

Living with losses: tips on using magnetic components

**MAGNETIC COMPONENTS POSE
TRICKIER DESIGN CHALLENGES
THAN OTHER PASSIVE
COMPONENTS.**



Closely matching standard bobbin-core sizes, DHV high-voltage inductors from DTMagnetics withstand 1500V core-to-winding voltage.

Of the three classes of passive components, resistors and capacitors are relatively docile elements. Their loss contributors—series inductance in resistors and effective series resistance and inductance in capacitors—are substantially linear and predictable in behavior. Magnetic components, on the other hand, harbor several complex loss-producing traits that you should take into account

when you design. Parameters such as frequency-dependent core losses, copper losses, and limited saturation currents can produce large deviations from your theoretical circuit-performance calculations.

Assume, for example, that you must design a dc/dc converter with a push-pull configuration. This topology requires an output transformer with center-tapped primary and secondary windings. Because you want to optimize performance,

you'll likely design your own transformer with core material and geometry and a turns ratio and core area tailored to your circuit's requirements. Alternatively, you could specify an off-the-shelf transformer, but that choice might entail unacceptable compromises in performance.

Magnetic components are the principal loss-producing culprits in high-frequency dc/dc-converter design. Provided the control circuitry delivers enough drive to rapidly charge and discharge the

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gate capacitance of the switching MOSFETs, the transient switching losses of the MOSFETs are negligible in comparison with the magnetic-component losses. Because of magnetic-device limitations, most dc/dc converters on the market use switching frequencies that are much less than 1 MHz.

Unitrode Integrated Circuits Corp describes the design of a 1.5-MHz, IC-controlled, 50W push-pull dc/dc converter in a paper that also goes into considerable detail about the design of the output transformer (**Reference 1**). Two-thirds of the document describes the transformer design; one-third covers the control and switching aspects of the converter. This ratio gives you an idea of the complexity of magnetic-component design, as opposed to, say, selecting capacitors or resistors (or even ICs). The arduous transformer-design process entails many critical and complex steps. **Figure 1** shows a skeletally simplified diagram of the converter's topology.

STEP-BY-STEP TRANSFORMER DESIGN

The first step in the transformer design is determining the primary-to-secondary turns ratio, N . This ratio is a function of

AT A GLANCE

- ▶ Magnetic materials exhibit complex behavior that requires careful analysis.
- ▶ Manufacturers' cookbook design procedures help remove the black magic from magnetic design.
- ▶ Magnetic design requires a lot of estimating, iterating and reiterating.
- ▶ Developments, such as new core materials and planar geometries, are significantly improving magnetic performance.

the minimum primary voltage, $V_{PRI(MIN)}$, and the minimum secondary voltage, $V_{SEC(MIN)}$. $V_{PRI(MIN)}$ is the input voltage minus the switching-transistor and sense-resistor voltage drops. $V_{SEC(MIN)}$ is the converter output voltage plus the rectifier and transformer losses. Assuming a 42V input and a total of 3V in transistor and sense-resistor drops, $V_{PRI(MIN)}$ is 39V. Assuming 5.1V output voltage and 0.8V total rectifier and transformer losses, $V_{SEC(MIN)}$ is 5.9V. The turns ratio is the quotient of these two values, multiplied by the maximum duty cycle: 0.85 for this

converter. For this example, $N=5.6$. Rounding to the nearest lower integral number, $N=5$.

The next step is to use a formula to determine the minimum cross-sectional area of the transformer core. In the design example from **Reference 1**, the formula is

$$A_{C(MIN)} = \frac{V_{PRI} \cdot DC \cdot 10^4}{2fNB}$$

where DC is the duty cycle, f is the switching frequency, and B is the flux-density swing in teslas. At first glance, you might assume that the core area that a 1.5-MHz converter requires would be 10 times smaller than the core area for a 150-kHz switcher. That assumption would be true if the flux density, number of turns, and core losses remained constant. However, the losses are a function of the frequency raised to an exponent between 1 and 2. As frequency increases, you must reduce the flux-density swing (ΔB) to keep temperature rise, or core losses, constant.

The core area is inversely proportional to the number of primary turns. For the 50W design in **Reference 1**, a five-turn primary yields a relatively large 1.5-cm² core area. With such a large cross-

GET THE L OUT OF THERE!

No, the headline for this sidebar is not a cockney oath—it means get rid of pesky inductors in your circuitry by using some electronic trickery. Compared with resistors, capacitors, and even op amps, inductors are

bulky and prone to the parametric quirks described in the text. By using a circuit configuration called a “gyrator,” you can use op amps, resistors, and capacitors to create “virtual inductors” that are compact, stable, and

predictable over wide temperature and frequency ranges.

Figure A, derived from **Reference A**, shows a floating virtual inductor. The input (inductive) impedance of the circuit is sR^2C , or $j\omega R^2C$ in non-Laplacian terms. The expression shows that you can create high-value inductors—of thousands of henries—by using resistors and capacitors with reasonable values. Real inductors of kilohenry values would be extremely bulky and heavy and prone to myriad frequency-related losses. **Figure A** shows a floating inductor. For a single-ended (grounded) inductor, you can simplify the circuit by eliminating the entire bottom section and grounding the resistor at the left.

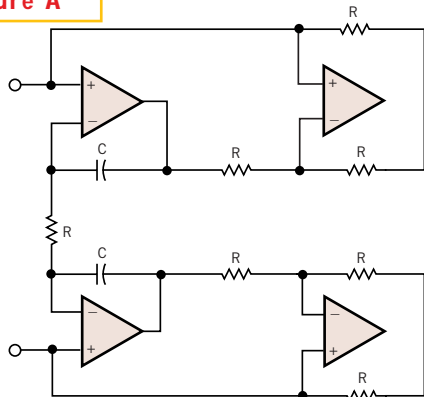
The virtual inductors really do work. I built an audio amplifier years ago and used seven variable virtual-inductor-based

bandpass filters for the equalizer section. It worked like a charm. Gyrators shine in variable-inductor circuits. It's much easier to procure and use variable resistors than variable-reactance (capacitance or inductance) elements. **Reference B** describes two practical gyrator circuits: one that turns a 3- μ F capacitor into a 10-kH inductor with a Q of 13 and one that provides a 12- μ H inductor with a maximum Q of 500 over 10 Hz to 1 MHz.

REFERENCES

- A. Faith, Donald, and Donald Christiansen, *Electronics Engineers' Handbook*, McGraw-Hill, 1982, pg 12 to 35.
- B. Markus, John, and Charles Weston, *Essentials Circuits Reference Guide*, McGraw-Hill, 1988, pg 119 and 302.

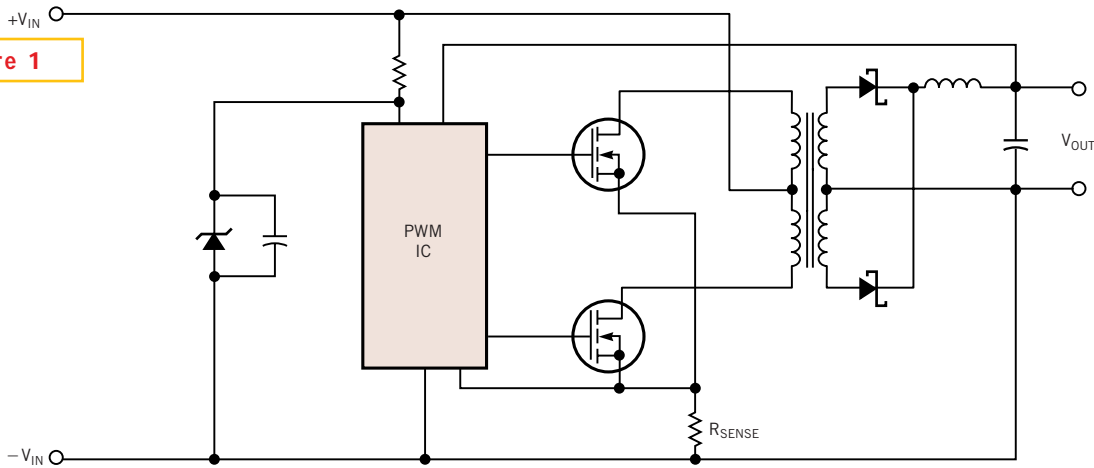
Figure A



When does $RC=L$? When an op-amp circuit “gyrates” the transfer function to provide a virtual inductive impedance at its inputs.



Figure 1



The transformer in a push-pull dc/dc converter is the most significant limiting factor in high-speed performance.

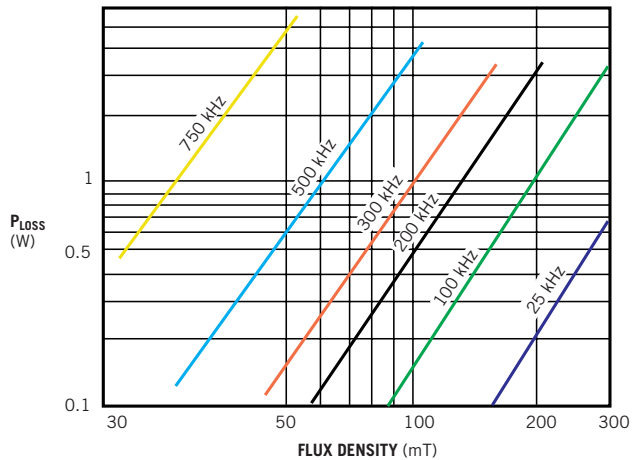
sectional area, much of the winding area would go unused. A better choice, a 10-turn primary, yields a more reasonable 0.68-cm² core area. As the switching frequency increases, you can generally reduce the core size or minimum number of turns. This statement is true, however, only to the point at which the increasing core losses prevent a further reduction of either size or the number of turns. This crossover point occurs at different frequencies for different ferrite materials.

To determine the crossover point, examine the manufacturer’s power-loss-versus-flux-density curves for various frequencies. **Figure 2** shows the curves for a typical ferrite-core material. You select an acceptable core loss (in milliwatts/centimeters cubed) and tabulate the flux density required for various frequencies from the intersection of the frequency curves and the power-loss axis. You can then plot a curve of flux density versus frequency (**Figure 3**). Next, calculate the minimum number of turns based on the tabulated values of frequency and flux density. Select an input voltage, cross-sectional area, and duty cycle. You can lump these terms together as a constant, K. The formula relating core area, number of turns, frequency, and flux density becomes

$$A_C \cdot N = \frac{K}{fB}$$

If you calculate and tabulate the data for various frequencies, you can plot a curve of required core area that resembles

Figure 2



Lacking manufacturer-provided equations describing a magnetic material’s frequency performance, you can extrapolate performance from these parametric curves.

the curve in **Figure 4**. The curve gives the minimum cross-sectional area with a fixed number of turns for various frequencies. The **figure** shows that the core size or required number of turns decreases from 50 to 500 kHz and then increases rapidly with frequency.

GEOMETRIES, WIRES, AND ASSEMBLY

Selecting the right core size for your transformer is a complex task, but it’s only the beginning of the transformer-design effort. You must next choose the optimum core shape for your design from among many available geometries. In this choice, you must consider safety-agency spacing requirements, shielding considerations, heat-transfer issues, physical dimensions, and assembly cost.

Next, and by no means an “oh,-by-the-way” consideration, you must choose the correct wire size and configuration. **Reference 1** avers that the single most difficult task in high-frequency magnetic design is minimizing eddy-current losses, or skin effect, while optimizing wire size.

In the skin-effect phenomenon, the penetration depth is the depth to which a wave will penetrate a conductor at a specific frequency. In a copper wire at 100 C, this depth (in centimeters) is 7.5/√f. At 750 kHz, the depth is 8.7 × 10⁻³ cm, or approximately the diameter of a #38 AWG wire. You can use larger wires, but ac current flows only in the penetration-depth area. **Table 1** gives the depth of penetration for several frequencies and the AWG wire that corre-

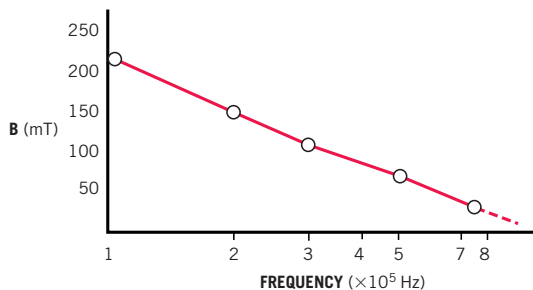


sponds most closely to the penetration depth.

For low-current windings, it is common practice to twist together several strands of fine wire to form a bundle. For example, seven strands twisted together closely approximate a round conductor with a net diameter that is three times the individual wire diameter. The twisting reduces the parasitics between wires at high frequencies. Medium- to high-current windings require the use of Litz wire (a bundle of many conductors), copper foil, or strapping. Industry practice is to operate at a current density of 450A rms/cm².

The assembly technique is another important consideration in transformer design. First, you should wind the primary and secondary windings as closely as possible to each other to maximize coupling. In the push-pull configuration, you should also try to closely couple the two secondary sections, because they simultaneously conduct during the circuit's off periods. Reference 1 and the design guides available from magnetic-materials manufacturers go into considerable

Figure 3



Data from Figure 2 yields this curve of flux density versus frequency for a given power loss.

detail about calculating wire type, length, and losses in transformer designs.

IT'S A LONG ROAD TO DESIGN

If you need a 100- μ F capacitor in your design, you simply choose a unit from a data book that has a low enough effective series resistance and an adequate voltage rating. If you need a transformer or an inductor, however, and you decide to design your own for optimum performance, you're in for an intricate design procedure. Allied Signal Inc gives cookbook-type design procedures for designing

magnetic components using the company's Metglas amorphous-silicon cores (references 2, 3, and 4). Allied maintains that the Metglas material overcomes the two main disadvantages of ferrites (high cost and the need for stacking) and offers lower core losses, volume, and weight than silicon-steel laminations.

Reference 2 gives a 12-step design procedure for switching-supply transformer design. It is beyond the scope of this article to give complete details of all the steps, but an overview of the procedure might be useful if you

contemplate designing your own transformers.

- **Step 1**—Determine the input data. Itemize voltage and current ratings, rectifier drop, maximum duty cycle, temperature limit, and estimated transformer and power-supply efficiencies.
- **Step 2**—Estimate the convective surface area. Give your best estimate of the surface area of the wound transformer, using your best-guess core for the application.
- **Step 3**—Estimate the power loss in the

MANUFACTURERS OF MAGNETIC MATERIALS AND PRODUCTS

For free information on magnetic materials and products such as those described in this article, circle the appropriate numbers on the postage-paid Information Retrieval Service card or use EDN's InfoAccess service. When you contact any of the following manufacturers directly, please let them know you read about their products in EDN.

Allied Signal Inc

Parsippany, NJ
1-973-581-7653
fax 1-973-581-7717
www.alliedsignal.com/metglas
Circle No. 307

MMG North America

Paterson, NJ
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fax 1-973-345-1172
www.mmгна.com
Circle No. 310

Prem Magnetics Inc

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fax 1-815-385-8578
www.premmag.com
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Unitrode Integrated Circuits Corp

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fax 1-603-424-3460
www.unitrode.com
Circle No. 316

DTMagnetics

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fax 1-603-742-9165
www.dtmagnetics.com
Circle No. 308

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www.newport-comps.com
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www.gowanda.com
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Philips Components

Saugerties, NY
1-914-246-2811
fax 1-914-246-0487
www.phillips.com
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Signal Transformer

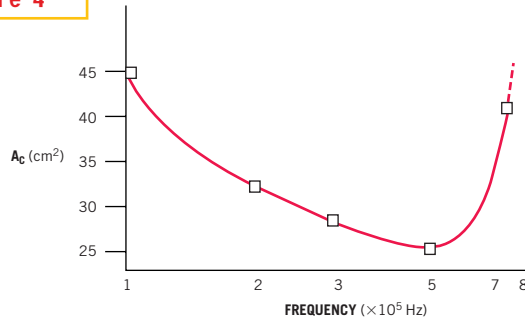
Inwood, NY
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transformer. This calculation uses the total power and the estimated transformer efficiency (say, 99%).

Figure 4

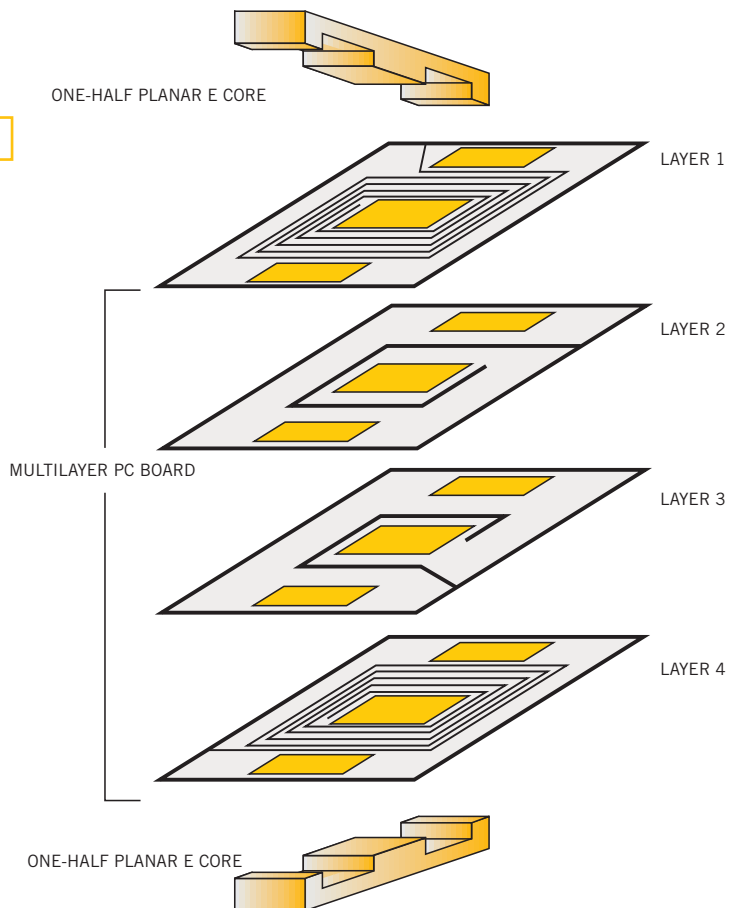


The ability to reduce core sizes as frequency increases reaches a point of diminishing returns. In this case, the inflection point is at 500 kHz.

- **Step 4**—Estimate the temperature rise. This step uses the empirical equation $\Delta T = (P_{TO-TAL} / (\text{surface area})) 0.833 \text{ C}$.
- **Step 5**—Estimate the required core size. This estimate uses the total power loss and the maximum temperature rise, along with temperature-rise data from Metglas data sheets.
- **Step 6**—Determine the flux swing (ΔB). This calculation uses the core loss per kilogram and the switching frequency.

Reference 1 gives the procedure to relate these two quantities by extrapolating from manufacturers' curves, as in figures 2, 3, and 4. Allied makes the calculation more convenient by providing an empirical equation: $P_{CORE} / \text{kg} = 6.5 f^{1.51} \Delta B^{1.74} \text{ W/kg}$. The loss is an exponential function of frequency.

Figure 5



Planar-transformer technology not only reduces size, but also provides better thermal characteristics, low leakage inductance, and excellent unit-to-unit repeatability.

- **Step 7**—Calculate the number of primary and secondary turns. This calculation uses the input and output voltages, duty cycle, flux swing, rectifier loss, and transformer-saturation data.
- **Step 8**—Calculate the primary and secondary current densities. The goal is to impart the same current density to all windings, which eliminates field gradients and resulting hot spots.
- **Step 9**—Select suitable conductors and estimate total copper losses. Use skin-effect data to choose conductors. For high-frequency designs, you will likely specify Litz wire or copper foil.
- **Step 10**—Calculate actual core hysteresis losses. This step uses the ΔB and P_{CORE} data from Step 6 to determine the total core loss.
- **Step 11**—Calculate the estimated temperature rise. This calculation reverts to the estimated figure in Step 4 and assumes that the transformer design equalizes core and copper losses.
- **Step 12**—Equalize the core and copper losses. Manipulate the conductor size and number of turns to make the core and copper losses equal. This step ensures that the power losses per unit volume are uniform.

As you can see, designing with mag-

netic materials is not a piece of cake. However, the step-by-step procedures available from materials manufacturers remove much of the black magic from the task. You might think that designing an inductor would be a laughable chal-

lenge, compared with the transformer-design procedure just described. Not so. The procedure in Reference 3 for power-factor-correction inductors entails 16 steps, four more than the transformer procedure requires. If you have a handful of op amps, resistors, and capacitors available, you can forget about magnetic-materials' idiosyncrasies in certain applications by configuring "virtual inductors" (see sidebar "Get the L out of there!").

TRANSFORMERS GO PLANAR

In keeping with the trend toward miniaturization, and espe-

cially diminishing board-to-board spacing, magnetic-materials manufacturers are offering their cores and finished products in planar configurations. This topology uses two thin E-shaped cores that enclose stacked printed layers of windings,

or printed windings, in a multilayer pc board (Figure 5). Planar magnetics offer a low profile, good thermal characteristics, low leakage inductance, and good repeatability of properties. Philips Components gives the design procedure for using the company's E cores (Reference 5). The procedure, which is somewhat simpler than the procedure for C cores, uses equations that are full of data-sheet constants and exponents that Philips derived from empirical curve-fitting for its core materials.

Philips uses its planar-core technology to produce the integrated-inductive-component (IIC) line of IC-like, surface-mountable inductor and transformer products. The data brochure for the IIC line gives complete curves of all magnetic parameters, including the frequency performance (Figure 2). Reference 6 provides a design procedure for specifying IIC magnetics for a 10W dc/dc converter.

Signal Transformer Co also uses planar E cores to produce a line of off-the-shelf

TABLE 1—SKIN-EFFECT PENETRATION DEPTH

Frequency (kHz)	Depth (cm × 10 ⁻³)	AWG gauge
50	33.5	27
100	23.7	31
200	16.8	33
300	13.7	35
400	11.9	36
500	10.6	38
750	8.7	39

and custom planar transformers. References 7 through 10 give procedures for specifying and designing transformers for various power-supply configurations. The SHF 2000 Series from Signal is available in six configurations with four isolated six- or eight-turn windings. You can connect the windings to accommodate a range of input voltages. The transformers can deliver as much as 200W dc at switching frequencies to 1 MHz with input voltages of 120 to 375V ac.

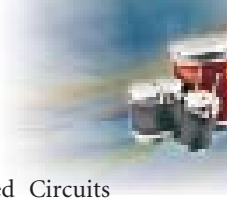
FINISHED PRODUCTS

The previous sections on rolling your own magnetic components notwithstanding, in many applications you can save design effort and time by selecting

an off-the-shelf inductive component. You have a vast array of choices; the world teems with magnetic-component manufacturers. The following section discusses a few recent offerings. Many newly announced inductors and transformers are surface-mountable for automated assembly. Gowanda

Electronics claims its SMP3013 surface-mount power inductors are the industry's smallest. They measure 0.3 × 0.11 × 0.125 in. and offer inductance values of 1 to 1000 μH in operating-current ratings of 0.105 to 0.78A. The SMP3013 data sheets give tightly specified inductance, dc-resistance, and saturation-current figures.

A range of surface-mount inductors and transformers is also available from Pulse. The line covers power inductors, quad-port T1/E1 LAN transformers, and 10/100 BaseT single-port LAN transformer modules, for example. The 10/100 BaseT product interfaces with National Semiconductor's (www.national.com) DP (Phyter) single-port, 10/100-Mbps physical-layer device. Newport Compo-



nents Inc also offers specialized, surface-mount transformers for particular applications. The 766 Series is suitable for line coupling, matching, and isolation in data-processing systems. The 7825 Series targets isolated RS-232C and RS-485 interfaces based on Maxim Integrated Products' (www.maxim.com) chip sets. More surface-mount transformers are available from Prem Magnetics. The company recently announced the SPT-048 telecomm transformer, which has the British Standards Institute's approval for use in V.34 modems.

You can obtain inductors for any specialized application. One area of specialization is high voltage. The DHV Series of high-voltage inductors of DTMagnetics uses bobbin cores to provide inductance from 10 μ H to 10 mH. A solder-wave-resistant insulation layer between the coil and the core allows you to ground the core with continuous coil voltages as high as 1500V dc, 600V ac.

Returning briefly to the concept of rolling your own magnetic components,

MMG North America has announced a low-profile cup-and-tack assembly that combines a bobbin core and a shielding sleeve in one 1.55 \times 3.53-mm unit. You can obtain the cores in nickel-zinc or manganese-zinc ferrite material. The tack portion of the assembly is fully enclosed by the cup and can accommodate an air-core winding. Applications for the assemblies include boost inductors in dc/dc converters, filter inductors, and broadband matching devices.

Designing and specifying magnetic materials and devices is not easy. These passive components have many more critical design parameters than resistors and capacitors, and their highly frequency-dependent behavior complicates the design of high-speed systems. Thankfully, magnetic-materials manufacturers provide a lot of cookbook design advice and procedures for specifying the materials and devices. \square

References

1. Andreyckak, William, "1.5-MHz, Current-Mode, IC-Controlled 50W Power

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2. "High-Power/ High-Frequency Switched-Mode Transformer Design Using Powerlite C-Cores," Allied Signal Inc.

3. "Power-Factor-Correction Inductor Design for Switched-Mode Power Supplies Using Powerlite C-Cores," Allied Signal Inc.

4. "Design of High-Frequency Output Inductors Using Metglas Amorphous Choke Cores," Allied Signal Inc.

5. "Design of Planar Power Transformers," application note, Philips Components.

6. "10-Watt DC/DC Converter using IIC Magnetics," application note, Philips Components.

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9. "PC Layout Guidelines," application note, Signal Transformer Co.

10. "Using Planar Transformers in Low-Profile Switch-Mode Power Supplies," application note, Signal Transformer Co.



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