Power Supply Magnetics

Part I: Selecting Transformer/Inductor Core Material

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A review of magnetic theory and core applications from DC to 100 kHz. Materials covered include silicon steel, powdered iron, Sendust, Molypermalloy, Hi-Flux™ and ferrites.

Switching power supply transformers and inductors require the use of high performance core materials. Choosing the proper core material for a specific application requires an understanding of core material technology.

A good start on understanding the magnetic principles is magnetizing force, which is proportional to the number of ampere turns and inversely proportional to the magnetic path length. The cgs and mks units of magnetizing force are the oersted and the ampere-turn/meter, respectively. The cgs and mks units of magnetic flux density are, respectively, the gauss and tesla. One oersted is 79.6 ampere turns/meter ($250/\pi$) and 10,000 gauss is one tesla. Magnetic flux density is designated by B, and magnetizing force is designated by H. A flux change of 1 gauss per second over a square Cm of core area will induce 10^{-8} volts/turn. One volt will be induced in a turn of wire that encloses one square meter of core area and whose flux density changes at one tesla/sec.

In non-magnetic materials, B is a linear function of H. Iron, cobalt and nickel, are examples of ferromagnetic materials. When subjected to a magnetizing force, such materials and many of their alloys have much higher internal flux levels than non-magnetic materials. As the magnetizing force varies, these materials have varying degrees of nonlinear flux levels.

The flux level of magnetic materials reaches a limit called saturation, regardless of how much magnetizing force is applied. Saturation flux levels of various magnetic materials
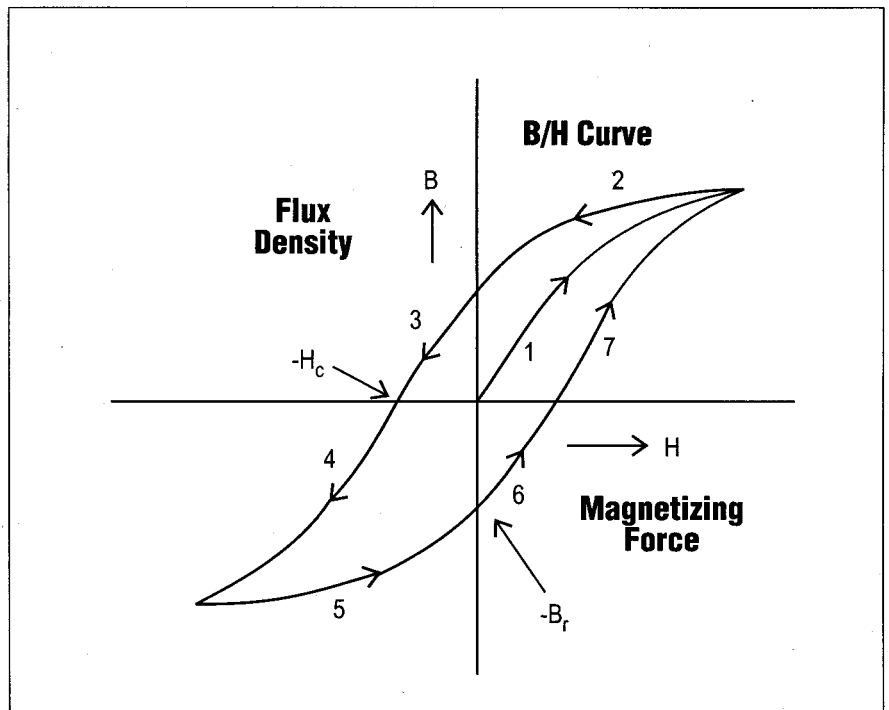

Figure 1. B/H Curve of Typical Soft Magnetic Material With Large Hysteresis.

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may range up to 24.5 kilogauss. Ferromagnetic materials also retain various amounts of permanent magnetism after removal of the magnetizing force.

Magnetic materials vary in their ease of magnetization, which is described by their permeability and designated by μ. Material maximum permeabilities can range up to 1,000,000 for Supermalloy. The permeability of air is 1.0000004 and may be considered equal to that of a vacuum. As long as a magnetic material remains linear and has no appreciable hysteresis, $B = H \times \mu$. For large magnetizing forces, ferromagnetic materials cease to obey this relationship and become nonlinear.

Any permanent magnetism in a material constitutes stored energy. An amount of demagnetizing force called the coercive force (Hc) is necessary to demagnetize the material. Materials that are easy to magnetize and demagnetize are referred to as soft magnetic materials. Residual flux in soft magnetic materials ranges from 50 to 15,000 gauss and coercive force varies from .01 to 10 oersteds.

B/H Curves

Figure 1 shows the behavior of a typical soft magnetic material having a large amount of hysteresis. It was initially demagnetized and then driven with an AC signal for a magnetizing force. Flux density in gauss is shown along the vertical or B axis and magnetizing force in oersteds is shown along the horizontal or H axis.

The relationship of energy storage to B/H curves will now be explained with winding resistance effects ignored. The voltage (E) across a winding is proportional to the product of the rate of change of flux (dB/dt), the core area (A), and the number of turns (n). The current (I) required to drive a given core is proportional to the desired magnetizing force (H) and path length (l), and inversely proportional to n.

The power (P) supplied to the core at any given instant is the product of the voltage and the current. When multiplying the expressions for E and I together we find that the number of turns cancel and that P is the product of A, l, H, and dB/dt. The product of A and l is the core volume or v. To find the energy provided to the core (U), we must integrate P over the time interval during which the core flux is changing. P=vH(dB/dt), so Pdt=vHdB, where v is a constant. The integral of Pdt or U is the energy expended by the drive source and the integral of vHdB is the energy stored in the inductor. The integral of HdB is the energy stored in the core per unit volume.

Figure 2 shows the current increasing from zero to a positive value, which makes the voltage positive. The energy stored per unit volume is not the area under the curve but rather the area to the left of the curve. This is because B is plotted vertically and H is plotted horizontally. If it was a H/B curve, the energy stored would be the area under it.
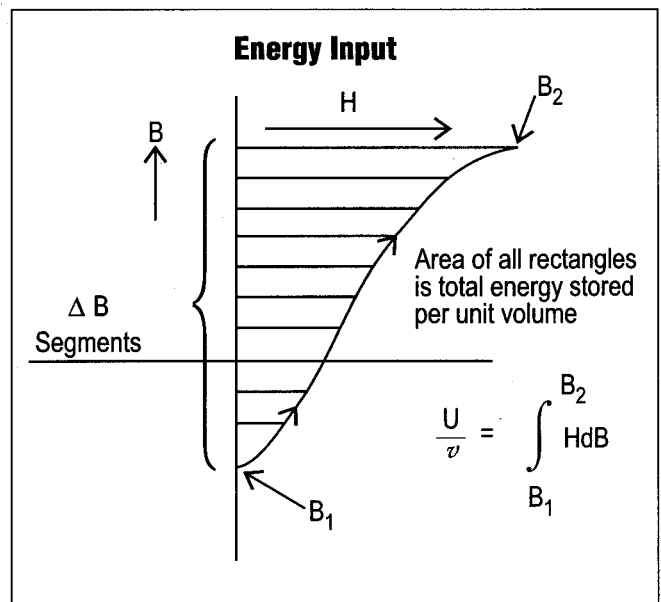

Figure 2. *Area of All Rectangles Is Total Energy Stored Per Unit Volume.*

If the current decreases from a positive value toward zero, the voltage will be negative while the current is still positive. Note that *Figure 3* is elevated above the curve in *Figure 2*. This is a region where the core delivers energy back to the drive circuit. The area to the left of the curve is now the energy returned from the core to the drive circuit. Due to hysteresis loss, eddy current loss, and residual loss this energy is always somewhat less than the energy originally stored in the core.

The difference between the area left of the curves of *Figure 2* and *Figure 3* is simply the area inside the right half of the loop of *Figure 1*. This is also the hysteresis energy lost in a half cycle of operation. The energy lost in the other half cycle of operation is the area inside the left half of the loop of *Figure 1*. Therefore, the total core loss per unit volume for a complete cycle of operation is the area inside the B/H loop plus the eddy current loss and residual loss.

Saturation

No satisfactory standard exists in the industry for definition of the saturation point of core materials. Identical core materials made by different manufacturers are advertised with different saturation levels because of different measurement methods. These published figures are much higher than practical operating levels and are frequently misleading to the designer.

In *Figure 4*, at point 1 the AC permeability is 60, at point 2 it has decreased to 30, and at point 3 it has been reduced to 15. Most filter choke applications have enough DC present at the operating point to reduce the AC μ to one-half its low level value. By the time the AC permeability has dropped to one-eighth of its initial value, the core may usually be considered saturated. AC permeability is the main parameter of interest to filter choke designers.

The DC permeability of a core may be defined as B/H. This ratio is the same as μ at low drive level and becomes progressively smaller as the drive level is increased. Here, saturation is defined as the point where the DC permeability has dropped to half its initial value or μ/2. The DC permeability is the average AC permeability over the entire drive range.

At point 4 in *Figure 4*, the DC permeability is 30 and the core meets the definition of saturation. Here H is 227 oersteds and B is 6700 gauss. The AC permeability is about 7.5 at this point. Core properties at higher drive levels are of academic interest only. Using this method of saturation measurement allows ready comparison of the energy storage capability of any permeability core.

DC permeability is easy to measure and is repeatable. At the point of saturation, AC permeability will usually have decreased to about one-eighth of its initial value. This depends to some extent upon the exact shape of the B/H curve, but the saturation figure coincides with the end of usefulness of the core in practice. The proper operating level of a core depends upon its specific application, but meaningful saturation specifications greatly assist in its determination. Saturation levels given here
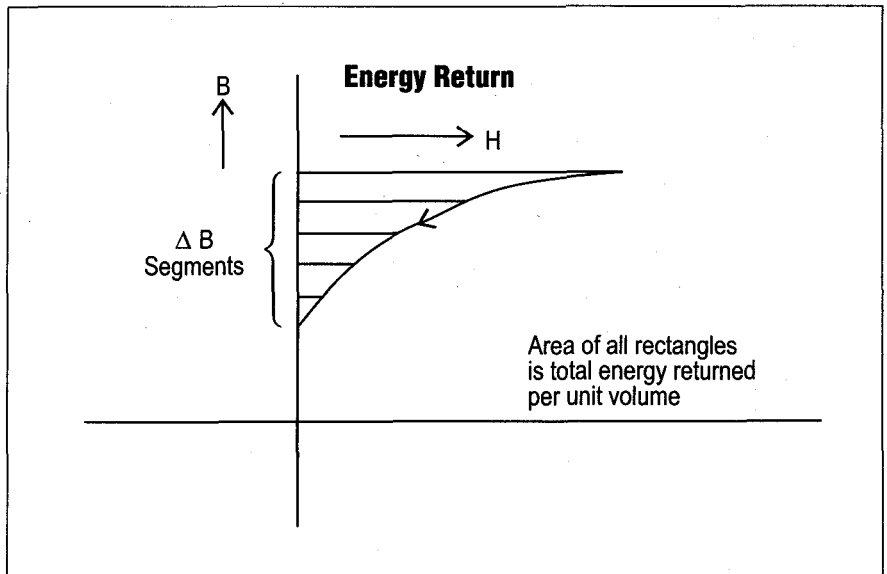

Figure 3. *Area of All Rectangles Is Total Energy Returned Per Unit Volume.*

may not agree with published data of manufacturers using different measurement methods.

Losses in Magnetic Components

For a given core size and frequency, the maximum power output of a transformer is proportional to its operating flux density. As the flux density is raised by removing turns or increasing operating voltage, core losses increase. A transformer is most efficient if it is loaded so that its copper loss is equal to the core loss.

The same rule applies for maximizing Q in an inductor.

Figure 5 shows the derivation of this rule.

Efficiency =

$$\frac{R_L (R_p)^2}{[R_L (R_P + R_S) + R_P R_S][R_L + R_P]}$$

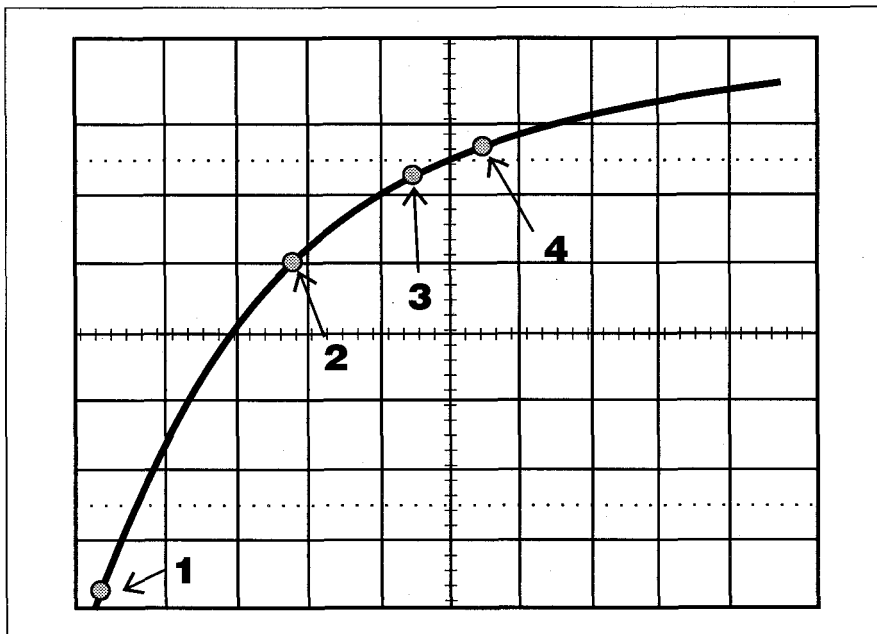If $R_L = \sqrt{\dfrac{R_p^2 R_s}{R_p + R_s}}$, efficiency is maximized.

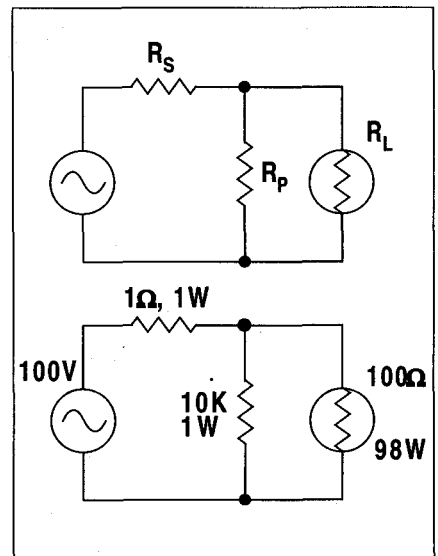**Figure 4.** *MPP Core: 1.06 in., μ=60; Vertical: 1 kgauss/div., Horizontal: 40 Oersteds/div.*

Figure 5. *Transformer Model Showing Winding Resistance, Core Loss and Load.*

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If $R_P >> R_S$ then $R_L \approx \sqrt{R_p \times R_s}$

$R_L = \sqrt{10K \times 1\Omega} = 100\Omega$

R_p represents the core loss and dissipates nearly the same power regardless of load. R_s represents the total series resistance of the transformer windings corrected for turns ratio effects. The power in R_s is proportional to the power in the load. Efficiency is the power in R_L divided by the total power in R_L, R_p, and R_s.

By differentiating the expression for efficiency and solving it for zero, we can find the maximum. Ordinarily we may assume that R_p is much greater than R_s. We find that efficiency is maximum when R_L is the geometric mean of R_p and R_s. In that case, the power in R_p is equal to the power in R_s. A typical example is shown for a 10k R_p, a 100Ω R_L, and a 1Ω R_s where the maximum efficiency is 98%. *Figure 6* shows that the efficiency drops to 96% with both a 25Ω and 400Ω load.

For the heavy load:

$$\text{Efficiency} \approx \frac{370}{370 + 1 + 15} = 96\%$$

For the light load:

$$\text{Efficiency} \approx \frac{25}{25 + 1 + 1/15} = 96\%$$

Figure 7 shows the model for an inductor that considers winding resistance and core loss.

$$Q \approx \frac{X_L R_p^2}{R_P^2 R_S + X_L^2 (R_P + R_S)}$$

$$\approx \frac{X_L R_P}{R_P R_S + X_L^2}$$

If $X_L = \sqrt{\dfrac{R_p^2 R_s}{R_p + R_s}}$

Q is maximum and

$$X_L \approx \sqrt{R_p \times R_s}$$

Q of an inductor is the tangent of the phase angle between voltage and current. Ordinarily, R_p is much greater than R_s. At very low frequencies the reactance of the inductor is near zero and the finite resistance of R_s gives a Q near zero. At very high frequencies, the inductive reactance is very high and the effect of R_s is very small. However, the shunt loss of R_p causes the Q to again approach zero. The Q will reach a maximum at some intermediate frequency.

The expression for Q in *Figure 7* may be solved for a maximum as well, by setting its derivative equal to zero. As in the transformer case, R_p can be assumed to be much greater than R_s. Much like the efficiency of a transformer load, the Q of an inductor is maximized when its reactance is the geometric mean of R_p and R_s. This shows that inductor Q is maximized at a frequency where the core loss is equal to the copper loss.

A given core has a unique L/R time constant associated with any full winding. Take for example a winding of any convenient number of turns for reference. If a winding is made with twice as many turns from wire of half the cross section, its resistance will be four times greater than the reference winding. This is because its length is twice as great and its resistance per unit length is also twice as great. Accordingly, a winding of half the turns made from wire of twice the cross section will have one fourth the resistance. Therefore, the winding resistance is proportional to the square of the number of turns. Because the inductance of a core is also proportional to the square of the number of turns, their ratio is a constant. This L/R ratio is equal to the inductive time constant. It can range from several milliseconds to several seconds, depending upon core permeability and size.

Silicon Steel Cores

Silicon steel is used to make laminated cores and tape wound cores for power transformers in the 50 to 400 Hz range. Many different alloys are used but 97% iron and 3% silicon is the most common. Silicon steel saturates at about 17 kilogauss and has a permeability of about 10,000. Because iron is highly conductive, solid steel cores act as shorted turns and cause huge circulating currents. These current paths resemble the eddies in a stream bed and are called eddy currents. Such eddy currents can cause very large losses.

Eddy current loss can be greatly reduced by dividing the core into a large number of thin slices that are insulated from each other by an oxide film. These slices are called laminations and are shown in *Figure 8*. If they are sufficiently thin, losses can be reduced to acceptable levels. There is a racetrack-shaped eddy current path in the lamination, but the voltage around it is very low due to its small area. The resistance around the path is relatively large because it is so thin.

The combination of low voltage and high resistance results in a low power loss per lamination. Thinner laminations further reduce eddy current losses but increase the cost of manufacture. Diminishing returns require a certain minimum thickness. In most transformer designs, the total eddy current loss is negligible compared to hysteresis.

Toroids are rolled up like a tape spool from a long thin steel strip coated with an oxide film. Individual layers of the tape core serve the function of laminations as shown in *Figure 9*. Toroids are the most efficient magnetically but require special machines to install their windings. This increases the total cost of manufacture to some extent.

Even when eddy current losses are made negligible, large hysteresis losses remain. At the 60 Hz power line frequency these losses are manageable and silicon steel is the most efficient core material. Its efficiency

is greatest at about 400 Hz provided the laminations are sufficiently thin. Aircraft electrical systems operate there for maximum efficiency and reduced weight. Flux densities must be reduced above that frequency to prevent core overheating. For transformers at 1 kHz and above, ferrites are generally more efficient than silicon steel.

Insulated bobbins are used to prevent lamination-to-winding shorts. Tape wound toroids are coated with an insulating material to prevent core-to-winding shorts. An air gap may be inserted in steel cores by sawing them in two, but burrs from the sawing process cause the laminations or tape layers to short together. These shorts must be removed by chemical means or polishing to prevent severe eddy current losses at the edge of the gap.

Litz Wire and Fringing Fields

At high frequencies, magnetic effects cause current to flow only near the surface of large diameter conductors. This phenomenon, called skin effect, makes the effective resistance of such conductors much greater than might be expected. Skin depth of a material is inversely proportional to the square root of the products of frequency, conductivity, and permeability [1].

Copper has the same permeability as a vacuum and its skin depth is only 0.33 in. at 60 Hz. Silicon steel laminations have only 1/35th the conductivity of copper and a permeability 10,000 times greater. The skin depth at 60 Hz in laminations is therefore only about 20 mils. The 20 kHz skin depths in copper and laminations are 18 mils and 1 mil, respectively.

Magnetic conductors are essentially useless at high frequencies. If large amounts of high frequency current are passed through steel screws, they will run very hot. Brass or aluminum screws should be used instead. Solid conductors are inefficient if their diameter is greater than two skin depths at their operating frequency. They work poorly, weigh
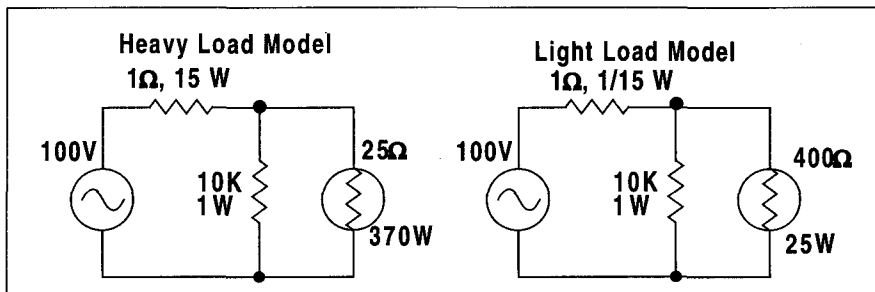
Figure 6. Efficiency for Heavy and Light Loads.

more, and cost more. Hollow conductors that are about one skin depth thick are the most economical. At 20 kHz, use of solid wire larger than #14 is a waste of money because of its excessive losses. At 200 kHz, #24 starts to become ineffective.

Thin foil may be used in high frequency transformers to avoid most skin effect losses but it is difficult to insulate properly during winding. Flexible printed circuit material consisting of copper foil on Kapton has been successfully used to wind high frequency transformers. As an ex-
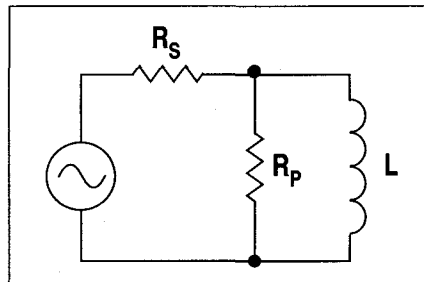
Figure 7. Inductor Model Considering Winding Resistance and Core Loss.

ample, 2-in. wide Kapton laminated with 4-mil thick copper foil is equivalent to #10 solid copper with very low skin effect losses.
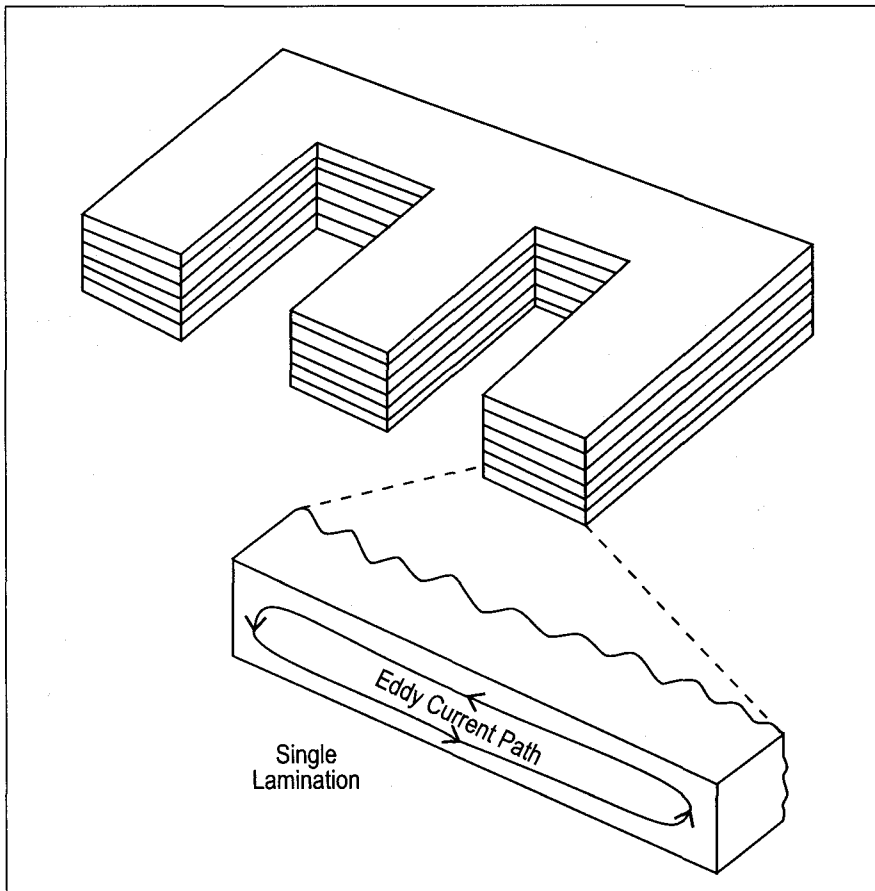
Figure 8. Laminated Transformer Stack.

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If large wire is made up from many small strands that are insulated from each other, skin effect losses are greatly reduced. This type of wire is called Litz wire. Strands near the center are gradually brought out to the edge of the bundle and vice versa. This prevents current from flowing only on some strands. As an example, 200 strands of #38 insulated magnet wire are used to make Litz wire that is the DC equivalent of #15 solid wire.

When a gap is introduced into a magnetic path, the field fans out around its edges as shown in *Figure 10*. This fringing field causes eddy current losses in any windings into which it extends. The larger the diameter of the wire in a fringing field, the greater its eddy current losses will be. Because copper conducts about 35 times better than silicon steel, the losses in large copper conductors are much more severe than in thick laminations. Litz wire is effective in reducing the eddy current losses in windings subjected to a fringing field. At 20 kHz, Litz wire can make a 10:1 improvement in the Q of gapped ferrite inductors.

Litz wire is effective in reducing fringing field loss, but its diameter is about 40% greater than solid wire of the same cross sectional conductor area. This is a result of its complex twisted structure and the presence of a large amount of internal insulation. A substantial winding area penalty must be paid to use Litz wire. The DC resistance of a winding using it will be about twice as great as if it were done with solid magnet wire. It is also hard to strip and about three times more expensive than solid magnet wire. Powdered cores usually do not need Litz wire because they have no fringing field. Litz wire is ordinarily used only where absolutely necessary.

Powdered Cores

Powdered cores are made from a variety of materials. A magnetic al-
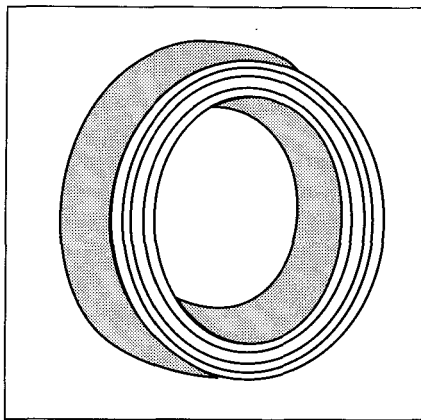
Figure 9. Tape Wound Core.

loy is made into fine particles from 5 to 200 microns in diameter. The particles are coated with an insulating material from 0.1 to 3 microns thick. The powder is then forced by punches into dies at pressures up to 300,000 psi. The high pressures require the use of tungsten carbide tooling. A variety of shapes are possible but toroids are the most common shape produced. Small toroids are coated with parylene and larger toroids are coated with an epoxy paint. This coating is necessary to prevent shorts during the winding process or later during operation.

Adjusting the size of the magnetic particles, thickness of their coating, and pressure allows the permeability to be varied from 4 to 550. The final product is then annealed at high temperature in a hydrogen atmosphere. This relieves stresses created during pressing, prevents oxidation, and improves the final magnetic properties of the core.

Because the individual magnetic particles are insulated from each other, eddy current losses are minimized. The insulating material provides the equivalent of a distributed air gap that lowers the permeability and allows the core to store substantial amounts of energy. The lack of a localized air gap eliminates the nuisance of a fringing field and its associated losses in nearby windings.

Powdered iron is made in permeabilities from 4 to 85 and is the least costly of all powdered cores. Because of manufacturing difficulties, the higher permeabilities avail-
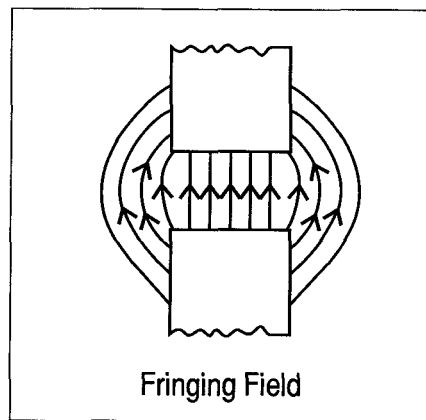
Figure 10. Gapped Core.

able with other core types are impractical. In addition to toroids, it is made in slug form for variable inductors as well as E and I cores. It saturates at about 10 kilogauss but has very high losses.

While the losses in powdered iron are high, they are less than in silicon steel. It is useful in slug form at very low flux levels for tuned transformer adjustment. Powdered iron is also used in lower performance filters where large amounts of DC must be tolerated and cost is very critical. It runs very hot when subjected to substantial amounts of high frequency ripple current.

MPP stands for molybdenum permalloy powder and is also known as molypermalloy[2]. MPP cores are manufactured from a powdered alloy of 2% molybdenum, 81% nickel and 17% iron. Molypermalloy toroids are made in permeabilities from 14 to 350. This is the widest range available with any type of powdered core. MPP saturates at 7 kilogauss and has the lowest loss of all powdered cores. It has the best temperature stability as well as the smallest permeability change from low to medium drive level. The high nickel content of MPP makes it the most expensive of all powdered cores, but its superior performance usually more than offsets its cost.

MPP is the best performing material for audio frequency filters, low level medium frequency tuned circuits, and smoothing chokes in switching power supplies. MPP cores may be used in high energy resonant

circuits operating near saturation up to about 3.5 kHz. When driven to saturation above that frequency, they tend to run too hot. It is the best material for current transformers in the 10 kHz to 1 MHz range, especially when large amounts of DC must be tolerated. Where short high energy voltage spikes must be isolated from a power bus, its low eddy current loss provides the most efficient solution.

Sendust was invented in Japan prior to World War II. It is made from an alloy of 6% aluminum, 9% silicon, and 85% iron in powdered form. Sendust toroids come in permeabilities from 60 through 125 under the trade name of Super-MSS™. This material has DC bias properties similar to MPP and has lower losses than powdered iron but more than MPP. High performance audio tape heads have been made from SENDUST. MSS is well suited to medium performance line and output filters.

Although it is somewhat more expensive than powdered iron, its greatly reduced losses of MSS almost always justify its cost premium. In switching power supply applications where loss is not too critical, Super MSS™ can usually be substituted for MPP. Because Super-MSS™ costs substantially less than MPP, it frequently is the best choice.

High Flux (HF) cores are made from a powdered alloy of 50% nickel and 50% iron. These HF toroids are available in permeabilities from 14 through 200. With a practical saturation of 11 kilogauss, it can store more than four times as much energy as MPP or MSS for the same core size and permeability. HF costs slightly less than MPP and is the best choice where maximum energy must be stored and core loss is only moderately critical. Its loss is much less than powdered iron but higher than MPP or Super-MSS™. HF is the best choice for low frequency high energy resonant circuits, and chokes that must pass large amounts of DC. It makes possible RFI filters for line operated supplies that can lengthen rectifier pulses and improve the power factor.

References

1. R. Landee, D. Davis, and A. Albrecht, Electronic Designers Handbook, McGraw-Hill, N.Y., 1957 page 20-11
2. MPP Core Catalog PC-104J, 8/94 Arnold Engineering Co., Marengo, IL
3. Super-MSS™ Core Catalog, Arnold Engineering Co., Marengo, IL
4. Hi-Flux™ Powder Core Catalog, PC 1461 Arnold Engineering Co., Marengo, IL

Power Supply Magnetics -
Part II: Selecting Core Material

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A review of magnetic theory and core applications from DC to 100 kHz. Materials covered include silicon steel, powdered iron, Sendust, Molypermalloy, Hi-Flux™ and ferrites.

Ferrites are a ceramic material made from iron oxide with manganese, zinc, or other metallic oxide. The components are mixed in powder form, then cast and fired. The result is a hard, brittle, glass-like substance. Common permeabilities range from 750 to 10,000. Saturation flux densities range from 3 to 5 kilogauss. Ferrites may be manufactured in any convenient shape such as toroids, E and I cores, pot cores, and slugs.

Ferrites have the lowest losses of all commonly available core materials due to their very low residual flux and high resistivity. They are the best choice for transformers from 1kHz to 1MHz. Unless gapped, they are not well suited to many high frequency applications. This stems from their large change in permeability and Q with temperature and drive level.

The resistivity of ferrites is typically a million times greater than most magnetic alloys. Although this gives them the lowest eddy current loss of any solid material, their windings must still be insulated from the core. Ferrite edges are razor sharp and will readily cut through the insulation of ordinary magnet wire during the winding process.

An insulating coating of parylene or epoxy paint is applied to ferrite toroids to prevent shorts during winding. Even though ferrites have a rather high resistance, eddy current losses are sometimes a problem. This occurs in high volts per turn applications, such as chokes with low frequency high voltage pulses. In such cases, MPP (molybdenum permalloy powder) is a better choice.

Core Shapes

Toroids are the least costly to manufacture and the most efficient magnetic shape. With a large number of turns they require special machines to install their windings, which offsets their lower cost to some extent. Gaps are not ordinarily used for toroids because of the difficulty of fastening the sections together.

E and I cores are more costly to manufacture because of the need to have the two core pieces fit together precisely. If they do not mate well, the sections will crack when clamped together. A gap is usually installed when required by grinding down the middle leg of an E core. Sometimes two such cores are mated together to double the gap length. The bobbin for E and I cores can be wound at high speed on inexpensive machines, which compensates for some of the higher core cost.

Pot cores are the most expensive core to manufacture. They are manu- factured in two halves which are designed to enclose a bobbin and are carefully mated together. The core sections almost completely surround the bobbin, except for small openings for the wires to the windings. Almost all of the magnetic field of the windings is confined to the interior of the core. If the core is grounded, the moderately conductive ferrite acts as an electrostatic shield for the windings as well. Pot cores are therefore the best shielded of all core types.

Pot cores suffer from heat buildup because their windings are surrounded by core material that does not conduct heat well and prevents air circulation. As with E cores, one or both pot core center legs may be ground down to provide an air gap. When a pot core is gapped, heating effects from fringing fields in the windings cause even greater temperature rise. Gapped pot cores may be equipped with threaded slugs in their center posts to adjust their inductance for tuned applications.

With an air gap, ferrites can be made stable in permeability as well as capable of storing substantial amounts of energy. While their permeability is reduced, they can tolerate large amounts of DC without saturation. Gapping is ordinarily accomplished by grinding down the middle leg of an E core or pot core. Litz wire is necessary to minimize

the fringing field losses due to the gap at higher frequencies.

In the early 1980s, Stackpole Carbon Company manufactured a low loss ferrite toroid with low permeability for energy storage. For reasons not known to the author, the product line was abandoned. The industry is in dire need of a resurrection of that or a similar technology.

Core Loss

The true cost to the end user of core loss is often overlooked. For example, consider a core that wastes 1 watt, when electricity costs 10 cents per kilowatt hour. In a year of continuous operation, this core wastes 88 cents of electricity. This heat may be useful in Marengo, IL in January but it is useless in Phoenix, AZ in July. In the latter case an additional dollar must be expended for air conditioning to pump that dollar's worth of wasted energy outside.

When the dimensions of a core are doubled, its surface area is quadrupled and its volume and loss increase eightfold. Temperature rise is proportional to the core loss to surface area ratio, which is twice as great in the larger core. Small cores run cooler than large cores and may be operated above efficient flux levels without overheating. On the other hand, cores over 3-in. diameter may require forced air cooling even though flux levels are reasonable.

Core loss causes an increase in winding temperature. Copper increases in resistance by 0.4%/°C. Therefore, a 30°C increase in temperature due to core loss increases copper loss by 12%, which increases the operating temperature even more. High core temperatures also cause deterioration of winding insulation and cause thermal stresses that may lead to winding shorts.

Most heat-driven chemical reactions roughly double in speed for each 10°C increase. The wear out mechanisms in most electronic components are heat-driven and their life is therefore halved for each 10°C increase in operating temperature.

$$E_L = \frac{dB}{dt} \times n \times A$$

$$Y = \int \frac{E_L \, dt}{RC} = \frac{BnA}{RC}$$

$$B = \int \frac{E_L \, dt}{nA} = Y\left(\frac{RC}{nA}\right)$$

$$X = 0.1 \times I$$

$$H = \frac{n(0.4\pi)}{0.1} \times X$$
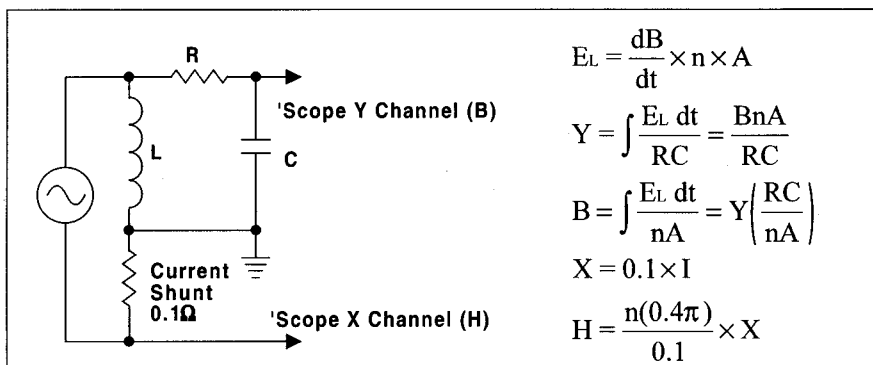
Figure 11. B/H Curve Test Circuit.

Wasted power inside electronic equipment raises internal temperatures and results in reduced component life.

Excessive heat can slowly oxidize and embrittle solder joints and char circuit boards. Long-term exposure of electrolytic capacitors to high temperatures will dry them out and shorten their life. Resistors change value when operated at elevated temperatures for long periods. High temperature operation of semiconductors tends to redistribute their carefully placed impurities and increase leakage. This leakage can elevate their temperature even more resulting in thermal runaway.

To maintain the reliability of semiconductors at higher ambient temperatures, larger and more expensive heat sinks are required. In many cases the additional heat sink cost will exceed the savings on a cheap core that runs hot. Frequently, a cooler running more expensive core can eliminate a fan and reduce overall cost.

An apparent saving of $1 on a core may cost the customer $100 if a power supply has to be repaired in the field. The author uses a figure of $1/W to justify a higher parts cost for saving energy and reducing operating temperature. The customer will accept this increased cost if the benefits are properly explained.

Measuring B/H Curves

Flux densities inside cores can only be determined indirectly by measuring the voltage induced in a winding by changes in the current through it. B/H curves are produced from these measurements, which allow cores to be evaluated for linearity, saturation, permeability, and hysteresis. A core is tested by driving it from an AC power source and monitoring its current and drive voltage on an oscilloscope.

Figure 11 shows the test setup for displaying B/H curves. A 0.1Ω current shunt allows measurement of the core drive current. The current required per oersted of magnetizing force, H, is:

$I = \ell H / 0.4\pi n$

Where:

ℓ = Magnetic path length of the core in centimeters

n = Number of turns.

This sets the calibration factor for the H axis or x channel of the 'scope in oersteds/amp.

The voltage produced across the winding is:

$$E_L = \frac{dB}{dt} \times n \times A$$

Where:

$\frac{dB}{dt}$ = Rate of change of flux density

A = Core cross sectional area

As shown in *Figure 11*, the driving voltage is applied to an RC integrator whose time constant is much longer than the period of the driving AC source. The winding essentially differentiates the core flux level and the RC network integrates the differentiated flux level. The original flux level is thereby restored except for any DC flux present, which is lost.
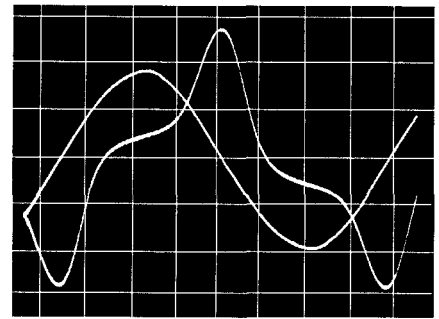
If the time constant of the core inductance with its associated wind-
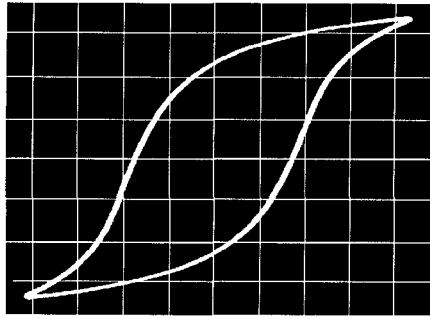
Figure 12. *Powerstat 1256 DU 40cm² x 80cm Tape Core, 70% of Normal Drive 60Hz Sine Wave.*
Vert. = 2 kgauss/div., Horiz. = 0.2 oersted/div.
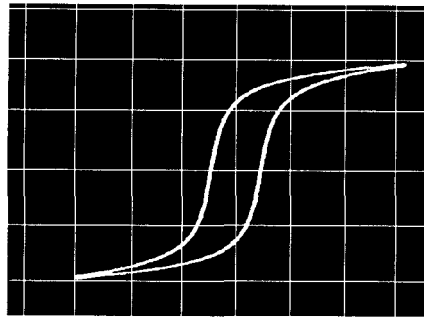
Figure 13. *Powerstat 1256 DU Normal Drive 60Hz Sine Wave (Shows Saturation).*
Vert. = 5 kgauss/div., Horiz. = 1oersted/div.

Figure 14. *Powerstat 1256 DU Sinewave Drive Waveforms.*
Vert. = 50V/div., Sine Wave, Vert. = 1A/div. Peaked Wave, Horiz. = 2msec/div.

ing resistance is equal to that of the integrator, errors due to voltage drops in winding resistance will cancel. The output of the integrator is applied to the scope Y channel or B-axis. The time constant sets the calibration factor with the core area and number of turns. A classic B/H curve is shown in *Figure 12*. The core shown is a Superior Electric Powerstat model 1256DU. It is a tape wound core of 40 square cm cross section and 80 cm path length. This curve shows the large hysteresis associated with high permeability silicon steel.

This drive level is only about 70% of normal and was chosen to better display the hysteresis and avoid saturation. The 64-turn tap was driven, resulting in one oersted per amp of drive current. Voltage at the tap was integrated by a 100kΩ resistor and 2.56F capacitor, which yields an output of 10 kgauss/V.

Figure 12 shows that the residual flux of the core is about 4.6 kgauss in each direction after the drive current has returned to zero. It further requires a coercive force of about 0.35 oersteds to demagnetize the core in each direction. Maximum core flux level is about 6.5 kgauss. As the drive level is increased, the B/H loop enlarges in a corresponding manner as saturation is approached.

In *Figure 13*, the core is being operated at normal drive levels. Note that the scale factors have been changed and saturation is now evident at about 10 kgauss. The area

inside the loop is about 3 square divisions or 15,000 oersted x gauss. Converting to mks units yields 119 (amperes/meter) x teslas. This loop area is the hysteresis loss per cycle in joules for each cubic meter of core material. The result is 23W, which is approximately the actual power loss.

The actual drive voltage and current for this test are shown in *Figure 14*. The voltage is a sine wave of ±99V but the current is distorted due to saturation effects. The current peaks at ±2.9A when the voltage passes through zero. Most 60 Hz power transformers operate well into saturation.

B/H Curves

Each of the following cores was driven by a square wave from a power MOSFET switch as shown in Figure 15[3]. Such a switch handles reverse current just as well as forward current. Current shunt resistors, integrator time constants, and scale factors were changed for the following cores as required. Figure 16 shows the B/H curve on a Philips (formerly Ferroxcube) ungapped EC70/70 (3C81) ferrite with 200 turns of #17 on its bobbin[4].
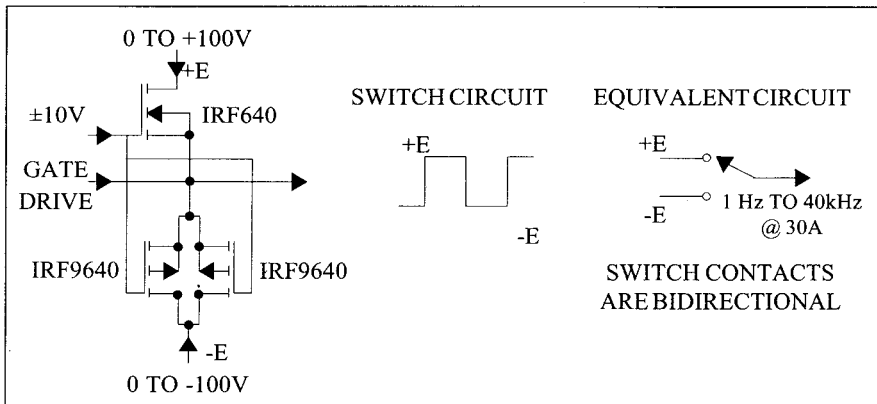
The inductance is about 160mH at 90mA. The residual flux is only 12% and the coercive force is only 45% as great as the silicon steel tape core in *Figure 13*. This means the ferrite has a much lower hysteresis loss. The same ferrite core is shown in *Figure 17* with an air gap (EC70/70G) in-

stalled in the center post. The drive voltage remains the same as before but the current is greatly increased. The inductance is now 4mH at 3.5A and the scale factors have been changed. The B/H curve has been made dramatically more linear by the 190 mil air gap, but saturation still occurs at 3 kgauss. An air gap, whether distributed or not, merely expands the H axis of a B/H curve. It does not affect core loss for the same frequency and flux level.
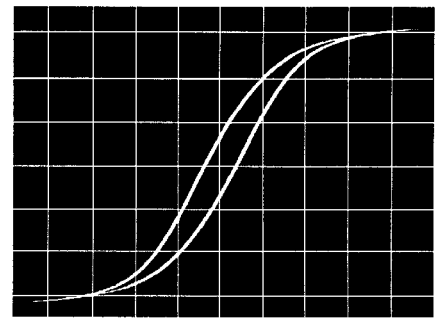
MPP, Super-MSS™, Hi-Flux™, and powdered iron cores were each used to wind 4mH chokes. The cores were 1.84-in. diameter toroids with 60 permeability wound with 172 turns of #17. *Figure 18* shows the curve for the MPP core. It is apparent that the gapped ferrite is much more linear than the MPP core before the onset of saturation.

The B/H curve for the powdered iron core is shown in *Figure 19*. Its hysteresis loop is massive compared to the MPP core and its loss is comparably huge. Many types of powdered iron have a large permeability drop at very low drive levels. The core used here had about a 45% drop in permeability under low drive. The core ran very hot during these tests and demonstrated an objectionable sound level.

Figure 20 is a triple exposure of the right half of the B/H curves of the Super-MSS™, MPP, and Hi-Flux™ cores. The Super-MSS™ and MPP cores have nearly identical curves except for the greater hysteresis of

Figure 15. Drive Switch Circuit.

Figure 16. EC70/70 Ungapped Ferrite, 3C81 Material E/E Core 2.79cm² x 14.4cm, 1.2kHz Square Wave Drive.
Vert. = 1kgauss/div., Horiz. = 0.5 oersted/div.

the MSS. These two cores saturate at about 7 kgauss. The Hi-Flux™ core has more hysteresis than the Super-MSS core and saturates more gradually at a much higher 11 kgauss level. The losses are so low for MPP and Super-MSS™ that their B/H curves appear to be nearly a straight line. The loop for HF can be seen but is nowhere near as prominent as the one for powdered iron in *Figure 19*.
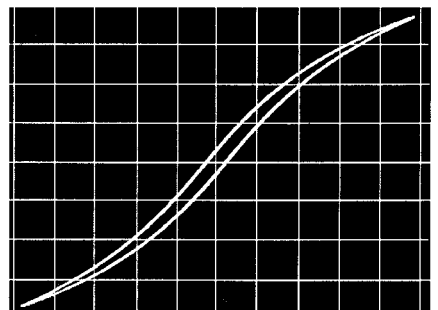
Naked ear listening tests during 1 kHz B/H curve testing demonstrate that powdered iron and gapped ferrites are equally offensive as far as producing audible noise. Hi-Flux™ was about 3db quieter. Super-MSS™ and MPP were both about 6 db lower in sound level than the powdered iron. Even these cores will cause objectionable sound when operated near 3 kHz at maximum flux levels. Because Hi-Flux™ cores ordinarily operate at very high flux levels, their noise can be objectionable. Passing a
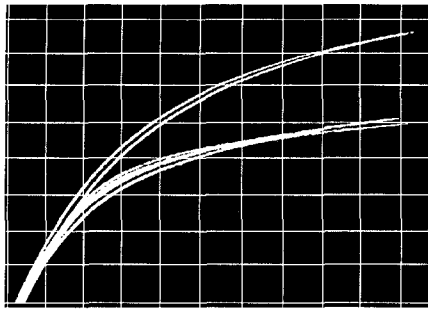
low frequency square wave current through them produces an irritating buzz. They are sufficiently quiet for line filters.

AC of several hundred volts or more are present on cores during many typical operating conditions. Arcs will occur in unprotected cores at these voltage levels. Copper magnet wire has burrs that sometimes penetrate the insulating coating. When high voltage is present, corona develops at burrs and cracks, which produces ozone. It causes rapid deterioration of organic materials and insulation and oxidizes exposed copper surfaces. An arc can be expected within an hour of smelling ozone.

After winding, high voltage cores can be dipped in Wacker T-84 liquid silicone. This material is thin enough to penetrate and fill the small cracks in insulation and prevent arcs. This type of coating does not become brittle with age and allows easy unwinding if the core must be reused.

Measuring Core Loss

Idealized waveforms associated with a 4mH choke driven to saturation by a 1 kHz square wave are shown in *Figure 21*. When current flows in the inductor, the energy stored in it is proportional to the inductance value and the square of the current. When the current becomes smaller, the inductor returns energy to the drive circuit. Energy is stored in the inductor when the drive voltage is positive and the current increases from zero to maximum. This is a time of positive power because energy flows from the drive source to the inductor.

When the drive voltage suddenly goes negative, energy is now returned from the inductor to the drive source. The power becomes negative at this point because the voltage is negative and current is positive. When the current crosses through zero and goes negative, energy now starts to flow

Figure 17. EC70/70 0.19-in. GappedFerrite, 3C81 Material E/E Core 2.79cm² x 14.4cm, 1.2kHz Square Wave Drive.
Vert. = 1kgauss/div., Horiz. = 25 oersted/div.

Figure 18. 1.84-in. MPP Toroid, 1.95cm² x 11.2 cm, μ=60, 1kHz Square Wave Drive.
Vert. = 2kgauss/div., Horiz. = 40 oersted/div.

Figure 19. 1.84-in. Powdered Iron Toroid, 1.95cm² x 11.2 cm, μ=60, 1kHz Square Wave Drive.
Vert. = 2kgauss/div., Horiz. = 40 oersted/div.

Figure 20. Triple Exposure HF/MPP/MSS 1.84-in. Toroid Material Comparison 1.95cm² x 11.2cm, μ=60, 1 kHz Square Wave Drive. Vert. = 2kgauss/div., Horiz. = 50 oersted/div.

into the inductor and power again becomes positive. At this point, voltage and current are both negative.

When the drive voltage becomes positive at the next square wave transition, drive power again becomes negative with energy being returned from the inductor to the drive source. Here the voltage is positive and the current is negative. Finally, the current crosses zero in a positive direction and power becomes positive again. The power waveform is a 2kHz sawtooth with a 5W DC offset corresponding to the core loss. To measure this small core loss in the presence of ±400W of reactive power requires a very accurate multiplying wattmeter.

An easy way to get around the above problem is to resonate the inductor under test with an appropriately rated capacitor. As shown in Figure 22, a 6F capacitor resonates
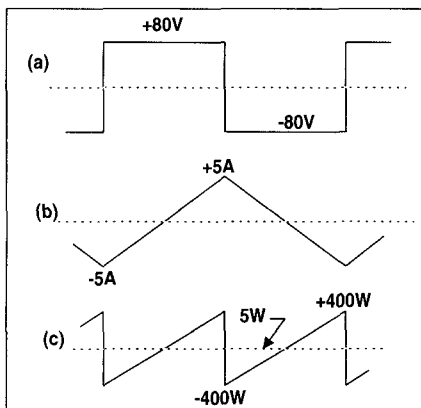
Figure 21. Core Loss Waveforms - (a) Drive Voltage, (b) Drive Current (c) Drive Power.

with a powdered iron inductor at 1 kHz. Figure 23 shows the switch output voltage and current. The capacitor voltage versus current is shown in Figure 24. Square wave drive requirements are now greatly reduced to ±10V and drive current is an in-phase ±6A sine wave.

A good capacitor has a negligible effect on the overall circuit Q. Circuit Q is the ratio of the capacitor voltage to the sine wave component (4/π x peak square wave voltage) of the driving square wave. The Q is about 9.4 and $X_c=X_L = 26\Omega$. The effective resistance of the inductor is the capacitive reactance divided by Q or 2.7Ω. Because the winding resistance is under 0.2Ω, the bulk of the Equivalent Series Resistance (ESR) is due to core loss. The low Q

is indicative of the excessive loss of powdered iron.

In a typical switching power supply application, the peak-to-peak smoothing choke ripple current is a function of the size of the core used. To make this ripple small requires a large and expensive core. If a smaller core is used to save money, the ripple current is higher. A higher ripple causes more heating from hysteresis loss and requires a larger filter capacitor. The most economical compromise is to use a core with a peak-to-peak ripple current of about one-fourth of the load current.

Hysteresis losses due to ripple current frequently dominate copper loss. A useful index of inductor performance in switching power supply service is a high level Q measurement at 40 kHz. This allows the ESR of the inductor to be determined. Q measurements were done for MPP, Super-MSS™, Hi-Flux™, powdered iron, and ferrite cores. Their DC ratings (with a 50% inductance drop) were 6, 6, 15, 13, and 3.5A. The permeability of the powdered cores was 60. Q at 40 kHz was tested at 2A peak-to-peak for the powdered cores and 1A peak-to-peak for the ferrite. The winding resistance was about 0.18Ω for the powdered cores and about 0.28Ω for the ferrite.

Figure 25 shows the drive circuit for the Q measurements. Resonance allows the core to be tested at a much higher voltage than can be readily
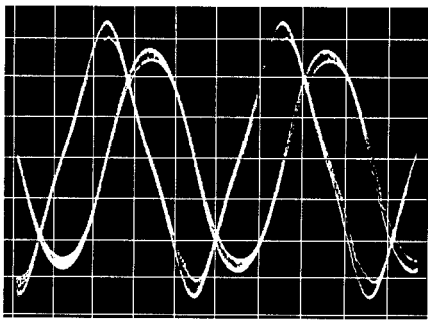

$$X_L = X_c = 26 \ \Omega \qquad \text{Resonance} = 1kHz$$
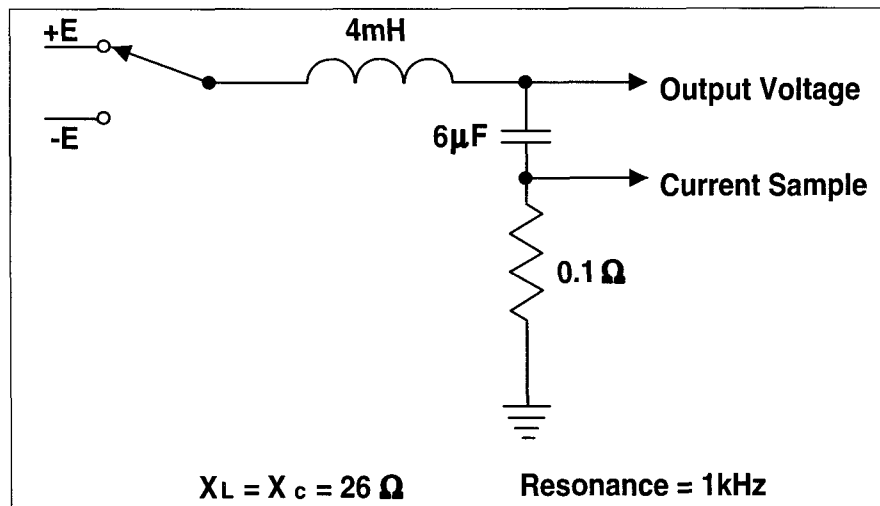
Figure 22. Resonance Core Tester
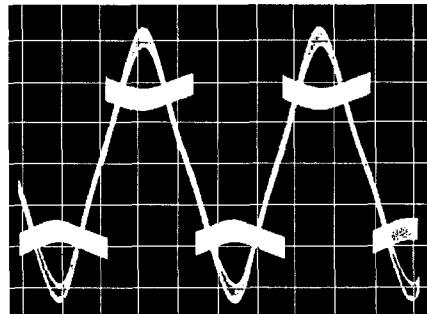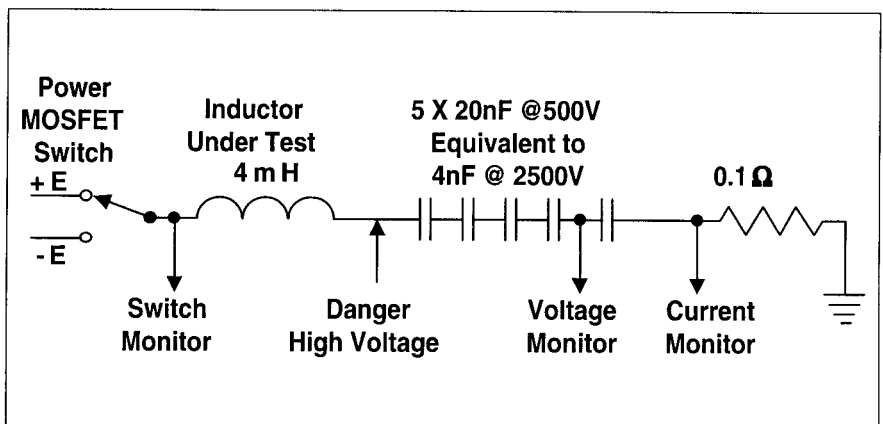

Figure 23. Powered Iron 1.84-in. Toroid, Resonant Drive With 6μF Capacitor Square Wave Drive.
Vert. = Square Wave 5v/div., Vert. = Sine Wave 2A/div., Horiz. = 200 sec/div.

Figure 24. Powered Iron 1.84-in. Toroid, Resonant Drive With 6μF Capacitor Square Wave Drive. Vert. = Capacitor Voltage (Inside) 50V/div., Vert. = Capacitor Current (Outside) 2A/div., Horiz. = 200μsec/div.

Figure 25. 40kHz Q Tester Circuit.
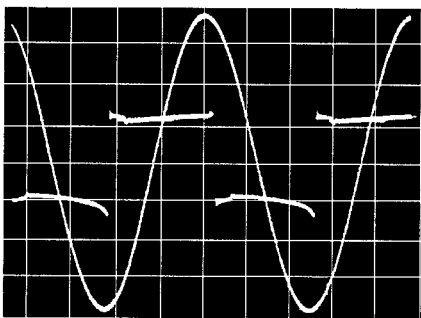
Figure 26. 40kHz Drive MPP Core With 4nF Capacitor. Vert.=Sine Wave 50v/div., Vert.=Sq. Wave 10v/div., Horiz.=5μ sec/div.

generated by an electronic switch. The square wave amplitude is set to give the desired core drive level and a current sampling resistor is provided. *Figure 26* shows the waveform for the MPP core. Note that the sine wave of ±1,000V presents the danger of 'scope probe destruction and an RF burn to an exposed finger. Except for the ferrite, almost all of the losses are from hysteresis. *Table 1* gives the results of the Q measurements and compares the power output and cost of the various cores. The ESR is so large that the effect of winding resistance is negligible.

The bulk of the losses of the ferrite originate from copper loss in its fringing field. Of course, the ferrite would have done much better if it had been wound with Litz wire or thin foil. A Q of 100 has been observed with

3C81 ferrite in this frequency range and drive level using Litz wire. Its power capability is still only 34% of MPP or Super-MSS™.

Q measurements show that for HF and powdered iron, the hysteresis loss from ripple current dwarfs the loss from DC resistance. MPP is clearly the star performer but Super-MSS™ runs a respectable second. Super-MSS™ cores cost between one-half and one-third as much as MPP and can frequently substitute for it in power applications. If the next larger core size can be used, Super-MSS™ will usually exhibit similar loss and lower overall core cost.

References

3. D. Pauly, "High Fidelity Switching Audio Amplifiers Using TMOS Power MOSFETs, AN1042 1989," Motorola Technical Information Center, Phoenix, AZ.
4. "Ferrite Materials and Components Catalog, Eighth Edition," 1994 Philips Components, Saugerties, NY, pp 1-12, 3-3, 4-24 through 4-33.

Core Type	Rated Current (A)	Output Voltage (V)	Output Power (kW)	Power Loss (W)	Power Loss (%)	Q (40kHz)	ESR	1k Price
MPP	6	320	1.92	11	0.6	74	14	$10.97
Super-MSS™	6	320	1.92	16	0.8	34	30	$3.02
Hi-Flux™	15	800	12	157	1.3	19	56	$10.68
Powdered Iron	13	693	9	131	1.5	16	64	$0.64
Gapped Ferrite	3.5	187	0.65	17	2.6	8	119	$12.00
Ferrite / Litz	3.5	187	0.65	8	1.2	100	10	$13.00

Table1. All Cores 1.84-in. diameter Toroids Except Ferrite That Is Gapped EC70/70G. Inductors Are 4.0mH. Ripple Current Is a 40kHz Triangle Wave at a Peak-to-Peak Level of 33% of Rated Current.

Power Supply Magnetics –
Part III: Selecting High Frequency Core Material

Donald E. Pauly, Consultant
Sponsored by The Arnold Engineering Company, Marengo, Illinois

> *Many trade-offs are possible in selecting the optimum HF core material for a specific application. Factors like volume, sound level, and cost enter into the decision.*

Hi-FLUX™ (HF) cores

can store very large amounts of energy for their size. They can operate up to 14 kilogauss with moderate loss. Among their many applications are test sets, RFI power line filters, and battery line filters.

For capacitor testing, high voltage polypropylene capacitors can be resonated with a high frequency inductor that employs an HF core. These capacitors have very low losses and are commonly rated at 480VRMS and 80A. Such capacitors find application in welders using SCR converters and as snubbers in GTO SCR power supplies for locomotive drives. The testing must be done with AC rather than DC to discover problems with lead attachments and end termi-

nations. The test circuit must be able to survive the shorting of the capacitor under test.

For testing, an over height version of the 26 permeability 5.2 inch diameter Arnold Hi-FLUX™ core was resonated with a 12μF polypropylene capacitor. This combination was driven by a power MOSFET switch operating from ±25V variable supplies. From low to high drive levels, resonance shifts from 190 Hz to 235 Hz.

Figure 28 shows that the inductor current departs considerably from a sine wave due to reduction in AC permeability at high peak currents. The core weighs about 3lbs and the

winding weighs about 5lbs. At full output, the Q is about 30. The operating frequency is picked below the frequency of maximum Q to keep the capacitor current low. About 210W is required to drive the capacitor to 600VRMS. Forced air cooling for the windings is required for more than a few minutes operation. After dipping in T-84 silicone, the core withstood ±3kV at 10kHz with no signs of corona.

To test this type of capacitor with 60 Hz AC would require a variac and transformer combination weighing about 30lbs. Furthermore, a circuit breaker would be too slow to prevent capacitor destruction from momen-
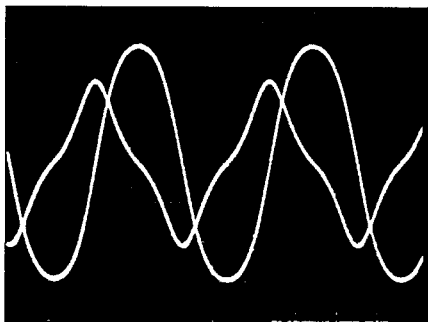

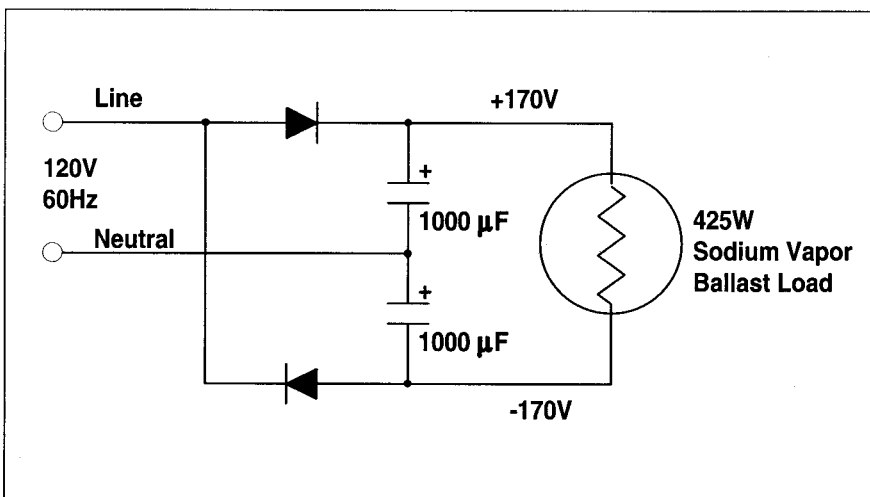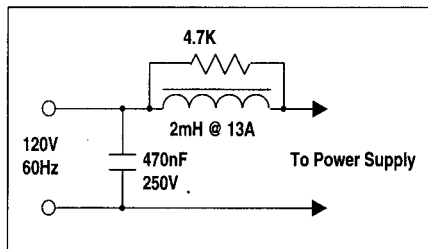*Figure 28.* 60nH @15A 5.2-in. HF Core Sinewave: 300V/div., Triangle Wave: 10A/div.; Sweep: 2:msec/div.

Figure 29a. Sodiium Vapor Lamp. 425W Ballast Load.

Power Supply Magnetics

Figure 29b. *Sodiium Vapor Lamp Line Filter.*

tary internal arcs. Use of a resonant circuit allows the arcs to extinguish before severe internal damage takes place.

Power Line Filters

Hi-FLUX™ cores are the best choice for EMI/RFI filters and power factor correction in line operated rectifier supplies. *Figure 29a* shows the DC power supply used in a 400W electronic sodium vapor ballast. The switching amplifier in the ballast causes spikes of several volts at 20kHz on both supply polarities [3]. Without a filter in the line high side, these spikes pass onto the power line during rectifier conduction and cause radio frequency interference. To be effective at high frequencies, the line filter must not saturate during the large 60Hz rectifier pulses. Because Hi-FLUX™ cores have the highest saturation level of all powdered types, they allow use of the largest inductance for a given core size.

With the shortest possible wiring run to the line circuit breaker and no choke, peak rectifier pulses can reach 23A. *Figure 30* shows these pulses and the line distortion they produce. The di/dt is about 50,000A/sec for the rise of the rectifier pulses and 30,000A/sec for their fall. Contrary to popular belief, line distortion is primarily caused by leakage inductance of the utility company power transformer but not its resistance. The line voltage drop in this resistance from the rectifier pulses is only about 280mV. This is invisible on an oscilloscope without special techniques.

Leakage inductance of a utility transformer is on the order of 60μH. When the rectifier pulse starts, the rapid current increase through that inductance lowers the line voltage near its peak by 3.0V. At the end of the pulse, the slower current decrease raises the line voltage just after the peak by 1.8V. These voltage disturbances are easily visible in *Figure 30* and pollute the line for other users. Personal computer power supplies commonly cause this problem.

A line filter using a 2mH HF choke is shown in *Figure 29b*. *Figure 31* shows that pulse currents are reduced to 18A and that all visible line distortion is eliminated. Rms line current is reduced from 6A to 5A. The choke is a 1.84-in. diameter Hi-FLUX™ core of 125 permeability, which costs 1W power loss. The choke operates slightly beyond saturation at the rectifier pulse peaks but still retains about 600μH inductance there. The 470nF capacitor is considerably larger than usual to reduce interference to WWVB and Loran C

receivers at 60 and 100 kHz, respectively. At 100kHz, this filter attenuates line interference by more than 40 db. At 1MHz, its attenuation is greater than 60 db.

Figure 30 shows that the current pulses without a choke reach a maximum of 23A about 15 degrees before the line voltage peak. With the choke, the current peak of 18A is reached about 15 degrees after the peak, as shown in *Figure 31*. If the choke is reduced to about 240μH, the current and voltage peaks will coincide. However, the distortion on the line voltage by the current pulses is still readily apparent and the rms line current is nearly as great as for no choke. The 4.7kΩ resistor across the choke is used to damp its ringing when the rectifier pulses end. The high Q self resonance of the choke occurs at approximately 200kHz due to its distributed winding capacitance. Without the resistor, bursts of narrow band RFI appear at that frequency with a 120Hz repetition rate.

Figure 32 shows the voltage across the choke and the current through it. Just before the positive peak of the line, the voltage across the choke rapidly rises to about +18V and then falls to zero. Here the choke is holding off a sudden increase of line current. During this time the current increases to 19A. The voltage now starts to reverse across the choke and the current starts to fall to zero. Here the line voltage has dropped below the voltage on the positive filter capacitor and the choke is returning energy to the capacitor.
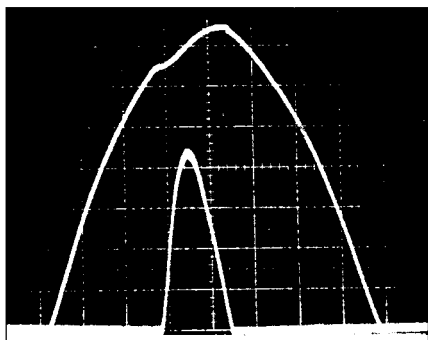
Figure 30. *Line Distortion and Rectifier Pulse Line (170V Peak): 20V/div., Pulse: 5A/div.; Sweep: 1msec/div.*
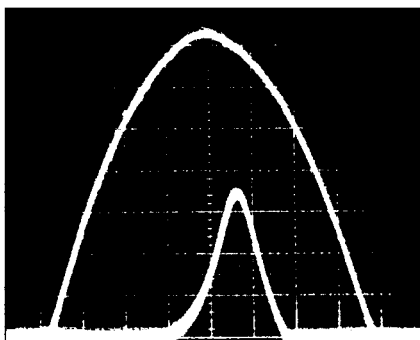
Figure 31. *Undistorted Line and Rectifier Pulse Line (170V Peak): 20V/div., Pulse: 5A/div.; Sweep: 1msec/div.*
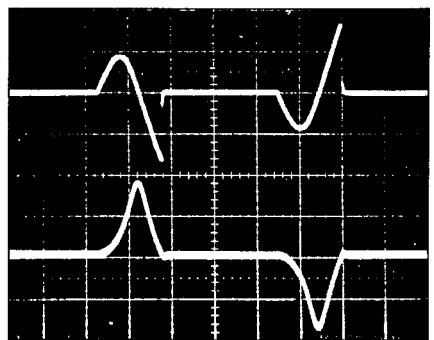
Figure 32. *Filter Choke Voltage and Current Choke Voltage: 20V/div., Choke Current: 10A/div.; Sweep: 2msec/div.*
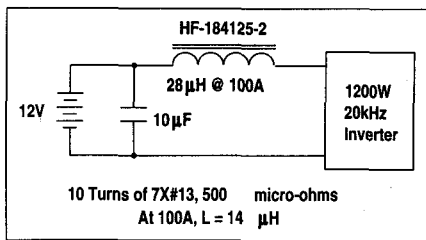
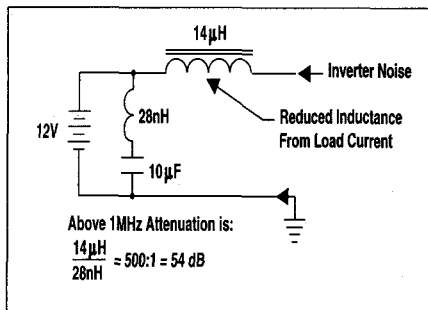**Figure 33a.** *Two-Pole Battery Line Filter, Battery Circuit.*

Figure 33b. *Two-Pole Battery Line Filter, Capacitor Inductance Model.*
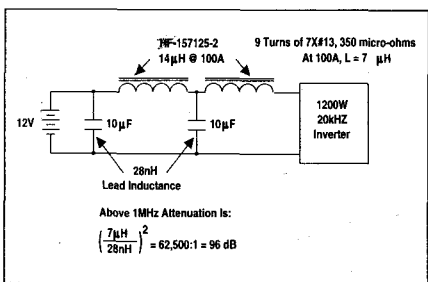

Figure 34. *Four-Pole Battery Line Filter.*

The capacitor continues to charge briefly even though the line voltage is now less than the capacitor voltage. This prevents a sudden decrease in line current. The choke voltage reaches a maximum negative voltage of about -32V at the moment the current reaches zero. The maximum di/dt during the rise of the current pulses is 9,000A/sec and the maximum di/dt during their fall is 16,000A/sec. This is a small fraction of their maximum rates without a choke.

An additional benefit of a large line choke is the reduction of the initial turn on current surge. With no line choke, it can be as high as 400A for 500μsec. This imposes severe stresses on rectifier diodes and will weld ordinary switch contacts together in short order. Negative temperature coefficient surge resistors can be used to limit this surge. After the initial surge, these resistors heat up and greatly decrease in value. Their remaining resistance causes substantial power losses after turn on. After loss of power, they require several seconds to recover their high resistance. For brief power interruptions, they provide no protection upon the reapplication of power.

Without the HF choke, the turn on surge was 400A and the di/dt was 2,800A/msec. With the choke, the surge was reduced to 250A and the di/dt was reduced to 85A/msec. The choke will be deeply saturated during the later part of the surge, but by then the switch contacts will have had time to close. This prevents most switch damage and reduces rectifier stress. In many cases, an HF choke will eliminate the need for a surge resistor.

Line Filter Precautions

A detailed treatment of RFI filter design is beyond the scope of this article, but a word of caution is in order. It is easy to build a good low pass line filter and then contaminate its input with switching spikes. The capacitor and inductor must not be close to rapidly changing voltages or currents. An RFI filter able to correct line power factor is considerably larger than common line filters. It should be in a separate shielded compartment where the line cord enters the chassis.

It is common to find a dv/dt of 3,000V/μsec or more in switching power supplies. A 10pF stray capacitance to such a circuit will have a 30mA displacement current. A 340V transition can then couple a 7mV spike into a 470nF line capacitor. The induced spikes are about the same size as the residual noise passing through the line filter. Shielding will prevent this.

Transformer primaries in switching supplies can have a di/dt of 100A/μsec or more. If one of the line conductors with a loop area of 1cm^2 is separated by 5cm from such a primary wire, every current transition will induce a 40mV spike. This is easily an order of magnitude greater than the rms voltages passing through the filter.

External magnetic fields can be dramatically reduced by pairing high di/dt leads with their return lines and twisting them together where possible. RFI will be minimized if circuit dv/dt and di/dt are designed to be as low as possible.

Battery Line Filters

Hi-FLUX™ cores are ideally suited for battery filter chokes in high power low voltage inverters. Such a choke prevents switching spikes from leaving the inverter through the battery input leads. *Figure 33a* shows a 1.84-in. 125 permeability core used as a filter for a 12V 100A inverter operating at 20kHz. A 28μH inductor rated at 100A was wound with 10 turns of seven twisted strands of #13 copper wire. The DC loss is about 5W at 100A.

A 10μF polypropylene capacitor across the battery terminals completes the filter. The load current reduces the effective inductance of the choke to about 14μH. The filter produces about 19 db attenuation at 40kHz and 52 db attenuation at 200kHz. Self inductance of about 28nH in the 10μF capacitor causes the attenuation to level off at 54 db above 300kHz *(Figure 33b)*.

Low impedance filters pose special challenges because of capacitor self resonance. A four-pole filter *(Figure 34)* overcomes this problem. By using two cores each of half the size and inductance, the total core cost is about the same. A second capacitor costs very little compared to the inductors. The effective inductance of each core is reduced to 7μH because of load current. For very little increase in cost, attenuation is now 26 db at 40kHz and 92 db at 200kHz. Well above 300kHz, attenuation approaches 96 db. Total DC losses are about 7W.
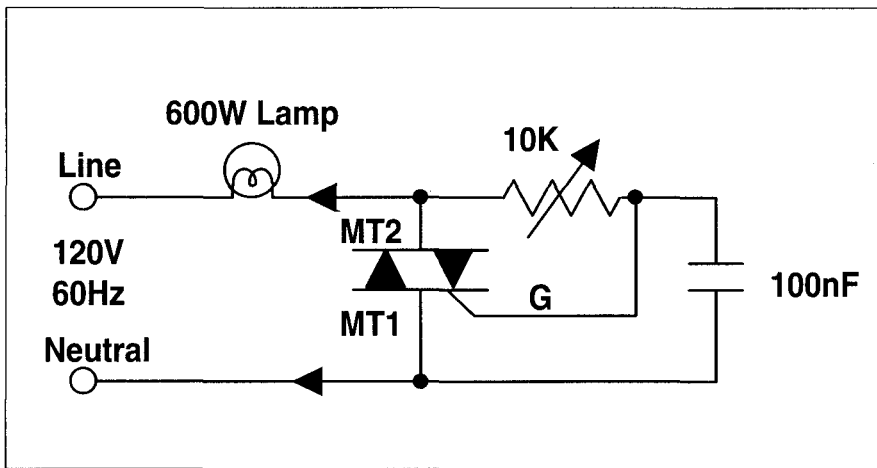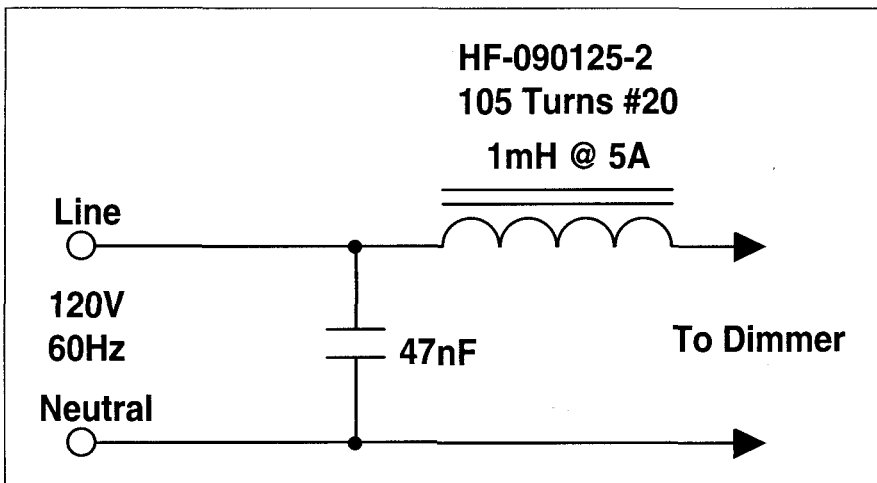
Figure 35a. *Triac Dimmer, Dimmer Circuit*

Figure 35b. *Triac Dimmer, Line Filter*

Light Dimmer Filters

Triac lamp dimmers are infamous for causing AM radio interference. The circuit for a typical 600W incandescent lamp dimmer is shown in *Figure 35a*. These dimmers control the average power supplied to the lamps in the load by varying the duty cycle of the triac switch. *Figure 36* shows the voltage across the switch for a 50% duty cycle.

At turn on, a voltage transient as great as 170V occurs across the terminals of the dimmer. The dv/dt of this transient is about 1,700V/μsec. The current rise time is limited only by the leakage inductance of the utility transformer. With leakage inductance of 60μH, the di/dt is 2.8A/usec. The high dv/dt and di/dt both radiate RFI from the lamp wiring.

Most commercial dimmers have a small inductor of about 60μH in series with their terminals but this is far too small to be effective against RFI. Usually, they have no shunt capacitor. *Figure 35b* shows an effective RFI filter for a 600W dimmer. An 0.90-in. diameter 125 permeability HF core is used as a line choke and is rated at 1.0mH at 5A. The dimmer terminals are shunted with a 47nF capacitor.

The capacitor reduces the dv/dt across the dimmer terminals at turn on to 2V/μsec. Overall voltage rise time is increased from 0.1μsec to 50μsec. The choke reduces the di/dt during turn on to 0.17A/μsec. Due to the reduced di/dt, the triac can completely turn on before full current is applied. This increases its reliability during the high surge currents from cold lamp filaments. The choke saturates at the 7A peak of the cycle, but the di/dt of the line current has been adequately controlled before this happens. The choke has a 2W loss at the dimmer's rated current.

Smoothing Chokes

MPP is the best choice for smoothing chokes in switching power supplies where the ripple current is small compared to DC current. Where the ripple current is large, as in high power flyback type supplies, MPP losses may be too large. In that case, gapped ferrites with Litz wire are better because of their lower hysteresis loss. For lower power flyback supplies, MPP cores are better than gapped ferrites.

Figure 37a shows a forward converter 5V, 30A supply operating at 20kHz. The waveforms shown in *Figure 37b* assume that the input voltage could drop by approximately 50% before regulation would be lost. The filter choke inductance will drop about 50% due to the 30A DC. This results in about 5A peak-to-peak triangle current in the output filter capacitor (1.4Arms). This relatively low ripple current reduces heating in the output filter capacitor. The capacitor should be selected for lowest possible ESR to minimize output ripple.

For small load currents, the inductance returns to 22μH, which reduces the ripple current to 2.5A peak-to-peak. The lower ripple allows the load current to go down to about 1.25A before regulation is lost. An MPP design for the filter choke is compared to a ferrite design. The cores used were a 1.06-in. diameter
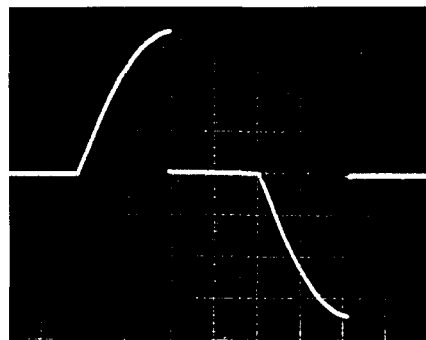

Figure 36. *Triac Voltage at 50% Duty Cycle Voltage: 50V/div.; Sweep: 2 msec/div.*

MPP core with a permeability of 60 and a EC41/41G gapped ferrite core with a permeability of 74. The MPP core is rated at 22μH at 31A and the ferrite is rated at 21μH at 26A.

The ferrite has a gap of 60 mil and a cross section of 1.21 cm². Its actual core constant is 125mH/kturn squared. If all the field of the ferrite was within the gap, its core constant would be 200mH/kturn squared. The ratio of the actual to the theoretical value is the portion of the total flux that is actually within the gap. This turns out to be only 63%. The remaining 37% constitutes the fringing field within the winding area.

To prevent excessive losses from the fringing field, the ferrite must be wound with Litz wire. Reducing the gap will reduce the fringing field but will also reduce the energy storage capability. The losses in gapped ferrites wound with solid wire smaller than #22 are usually acceptable. In some cases, thin foil windings have reduced losses in gapped cores to acceptable levels. The ferrite core is wound with 13 turns of #13 litz.

The MPP core is wound with 17 turns of #12. If available, rectangular #12 could be used on the MPP core for greater winding efficiency. The superiority of the MPP core is evident from its 31A rating compared to 26A for the ferrite. The DC resistance of the MPP is 4mΩ and the ferrite is 5mΩ. The DC loss of the cores at 26A is 2.7W in the MPP vs 3.4W in the ferrite.

In summary, the MPP core is capable of 19% more current, has 20% less DC loss at the same current, and weighs 31% less. Actual circuit losses in the MPP and ferrite are about 50% greater due to hysteresis and fringing field, respectively.

Low-Level Resonant Circuits

MPP cores are the best choice for inductors in fixed tuned resonant circuits operating at low level up to 200kHz. Ungapped ferrites might appear to be better because they need fewer turns. However, their inductance changes considerably with temperature. Over the 0 to 60°C range,
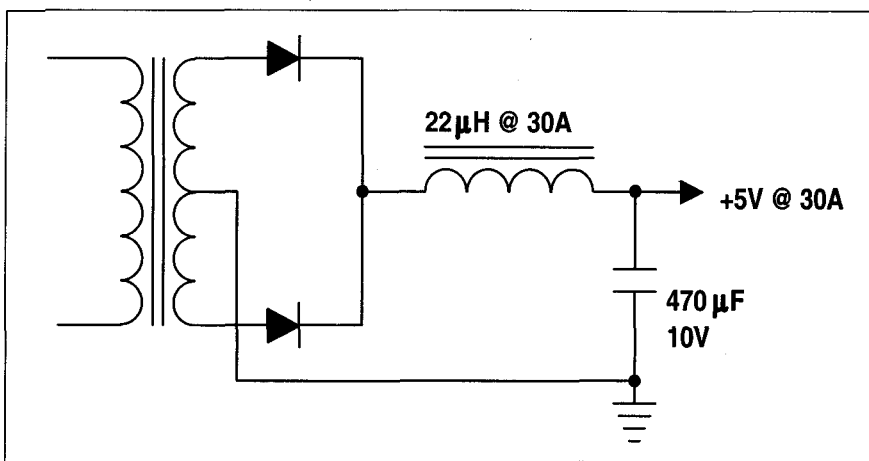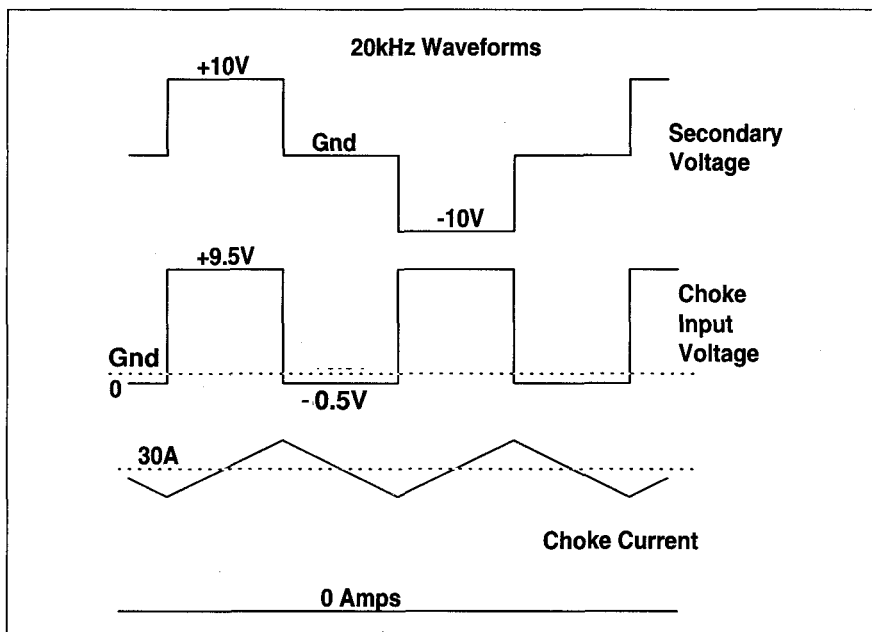

Figure 37a. Forward Converter, Circuit.
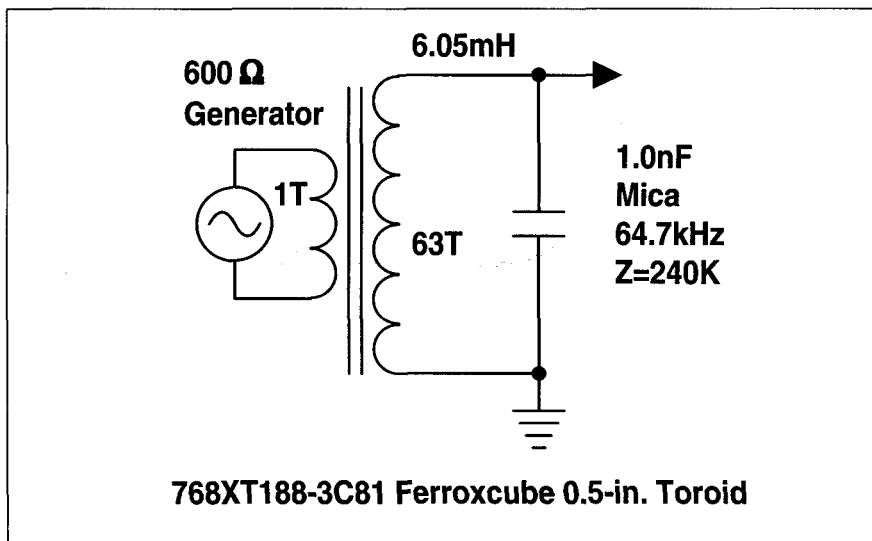

Figure 37b. Forward Converter, Waveforms.

Figure 38a. Ferrite Tuned Circuit.

Figure 39a. *MPP Tuned Circuit.*

600 Ω Generator

4T

164T

1.5mH

4.7nF
Mica
60.7kHz
Z=22K
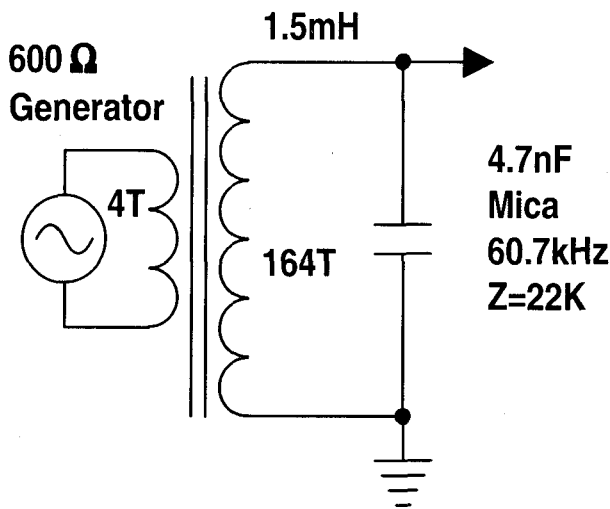
A-050056-2 Arnold MPP 0.5-in. Torid

the permeability of 3C81 ferrite will change by 25% [4]. With increased drive level, Q drops dramatically and permeability increases somewhat.

The gain of tuned amplifier stages using ungapped ferrite transformers for collector or drain loads will decrease with increased signal level. This results in nonlinearity and intermodulation distortion. *Figure 38a* shows a 60kHz tuned circuit

using 3C81 ferrite. From 0 to 115 gauss, permeability increases by about 9%. This causes the circuit to detune as signal level increases. Over the same range, Q decreases by 73% *(Figure 38b)*.

Figure 39a shows a 60 kHz tuned circuit using 125 permeability MPP. From 0 to 138 gauss, permeability increases by only 1% and Q goes down by only 19% *(Figure 39b)*. As

the drive level increases, MPP cores minimize detuning and nonlinearity effects. The permeability of MPP cores is more stable with temperature, typically within ±1% at the room temperature value from 0 to 60°C

Most of the problems associated with ungapped ferrites in tuning applications can be eliminated by using pot cores with a gap. Compared to MPP, they are more expensive and fragile. Both have comparable Q and require essentially the same number of turns. For low frequency fixed tuned applications MPP cores are superior.

Conclusions

Tables 1 and 2 summarize the properties of the various core types covered in these articles. The losses of MPP, SUPER-MSS™ and HI-FLUX™ cores are dramatically lower than powdered iron. Best performance in switching power supply filter chokes is achieved when MPP cores are used in place of powdered iron.

However, SUPER-MSS™ cores can be substituted for MPP cores in many high power applications with nearly the same level of performance

Output Level (±)	Flux Gauss (±)	Q	ΔQ	Δf
40mV	1.3	73	0	0
200mV	6.6	69	-6%	-0.1%
500mV	16	47	-42%	-4.6%
1.0V	33	35	-58%	-1.1%
2.0V	66	30	-66%	-2.6%
3.5V	115	24	-73%	-4.0%

Figure 38b. *Ferrite Performance.*

Output Level (±)	Flux Gauss (±)	Q	ΔQ	Δf
100mV	1.4	37	0	0
200mV	2.8	35	-5.4%	-0.02%
400mV	5.5	33	-11%	-0.03%
2.0V	28	32	-13%	-0.19%
4.0V	55	31	-16%	-0.31%
10V	138	30	-19%	-0.52%

Figure 39b. *MPP Performance.*

Core Type	Current Rating (A)	Output Voltage (V)	Output Power (kW)	Power Loss (W)	Power Loss (%)	Q@ 40kHz	ESR (Ω)	1K Price ($)
Mollypermalloy	6	320	1.92	11	0.6	74	14	10.97
MSS™	6	320	1.92	16	0.8	34	30	3.02
High Flux™	15	800	12.0	157	1.3	19	56	10.68
Powered Irom	13	693	9.0	131	1.5	16	64	.64
Gapped Ferrite	3.5	187	0.65	17	2.6	8	119	12.00
Ferrite/Litz	3.5	187	0.65	8	1.2	100	10	13.00

Table 1. All Cores Are 1.84-in. Diameter Toroids, Except For the Ferrite that Is a Gapped EC70/70G. Inductors Are 4.0mH. Ripple Current Is a 40kHz Triangle Wave With a Peak-to-Peak Level of 33% of Rated Current. Output Power Is Theoretical and Ignores High Voltage and Core Dissipation Problems.

Core Type	Flyback Circuits	Line Filter	High Energy Circuit	Saturation Frequency Limit (kHz)	Filter Chokes	Precision Filters	Price
Molypermalloy	Good	Poor	Poor	3.5	Best	Good	High
MSS™	Fair	Good	Fair	1.6	Good	Fair	Low
High-Flux™	Poor	Best	Best	1.0	Fair	Poor	Medium
Powdered Irom	Worst	Fair	Good	0.8	Worst	Worst	Lowest
Ferrite/Litz	Best	Worst	Worst	20.0	Poor	Best	Highest

Table 2. Ranking of Cores Is the Author's Opinion of the Typical Suitability of the Cores for the Given Application. Generally, Smaller Cores May Be Operated at Higher Frequencies and Flux Levels and Vice Versa for the Larger Cores.

and lower cost. Where cores operate with high flux levels at a few kilohertz, MSS and MPP cores are the quietest and powdered iron and gapped ferrites are the loudest. HF cores can make an irritating buzz when filtering a low frequency square wave.

The most dramatic improvement in performance as a function of cost is achieved when MSS cores are used to replace powdered iron cores. HF cores are best for high energy resonant circuits below 1 kHz. They also make the best RFI chokes that must pass DC or line voltage. Powdered iron cores are best only where cost is extremely critical and losses are unimportant.

Iron laminated or tape cores are the best choice for transformers operating below 1 kHz. Ferrites are preferred for transformers from 1 kHz to 1 MHz. MPP cores and gapped ferrites are both suited to low level filter and fixed tuned applications below 100 kHz. If variable tuning is needed, a gapped ferrite pot core with an adjustable center post is best.

The increase in operating frequencies and densities of switching power supplies has resulted in an urgent need for improved core materials. Resonant power supplies operating at 100 kHz or more place severe demands on the Q of high power resonant inductors. That demand is not well served by gapped ferrites because of the losses in their fringing field. New ungapped low permeability materials are needed that can operate at high flux densities in that frequency range. They should be usable above 100kHz with a Q over 200.

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4. © Ferrite Materials and Components Catalog, Eighth Edition,[a] 1994 Philips Components, Saugerties, NY, pp 1-12, 3-3, 4-24 through 4-33.

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