Practical Design of a Wideband EMI Injection Transformer

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***Abstract*—In the area of electromagnetic compatibility (EMC) testing, a wideband injection transformer is often an engineer's best friend. Unfortunately, these are often inaccessible both in terms of cost and scalability, creating the need for a custom solution. This report serves as a design outline for a wideband EMI injection transformer, as well as a general guide for the rapid prototyping of practical custom magnetics.**

I. INTRODUCTION

Electromagnetic interference (EMI) and parasitic noise coupling can lead to potential critical failures when conducted susceptibility and EMC testing are not performed. A common testing method is performed using a noise generator paired with an injection transformer to maintain galvanic isolation with the unit under test (UUT). Often times, the same signal must be injected over several points in the UUT (e.g. a battery stack), making single 1:1 commercial injection transformers unusable. The following sections outline the simplified theory and design guidelines for a custom solution, maintaining similar frequency bandwidths (30Hz to 1MHz) to a commercially available product.

II. CORE SELECTION

A. Transformer Design Equation

A transformer design equation to guide core selection can be derived starting with Faraday's law of induction:

$$V = N\frac{d\phi}{dt} = NA\frac{dB}{dt}$$

Since the injected voltage and therefore current is sinusoidal, flux may be represented as such:

$$B = B_{max}\sin(\omega t)$$

Following this:

$$\frac{dB}{dt} = B_{max}\omega\cos(\omega t)$$

Since the limiting factor is the maximum saturation flux, the cosine can be dropped:

$$\frac{dB}{dt}_{max} = B_{max}\omega$$

Substituting this back into Faraday's equation results in:

$$V_{pk} = NAB_{max}\omega$$

Since V_{pk} is the maximum output voltage and voltage is sinusoidal, we can substitute $V_{pk} = \sqrt{2}V_{rms}$:

$$\sqrt{2}V_{rms} = NAB_{max}\omega = NAB_{max}2\pi f$$

$$(V_{rms})\left(\frac{1}{f}\right) = NAB_{max}\frac{2\pi}{\sqrt{2}}$$

To account for a step-down factor, a turns ratio is added. This may be used to construct the following transformer design equation:

$$N_p \geq \frac{(turns\ ratio)\cdot\sqrt{2}\cdot V_{rms}}{2\pi f_{min}A_eB_{max}} \tag{1}$$

where
N_p – turns on the primary winding
V_{rms} – RMS voltage of the output secondaries
f_{min} – minimum -3dB frequency cutoff
A_e – effective core area
B_{max} – maximum flux limit (50% B_{sat})

B. Magnetizing & Self Inductances

Calculating the inductance of a transformer is important for both simulation and leakage optimization. Magnetizing inductance of the transformer can be approximated from the primary self-inductance (L_p):

$$L_p = A_LN_p^2 = \frac{\mu_r\mu_0N_p^2A_e}{l_e} \tag{2}$$

where
A_L – inductance coefficient
μ_r – relative permeability
μ_0 – vacuum permeability
l_e – effective path length

The magnetizing inductance (L_{mag}) can be modeled using an estimated leakage inductance:

$$L_{mag} = L_p - L_{leakage}$$

The secondary self-inductance (L_s) can be calculated using an inductance ratio:

$$\frac{L_p}{L_s} = \left(\frac{N_p}{N_s}\right)^2$$

C. Parasitic Elements

Depending on the application of the transformer, practical transformer design often results in a trade off between the different non-ideal parasitics of the magnetics.

1) Winding Resistance: Winding resistances may serve to damp the frequency response of a transformer. Although highly dependent on winding geometry, primary and secondary winding resistances are approximated using the following equation:

$$R_{wind} \approx N \cdot \frac{\rho \pi d_{wind}}{A_{wire}}$$

where N is the number of winds, d_{wind} is the winding diameter, ρ is the material resistivity of the wire ($\rho_{cu} \approx 1.72 \cdot 10^{-8}$), and A_{wire} is the cross-sectional area of the wire. Given the wire gauge in AWG, this can be found using:

$$A_{wire}\,(\mathrm{mm}^2) = 0.012668\,(\mathrm{mm}^2) \times 92^{(36-\mathrm{AWG})/19.5}$$

2) Leakage Inductance: Frequently, the limiting factor to a transformer's upper frequency bandwidth is the leakage inductance, particularly if a transformer has higher windings. Leakage inductance occurs from uncoupled inductance between the primaries and secondaries. Capacitive loads reduce frequency bandwidth when resonating with magnetizing and leakage inductance at higher frequencies. Leakage may be minimized by using an interleaved winding geometry to maximize coupling symmetry. Leakage inductance can often vary between 0.01% to 0.001% of L_{mag} depending on the winding configuration.

3) Winding Capacitance: Capacitive coupling between windings provide another array of parameters to mitigate. Winding geometries play a major factor in capacitance; sparser winds help minimize capacitance, and sector wound primaries and secondaries decrease inter-winding capacitance.

III. Transformer Modeling

Simulating a transformer in SPICE can give valuable insight before final core selection. An equivalent model for an injection transformer can be found in Fig. 1, where the magnetizing inductances and parasitics can be calculated from the above sections. In most cases, such a model is sufficient for core selection purposes.

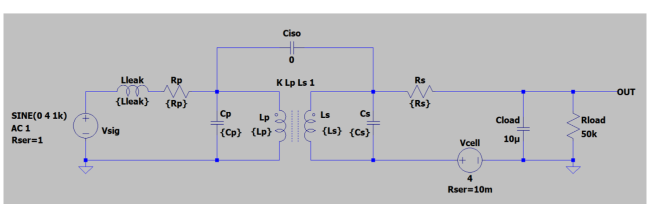

Fig. 1. Transformer (1:1) equivalent SPICE model with a capacitive load.

Ultimately, the main factors influencing frequency bandwidth is the saturation frequency (for lower cutoff) and leakage inductance in conjunction with any capacitance at the load (for upper cutoff). As shown in Fig. 2, our upper frequency cutoff is too low for our target specification (30Hz - 1MHz).

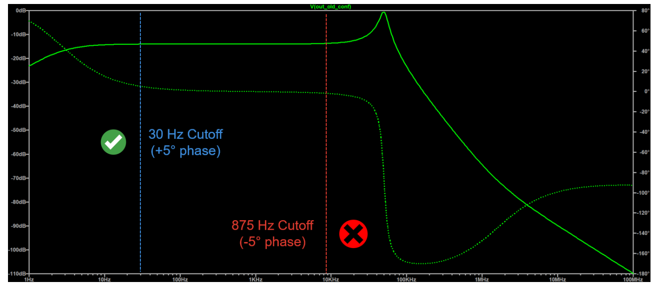

Fig. 2. Initial frequency response showing an upper cutoff frequency far below our target 1MHz cutoff.

This is caused by the leakage inductance from the transformer resonating with the $10\mu F$ load capacitor, as visualized in Fig. 3.

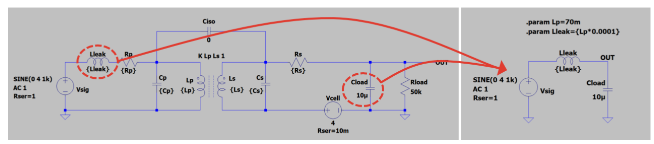

Fig. 3. Leakage inductance from the transformer resonates with the $10\mu F$ capacitive load and limits the higher frequency band.

An equivalent LC resonant circuit is simulated in Fig. 4 to show the dominant pole response caused by this capacitive loading. This cutoff frequency f_0 can be modeled by:

$$f_0 = \frac{1}{2\pi\sqrt{L_{leak}C_{load}}} \tag{3}$$

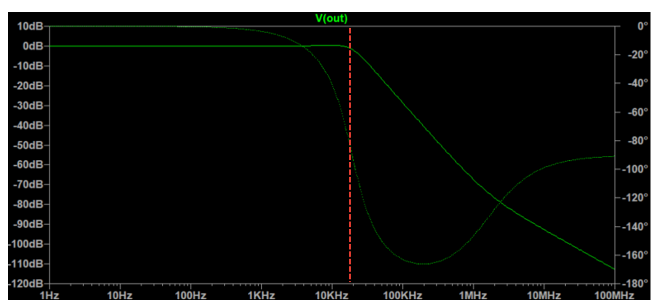

Fig. 4. Frequency response of the LC resonance between the leakage inductance and capacitive load. Compare with Fig. 2 and note the matching in upper cutoff.

As seen in Equation 3, the solution for a higher bandwidth is to either reduce the load capacitance, or to reduce the leakage. Reducing the load capacitance may be impractical if the load is fixed, meaning reducing parasitic inductance is the most practical option.

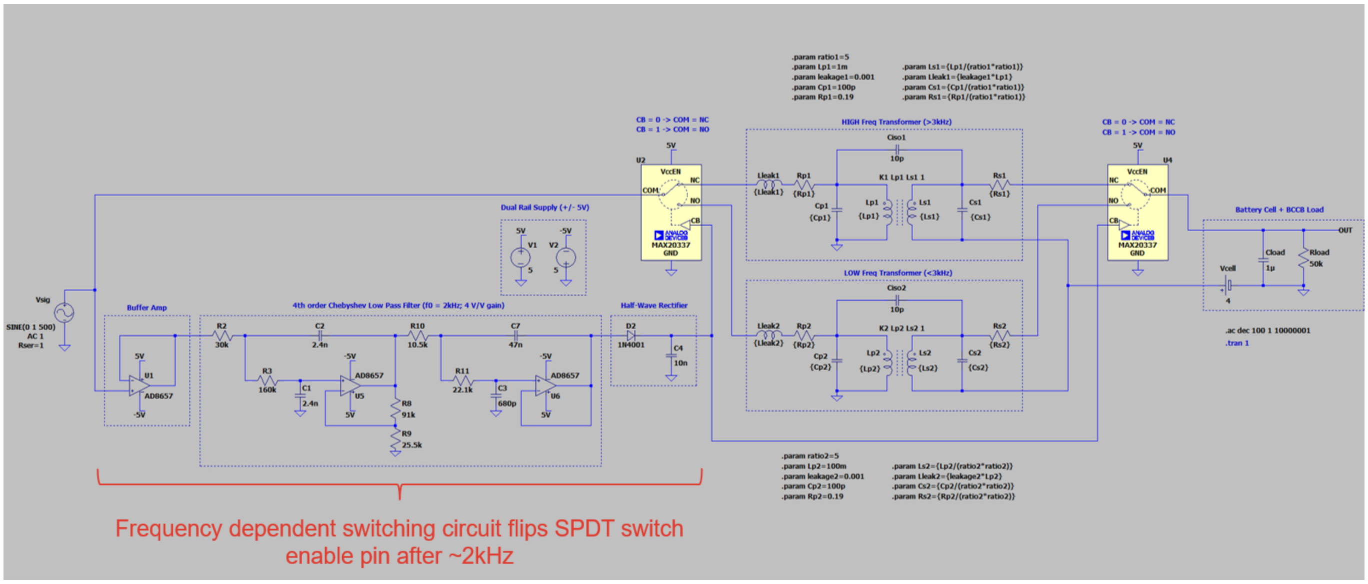

Fig. 5. Proposed transformer switching circuit.

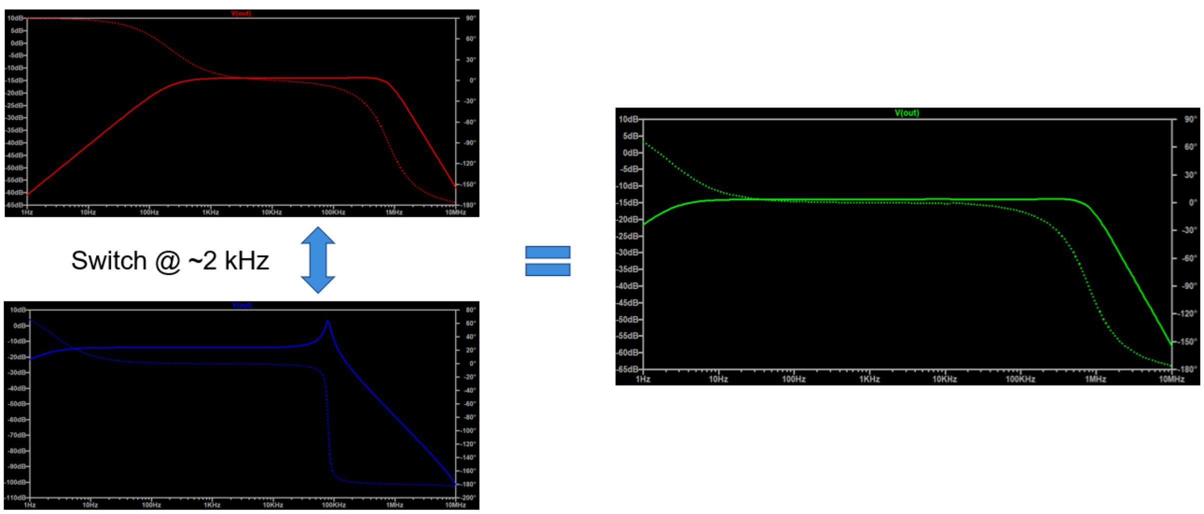

Fig. 6. Switching between a dedicated high frequency and low frequency transformer can broaden the bandwidth to a wider EMI input range.

IV. Switched Transformers

A solution for maintaining a high bandwidth could involve a switched transformer setup. A transformer built for high frequencies may use minimal turns to lower the magnetizing and therefore leakage inductances, pushing a higher frequency cutoff since the LC resonant frequency would be higher. A low frequency transformer would use more turns in order to gain a higher magnetizing inductance (at the cost of higher leakage) to get the lower frequency cutoff due to being more saturation resistant. A setup which could autonomously switch between the two would lengthen the net bandwidth. An outline for an implementation of this circuit is shown in Fig. 5. The combined frequency response is shown in Fig. 6.

V. Conclusion

The theory behind designing custom magnetics may seem esoteric at a low level. However, most practical designs do not require more than the simple design equations presented in this outline. The goal is for this to serve as a resource for those looking where to start.