Resonant Converters with Phase Shift Control

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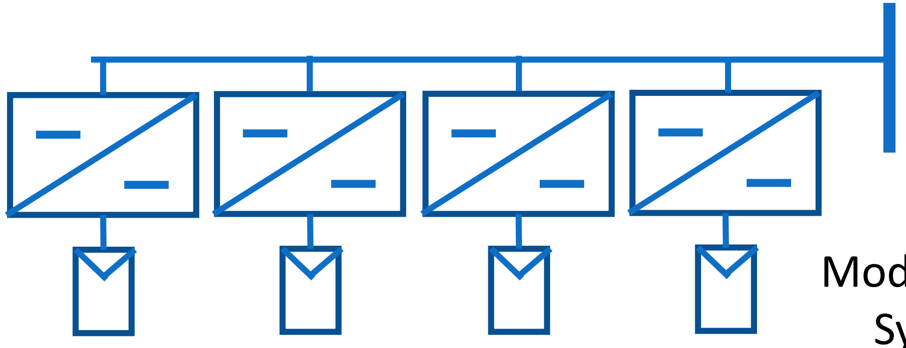
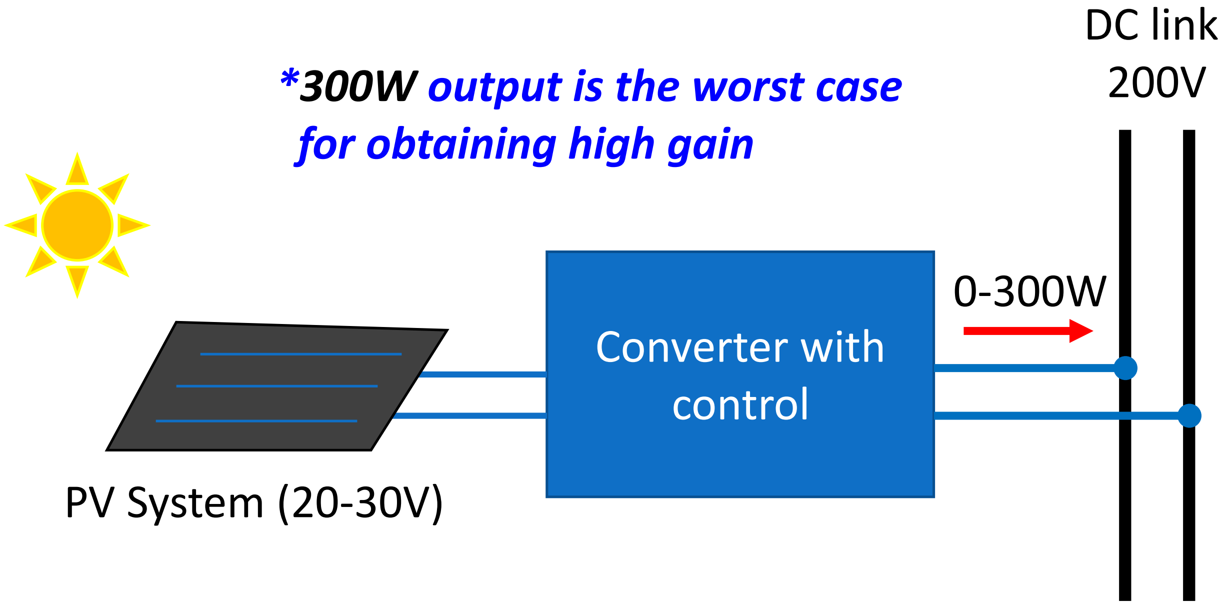
Back to the story…

- Build a (DC/DC, **boost**) converter
- Connecting small PV systems (**30V, 300W**) to DC link **200V**
- With **fluctuating input and load**
- **High efficiency**
- **Small** in size converter is needed



Modularized System

300W output is the worst case for obtaining high gain



Specification

- Input voltage: **20 - 30V** (20V depends on the design max gain of converter)
- Input power: **0 - 300W**
- Switching frequency: **100kHz**
  (at large as possible to reduce the size of components;
  not too large so that parasitic components and sensing error are still negligible;
  for PV small power (around 200-400W) applications, 100kHz – 350kHz is common
- Output Voltage: **200V (for American System)**

Resonant Converter

The design of resonant converter can be divided into **3 stages**.

Considerations:

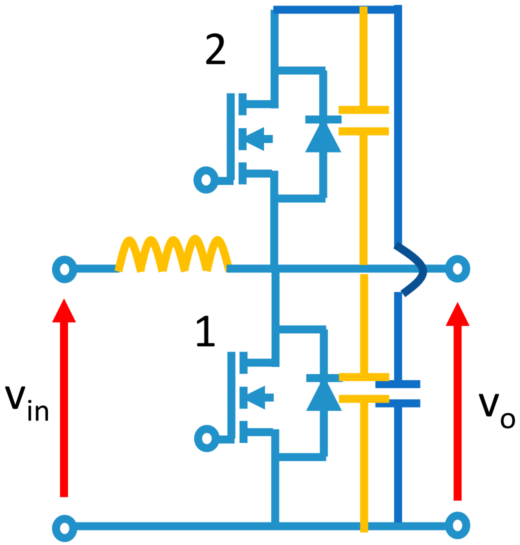
- **Gain** contributed in each process.
- **Losses and Efficiency** in each network.
  (Resonant Converter has soft switching property so there is no switching loss.)
- **Frequency range** and **size** of the converters

More to think of:

- overvoltage/ short circuit protection
- ZVDS (reaching dv/dt = 0)
- Avoid DC injection
- Multiple functions

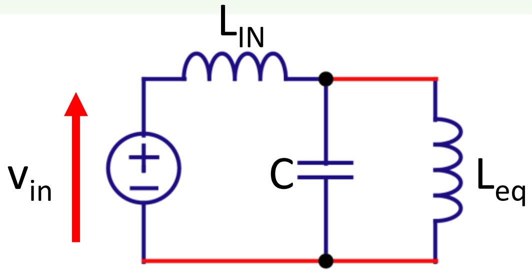
A. Switch Network

Half-Bridge Boost Topology

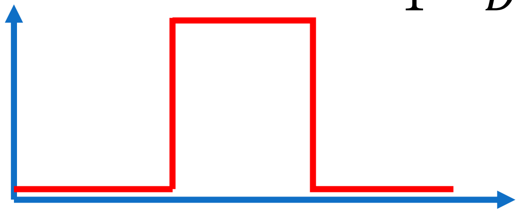


Q1 ON, Q2 OFF: $$V_{in} = V_L = L\frac{\Delta i}{DT}$$

Q1 OFF, Q2 ON: $$V_{in} + L\frac{\Delta i}{(1-D)T} = V_o$$



$$v_o = \frac{1}{1-D}V_{in}$$

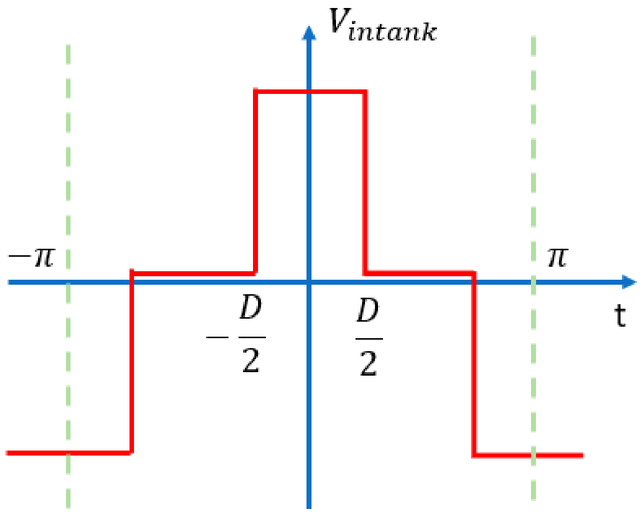


For Full-bridge Phase shift:

$$v(t) = \begin{cases} \frac{1}{1-D} & 0 < t < \frac{D}{2} \\ -\frac{1}{1-D} & \pi - \frac{D}{2} < t < \pi \\ 0 & \text{otherwise} \end{cases}$$

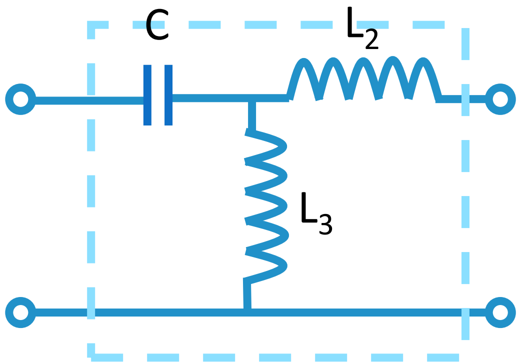
$$v_i = \sum_{n=0}^{\infty} \frac{2\sin(2i-1)\pi D}{\pi(1-D)(2i-1)} \cos it$$

$$\boxed{|v_1| = \frac{2\sin \pi D}{\pi(1-D)}}$$



B. Resonant Tank Design

To minimize component size, an LLC (T circuit) is selected.
The followings are satisfied to obtain desired topology.



$$\omega_s = \sqrt{\frac{1}{L_2 C} + \frac{1}{L_3 C}} = 2\pi f_s$$

$$G = \left|\frac{U_{OUT}}{U_{IN}}\right| = \left|\frac{1}{1 - \frac{1}{\omega_s^2 C L_3}}\right|$$

2 equations 3 unknowns !!

The last equations is rather flexible, can be load factor (for high harmonics attenuation), phase control (for harmonics accumulation), or bandwidth.

$$\frac{U_{IN}}{U_O} = \left(1 - \frac{1}{\omega^2 C L_3}\right) + j\frac{\sqrt{\frac{L_2}{L_3}}}{R_L}\left(\frac{\omega_n^2}{\omega^2} - 1\right)$$

For a 100kHz tank with gain 1.5,
assume a capacitance with C = 150nF,
L_2 = 25uH
L_3 = 50uH

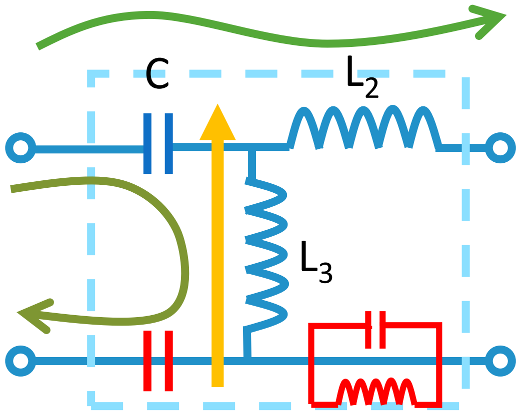
** Make sure the resonant frequency is on the **left** of the switching frequency.*

B. Implications on the Equations

Requirement for load independence

$$\Delta = \frac{Z_1Z_2 + Z_2Z_3 + Z_1Z_3}{Z_3} = 0$$

→ Provide a switching frequency with Load independent tank
(other modulus is load independent?)
→ Need 1 cap to 2 inductors, or 2 cap to 1 inductors
→ Z can have **more components**
→ Z_1 can be distributed to **lower arm**



CLL T-Tank Design Equation

$$\omega_r = \frac{1}{CL_2}\sqrt{1+\frac{L_2}{L_3}} = 2\pi f_s$$

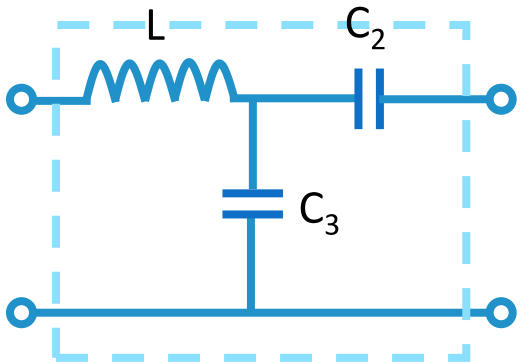
$$G = \left|\frac{U_{OUT}}{U_{IN}}\right| = \left|\frac{1}{1-\frac{1}{\omega_r^2 CL_3}}\right|$$

→ fs relies more on $\mathbf{L_2}$, yet tuning L_2 doesn't react much
G relies more on $\mathbf{L_3}$
→ Having too small L_3, we have a high gain, yet the circulating current will have **higher loss**
→ Let $L_3 = kL_2$

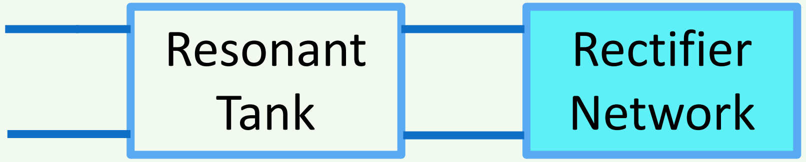
$$G = \frac{(1+k)L_3}{(1+k)L_3 - C}$$

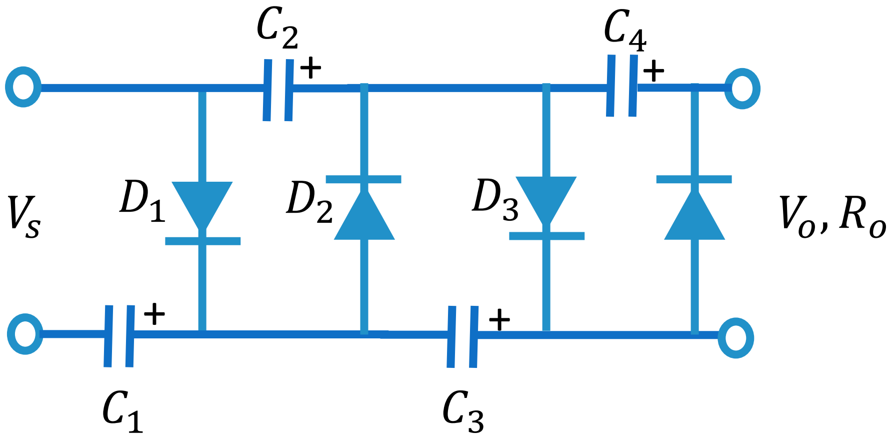
efficiency <50%

If **k > 1,** with C ~ 10^{-7}, L~10^{-5}, $\boldsymbol{G \approx 1}$
→ In reality, L_2 must have an inherited **resistance**



C. Cockcroft Walton Multiplier



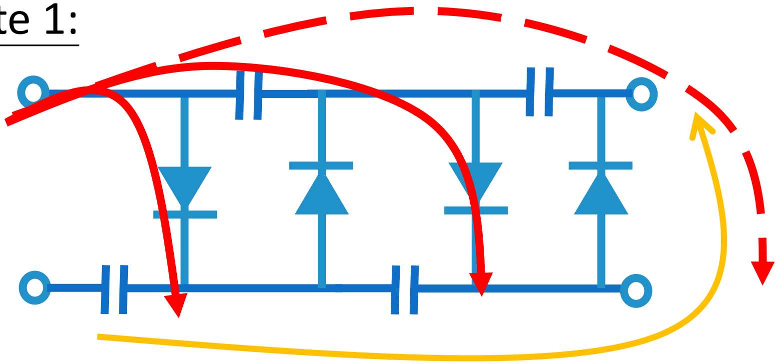


$$V_s = V_{smax} \sin(2\pi f_s t)$$

With first harmonic approximation (FHA) and without diode voltage drop,

$$\boxed{\frac{V_{smax}^2}{2} \frac{1}{R_{eq}} = \frac{V_o^2}{R_o}}$$

State 1:



When V_s reaches V_{smax}

Diode D_1 D_3 … D_{2n-1} conducts

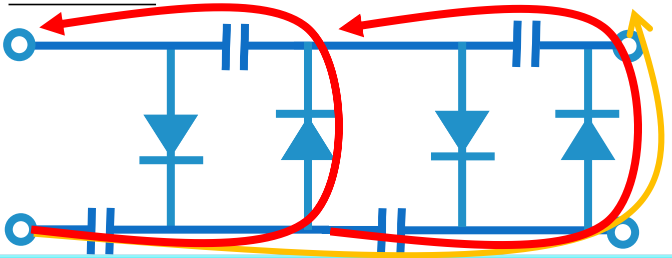
C_1 C_3 … C_{2n-1} are charged.

C_2 C_4 … C_{2n} are discharged.

$$V_{smax} = V_{c1}$$
$$V_{smax} + V_{c2} = V_{c1} + V_{c3}$$
$$V_{smax} + \sum_{k=1}^{(n-1)/2} V_{c2k} = V_{c1} \sum_{k=1}^{(n-1)/2} V_{c2k+1}$$
$$V_{smax} + \sum_{k=1}^{(n-1)/2} V_{c2k} = V_o$$

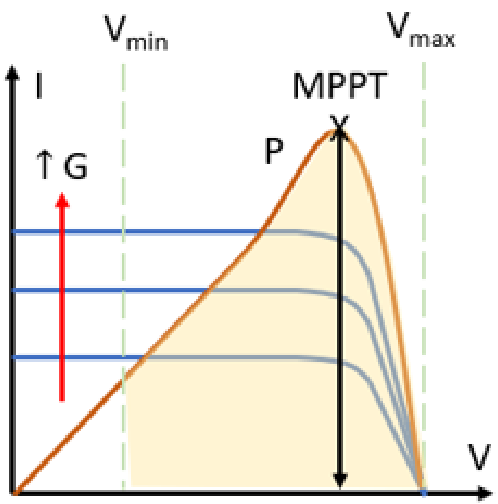
(opposite)

State 2:

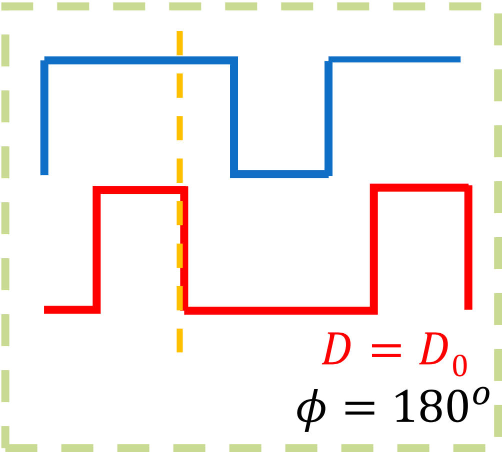


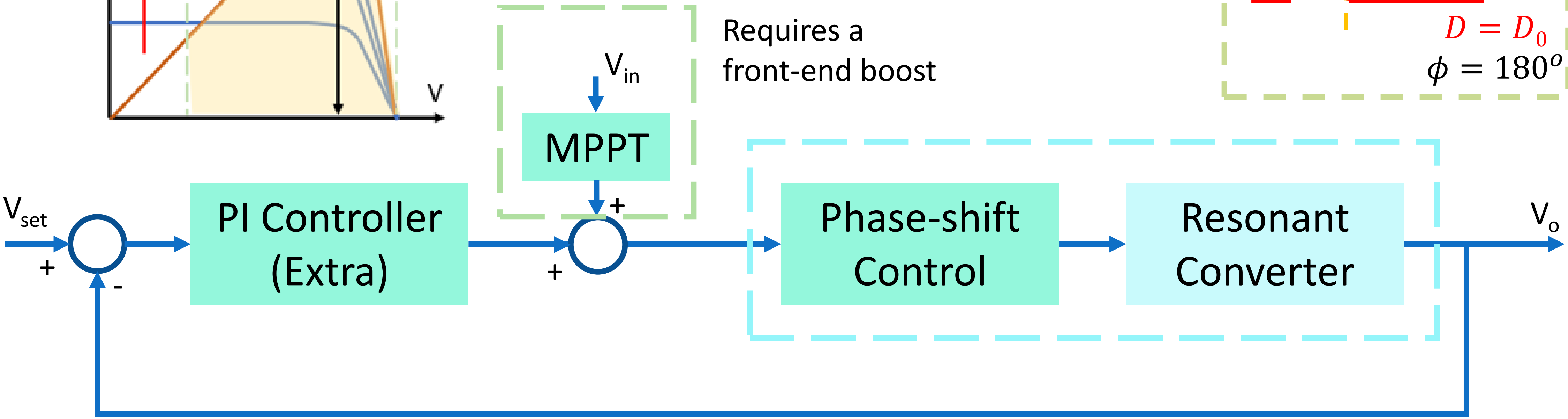
$$V_{smax} + V_{c1} = V_{c2}$$
$$V_{smax} + V_{c1} + V_{c3} = V_{c2} + V_{c4}$$
$$V_{smax} + \sum_{k=1}^{(n-1)/2} V_{c2k+1} = V_{c1} \sum_{k=1}^{(n-1)/2} V_{c2k}$$
$$V_{c1} + \sum_{k=1}^{(n-1)/2} V_{c2k+1} = V_0$$

D. Control Loops implemented



- More stable to control voltage gain
- Not affect ZVS/ZCS





Overall Topology

$$G_{sw} = \frac{2\sin\pi D}{\pi(1-D)}$$

- Variable Gain
- Ability to boost

$$G_{tank} = \left|\frac{1}{1-\frac{1}{\omega_r^2 C L_3}}\right| = 1.5$$

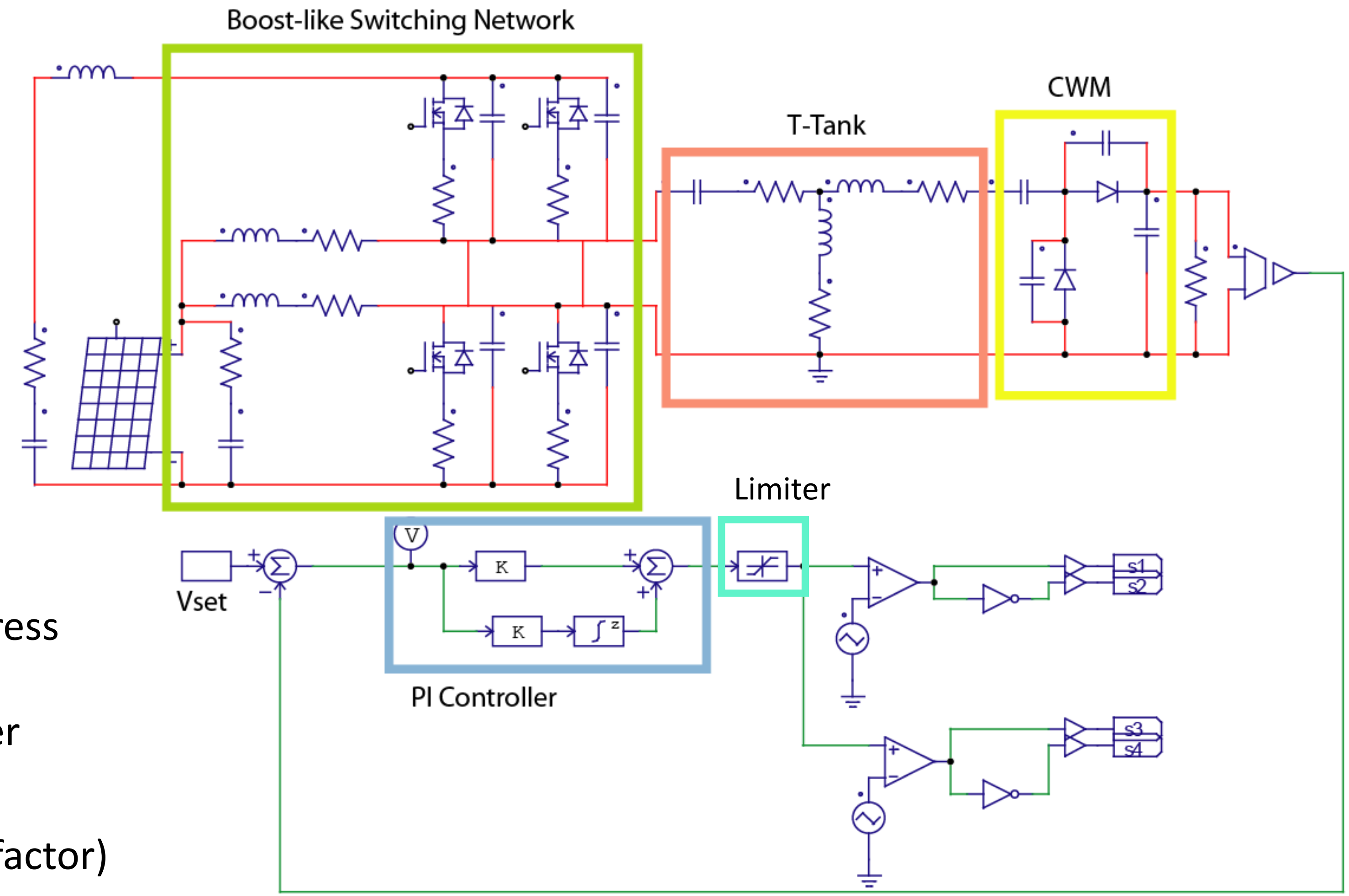
- Load independent Gain
- Soft Switching

$$G_{CWM} = \frac{\pi}{4}N$$

- Simple Structure
- Reduced Voltage Stress
- With Gain
- Without Transformer

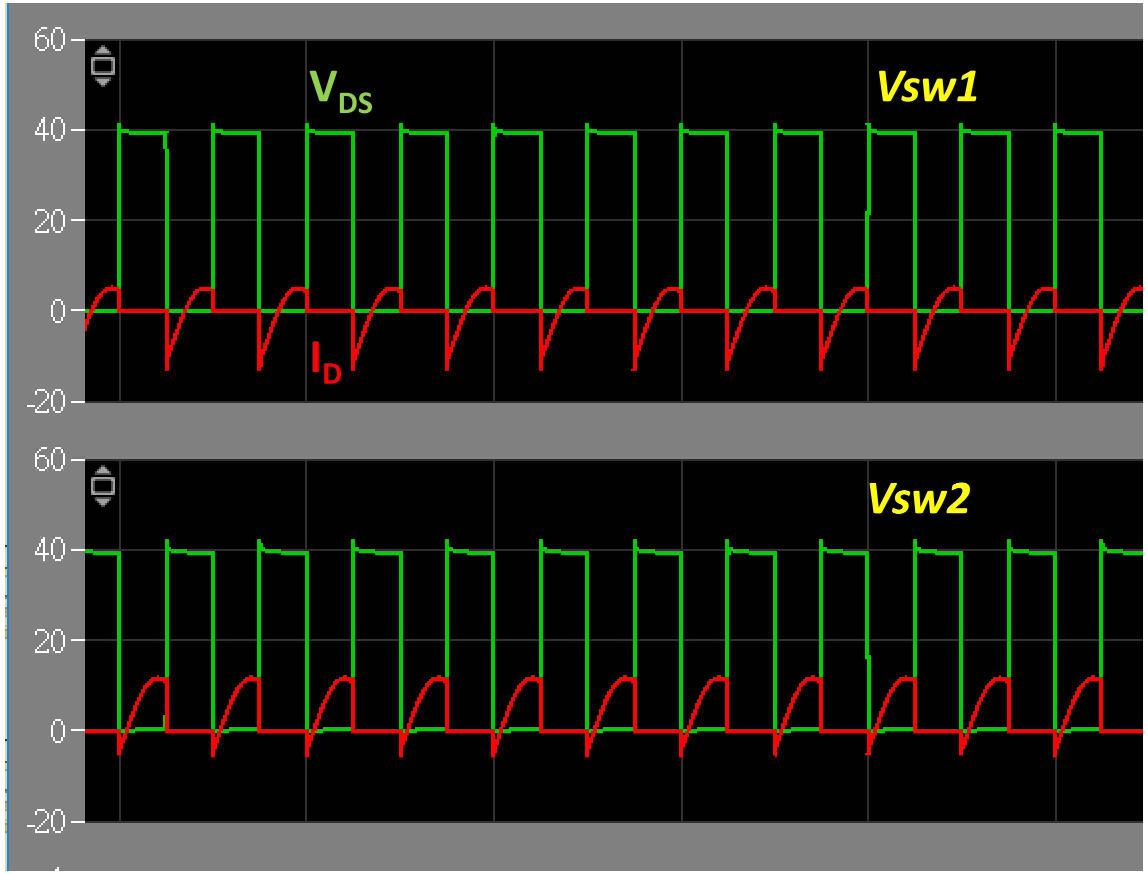
$$G_{overall} = \frac{3\varepsilon\sin\pi D}{2(1-D)}$$

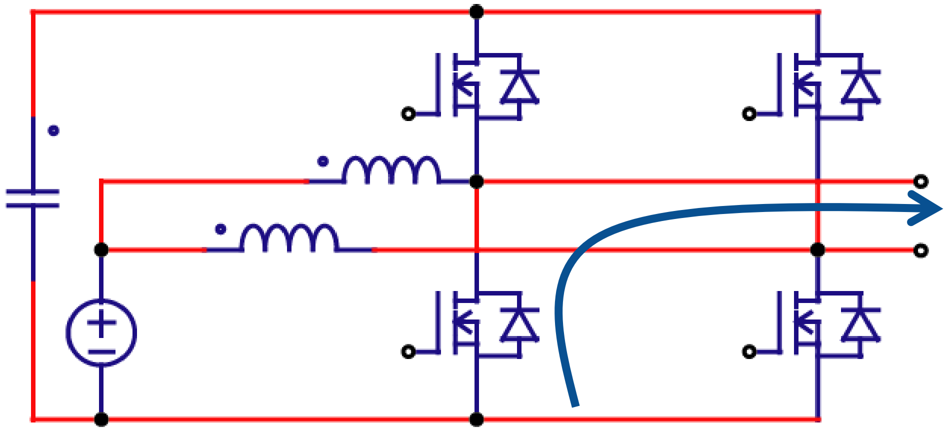
$\varepsilon \sim 0.9$ (correction factor)



Simulation (Expected Results)

Hard Switching / Soft Switching

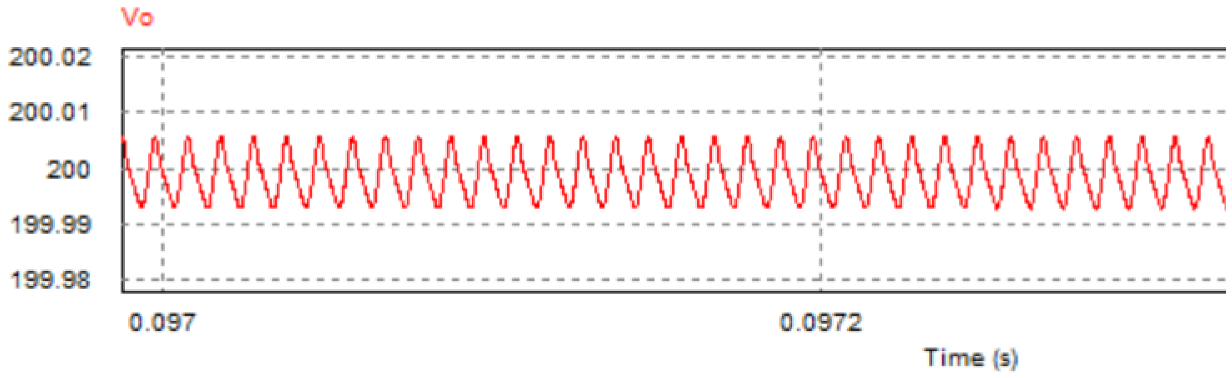


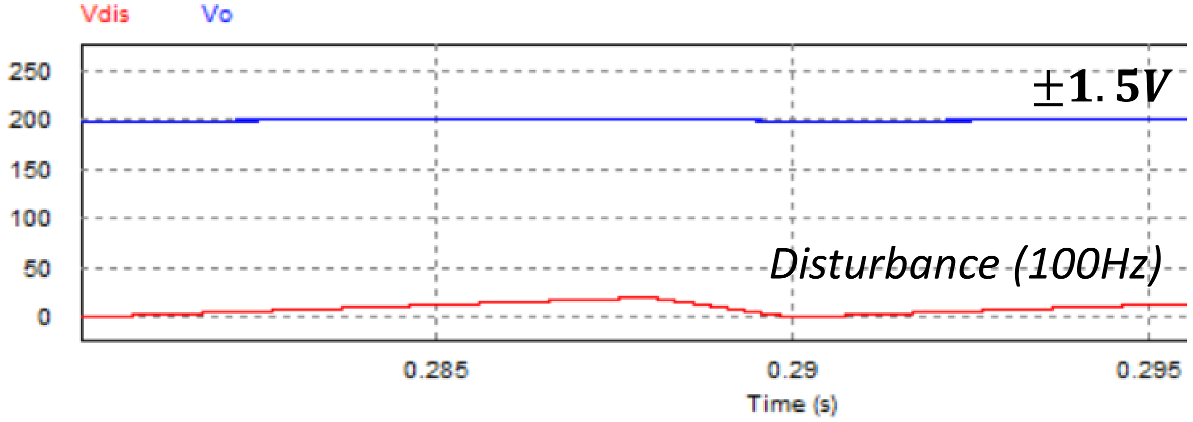


- V_{DS} falls when turn ON. At that time, diode conducts Hence the overlapping region (switching loss) becomes body **diode conduction loss**.
- In practice, **only V_{DS} and V_{GS} are tested.** $V_{GS} > V_{GS(th)}$, current starts the rise V_{cap2} swings above V_G and V_{DS} becomes negative before current passes through the switch

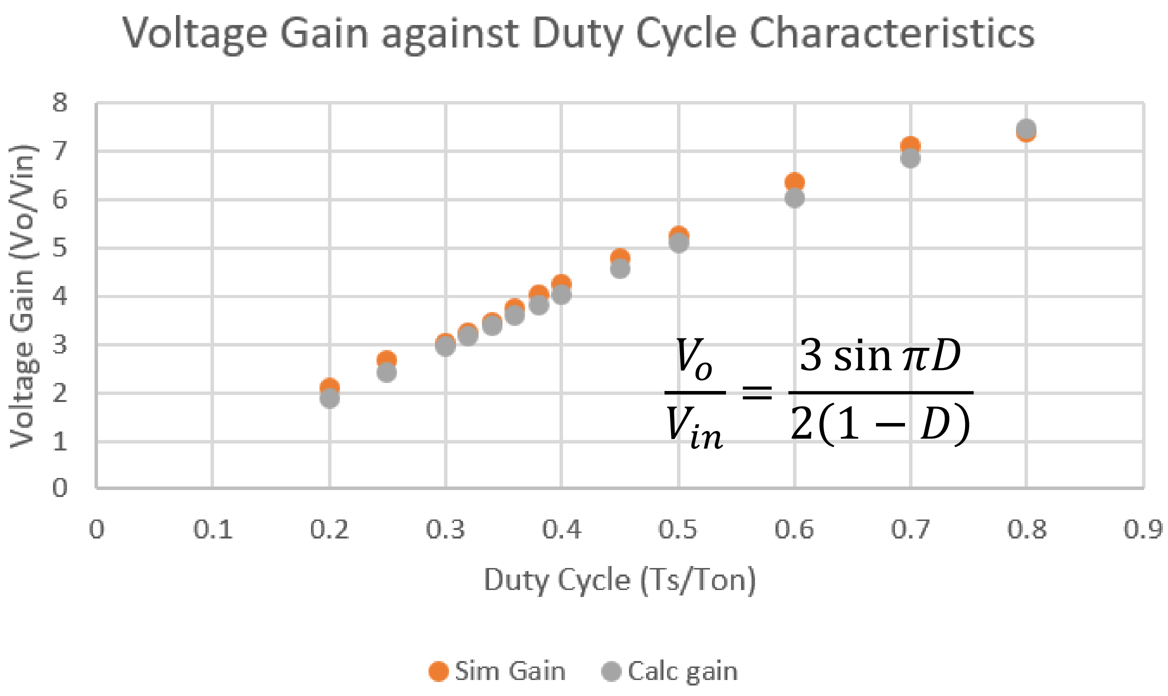
Simulation (Expected Results)

Output Voltage with Disturbance



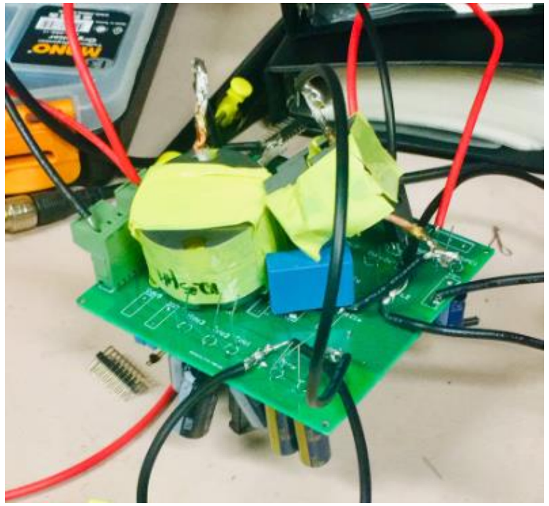


Voltage Gain against Duty Cycle

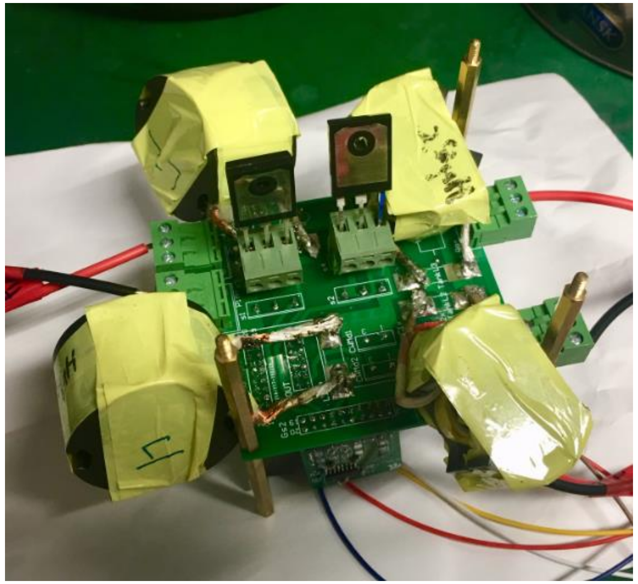


Hardware Implementation

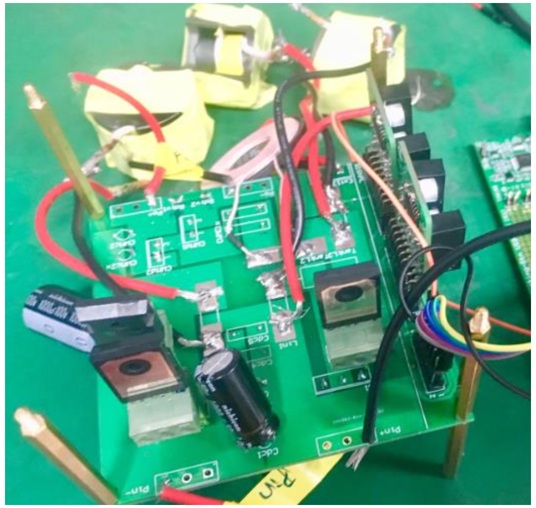
Version 1



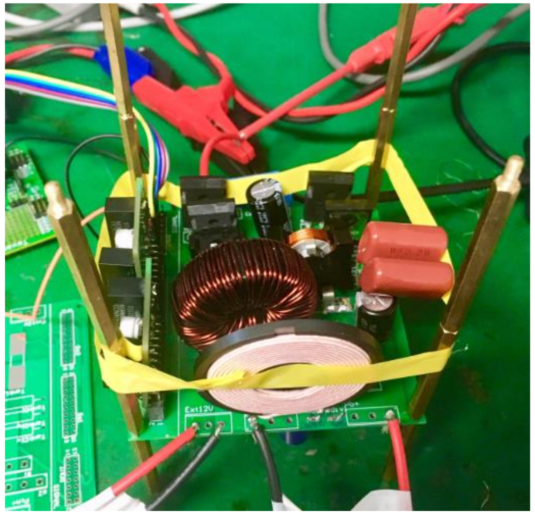
Version 2

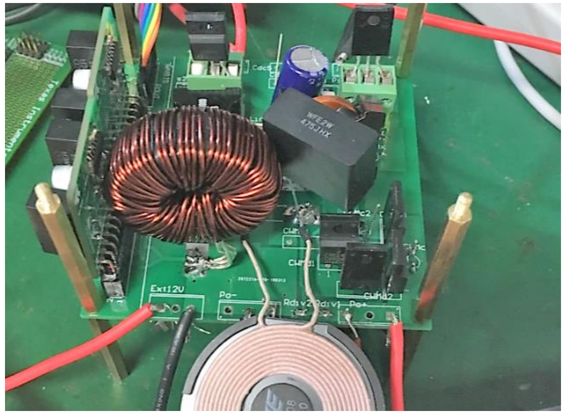


Version 3



Version 4





Version 5

Discussion:
- Components Selection
- Output Waveform
- Problem Solving

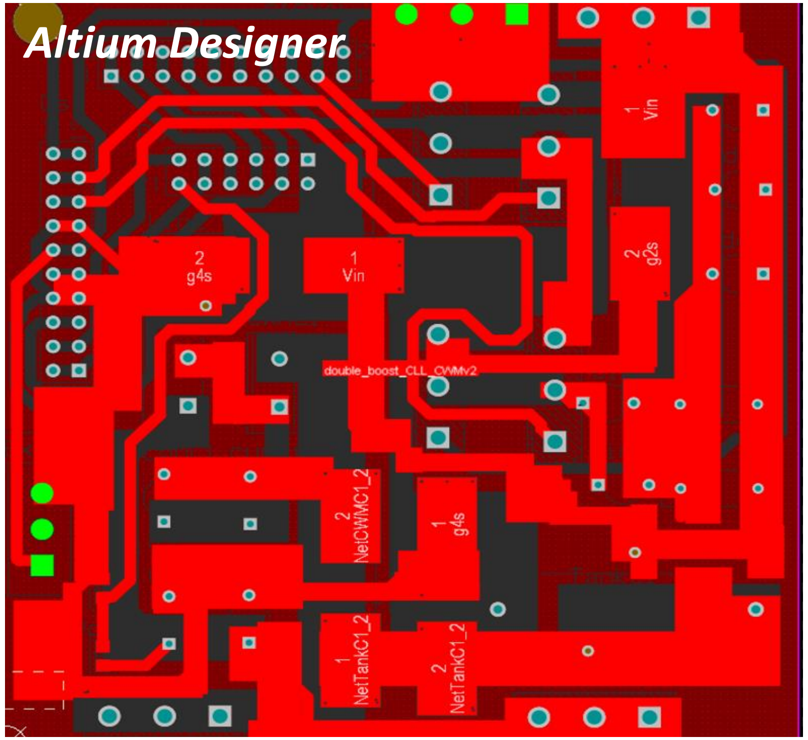
Hardware Implementation

Components Selection

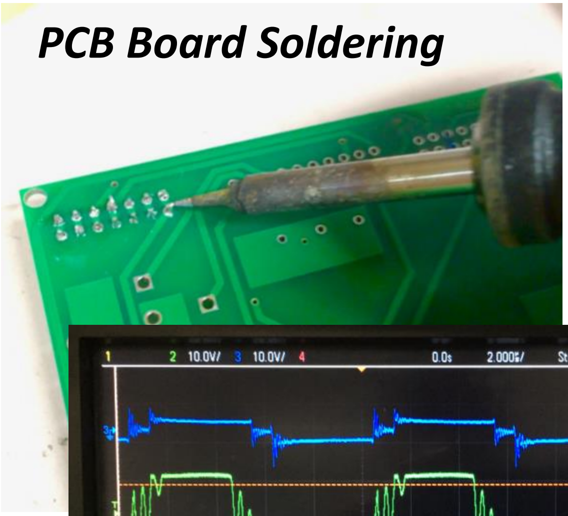
說明	IRF520NPBF N-Channel MOSFET, 9.7 A, 100 V HEXFET, 3-Pin TO-220AB Infineon	IRFP9140NPBF P-Channel MOSFET, 23 A, 100 V HEXFET, 3-Pin TO-247AC Infineon	IRF9540PBF P-Channel MOSFET, 19 A, 100 V, 3-Pin TO-220AB Vishay
	541-1180	542-9816	708-5152
Brand	Infineon	Infineon	Vishay
Manufacturer Part No	IRF520NPBF	IRFP9140NPBF	IRF9540PBF
技術 資訊			
Typical Gate Charge @ Vgs	25 nC @ 10 V	97 nC @ 10 V	61 nC @ 10 V
Category	Power MOSFET	Power MOSFET	Power MOSFET
Maximum Gate Source Voltage	-20 V, +20 V	-20 V, +20 V	-20 V, +20 V
Transistor Material	Si	Si	Si
Number of Elements per Chip	1	1	1
Maximum Drain Source Voltage	100 V	100 V	100 V
Channel Mode	Enhancement	Enhancement	Enhancement
Transistor Configuration	Single	Single	Single
Typical Turn-On Delay Time	4.5 ns	15 ns	16 ns
Channel Type	N	P	P

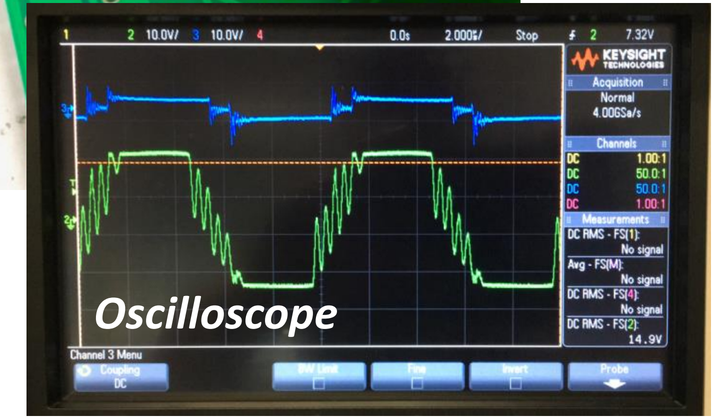
RS – Online Mouser

PCB Sketching



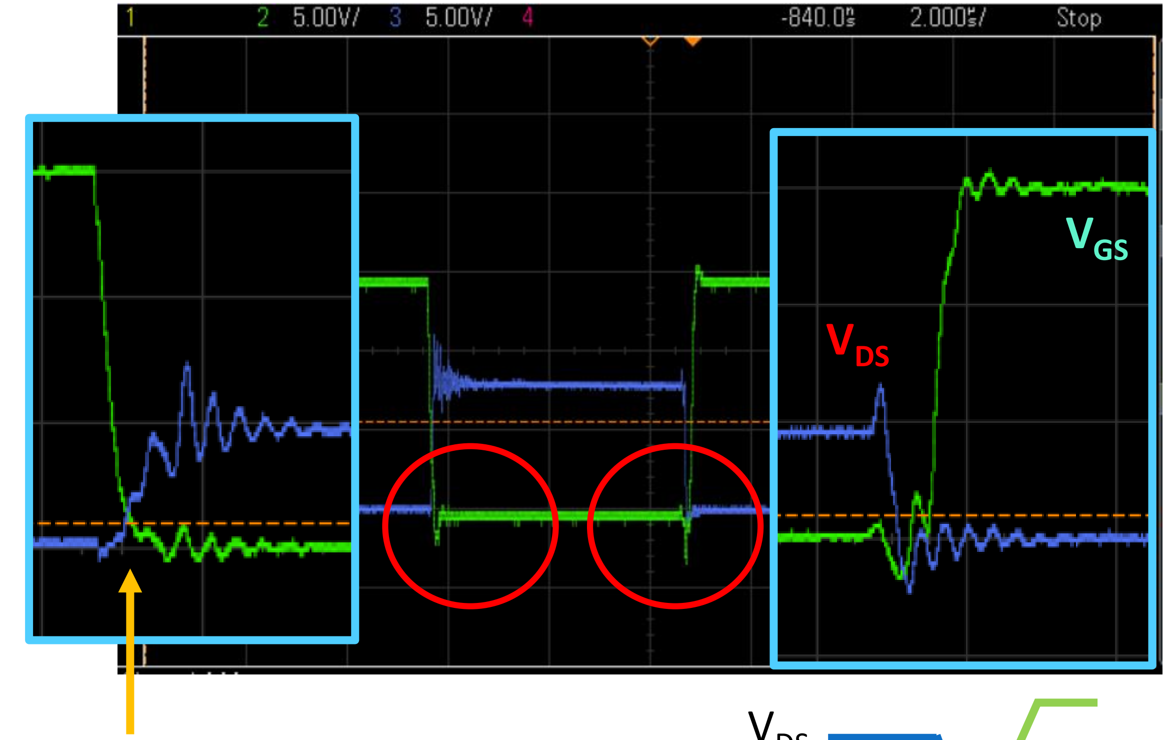
Solder and Test

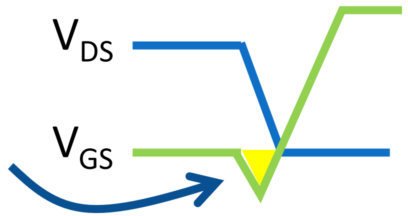




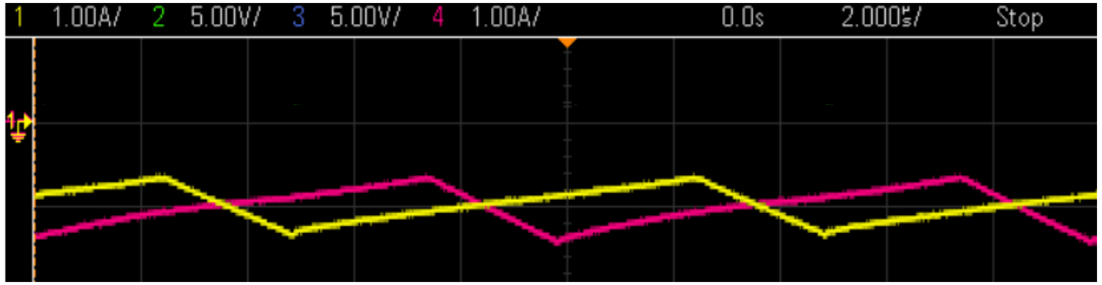
Results

Zero Voltage Switching (Turn ON)





Input Current Waveform

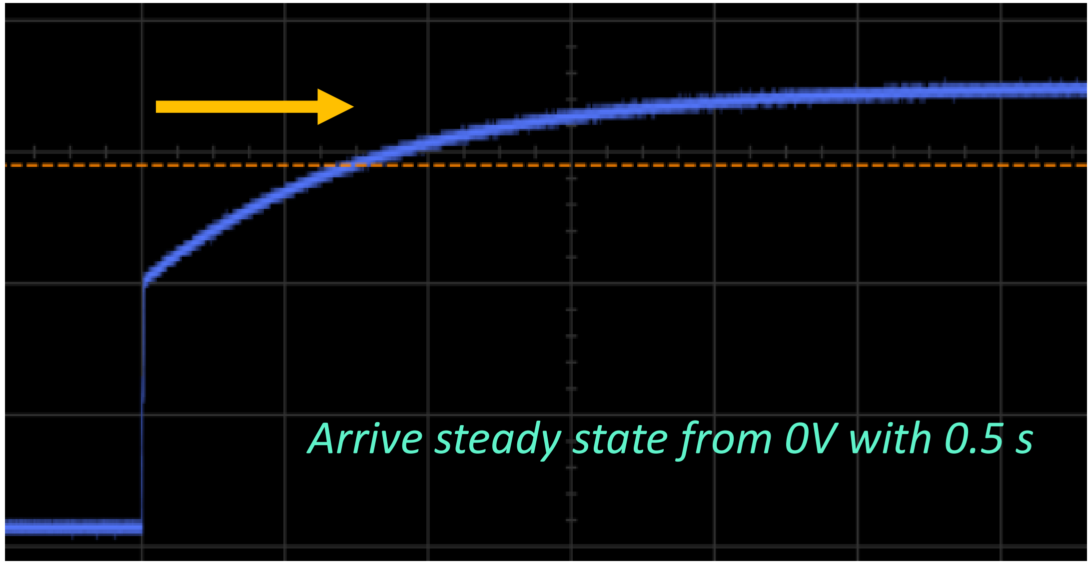


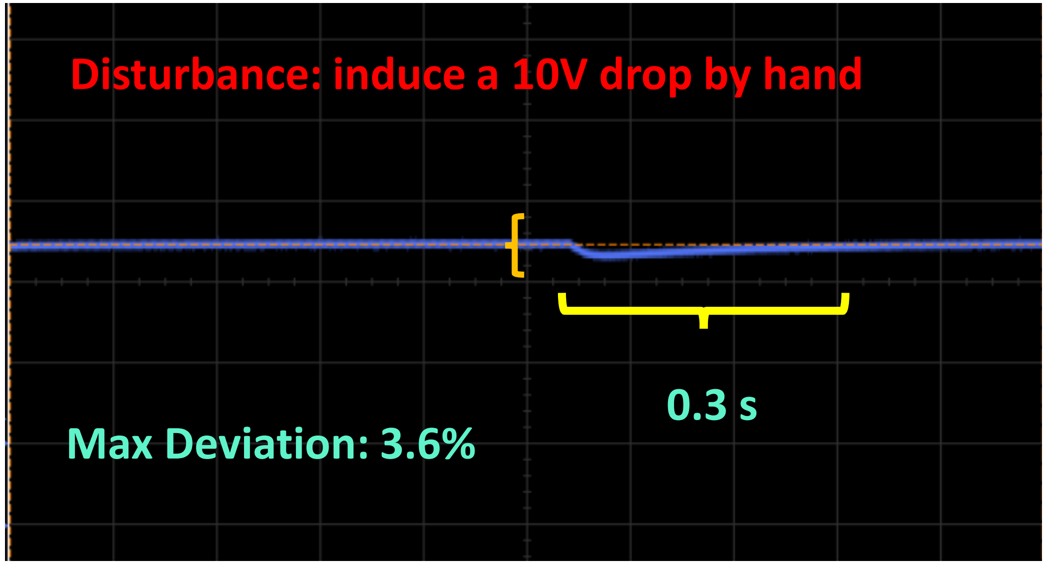
Tank Input and Output voltage

Results

- Factor of speed variation: PI controller
  Limiter
  Output Capacitor

<u>Output Voltage under Disturbance</u>





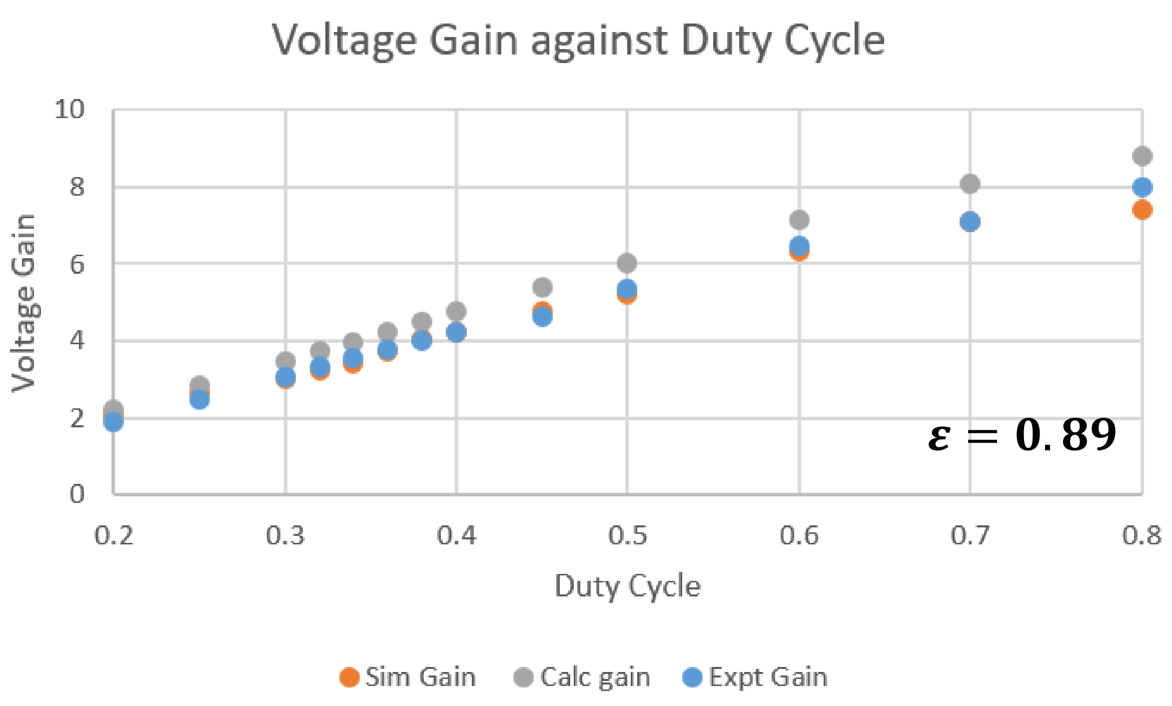
Is 0.3s and 0.5s too long, or is 3.6% voltage drop too large?

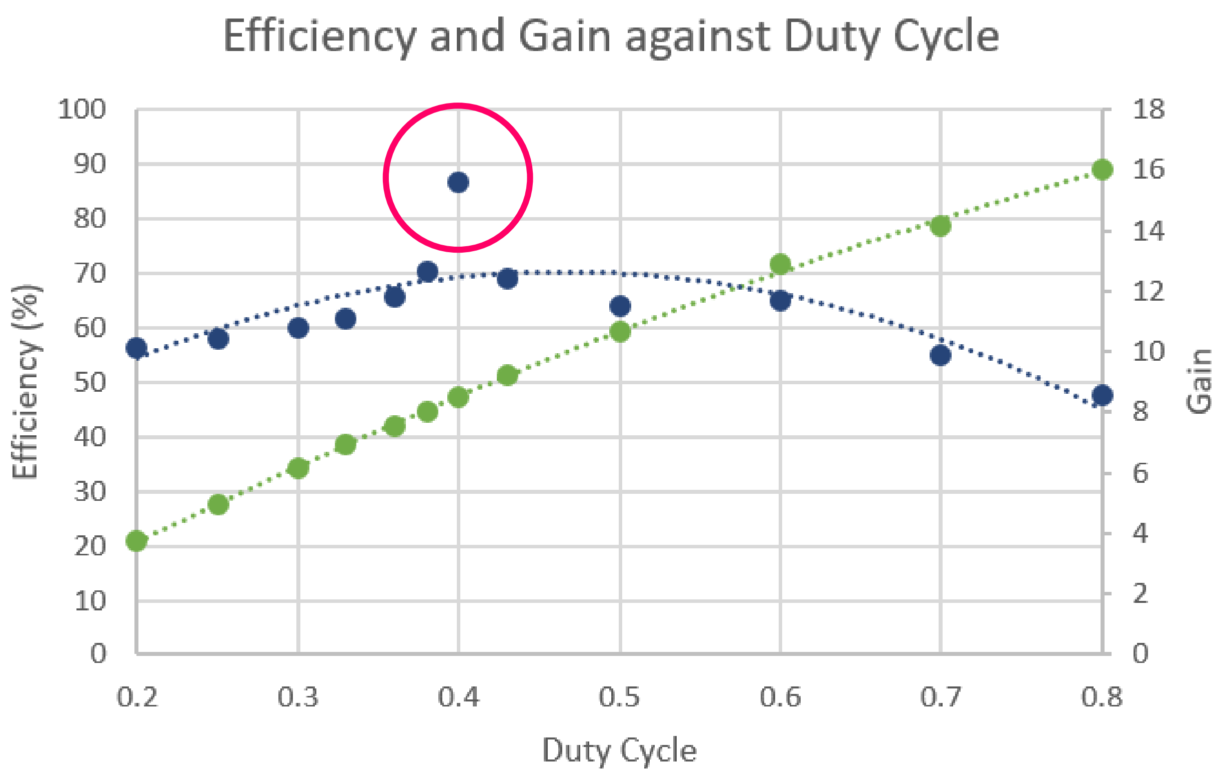
- Allowable voltage variation: 6%
- Fault Automatic Restoration: 0.2s

Result -- Voltage Gain

Note: The simulation has included loss.

Characteristics of Gain, Efficiency with Duty Cycle





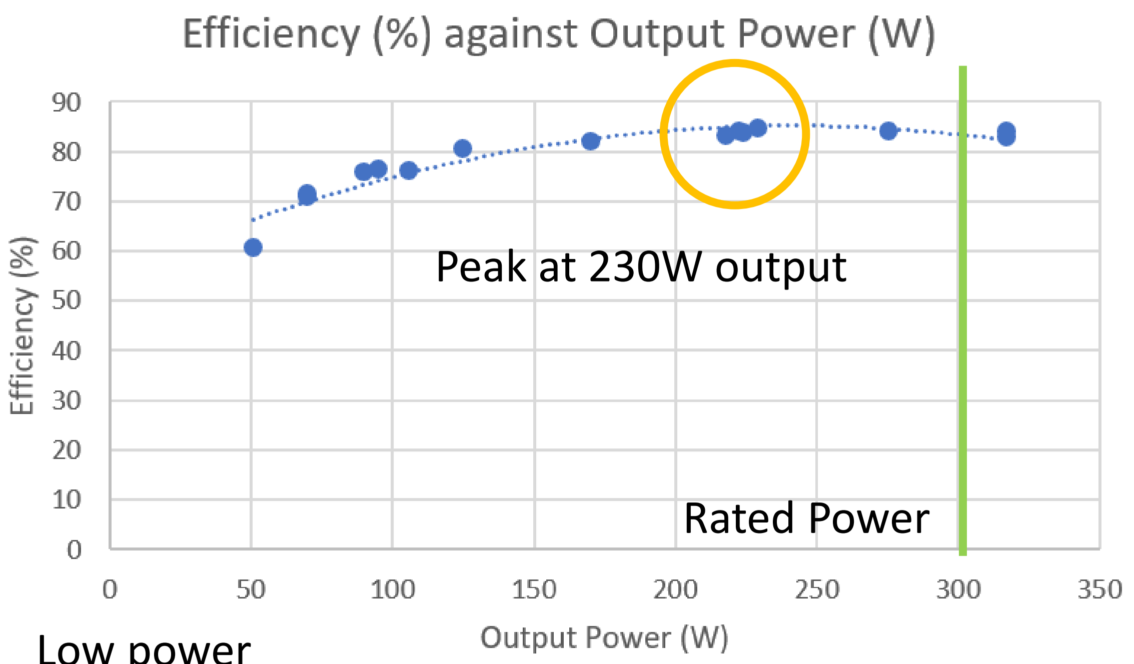
- Voltage drop from resistance, forward voltage
- With First Harmonic Approximation

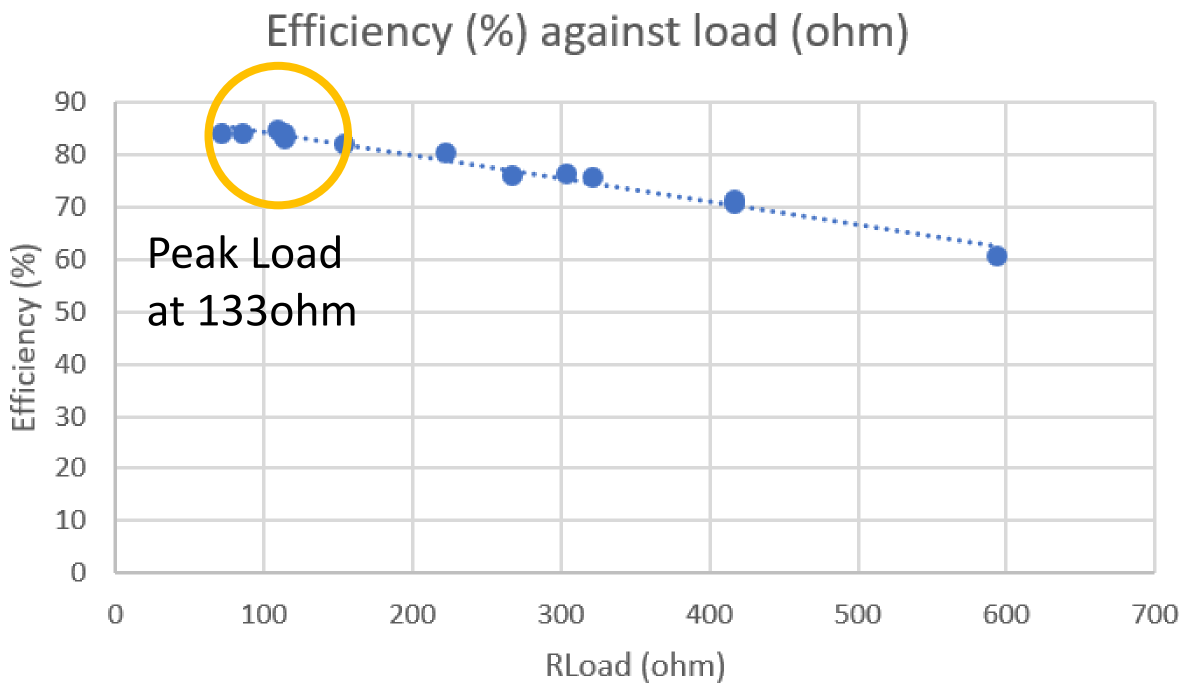
*A surge of efficiency in D= 0.4

Result -- Efficiency

The efficiency of the converter reaches ***84.6%*** at ***P_O = 230W***. → **How to improve Efficiency? (5)**

The following graphs are based on constant V_{IN} = 30V, closed-loop, but varying load.





Low power
-> circulating current
-> efficiency ↓

High output current
-> diode conduction loss ↑

Way Forward…

- Implementation with MPPT, Hysteresis Control, Sliding Mode Control
- Multi-resonant (Harmonic Boost, Short Circuit Protection)
- Coupled-Inductor/ Cross Transformer in phases
- Small-Signal Model
- Grid Connection Control Architecture
- Under Varying Load (*Constant Power Load)
- Hardware Separation (Fault Resilience)
- Multiport-Multilevel-Modular (3M) Converter
- DC Earthing Configuration with Converter

Conclusion

We have designed a new resonant converter,
which decouples the load and source,
provides voltage regulation, and gives good efficiency.

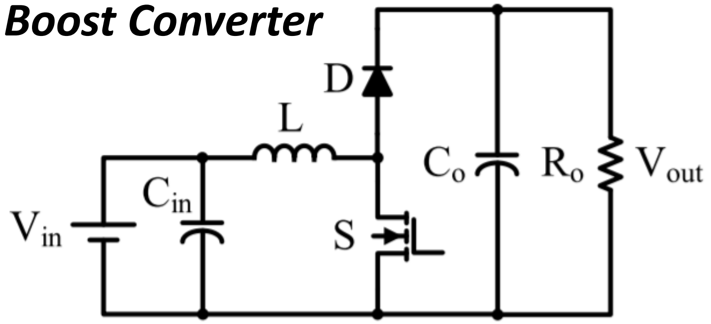
I appreciate the help and support from my supervisor Prof. S.C. Tan & Mr. Kerui Li.

Appendix

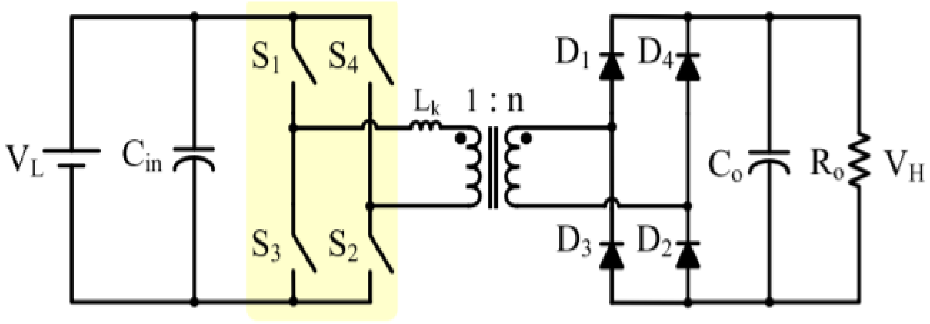
- Possible Designs
- Traditional Tank Design and Analysis, Soft Switching
- Current Flow Diagram
- Difficulties and Failure
- Discussion: Gate Driver, Components Selection, Ringing Effect
  Circulating Current, Loss Analysis, Grounding
  Control Scheme, PV Simulation
- Appendix: Benchmarking of Converters, DC Link, PV Modelling

What are the possible converters to be used?

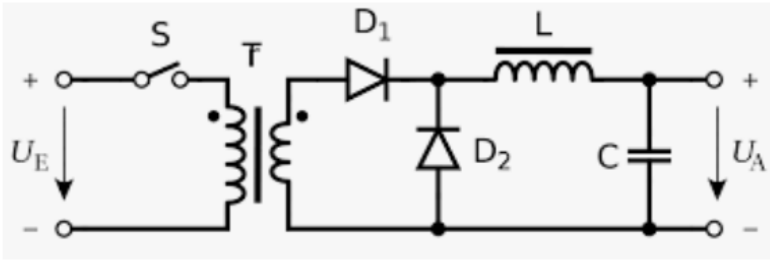
Boost Converter



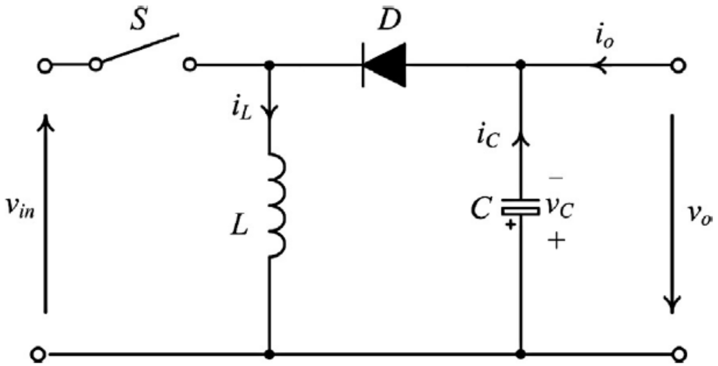
DC/AC, AC/AC, AC/DC (Mainly Step up by Tx)



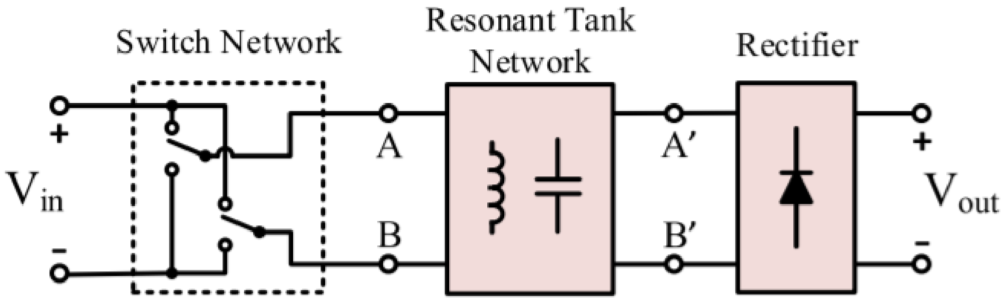
Forward Converter



Buck Boost Converter



Resonant Converter

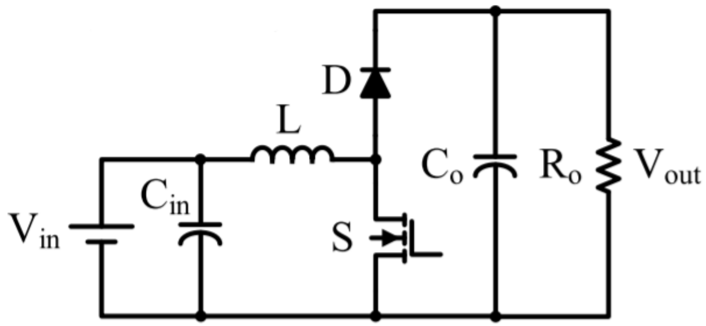


Consideration:

- **Uni-direction**/ Bi-direction
- **High Gain**
- **Less losses** (**switching**/ conduction)
- **Isolation** (Not Preferred)
- Size
- Voltage/**Current Fed**

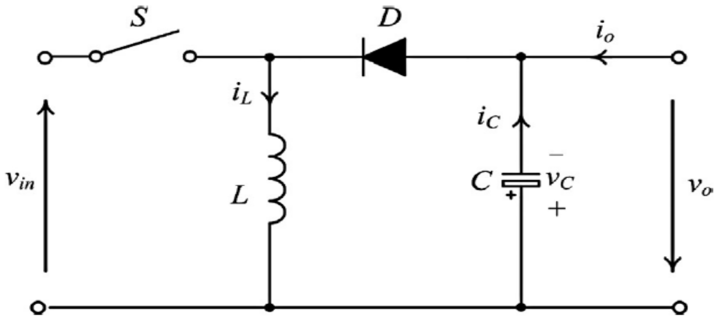
Comparison of Converter Topologies

Boost Converter



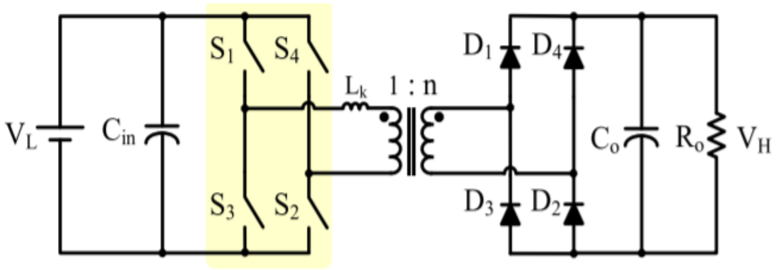
Requirement	
Uni-directional	✓
Without Isolation	✓
Able to Achieve High Gain	x 5
High Efficiency	
No Reversed Voltage Output	✓
Power can be cut off	

Buck-Boost Converter



Requirement	
Uni-directional	✓
Without Isolation	✓
Able to Achieve High Gain	x 5
High Efficiency	
No Reversed Voltage Output	
Power can be cut off	✓

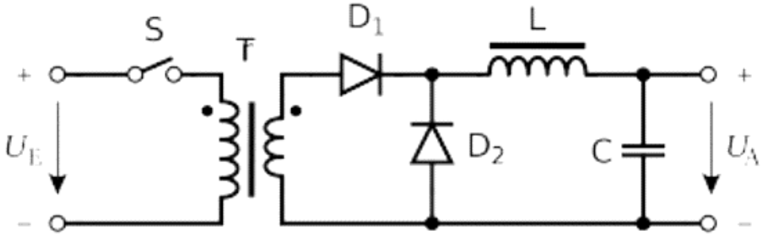
Transformer-based



Requirement	
Uni-directional	✓
Without Isolation	
Able to Achieve High Gain	x n
High Efficiency	
No Reversed Voltage Output	✓
Power can be cut off	✓

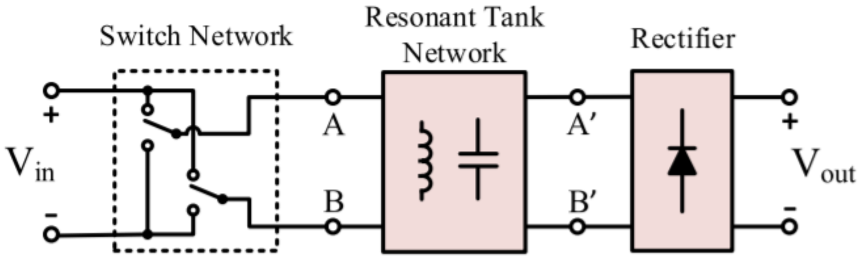
Comparison of Converter Topologies

Forward Converter



Requirement	
Uni-directional	✓
Without Isolation	
Able to Achieve High Gain	x n
High Efficiency	
No Reversed Voltage Output	✓
Power can be cut off	✓

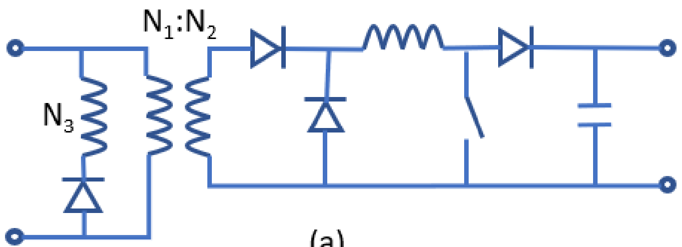
Resonant Converter



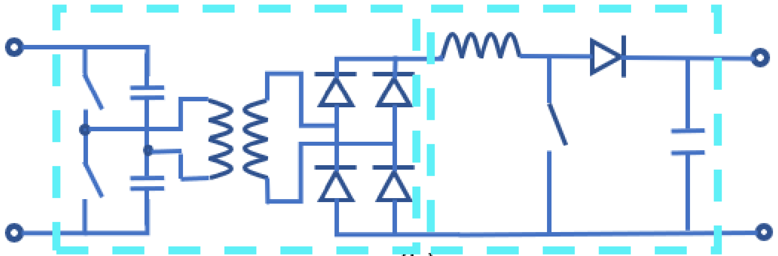
Requirement	
Uni-directional	?
Without Isolation	?
Able to Achieve High Gain	✓
High Efficiency	✓
No Reversed Voltage Output	✓
Power can be cut off	✓

Hossain M.Z. et al. (2018), Recent Progress & Development on Power DC-DC Converter Topology, Control, Design & Application: Review

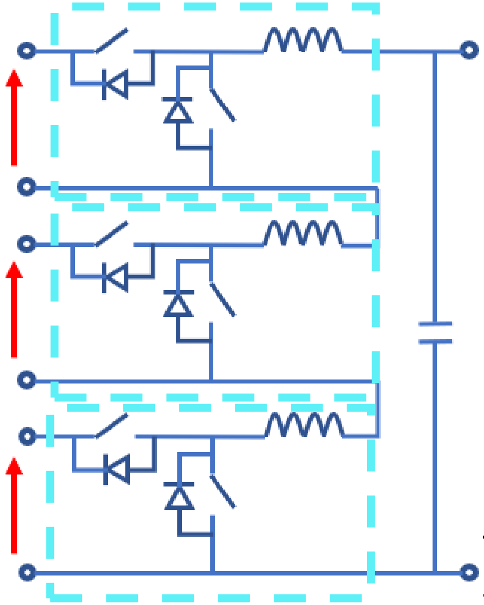
Other Possible Converter Topologies



(a)

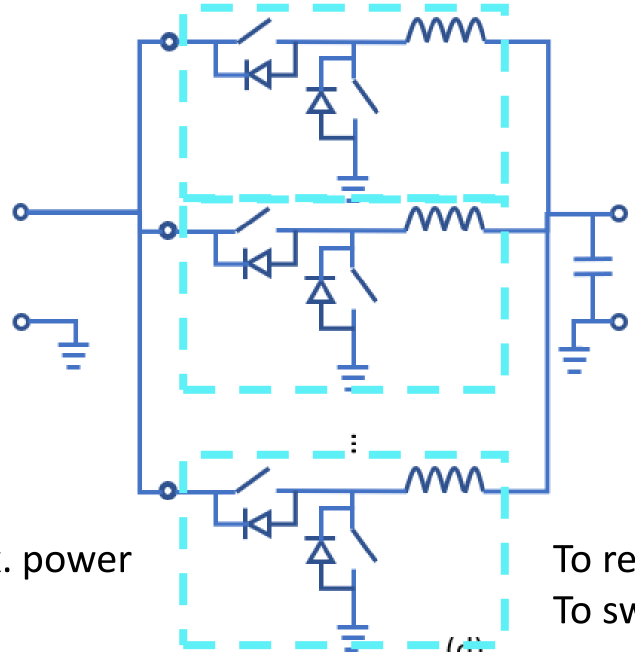


(b)



(c)

To harvest max. power from PV



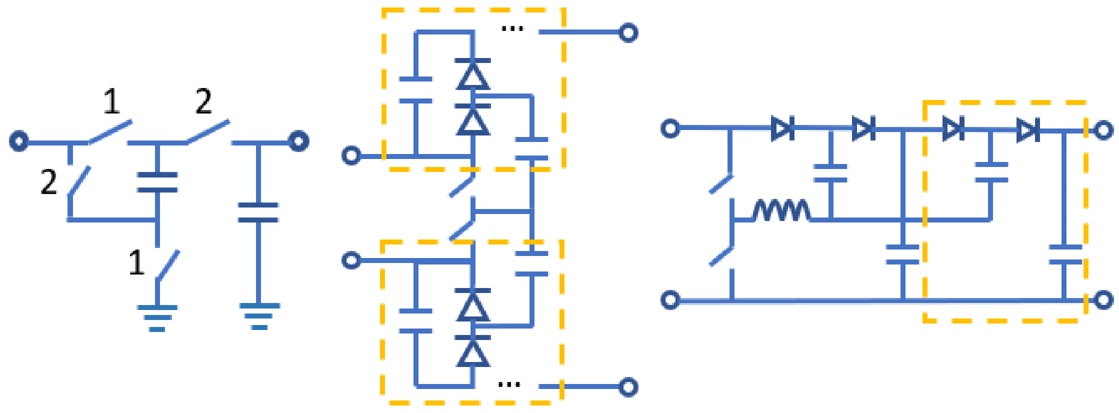
(d)

To reduce current stress at branches
To switch in different phases

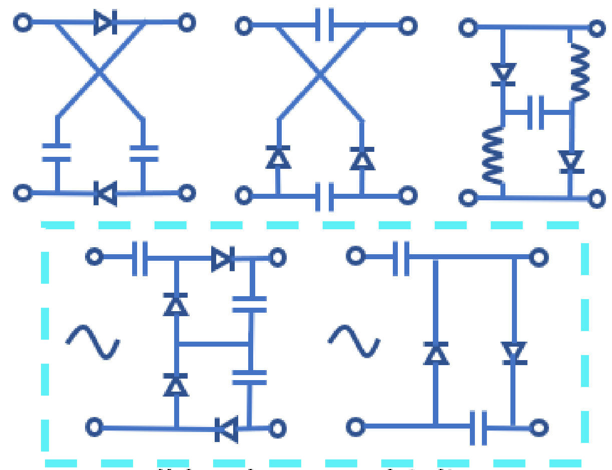
Based on ***cascading/ parallel*** different topologies

Forouzesh M. et al. (2017), Step-Up DC-DC Converters: A Comprehensive Review of Voltage Boosting Techniques, Topologies, and Applications

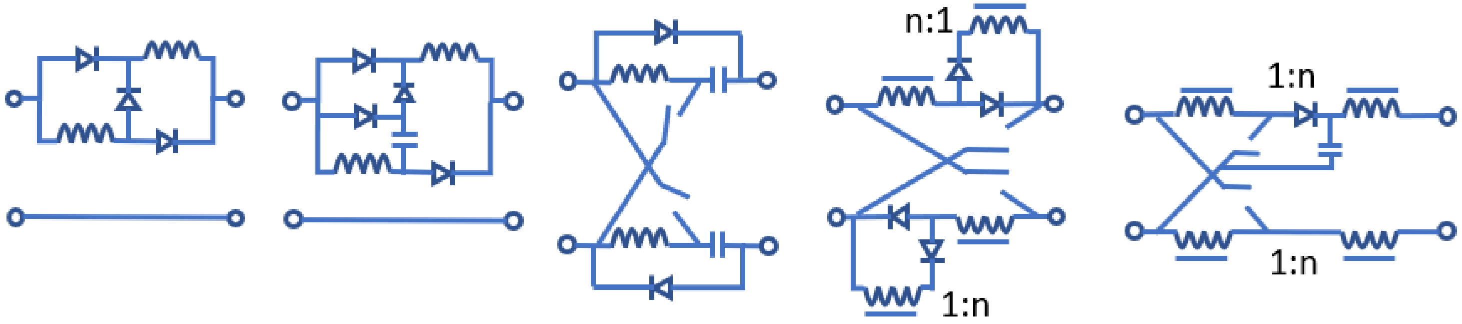
Possible Modular Topologies for Higher Gain



(a) Charge Pump (Switched Capacitor)



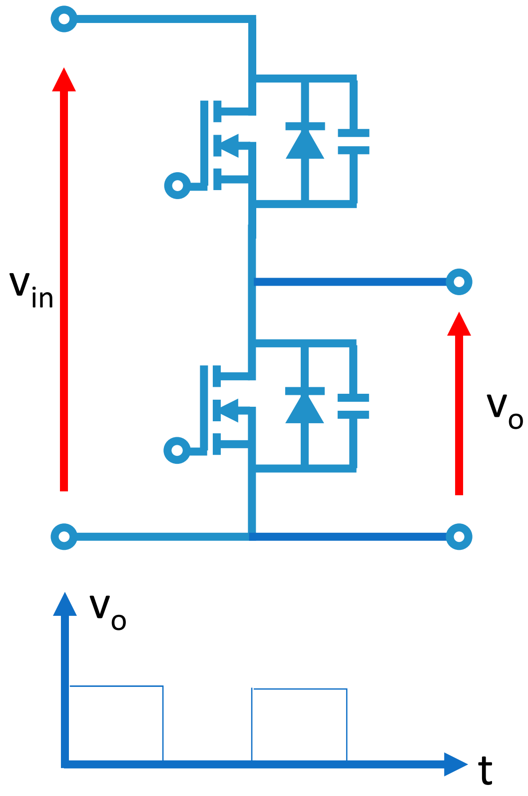
(b) Voltage Multiplier



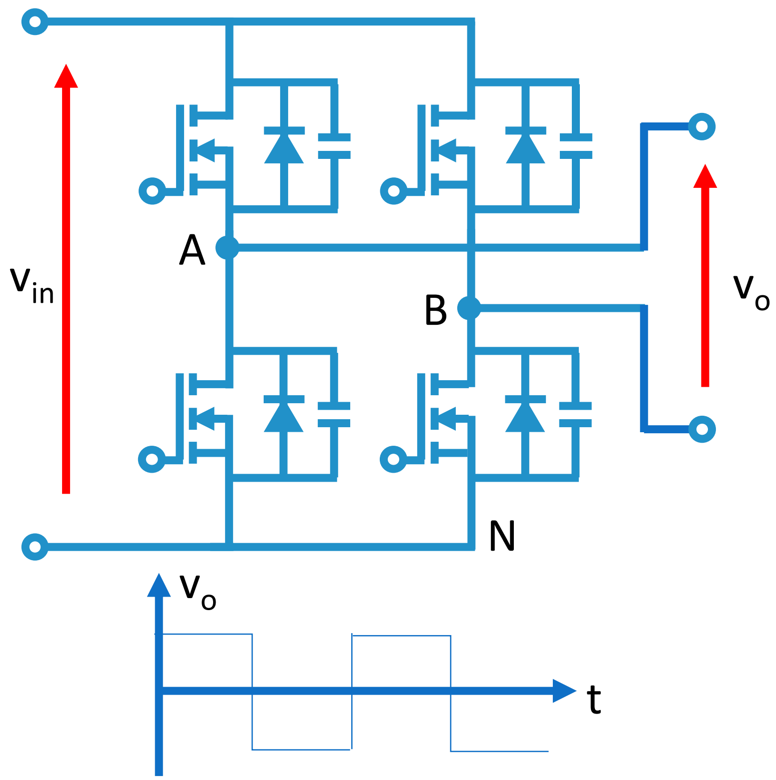
(c) Voltage Lift / Switched Inductance

Switch Network

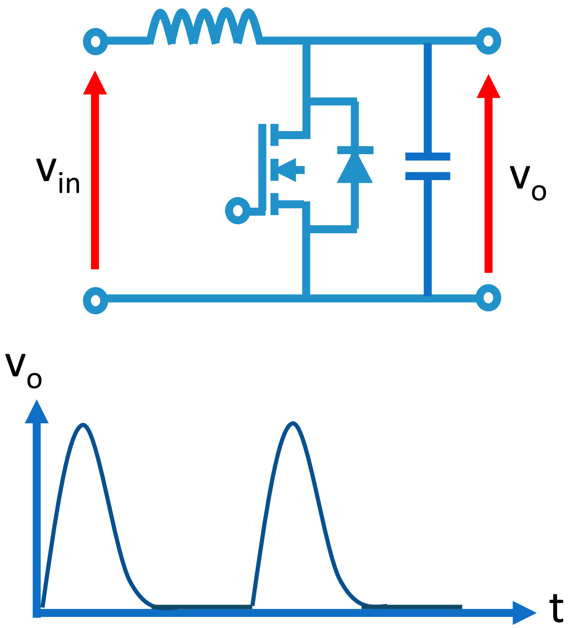
Half Bridge



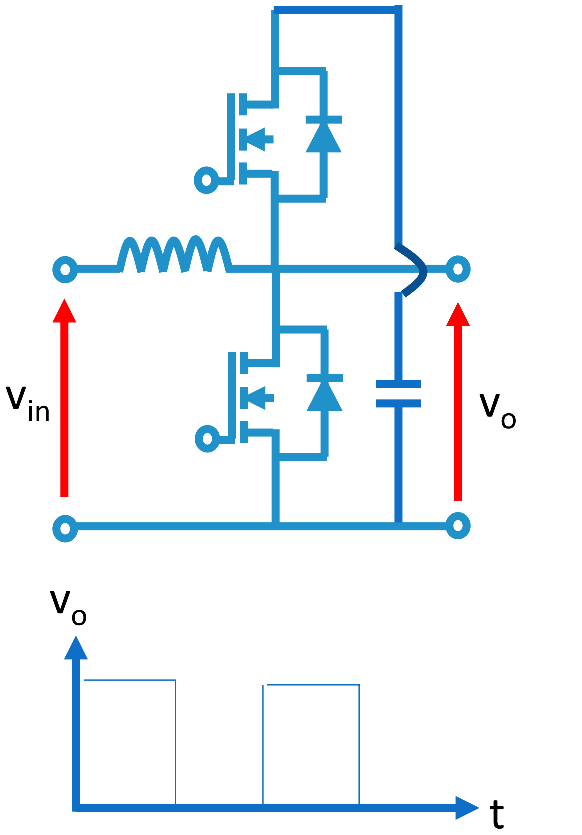
Full Bridge



Class E Topology



Half-Bridge Boost Topology



Consideration: Ease in modelling, Output Waveform, Number of Components, Input Independent, Boosting ability

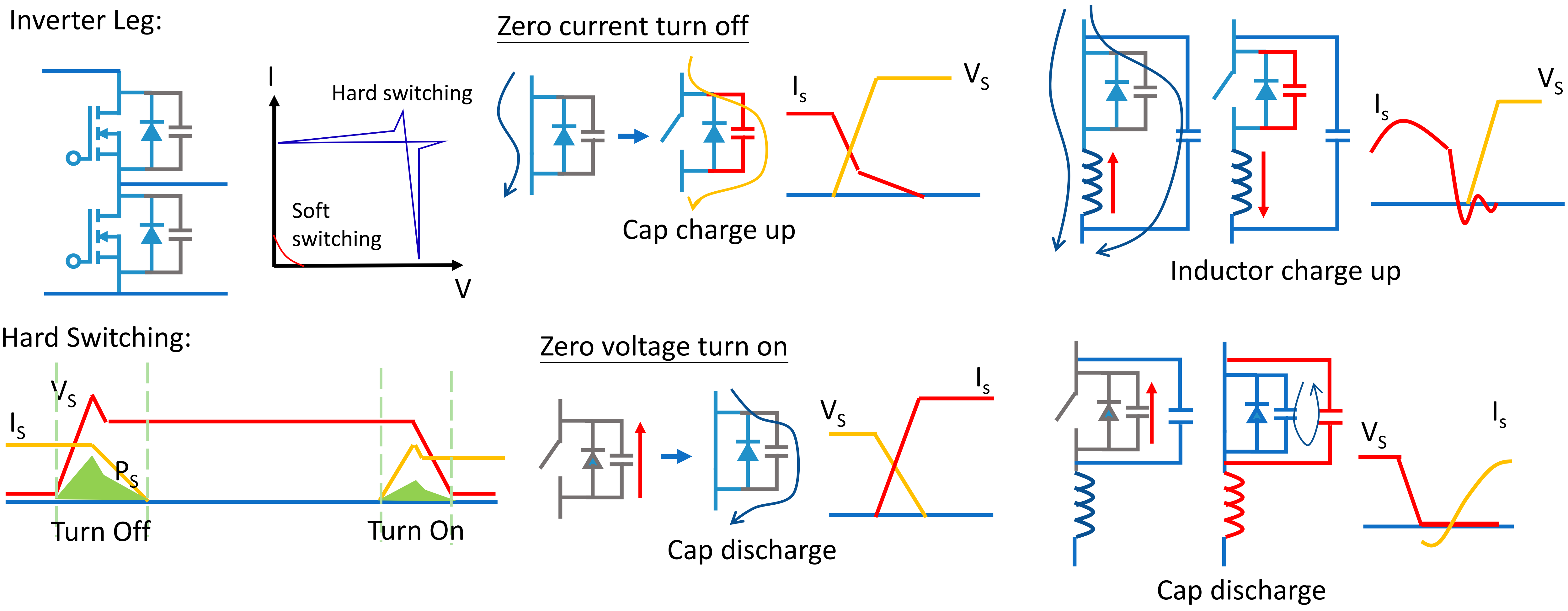
Switching loss

Inverter Leg:

Zero current turn off

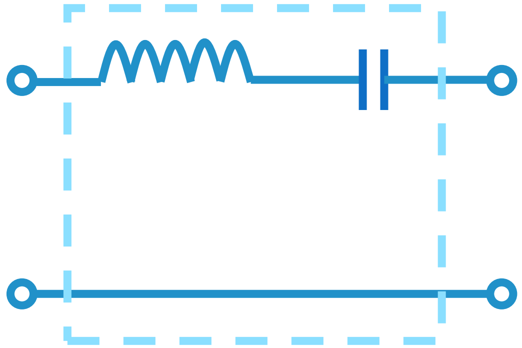
Hard Switching:

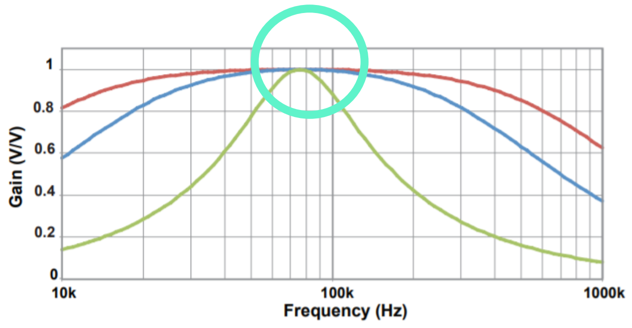
Zero voltage turn on



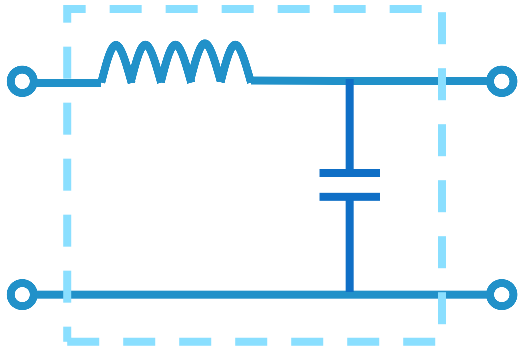
Resonant Tank

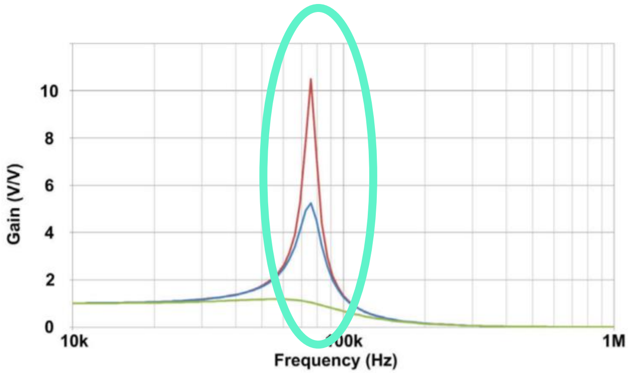
Series Resonant Tank



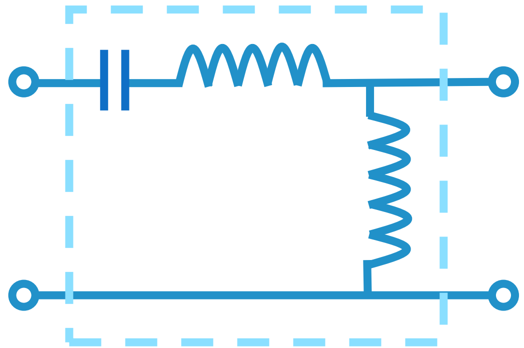


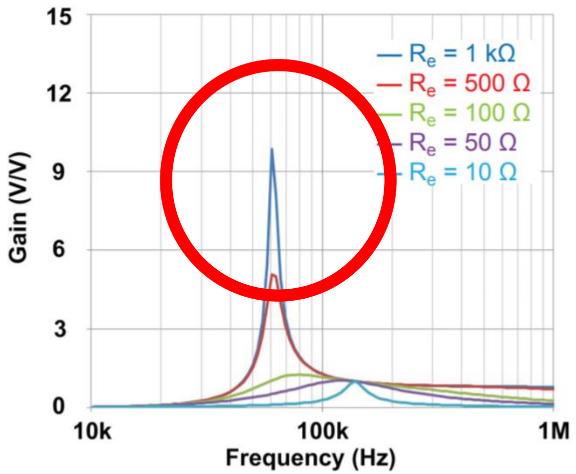
Parallel Resonant Tank



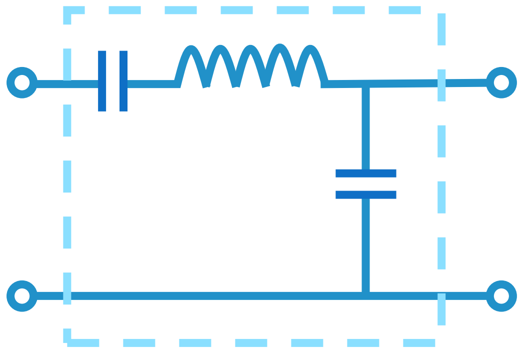


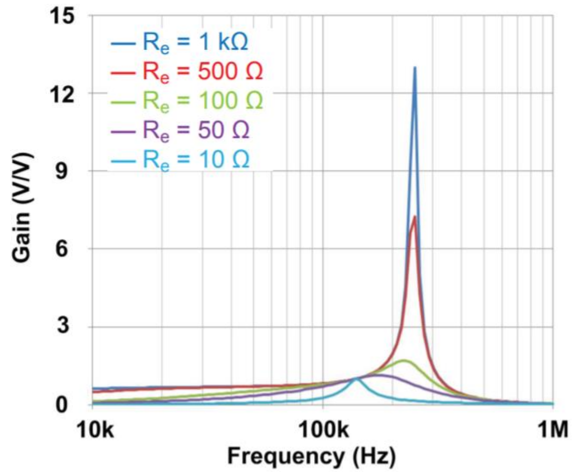
LLC Resonant Tank





LCC Resonant Tank

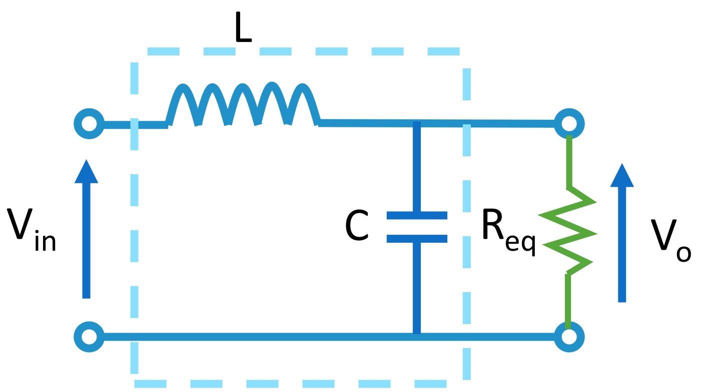




Texas Instruments (2018), Survey of Resonant Converter Topology,
2018 Texas Instruments Power Supply Design Seminar SEM2300, Topic 1

Targeted Resonant Tank

Parallel Resonant Tank :



Voltage Gain:

$$\frac{V_o}{V_{in}} = \frac{\frac{1}{j\omega C} \parallel R}{\frac{1}{j\omega C} \parallel R + j\omega L} = \frac{1}{(1-\omega^2 LC) + \frac{j\omega L}{R_{eq}}}$$

$$\left|\frac{V_o}{V_{in}}\right| = \frac{1}{\sqrt{(1-\omega^2 LC)^2 + \left(\frac{\omega L}{R_{eq}}\right)^2}}$$

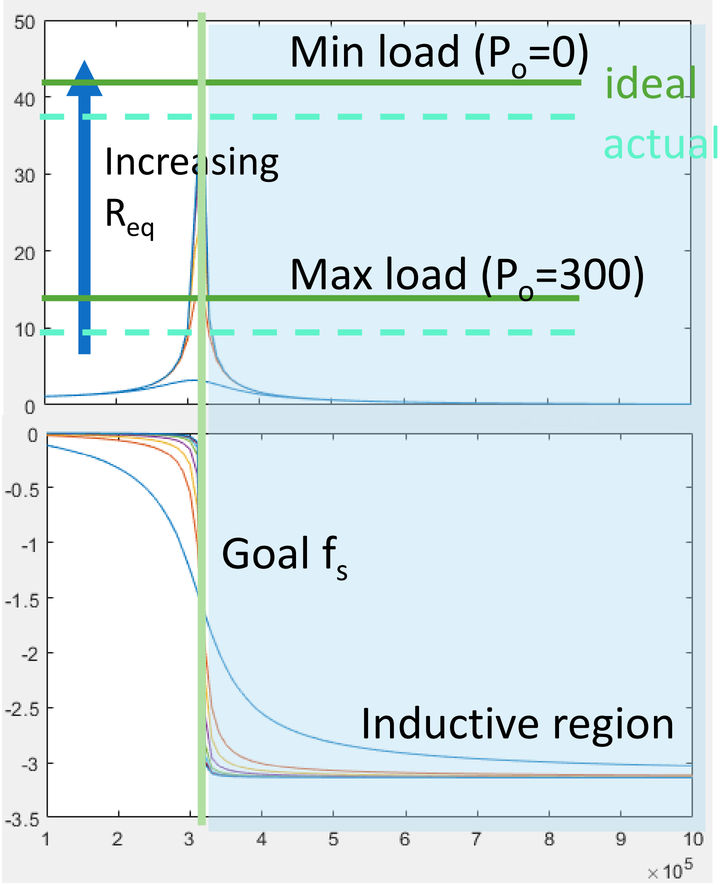
With $\omega \sim 10^5, R_{eq} \sim 10^3, L \sim 10^{-6}$, the term $\frac{\omega L}{R_{eq}}$ is negligible.

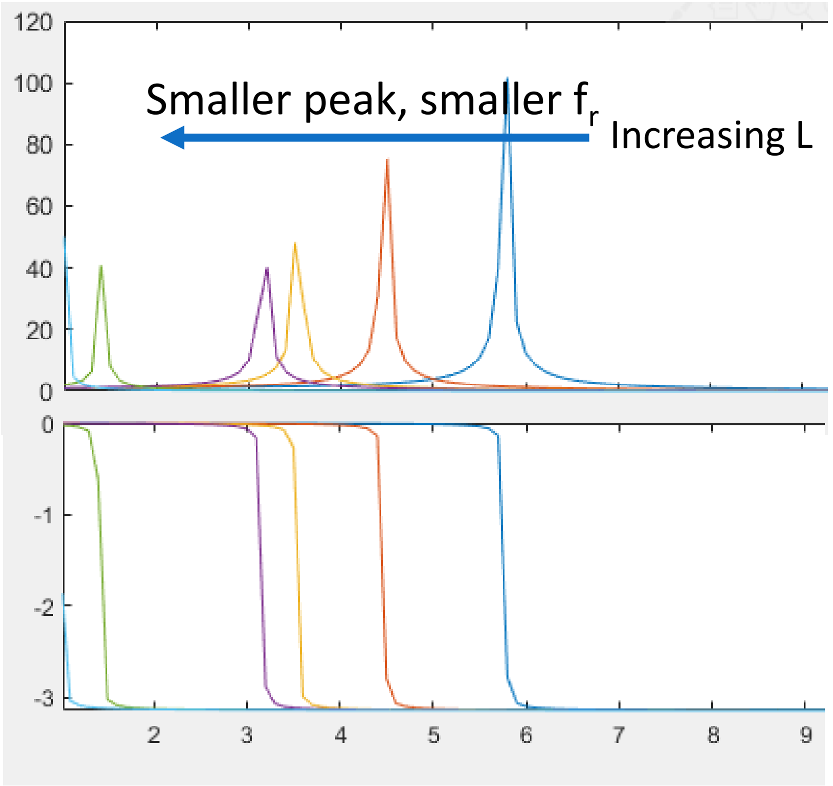
Hence, the resonant frequency is located at:

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

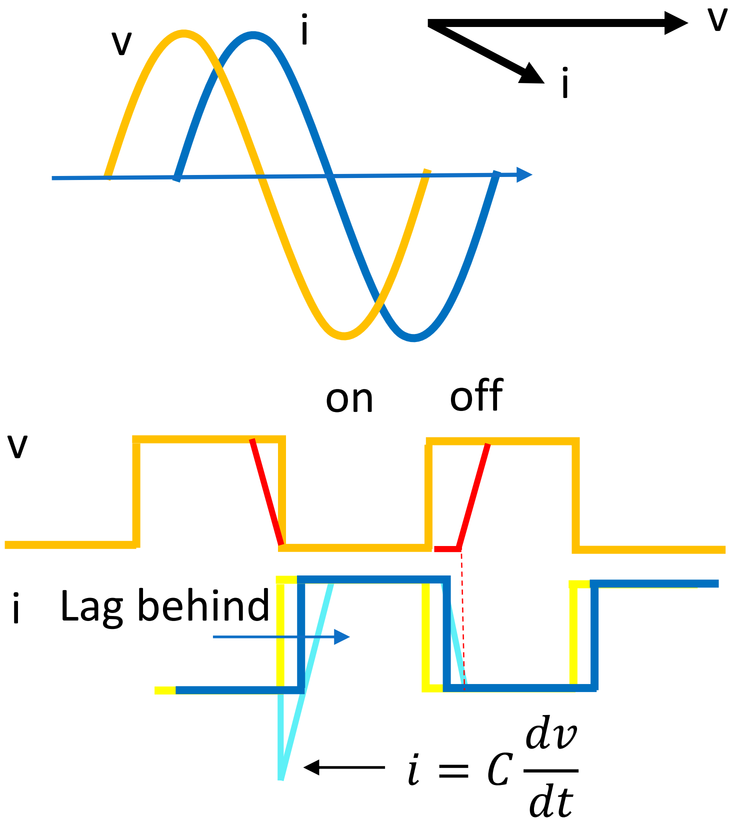
Targeted Resonant Tank

Parallel Resonant Tank (Bode Plot):





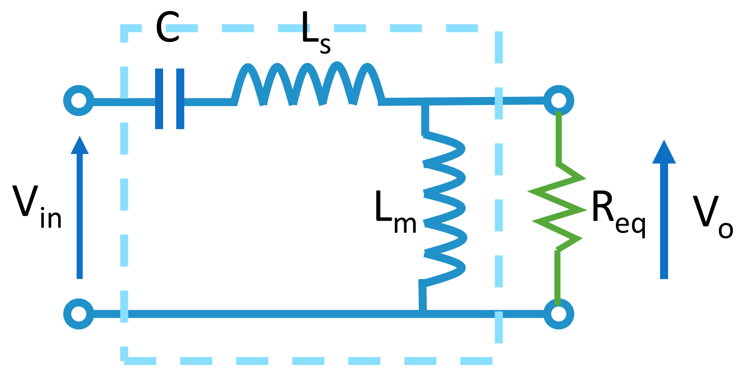
Remarks:



In ZVS, the current delays to rise. After current discharged, voltage starts to rise.

Targeted Resonant Converter

LLC Resonant Tank :



Voltage Gain:

$$\frac{V_o}{V_{in}} = \frac{j\omega L_m \parallel R_{eq}}{\frac{1}{j\omega C} + j\omega L_s + j\omega L_m \parallel R_{eq}} = \frac{1}{\left(1 + \frac{L_s}{L_m} - \frac{1}{\omega^2 L_m C}\right) + j\frac{1}{R_{eq}}\left(\omega L_s - \frac{1}{\omega C}\right)}$$

$$\left|\frac{V_o}{V_{in}}\right| = \frac{1}{\sqrt{\left(1 + \frac{L_s}{L_m} - \frac{1}{\omega^2 L_m C}\right)^2 + \frac{1}{R_{eq}^2}\left(\omega L_s - \frac{1}{\omega C}\right)^2}}$$

If $L_m \to \infty$, It is a series LC tank.
Resonant frequency:

$$f_r = \frac{1}{2\pi\sqrt{L_s C}}$$

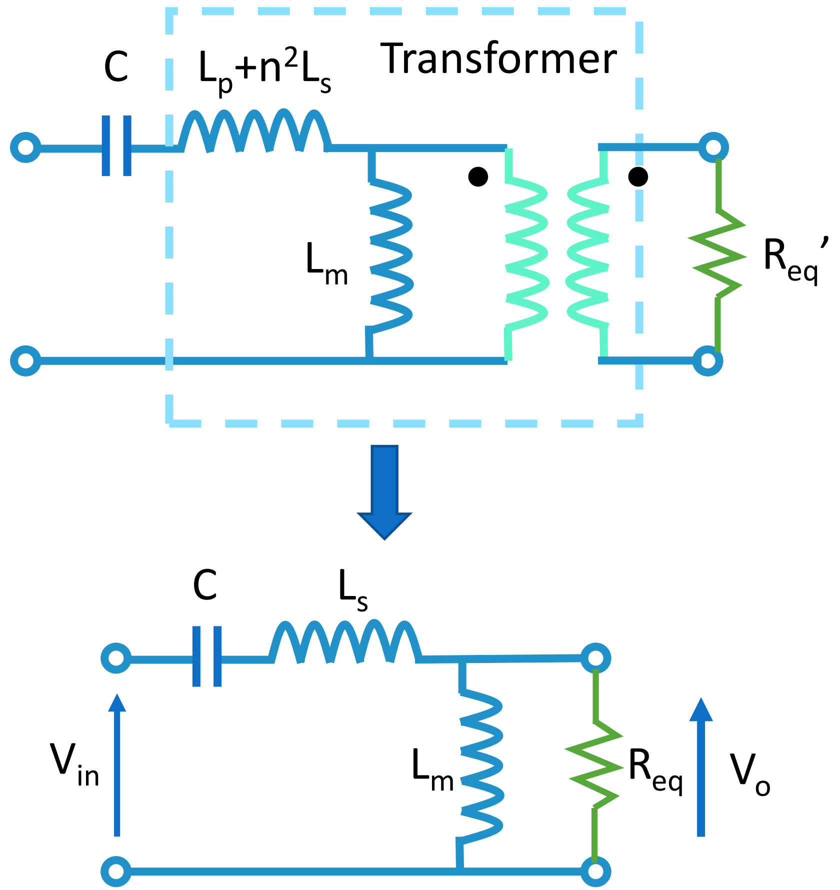
If $L_s \to 0$, i.e. Ls is negligible, it is a parallel LC tank.
Resonant frequency:

$$f_r = \frac{1}{2\pi\sqrt{L_m C}}$$

* Either it is load dependent or it is with low gain

Targeted Resonant Converter

Lr and Lm are normally obtained from a **transformer**.



Let $k = \frac{L_m}{L_s}$, $\omega_{fixed} = \frac{1}{\sqrt{L_s C}}$, $\omega_p = \frac{1}{\sqrt{L_m C}}$.

$$\left|\frac{V_o}{V_{in}}\right| = \frac{1}{\sqrt{\left(1 + k - \frac{\omega_p^2}{\omega^2}\right)^2 + \frac{\omega^2 L_s^2}{R_{eq}^2}\left(1 - \frac{\omega_{fixed}^2}{\omega^2}\right)^2}}$$

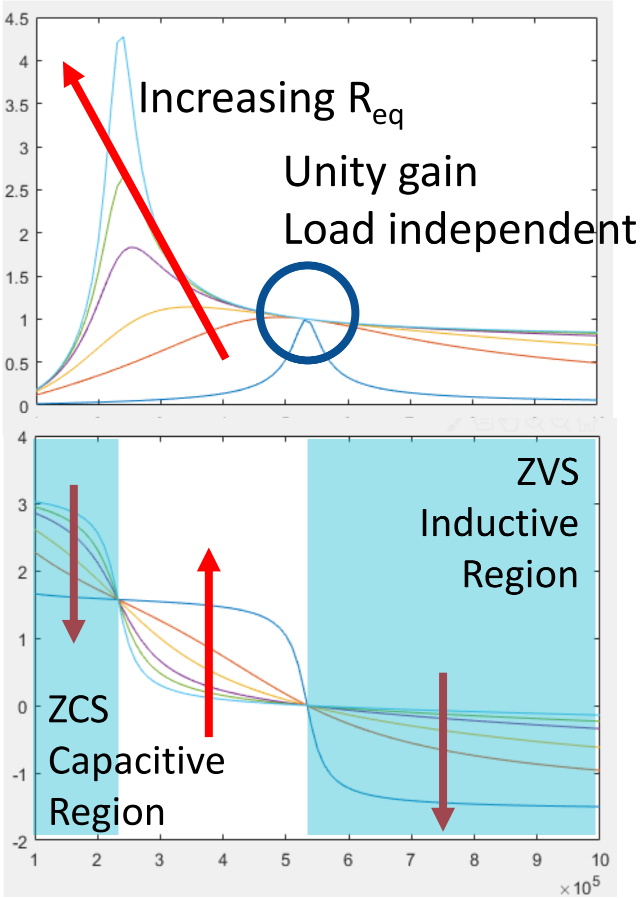
If $\omega = \omega_{fixed}$ and $L_m \to \infty$, it has unity gain: $\left|\frac{V_o}{V_{in}}\right| = 1$

If $\omega = \omega_p$, it has gain: $\left|\frac{V_o}{V_{in}}\right| = \frac{1}{\sqrt{k^2 + \frac{\omega_p^2 L_s^2}{R_{eq}^2}(1-k)^2}}$.

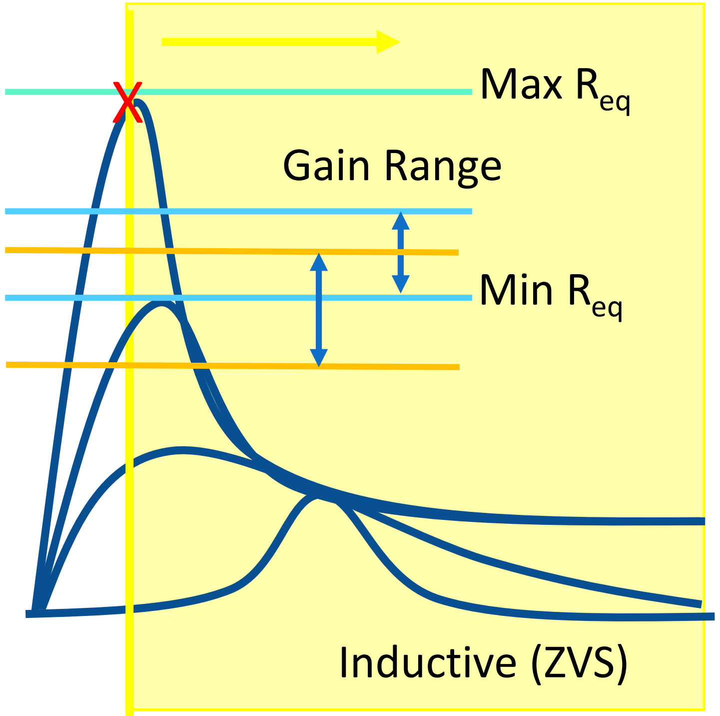
If $\frac{\omega_p L_s}{R_{eq}} \ll 1$, the gain becomes 1/k, or Ls/Lm

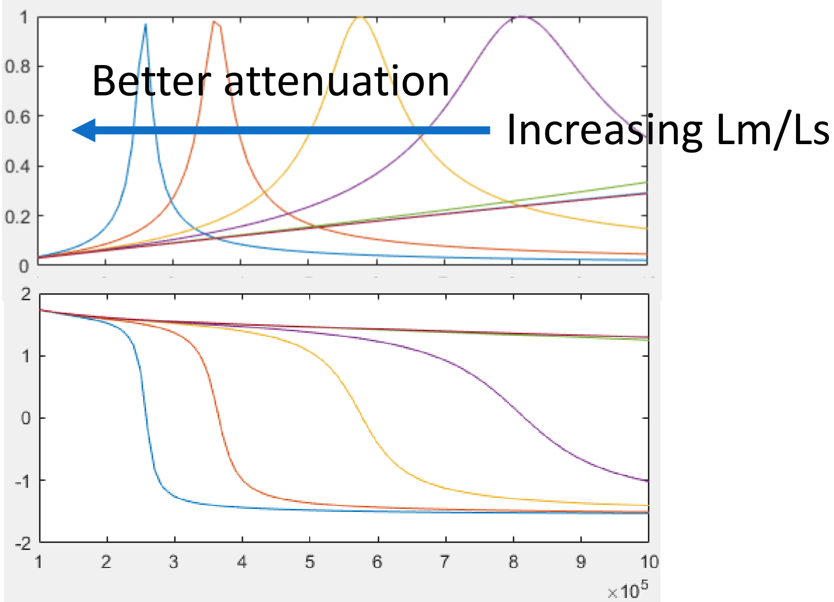
Target Resonant Converter

LLC Resonant Tank (Bode Plot): // Changing Lm/Ls ratio, Changing load



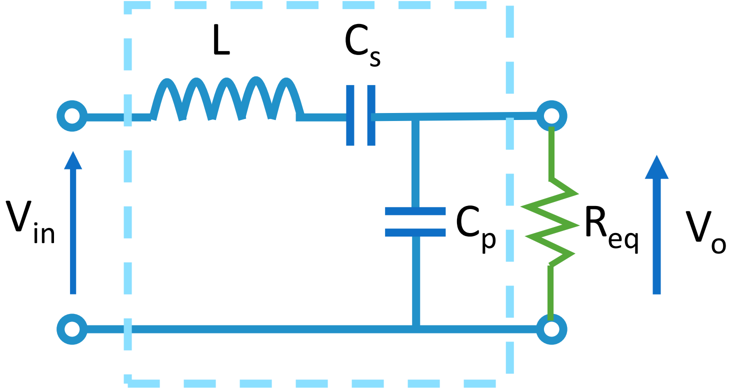
Design according to plot:





Target Resonant Converter

LCC Resonant Tank :



$$\frac{V_o}{V_{in}} = \frac{\frac{1}{j\omega C_p} \parallel R_{eq}}{\frac{1}{j\omega C_s} + j\omega L + \frac{1}{j\omega C_p} \parallel R_{eq}} = \frac{1}{\left(1 + \frac{C_p}{C_s} - \omega^2 L C_p\right) + j\frac{1}{R_{eq}}\left(\omega L - \frac{1}{\omega C_s}\right)}$$

$$\left|\frac{V_o}{V_{in}}\right| = \frac{1}{\sqrt{\left(1 + \frac{C_p}{C_s} - \omega^2 L C_p\right)^2 + \frac{1}{R_{eq}^2}\left(\omega L - \frac{1}{\omega C_s}\right)^2}}$$

For typical 300W, with V_0 = 66.7V,
R_{eq} = 14.8Ω → very small
With gain << 1

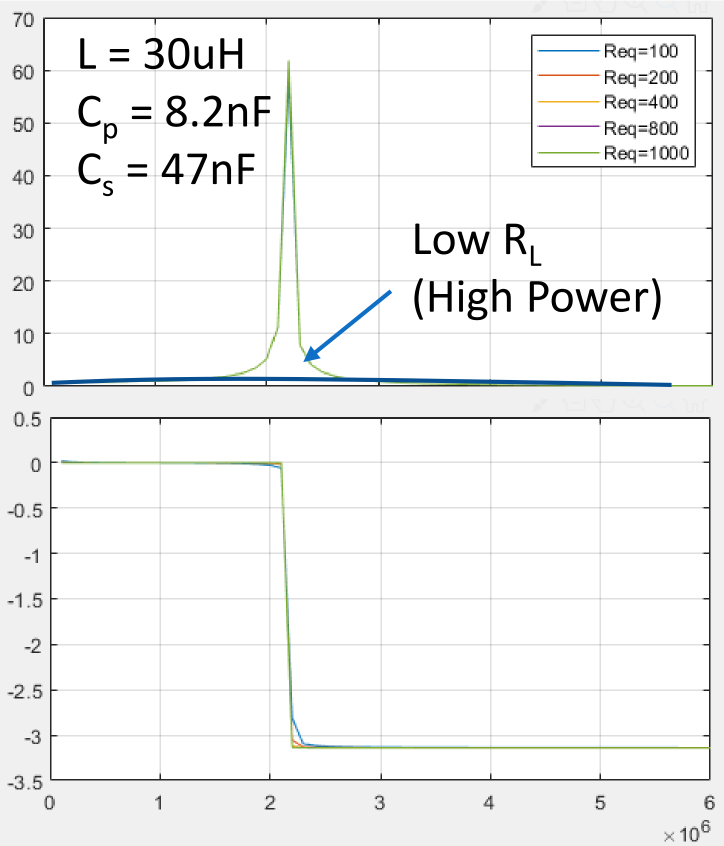
If R is large enough, With $\frac{1}{R_{eq}}\left(\omega L - \frac{1}{\omega C_s}\right) << 1$, $f_r = \frac{1}{2\pi}\sqrt{\frac{1 + \frac{C_p}{C_s}}{LC_p}}$

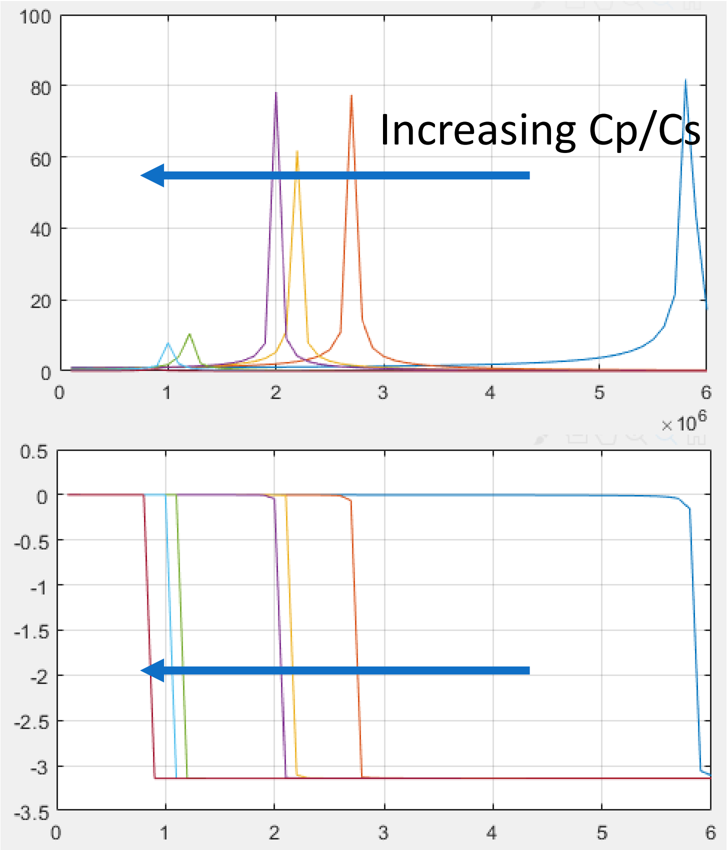
Not good for our design,
For low power only!!

If R is small enough, to make $\frac{1}{R_{eq}}\left(\omega L - \frac{1}{\omega C_s}\right) << 1$, $f_r = \frac{1}{2\pi}\frac{1}{\sqrt{LC_s}}$, Gain = 1

Target Resonant Converter

LCC Resonant Tank (Bode Plot): // Changing Cp/Cs ratio, Changing load





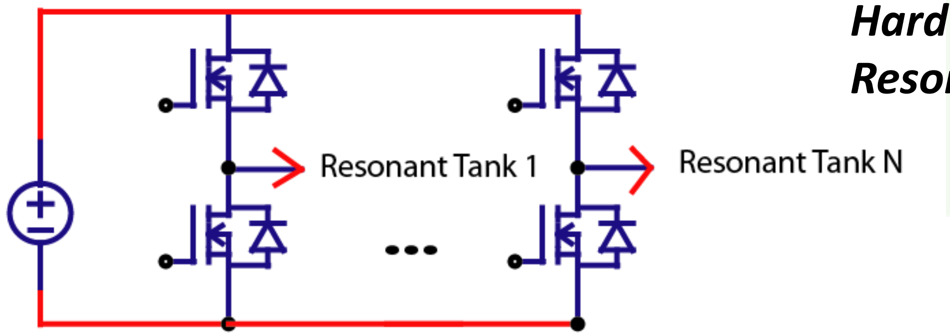
If we give up the properties of load independent,

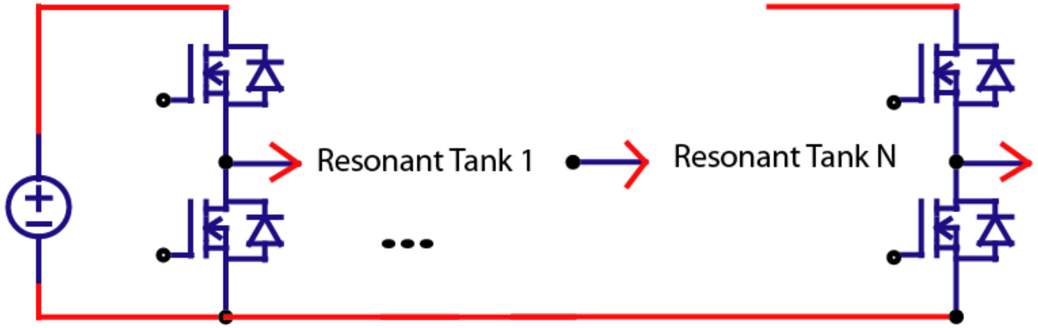
Given that $f_s = \frac{1}{2\pi}\sqrt{\frac{1+\frac{C_p}{C_s}}{LC_p}}$

$$\left|\frac{V_o}{V_{in}}\right| = \sqrt{\left(\frac{1}{C_p}+\frac{1}{C_s}\right)L} - \sqrt{\frac{LC_p}{C_s\left(1+C_p\right)}}$$

Other Possible Tanks

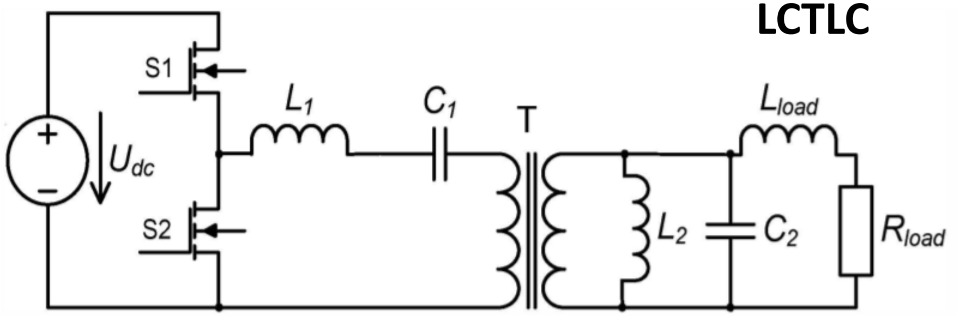
Hardware/Software Resonant?



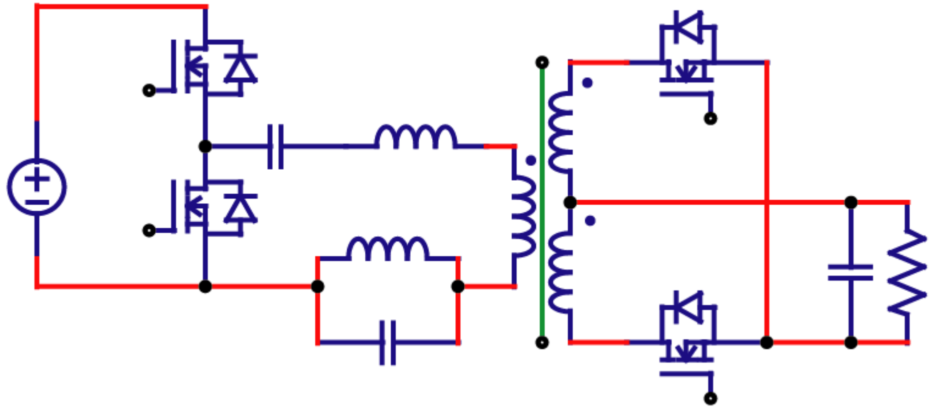


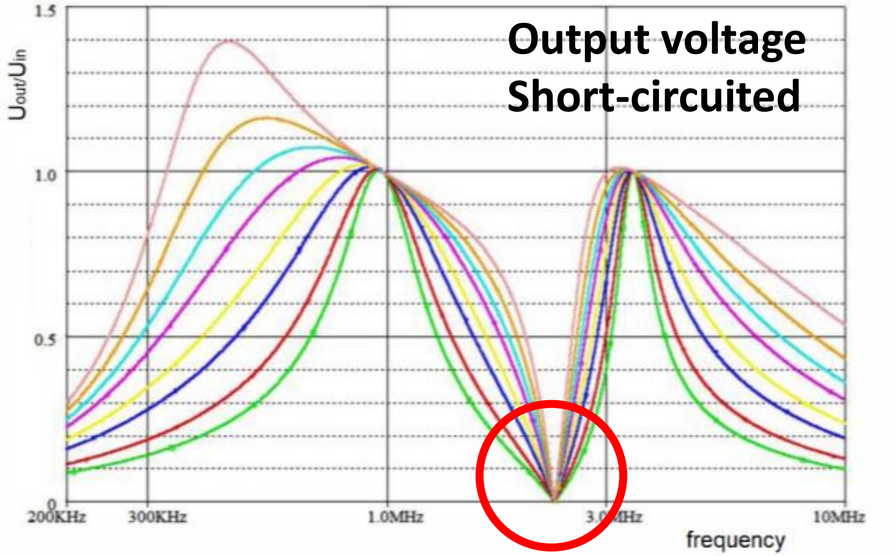
Multi-resonant Tank

LCTLC



LCLCL





Tank Design (LC Selection), for LLC tank

For Req, assume ΔV is negligible and power in = power out, $\left(\frac{V_{s(p-p)}}{\sqrt{2}}\right)^2 \frac{1}{R_{eq}} = \frac{\mathrm{V_o^2}}{\mathrm{R_o}} = \frac{m^2 V_{s(p-p)}^2}{R_o}$ $\quad R_{eq} = \frac{R_o}{2m^2}$

Select a switching frequency ω_s = 350kHz, to switch at the parallel resonant frequency, desired rated gain G = 10, a load factor Q to obtain such a gain.

$$f_{series} = \frac{1}{2\pi\sqrt{L_s C}} \qquad \boxed{f_{parallel} = \frac{1}{2\pi\sqrt{L_m C}}}$$

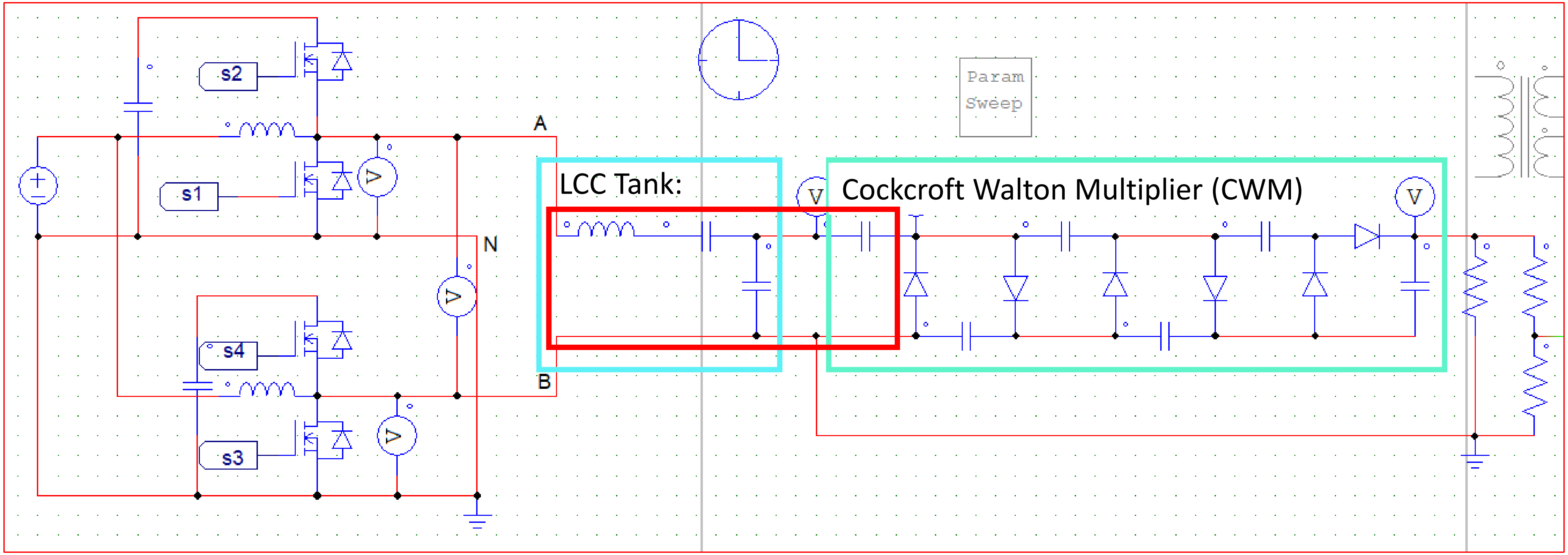
$$G = \frac{1}{\sqrt{\left(\frac{L_s}{L_m}\right)^2 + \frac{1}{R_{eq}^2}\left(\omega L_s - \frac{1}{\omega C}\right)^2}} = 10 \quad \ldots(1)$$

$$f_p = \frac{1}{2\pi\sqrt{L_m C}} = 350kHz \qquad \ldots(2)$$

$$Q = \frac{L_s/C}{R_{eq}} = (\qquad) \qquad \ldots(3)$$

* The load factor Q is to determine the attenuation ability for non-fundamental frequencies.

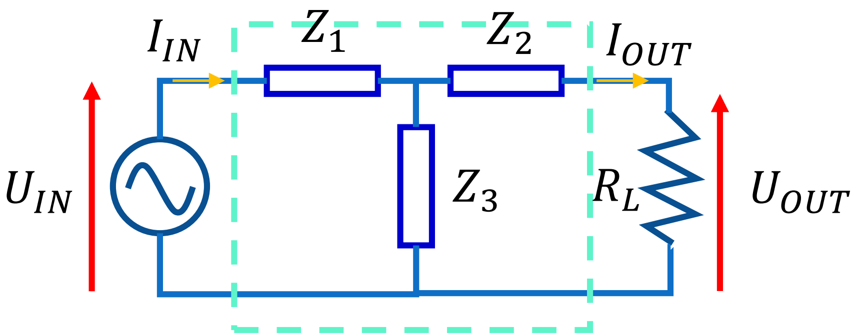
Difficulties in designing an LCC with CWM



* The **resonant frequency** and the **tank properties** are altered with the first capacitor in the Cockcroft Walton Multiplier. There is no Thevenin or Equivalent Circuit for the CWM so that the actual capacitance can be determined.

Load independent Resonant Tank Design

Designing a resonant tank with properties of load independence, high efficiency and avoiding bifurcation resistance is the critical task for wireless power transfer. As given in **Zhang & Mi (2016)**, the following method is used for designing load independent resonant tank.



By Ohm's Law and KVL:

$$U_{OUT} = I_{OUT} R_L$$
$$U_{IN} = I_{IN} Z_1 + (I_{IN} - I_{OUT}) Z_3$$
$$U_{IN} = I_{IN} Z_1 + I_{OUT} Z_2 + U_{OUT}$$

Simplifying,

$$U_{IN} = \left(1 + \frac{Z_1}{Z_3}\right) U_{OUT} + \frac{\Delta}{R_L} U_{OUT}$$

To make the system load independent,

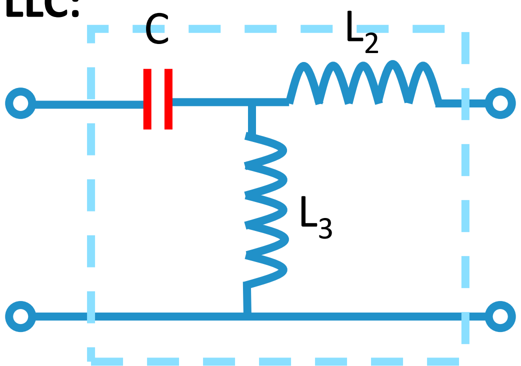
$$\Delta = \frac{Z_1 Z_2 + Z_2 Z_3 + Z_1 Z_3}{Z_3} = 0$$

* Zhang & Mi (2016), Compensation Topologies of High Power Wireless Power Transfer System, IEEE Transactions of Vehicular Technology, Vol.65, No.6, pp. 4768-4778

Load independent Resonant Tank Design

To inherit the properties from the LC series and parallel resonant tank, the following LLC and LCC tank is studied.

LLC:

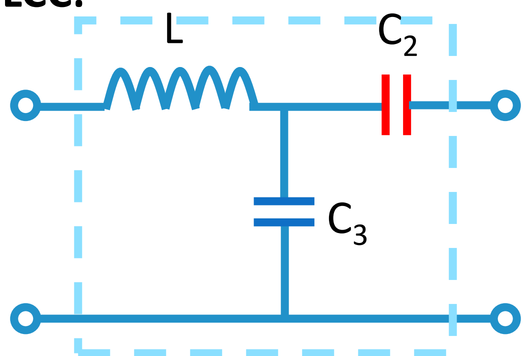


$$\Delta = \frac{Z_1 Z_2 + Z_2 Z_3 + Z_1 Z_3}{Z_3} = \frac{\frac{L_2 + L_3}{C} - \omega^2 L_2 L_3}{j\omega L_3} = 0$$

$$\boxed{U_{IN} = \left(1 + \frac{Z_1}{Z_3}\right) U_{OUT}}$$

The requirement is satisfied with $\omega_r = \sqrt{\frac{1}{L_2 C} + \frac{1}{L_3 C}}$. The property of the tank is then determined by $U_{IN}(1 - \frac{1}{\omega^2 C L_3}) = U_{OUT}$, which is inherently a parallel LC tank with high gain.

LCC:



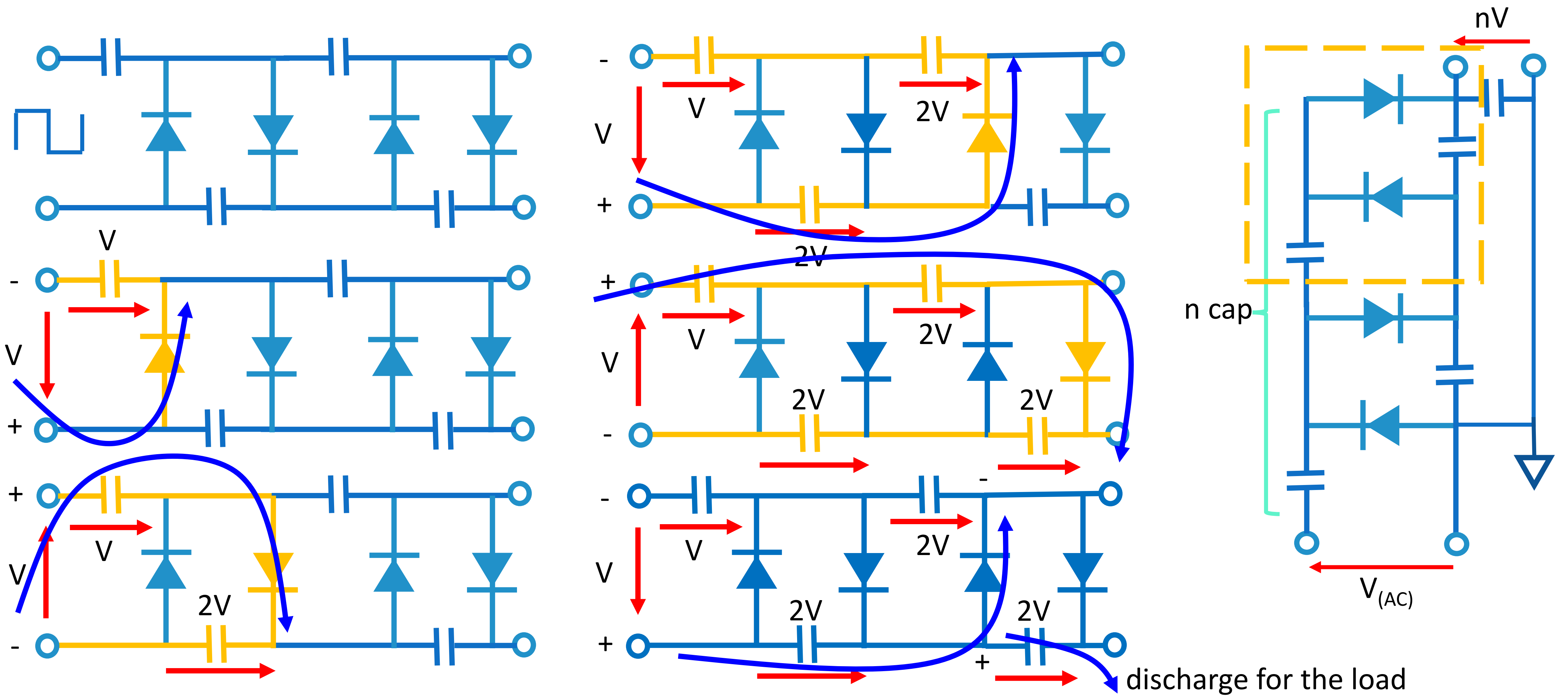
$$\Delta = -\left(\frac{L}{C_2} + \frac{L}{C_3} - \frac{1}{\omega^2 C_2 C_3}\right)\frac{1}{\omega C_3} \rightarrow \omega_r = \frac{1}{\sqrt{LC_2 + LC_3}}$$

$$U_{IN} = \frac{1}{(1 - \omega^2 L C_3)} U_{OUT}$$

Consideration:

- Component Size (efficiency)
- Separation from latter circuit
- Voltage gain

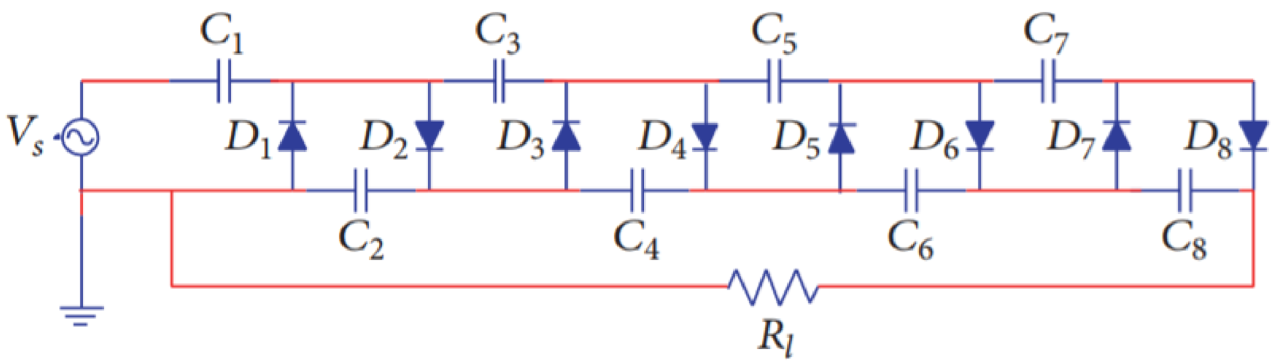
Cockcroft Walton Multiplier



[1] Mohsen Ruzbehani, A Comparative Study of Symmetrical Cockcroft-Walton Voltage Multipliers, Journal of Electrical and Computer Engineering Volume 2017

[2] I.C. Kobougias and E.C. Tatakis , Optimal Design of a Half-Wave Cockcroft-Walton Voltage Multiplier with Minimum Total Capacitance , IEEE Transaction of Power Electronics, 2010

Cockcroft Walton Multiplier



Voltage Drop:

C_2 loses nq charges, and C_1 has to retake it.

Hence C_1 and C_2 are charged up to $V_s - \frac{nq}{C_1}$ and $2V_s - \frac{nq}{C_2}$

$$\Delta V_2 = \frac{q}{C} n \qquad \Delta V_4 = \frac{q}{C}(2n + (n-1)) \quad \dots$$

q = charge from C_{2n} to the load

(peak to peak -> 2xδV)

$$2\delta V_{2n} = \frac{q}{C_{2n}}$$

q to the load; q to C_{2n-1}

$$2\delta V_{2n-2} = \frac{2q}{C_{2n-2}}$$

By intuition:

$$2\delta V_{2n-2i} = \frac{(i+1)q}{C_{2n-2i}}$$

Total ripple:

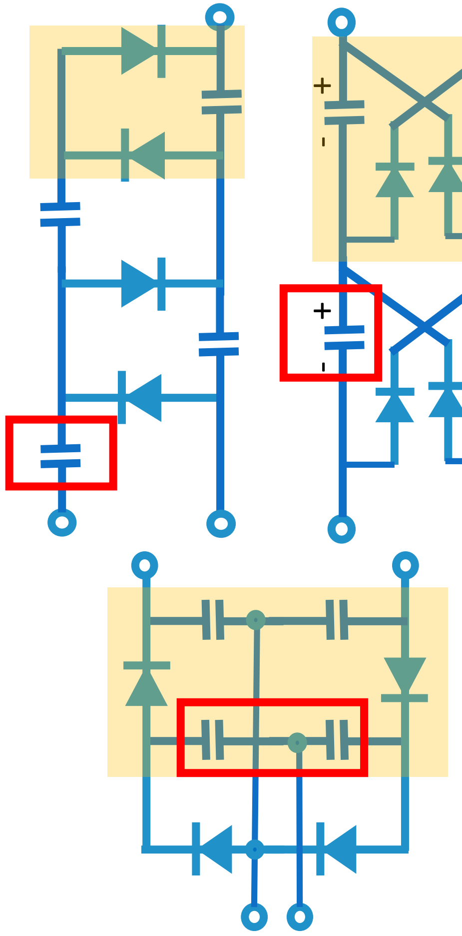
$$2\delta V = q\left(\frac{1}{C_{2n}} + \frac{1}{C_{2n-2}} + \cdots + \frac{1}{C_2}\right) = \frac{I}{fC}\left(\frac{n(n+1)}{2}\right)$$

$$\Delta V_{tot} = \frac{I}{fC}\left(\frac{2}{3}n^3 + \frac{1}{2}n^2 - \frac{n}{6}\right)$$

max $V_o = 2nV_s - \frac{I}{fC}\left(\frac{2}{3}n^3 + \frac{1}{2}n^2 - \frac{n}{6}\right)$

* Output ripple can be filtered by large C_o
* max V_o depends on load current
* q requested by load may not large enough to charge the capacitor to enough voltage

Load independent Resonant Tank Design



Possible AC/DC with boosting

In the latter stage, it is a voltage multiplier or a transformer.

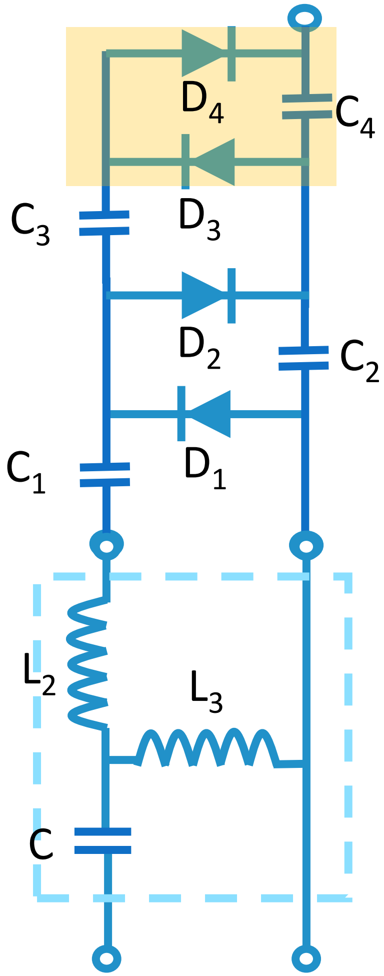
There is often capacitor connected.

$$\Delta = \frac{Z_1 Z_2 + Z_2 Z_3 + Z_1 Z_3}{Z_3} = 0 \quad \text{(only } Z_2 \text{ is affected)}$$

For LLC: $Z_2 = j\omega L + \frac{1}{j\omega C_{eq}}$

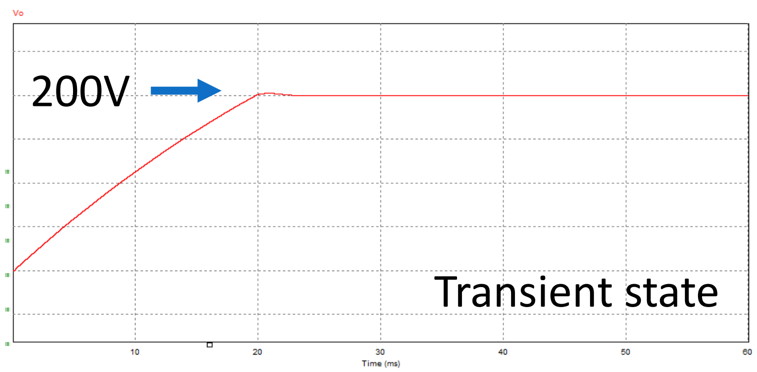
For LCC: $Z_2 = \frac{1}{j\omega C_2} + \frac{1}{j\omega C_{eq}}$

To make the CWM capacitance negligible, one can increase C_{eq} to 10 times larger than the original part.

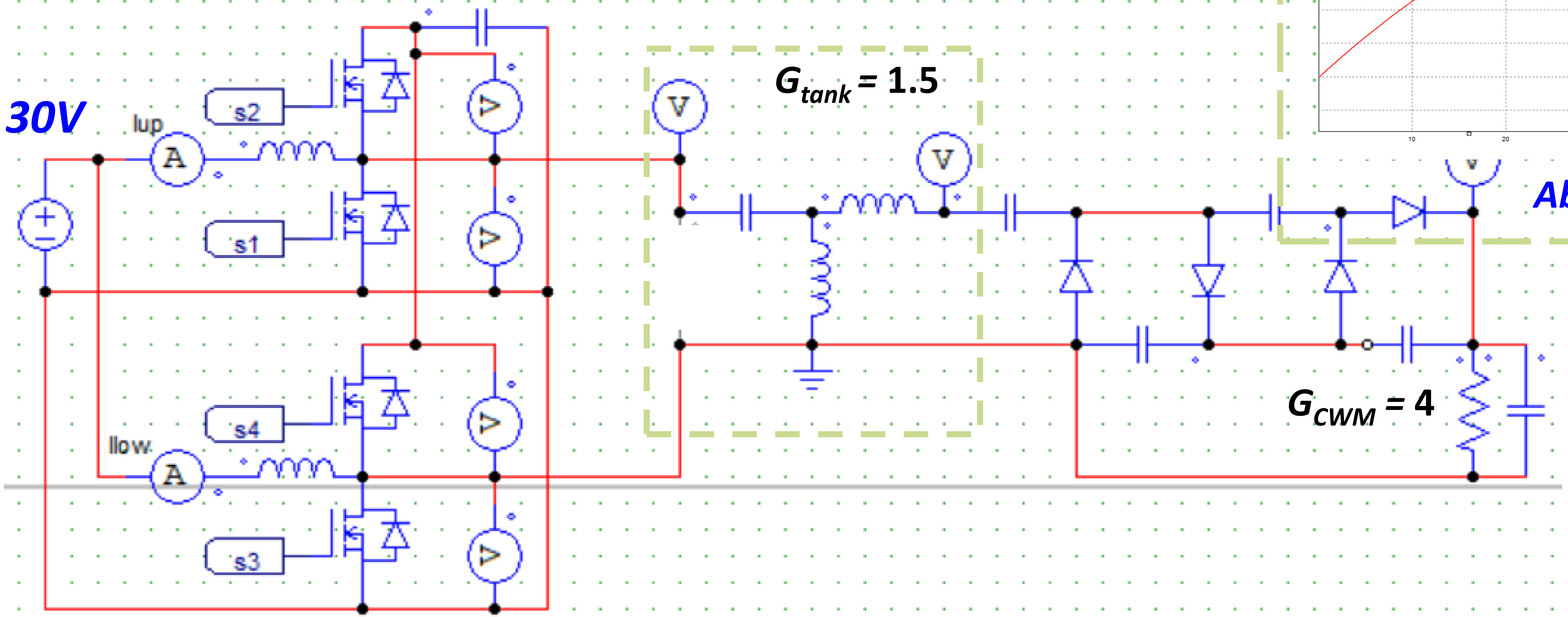


Simulation

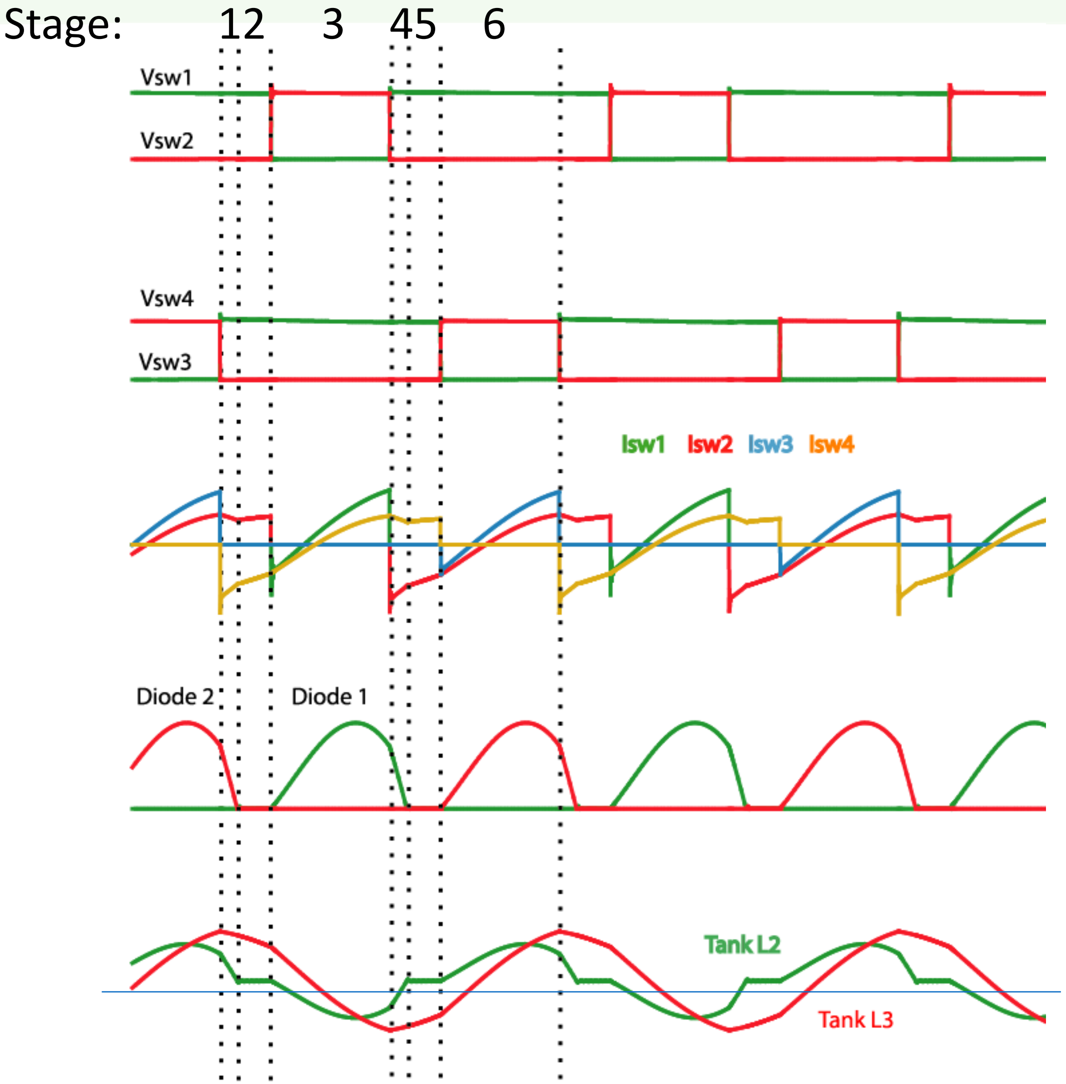
Ideal situation,
*yet in hardware, a **90V output in simulation** will result in a **45V output in hardware**.*

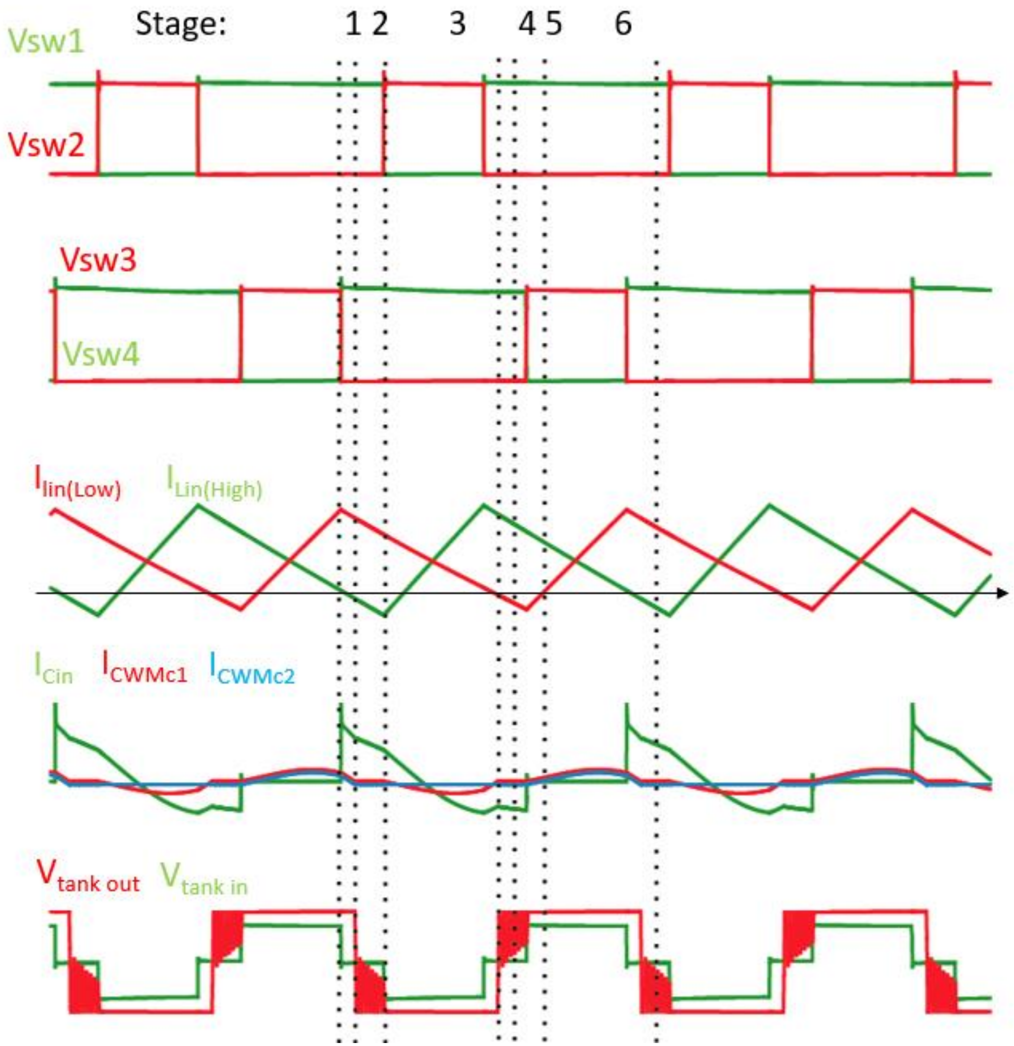


Able to achieve 200V

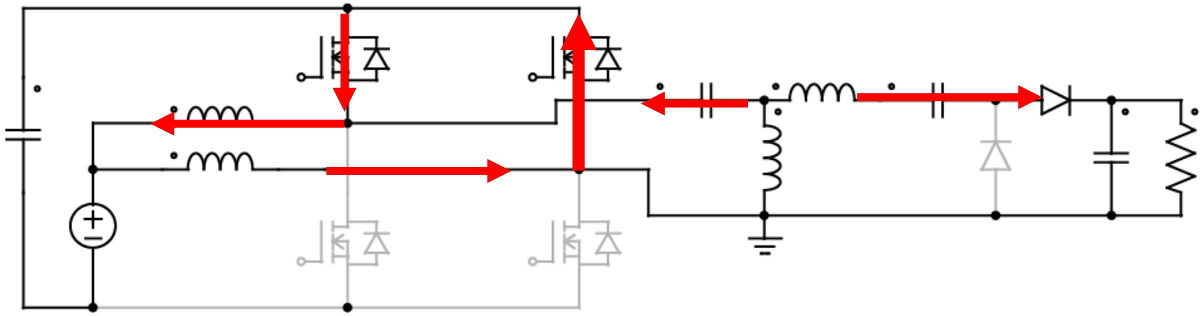


Current Flow Diagram (steady state)



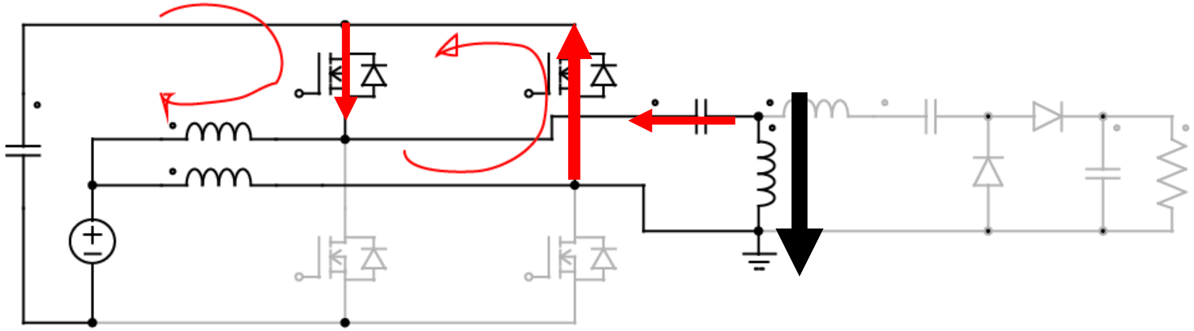


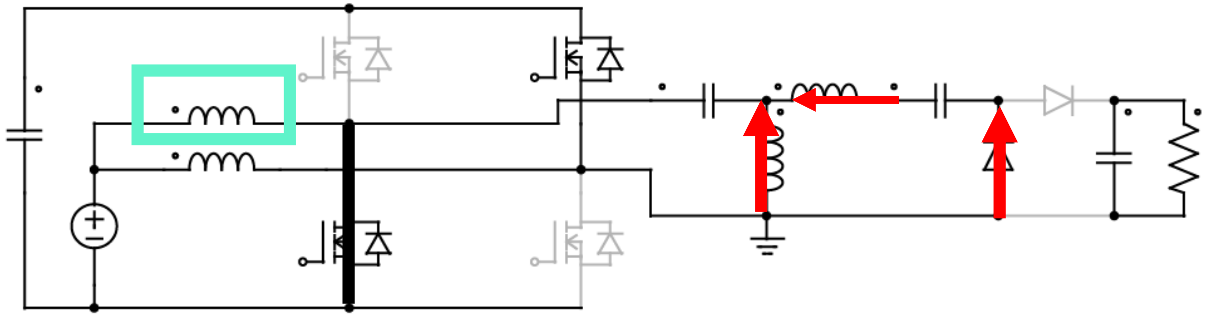
Current Flow Diagram



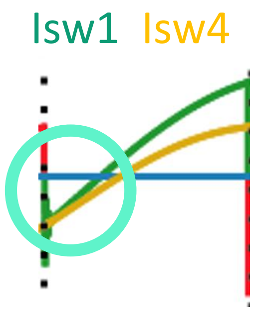
Stage 1:
Switch 2 & 4, diode 2 are ON. The residue current in the load side is decreasing with higher voltage in load side.

Stage 2:
Diode 2 is OFF. The differential current of the two branches circulates to obtain high voltage in the tank.



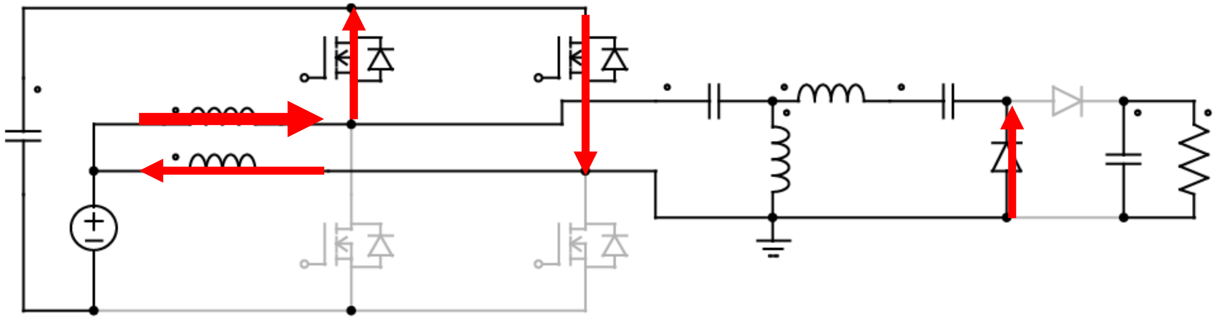


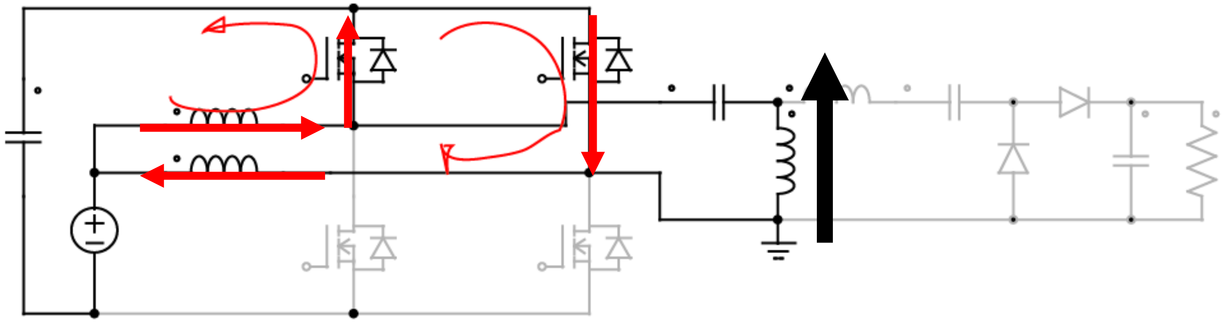
Stage 3:
Switch 1 & 4 are ON. The upper branch inductor is storing energy.



Current Flow Diagram

Stage 4:
Switches 2 & 4 are ON. The current flowing through diode 1 is reducing with negative voltage.

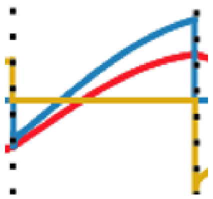


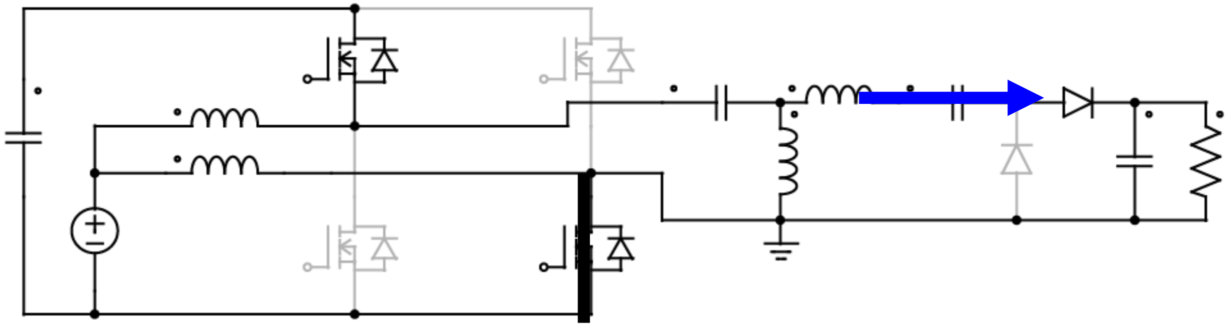


Stage 5:
Switches 2 & 4 are ON. The current circulates the tank and reaches a high voltage. The upper inductor clamped to the capacitor voltage.

Stage 6:
The tank receives higher voltage than the load, hence transferring energy to the load. The lower inductor charges with Switch 3 turns ON.

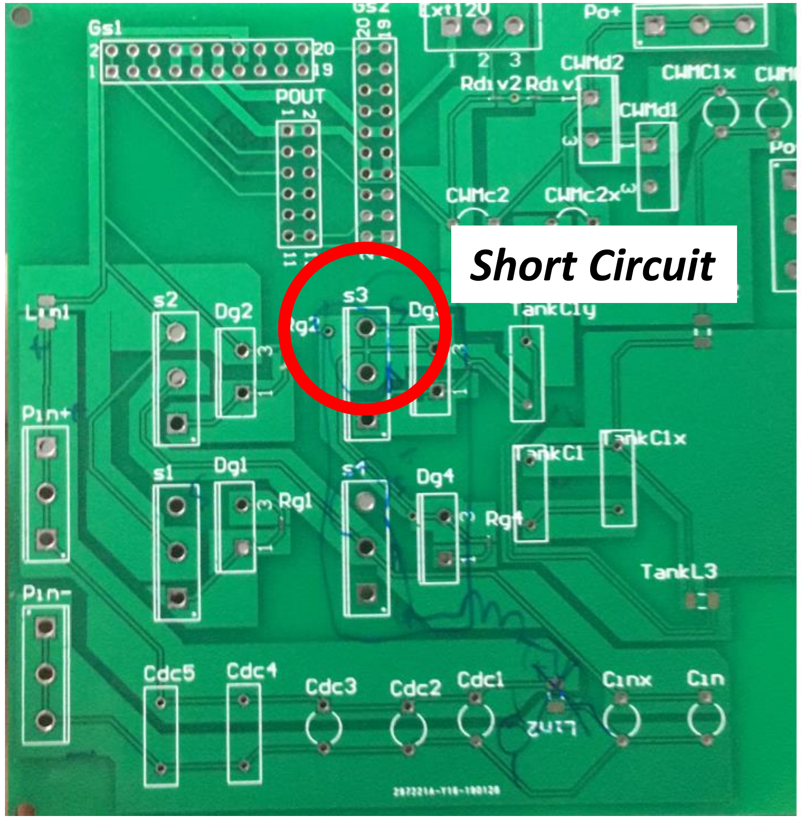
Isw2 Isw3



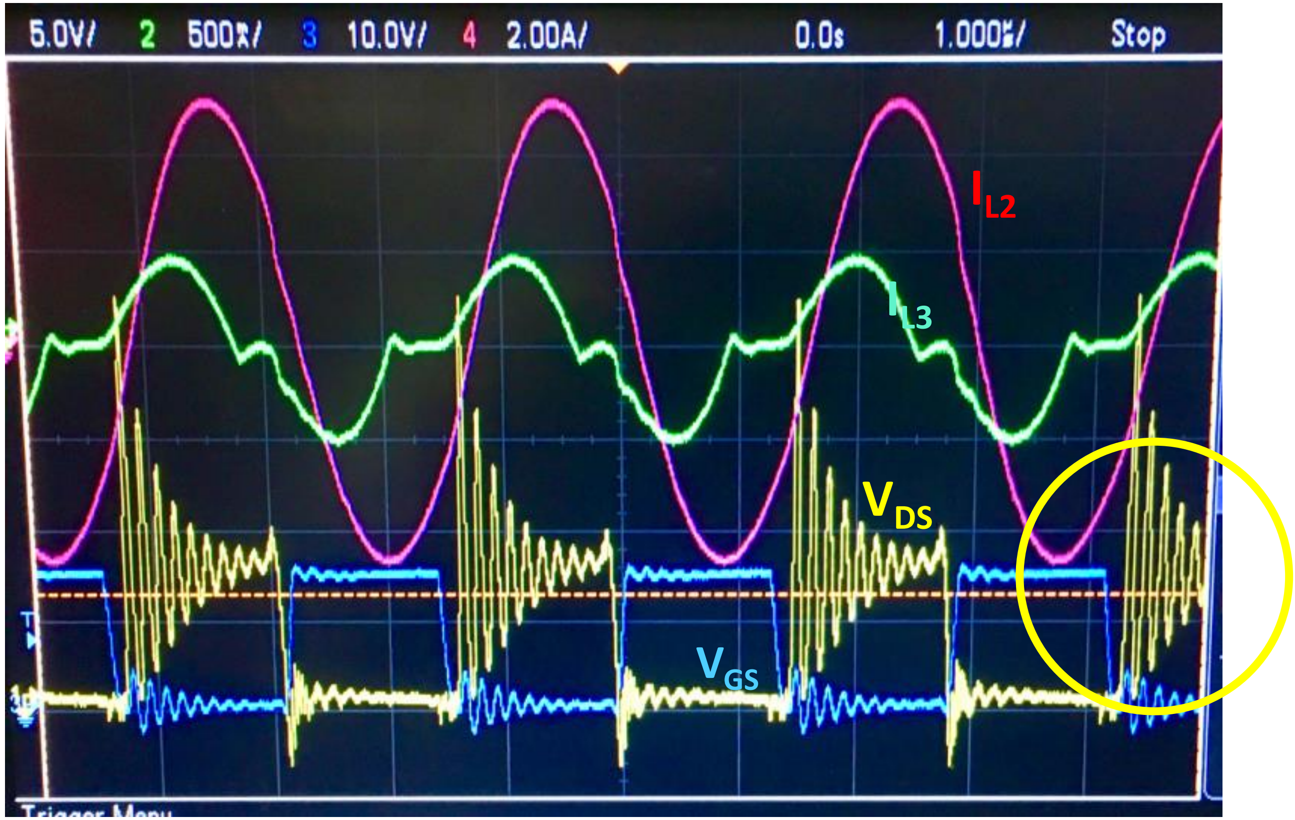


Results (Failure)

Version 1: Failure on Short Circuit



Yet PWM can still apply to test the performance of latter part.



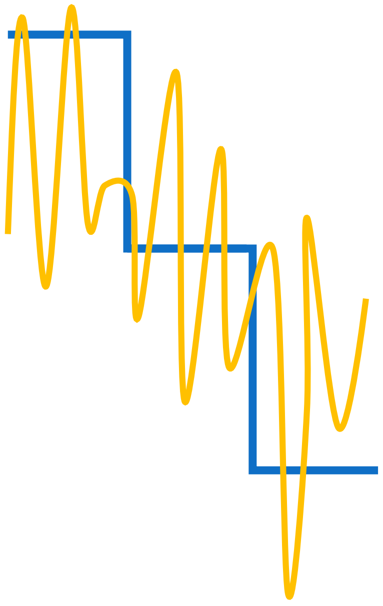
Large Ringing due to resonant

Results (Failure)

Version 5: Separated Inverter & CWM
Version 6: Common Grounding between input & output

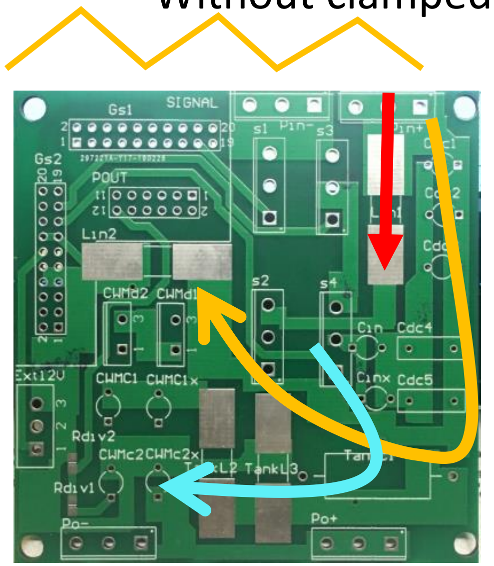
Version 2: Owing to the twisting of copper wire, **waveform distortion** is excessive large. (Aim to reduce size)
Version 3: **Clamped capacitor** is separated into two, voltage and current of upper branch and lower branch is different. (Make sure no short circuit during switching)
Version 4: **Grounding Issue,** Voltage fluctuation pollutes the grounding voltage of PWM signal (Reconnect the clamp cap)

Expected Vs Actual



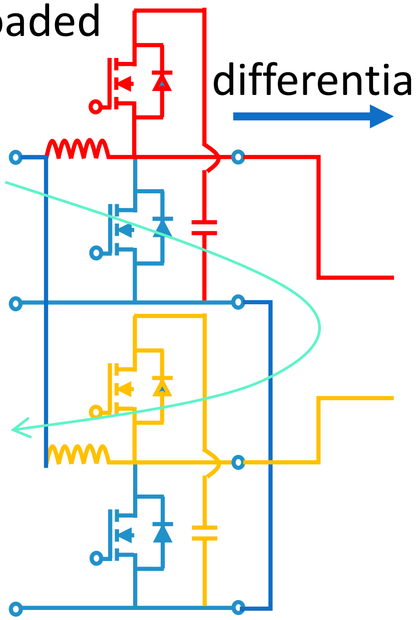
Clamped

Without clamped



Heavily loaded

differential

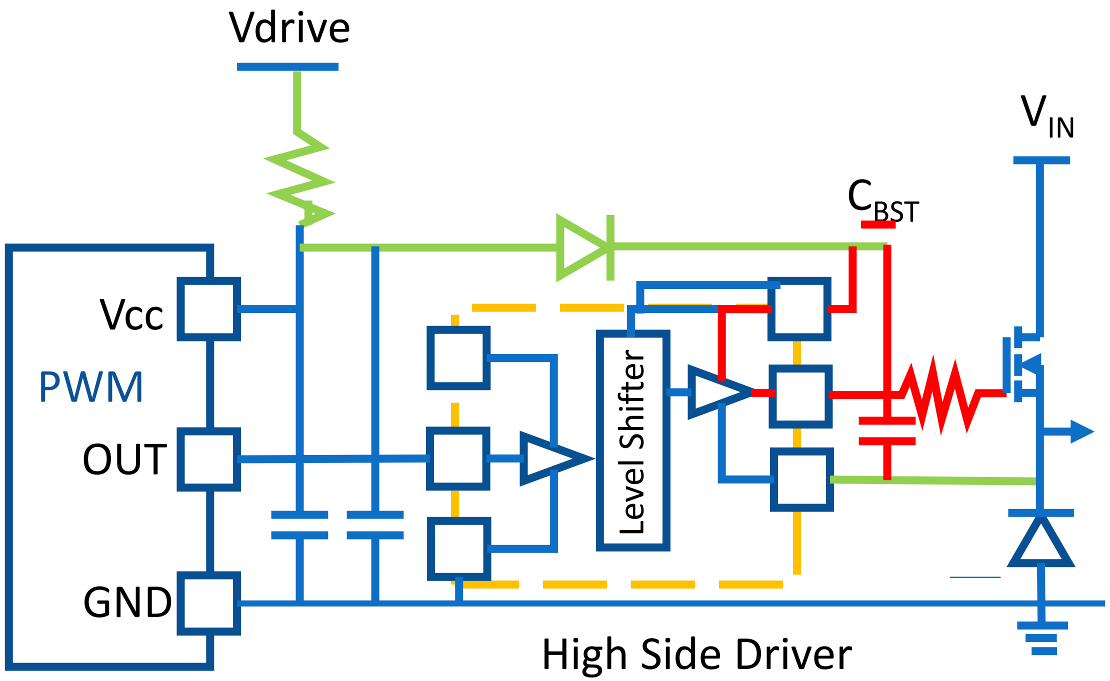


Less heavy

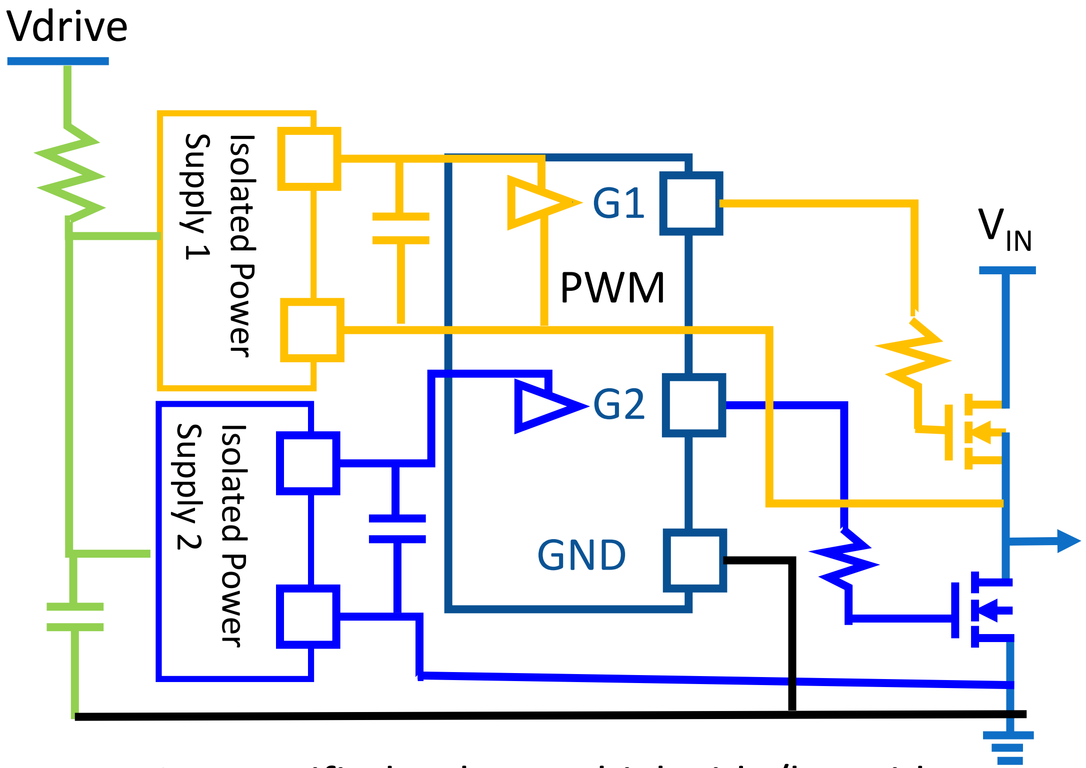
Distortion

1. Gate Driver

Traditional Bootstrap Gate Driver



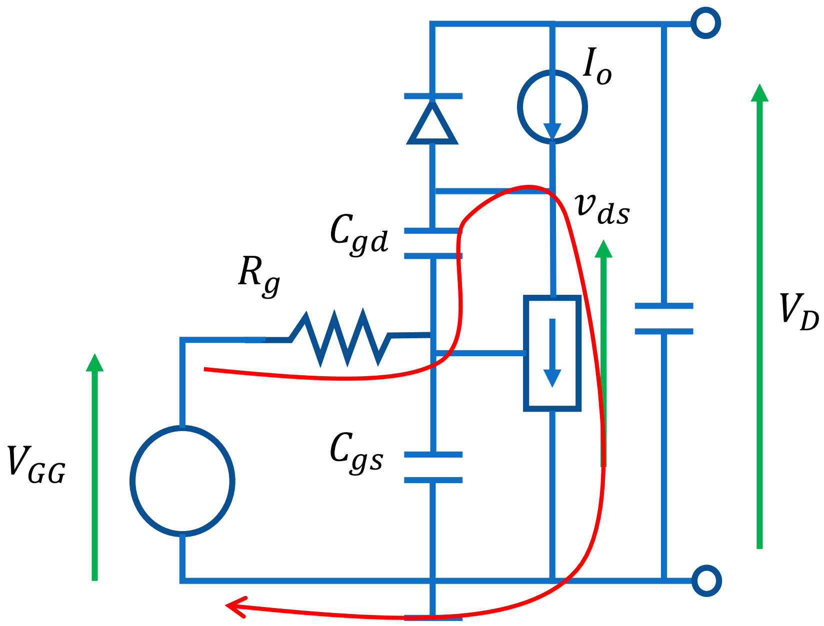
Isolated Power Supply Gate Driver



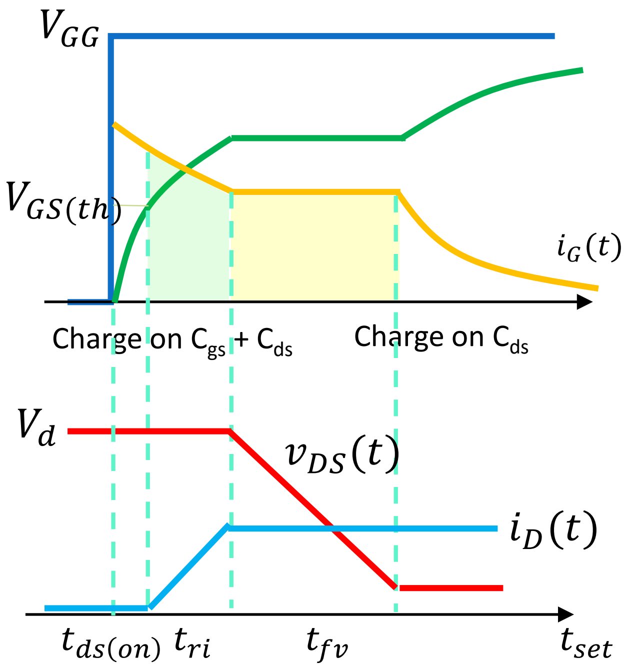
Not specified to have a high side/low side

2. Components Selection

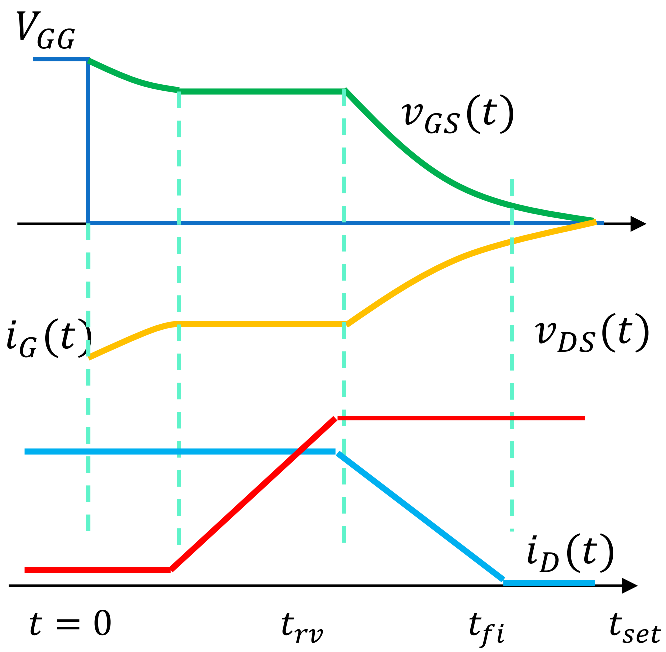
MOSFET Equivalent Circuit



Turn ON transient



Turn OFF transient

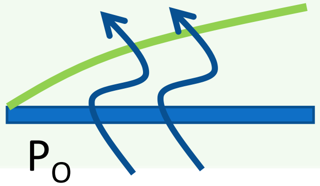


Implication: 1. v_{GS} (t) is similar to i_{DS}(t)

2. C_{gs} controls the speed of rise and fall, and switching loss;
There are also gate loss ($3/2\ CV^2$) and conduction loss

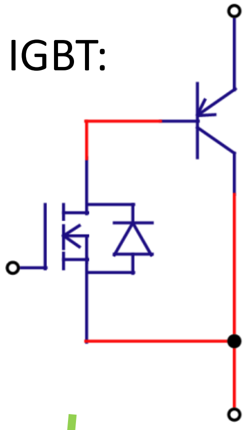
2. Components Selection

$Nu = 0.13(Gr_L Pr)^{1/3}$
$h = 0.453 Re_x^{1/2} Pr^{1/3}$



Switch Selection: MOSFET Vs IGBT: MOSFET for high frequency application
IGBT for high power application

IGBT:

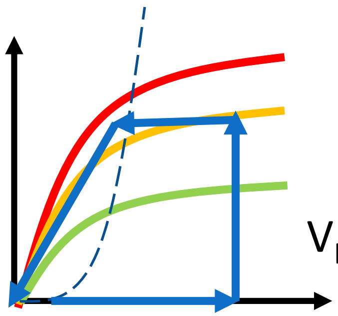


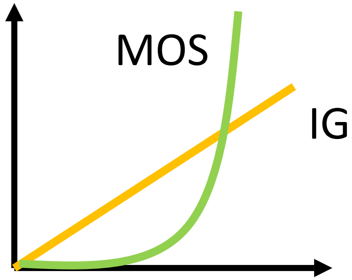
***Steady State**

Preference:
1. Voltage stress & Current stresses the switches takes

$V_{sw} = 60V$ (clamped)
$I_{sw} = 50A$ (designed max)

2. $R_{ds(on)}$ = 20mohm
   * Conduction loss = $i^2 R_{ds(on)}$
   * Lower when operated at Linear Region
     -> dependent on T_{cell} and V_{GS}
   * Parallel MOS

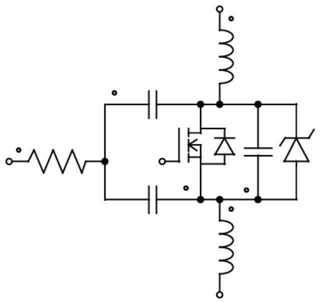


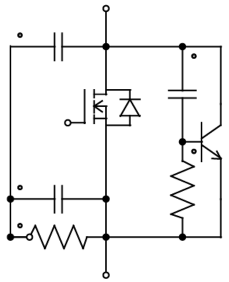


3. Speed to turn ON
   * Depends on gate-drive circuit
   * R_G and (C_{GS}+C_{DS}) for hard switching
   * Low impedance drive

-> Switching loss

4. Cost





5. dv/dt and di/dt
   * dv/dt: parasitic npn turns ON
   * di/dt: Wiring inductance -> overvoltage

6. Ability to damp Avalanche Energy (Avalanche Ruggedness)

2. Components Selection

Possible Solutions:
- Apply active rectification
- Use a GaN transistor with gate/source shorted.

Diode Selection:

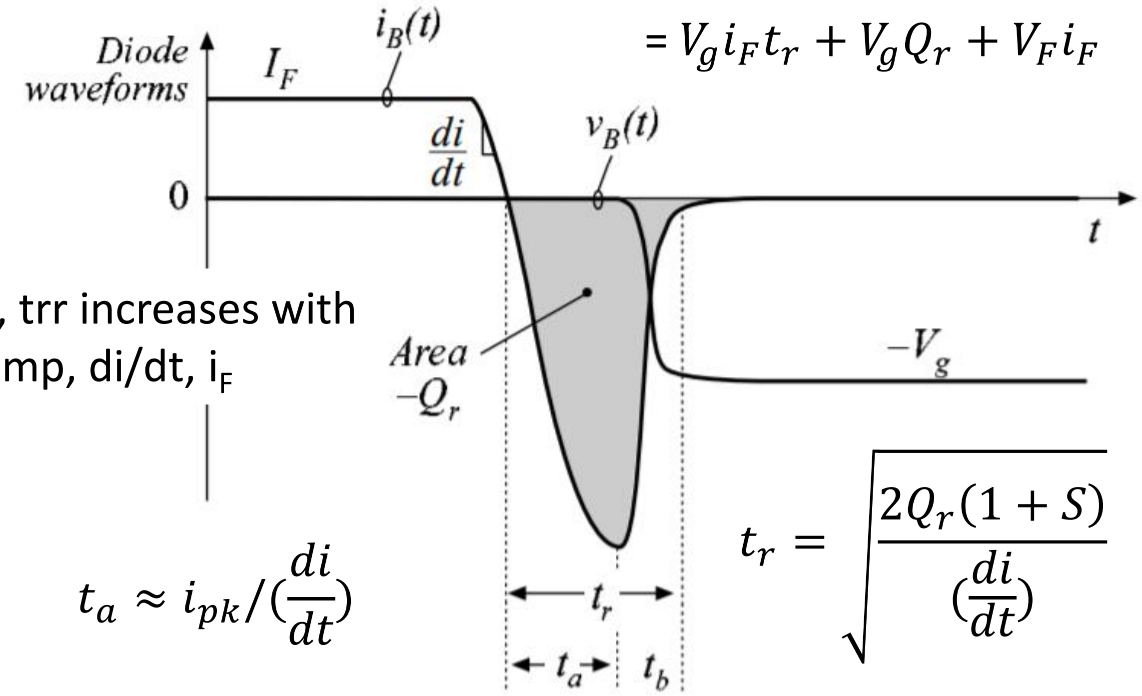
1. Voltage Stress & Current Stress

2. Conduction loss & Reverse Recovery Loss = $v_F i_F + Q_r V_g f_{sw}$

Consideration:

Si Diode, Schottky Diode (Fast Recovery),
SiC Diode (High power)
Optimization between **V_F and Q_{rr}**.

$= V_g i_F t_r + V_g Q_r + V_F i_F$

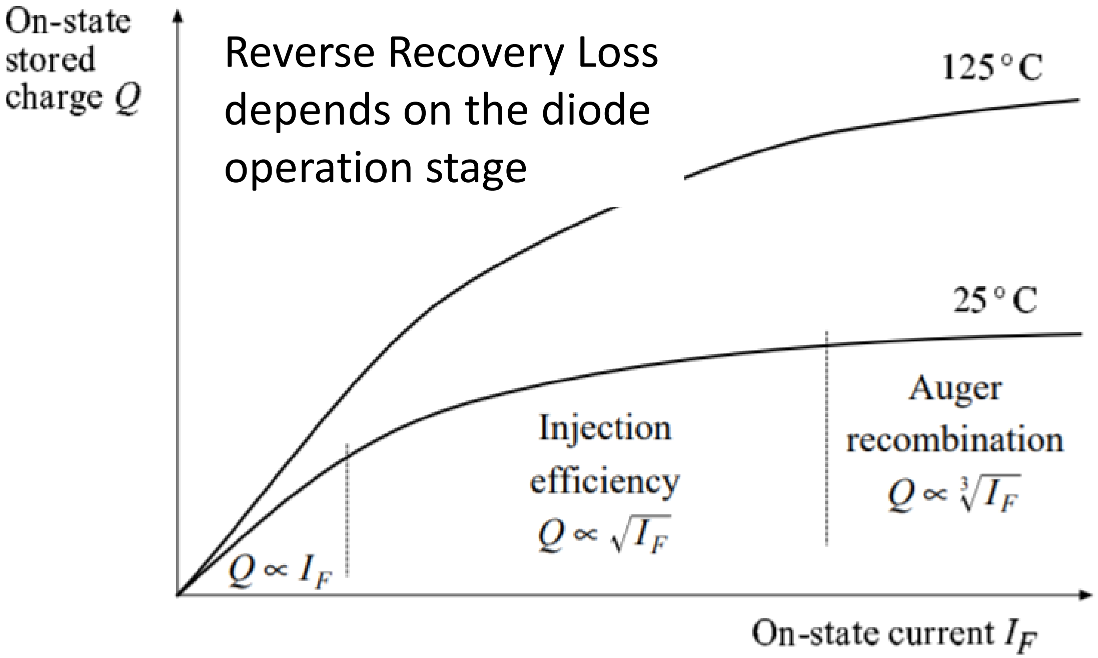


Ipk, trr increases with
- temp, di/dt, i_F

$t_a \approx i_{pk}/(\frac{di}{dt})$

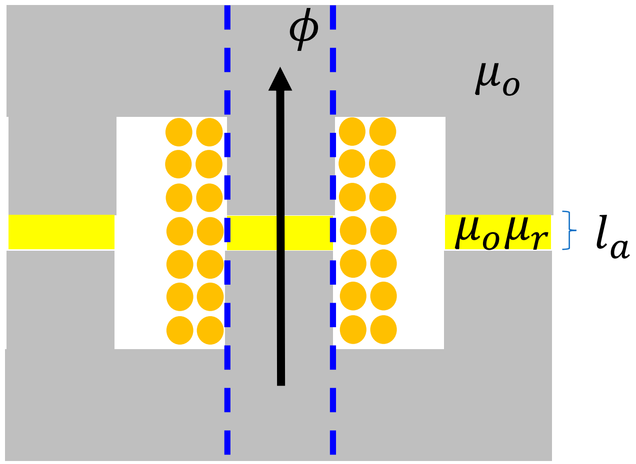
$t_r = \sqrt{\frac{2Q_r(1+S)}{(\frac{di}{dt})}}$

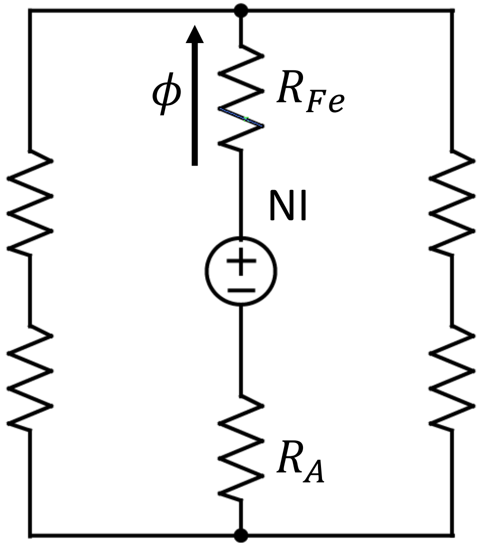
Reverse Recovery Loss depends on the diode operation stage



2. Components Selection

Magnetics and Inductance



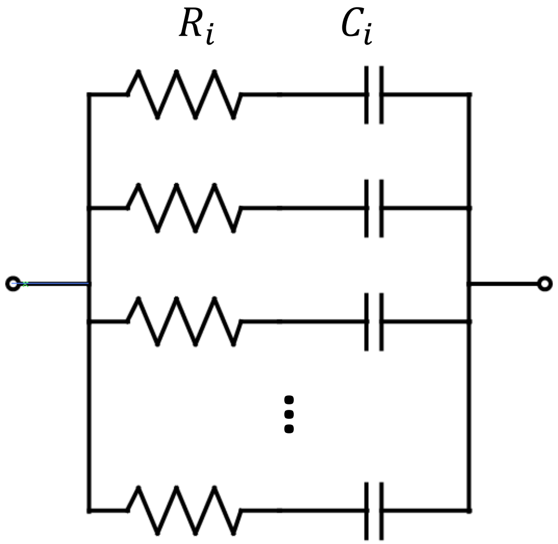


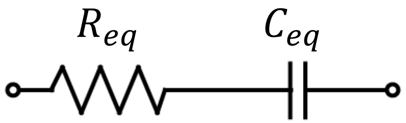
$NI = \phi\, R_{eq}$ $\quad R_i = \dfrac{l_i}{\mu_i A_i}$ $\quad \phi = \boldsymbol{L}i$

$$V_L = \boldsymbol{N}\frac{d\left(\boldsymbol{N}i\frac{\boldsymbol{\mu}A}{l}\right)}{dt}$$

Control: # of turns, air gap
magnetic coupling

Capacitance and Equivalent Series Resistance (ESR)





$$\left(R_{eq} + \frac{1}{j\omega C_{eq}}\right)^{-1} = \sum_{i=1}^{N}\left(R_i + \frac{1}{j\omega C_i}\right)^{-1}$$

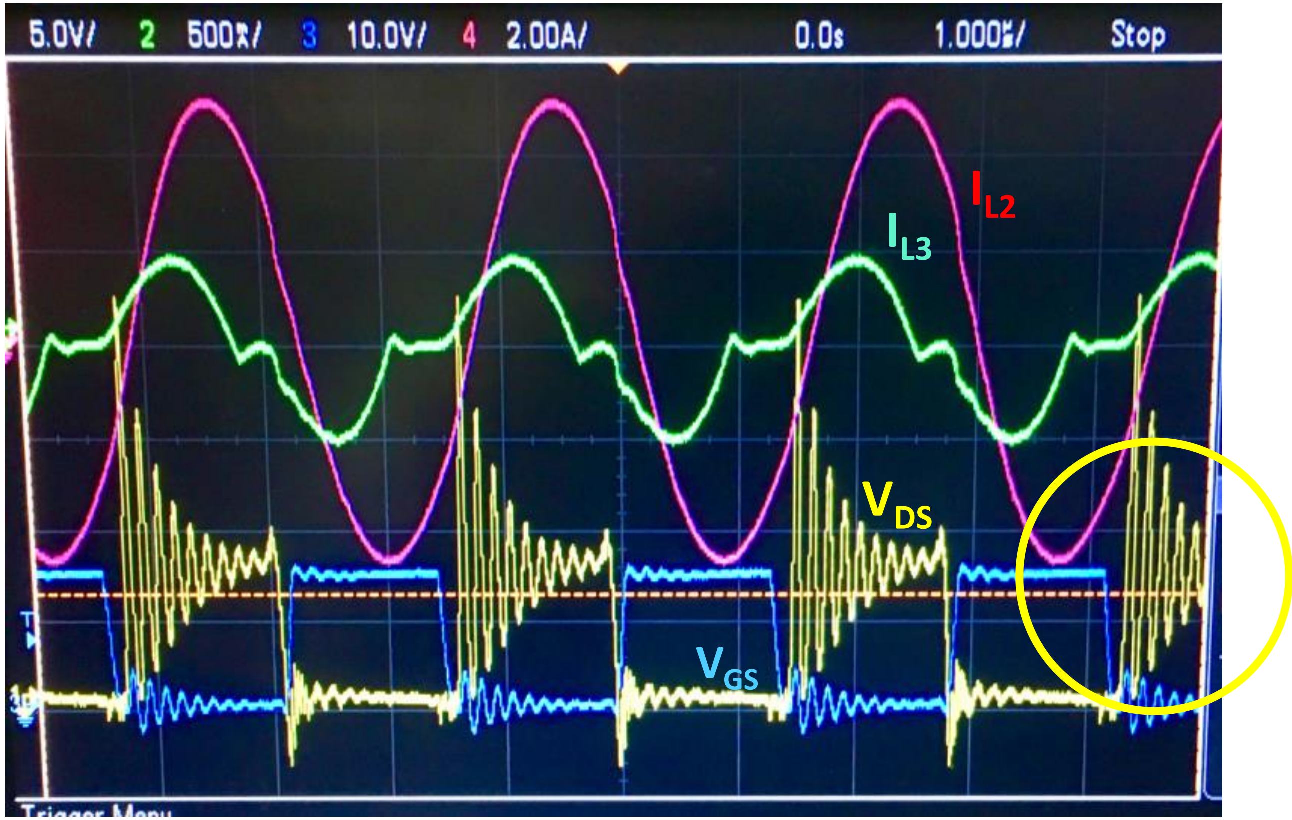
2. Component Selection

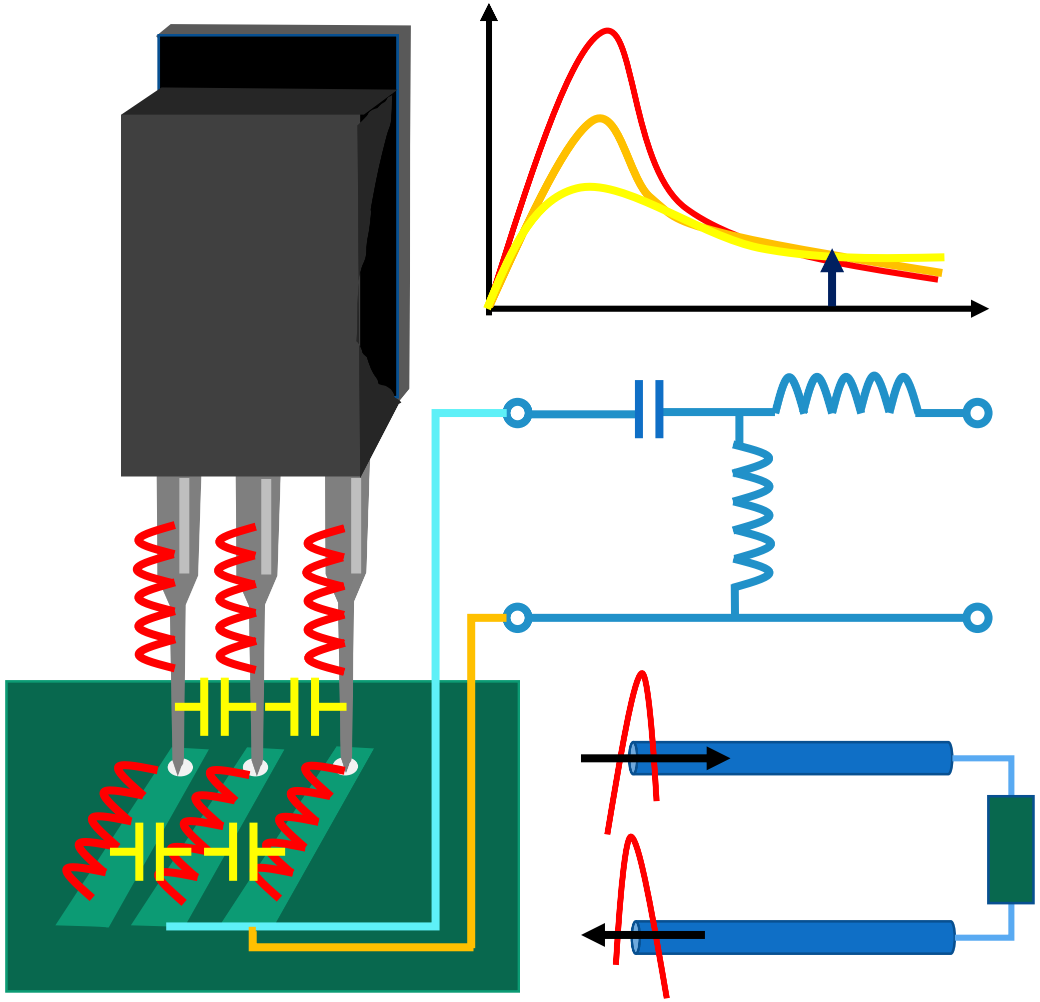
Utilized Components and Parameters of Prototype

Components	Parameters
IXFH96N15P N-Channel MOSFET, HiperFET	96A, 80V
C_{IN} EPCOS Polyester Capacitor PET B32529	1uF, 63Vdc
C_{IN} Nichicon Aluminium Electrolytic Capacitor	1mF, 50Vdc
$C_{Clamped}$ Nichicon Aluminium Electrolytic Capacitor	18uF, 100V dc
C_{CWM2} Panasonic Aluminium Electrolytic Capacitor	10uF, 450V dc
C_{TANK} EPCOS Polyester Capacitor PET B81141	100nF, 440V ac
Diode(CWM) C3D02060F Wolfspeed Diode	600V 4A
C_{CWM1} Panasonic Aluminium Electrolytic Capacitor	10uF, 250V dc
Wurth Ferrite Toroidal Inductor	26uH, 10.3A dc, 15mohm
Tamura Ferrite Coil Inductor	50uH, 20A dc, 7mohm
Wurth Manganese Zinc Ferrite Power Inductor	10uH, 36A dc, 1.31mohm

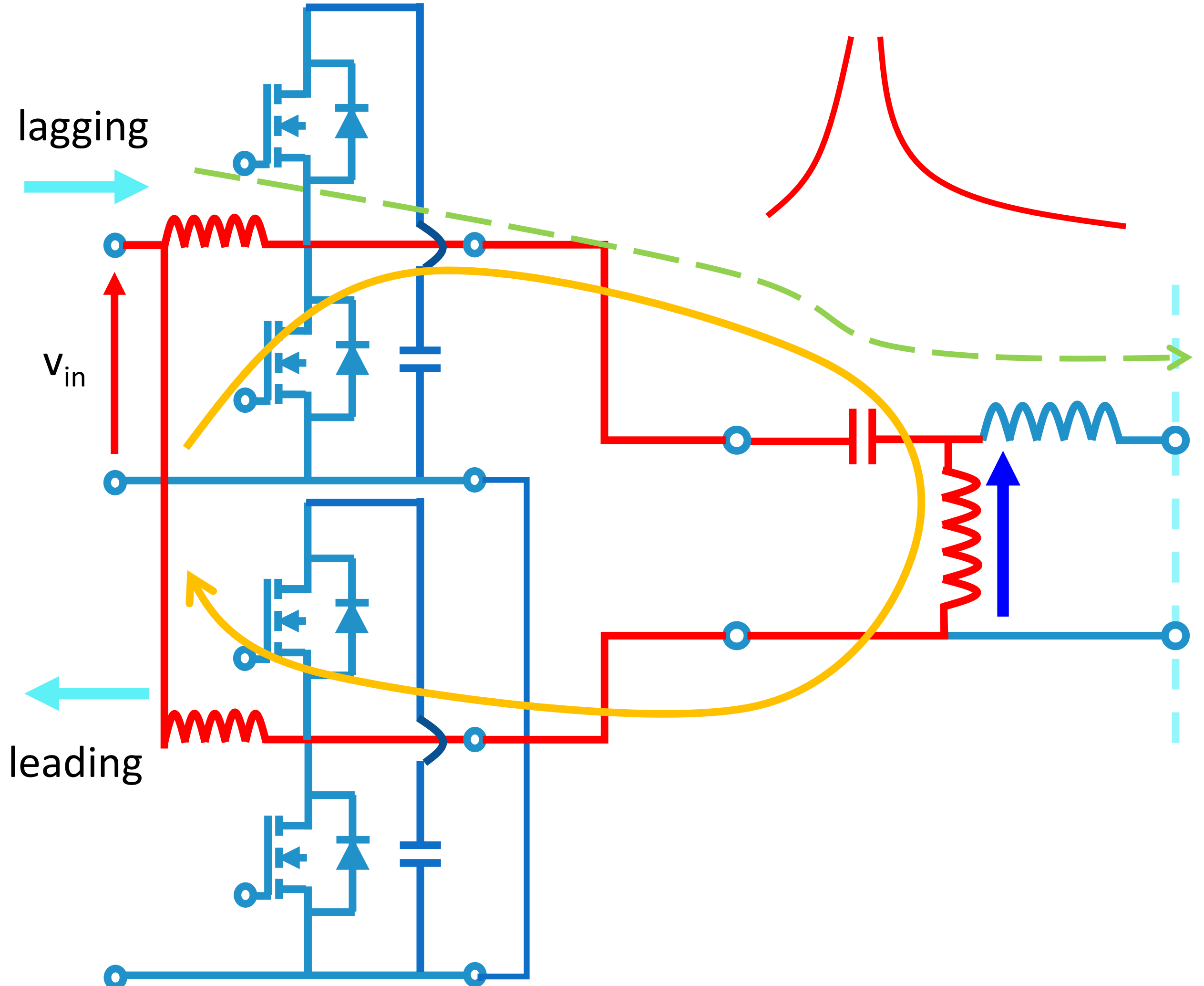
3. Ringing Effect

1. Parasitic Capacitance & Inductance
2. Cutoff frequency of resonant tank
3. Transmission lines reflection



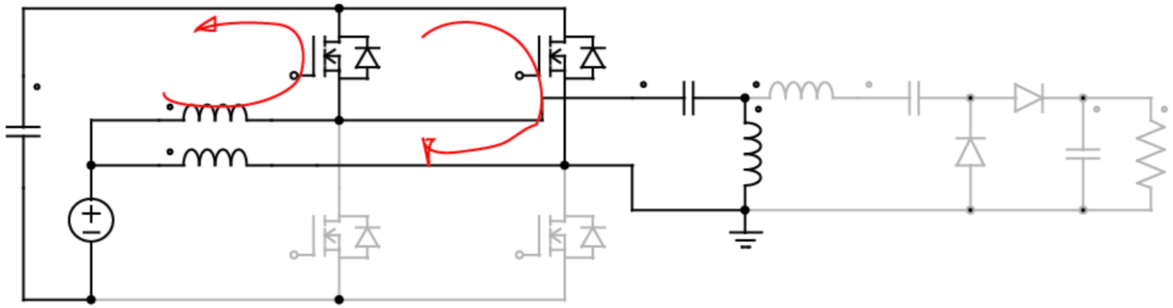


4. Circulating Current

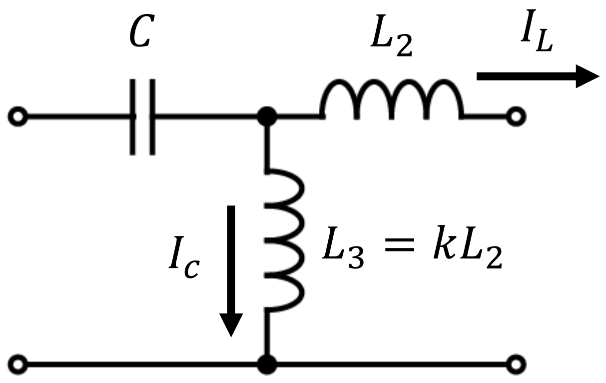


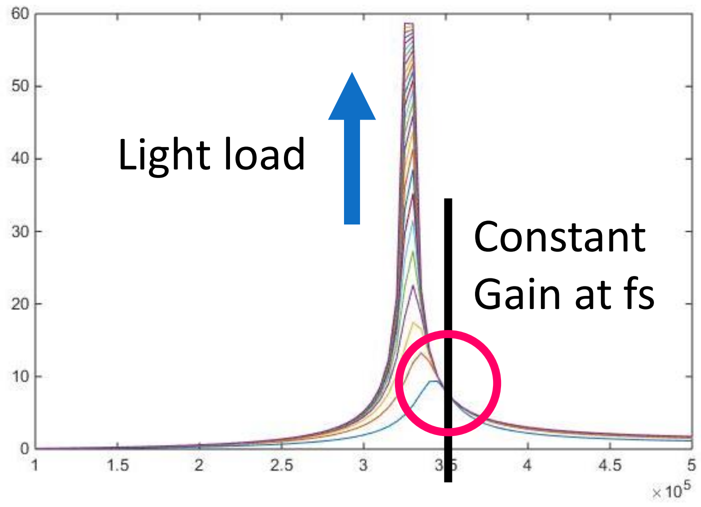
As seen in the current flow diagram,
Current circulation in the source side leads to high voltage.

Yet, in this path, most of the components are inductance, and inherited resistance. The efficiency could be low with **high copper loss**.



5. Revisit of T-Tank Design





Consider the resonant tank with $L_3 = kL_2$

$$G = \left|\frac{U_{OUT}}{U_{IN}}\right| = \left|\frac{1}{1-\frac{1}{\omega_s^2 C L_3}}\right| \qquad \omega_s = \frac{1}{L_2 C}\sqrt{1+k} = 2\pi f_s$$

$$= \frac{k+1}{k+1-L_2 C} \geq 1$$

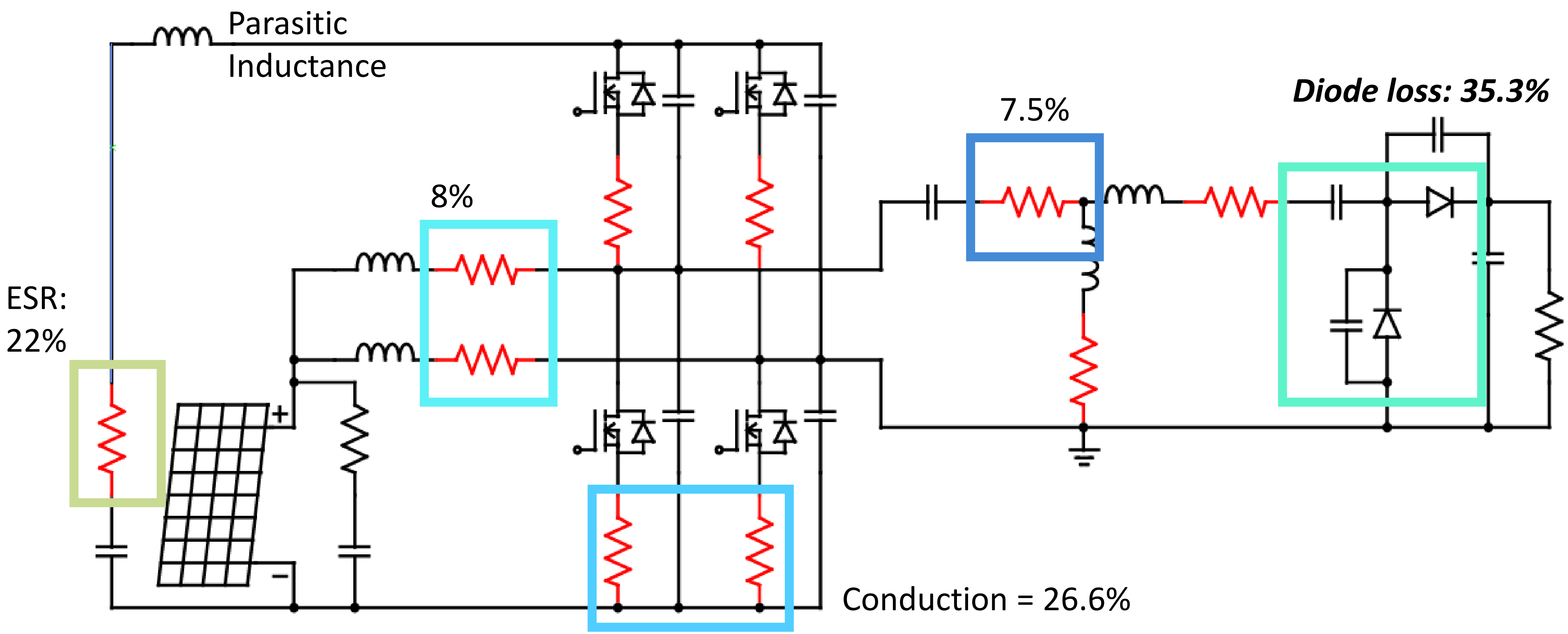
→ again, L_2 controls the gain
→ Inherently, the gain is larger than 1.

$I_L > I_C$: $\quad j\omega L_3 > j\omega L_2 + R_L$ → k must be larger than 1 ($L_3 > L_2$)

Physical meaning: Design a low gain tank.

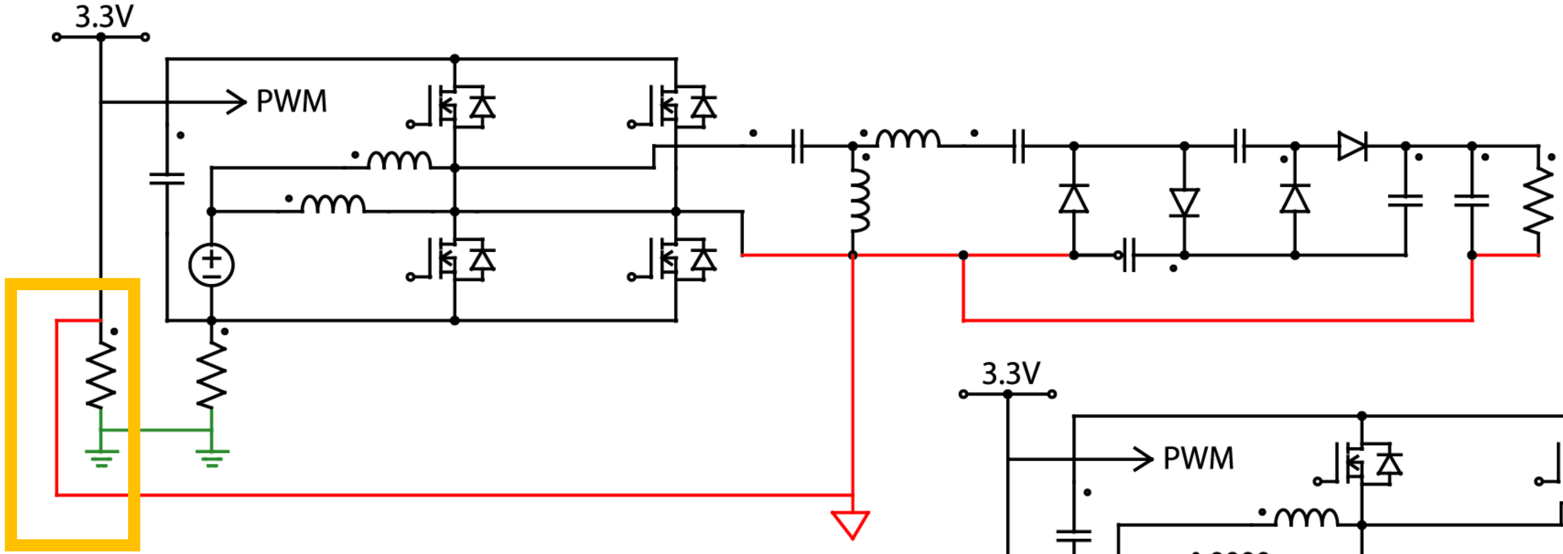
6. Loss Analysis

The losses are distributed as follows.



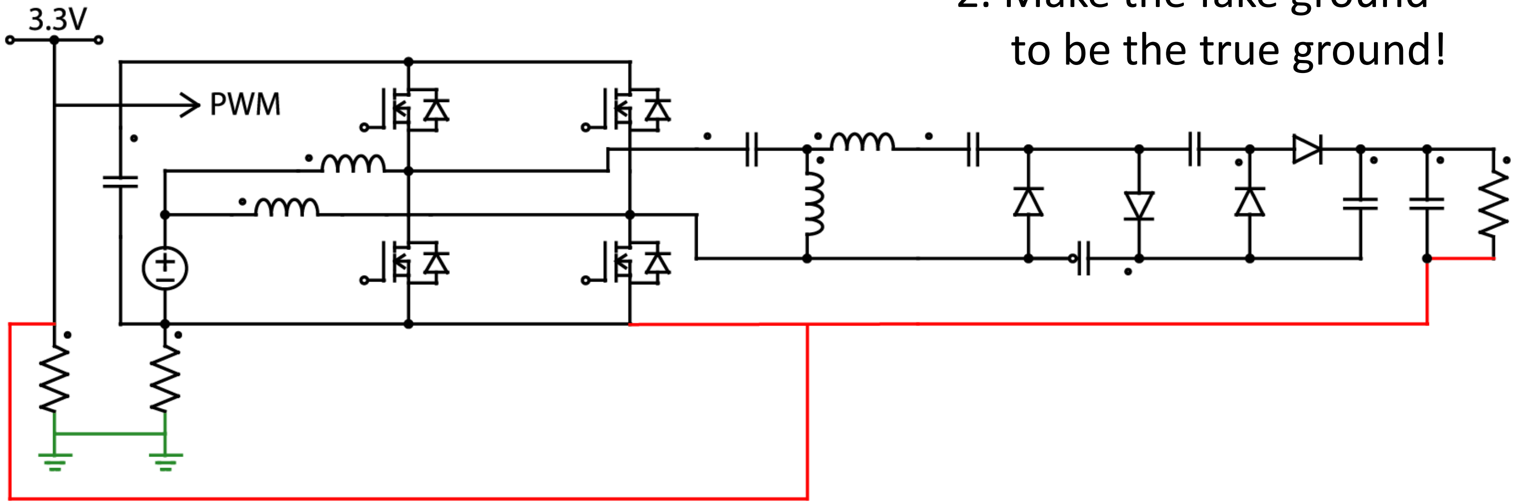
7. Grounding

A 3.3V obtained from oscilloscope and a 30V obtained from DC source are used to drive the converter.
Yet, due to short circuiting true ground of 3.3V to the fake ground in converter output, the PWM signal is interfered.



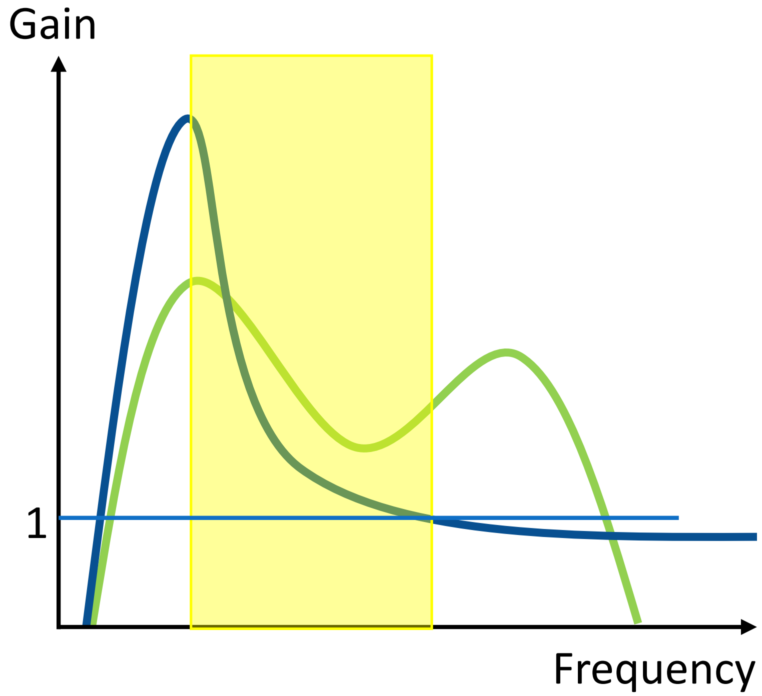
1. Disconnect the power source of controller from ground, or use high impedance grounding

2. Make the fake ground to be the true ground!



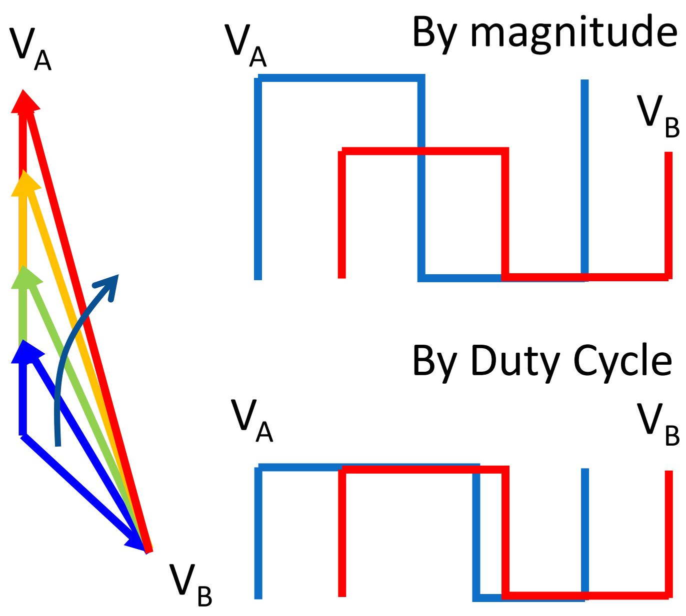
8. Possible Control Methodologies

Frequency Control



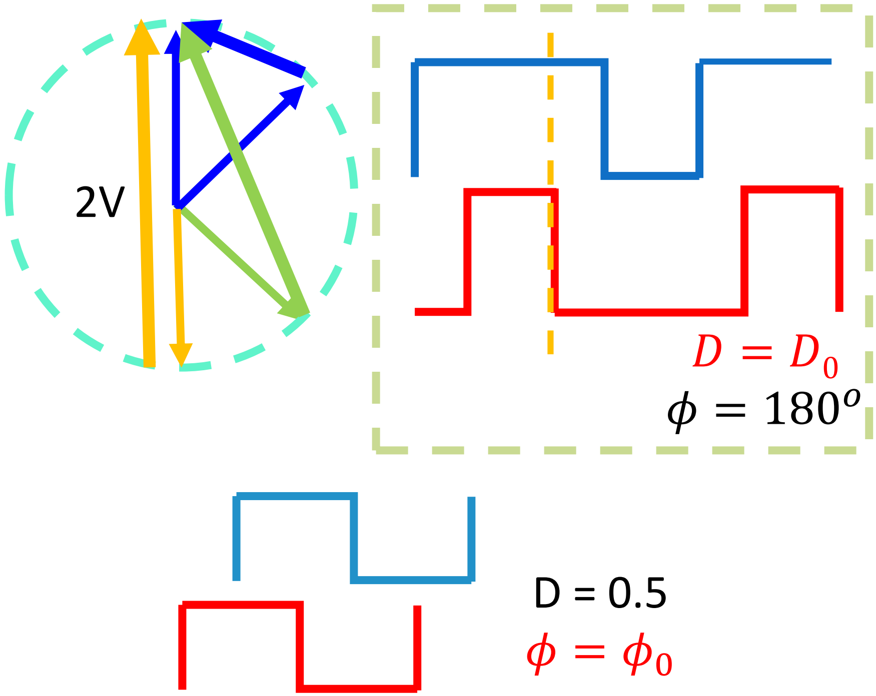
- Require Search Algorithm
- Varying Component Parameters
- ZVS/ZCS may be affected

Voltage Control



- Need to take care of CCM/DCM
- High harmonic pollutant
- May need extra components

Phase-Shift Control

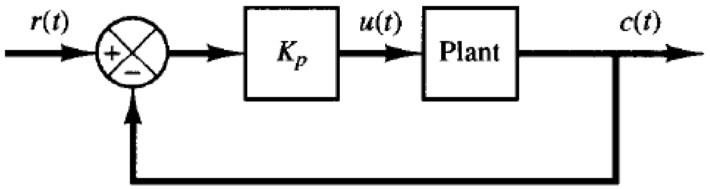


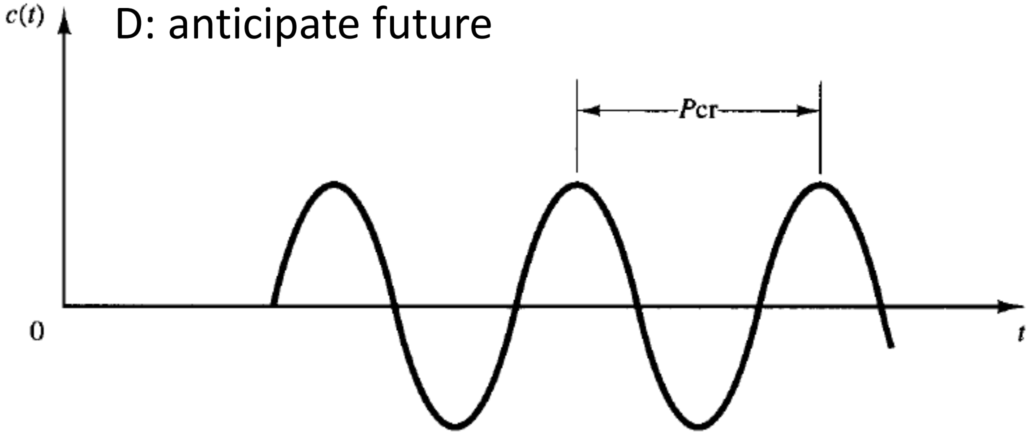
- More stable to control voltage gain (0-2)
- Not affect ZVS/ZCS

8. Control Scheme, a general pic

Tuning: N-Z Method

P: speed/overshoot
I: accumulate past
D: anticipate future

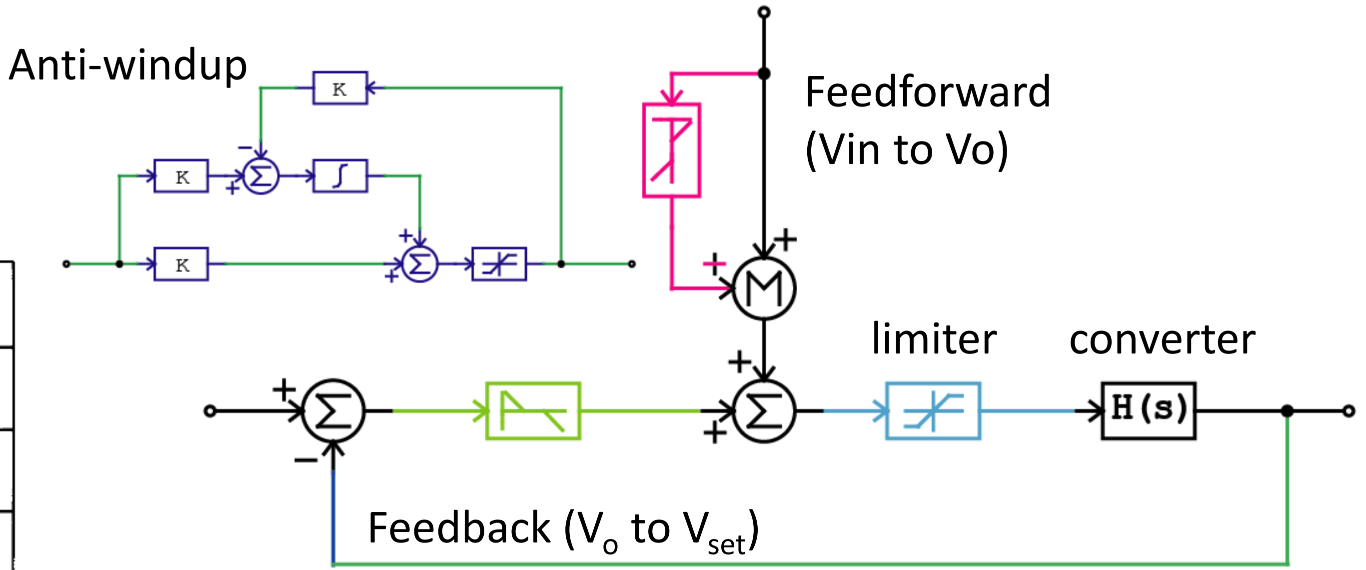




Type of Controller	K_p	T_i	T_d
P	$0.5K_{cr}$	∞	0
PI	$0.45K_{cr}$	$\frac{1}{1.2}P_{cr}$	0
PID	$0.6K_{cr}$	$0.5P_{cr}$	$0.125P_{cr}$

Rule-based Tuning:

CL RESPONSE	RISE TIME	OVERSHOOT	SETTLING TIME	S-S ERROR
Kp	Decrease	Increase	Small Change	Decrease
Ki	Decrease	Increase	Increase	Decrease
Kd	Small Change	Decrease	Decrease	No Change



8. Control Scheme

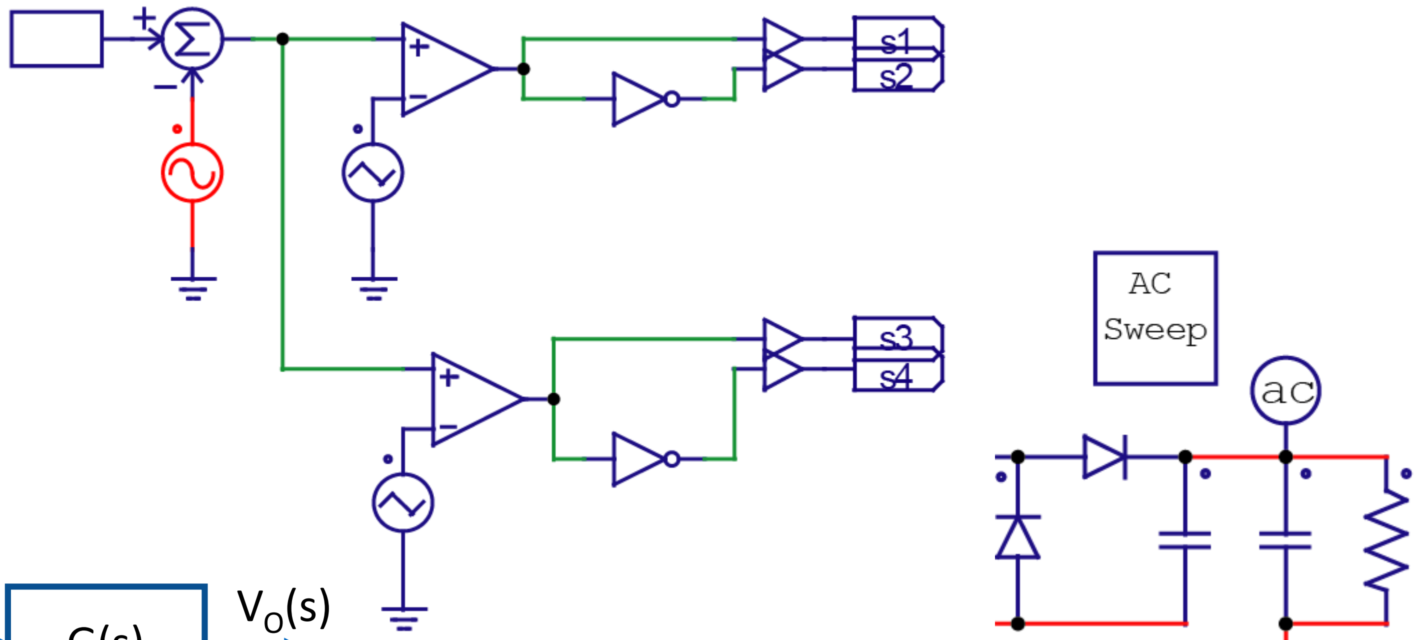
[1] A.K.S. Bhat, A Generalized Steady-State Analysis of Resonant Converters using Two port model and Fourier-Series Approach, IEEE Transactions on Power Electronics, Vol. 13, No. 1, 1998
[2] Y. Zhang & P.C. Sen, D-Q Models for Resonant Converters, 2004 35th IEEE Power Electronics Specialist Conference

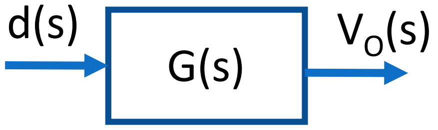
To provide output voltage regulation, the converter should be in **closed-loop control**.
Step 1: Know the converter transfer function.

$$GSSA: \dot{x} = \left(\sum_{i=1}^{k} d_i A_i\right) x + \frac{1}{T}\sum_{i=1}^{k} \int_{t_{i-1}}^{t_i} B_i(\lambda)d\lambda$$

As the converter is non-linear, *Generalized State-Space Averaging*[1], Or *D-Q Model*[2], the transfer function can be obtained. (Off-scope)

To simplify the process, **numerical method** is used.





converter

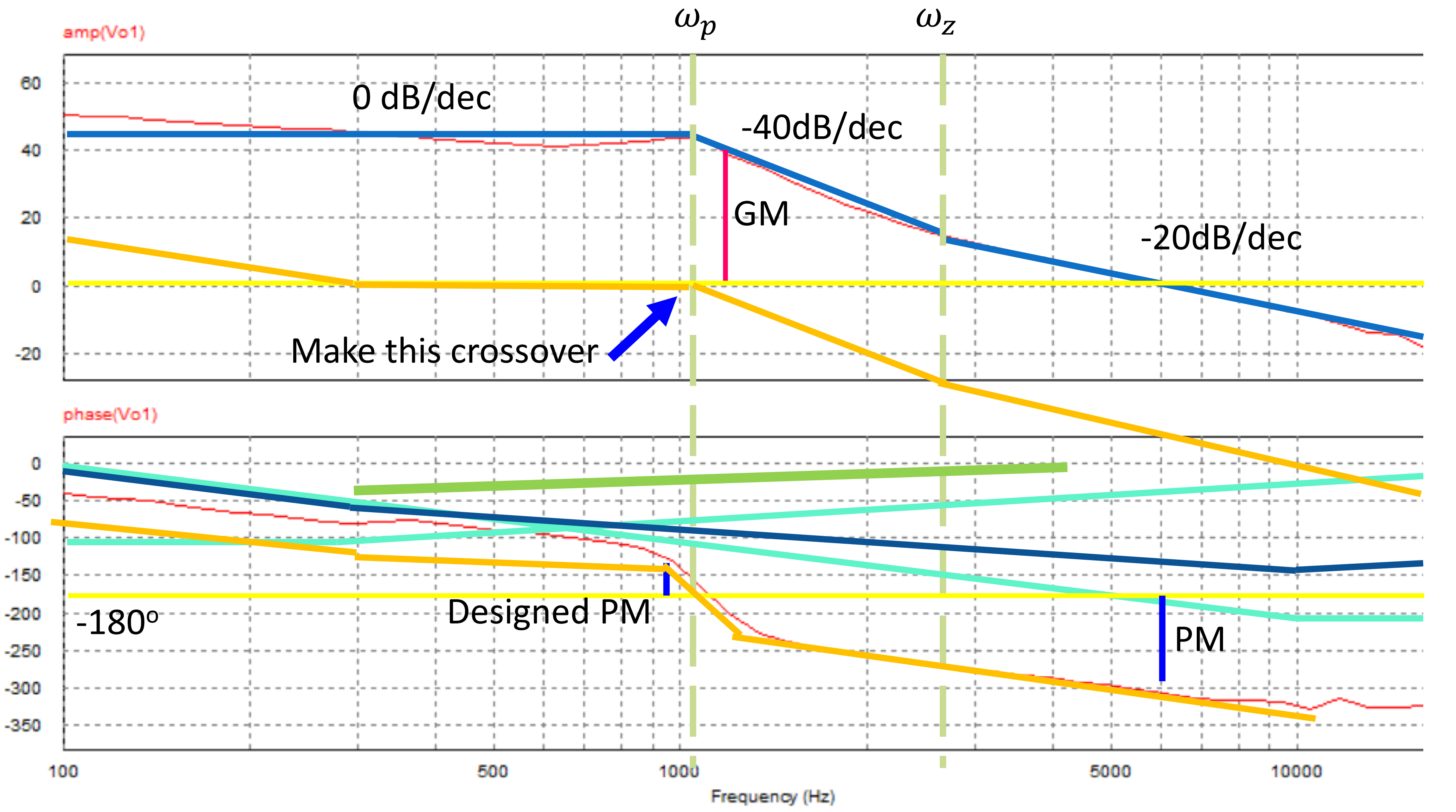
8. Control Scheme

The control function is approximately:

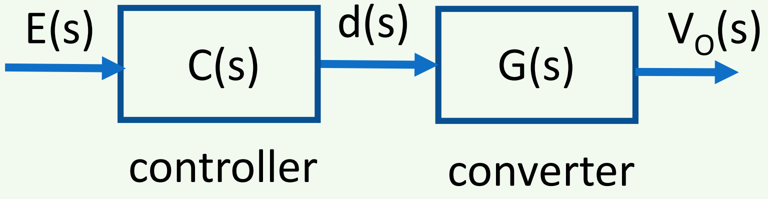
$$G(s) = k\frac{s+\omega_z}{s^2+2\zeta\omega_p s+\omega_p^2}$$

The goal:
1. High gain at low frequency
2. Low gain at high frequency
3. Cutoff frequency
4. Enough Phase Margin

G(s) depends on:
- input (CCM/DCM mode)
- Load*



8. Control Scheme



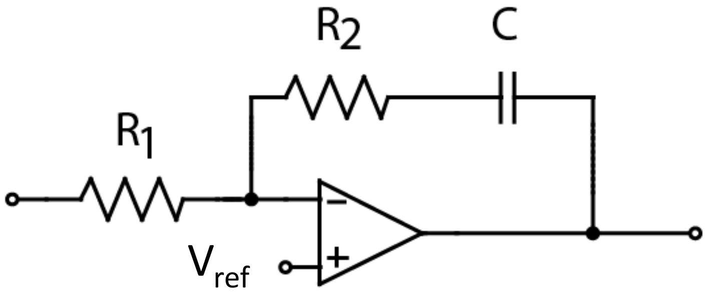
Step 2: Add in the controller

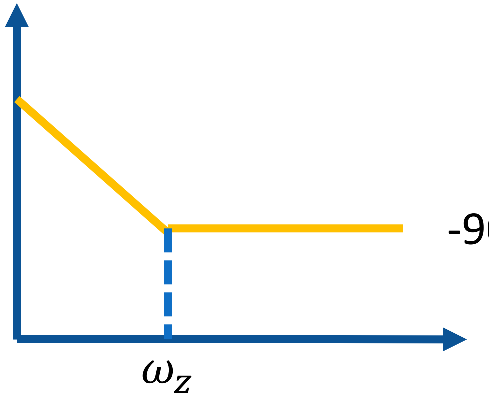
Gain Plot

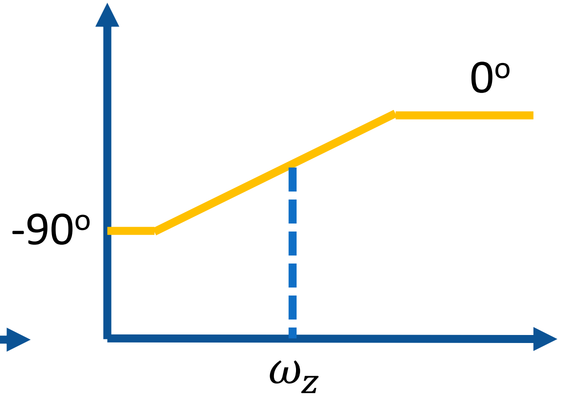
Phase Plot

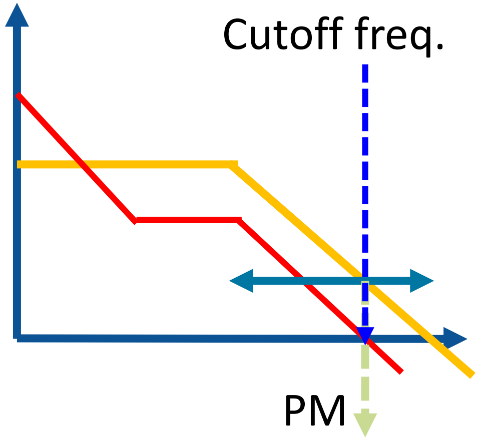
Updated Gain

PI controller

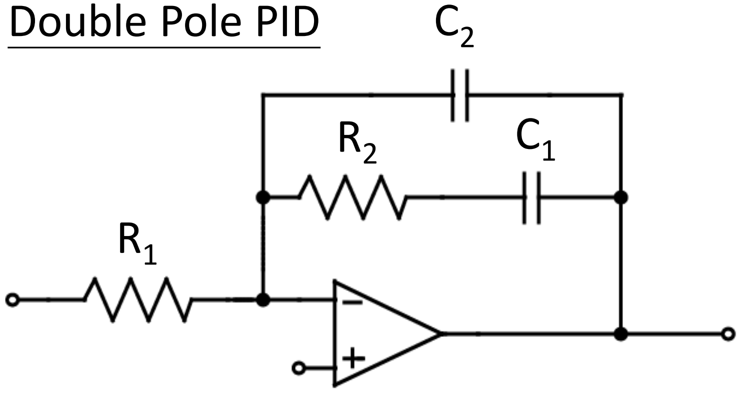


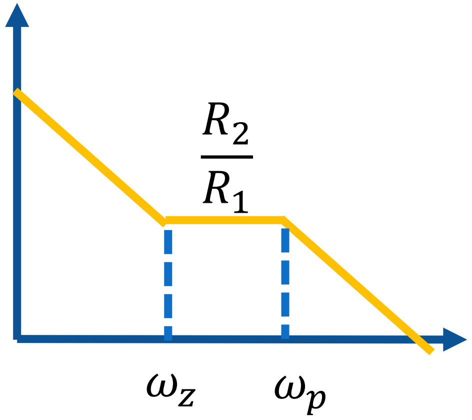


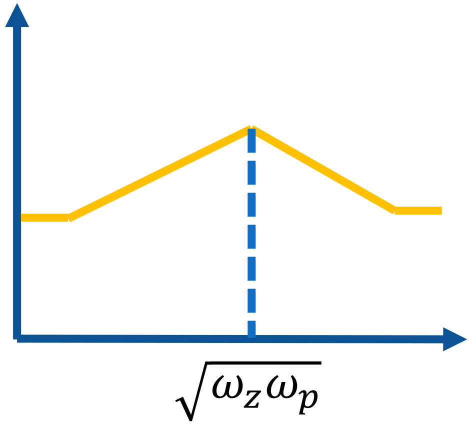


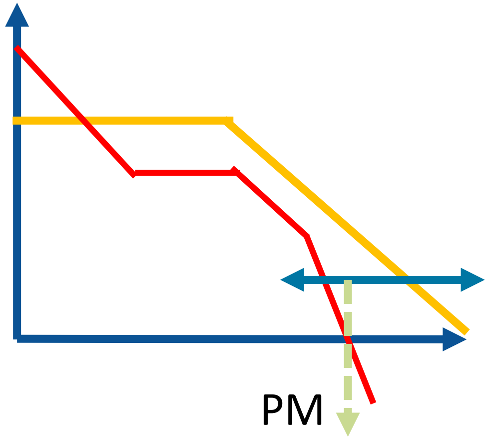


Double Pole PID









8. Control Scheme

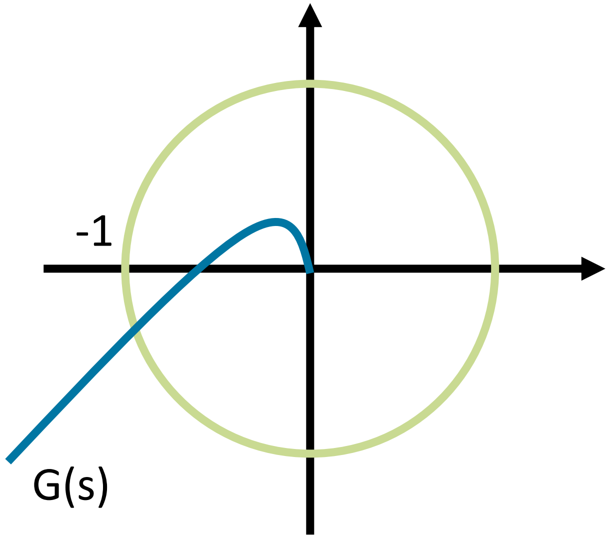
(f) Virtual Inertia

$$J\frac{d\omega}{dt} = T_m - T_e - D(\omega - \omega_0)$$

$$\mathfrak{I}\{f(t-t_0)\} = e^{-j\omega t_0}F(s) \longrightarrow$$

Time delay = phase shift in frequency domain

Nyquist Stability Criterion

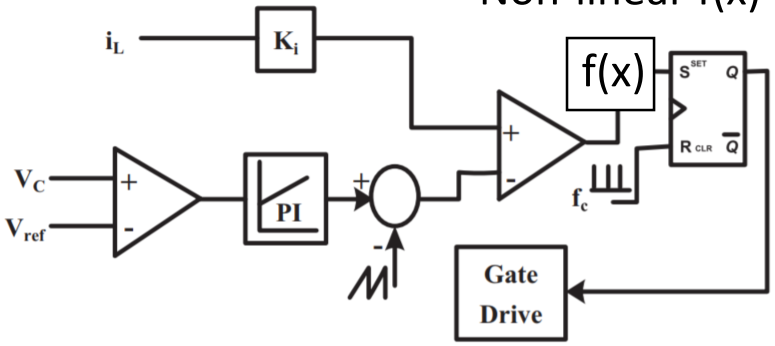


$\boxed{Z = P + N}$

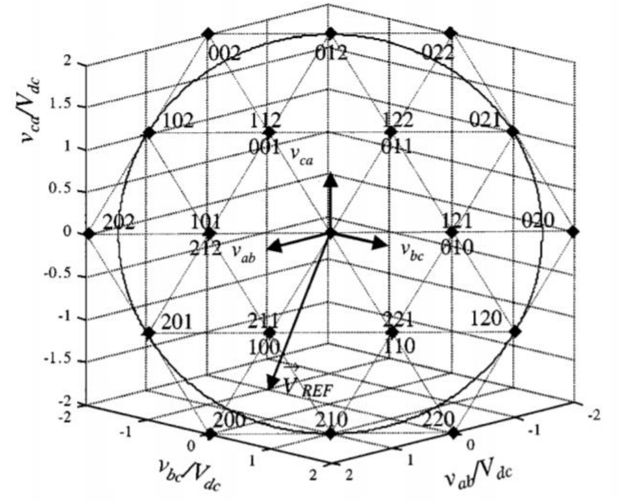
Number of closed-loop unstable pole = number of open loop unstable pole + encirclement to (-1, j0)

Other Possible Control Components

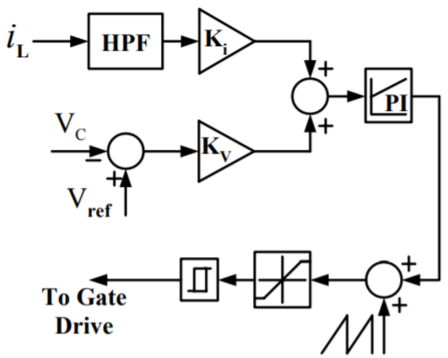
Non-linear f(x)



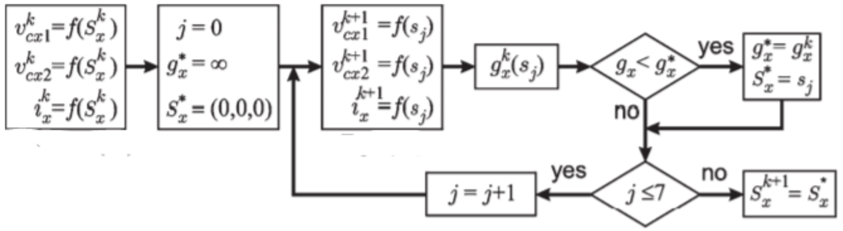
(a) Classic Current Mode Control



(b) Space Vector Modulation



(c) Sliding Mode/ Hysteresis Control



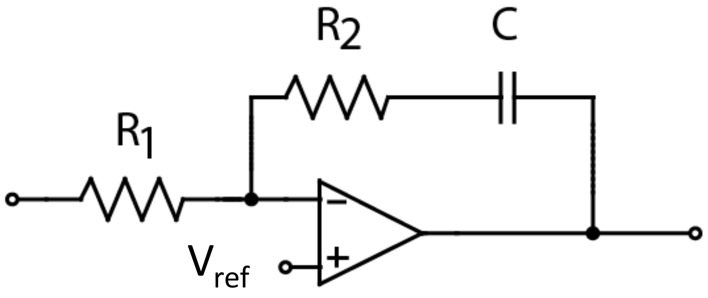
(d) Adaptive Predictive Control

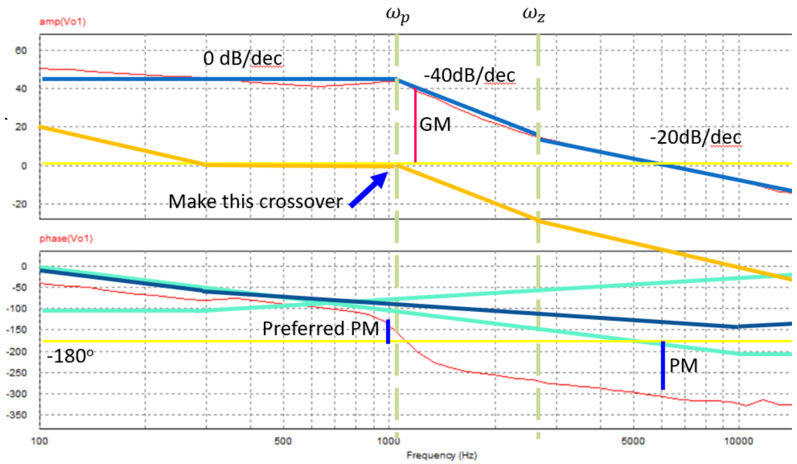
(e) Deadbeat Control/ Pole Placement

8. Control Scheme

Step 3: Find the components value/ Transfer back to z-domain

PI controller





Select PI controller for its simplicity.

$$C(s) = -\frac{R_2}{R_1}\frac{s+\omega_z}{s} = -(K_p + \frac{K_I}{s}) \qquad K_p = \frac{R_2}{R_1} \qquad K_I = \frac{R_2}{R_1}\omega_z = \frac{1}{R_1 C}$$

Transferring to z-domain, it is either by $G(z) = (1-z^{-1})Z\left\{\frac{G(s)}{s}\right\}$

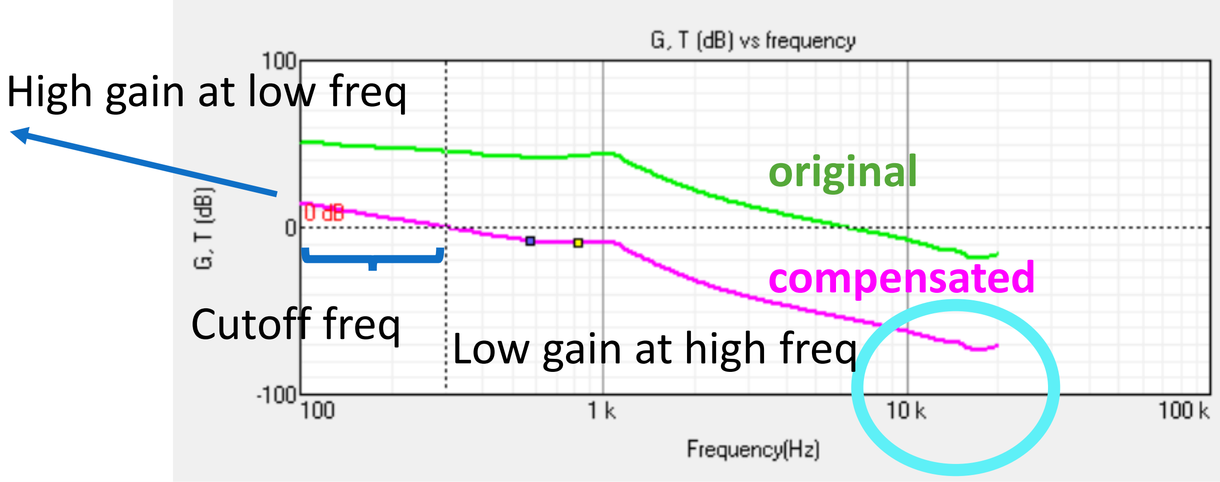
or simply by $K_{P(z)} = K_{P(s)}$, $K_{I(z)} = K_{I(s)}T$ where T is the sampling time.

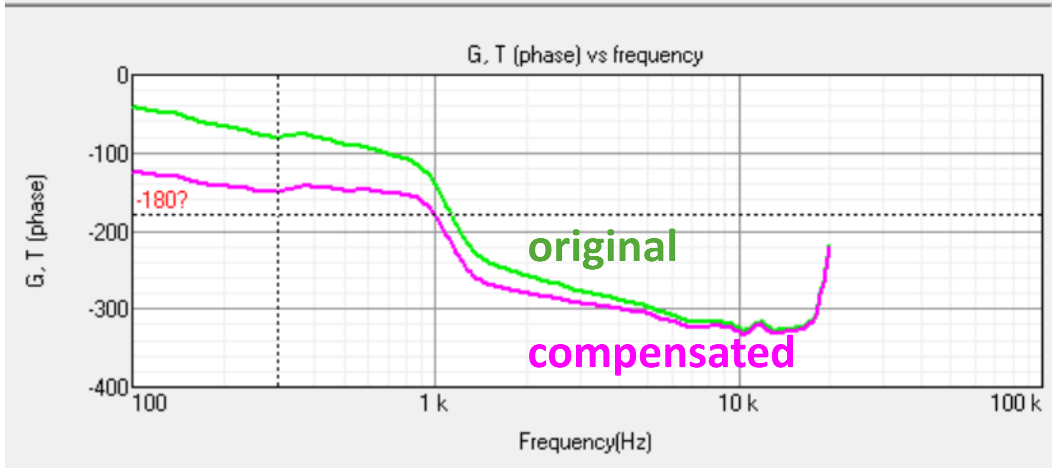
From the graph, we put the zero at f_Z = 300Hz and the constant gain has to compensate for the converter and controller gain at $\omega = 1$ rad/s, i.e. 85dB.

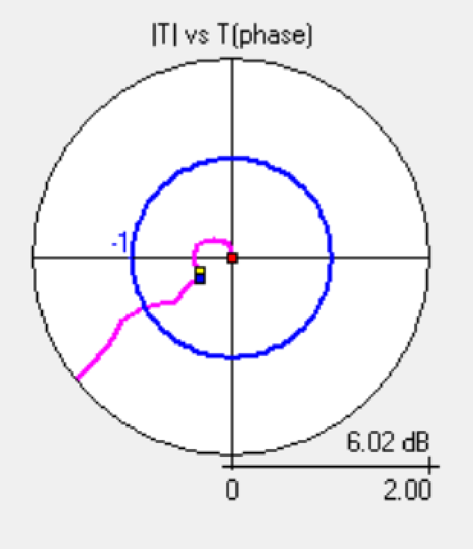
Hence, $K_P = 10^{-\frac{85}{20}} = \frac{R_2}{R_1}$ $\qquad K_{I(z)} = 300(2\pi)T = \frac{T}{R_1 C}$

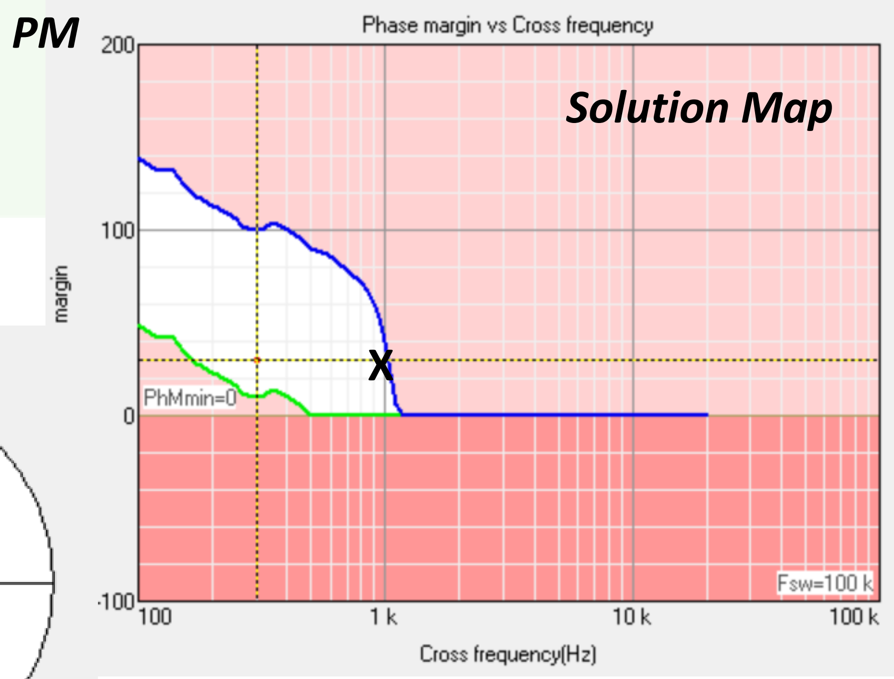
8. Control Scheme

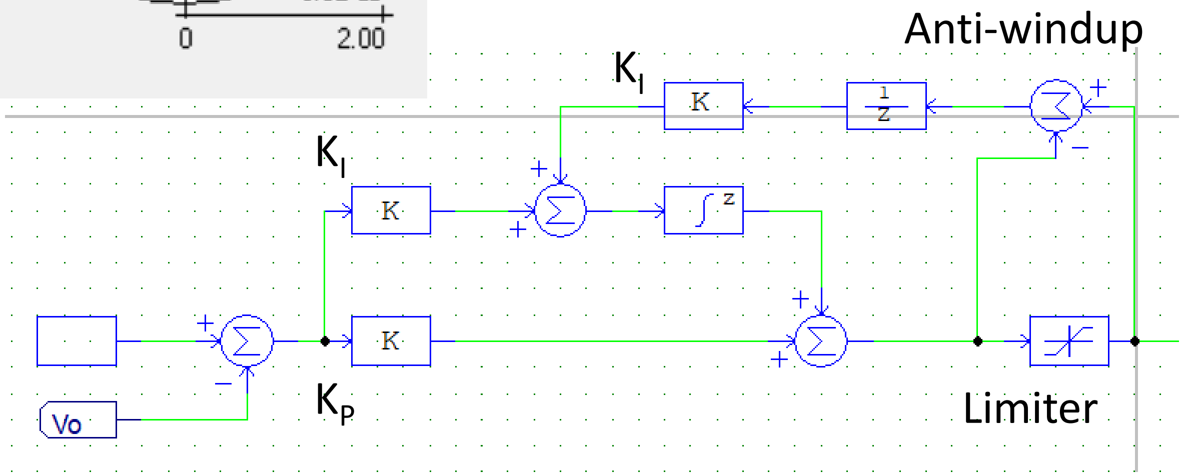
Yet the controller can be designed directly using SmartCtrl @ PSIM.



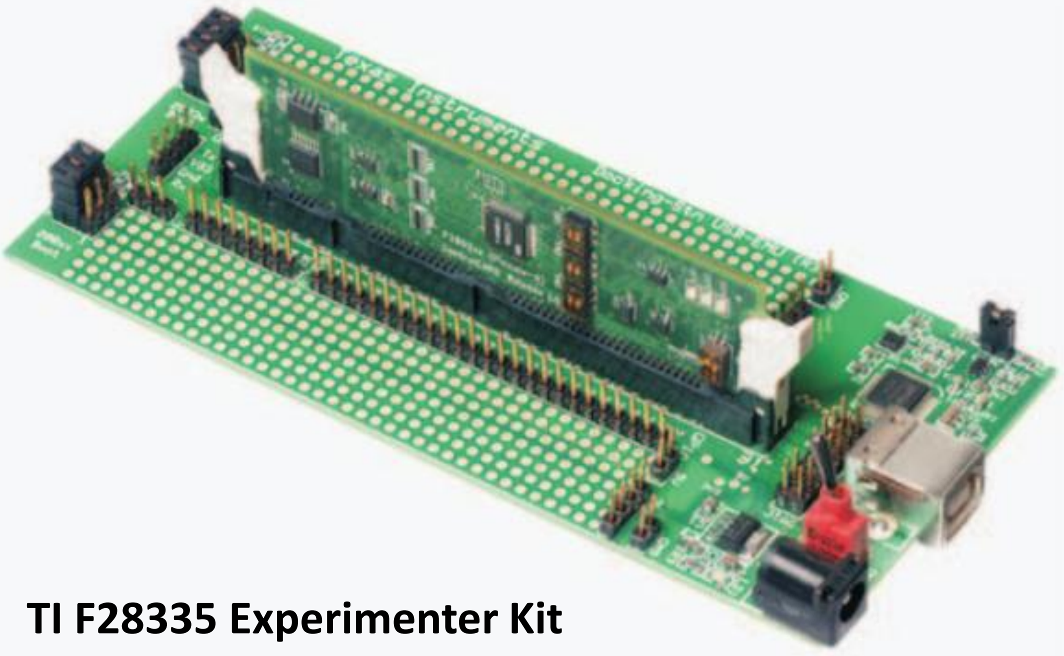








8. Control Scheme (Controller used)



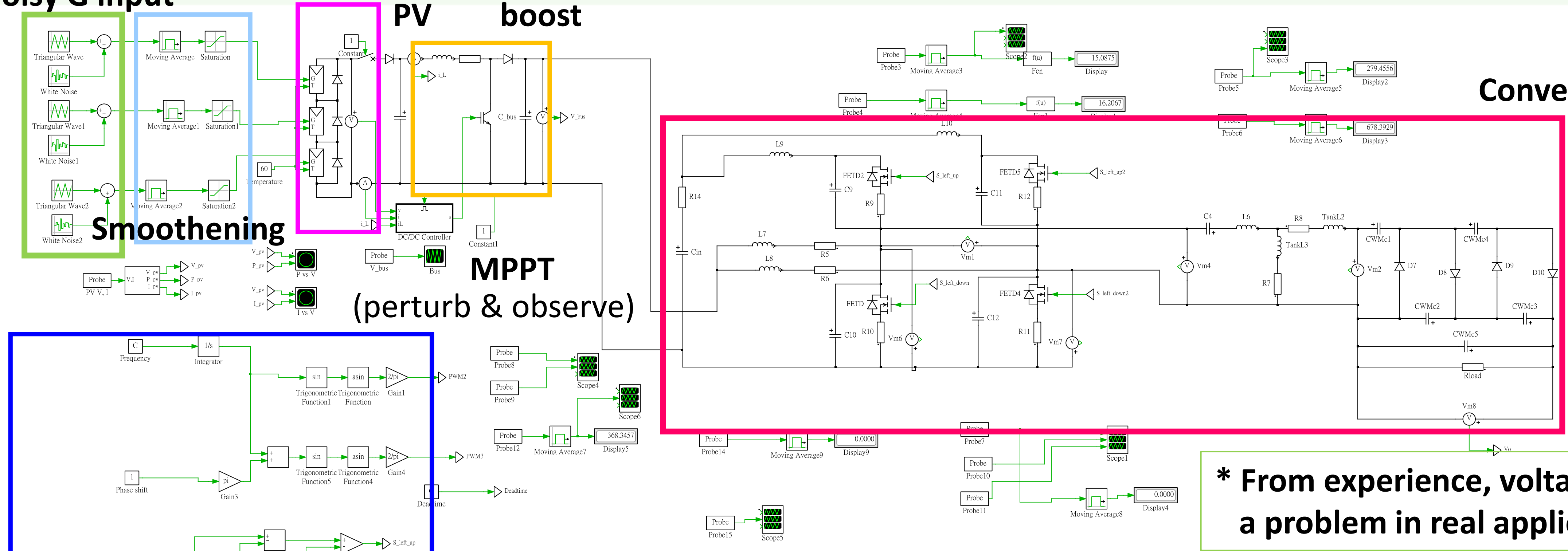
TI F28335 Experimenter Kit

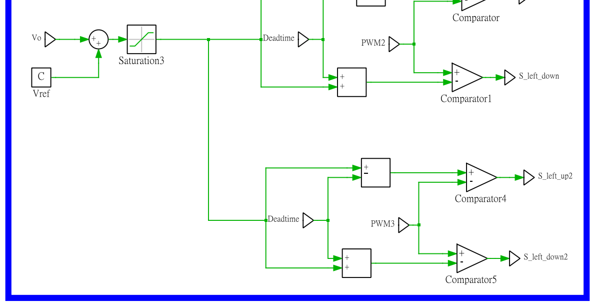
1. Digital Controller:
   - can have pre-processing in data (e.g. smoothening)
   - can have complex control loop (e.g. neural networks)
   - can have storage (i.e. more accurate in forecast)

2. Features of the Controller:
   - can make its program with PSIM (i.e. graphical programming)
   - provide many input & output pins
   - in-built analog-to-digital converter (for sensing) with clamped diode protection
   - fs up to 150MHz (enable high-frequency switching & data sampling)
   - have motor control/ power system control libraries
   - Fast interrupt and Self-Protection

9. Connection with PV, Simulation

Difficulties in time:
A 10s simulation takes me a night!





*** From experience, voltage flicker is not a problem in real application.**

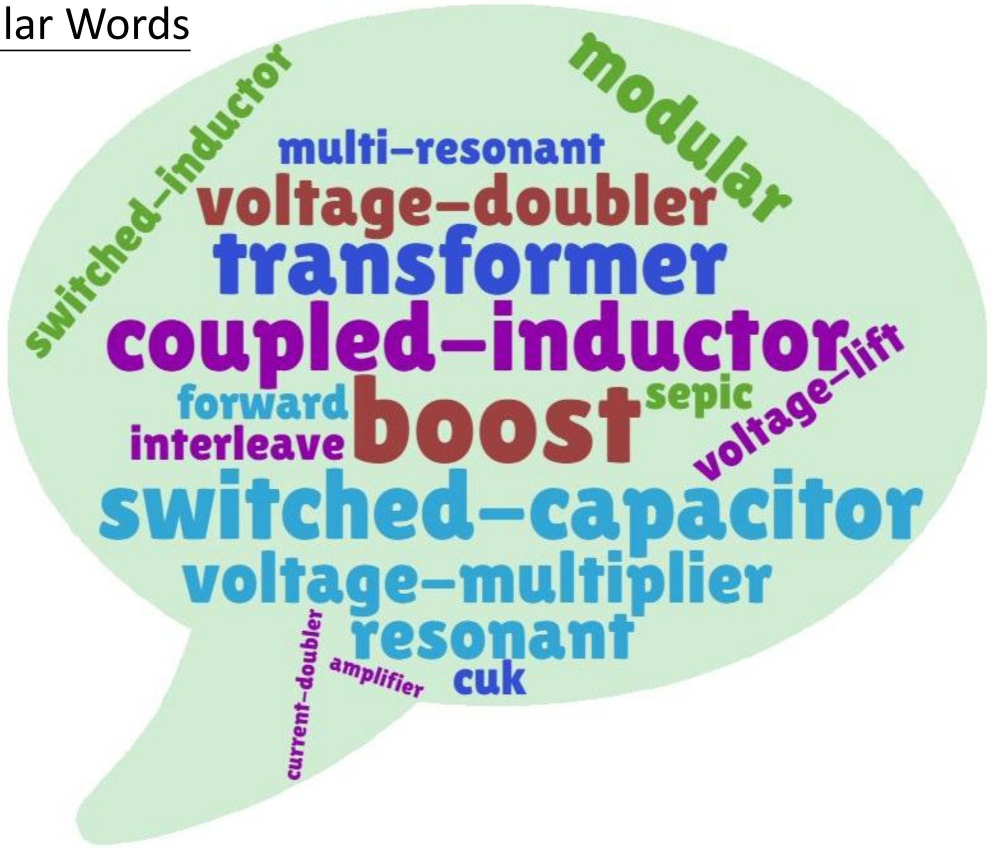
A PLECS simulation model is built on PV – converter interaction.

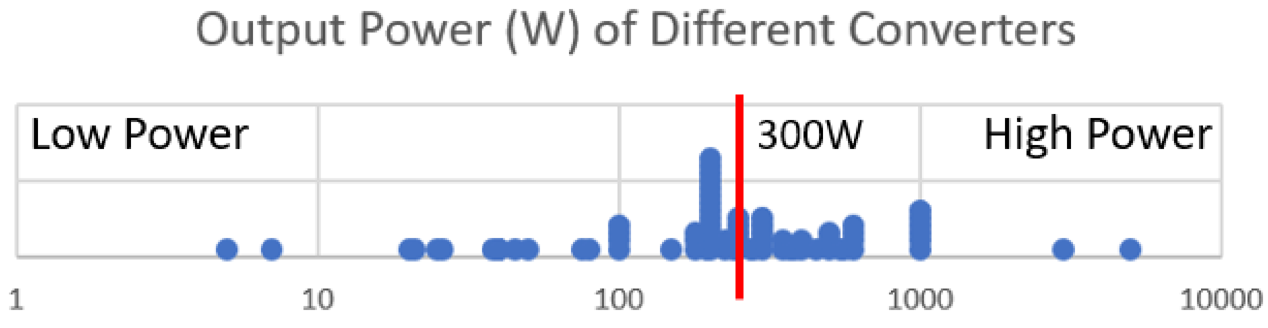
It is observed that:
- it needs a much longer (>2s) time to stabilize the system at start.
- it is hard to regulate input voltage of converter
- it is not sensitive to fluctuation > 1Hz, slow fluctuation does not deviate the output much.

Appendix: Benchmark with PV Converters

Among 81 DC/DC Converters, which is possible for PV applications,

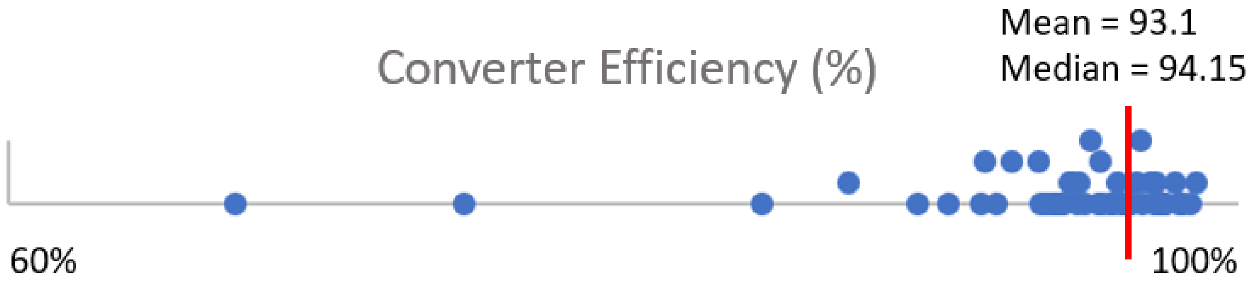
Popular Words





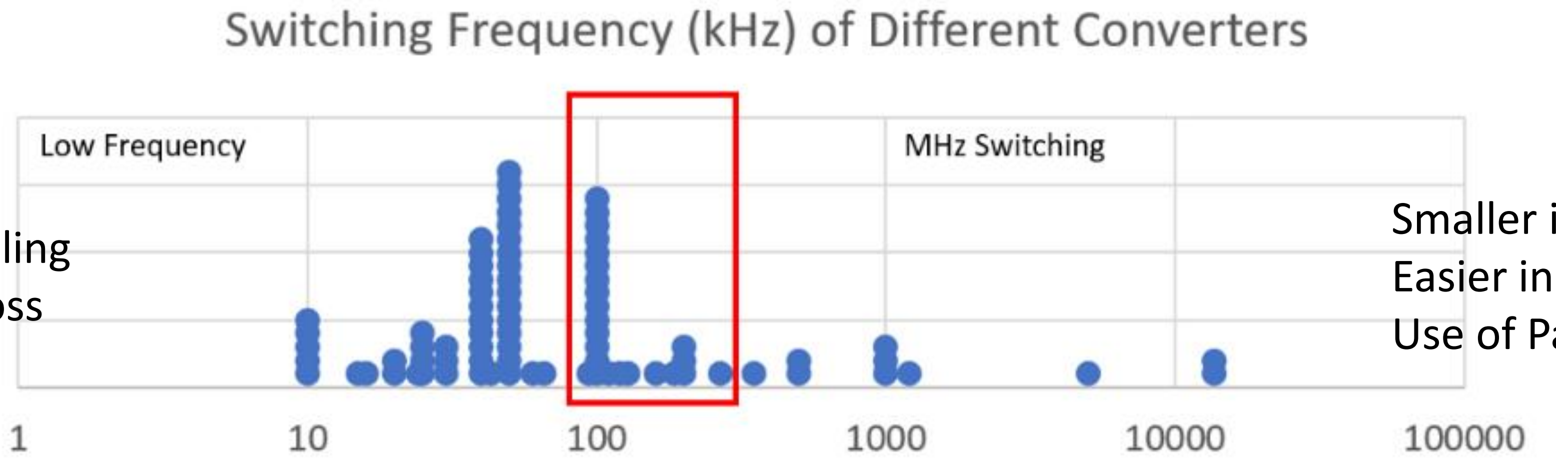
Loss becomes a main concern,
Easier for resonant

Critical in stresses to
active components



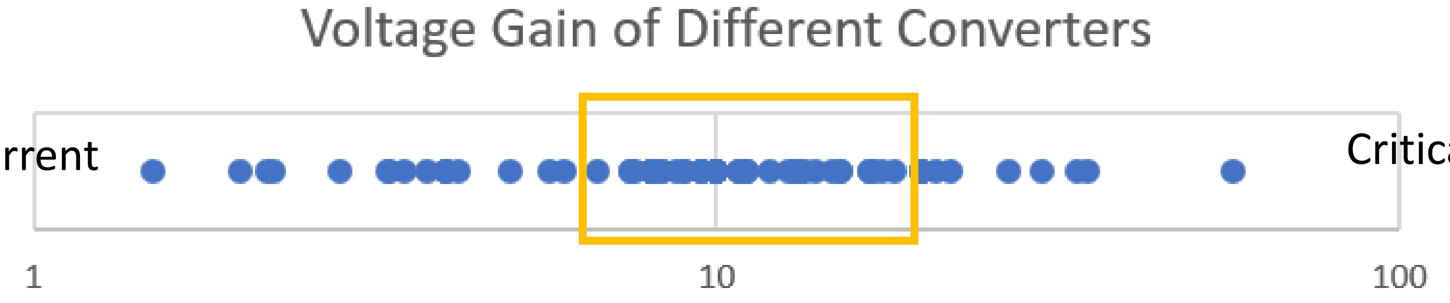
Difficult to achieve high gain with low loss

Appendix: Benchmark with PV Converters



Easier for modelling
Low switching loss

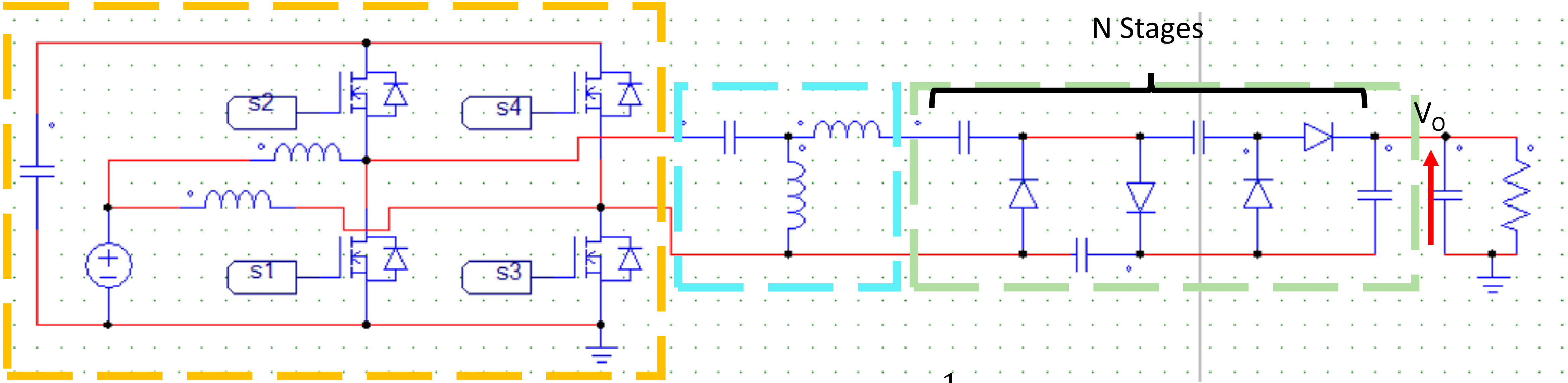
Smaller in size
Easier in transmission
Use of Parasitic Components



Low Gain (Easier)
Less Circulating Current

Critical in lossy components

Appendix: Component Sizing



Input inductance:

$$\Delta i_L = \frac{V_{in}T_s}{4L} \leq \frac{P_o}{2\eta V_{in}} = I_L$$

MOSFET, Clamped Capacitance:

$$V_{MOS} = V_{clamped} = 2V_{in}$$

Resonant Tank:

$$f_r = \frac{1}{2\pi\sqrt{L_3C}} < f_s$$

$$\omega_s = \frac{1}{CL_2}\sqrt{1 + \frac{L_2}{L_3}} = 2\pi f_s$$

$$G = \left|\frac{U_{OUT}}{U_{IN}}\right| = \left|\frac{1}{1 - \frac{1}{\omega_r^2 CL_3}}\right|$$

Output Capacitor:

$$I_o = C\frac{f_s \Delta V_o}{D}$$

CWM Capacitor Voltage Stress:

$$V_{cap} = \frac{2V_o}{N}$$

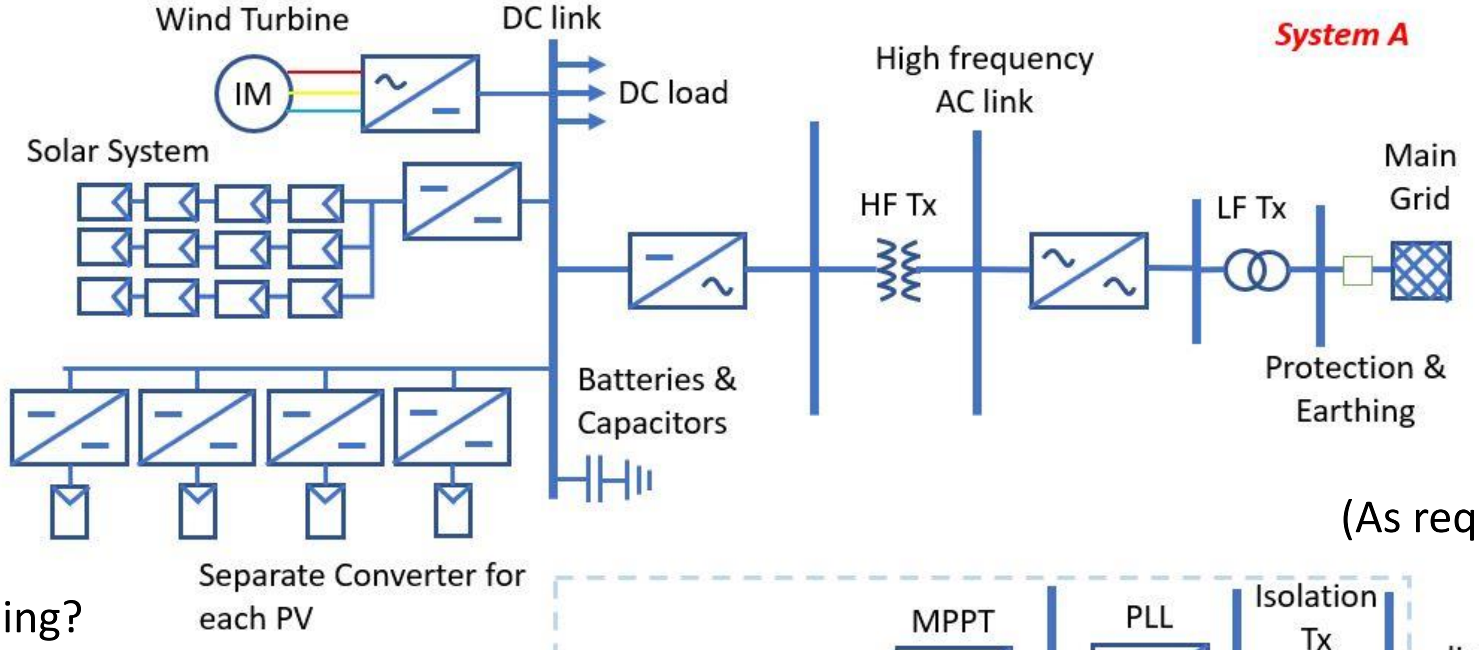
Appendix: Configuration of PV and Grid

-Configuration of System?

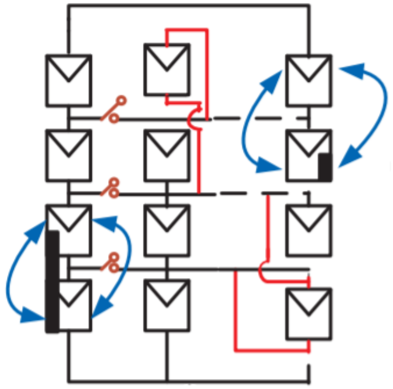
- Time domain
- Space domain

-Modular
-Multilevel
-Multiphase

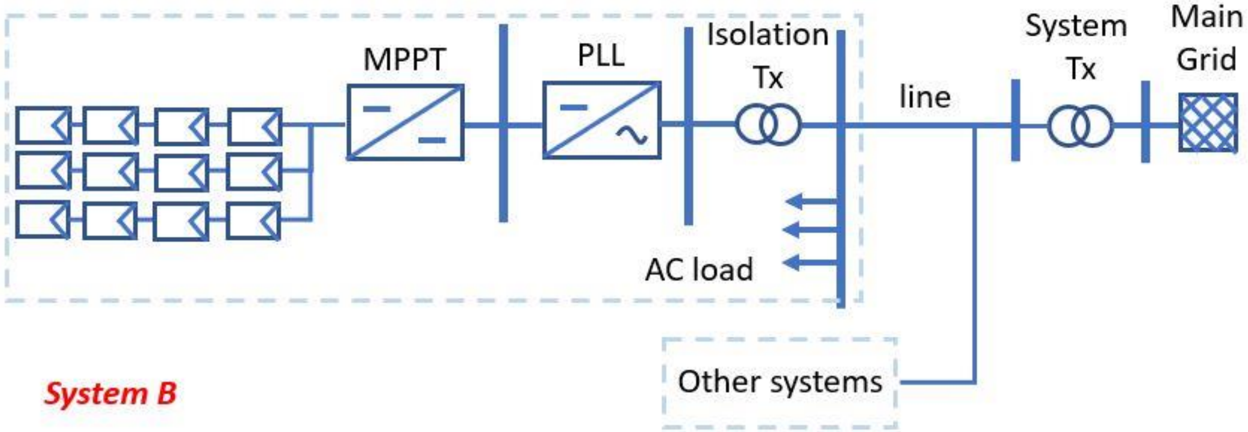
Earthing?



-Reconfigurable?



(As required in HKGOV)



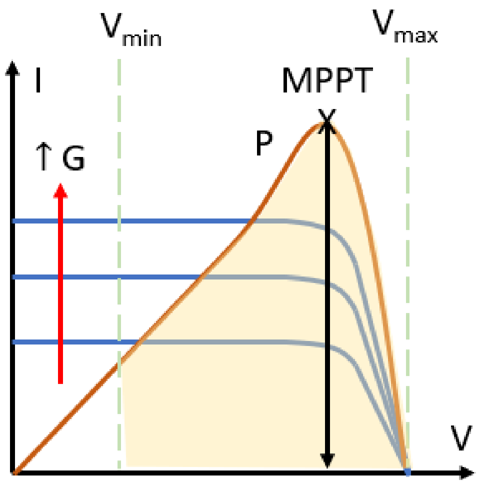
Three layers of control:

- Primary: Droop Control -> V_{DC}
- Secondary: Centralized -> Power Flow
- Tertiary: Exchange of Power with Grid

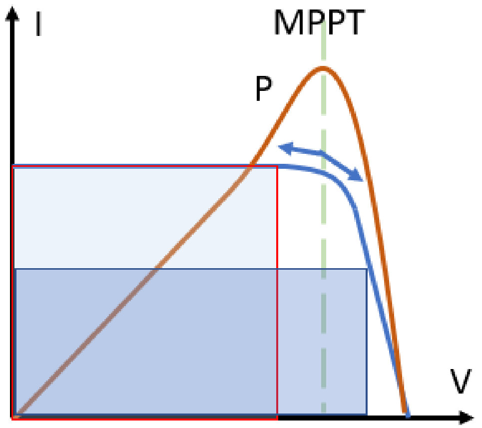
[1] T. Khatib, W. Elmenreich, *Modeling of Photovoltaic Systems Using Matlab: Simplified Green Codes*, John Wiley & Sons, 2016
[2] W. Xiao, *Photovoltaic Power Systems: Modeling, Design and Control*

Appendix: PV Modelling

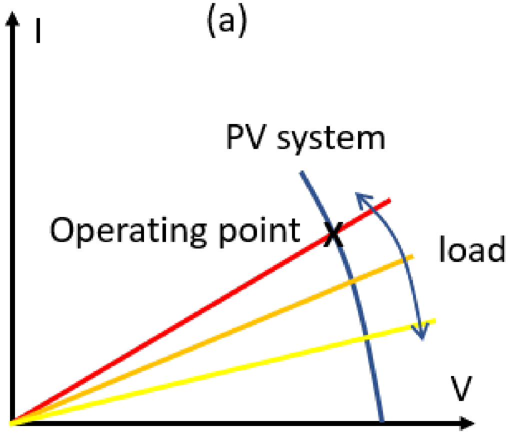
(a) I_{SC} depends more on G, V_{OC} depends more on T
I-V characteristics are interpolated,
MPPT is to harvest most energy with D in boost
(b) In AC system, to mitigate voltage rise, one may apply active power curtailment, i.e. shift point of operation
(c) Source and Load reach the equilibrium point
(d) PV 1D2R model
(e) Source and Load nature
(f) Partial Shading, solved with MPPT/Reconfig.



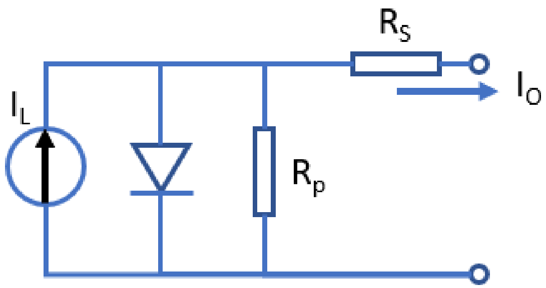
(a)



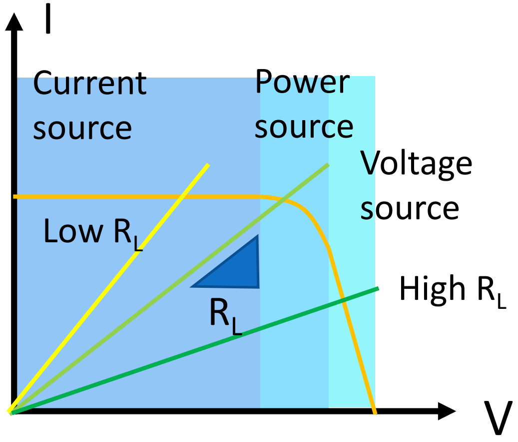
(b)



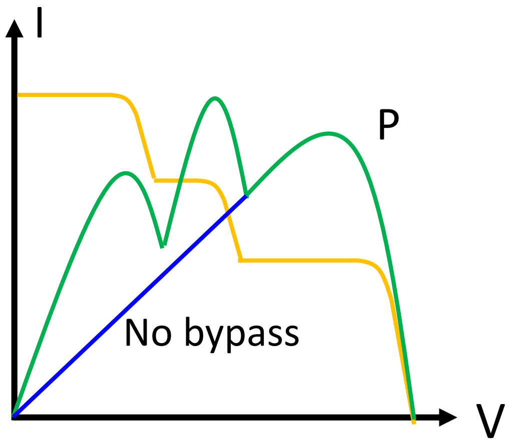
(c)



(d)



(e)

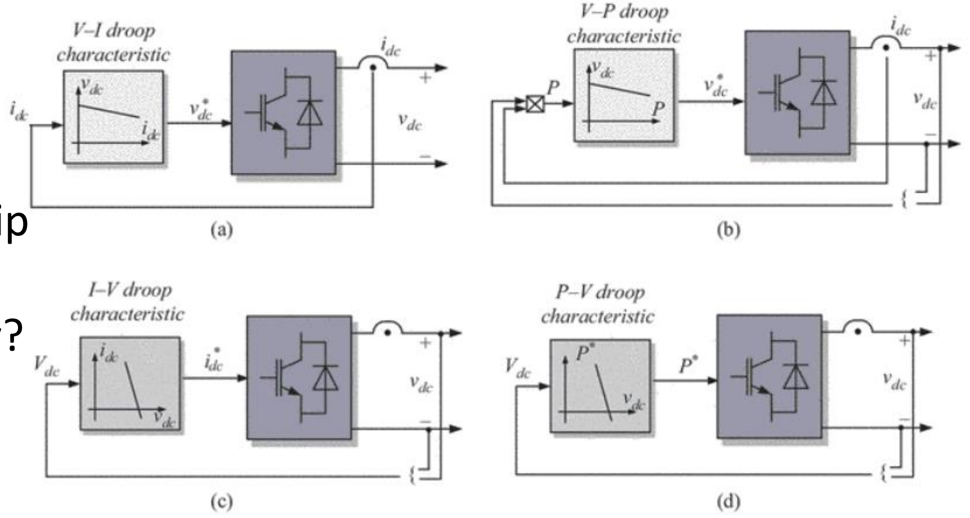


(f)

[1] Dragicevic T. et al (2018), DC Distribution Systems and Microgrids, IET

Appendix: DC link Configuration & Problems

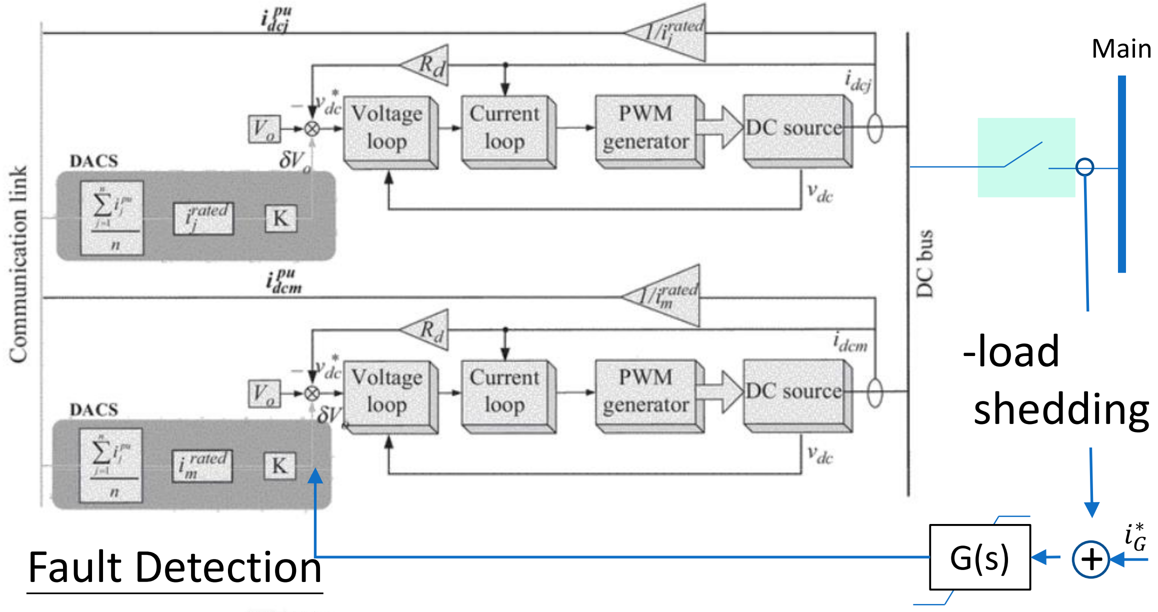
Droop Control:



-relationship
-non-linear
-Frequency?
-Deadband

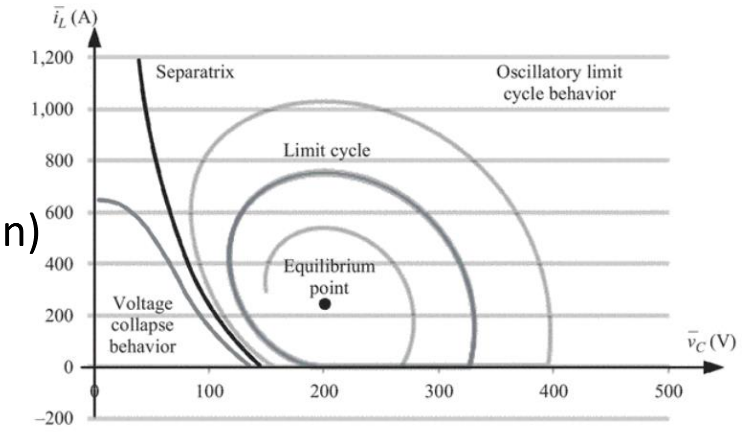
Three-Layer Structure

-consensus



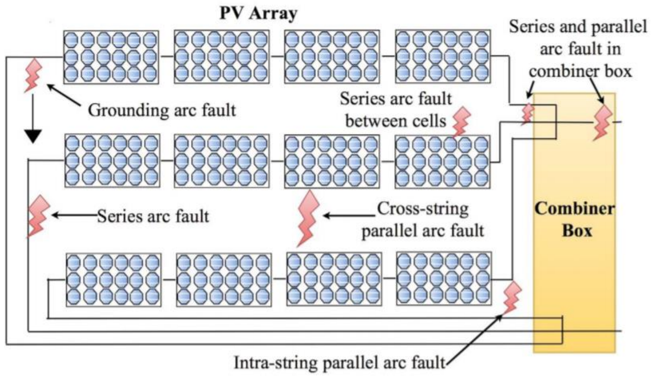
-load shedding

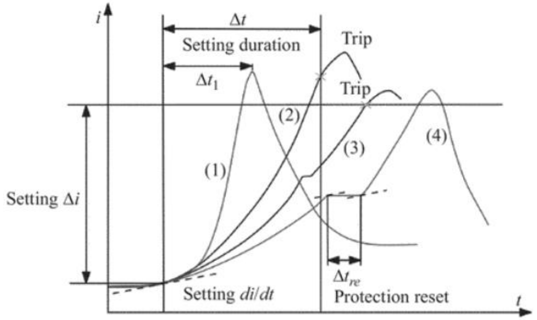
Constant Power Stability Issue:



-add in bulky cap
-PD controller (X regulation)
-virtual impedance
-hysteresis/geometric controller

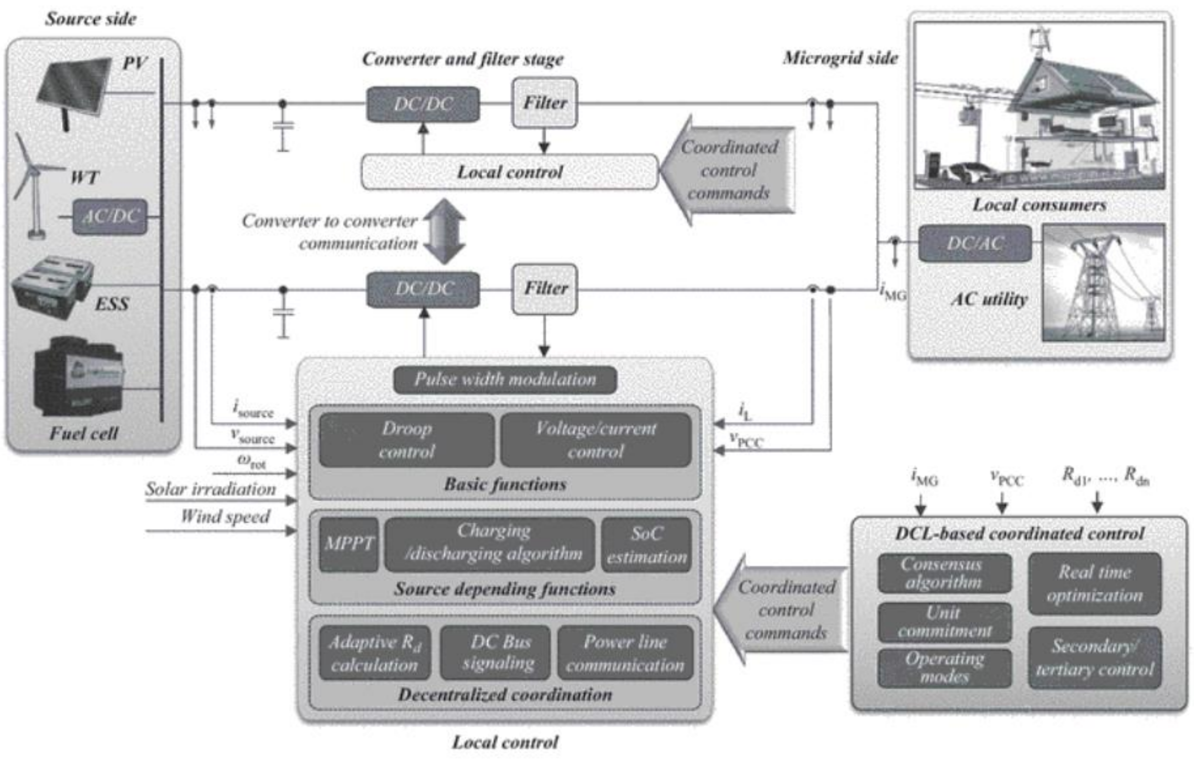
Fault Detection





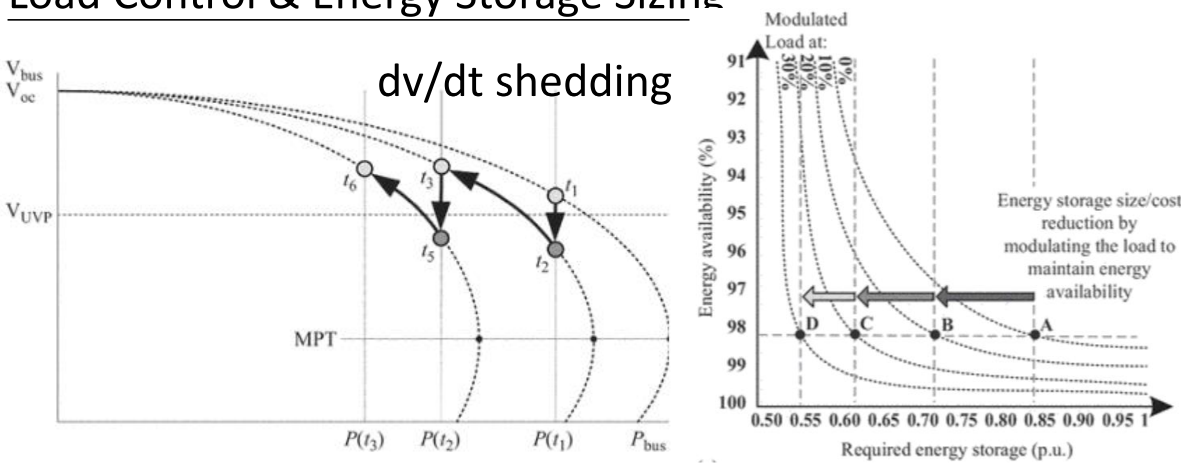
Appendix: DC link Configuration & Problems

Energy Management

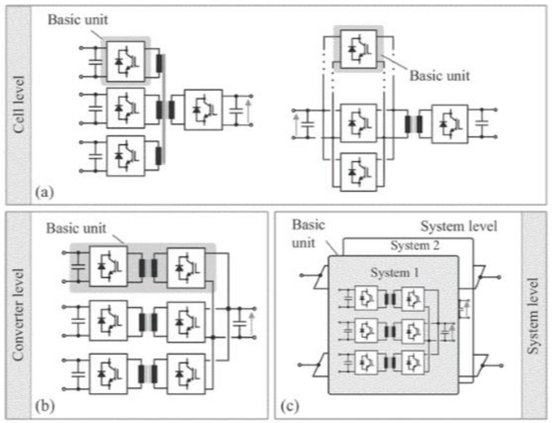


- Centralized system/Decentralized system
- Active Current Sharing (Master and Slave)

Load Control & Energy Storage Sizing



Power Converter Topology/ Solid-State Transformer (SST)



- Power flow control
- DC breaker function
- Modularity
- Degree of Decouple
- Voltage/Current Stress
- Stability