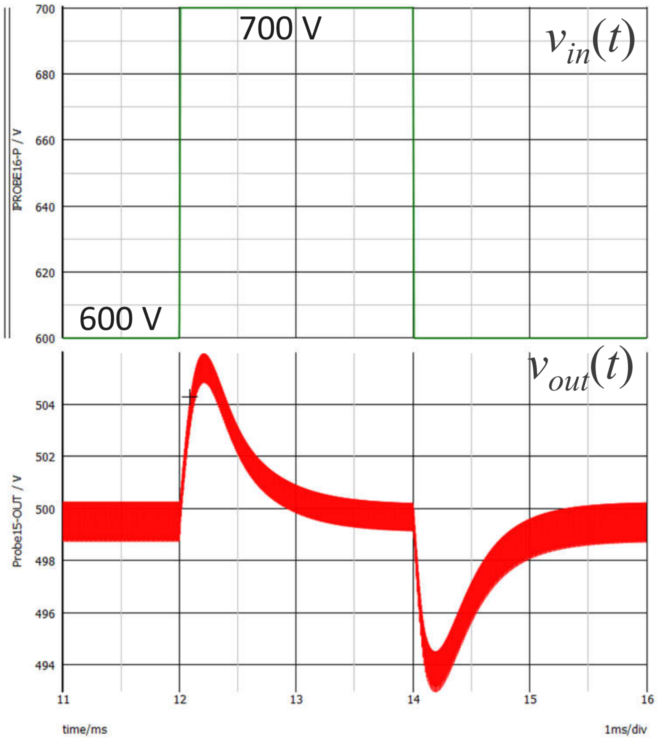


Curve label	Name	Value
Gain	Gain Crossover Frequency	1.0309205kHz
Gain	Gain Margin	29.722028dB
Phase	Phase Margin	58.675116degrees

Step-Load Response of the DAB Converter

- Output power is stepped from 7.5 kW to 10 kW from a 600-V input source
- The input voltage is then stepped from 500 to 600 V with a 10-kW load





Conclusion

- A DAB converter implements two phase-shifted full bridges:

✓ one in the primary and a second one for controlled rectification in the secondary

- A transformer scales the voltage between primary and secondary

✓ A series inductance – usually the leakage term – adjusts the maximum power transfer

- Power is controlled by driving the phase between the two bridges working at 50% duty ratio

✓ Depending on the phase polarity (lead or lag), power can flow either way: bidirectionality

✓ Different modulation schemes exist to minimize the circulating rms current

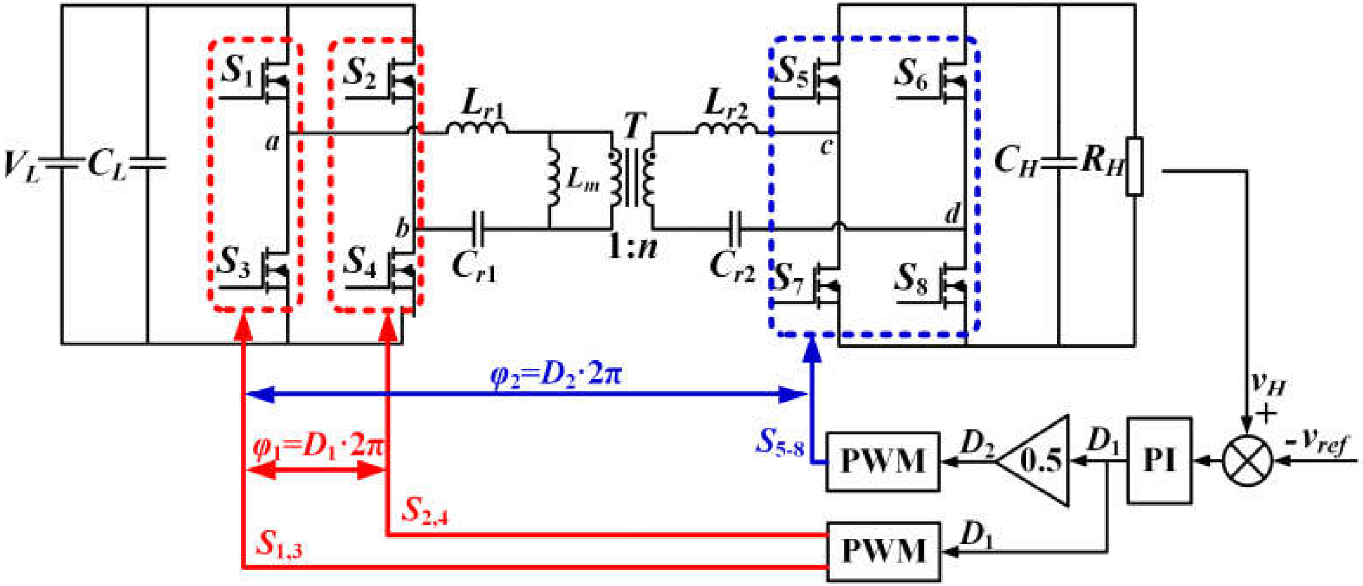
- It is also very important to check for the absence of dc shift in the operating transformer

- In voltage-mode control, the control-to-output transfer function is that of a 1st-order system

Backup slides

The CLLC Converter

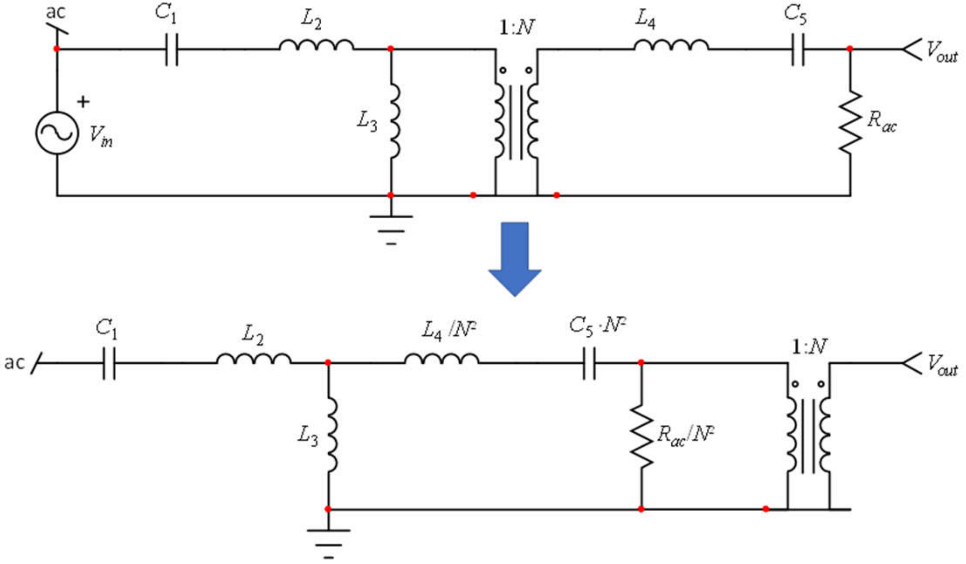
- This is a resonant converter authorizing a bi-directional power flow
- Control can be by frequency modulation or phase shift
- Light-load regulation can be a problem with frequency modulation (flat gain curves)

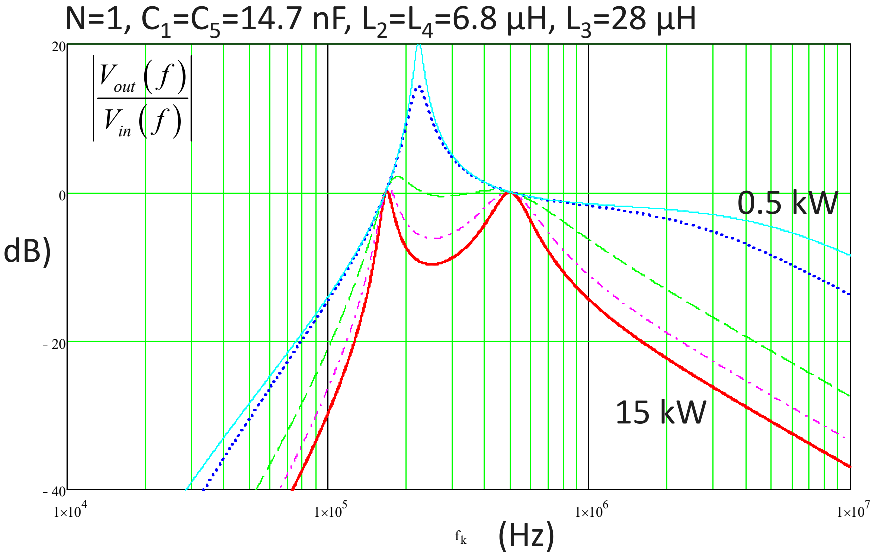


Zhu, T. et al., *Optimization of Extended Phase-Shift Control of Full-Bridge CLLC Resonant Converter with Improved Light-Load Efficiency*, *IEEE Transactions on Power Electronics* , 2020

Dc Transfer Characteristic of the CLLC

- The dc transfer characteristics link the input and output voltages
- Use first-harmonic approximation to derive an equivalent model



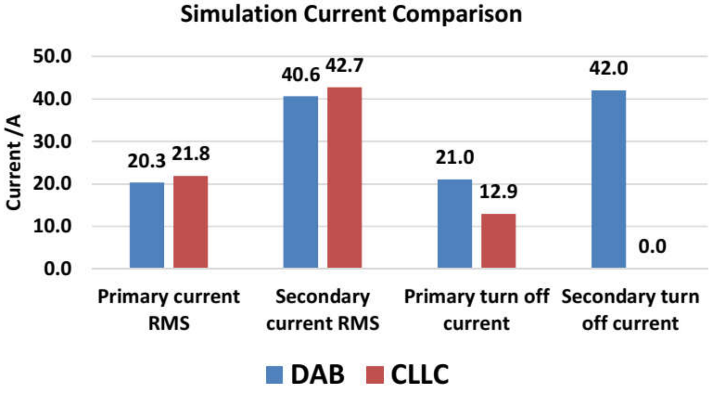

N=1, C_1=C_5=14.7 nF, L_2=L_4=6.8 µH, L_3=28 µH

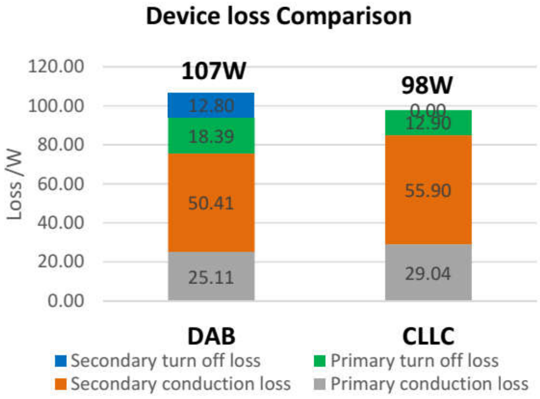
$$H(s) = N\frac{s^3 R_{ac} L_3 C_5 C_1}{1 + sR_{ac}C_5 + s^2\left[C_1\left(L_2+L_3\right)+C_5\left(L_3+L_4\right)\right] + s^3 C_1 C_5 R_{ac}\left(L_2+L_3\right) + s^4 C_1 C_5\left(L_2L_3+L_2L_4+L_3L_4\right)}$$

Comparison between DAB and CLLC

- ■ Fixed-frequency operation
- ✓ 1 or several phase-shift/duty ratio control strategies
- ■ ZVS depends on leakage inductance
- ✓ Increasing leakage for wider ZVS increases rms current
- ■ Magnetizing inductance value is irrelevant
- ■ Turn-off of switches occurs at maximum peak primary current, large E_{off}
- ■ Equal gains in direct and reverse modes

- ■ Variable frequency or phase-shift control
- ✓ Direct or indirect drive via charge control
- ■ ZVS depends on magnetizing inductance
- ✓ ZVS is obtained in a large range
- ✓ Dead-time is selected accordingly
- ✓ Secondary-side switches operate in ZCS
- ■ Turn-off of switches occurs at peak magnetizing current, smaller E_{off}
- ■ Unequal gains in direct and reverse modes





V_{in} = 800 V, V_{out} = 400 V, F_{sw} = 500 kHz, P_{out} = 15 kW, turns ratio 1:0.5, l_{leak} = 2 µH, l_{leak} sec. = 0.5 µH, L_{mag} LLC = 28 µH

S. Zhao, High Frequency Isolated Power Conversion from Medium Voltage Ac to Low Voltage Dc, Thesis in MSEE, Vtech 2017