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Designing An Open-Source Power Inverter (Part 12): Sizing The Converter Magnetics

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As this series on designing the Volksinverter[1-11] continues, we address another aspect of the battery converter stage. Specifically, we delve into the design of the magnetics in the BCV402 power transfer circuit (Fig. 1).

The power-transfer circuit, shown in Fig. 2, has two magnetic components: a coupled inductor and transformer. It operates normally in CA or boost push-pull (BPP) mode whenever $V_s' > V_g$ but otherwise reverts to CP (buck) mode when output current or voltage is low. (In a later part of this series the design of converter control considers behavior over load (output) current and voltage ranges.) Transformer design is based on the CA mode whereby the transformer transfers the full-scale (f_s) output power and is the normal mode of operation.

In this application, magnetics sizing—determining the power rating of the magnetic components—is a basic design consideration complicated by battery input voltage V_g range. The Volksinverter power source is a battery bank of (preferably) NiFe or (alternatively) $LiFePO_4$ batteries. As batteries of all chemistries charge and discharge (and also overcharge and undercharge), battery voltage varies significantly. Battery converter design consequently requires specification of a range for V_g for acceptable operation.

Design often proceeds based on a nominal or extreme value and then is checked to be within the rating for other values in the range. The Volksinverter design is based on design formulas optimized over the specified range of V_g and not at a nominal or other single value.

This article begins with a discussion of turns ratio in the boost push-pull circuit and how the selection of this parameter is influenced by input voltage range and the design power rating of the transformer's primary. Working formulas for the inductor and transformer's design power ratios (which correspond to magnetics sizing) are given, and values for these ratios are calculated for different turns ratios. Ultimately, the turns ratio is selected to optimize magnetics sizing.

Then the choice of converter operating point is discussed. Use of the form factor leads us to an operating point that optimally equates losses of the active and passive switches to minimize switching losses. However, this equal-losses condition occurs at a specific value of V_s', and V_s' is a term that also determines magnetics design power.

So there is a tradeoff to be made between equalizing switch losses (leading to lower switching losses) and minimizing power requirements of the magnetics. As we'll see, two of the commercially available inverters have chosen different turns ratios, which gives them different magnetics design power ratios, suggesting that one design is potentially more optimized for smaller magnetics, while the other leans more toward lower switching losses.

However, as our discussion proceeds with a turns ratio decided for the Volksinverter, we then consider options for implementing the magnetics using multiple paralleled components based on circuit operating conditions, and then calculate the impact of these choices on various operating parameters. We also determine the condition of maximum magnetizing current ripple which corresponds to maximum core loss.

Finally, the relationships between switching time and control loop delay on the one hand and switching losses are discussed, and we see that these do not have a major impact on the selection of operating conditions for the BPP circuit, but would be more of a consideration with alternative topologies for the battery converter stage. Ultimately, the formulas derived here based on circuit conditions will be used in completing the design of the coupled inductor and transformer in the BPP stage in subsequent parts of the series.

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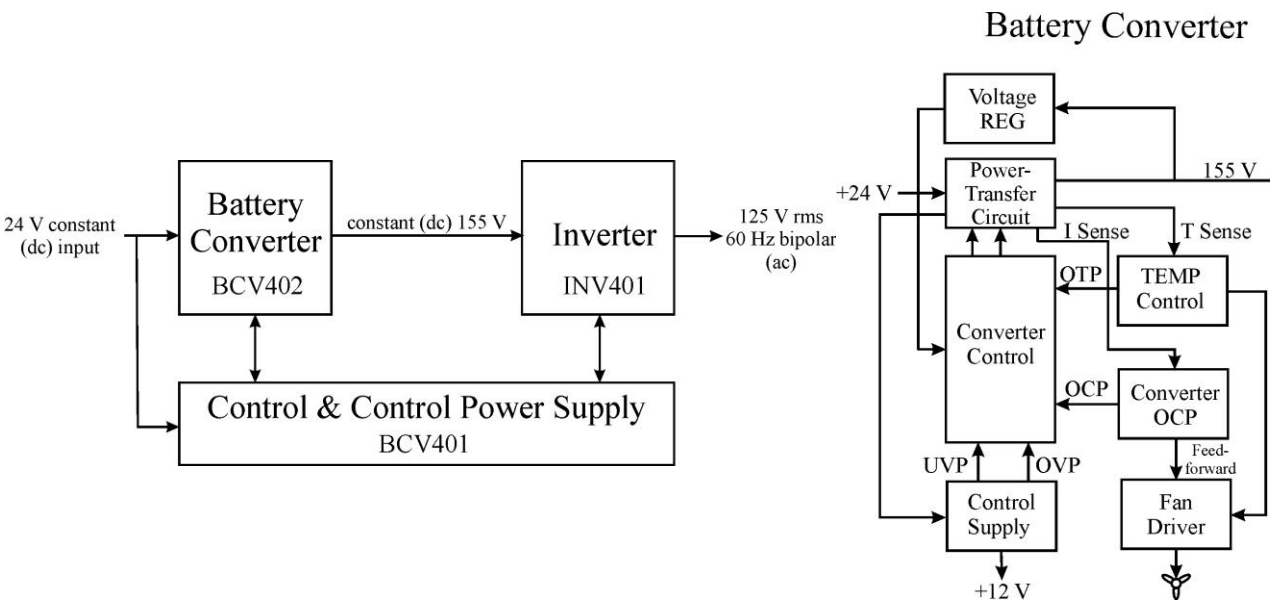


Fig. 1. The Volksinverter's system block diagram (left) and the BCV402 battery converter stage block diagram (right). This part addresses sizing of the magnetics in this stage's power transfer circuit.

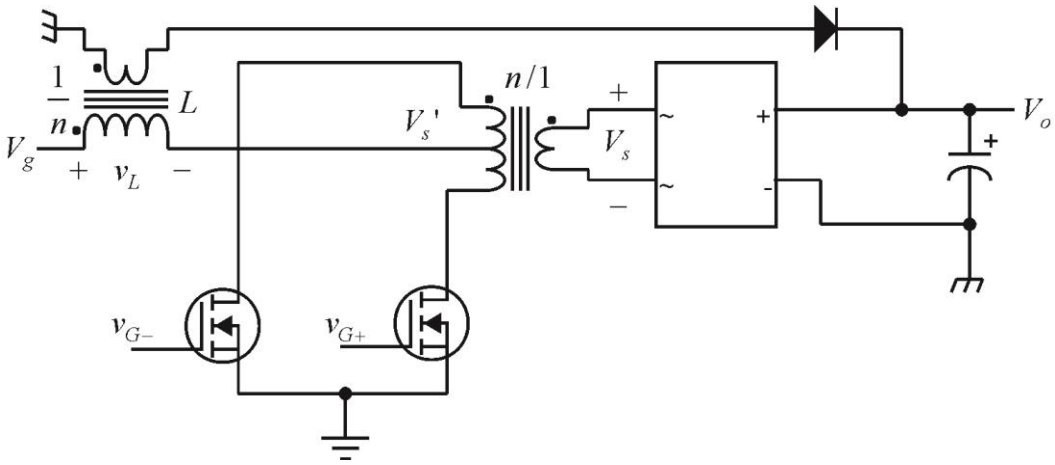


Fig. 2. Volksinverter power-transfer circuit. Transformer turns ratio n determines V_s' during off-time.

Power-Transfer Circuit Turns Ratio Optimization

How is turns ratio $n = N_p/N_s < 1$ optimized for a boost push-pull (BPP) circuit? Voltage V_s' at the primary-side center-tap must exceed input voltage V_g, or $V_s' > V_g$. Otherwise, circuit flux λ_L of input inductor L during on- and off-times is not of opposite polarity, and over a switching cycle, $\Delta\lambda_L \neq 0$ V·s.

No steady-state operation is possible without flux balance ($\Delta\lambda_L = 0$ V·s) over a switching cycle, and inductor current *ratchets*. Therefore, the lowest allowable BPP value of $V_s' = n \cdot V_s \geq V_{g\max}$. Volksinverter input voltage is specified over the range of $V_{g\min}$ = 20 V to $V_{g\max}$ = 30 V.

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Thus, the choice of *n* sets V_s' but is also constrained by the design ratings of inductor power $\overline{P}_{Ld}$ and primary winding power $\overline{P}_{pd}$ relative to input power $\overline{P}_g$. [12] For the inductor,[13] the ratio is

$$\frac{\overline{P}_{Ld}}{\overline{P}_g} = \frac{1}{4} \cdot \frac{V_s'}{V_{g\min}}$$

The BPP transformer primary-referred design-power ratio for the primary windings is [14]

$$\frac{\overline{P}_{pd}}{\overline{P}_g} = \frac{\tilde{v}_{p\max} \cdot \tilde{i}_{p\max}}{V_{g\min} \cdot I_{g\max}} = 2 \cdot \frac{(\sqrt{\frac{1}{2} \cdot D'_{\max}} \cdot V_s') \cdot (\sqrt{1 + D'_{\min}} \cdot I_{g\max}/2)}{V_{g\min} \cdot I_{g\max}} = \frac{1}{\sqrt{2}} \cdot \sqrt{\frac{V_{g\max}}{V_{g\min}} \cdot \left(1 + \frac{V_s'}{V_{g\min}}\right)}$$

These ratios give the factor of how much greater than the maximum input power must be the part design rating. From the BPP transfer ratio, the bounding values of *D'* are

$$D'_{\min} = \frac{V_{g\min}}{V_s'} \;\; ; \;\; D'_{\max} = \frac{V_{g\max}}{V_s'}$$

Substituting values, the working formulas for the Volksinverter magnetics sizing are

$$\boxed{\frac{\overline{P}_{Ld}}{\overline{P}_g} \approx \frac{V_s'}{80\,\text{V}} \;\; ; \;\; \frac{\overline{P}_{pd}}{\overline{P}_g} \approx (0.866) \cdot \sqrt{1 + \frac{V_s'}{20\,\text{V}}}}$$

The choice of V_s = 160 V exceeds the peak 3HSW inverter output waveform of 155 V by a 5 V margin to allow for secondary winding and rectifier diode (two) voltage drops, inverter power-switch drops (two), and output inductor drop. Over an input voltage range the inductor and transformer must be able to handle both maximum flux and current, resulting in the design power ratings of Table 1 for different choices of 1/*n* or V_s'.

Table 1. BVC402 1/*n* choices for optimizing inductor and transformer power ratings.

Note	1/*n*	$\overline{P}_{Ld}/\overline{P}_g$	$\overline{P}_{pd}/\overline{P}_g$	$D'_{\max}$	$D'_{\min}$	V_{gopt}, V	V_s', V
min-V_s' design	5.33	0.375	1.369	1	0.667	18.5	30.0
	5.16	0.3875	1.383	0.968	0.645	19.16	31.0
	5	0.400	1.396	0.938	0.625	19.8	32.0
	4.7	0.425	1.423	0.882	0.588	21.0	34.0
Sumry 3 kW*	4.3	0.465	1.465	0.8065	0.538	23.0	37.2
	4.13	0.484	1.485	0.774	0.516	23.95	38.75
Volksinverter	**4**	**0.500**	**1.500**	**0.750**	**0.500**	**24.72**	**40**
	3.92	0.510	1.510	0.736	0.490	25.21	40.79
Statpower 500*	3.8	0.526	1.526	0.713	0.475	26.0	42.1

*Commercially available power inverters.

Operating-Point Selection

Active and passive switch form-factors are optimized for $\kappa_Q = \kappa_D$ at $D'_{opt} \approx 0.618$ from part 11.[11] Corresponding to this optimal operating-point is V_{gopt} from the transfer function;

$$\frac{V_s'}{V_g} = \frac{n \cdot V_s}{V_g} = \frac{1}{D'} \Rightarrow V_{gopt} = D'_{opt} \cdot V_s'$$

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A battery-bank as a V_g source has a plateau of nearly constant voltage over a wide discharge range that spans across V_{gmid} = 25 V. Below V_{gmin} the voltage-charge curve falls off abruptly.

When the battery-bank is fully charged, its voltage at V_{gmax} decreases quickly with discharge so that over its charge range, the source includes D'_{opt} = 0.618 where V_{gopt} = 24.72 V ≈ V_{gmid} = 25 V. At midrange, D'_{mid} ≈ 0.672—slightly above D'_{opt} and where $\kappa_Q \approx \kappa_D$. Fig. 3 plots κ of the secondary diodes κ_D, primary MOSFETs κ_Q, and their product κ_{QD}.[15]

From Table 1, both inductor and transformer power ratings decrease as V_s' decreases and 1/n increases. Consequently, power ratings are reduced by making V_s' as close to V_g as is tolerable. The tradeoff is both inductor flux imbalance and also switch-loss imbalance between diodes and MOSFETs. A low V_s' choice of 1/n = 5 is attractive for minimizing both magnetics design power and switch loss as κ_{QD}, but it has only 2 V of margin for balanced inductor flux above V_{gmax} and the primary-side switches have higher form factor and conductive loss.

Commercial inverters from Sumry (China) have V_{gopt} = 23 V at 1/n = 4.3 and Statpower (Samlex, Taiwan) has V_{gopt} ≈ 26 V at 1/n = 3.8. Statpower requires larger design power than that of Sumry by 13% for the inductor and 4.2% for the transformer.

The Volksinverter choice of 1/n = 4 is between these two and has a design-power ratio relative to the min-V_s' design of 33.3% greater for the inductor and 9.6% for the transformer with V_{gopt} = 24.72 V, near the V_g = 25 V mid-range value. (With an integer 1/n = 4, most Volksinverter numbers in the Table 1 entry are conveniently round numbers.)

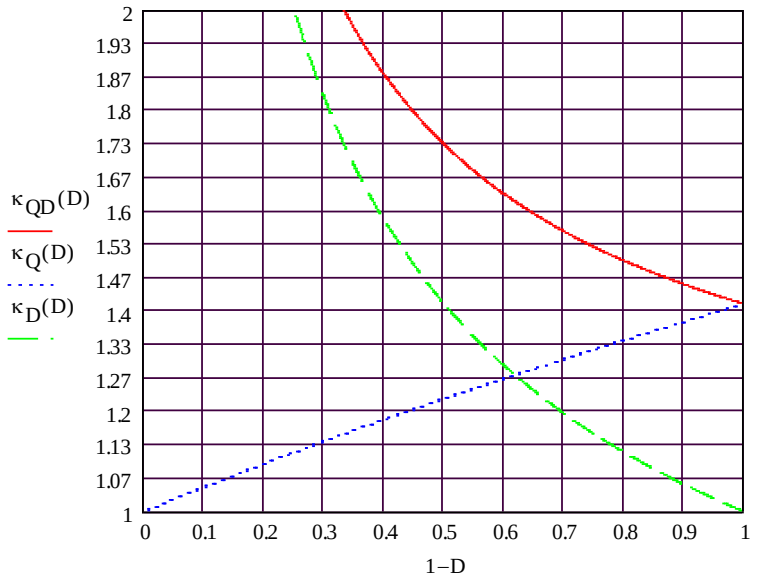


Fig. 3. Graph of form-factor κ (rms/avg) plots: passive, diode (dash); active, MOSFET (dot), and product (solid).

Circuit Specifications For Magnetics Design

The design choice of 1/n = 4 (N_s = 4·N_p) has $\overline{P}_{Ld} = 0.5 \cdot \overline{P}_g$ and $\overline{P}_{pd} = 1.5 \cdot \overline{P}_g$ with V_s' = 40 V—an ample 10 V of flux-balance design margin. The optimal switch-loss operating-point is near the middle of the V_g range, at V_{gopt} = 24.72 V ≈ 25 V and D'_{opt} ≈ 0.618 ≈ D'(25 V) = 0.625, a 1.1% difference.

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A total inverter design is scaled up from multiple units having a full-scale rating of $\overline{P}_g$ = 1 kW. The 1-kW building block requires three parallel transfer circuits of $\overline{P}_g$ = 333 W. Instead of one large transformer and inductor, distributing magnetics to three smaller components results in a core transfer-power density advantage over a single larger component of $3^{2/3} \approx 2.08$ times.[16]

For 1/*n* = 4, transformer design power is $\overline{P}_{pd} \approx (1.5)\cdot \overline{P}_g$ = (1.5)·(333 W) = 500 W. Two transformers instead result in a $2^{2/3} \approx 1.59$ times thermal advantage with $\overline{P}_{pd} \approx (1.5)\cdot \overline{P}_g$ = (1.5)·(500 W) = 750 W. The 333-W design is presented here, though the 500-W alternative is later reviewed retrospectively.

Derivations of transfer-circuit design formulas are given in reference [16]. The FW diode bridge (FWB) has two alternating branches of conduction on successive half-cycles, and $\kappa_S \neq \kappa_D$. Variables given in Table 2 are based on the parameters:

$$f_s = 150\ \text{kHz} \Rightarrow T_s = 6.67\ \mu\text{s};\ n = 1/4;\ V_s\text{'} = 40\ \text{V};\ V_s = 160\ \text{V};\ \overline{P}_g = 333\ \text{W};\ \overline{P}_{gd} = 500\ \text{W}$$

A 20-A fuse has a 20% current tolerance when $I_{g\max}$ =16.66 A. A ×1.20 overcurrent is tolerable because transient peak-to-average current ratings of Si switches for the time it takes a fast-acting fuse to blow is typically around ×10.

Magnetic design limits are calculated at the specified full-scale input power over the specified range of V_g. One of the limits is the primary-winding flux change $\Delta\lambda_p$ which varies with V_g. The transfer ratio for the BPP is

$$\frac{V_s}{V_g} = \frac{1}{n}\cdot\frac{1}{D'} \Rightarrow \frac{V_s'}{V_g} = \frac{1}{1-D} \Rightarrow D = \frac{V_s' - V_g}{V_s'}$$

The flux change of the primary winding is

$$\Delta\lambda_p = V_g\cdot(D\cdot T_s) = V_g\cdot\left(\frac{V_s' - V_g}{V_s'}\right)\cdot T_s = \left(V_g - \frac{V_g^2}{V_s'}\right)\cdot T_s$$

The primary-referred V_s through turns ratio $n = N_p/N_s$ is $V_s' = n\cdot V_s$ and $\Delta\lambda_p$ for the transformer is maximum at the V_g value for which $d\Delta\lambda_p/dV_g = 0$. Differentiating and solving,

$$\frac{d}{dV_g}\left(V_g - \frac{V_g^2}{V_s'}\right) = 0 \Rightarrow V_g = \frac{V_s'}{2} = \frac{n\cdot V_s}{2},\ \max\Delta\lambda_p \Rightarrow D = D' = 0.5,\ \max\Delta\lambda_p$$

Magnetizing-current ripple is largest at $D'_{\min}$ for the Volksinverter BPP at $V_g = V_{g\min}$ = 20 V. Maximum current ripple corresponds to maximum magnetic (core) power loss.

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Table 2. BVC402 circuit design parameters and formulas.

V_g (V)	20	25	30	Design formula
I_g (A)	16.66	13.32	11.1	$\bar{i}_g = I_g = \bar{P}_g / V_g$
D'	0.500	0.625	0.750	$D' = V_g / (n \cdot V_s)$
D	0.500	0.375	0.250	$D = 1 - D'$
$\bar{i}_Q$ (A)	12.5	10.0	8.33	$\bar{i}_Q = I_g / 2$
$\kappa_Q = \kappa_p$	1.225	1.275	1.323	$\kappa_p = \kappa_Q = \sqrt{1 + D'}$
$\tilde{i}_Q$ (A)	15.31	12.75	11.02	$\tilde{i}_p = \tilde{i}_Q = \kappa_Q \cdot (I_g / 2)$
$\bar{i}_s$ (A)	3.125	3.125	3.125	$\bar{i}_s = D' \cdot n \cdot I_g$
κ_s	1.414	1.265	1.155	xfmr $\kappa_s = 1 / \sqrt{D'}$
$\tilde{i}_s$ (A)	4.42	3.95	3.61	$\tilde{i}_s = \kappa_s \cdot \bar{i}_s = n \cdot I_g \cdot \sqrt{D'}$
$\bar{i}_D$ (A)	1.56	1.56	1.56	FWB $\bar{i}_D = \bar{i}_s / 2$
FWB κ_D	2.00	1.79	1.63	FWB $\kappa_D = \sqrt{2 / D'}$
FWB $\tilde{i}_D$ (A)	3.12	2.79	2.54	FWB $\tilde{i}_D = \kappa_D \cdot \bar{i}_D$
FWB κ_{QD}	2.45	2.28	2.16	FWB $\kappa_{QD} = \kappa_Q \cdot \kappa_D$
$V_s' - V_g$ (V)	20	15	10	$n \cdot V_s - V_g$
t_{on} (µs)	3.33	2.50	1.67	$t_{on} = D \cdot T_s$
$\Delta\lambda_L$ (V·µs)	66.7	62.5	50.0	$\Delta\lambda_L = V_g \cdot (D \cdot T_s)$
$\bar{P}_L$ (W)	166.5	125	83.25	$(V_s' - V_g) \cdot D' \cdot I_g = D \cdot \bar{P}_g$

Switching Time And Control Loop Delay Limitations

A circuit tradeoff in having V_s' too close to V_{gmax} is that D_{max} approaches zero and can cause switch-loss problems from excessive switching time and also from control-loop delay. The active power switches switch an inductive load which has switch loss while shutting off. The minimum $D = D_{min} = 0.25$ corresponds to an on-time of $t_{on} = 1.67$ µs, or 10% of the switching time $t_{sw} \approx 167$ ns of commodity MOSFET gate drivers. Thus switching loss places a constraint on maximum f_s.

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Peak-current control-loop delay time causes commanded peak current to exceed the comparator value as shut-off is delayed by the loop. This has the effect of increasing RMS and average current and if in excess of load current can be controlled by the same method in control design of controlling low-current loading and low current whenever $V_g < V_{gmin}$ during power-on.

The BPP $D_{opt} \approx 0.618$ is well away from the extremes of zero and one, and this lessens design demand of switching time and loop delay. We will find in the comparison of BPP with full-bridge CP (buck) transfer circuits that the buck alternative has an optimal D at an extreme, which places greater design demands on switching time and loop delay than does the BPP circuit.

A derivation of $1/n$ or V_s' that minimizes both magnetics design power and switching loss is left as an exercise for the reader. The BPP optimization of equal passive and active switch loss is applied in the Volksinverter design for setting the circuit operating-point.

Specification of design power for parts over their specified operating ranges is an important (though easily neglected) step in design. Now that we have power ratings from circuit-applied constraints, we will use the values from the above table in subsequent magnetic component design. The general inductor and transformer design-power formulas for BPP circuits over a V_g range as given previously can be applied to other designs.

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

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