Experiment 4: inverter system

- 12 VDC
- Battery
- DC-DC converter
  - Isolated flyback
- DC-AC inverter
  - H-bridge
- AC load
  - 120 Vrms 60 Hz

Step-up dc-dc converter with isolation (flyback)

DC-AC inverter (H-bridge)

Feedback controller to regulate HVDC
Due dates

Right now:
  Prelab assignment for Exp. 3 Part 3 (one from every student)
  Due within five minutes of beginning of lecture

This week in lab (Feb. 19-21):
  Nothing due. Try to finish Exp. 3.

Next week in lecture (Feb. 26):
  Prelab assignment for Exp. 4 Part 1 (one from every student)

Next week in lab (Feb. 26-28):
  Definitely finish Exp. 3, and begin Exp. 4

The following week in lab (Mar. 4-6):
  Exp. 3 final report due
Goals in upcoming weeks
Exp. 4: A three-week experiment

Exp. 4 Part 1:
- Design and fabrication of flyback transformer
- Snubber circuit
- Demonstrate flyback converter power stage operating open loop

Exp. 4 Part 2:
- Design feedback loop
- Measure loop gain, compare with simulation and theory
- Demonstrate closed-loop control of converter output voltage
Exp. 4, Part 3
H-bridge inverter, off grid

- Filtering of ac output not explicitly shown
- IR 3101 half-bridge modules with integrated drivers
- Grid-tied: control $i_{ac}(t)$
- Off-grid: control $v_{ac}(t)$

**Exp. 4 Part 3: off-grid inverter**
- Demonstrate modified sine-wave inverter (required)
- Demonstrate PWM inverter (extra credit)
"Modified Sine-Wave" Inverter

$v_{ac}(t)$ has a rectangular waveform.
Inverter transistors switch at 60 Hz,
$T = 8.33$ msec.

RMS value of $v_{ac}(t)$ is:

$$V_{ac,RMS} = \sqrt{\frac{1}{T} \int_0^T v_{ac}^2(t) \, dt} = \sqrt{D} \, V_{HVDC}$$

- Choose $V_{HVDC}$ larger than desired $V_{ac,RMS}$.
- Can regulate value of $V_{ac,RMS}$ by variation of $D$.
- Waveform is highly nonsinusoidal, with significant harmonics.
PWM Inverter

Average $v_{ac}(t)$ has a sinusoidal waveform
Inverter transistors switch at frequency substantially higher than 60 Hz

- Choose $V_{HVDC}$ larger than desired $V_{ac,peak}$
- Can regulate waveshape and value of $V_{ac,RMS}$ by variation of $d(t)$
- Can achieve sinusoidal waveform, with negligible harmonics
- Higher switching frequency leads to more switching loss and need to filter high-frequency switching harmonics and common-mode currents
The buck-boost converter

Switch in position 1: $V_g$ charges inductor

Switch in position 2: energy stored in inductor is transferred to output

Conversion ratio:

$$\frac{V}{V_g} = -\frac{D}{1-D}$$
The flyback converter:
A transformer-isolated buck-boost converter

See also:
supplementary notes on Flyback converter, Exp. 4 web page
Derivation of flyback converter, cont.

Isolate inductor windings: the flyback converter

Flyback converter having a 1:\textit{n} turns ratio and positive output:
A simple transformer model

**Multiple winding transformer**

\[ v_1(t) - i_1(t) \quad n_1 : n_2 \quad i_2(t) + \]

\[ v_2(t) - i_3(t) \quad \]  

\[ v_3(t) - : n_3 \]

**Equivalent circuit model**

\[ v_1(t) - i_1(t) \quad i_1'(t) \quad i_2(t) + \]

\[ v_2(t) - i_M(t) \quad L_M \]

\[ v_3(t) - : n_3 \]

\[ \frac{v_1(t)}{n_1} = \frac{v_2(t)}{n_2} = \frac{v_3(t)}{n_3} = ... \]

\[ 0 = n_1 i_1'(t) + n_2 i_2(t) + n_3 i_3(t) + ... \]
The magnetizing inductance $L_M$

- Models magnetization of transformer core material
- Appears effectively in parallel with windings
- If all secondary windings are disconnected, then primary winding behaves as an inductor, equal to the magnetizing inductance
- At dc: magnetizing inductance tends to short-circuit. Transformers cannot pass dc voltages
- Transformer saturates when magnetizing current $i_M$ is too large

**Transformer core B-H characteristic**

\[
B(t) \propto \int v_1(t) \, dt
\]

\[
H(t) \propto i_M(t)
\]

\[
slope \propto L_M
\]

\[
saturation
\]
The magnetizing inductance is a real inductor, obeying

\[ v_1(t) = L_M \frac{di_M(t)}{dt} \]

Integrate:

\[ i_M(t) - i_M(0) = \frac{1}{L_M} \int_0^t v_1(\tau) \, d\tau \]

Magnetizing current is determined by integral of the applied winding voltage. The magnetizing current and the winding currents are independent quantities. Volt-second balance applies: in steady-state, \( i_M(T_s) = i_M(0) \), and hence

\[ 0 = \frac{1}{T_s} \int_0^{T_s} v_1(t) \, dt \]
The “flyback transformer”

- A two-winding inductor
- Symbol is same as transformer, but function differs significantly from ideal transformer
- Energy is stored in magnetizing inductance
- Magnetizing inductance is relatively small

- Current does not simultaneously flow in primary and secondary windings
- Instantaneous winding voltages follow turns ratio
- Instantaneous (and rms) winding currents do not follow turns ratio
- Model as (small) magnetizing inductance in parallel with ideal transformer
Subinterval 1

CCM: small ripple approximation leads to

\[
\begin{align*}
  v_L &= V_g \\
  i_c &= -\frac{V}{R} \\
  i_g &= i
\end{align*}
\]

Q₁ on, D₁ off
Subinterval 2

CCM: small ripple approximation leads to

\[ v_L = -\frac{V}{n} \]
\[ i_C = \frac{I}{n} - \frac{V}{R} \]
\[ i_g = 0 \]

Q₁ off, D₁ on
CCM Flyback waveforms and solution

Volt-second balance:
\[ \langle v_L \rangle = D\left( V_g \right) + D'\left( -\frac{V}{n} \right) = 0 \]

Conversion ratio is
\[ M(D) = \frac{V}{V_g} = n \frac{D}{D'} \]

Charge balance:
\[ \langle i_C \rangle = D\left( -\frac{V}{R} \right) + D'\left( \frac{I}{n} - \frac{V}{R} \right) = 0 \]

Dc component of magnetizing current is
\[ I = \frac{nV}{D'R} \]

Dc component of source current is
\[ I_g = \langle i_g \rangle = D(I) + D'(0) \]
Equivalent circuit model: CCM Flyback

\[
\langle v_L \rangle = D(V_g) + D'(\frac{-V}{n}) = 0
\]

\[
\langle i_C \rangle = D\left(-\frac{V}{R}\right) + D'\left(\frac{I}{n} - \frac{V}{R}\right) = 0
\]

\[
I_g = \langle i_g \rangle = D(I) + D'(0)
\]
Step-up DC-DC flyback converter

Need to step up the 12 V battery voltage to HVDC (120-200 V)

How much power can you get using the parts in your kit?

Key limitations:

- MOSFET on-resistance (90 mΩ)
- Input capacitor rms current rating:
  - 25 V 2200 μF: 2.88 A
  - 35 V 2200 μF: 3.45 A

Snubber loss

Need to choose turns ratio, as well as $D$, $f_s$, to minimize peak currents

Possible project for expo: build a better (and more complex) step-up dc-dc converter
Design of CCM flyback transformer
Approach

Use your PQ 32/20 core

Choose turns ratio $n_2/n_1$, $L_M$, $D$, and $f_s$ (choose your own values, don’t use values in supplementary notes)

Select primary turns $n_1$ so that total loss $P_{\text{tot}}$ in flyback transformer is minimized:

$$P_{\text{tot}} = P_{\text{fe}} + P_{\text{cu}} = \text{core loss plus copper loss}$$

Determine air gap length

Determine primary and secondary wire gauges

Make sure that core does not saturate
Core loss
CCM flyback example

B-H loop for this application:

The relevant waveforms:

\[ \frac{dB(t)}{dt} = \frac{v_M(t)}{n_1 A_c} \]

For the first subinterval:

\[ \frac{dB(t)}{dt} = \frac{V_g}{n_1 A_c} \]

Solve for \( \Delta B \):

\[ B = \frac{V_g DT_s}{2n_1 A_c} \]
Calculation of ac flux density and core loss

Fitting an equation to the plot at right

\[ P_{fe} = K_{fe}(\Delta B)^{\beta} A_c \ell_m \]

\( \beta \) = slope

\( K_{fe} \) = constant that depends on \( f_s \)

\( A_c \ell_m \) = core volume

At 60°C:

\( \beta = 2.6 \)

\( K_{fe} = 16 \) (50 kHz), 40 (100 kHz)

with \( P_{fe} \) in watts, \( A_c \ell_m \) in cm\(^3\), \( \Delta B \) in Tesla

From previous slide:

\[ B = \frac{V_g DT_s}{2n_1 A_c} \]

More turns \( \rightarrow \) less \( \Delta B \) \( \rightarrow \) less core loss
Copper loss
Power loss in resistance of wire

Must allocate the core window area between the various windings

Winding 1 allocation \[ \alpha_1 W_A \]
Winding 2 allocation \[ \alpha_2 W_A \]

\[ \alpha_1 + \alpha_2 + \cdots + \alpha_k = 1 \]

Total window area \( W_A \)

Optimum choice:

\[ \alpha_m = \frac{n_m I_m}{\sum_{n=1}^{\infty} n_j I_j} \]

(leads to minimum total copper loss)

The resulting total copper loss is:

\[ P_{cu} = \frac{\rho (MLT) n_1^2 I_{tot}^2}{W_A K_u} \quad \text{with} \quad I_{tot} = \sum_{j=1}^{k} \frac{n_j}{n_1} I_j \]

Choose wire gauges:

\[ A_{w1} \leq \frac{\alpha_1 K_u W_A}{n_1} \]
\[ A_{w2} \leq \frac{\alpha_2 K_u W_A}{n_2} \]
\[ \vdots \]

More turns \( \rightarrow \) more resistance
\( \rightarrow \) more copper loss
Total power loss

\[ P_{tot} = P_{cu} + P_{fe} \]

There is a value of \( \Delta B \) (or \( n_1 \)) that minimizes the total power loss

\[ P_{tot} = P_{fe} + P_{cu} \]

\[ P_{fe} = K_{fe}(\Delta B)\beta A_c \ell_m \]

\[ P_{cu} = \frac{\rho(MLT)n_1^2 I_{tot}^2}{W_A K_u} \]

\[ B = \frac{V_g D T_s}{2 n_1 A_c} \]

**Prelab assignment for next week**: use a spreadsheet or other computer tool to compute \( P_{tot} \) vs. \( n_1 \), and find the optimum \( n_1 \).

Then design your flyback transformer.
Effect of transformer leakage inductance

- Leakage inductance $L_l$ is caused by imperfect coupling of primary and secondary windings.

- Leakage inductance is effectively in series with transistor $Q_1$.

- When MOSFET switches off, it interrupts the current in $L_l$.

- $L_l$ induces a voltage spike across $Q_1$.

If the peak magnitude of the voltage spike exceeds the voltage rating of the MOSFET, then the MOSFET will fail.
Protection of Q1 using a voltage-clamp snubber

- Snubber provides a place for current in leakage inductance to flow after Q1 has turned off
- Peak transistor voltage is clamped to $V_g + v_s$
- $v_s > V/n$
- Energy stored in leakage inductance (plus more) is transferred to capacitor $C_s$, then dissipated in $R_s$

Usually, $C_s$ is large

Decreasing $R_s$ decreases the peak transistor voltage but increases the snubber power loss

See supplementary flyback notes for an example of estimating $C_s$ and $R_s$