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Designing An Open-Source Power Inverter (Part 7): Kilowatt Inverter Magnetics

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The last two parts of this series,[1-6] described the design of circuitry for the inverter stage (INV401) of the 1.2-kW Volksinverter design. This part shows how to design the output inductor for this stage. Specifically, it is a coupled inductor having two identical windings, and forms an LC filter with an output capacitor of 0.2 µF, 200 V as depicted by X1 and C7 in Fig. 1. This coupled inductor rejects noise and prevents overcurrent of the output waveform.

This design begins with a set of min and max constraints on the range of inductance values for this component, as established earlier in the series. From there a saturation model derived by the author enables us to identify the most suitable core material based on cost and size considerations. Core material selection then leads to details of core size selection and winding design including number of turns, wire size and wire bundling, number of layers and required wire length. This part 7 concludes with an explanation of assumptions made in the design of the inductor, which may differ from what might be expected in other power supply designs.

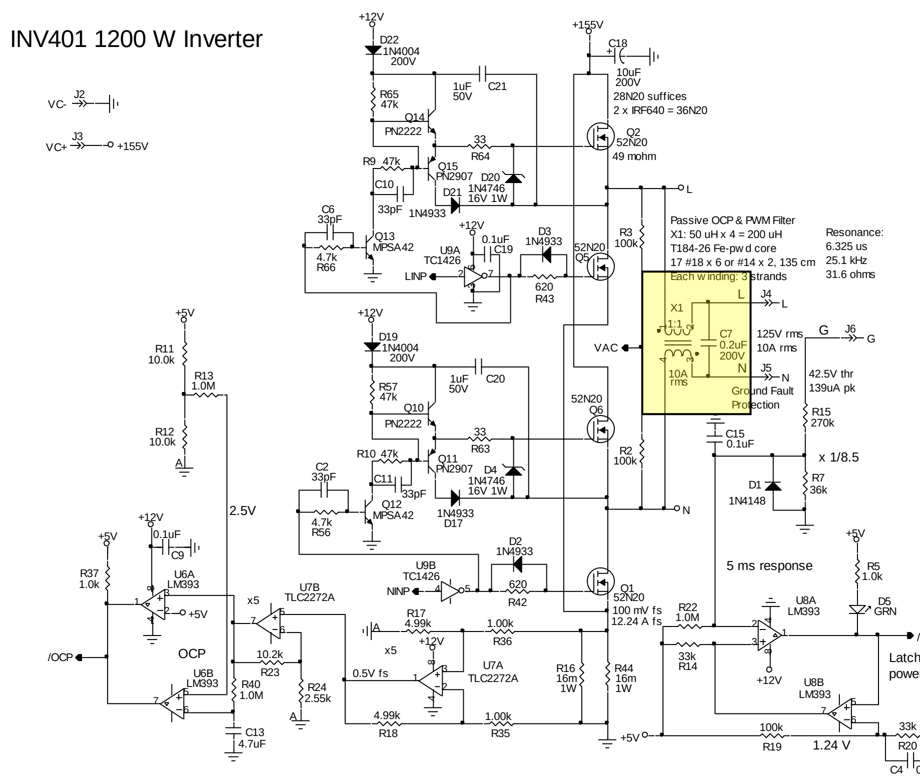

Fig. 1. INV401 inverter power-transfer circuitry. The coupled inductor X1 forms an LC filter with output capacitor C7 that rejects noise and provides overcurrent protection.

Using The Saturation Model

In part 5, it was determined that the inductor value at 25 A should be no less than 40 µH, though the absolute value as calculated was 36.44 µH. To keep attenuation of the output waveform to ≤0.1%, inductance should be no more than 40 mH. The steady-state full-scale RMS current is 10 A. Designs for several possible core choices demonstrate the design procedure.

Power inductors have inductance that decreases with current because of saturation. The Innovatia saturation model is shown in Fig. 2.[7] To determine the zero-current (maximum) inductance, we want to choose a k_{sat} that maximizes the inductance at 25 A, where

$$k_{sat} = \frac{\mu}{\mu_i} = \frac{\mathcal{L}}{\mathcal{L}_0} = \frac{L(I)}{L(0\,\mathrm{A})}$$

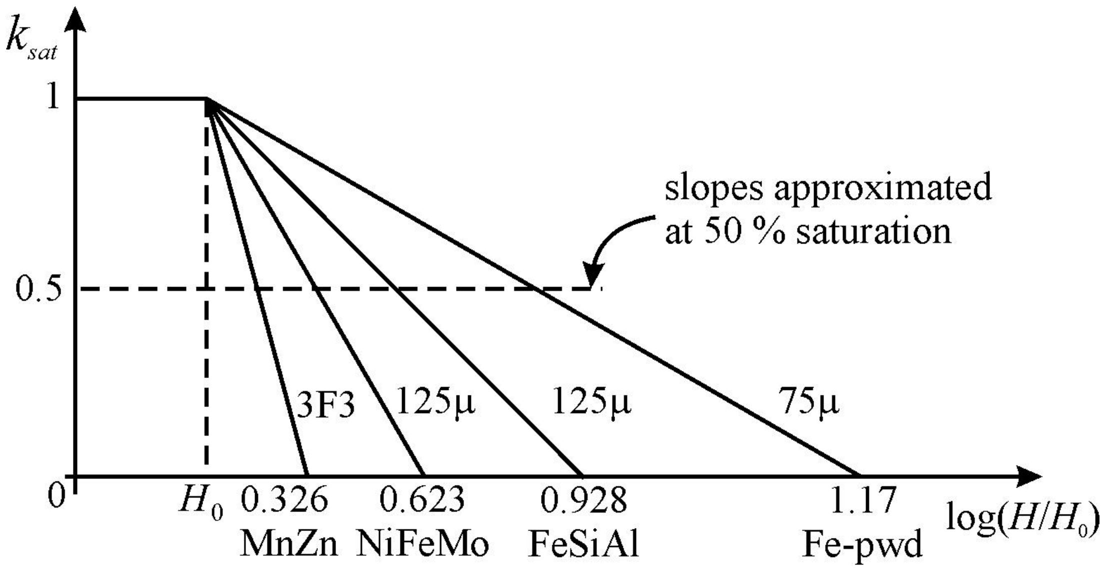

Fig. 2. Semi-log graph plotting the fractional saturation (k_{sat}) curves for various materials approximated around k_{sat} = 0.5, a typical minimum operating-point for power inductors. Fe-pwd is Micrometals 26 material.

In the model, the values shown at k_{sat} = 0 for each curve represent the H_T values for each of the materials where the core is so completely saturated that it has no inductance. This is an extrapolated model parameter in that real inductors always have some inductance.

H_0 is another extrapolated parameter, the asymptotic breakpoint at the onset of saturation. H_0 is found on the graph by linearly extrapolating a line tangent to the plot at k_{sat} = 0.5. Where it intersects k_{sat} = 1 is H_0 as shown in Fig. 3 on the catalog curve for Micrometals 26 material. For $H < H_0$, k_{sat} = 1 in the saturation model. It is like a breakpoint on a frequency-response plot.

To keep the cost low, a minimum-cost core material is chosen: the Micrometals iron-powder (Fe-pwd) 26 material. (The 26 cores are color coded yellow and white.) By choosing this material, cost is minimized at the expense of volume. Kilowatt-size off-grid inverters are seldom constrained by size, making this a preferred tradeoff.

Board area for the component, however, might constrain it, as we will see. The catalog 26 material saturation graph is copied in Fig. 3 with the saturation-model line constructed on it, to show how the Fig. 2 plots relate to catalog curves. From the graph,

$$k_{sat} = 1 - \frac{\log H - \log H_0}{\log H_T - \log H_0} = 1 - \frac{\log(H/H_0)}{\log(H_T/H_0)} = \frac{\log(H_T/H)}{\log(H_T/H_0)}\ ,\ H \geq H_0$$

Some of the model parameters for various core materials are given in Table 1 (from reference 8).

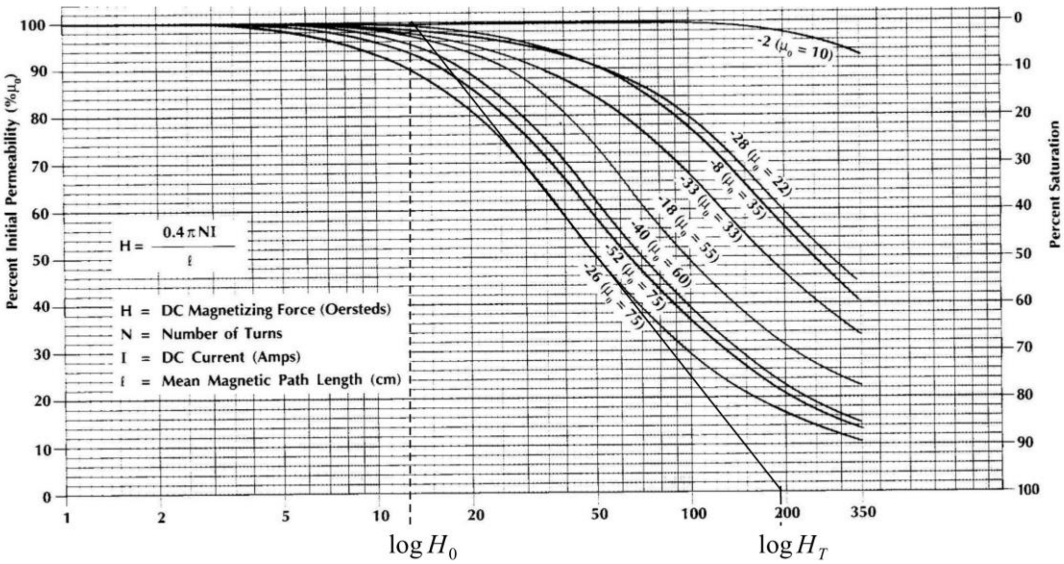

Fig. 3. Micrometals iron-powder core material fractional saturation, $k_{sat}(H)$ graph from a Micrometals catalog. k_{sat} is the quantity graphed on the left side, in percent. (The right side is 1 – k_{sat} in percent.) The asymptotic value at k_{sat} = 1 (no saturation) has its breakpoint at H_0 and intersects k_{sat} = 0 (full saturation) at H_T. The H values on the horizontal axis are in oersteds, where 1 Oe ≈ 80 A/m.

Table 1. Saturation model parameters for various core materials.

Material	μ_r	H_0, A/m	$H(0.5)$, A/m	H_T, A/m	$\log\left(\frac{H_T}{H_0}\right)$
Fe-pwd (26)	75	1035	3980	15305	1.17
FeSiAl	125	1100	3200	9309	0.928
NiFeMo	300	800	1740	3785	0.675
	125	2100	4300	8805	0.623
MnZn 3F3	2000	24.5	35.7	52.02	0.326

Maximization of inductance is the driving design parameter. The design equation[8] gives the number of winding turns that maximizes inductance at a given current operating-point of I with core magnetic path length l;

$$\boxed{N_{opt} = \frac{H_T \cdot l}{I \cdot \sqrt{e}} \approx 0.6065 \cdot \frac{H_T \cdot l}{I}}$$

where H_T is given in Table 1 as found from a Fig. 3 asymptotic plot. The maximized inductance is

$$\boxed{L_{max} = \left(\frac{H_T \cdot l}{I}\right)^2 \cdot \frac{\mathcal{L}_0}{\log(H_T / H_0)} \cdot \frac{\log \sqrt{e}}{e} \approx \left(\frac{H_T \cdot l}{I}\right)^2 \cdot \frac{\mathcal{L}_0}{\log(H_T / H_0)} \cdot (0.07988)}$$

For the inductor design, we have L_{max} = 40 µH at $\hat{i}$ = 25 A as given, and H_T = 15305 A/m, $\log(H_T/H_0)$ = 1.17 from Table 1. The core field inductance $\mathcal{L}_0$ and path length l are dependent on the core size. The design formula for saturation reduces to

$$\boxed{k_{sat} = \frac{\log\sqrt{e}}{\log(H_T/H_0)}}$$

Substituting known values,

$$k_{sat} \approx \frac{0.217}{1.17} \approx 0.186$$

In ordinary design practice, this is a low value of k_{sat}. The typical design range for k_{sat} is 0.5 to 0.7. At the optimal k_{sat},

$$N_{opt} \approx 0.6065 \cdot \frac{H_T \cdot l}{I} = (0.6065) \cdot \frac{(15305\ \text{A/m})}{25\ \text{A}} \cdot l = \frac{l}{2.693\ \text{mm}}$$

$$L_{\max} \approx (0.07988) \cdot \left(\frac{H_T \cdot l}{I}\right)^2 \cdot \frac{\mathcal{L}_0}{\log(H_T/H_0)} = (25588\ \text{m}^{-2}) \cdot (\mathcal{L}_0 \cdot l^2)$$

Core size also constrains winding area of the winding design. From a wire table[9], a wire with the ampacity to handle the full-scale RMS current of 10 A rms is 14 AWG (or #14), or #18 × 3 (3 strands) or #21 × 5. The turn area, taking into account the enlargement of the packing factor, is $A_{cwp} = A_c/k_p$ where A_c is the conductor area of the wire or bundle and k_p = packing factor. The choice of #14 wire has A_{cwp} = 3.021 mm^2 and there are two identical windings (turns ratio, n = 1) totaling 6.042 mm^2.

For field containment, low cost, and low thermal resistance, the toroid core shape is chosen. The next decision is how much of window area A_w can be winding area A_{ww}. The factor to choose is

$$k_{ww} = A_{ww}/A_w \leq 0.75$$

Keeping k_{ww} at or below this maximum makes winding the core a task, not a trial. Some different core sizes and their parameters are given in Table 2. The T106 core alternative stacks three T106-26 cores in alignment and joins them with cyanoacrylate glue. N_w is the maximum number of turns that can be placed in the window for turns comprised of #18 × 3 except the bottom entry.

Table 2. Toroid Fe-pwd 26 cores, parameters and design values.

Core Size	A_w, mm^2	l, mm	$\mathcal{L}_0$, nH	N_w ($k_{ww} \leq$ ¾), #18 × 3	N_{opt}	L_0, µH	L_{max}, µH, 25 A	@ k_{sat}
T131	209	77.2	116	23	29	98	17.7	0.186
T157	456	101	100	50	38	144	26.1	0.186
T184	**456**	**112**	**169**	**50**	**42**	**298**	**54.2**	**0.186**
T201	456	118	242	50	44	469	86.2	0.186
3 × T106	165	64.9	279	55 (#14)	24	160	30.1	0.186

While single wires have an insulated radius of r_{cw}, a bundle of twisted wires requires calculation of its *bundle radius* r_{bw}, the radius to the perimeter of the circle twisted out in cross-section by the bundle as it twists towards us. Then in the winding formulas, r_{cw} of a single wire is replaced by r_{bw}.

What is needed is the packed cross-sectional area, A_{cwp} of a bundle.[10] How tightly the wire is twisted affects bundle diameter. A typical length of bundle twist, or *pitch* of $p = 30 \cdot r_{bw}$ results in an expansion of the bundle area from twisting by × 1.022 (about 2.2%) so that twist packing, k_{tw} = 1/1.022 = 0.9785.[11] Additionally, the bundle radius of three wires, when twisted (′), is

$$r_{bw}' = 1.886 \cdot r_{cw}$$

From a wire table, the #18 wire insulated radius is r_{cw}(#18) = 0.559 mm. Thus, r_{bw}' = 1.054 mm, about the radius of a #13 wire. Then $A_{bw}' = 2 \cdot \pi \cdot r_{bw}'^2 \cdot k_{tw}$ = 6.83 mm^2.

The implemented choice from Table 2 is the T184 core. It has sufficient inductance (54 µH > 40 µH). The circuit range constraints on inductance are wide—about three decades, or 33.6 mH/36.44 µH = 922 ≈ 3 decades. The L_{max}/L_0 saturation range is $1/k_{sat}$ ≈ 5.4 and is easily within the acceptable range. Thus, the optimal choice of design solution has L_{max} ≥ 40 µH but less than 33.6 mH. The T184 choice meets the conditions. Then L_0/33.6 mH ≈ 0.12% with the goal of 0.1%, and attenuation is negligible at both 60 Hz and its third harmonic.

Determining Winding Lengths

To wind toroids, the winding is cut to length before attempting to wind because the entire length of wire must pass through the core window every turn. An accurate winding length formula[12] requires the following T184 core geometric parameters and the twisted bundle radius:

$$r_i = 12.05 \text{ mm},\ w = r_o - r_i = 11.30 \text{ mm},\ h = 18.0 \text{ mm},\ w + h = 29.3 \text{ mm},\ A_w = 456 \text{ mm}^2,\ r_{cw} \rightarrow r_{bw}' = 1.054 \text{ mm} \Rightarrow$$

the spiral length of turns of a layer advancing over full layer length = $2 \cdot \pi \cdot M \cdot (r_i + w/2) = 2 \cdot \pi \cdot M \cdot (17.70 \text{ mm})$

while

max layers = $\hat{M} = \dfrac{r_i}{(1+\sqrt{3}/2) \cdot r_{cw}} \approx \dfrac{r_i}{(1.866) \cdot r_{cw}}$ = 6.127

full-window turns = $N_w = \pi \cdot \hat{M}^2$ = 117.9

layers = $M = \hat{M} \cdot \left(1 - \sqrt{1 - \dfrac{N}{N_w}}\right)$ = (6.127)·(1 − $\sqrt{1 - 42/117.9}$) = 1.211

winding length = $l_w = 2 \cdot \pi \cdot M \cdot [(2 \cdot (h+w) + 8 \cdot M \cdot r_{cw}) \cdot (\hat{M} - M/2) + \frac{4}{3} \cdot r_{cw} \cdot (1 - M^2) + (r_i + w/2)]$ = 3.021 m

The intermediate values given above are stored in calculator memory (or computed) and length is calculated from them. The resulting l_w does not include extra lead length required to connect to the bobbin terminals, and must be added to l_w for both ends.

Each winding is half the turns and half this length, and the two bundles are bifilar wound (as shown in Fig. 4). The two windings are coupled and behave as a single inductor with 42 turns. Each winding thus has half this or 21 turns. The length, l_w is for both windings and is cut in half for each winding. A total length is chosen of about 3.06 m/2 = 1.53 m per winding.

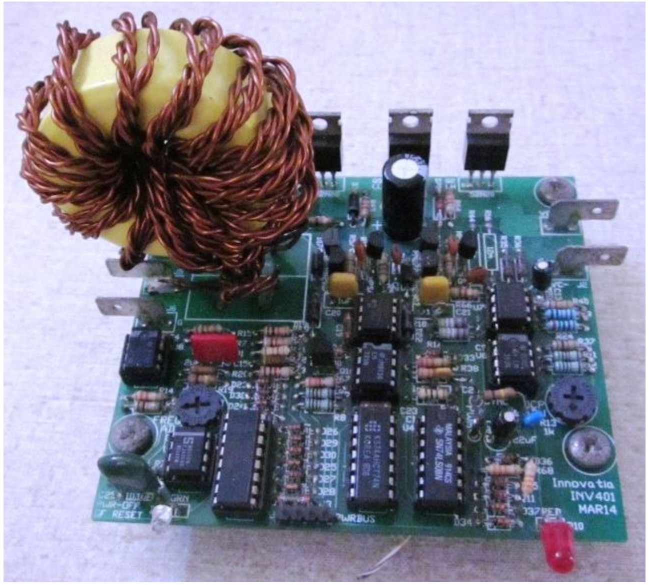

Fig. 4. INV401 inverter stage with output transductor, showing bifilar winding. The six strands of 3 strands × 2 windings are constructed here as a twisted four-strand bundle with a two-strand bundle for easier construction.

To drive an inductor to a saturation of k_{sat} = 0.186 is unusual and is not within the bounds of what is ordinarily considered optimal design practice. In magnetics designs for closed-loop converters where peak current is detected by a comparator, the superlinear current waveform at this saturation can result in loss of feedback control. Furthermore, for power transfer through a coupled inductor, saturation below $k_{sat} \approx 0.5$ is an exercise in diminishing advantage for achieving extra power transfer.

In most converter applications, this reasoning is quite valid, but neither argument applies here; there is no closed control loop and the inverter output inductor does not have the purpose of transferring energy. Indeed, it has a purpose instead of constraining output current until the active OCP loop can shut off the H-bridge switches. The worst-case condition is at maximum output current at which the inductance must be above some minimum to protect the H-bridge.

Its second purpose, as a low-pass filter of harmonics, limits the maximum inductance to avoid filtering out the power waveform itself. The minimum and maximum values are widely separated and allow the design to have a wide saturation range.

What is the lesson to be learned? Beware of the usual assumptions brought to magnetics (or other) design. Do they apply? If not, do not let design habits prevent a design from going "outside the box" to be optimal. This one does and exemplifies the need to remain flexible in design thinking.

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About The Author

Dennis Feucht has been involved in power electronics for 40 years, designing motor-drives and power converters. He has an instrument background from Tektronix, where he designed test and measurement equipment and did research in Tek Labs. He has lately been working on projects in theoretical magnetics and power converter research.

For further reading on inverter design, see the How2Power Design Guide, locate the Power Supply Function category and select "DC-AC power inverters."