Analysis And Applications of Power Electronics Converters

(EE-332)

Analysis & App. Of PE Converters (S.K. Singh)
(EE-332)

Date __/__/__ Page No.: ____

Power Electronics

→ Power Electronic Devices
- → Diodes — PiN Diodes → transient Voltage & current
- → Controlled transistors
  - IGBT, MOSFET — Voltage Controlled Devices. → (Low to Medium Power App.)
  - BJT, Thyristors — Current Controlled Devices. ↳ Very high power App. (HVDC/FACTS)

→ AC - DC Controlled Rectifiers

↳ DC - DC Choppers

↳ DC - AC Inverters

↳ AC - AC Converters

Applications—
- Photovoltaic Inverters.
- Battery Chargers
- HVDC Transmission
- FACTS Devices
- Electric Vehicles

Syllabus

★ Unit-I - Analysis of Phase Controlled Converters, effect of source inductance, Analysis Of Inverters, PWM techniques, harmonic reduction, resonant D.C. link Inverter. Analysis Of Resonant Converters.

★ Unit-II - DC-DC Switched Mode Converters, Buck, Boost, Buck-Boost, Cuk and bridge converters, Convertors with Isolation, flyback, forward, push-pull

Unit III:- Switch Mode Power Supplies Configuration, Regulation, Control circuits, EMI, HF transformers, rectifiers & filters. High frequency cycloconverter, PWM Rectifier

Unit IV:- Introduction to TCR, TSC, VAR Compensation, electric utility applications, residential industrial applications

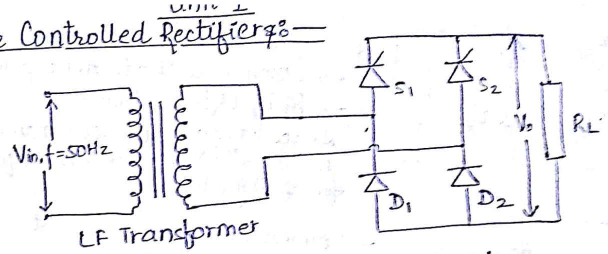
→ Books:-

1. Power Electronics Converter Application & Design - by Ned Mohan, Undeland and Robbins (Unit II)
2. Power Electronics Circuits, Devices & Applications - by M.H.Rashid.
3. Fundamentals of Power Electronics - by Robert W. Erickson & Dragan Maksimovic (Control circuits)
4. Power electronics by P.C.Sen. (Unit I)

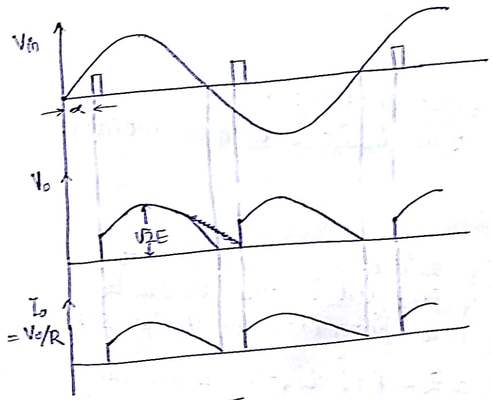
→ Evaluation Pattern:-

- Midterm - 30%
- Assignments - 10%
- End Sem - 45%
- Lab - 15%

Phase Controlled Rectifiers:-



Half Controlled Bridge Rectifier



Centre Tapped Limitation:
1. PIV
2. More no. of turns
3. Asymmetry of transformer windings - Harmonics increased.

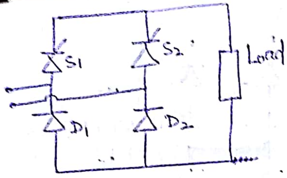
$$V_{o,avg} = \frac{2\sqrt{2}E}{2\pi}\int_{\alpha}^{\pi} \sin\omega t \cdot d(\omega t) = \frac{\sqrt{2}E}{\pi}[1+\cos\alpha]$$

$$V_{o,rms} = \left[\frac{1}{\pi}(\sqrt{2}E)^2\int_{\alpha}^{\pi} \sin^2\omega t \cdot d(\omega t)\right]^{1/2} = E\left[\frac{\pi-\alpha+\frac{\sin 2\alpha}{2}}{\pi}\right]^{1/2}$$

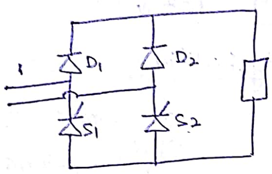
Input Power, $P_{in} = E.I_{rms}$ → Input Rms Current

Output Power, $P_o = V_{orms}.I_{orms}$

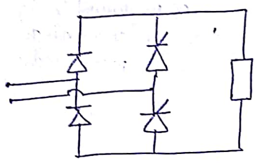
$$\text{Power factor} = \frac{P_o}{P_{in}} = \frac{V_{orms}.I_{orms}}{E.I_{rms}}$$



→ Common Cathode Arrangement
→ Firing ckt. is simple.
Since they have common cathode, no isolated firing ckt. is required for 2 thyristors.



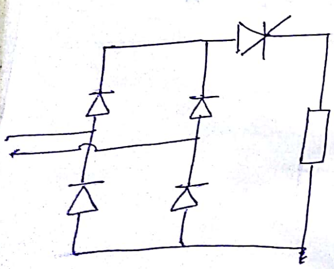
→ Common Anode Arrangement
→ Isolated firing ckt.



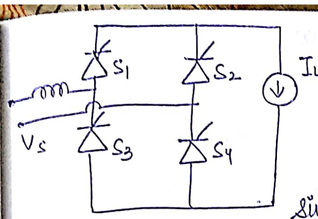
→ Isolated firing ckt.
This circuit should be avoided.

For resistive load, the O/P will be same, but w.r.t. the firing ckt., the I ckt. is prefered because it can be triggered by a common firing ckt. for the two SCR's.

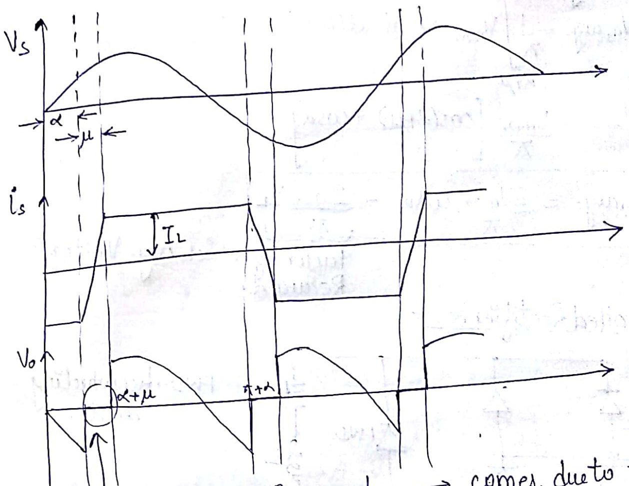
Now, in the case of load $Z = R + j\omega L$, the I & II ckt. will have the similar problem; due to the inductive load so the heavy current may force the thyristor to conduction.



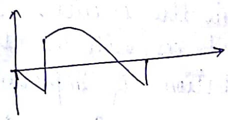
This is a simple one with only one triggering circuit. The resistor load works fine. But for inductive load, we can lose the control & current will circulate. So, not used for inductive load.



L_s = Source Inductance of Transformer
I_L = (load) current source.
μ = overlapping angle, when all the 4 thyristors conduct simultaneously, but there is no short circuit condition.



due to overlapping angle. → comes due to the source inductance L_s. If there is no L_s, then the avg. value will increase & V_o will be like —



$V_o = 0$; $\alpha \le \omega t \le \alpha+\mu$

During commutation,

$$L_s \frac{di_s}{dt} = V_s = V_{max} \sin\omega t$$

Integrating over the commutation interval—

$$\int_{-I_L}^{+I_L} di_s = \frac{V_{max}}{L}\int_{\alpha/\omega}^{\frac{\alpha+\mu}{\omega}} \sin\omega t\,.dt$$

$$\cos(\alpha+\mu) = \cos\alpha - \frac{2\omega L}{V_{max}} I_L$$

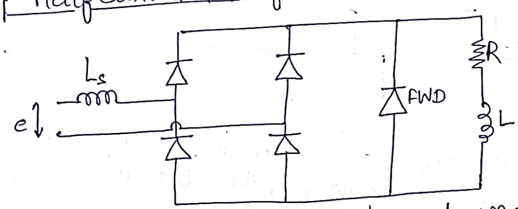
Avg. o/p Voltage—

$$V_{o,avg.} = \frac{1}{\pi}\int_{\alpha+\mu}^{\pi+\alpha} V_{max}\sin\omega t\; d(\omega t)$$

$$= \frac{V_{max}}{\pi}\left[\cos(\alpha+\mu) + \cos\alpha\right]$$

$$\Rightarrow \boxed{V_{o,avg.} = \frac{2V_{max}}{\pi}\cos\alpha - \underbrace{\frac{2\omega L_s}{\pi} I_L}_{\text{Factor By which Avg. Voltage is Reduced.}}}$$

Half Controlled Rectifier:—



FWD → freewheeling diode

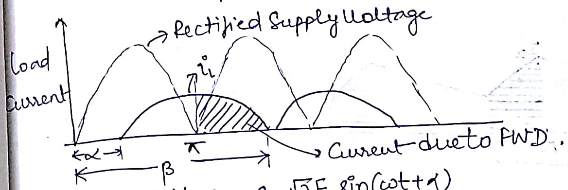
Voltage drop of FWD should be much smaller as compared to other diodes in the circuit. So, FWD takes care of inductive load as well as the effects due to L_s. Voltage Regulation is improved.

Disadv. → An extra diode is required.

Effect of Load Power factor:—

Case (i) – Discontinuous load current:—

$\alpha > \phi$ & $\beta - \alpha < \pi$



Let supply voltage, $e = \sqrt{2}E\sin(\omega t+\alpha)$
Power factor, $\tan\phi = \frac{\omega L}{R}$.

$$L\frac{di}{dt} + iR = e.$$

$$i = \underbrace{\frac{\sqrt{2}E}{Z}\sin(\omega t+\alpha-\phi)}_{\text{Steady State Component}} + \underbrace{Ce^{-\omega t/\tan\phi}}_{\text{Transient Component}}$$

For calculating C, the boundary condition is $i=0$ at $\omega t=0$.

$$C = -\frac{\sqrt{2}E}{R}\cos\phi\sin(\alpha-\phi).$$

To find avg. current during α to π; just integrate.
To find the avg. FWD current (during π to β). Find the total avg. voltage and divide by R to find total avg. current & subtract the bridge current to find FWD current.

Avg. load current, $I_o = \frac{\sqrt{2}E[1+\cos\alpha]}{\pi R}$.

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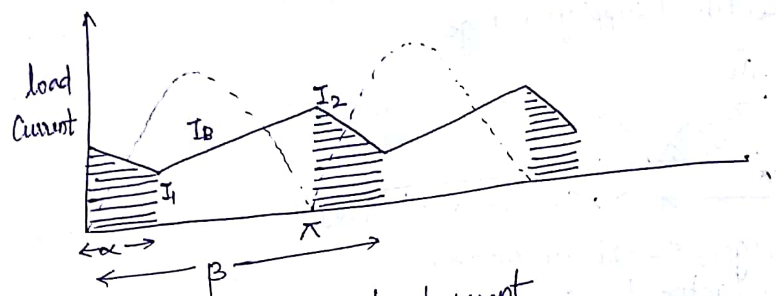
Case (ii:) Continuous load Current [$\alpha < \beta$ and $\beta - \alpha \geq \pi$]



I_1 = ~~Max.~~ Minimum load current
I_2 = Max. load Current
I_B = Bridge Current (Avg.)
I_D = FWD Current (Avg.)

Load Current –

$$i = \frac{\sqrt{2}E\cos\phi}{R}\sin(\omega t+\alpha-\phi) + Ce^{-\omega t/\tan\phi}$$

Boundary Conditions –

$$I_1 = i|_{\omega t=0} = \frac{\sqrt{2}E\cos\phi}{R}\sin(\alpha-\phi)+C \quad \text{— (1)}$$

$$I_2 = i|_{\omega t=\pi-\alpha} = \frac{\sqrt{2}E\cos\phi}{R}\sin(\pi-\phi) + Ce^{-\frac{\pi-\alpha}{\tan\phi}} \quad \text{— (2)}$$

$$I_1 = I_2 e^{-\alpha/\tan\phi} \quad \text{— (3)}$$

from eqⁿ (1), (2) & (3) –

$$C = \frac{\sqrt{2}E}{R}\cos\phi\left[\frac{\sin\phi\, e^{-\alpha/\tan\phi} - \sin(\alpha-\phi)}{1-e^{-\pi/\tan\phi}}\right]$$

Bridge Current,

$$i_B = \frac{\sqrt{2}E}{R}\cos\phi\left[\sin(\omega t+\alpha-\phi)+\text{------}\right]$$

Avg. Bridge Current,

$$I_B = \frac{2}{2\pi}\int^{\pi-\alpha} i_B\, d(\omega t).$$

Avg. load Current, $I_o = \frac{\sqrt{2}E}{\pi R}[1+\cos\alpha]$

$I_D = I_o - I_B$

Avg. Thyristor Current $= \frac{I_B}{2}$.

Higher Avg. Thyristor Current — Without freewheeling diode

But we prefer with FWD.

Q. A single phase half Controlled (Semi-Converter) is connected to 220V, 50Hz Supply. A load R=10Ω is connected in series with a large inductance and load current is ripple free. If the firing angle of converter is 60°, determine –

a.) Avg. O/P Voltage,

$$V_o = \frac{\sqrt{2}E}{\pi}[1+\cos\alpha] = 148.62\text{V}.$$

b.) Avg. load Current,

$$I_o = V_o/R = 14.862\text{A}.$$

c.) DC O/P Power $= V_o I_o = 2208.79$ W.

d.) RMS O/P ~~Current~~ Voltage,

$$V_{rms} = E\left[\frac{\frac{1}{\pi}[(\pi-\alpha)+\frac{\sin 2\alpha}{2}]}{\pi}\right]^{1/2} = 177\text{V}.$$

*** e.) $\boxed{\text{RMS O/P Current, } I_{rms} = I_o = 14.862\text{A.}}$

Because of Ripple free (AC Component is Zero).

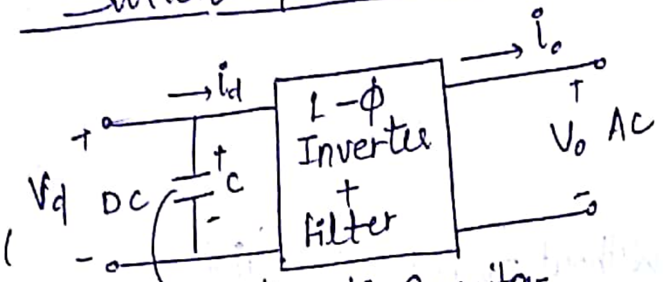
f.) AC O/P Power $= V_{rms} I_{rms} = 2630.57$ W.

g.) Rectification efficiency $= \frac{P_{dc}}{P_{ac}} = \frac{2208.79}{2630.57} \simeq 84\%$.

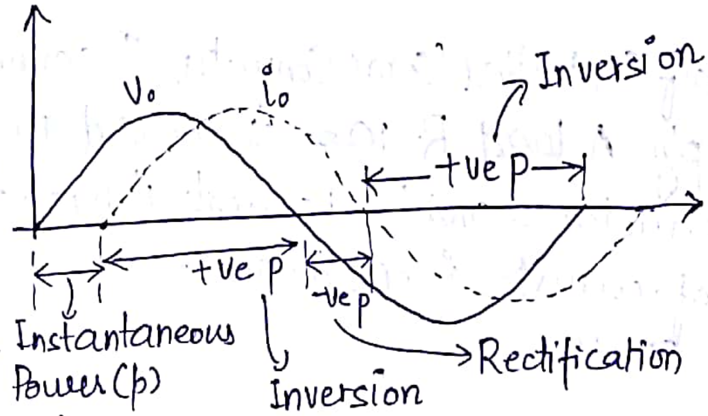
h.) form factor $= \frac{V_{rms}}{V_o} = \frac{177}{148.62} = 1.1909$

i.) Ripple factor, $RF = \sqrt{FF^2 - 1}$

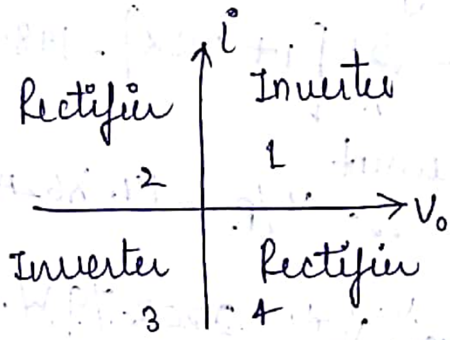
Switch Mode Inverters :-

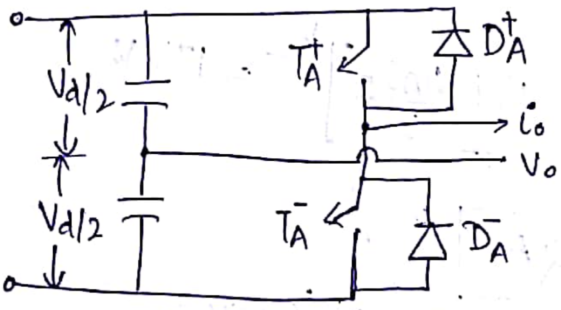


Electrolytic Capacitor polarity fixed.



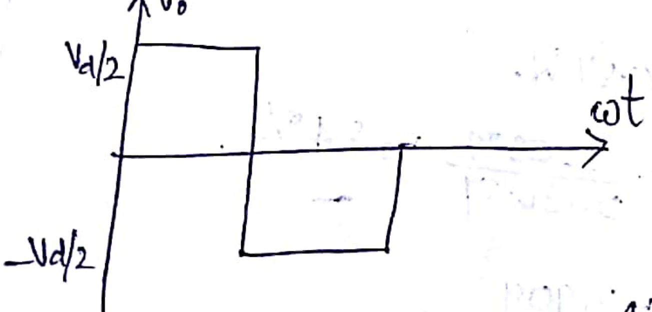
Rectification mode as power flows from O/P to I/P AC to DC.





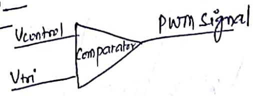
50% time T_A^+ ON & rest T_A^-.

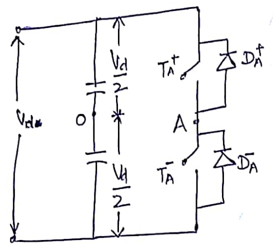
So, square wave is generated at O/P.



1) More harmonics
2) Energy of square wave > sinewa[ve]
3) Switching losses small due to less freq.
4) Main problem is filter ~~size~~ design (size) & heat sinks.

Switch Mode Inverters:-





Effect of Modulation Index

$m_a = \frac{V_{control}}{V_{tri}}$

$m_a < 1 \rightarrow$ linear operating Region.

$m_a > 1 \rightarrow$ Non-linear op. Region

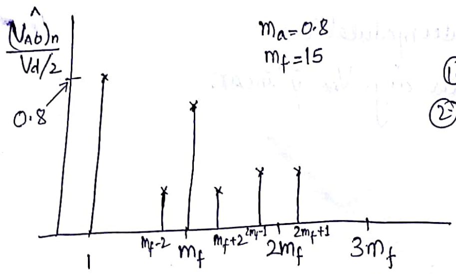
m_f = frequency Modulation Index

$m_f = \frac{f_{sw}}{f_{control}}$

$f_{sw} \Rightarrow$ Triangular Carrier Signal

m_f — Odd (3, 5, ----, 21, ----)

m_f — Even(X) (Because DC Component will appear. → losses will increase.



$m_a = 0.8$
$m_f = 15$

1. Device Rating
2. Switching losses

} Disadv. of higher m_f.

$m_f < 21 \rightarrow$ Sub-harmonic components
(Function of fundamental freq.)
↳ Synchronous PWM

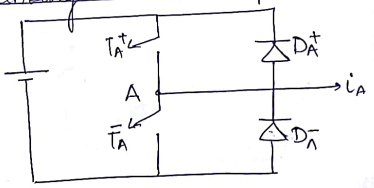
∴ $m_f > 21 \rightarrow$ Sub-harmonic negligible
* Asynchronous PWM

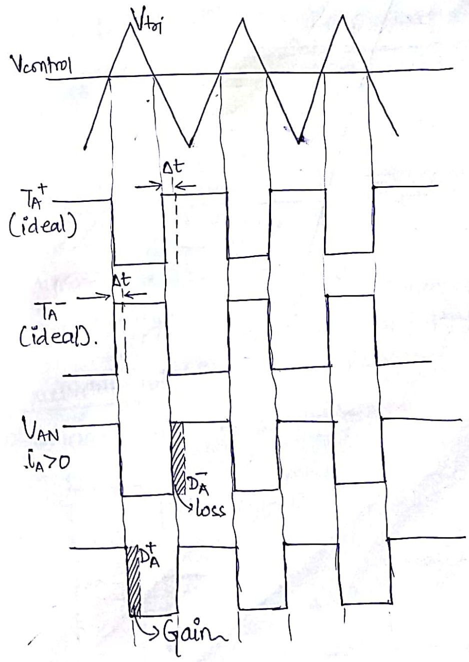
Dead Time / Blanking time

During this time, the ckt. is dead.

Time required for the switch to go from ON to Off state.

Effect of Blanking time (Δt) ON Voltage Source Inverter (VSI):-





Effect of blanking time on Voltage Source Inverter (VSI):—

Change in V_{AN} as compared to the ideal case—

$$V_{AN(ideal)} - V_{AN(\text{Blanking time})} = \Delta V_{AN} = +\frac{\Delta t}{T_{sw}} \cdot V_i \quad ; i_A > 0$$

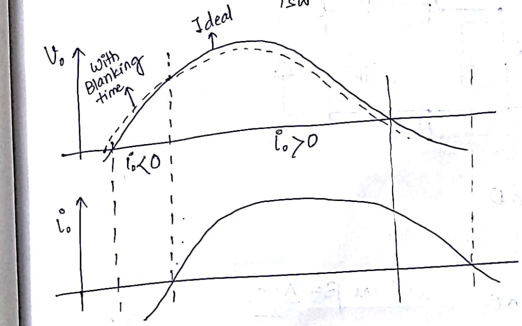
$$= -\frac{\Delta t}{T_{sw}} \cdot V_i \quad ; i_A < 0.$$

→ Numericals.

Change in V_o—

$$V_{Ao(Ideal)} - V_{Ao(\text{Blanking time})} = \Delta V_o = \frac{2\Delta t}{T_{sw}} \cdot V_d \quad ; i_o > 0$$

$$= -\frac{2\Delta t}{T_{sw}} \cdot V_d \quad ; i_o < 0.$$

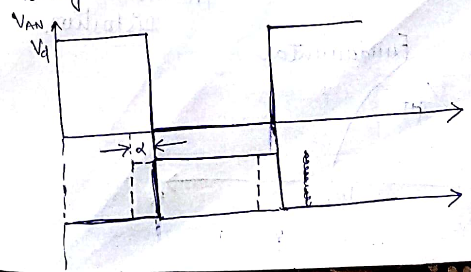


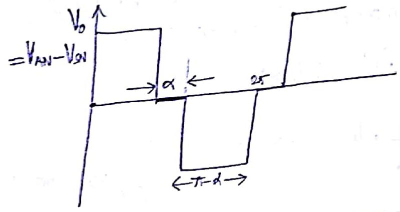
Harmonics $\propto \Delta t$

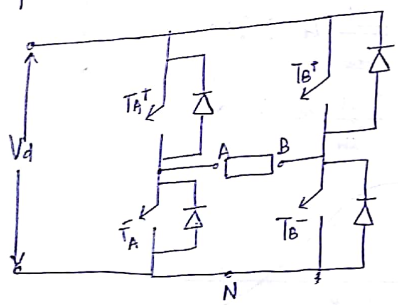
So, due to blanking time, some harmonics are introduced.

SquareWave Switching:—

Voltage Cancellation Technique (1-ϕ inverter).





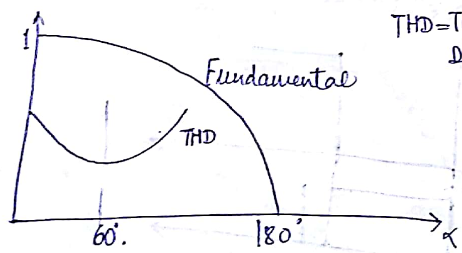


$$(\hat{V_o})_n = \frac{2}{\pi}\int_{-\pi/2}^{+\pi/2} V_o \cos(n\theta)\,d\theta$$

$$= \frac{2}{\pi}\int_{-\beta}^{\beta} V_d \cos(n\theta)\,d\theta \qquad \text{where } \beta = \frac{\pi-\alpha}{2}$$

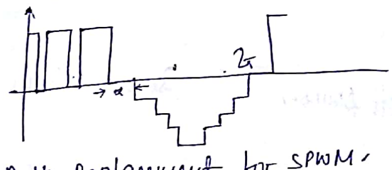
$$= \frac{4}{n\pi} V_d \sin(n\beta)$$

$$(\hat{V_o})_n = \frac{4}{n\pi} V_d \sin\left[n\left(\frac{\pi}{2} - \frac{\alpha}{2}\right)\right]$$



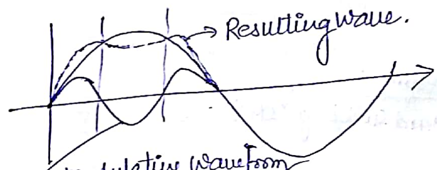
THD = Total Harmonic Distortion.

Notching Technique →

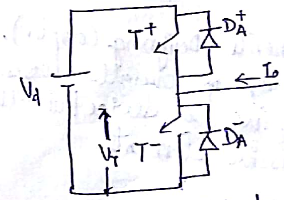


Better Replacement for SPWM.
Thyristor based power system SPWM can't be used with thyristor.

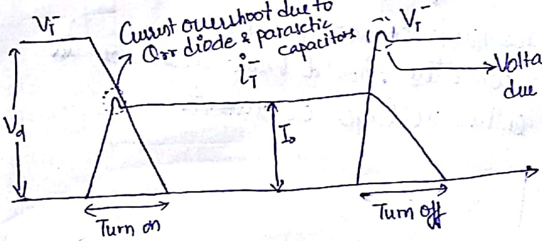
THIPWM ⇒ 3-ϕ Inverter.
Third Harmonic Injection PWM.



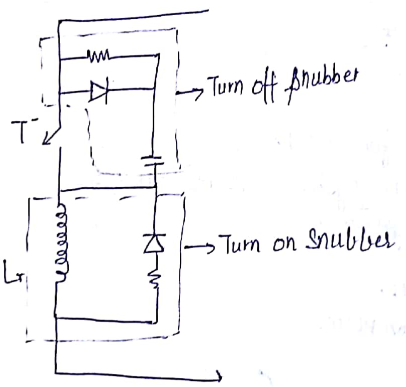
Resonant Converters :-



→ High Switching losses.
→ Higher Switching stress.
→ Huge $\frac{di}{dt} + \frac{dv}{dt}$ ⇒ Electromagnetic Switching Noise (EM Induction or EM compatibility)

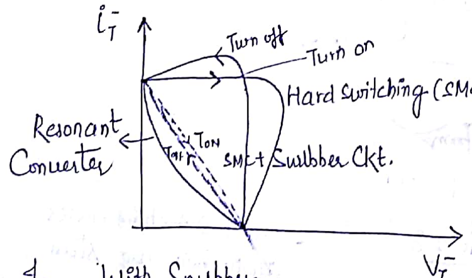


Earlier Snubber Ckt. was proposed ...



Prevent spikes

Snubber prevents switching stress.



Issues With Snubber —

- More no. of electrical components (Designing Complex).
- Don't get benefit in terms of overall losses. Although loss is appearing in device but the loss appears in passive elements.

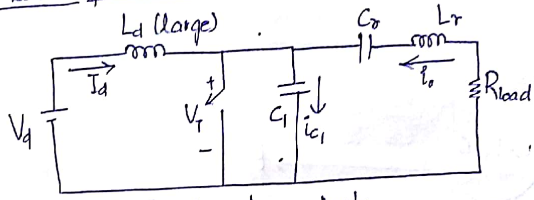
So, idea for resonant converter comes —
Resonant converters should have resonance in either voltage or current.

— Resonant Converter
- → load
- → switch (ZCS/ZVS)
- → DC link

} Based on topology.

ZC/VS → Zero Current/Voltage switching.

Load Resonant : Class E Inverter :—

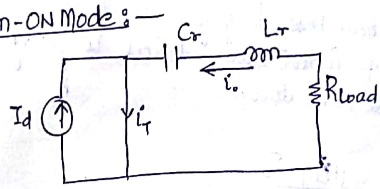


Used in electronic ballast.

I_d is almost constant.
L_d used to convert voltage source to current source.

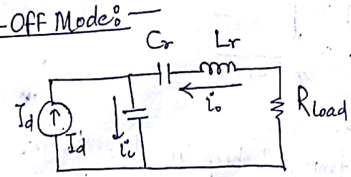
Two Modes Of Operation Based On Switch op. :—

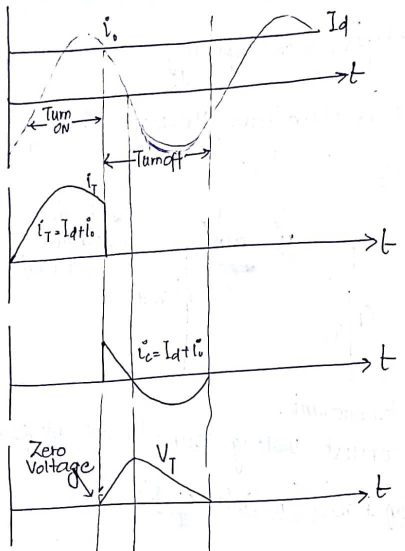
① Turn-ON Mode :—



$i_T = i_o + I_d$

② Turn-Off Mode :—





* f_{sw} is slightly bigger than resonating freq.
* The voltage & current of individual devices is going up. Switching losses will come down.

Now how to control o/p voltage?

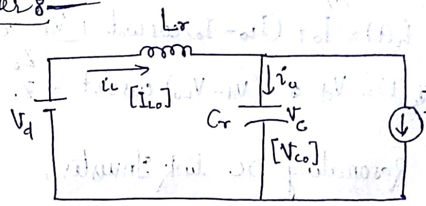
$f_{sw} \to$ Switching frequency

$f_r \to$ Resonating freq. $= \frac{1}{2\pi\sqrt{L_r C_r}}$

f_{sw} is slightly less than f_r.
So, we are controlling f_{sw}, & keep it close to f_r, so that current changes & hence gain & then voltage changes.

We are using current source, so that it can boost the o/p Voltage

DC Link Inverter:-



$V_{co} \to$ initial cap. voltage at $t=0$.
$I_{Lo} \to$ initial current through inductor at $t=0$.

$$v_c = V_d - L_r \frac{di_L}{dt} \quad \text{—(1)}$$

$$i_L - i_c = I_o \quad \text{—(2)}$$

Differentiate (1) −

$$i_L = C_r \frac{dv_c}{dt} = -L_r C_r \frac{d^2 i_L}{dt^2} \quad \text{—(3)}$$

Substituting i_c from (3) to (2) −

$$i_L + C_r L_r \frac{d^2 i_L}{dt^2} = I_o$$

$$\frac{d^2 i_L}{dt^2} + \frac{i_L}{L_r C_r} = \frac{I_o}{L_r C_r}$$

or $$\frac{d^2 i_L}{dt^2} + \omega_0^2 i_L = \omega_0^2 I_o$$

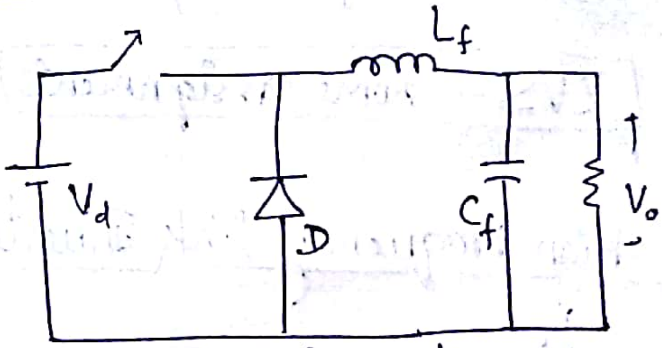
where, $\omega_0^2 = \frac{1}{L_r C_r}$; $Z_0 = \sqrt{\frac{L_r}{C_r}}$

Characteristic Impedance

Resonating frequency

Resonant Switch Converters:—

Zero Current Switching (ZCS):—

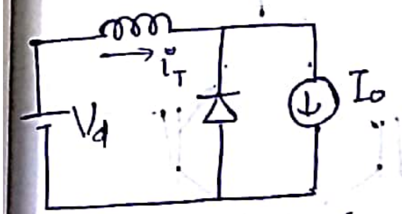


Buck Converter

Initially $i_T = 0$, $V_c = 0$, load current freewheels through the diode D.

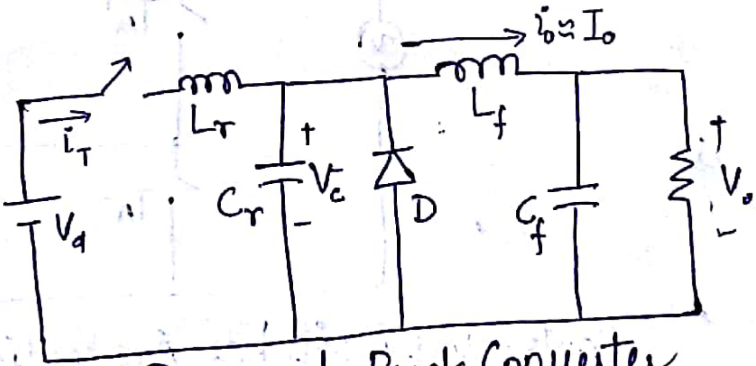
$t_0 < t < t_1$ —

At $t = t_0$, switch is turned ON.

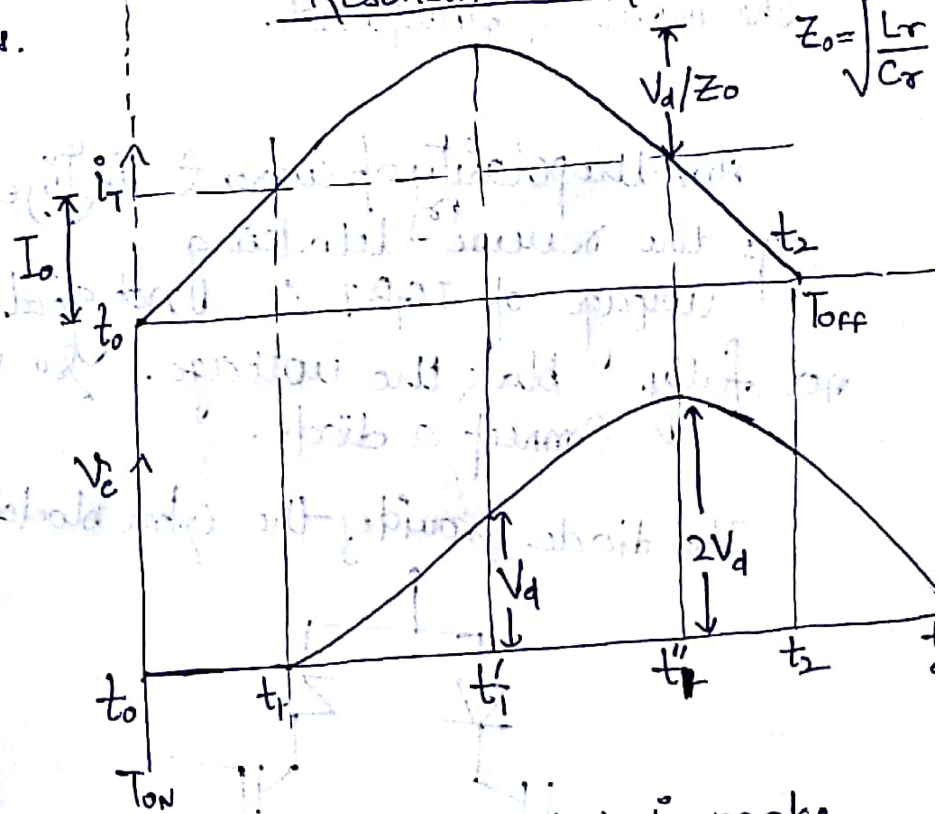


→ i_T linearly increases.

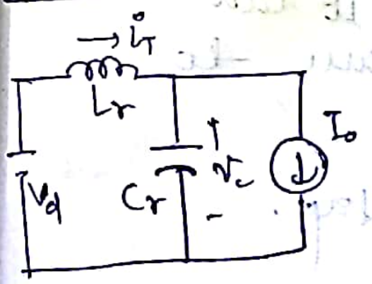
→ $V_c \simeq 0$.



Resonant Buck Converter



$t_1 < t < t_2$ —

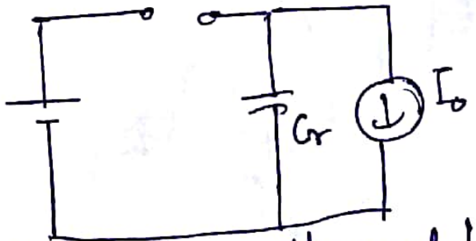


$t_1 = t_1'$ $V_c = V_d$; i_T peaks

$t_1 = t_1''$ $i_T = I_o$; $V_c = 2V_d$

$t_1 = t_2$ $i_T = 0$.

$t_2 < t < t_3$ —

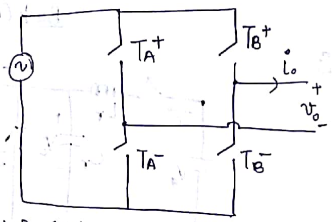


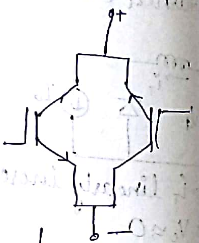
→ V_c discharges through load.

T_{sw} - time period from one ton to another.

ZVS — home Assignment Nedmohan

High Frequency link Inverter (CycloConverter):—

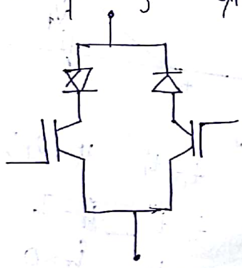




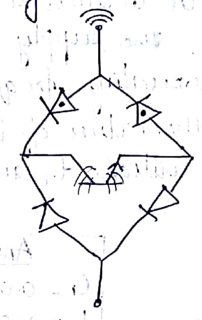
Controlled switches Are Bidirectional.
The input is sinusoidal.

Here the polarity of current is fixe.
As the reverse blocking voltage of IGBT is not good so it can not fully block the voltage. So we have to Connect a diode

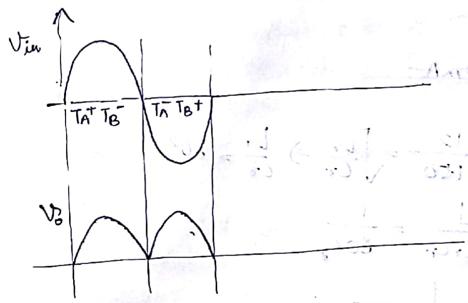
The diode provides the extra blocking voltage

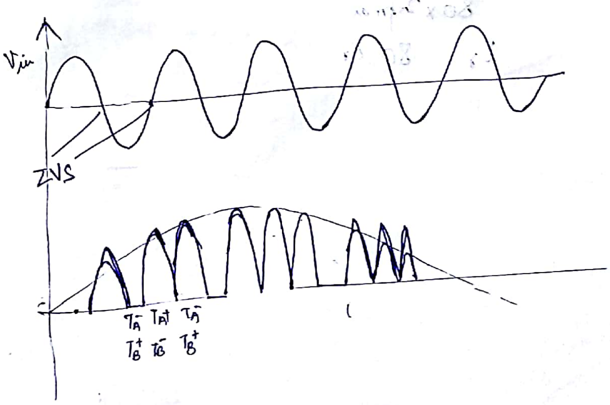


Alternatively, we can design as —



* Less no. of gate driver.
* Switching losses are less.
* Higher conduction losses.





Q. The ZCS Resonant Converter delivers a maximum power $P_L = 400mW$ at $V_o = 4V$. The supply voltage $V_4 = 12V$. The max. operating frequency $f_{max} = 50KHz$. Determine the values of L_r and C_r. Assume that the initial t_1 and $(t_2 - t_1')$ are negligible and $\frac{I_m}{I_o} = 1.5$.

Ans.
$C_r = 0.0407 \mu F$
$L_r = 260.52 \mu H$

Soln

$400 = 4 . I_o$

$I_o = 100 mA$

$\frac{I_m}{I_o} = 1.5 \Rightarrow I_m = 150 mA$

$150 = \frac{12}{Z_o} \Rightarrow Z_o = \frac{12}{150} = \sqrt{\frac{L_r}{C_r}} \Rightarrow \frac{L_r}{C_r} = 80^2$

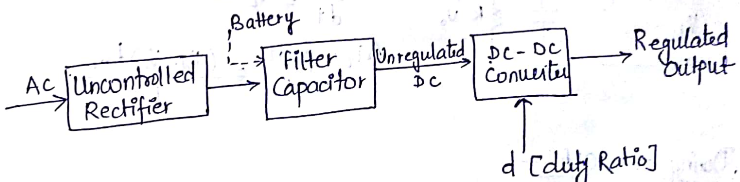
$2\pi f_{max} = W = \frac{1}{\sqrt{L_r C_r}} = \frac{1}{80 C_r}$

$C_r = \frac{1}{80 \times 2\pi f_{max}}$

$L_r = 80^2 . C_r$

DC-DC SWITCH MODE CONVERTER:-

Non-Isolated — High frequency / low freq. transformer is absent.

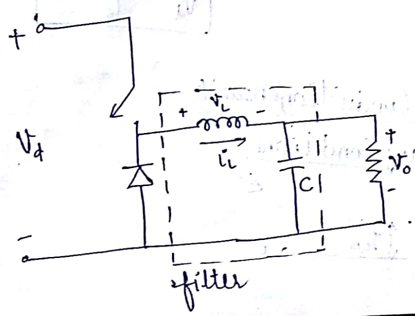


Regulated O/P ⇒ it can be varied as per need.

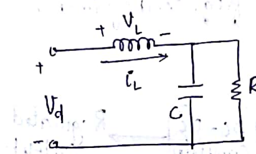
For the constant O/P Voltage & to accomodate change in loads, we control the duty ratio.

→ Non-isolated—
- Buck
- Boost
- Buck-Boost
- Cuk
- Full Bridge

Buck Converter:-



During turn ON:—

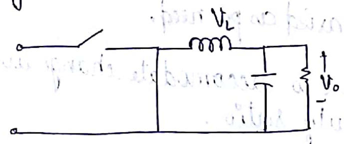


$V_L = V_d - V_o$.

$L\frac{di_L}{dt} = V_L$

$\Delta i_L = \frac{V_L dt}{L} = \frac{(V_d - V_o)\,t_{ON}}{L}$

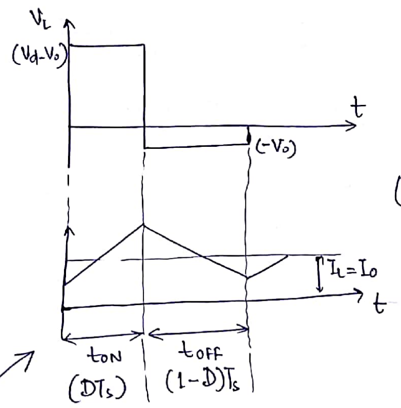
During turn OFF:—



$V_L = -V_o$

$L\frac{di_L}{dt} = V_L$

$\Delta i_L = \frac{V_L t_{OFF}}{L} = -\frac{V_o t_{OFF}}{L}$.



Avg. Voltage across inductor $= 0$.

$(V_d - V_o)t_{ON} + (-V_o)t_{OFF} = 0$.

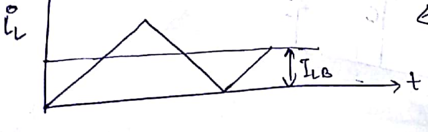
$(V_d - V_o)t_{ON} = V_o t_{OFF}$.

$V_d t_{ON} = V_o T_s$.

$$\boxed{\frac{V_o}{V_d} = \frac{t_{ON}}{T_s} = D}$$

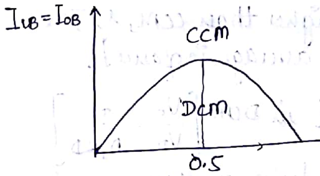
CCM (Continuous Conduction mode):—

Boundary Conditions

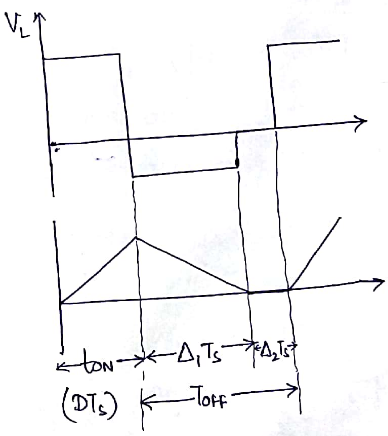


$I_{OB} = I_{LB} = \frac{1}{2}\hat{i}_L = \frac{(V_d - V_o)DT_s}{2L} = \frac{V_d T_s}{2L}D(1-D)$

$I_{LB,max} = \frac{V_d T_s}{8L}$.



DCM (Discontinuous Conduction Mode):—



$(V_d - V_o)t_{ON} + (-V_o)\Delta_1 T_s = 0$.

$(V_d - V_o)DT_s = V_o \Delta_1 T_s$.

$$\boxed{\frac{V_o}{V_d} = \frac{D}{D + D_1}}$$

Scanned by CamScanner

Power loss negligible,

$$\boxed{\frac{I_o}{I_D} = \frac{1}{D}} \quad , \quad \boxed{\frac{I_o}{I_D} = \frac{D + D_1}{D}}$$

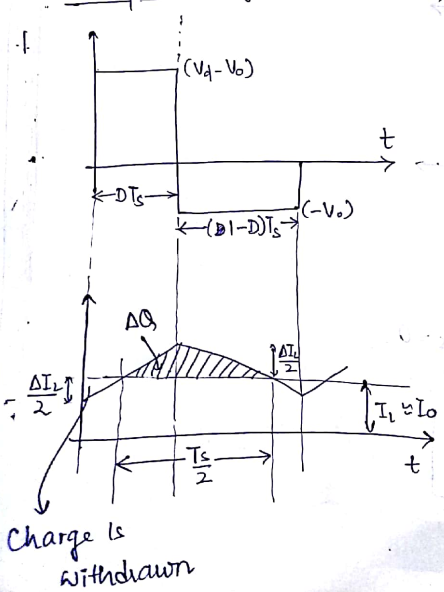
CCM   DCM

→ Under DCM, peak current is higher than CCM, so the current rating is high. [when average is same].

→ Some non-linearity is introduced in DCM $\left[\frac{V_o}{V_D} = \frac{D}{D_1 + D} \right]$ because of the above eqn. So, design problem.

H.W. Derive the same for boost & buck-boost converter.

Output Voltage Ripple :—



During Turn-OFF :—

$$\frac{L \Delta I_L}{(1-D)T_s} = V_o$$

$$\boxed{\Delta I_L = \frac{V_o (1-D) T_s}{L}}$$

$$\Delta V_o = \frac{\Delta Q}{C} = \frac{1}{2C} \frac{T_s}{2} \frac{\Delta I_L}{2}$$

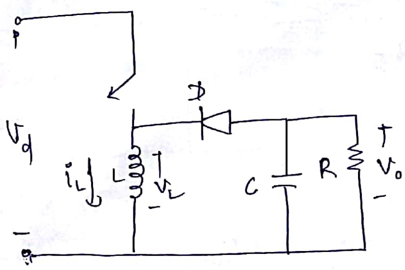
$$\Rightarrow \frac{\Delta V_o}{V_o} = \frac{1}{2} \frac{T_s}{2} \frac{(1-D)T_s}{2LC}$$

$$\Rightarrow \boxed{\frac{\Delta V_o}{V_o} = \frac{\pi^2}{2}(1-D)\left(\frac{f_c}{f_s}\right)^2} \; ; \; f_c = \frac{1}{2\pi\sqrt{LC}} \; ; \; f_s = \frac{1}{T_s}.$$

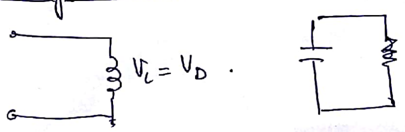
Cut-off frequency (freq. at which gain comes down by 63.7%).

So, during designing, f_s is constant ; f_c & D are varied according to requirement.

Buck-Boost :—



During Turn ON :—

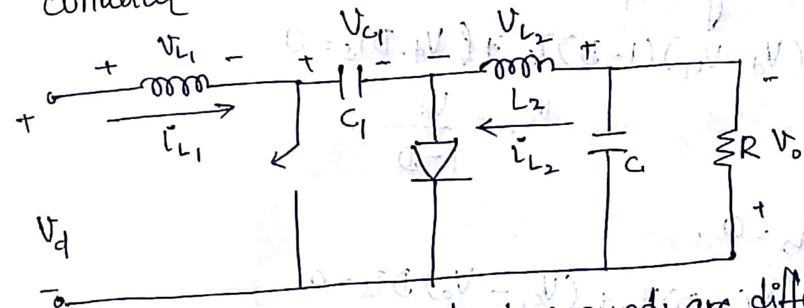


$$\Delta i_L = \frac{V_L \, dt}{L} = \frac{V_d \, t_{ON}}{L}.$$

Ĉuk Converter (~Buck-Boost Converter):—

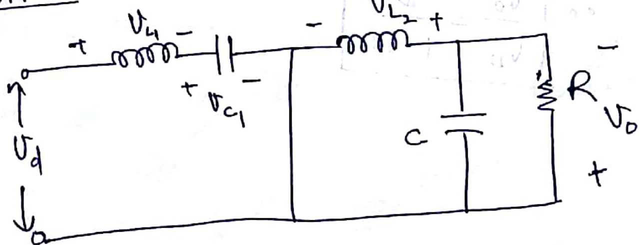
Problems with Buck-Boost:—

1. Input & output grounds are different.
2. Initial charging of the capacitor is required.
3. Because of the switch, the input current is discontinuous. So, we require a pulsating input current which reduces the life of the source & also the converter.



Here, also the input & output grounds are different.

Turn-OFF:—



$V_{L_1} = V_d - V_{C_1}$

$V_{L_2} = -V_o$

Initially, avg. voltage across L_1 & $L_2 = 0$.

But, we have some voltage across C_1.

$V_{C_1} = V_d + V_o$.

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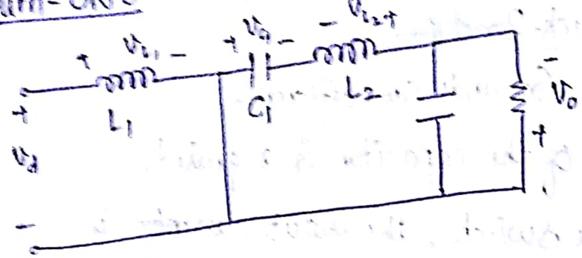
Turn-ON:-



$v_{L_1} = V_d$.

$V_{L_2} = V_{C_1} - V_o$.

Avg. $V_{L_1} = 0$.

$\Rightarrow (V_d - V_{C_1})(1-D)T_s + V_d . DT_s = 0$

$V_{C_1} = \frac{V_d}{1-D}$

Avg. $V_{L_2} = 0$.

$\Rightarrow (-V_o)(1-D)T_s + (V_{C_1} - V_o) DT_s = 0$

$\Rightarrow V_{C_1} = \frac{V_o}{D}$.

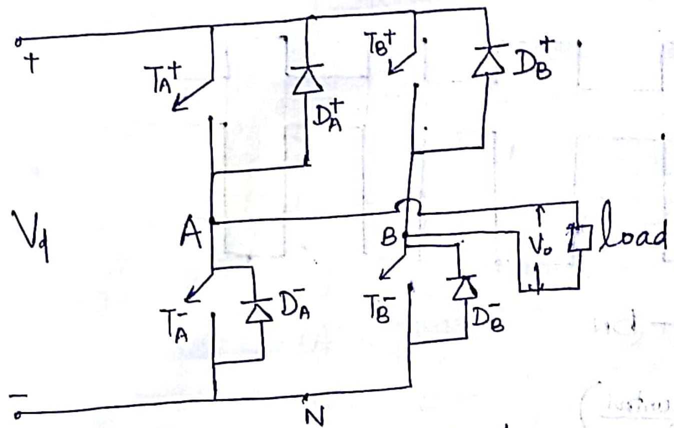
$\therefore \boxed{\frac{V_o}{V_D} = \frac{D}{1-D}}$

Ćuk Converter:-

Characteristics- (Buck-Boost)

* Continuous input and output current. (+)
* Output is of negative polarity. (-).
* C_1 acts as a primary means of storing & transferring energy. It should carry large ripple current.
* Less Ripples in i_{L_1} and i_{L_2}, hence reduced filtering is required.

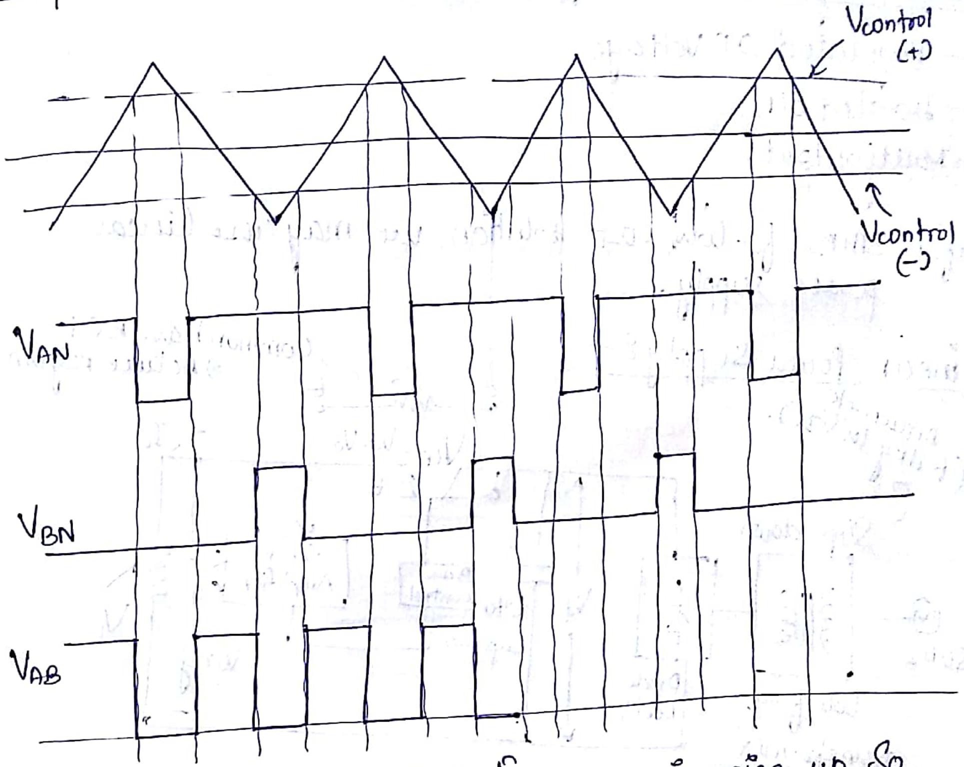
Full Bridge DC-DC Converter:-



It's a voltage source converter bcoz no inductor at source. It can not be a boost converter, bcoz we do not have energy storing elements.

It is bidirectional in terms of both current & voltage.
"Four Quadrant Application"

Unipolar PWM :—



The change is that the ripple frequency is going up. So, filter requirement should decrease.

$$D_1 = \frac{1}{2}\left(1 + \frac{V_{control}}{\hat{V}_{tri}}\right)$$

$$D_2 = 1 - D_1$$

$$\boxed{V_o = (2D_1 - 1) V_d}$$

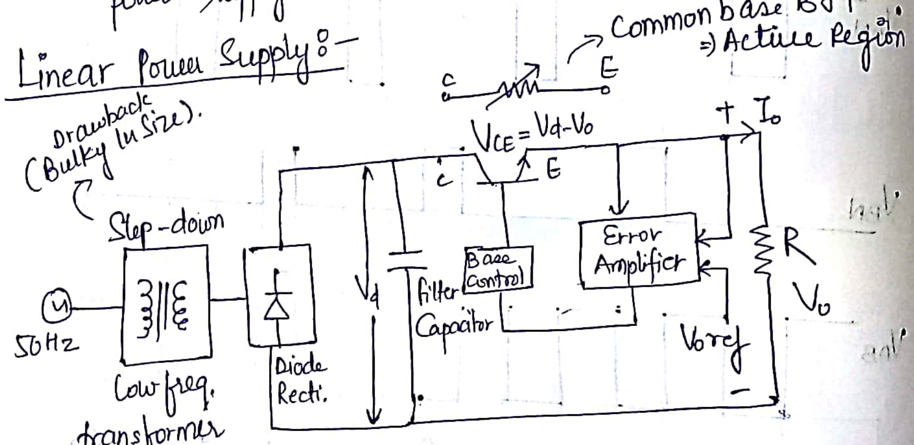
Switching DC Power Supplies :-

→ Regulated O/P Voltage.
→ Isolated O/Ps.
→ Multioutput.

Before SMPS, for low cost solution, we may use linear power supply.

Linear Power Supply :-

Drawback (Bulky in size).



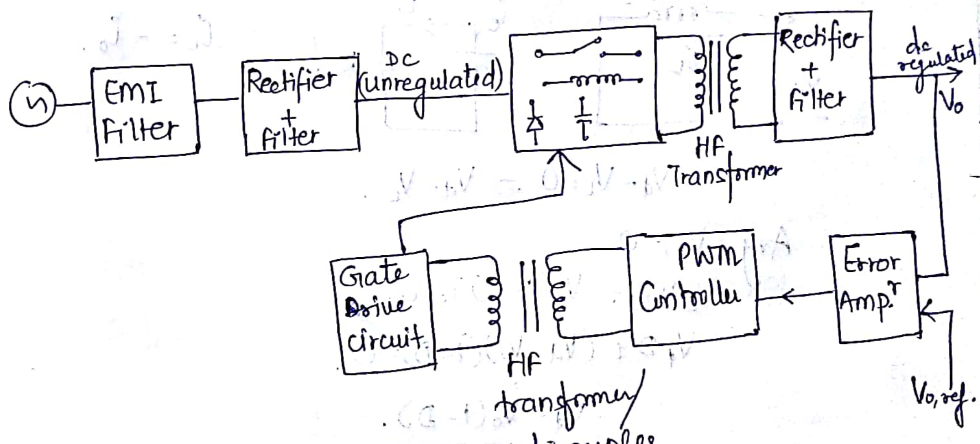
* Bulky Power Supply.
* 30–50% efficiency.

Advantages-
* EMI Issues Are negligible
* Diodes are not contributing in hard switching.

Applications-
* Medical Equipments (noise is avoided).
* Audio Amplifier.

low freq. transformer → ferromagnetic core
Permeability is High.
$\mu_r(\text{low freq.}) \simeq 3000$.
$\mu_r(\text{high freq.}) << \mu_r(\text{low freq.})$

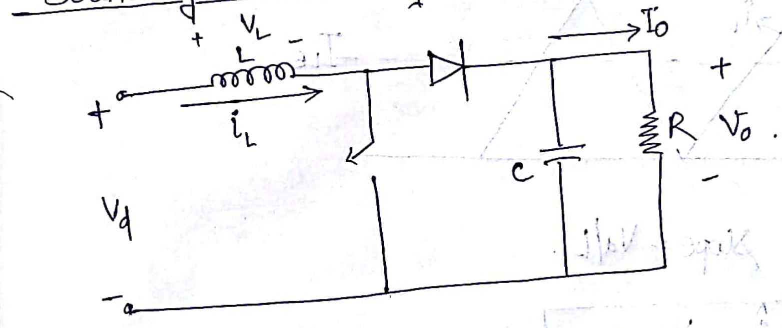
Switch Mode Power Supply (SMPS) :-



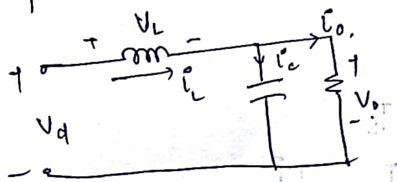
Advantages-
* 70–90% efficiency.
* HF transformer is compact.
* Heat sink of Reduced size.

Limitations-
* EMI Problem
* More no. of elements → life of system gets affected (Reliability Affected)

Boundary Conditions Of Boost Converter :-

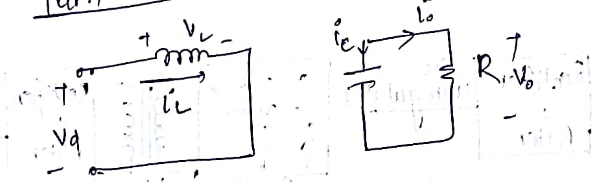


During Turn Off :-



$V_d - V_L - V_o = 0$
$\Rightarrow V_L = V_d - V_o$.
$i_L = i_c + i_o$

Turn ON :-



$i_c = -i_o$.

$V_d - V_L = 0 \Rightarrow V_d = V_L$.

Avg. $V_L = 0$

$V_L D T_s + V_L(1-D)T_s = 0$

$V_d D + (V_d - V_o)(1-D) = 0$.

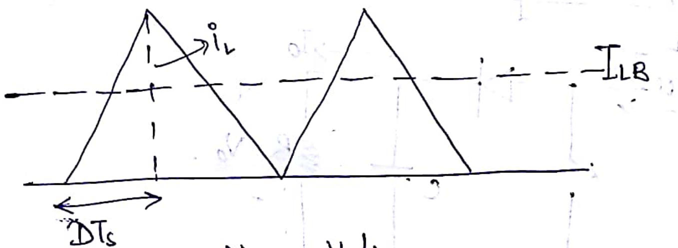
$V_d = V_o(1-D)$.

$$\boxed{\frac{V_o}{V_d} = \frac{1}{1-D}}$$

$i_c D T_s + i_c(1-D)T_s = 0$.

$-i_o D + (i_L - i_o)(1-D) = 0$.

$$\boxed{\frac{i_o}{i_L} = 1-D}$$



Slope $= V_d/L$.

$$\boxed{\hat{i}_L = \frac{V_d}{L} DT_s}$$

By averaging-

$I_{LB} \cdot T_s = \frac{1}{2} \hat{i}_L \cdot T_s$

$I_{LB} = \frac{1}{2} \hat{i}_L$.

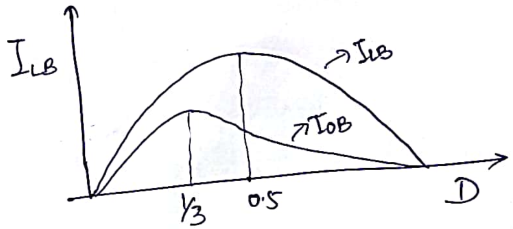
$I_{LB} = \frac{V_d}{2L} DT_s$.

$$\boxed{V_d = V_o(1-D)}$$

$I_{LB} = \frac{V_o D(1-D)T_s}{2L}$

$I_{OB} = (1-D) I_{LB} \longrightarrow$ By Avg.

$$\therefore \boxed{I_{OB} = \frac{T_s \cdot V_o}{2L} D(1-D)^2}$$



Q. In a boost converter, the duty ratio is adjusted to regulate the output voltage V_o at 48V. The input voltage varies from 12 to 36V. The maximum power output is 120W. The converter operation in DCM at $f_{sw} = 50kHz$. Assuming ideal components & C very large, calculate the maximum value of L that can be used.

$I_o = \frac{120}{48} = 2.5A$.

Sol^n $I_{OB} = \frac{T_s \cdot V_o}{2L} D(1-D)^2$.

$V_o/V_d = 1/1-D$.

$\therefore D = \left(\frac{1}{4} \leftrightarrow \frac{3}{4}\right)$.

$T_s = \frac{1}{50 \times 10^3} = 2 \times 10^{-5} s$

Point C will be chosen to be in DCM to make 2.5A value.

$\therefore D = 0.75$.

$L = \frac{2 \times 10^{-5} \times 48}{2 \times 2.5} \times \left(\frac{3}{64}\right) = 9 \mu H$.