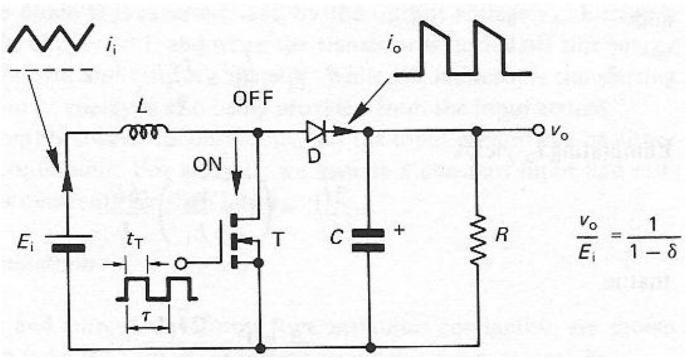
EEE307 / EEE6023 Answers – Summer 2012

Qu 1.

a. The dc-dc converter is a boost converter. The output is always greater than or equal to the input.



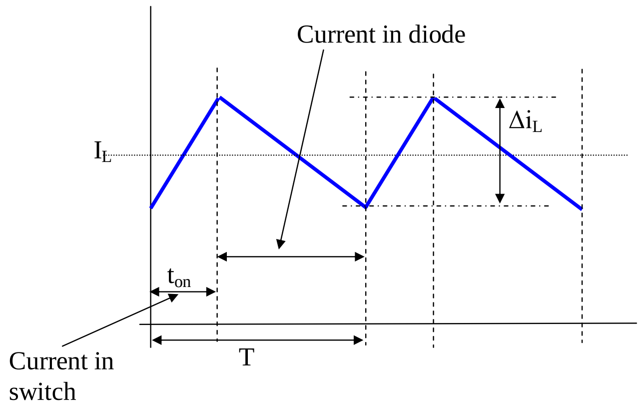
When switch T is on, current in the inductor increases at a rate:

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

When switch T is off the current in the inductor decreases at a rate:

$$\frac{di_L}{dt} = \frac{V_o - E}{L}$$

Operation in Continuous Inductor (Choke) current:



For time switch is on, the current increase is:

$$\Delta i_L = \frac{E}{L}.t_{on}$$

For the time the switch is off, the current decrease is:

$$\Delta i_L = \frac{(V_o - E)}{L}.(T - t_{on})$$

In steady state, the current increase equals the current decrease, therefore by equating the two equations for the ripple we get:

$$\frac{V_o}{E} = \frac{1}{(1-\delta)}$$

where $\delta = t_{on}/T$, and δ can take values between 0 and 1 inclusive.

b. If the input voltage varies between 90V and 250V dc, and the output is 400V with an output current of between 0.1A and 6A, operating at 50kHz, the inductor value is given from:

$$I_o \geq \frac{E}{2L}.T.\delta.(1-\delta).$$

The duty cycle varies from 0.775 at 90V to 0.375 at 250V, and for the 2 conditions at the lowest input voltage the inductor has to be the greater of 1.57mH and 5.86mH, therefore:

L=5.86mH

For the capacitor:

$$\Delta V_o = \frac{\Delta Q}{C} = \frac{I_o \delta T}{C}$$

Therefore with the ripple = 5% of 400V = 20V, at maximum output current (6A), the capacitor will be the greater of 2.25uF and 4.65uF, therefore a preferred value of

C=4.7uF would be chosen.

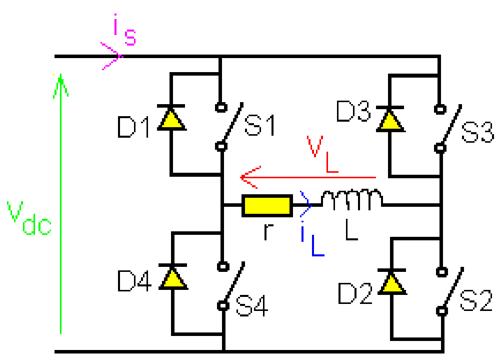
The MOSFET should be have a breakdown voltage >400V, therefore perhaps choose a 500V or 600V device, with a current rating of greater than

$I_L = \frac{I_o}{(1-\delta)}$, so for a 6A output with a duty cycle of 0.775, the MOSFET current will be > 26.7A, so choose a current rating of 30A or greater, maybe 40A

The diode will also need to withstand >400V, so choose a 600V device, with a current rating equal to that of the MOSFET, 30A.

C. The key features of D1 and C1 are that D1 should behave a current and voltage rating able to withstand the maximum voltage and current within the circuit and also have a fast reverse recovery time. C1 should be sufficiently large to give an output ripple voltage below the specification, and also have a very low equivalent series resistance, which will affect the output voltage ripple as the current ripple into and out of the capacitor has to pass through this series impedance, and it gives rise to heat in the capacitor due to I^2R heating within the device.

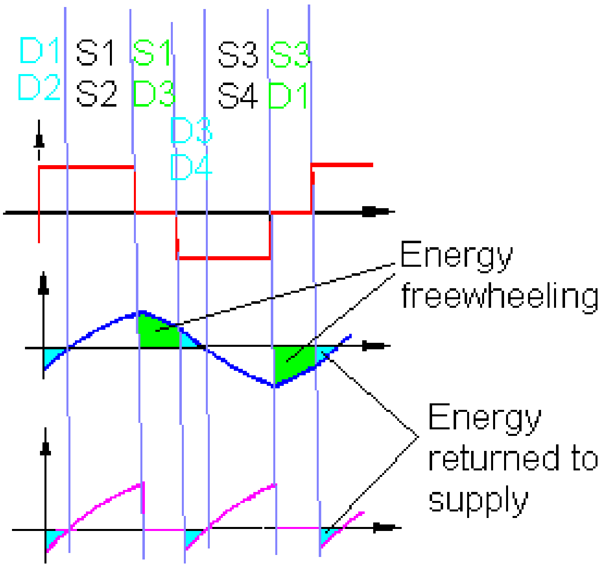
Qu 2.



a. Single Phase Voltage Source Inverter.

Switches are usually MOSFET's, BJT's or IGBT's, inthis case they are MOSFETs..

In operation, each end of the load is switched alternately from the +ve rail of the supply to the -ve



rail. In quasi square wave operation, the inverter legs are switched out of phase by less than 180 degrees. This leads to a time in every cycle when there is no voltage developed across the load, as both ends of the load are connected to the same supply rail, and the load current freewheels through a diode and a switch. This switching strategy leads to a stepped output voltage waveform with lower harmonic content than the square wave output. The device references are with respect to the diagram in the answer.

b. If the load resistor R1=40Ω, and the inductance is zero, we have a resistive load. Under this case of operation, the diodes do not conduct, therefore there will be no losses in the diodes. With a purely resistive load the switching loss can be shown to be

$$P_{AVE\,(switching)} = \frac{V_{CE}.I_C.f.(t_{on}+t_{off})}{6}$$

Therefore the switching loss per device will be:

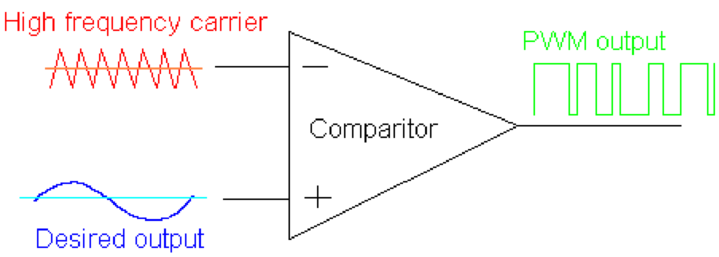
I_c = 400V/40 = 10A, P_{loss} = 6.7W

The conduction loss will be I^2R = 100 × 0.05 = 5W per device × 50% duty cycle per device = 2.5W

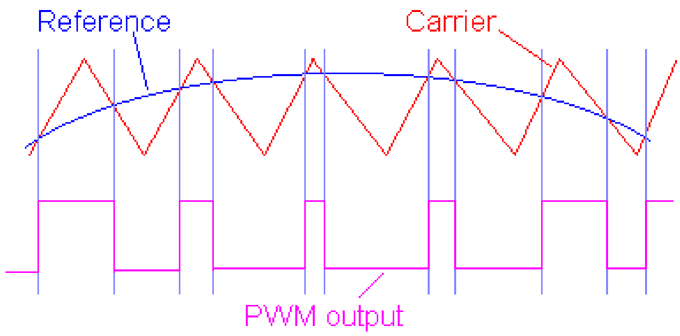
Total loss for 4 devices is therefore (6.7 + 2.5) x 4 = 36.8W

With a 60°C temperature rise for 36.8W, we need a heatsink with R_{th} < 1.63 °C/W

c. To lower the output harmonics further than in a quasi-square wave inverter, pulse width modulation techniques are employed. The basic idea is to switch both ends of the load fast enough for the inductance of the load to filter out the high frequency harmonics. The widths of the pulses are then modulated and produce a sinusoidal current in the load.

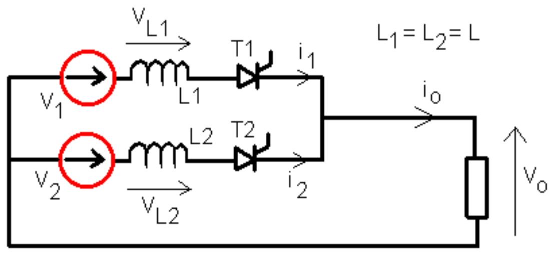


In sinusoidal natural sampling, the required pulse widths are produced by comparing a high frequency triangular carrier wave with a low frequency sinusoidal signal, which is a representation of the desired load current.

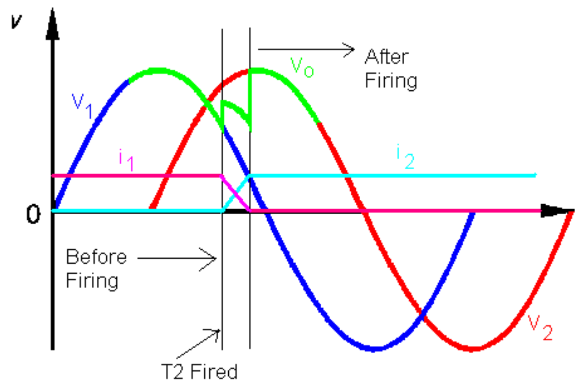


A high ratio of carrier frequency to reference frequency leads to a better the reproduction of the desired current waveform in the load. However, too high a carrier frequency leads to unacceptably high switching losses in the inverter. Therefore a compromise is needed. In commercial inverters, frequencies in the range of 5kHz to 20kHz are used, dependant on the power rating of the drive and the type of switching device used. There are many different PWM modulation strategies in use. Many are now based on digital signal processing (DSP) chips to allow the real-time elimination of certain harmonics in the motor / drive systems.

Qu 3. Commutation Overlap.



a. If we consider two arbitrary phases of a multiphase system, the Current in the supply may only change at a rate dependant on the inductance of the lines, L. Therefore both of the thyristors must conduct for the interval of commutation. This is referred to as commutation overlap.



If the output current is assumed to be constant during the commutation period, then the output current equals the sum of the phase currents:

$$i_o = i_1 + i_2$$

Differentiating we get:

$$0 = \frac{d\,i_1}{d\,t} + \frac{d\,i_2}{d\,t}$$

Therefore,

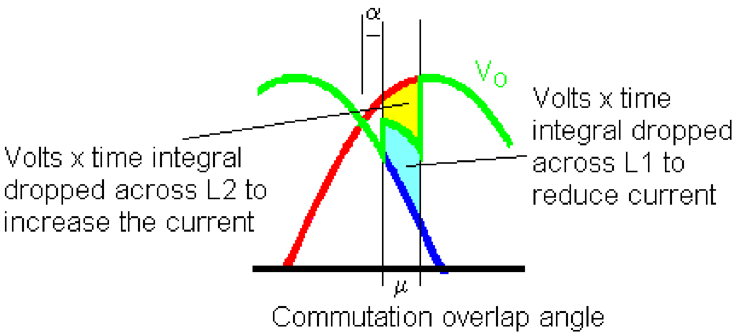
$$\frac{d\,i_1}{d\,t} = -\frac{d\,i_2}{d\,t}$$

The voltage at the output during the commutation event may now be found from both:

$$V_o = V_1 + L\frac{d\,i_1}{d\,t}$$ and $$V_o = V_2 - L\frac{d\,i_1}{d\,t}$$

This leads to the output voltage being the average of the two phase voltages during the commutation event, as shown in the diagram above.

$$V_o = \frac{V_2 + V_1}{2}$$



There is now a corresponding volts x time integral dropped across each phase line impedance to change the current within the respective phase.

Both thyristors continue to conduct until the volts x time integral (shown shaded) is large enough to reduce the current to zero. i.e. as

$$e = L\frac{di}{dt}$$

then

$$L\left(i_1 + i_2\right) = \int_{t1}^{t2} V\,dt$$

Therefore the voltage dropped across the source inductor equals half the line voltage, for a period

$$t_1 = \frac{\alpha}{\omega} \text{ to } t_2 = \frac{\alpha + \mu}{\omega}$$

therefore

$$L_s I_o = \int_{\frac{\alpha}{\omega}}^{\frac{(\alpha+\mu)}{\omega}} \frac{\sqrt{2}V_{line}}{2}\sin(\omega t)\,dt$$

This may be solved and re-arranged to give

$$\mu = \left(\cos^{-1}\left(\cos(\alpha) - \frac{2\omega L_s I_o}{\sqrt{2}V_{line}}\right)\right) - \alpha$$

b. 12V 50Hz, with a supply inductance of 100μH. $R_L = 0.05\Omega$, $I_L = 150A$.

The average output voltage is therefore 150 × 0.05 = 7.5V without the commutation volts drop. Including the commutation volts drop:

$$V_o' = V_o - N\,f\,L_s\,I_o$$ with N=2 for single phase,

So the average output voltage = 7.5 + (2×50×100μH×150) = 9V

For the single phase full converter shown, in continuous conduction:

$$V_o = \frac{2\sqrt{2}V_s}{\pi}\cos(\alpha)$$

Therefore from this the **firing angle = 33.6°**

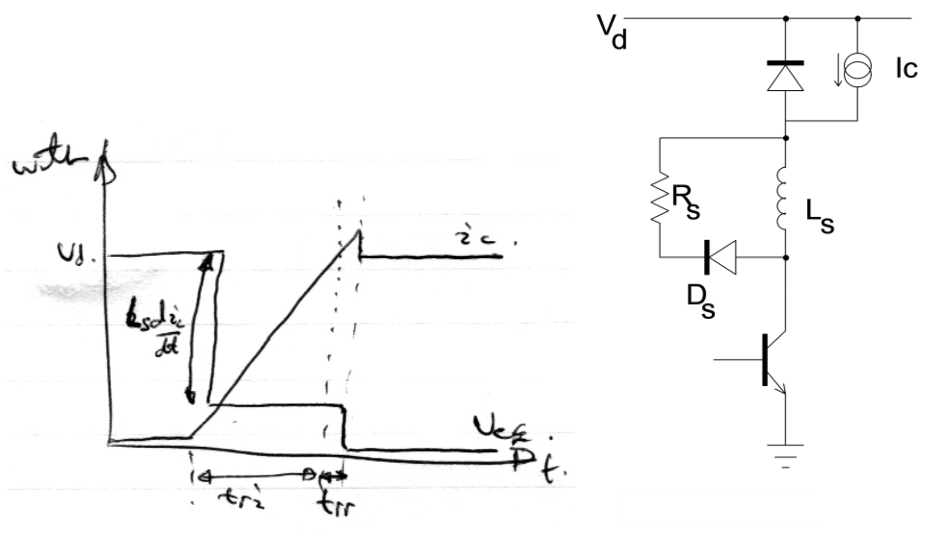
c. If the Thyristors were replaced by diodes in the full converter, we would have a single phase rectifier. The output voltage without the commutation volts drop and zero degrees firing angle (rectifier) becomes 10.8V.

The commutation volts drop is dependent on the output current, which is linked to the average output voltage to rearranging equations gives

$$\bar{V_o}'\left(1 + \frac{NfL_s}{R}\right) = 10.8V$$

Therefore the actual average output voltage = 9V, and the output current = 180A

Qu 4 a. A turn on snubber network is shown below, and the operating waveforms are:



The aim of the turn on snubber circuit is to reduce the voltage seen by the switching during the turn-on event, thus lowering the losses in the device during this period. Without the snubber circuit, the

BJT 'sees' the full supply voltage whilst the current in the device rises to the full load current level. At this point the diode can commutate off, as the current flowing in it is zero, and the voltage across the BJT can then fall. The snubber circuit consists of an inductor, a resistor and a diode. The inductor is in series with the switching device, and drops some of the voltage seen by the switching device whilst the current in it is rising. The voltage dropped across the inductor L_s whilst the current is rising is

$$\Delta V_{ce} = \frac{L\, I_o}{t_{ri}}$$

During the off time, the energy stored in Ls ($0.5L_sI_o^2$) is dissipated in R_s with a time constant for the current decay of $\tau=L_s/R_s$.

The choice of R_s therefore becomes:

A) During turn off, the snubber creates an over-voltage of $\Delta V_{ce(max)} = R_sI_o$

B) During the off period, the current in the snubber inductor L_s must decay to a low value (~$0.1 \times I_o$) such that the snubber circuit may be effective at the next turn-on event.

$$\therefore t_{off\ state\min imum} \geq \frac{5L_s}{R_s}$$

b. Resistor value = 40V/10A = 4Ω. The inductor is given as above and comes out at L<1.6μH. The power in the resistor at 40kHz is the energy in the inductor per cycle, therefore P=0.32W, therefore a 0.5W resistor would be adequate.

c. The choice of a BJT as the switching device at 40kHz may be wrong, as this would be slightly above what would normally be expected to be the operating frequency range for BJT's in power applications due to excessive turn-off times normally associated with the devices. A MOSFET would be a more normal choice. If this were the case, a MOSFET would not need a turn-on snubber network. The MOSFET has a square safe operating area and also, it has a transient current rating which far exceeds its steady state current rating and therefore won't require a turn on snubber network.