Design Considerations for an LLC Resonant Converter

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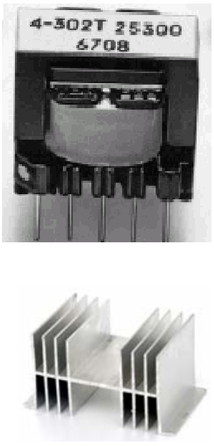
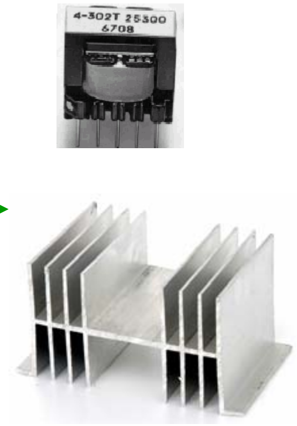
Power Conversion Team

1. Introduction

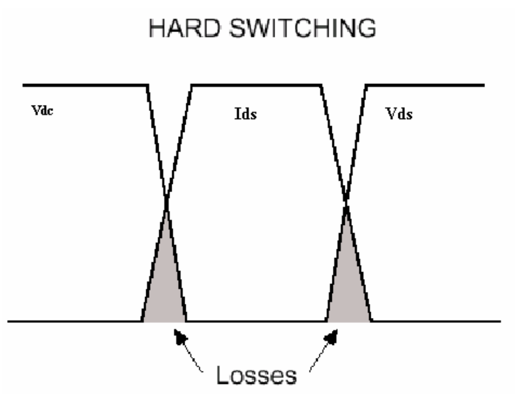
- Growing demand for higher power density and low profile in power converter has forced to increase switching frequency
- However, **Switching Loss** has been an obstacle to high frequency operation



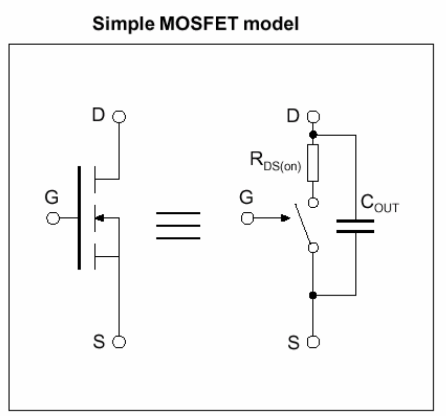
High frequency operation



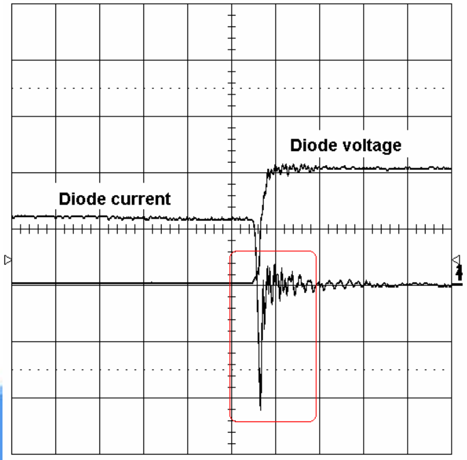
Overlap of voltage and current



Capacitive loss

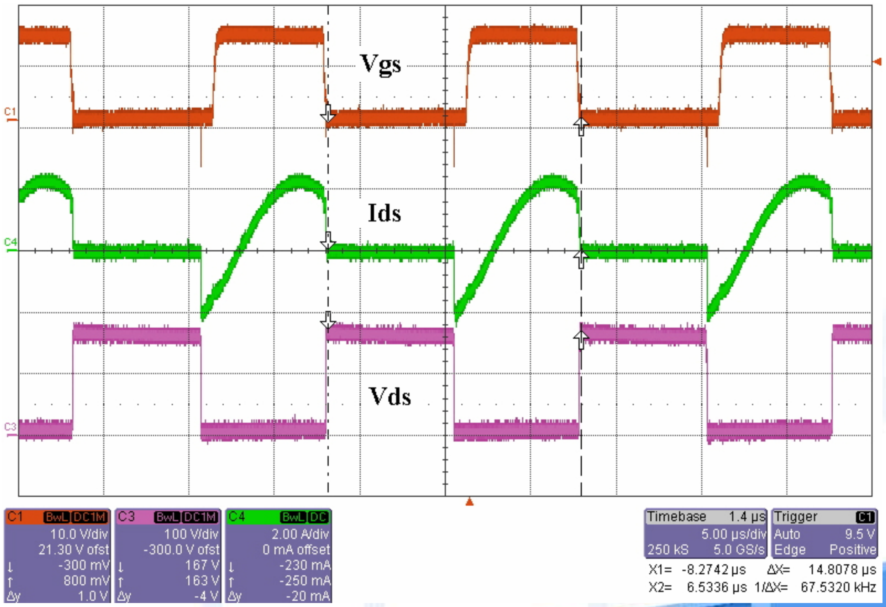


Reverse recovery loss



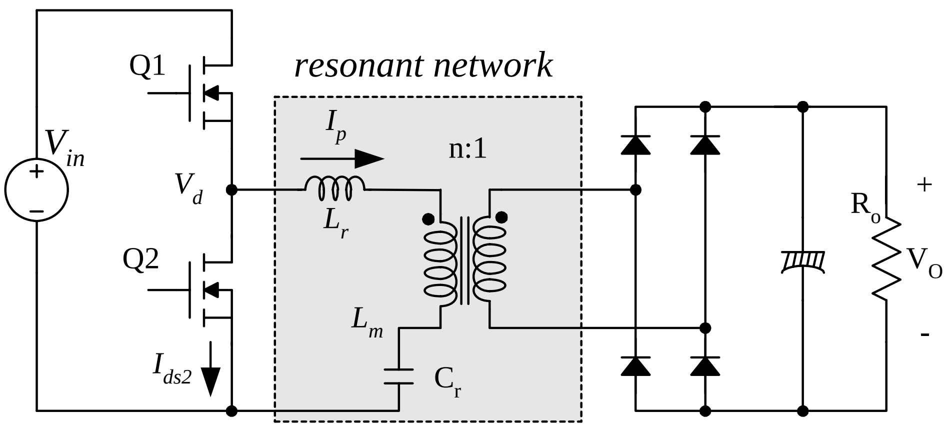
1. Introduction

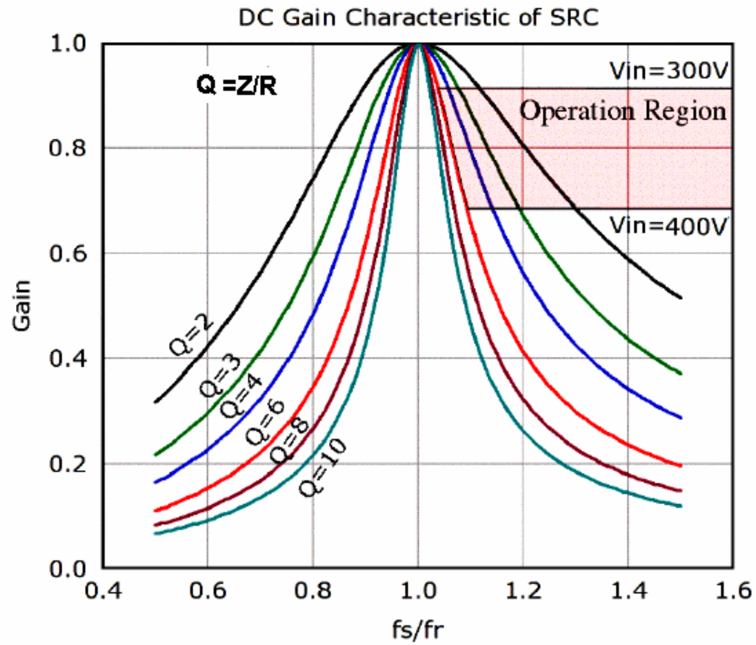
- **Resonant converter: processes power in a sinusoidal manner and the switching devices are softly commutated**
  - ✓ Voltage across the switch drops to zero before switch turns on (ZVS)
    - Remove overlap area between V and I when turning on
    - Capacitive loss is eliminated
- **Series resonant converter / Parallel resonant converter**



1. Introduction

Series Resonant (SR) converter





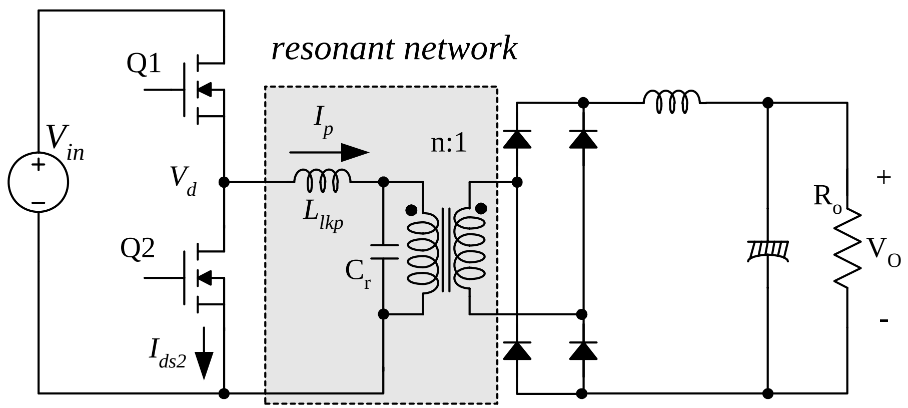
- The resonant inductor (Lr) and resonant capacitor (Cr) are in series
- The resonant capacitor is in series with the load
  - ✓ The resonant tank and the load act as a voltage divider→ DC gain is always lower than 1 (maximum gain happens at the resonant frequency)
  - ✓ The impedance of resonant tank can be changed by varying the frequency of driving voltage (V_d)

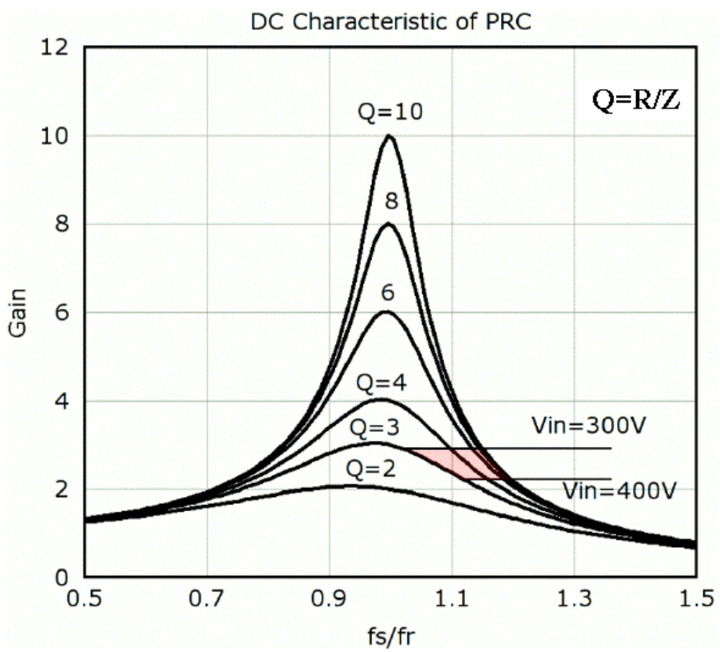
1. Introduction

Series Resonant (SR) converter

- Advantages
  - ✓ Reduced switching loss and EMI through ZVS → Improved efficiency
  - ✓ Reduced magnetic components size by high frequency operation
- Drawbacks
  - ✓ Can optimize performance at one operating point, but not with wide range of input voltage and load variations
  - ✓ Can not regulate the output at no load condition
  - ✓ Pulsating rectifier current (capacitor output): limitation for high output current application

Parallel Resonant (PR) converter





- The resonant inductor (Lr) and resonant capacitor (Cr) are in series
- The resonant capacitor is in parallel with the load
  - ✓ The impedance of resonant tank can be changed by varying the frequency of driving voltage (V_d)

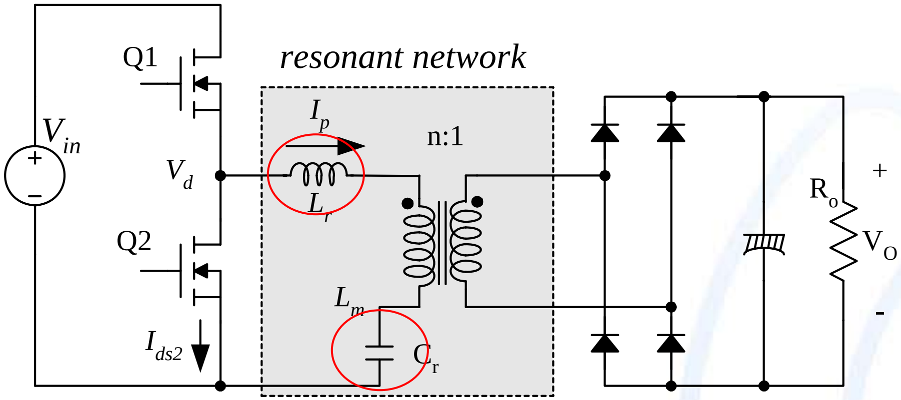
Parallel Resonant (PR) converter

- Advantages
  - ✓ No problem in output regulation at no load condition
  - ✓ Continuous rectifier current (inductor output): suitable for high output current application
- Drawbacks
  - ✓ The primary side current is almost independent of load condition: significant current may circulate through the resonant network, even at the no load condition
  - ✓ Circulating current increases as input voltage increases: limitation for wide range of input voltage

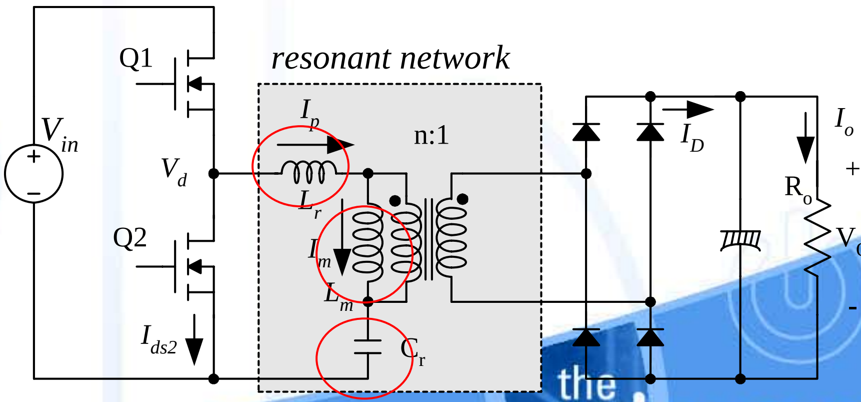
1. Introduction

- **What is LLC resonant converter?**
  - ✓ Topology looks almost same as the conventional LC series resonant converter
  - ✓ Magnetizing inductance (L_m) of the transformer is relatively small and involved in the resonance operation
  - ✓ Voltage gain is different from that of LC series resonant converter

LC Series resonant converter



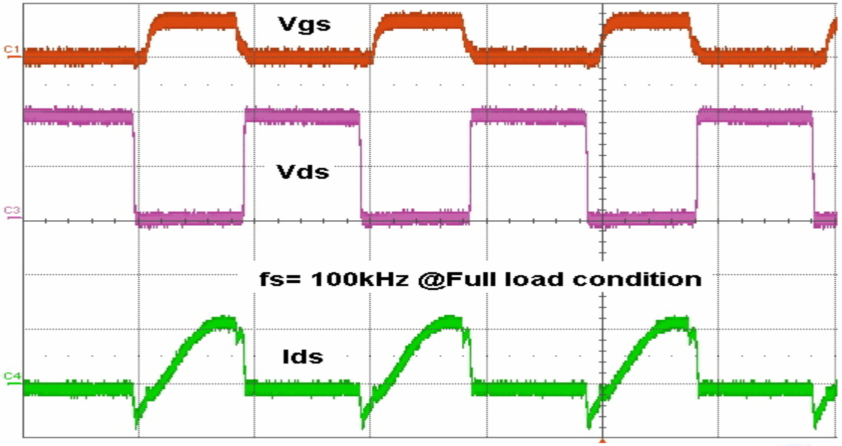
LLC resonant converter

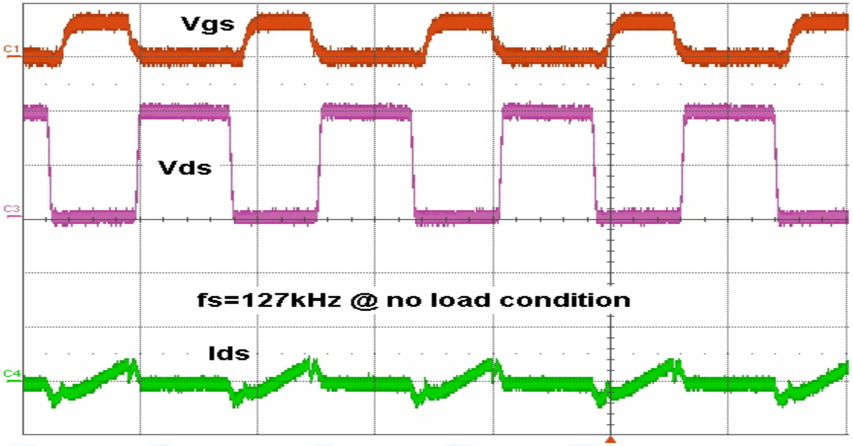


1. Introduction

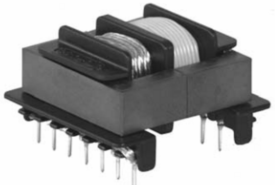
- **Features of LLC resonant converter**

  - Reduced switching loss through ZVS: Improved efficiency
  - Narrow frequency variation range over wide load range
  - Zero voltage switching even at no load condition



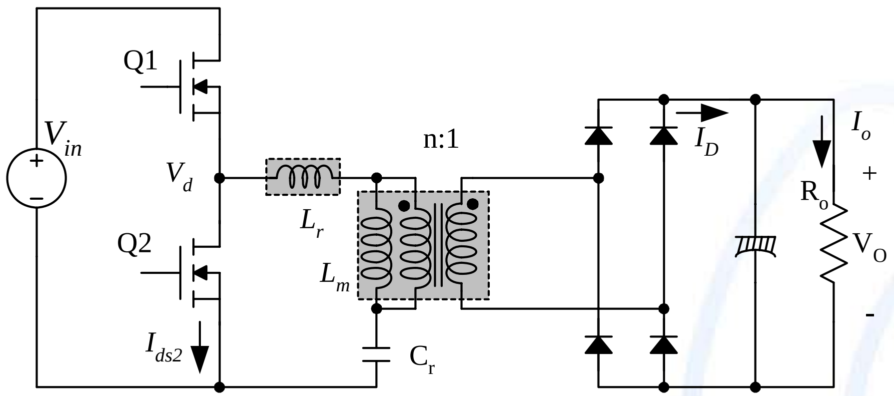


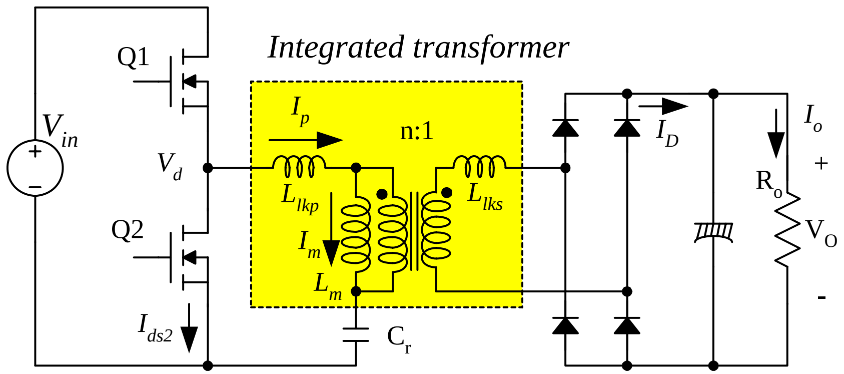
- Typically, **integrated transformer** is used instead of discrete magnetic components



1. Introduction

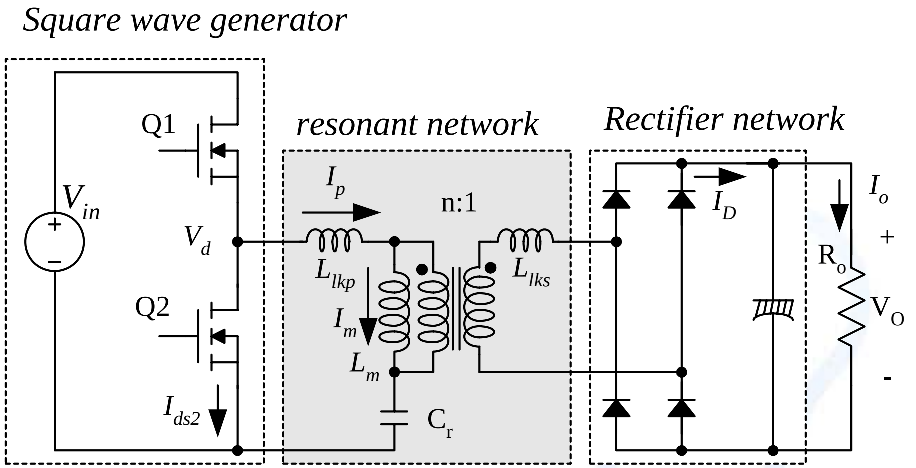
- **Integrated transformer in LLC resonant converter**
  - ✓ Two magnetic components are implemented with a single core (use the primary side leakage inductance as a resonant inductor)
  - ✓ One magnetic components (Lr) can be saved
  - ✓ Leakage inductance not only exists in the primary side but also in the secondary side
  - ✓ Need to consider the leakage inductance in the secondary side

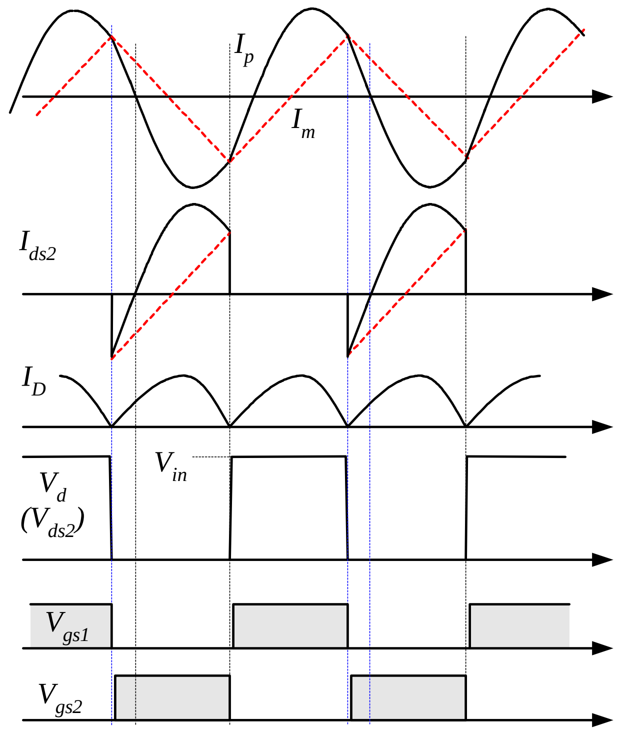




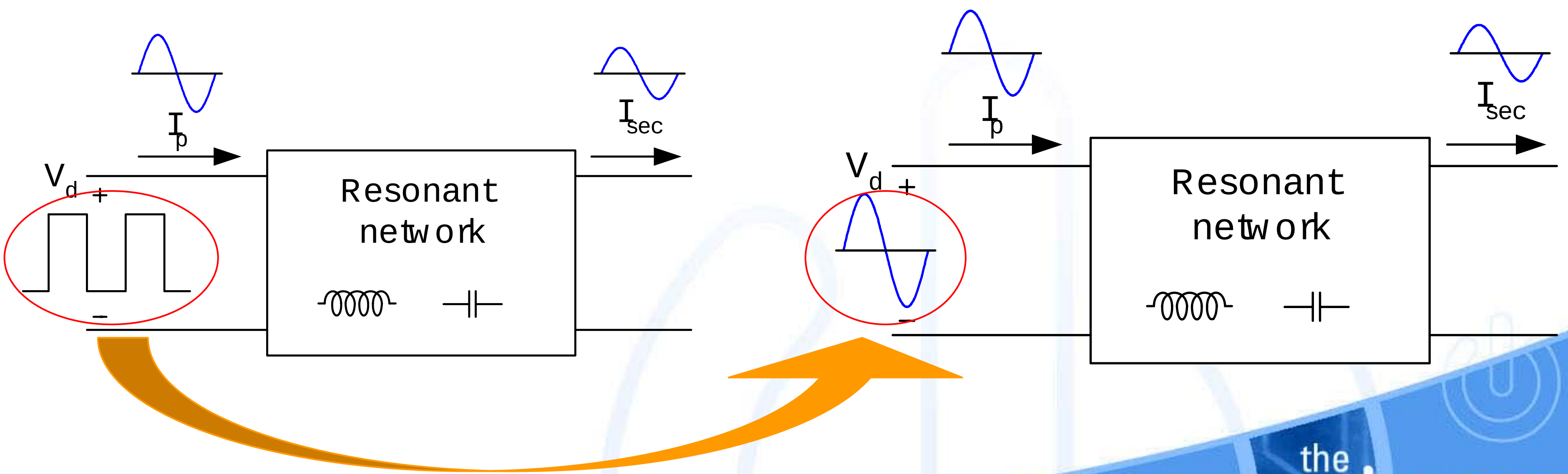
2. Operation principle and Fundamental Approximation

- Square wave generator: produces a square wave voltage, V_d by driving switches, Q1 and Q2 with alternating 50% duty cycle for each switch.
- Resonant network: consists of L_{lkp}, L_{lks}, L_m and C_r. The current lags the voltage applied to the resonant network which allows the MOSFET's to be turned on with zero voltage.
- Rectifier network: produces DC voltage by rectifying AC current





- The resonant network filters the higher harmonic currents. Thus, essentially only sinusoidal current is allowed to flow through the resonant network even though a square wave voltage (V_d) is applied to the resonant network.
- Fundamental approximation: assumes that only the fundamental component of the square-wave voltage input to the resonant network contributes to the power transfer to the output.
- The square wave voltage can be replaced by its fundamental component

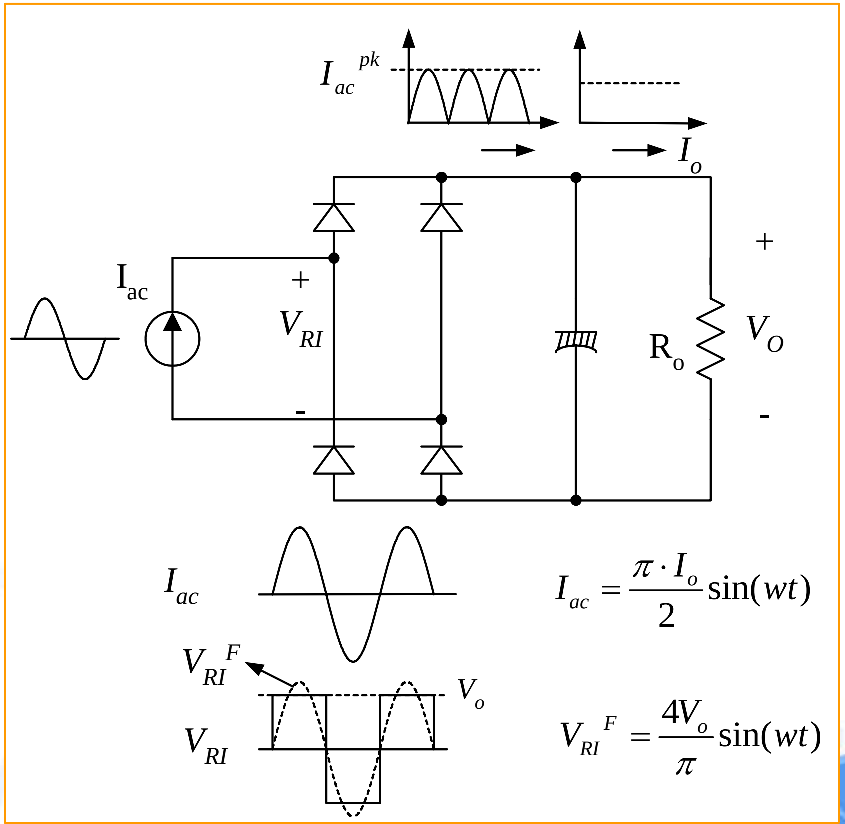


2. Operation principle and Fundamental Approximation

- Because the rectifier circuit in the secondary side acts as an impedance transformer, the equivalent load resistance is different from actual load resistance.

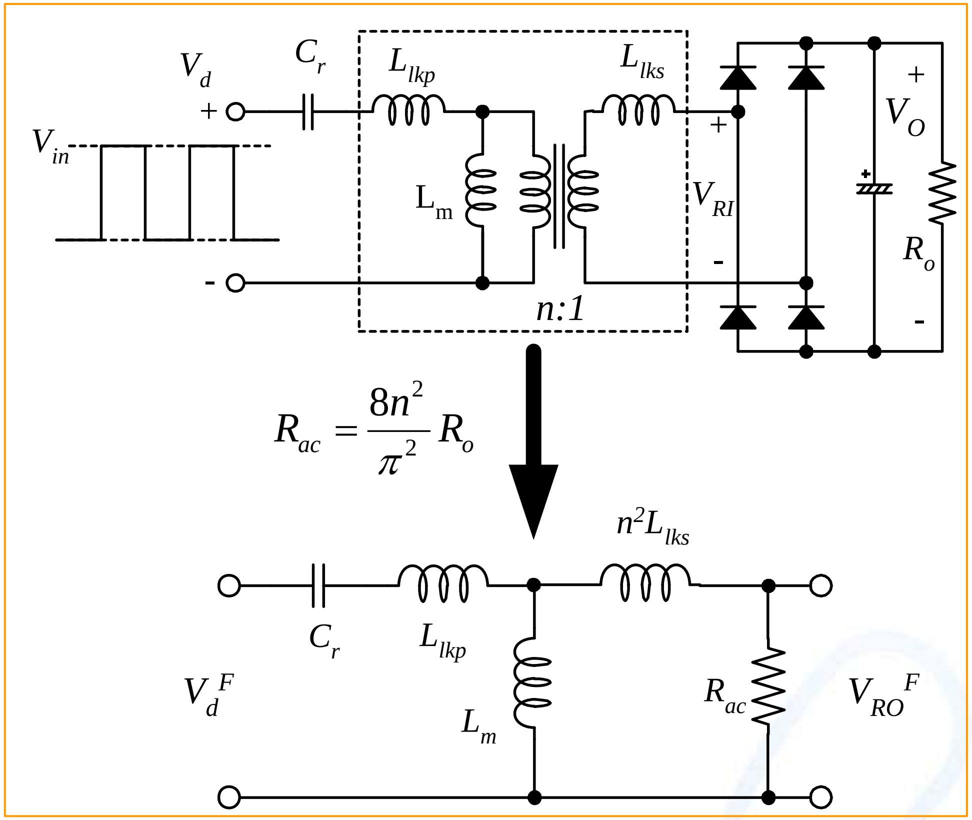
- The primary side circuit is replaced by a sinusoidal current source (I_{ac}) and a square wave of voltage (V_{RI}) appears at the input to the rectifier.
- The equivalent load resistance is obtained as

$$R_{ac} = \frac{V_{RI}{}^{F}}{I_{ac}{}^{F}} = \frac{V_{RI}{}^{F}}{I_{ac}} = \frac{8}{\pi^2}\frac{V_o}{I_o} = \frac{8}{\pi^2}R_o$$



2. Operation principle and Fundamental Approximation

▪ AC equivalent circuit (L-L-L-C)



$$M = \frac{V_{RO}^F}{V_d^F} = \frac{n \cdot V_{RI}^F}{V_d^F} = \frac{\frac{4n \cdot V_o}{\pi}\sin(\omega t)}{\frac{4}{\pi}\frac{V_{in}}{2}\sin(\omega t)} = \frac{2n \cdot V_o}{V_{in}}$$

$$= \left| \frac{\omega^2 L_m R_{ac} C_r}{j\omega \cdot (1 - \frac{\omega^2}{\omega_o^2}) \cdot (L_m + n^2 L_{lks}) + R_{ac}(1 - \frac{\omega^2}{\omega_p^2})} \right|$$

$$R_{ac} = \frac{8n^2}{\pi^2} R_o$$

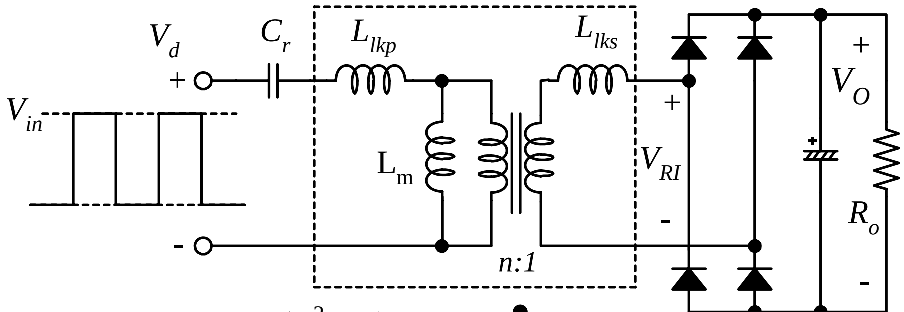
$$\omega_o = \frac{1}{\sqrt{L_r C_r}}, \quad \omega_p = \frac{1}{\sqrt{L_p C_r}}$$

$$L_p = L_m + L_{lkp}, \quad L_r = L_{lkp} + L_m // (n^2 L_{lks})$$

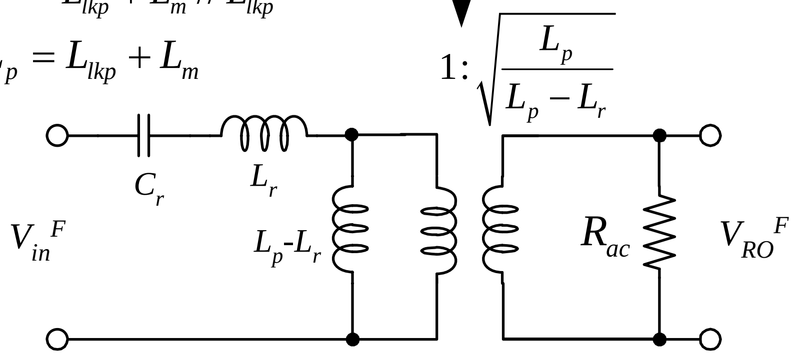
- Lr is measured in primary side with secondary winding short circuited
- Lp is measured in primary side with secondary winding open circuited

2. Operation principle and Fundamental Approximation

▪ Simplified AC equivalent circuit (L-L-C)



$$L_r = L_{lkp} + L_m // (n^2 L_{lks})$$
$$= L_{lkp} + L_m // L_{lkp}$$
$$L_p = L_{lkp} + L_m$$



- Lr is measured in primary side with secondary winding short circuited

- Lp is measured in primary side with secondary winding open circuited

Assuming $L_{lkp}=n^2L_{lks}$

$$M = \frac{2n \cdot V_O}{V_{in}} = \left| \frac{\left(\frac{\omega^2}{\omega_p^{\;2}}\right)\frac{k}{k+1}}{j\left(\frac{\omega}{\omega_o}\right)\cdot\left(1-\frac{\omega^2}{\omega_o^{\;2}}\right)\cdot Q\frac{(k+1)^2}{2k+1} + \left(1-\frac{\omega^2}{\omega_p^{\;2}}\right)} \right|$$

$$Q = \frac{\sqrt{L_r / C_r}}{R_{ac}} \qquad k = \frac{L_m}{L_{lkp}}$$

Expressing in terms of L_p and L_r

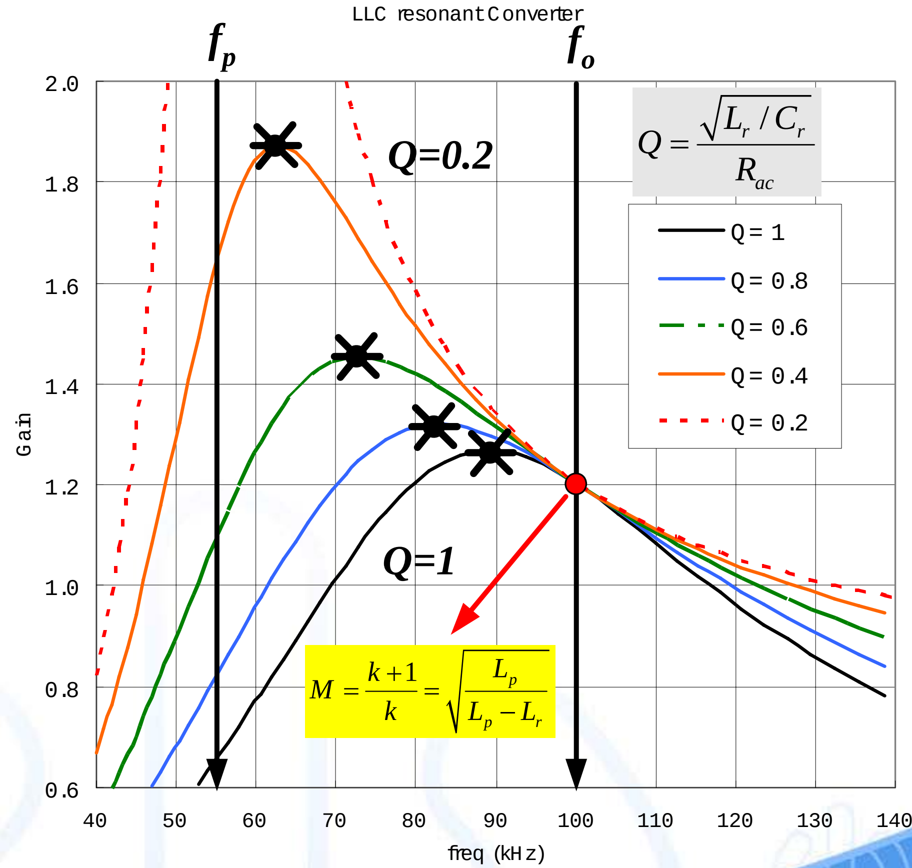
$$M = \frac{2n \cdot V_O}{V_{in}} = \left| \frac{\left(\frac{\omega^2}{\omega_p^{\;2}}\right)\sqrt{\frac{L_p - L_r}{L_p}}}{j\left(\frac{\omega}{\omega_o}\right)\cdot\left(1-\frac{\omega^2}{\omega_o^{\;2}}\right)\cdot Q\frac{L_p}{L_r} + \left(1-\frac{\omega^2}{\omega_p^{\;2}}\right)} \right|$$

2. Operation principle and Fundamental Approximation

- Gain characteristics
  - ✓ Two resonant frequencies (f_o and f_p) exist
  - ✓ The gain is fixed at resonant frequency (f_o) regardless of the load variation

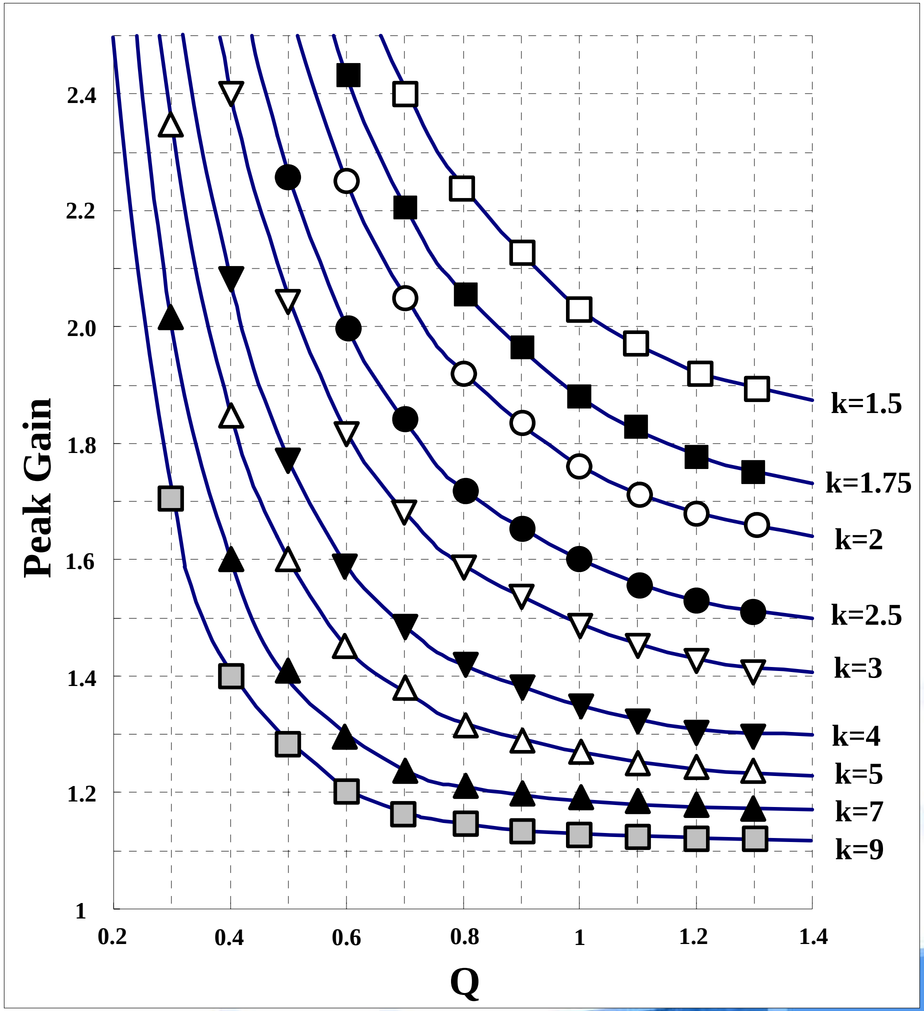
$$M_{@\omega=\omega_o}=\frac{k+1}{k}=\sqrt{\frac{L_p}{L_p-L_r}}$$

  - ✓ Peak gain frequency exists between f_o and f_p
  - ✓ As Q decreases (as load decreases), the peak gain frequency moves to f_p and higher peak gain is obtained.
  - ✓ As Q increases (as load increases), peak gain frequency moves to f_o and the peak gain drops



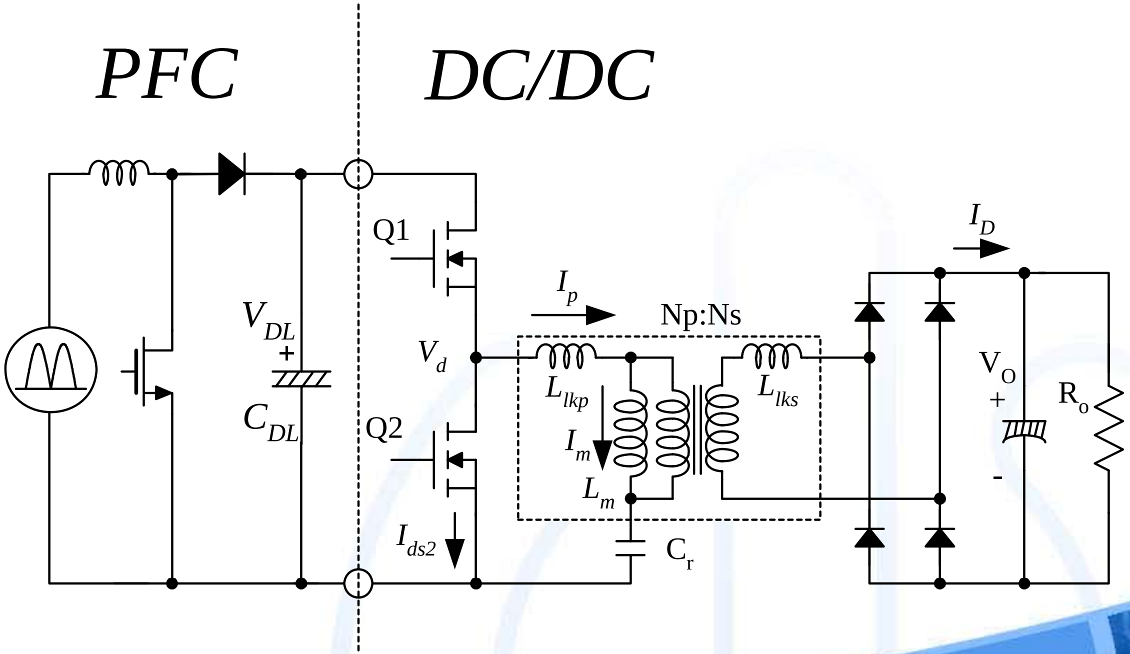
2. Operation principle and Fundamental Approximation

- Peak gain (attainable maximum gain) versus Q for different k values



3. Design procedure

- Design example
  - Input voltage: 380Vdc (output of PFC stage)
  - Output: 24V/5A (120W)
  - Holdup time requirement: 17ms
  - DC link capacitor of PFC output: 100uF



3. Design procedure

[STEP-1] Define the system specifications

- ✓ Estimated efficiency (*E*ff)
- ✓ Input voltage range: hold up time should be considered for minimum input voltage

$$V_{in}^{\min} = \sqrt{V_{O.PFC}^{\;2} - \frac{2P_{in}T_{HU}}{C_{DL}}}$$

(Design Example) Assuming the efficiency is 95%,

$$P_{in} = \frac{P_o}{E_{ff}} = \frac{120}{0.95} = 126W$$

$$V_{in}^{\min} = \sqrt{V_{O.PFC}^{\;2} - \frac{2P_{in}T_{HU}}{C_{DL}}}$$

$$= \sqrt{380^2 - \frac{2\cdot 126\cdot 17\times 10^{-3}}{100\times 10^{-6}}} = 319V$$

$$V_{in}^{\max} = V_{O.PFC} = 380V$$

3. Design procedure

[STEP-2] Determine the maximum and minimum voltage gains of the resonant network by choosing k $(k = L_m / L_{lkp})$

- it is typical to set k to be 5~10, which results in a gain of 1.1~1.2 at f_o

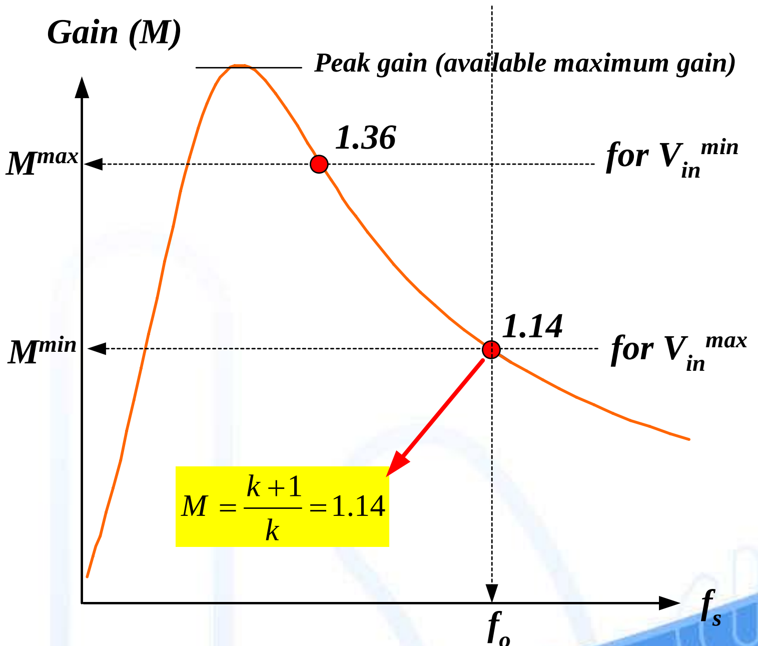
$$M^{\min} = \frac{V_{RO}}{\frac{V_{in}^{\max}}{2}} = \frac{L_m + n^2 L_{lks}}{L_m} = \frac{L_m + L_{lkp}}{L_m} = \frac{k+1}{k}$$

$$M^{\max} = \frac{V_{in}^{\max}}{V_{in}^{\min}} M^{\min}$$

(Design Example) The ratio (k) between L_m and L_{lkp} is determined as 7, which results in the minimum gain as

$$M^{\min} = \frac{V_{RO}}{\frac{V_{in}^{\max}}{2}} \cong \frac{k+1}{k} = \frac{7+1}{7} = 1.14$$

$$M^{\max} = \frac{V_{in}^{\max}}{V_{in}^{\min}} M^{\min} = \frac{380}{319} \cdot 1.14 = 1.36$$



3. Design procedure

[STEP-3] Determine the transformer turns ratio (n=Np/Ns)

$$n = \frac{N_p}{N_s} = \frac{V_{in}^{\max}}{2(V_o + V_F)} \cdot M^{\min}$$

(Design Example)

$$n = \frac{N_p}{N_s} = \frac{V_{in}^{\max}}{2(V_o + V_F)} \cdot M_{\min} = \frac{380}{2(24+1.2)} \cdot 1.14 = 8.6$$

[STEP-4] Calculate the equivalent load resistance (Rac)

$$R_{ac} = \frac{8n^2}{\pi^2}\frac{V_o^2}{P_o}$$

(Design Example)

$$R_{ac} = \frac{8n^2}{\pi^2}\frac{V_o^2}{P_o} = \frac{8 \cdot 8.6^2 \cdot 24^2}{\pi^2 \cdot 120} = 288\Omega$$

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3. Design procedure

[STEP-5] Design the resonant network

- With k chosen in STEP-2, read proper Q from gain curves

$$k = 7,\ M^{\max} = 1.36$$

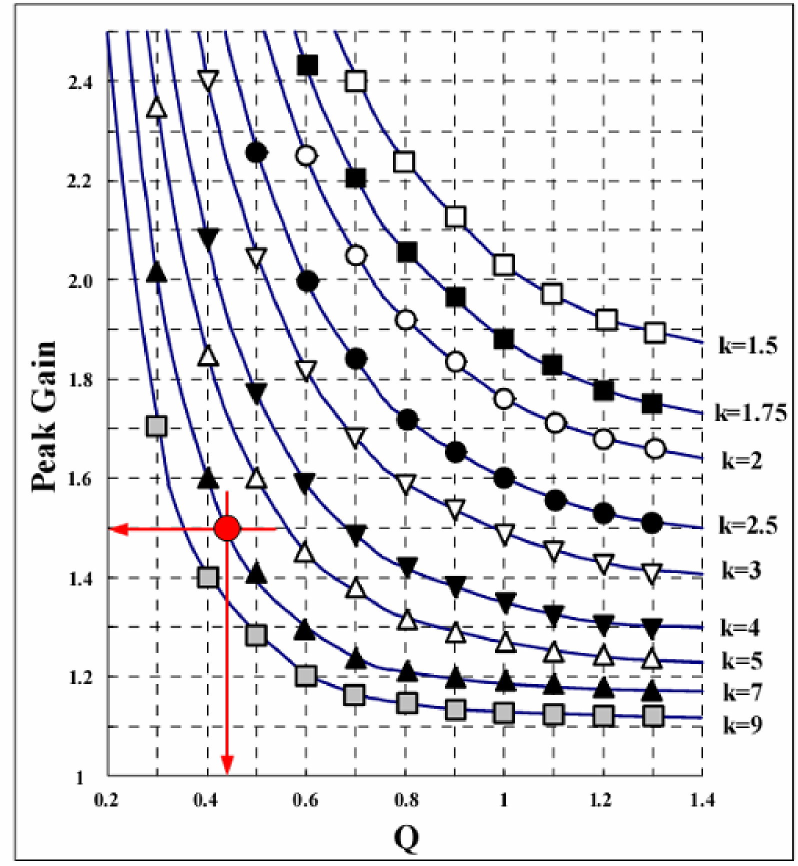
$$peak\ gain = 1.36 \times 110\% = 1.5$$

(Design Example)
As calculated in STEP-2, the maximum voltage gain (M^{max}) for the minimum input voltage (V_{in}^{min}) is 1.36. With 10% margin, a peak gain of 1.5 is required. k has been chosen as 7 in STEP-2 and Q is obtained as 0.43 from the peak gain curves in Fig. 12. By selecting the resonant frequency as 85kHz, the resonant components are determined as

$$C_r = \frac{1}{2\pi Q \cdot f_o \cdot R_{ac}} = \frac{1}{2\pi \cdot 0.43 \cdot 85\times10^3 \cdot 288}$$
$$= 15nF$$

$$L_r = \frac{1}{(2\pi f_o)^2 C_r} = \frac{1}{(2\pi \cdot 85\times10^3)^2 \cdot 15\times10^{-9}}$$
$$= 234uH$$

$$L_p = \frac{(k+1)^2}{(2k+1)} L_r = 998uH$$



3. Design procedure

[STEP-6] Design the transformer

- Plot the gain curve and read the minimum switching frequency. Then, the minimum number of turns for the transformer primary side is obtained as

$$N_p^{\min} = \frac{n(V_o + V_F)}{2 f_s^{\min} \cdot \Delta B \cdot A_e}$$

(Design Example) EER3541 core (A_e=107mm^2) is selected for the transformer. From the gain curve of Fig .12, the minimum switching frequency is obtained as 66kHz. Then, the minimum primary side turns of the transformer is given as

$$N_p^{\min} = \frac{n(V_o + V_F)\times 10^6}{2 f_s^{\min} \Delta B \cdot A_e}$$

$$= \frac{8.6 \times 25.2 \times 10^6}{2 \cdot 66 \times 10^3 \cdot 0.3 \cdot 107} = 51.1\ turns$$

$$\therefore N_p = n \cdot N_s = 8.6 \times 6 = 51.6 > N_p^{\min}$$

Choosing Ns as 6 turns, Np is given as

$$N_p = n \cdot N_s = 8.6 \times 6 = 51.6 \Rightarrow 52 > N_p^{\min}$$

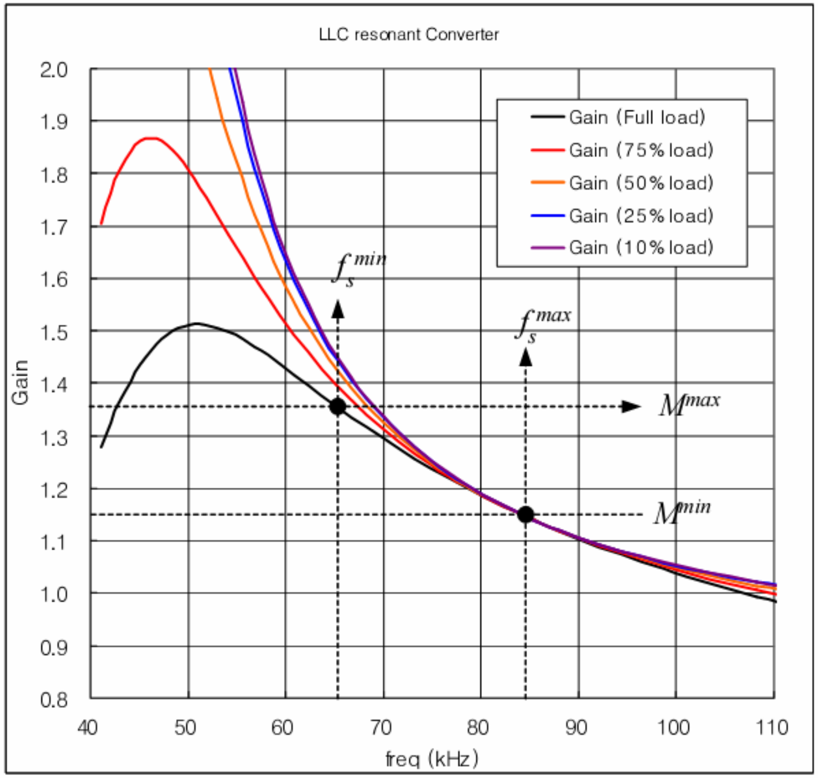
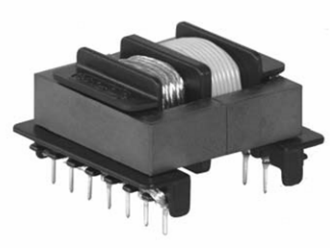


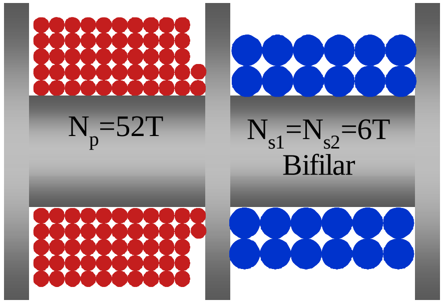
Fig. 12 Gain curve

3. Design procedure

[STEP-7] Transformer Construction

- Since LLC converter design results in relatively large L_r, usually sectional bobbin is typically used
- # of turns and winding configuration are the major factors determining Lr
- Gap length of the core does not affect Lr much
- Lp can be easily controlled with gap length





Gap length	L_p	L_r
0.0 mm	5,669 μH	237 μH
0.05 mm	2,105 μH	235 μH
0.10 mm	1,401 μH	233 μH
0.15 mm	**1,065 μH**	**230 μH**
0.20 mm	890 μH	225 μH
0.25 mm	788 μH	224 μH
0.30 mm	665 μH	223 μH
0.35 mm	623 μH	222 μH

Design value: L_r=234uH, L_p=998uH

3. Design procedure

[STEP-8] Select the resonant capacitor

$$I_{C_r}^{RMS} \cong \sqrt{[\frac{\pi I_o}{2\sqrt{2}n}]^2 + [\frac{n(V_o + 2\cdot V_F)}{4\sqrt{2} f_o L_m}]^2} \qquad V_{C_r}^{\max} \cong \frac{V_{in}^{\max}}{2} + \frac{\sqrt{2}\cdot I_{Cr}^{RMS}}{2\cdot\pi\cdot f_o \cdot C_r}$$

(Design Example)

$$I_{C_r}^{RMS} \cong \sqrt{[\frac{\pi I_o}{2\sqrt{2}n}]^2 + [\frac{n(V_o + 2\cdot V_F)}{4\sqrt{2} f_o L_m}]^2}$$

$$= \sqrt{[\frac{\pi\cdot 5}{2\sqrt{2}\cdot 8.6}]^2 + [\frac{8.6\cdot(24+1.2)}{4\sqrt{2}\cdot 873\times10^{-6}\cdot 85\times10^{3}}]^2}$$

$$= 0.87A$$

$$V_{C_r}^{\max} \cong \frac{V_{in}^{\max}}{2} + \frac{\sqrt{2}\cdot I_{Cr}^{RMS}}{2\cdot\pi\cdot f_o\cdot C_r}$$

$$= \frac{380}{2} + \frac{\sqrt{2}\cdot 0.916}{2\cdot\pi\cdot 85\times10^{3}\cdot 15\times10^{-9}} = 343V$$

4. Conclusion

- Using a fundamental approximation, gain equation has been derived
- Leakage inductance in the secondary side is also considered (L-L-L-C model) for gain equation
- L-L-L-C equivalent circuit has been simplified as a conventional L-L-C equivalent circuit
- Practical design consideration has been presented

Appendix - FSFR-series

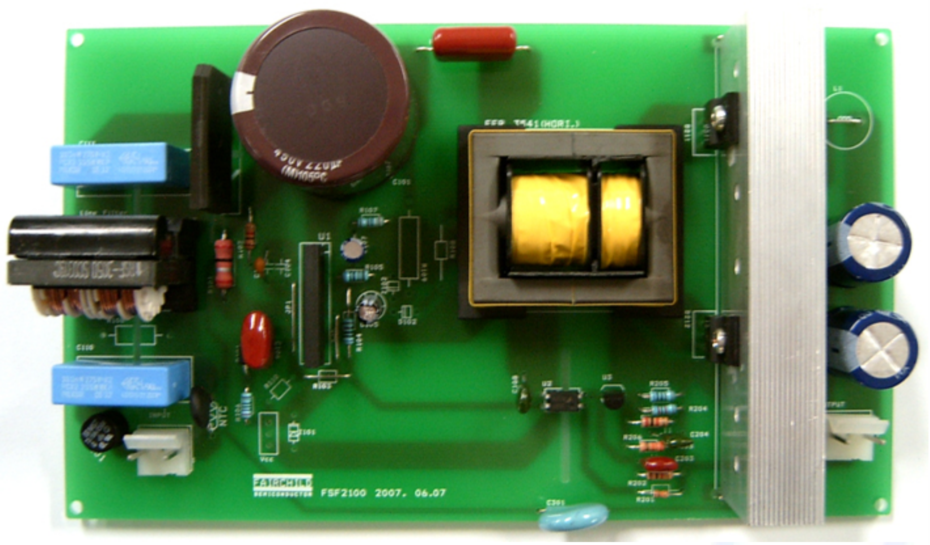
- Variable frequency control with 50% duty cycle for half-bridge resonant converter topology
- High efficiency through zero voltage switching (ZVS)
- Internal Super-FETs with Fast Recovery Type Body Diode (trr=120ns)
- Fixed dead time (350ns)
- Up to 300kHz operating frequency
- Pulse skipping for Frequency limit (programmable) at light load condition
- Simple remote ON/OFF control
- Various Protection functions: Over Voltage Protection (OVP), Over Current Protection (OCP), Abnormal Over Current Protection (AOCP), Internal Thermal Shutdown (TSD)

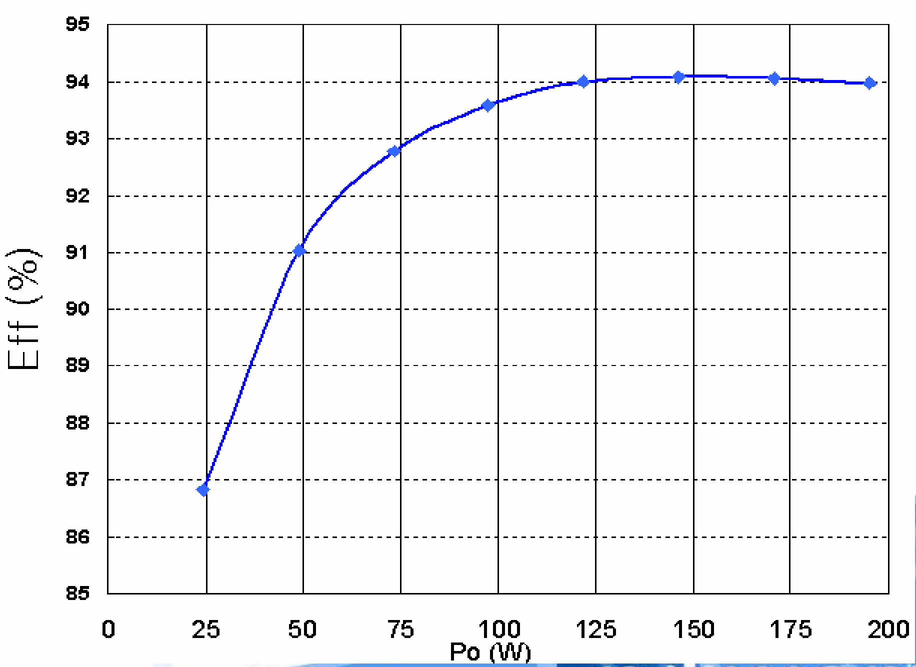
Part Number	Package	Operating Ambient Temperature Ranges	$R_{DS(ON)}$ (MAX)	Maximum Output Power without heat sink (Vin=350~400V) (1) (2)	Maximum Output Power with heat sink (Vin=350~400V) (1) (2)
FSFR2100	9-SIP	-40 to +85 °C	0.38Ω	200W	450W
FSFR2000	9-SIP	-40 to +85 °C	0.6Ω	160W	350W
FSFR1900	9-SIP	-40 to +85 °C	0.8Ω	140W	300W

Appendix - FSFR-series demo board

Application	FPS device	Input voltage range	Rated output power	Output voltage (Rated current)
LCD TV	FSFR2100	Vin nominal : 390Vdc* (340~400Vdc) Vcc supply: 16~20V	192W	24V-8A

* 20ms hold up time for Vin=390Vdc





Appendix - FSFR-series demo board

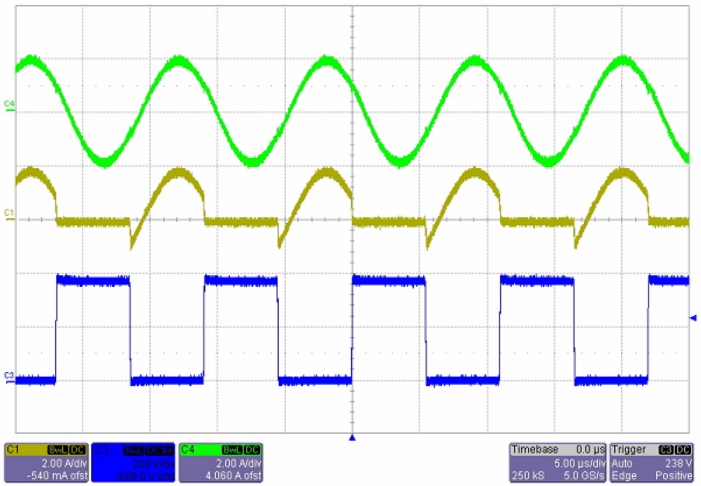


Figure 6. Operation waveforms at nominal input voltage [Vin=390Vdc, Po=192W (24V/8A)]
C4: Transformer Primary side current (2A/div), **C1**: Low side MOSFET current (2A/div)
C3: Low side MOSFET Vds (200V/div), time: 5us/div

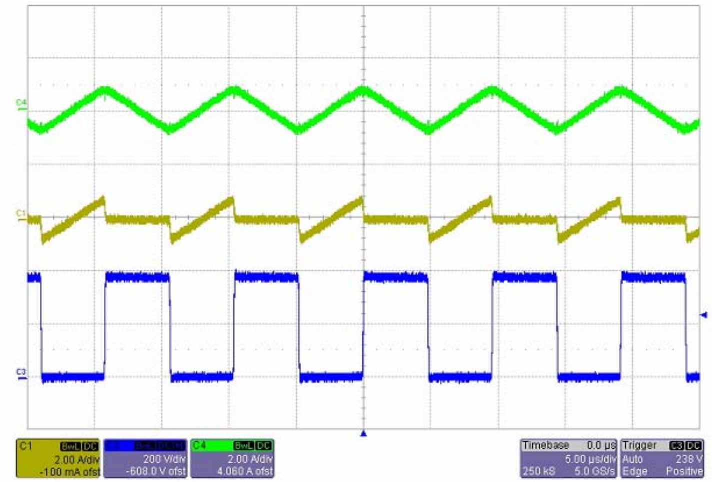


Figure 7. Operation waveforms at nominal input voltage [Vin=390Vdc, Po=0W (24V/0A)]
C4: Transformer Primary side current (2A/div), **C1**: Low side MOSFET current (2A/div)
C3: Low side MOSFET Vds (200V/div), time: 5us/div

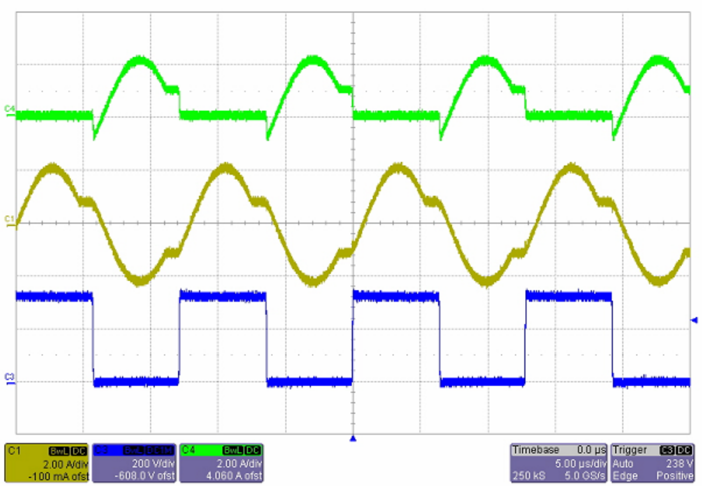


Figure 8. Operation waveforms at minimum input voltage [Vin=340Vdc, Po=192W (24V/8A)]
C4: Transformer Primary side current (2A/div), **C1**: Low side MOSFET current (2A/div)
C3: Low side MOSFET Vds (200V/div), time: 5us/div

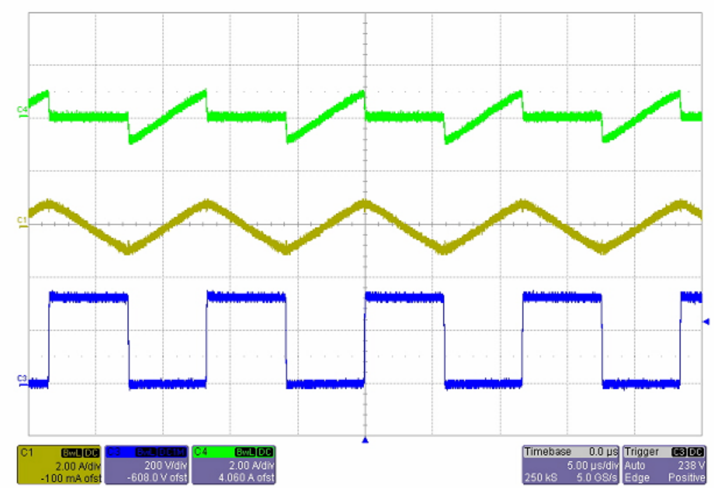


Figure 9. Operation waveforms at minimum input voltage [Vin=340Vdc, Po=0W (24V/0A)]
C4: Transformer Primary side current (2A/div), **C1**: Low side MOSFET current (2A/div)
C3: Low side MOSFET Vds (200V/div), time: 5us/div