



**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:

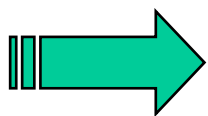
- Output power from 150 W up to 600 W
- Universal mains
- Active or passive PFC (given by needed power)
- Limited width and space, no fan: limited air flow
- Low standby power consumption
- Consumer Electronics market: fierce competition

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

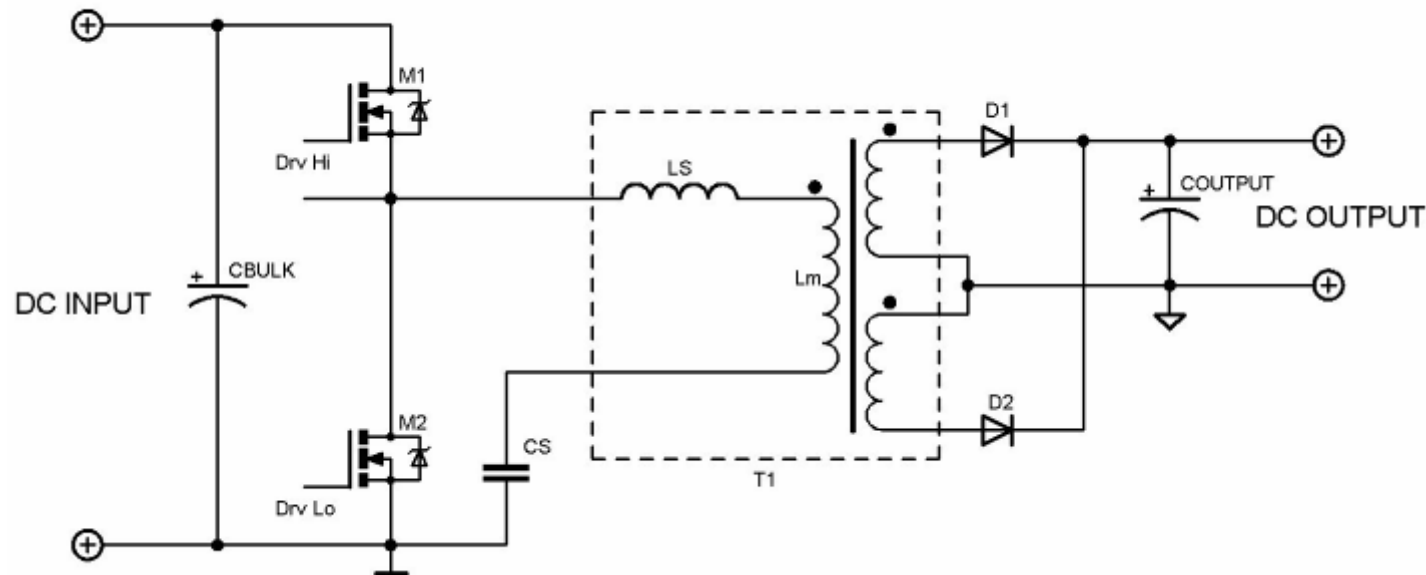


Cost effective, highly efficient and EMI friendly solution for high and medium output voltage converters

# Agenda

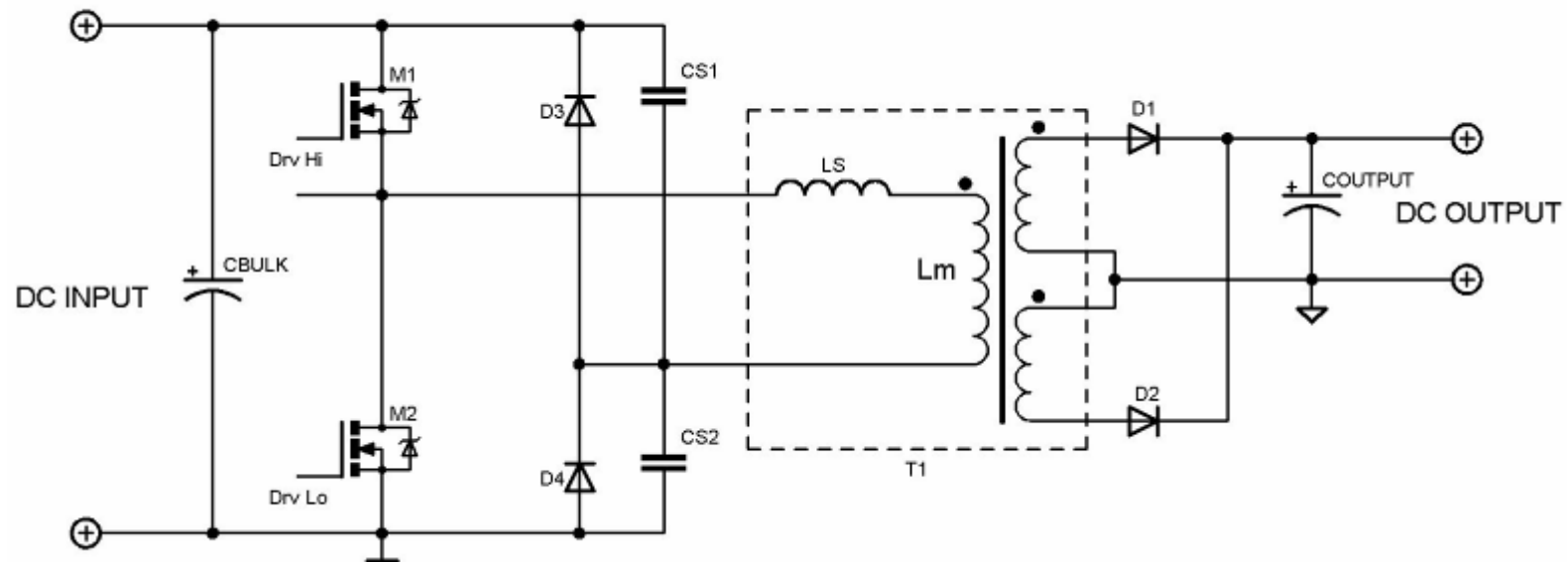
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# Configurations of an HB LLC – single res. cap



- 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:

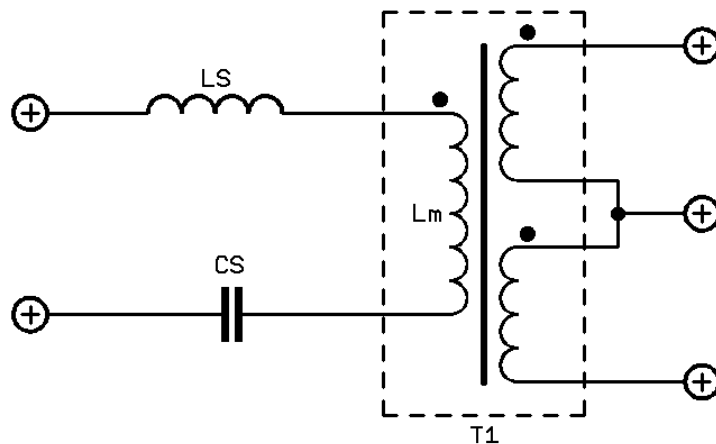
- Lower input current ripple and RMS value ( $1/\sqrt{2}$ )
- Resonant capacitors handle half RMS current
- Capacitors with half capacitance are used
- Low voltage ratings (450 V) for the resonant capacitors when clamping diodes D3, D4 perform an easy and cheap overload protection

## 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

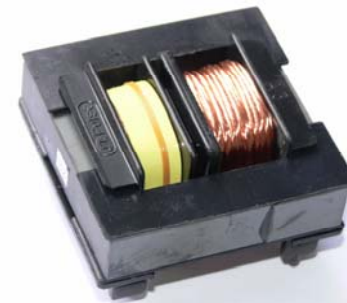
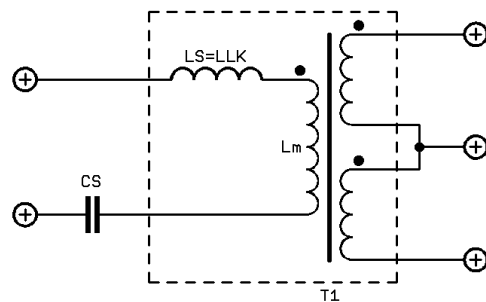


# Resonant tank configurations – integrated solution

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  $L_s$  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

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# Operating states of the LLC converter

## Discrete resonant tank solution

Two resonant frequencies can be defined:

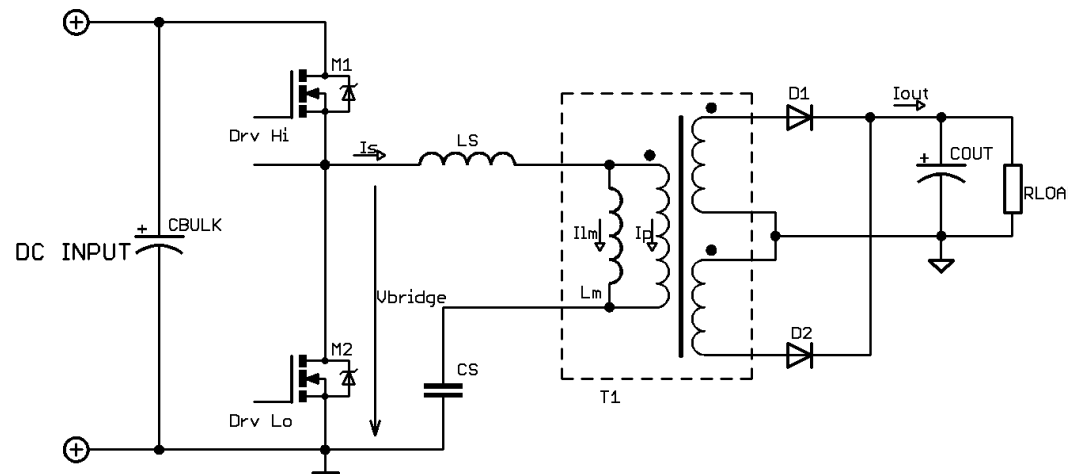
$$F_s = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot L_s}}$$

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

LLC converter can operate:

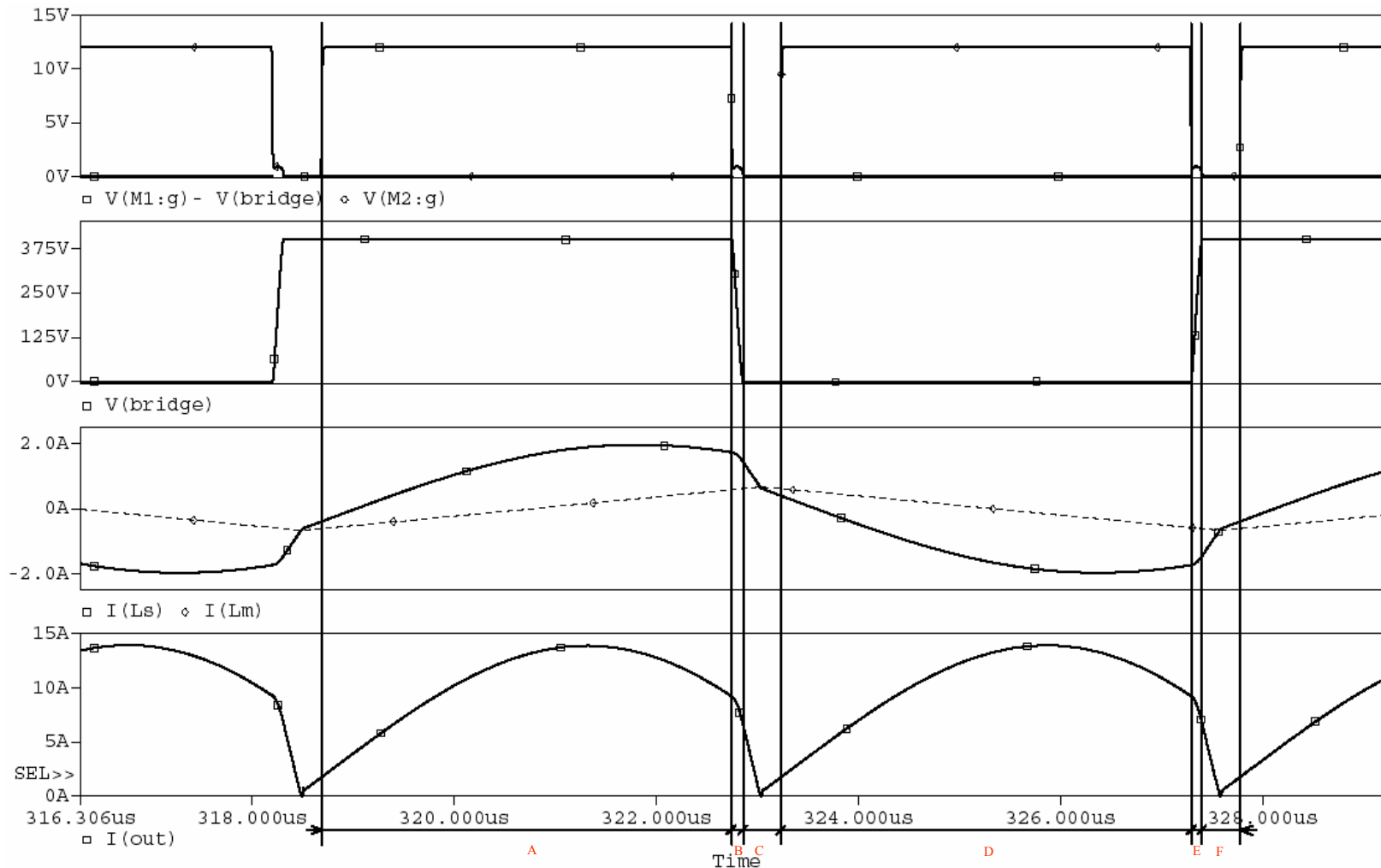
- a) between  $F_{\min}$  and  $F_s$
- b) direct in  $F_s$

- c) above  $F_s$
- d) between  $F_{\min}$  and  $F_s$  - overload
- e) below  $F_{\min}$



# Operating states of the LLC converter

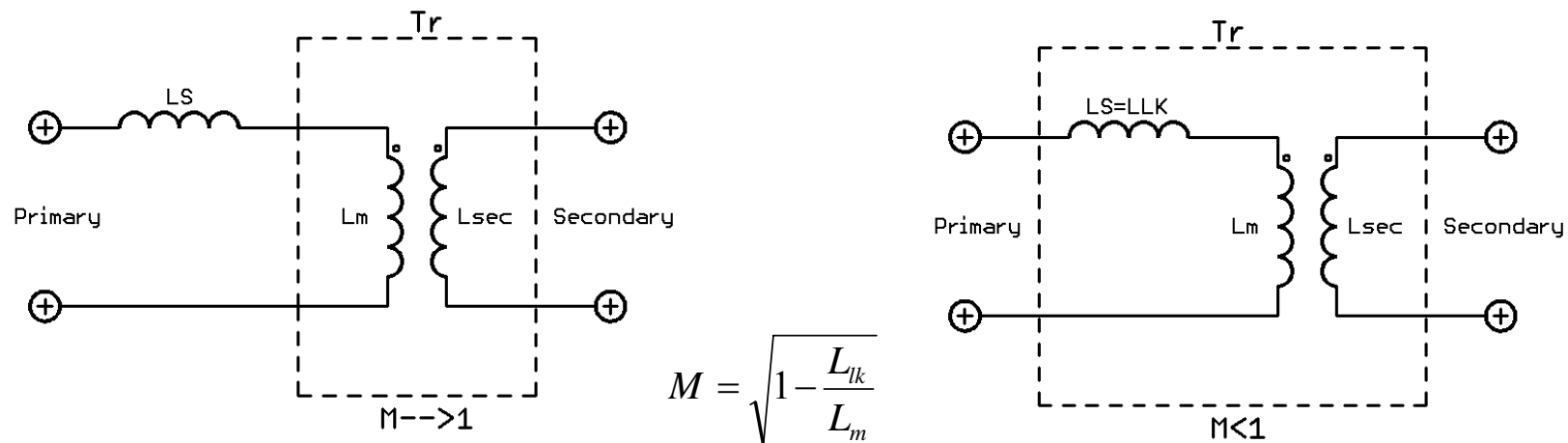
c) Operating waveforms for  $F_{op} > F_s$  Discrete resonant tank solution



# 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



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:

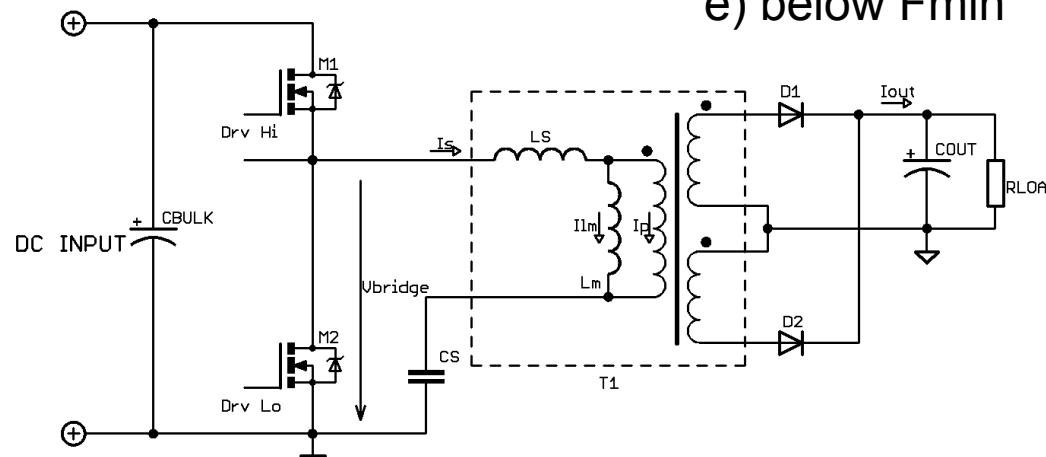
$$F_s = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot L_s}}$$

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

LLC converter can again operate:

- a) between  $F_{\min}$  and  $F_s$
- b) direct in  $F_s$

- c) above  $F_s$
- d) between  $F_{\min}$  and  $F_s$  – overload
- e) below  $F_{\min}$

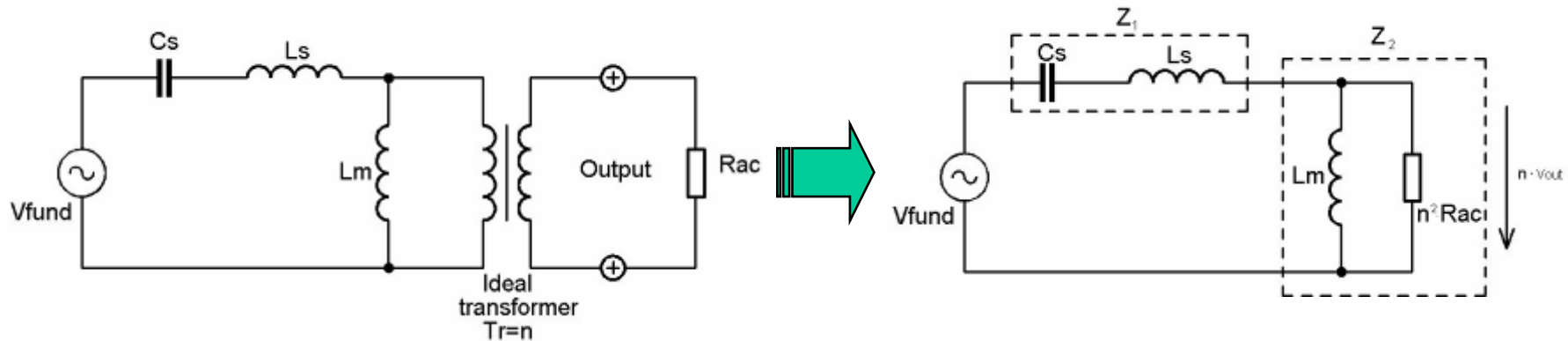


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# LLC converter modeling – equivalent circuit

LLC converter can be described using first 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:

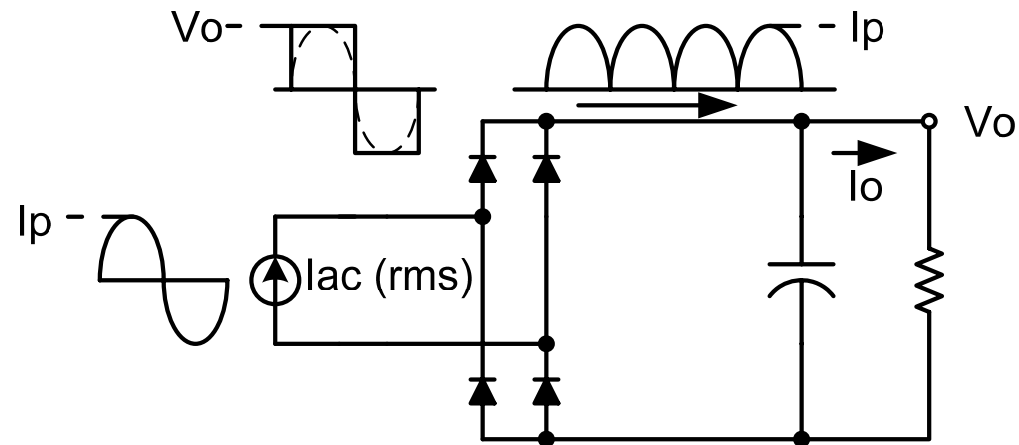
$$I_{ac\_RMS} = \frac{\pi}{2\sqrt{2}} I_O$$

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  $R_{ac}$  can 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:

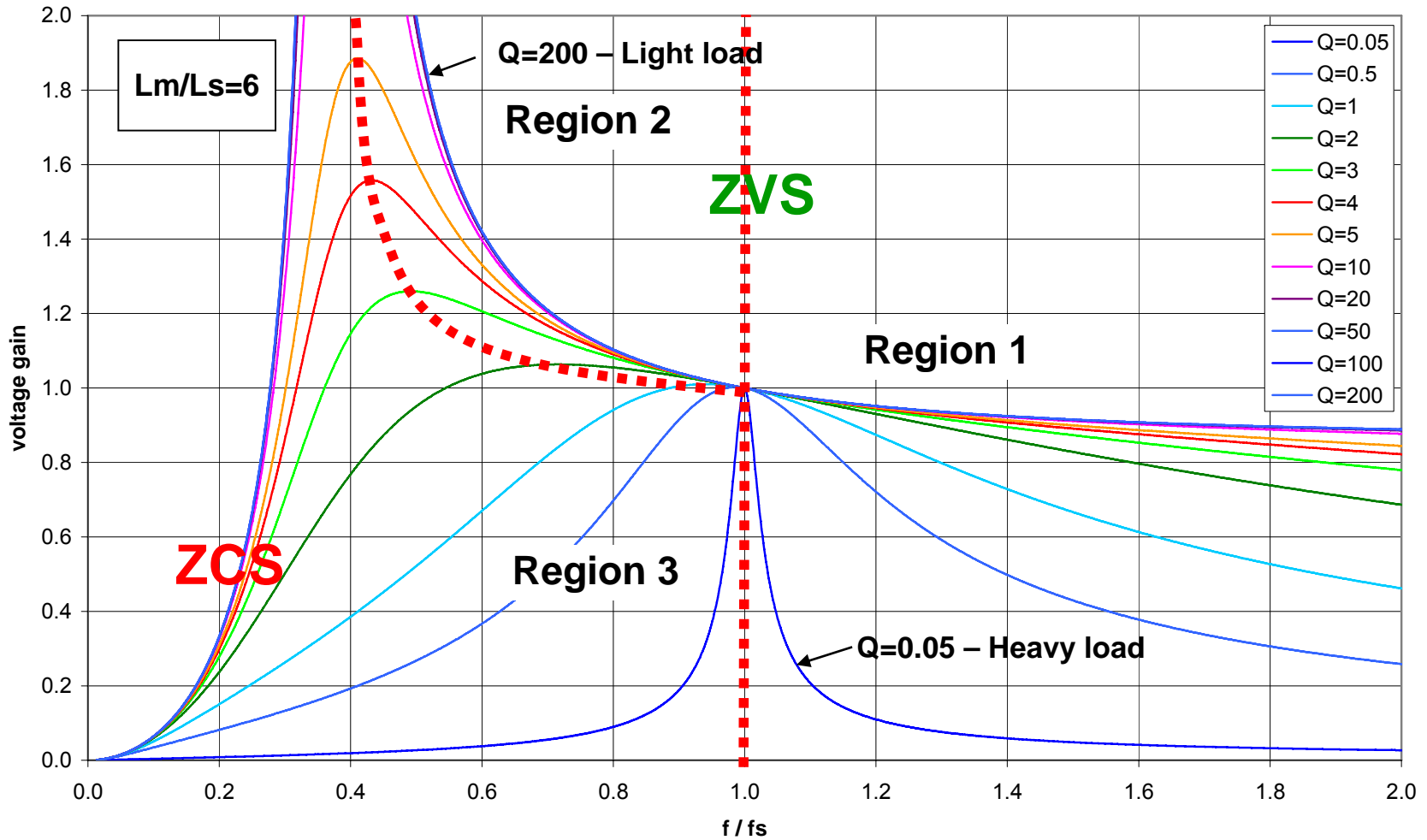
$$F_s = \frac{1}{2 \cdot \pi \cdot \sqrt{C_s \cdot L_s}}$$

Minimum resonant frequency:

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



# Normalized gain characteristic



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).

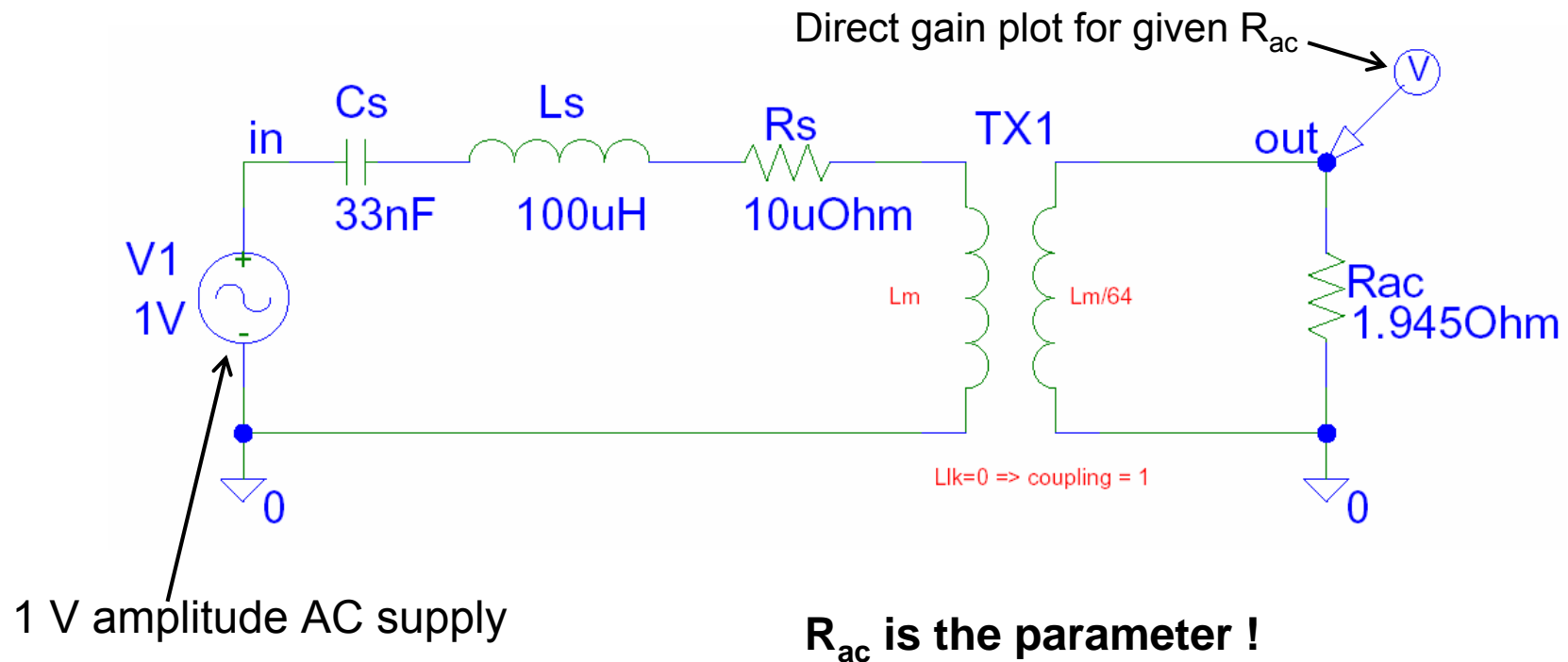
- Gain of the LLC converter, which operates in the  $f_s$  is 1 (for discrete resonant tank solution) - i.e. is given by the transformer turns ratio. This operating point is the most attractive from the efficiency and EMI point of view – sinusoidal primary current, MOSFETs and secondary diodes optimally 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:

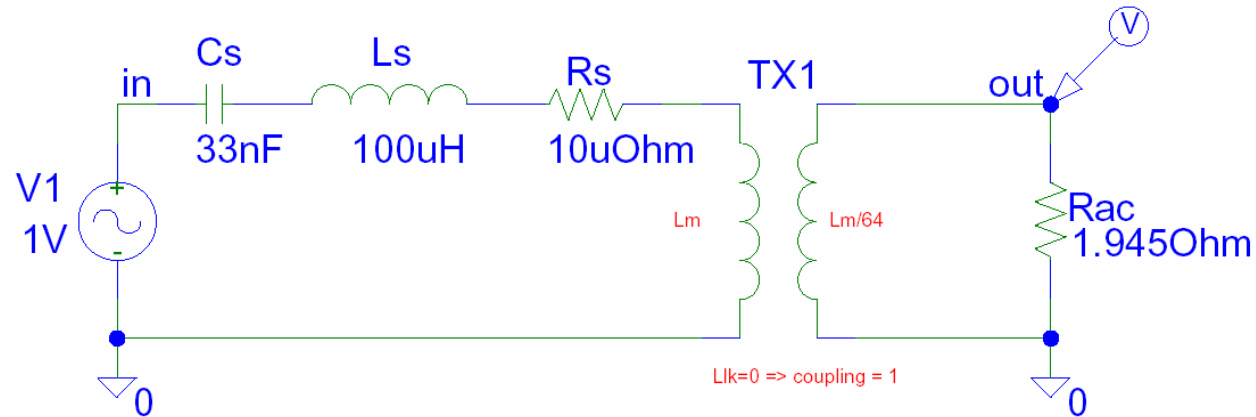
- $L_m/L_s$  ratio
- Characteristic impedance of the resonant tank
- Load value
- Transformer turns ratio

## How to obtain gain characteristics?

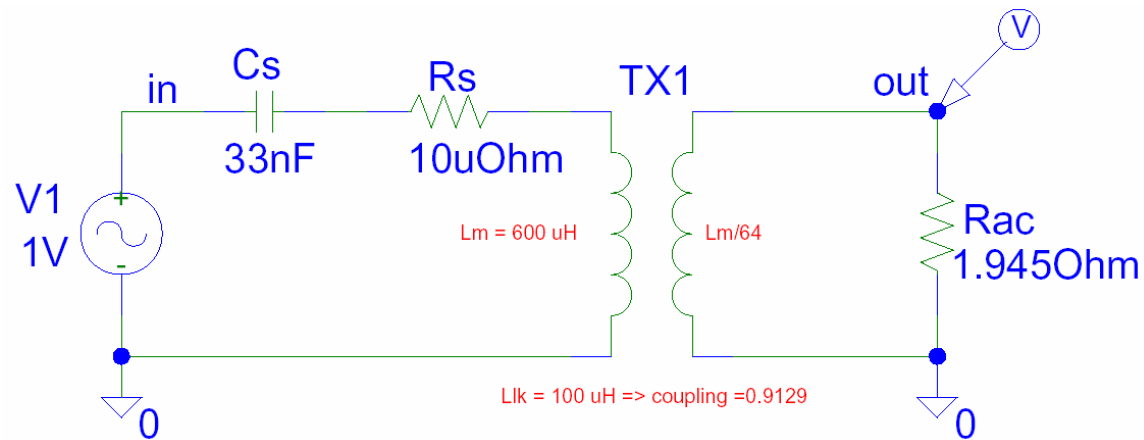
Use fundamental approximation and AC simulation in any simulation software like PSpice, Icap4 etc..



# 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:

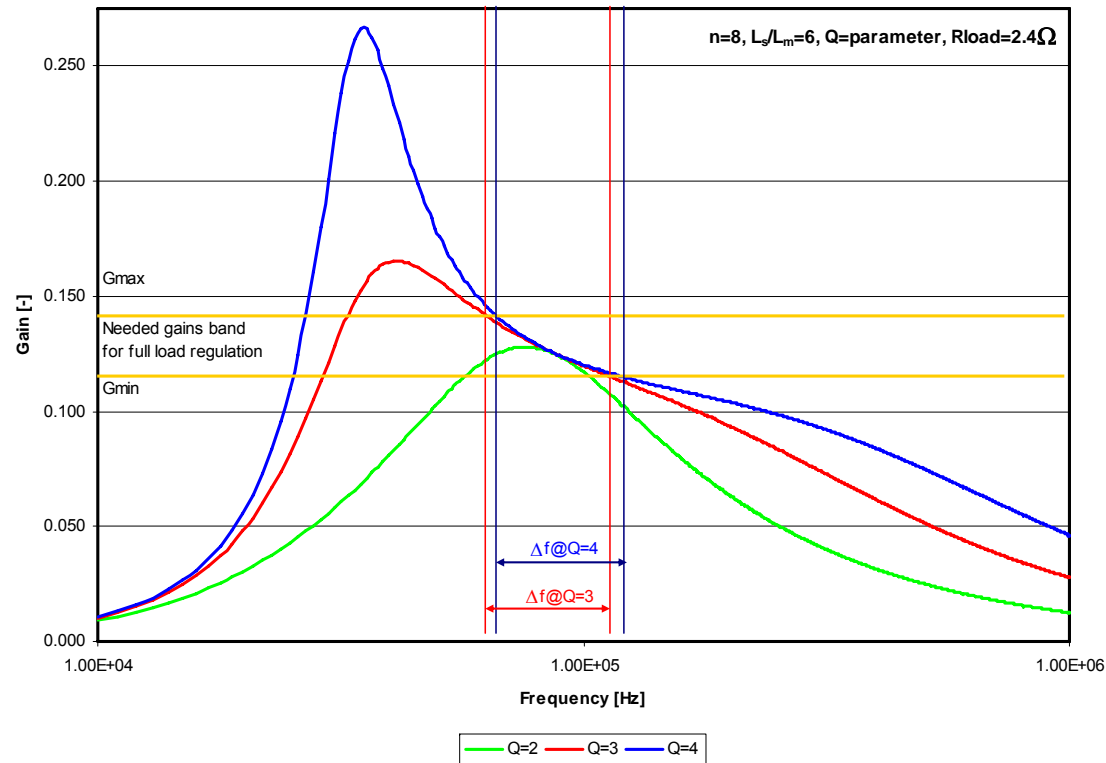
- Needed operating frequency range for output voltage regulation
- Line and load regulation ranges
- Value of circulating energy in the resonant tank
- Efficiency of the converter

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:

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

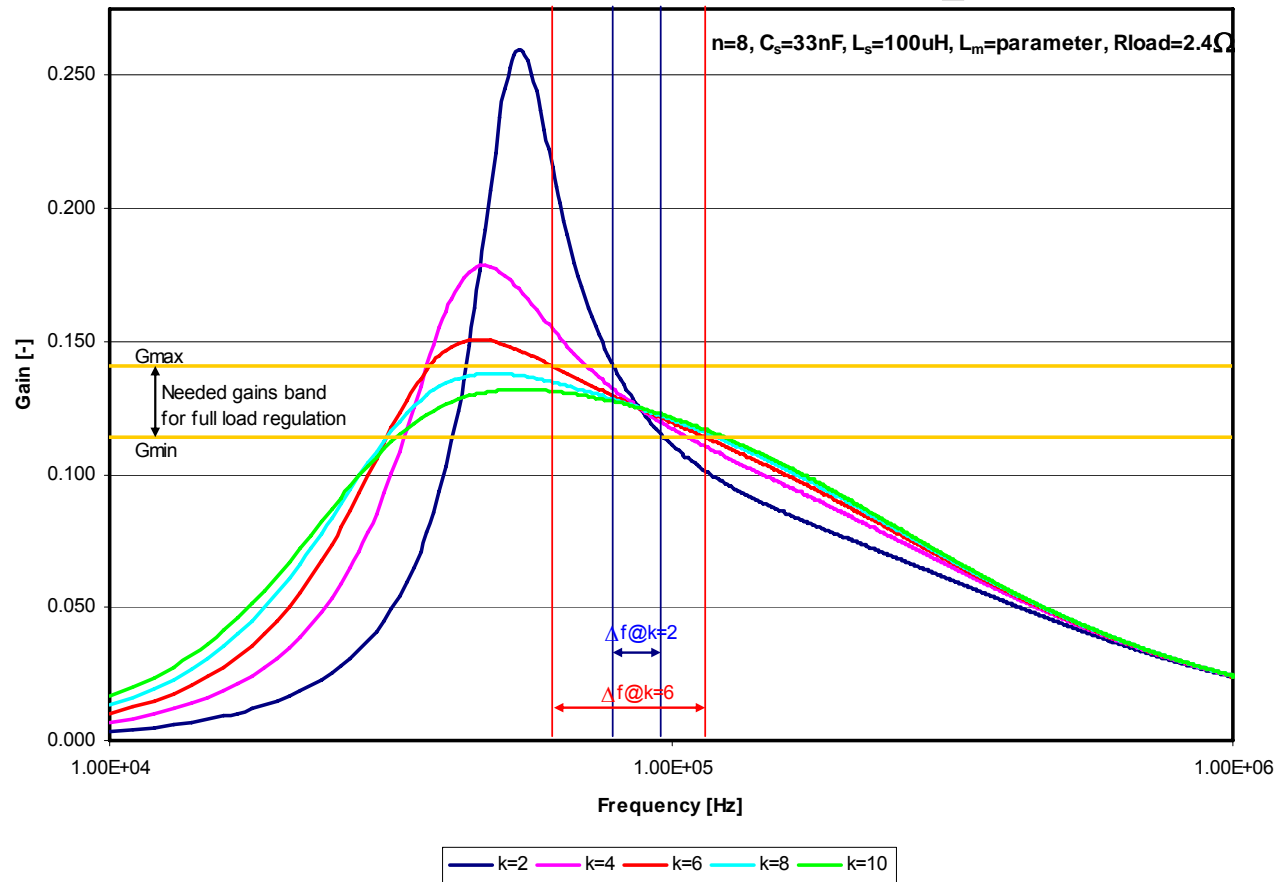
# Full load Q and k factors optimization



- 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 lower 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  $L_s$  (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  $f_s$ .

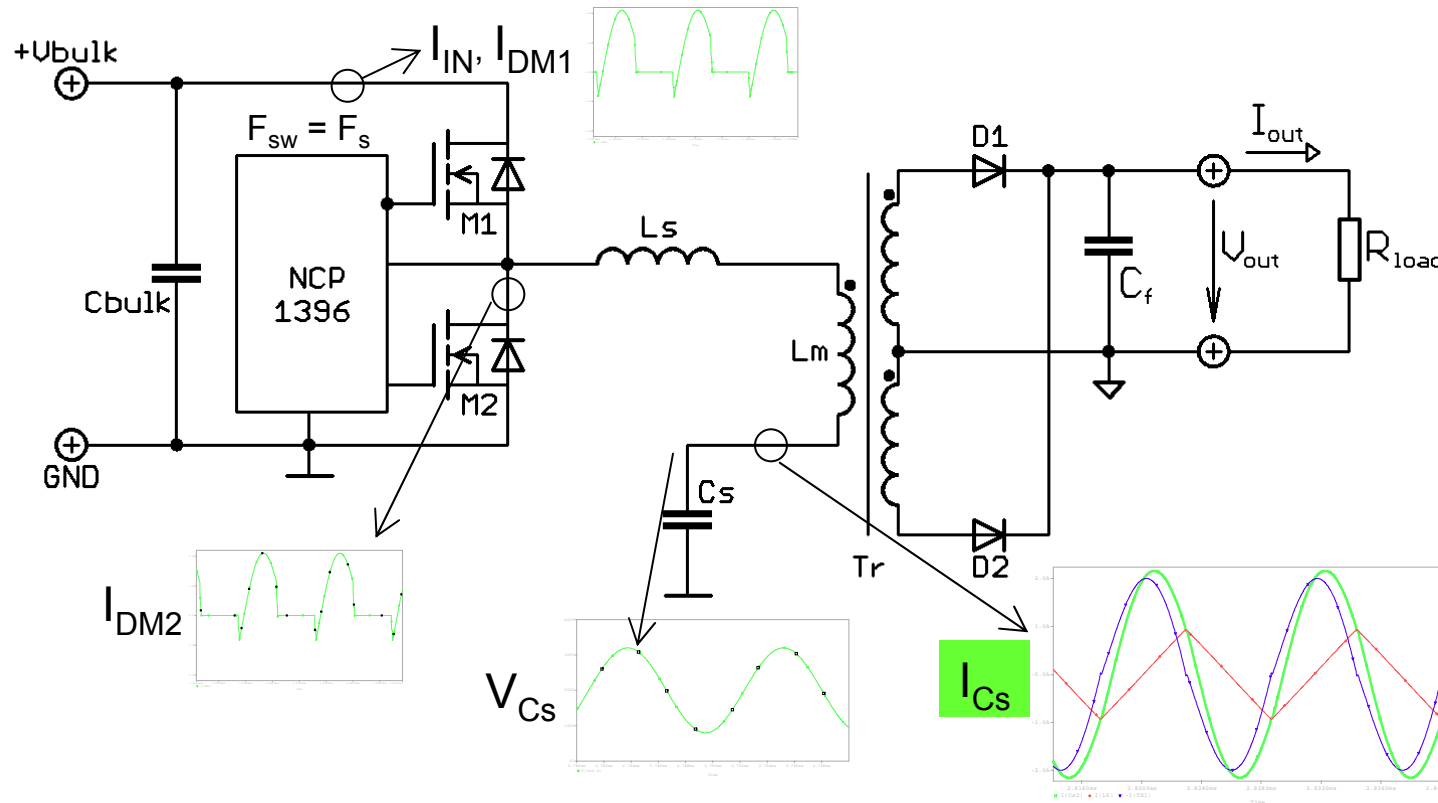
The  $k$  factor has to be then calculated to assure gains needed for the output voltage regulation (with line and load changes).

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

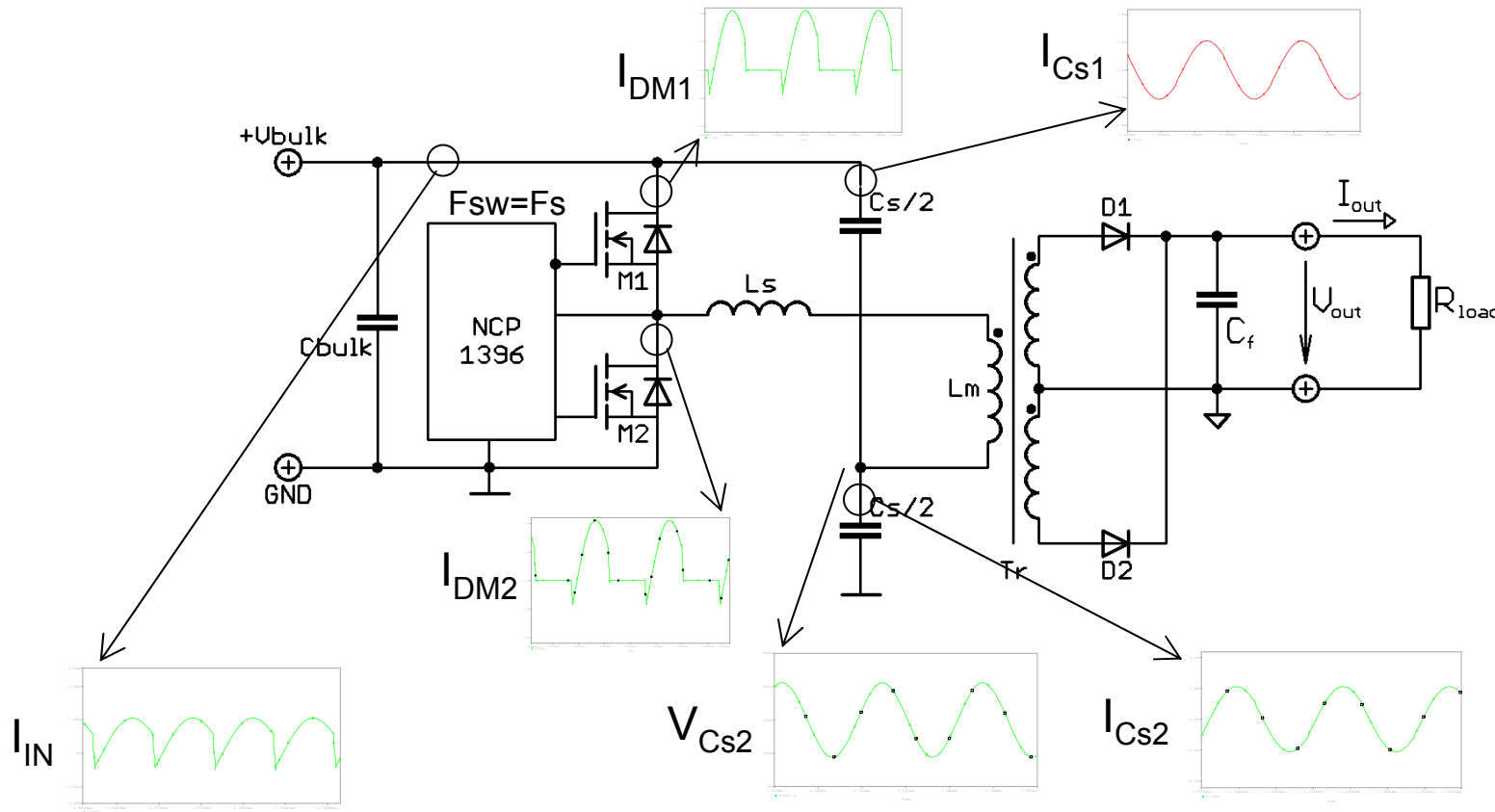
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# Primary currents – single resonant cap



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

# Primary currents – split resonant cap



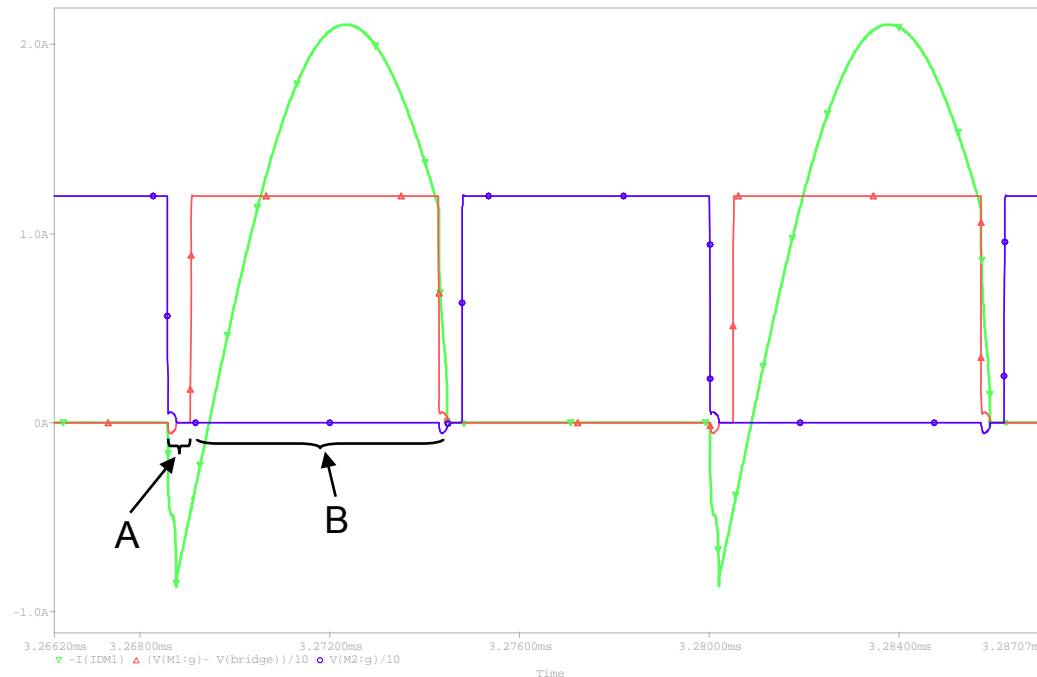
## Comparison of Primary Currents

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

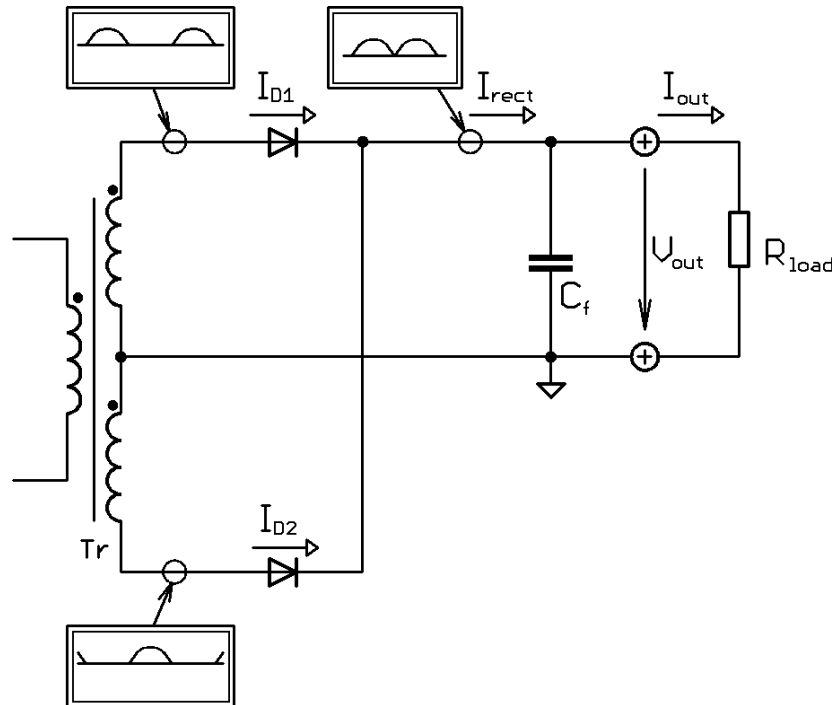
$$P_{switch\_total} \approx I_{switch\_RMS}^2 \cdot R_{dsON} + P_{OFF}$$



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## 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<br>$I_{D\_RMS} = I_{out} \cdot \frac{\pi}{4}$ | $I_{D\_RMS} = 7.85A$ | $I_{D\_RMS} = 15.7A$ |
| AVG diode current<br>$I_{D\_AVG} = \frac{I_{out}}{2}$           | $I_{D\_AVG} = 5A$    | $I_{D\_AVG} = 10A$   |
| Peak diode current<br>$I_{D\_PK} = I_{out} \cdot \frac{\pi}{2}$ | $I_{D\_PK} = 15.7A$  | $I_{D\_PK} = 31.4A$  |

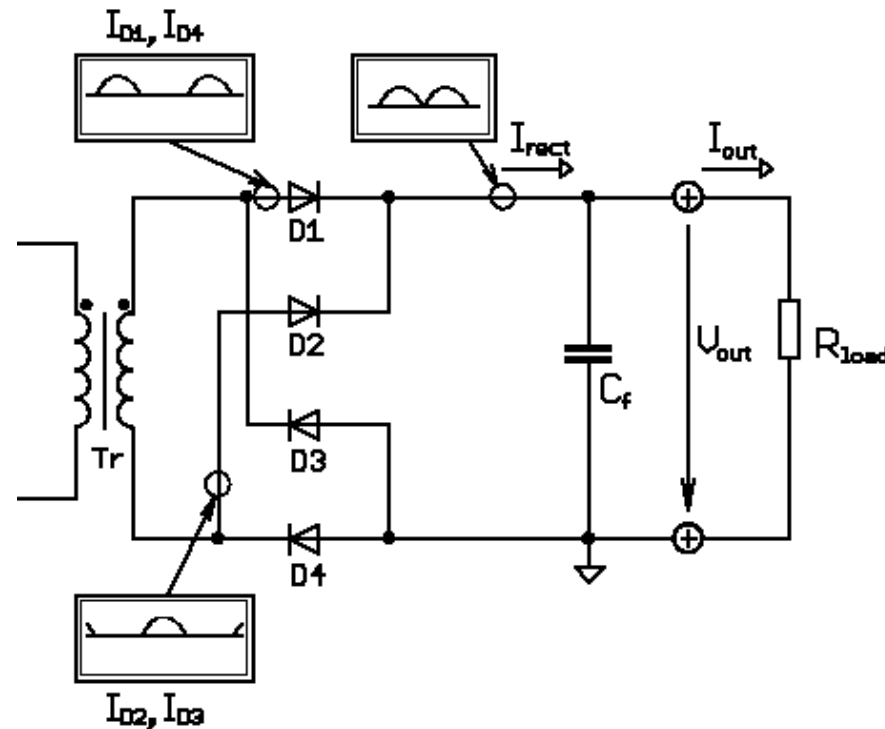
- To simplify calculations, assume sinusoidal current and  $F_{op} = F_s$

## Rectifier Losses – Push-Pull

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

| 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

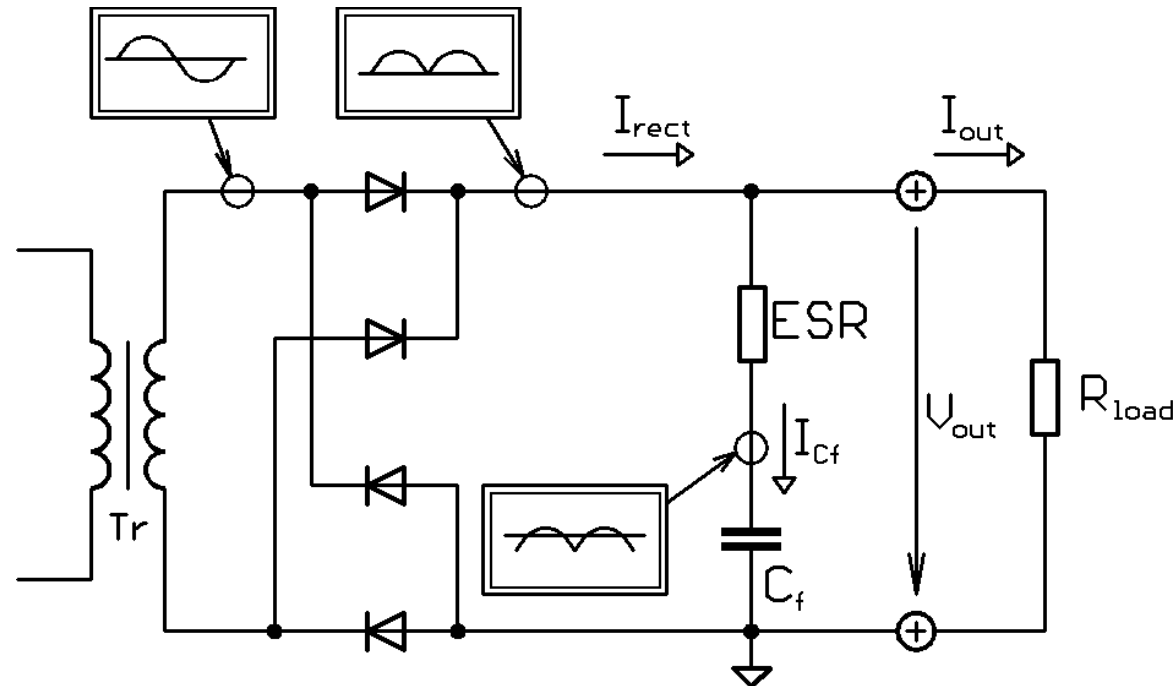
## Secondary Rectifier Design Procedure

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



## 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:<br>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:<br>Cf=100 uF, Fop=100 kHz<br><br>$V_{out\_ripple\_cap} = 104 \text{ mV}$ |

## Filter Capacitor Design Procedure

1. Calculate peak and rms rectifier and capacitor currents based on  $I_o$  and  $V_{out}$
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.  
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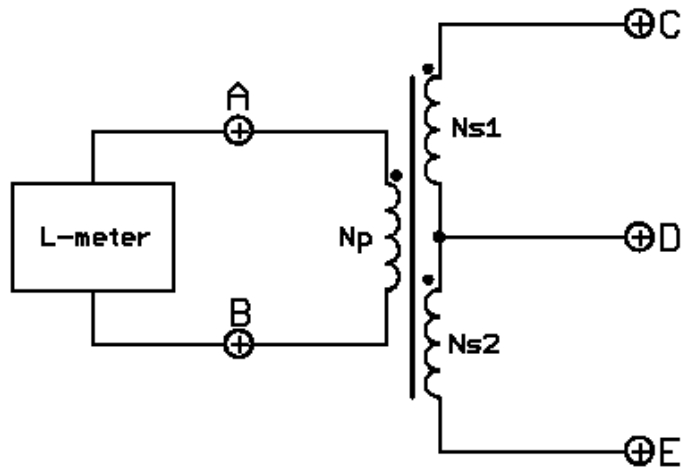
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# Resonant inductance balance

## 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<br>D-E open        |
| $L_{lk(p-s2)}$  | A-B                   | C-D open<br>D-E short        |
| $L_{lk(total)}$ | A-B                   | C-D short<br>D-E short       |
| $L_m$           | A-B                   | C-D open<br>D-E open         |

Example:  $L_{lk(p-s1)} = 105 \mu H$

$L_{lk(p-s2)} = 115 \mu H$

$\Delta L_{lk} = 10 \mu H$

$L_{lk(total)} = 100 \mu H$

$L_m = 600 \mu H$

$C_s = 33 \text{ nF}$

$f_{s1} = 85.5 \text{ kHz}$

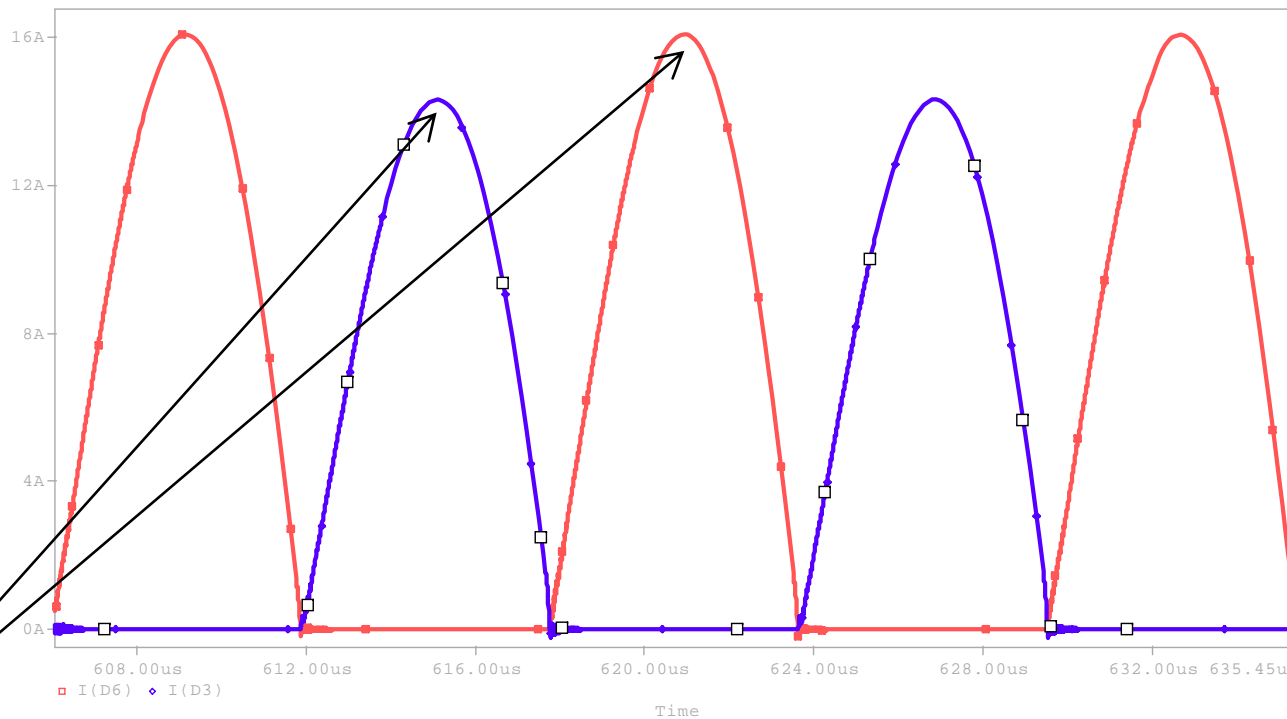
$f_{s2} = 81.7 \text{ kHz}$



**5 % difference**

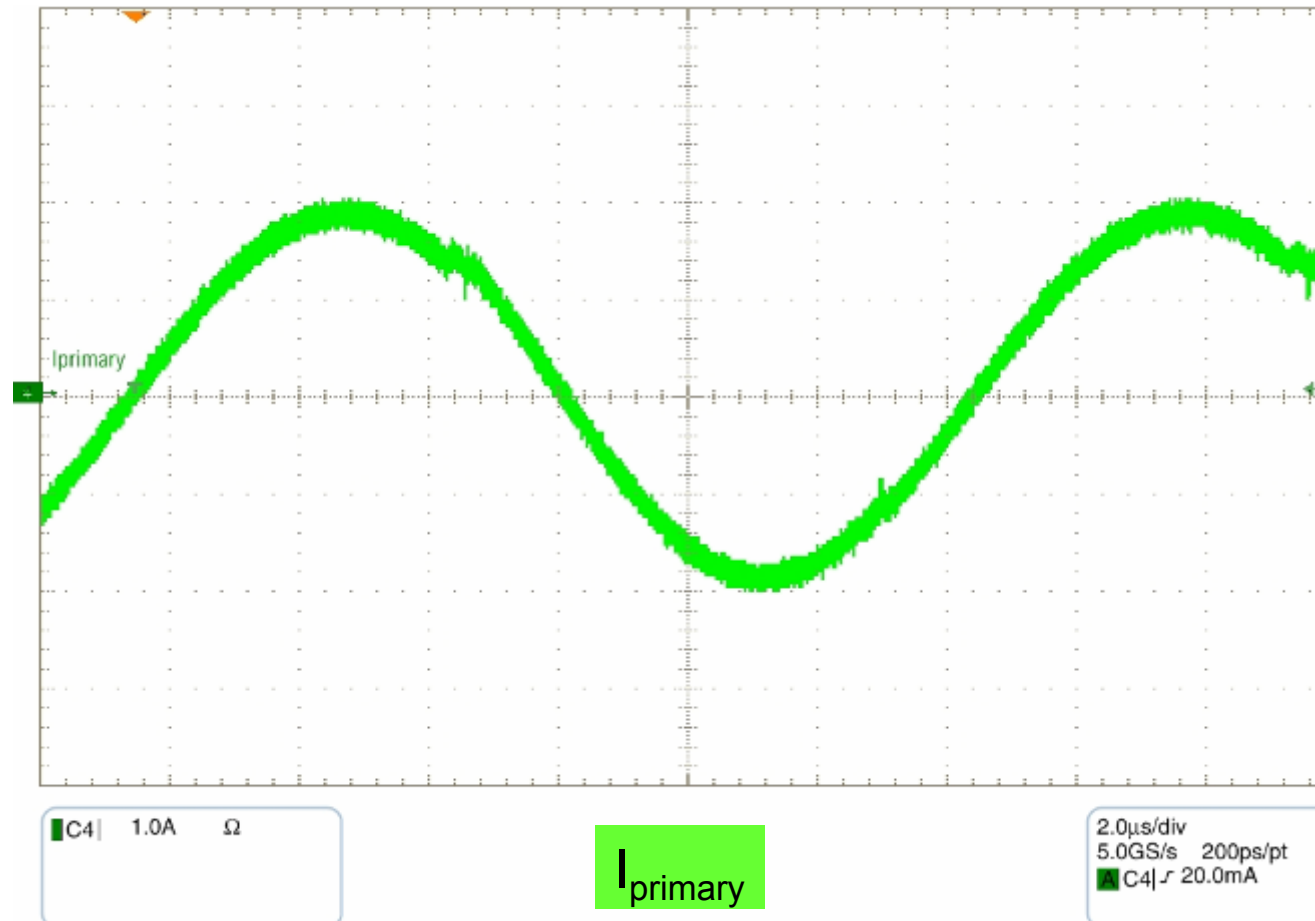
## Resonant inductance balance

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.

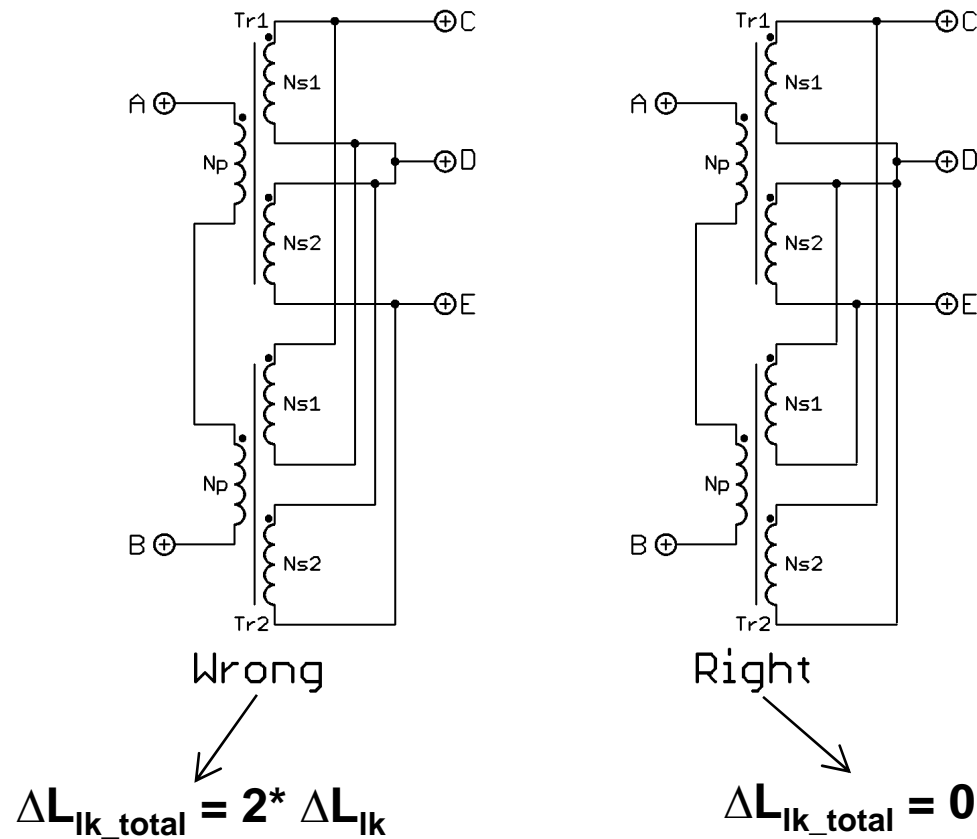
## Resonant inductance balance



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

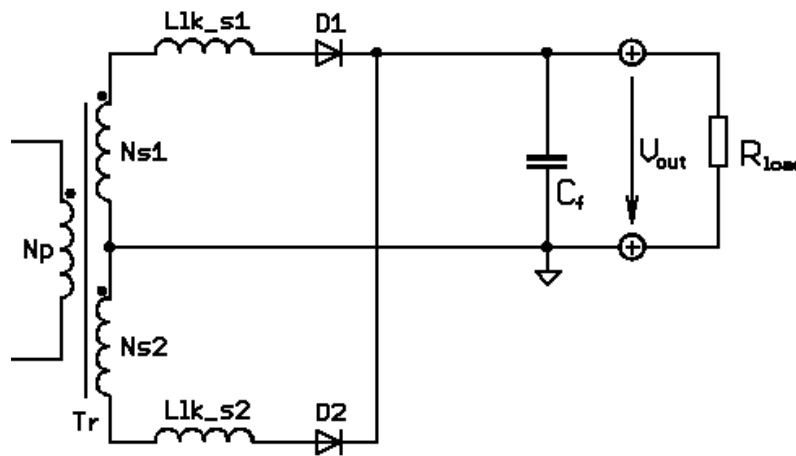
For high power app. it is attractive to connect primary windings in series and secondary windings in parallel. There is possibility to compensate transformer leakage imbalance by appropriate connection of the secondary windings:





## 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:**

$$N_p = 35 \text{ turns}$$

$$L_{lk\_s1} = 100 \text{ nH}$$

$$N_s = 2 \times 2 \text{ turns}$$

$$L_{lk\_s2} = 150 \text{ nH}$$

$$n = N_p / N_s = 17.5$$

$$L_s = 110 \text{ uH}$$

$$L_m = 630 \text{ uH}$$

$$\Delta L_{lk\_s} = 50 \text{ nH} \quad \longrightarrow \quad \Delta L_s = \Delta L_{lk\_s} \cdot n^2 = 15.3 \mu\text{H}$$

**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.

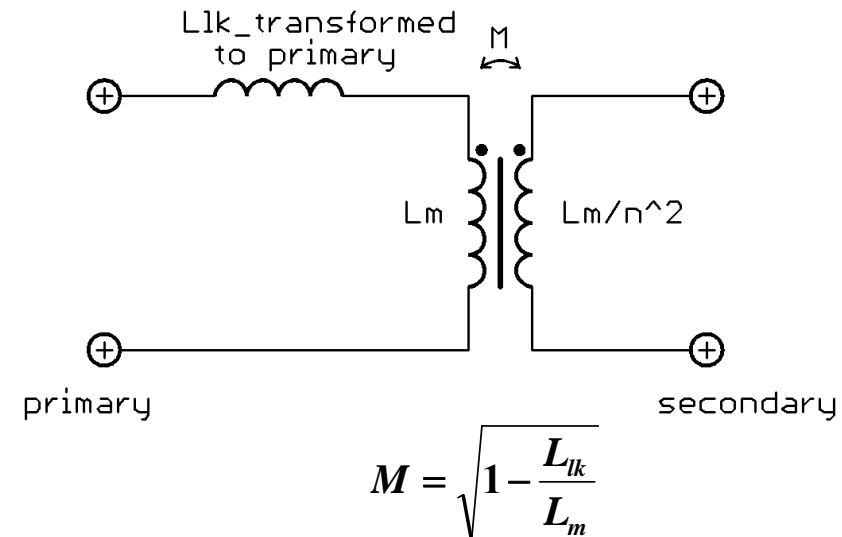
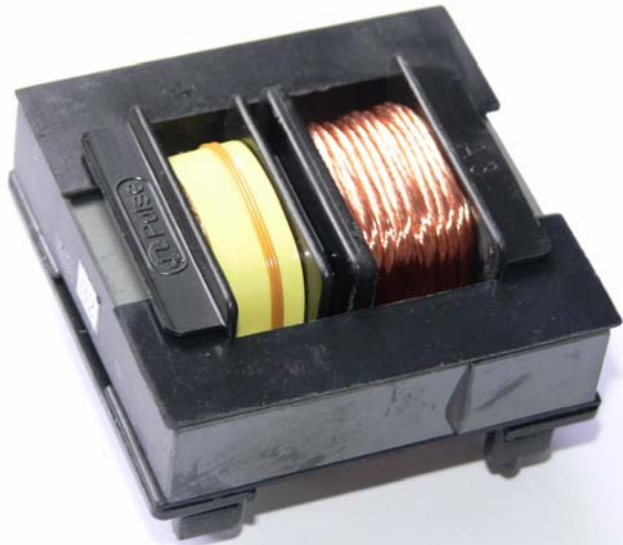
$$I_{primary\_RMS} \approx \sqrt{\frac{1}{8} \cdot \left( \frac{I_{out}^2 \cdot \pi^2}{n^2} + \frac{V_{bulk}^2}{16 \cdot L_m^2 \cdot f_{sw}^2} \right)}$$

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

- The skin effect and mainly proximity effect decreases effective copper 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

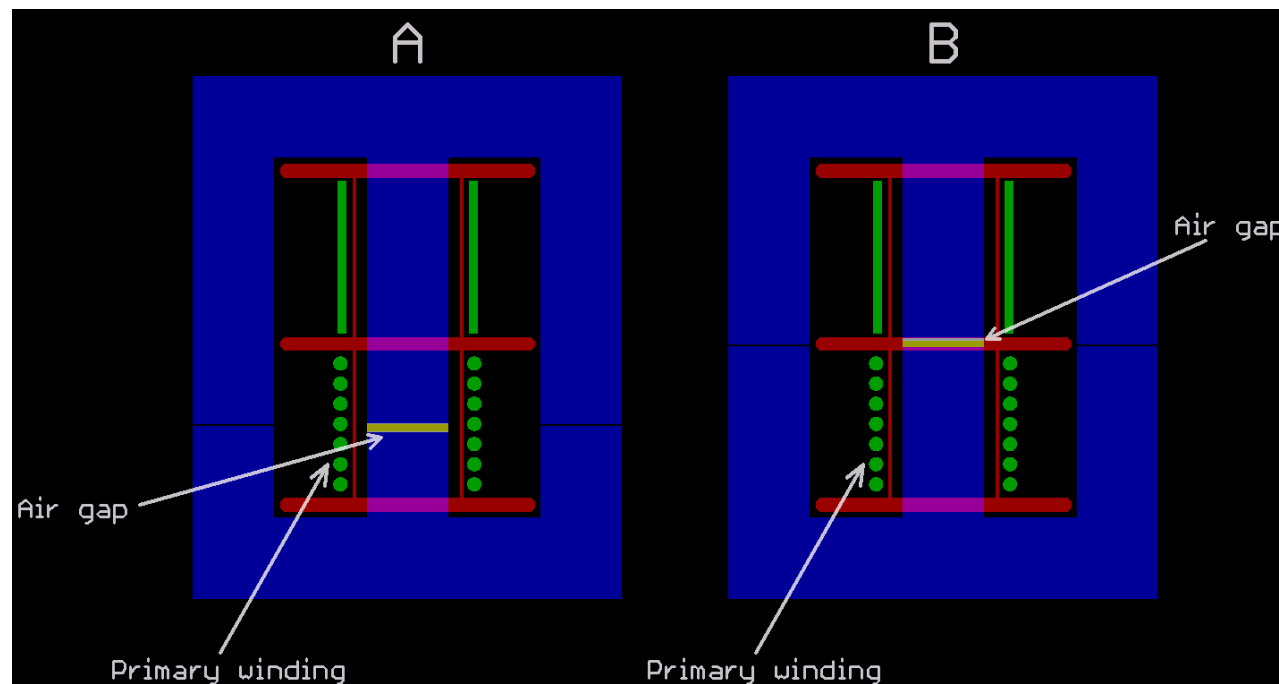
- For the standard transformer with good coupling ( $L_{lk} < 0.1 * L_m$ ) is the leakage inductance independent on the air gap thickness and position



- 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.:  $L_m = n^2 \cdot A_l$
- 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!