E3.14
AO7

IMPERIAL COLLEGE LONDON

DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING
EXAMINATIONS 2010

MSc and EEE PART III/IV: MEng, BEng.and ACGI

POWER ELECTRONICS AND MACHINES

Tuesday, 18 May 10:00 am

Time allowed: 3:00 hours

There are SIX questions on this paper.

Answer FOUR questions.

All questions carry equal marks.

Any special instructions for invigilators and information for candidates are on page 1.

Examiners responsible	First Marker(s) :	T.C. Green, P.D. Mitcheson
	Second Marker(s) :	P.D. Mitcheson, T.C. Green

Answers

1.

Figure Q1 shows a double-switched flyback switch-mode power supply (SMPS).

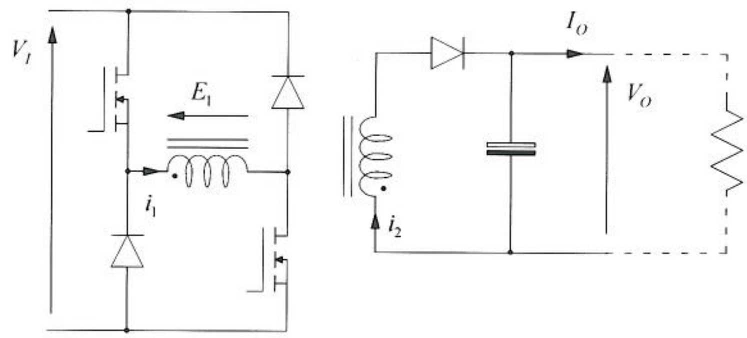


Figure Q1.1 A double-switched flyback SMPS

Comment: This question as a whole was poorly handled by the class despite the topic appearing in the last two exams and in the coursework

(a)

(i) Discuss the choice of the double-switched flyback SMPS over the single-switched version. [3]

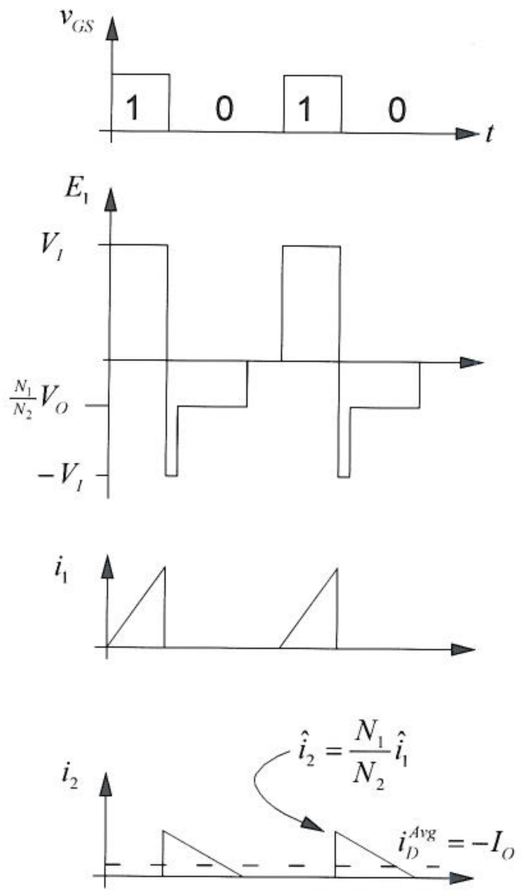
[Interpretation of material in notes]
Mention to be made of primary leakage inductance preventing instantaneous commutation of current from primary to secondary and the trapped energy in the leakage inductance causing overshoot of the switch voltage at turn-off. Doubled-switched circuit provides diode path for recovery of this energy. Circuit uses two switches rated that the input voltage. Single-switch circuit has switch rated at input plus reflected output voltage.

Comment: the answers were sometimes very poorly expressed. The most common technical error was in saying there was no path for the leakage flux in the single-switched version. There is path for the flux; what there isn't is a path for a corresponding current once the MOSFET is switched off.

(ii) Sketch waveforms of primary current, secondary current and primary voltage for discontinuous operation. [4]

[Bookwork but extended to double-switched will mean difference in primary voltage]

[3.14]



Comment: surprising poor diagrams drawn, perhaps because there isn't an exact example in the notes and some interpretation was needed. The glaring errors were to show trapezoid currents that one expects for continuous operation rather than the triangles of discontinuous operation. It must also be obvious that i2 must reach zero before the end of the off-time. Most answers showed the primary voltage as uni-polar; it must be bi-polar if the core is to be both charged and discharged.

(iii) Discuss the choice of operation in discontinuous mode rather than continuous conduction mode. [3]

Discontinuous operation allows the design to use a lower stored energy in the core and hence a physically smaller core. This is because all of the stored energy is transferred to the output at each cycle whereas in continuous mode, only a fraction is and to maintain the same throughput, more energy must be stored. Discontinuous mode also means that the MOSFET turn on into a zero-current condition which means very low turn-on switching loss and no diode reverse recovery power loss.

(b) A flyback SMPS has been designed with a primary inductance of 20 μH and a secondary inductance of 100 μH and specified to operate at an input voltage of 24 V with duty-cycles up to 40% and a switching frequency of 25 kHz.

(i) Calculate the power throughput can be achieved at maximum duty-cycle. [3]

[3.14]

[Calculation from known equations]

$$\frac{di}{dt} = \frac{V_1}{L_1}$$

$$\hat{i} = \frac{V_1}{L_1} t_{on}$$

$$E = \tfrac{1}{2} L_1 \hat{i}^2 = \tfrac{1}{2} \frac{V_1^2}{L_1} t_{on}^2$$

$$P = f_{Sw} E = \tfrac{1}{2} \times 25 \times 10^3 \times \frac{24^2}{20 \times 10^{-6}} \times \frac{0.4^2}{\left(25 \times 10^3\right)^2} = 92.16W$$

Comment: most answers were accurate. Common mistakes were to confuse the peak current and average current and therefore incorrectly calculate power.

(ii) Calculate the maximum allowable ESR for an output voltage ripple of 0.1V with $V_O = 40$V and duty-cycle at 40% assuming the ripple across the capacitance to be negligible. [3]

[Development from known equations]

Peak-to-peak voltage ripple will be the peak-to-peak current times resistance. Since the output current is DC, the ripple all comes from the diode (secondary) current.

$$\hat{i}_2 = \frac{N_1}{N_2} \hat{i}_1 = \frac{N_1}{N_2} \frac{V_I}{L_1} t_{on} = \sqrt{\frac{L_1}{L_2}} \frac{V_I}{L_1} \frac{\delta}{f_{Sw}}$$

$$\Delta v_{ESR} = \hat{i}_2 R_{ESR}$$

$$R_{ESR} < \Delta v_{ESR} \sqrt{\frac{L_2}{L_1}} \frac{L_I}{V_I} \frac{f_{Sw}}{\delta} = 0.1 \times \sqrt{5} \times \frac{20 \times 10^{-6}}{24} \frac{25 \times 10^3}{0.4} < 11.6 m\Omega$$

Comment: most common error was to assume continuous operation and calculate current from that. Another common error was to work with average current not peak current.

(iii) Calculate the minimum output voltage that can be used for the circuit to remain in discontinuous conduction. [4]

[Development from known equations]

Discharge of stored energy through secondary governed by applied output voltage. This must complete in less than 60% of the period to make conduction discontinuous.

[3.14]

$$\frac{di_2}{dt} = \frac{-V_O}{L_2}$$

$$V_O = \frac{\hat{i}_2}{t_D} L_2 = \frac{f_{Sw}\hat{i}_2}{(1-\delta)} L_2$$

$$\hat{i}_2 = \sqrt{\frac{L_1}{L_2}} \frac{V_I}{L_1} \frac{\delta}{f_{Sw}}$$

$$V_O = \frac{f_{Sw}}{(1-\delta)} \sqrt{\frac{L_1}{L_2}} \frac{L_2}{L_1} V_I \frac{\delta}{f_{Sw}}$$

$$= \frac{\delta}{(1-\delta)} \sqrt{\frac{L_2}{L_1}} V_I$$

$$= \frac{0.4}{0.6} \sqrt{\frac{100}{20}} \times 24$$

$$= 35.77\,V$$

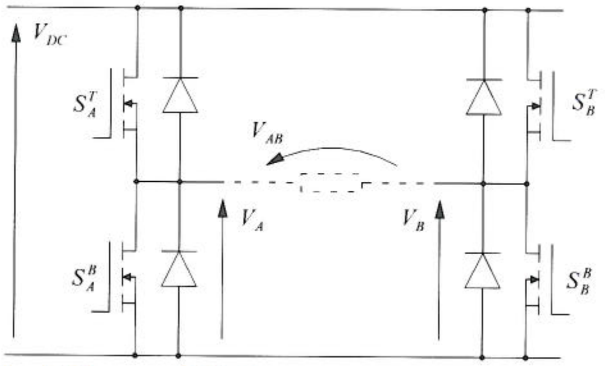
Comment: not everyone attempted this part but, those that did, generally did it well.

[3.14]

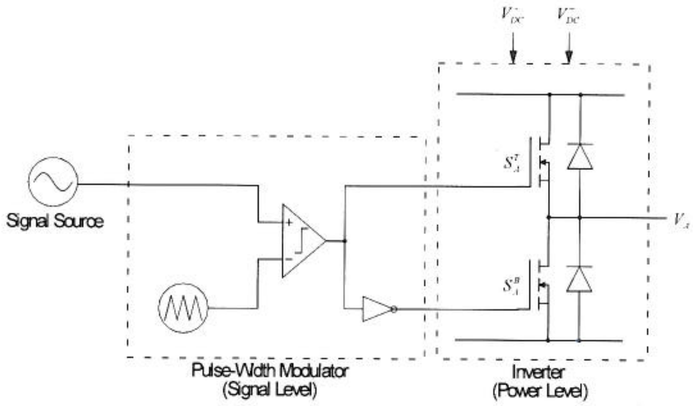
2. Comment: the question is all bookwork and was answered well in general. Some candidates skipped some sections and showed that their revision had been selective.

(a) Sketch the circuit of a single-phase DC to AC converter (inverter) and explain how pulse-width modulation and an appropriate filter can be used to achieve approximately sinusoidal output voltage. [7]

[Bookwork]

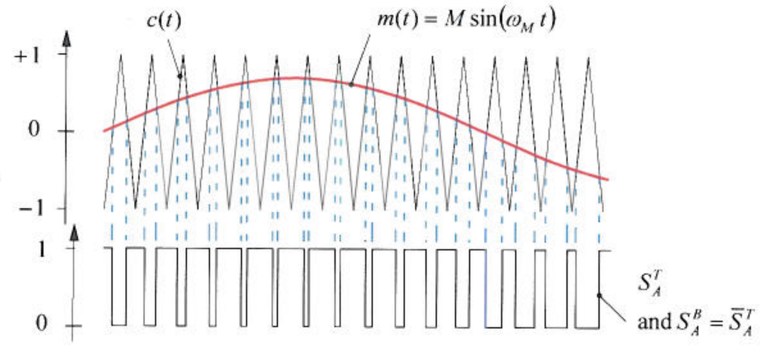


Circuit Diagram of H-bridge *[2]*

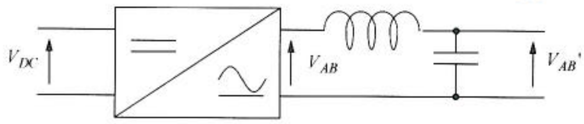


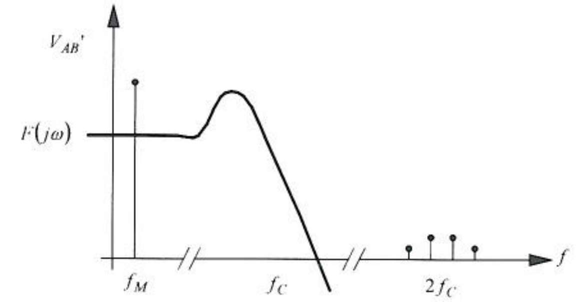
Pulse-width modulator formed of triangle wave source and comparator creates pulse of width proportion to modulating signal, which will be a sinewave. Two halves of the H-bridge run with separate modulators and signal sources in anti-phase. Output is then the difference between the mid-point voltage of the two sides. *[2]*

[3.14]



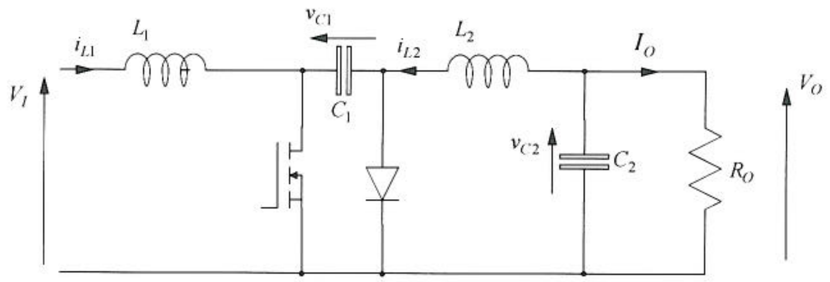
PWM signals from one modulator. Output voltage is simply a pulse-train amplified to VDC. The cycle-by-cycle average is the required sine-wav and imposed on top are carrier and sideband terms at much higher frequencies that can be removed with a filter *[3]*





(b) Sketch a Ćuk SMPS and derive the ratio of output voltage to input voltage for continuous conduction. Explain why a Ćuk SMPS might be preferred over a Buck-Boost SMPS [7]

[Bookwork and interpretation]



[2]

Comment: about 30% of answers showed the diode connected in the opposite sense. This, of course, prevents the flow of current in the expected loops. However, most people went on to give the derivation of the output voltage for the correct circuit even though it wasn't applicable to the circuit drawn. Clearly the derivation was memorised rather than thought through from the circuit diagram.

We assume that the capacitors are sufficiently large such that the voltages across them do not change significantly during a switching cycle. We further assume that the inductors are in continuous conduction and that resistive and semiconductor voltage drops are negligible. The rise and fall of current in the inductors can be found and the assumption of steady-state applied.
Comment: a lack of description of the assumptions made and the basis of the derivation was the most common cause of lost marks.

$$\Delta i_1(on) = \frac{V_I}{L_1}\cdot\frac{\delta}{f} \qquad \Delta i_1(off) = \frac{V_I - V_{C1}}{L_1}\cdot\frac{1-\delta}{f}$$
$$\Delta i_1(on) + \Delta i_1(off) = 0$$
$$\frac{V_{C1}}{V_I} = \frac{1}{1-\delta}$$

$$\Delta i_2(on) = \frac{V_O + V_{C1}}{L_2}\cdot\frac{\delta}{f} \qquad \Delta i_2(off) = \frac{V_O}{L_2}\cdot\frac{1-\delta}{f}$$
$$\Delta i_2(on) + \Delta i_2(off) = 0$$
$$\frac{V_O}{V_{C1}} = -\delta$$

$$\frac{V_O}{V_I} = \frac{-\delta}{1-\delta}$$

[3]

The Cuk and Buck-boost SMPS have the same output voltage characteristic but the Cuk has an extra inductor and an extra capacitor. The advantage these bring is that both the input and output currents flow in inductors and thus if the inductor currents are maintained in continuous conduction then the input and output currents can be designed to have very low ripple values. For EMC compliance and low output voltage ripple the Cuk converter has advantages. The buck-boost converter is a lower cost circuit. [2]

(c) Explain why the method of state-space averaging is useful when designing feedback controllers for SMPS and outline the steps in applying the method. It is not necessary to derive equations. [6]

[Bookwork]

- *The switching action of an SMPS means that they can not be treated as linear and time-invariant.*
- *However, in the on-state and off-state individually, the SMPS is LTI and so overall is piece-wise LTI*
- *If the rate-of-change of the state variables are slow compared to the period of the PWM, then the switching action can be ignored in the control design.*
- *An average model can be formed in which the state evolution over one PWM period is the weighted (by duty-cycle) sum of rates-of-change from the on- and off-states. Thus the A, B, C, D matrices are weighted by duty cycle.*
- *This average model is non-linear because of the product of x and/or u with δ.*
- *The model is linearised about an operating point by expressing each variable as the sum of a steady-state term and a perturbation.*
- *Steady-state only terms and perturbation terms are separated into two models*
- *Products of perturbation are ignored.*

[3.14]

Comment: some answers failed to explain why SMPS are not LTI so the point of SSA was lost. Often the fact that the averaged model is non-linear was not explained and the perturbation analysis was either not discussed or not given proper context.

3. Comment: this question followed a familiar format with a classic example calculation; well-prepared students did well but others did not know the calculation formulae.

(a) Explain why a diode rectifier of the type shown in Figure Q3.1 has a low power factor. [3]

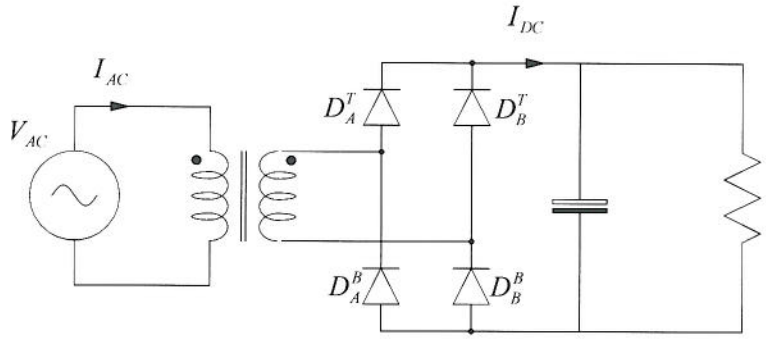


Figure Q3.1 A single-phase diode rectifier

[Bookwork]
The diodes only begin conduction once the AC-side input voltage exceeds the DC-side capacitor voltage. With a large valued capacitor, holding a near constant voltage, this can mean very brief conduction periods of a high current amplitude. Such currents contain harmonics of high amplitude which increase the RMS value of the current above the RMS of the fundamental term. Since only the fundamental current transfers power, the harmonics add to the apparent power but not the real power and thus lower the power factor.

(b) Explain why rectifiers with low power factors are a concern to both the network operators and other consumers on the same network. [3]

[Bookwork]
The harmonic components of current do not transfer real power but do contribute to losses in the supply lines and transformers of the distribution system leading to inefficient operation and this is a concern to network operators.

The harmonic current flows cause harmonic voltage drops in the distribution system and cause other consumers to be supplied with a distorted voltage waveform. This may cause mal-operation of equipment operated by other consumers. EMC regulations, such as EN 61000, prohibit equipment from drawing harmonic current. Network operators have a duty to maintain low amplitudes of harmonic voltage.

(c) A diode rectifier design has been built with two different values of DC-side capacitor. The measured AC-side input current for the two designs is composed of pulses which can be approximated as shown in Figure Q3.2. The circuits were tested with a pure cosine wave voltage applied. Determine the displacement factor and distortion factor with which the two designs operate. Comment on the power factor, real power and EMC properties of the two designs. [10]

[3.14]

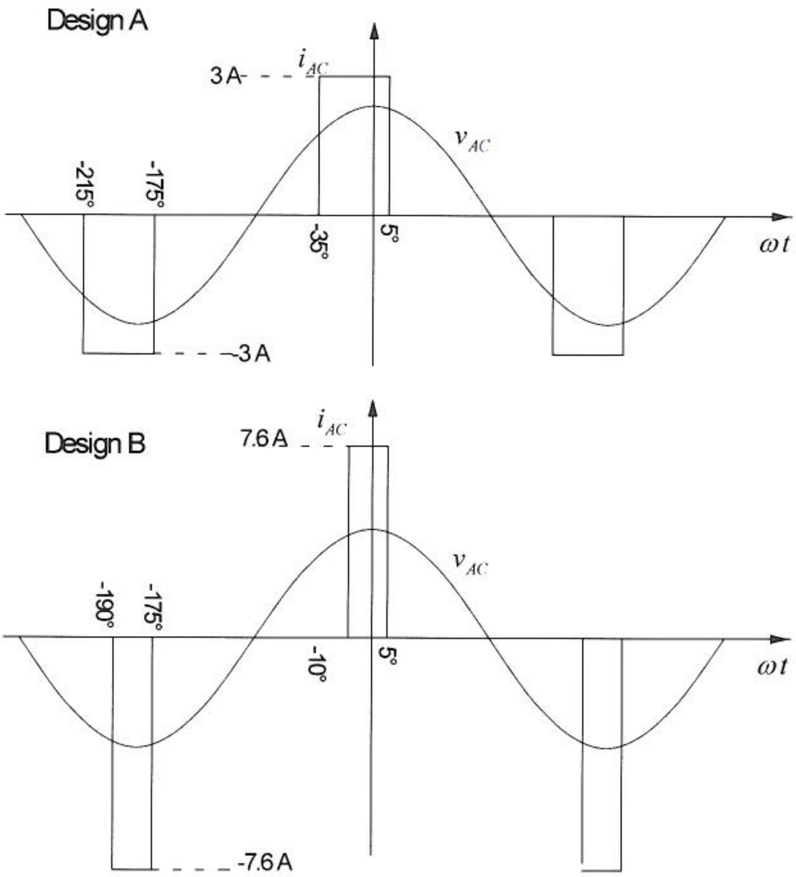


Figure Q3.2 Current waveforms from two designs of single-phase diode rectifier

[Calculation following established method]

Power factor is product of distortion factor and displacement factor.
First task is to find the RMS current and the angle and magnitude of the fundamental

Design A
[2 for RMS current; 2 for Fourier components; 2 for displacement and distortion factors]

RMS Current

$$I_{RMS} = \sqrt{\frac{1}{2\pi}\int_{-\pi}^{\pi} i^2(\omega t).d\omega t} = \sqrt{\frac{I^2}{\pi}\left[\omega t\right]_{-35^\circ}^{5^\circ}}$$

$$= \sqrt{\frac{3^2}{\pi}\left(5 - {}^{-}35\right)\frac{\pi}{180}} = 1.414$$

Fundamental Current

[3.14]

$$I_{A1} = \frac{1}{\pi}\int_{-\pi}^{\pi} i(\omega t)\cos(\omega t).d\omega t = \frac{2}{\pi}\int_{-35°}^{5°} I\cos(\omega t).d\omega t$$

$$= \frac{2}{\pi} I[\sin(\omega t)]_{-35}^{5} = \frac{2\times 3}{\pi}[\sin(5°) - \sin(-35°)] = 1.262\,A$$

$$I_{B1} = \frac{1}{\pi}\int_{-\pi}^{\pi} i(\omega t)\sin(\omega t).d\omega t = \frac{2}{\pi}\int_{-35°}^{5°} I\sin(\omega t).d\omega t$$

$$= \frac{2}{\pi} I[-\cos(\omega t)]_{-35}^{5} = \frac{2\times 3}{\pi}[-\cos(5°) + \cos(-35°)] = -0.338\,A$$

$$I_1 = 1.306\angle -15.0°$$

$$I_{1RMS} = \frac{1.306}{\sqrt{2}} = 0.924 \quad A$$

Displacement factor is $\cos(-15.0°) = 0.966$

Distortion factor is $\frac{I_1}{I_{RMS}} = \frac{0.924}{1.414} = 0.653$

Power factor is 0.631

Design B

RMS Current

$$I_{RMS} = \sqrt{\frac{7.6^2}{\pi}(5 -^{-}10)\frac{\pi}{180}} = 2.194$$

Fundamental Current

$$I_{A1} = \frac{2\times 7.6}{\pi}[\sin(5°) - \sin(-10°)] = 1.262\,A$$

$$I_{B1} = \frac{2\times 7.6}{\pi}[-\cos(5°) + \cos(-10°)] = -0.055\,A$$

$$I_1 = 1.263\angle -2.5°$$

$$I_{1RMS} = \frac{1.263}{\sqrt{2}} = 0.893 \quad A$$

Displacement factor is $\cos(-2.5°) = 0.999$

Distortion factor is $\frac{I_1}{I_{RMS}} = \frac{0.893}{2.194} = 0.407$

Power factor is 0.407

[1 for RMS; 1 for Fourier components; 1 for displacement and distortion factors]]

Design A has the higher power factor. Its distortion factor is much higher (better) although its displacement factor slightly lower. Design A would be better from an EMC point of view because of its lower harmonic distortion.

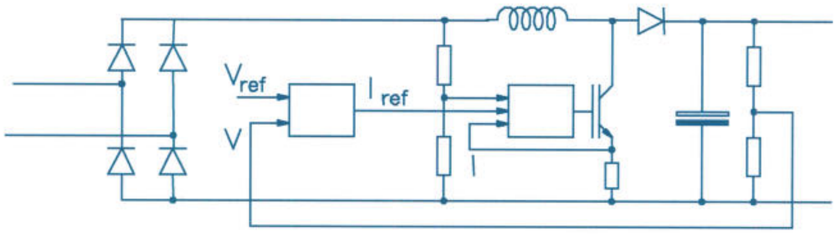
The in-phase fundamental current is the same in both cases indicating they consume the same power from the same source voltage.
[1 for discussion]

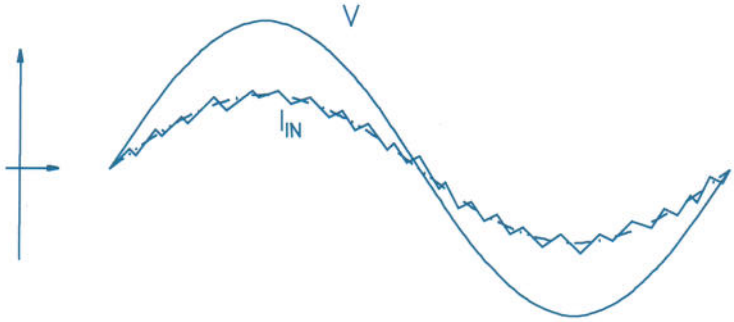
Comment: most common errors were to confuse degrees and radians in the integrations and to divide the Fourier integral by 2π rather than π.

The discussion was generally poor. Points to note are that higher power factor does not directly imply higher power, it depend also on the apparent power. In this example Design A has a higher power factor and the same real power with the implication that the apparent power is lower. The distortion factor is a confusing name since a low distortion factor, in this definition, is a bad thing since it implies that the fundamental is low compared with the total current.

(d) Describe the circuit and control system of a single-phase rectifier able to operate with a near sinusoidal AC-side input current. [4]

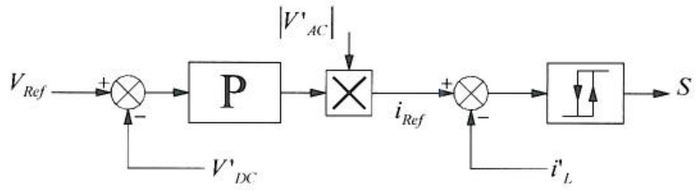
A boost type SMPS can be arranged to draw a controlled current from a varying voltage. A diode bridge is used to rectify the AC voltage and a control loop set to force the current to be a full wave rectified sinewave also (a simple sinewave will be drawn from the AC supply). The magnitude of the sinewave is set by a second controller according to the error between the DC output voltage and its reference value.

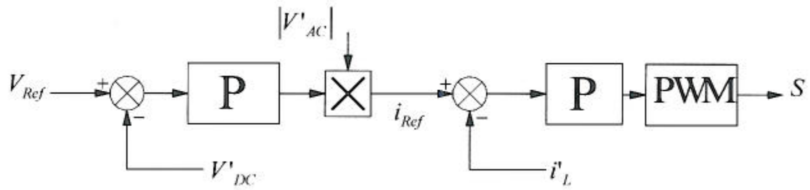




Two variants of control exist. The first uses a hysteresis controller for the current loop and results in fast response but a varying switching frequency. The second uses fixed-frequency pwm. The spread spectrum resulting from a varying switching frequency can make filter *design difficult but it does have the advantage of spreading emissions thinly over a range rather than producing high level emissions concentrated at particular frequencies.*

The control loop for the output voltage must be designed with a limited bandwidth typically around 5 Hz. The energy drawn from a single-phase supply will vary at twice line frequency even if sinusoidal current wave-shape is achieved. Thus, there will be an unavoidable output voltage ripple at twice line frequency which the control loop should not attempt to reject.





4.

(a)

i) One of the fundamental requirements in the manufacture of power semiconductor devices is that maximum use is made of the silicon wafer. Explain how this is achieved in terms of device geometry and doping.

[4]

[bookwork]

Most signal devices are made laterally and only use the surface of the wafer. A power device is built vertically through the wafer, thus allowing the whole volume of the wafer to be used rather than just the surface.

[1]

In order to achieve maximum voltage blocking for a minimum distance between terminals, intrinsic silicon regions are used which allow a flat-topped electric field profile and allow a significant depth of silicon to operate at near the maximum field strength. In contrast, a standard signal device would have a triangular field profile and thus only an infinitely small depth of silicon opera rates at the maximum (before breakdown) field strength.

[2]

In addition, power devices are often created as many individual parallel cells. This allows the gate area to be maximised thus reducing on-state resistance.

[1]

ii) Explain the differences between a power MOSFET and an IGBT. Why is the MOSFET a better device for low blocking voltages and the IGBT a better device at high blocking voltages?

[4]

[bookwork]

The main physical difference between an IGBT and a MOSFET is that an IGBT has an additional p+ layer on the backside of the wafer.

[1]

A power MOSFET is a majority carrier device and thus looks resistive between drain and source when turned on (assuming it is not saturated).

[1]

As the voltage blocking capability of the device increases, the resistance between the drain and source increases as the device must get longer and the doping density of the drift region must decrease. Although the additional p+n junction in the IGBT causes an additional junction drop, conductivity modulation of the drift region causes the resistance of the device to fall.

[1]

Thus, the trade off is for a junction drop plus a small resistance for the IGBT versus a larger resistance for the MOSFET, but without the junction drop.

[1]

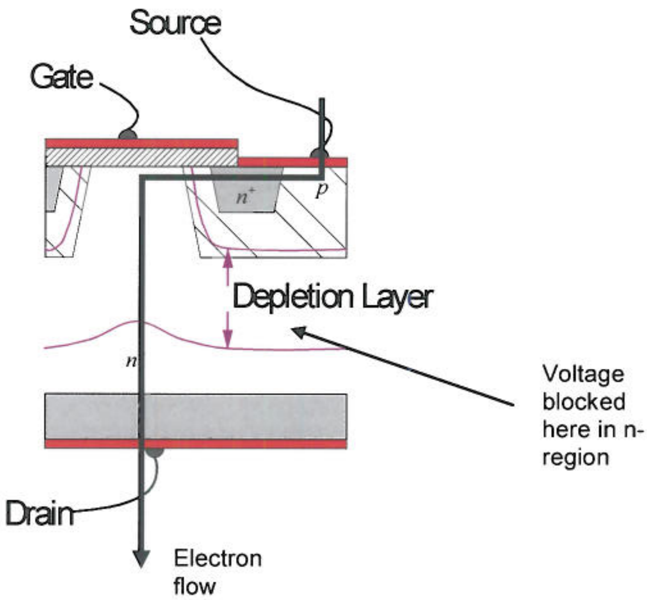
b) A power MOSFET is to be designed to be capable of conducting 100 A and blocking 200 V.

i) Draw a cross section through an *n*-channel power MOSFET and label the region where most of the voltage is blocked when the device is in the off-state and identify this region as either *n*-type or *p*-type. Also, indicate the path of charge carriers when the device is conducting.

[3]

[bookwork]

A cross section through part of a typical Power MOSFET is shown below:



ii) Calculate the length and doping density for the region identified in part (b)(i) so that the device blocks the rated voltage without exceeding the breakdown field strength of silicon.

[3]

[calculation]

For the triangular shaped field profile in the MOSFET, we have:

$$V_{block} = -\int E dx$$

$$= \frac{1}{2} E_{\max} L$$

Thus, L= 20μm.

[3.14]

We must then use Poisons equation to calculate the doping density:

$$E = -\frac{1}{\varepsilon_0 \varepsilon_r} \int \rho dx$$

Thus,

$$\rho = \varepsilon_0 \varepsilon_r \frac{dE}{dx}$$

$$\rho = 8.85 \times 10^{-12} \times 11.7 \times \frac{2 \times 10^7}{20 \times 10^{-6}} = 103.5 C/m^3$$

Therefore the donor doping density is given by:

$103.5/1.6e-19 = 6.47e20/m^3$

or $6.47e14/cm^3$

c) i) A *pin* diode is to be designed to block a voltage of 200 V. What is the minimum required length of the intrinsic region? Explain your answer.

[2]

The field profile in a pin diode in blocking mode is rectangular so only half the length is required than in the MOSFET case where the field profile is triangular. Thus, the length required is 10μm

ii) Explain why the ideality factor of a p^+n diode may approach 2 when used in a power application.

[4]

- A p^+n diode is heavily doped on the p side and therefore there are few electrons in the n side compared to holes in the p side when in steady state
- When in forward bias there is normally electron and hole diffusion away from the junction allowing forward current
- In a power application, high current density is required meaning that the weakly doped n side of the device enters high level injection as the number of injected (minority) holes quickly approaches the number of majority electrons.
- As there is almost no electron current in the p-side of the material (due to poor injection from the low n doping density) there diffusion current of holes must be cancelled by a drift of electrons
- In the limit of high level injection (hole conc = electron conc) the voltage required to cancel the electron diffusion current is the same as that across the junction. Therefore, the voltage creating the concentration gradient for the holes (which are the majority charge carrier) is divided by 2, which is represented by an increase in the ideality factor from 1 (in low level injection) to 2 (in high level injection).

5. a) The circuit of Figure 5.1 shows an IGBT switching an inductive diode-clamped load. The IGBT is switched at 50 kHz with a duty cycle of 1/3.

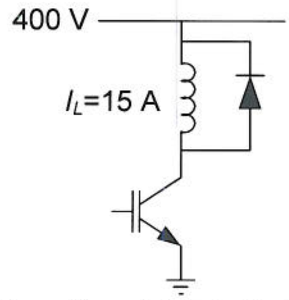


Figure Q5.1 An IGBT used to switch inductive load with diode clamp

The following data is taken from the device datasheet. In the datasheet, the device was switching an switching an inductive load under the same conditions as in Figure 5.1

Voltage fall time [ns]	50
Voltage rise time [ns]	45
Current fall time [ns]	29
Current rise time [ns]	28

i) Calculate the turn-on and turn-off losses of the IGBT in the circuit of Figure Q5.1.

[3]

[calculation]

With an inductive diode clamped load the current/voltage must rise/fall before the voltage/current can fall/rise. Thus, the losses are:

Ploss-on = 0.5(Tir+Tvf)*15*400*20000 = 0.5*(28n+50n)*15*400*50000 = 11.7 W*

Ploss-off = 0.5(Tvr+Tif)*15*400*20000 = 0.5*(45n+29n)*15*400*50000 = 11.1 W*

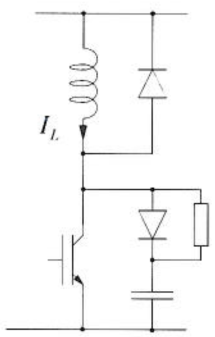
ii) Draw a circuit diagram and calculate component values for a snubber such that the turn-off loss in the IGBT with the inductive load are reduced by a factor of 5.

[5]

[calculation]

The circuit with a turn off snubber added is shown below:

[3.14]

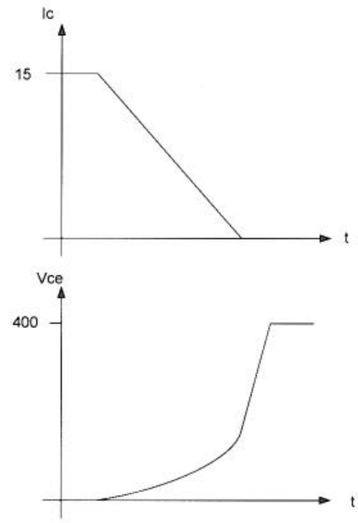


A suitable value of R and C must be calculated:

The IGBT turn off loss must be reduced to 11.1/5 = 2.22 W.

At turn off we assume that the current through the IGBT falls linearly, over 29ns.

The current and voltage graphs are as follows:



From the start of turn-off at t=0, the current falls as:

$$I_{CE} = 15 - 5.17\times10^{8}t$$

The current into the capacitor is:

$$I_{CAP} = 5.17\times10^{8}t$$

Therefore:

$$V_{CE} = \frac{1}{C}\int 5.17\times10^{8}t\,dt = \frac{1}{C}2.59\times10^{8}t^{2}$$

The energy dissipated in the IGBT during turn off is therefore:

$$E = \frac{1}{C}\int(2.59\times10^{8}t^{2})\times(15-5.17\times10^{8}t)dt$$

$$E = \frac{1}{C}\int(3.89\times10^{9}t^{2}-1.34\times10^{17}t^{3})dt$$

$$E = \frac{1}{C}\left[1.30\times10^{9}t^{3}-3.35\times10^{16}t^{4}\right]$$

Evaluated at t=29ns, this gives:

$$E = \frac{8.01\times10^{-15}}{C}$$

Giving a power loss of

$$P = \frac{4\times10^{-10}}{C}$$

If this is to be reduced by a factor of 5, the turn-off loss must be reduced to:

For the factor of 5 reduction, this loss should equal 2.22 W, which gives:

C = 1.59e-9/2.22 = 0.18 nF

The snubber resistor must allow a reset during the on-time of the IGBT. The on-time is 0.333/50000 = 6.7μs

Allowing 5 time constants, this gives 5RC = 6.7μs, or R~7.5 kΩ

iii) What are the total circuit switching losses now that the snubber has been added?

[2]

[calculation]
The turn on loss in the IGBT is the same, i.e. 11.7 W.
The turn off loss in the IGBT is 11.1/5 = 2.22 W
The loss in the snubber resistor due to the discharge of the snubber capacitor is:
*50000*0.5*0.18n*400^2 = 0.7 W (assume charging is lossless as it is through inductor)*

Thus the total circuit switching losses are now 14.6 W.

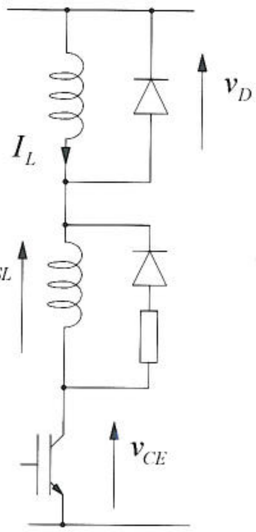
b) A turn–on snubber is now to be integrated into the design.

i) Draw a turn-on snubber circuit and explain the principle of operation by sketching a graph of collector-emitter voltage and collector current during a turn-on event. You may assume that the fall of the collector-emitter voltage is linear and the transistor does not limit the rate-of-rise of collector current.

[3]

[3.14]

[bookwork]
The inductive switched load with a turn on snubber circuit added is shown below



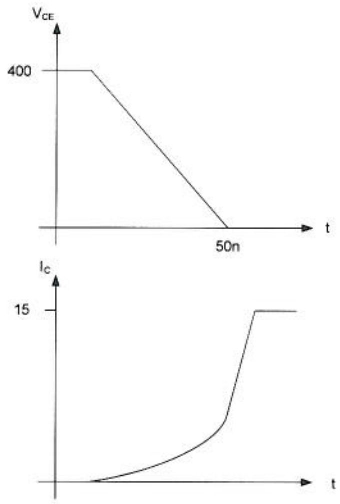
The principle of operation is that with the turn-on snubber present, the voltage across the IGBT can fall without the current rising to a large value. The voltage can thus completely collapse before the current rises to an appreciable value, thus limiting the switching loss in the IGBT.

(ii) Design a turn-on snubber such that turn-on losses in the device are reduced by a factor of 5. [5]

[calculation]
We must now design the inductor and resistor so that the turn-on power loss is reduced to 11.7/5=2.34 W.

The voltage across the IGBT falls linearly and thus the current rises proportional to t^2 until V_{CE} has collapsed at which point the current rises linearly until reaching 15 A:

[3.14]



Taking the start of the turn on time at t=0, the collector emitter voltage is given by:

$$V_{CE} = 400 - 8\times10^{9} t$$

Therefore, the voltage across the snubber inductor is given by:

$$V_{SL} = 8\times10^{9} t$$

And the current through it (and thus through the IGBT) is given by:

$$I_L = \frac{1}{L}\int 8\times10^{9} t\,dt = \frac{1}{L}4\times10^{9} t^2$$

Thus, the power loss in the IGBT is given by:

$$E = \frac{1}{L}\int (4\times10^{9} t^2)\times(400 - 8\times10^{9} t)dt$$

$$E = \frac{1}{L}\int (1.6\times10^{12} t^2 - 3.2\times10^{19} t^3)dt$$

$$E = \frac{1}{L}\left[5.33\times10^{11} t^3 - 8\times10^{18} t^4\right]$$

Evaluated at t=50ns, this gives:

$$E = \frac{1.663\times10^{-11}}{L}$$

Giving a power loss due to turn off of:

$$P = \frac{8.313\times10^{-7}}{L}$$

Therefore ,to reduce the turn on loss to 2.3W4, we must ensure that the snubber inductor has a value of:

[3.14]

L=2.313e-7/2.34=364nH

In order to reset the snubber within 5 time constants, we must have 5L/R=13.3μs:

R=0.136Ω.

It may be necessary to make R larger than this so that the resistor dissipates the energy rather than the .freewheel snubber diode.

iii) What are the total switching losses now that the turn-on snubber has been added?

[2]

[calculation]

The turn off losses stay the same as with the turn-off snubber, i.e. 5.02 W.

Turn on losses in the IGBT have been reduced to 11.5 /5=2.34 W

The additional loss from the energy stored in the snubber inductor is:

*50000*0.5*364e-9*15^2=2.05 W*

Therefore the total switching losses are now:

2.05+2.3+5.02=9.41 W.

6.

(a) Figure Q6.1 shows graphs of the required stator voltage against electrical supply frequency and the torque limit as a function of rotor speed for an induction machine drive.

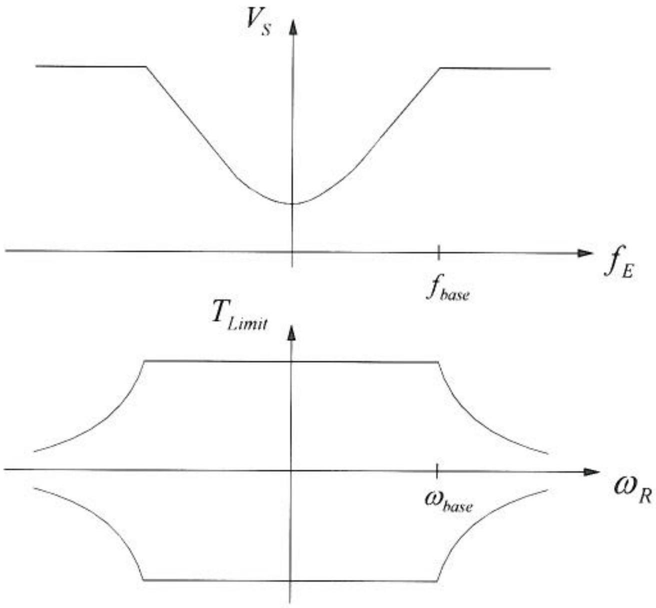


Figure Q6.1 Required stator voltage and maximum torque achieved as functions of speed

(i) Explain why all four quadrants of the torque/speed characteristic may need to be used for a particular application. [2]

An example application is a position servo. Moving from A to B requires first acceleration in the positive direction (Quadrant 1) under positive torque and the deceleration in the forward direction to stop at B under negative torque (Quadrant 4). To return to A requires acceleration in the negative direction under negative torque (Quadrant 3) followed by deceleration under positive torque (quadrant 2).

(ii) Explain how negative torque is achieved at positive speed [3]

Negative torque requires the rotor current to lead the flux vector not lag it, ideally by 90 deg. The phase of the current will reverse if the slip, s, is negative since the resistance term will turn negative. This is achieved by having the rotor speed greater than the synch speed which in turn is achieved by dropping the electrical supply frequency. This gives a retarding torque on the rotor and the negative resistance term indicated that power is (re)generated in the electrical system.

Comment: most common error was to say that reverse torque required reverse rotation of the field.

(iii) Explain the shape of the stator voltage characteristic. [2]

Stator voltage can be expressed as:

$$V_S = R_S I_S + j\omega_e L_S I_S + j\omega_e \psi_{AG}$$

If the objective is to hold the flux linkage constant then a voltage proportional to frequency is required across LM. The voltage drop across the stator winding impedance needs to be added but this is only significant at low frequency. At high frequency, it may not be possible to achieve sufficient voltage from *the inverter because of the limit imposed by the DC-link voltage so the voltage is simply held constant at its maximum value.*

Comment: the use of the equation to explain the shape was important and the main feature of the central section (V proportional to f) needed to be highlighted.

(iv) Explain the shape of the torque characteristic. [3]

Below the limit imposed by the inverter, the flux is constant and this flux multiplied by the maximum rotor current determines the maximum torque available. This is therefore independent of speed an available as positive or negative torque by imposition of suitable slip. At speeds above the inverter voltage limit, the flux-linkage is inversely proportional to frequency/speed and so too is the maximum torque that can be achieved. All of this applies equally to negative frequency/speed.

(b) A 3-phase induction machine with the equivalent circuit shown in Figure Q6.2 is to be operated from a 3-phase inverter with a DC-link voltage of 650 V.

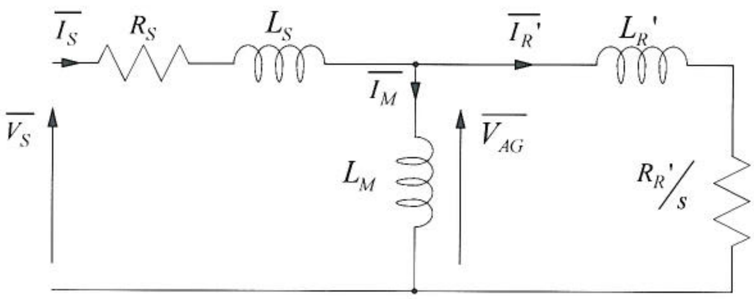


Figure Q6.2 Equivalent circuit of an induction machine.

The circuit parameters of the machine are:

$L_M = 0.04$ H;
$L_S = 0.001$ H;
$L'_R = 0.001$ H;
$R_S = 0.08\ \Omega$;
$R_R = 0.1\ \Omega$;
$P = 1$;

The machine has been designed to operate with an RMS flux-linkage of Ψ(design) = 0.75 Wb and a maximum (referred) rotor current of I'_R = 25 A.

(i) It is suggested that when calculating rotor current and torque at low values of slip, the rotor leakage inductance can be neglected. Comment on this approximation. [2]

At low values of slip, perhaps 4% at rated speed, the rotor branch will be resistance dominated. (R = 0.1/ 0.04; X = 0.001 2 pi * 50) and so the current magnitude and angle are little affected by the inductance. Sample applies at other frequencies (at lower frequency X reduces but percentage slip increases to cancel)*

(ii) Express the rotor current as a function of the slip-frequency ($\omega_{Slip}=\omega_S - \omega_R$) and hence calculate the slip-frequency required (when no field weakening occurs) to achieve maximum rotor current. Calculate the torque that results. [4]

[3.14]

$$\overline{V}_{AG} = j\omega_E \overline{\psi}_{AG}$$

$$I_R = \frac{\overline{V}_{AG}}{R_R} s = \frac{j\omega_E \overline{\psi}_{AG}}{R_R} \frac{\omega_S - \omega_R}{\omega_S}$$

$$\omega_E = P\omega_S = \omega_S$$

$$|I_R| = \frac{P\psi_{AG}}{R_R}(\omega_S - \omega_R)$$

Comment: this was a simple re-arrangement of equations from the notes and was straightforward for well-prepared students.

Assume current and flux-linkage are at 90deg with no inductance in rotor branch

$$(\omega_S - \omega_R) = \frac{I_R R_R}{\psi_{AG}} = \frac{25 \times 0.1}{0.75} = 3.33 rad/s \quad or \quad 31.8 rpm$$

$$T = 3PI_R \psi_{AG} \sin(\delta)$$

$$= 3 \times 1 \times 25 \times 0.75 \times 1 = 56.25 Nm$$

(iii) Calculate the electrical supply frequency at which field weakening begins. [2]

First find maximum phase voltage from inverter in RMS form.

$$V_S = \frac{\frac{1}{2} V_{DC}}{\sqrt{2}} = \frac{650}{2\sqrt{2}} = 229.8 V$$

$$\overline{V}_S \approx j\omega_E \overline{\psi}_{AG}$$

$$\omega_E(base) = \omega_E \approx \frac{V_S}{\psi_{AG}} = \frac{229.8}{0.75} = 306.4$$

(iv) Calculate the slip-frequency required to achieve maximum rotor current at a supply frequency which is twice that found in (iii) and calculate the torque that results. [2]

$$\psi_{AG} = \frac{V_S}{2\omega_E(base)} = \frac{\psi_{AG}(design)}{2} = 0.375 Wb$$

$$(\omega_S - \omega_R) = \frac{I_R R_R}{\psi_{AG}} = \frac{25 \times 0.1}{0.375} = 6.67 rad/s \quad or \quad 63.6 rpm$$

$$T = 3PI_R \psi_{AG} \sin(\delta)$$

$$= 3 \times 1 \times 25 \times 0.375 \times 1 = 28.125 Nm$$

Flux linkage halves, slip doubles and torque halves as expected