Use of Ferrites in Broadband Transformers

Introduction

Most of the magnetic information in this catalog is data obtained from cores wound with a single multi-turn-winding which forms an inductor. When a second winding is added on the core, the inductor becomes a transformer. Depending on the requirements, transformers can be designed to provide dc isolation, impedance matching and specific current or voltage ratios. Transformer designed for power, broadband, pulse, or impedance matching can often be used over a broad frequency spectrum.

In many transformer designs ferrites are used as the core material. This article will address the properties of the ferrite materials and core geometries which are of concern in the design of low power broadband transformers.

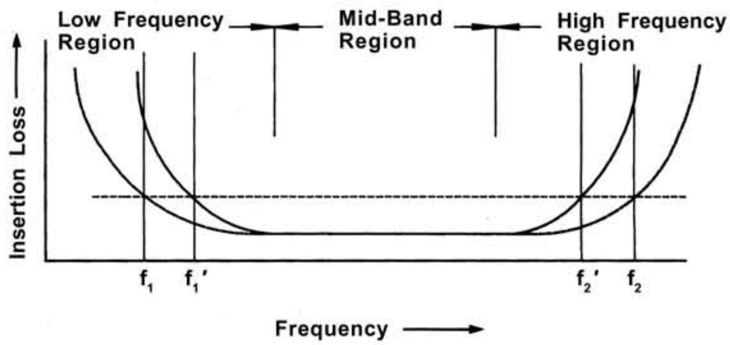

Figure 1 Typical Characteristic Curve of Insertion Loss vs. Frequency for a broadband transformer.

Brief Theory

Broadband transformers are wound magnetic devices that are designed to transfer energy over a wide frequency range. Most applications for broadband transformers are in telecommunication equipment where they are extensively used at a low power levels.

Figure 1 shows a typical performance curve of insertion loss as a function of frequency for a broadband transformer. The bandwidth of a broadband transformer is the frequency difference between f_2 and f_1, or between f_2' and f_1', and is a function of the specified insertion loss and the transformer roll-off characteristics.

It can be seen that the bandwidth is narrower for transformers with a steep roll-off (f_2'- f_1') than those with a more gradual roll-off (f_2- f_1). Also in Figure 1, the three frequency regions are identified.

The cutoff frequencies are determined by the requirements of the individual broadband transformer design. Therefore, f_1 can be greater than 10 MHz or less than 300 Hz. Bandwidths also can vary from a few hundred hertz to hundreds of MHz. A typical broadband transformer design will specify for the mid frequency range a maximum insertion loss and for the cutoff frequencies, f_1 and f_2 maximum allowable losses. Figure 2 is a schematic diagram of the lumped element equivalent circuit of a transformer, separating the circuit into an ideal transformer, its components and equivalent parasitic resistances and reactances. The secondary components, parasitics and the load resistance have been transferred to the primary side and are identified with a prime.

To simplify this circuit, the primary and secondary circuit elements have been combined and the equivalent reduced circuit is a shown in Figure 3. The physical significance of the parameters are listed below the equivalent circuits. In the low frequency region the roll-off in transmission characteristics is due a lowering of the shunt impedance. The shunt impedance decreases when the frequency is reduced, which results in the increases level of attenuation. The impedance is mainly a function of the

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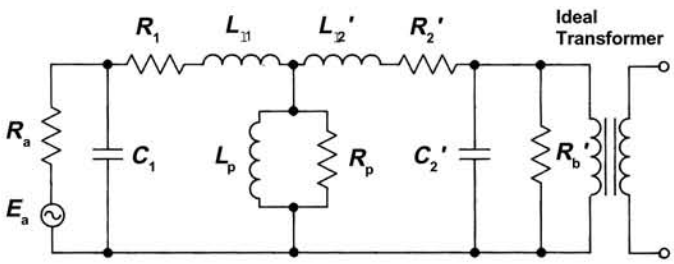

Figure 2 Lumped equivalent of a transformer.

E_a = source EMF
R_a = source resistance
C_1 = primary winding capacitance
R_1 = resistance of primary winding
$L_{\ell 1}$ = primary leakage inductance
L_p = open circuit inductance of primary winding
R_p = shunt resistance that represents loss in core

Secondary parameters reflected to the primary side.

C_2' = secondary winding capacitance
R_2' = resistance of secondary winding
$L_{\ell 2}'$ = secondary leakage inductance
R_b' = load resistance

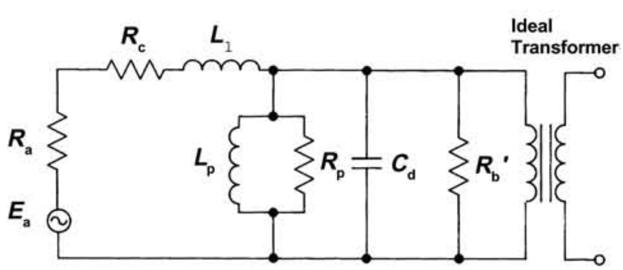

Figure 3 Simplified equivalent transformer circuit

$C_d = C_1 + C_2'$
$R_c = R_1 + R_2'$
$L_\ell = L_{\ell 1} + L_{\ell 2}'$

For other circuit parameters see Figure 2.

primary reactance X_{LP} with a negligible contribution of the equivalent shunt loss resistance R_p. The insertion loss may therefore be expressed in terms of the shunt inductance:

$$A_i = 10 \log_{10} \left(1 + \left(\frac{R}{\omega L_p}\right)^2\right) \text{ dB}$$

Where $R = R_a \times R_b' / R_a = R_b'$

For most ferrite broadband transformer designs, the only elements that are likely to effect the transmission at the mid-band frequency range are the winding resistances. The insertion loss for the mid-band frequency region due to the winding resistance may be expressed as:

$$A_i = 20 \log_{10} \left(1 + \frac{R_c}{R_a + R_b'}\right) \text{ dB}$$

Where $R_c = R_1 + R_2'$

In the higher frequency region the transmission characteristics are mainly a function of the leakage inductance or the shunt capacitance. It is often necessary to consider the effect of both of these reactances, depending upon the circuit impedance. In a low impedance circuit the high frequency droop due to leakage inductance is:

$$A_i = 10 \log_{10} \left(1 + \left(\frac{\omega L_\ell}{R_a + R_b'}\right)^2\right) \quad \text{dB}$$

This high frequency droop in a high impedance circuit, due to the shunt capacitance, is as follows:

$$A_i = 10 \log_{10} \left(1 + \left(\omega CR\right)^2\right) \quad \text{dB}$$

Reviewing the insertion loss characteristics for the three frequency regions, it can be concluded that the selection of ferrite material and core shape should result in a transformer design that yields the highest inductance per turn at the low frequency cutoff f_1. This will result in the required shunt inductance for the low frequency region with the least number of turns. The low number of turns are desirable for low insertion loss at the mid-band region and also for low winding parasitics needed for good response at the high frequency cutoff f_2.

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Low and Medium Frequency Broadband Transformers

For broadband transformer applications the optimum ferrite is the material that has the highest initial permeability at the lower cutoff frequency f_1. Manganese zinc ferrites, such as Fair-Rite 77 or 78 material, are very suitable for low and medium frequency broadband transformers designs. As stated before, the transformer parameter that is most critical is the shunt reactance (ωL), which will increase with frequency as long as the material permeability is constant or diminishing at a rate less than the increase in frequency. This holds true even if a transformer is designed using a manganese zinc ferrite where f_1 is at the higher end of the flat portion of the permeability vs. frequency curve. Although the whole bandpass lies in the area where the initial permeability is decreasing, yet the bandpass characteristics will be virtually unaffected. For broadband transformers that use a manganese zinc ferrite material the core geometry should be such as to minimize the R_{dc}/L ratio. In other words, the ratio of dc resistance to the inductance for a single turn should be a minimum. The range of pot cores, standardized by the International Electrotechnical Commission in document IEC 60133, has been designed for this minimum R_{dc}/L ratio. Other core shapes can also be used in the design of these broadband transformers. Often the final core selection will also be influenced by such considerations as ease of winding, terminating and other mechanical design constraints of the transformer.

Broadband Transformers with a Superimposed Static Field

In transformer designs that have a superimposed direct current, gapped cores can be employed to overcome the decrease in the shunt inductance. Hanna curves can be used to aid in the design of inductive devices that carry a direct current. For more information see section "The Effect of Direct Current on the Inductance of a Ferrite Core".

High Frequency Broadband Transformers.

Although there is no clear division between the frequency regions, for this article it is assumed that the high frequency broadband transformer designs use nickel zinc ferrites as the preferred core material. This will typically occur for transformer designs where the bandpass lies wholly above 500 kHz. At these higher operating frequencies it becomes more important to consider the complex magnetic parameters of the core material, rather than use the simple core constants, such as A_L, recommended for low frequency designs.

Another important consideration is that high frequency transformers are generally used in low impedance circuits, which means that these designs require low shunt impedances. This can often be accomplished with a few turns, hence winding resistances are no longer an issue, and the design concept of minimizing R_{dc}/L is no longer required. The design will instead become focused on core shape and material for the required shunt impedance at f_1 along with reducing leakage inductance of the winding. Since the material characteristics permeability and losses affect the shunt impedance these parameters need to be considered in high frequency broadband transformer designs. Figures 4, 5 and 6 are typical curves of impedance Z, equivalent parallel reactance X_p and equivalent parallel loss resistance R_p as a function of frequency. They are measured on the same multi-aperture core 28—002302, in 73, 43, 61 & 67 material, wound with a single turn through both holes. For high frequency broadband transformers the toroidal core shape becomes an attractive core geometry. The few turns that are often required can easily be wound on the toroid. However, windings that require only a few turns may give rise to problems in obtaining the desired impedance ratios. To minimize leakage inductance it is suggested that the primary and secondary windings be tightly coupled and where possible a bifilar winding be used.

An improvement in core performance over toroids can be obtained by the use of multi-aperture cores, which can be considered as two toroidal cores side by side. This core shape has a lower single turn winding length than the equivalent toroidal core with the same core constant C_1, and will result in a wider bandwidth of the transformer design. Many broadband transformers have been designed utilizing nickel zinc ferrite toroids with good results. If bandwidth requirements cannot be met using toroids, multi-aperture nickel zinc cores should be considered.

The multi-aperture cores are available in the nickel zinc ferrite materials 67, 61 and 43 as well as the manganese zinc ferrite 73 material.

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Summary

The low cutoff frequency f_1 is the single most important factor in the ferrite material selection. The material with the highest initial permeability at f_1 is the recommended choice.

Manganese zinc ferrites, 77 and 78, can be used to a cutoff frequency f_1 of 500 kHz. Above this frequency use a nickel zinc ferrite, again depending upon the frequency f_1, select 43, 61 or 67 material.

For low and medium frequency transformers the optimum core shape should provide the lowest DC resistance per unit of inductance. If there is a superimposed dc present the use of gapped cores and Hanna curves is suggested. For high frequency designs, use nickel zinc ferrite. The toroidal and multi-aperture cores are the recommended core configurations.

The number of turns should be kept to a minimum to reduce leakage inductance and self-capacitance of the windings. Wind primary and secondary windings tightly coupled or as bifilar windings to lower leakage inductance.

The "Multi-Aperature Core Kit", (part number 0199000036), contains a variety of components suited for broadband transformer design evaluations.

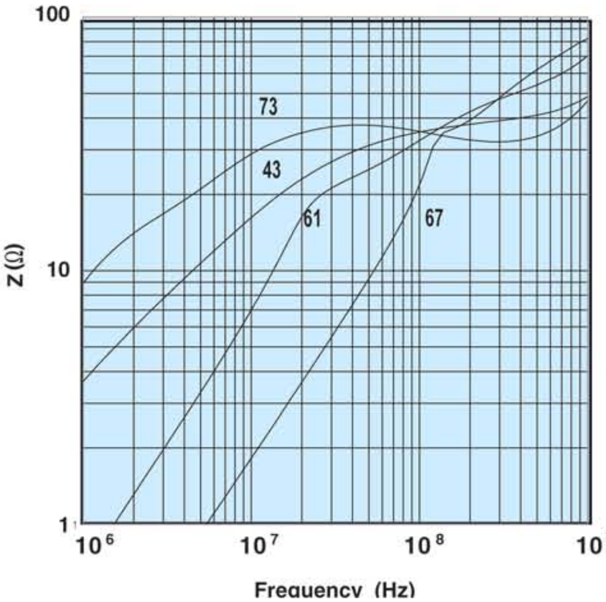

Figure 4 Impedance vs. frequency for part number 28—002302 in 73, 43, 61 & 67 material.

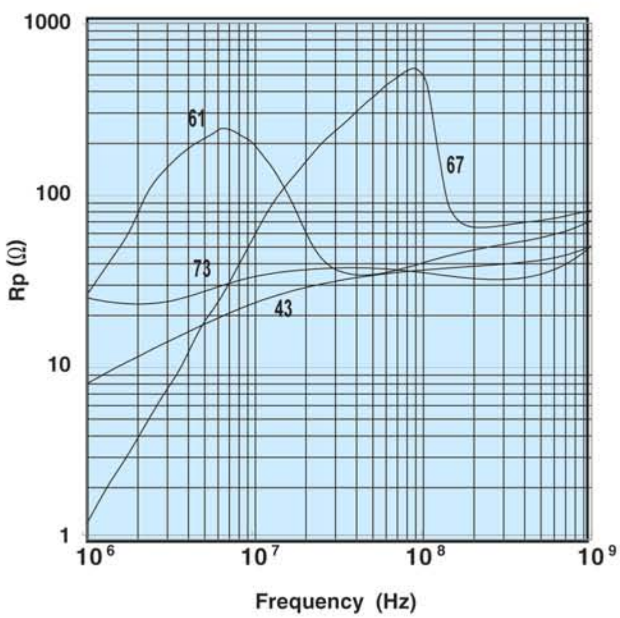

Figure 5 Parallel resistance vs. frequency for part number 28—002302 in 73, 43, 61 & 67 material.

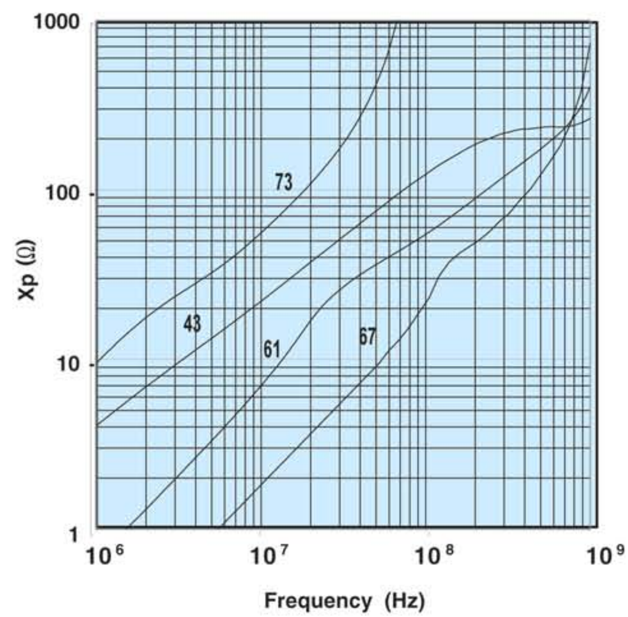

Figure 6 Parallel reactance vs. frequency for part number 28—002302 in 73, 43, 61 & 67 material.

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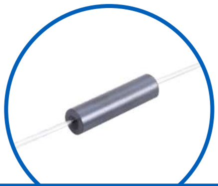
SUPPRESSION COMPONENTS OVERVIEW

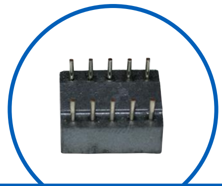
EMI SUPPRESSION BEADS

Fair-Rite offers a broad selection of ferrite EMI suppression beads with guaranteed minimum impedance specifications.



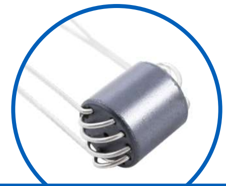
BEADS-ON-LEADS

Beads-on-leads are ferrite suppression devices supplied assembled on tinned copper for automted circut board assemby.



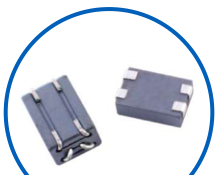
PC BEADS

Multiple single-turn or multi-turn printed circut board beads are available in 44 or 52 materials.



WOUND BEADS

Wound beads available in 6 or 11 hole configurations, both with or without windings, in 44 and 61 material.



SM BEADS (Common Mode)

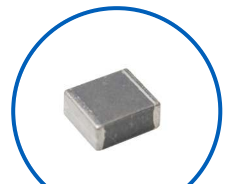
Available in both broadband and high-frequency materials and various sizes for suppression of EMI noise.

SM BEADS (Differential Mode)

The rugged construction of these SM beads lowers the DC resistance and increases current carrying capacity compared to plated beads.

MULTI-APERTURE CORES

Can be used in a variety of applications including suppression, balun (balanced – unbalanced) designs, common-mode chokes, and other broadband transformers.



CHIP BEADS

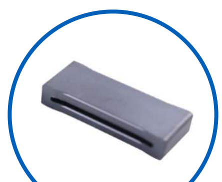
100% tested for impedance and DC resistance. They are available in standard, high and GHz signal speeds.

ROUND CABLE CORES

Offered in several different materials, these mitigate both differential and common-mode conducted EMI from 200 kHz up through 1 GHz.

ROUND CABLE SNAP-ITS

Available in several materials to suppress differential or common-mode conducted EMI from 200 kHz up through 1 GHz.



FLAT CABLE CORES

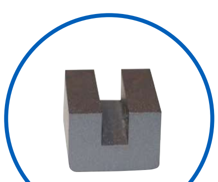
Can accommodate multi-conductor flat cables in widths from 8.9mm (0.35") up to 77mm (3.02").

FLAT CABLE SNAP-ITS

Can accommodate multi-conductor flat cables in up to 64mm (2.52") wide. These are available in two materials to reduce broadband conducted EMI from one to hundreds of MHz.

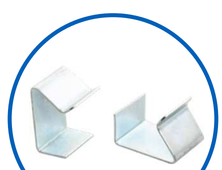
CONNECTOR PLATES

All connector plates are supplied in our NiZn 44 material grade ideally suited for this application with its high impedance and high resistivity.



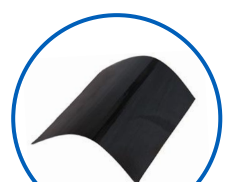
MISCELLANEOUS CORES

Fair-Rite Products offers a selection of ferrite cores in special geometries. Cores are tooled and manufactured with our 43 or 77 materials.



FLAT CABLE CORES ASSEMBLY CLIPS

Fair-Rite Products offers several securing clips to accommodate the assembly of our split flat cable suppression cores.



FLEXIBLE FERRITE

NiZn flexible ferrite available in six initial permeabilities. Useful in the mitigation of radiated emissions, sheets are available in thicknesses ranging from 0.1mm to 0.5mm .

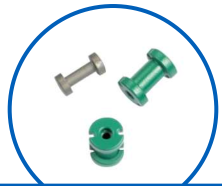
POWER & INDUCTIVE COMPONENTS OVERVIEW

RODS

Used extensively in high energy storage designs due to their high DC saturation levels. These are also used for inductive components that require temperature stability and high Q.

ANTENNA/RFID RODS

Available in three materials to cover the frequency range from 50kHz to 25MHz: 78 for less than 200kHz, 61 for 0.2 to 5MHz and 67 for greater than 5MHz



BOBBINS

Fair-Rite Products offers bobbins made in our 44 material for higher frequency designs and our 77 material for higher power designs.

TOROIDS

Fair-Rite offers this versatile geometry in various materials, from our lowest to our highest permeability.

COATED TOROIDS

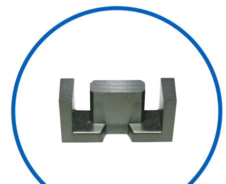
Smaller Toroids (<9.5mm) can be supplied Parylene C coated. Larger Toroids (>9.5mm) can be supplied with a uniform coating of thermo-set plastic.

POT CORES

Commonly used in high Q inductors, tuned circuits, and wide band transformers, pot cores are available in 78 and 95 materials for operating frequencies up to 200kHz.

E CORES

Offer an economical design approach for inductive applications in a variety of designs. They are widely applied in differential-mode inductors and power converter transformers.



EFD CORES

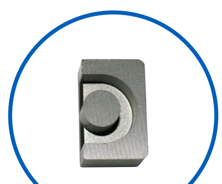
Designed to maximize volume in a low profile package providing improved heat dissipation.

ETD CORES

Makes optimal use of volume area for maximum power throughout and increased efficiency, specifically for forward power converter transformers.

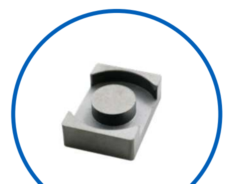
EER CORES

Widely used in switched-mode power supplies and permit off-line designs where IEC and VDE isolation requirements must be met.



EP CORES

Reduce the effect of residual air gap upon the effective permeability of the core thereby minimizing coil volume for a given inductance while providing excellent shielding.

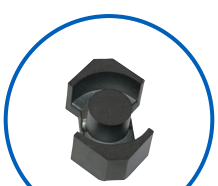


PLANAR CORES

Ideal for integrating with PCB designs since they have a smaller footprint than other power geometries. Available in our 95 and 78 materials.

PQ CORES

Developed for use in power applications, the large surface area of PQ cores aid in heat dissipation. Available in 78 and 95 materials



RM CORES

Available in 78 and 95 materials for operating frequencies up to 200kHz, typical applications for RM cores include power chokes and broad band transformers.

U CORES

Offers an economical core design with a nearly uniform cross-sectional area. Long legs can provide high voltage isolation and low leakage inductance. Available in 77 material.

For more information on these and other products visit

FAIR-RITE.COM

Transformer Design

Magnetics offers two methods to select a ferrite core for a power application.

CORE SELECTION BY POWER HANDLING CAPACITY

The Power Chart characterizes the power handling capacity of each ferrite core based upon the frequency of operation, the circuit topology, the flux level selected, and the amount of power required by the circuit. If these four specifics are known, the core can be selected from the Power Chart on page 68.

CORE SELECTION BY WaAc PRODUCT

The power handling capacity of a transformer core can also be determined by its WaAc product, where Wa is the available core window area, and Ac is the effective core cross-sectional area. Using the equation shown below, calculate the WaAc product and then use the Area Product Distribution (WaAc) Chart to select the appropriate core.

$$WaAc = \frac{P_o D_{cma}}{K_t B_{max} f}$$

WaAc = Product of window area and core area (cm^4)

P_o = Power Out (watts)

D_{cma} = Current Density (cir. mils/amp) Current density can be selected depending upon the amount of heat rise allowed. 750 cir. mils/amp is conservative; 500 cir. mils is aggressive.

B_{max} = Flux Density (gauss) selected based upon frequency of operation. Above 20 kHz, core losses increase. To operate ferrite cores at higher frequencies, it is necessary to operate the core flux levels lower than ± 2 kG. The Flux Density vs. Frequency chart shows the reduction in flux levels required to maintain 100 mW/cm^3 core losses at various frequencies, with a maximum temperature rise of 25°C for a typical power material, Magnetics P material.

A_c = Core area in cm^2	V = Voltage
f = frequency (hertz)	I_p = Primary current
K_t = Topology constant (for a space factor of 0.4)	I_s = Secondary current
	N_p = Number of turns on the primary
	N_S = Number of turns on the secondary

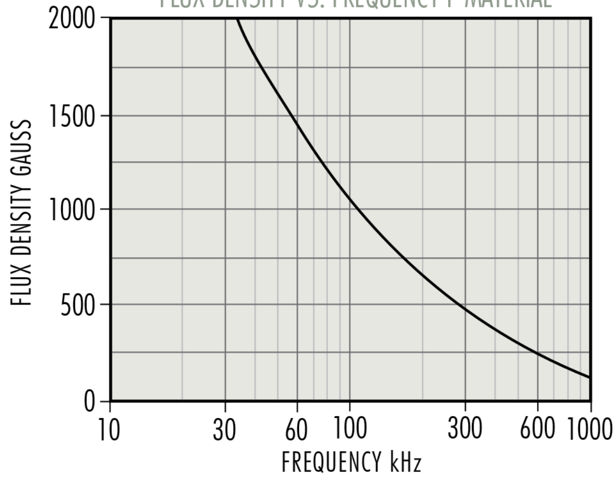
TOPOLOGY CONSTANTS K_t

Forward converter = 0.0005	Push-Pull = 0.001
Half-bridge = 0.0014	Full-bridge = 0.0014
Flyback = 0.00033 (single winding)	Flyback = 0.00025 (multiple winding)

For individual cores, WaAc is listed in this catalog under "Magnetic Data."

The WaAc formula was obtained from derivations in Chapter 7 of A. I. Pressman's book, "Switching Power Supply Design. Choice of B_{max} at various frequencies, D_{cma} and alternative transformer temperature rise calculations are also discussed in Chapter 7 of the Pressman book.

FLUX DENSITY VS. FREQUENCY P MATERIAL



Once a core is chosen, the calculation of primary and secondary turns and wire size is readily accomplished.

$$N_p = \frac{V_p \times 10^8}{4BA_c f} \qquad N_s = \frac{V_s}{V_p} N_p$$

$$I_p = \frac{P_{in}}{V_{in}} \qquad I_s = \frac{P_{out}}{V_{out}}$$

$KWa = N_p A_{wp} + N_s A_{ws}$

Where

A_{wp} = primary wire area A_{ws} = secondary wire area

Assume K = .4 for toroids; .6 for pot cores and E-U-I cores

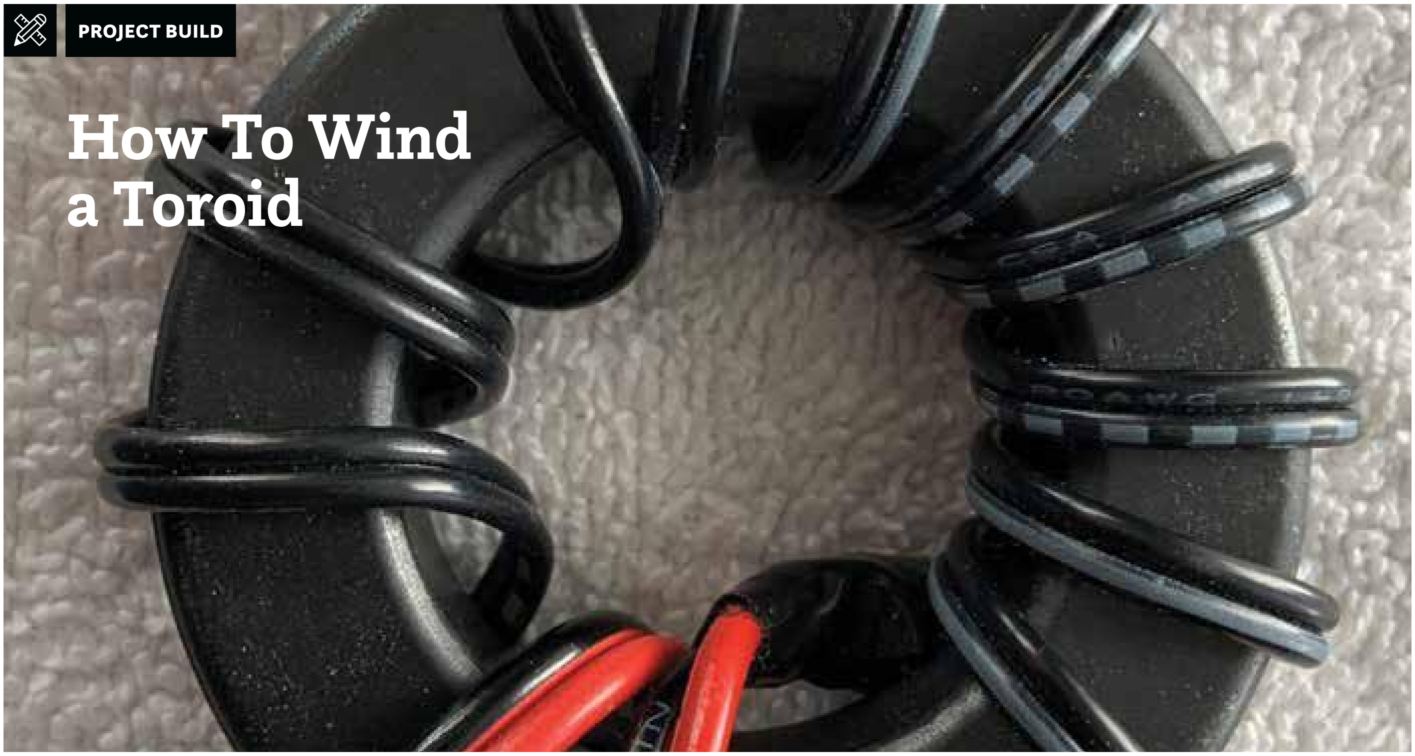
Assume $N_p A_{wp}$ = 1.1 $N_s A_{ws}$ to allow for losses and feedback winding

$$\text{efficiency } e = \frac{P_{out}}{P_{in}} = \frac{P_{out}}{P_{out} + \text{wire losses} + \text{core losses}}$$

$$\text{Voltage Regulation (\%)} = \frac{V_{no\ load} - V_{full\ load}}{V_{full\ load}} \times 100$$

PROJECT BUILD

How To Wind a Toroid



When you think of a coil of wire, better known in electronics as an *inductor*, you probably imagine something similar to the coil shown in Figure 1. Inductors store energy in magnetic fields when electric current flows through them. They are used in radio circuits to reduce or block signals at certain frequencies. When combined with capacitors, inductors can be used to create so-called *tuned circuits*.

Inductors come in many forms, but among the most common are those that use wires wound around a donut-shaped form called a *core*, which is made of powdered iron or other materials. When you wrap wires around a core, the result is a *toroidal inductor*, or what most hams simply call a *toroid*.

Toroids have certain advantages over traditional coils:

- Toroids are compact. They usually take up less room than an air-wound coil and offer the same amount of inductance.
- Toroids offer superior electrical performance. The advantage of a toroid's shape is that, due to its symmetry, the magnetic fields are confined mostly to the core. This means a toroid doesn't interfere with surrounding circuitry.

Core Types

When you're using a toroid inductor in a radio frequency (RF) circuit, not any old core will do. Cores are sold in a variety of sizes, but they are also classified according to their *type*. Loosely defined, a core's type (sometimes called a *mix*) dictates how the inductor will behave in a circuit in a given range of frequencies. The types you're most likely to encounter are 43, 52, 61, 67, 68, 75, and 76. It is important to know which type you are using, which we'll explain later.

Why You Should Care About Toroids

You'll find toroid inductors frequently in amateur radio equipment, so it helps to know a little about what they are and how they work. Some toroids are simply inductors made with a core and a single strand of wire (see Figure 2). Other toroids use multiple wires and can be somewhat complex, such as the transformer shown in Figure 3.

If you want to try your hand at building a receiver or transmitter kit, chances are the kit will require you to assemble one or more toroids. The instructions will include steps such as, "Take the yellow Type 67 core and wind 20 turns of number 18 wire."

Also, if you're running into interference problems, such as when your radio signal is getting into your stereo speakers, you can use a toroid core to create a *choke* that will block the energy from entering the speaker. Chokes like these can also be used to suppress RF energy that may be traveling down your antenna system cables and causing problems in your station.

Figure 1: When you think of a coil, an image like this probably comes to mind. You'll see coils like these in ham equipment such as power supplies, manual antenna tuners, transceivers, and more. Regardless of how they may look, the function of a coil is always the same: to oppose changes in the current flowing through it.

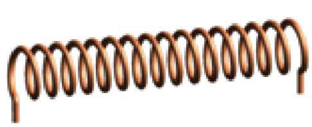


Figure 2: A simple toroid is little more than a length of wire that has been wound around a core of powdered iron or other materials. Some hams refer to cores as *ferrite donuts*.

Creating a Toroid Inductor

Some hams find toroids intimidating because they don't know how to wind them properly. Yes, toroids with multiple wires can be a bit tricky if you don't follow the instructions carefully. However, simple toroids wound with single wires or cables are extremely easy to build, and these are the ones you're most likely to use.

Step 1:

If you're building a kit, the instructions will tell you which core to use and how many turns of wire you need to apply. When winding a toroid, the first time the wire passes through the center of the core counts as the *first* turn — even though you haven't actually turned or bent the wire yet. This is a fact that confuses some hams (see ①).

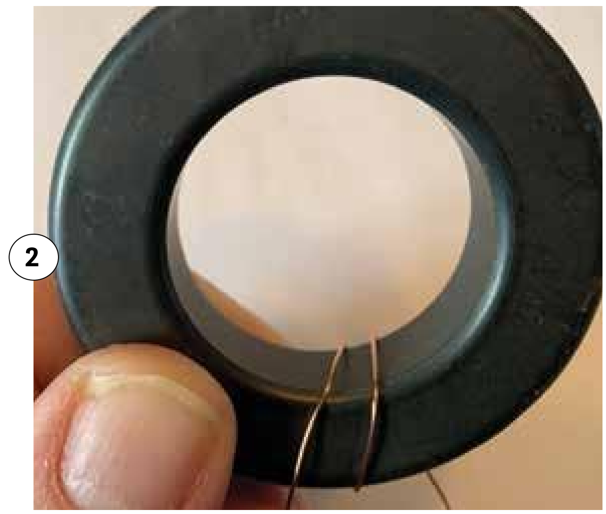
Step 2:

Once you've passed the wire through the center of the core (turn number 1, as described in Step 1), bring the wire up and over the outside edge of the core's "ring," and then send it back through the center again and pull it tight. This will be turn number *two* (see ②).

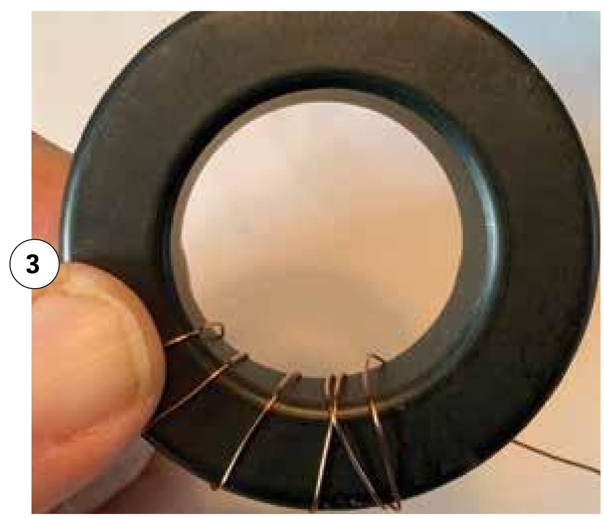
Step 3:

Repeat Step 2 over and over, spacing the turns of wire evenly around the core as you go. Each time you thread the wire through the center counts as another turn. Keep wrapping until you reach the required number of turns. Be careful not to overlap turns (see ③) and make sure they are evenly spaced. You can adjust the spacing after you wrap the last turn.

We count turns beginning with the first time the wire passes through the center of the core. This photo shows turn number 1.



Once the wire has passed through the center of the core, bring it back around, pass it over the edge of the core, and then send it back through the center once again. This will be the *second* turn.



Beware of overlapping turns when winding wire around a toroid core. This photo shows an example of what *not* to do!

Figure 3: Some toroids can be complicated, such as this toroid transformer that has several different windings. In a transformer application, one winding is often larger or smaller than the other. The fields around the windings interact, allowing the transformer to boost an ac voltage (including a signal) to a higher value, or vice versa.

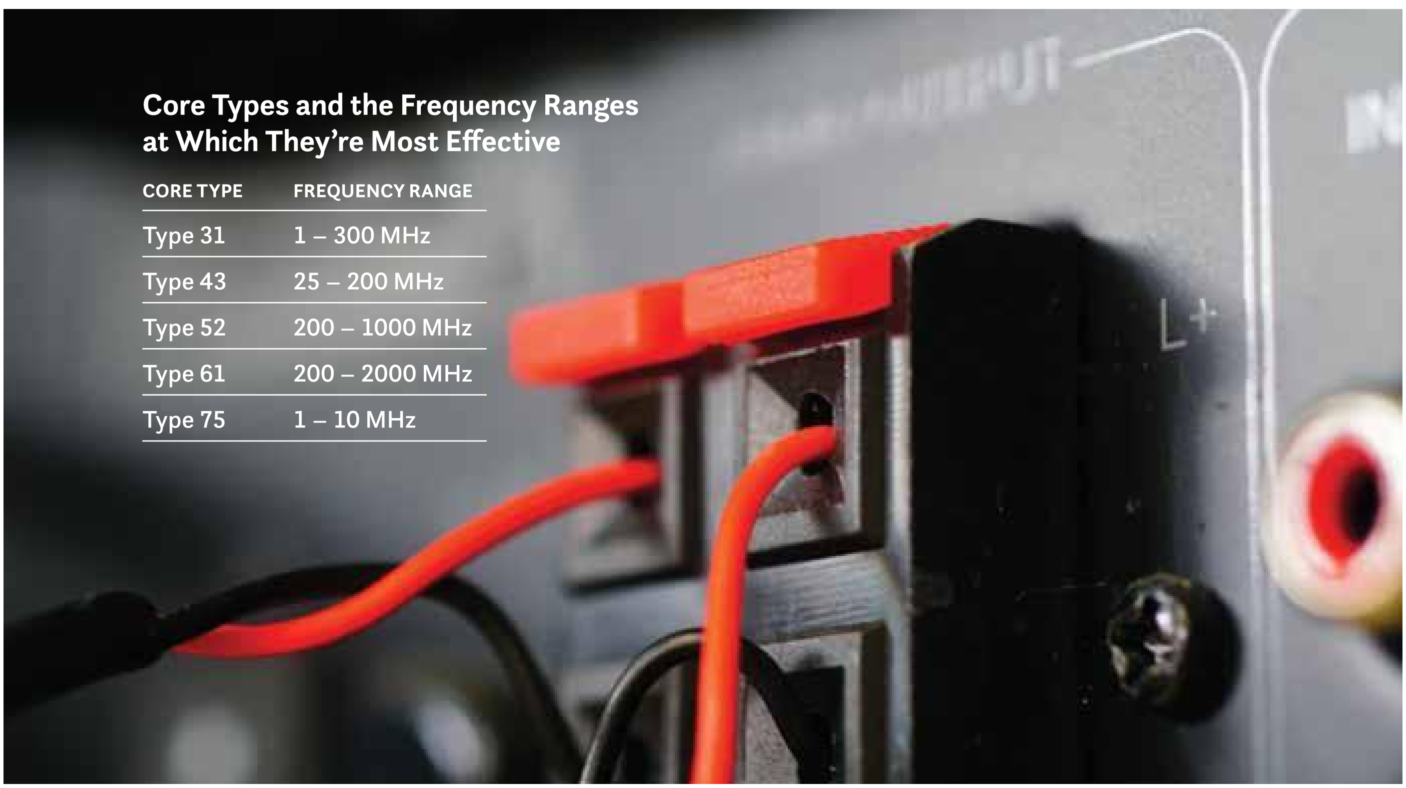
Making an RF Choke from Coaxial Cable

The toroid shown in (4) is made by using coaxial cable rather than wire. This is a type of choke that we discussed earlier, and it is used to reduce any RF energy on the antenna feed line

For applications such as blocking RF signals from entering audio amplifiers through various wires (such as speakers or ac power lines), you can create a choke by wrapping 7 or 8 turns of the troublesome wire through a toroid core (see the facing page). This is where the core type becomes important. See the table of "Core Types and the Frequency Ranges at Which They're Most Effective." As you might guess, Type 31 is popular for ham applications because it can be used to create effective chokes at frequencies amateurs use most often.

4

You can make an effective RF choke for an antenna system by wrapping coaxial cable through a sizable core.



Core Types and the Frequency Ranges at Which They're Most Effective

CORE TYPE	FREQUENCY RANGE
Type 31	1 – 300 MHz
Type 43	25 – 200 MHz
Type 52	200 – 1000 MHz
Type 61	200 – 2000 MHz
Type 75	1 – 10 MHz

A toroid core can be used as a choke to block RF signals from entering other devices. In this example, a dc power cable is wound through a core to prevent interference to a powered device. Unlike toroids used in circuitry, there is no hard-and-fast rule for the number of turns used to create a choke. Generally, the more the better. More turns means more inductance and a greater ability to “choke off” RF energy that might be traveling on the line and causing problems. In this example, 12 turns were wrapped around the core. We've numbered the turns to show the correct way to count them.

Multi-Aperture Cores

Multi-aperture cores are used in balun (balance-unbalance) transformers and find wide application as broadband transformers in communication and CATV circuits. They are also employed in auto air bag circuits to guard against accidental activation.

- *All multi-aperture cores are supplied burnished.*
- *For additional technical information on the use of these cores, see section "Use of Ferrites in Broadband Transformers" found on page 170.*
- *Multi-aperture cores are controlled for impedance limits only. They are tested for impedance with a single turn through two holes, using the Hewlett Packard HP 4193A Vector Impedance Meter.*
- *For impedance vs. frequency curves for these parts, see Figures 4-37.*
- *For any multi-aperture core requirement not listed in the catalog, please contact our customer service group for availability and pricing.*

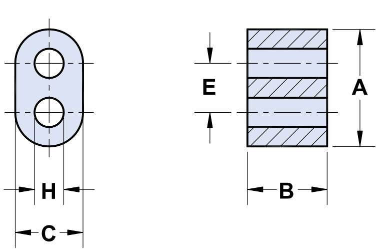


Figure 1

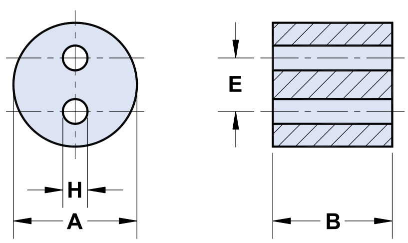


Figure 2

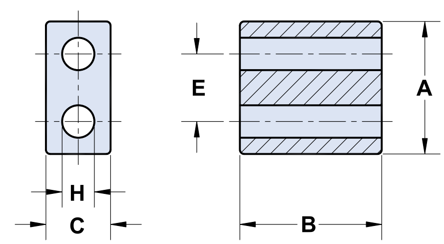


Figure 3

Dimensional letter designations have been changed from the 13th edition catalog and are now in accordance to the MMPA SFG-96.

Dimensions (**Bold numbers are in millimeters**, light numbers are nominal in inches.)

Part Number**	Fig.	A	B*	C	E	H	Wt (g)	Typical Impedance(Ω) [1]		Z, R_s, X_L vs. Frequency Curve
								25 MHz	100 MHz	
2873002302	1	**3.45±0.25** .136	**2.35±0.25** .093	**2.0±0.15** .079	**1.45±0.1** .057	**0.75+0.25** .034	.1	44	–	Figure 4
2843002302	1	**3.45±0.25** .136	**2.35±0.25** .093	**2.0±0.15** .079	**1.45±0.1** .057	**0.75+0.25** .034	.1	–	44	Figure 5
2861002302	1	**3.45±0.25** .136	**2.35±0.25** .093	**2.0±0.15** .079	**1.45±0.1** .057	**0.75+0.25** .034	.1	–	38	Figure 6
2873002702	1	**7.0±0.25** .276	**3.1±0.25** .122	**4.2 - 0.25** .160	**2.9±0.1** .114	**1.7+ 0.2** .071	.3	38	–	Figure 7
2843002702	1	**7.0±0.25** .276	**3.1±0.25** .122	**4.2 - 0.25** .160	**2.9±0.1** .114	**1.7+ 0.2** .071	.3	–	50	Figure 8
2861002702	1	**7.0±0.25** .276	**3.1±0.25** .122	**4.2 - 0.25** .160	**2.9±0.1** .114	**1.7+ 0.2** .071	.3	–	44	Figure 9
2873002402	1	**7.0±0.25** .276	**6.2±0.25** .244	**4.2 - 0.25** .160	**2.9±0.1** .114	**1.7+ 0.2** .071	.5	75	–	Figure 10
2843002402	1	**7.0±0.25** .276	**6.2±0.25** .244	**4.2 - 0.25** .160	**2.9±0.1** .114	**1.7+ 0.2** .071	.5	–	100	Figure 11
2861002402	1	**7.0±0.25** .276	**6.2±0.25** .244	**4.2 - 0.25** .160	**2.9±0.1** .114	**1.7+ 0.2** .071	.5	–	88	Figure 12
2873001802	2	**6.35±0.25** .250	**6.15±0.25** .242	–	**2.75±0.2** .108	**1.1 + 0.3** .050	.8	106	–	Figure 13
2843001802	2	**6.35±0.25** .250	**6.15±0.25** .242	–	**2.75±0.2** .108	**1.1 + 0.3** .050	.8	–	131	Figure 14

** This dimension may be modified to suit specific applications.*
*** Bold part numbers designate preferred parts.*

[1] Guaranteed Z Min is Z Typ -20%

Multi-Aperture Cores

Dimensions (**Bold numbers are in millimeters**, light numbers are nominal in inches.)

Part Number**	Fig.	A	B*	C	E	H	Wt (g)	Typical Impedance(Ω) [1]		Z, R_s, X_L vs. Frequency Curve
								25 MHz	100 MHz	
2861001802	2	**6.35±0.25** .250	**6.15±0.25** .242	–	**2.75±0.2** .108	**1.1 + 0.3** .050	.8	–	119	Figure 15
2873001702	2	**6.35±0.25** .250	**12.0±0.35** .471	–	**2.75±0.2** .108	**1.1 + 0.3** .050	1.6	200	–	Figure 16
2843001702	2	**6.35±0.25** .250	**12.0±0.35** .471	–	**2.75±0.2** .108	**1.1 + 0.3** .050	1.6	–	256	Figure 17
2861001702	2	**6.35±0.25** .250	**12.0±0.35** .471	–	**2.75±0.2** .108	**1.1 + 0.3** .050	1.6	–	230	Figure 18
2873001502	1	**13.3±0.6** .525	**6.6±0.25** .260	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	1.7	50	–	Figure 19
2843001502	1	**13.3±0.6** .525	**6.6±0.25** .260	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	1.7	–	88	Figure 20
2861001502	1	**13.3±0.6** .525	**6.6±0.25** .260	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	1.7	–	69	Figure 21
2873000302	1	**13.3±0.6** .525	**10.3±0.3** .407	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	2.6	75	–	Figure 22
2843000302	1	**13.3±0.6** .525	**10.3±0.3** .407	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	2.6	–	130	Figure 23
2861000302	1	**13.3±0.6** .525	**10.3±0.3** .407	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	2.6	–	106	Figure 24
2873000102	1	**13.3±0.6** .525	**13.4±0.3** .528	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	3.5	94	–	Figure 25
2843000102	1	**13.3±0.6** .525	**13.4±0.3** .528	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	3.5	–	175	Figure 26
2861000102	1	**13.3±0.6** .525	**13.4±0.3** .528	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	3.5	–	138	Figure 27
2873000202	1	**13.3±0.6** .525	**14.35±0.5** .565	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	3.7	106	–	Figure 28
2843000202	1	**13.3±0.6** .525	**14.35±0.5** .565	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	3.7	–	180	Figure 29
2861000202	1	**13.3±0.6** .525	**14.35±0.5** .565	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	3.7	–	150	Figure 30
2873006802	1	**13.3±0.6** .525	**27.0±0.75** 1.062	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	7.0	180	–	Figure 31
2843006802	1	**13.3±0.6** .525	**27.0±0.75** 1.062	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	7.0	–	300	Figure 32
2861006802	1	**13.3±0.6** .525	**27.0±0.75** 1.062	**7.5±0.35** .295	**5.7±0.25** .225	**3.8±0.25** .150	7.0	–	280	Figure 33
2843010402	3	**19.45±0.4** .765	**12.7±0.5** .500	**9.5±0.25** .375	**9.9±0.25** .390	**4.75±0.2** .187	7.5	–	200	Figure 34
2843010302	3	**19.45±0.4** .765	**25.4±0.7** 1.000	**9.5±0.25** .375	**9.9±0.25** .390	**4.75±0.2** .187	18	–	400	Figure 35
2843009902	3	**28.7±0.6** 1.130	**28.7±0.7** 1.130	**14.25±0.3** .560	**14.0±0.3** .550	**6.35±0.15** .250	48	–	500	Figure 36
2861010002	3	**30.2±0.6** 1.190	**28.7±0.7** 1.130	**15.0±0.4** .590	**14.6±0.4** .575	**6.8±0.2** .268	46	–	600	Figure 37

** This dimension may be modified to suit specific applications.*

*** Bold part numbers designate preferred parts.*

[1]Guaranteed Z Min is Z Typ -20%

Multi-Aperture Cores

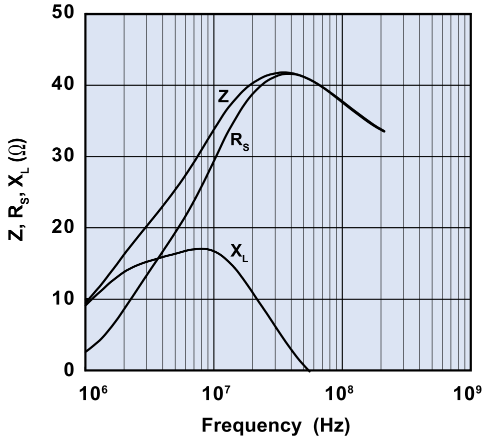


Figure 4 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873002302.

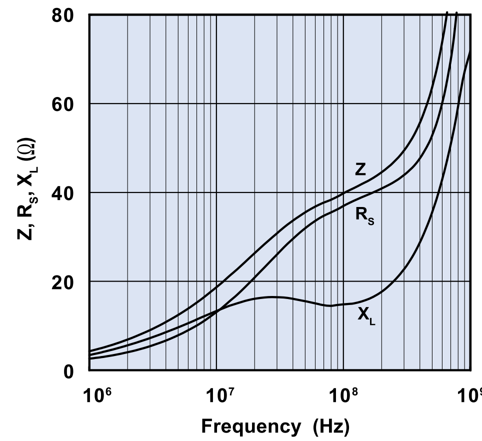


Figure 5 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843002302.

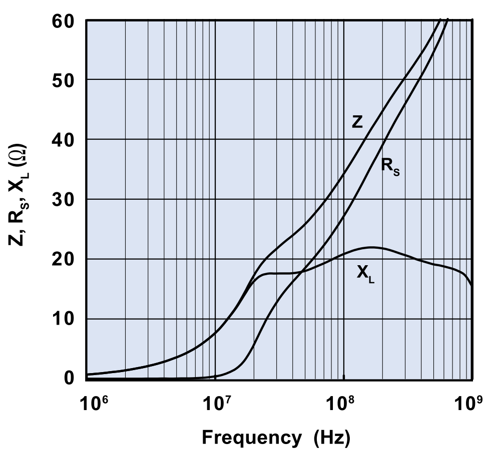


Figure 6 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861002302.

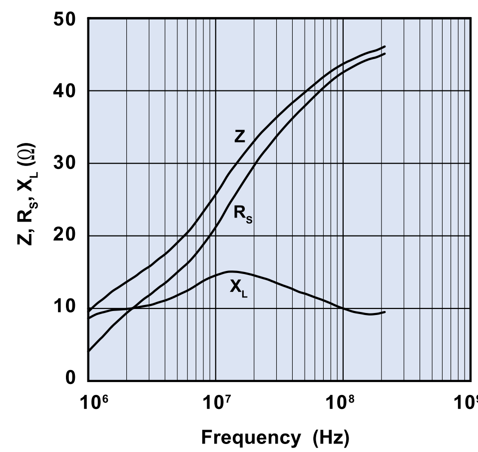


Figure 7 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873002702.

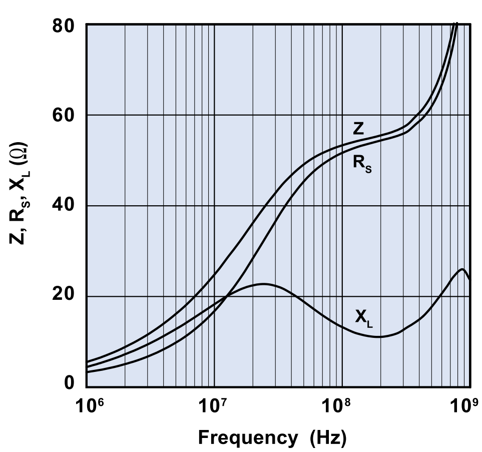


Figure 8 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843002702.

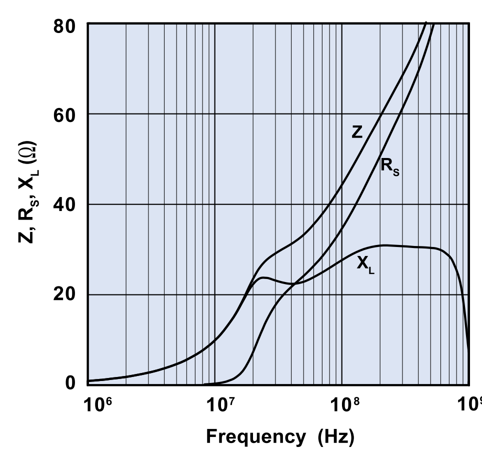


Figure 9 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861002702.

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Multi-Aperture Cores

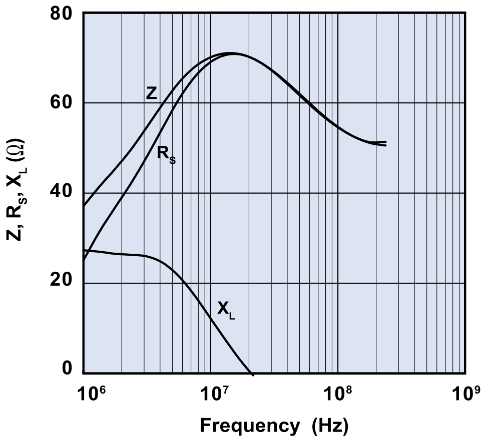

Figure 10 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873002402.

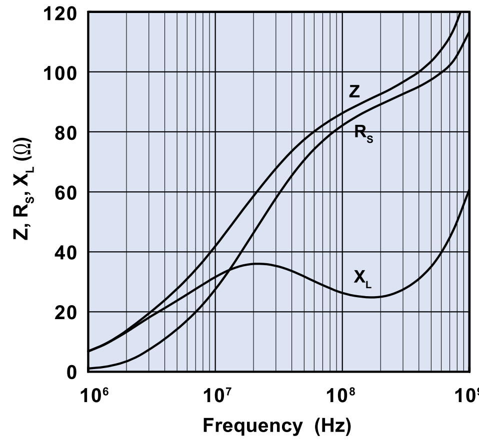

Figure 11 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843002402.

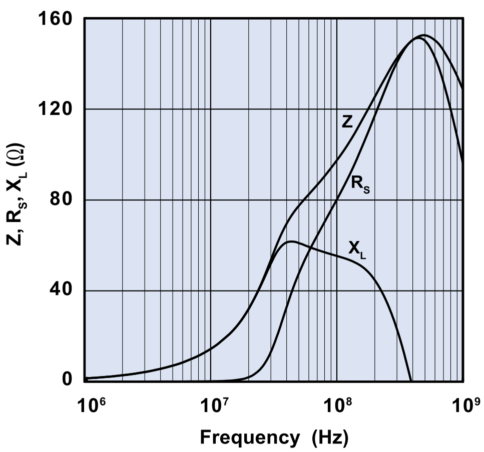

Figure 12 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861002402.

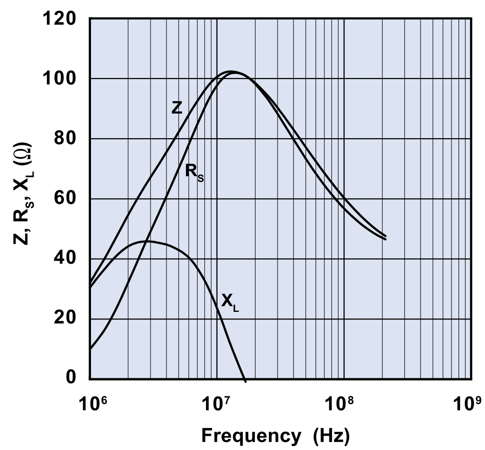

Figure 13 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873001802.

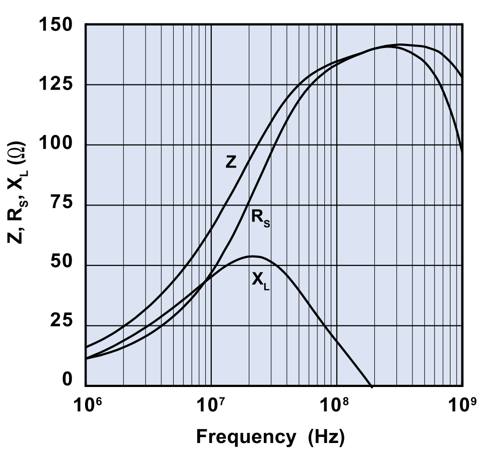

Figure 14 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843001802.

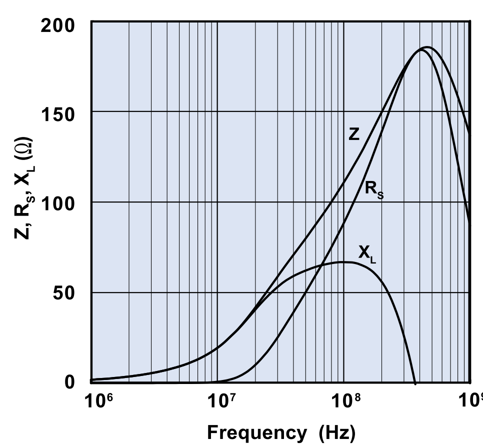

Figure 15 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861001802.

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Multi-Aperture Cores

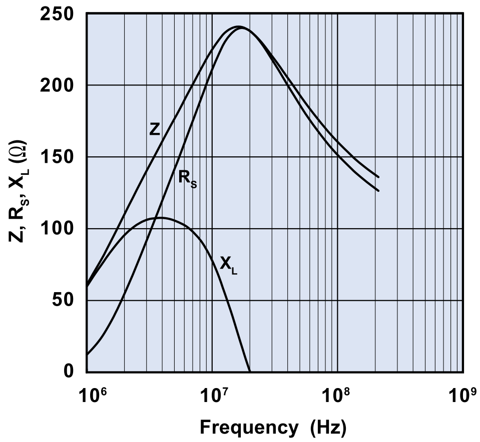

Figure 16 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873001702.

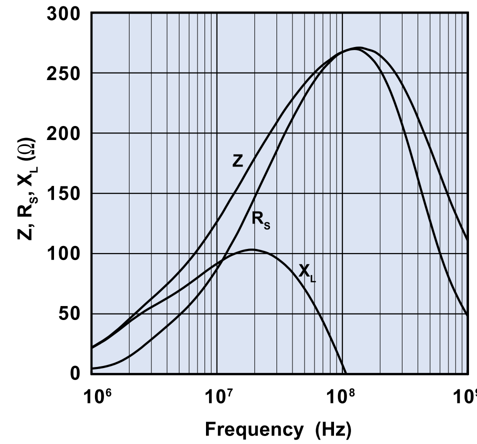

Figure 17 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843001702.

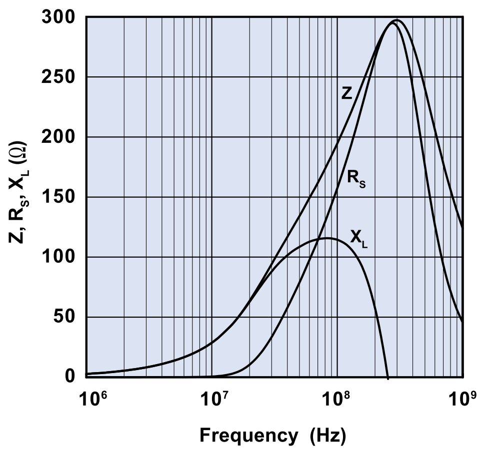

Figure 18 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861001702.

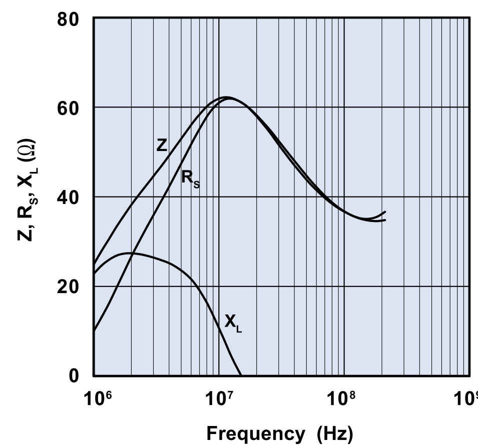

Figure 19 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873001502.

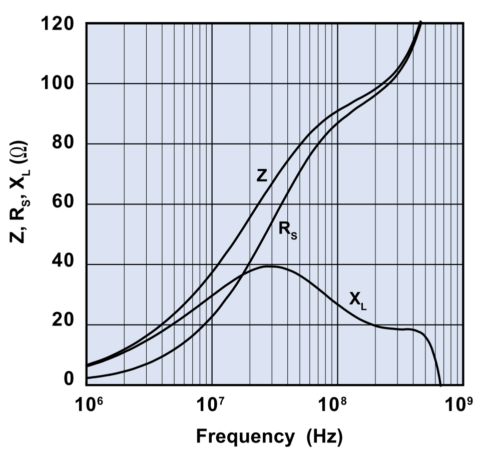

Figure 20 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843001502.

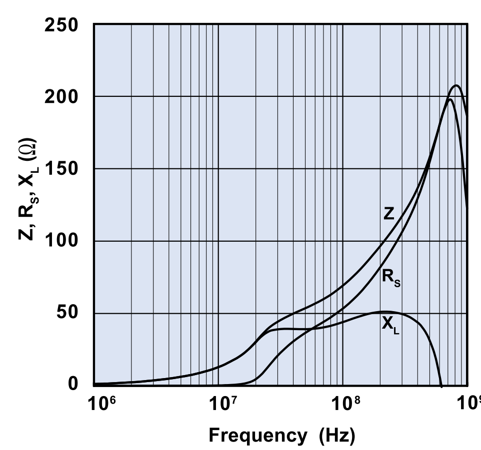

Figure 21 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861001502.

Multi-Aperture Cores

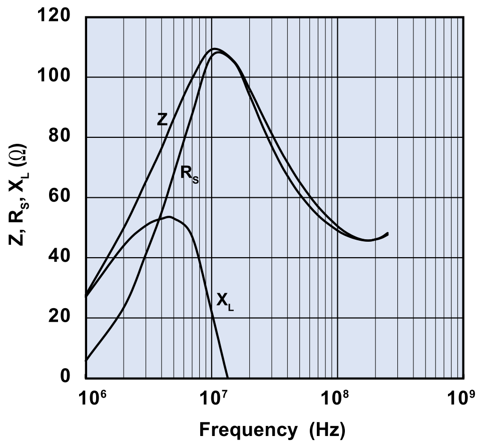

Figure 22 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873000302.

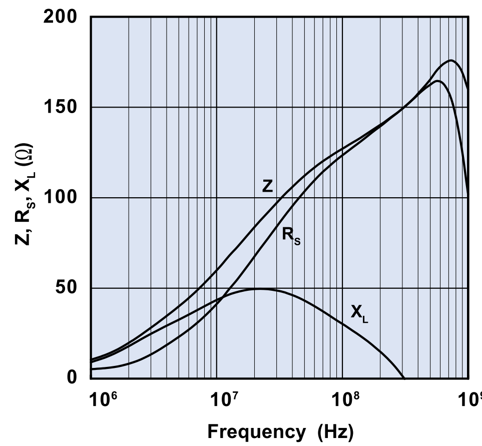

Figure 23 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843000302.

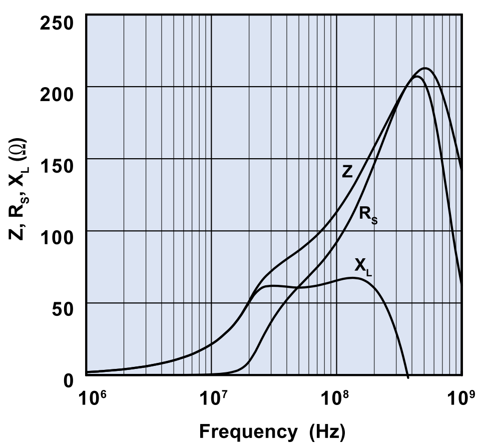

Figure 24 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861000302.

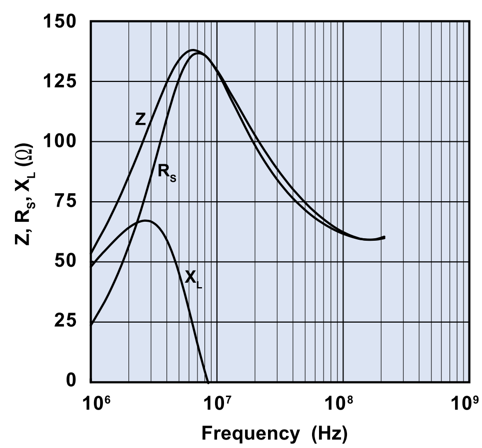

Figure 25 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873000102.

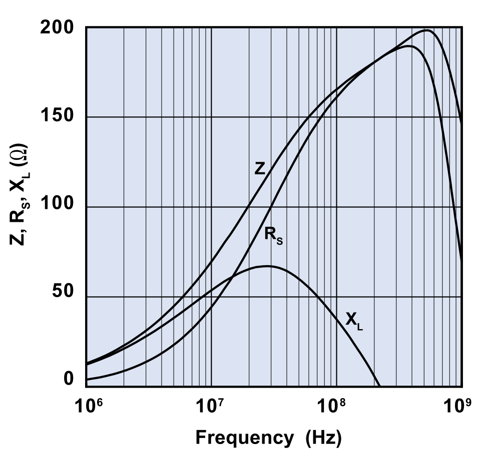

Figure 26 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843000102.

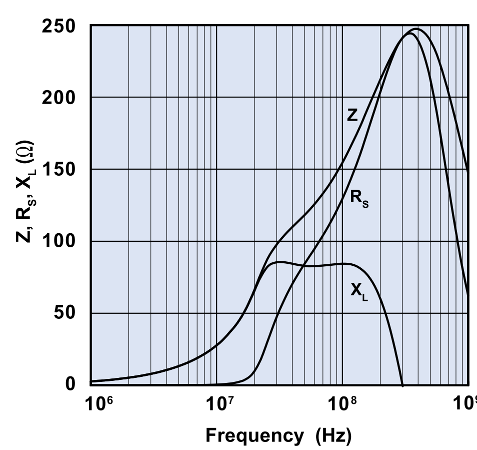

Figure 27 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861000102.

Multi-Aperture Cores

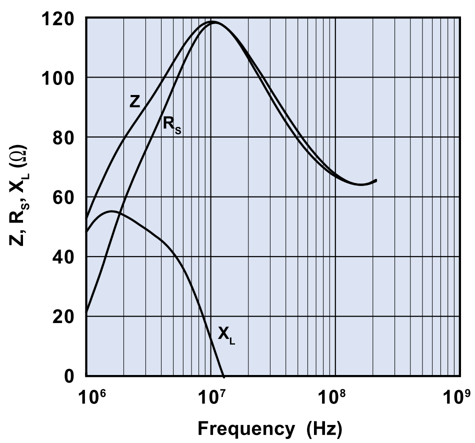

Figure 28 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873000202.

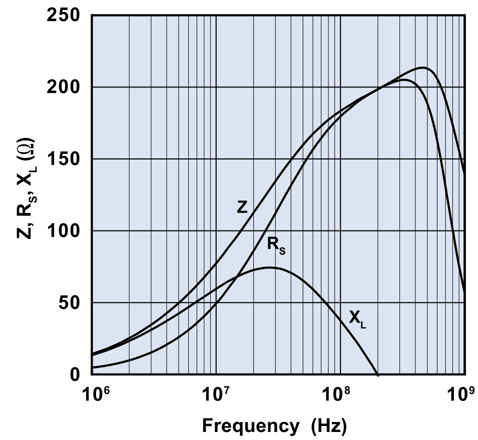

Figure 29 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843000202.

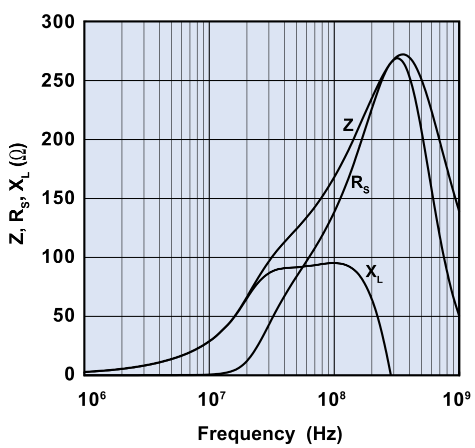

Figure 30 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861000202.

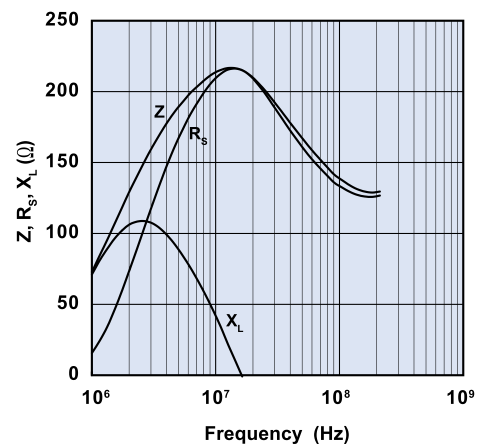

Figure 31 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2873006802.

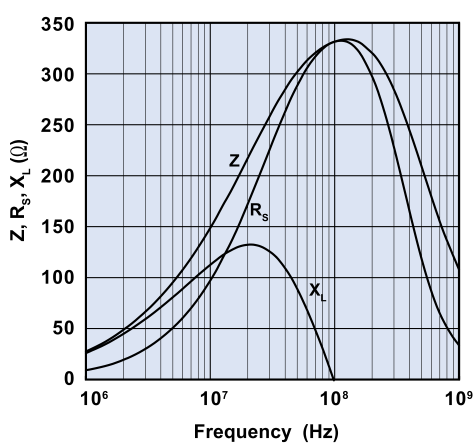

Figure 32 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843006802.

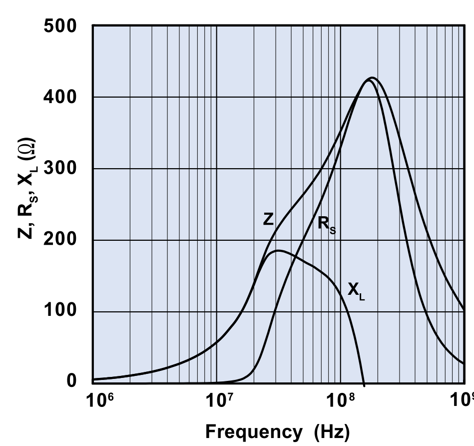

Figure 33 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861006802.

Multi-Aperture Cores

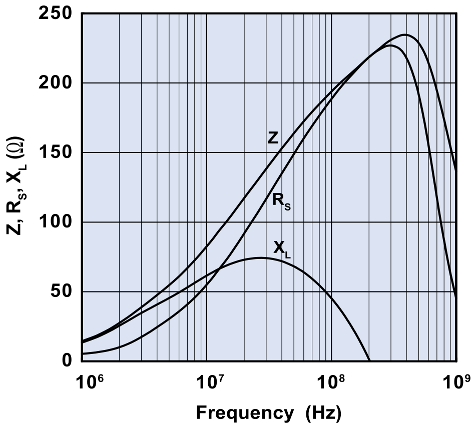

Figure 34 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843010402.

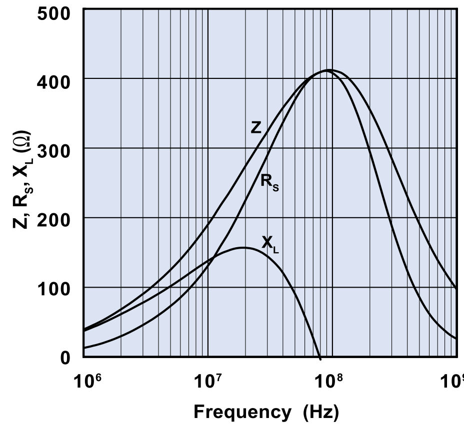

Figure 35 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843010302.

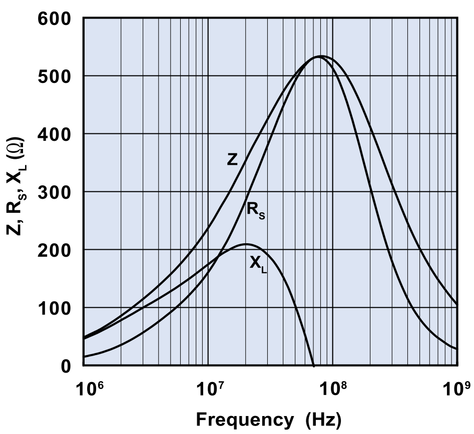

Figure 36 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2843009902.

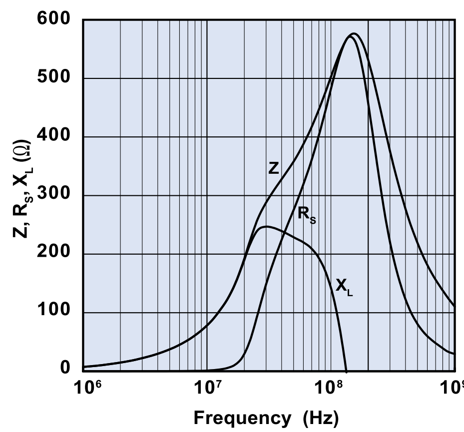

Figure 37 Impedance, reactance, and resistance vs. frequency curve for multi-aperture core 2861010002.

Fair-Rite Products Corp.
Your Signal Solution®

Toroids (5943007601)

Part Number: 5943007601

43 TOROID

Explanation of Part Numbers:
– Digits 1 & 2 = Product Class
– Digits 3 & 4 = Material Grade
□– 9th digit 1 = Parylene Coating, 2 = Thermo- Set Plastic Coating

A ring configuration provides the ultimate utilization of the intrinsic ferrite material properties. Toroidal cores are used in a wide variety of applications such as power input filters, ground- fault interrupters, common- mode filters and in pulse and broadband transformers.

□All toroidal cores are supplied burnished to break sharp edges.

Coating Options:
□□– Toroids with an outside diameter of 9.5 mm (0.375″) or smaller can be supplied Parylene C coated. The Parylene coating will increase the "A" and "C" dimensions and decrease the "B" dimension a maximum of 0.038 mm (0.0015″). The ninth digit of a Parylene coated toroid part number is a "1". See reference tables for the material characteristics of Parylene C. Parylene C coating is RoHS compliant.
□– Toroids with an outside diameter of 9.5 mm (0.375″) or larger can be supplied with a uniform coating of thermo- set plastic coating. This coating will increase the "A" and "C" dimensions and decrease the "B" dimension a maximum of 0.5 mm (0.020″). The 9th digit of the thermo- set plastic coated toroid part number is a "2". Thermo- set plastic coating is RoHS compliant.
□– Thermo- set plastic coated parts can withstand a minimum breakdown voltage of 1000 Vrms, uniformly applied across the "C" dimension of the toroid.

□For any toroidal core requirement not listed in the catalog, please contact our customer service department for availability and pricing.

The □C□ dimension may be modified to suit specific applications.

Weight: 15 (g)

Dim	mm	mm tol	nominal inch	inch misc.
A	22.1	±0.40	0.87	_
B	13.7	±0.30	0.54	_
C	12.7	±0.45	0.5	_

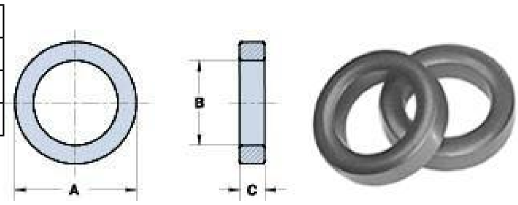

Chart Legend
Σl/ A : Core Constant, l_e : Effective Path Length, A_e : Effective Cross- Sectional Area, V_e : Effective Core Volume
A_L : Inductance Factor

Electrical Properties	
A_L(nH)	970 ±20%
Ae(cm^2)	0.52
Σl/ A(cm^{-1})	10.3
l_e(cm)	5.4
V_e(cm^3)	2.83

Toroids are tested for A_L values at 10 kHz.

Fair- Rite Products Corp. • One Commercial Row, Wallkill, New York 12589-0288
888-324-7748 • 845-895-2055 • Fax: 845-895-2629 • ferrites@fair- rite.com • www.fair- rite.com

Product Focus on

EMC Components

Specifying a Ferrite for EMI Suppression

by Carole U. Parker, Fair-Rite Products

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Our past article (see "How to Choose Ferrite Components for EMI Suppression," *Conformity*, June 2002) was intended to help design engineers optimize the performance of ferrite materials by analyzing the effects of frequency, field strength, temperature and core geometry. In our ideal world, safety (including effect on environment), quality and performance are paramount. However, the cost of the final component (which includes the ferrite) has, in many cases, become the deciding factor.

This article is written as an aide to the design engineer looking for alternative ferrite materials as a means to reduce cost.

A Review of Ferrite Applications

The following are three major applications for soft ferrite:

1. Low signal level
2. Power
3. EMI

The required intrinsic material characteristics and core geometry are dictated by each specific application. The intrinsic characteristics controlling the performance of low signal level applications are permeability (particularly with temperature), low core loss, and good magnetic stability with time and temperature. Applications include high Q inductors, common mode inductors, wideband, matching and pulse transformers, antenna elements for radios and both active and passive transponders. For power applications, the desirable characteristics are high flux density and low losses at the operating frequency and temperature. Applications include switchmode power supplies, magnetic amplifiers, dc-dc converters, power filters, ignition coils, and transformers for battery charging of electrical vehicles.

The intrinsic characteristic that most influences the performance of soft ferrite in suppression applications is the complex permeability, which is directly proportional to the core's impedance. There are three ways to use ferrites as suppressors of unwanted signals, conducted or radiated. The first, and least common, is as actual shields where ferrite is used to isolate a conductor, component or circuit, from an environment of radiated stray electromagnetic fields. In the second application, the ferrite is used with a capacitive element to create a low pass filter that is inductance-capacitance at low frequencies and dissipative at higher frequencies.

The third, and most common use, is when ferrite cores are used alone on component leads or in board level circuitry. In this application, the ferrite core prevents any parasitic oscillations and/or attenuates unwanted signal pickup or transmission that might travel along component leads or interconnected wires, traces, or cables.

In both the second and third applications, the ferrite core suppresses the conducted EMI by eliminating or greatly reducing the high frequency currents emanating from the EMI source. The introduction of the ferrite provides a sufficiently high frequency impedance that results in the suppression of the high frequency currents. Theoretically, the ideal ferrite would provide a high impedance at EMI frequencies, and zero impedance at all other frequencies. In reality, ferrite suppresser cores provide a frequency-dependent impedance, that is, low at frequencies below 1 MHz, and (depending upon the ferrite material) the maximum impedance that can be obtained between 10 MHz to 500 MHz.

Complex Permeability

As is consistent with electrical engineering principles in which alternating voltages and currents are denoted by complex parameters, so the permeability of a material can be represented as a complex parameter consisting of a real and an imaginary part. This is evidenced at high frequencies where the permeability separates into two components. The real component (μ') represents the reactive portion, and is in phase[1] with the alternating magnetic field, whereas the imaginary component (μ'') represents the losses, and is out of phase with the alternating magnetic field.

These may be expressed as series components (μ_s' μ_s'') or parallel components (μ_p' μ_p''). The graphs in Figures 1, 2 and 3 show the series components of the complex initial permeability as a function of frequency for three ferrite materials. Material type 73 is a manganese zinc ferrite with an initial permeability of 2500. Material type 43 is a nickel zinc ferrite with an initial permeability of 850. Material type 61 is a nickel-zinc ferrite with an initial permeability of 125.

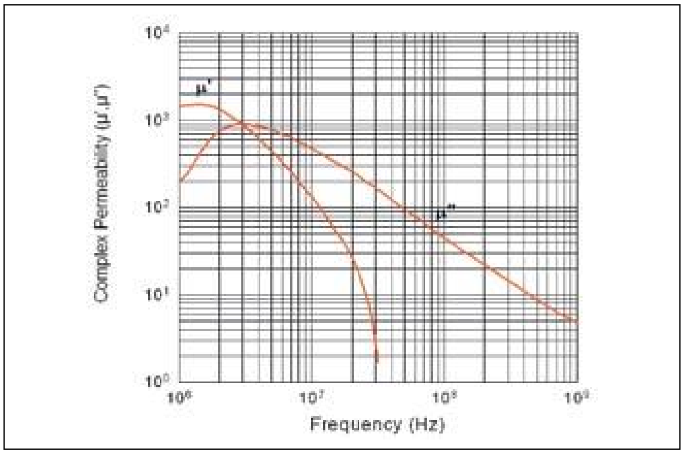


Figure 1

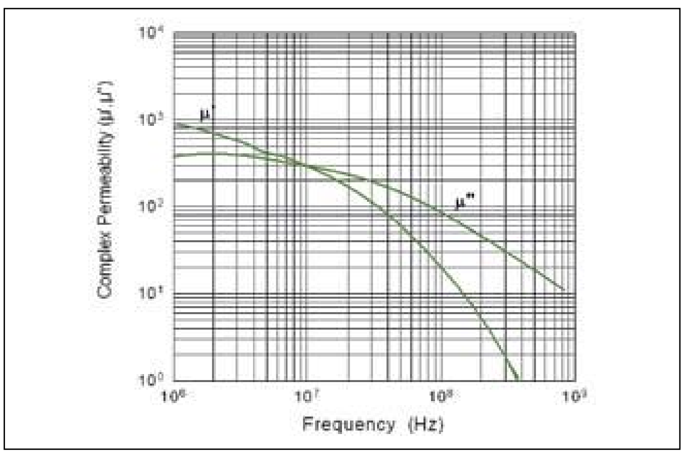


Figure 2

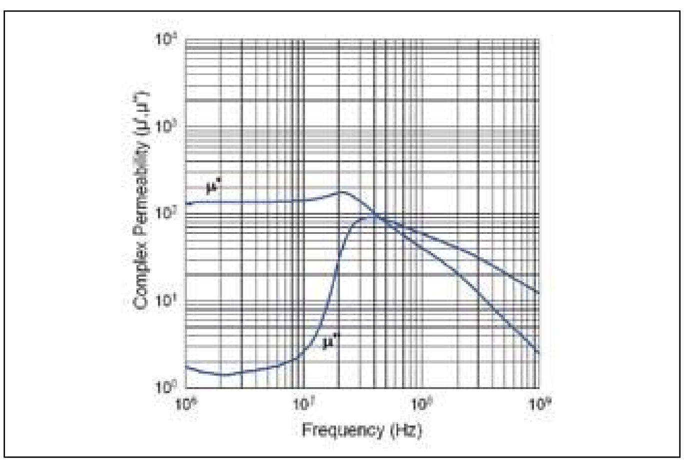


Figure 3

Concentrating on Figure 3, the series components of 61 type material, we see that the real part of the permeability, μ_s', remains constant with increasing frequency until a critical frequency is reached and then decreases rapidly. The losses, or μ_s'', rise, then peak as μ_s' falls. This decrease in μ_s' is due to the onset of ferrimagnetic resonance[2]. It should be noted that the higher the permeability, the lower the frequency at which this occurs. This inverse relationship was first observed by Snoek and given the following formula:

$$f_{res} = \frac{\gamma M_{sat}}{3\pi (\mu_i - 1)} \text{ Hz} \qquad \text{Equation 1}$$

Where:

f_{res} = frequency at which μ_s'' is maximum
γ = gyromagnetic ratio=0.22 x 10^6 A^{-1} m
μ_i = initial permeability
M_{sat} = 250-350 Am^{-1}

This same Equation can be approximated by:

$$f_{res} = B_{sat}/\mu_i \text{ MHz} \qquad \text{Equation 2}$$

Since ferrite cores used in low signal level and power applications are concerned with magnetic parameters below this frequency, rarely does the ferrite manufacturer publish data for permeability and/or losses at higher frequencies. However, higher frequency data is essential when specifying ferrite cores used in the suppression of EMI.

Relationship Between Complex Permeability and Impedance

The characteristic that is specified by most ferrite manufacturers for components used in EMI suppression is impedance. Impedance is easily measured on readily available commercial analyzers with direct digital readouts. Unfortunately, impedance is usually specified at particular frequencies and is the scalar quantity representing the magnitude of the complex impedance vector. Although this information is valuable, it is often not sufficient, especially when modeling the ferrite's circuit performance. In order to accomplish this, the impedance value and phase angle for the components, or the complex permeability for the specific material, must be available.

But even before beginning to model the performance of a ferrite component in a circuit, the designer should know the following:

- The frequency of unwanted signals;
- The source of the EMI (radiated/conducted);

- Operating conditions (environment);
- If high resistivity is required because of multiple turns, conductor pins in a connector filter plate or position in the circuit;
- The circuit impedance, source and load;
- How much attenuation is required;
- The allowable space on the board.

The design engineer can then compare materials at the relevant frequencies for the complex permeability, heeding effects of temperature and field strength. After this, core geometry can be selected, from which the inductive reactance and series resistance can be defined.

The Equations

The impedance of a ferrite core in terms of permeability is given by:

$$Z = j\omega\mu\, L_o \qquad \text{Equation 3}$$

And:

$$\mu = \mu' - j\mu'' = (\mu_s'^2 + (j\mu_s'')^2)^{1/2} \qquad \text{Equation 4}$$

Where:

μ'= real part of the complex permeability
μ''= imaginary part of the complex permeability
j = unit imaginary vector
L_o= the air core inductance

Therefore:

$$Z = j\,\omega L_o\,(\mu' - j\mu'') \qquad \text{Equation 5}$$

The impedance of the core is also considered to be a series combination of inductive reactance (X_L) and the loss resistance (R_s), both of which are frequency dependent. A loss free core would have an impedance that would be given by the reactance:

$$X = j\,\omega L_s \qquad \text{Equation 6}$$

A core that has magnetic losses may be represented as having an impedance:

$$Z = R_s + j\,\omega L_s \qquad \text{Equation 7}$$

Where:

R_s= total series resistance $=R_m + R_e$
R_m = Equivalent series resistance due to the magnetic losses
R_e = equivalent series resistance for copper losses

At low frequencies, the impedance of the component is primarily the inductive reactance. As frequency increases, the inductance decreases at the same time the losses increase and the total impedance increases. Figure 4 is a typical curve of X_L, R_s and Z vs. frequency for our a medium permeability material.

Knowing that, the magnetic quality factor:

$$Q = \mu'/\mu'' = \omega L_s/R_s \qquad \text{Equation 8}$$

Then, the inductive reactance is made directly proportional to the real part of the complex permeability by L_o, the air core inductance:

$$j\,\omega L_s = j\,\omega L_o \mu_s'$$

The loss resistance is also made directly proportional to the imaginary part of the complex permeability by the same constant:

$$R_s = \omega L_o \mu_s''$$

Substituting into Equation 7 for impedance:

$$Z = \omega L_o \mu_s'' + j\,\omega L_o \mu_s'$$

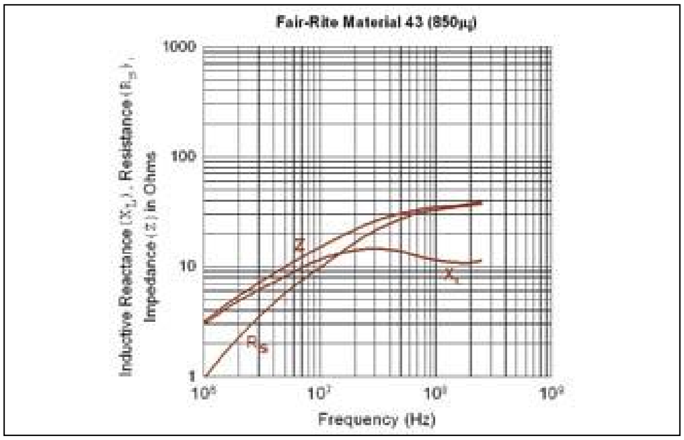


Figure 4

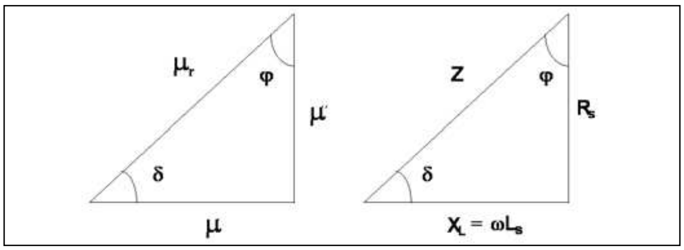


Figure 5

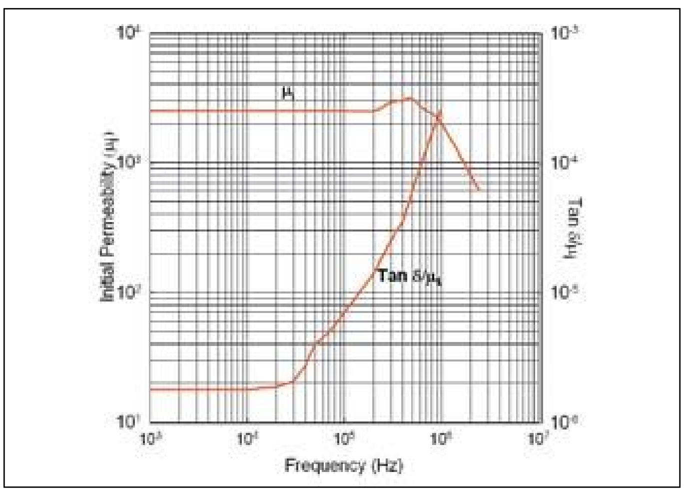


Figure 6

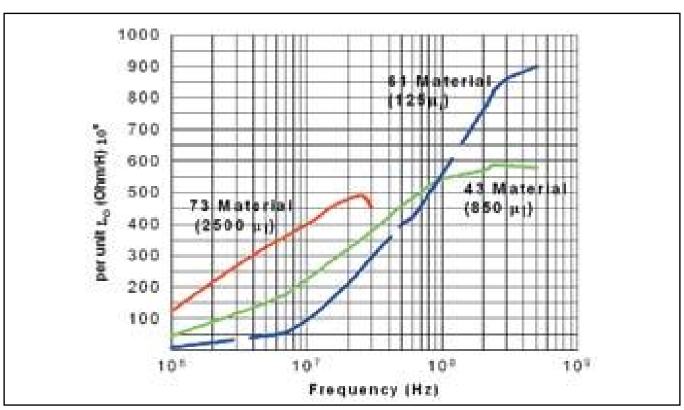


Figure 7

And factoring:

$Z = j\,\omega L_o\,(\mu_s' - j\,\mu_s'')$ Equation 9

In Equation 9, the core material is given by μ_s' and μ_s'', and the core geometry is given by L_o. Thus, knowing the complex permeability for different ferrites, a comparison can be made to obtain the most suitable material at the desired frequency or frequency range. After the optimum material is chosen, the best size component can be selected. The vector representation for both the complex permeability and impedance is found in Figure 5.

If the manufacturer supplies graphs of complex permeability vs frequency for the ferrite materials recommended for suppression applications, then a comparison of cores shapes and core materials for optimizing impedance is straight forward. Unfortunately, this information is rarely made available. However, most manufacturers supply curves of initial permeability and losses vs. frequency. From this data, a comparison of materials for optimizing core impedance can be obtained.

Material Selection--Example

Referring to Figure 6 (Initial Permeability & Loss Factor[3] vs. Frequency for Fair-Rite 73 material), suppose a designer wants to guarantee maximum impedance between 100 and 900 kHz. 73 material is chosen. For purposes of modeling, the designer also needs to know the reactive and resistive portions of the impedance vector at 100 kHz (105 Hz) and 900 kHz. This information can be derived from the graphs as follows:

At 100kHz, $\mu_s' = \mu_i = 2500$

And:

$(\tan\delta/\mu_i) = 7 \times 10^{-6}$

Since:

$\tan\delta = \mu_s''/\mu_s'$

Then:

$\mu_s'' = (\tan\delta/\mu_i) \times (\mu_i)^2 = 43.8$

Calculating for the complex permeability:

$\mu = \mu' - j\mu'' = ((\mu_s'^2 + (j\mu_s'')^2)^{1/2} = 2500.38$

It should be noted that, as expected, μ" adds very little to the total permeability vector at this low frequency. The impedance of the core is primarily inductive.

However at 900 kHz, μ_s'' has become a significant contributor:

$\mu_s' = 2100$, $\mu_s'' = 1014$ $\mu = 2332$

Core Selection

The designer knows that the core must accept a #22 wire, and fit into a space of 10 mm by 5 mm. The inside diameter will be specified as .8 mm. Solving for the estimated impedance and its components, first a bead with an outside diameter of 10 mm and a height of 5 mm is chosen:

at 100 kHz

Since:

$Z = \omega L_o\, \mu$ and Toroidal $L_o = .0461\ N^2 \log^{10} (OD/ID)\ Ht\ 10^{-8}$ (H)

Then:

$Z = \omega L_o (2500.38) = (6.28 \times 10^5) \times .0461 \times \log^{10} (10/.8) \times 5 \times (2500.38) \times 10^{-8} = 3.97$ ohms

Where:

$R_s = L_o\, \omega\mu_s'' = .069$ ohms
$X_L = L_o\, \omega\mu_s' = 3.97$ ohms

at 900 kHz

$Z = 33.3$ ohms, $R_s = 14.48$ ohms, $X_L = 30.0$ ohms

Then a bead with an outside diameter of 5 mm and a length of 10 mm is selected:

at 100 kHz

$Z = \omega L_o (2500.38) = (6.28 \times 10^5) \times .0461 \times \log^{10} (5/.8) \times 10 \times (2500.38) \times 10^{-8} = 5.76$ ohms

Where:

$R_s = L_o\, \omega\mu_s'' = .100$ ohms
$X_L = L_o\, \omega\mu_s' = 5.76$ ohms

at 900 kHz

$Z = 48.1$ ohms, $R_s = 20.9$ ohms, $X_L = 43.3$ ohms

In this instance, as in most, maximum impedance is achieved by using a smaller OD with the longer length. If the ID were larger (for instance 4 mm), the reverse should have been true.

This same approach can be used if graphs of Impedance per unit Lo and phase angle vs. frequency are supplied. Figures 9, 10 and 11 are representative of such curves for the same three materials used throughout this article.

Example

The designer wants to guarantee maximum impedance for the frequency range of 25 MHz to 100 MHz. The available board space is again 10 mm by 5 mm and the core must accept a #22 awg wire. Referring to Figure 7, Impedance per unit L_o for three ferrite materials, or Figure 8, complex permeability for the same three materials, an 850 μ_i material is chosen[4]. Using the graph of Figure 9, Z/L_o for the medium permeability material, at 25 MHz, is 350×10^8 ohm/H.

Solving for the estimated impedance:

$Z = 350\ 10^8 \times .0461 \times \log^{10} (5/.8) \times 10 \times 10^{-8}$

$Z = 128.4$ ohm $\Phi = 30$ degrees
$X_L = Z \sin \Phi = 126.8$ ohms
$R_s = Z \cos \Phi = 19.81$ ohms

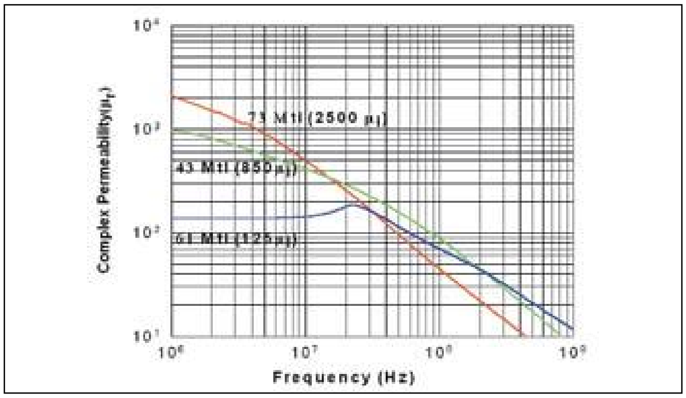


Figure 8

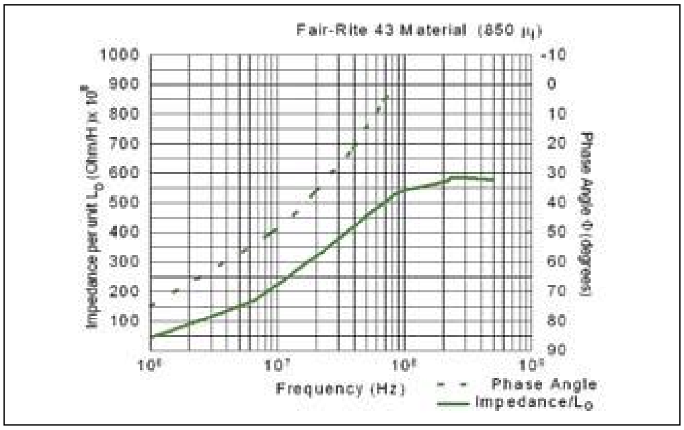


Figure 9

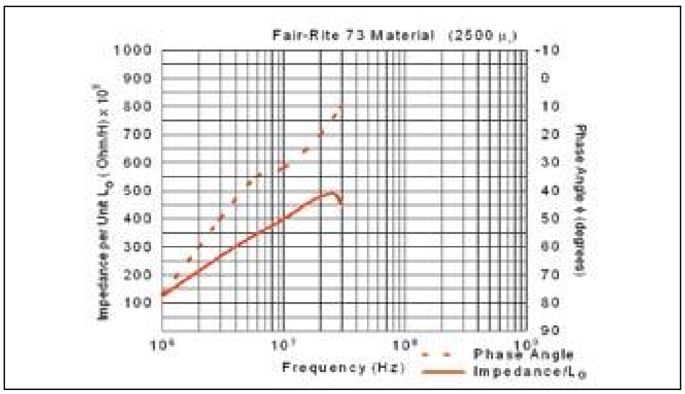


Figure 10

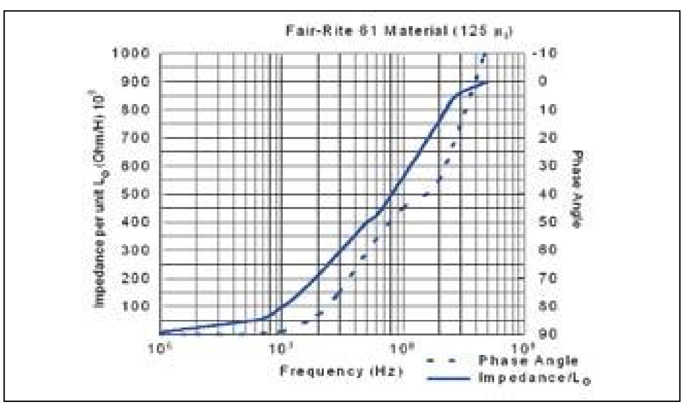


Figure 11

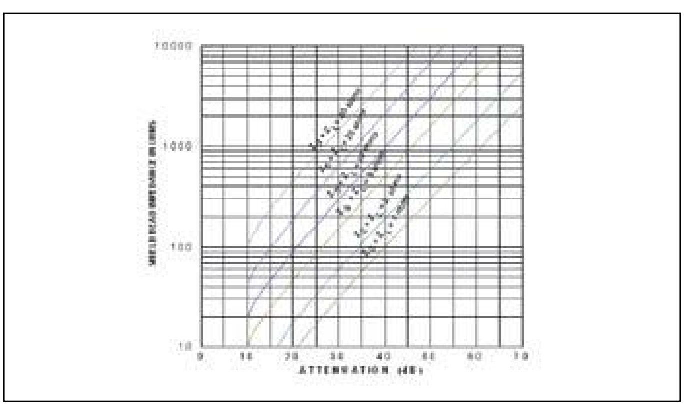


Figure 12

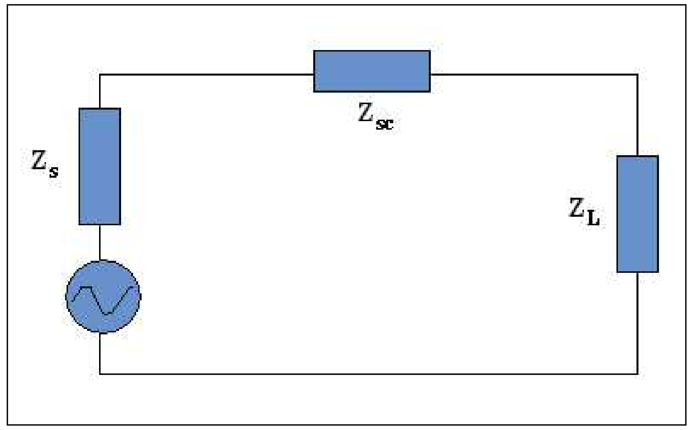


Figure 13

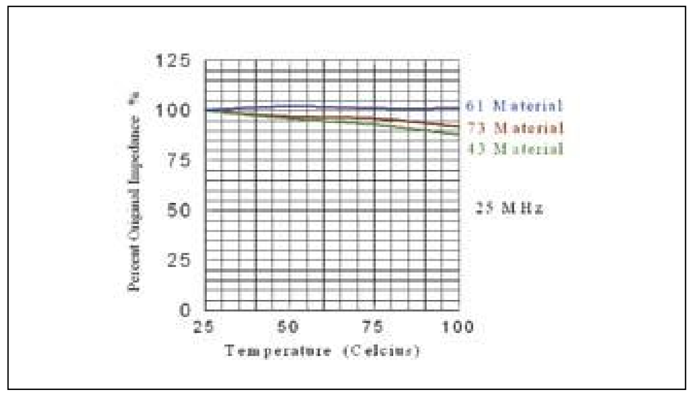


Figure 14

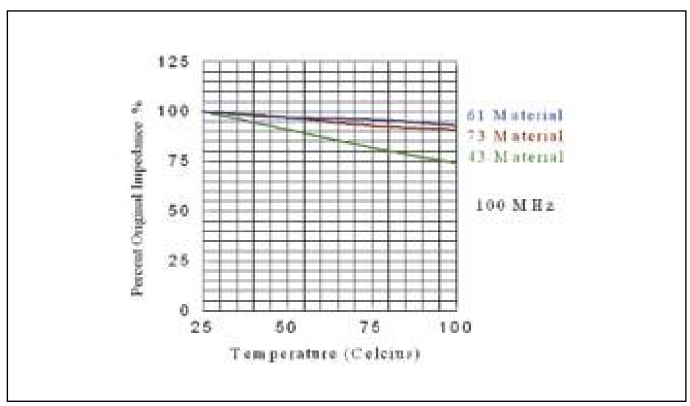


Figure 15

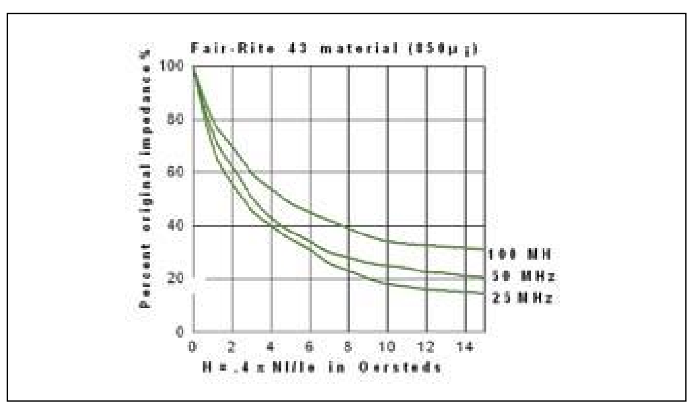


Figure 16

And, at 100 MHz

$Z = 179.8$ ohms $\Phi = 0$
$X_L = 0$ ohms $R_s = 179.8$ ohms

The same approach may be used for different materials, dimensions, and frequencies.

The previous discussion assumed that the core of choice was cylindrical. If the ferrite core being used is for flat ribbon, or bundled cable, or a multi-hole plate, the calculation for the L_o becomes more difficult, and fairly accurate figures for the cores path length and effective area must be obtained in order to calculate the air core inductance. This can be done by mathematically sectioning the core and summing the calculated path length and magnetic area for each section. In all cases though, an increase or decrease in impedance will be directly proportional to an increase or decrease in the height/length of the ferrite core.[5]

Relationship Between Impedance and Attenuation

As stated, most manufacturers are specifying cores for EMI applications in terms of impedance, but often the end user needs to know the attenuation. The relationship that exists between these two parameters is:

$$\text{Attenuation} = 20 \log^{10} ((Z_s + Z_{sc} + Z_L) / (Z_s + Z_L)) \text{ dB}$$

Where:

Z_s = Source impedance
Z_{sc} = Suppressor core impedance
Z_L = Load impedance

The relationship is dependent on the impedance of the source generating the noise and the impedance of the load receiving it. These values are usually complex numbers that can be infinite in scope and not easily obtained by the designer. Selecting a value of one ohm for both the load and the source impedance (as may be the case when the source is a switch mode power supply and the load many low impedance circuits) simplifies the equation and allows comparison of ferrite cores in terms of attenuation.

Under these conditions, the equation reduces to:

$$\text{Attenuation} = 20 \log^{10} (Z_{sc}/2) \text{ dB}$$

The graph in Figure 12 is a family of curves that show the relationship between the shield bead impedance and the attenuation for a number of commonly used values of the load plus the generator impedance.

Figure 13 is the equivalent circuit of an interference source with an internal impedance of Z_s, generating an interference signal through the series impedance of the suppressor core Z_{sc} and the load impedance Z_L.

The Environment

Temperature

As stated, ferrite's magnetic parameters can be affected by temperature and field strengths.

Figures 14 and 15 are graphs of impedance vs. temperature for the same three ferrite materials. The most stable of these materials is the 61 material, with a decrease in impedance of 8% at 100° C and 100 MHz. This is compared to a 25% drop in impedance for the 43 material at the same frequency and temperature. These curves, when supplied, may be used to adjust the specified room temperature impedance if desired attenuation is to be at elevated temperatures.

Field Strength

As in the case of temperature, dc and 50 or 60 Hz power current will also affect the same intrinsic ferrite characteristics which, in turn, will result in lowering of the impedance of the core. Figures 16, 17 and 18 are typical of curves that illustrate the effect of biases on impedance a ferrite material. The curve depicts the degradation of impedance as a function of field

Definitions/Equations

Air Core Inductance - L_o (H)
The inductance that would be measured if the core had unity permeability and the flux distribution remained unaltered.
General formula $L_o = 4\pi N^2 10^{-9}$ (H) C^1
Toroidal $L_o = .0461 N^2 \log^{10}$ (OD/ID) Ht 10^{-8} (H)
Dimensions in mm's

Attenuation - A (dB)
The decrease in signal magnitude in transmission from one point to another. It is a scalar ratio of the input magnitude to the output magnitude in decibels.

Core Constant - C^1 (cm^{-1})
The summation of the magnetic path lengths of each section of a magnetic circuit divided by the corresponding magnetic area of the same section.

Core Constant - C^2 (cm^{-3})
The summation of the magnetic path lengths of each section of a magnetic circuit divided by the square of the corresponding magnetic area of the same section.

Effective Dimensions of a Magnetic Circuit
Area Ae (cm^2), Path Length le (cm) and Volume Ve (cm^3)
For a magnetic core of given geometry, the magnetic path length, the cross-sectional area and the volume that a hypothetical toroidal core of the same material properties should possess to be the magnetic equivalent to the given core.

Field Strength - H (oersted)
The parameter characterizing the amplitude of the field strength.
$H = .4 \pi NI/le$ (oersted)

Flux Density - B (gauss)
The corresponding parameter for the induced magnetic field in a area perpendicular to the flux path.

Impedance - Z (ohm)
The impedance of a ferrite may be expressed in terms of its complex permeability.
$Z = j\,\omega L_s + R_s = j\,\omega L_o(\mu_s' - j\,\mu_s'')$ (ohm)

Loss Tangent - tan δ
The loss tangent of the ferrite is equal to the reciprocal of the Q of the circuit.

Loss Factor - tan δ/μ_i
The phase displacement between the fundamental components of the flux density and the field strength divided by the initial permeability.

Phase Angle - Φ
The phase shift between the applied voltage and current in a inductive device.

Permeability - μ
The permeability obtained from the ratio of the flux density and the applied alternating field strength is:

Amplitude Permeability, μ_a - when stated values of flux density, are greater than that used for initial permeability.

Effective Permeability, μ_e - when a magnetic circuit is constructed with an airgap or airgaps, and then the permeability is that of a hypothetical homogeneous material which would provide the same reluctance.

Incremental Permeability, $\mu\Delta$ - when a static field is superimposed.

Initial Permeability, μ_i - when the flux density is kept below 10 gauss.

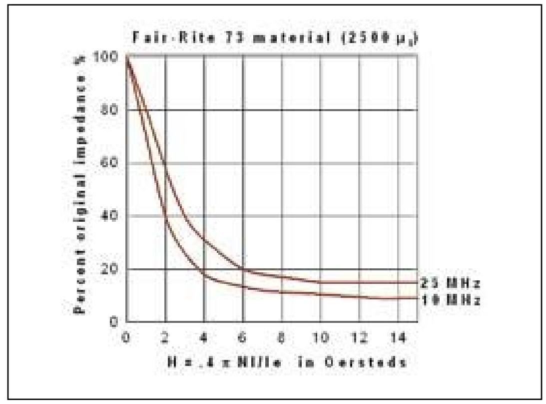


Figure 17

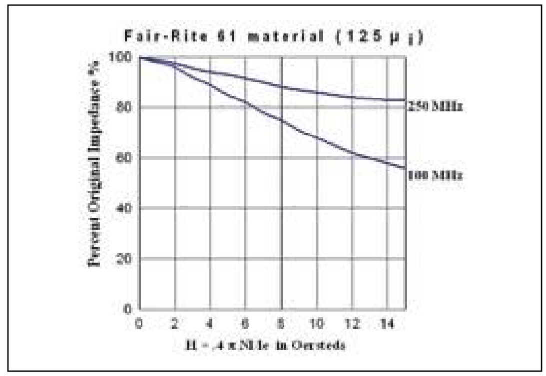


Figure 18

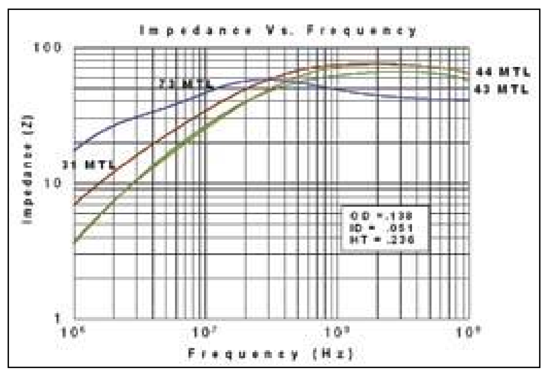


Figure 19

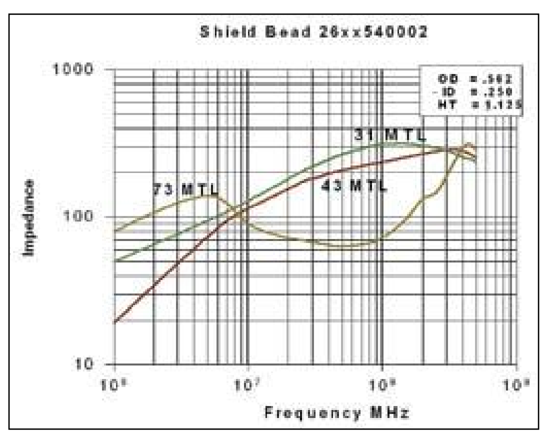


Figure 20

strength for a specific material as a function of frequency. It should be noted that, as frequency increases, the effect of biases diminishes.

New Materials

Since the compilation of this data, our company has introduced two new materials. Our 44 is a nickel zinc, medium permeability material, while 31 is a manganese zinc high permeability material.

Figure 19 is a plot of impedance vs. frequency for the same size bead in 31, 73, 44 and 43 materials. 44 material is an improved 43 material with higher dc resistivity, 10^9 ohm cm, better thermal shock characteristics, temperature stability and higher curie temperature (T_c). When compared to our 43 material, 44 material has slightly higher impedance vs. frequency characteristics.

Still, material 31 exhibits higher impedance than either 43 or 44 throughout the measured frequency range. Designed to alleviate the problem of dimensional resonance that affects low frequency suppression performance of the larger manganese-zinc cores, 31 has found successful applications as cable connector suppressor cores and large toroidal cores. Figure 20 is a curve of impedance vs frequency for 43, 31, and 73 material for a Fair-Rite core with an OD of .562", ID of .250 and a HT of 1.125.

When comparing Figure 19 to Figure 20, it should be noted that for the smaller core in Figure 19, for frequencies up to 25 Mhz, 73 material is the optimum suppression material. However, as the core cross section increases, the maximum frequency decreases, as can be shown by the data in Figure 20, where the highest frequency where 73 is optimum is 8 Mhz.

Also noteworthy is that 31 material is superior from 8 MHz to 300 MHz. However, being a manganese zinc ferrite, 31 material has a much lower volume resisitivity of 102 ohm-cm and exhibits greater changes in impedance with extreme temperature variation. ☐

Carole U. Parker *is the president of Fair-Rite Products Corporation, and can be reached at parkerc@fair-rite.com.*

References

- Soft Ferrites, Properties and Applications, 2nd Edition 1988, Snelling E. C., Butterworth, Boston MA
- Magnetic Materials and their Applications, 1974, Heck, C. Crane, Russak and Company, New York, NY
- "Prayer Beads" Solve Many of Your EMI Problems, Parker C., Tolen, B., Parker, R., EMC Technology, Vol. 4, No. 2, 1985
- Fair-Rite Products Soft Ferrite Catalog, 15th Edition, 2007

Notes

1. In phase is when the maxima and minima of the magnetic field, H, and those of the induction B, coincide. Out of phase is when the maxima and minimum is displaced by 90°.
2. Ferromagnetic Resonance is also called spin precession resonance.
3. It should be noted that the impedance for each ferrite material is optimum over a specific frequency range. As a rule of thumb, the higher the permeability, the lower the frequency range.
4. It should be noted that the impedance for each ferrite material is optimum over a specific frequency range. As a rule of thumb, the higher the permeability, the lower the frequency range.
5. This remains true so long as the increase in height/length does not cause the core to be in dimensional resonance.

FAST Link www.conformity.com/1870

Design Considerations for High Frequency Magnetic Materials

PRESENTED BY JOHN LYNCH
DIRECTOR OF ENGINEERING
FAIR-RITE PRODUCTS CORP

Agenda

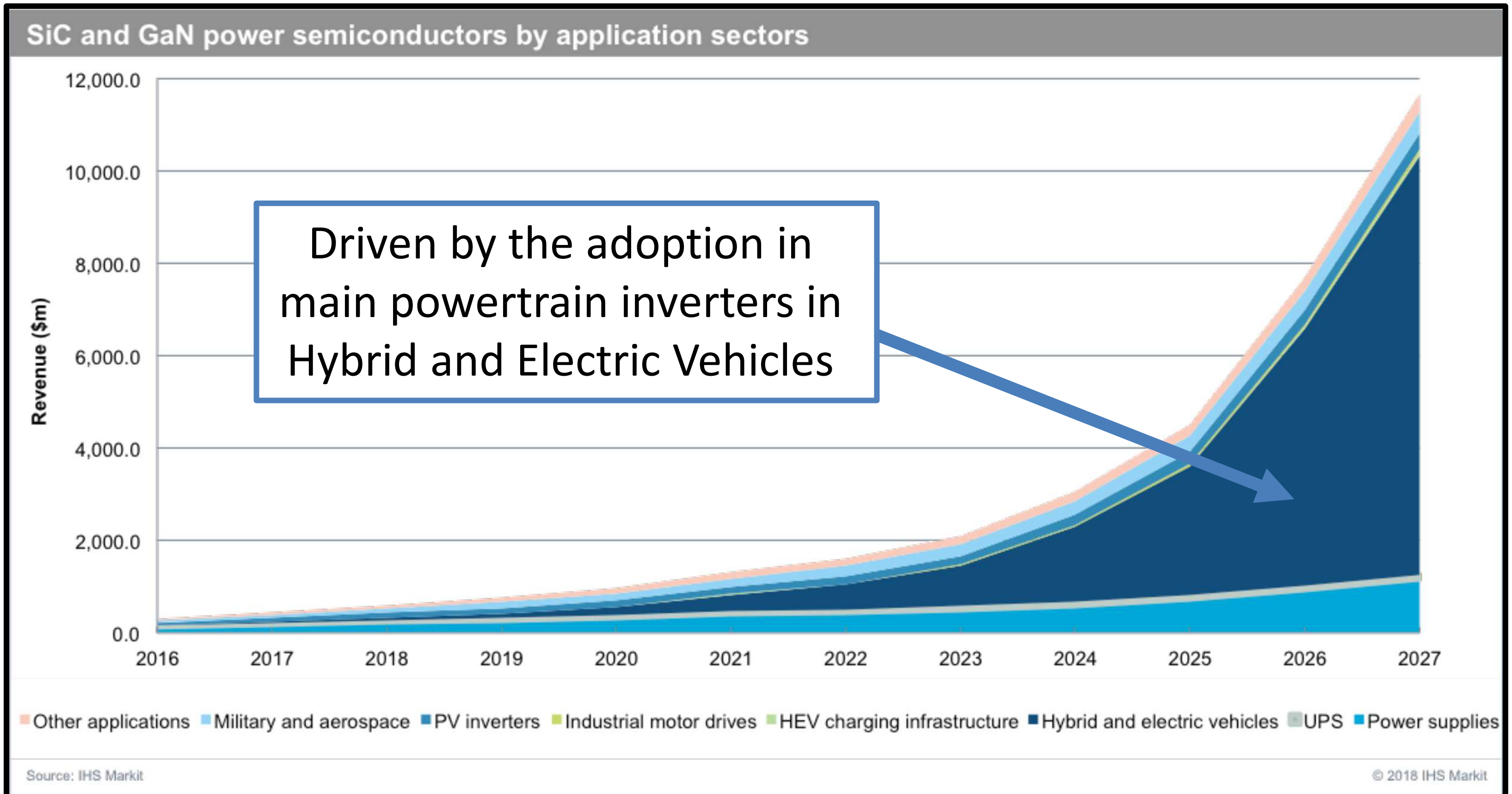
- Market Motivation
- Measurement of high frequency power loss.
- Leveraging Existing Materials in Emerging Application
- New Material Development
- The advantages of ferrite materials in power supply designs
- Important parameters of ferrite materials and their impact on performance, such as:
  - Permeability
  - Performance factor
  - Power Loss characteristics
- Guidelines to selecting the appropriate ferrite material
  - Operating conditions
  - Environmental factors
- Optimizing core configuration based on design limitations
  - Effects of size and geometry
  - Effects of DC and air gaps

Market Motivation

- Miniaturization is a driving force in electronics design.
  - Magnetics are typically the largest component in power supplies.
  - In order to minimize power supply footprints, operating frequency has been increasing.
    - Power loss of magnetic components incorporated into these designs can cause issues with efficiency and heat management
- Increased efficiency
  - GaN has no reverse recovery charge which enables topologies such as the totem-pole PFC to improve efficiency

Markets for GaN and SiC

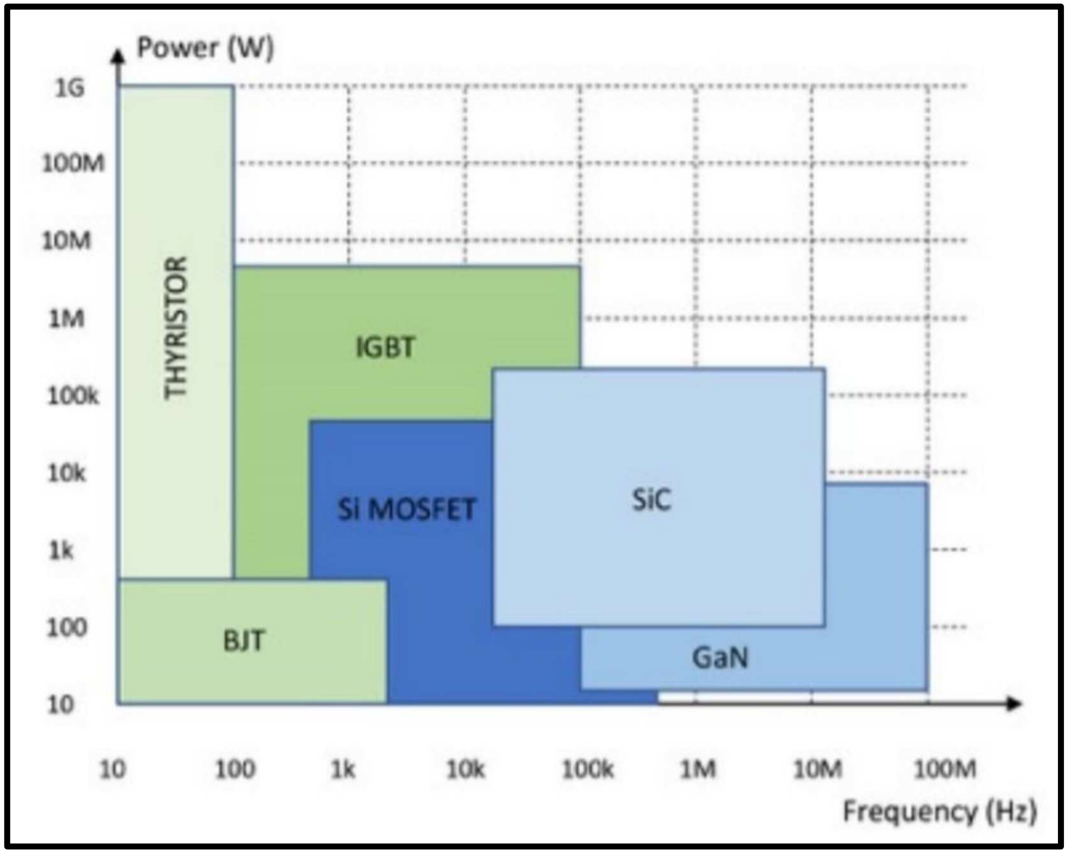
- GaN + SiC likely to reach ~$10B by 2027[1]
  - GaN is expected to achieve price parity with Si MOSFETS and IGBTs by 2020, leading to a CAGR = 21.5% from 2017 to 2025[2]



IHS Markit | Technology[1]

Increasing Operating Frequencies

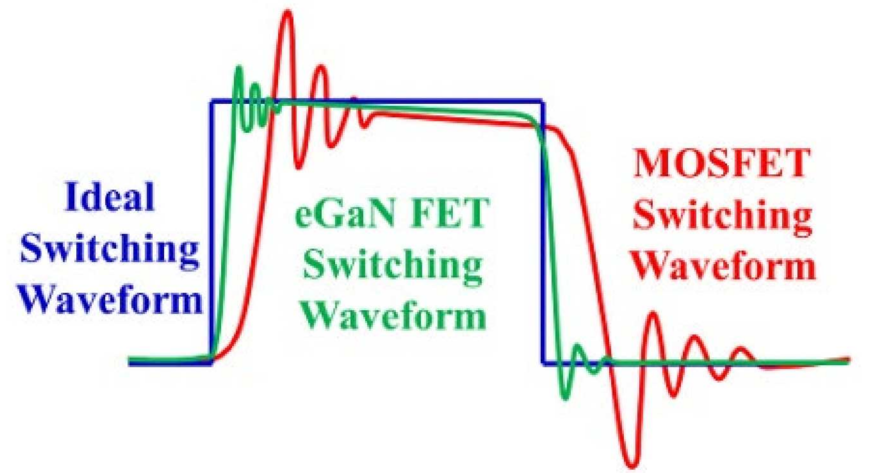
- EPC: "Cutting new ground for power transistors, these devices have switching transition speeds in the sub nano-second range, making them capable of hard switching applications **above 10 MHz**."[1]
- GaN Systems: "We have many customers using our devices from **hundreds of kHz, to 13.56MHz, and even some above 50MHz**."[2]
- Cree: "Each gate drive circuit … can comfortably switch the SiC MOSFETs at **up to 3MHz.**"[3]



Possible future scenario presented by UnitedSiC

Semiconductor Development

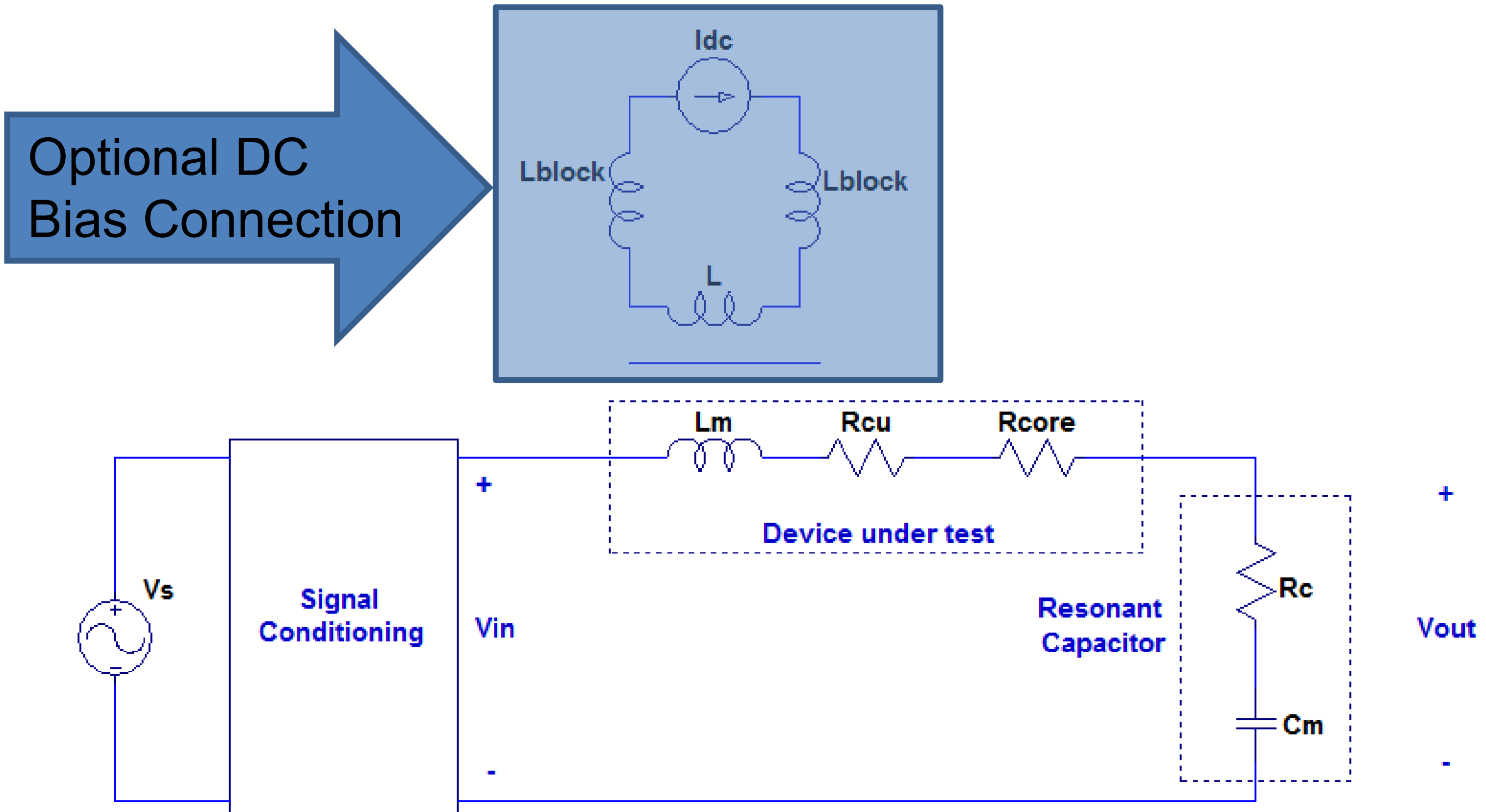
- "The rapid progress in GaN and SiC power semiconductors will lead to a further miniaturization of power electronic assemblies and subsystem…
  - **The drastically increased frequency requires improved ferrite materials with lowest losses."**



Jungwirth,, H., Schmidhuber, M., Baumann, M., Schmeller, M.
"A new high frequency ferrite material for GaN applications", PCIM Europe 2016.
Lidow, Alex. "How to GaN: eGaN® FETS in High Performance Class-D Audio Amplifiers." EE Web. February 19. 2014.

Measurement Setup for Power Loss

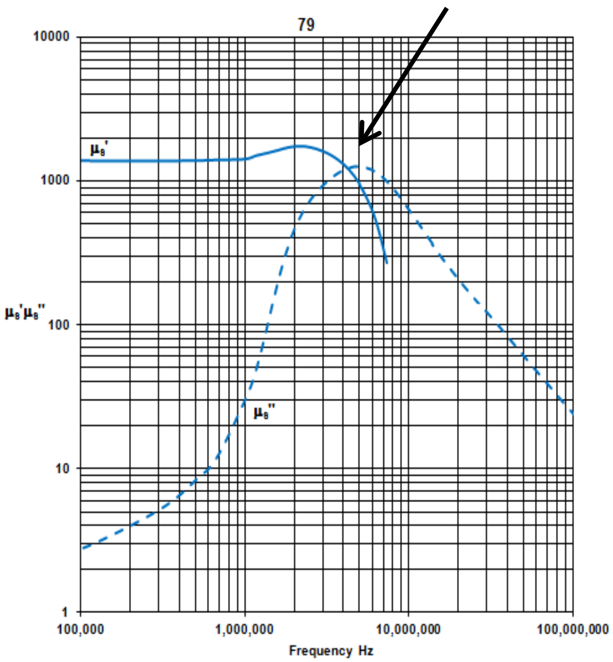
- Fair-Rite utilizes the "resonant Q" method developed by MIT to conduct measurements.
  - This system has been replicated at Fair-Rite with MIT's assistance.
- This method removes the reliance on phase angle as part of the measurement.



(1) Han, Y; Cheung, G; Li, A; Sullivan, C.R.; Perreault, D.J.; "Evaluation of Magnetic Materials for Very High Frequency Power Applications" in Power Electronics, IEEE Transactions on , vol. 27, no.1, pp.425-435

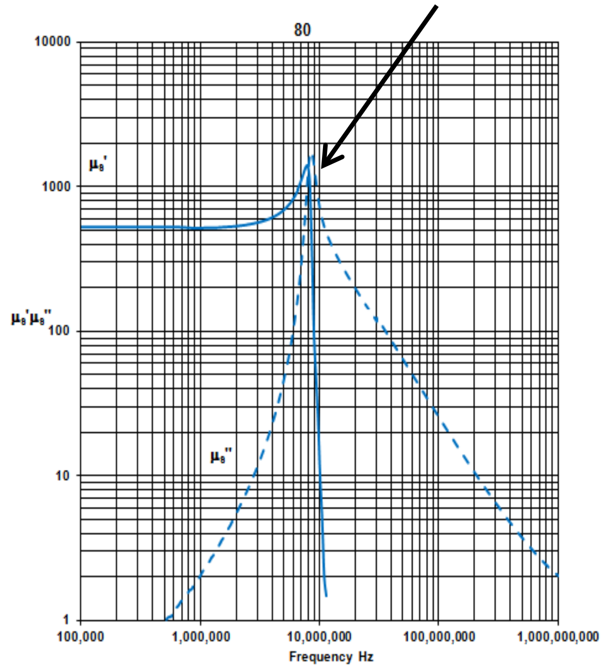
79 material

cut-off frequency 4MHz



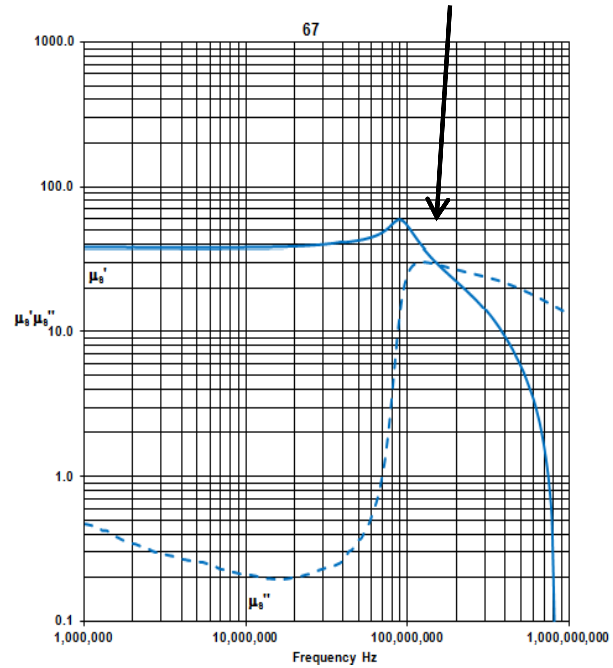
80 material

cut-off frequency 8MHz



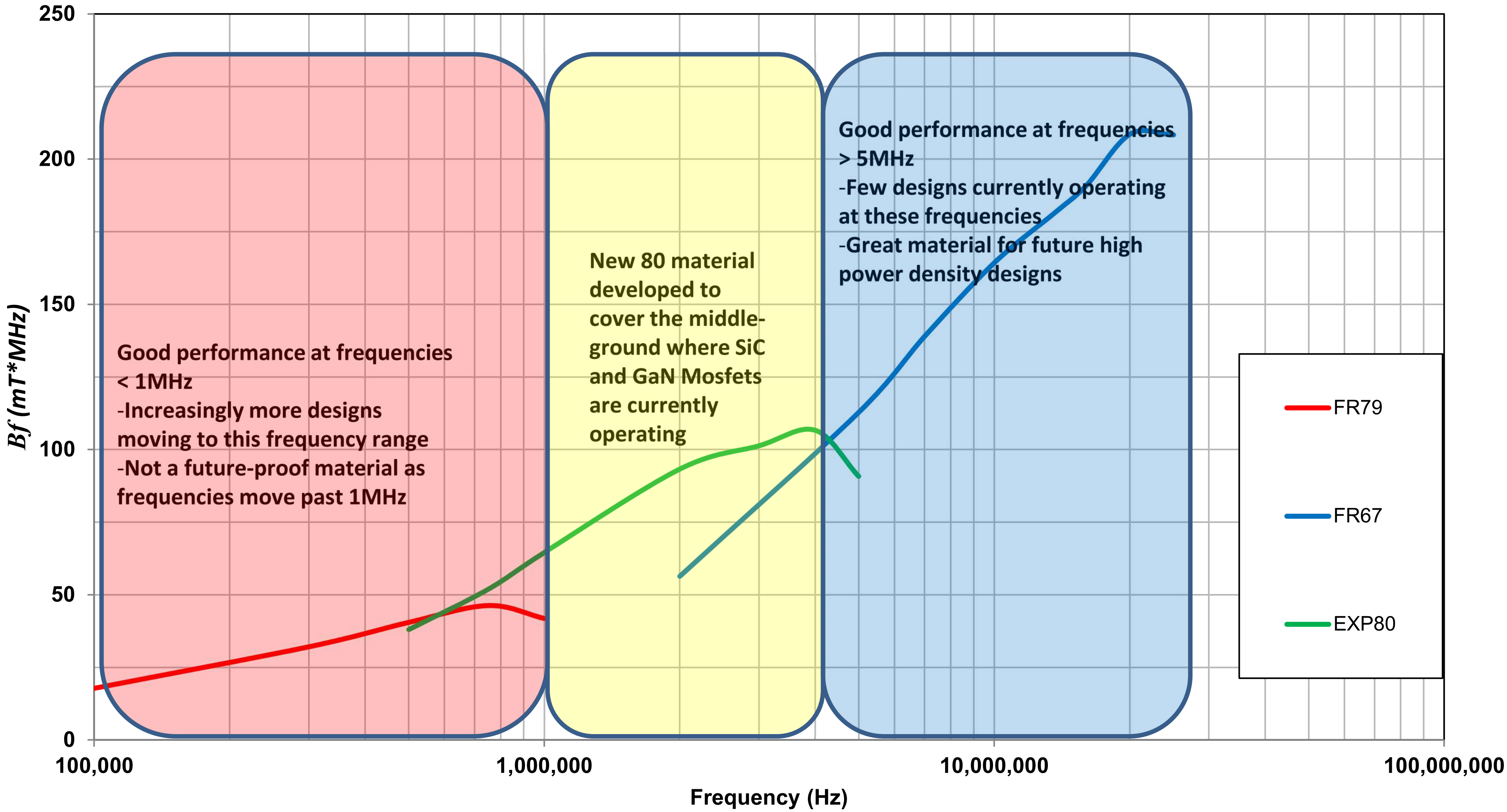
67 material

cut-off frequency 150MHz



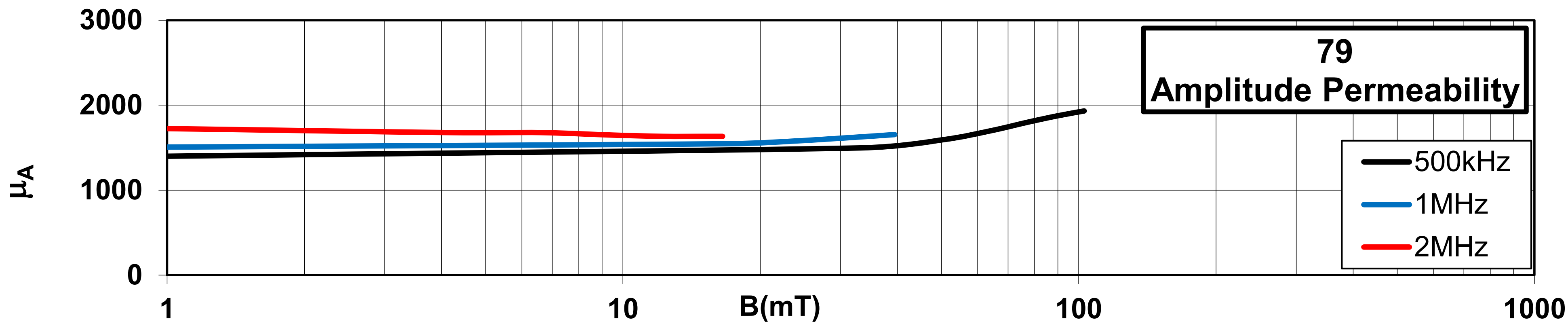
- Loss factor is the principal loss parameter at low flux density (u''/u'^{2})
- Typically lowest loss factor will have lowest core loss
- Higher cut-off frequency typically means lower core loss to higher frequencies

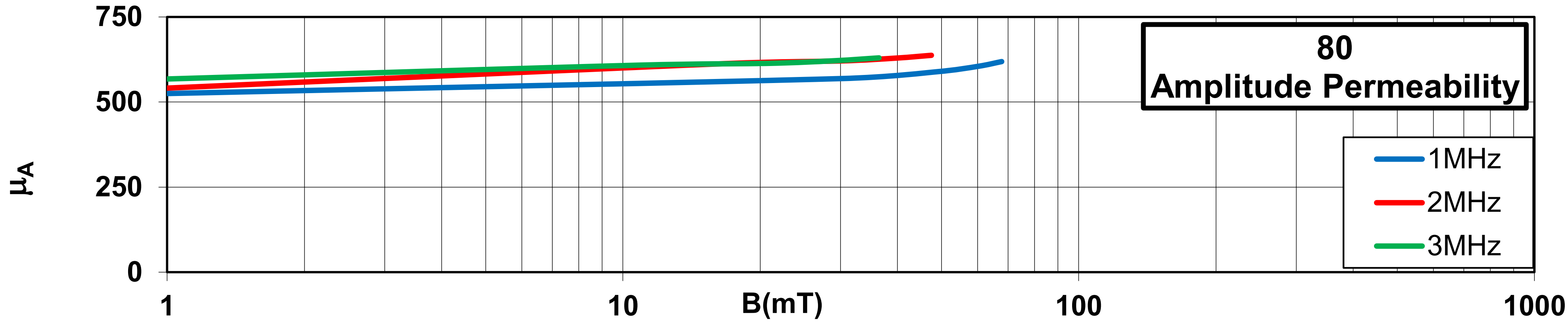
Performance Factor(500mW/cc) for Fair-Rite's Materials

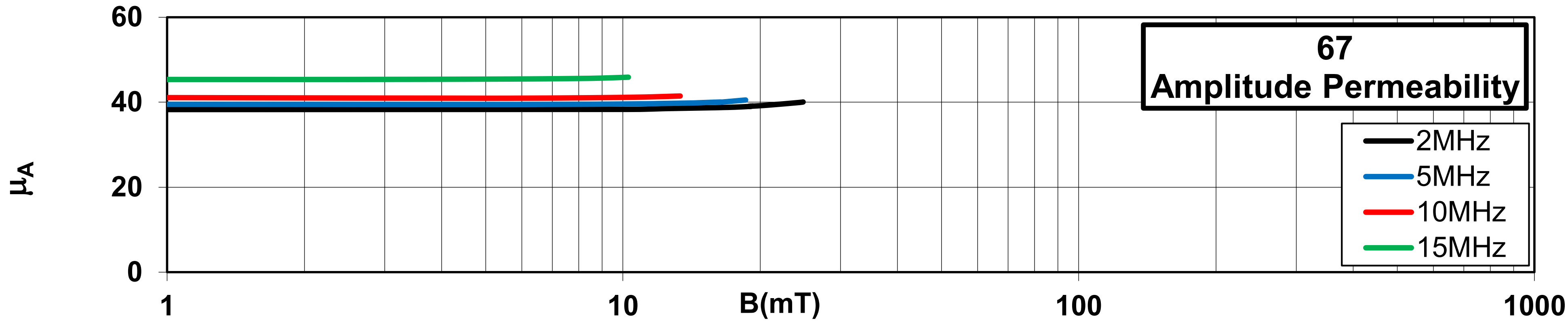


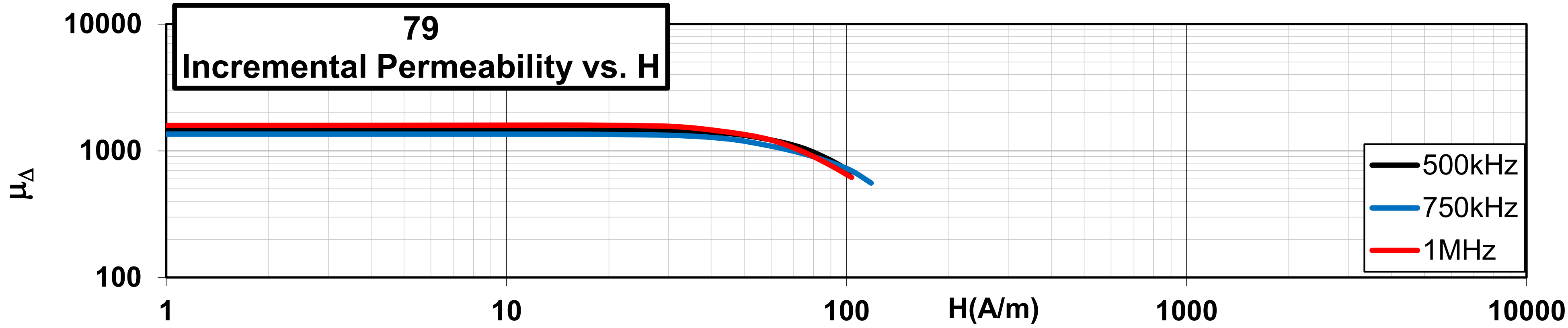
Considerations for selecting appropriate material and core configuration

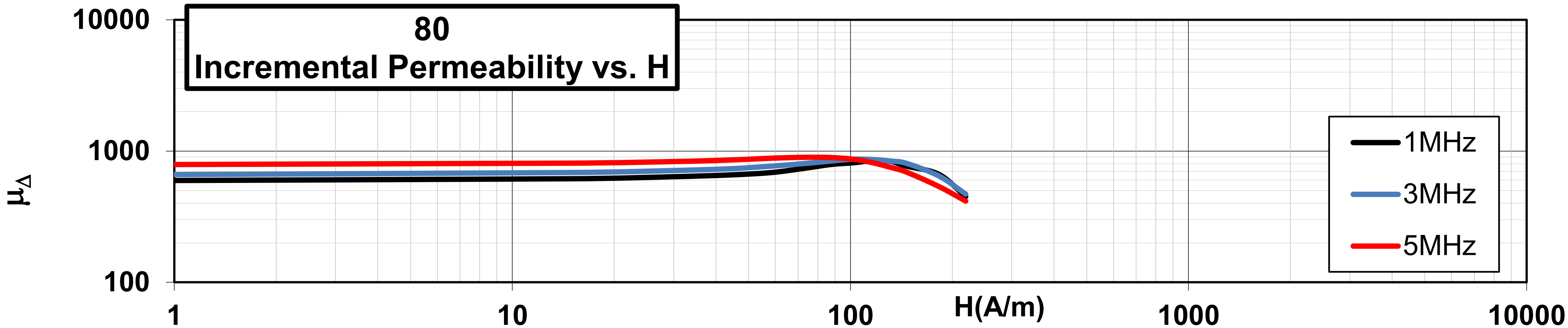
- Permeability with:
  - Elevated flux densities
  - DC bias over frequency
- Power loss density
  - At operating frequency
  - Over temperature
- Optimal core size considering operating frequency.
- Geometry selection based on design requirements.
- Effects of DC bias on performance.
- Impacts of adding an air gap.

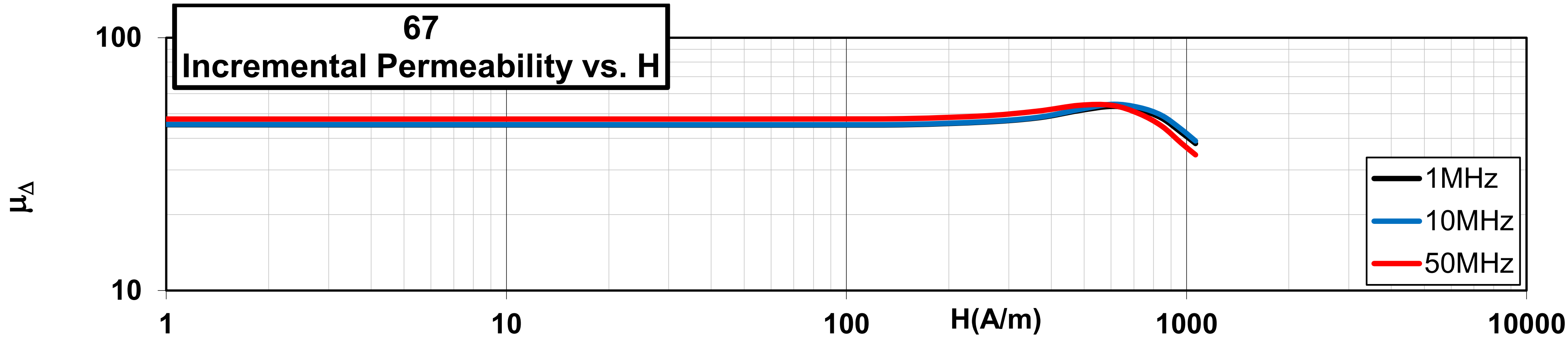

79
Amplitude Permeability
μA
3000
2000
1000
0
1
10
B(mT)
100
1000
500kHz
1MHz
2MHz


80
Amplitude Permeability
μA
750
500
250
0
1
10
B(mT)
100
1000
1MHz
2MHz
3MHz

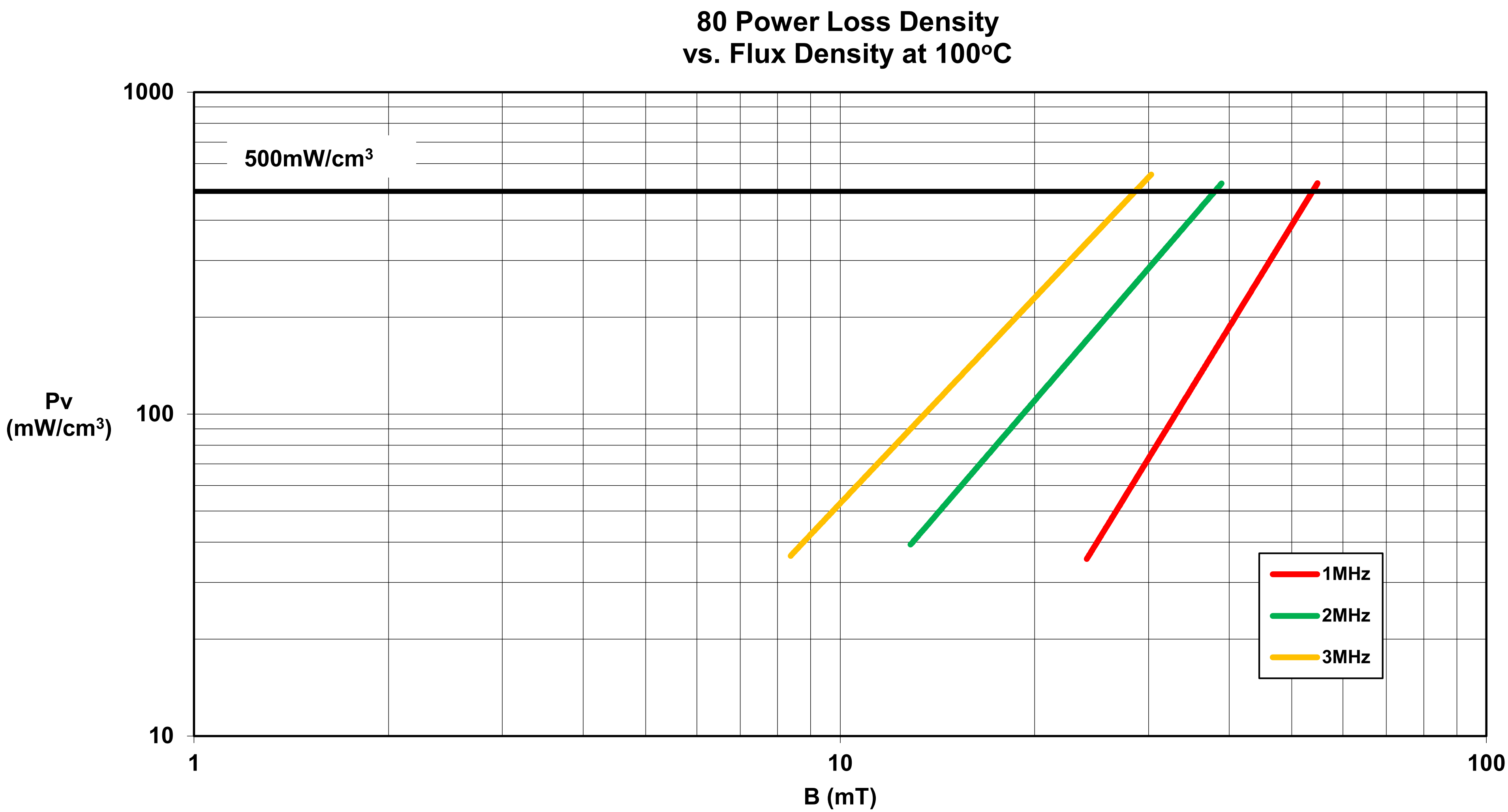

67
Amplitude Permeability
μA
60
40
20
0
1
10
B(mT)
100
1000
2MHz
5MHz
10MHz
15MHz


79
Incremental Permeability vs. H
10000
1000
100
μΔ
1
10
100
H(A/m)
1000
10000
500kHz
750kHz
1MHz


80
Incremental Permeability vs. H
10000
1000
100
μΔ
1
10
100
H(A/m)
1000
10000
1MHz
3MHz
5MHz

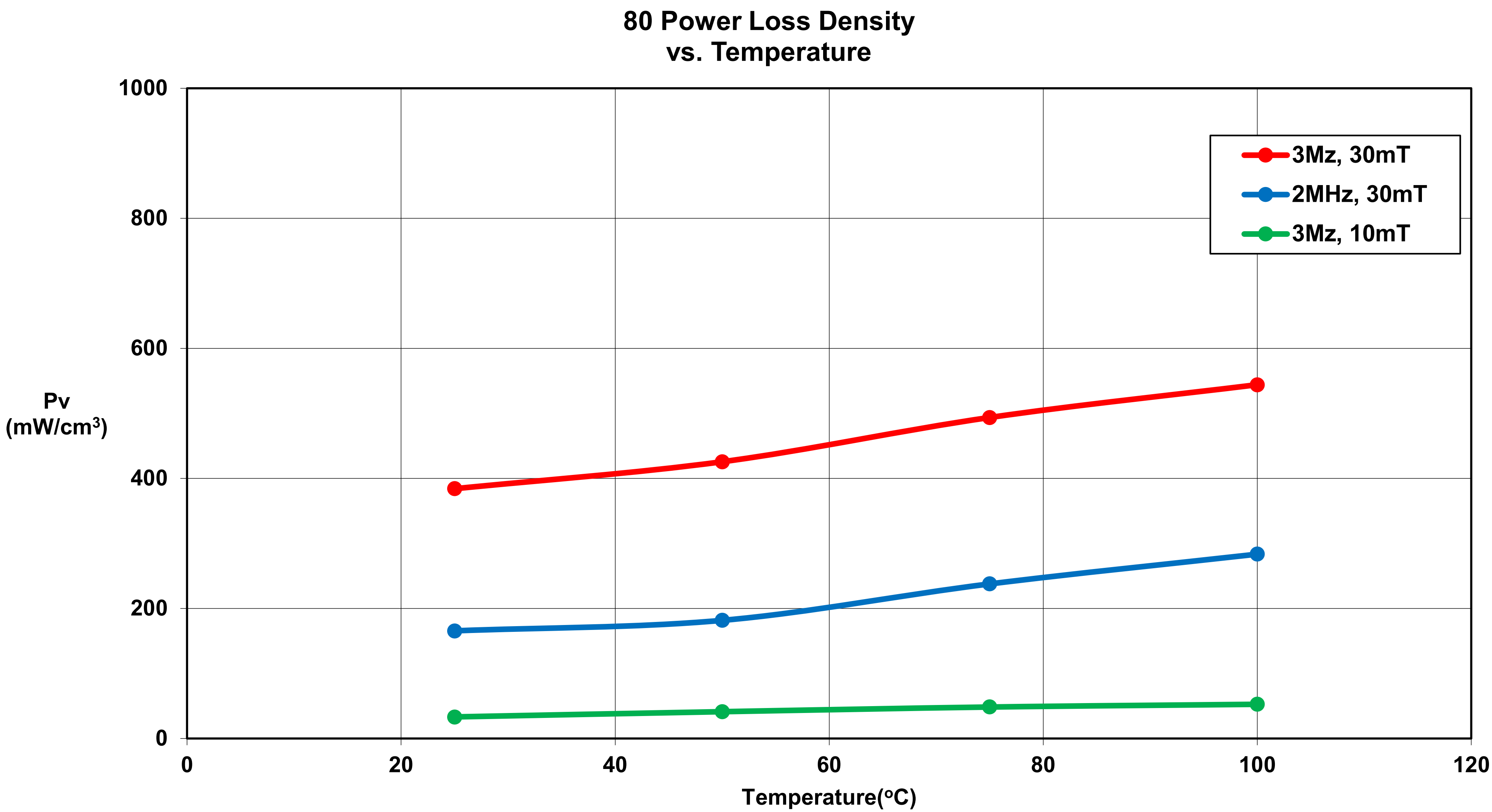

67
Incremental Permeability vs. H
100
10
μΔ
1
10
100
H(A/m)
1000
10000
1MHz
10MHz
50MHz

HF Power Loss Curves @ 100°C

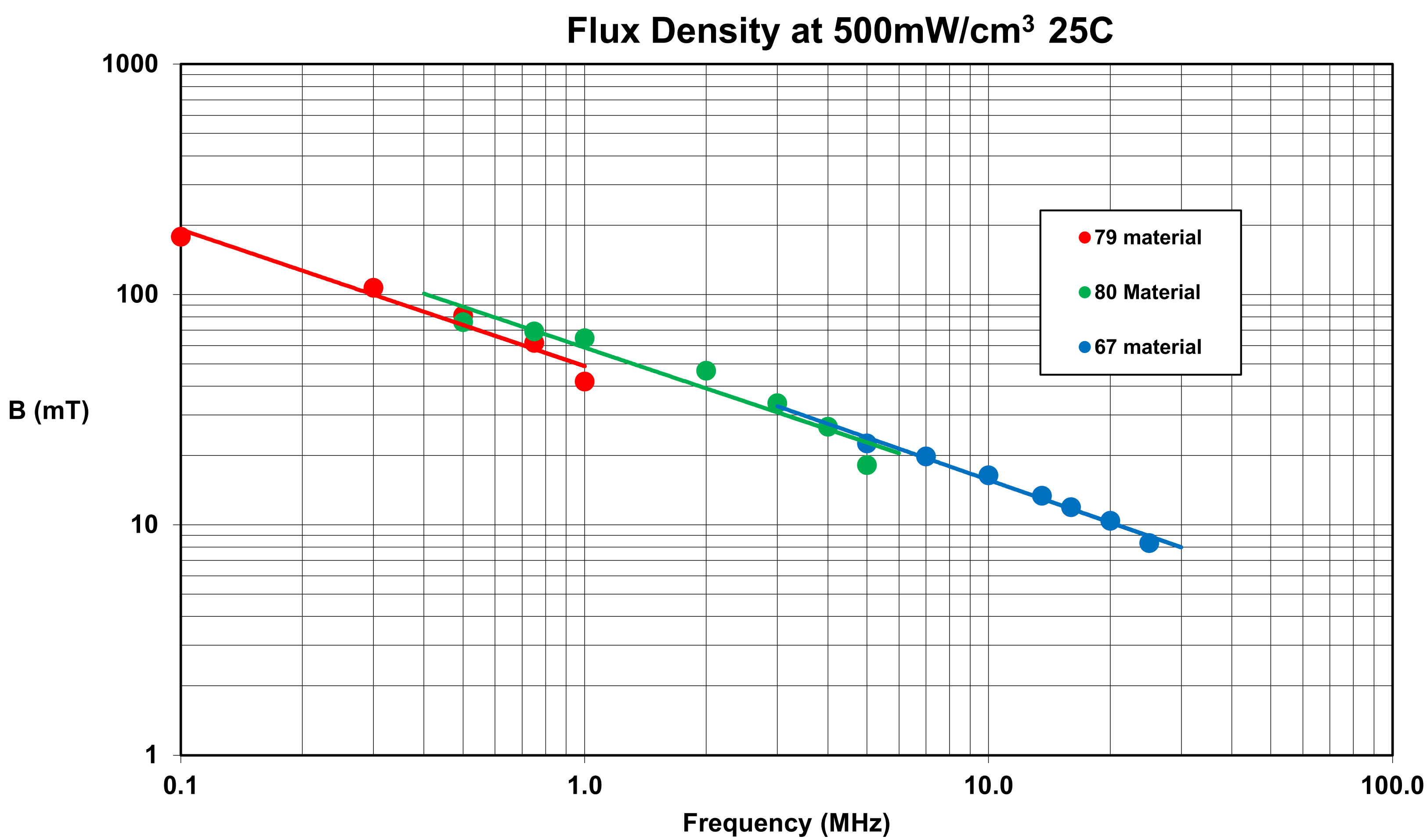


Measured on a 22.1mm/13.7mm/6.35mm toroid at100° C.

Power Loss vs. Temperature



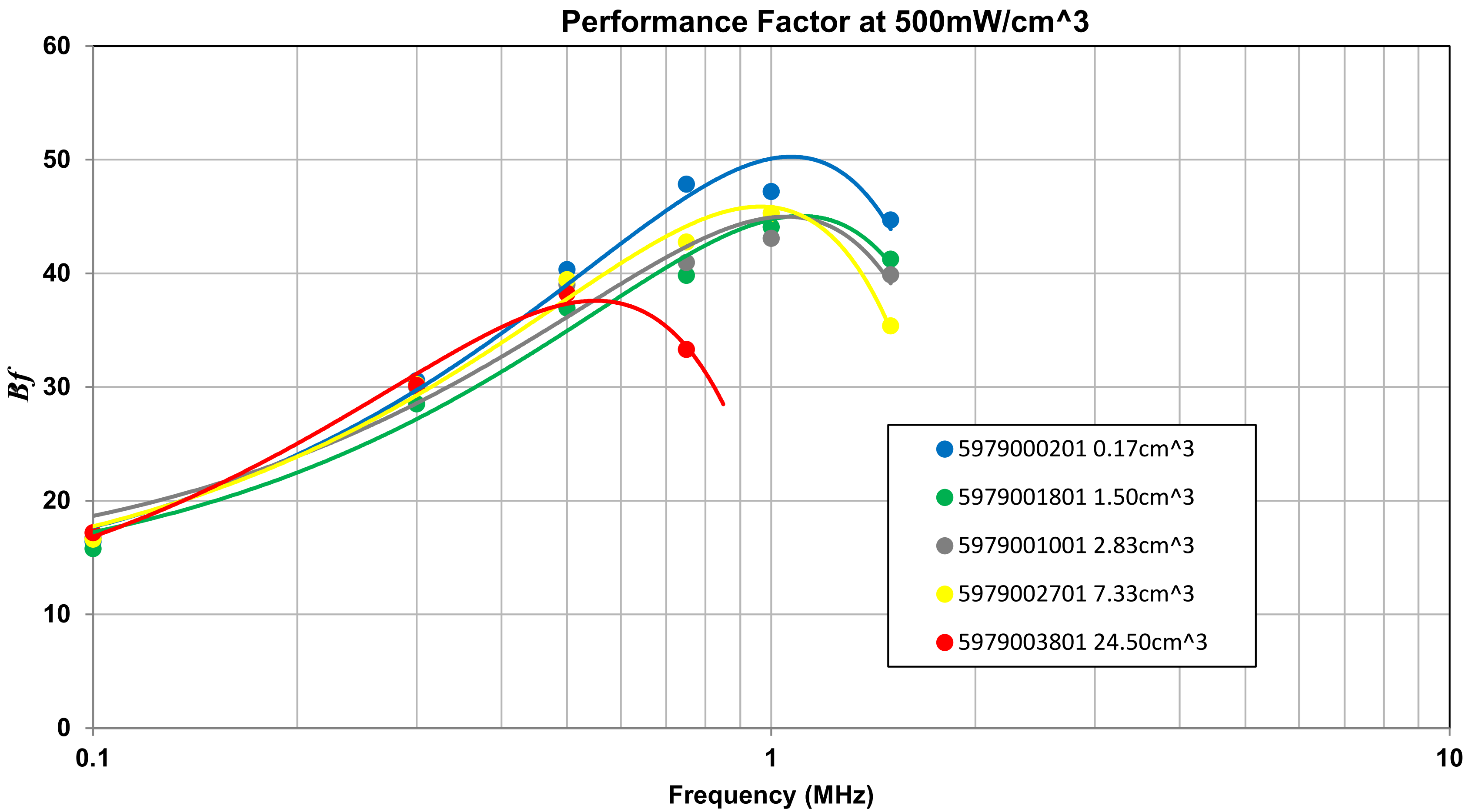
Measured on a 22.1mm/13.7mm/6.35mm toroid .



Measured on a 12.7mm/7.9mm/6.35mm toroid at 25° C.

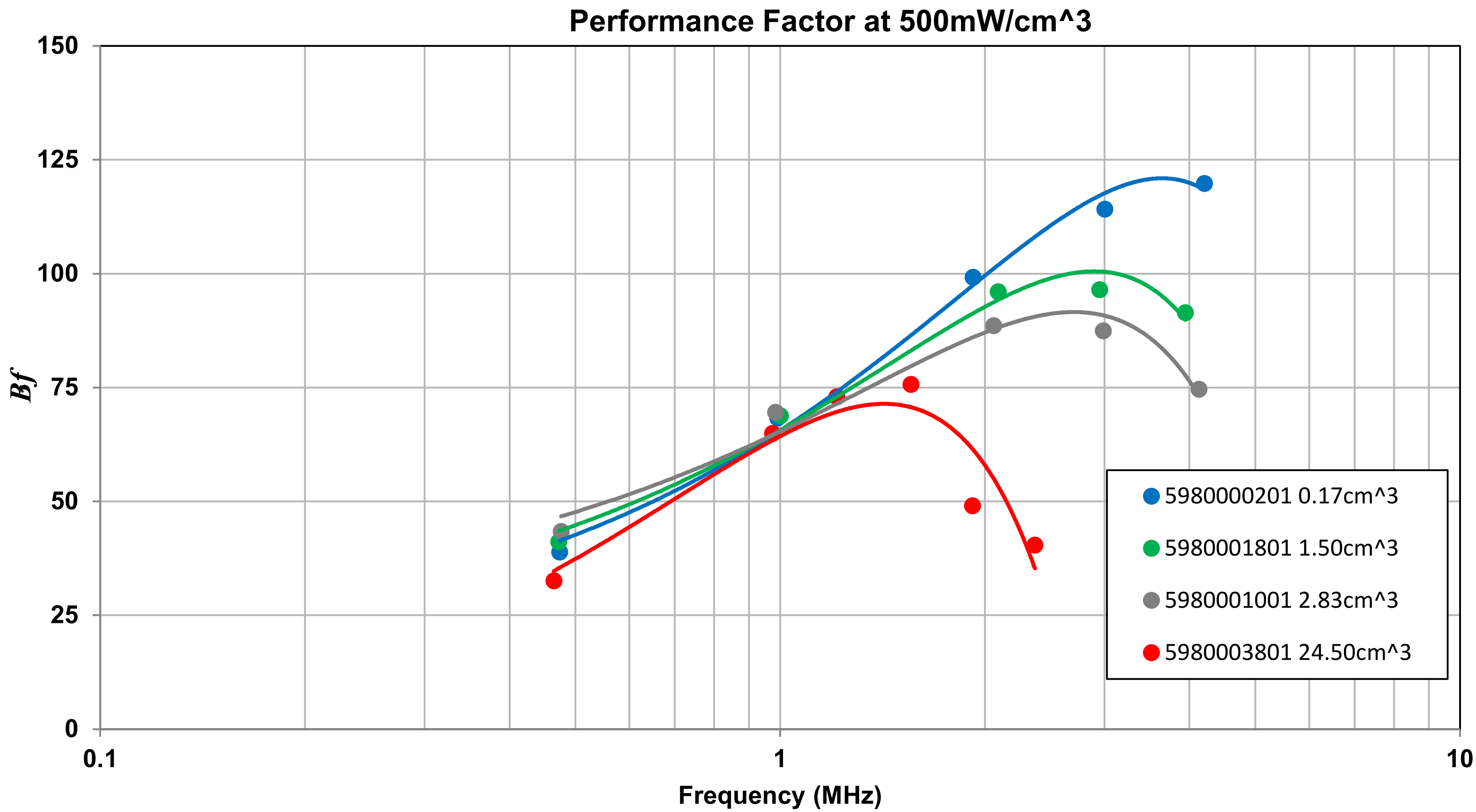
- For Power Conversion purposes operation should be limited to 500 mW/cc.
- For more conservative limiting of heat rise: 300 mW/cc is a better design constraint.
- At higher frequencies: operation is "loss limited" as opposed to flux density limited at low frequency.

Performance Factor Curves for different size 79 material toroids



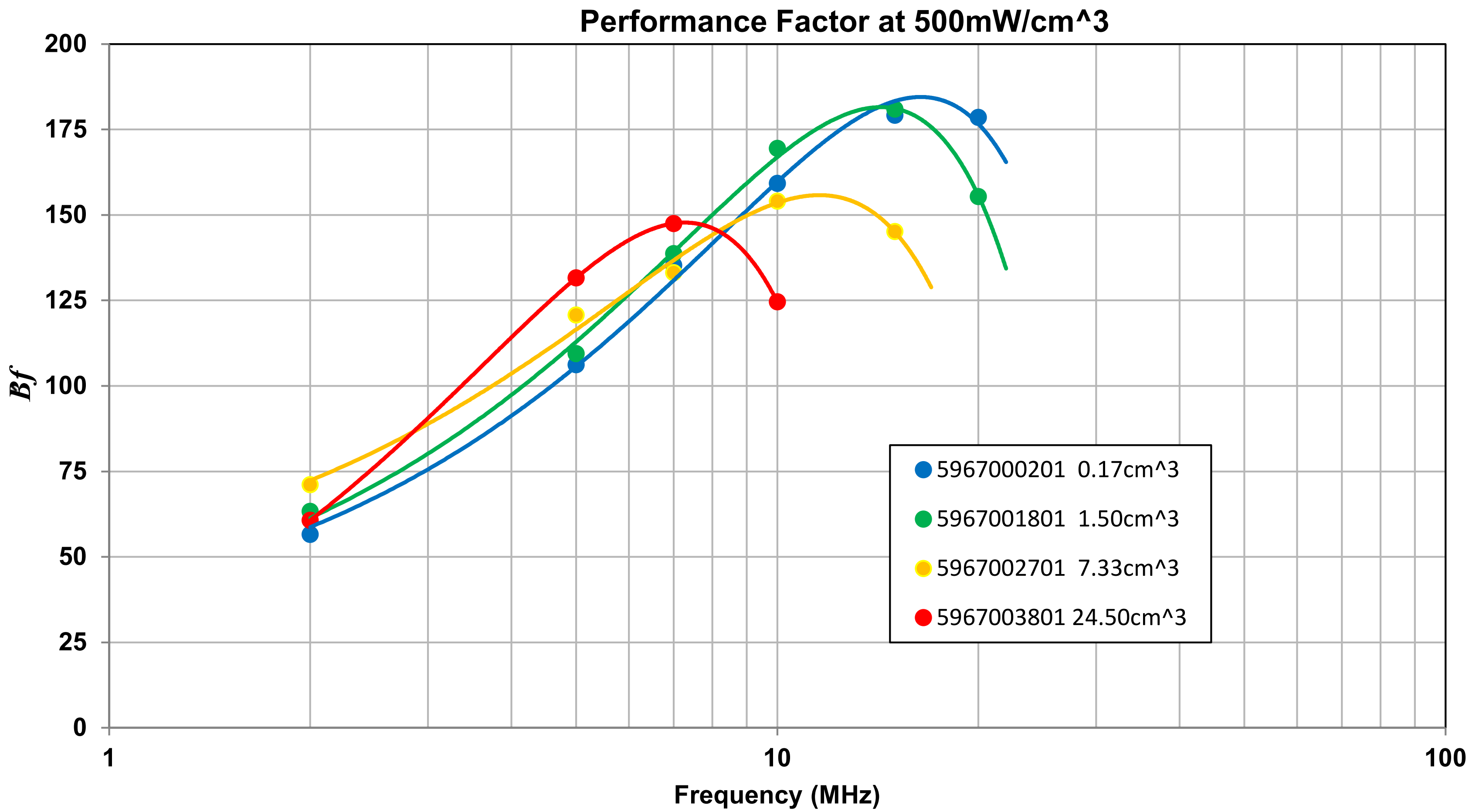
- Optimal operating frequency decreases as core size increases

Performance Factor Curves for different size 80 material toroids



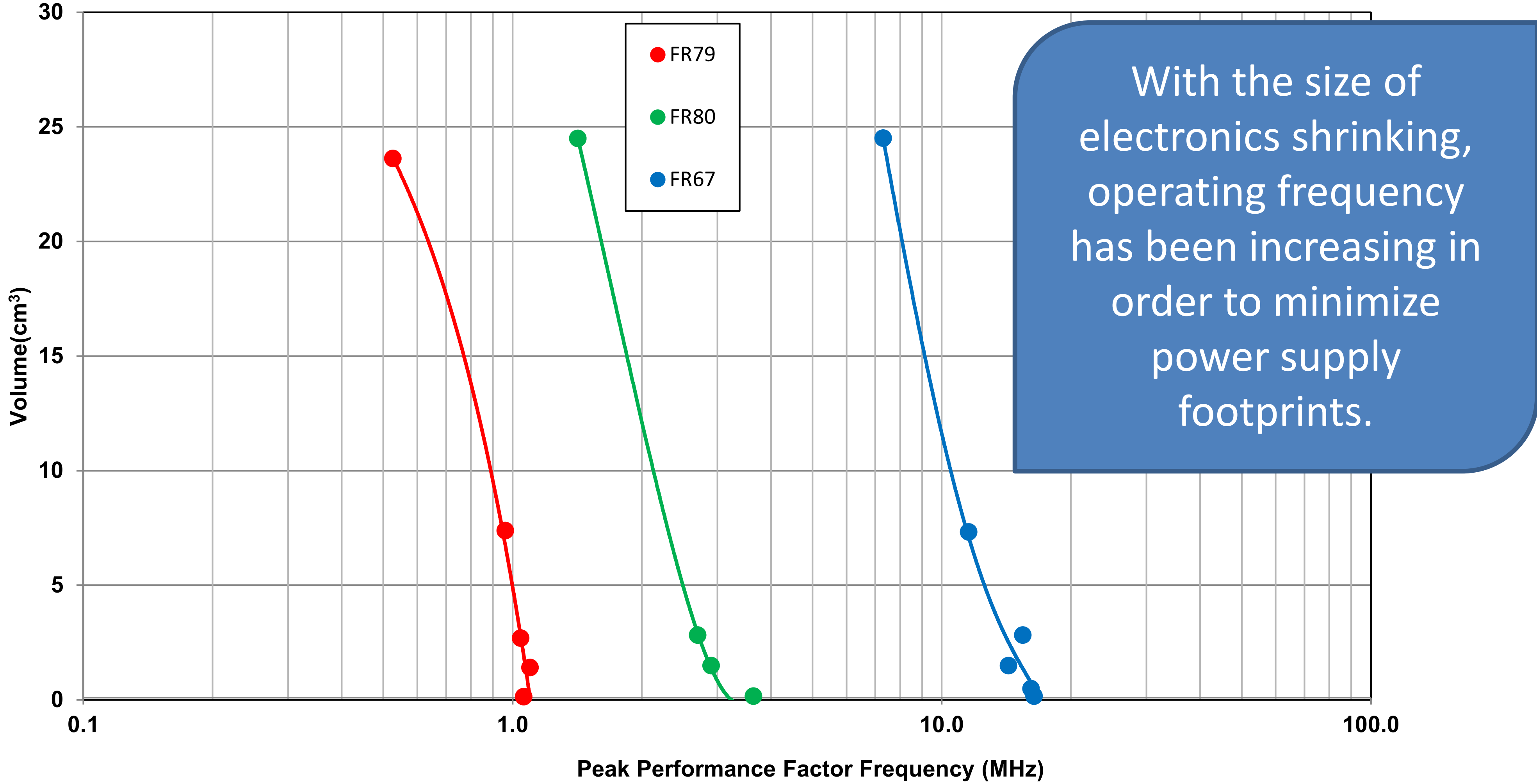
- Optimal operating frequency decreases as core size increases

Performance Factor Curves for different size 67 material toroids



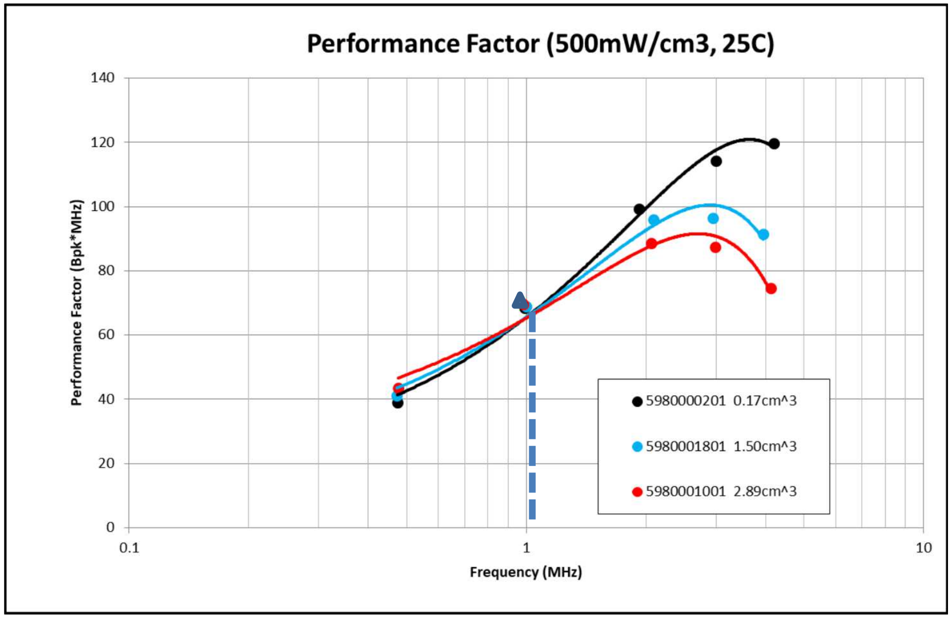
- Optimal operating frequency decreases as core size increases

Peak Performance Factor Frequency for Toroids



- Want to operate where performance factor is highest, but smaller cores cannot handle as high a power level.

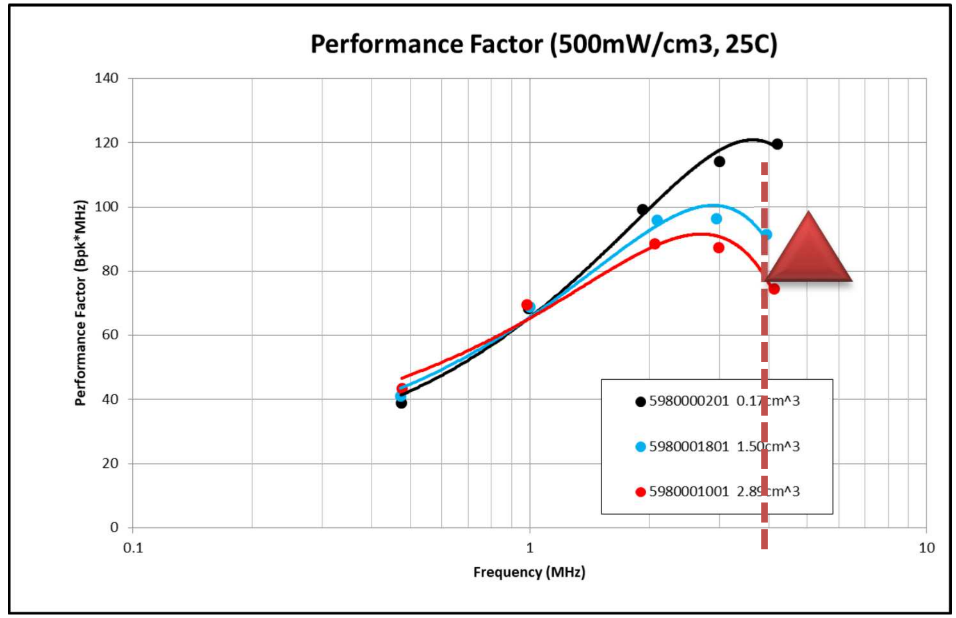
Comparison #1



Max Temperature	
46.3°C	53.3°C

Maximum Temperature Difference = 7°C.

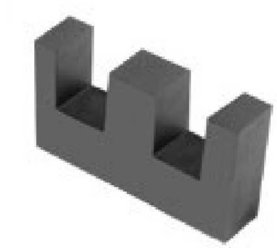
Comparison #2



Max Temperature	
50.3°C	77.1°C

Maximum Temperature Difference = 26.8°C.

Considerations for Core Geometry Selection



- Power handling limited by temperature rise due to:
  - Copper losses of the windings
  - Power loss of the core
- Core size limited by available space on board
  - Low profile vs. large volume
- Magnetic shielding to limit EMI
- Cost and manufacturability

Desired Geometry Characteristics

Feature	Relation	Desired Ratio
inductance / length of wire	Lo/MTL	High
power-to-volume density	AeAw/FPtotal	High
uniform cross section	Amin/Ae	High
magnetic shielding	Exposed Winding Area/Footprint	Low
heat dissipation	Exposed Winding Area/Footprint	High

Case study

- *Geometries with largest physical dimension [A] in range of 20-30 mm and throughput power of 30-60 watts in traditional designs @ 100 kHz, < 200 mT*

Comparison of the Different Geometries

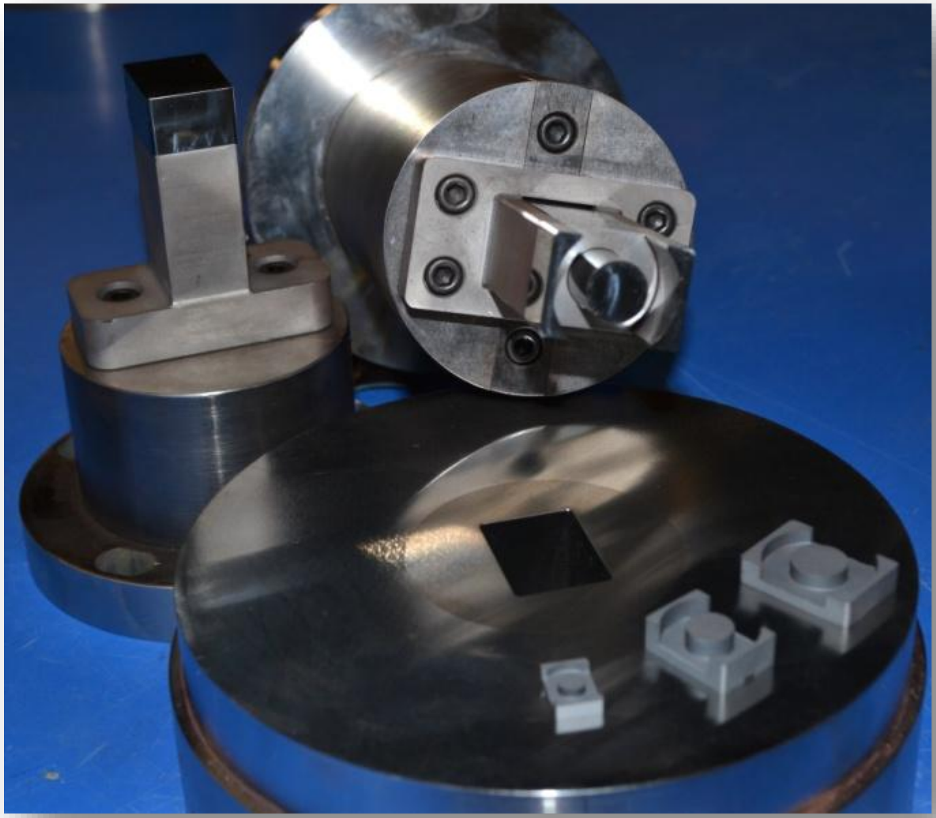
feature	T	std UU	std EE	P	EP	EFD	PQ	RM	ETD/ EER	pEE	pE I	pEER/ pEEQ	pERI/ pEQI
inductance/wire length	2	1	1	3	4	1	2	2	1	4	5	3	4
power to volume density	5	5	5	2	2	3	3	3	4	3	2	3	2
off board height	3	1	2	3	1	4	2	2	2	4	5	3	5
round center post	5	1	1	5	5	1	5	5	5	1	1	5	5
uniform cross section	5	5	5	3	3	5	5	4	5	5	5	4	4
magnetic shielding	4	1	1	5	5	1	3	4	1	2	2	3	3
heat dissipation	4	5	5	1	1	4	4	2	5	4	3	4	3
core standardization	no	no	IEC	IEC	IEC	IEC	IEC	IEC	IEC	IEC	IEC	de facto	de facto
totals :	28	19	20	22	21	19	24	22	23	23	23	25	26
rating :	0.80	0.54	0.57	0.63	0.60	0.54	0.69	0.63	0.66	0.66	0.66	0.71	0.74

Rating: 1 to 5, where 1 is least and 5 is best.
Note: Small 'p' indicated planar shape.

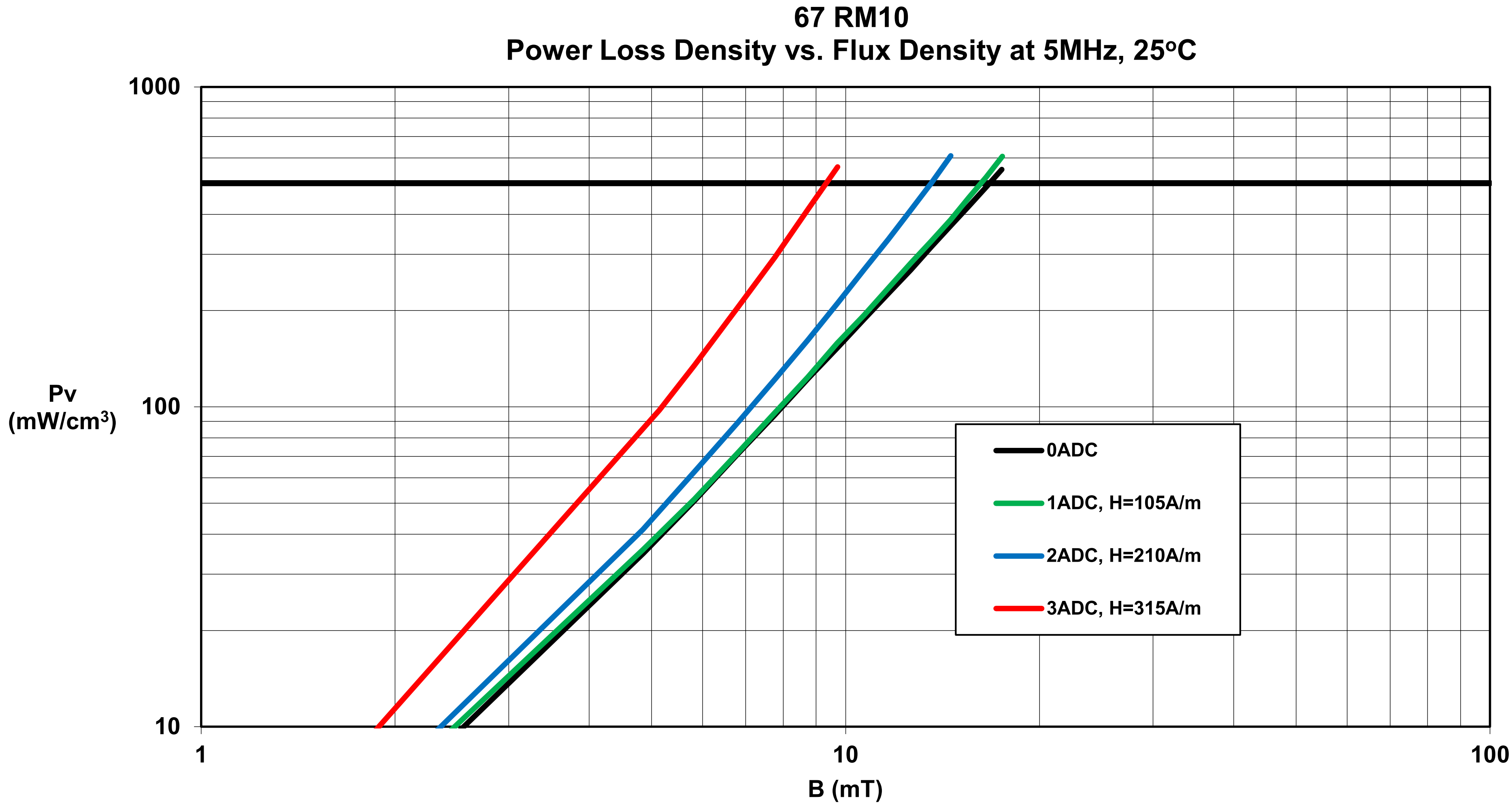
Geometry Conclusion

- Tradeoff of heat dissipation versus magnetic self shielding
- Center posts set back from the outer edge is most desirable for self shielding
- Maximum Lo/MTL is most desirable for lowest AC and DC copper losses
- Low Aw to Ae ratio pushes ratio of overall losses toward the ferrite instead of the winding
- Higher frequency operation requires less winding area because less inductance required

<u>Toroids, Planar ER and EQ designs achieve best tradeoff for high frequency geometries</u>

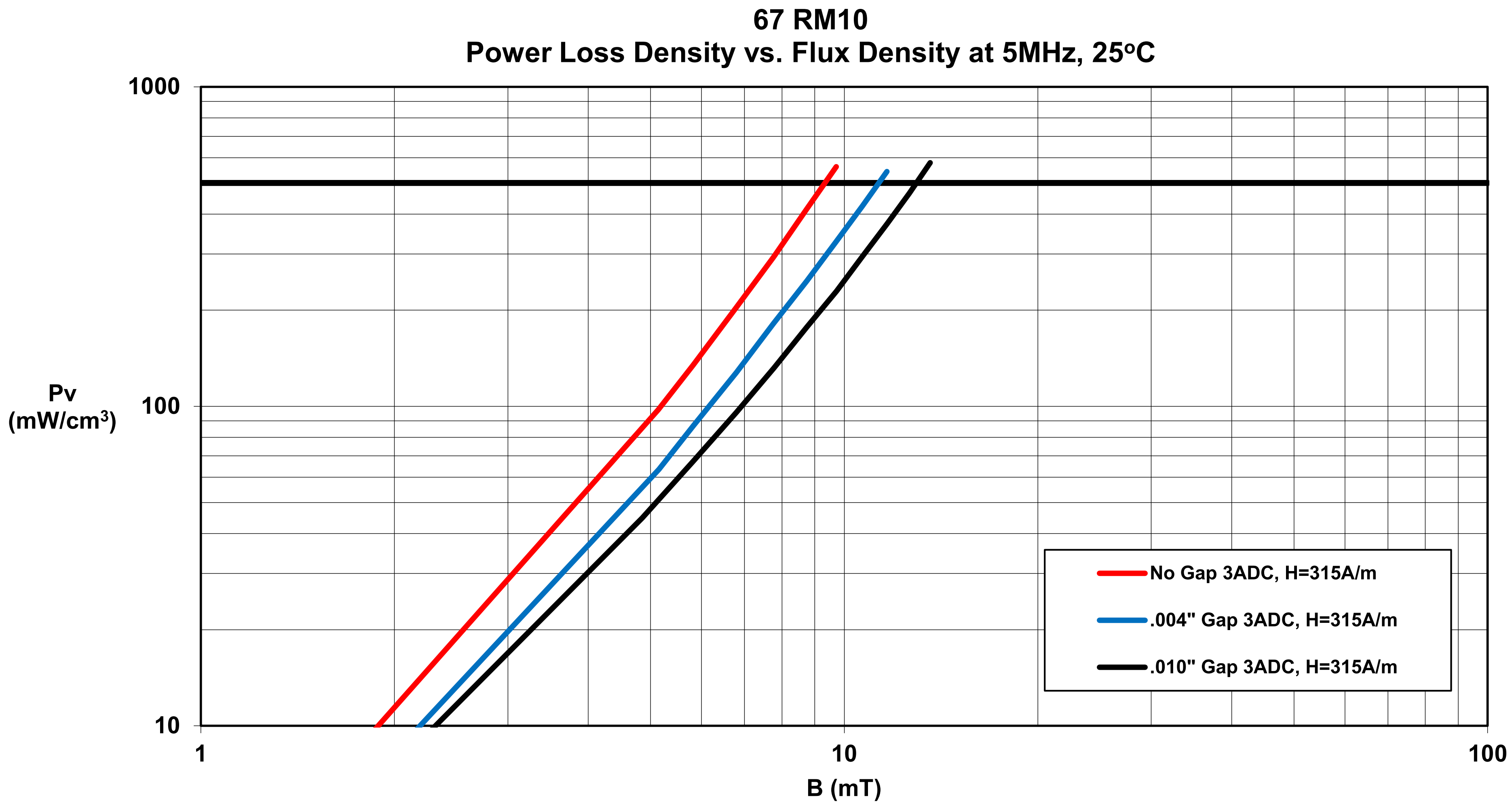


Effects of DC Bias



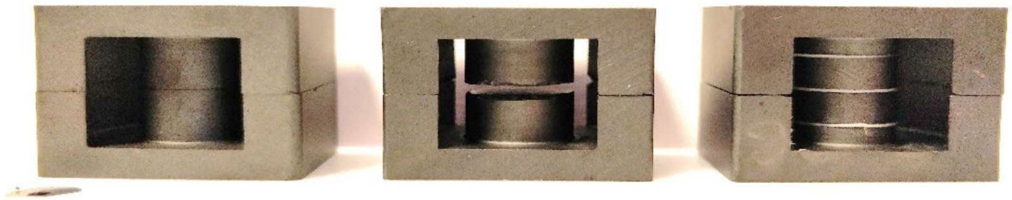
N=5

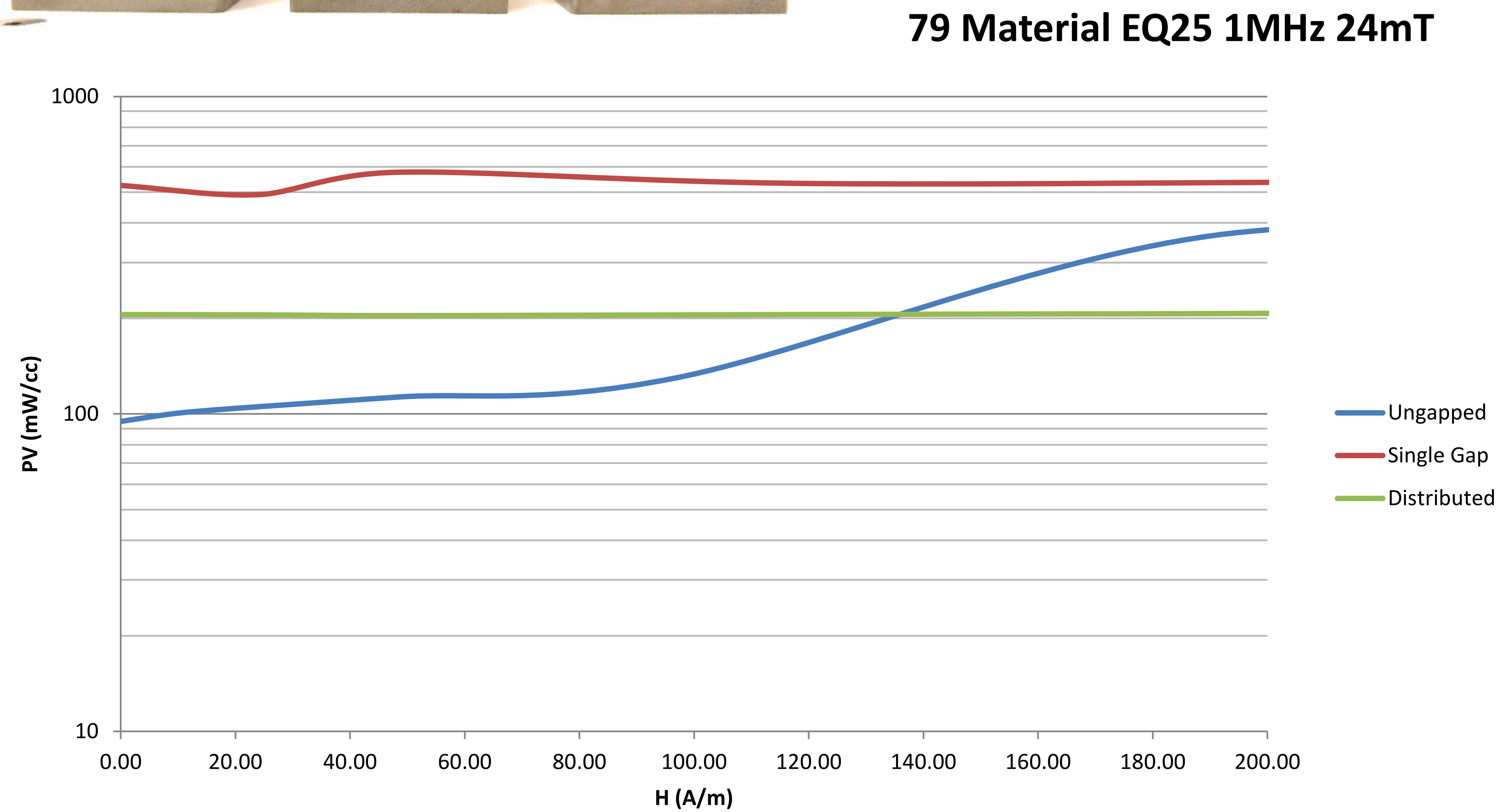
Effects of DC Bias with a Gap



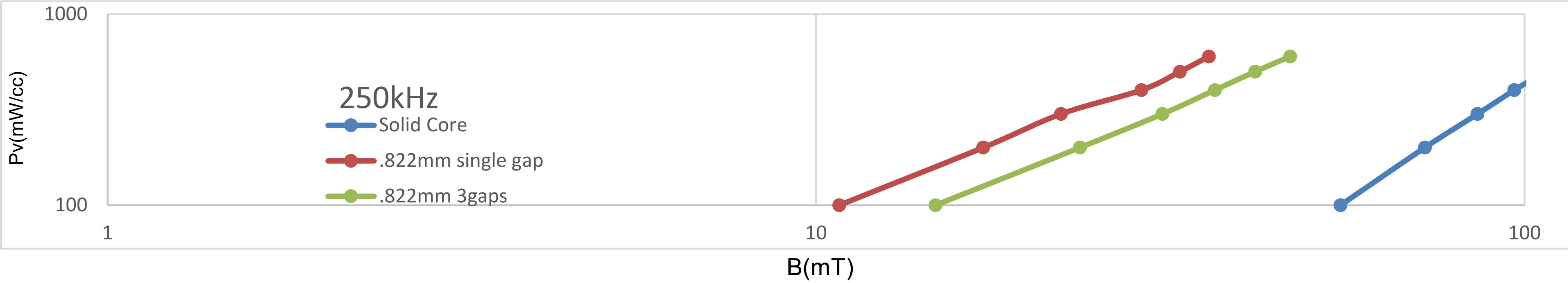
N=5

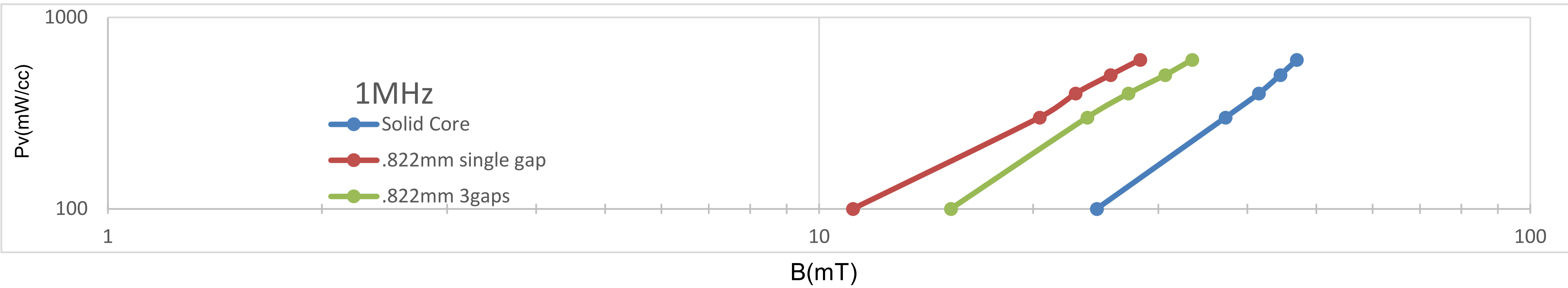
Effect of Bias

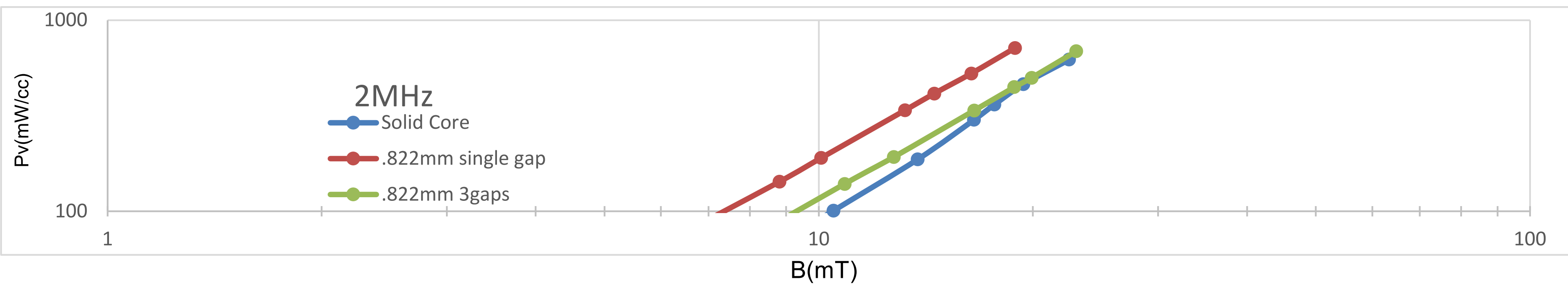




Power Loss 79 EQ25 Gapped vs Ungapped



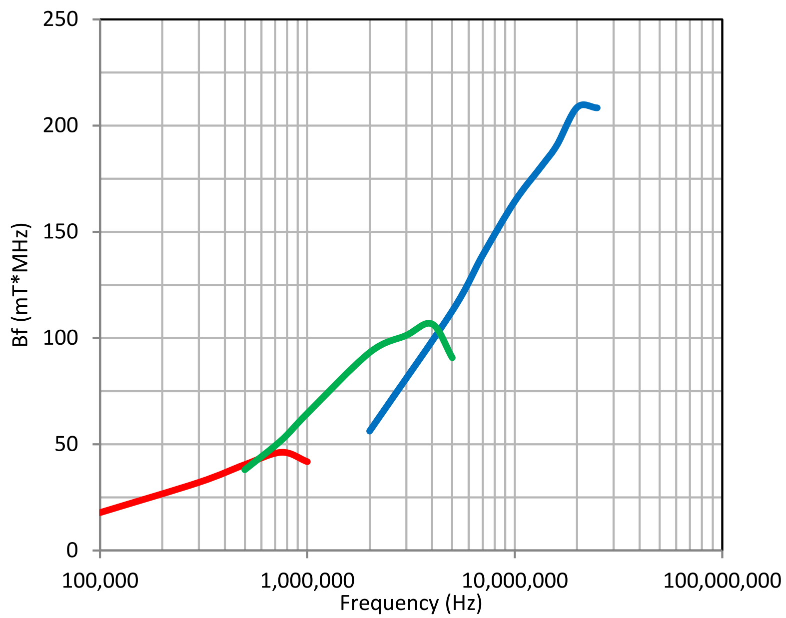




As frequency increases, power loss due to gap becomes less significant

Summary

- Market moving toward higher frequency with smaller core sizes.
- Materials developed to cover the higher frequencies:
  - 79 material (f<1MHz)
  - 80 material (1-5MHz), newly developed
  - 67 material (f>5MHz), optimized
- 79, 80, and 67 offer:
  - Stable permeability with increased flux densities and DC currents over frequency.
  - Low power loss density at higher frequencies covering a range from 500kHz to over 10MHz.
  - Stable power loss densities over temperature.
- Fair-Rite currently offers toroids and EQ cores and continues to add new parts.
- Custom parts and evaluation kits available upon request.



Thank you!

Stay current on our latest developments at Fair-Rite.com!