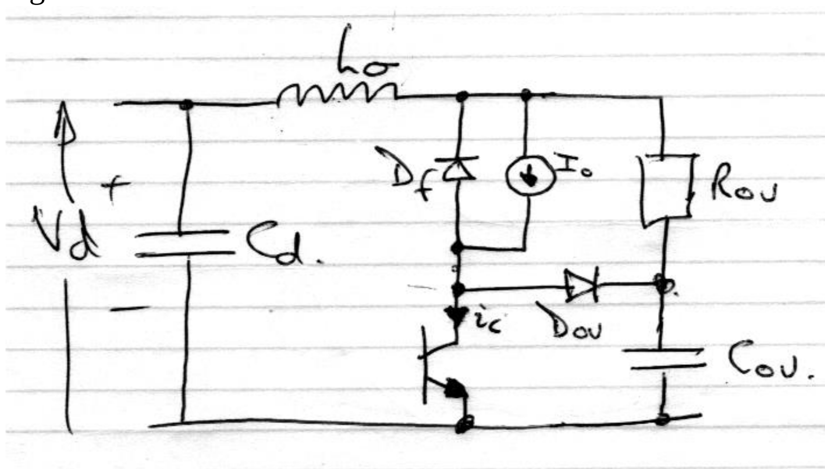
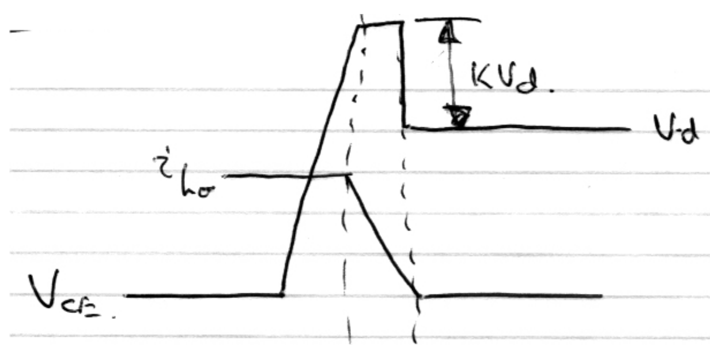
EEE307 2014 / 2015 answers - CRG

Qu 1.

a). Over-voltage snubber network.



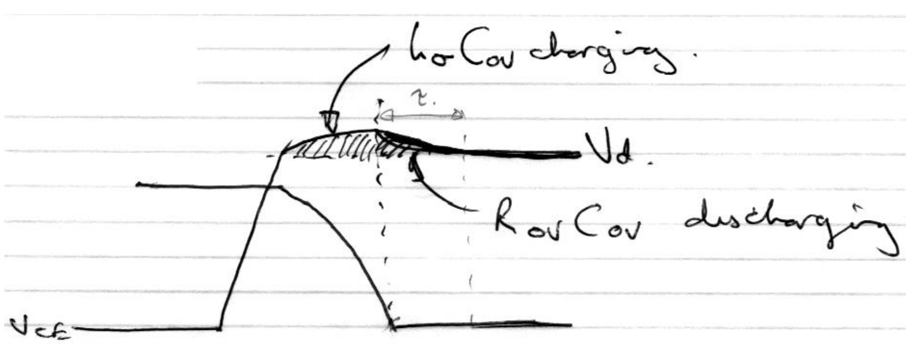
Without the over-voltage snubber:



Without the over-voltage snubber, the voltage across the switching device rises to $(1+k)V_d$, during switch off. This may cause device destruction is the $V_{ce(max)}$ is exceeded for the device.

$\mathbf{i_{L\sigma}}$ = the current in the lumped circuit stray inductance.

With the over-voltage snubber:



Initially the BJT is on, and the voltage across C_{ov} is V_d, the diode D_{ov} is reverse biased. At turn off of the BJT, the current in L_σ is essentially I_o. the BJT current i_c decreases to 0 and the load current free-wheels though the freewheel diode D_f. The energy stored in L_σ is now transferred via D_{ov} to C_{ov} resulting in a small increase in the voltage on the capacitor. (D_fI_o appears as a short circuit and the switch appears as an open circuit to this energy transfer.)

The change in capacitor voltage is then given by an energy balance:

$$\frac{L_\sigma I_o^{\,2}}{2} = \left(\frac{C_{ov} V_{final}^{\,2}}{2} - \frac{C_{ov} V_d^{\,2}}{2} \right)$$

Once the current in the stray inductance falls to zero, and all the energy it contained has been transferred to the capacitor, the final voltage on the capacitor exceeds the supply voltage, and the energy is dissipated harmlessly in the snubber resistor R_{ov} IN this way the energy stored in the stray inductance does not damage the switch. The power rating of the resistor may be found by calculating the energy dissipated in it in each cycle. As the voltage has to have returned to the supply voltage level by the start of the next turn-off event, it is usual to make the time constant τ = 2.2RC less than 1/10th of the time of the switching half cycle.

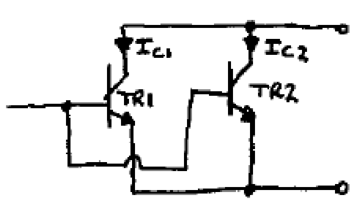
b). From the energy balance equation:

$$\frac{L_\sigma I_o^{\,2}}{2} = \left(\frac{C_{ov} V_{final}^{\,2}}{2} - \frac{C_{ov} V_d^{\,2}}{2} \right)$$

C_{ov} > 640pF.

Also, from time constant = 2.2RC, and 40kHz switching frequency. The time for a half cycle is 12.5µs, therefore the time constant needs to be less than 1.25µs (i.e. one tenth of half period) to ensure discharge of the snubber network. This leads to R<890Ω. Therefore C=680pF and R=820Ω may be suitable preferred values (time constant = 1.23µs).

c). The parallel operation of BJT's should be approached with caution. The high standards of today's construction leads to the possibility of being able to parallel BJT devices by direct connection of their terminals. Unfortunately, in most cases, this may lead to thermal runaway in one or more of the paralleled devices if further components aren't used to ensure static and dynamic current sharing between the devices.



If I_{C1} is greater than I_{C2}, TR1 gets hotter than TR2. as the conductivity of BJT's increase with temperature, I_{C1} increases giving a positive feedback mechanism, and the device overheats. This is termed thermal runaway. Therefore we need to closely couple the devices on the same heatsink, and externally share the current using emitter resistors to ensure sharing of the current.

Qu 2.

From: $$\dot{i}_L = \frac{dv_i}{L} - \frac{Ri_L}{L} + \frac{CR\dot{v}_o}{L}$$ $$\dot{v}_o = \frac{i_L}{C} - \frac{v_o}{CR}$$ if we perturb the input voltage, the output voltage and the inductor current we get $\dot{I}_L + \dot{i}_L = \frac{dV_i}{L} + \frac{d\tilde{v}_i}{L} - \frac{RI_L}{L} - \frac{R\tilde{i}_L}{L} + \frac{CR\dot{V}_o}{L} + \frac{CR\dot{\tilde{v}}_o}{L}$

Now, splitting into steady state and transient terms, the transient terms give:

$$\boxed{\dot{i}_L = \frac{d\tilde{v}_i}{L} - \frac{R\tilde{i}_L}{L} + \frac{CR\dot{\tilde{v}}_o}{L}}$$

Also, $\dot{V}_o + \dot{\tilde{v}}_o = \frac{I_L}{C} + \frac{\tilde{i}_L}{C} - \frac{V_o}{CR} - \frac{\tilde{v}_o}{CR}$, so extracting the transient terms from this gives:

$\boxed{\dot{\tilde{v}}_o = \frac{\tilde{i}_L}{C} - \frac{\tilde{v}_o}{CR}}$ which may be transformed into the Laplace domain to give:

$$si_L = \frac{d\tilde{v}_i}{L} - \frac{R\tilde{i}_L}{L} + \frac{sCR\tilde{v}_o}{L}$$

Now: $s\tilde{v}_o = \frac{\tilde{i}_L}{C} - \frac{\tilde{v}_o}{CR} \quad = v_o\left(s + \frac{1}{CR}\right) = \frac{i_L}{C}$

$$\therefore v_o C\left(s + \frac{1}{CR}\right).\left(s + \frac{R}{L}\right) = \frac{dv_i}{L} + \frac{sCRv_o}{L}$$

$$\therefore v_o LC\left(s^2 + \frac{s}{CR} + \frac{sR}{L} + \frac{1}{LC}\right) = dv_i + sCRv_o$$

$$\therefore v_o\left(s^2 LC + \frac{sL}{R} + sRC + 1\right) - sCRv_o = dv_i$$

$$\therefore v_o LC\left(s^2 + \frac{sL}{RLC} + \frac{1}{LC}\right) = dv_i$$

Giving what was required:

$$\boxed{\frac{v_o}{v_i} = \frac{d}{LC\left(s^2 + \frac{s}{CR} + \frac{1}{LC}\right)}}$$

b) The value of the inductance can be found from the limit of discontinuous current operation (lowest output current) using:

$$L > \frac{E(1-\delta)}{2I_o}\delta T \qquad (\delta = \text{Vo/E})$$

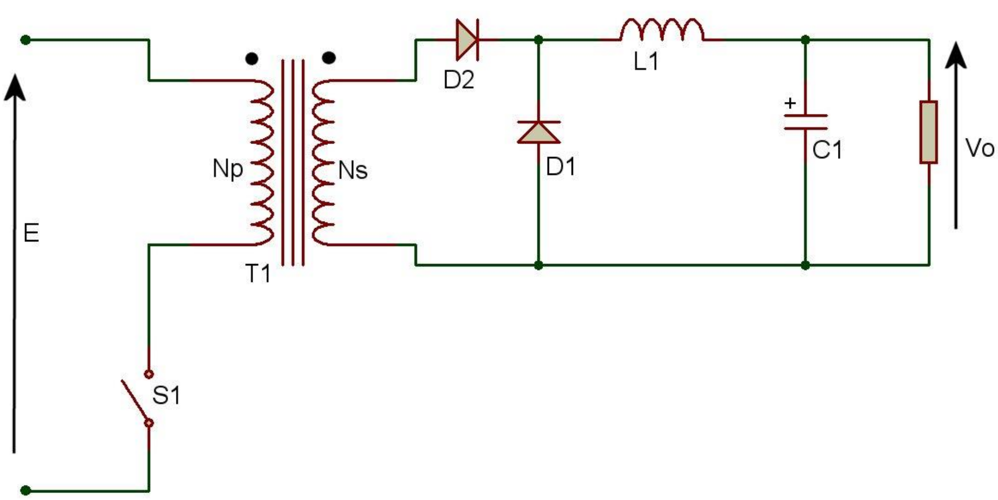
Therefore L > 400µH (36V) or > 600µH (48V), so must be L=620µH (+10% allowable) as 400µH at 48V gives wrong inequality for 0.5A.

The capacitance can be found from:

$$\frac{\Delta v_o}{V_0} = \frac{1}{8}T^2\frac{(1-\delta)}{LC}$$

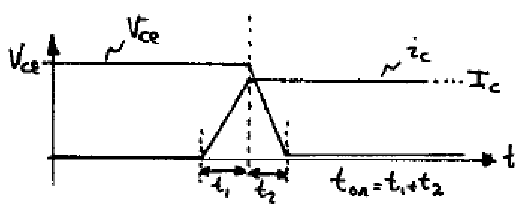
Therefore if L = 620uH, C = 16.8µF (36V) or 25.2µF (48V), so must be 27µF preferred (33µF allowable) or ripple >1%.

c) By introducing an ideal transformer after the switch and before Diode D1 an ideal Forward converter may be realised from a Buck converter. Diode D2 is now required to prevent large circulating currents through D1 when the switch turns off, and it provides a DC output



Qu 3.

a) ***Inductive load:***



An inductive load is more typical of power electronic circuits.
Here, the inductive load keeps the current flowing, via the freewheel diode. The diode cannot start conducting when the BJT switches off until the voltage across the BJT has risen above V_d and forward biased the diode. Until the diode conducts, it cannot provide an alternate path for the load current, so the BJT must carry the full load current whilst the voltage across the BJT rises to forward bias the diode at switch off of the BJT. Similarly, at turn-on of the BJT, the diode cannot

switch off until the current flowing through it falls to zero, therefore the voltage across the BJT is clamped at the power supply rail until the current through the BJT rises to the load current level. Here then, the energy per turn on switching event can be calculated from the following, assuming a linear characteristic:

$$E_{on} = V_{CE}.\int_0^{t_1} \frac{I_C.t}{t_1} dt + I_C.\int_0^{t_2} \left(V_{CE} - \frac{V_{CE}.t}{t_2} \right) dt$$

$$E_{on} = \frac{V_{CE}.I_C}{2}.(t_1 + t_2)$$

$$E_{on} = \frac{V_{CE}.I_C}{2}.t_{on}$$

Now when operating at switching frequency f:-

$$P_{AVE\,(on)} = \frac{V_{CE}.I_C.t_{on}.f}{2}$$

similarly,

$$P_{AVE\,(off)} = \frac{V_{CE}.I_C.t_{off}.f}{2}$$

Average switching loss becomes

$$P_{AVE\,(switching)} = \frac{V_{CE}.I_C.(t_{on} + t_{off}).f}{2}$$

b.) Top Left –33% - Switch conducting.
Bottom Left – 67 % - Diode conducting.
Top Right – 67% - Diode conducting.
Bottom right – 33% - Switch conducting.

During Top-Left and Bottom-Right switching operation as above:
Losses in a single switch at 33% are:
Switching loss = 0.5*40*5*(200ns+300ns)*40kHz = 2W
On-state loss = 5 * 5 * 0.5 * 33% = 4.125W

Single diode on-state loss at 67% = 5 * 0.5 * 67% = 1.675W

During Top-Right and Bottom-Left switching operation:
Losses in a single switch at 67% are:
Switching loss = 0.5*40*5*(200ns+300ns)*40kHz = 2W
On-state loss = 5 * 5 * 0.5 * 67% = 8.375W

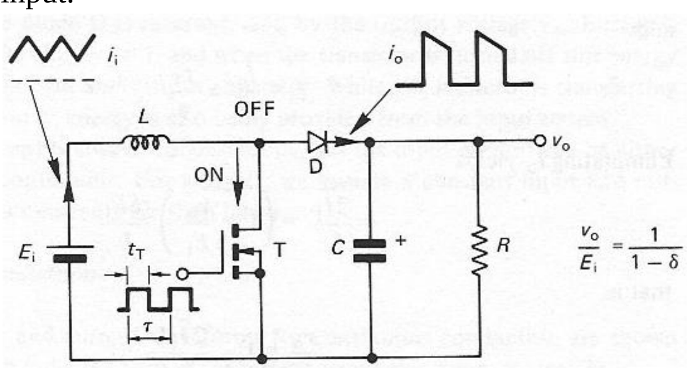
Single diode on-state loss at 33% = 5 * 0.5 * 33% = 0.825W

Total losses in all 8 devices = 2 * (2 + 4.125 + 1.675) + 2 * (2 + 8.375 + 0.825) = 38W

Given a 22°C rise in temperature for 15.6W, the thermal resistance of the heatsink required is < 22/38 = **0.58°C/W**.

c) Given that 100Hz is within the switching speed of a thyristor, the first answer will be yes, HOWEVER, as a thyristors cannot be turned off without a commutation circuit to force the current through the thyristor to zero, then the answer must be NO. Appreciation must be made of the higher switching losses in thyristors over MOSFETs.

Qu 4.
a) The dc-dc converter is a boost converter. The output is always greater than or equal to the input.



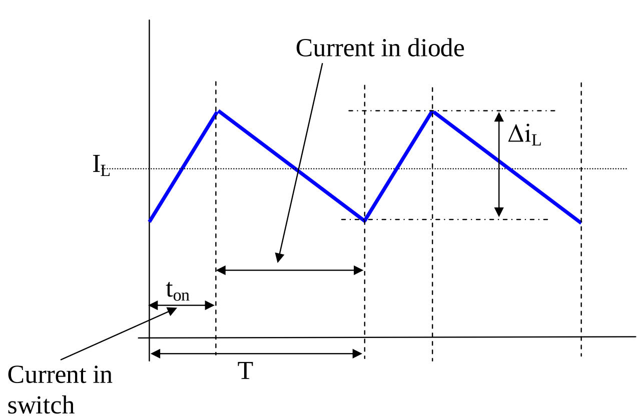
When switch T is on, current in the inductor increases at a rate:

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

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

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

Assuming operation with 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 120V and 300V dc, and the output is 400V with an output current of between 0.1A and 7A, operating at 100kHz, the inductor value is given from:

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

The duty cycle varies from 0.7 at 120V to 0.25 at 300V, and for the 2 conditions at the lowest load current the inductor has to be the greater of 1.26mH and 2.81mH for inequality, therefore:
L=2.81mH (+10% is allowable).
For the capacitor:

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

Therefore with the ripple = 2% of 400V = 8V, at maximum output current (7A), the capacitor will be the greater of 2.19uF and 6.13uF, therefore a preferred value of
C = 6.8uF (preferred) would be chosen to achieve <1% ripple.
The MOSFET should have a breakdown voltage >400V, therefore perhaps choose a 600V device, with a current rating of greater than $I_L = \frac{I_o}{(1-\delta)}$, so for a 7A output with a duty cycle of 0.7, the MOSFET current will be > 23.33A, so choose a current rating of 30A or greater.

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 have a current and voltage rating able to withstand the maximum voltage and current within the circuit (600V and 30A ratings) 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.