Core loss measurement techniques to address impact of signal and frequency ranges

Mike Ranjram – Arizona State University
Alex Hanson – University of Texas at Austin

ISO1 Core Loss Measurements for Different Materials and Excitations
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Mike K. Ranjram is an Assistant Professor in Electrical, Computer, and Energy Engineering at Arizona State University.

He received the B.A.Sc. And M.A.Sc. Degrees in electrical engineering from the University of Toronto in 2013 and 2015, respectively, and the Ph. D. degree from the Massachusetts Institute of Technology in 2021. His main research interest is in making power electronics smaller, more efficient, and more capable, and in leveraging these improvements to enable the next generation of sustainable systems and devices. A current focus is on high-frequency power conversion and on techniques for reducing the cost, volume, weight, and loss of power magnetic components.

He is a recipient of the IEEE Transactions on Power Electronics Prize Paper Award.

Arizona State University

Alex J. Hanson is an Assistant Professor and a Fellow of the Jack Kilby/Texas Instruments Endowed Faculty Fellowship in Computer Engineering in the Department of Electrical and Computer Engineering at The University of Texas at Austin.

He received the B.E. degree in electrical engineering from Dartmouth College in 2014 and the S.M. and Ph.D. degrees from the Massachusetts Institute of Technology in 2016 and 2019 respectively. His research interests include component- and system-level power electronics with emphasis on high-frequency magnetics and circuits. He is also interested in leveraging power and energy in diverse applications like medicine, aerospace, etc.

He received the UT Austin ECE Junior Faculty Excellence in Teaching Award in 2022, the NSF CAREER award in 2022, the AFOSR Young Investigator Award in 2022, and the William Portnoy prize paper award from the IAS-PEDCC in 2016.

TEXAS
The University of Texas at Austin

A gapless inductor has imaginary impedance

$$Z = j\omega L$$

Where L depends on the permeability and geometry of the core:

$$L = \frac{N^2}{\mathcal{R}} = \frac{N^2}{\frac{l_c}{\mu_c A}} = \frac{N^2 A \mu'}{l_c} \qquad Z = j\frac{\omega N^2 A \mu'}{l_c}$$

An inductor with *linear* loss has complex impedance

$$Z = j\omega L + R$$

Linear power loss means that

$$P_{loss} \propto I^2$$

which, in an inductor, also means that

$$P_{loss} \propto B^2$$

An inductor with *linear* loss has complex impedance

$$Z = j\omega L + R$$

We can think of this impedance as two parts, an inductive part calculated above, plus an additional resistive part calculated separately. **Alternatively**, we could associate the loss entirely with the inductance and model the system as

$$Z = j\omega L_{complex}$$

Where $L_{complex}$ must have both real (inductive) and imaginary (resistive) parts. In this interpretation, $L_{complex}$ cannot depend on a real permeability μ'; instead, the "permeability" must be thought of as complex

$$L_{complex} = \frac{N^2 A(\mu' - j\mu'')}{l_c} \qquad Z = j\omega \frac{N^2 A\mu'}{l_c} + \omega \frac{N^2 A\mu''}{l_c}$$

Where $\mu \triangleq \mu' - j\mu''$ so that the resulting real resistance is positive. Both μ' and μ'' are functions of frequency, but must be invariant with drive level (otherwise not a linear system!)

This interpretation allows one to treat *linear* core loss as a material property, μ''

$$Z = j\omega L + R = j\omega L_{complex} = j\omega \frac{N^2 A\mu'}{l_c} + \omega \frac{N^2 A\mu''}{l_c}$$

The ratio of the imaginary part to the real part is the Quality Factor:

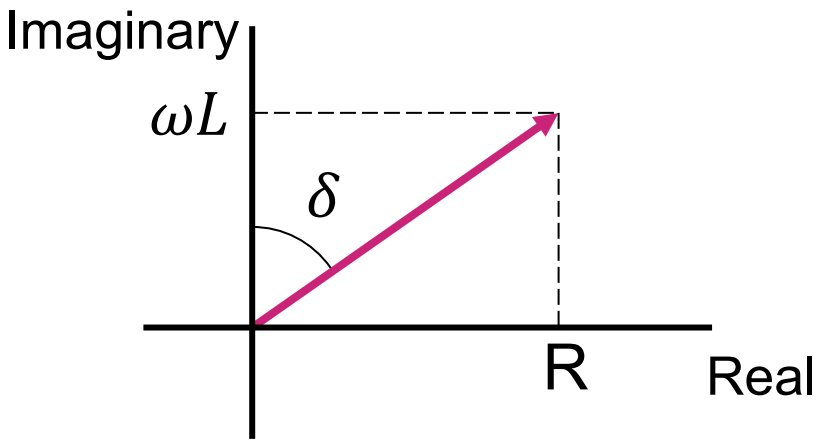
$$Q = \frac{\omega L}{R} = \frac{\mu'}{\mu''}$$

An equivalent expression is the **loss tangent**,

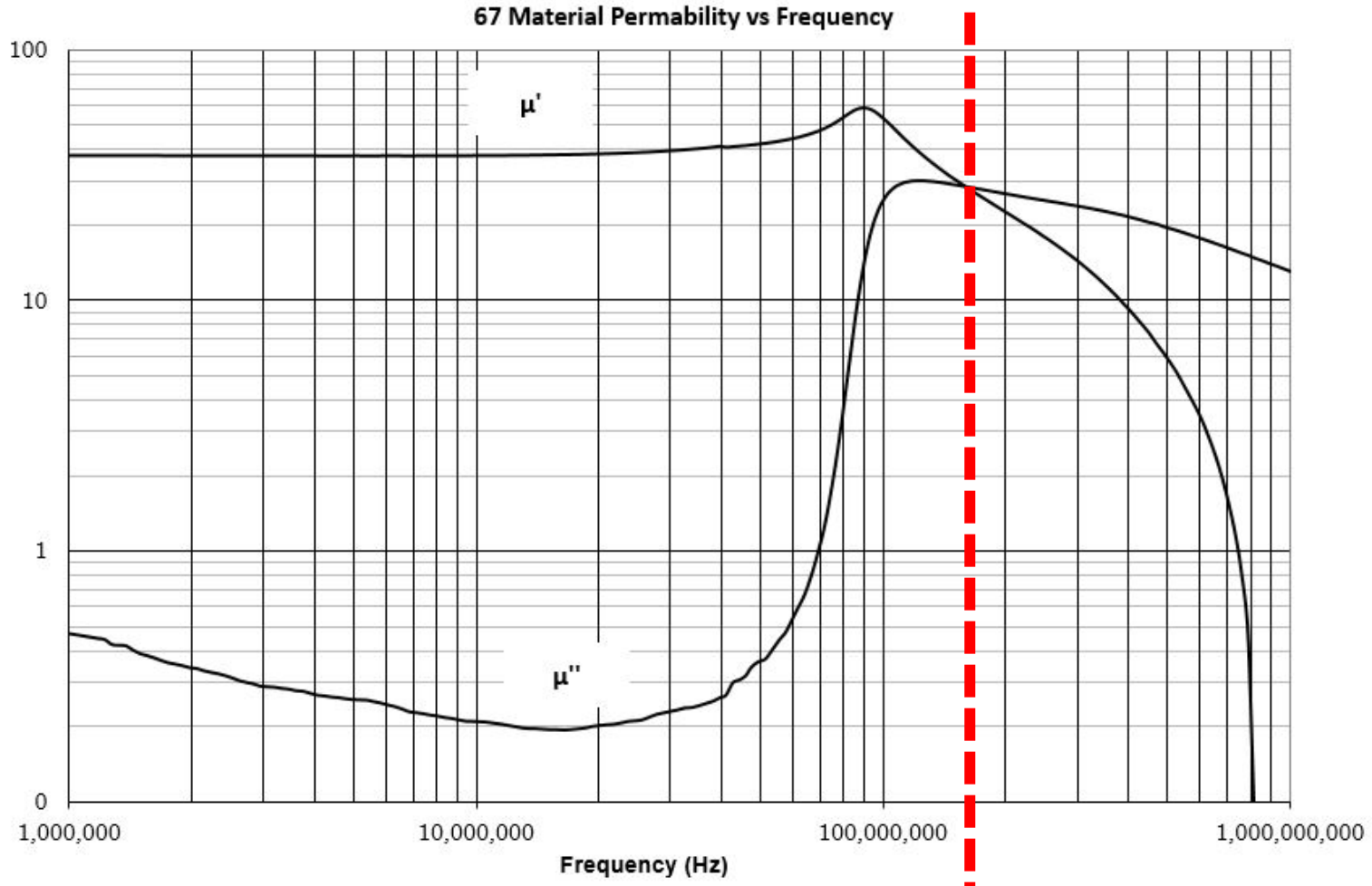
$$\frac{1}{Q} = \frac{R}{\omega L} = \frac{\mu''}{\mu'}$$

The loss tangent is so named because $R/\omega L$ is opposite/adjacent in a complex impedance triangle

$$\tan\delta = \frac{\mu''}{\mu'}$$



Typical Magnetic Material Complex Permeability



$\tan\delta$ very low

$\tan\delta$ rising rapidly

$\tan\delta = 1$

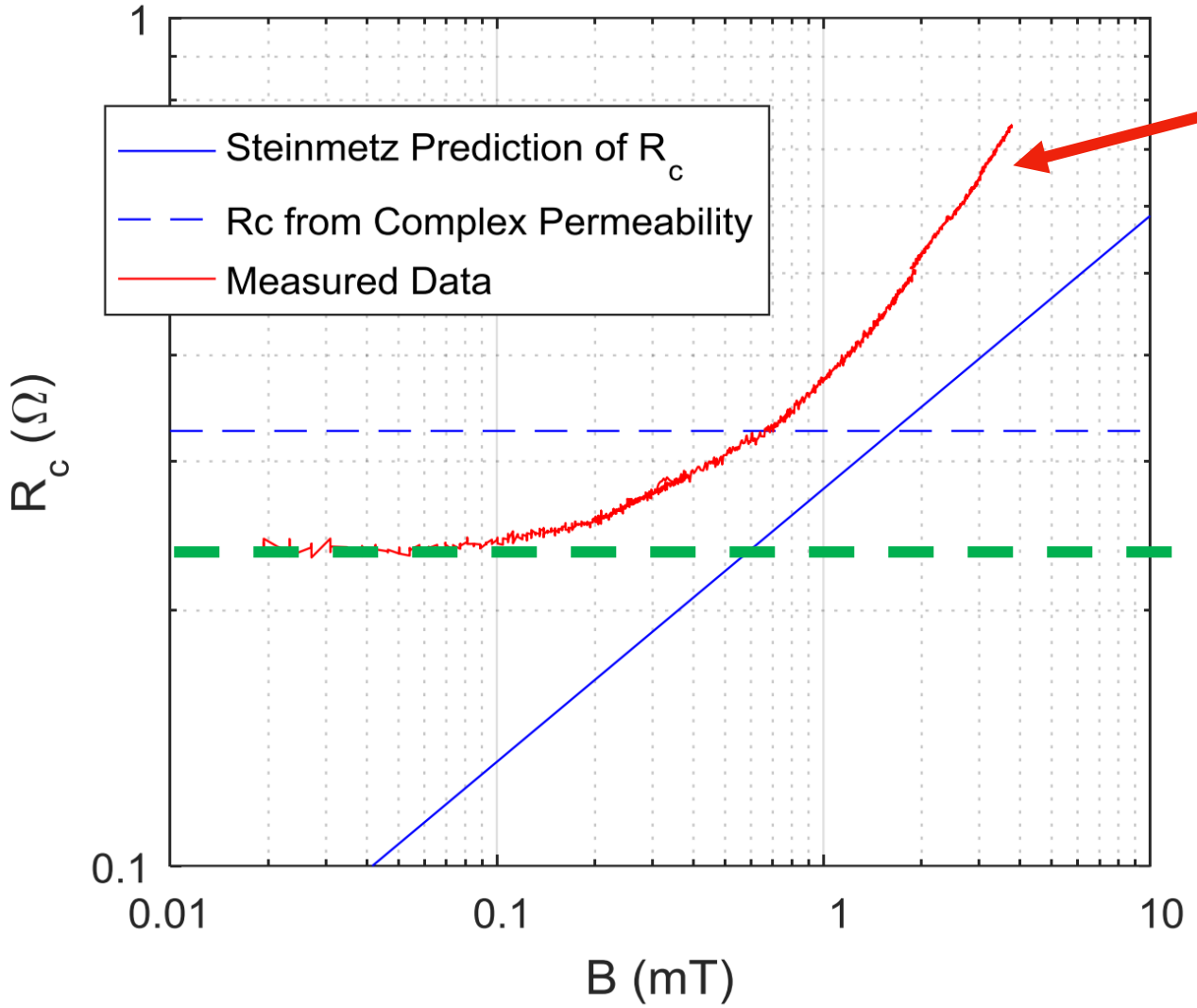
When loss is approximately linear, complex permeability is very convenient

a) Loss tangent ($\leftrightarrow R$) is a constant material property; does not depend on operating condition

b) R does not need to be measured at the true drive level and therefore is easily and quickly measured on an LCR meter/impedance analyzer/VNA, etc.

c) It is valid and convenient to include loss as a simple resistor in simulations

d) It is valid to break complicated excitations into individual Fourier components (fundamental, third harmonic, etc.), calculate loss individually for each one, and add the results back up



Actual core

If core were linear (on a P vs B plot, this would be $P \propto B^2$)

⇒ Core loss is only linear for very small ac drive levels (small-signal)

B. X. Foo, A. L. F. Stein and C. R. Sullivan, "A step-by-step guide to extracting winding resistance from an impedance measurement," *2017 IEEE Applied Power Electronics Conference and Exposition (APEC)*, Tampa, FL, USA, 2017, pp. 861-867, doi: 10.1109/APEC.2017.7930797.

Pet Peeves Regarding VNAs/LCR Meters:

1) You cannot measure loss with a VNA and expect results to work at high drive levels. Core loss is nonlinear.

2) Just because core loss decreases quickly with drive level does *not* mean you can use very low drives (like VNA) to measure copper loss alone. Core loss becomes linear for low enough drive and is often not negligible.

3) You cannot *remove* the core and measure copper loss alone on a VNA. HF current flows near regions of high H-field, and removing the core completely changes the H-field distribution near the wires.

Only very-low-ripple chokes/filters experience “small-signal” ac excitation. For most power applications, including buck chokes, modeling loss as linear is invalid. If loss is not linear, we lose a lot of convenient and powerful tools:

a) Loss now depends on drive level and we cannot extrapolate from the loss at one drive level to the loss at another without more information

b) It is not valid to include loss as a simple resistor in simulations

c) Loss must be measured at the true drive level and therefore is not easily or quickly measured. Small-signal tools (LCR meter, impedance analyzer, VNA) are all out of the question

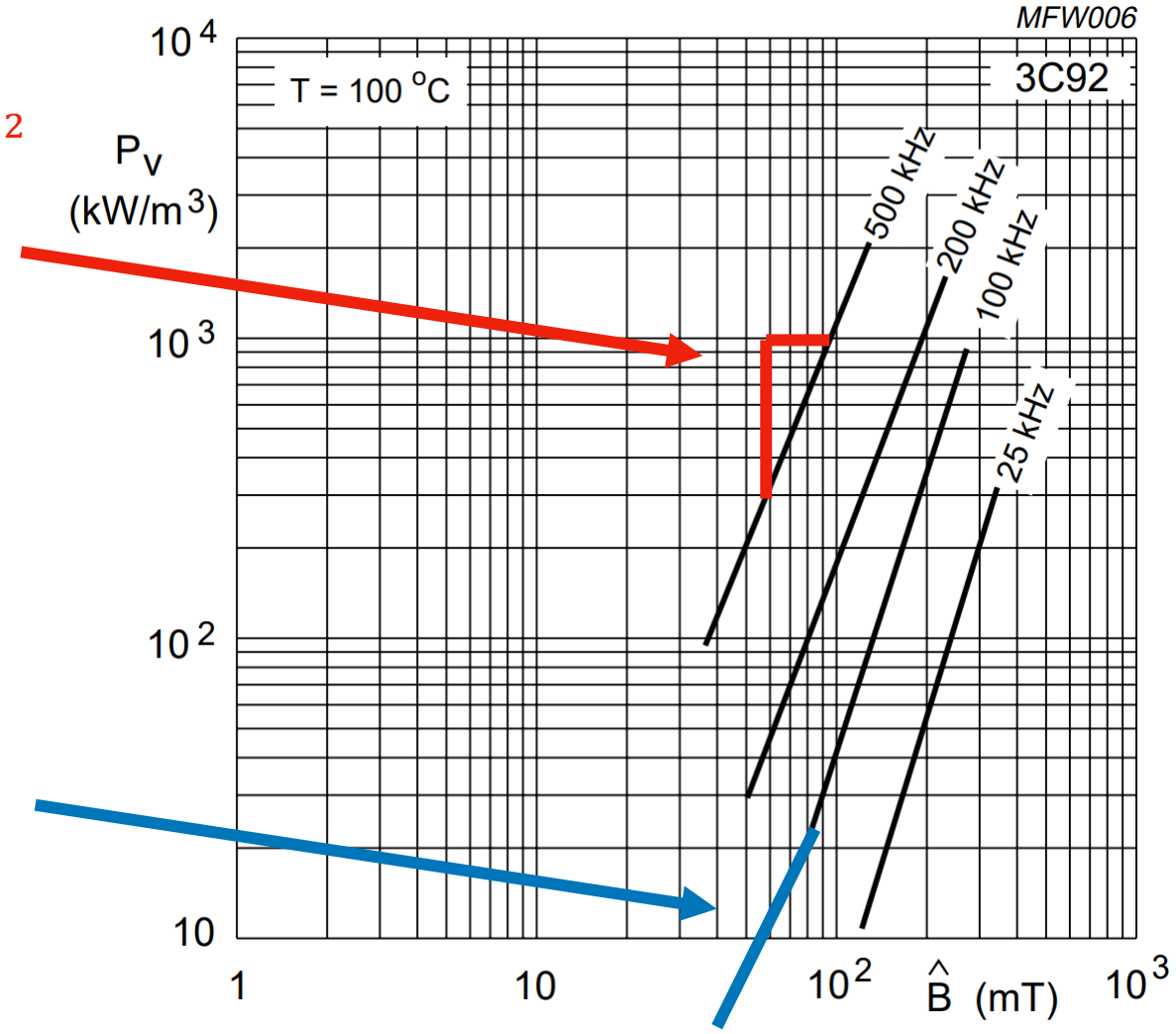
d) Loss must be measured under relevant excitations. It is not valid to break complicated excitations into individual Fourier components (fundamental, third harmonic, etc.), calculate loss individually for each one, and sum the results (though iGSE has been shown to work pretty well)

Large-signal power loss must be characterized experimentally across both frequency *and* drive level

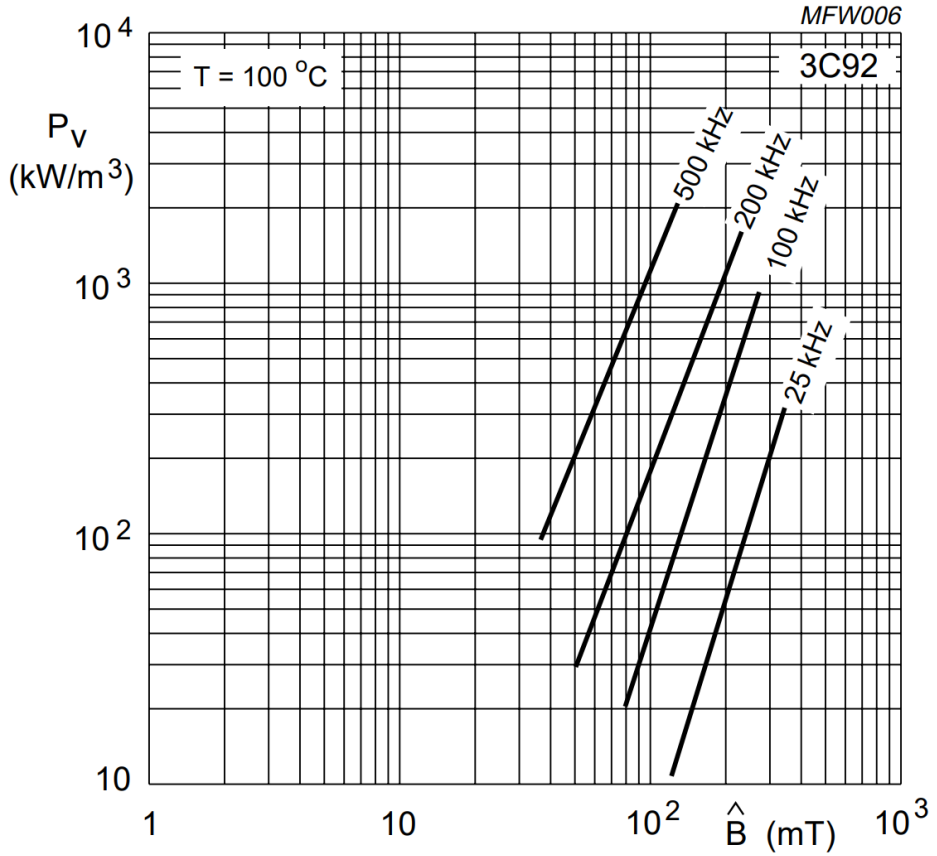
At high drive, P scales faster than B^2 ("resistance" not constant).

This is nonlinear

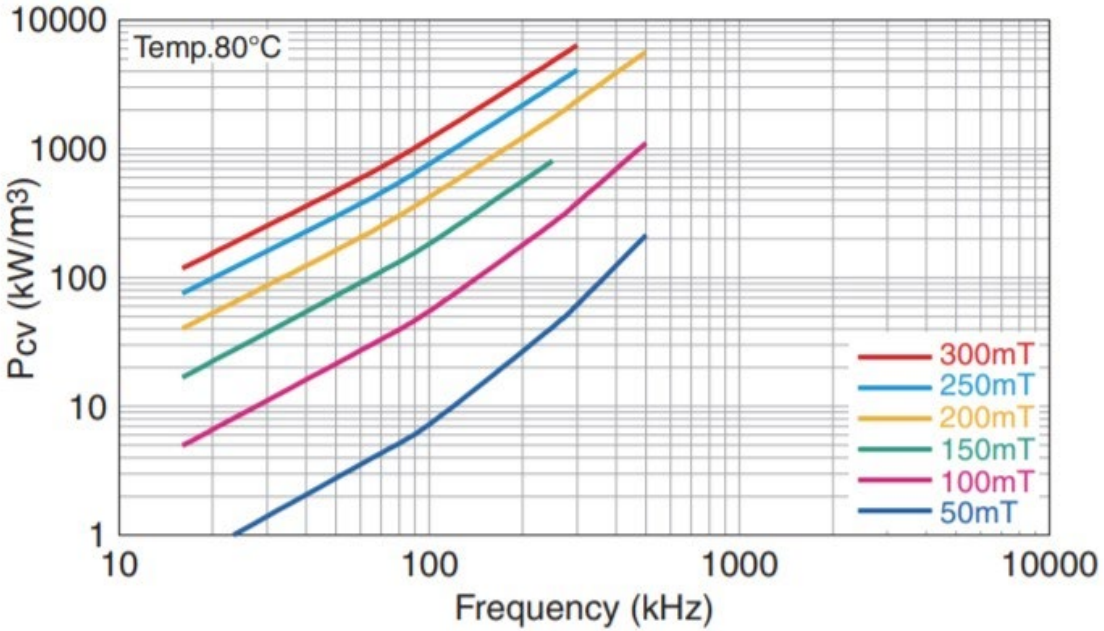
At a low enough drive level, we expect the curve to become $\propto B^2$ (constant resistance), as we showed earlier



Better and Worse Ways to Communicate Core Loss



Good: B field is at the discretion of the component designer, whereas frequency is dictated by circuit and (rightfully) chosen as a tradeoff among many factors. It makes sense to fix frequency and vary B.

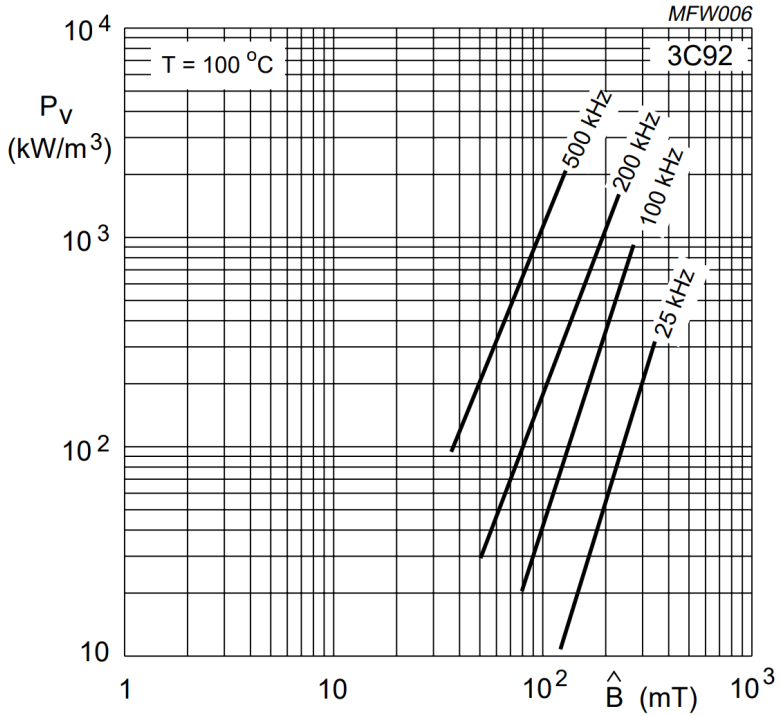


Bad: B field is rarely fixed by circuit and frequency is rarely at the discretion of the component designer. It makes little sense to fix B and vary frequency. Convenient for the manufacturer to measure but not for the user!

You may have observed that loss is not linear $\left(P \propto B^2\right)$, but it *does* appear to obey a power law relationship (straight lines on loglog plot, $P \propto B^{\beta \neq 2}$). This observation is known as the Steinmetz equation

$$P = k(f)B^{\beta(f)}$$

Where typically $2 < \beta(f) < 3$.



Note that this is an *observation* about a *common* trend in magnetic material losses. It is not an immutable law of the universe and is not derived from first principles

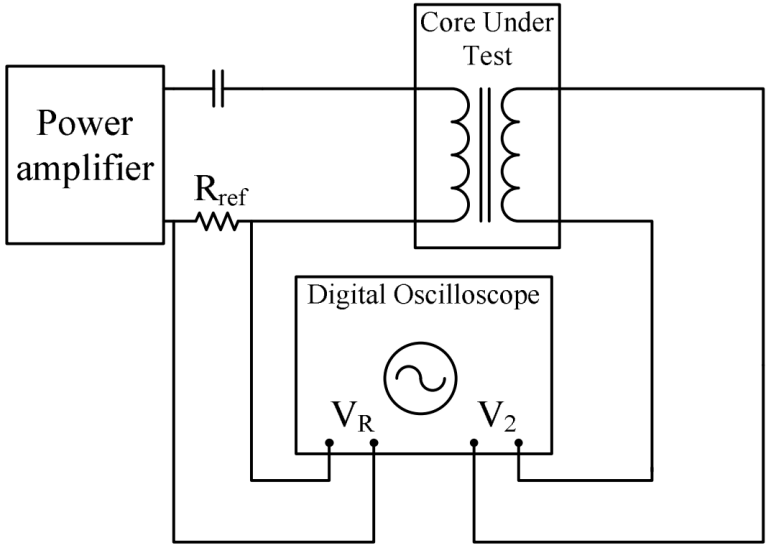
Less-reliably, β may be constant and P may obey a power law relationship with frequency: $P \propto kf^{\alpha}B^{\beta}$

The direct approach to measuring loss is to measure voltage and current across time and then average their product

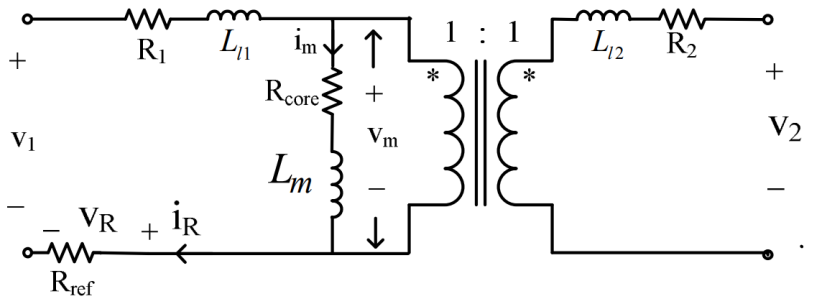
$$P_v = \frac{1}{VOL}\int_0^{2\pi} v(t)i(t)\,\mathrm{d}(\omega t)$$

While $v(t)i(t)$ includes both loss and energy storage, the average across a cycle is pure loss in steady state.

(in many cases, a second winding is added to try to isolate core loss from winding loss – the distinction is not important here)



(a) four-wire method for core loss measurement



(b) equivalent model

M. Mu, F. C. Lee, Q. Li, D. Gilham and K. D. T. Ngo, "A high frequency core loss measurement method for arbitrary excitations," *2011 Twenty-Sixth Annual IEEE Applied Power Electronics Conference and Exposition (APEC)*, Fort Worth, TX, USA, 2011, pp. 157-162, doi: 10.1109/APEC.2011.5744590.

Even for a nonlinear system, **if V is sinusoidal, then only the component of I which is at the same frequency (I_1) can contribute to real power**, therefore we can (and, in some sense, must) only focus on the fundamental component of I. If I is markedly non-sinusoidal, Fourier analysis can extract the fundamental component

$$P_v = \frac{1}{VOL}\int_0^{2\pi} v(t)i(t)d(\omega t) = \frac{1}{VOL}\int_0^{2\pi} V\sin(\omega t)\, I_1 \sin(\omega t + \phi_1)\, d(\omega t)$$

$$= \frac{VI_1}{VOL}\int_0^{2\pi} \sin(\omega t)\sin(\omega t + \phi_1)\, d(\omega t)$$

$$= \frac{VI_1}{VOL}\pi\cos\phi_1$$

For low-loss inductors, ϕ will be close to 90 degrees

$$\cos 88^\circ = 0.0349$$
$$\cos 89^\circ = 0.0174$$
$$\cos 90^\circ = 0.0000$$

If the true phase is 89 degrees and **if our measurement is off by 1 degree, our answer is either 100% too high or 100% too low**!

Quantifying sensitivity to phase angle for linear or quasi-linear systems:

Total Impedance: $z = R + j\omega L = Z\angle\theta$

Quality Factor: $Q = \frac{2\pi E_{store}}{E_{diss}} = \frac{\omega L}{R} = \tan\theta$

Phase Angle [deg]: $\theta = \frac{180}{\pi}\tan^{-1} Q$

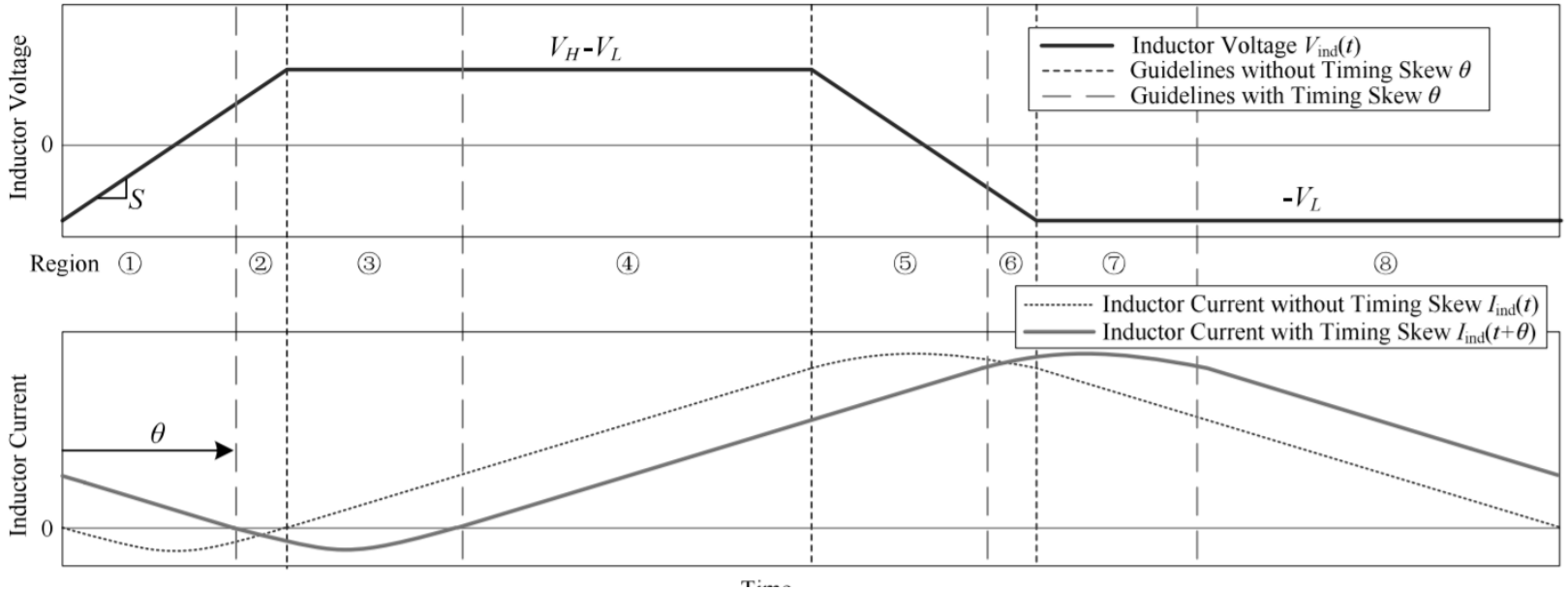
How fast does phase change with Q? (Phase Sensitivity) $\frac{d\theta}{dQ} = \frac{180}{\pi}\frac{1}{Q^2+1}$

For a given $\frac{\mathrm{dQ}}{\mathrm{Q}}$ (% error), what dθ is acceptable? $d\theta = \frac{180}{\pi}\frac{dQ}{Q^2+1} = \frac{180}{\pi}\frac{dQ/Q}{Q+\frac{1}{Q}} \approx \frac{180}{\pi}\frac{dQ/Q}{Q}$

Delay/Skew Error: $d\theta = \frac{t_d}{T} \times 360^\circ = t_d f \times 360^\circ$

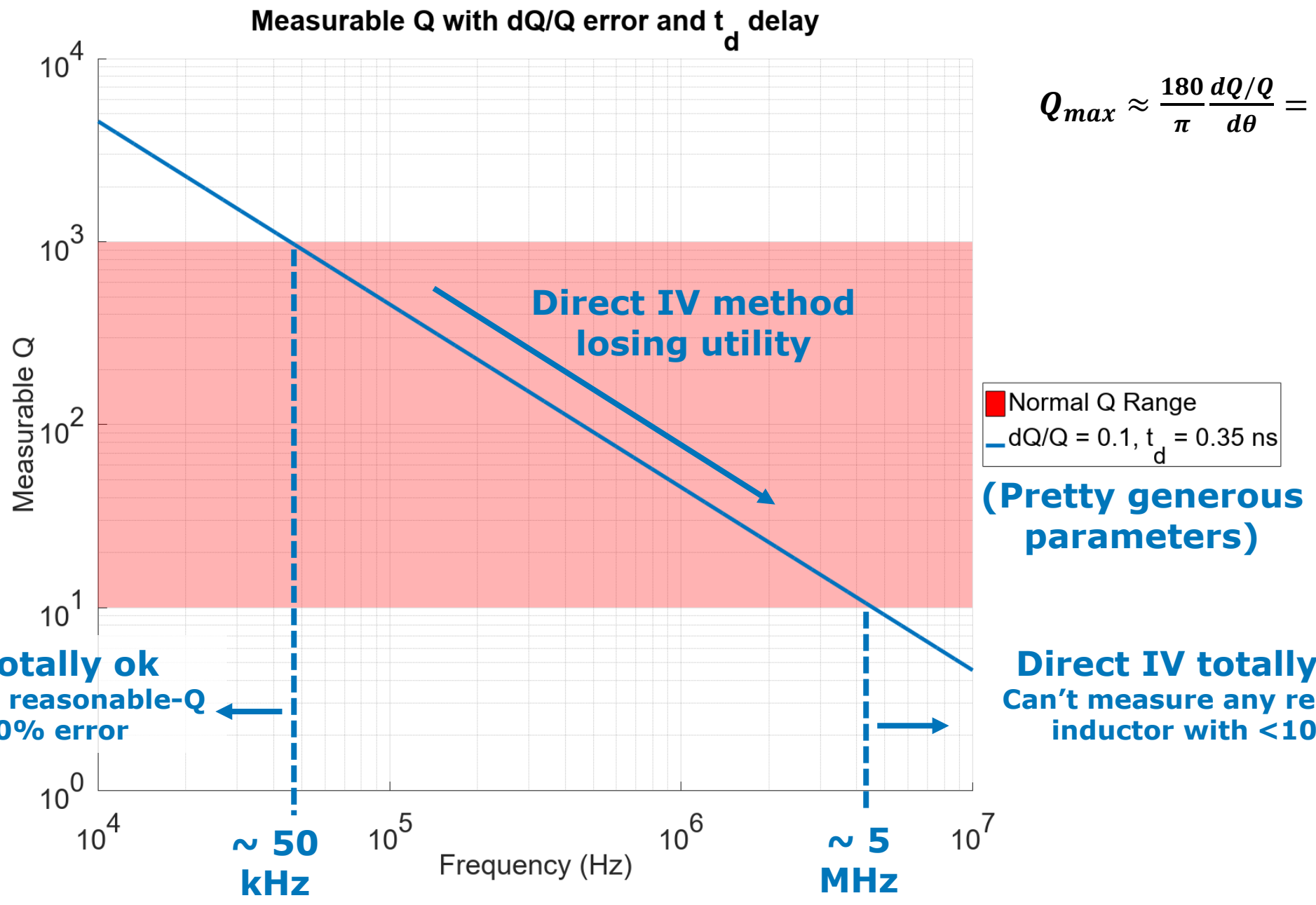
Max measurable Q for a target % accuracy and given delay/skew: $Q_{max} \approx \frac{180}{\pi}\frac{dQ/Q}{d\theta} = \frac{dQ/Q}{2\pi\, t_d f}$

The best work I know of on direct IV skew correction still has uncertainty ~ 1/3 ns



V_H(V)	θ_{real}(ns)	
	Mean (μ)	SD (1-σ)
20	-39.5047	0.3243
40	-39.2776	0.1897
60	-39.1504	0.2159

L. Yi and J. Moon, "In Situ Direct Magnetic Loss Measurement With Improved Accuracy for Lossier Magnetics," in *IEEE Transactions on Instrumentation and Measurement*, vol. 71, pp. 1-14, 2022, Art no. 6001414, doi: 10.1109/TIM.2022.3150877.



$$Q_{max} \approx \frac{180}{\pi}\frac{dQ/Q}{d\theta} = \frac{dQ/Q}{2\pi\, t_d f}$$

(Pretty generous parameters)

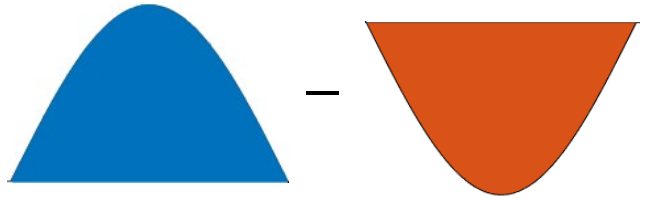
Direct IV totally ok
Can measure any reasonable-Q inductor with <10% error

Direct IV totally useless
Can't measure any reasonable-Q inductor with <10% error

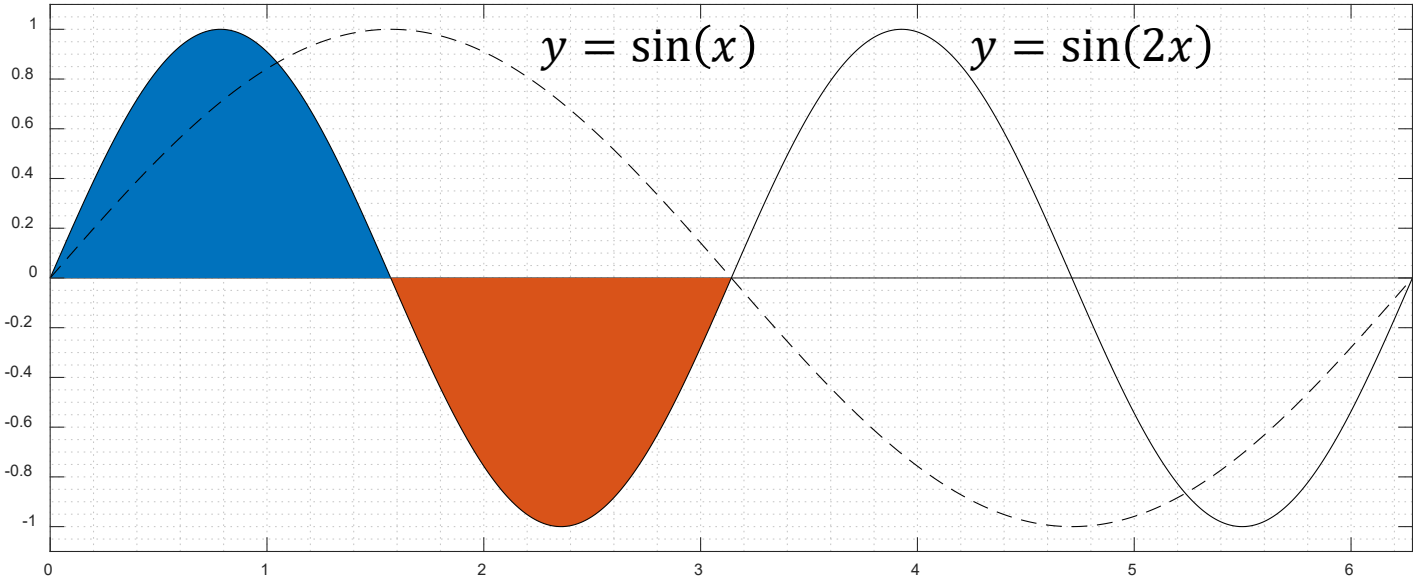
What is the fundamental problem with the direct IV measurement? We have to integrate the following:

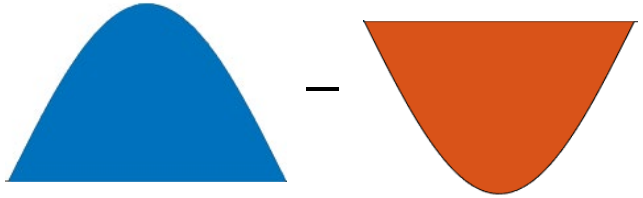
$$v(t)i(t) \approx \sin(x)\cos(x) = \frac{1}{2}\sin(2x)$$

$\frac{1}{2}\sin(2x)$ has zero average value but plenty of area to integrate.

$$\int \frac{1}{2}\sin(2x) =$$  $-$

While the blue and orange areas are individually large, their difference is small. Thus **minor errors** in measuring the large numbers results in **huge errors** in the difference



$$\int \sin(x)\cos(x) = \int \frac{1}{2}\sin(2x) =$$  $-$

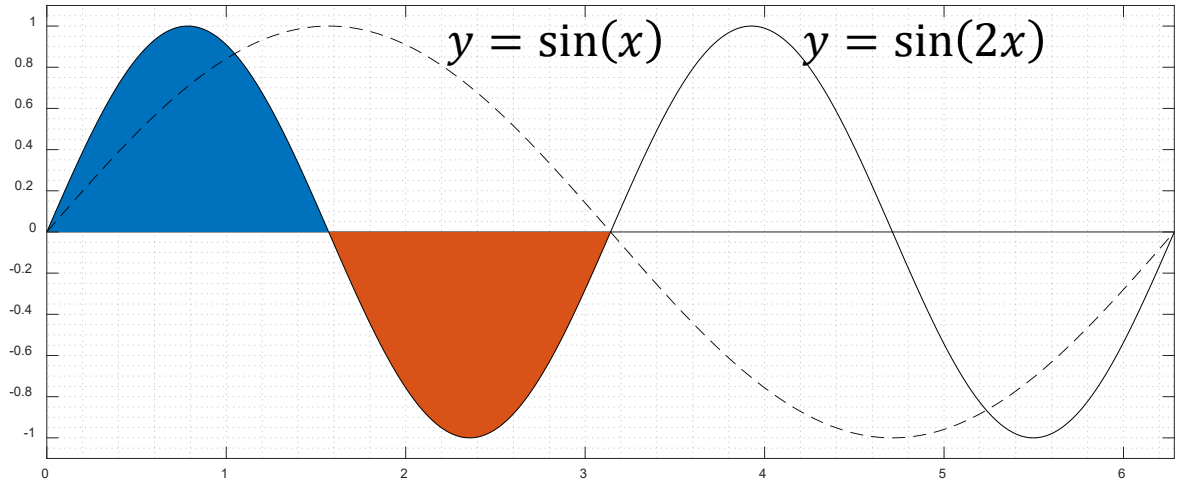
Why is there a small-difference-of-large-numbers problem?

Blue and orange areas are *large* because the *energy storage* is large $(j\omega L)$

The difference is *small* because the *loss* is small (R)

$\Rightarrow$ We're trying to find a tiny needle of loss in a haystack of energy storage

$\Rightarrow$ We're trying to find a tiny needle of real impedance in haystack of imaginary impedance



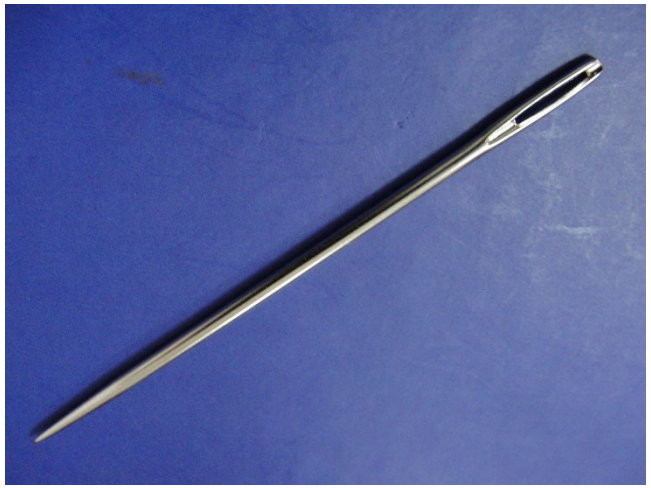
⇒ We're trying to find a tiny needle of loss in a haystack of energy storage

⇒ We're trying to find a tiny needle of real impedance in haystack of imaginary impedance

⇒ We're trying to find a tiny needle of loss in a haystack of energy storage

⇒ We're trying to find a tiny needle of real impedance in haystack of imaginary impedance

⇒ **What if we could get rid of the haystack and leave the needle? (get rid of the imaginary impedance and leave the real impedance?)**

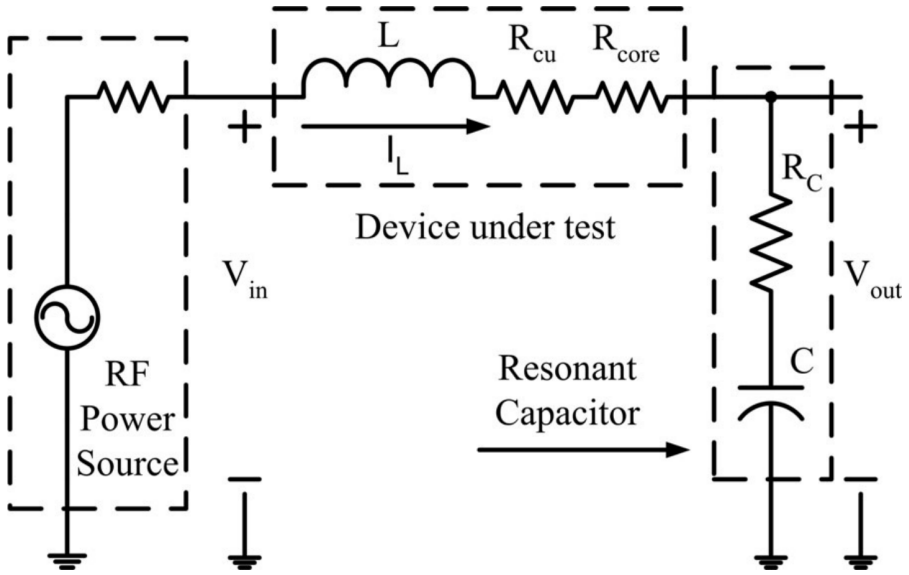


By operating at the resonant frequency between a capacitor and the inductor DUT, $j\omega L$ cancels with $1/j\omega C$ (burning down the haystack) and we are left with R alone (the needle).

Now $\frac{1}{2\pi}\int v(t)i(t) \approx \frac{1}{2\pi}\int V\sin\omega t \;\; I\sin\omega t = \frac{VI}{2}$ (no ϕ dependence at all!)

Before, small ϕ differences caused *huge* answer differences

Now, small ϕ differences cause no error at all (to first-order Taylor approximation)



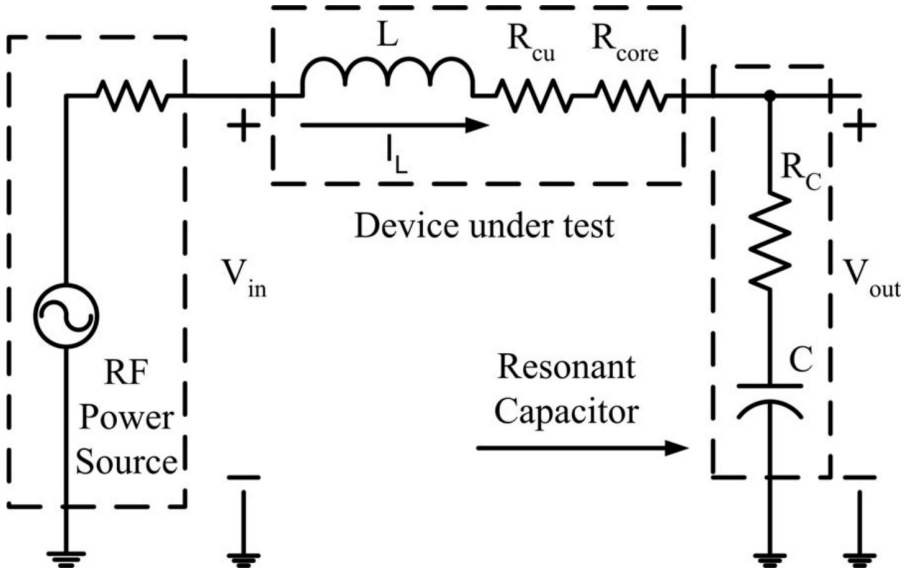
We can be even more clever and measure $V_C(t)$ instead of $i(t)$ at high frequency. If R_C is small, then it turns out that

$$\frac{|V_C|}{|V_{in}|} = Q_L = \frac{\omega L}{R}$$

$\Rightarrow$ Only voltage measurements

$\Rightarrow$ Only voltage *magnitudes*

Y. Han, G. Cheung, A. Li, C. R. Sullivan and D. J. Perreault, "Evaluation of Magnetic Materials for Very High Frequency Power Applications," in IEEE Transactions on Power Electronics, vol. 27, no. 1, pp. 425-435, Jan. 2012, doi: 10.1109/TPEL.2011.2159995.



$$\frac{|V_C|}{|V_{in}|} \approx Q_L = \frac{\omega L}{R}$$

Disadvantages:

1) For each frequency, need to change capacitor and usually fine-tune frequency
2) Low net impedance (low Vin, high I) – sometimes makes PA nonlinear
3) Need very low-loss capacitors, sometimes with up to a couple kV across them at HF
4) Result reflects *total* inductor loss. Must design experiment to ensure $P_{core} \gg P_{cu}$
   (Can also estimate P_{cu} to "back-out" P_{core})

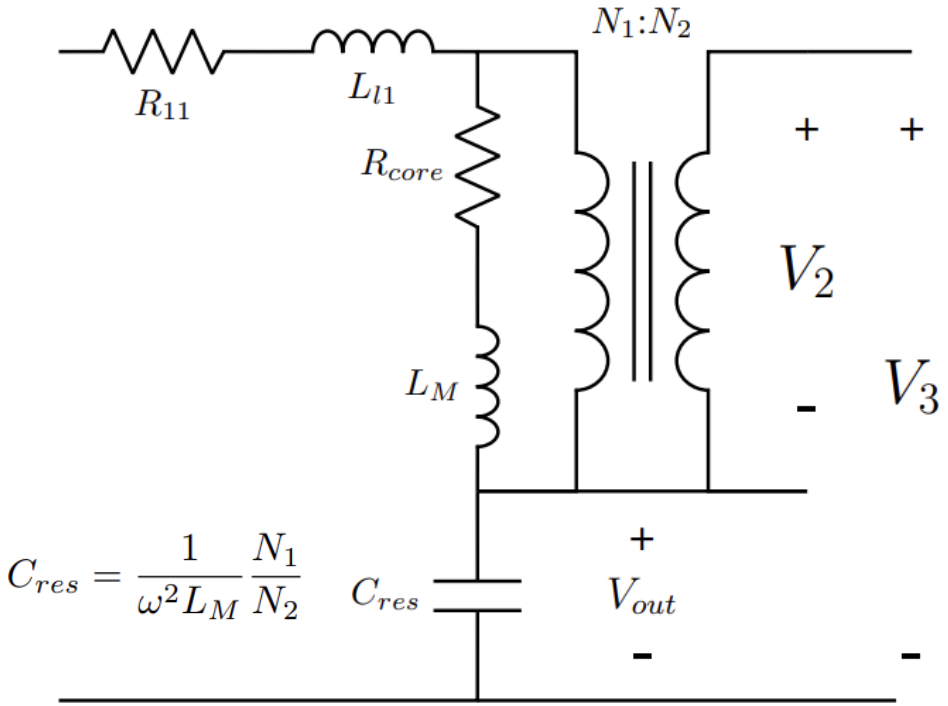
Y. Han, G. Cheung, A. Li, C. R. Sullivan and D. J. Perreault, "Evaluation of Magnetic Materials for Very High Frequency Power Applications," in IEEE Transactions on Power Electronics, vol. 27, no. 1, pp. 425-435, Jan. 2012, doi: 10.1109/TPEL.2011.2159995.

Another series-resonant circuit, originally from Virginia Tech:

Original method used a current-sense resistor and integrated $v(t) \times i(t)$

Slightly modified method measures the amplitudes of V_3 and V_{out}

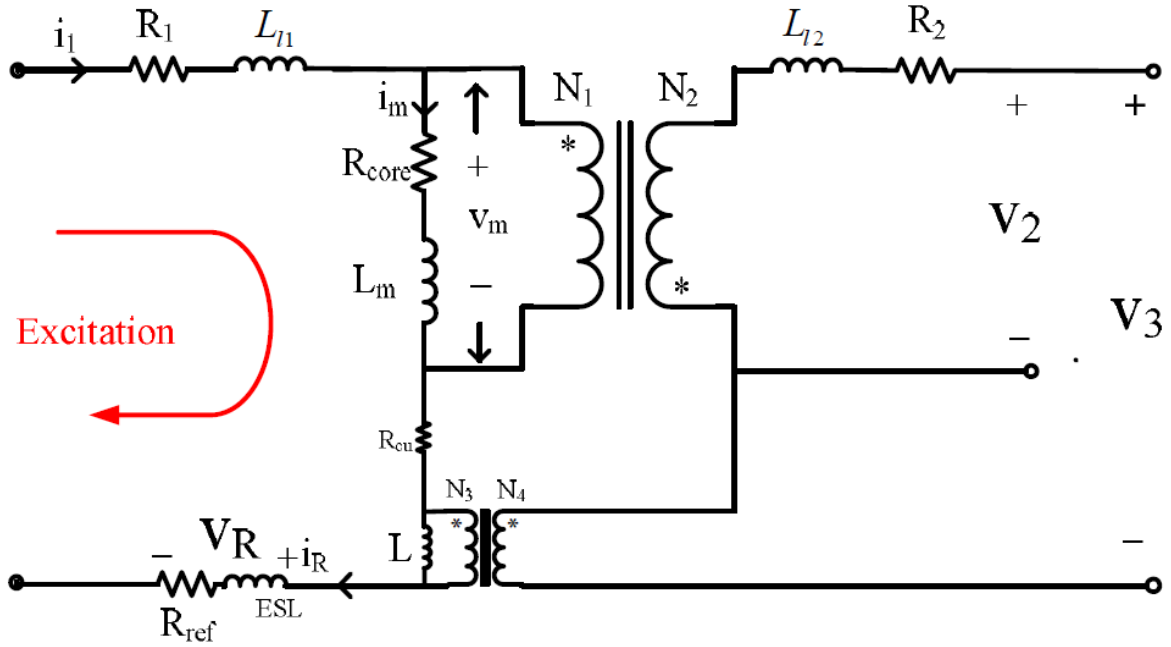
$$\frac{V_{3,pk}}{V_{out,pk}} = \frac{\left|\frac{N_2}{N_1} I_1 (R_{core} + j\omega L_M) + I_1 \frac{1}{j\omega c_{res}}\right|}{\left|I_1 \frac{1}{j\omega C_{res}}\right|} = \frac{\left|\frac{N_2}{N_1}(R_{core} + j(\omega L_M - \omega L_M))\right|}{\left|\frac{N_2}{N_1}\omega L_M\right|} = \frac{R_{core}}{\omega L_M}$$



M. Solomentsev, O. Okeke and A. J. Hanson, "A Resonant Approach to Transformer Loss Characterization," 2022 IEEE Applied Power Electronics Conference and Exposition (APEC), Houston, TX, USA, 2022, pp. 596-603, doi: 10.1109/APEC43599.2022.9773516.

Non-Resonant Method: "Partial Cancellation"

Another way to "burn the haystack" and avoid resonance



An air-core "twin" is used instead of a capacitor.

$$v_3 = i_1 R_{core} + (L_m - L)\frac{d}{dt} i_1$$

If $L = L_m$, then v_3 measures the "voltage across R_{core}."

\+ Works for arbitrary excitations

\- Relies on v*i measurement; still somewhat sensitive to skew

\- Need air core L equal to cored L_m, may not be achievable

M. Mu, F. C. Lee, Q. Li, D. Gilham, and K. D. T. Ngo, "A high frequency core loss measurement method for arbitrary excitations," in *2011 Twenty-Sixth Annual IEEE Applied Power Electronics Conference and Exposition (APEC)*, Mar. 2011, pp. 157–162. doi: 10.1109/APEC.2011.5744590.

Frontiers

- Automating core loss measurements at high frequency
- HF loss measurement for arbitrary waveforms – still elusive
- HF characterization of coupled-magnetic components
- Machine learning to predict behavior under un-tested conditions
- Core-shape dependent loss mechanisms (dimensional resonance)

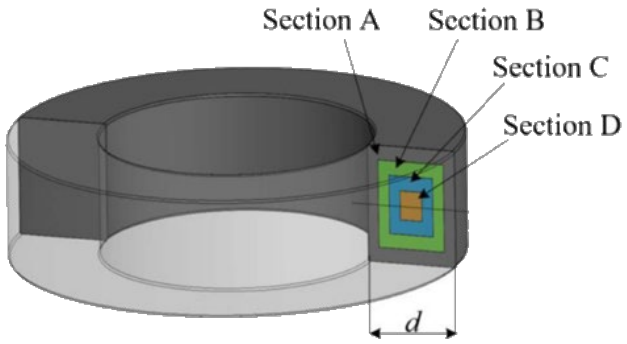


Fig. 29. Section arrangement in tested ring core.

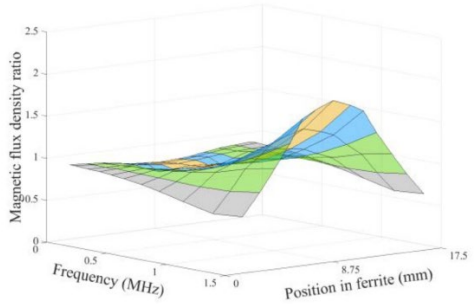


Fig. 32. Experimental magnetic flux density ratio in the each segment, frequency characteristic—material 3F36.

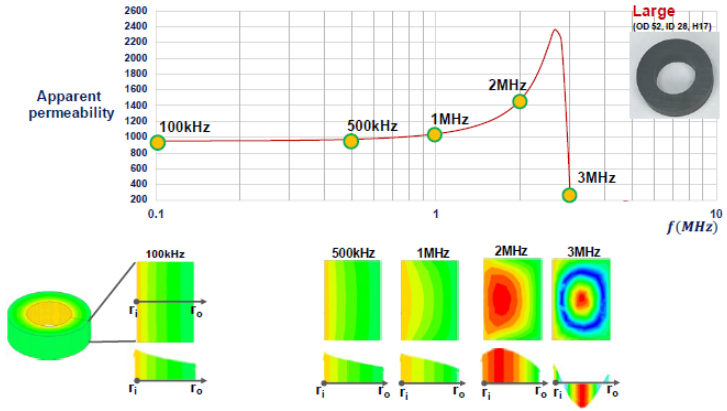


Fig.5. Tested apparent permeability of large toroid from ML91s, and corresponding simulated flux distribution with permeability $\mu = 1000 - j35$, and permittivity $\varepsilon = 30000 - j6000$.

M. Kącki, M. S. Ryłko, J. G. Hayes, and C. R. Sullivan, "Analysis and Experimental Investigation of High-Frequency Magnetic Flux Distribution in Mn-Zn Ferrite Cores," *IEEE Transactions on Power Electronics*, vol. 38, no. 1, pp. 703–716, Jan. 2023, doi: 10.1109/TPEL.2022.3200954.

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Summary

Understanding loss in a linear way (core loss “resistance,” complex permeability, loss tangent) only applicable for small excitations

Large-signal core loss characterized by experimental graphs and/or Steinmetz equation

Low-frequency core loss can be measured by direct VI measurements, but phase error makes this method untenable at high frequency (MHz+)

Resonant and other methods cancel energy storage (“reactance”) and make loss (“resistance”) easier to measure at high frequency

Mike Ranjram, Arizona State University Mike.Ranjram@asu.edu
Alex Hanson, University of Texas at Austin, ajhanson@utexas.edu