ON Semiconductor®

Design considerations for a Half-Bridge LLC resonant converter

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

Why an LLC series resonant converter ?

The LCD and PLASMA TV market is growing each year. These and also other markets call for an SMPS unit that can provide these features:

- Output power from 150 W up to 600 W
- Universal mains
- Active or passive PFC (given by needed power)
- Limited width and space, no fan: limited air flow
- Low standby power consumption
- Consumer Electronics market: fierce competition

The above requirements can be fulfilled with an SMPS that provides:

→ High power density
→ Smooth EMI signature
→ Cost effective solution with minimum component count

Benefits of an LLC series resonant converter

- Type of serial resonant converter that allows operation in relatively wide input voltage and output load range when compared to the other resonant topologies

- Limited number of components: resonant tank elements can be integrated to a single transformer – only one magnetic component needed

- Zero Voltage Switching (ZVS) condition for the primary switches under all normal load conditions

- Zero Current Switching (ZCS) for secondary diodes, no reverse recovery losses
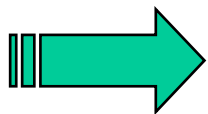
Cost effective, highly efficient and EMI friendly solution for high and medium output voltage converters

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

Configurations of an HB LLC – single res. cap
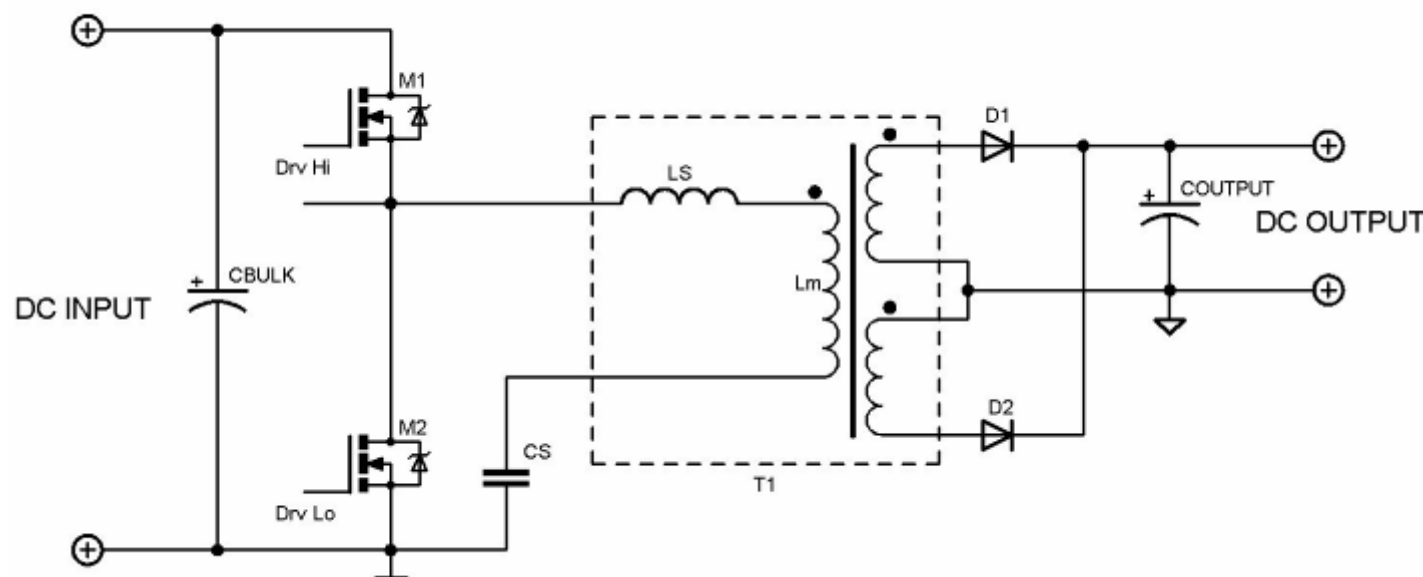
- Higher input current ripple and RMS value
- Higher RMS current through the resonant capacitor
- High voltage (600 – 1500 V) resonant capacitor needed
- Low cost
- Small size / easy layout

Configurations of an HB LLC – split res. cap
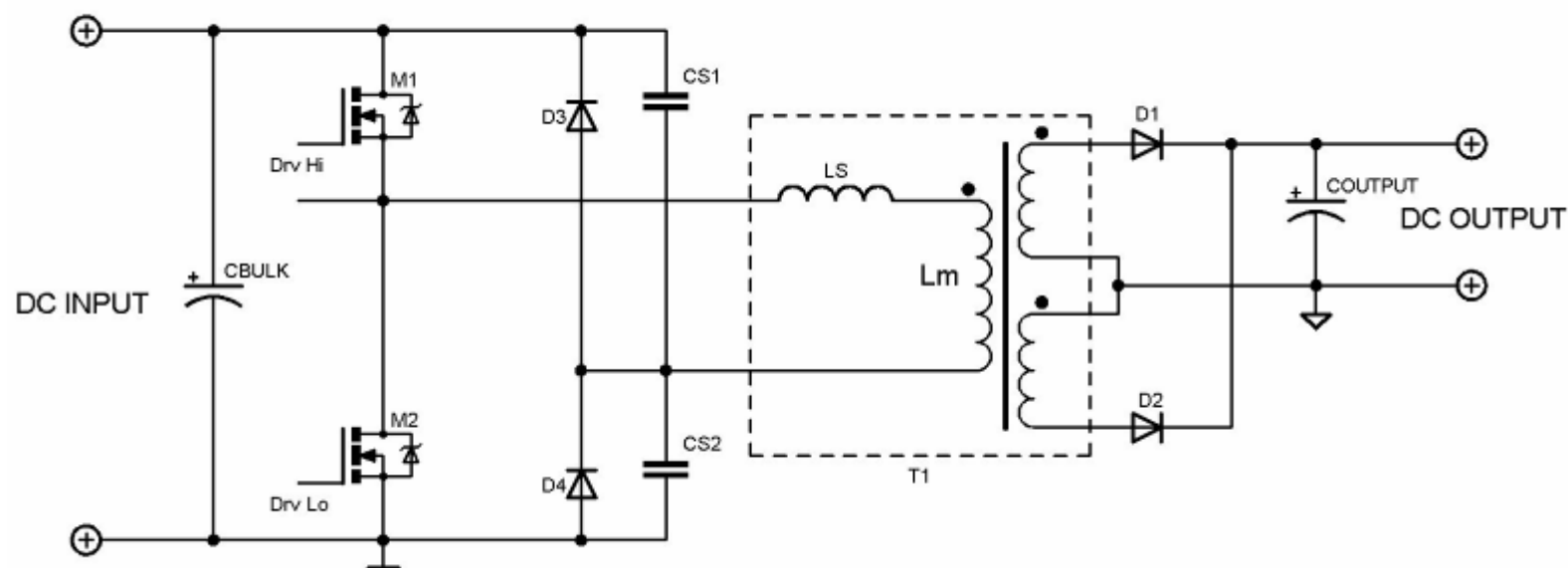
Compared to the single capacitor solution this connection offers:
- Lower input current ripple and RMS value (1/ √ 2)
- Resonant capacitors handle half RMS current
- Capacitors with half capacitance are used
- Low voltage ratings (450 V) for the resonant capacitors when clamping diodes
D3, D4 perform an easy and cheap overload protection

Resonant tank configurations – discrete solution

Resonant inductance is located outside of the transformer

Advantages:

- Greater design flexibility (designer can setup any L_s and L_m value)
- Lower radiated EMI emission
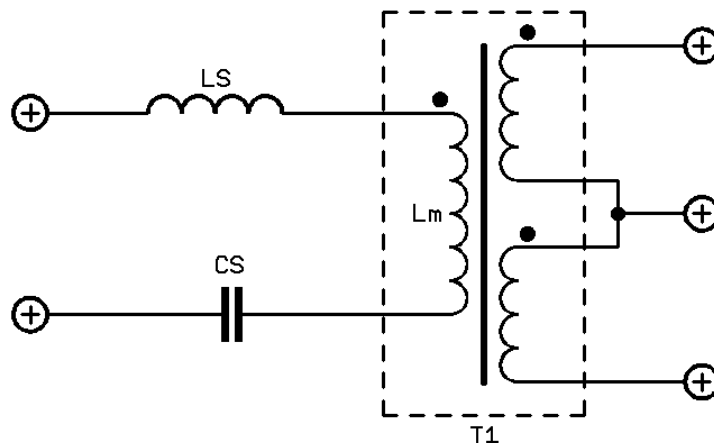


Disadvantages of this solution are:

- Complicated insulation between primary and secondary windings
- Worse cooling conditions for the windings
- More components to be assembled

Resonant tank configurations – integrated solution

Leakage inductance of the transformer is used as a resonant inductance.
Advantages:
- Low cost, only one magnetic component is needed
- Usually smaller size of the SMPS
- Better cooling conditions for transformer windings
- Insulation between primary and secondary side is easily achieved
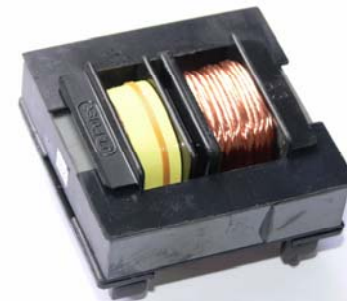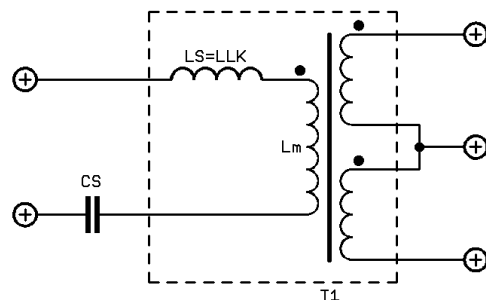
Disadvantages:
- Less flexibility (achievable Ls inductance range is limited)
- Higher radiated EMI emission
- LLC with integrated resonant tank operates in a slightly different way than the solution with discrete L_s
- Strong proximity effect in the primary and secondary windings

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

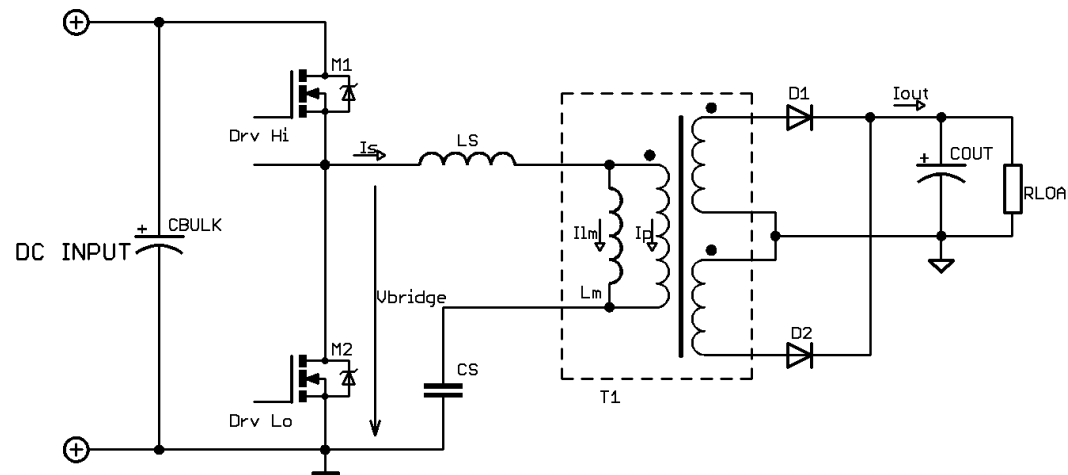# Operating states of the LLC converter

Discrete resonant tank solution

Two resonant frequencies can be defined:

$$F_s = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot L_s}}$$

$$F_{min} = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot (L_s + L_m)}}$$

LLC converter can operate:

a) between F_{min} and F_s

b) direct in F_s

c) above F_s

d) between F_{min} and F_s - overload

e) below F_{min}

Operating states of the LLC converter

c) Operating waveforms for $F_{op} > F_s$ Discrete resonant tank solution
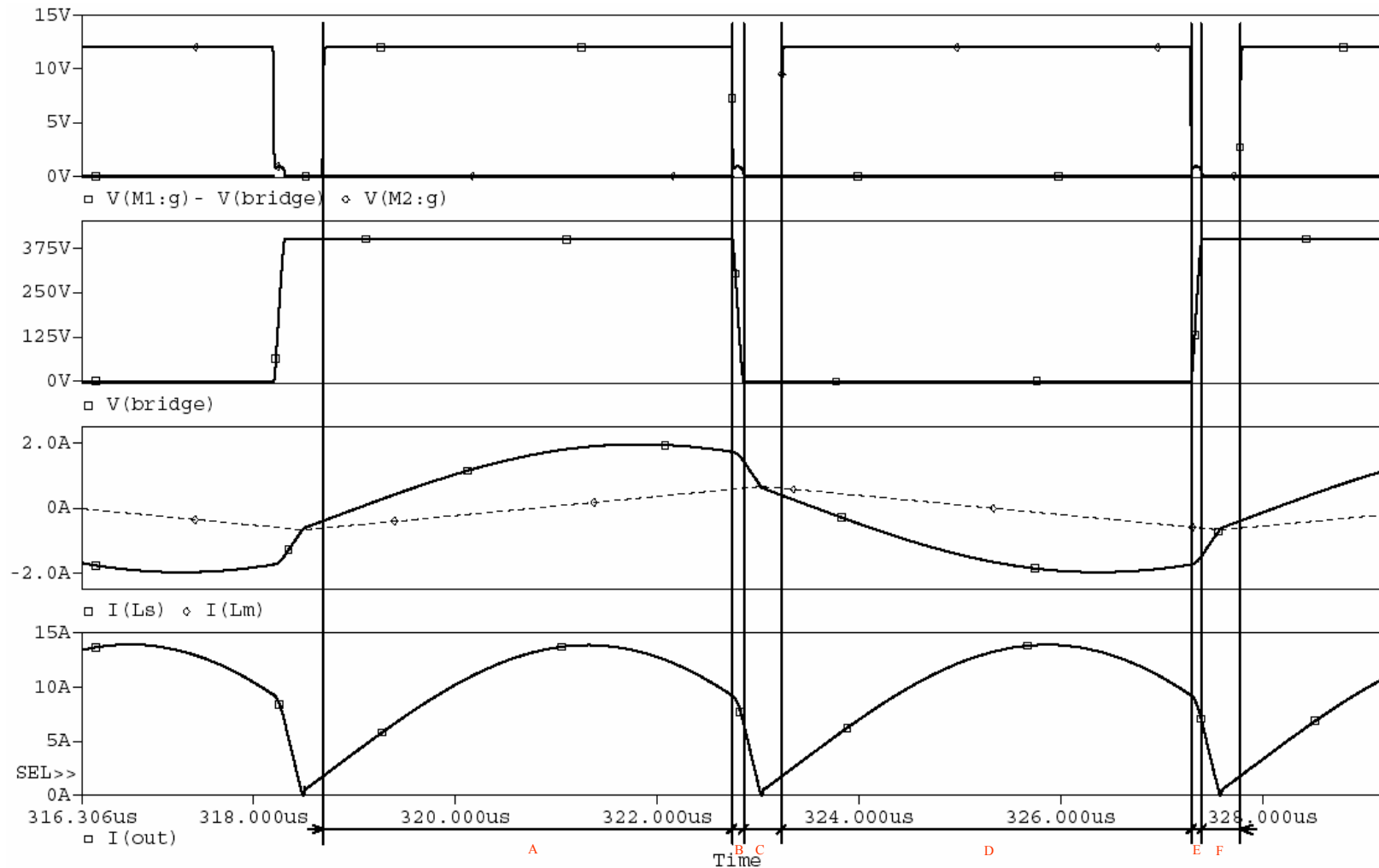
Operating states of the LLC converter

Integrated resonant tank solution

- Integrated resonant tank behaves differently than the discrete resonant tank
- leakage inductance is given by the transformer coupling
- L_{lk} participates only if there is a energy transfer between primary and secondary
- Once the secondary diodes are closed under ZCS, L_{lk} has no energy

$$M = \sqrt{1 - \frac{L_{lk}}{L_m}}$$

Secondary diodes are always turned OFF under ZCS condition in HB LLC.
The resonant inductance L_s and magnetizing inductance L_m never participate in the resonance together as with discrete resonant tank solution!

Operating states of the LLC converter

Integrated resonant tank solution

Two resonant frequencies can be defined:

$$F_s = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot L_s}}$$

$$F_{min} = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot L_m}}$$

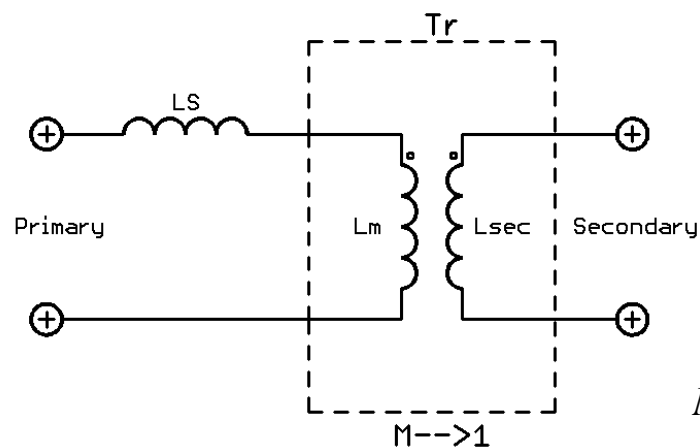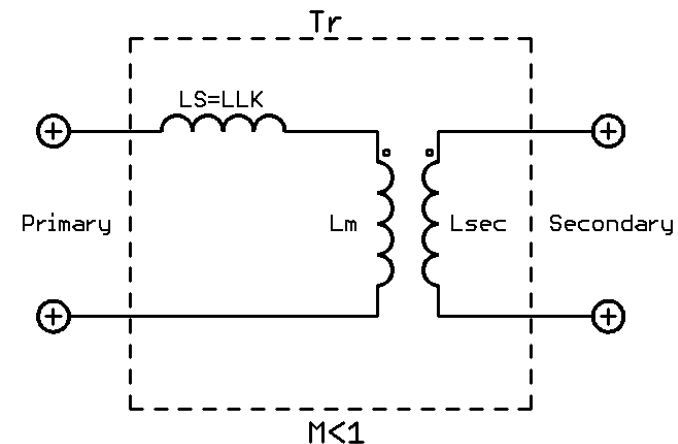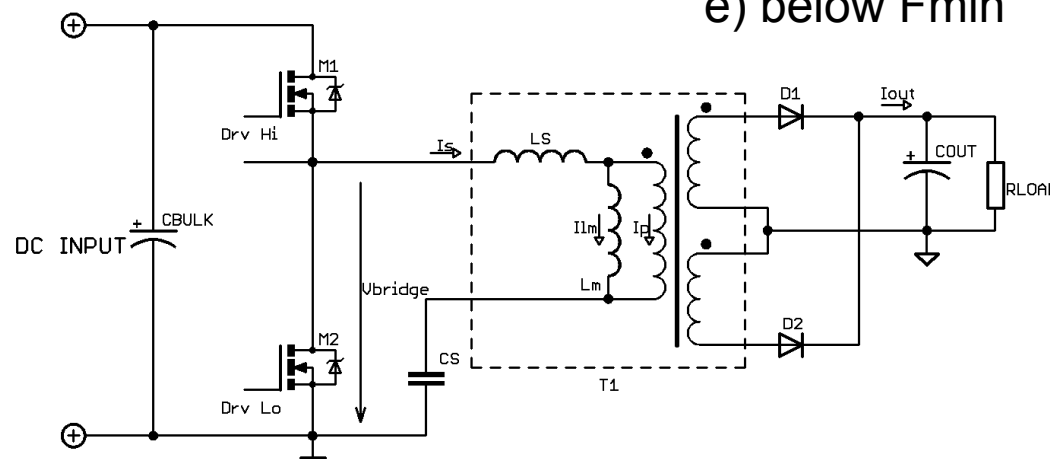**LLC converter can again operate:**

a) between Fmin and Fs

b) direct in Fs

c) above Fs

d) between Fmin and Fs – overload

e) below Fmin

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

LLC converter modeling – equivalent circuit

LLC converter can be described using firs fundamental approximation. Only approximation – accuracy is limited!! Best accuracy is reached around F_s.
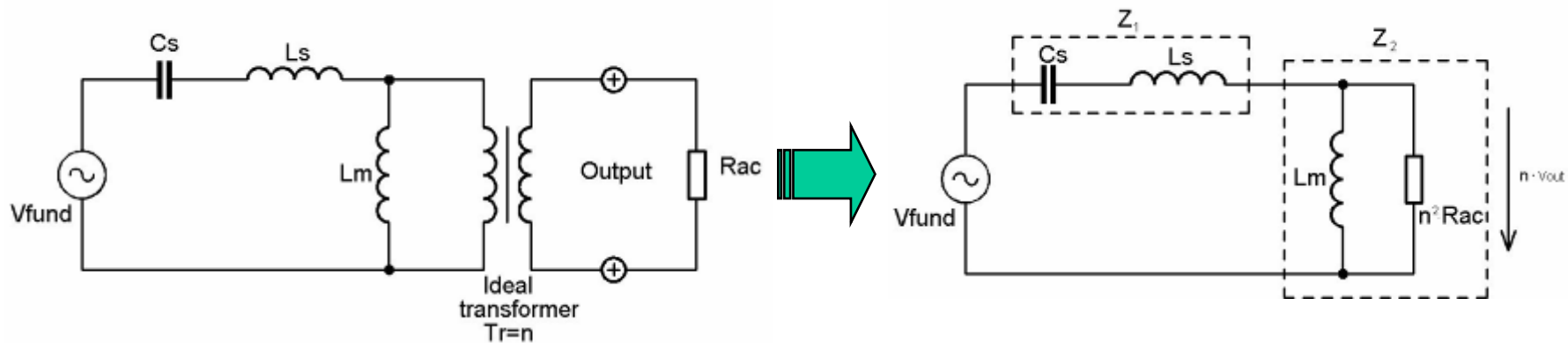
Transfer function of equivalent circuit:

$$G_{ac} = \frac{n \cdot V_{out}}{V_{in}} = \frac{Z_2}{Z_1 + Z_2}$$

Z_1, Z_2 are frequency dependent => LLC converter behaves like frequency dependent divider. The higher load, the L_m gets to be more clamped by R_{ac}. Resonant frequency of LLC resonant tank thus changes between F_s and F_{min}.

LLC converter modeling – equivalent circuit

Real load resistance has to be modified when using fundamental approximation because the real resonant tank is driven by square wave voltage.
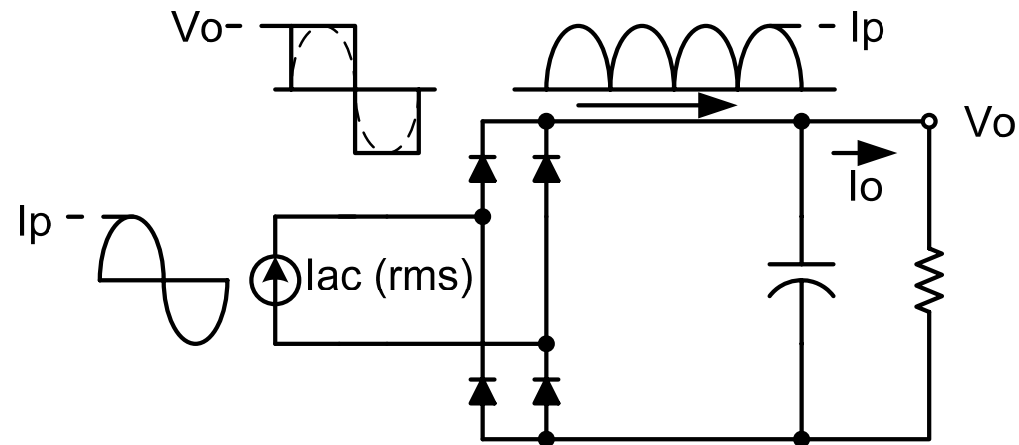
In a full-wave bridge circuit the RMS current is:

$$I_{ac\_RMS} = \frac{\pi}{2\sqrt{2}} I_O$$

Considering the fundamental component of the square wave, the RMS voltage is:

$$V_{ac\_RMS} = \frac{2\sqrt{2}}{\pi} V_O$$

The AC resistance R_{ac} ca be expressed as:

$$R_{ac} = \frac{V_{ac\_RMS}}{I_{ac\_RMS}} = \frac{8}{\pi^2} \frac{E_O}{I_O} = \frac{8}{\pi^2} R_L$$

Resonant tank equations

Quality factor:

$$Q = \frac{n^2 \cdot R_L}{Z_0}$$

← **Load dependent !**

Characteristic impedance:

$$Z_0 = \sqrt{\frac{L_s}{C_s}}$$

Lm/Ls ratio:

$$k = \frac{L_m}{L_s}$$

Gain of the converter:

$$G = \frac{2 \cdot (V_{out} + V_f)}{V_{in}}$$

Series resonant frequency:

$$F_s = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot L_s}}$$

Minimum resonant frequency:

$$F_{min} = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot (L_s + L_m)}}$$

Normalized gain characteristic

Q=200 – Light load

Region 2

Lm/Ls=6

ZVS

Region 1

voltage gain

ZCS

Region 3

Q=0.05 – Heavy load

f / fs

Q=0.05
Q=0.5
Q=1
Q=2
Q=3
Q=4
Q=5
Q=10
Q=20
Q=50
Q=100
Q=200
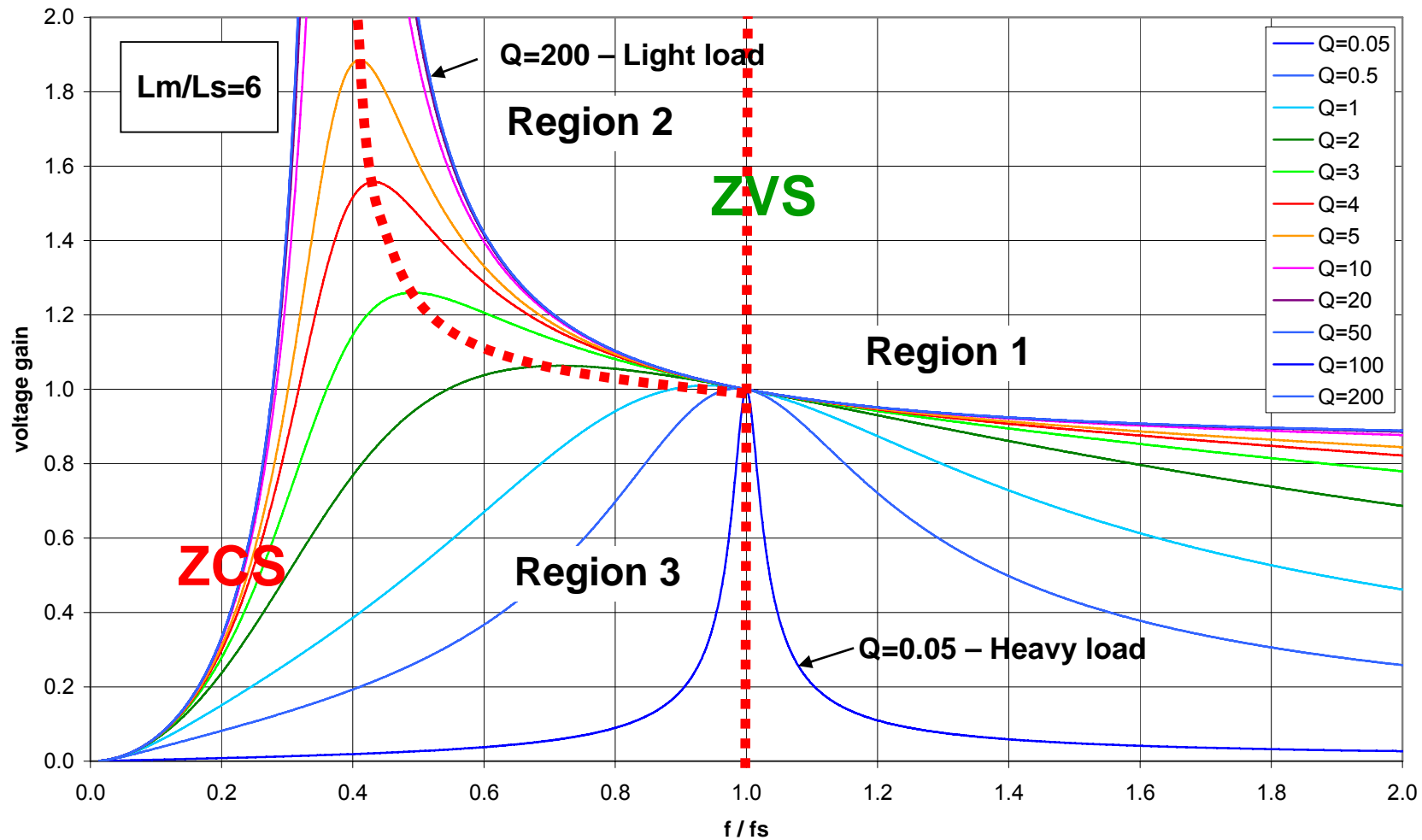
Region 3: ZCS region   **Region 1 and 2: ZVS operating regions**

Gain characteristic discussion

- The desired operating region is on the right side of the gain characteristic (negative slope means – ZVS mode for primary MOSFETs).

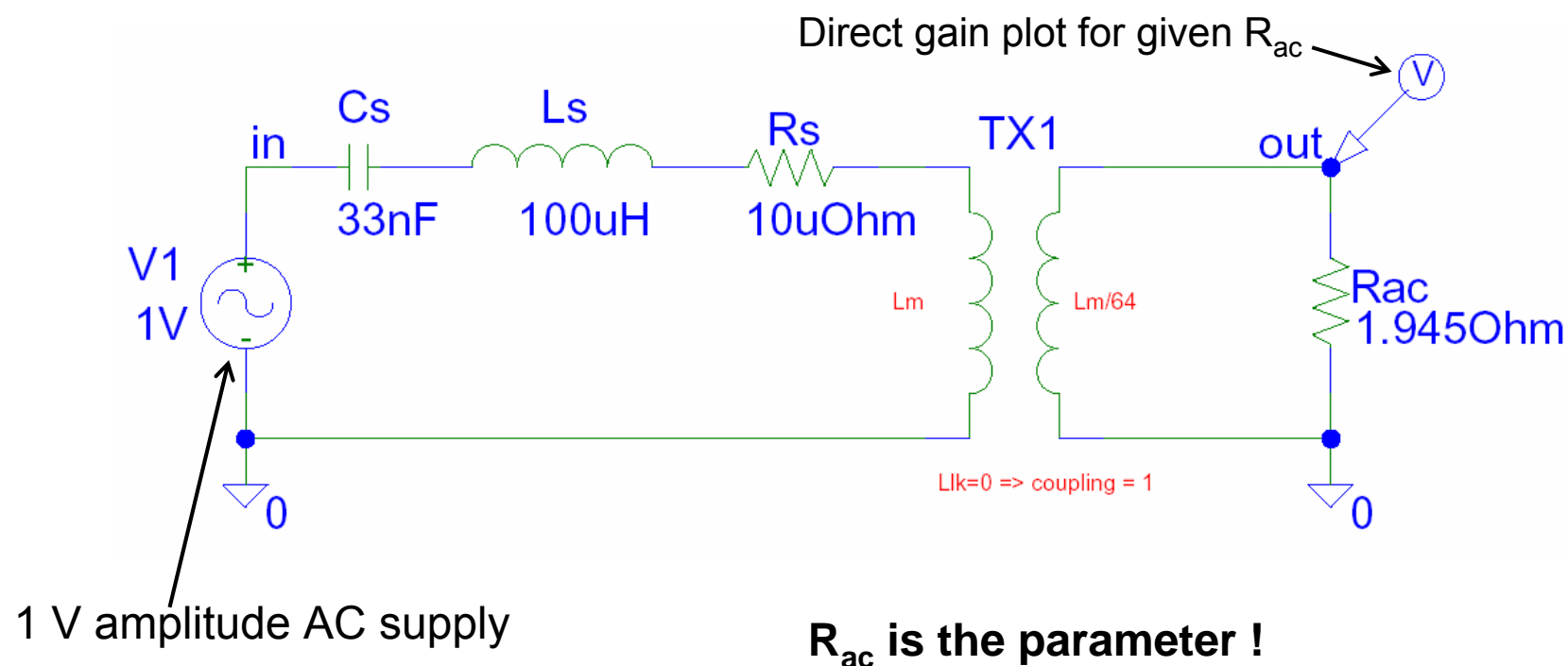-Gain of the LLC converter, which operates in the f_s is 1 (for discrete resonant tank solution) - i.e. is given by the transformer turns ratio. This operating point is the most attractive from the efficiency and EMI point of view – sinusoidal primary current, MOSFETs and secondary diodes optimally used. This operating point can be reached only for specific input voltage and load (usually full load and nominal V_{bulk}).

Gain characteristics shape and also needed operating frequency range is given by these parameters:
- L_m/L_s ratio
- Characteristic impedance of the resonant tank
- Load value
- Transformer turns ratio

How to obtain gain characteristics?

Use fundamental approximation and AC simulation in any simulation software like PSpice, Icap4 etc..

Direct gain plot for given R_{ac}

Cs 33nF
Ls 100uH
Rs 10uOhm
TX1
in
out
V1 1V
Lm
Lm/64
Rac 1.945Ohm
Llk=0 => coupling = 1
0
0

1 V amplitude AC supply

R_{ac} **is the parameter !**

Discrete and integrated tank gain differences
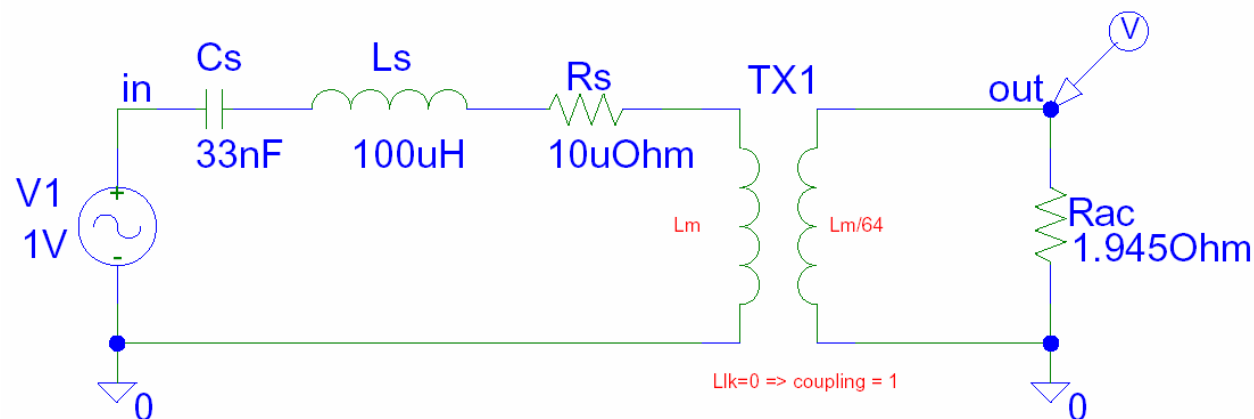
Simulation schematic for discrete solution
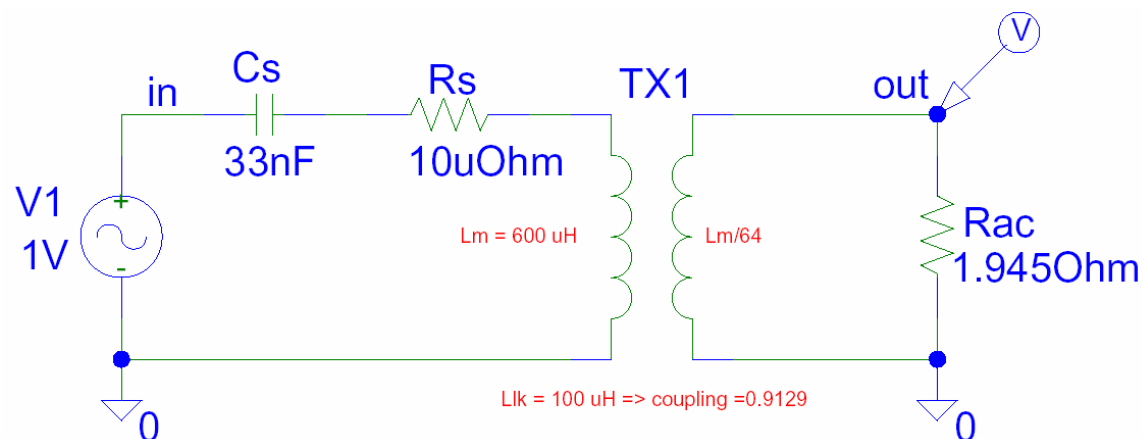
Simulation schematic for integrated solution

Full load Q and k factors optimization

Proper selection of these two factors is the key point for the LLC resonant converter design! Their selection will impact these converter characteristics:

- Needed operating frequency range for output voltage regulation
- Line and load regulation ranges
- Value of circulating energy in the resonant tank
- Efficiency of the converter

The efficiency, line and load regulation ranges are usually the most important criteria for optimization.

Quality factor Q directly depends on the load. It is given by the L_s and C_s components values for full load conditions:

$$Q = \frac{n^2 \cdot R_L}{\sqrt{\dfrac{L_s}{C_s}}}$$

Full load Q and k factors optimization
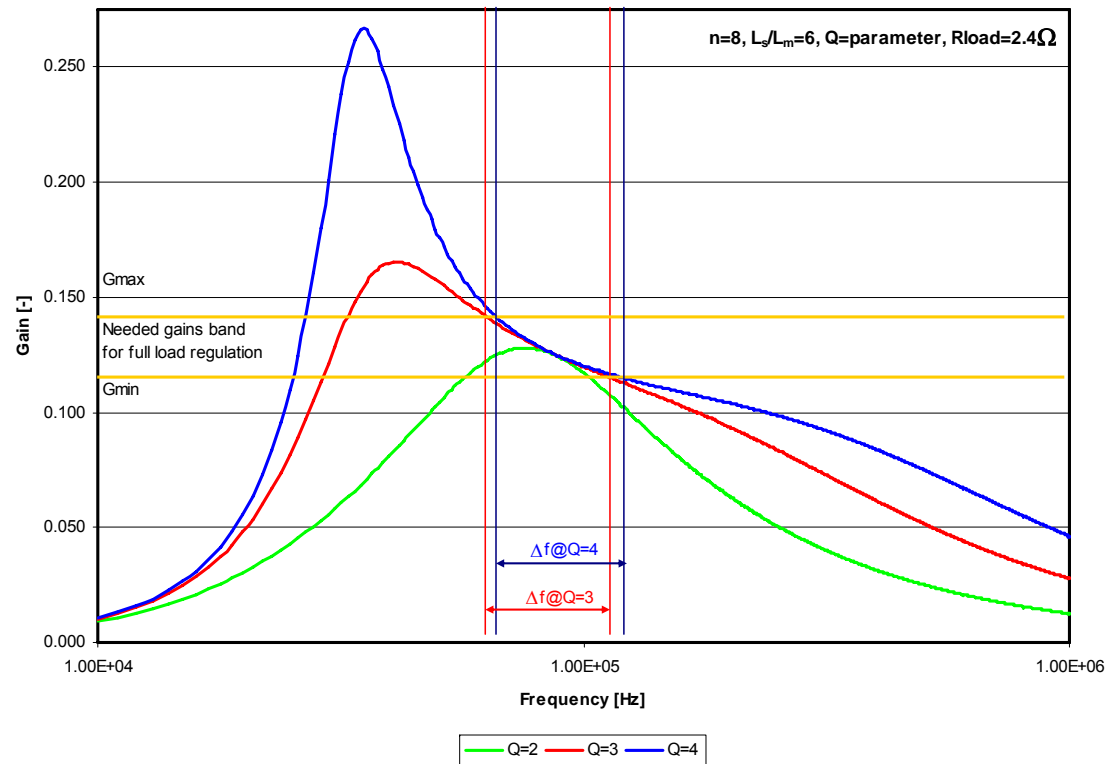
- Higher **Q factor** results in larger F_{op} range
- Characteristic impedance has to be lower for higher Q and given load => higher C_s
- Low Q factor can cause the loss of regulation capability!
- LLC gain characteristics are degraded to the SRC for very low Q values.

Full load Q and k factors optimization
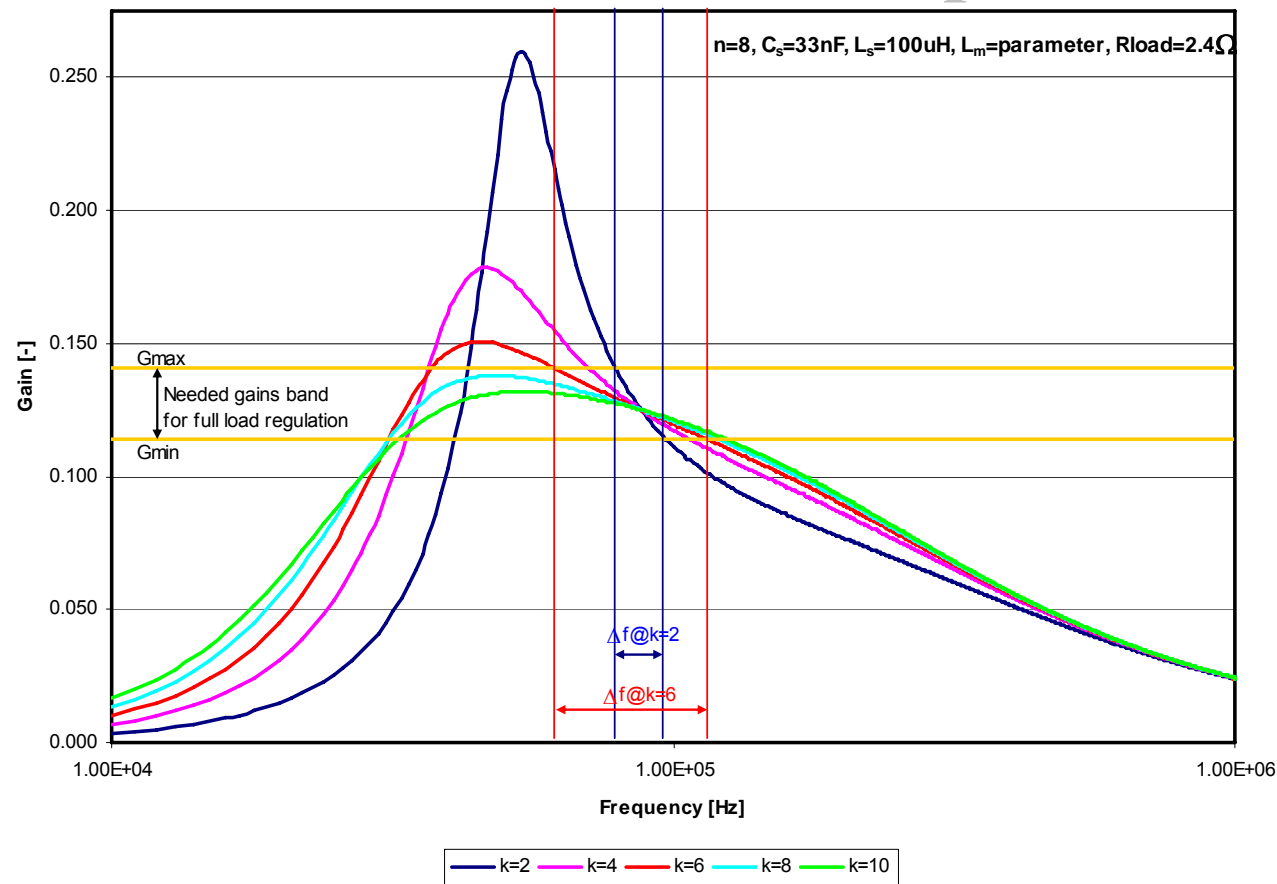
- The $k = L_m/L_s$ ratio dictates how much energy is stored in the L_m.
- Higher **k** will result in the lover magnetizing current and gain of the converter.
- Needed regulation frequency range is higher for larger k factor.

Full load Q and k factors optimization

Practically, the $\mathbf{L_s}$ (i.e. leakage inductance of the integrated transformer version) has only limited range of values and is given by the transformer construction (for needed power level) and turns ratio.

The **Q** factor calculation is then given by the wanted nominal operating frequency f_s.

The **k** factor has to be then calculated to assure gains needed for the output voltage regulation (with line and load changes).

The **k** factor can be set in such a way that converter wont be able to maintain regulation at light loads – skip mode can be easily implemented to lover no load consumption.

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

Primary currents – single resonant cap
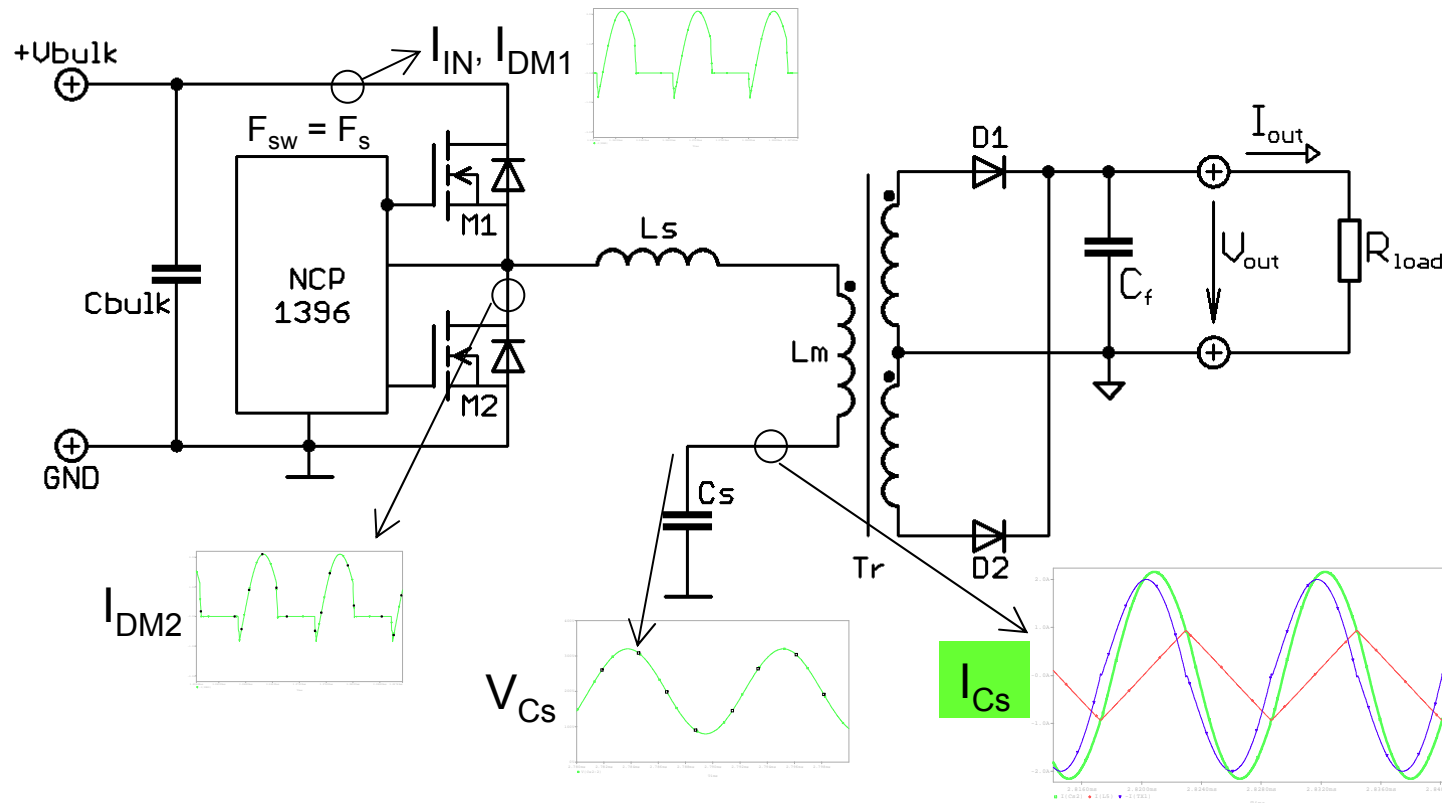
$$I_{C_s} = I_{primary} = \frac{I_{sec}}{n} + I_{L_m} \implies I_{C_s\_RMS} \approx \sqrt{\frac{1}{8} \cdot \left(\frac{I_{out}^2 \cdot \pi^2}{n^2} + \frac{V_{bulk}^2}{16 \cdot L_m^2 \cdot f_{sw}^2} \right)}$$

Primary currents – split resonant cap
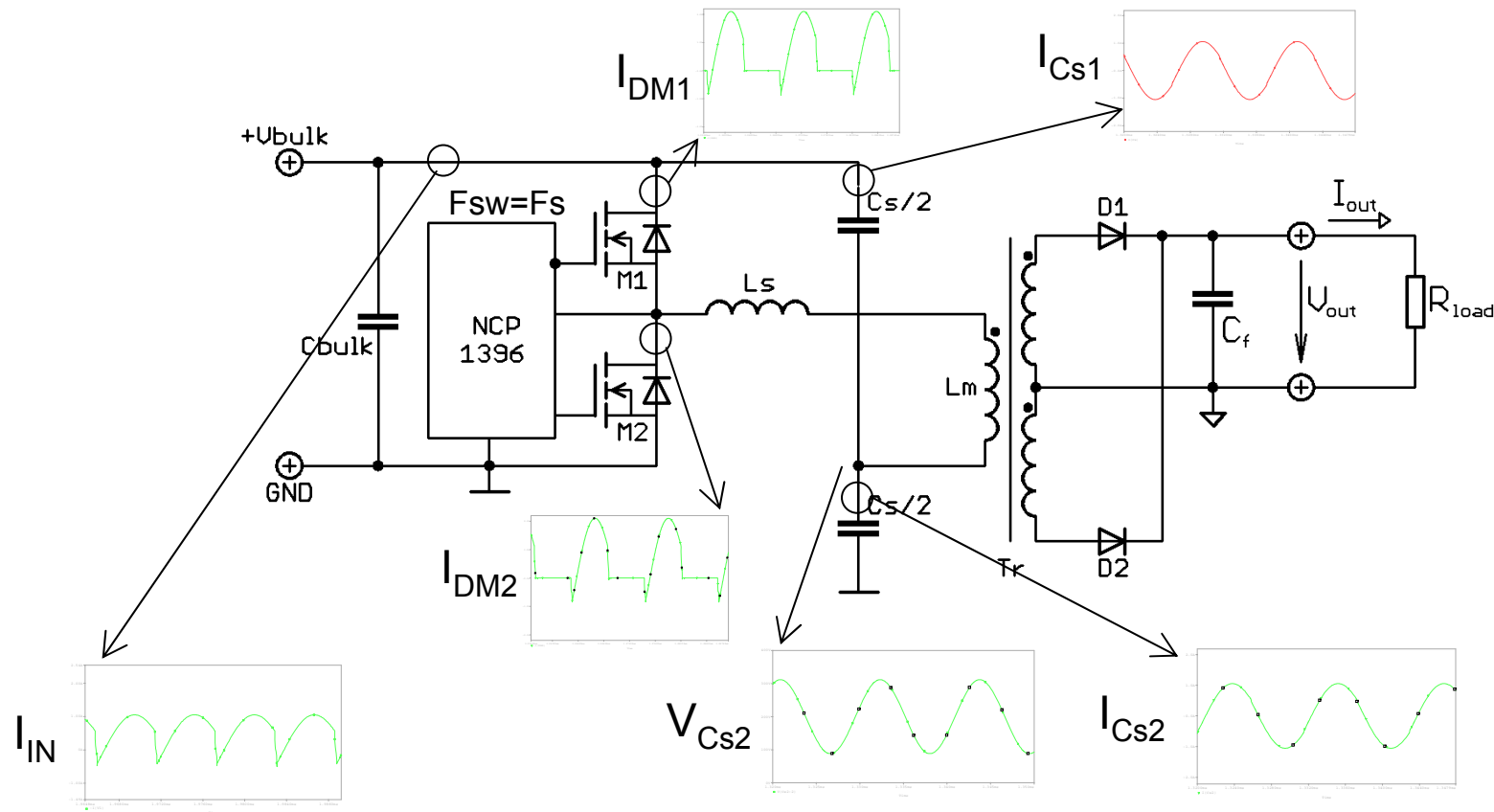
Comparison of Primary Currents

Single and split resonant capacitor solutions - 24 V / 10 A application

Parameter	Single Cap	Split Caps
I_{Cs_Pk}	2.16 A	1.08 A
I_{Cs_RMS}	1.52 A	0.76 A
I_{IN_Pk}	2.16 A	1.08 A
I_{IN_RMS}	1.07 A	0.76 A

- Split solution offers 50% reduction in resonant capacitor current and 30% reduction in input rms current

- Select resonant capacitor(s) for current and voltage ratings

Primary switches dimensioning

- Body diode is conducting during the dead time only (A)
- MOSFET is conducting for the rest of the period (B)
- Turn ON losses are given by Q_g (burned in the driver not in MOSFET)
- MOSFET turns OFF under non-zero current => turn OFF losses

Primary switches dimensioning

MOSFET RMS current calculation

- The body diode conduction time is negligible
- Assume that the MOSFET current has half sinusoid waveform

$$I_{switch\_RMS} \approx \sqrt{\frac{1}{16} \cdot \left(\frac{I_{out}^2 \cdot \pi^2}{n^2} + \frac{V_{bulk}^2}{16 \cdot L_m^2 \cdot f_{sw}^2} \right)}$$
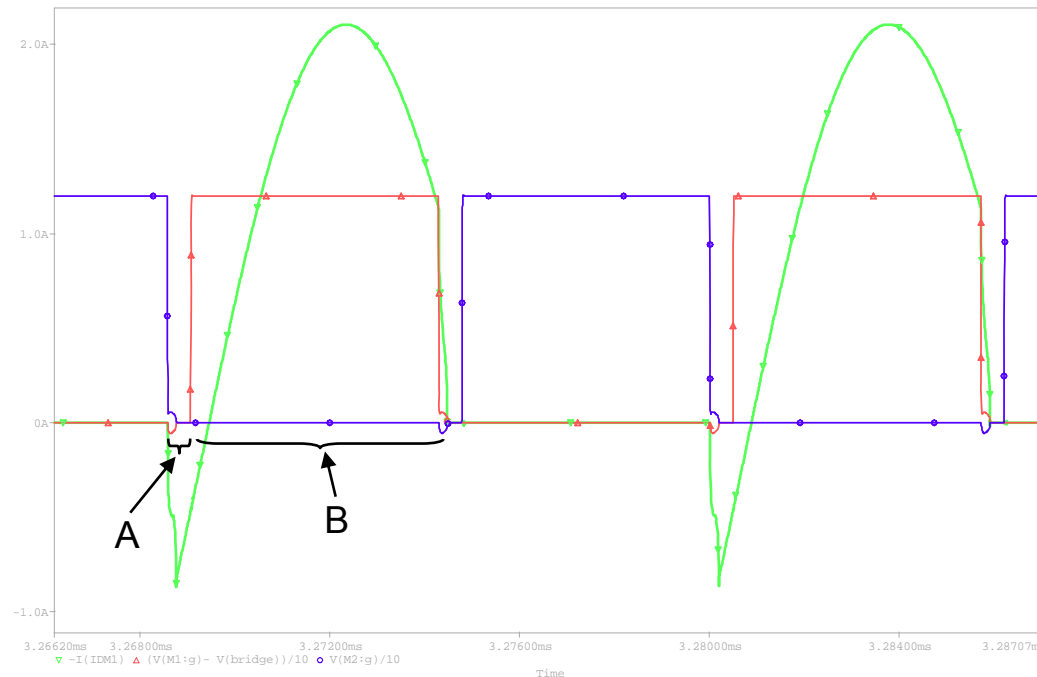
Turn OFF current calculation

- Assume that the magnetizing current increases linearly

$$I_{OFF} \approx \frac{V_{bulk}}{8 \cdot L_m \cdot f_{sw}}$$

- Turn OFF losses (E_{OFF} @ I_{OFF}) can be find in the MOSFET datasheet

$$P_{switch\_total} \approx I_{switch\_RMS}^2 \cdot R_{dsON} + P_{OFF}$$

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

Secondary Rectifier Design
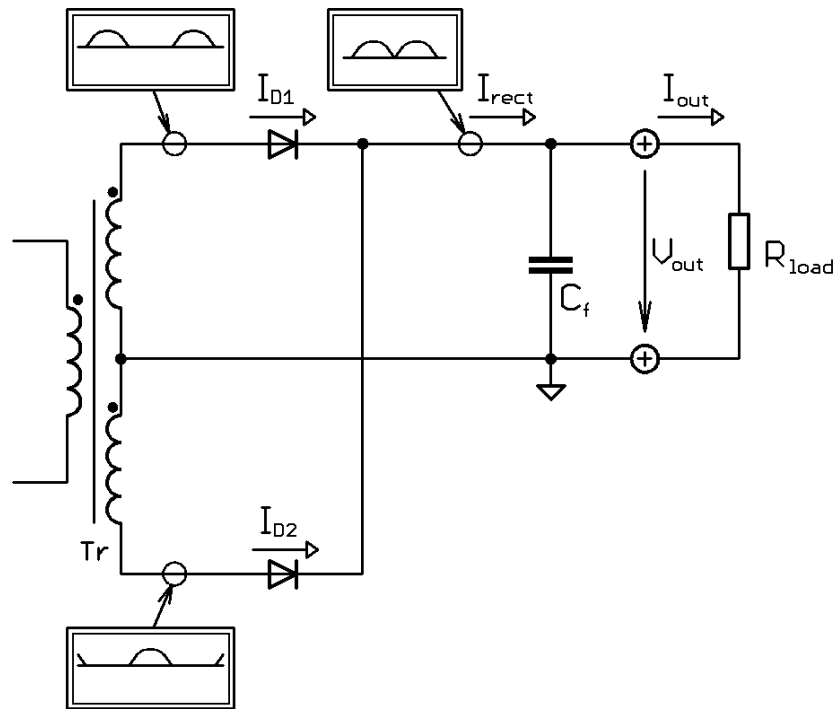
Push-Pull Configuration

- Secondary rectifiers work in ZCS
- Possible configurations:

  a) Push-Pull configuration – for low voltage / high current output
  b) Bridge configuration – for high voltage / low current output
  c) Bridge configuration with two secondary windings – for complementary output voltages

Advantages:
- Half the diode drops compared to bridge
- Single package, dual diode can be used
- Space efficient

Disadvantages:
- Need additional winding
- Higher rectifier breakdown voltage
- Need good matching between windings

Secondary Current Calculations – Push-Pull

| Equations	24 V/10 A output	12 V/20 A output
RMS diode current $I_{D\_RMS} = I_{out} \cdot \dfrac{\pi}{4}$	$I_{D\_RMS} = 7.85A$	$I_{D\_RMS} = 15.7A$
AVG diode current $I_{D\_AVG} = \dfrac{I_{out}}{2}$	$I_{D\_AVG} = 5A$	$I_{D\_AVG} = 10A$
Peak diode current $I_{D\_PK} = I_{out} \cdot \dfrac{\pi}{2}$	$I_{D\_PK} = 15.7A$	$I_{D\_PK} = 31.4A$

- To simplify calculations, assume sinusoidal current and $F_{op}=F_s$

Rectifier Losses – Push-Pull

Equations	24 V/ 10 A Vf=0.8 V, Rd=0.01 Ohm	12 V/ 20 A Vf=0.5 V, Rd=0.01 Ohm
Losses due to forward drop: $$P_{DFW} = \frac{V_F \cdot I_{OUT}}{2}$$	$$P_{DFW} = 4.0\ \text{W}$$	$$P_{DFW} = 5.0\ \text{W}$$
Losses due to dynamic resistance: $$P_{DRd} = \frac{R_d \cdot I_{OUT}^{\,2} \cdot \pi^2}{16}$$	$$P_{DRd} = 0.62\ \text{W}$$	$$P_{DRd} = 2.48\ \text{W}$$

Equation	24 V/ 10 A	12 V/ 20 A
$$P_{Rect\_total} = (P_{DFW} + P_{DRd}) \cdot n_{rect}$$	$$P_{Rect\_total} = 9.24\ \text{W}$$	$$P_{Rect\_total} = 15\ \text{W}$$

Secondary Rectifiers - Bridge Configuration
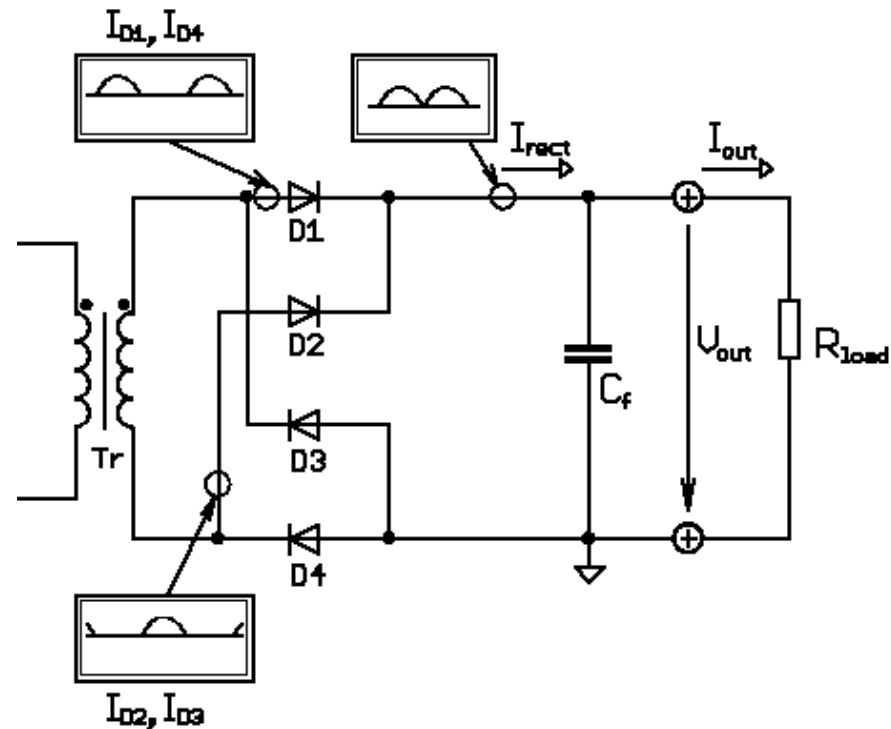
Advantages:

- Lower voltage rating
- Needs only one winding
- No matching needed for windings

Disadvantages:

- Higher diode drops
- Need four rectifiers

Secondary Rectifier Design Procedure

1. Select appropriate topology (push-pull or bridge)
2. Calculate rectifier peak, AVG and RMS current
3. Select rectifier based on the needed current and voltage ratings
4. Measure the diode voltage waveform in the application and design snubber to limit diode voltage overshoot and improve EMI signature (for LLC "weak" snubber is needed since diodes operate in ZCS mode)

Notes:

-   The current ripple increases for $f_{op} < f_s$, the current waveform is still half "sinusoidal" but with dead times between each half period
-   The peak current is very high for low voltage and high current LLC applications – example 12 V/ 20 A output: I_{peak} = 31.4 A and I_{RMS} = 9.7 A!! Each "mΩ" becomes critical - PCB layout. The secondary rectification paths should be as symmetrical as possible to assure same parameters for each switching cycle.

Output Capacitor Dimensioning
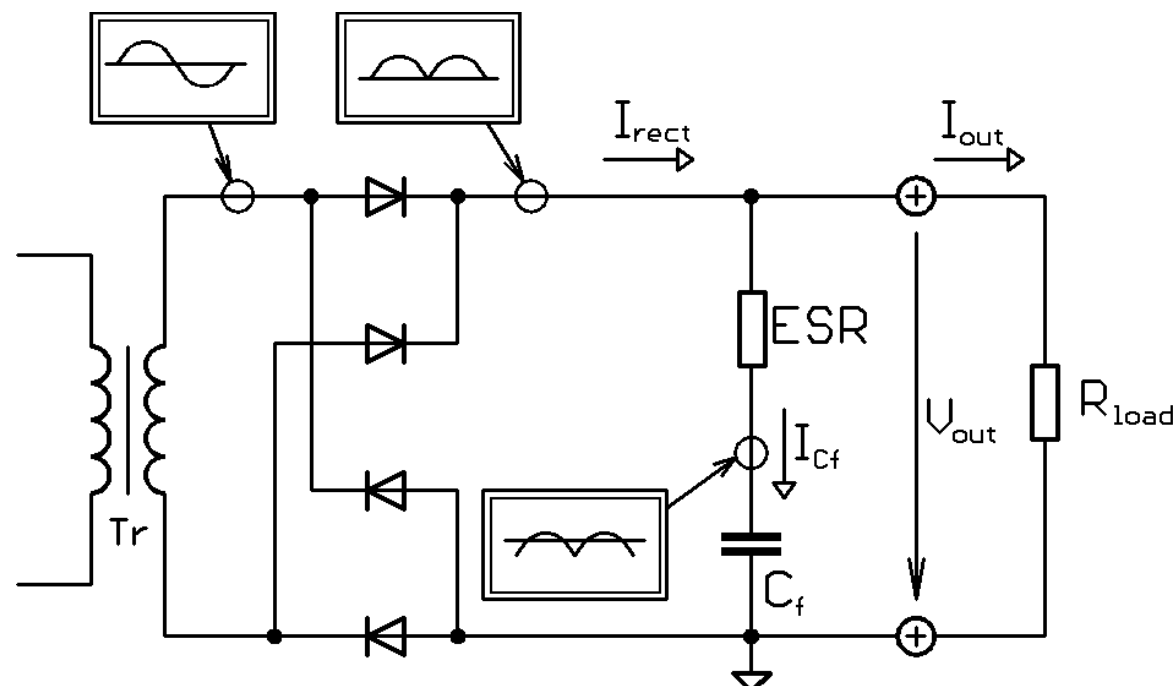
- Output capacitor is the only energy storage device
  - Higher peak/rms ripple current and energy
- Ripple current leads to:
  - Voltage ripple created by the ESR of output capacitor (dominant)
  - Voltage ripple created by the capacitance (less critical)

ESR Component of Output Ripple

- In phase with the current ripple and frequency independent
- Low ESR capacitors needed to keep ripple acceptable
  - Cost/performance trade-off (efficiency impact)

Equations:	24 V/ 10 A example: Cf=5000 uF, ESR=6 mΩ
Peak rectifier current $$I_{rect\_peak} = I_{out} \cdot \frac{\pi}{2}$$	$$I_{rect\_peak} = 15.7A$$
Output voltage ripple peak to peak $$V_{out\_ripple\_pk-pk} = ESR \cdot I_{rect\_peak}$$	$$V_{out\_ripple\_pk-pk} = 94\,mV$$
Capacitor RMS current: $$I_{Cf\_RMS} = I_{out} \cdot \sqrt{\frac{\pi^2}{8} - 1}$$	$$I_{Cf\_RMS} = 4.83\ A$$
ESR power losses $$P_{ESR} = I_{Cf\_RMS}{}^2 \cdot ESR$$	$$P_{ESR} = 140\,mW$$

Capacitive Component of Output Ripple

- Out of phase with current and frequency dependent
- Actual ripple negligible due to high value of capacitance chosen

Equation:	24 V/ 10 A output example: Cf=5000 uF, Fop=100 kHz
$$V_{out\_ripple\_cap\_pk-pk} = \frac{I_{out}}{2 \cdot \sqrt{3} \cdot \pi \cdot f_{op} \cdot C_f} \cdot (\pi - 2)$$	$$V_{out\_ripple\_cap} = 2.1\,\text{mV}$$
	24 V/ 10 A output example: Cf=100 uF, Fop=100 kHz
	$$V_{out\_ripple\_cap} = 104\,\text{mV}$$

ON Semiconductor® (ON)

Filter Capacitor Design Procedure

1. Calculate peak and rms rectifier and capacitor currents based on Io and Vout

2. Calculate needed ESR value that will assure that the output ripple will be lower than maximum specification

3. Select appropriate capacitor(s) to handle the calculated rms current and having calculated ESR or lower

4. Factor in price, physical dimensions and transient response

5. Check the capacitive component value of the ripple (usually negligible for high enough C_f)

Notes:

• The secondary rectification paths should be as symmetrical as possible to assure same parameters for each switching cycle

f1   This slide may be better off getting split into two slides and add some more notes. Let's discuss.
ffmrmw, 2007-9-3

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

Resonant inductance balance

Transformer leakage inductance

- Total L_s is always affected by the transformer leakage inductance
- Special case for transformer with integrated leakage inductance - $L_s = L_{lk}$
- Push pull and mult. output app. are sensitive to the leakage inductance balance
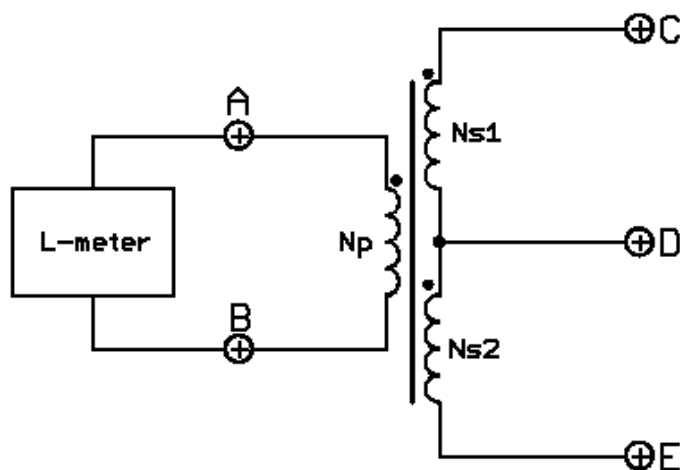
Parameter	Measured between pins	Secondary pins configuration
$L_{lk(p-s1)}$	A-B	C-D short D-E open
$L_{lk(p-s2)}$	A-B	C-D open D-E short
$L_{lk(total)}$	A-B	C-D short D-E short
L_m	A-B	C-D open D-E open

Example: $L_{lk(p-s1)}$ = 105 uH

$L_{lk(p-s2)}$ = 115 uH

ΔL_{lk} = 10 uH

$L_{lk(total)}$ = 100 uH

L_m = 600 uH

C_s = 33 nF
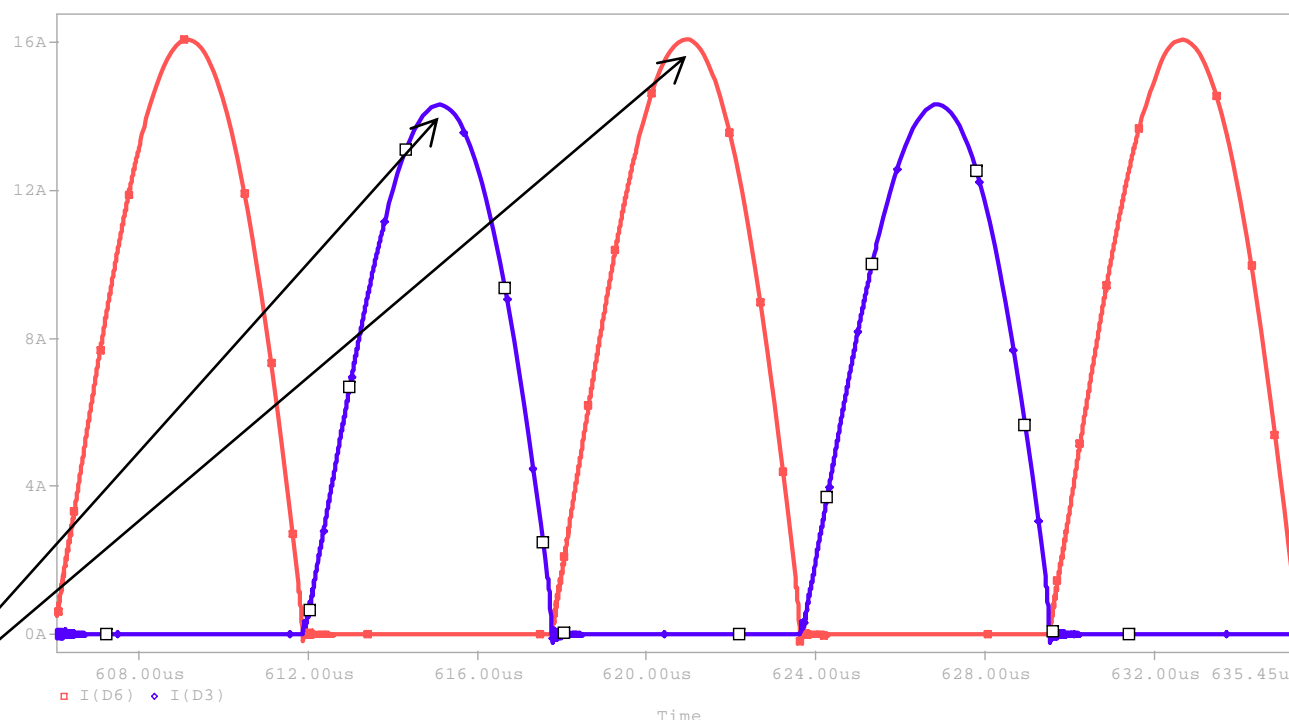
f_{s1} = 85.5 kHz

f_{s2} = 81.7 kHz

➡ **5 % difference**

Resonant inductance balance

Series resonant frequency differs for each switching half-cycle that results in primary and mainly secondary current imbalance.

3 A difference in the peak secondary current – the power dissipation is different for each rectifier from pair as well as for the secondary windings.
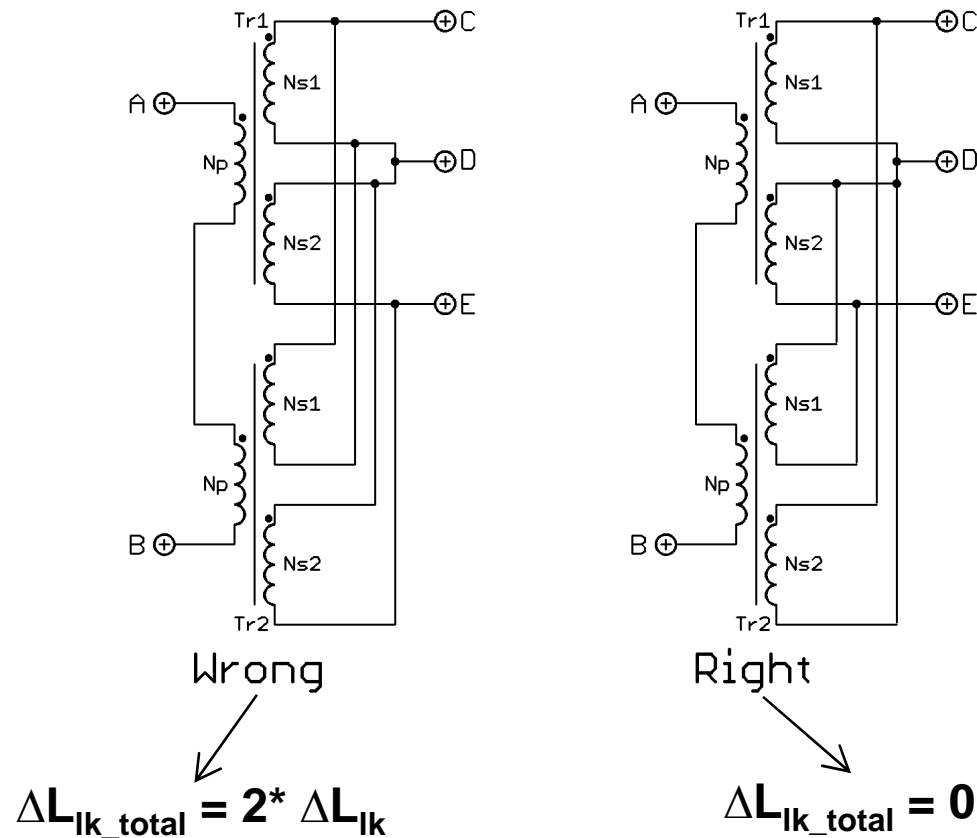
Resonant inductance balance
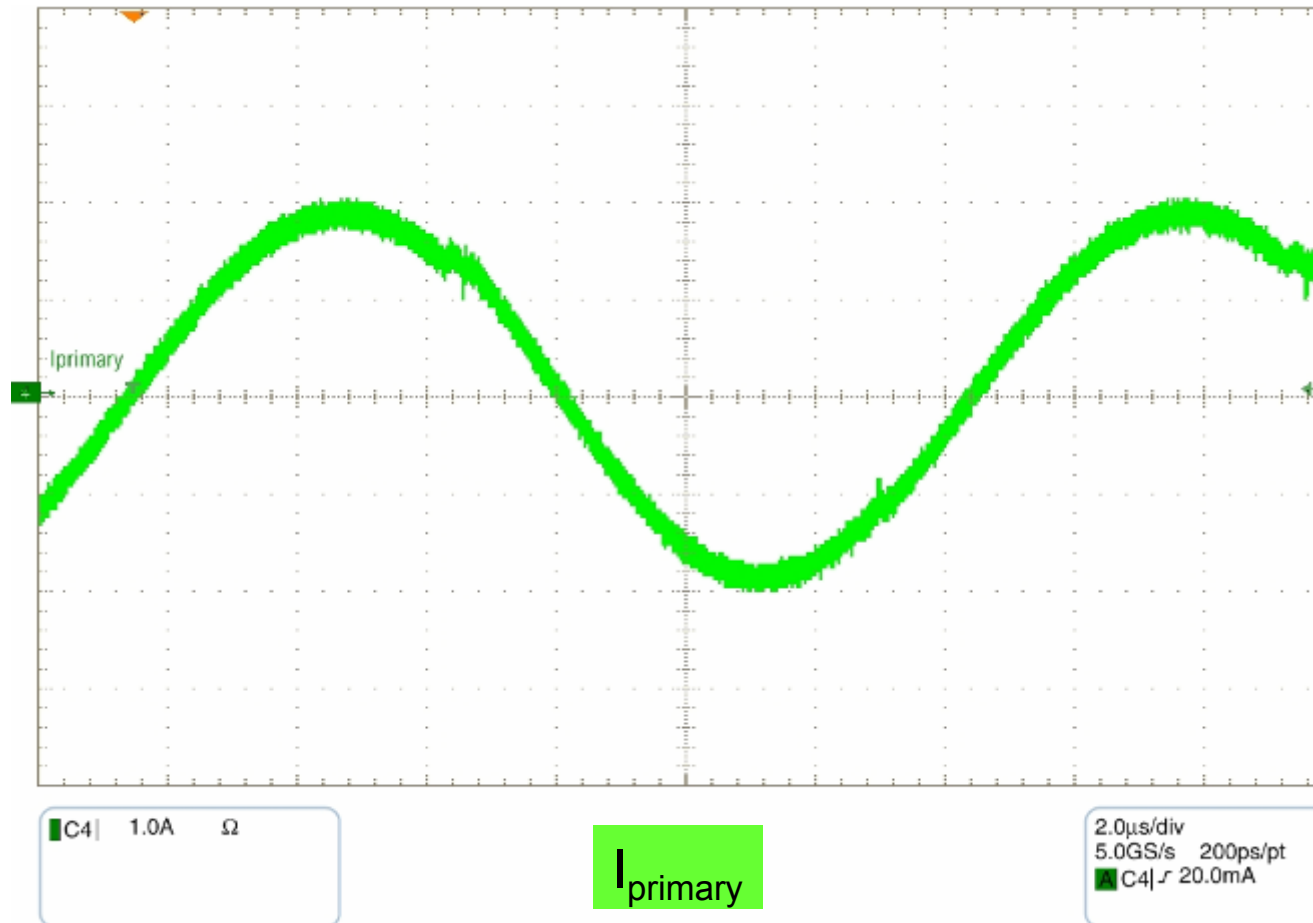
Converter works below series resonant frequency F_s for the one half of the switching cycle and in the F_s for the second half of the switching cycle.
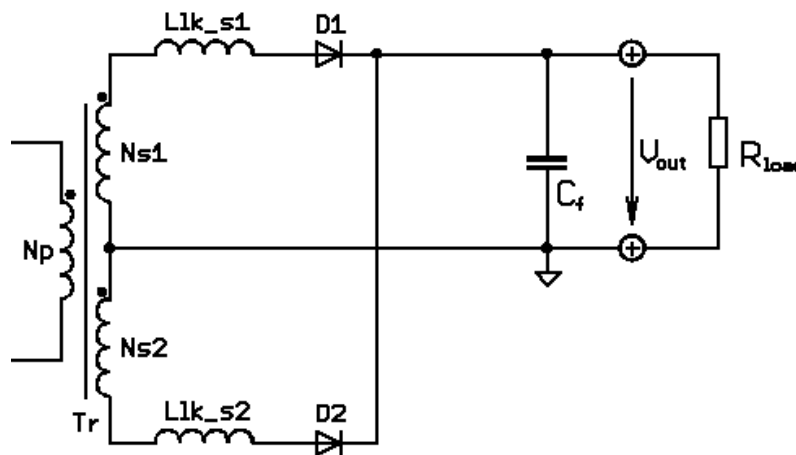
Resonant inductance balance

For high power app. it is attractive to connect primary windings in series and secondary windings in parallel. There is possibility to compensate transformer leakage imbalance by appropriate connection of the secondary windings:

Wrong

$\Delta L_{lk\_total} = 2* \Delta L_{lk}$

Right

$\Delta L_{lk\_total} = 0$

Resonant inductance balance

The secondary leakage inductance is transformed to the primary and increases the total resonant inductance value. Situation becomes critical for the LLC applications with high turns ratios.

12 V / 20 A application example:

N_p = 35 turns L_{lk_s1} = 100 nH

N_s = 2 x 2 turns L_{lk_s2} = 150 nH

$n = N_p/N_s = 17.5$

L_s = 110 uH

L_m = 630 uH

ΔL_{lk_s} = 50 nH   ⟹ $\Delta L_s = \Delta L_{lk\_s} \cdot n^2 = 15.3 \mu H$

50 nH difference on the secondary causes 14 % difference of L_s !!!

Resonant inductance balance

Transformer construction and secondary layout considerations:

- Resonant tank parameters can change each switching half cycle when push pull configuration is used. This can cause the primary and secondary currents imbalance.

- For the transformer with integrated resonant inductance, it has to be checked how the transformer manufacturer specifies the leakage inductance. Specification for all secondary windings shorted is irrelevant. The particular leakage inductance values can differ.

- When using more transformers with primary windings in series and secondary windings in parallel the leakage inductance asymmetry can be compensated by appropriate secondary windings connection.

- Secondary leakage inductance can cause significant resonant inductance imbalance in applications with high transformer turns ratio. Layout on the secondary side of the LLC resonant converter is critical in that case.

Agenda

- Why an HB LLC converter

- Configurations of the HB LLC converter and a resonant tank

- Operating states of the HB LLC

- HB LLC converter modeling and gain characteristics

- Primary currents and resonant cap dimensioning

- Secondary rectification design and output cap dimensioning

- Resonant inductance balance

- Transformer winding dimensioning and transformer construction

Transformer winding dimensioning

The primary current is sinusoidal for $F_{op}=F_s$. The secondary current is almost sinusoidal too – there is slight distortion that is given by the magnetizing current.

$$I_{primary\_RMS} \approx \sqrt{\frac{1}{8} \cdot \left(\frac{I_{out}^2 \cdot \pi^2}{n^2} + \frac{V_{bulk}^2}{16 \cdot L_m^2 \cdot f_{sw}^2} \right)}$$

$$I_{secondary\_RMS} \approx I_{out} \cdot \frac{\pi}{2 \cdot \sqrt{2}} \qquad \text{(single winding solution)}$$

- The skin effect and mainly proximity effect decreases effective cooper area.
- Proximity effect can be overcome by the interleaved winding construction (for discrete resonant tank solution)

- The proximity effect becomes critical for the transformer with integrated leakage
- Wires that are located to the center of the bobbin "feels" much higher current density than the rest of the windings even when litz wire used!

Transformer with integrated leakage

- For the standard transformer with good coupling ($L_{lk} < 0.1 * L_m$) is the leakage inductance independent on the air gap thickness and position

Llk_transformed
to primary

M

Lm

Lm/n^2

primary

secondary

$$M = \sqrt{1 - \frac{L_{lk}}{L_m}}$$
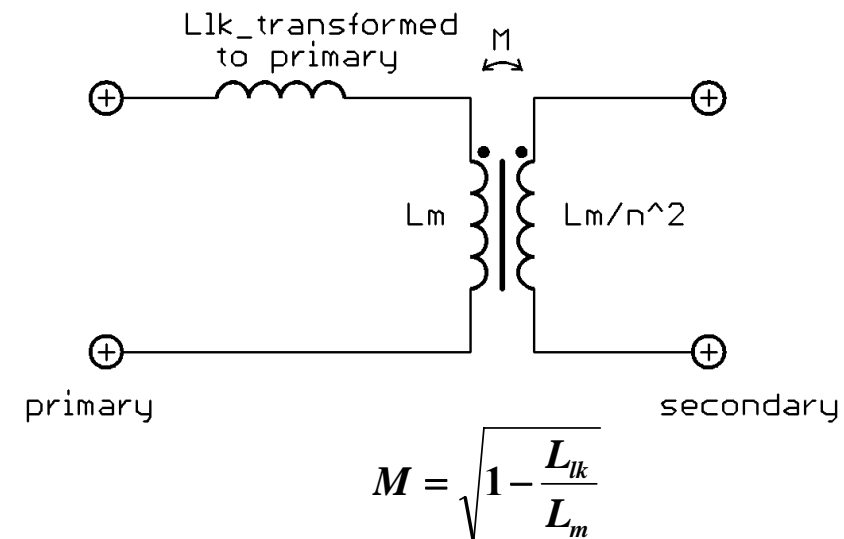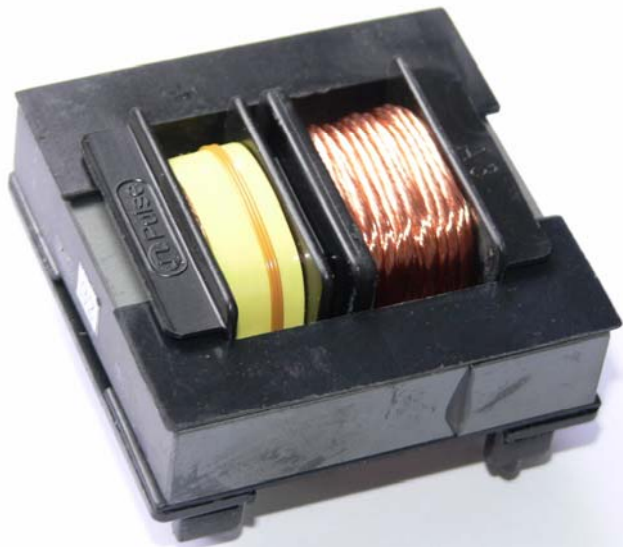
- Transformer with divided bobbin exhibits high leakage inductance
- Significant energy is related to the stray flux
- The L_{lk} is dependent on air gap thickness and position

Transformer with integrated leakage

- Small energy is stored in the L_m each switching cycle to prepare ZVS
- L_m is given by the magnetic conductivity of the magnetic circuit i.e.: $L_m = n^2 * A_l$
- Gap is used to absorb most of the magnetizing energy and to adjust L_m value
- Magnetizing energy is taken from the primary winding - it is advantageous to locate the air gap in the centre of the primary winding
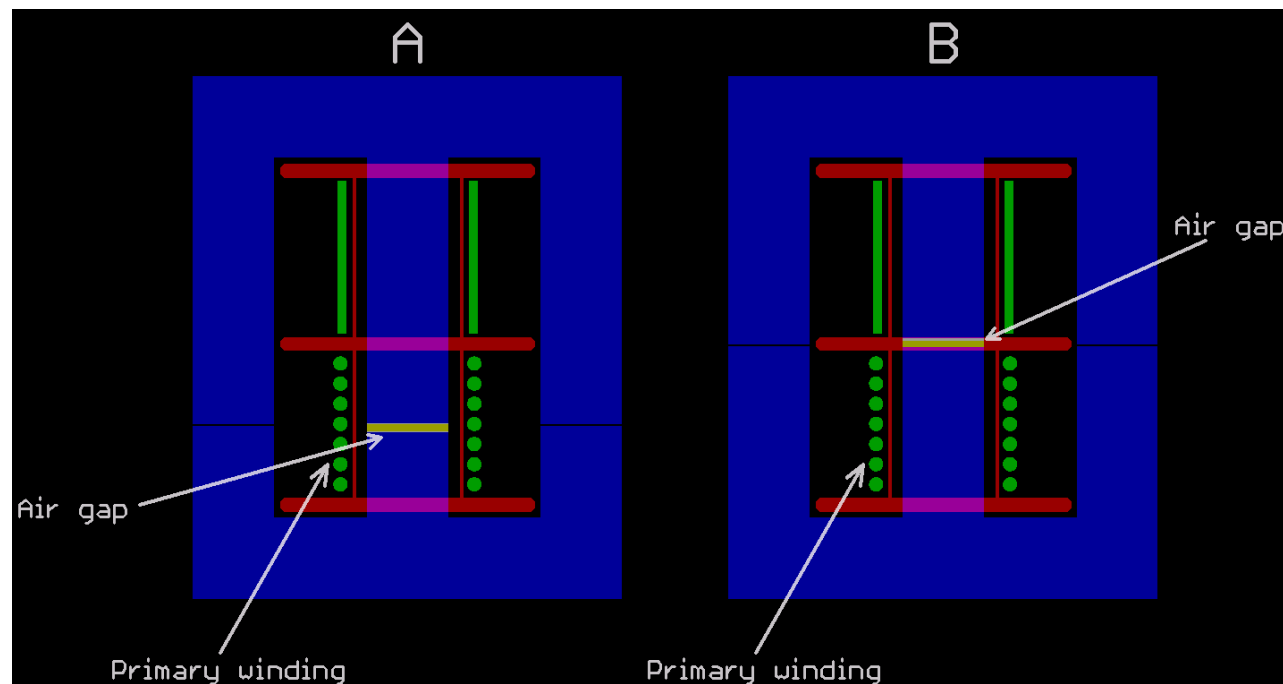
Conclusions

- HB LLC Converter provides a very good solution for high efficiency, low EMI power conversion requirements
- Design considerations for HB LLC converter are more complex than traditional topologies
  - Transformer design choice is critical for converter operation
  - Other passive components also exhibit high stresses
  - Frequency variation is used to maintain output regulation
- This presentation covered all power stage design considerations
  - More details available in future app. note
- ON Semiconductor offers full support for your designs using NCP1395/1396

Thank you for your attention!