Utilizing Full Saturation and Power Loss
To Maximize Power Transfer In Magnetic Components

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Power-electronics engineering is largely an exercise in how to convert power from one form to another efficiently and at minimum volume and cost. Switching power converters all work on the basic principle of storing input power in a field, then extracting energy from the field to deliver output power. In the usual scheme, the on-time part of a switching cycle inputs power to the field. Then, during the off-time power is delivered to the output.

To minimize converter size and thus maximize power density, the energy density of the field must be maximized. At present, electric fields of capacitors and magnetic fields of inductors are the two choices. (In the future, perhaps ion currents or transmission lines will be used to store energy.) Of the two existing energy storage options, magnetic materials are capable of sustaining a higher power density at a given cost. Thus, magnetic cores are used in power supplies to transfer power from input to output. Of these, ferrites (MnZn) and iron-powder (Fe-pwd) are the most commonly used. Other material choices such as NiFeMo, NiFe, and FeSiAl have parameters between those of ferrite and iron-powder cores.

To optimize magnetic component design, electrical (winding) and magnetic (core) power losses should be made equal to maximize power transfer from input winding to output winding. Utilization of the core is maximized when the winding window area is full of wire and the core is driven to near saturation while dissipating the desired maximum power loss. To achieve these optima, a multiplicity of design parameters must be just right including input and output voltages, number of turns, current ripple factor, field-referred inductance, and core power-loss density, p_c ($\hat{B}_\sim$, f_s). This article will describe how to determine these parameters in a power conversion application, so that maximum power transfer can be achieved for a given core material at the smallest possible core size. At the same time, this discussion will help designers understand how to optimize their choice of core material for a given application.

The ripple amplitude of the magnetic field density, $\hat{B}_\sim$, and switching frequency, f_s, determine core power loss. Furthermore, the power density of the converter is proportional to f_s because the more per-cycle energy that can be moved through the field from input to output per unit time, the higher the power. However, p_c increases with f_s placing an upper bound on f_s. That leaves $\hat{B}_\sim$. A Micrometals Inc. Fe-pwd 26 material has an approximate power loss of

$$\frac{\overline{p}_c}{100\,\mathrm{mW/cm^3}} = \left(\frac{f_s}{100\,\mathrm{kHz}}\right)\cdot\left(\frac{\hat{B}_\sim}{15\,\mathrm{mT}}\right)^2$$

That is, it dissipates 100 mW/cm^3 when driven at 100 kHz with a B-field ripple amplitude of 15 mT.

The value of the field flux ripple amplitude is

$$\hat{\phi}_\sim = \hat{B}_\sim \cdot A$$

where *A* is the magnetic path cross-sectional area.

An iron-powder (Fe-pwd) Micrometals T50B-26 0.5-inch (12.7-mm) OD core has an *A* of 14.8 mm^2. With equal winding loss and a 40°C rise, the core power-loss density is about 700 mW/cm^3. The corresponding $\hat{B}_\sim$ = 39 mT. Therefore, $\hat{\phi}_\sim$ = 0.577 μV·s.

For f_s = 100 kHz and duty ratio of D = 0.5 (which is the optimal value for many converter schemes), t_{on} = 5 μs. Given these conditions, the minimum number of turns for which $\hat{B}_{\sim} \leq$ 39 mT is

$$N_\lambda \geq \frac{\Delta\lambda}{\Delta\phi} = \frac{V_p \cdot t_{on}}{\Delta B \cdot A} = \frac{V_p \cdot (D/f_s)}{(2 \cdot \hat{B}_{\sim}) \cdot A} = \frac{V_p}{231\,\text{mV}}$$

In the design of the winding, the turns must not be less than N_λ or the core will overheat if 700 mW/cm^3 will raise it to its maximum allowable temperature. A 5 V input requires at least 22 turns.

For a ferrite core of about the same dimensions—a Magnetics Inc. 41306-TC, K material—the core-loss formula is

$$\frac{\overline{p}_c}{100\,\text{mW/cm}^3} = \left(\frac{f_s}{100\,\text{kHz}}\right) \cdot \left(\frac{\hat{B}_{\sim}}{110\,\text{mT}}\right)^2 .$$

Note the 7.3 times increase in $\hat{B}_{\sim}$ for the same p_c at the same f_s. This difference reduces N_λ to about 3 turns.

What about magnetic saturation? Saturation, another material limitation, is quantified as

$$k_{sat} = \frac{L(i_{sat})}{L(0\,\text{A})}$$

In other words, the fractional decrease of inductance at current i_{sat} relative to the zero-current inductance is k_{sat}. It is typically chosen to be in the range of 0.5 to 0.7. Any lower than 0.5 and the core is driven too deeply into saturation. The circuit effect is that the current for a fixed voltage across the winding increases superlinearly and can become increasingly uncontrollable, leading to overcurrent failures. A k_{sat} any higher than 0.7 does not sufficiently utilize the core for power transfer applications. That is to say, it does not store as much energy per cycle as could be achieved with greater drive current.

The saturation places an upper bound on turns as

$$N_i = \frac{N\bar{i}}{I_p}$$

where $N\bar{i}$ is the static operating-point field-referred current at the chosen k_{sat}, and I_p is the on-time average winding-current amplitude corresponding to the core average field intensity, $\overline{H}$, at operating-point ($\overline{H}$, $\overline{B}$).

Although the peak current will drive the core even farther into saturation, linear analysis of the magnetics around an operating point requires the small-ripple assumption: that the deviations from the operating point are small enough to assume linearity. In this case, $\overline{H} \approx \hat{H}$ and saturation can be treated as a static behavior of the core. By using the static value, the effect of saturation is separated from power loss.

N_i can also be limited by the winding area of the core, and it then sets the upper bound on N_i. For toroidal core sizes under 12.7 mm OD, winding area is the limiting factor for power transfer. Thus, we have bounds on N:

$$N_\lambda \leq N \leq N_i$$

Somewhere in this range is the optimum N and it occurs when the two bounds are equal, thus satisfying the conditions of both a maximum saturation and a maximum core-power dissipation:

$$N_\lambda = N_i = N = \frac{V_p \cdot t_{on}}{(2 \cdot \hat{B}_{\sim}) \cdot A} = \frac{N\bar{i}}{I_p}$$

This maximum power transfer condition can be solved for average transferred power:

$$\overline{p}_p = D \cdot V_p \cdot I_p = V_p \cdot \overline{i}_p = N\overline{i} \cdot (2 \cdot \hat{B}_{\sim}) \cdot A \cdot f_s = \Delta W_L \cdot f_s$$

Power as a function of time has a non-zero amplitude of $V_p \cdot I_p$ for *D* of the switching cycle. For the Fe-pwd core operated at a fractional saturation of k_{sat} = 0.6, average power is

$$\overline{p}_p(\text{Fe-pwd}) = [(98\text{ A}) \cdot (2 \cdot 39\text{ mT}) \cdot (14.8\text{ mm}^2)] \cdot (100\text{ kHz}) = 11.3\text{ W}$$

To achieve this transfer power, the circuit conditions must be such that

$$I_p = \frac{N\overline{i}}{N} = \frac{98\text{ A}}{22} = 4.46\text{ A}$$

The Fe-pwd cores can sustain a significantly higher value of static field current for comparable saturation to MnZn ferrites. For the above two cores, for k_{sat} = 0.6,

$$\frac{\overline{H}(\text{Fe}-\text{pwd})}{\overline{H}(\text{Mn}-\text{Zn})} = \frac{3064\text{ A/m}}{138\text{ A/m}} = 22.2$$

The Fe-pwd core can sustain field-referred current of over 20 times that of the ferrite core of similar geometry at the same saturation. For maximum core-transfer energy, the winding current will also be 22 times that of the comparable ferrite core. Nevertheless, equal power loss requires that the Fe-pwd $\hat{B}_{\sim}$ drive be only about 20% that of the ferrite core:

$$\frac{\hat{B}_{\sim}(\text{Fe-pwd})}{\hat{B}_{\sim}(\text{Mn-Zn})} = \frac{39\text{ mT}}{200\text{ mT}} = \frac{1}{5.1}$$

Because the incremental *μ* of ferrite is so much larger than Fe-pwd, the ripple field intensity amplitude, $\hat{H}_{\sim}$, of Fe-pwd must be larger than for ferrite to achieve the maximum *B* value:

$$\frac{\hat{H}_{\sim}(\text{Fe-pwd})}{\hat{H}_{\sim}(\text{Mn-Zn})} = \frac{690\text{ A/m}}{71\text{ A/m}} = 9.75$$

It takes more ripple current (and hence Δ*H*) to drive the Fe-pwd core to the same $\hat{B}_{\sim}$. Fe-pwd is superior to ferrite when the static (dc) current is much larger than the ripple current—yet by how much?

When numbers are put to these parameters, the optimal use for Fe-pwd cores are for "inductors", which means low average-ripple-factor applications. *Average ripple factor* is defined as

$$\gamma = \frac{\Delta x / 2}{\overline{x}} = \frac{\hat{x}_{\sim}}{\overline{x}}$$

or the ripple amplitude (peak ripple) over the average of *x*, where *x* is a current or voltage.

Magnetic saturation is a limit that applies to the average (or static) current while Δ*i*/2 is related to Δ*B*/2, the $\hat{B}_{\sim}$ on the horizontal axis of core manufacturer p_c graphs. Because Fe-pwd cores are weak on power loss, requiring low values of $\hat{B}_{\sim}$ for a given f_s, they are best applied with low-*γ* current waveforms. Ferrites are at the opposite extreme.

For ferrites, gapped cores extend optimization toward the low-μ Fe-pwd cores. Gapped cores, however, have their disadvantages: custom parts, field concentration around the gap causing excessive winding dissipation nearby caused by the proximity effect, and more noise around the part. The other materials such as NiFeMo

(Magnetics Inc MPP—molypermalloy) and FeSiAl (Kool-μ) cost more but provide intermediate loss and saturation between the ferrite and Fe-pwd extremes at higher energy density.

What is important to take away from the above exposition is that each material has an optimal γ and hence an optimal current waveshape for full core utilization. For a particular material, the highest energy transfer is achieved with a given average current relative to the current ripple amplitude. The average drives the core to the allowed amount of saturation (k_{sat}) and the ripple amplitude drives it to its full power loss (p_c). Together, they determine γ. As a general rule,

$$\bar{\gamma}_{opt}(\text{Fe-pwd}) \cong 0.1$$
$$\bar{\gamma}_{opt}(\text{Mn-Zn}) \cong 0.4$$

This is a broad rule of thumb based on the following ratios of saturation and ripple values of H for each material, and the relationship

$$\gamma_{opt} = \frac{\hat{H}_{\sim}}{\overline{H}} = \frac{(\hat{B}_{\sim} \cdot A)}{\mathcal{L} \cdot N\bar{i}}$$

To find the field current, $N\bar{i}$, two charts (on the same page) of the Micrometals catalog (Catalog 4, Issue G, page 30, labeled “DC Applications”) are used. The lower chart vertical axis gives percent saturation as 1 – k_{sat} (“30% saturation” is k_{sat} = 0.7). Choose a k_{sat} value and go over to the core curve to read off “DC Energy Storage”, then use it with the upper chart and same core to read “NI” from the vertical axis. This is $N\bar{i}\,(k_{sat})$. $\hat{B}_{\sim}$ is read from the horizontal axis of the power-loss plot for the given core material (page 24 of the same catalog for 26 material) for a chosen power-loss density (in mW/cm^3) and frequency, f_s, A and $\mathcal{L}(0)$ are found on the page for the given core; $\mathcal{L}(0)$ is the zero-current field-referred (per-turn-squared) inductance.

As an example, the T50B-26 Fe-pwd core, operating at 100 kHz and 700 mW/cm^3 with k_{sat} = 0.6 has an optimal

$$\gamma_{opt}(\text{Fe-pwd}) = \frac{\hat{H}_{\sim}}{\overline{H}} = \frac{(\hat{B}_{\sim} \cdot A)}{\mathcal{L} \cdot N\bar{i}} = \frac{(39\,\text{mT}) \cdot (14.8\,\text{mm}^2)}{(0.6) \cdot (43.5\,\text{nH}) \cdot (98\,\text{A})} = 0.225$$

The value of γ_{opt} should be calculated for each application. Then, for a given core material, the optimal current waveshape is determined for maximum energy transfer and minimum core size. Using the general optimum values as given above is better than not optimizing at all. When converter circuit design constrains γ, then this can affect which core material is chosen for an optimal magnetic design.

About the Author

Dennis Feucht has been involved in power electronics for 25 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 doing current-loop converter modeling and converter optimization.

For further reading on magnetics design, see the How2Power Design Guide, search the Design Area category and select the Magnetics subcategory.