Rectifier snubbing – background and Best Practices

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Look inside almost any piece of commercial equipment having a linear power supply and you will find small capacitors across the rectifier diodes. Those capacitors are there to prevent the combination of mains transformer leakage inductance and diode capacitance from ringing when the diode switches off. But would ringing really occur, and if snubbers do work, how well do they work?

To answer these questions, this article will first focus on investigating transformer leakage inductance, and then address semiconductor diode capacitance. The two results are then applied to simulations.

Transformer leakage inductance

Transformers first convert electric current into magnetic flux, then convert that flux back into a current – with the benefit of electrical isolation and convenience of current/voltage transformation. If the transformer was 100% efficient, all electrical power entering the primary would be available at the secondary, but no transformer is lossless, and one loss is due to imperfect magnetic coupling between primary and secondary. Imperfect coupling is a problem in audio not just because it represents power wastage but because leakage flux may induce currents in neighbouring circuitry leading to audible hum. Further, eddy currents in the core consume power. The combined effect of flux leakage and core losses is habitually represented by a series inductance known as the leakage inductance, as shown in **Figure 1**.

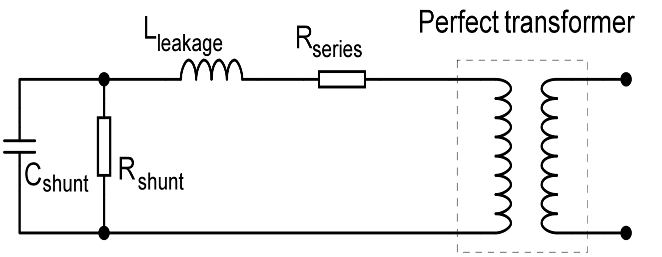

Figure 1 Simple high frequency transformer model.

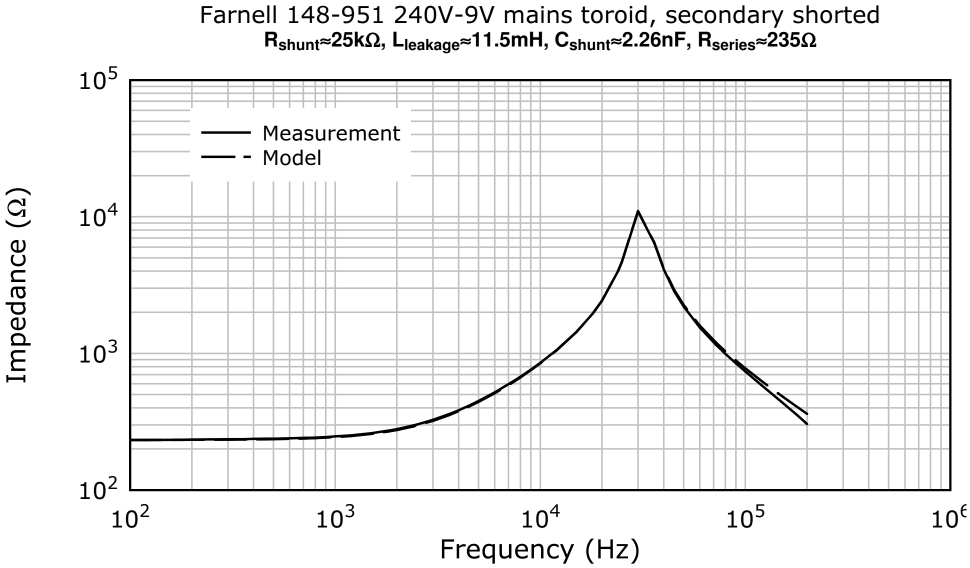

Figure 2 Deriving leakage inductance from a plot of impedance against frequency.

Discussions of leakage inductance blithely state that it is measured by measuring one winding's inductance with the other short-circuited – suggesting that an inductance measurement at a single frequency is valid. It isn't. Figure 1 implies a high frequency resonance, resulting in only a small region that could be described as being inductive, and until a full impedance against frequency graph has been plotted, you don't know where it is; see **Figure 2**.

Figure 2 and its derived values were obtained as follows: The transformer secondary was short-circuited, and primary impedance was measured for various frequencies from 100Hz to 200kHz using a Hameg 8118 LCR bridge reporting directly to a spreadsheet. Next, the calculated impedance of the four component model from Figure 1 was overlaid, and model component values iteratively adjusted to minimise the discrepancy between model and measurement, at which point, the leakage inductance of 11.5mH had been determined. It should be emphasised that this particular example was chosen because it had a good fit between measurement and simple model – most are worse.

Primarily considering small ferrite-cored transformers, Wilson [1] proposed splitting the single leakage inductance model into a pair of inductances, one having a parallel resistance. This model, shown in **Figure 3,** effectively adds a loss resistance whose tapping point along the low frequency leakage inductance can be varied, and has also proved useful for iron-cored transformers.

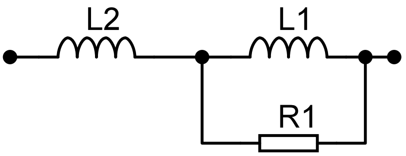

Figure 3 Tapping a transformer's leakage inductance to add loss resistance.

Because we are interested in the interaction between a rectifier and its transformer, we must measure the leakage inductance seen looking into the secondary rather than primary. To measure leakage inductance looking into a secondary, we simply reverse the previous procedure – we short-circuit the primary and measure secondary impedance, see **Figure 4**.

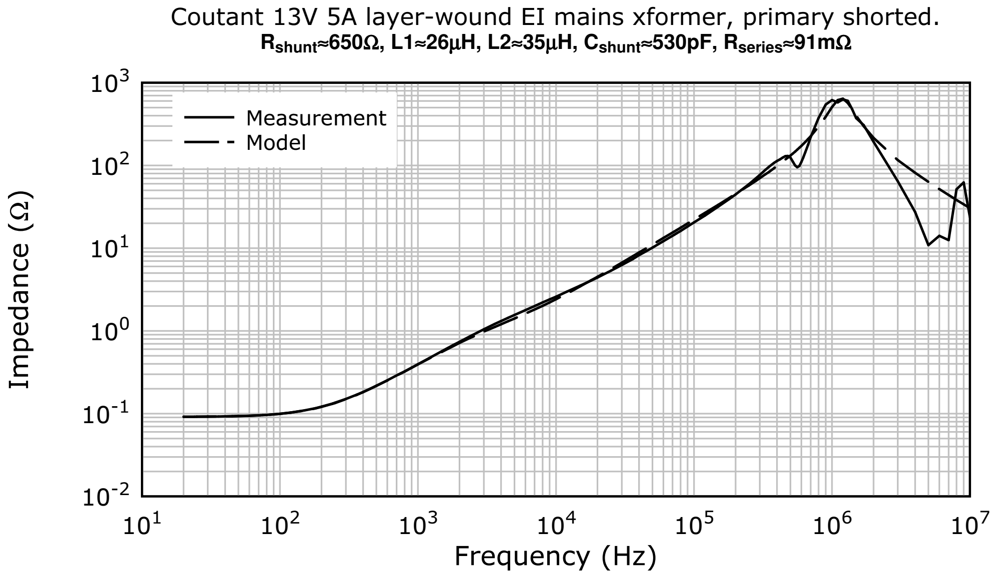

Figure 4 Impedance against frequency due to secondary leakage inductance.

If Figure 2 didn't demonstrate the futility of attempting to determine inductance from a single frequency measurement, Figure 4 should. Not only is there a main resonance at 1.1MHz followed by minor resonances, but there is a kink in the rising response centred around 6kHz that necessitated the six component tapped inductance model from **Figure 5**.

There are three common contemporary forms of construction for the small mains transformers used in audio:

- EI layer-wound
- Toroidal
- EI split bobbin

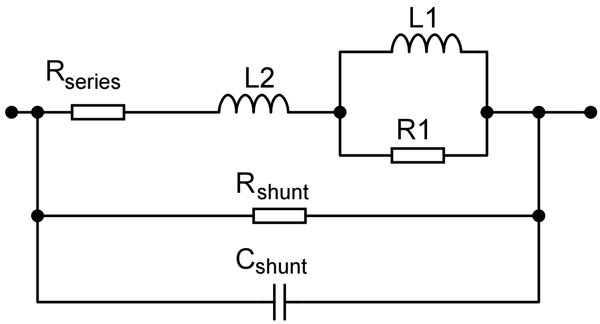

Figure 5 Combining tapped leakage inductance with the original four component model.

We have already seen an EI layer-wound transformer and might expect a toroidal transformer to have lower leakage inductance because of its good coil geometry and near-ideal construction, **Figure 6**.

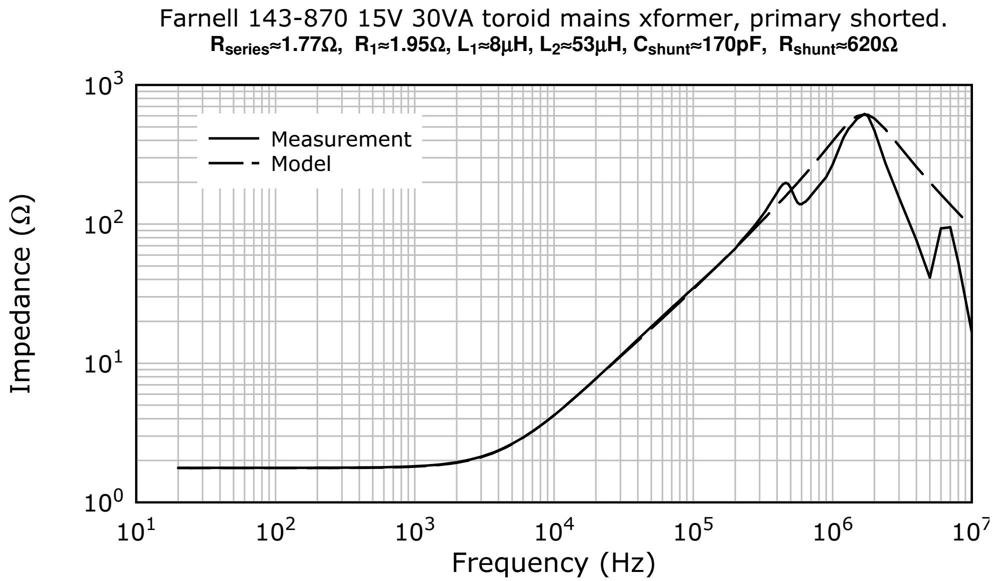

Figure 6 Toroidal mains transformer.

In practice, this toroid is surprisingly similar to the previous EI layer-wound transformer, with identical low frequency leakage inductance (61µH) and multiple resonances after its main 1.7MHz resonance.

We would expect a split bobbin EI transformer to have high leakage inductance due to its deliberately poor coil geometry, as shown in **Figure 7**.

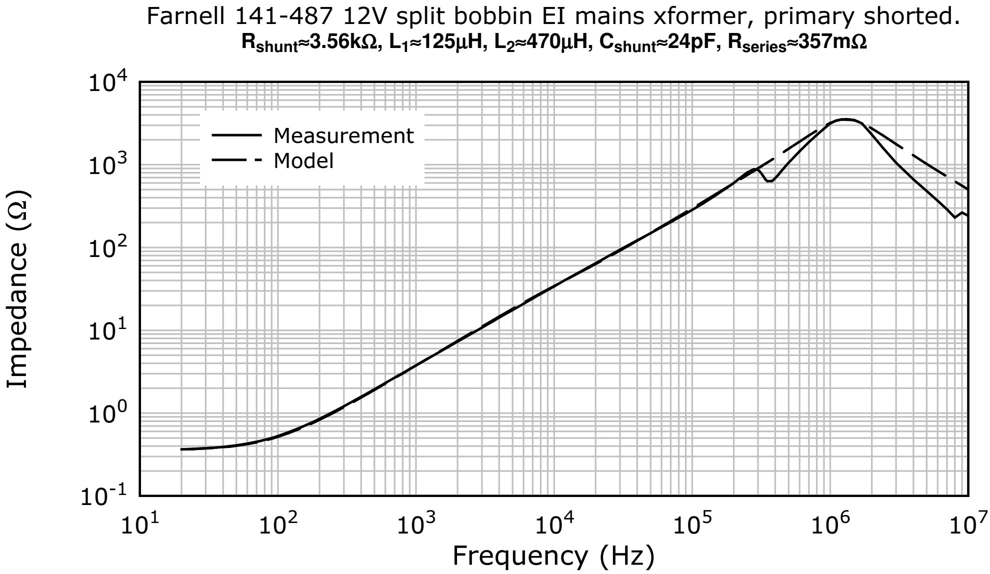

Figure 7 EI split bobbin.

The split bobbin EI transformer meets expectations by having low frequency leakage inductance (595µH) almost an order of magnitude higher than the preceding transformers. However, the enforced low capacitance to the primary significantly attenuates resonances beyond the 1.3MHz main resonance.

Surprisingly, these detailed measurements of the three constructions show that the simple model of leakage inductance and series resistance represents a low voltage winding quite well up to about 1MHz, despite wildly different values for leakage inductance and shunt capacitance. The source of the ≈400kHz inflection is unclear.

I had hoped that by measuring transformers of different constructions and secondary voltages I might be able to offer an empirical formula for estimating leakage inductance, but after measuring twenty transformers, nothing firm could be deduced from the results. If we need to quantify an iron-cored transformer's leakage inductance, only a series of impedance measurements will do.

Measuring impedance against frequency

Although a dedicated impedance analyser certainly speeds measurement, the cost of an analyser that measures >200kHz is prohibitive. Fortunately, a cheap function generator and digital oscilloscope having automated measurements can do the job quite well as is shown in **Figure 8**.

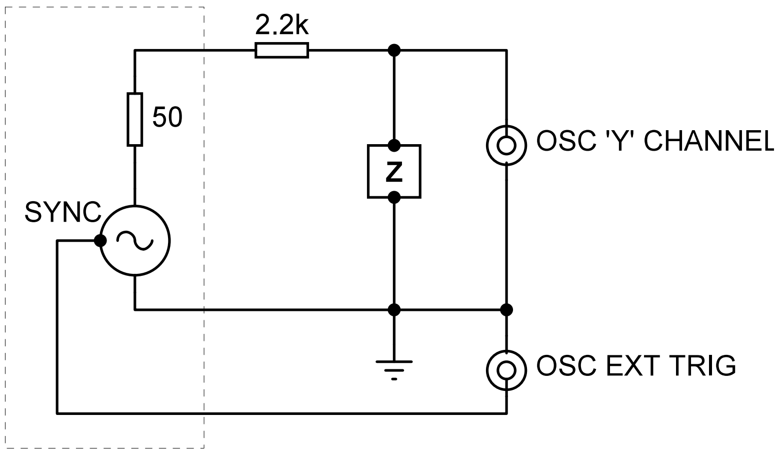

Figure 8 This simple set-up enables impedance to be measured over a wide range.

The function generator's open circuit sinusoidal output voltage V_{out} is set to its maximum and measured before connecting to the transformer.

The function generator drives the transformer's secondary via a 2.2kΩ series resistance. The oscilloscope is externally triggered from the generator's trigger output and measures V_{RMS} and frequency across the transformer secondary. The transformer's primary is short-circuited. The measurements are averaged over 32 traces to increase accuracy.

Amplitude measurements are then made at various frequencies and recorded in a spreadsheet as a function of frequency. The function generator's output resistance and 2.2kΩ series resistance form the upper leg of a potential divider, and because the generator voltage is known, the output voltage is known, and as the source resistance is known, the load impedance can be calculated.

Although 2.2kΩ is suggested as a series resistance, the real requirement is that the series resistance should cause the transformer voltage at resonance to be roughly half the generator's open-circuit output voltage. Derived result accuracy suffers as this voltage is exceeded because small changes in amplitude imply progressively larger impedance changes. Conversely, if the voltage at resonance is too small, derived result accuracy suffers as signals fall into the noise floor. The ideal methodology is to use a variable resistance, quickly sweep frequency whilst observing transformer voltage, adjust the variable resistance to achieve roughly half open-circuit amplitude across the transformer at resonance, replace the variable resistance with the nearest convenient standard value, and then perform the formal measurement sweep.

My wideband measurements were made <20kHz with my Hameg 8118 LCR bridge, but >20kHz measurements were achieved using the function generator and oscilloscope technique.

Leakage inductance summary

Although a useful theoretical concept, leakage inductance can only be determined accurately by fitting a model to measurement results of an impedance sweep against frequency.

From low frequencies to a little above resonance, the impedance deviation between optimum model and measurement using the four component model is usually within ±30%, and sometimes as little as ±5%, whereas the six component model can invariably be fitted to within ±10% and sometimes ±1%.

Determining transformer model values is not an end in itself – we require those values for an AC simulation. 30% impedance deviation corresponds to 3dB error, whereas 10% deviation corresponds to <1dB error. Thus, a carefully obtained six component model enables subsequent AC simulation having <1dB error, and therefore believable results.

Silicon diode depletion capacitance

When a semiconductor diode is reverse biased, a depletion region devoid of current carriers forms between the two contacts. The contacts have non-zero area and are either side of an insulator, creating capacitance ($\varepsilon_r \approx 12$ for silicon). It is well-known that the thickness of the depletion region is proportional to the square root of applied voltage, and this property is exploited in varicap (*vari*able *cap*acitance) diodes for tuning radios. Semiconductor manufacturer data sheets [2] commonly include graphs of capacitance against reverse voltage, as in **Figure 9**.

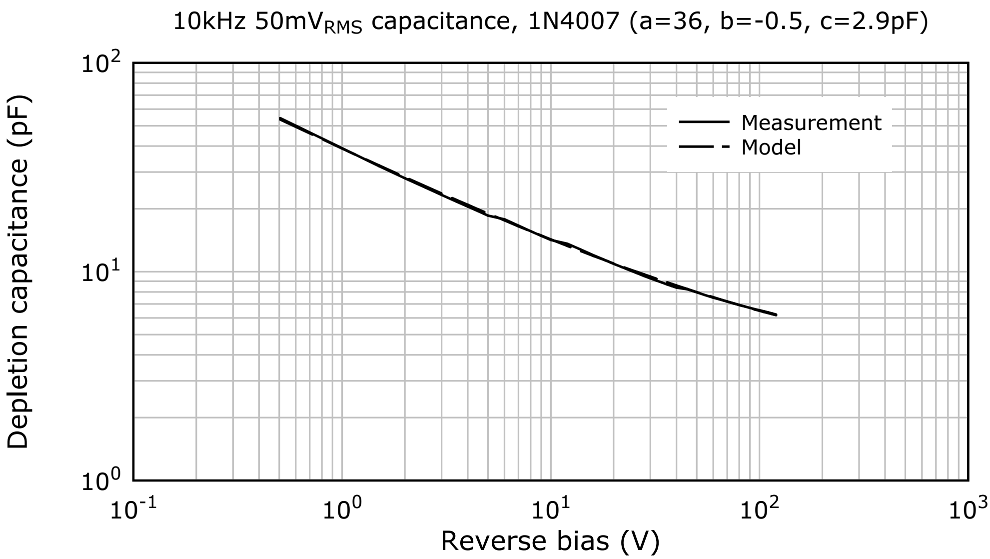

Figure 9 1N4007 capacitance against reverse bias.

Figure 9 shows measurement overlaid with modelled capacitance:

$$C_{(pF)} = \frac{36}{\sqrt{V_{reverse}}} + 2.9$$

The model fits measurement within ±2.5% over the plotted range; the 2.9pF constant is end-cap capacitance. However, neither Figure 9 nor semiconductor manufacturer data sheet graphs show the behaviour from small reverse voltages, through 0V, and up to conduction.

Figure 10 shows capacitance for as far as I could measure. At the same time, shunt resistance was measured, and taken together, the two graphs, Figure 10, and **Figure 11**, are enlightening.

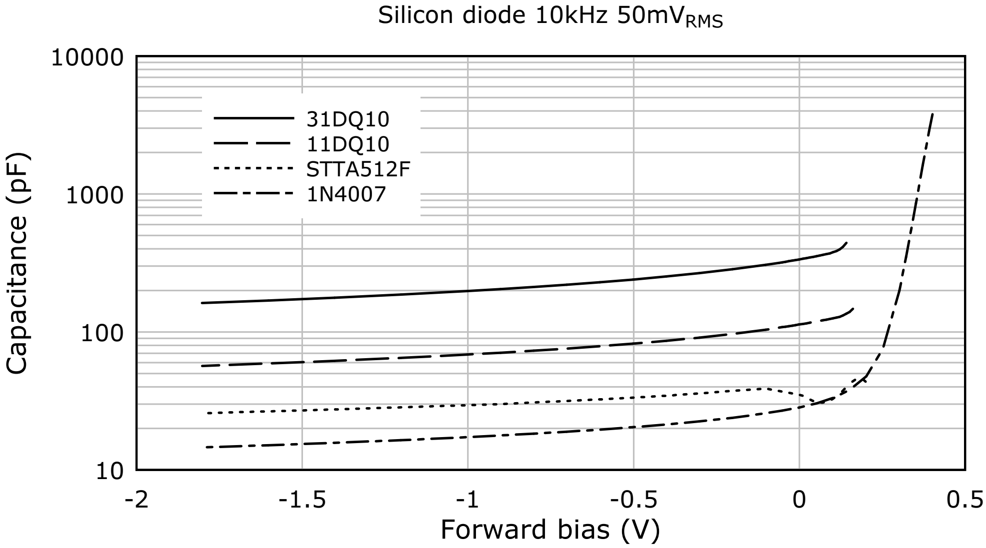

Figure 10 Capacitance comparison of Schottky (31DQ10, 11DQ10), soft recovery (STTA512F), and standard (1N4007) rectifiers.

We note that for all the diodes, capacitance increases and shunt resistance falls as conduction approaches, and that the STTA512F is quite leaky. The range of capacitance measurement towards conduction was limited not by absolute voltage but by falling shunt resistance, and because the 1N4007 has a much higher shunt resistance, its depletion region can be made much thinner (and therefore higher capacitance) before its shunt resistance falls sufficiently to render capacitance irrelevant. Thus, the 1N4007 effectively has a much larger (and more problematic) range of depletion capacitance.

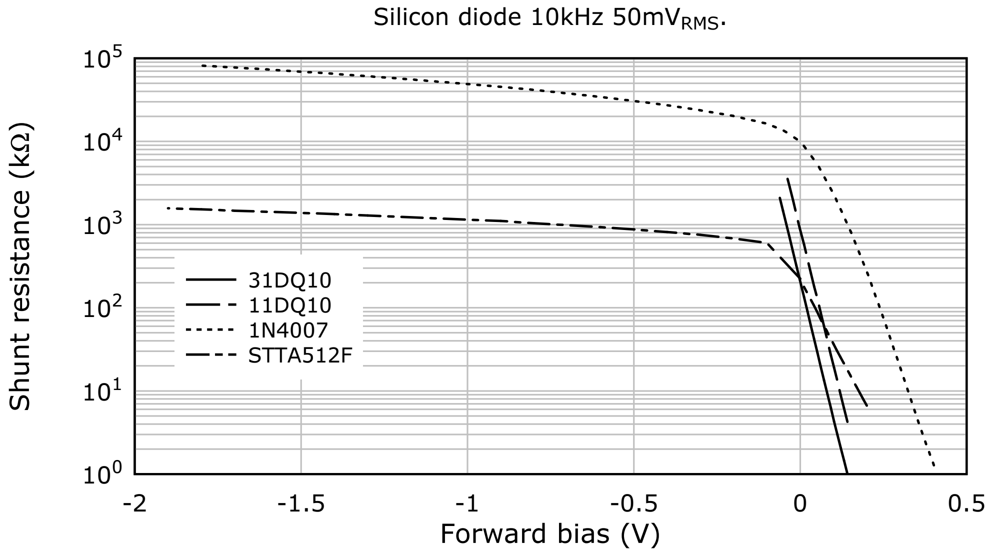

Figure 11 Shunt resistance comparison of Schottky (31DQ10, 11DQ10), soft recovery (STTA512F), and standard (1N4007) rectifiers.

Radio receivers, rectifier diode capacitance, and transformer leakage inductance

We have investigated transformer leakage inductance and semiconductor diodes. It is now time to take the results of those two investigations and combine them in an LTspice model. To do this, we initially make some simplifying assumptions:

- We are only interested in behaviour at the transition between the diode switching on and off.
- The transformer can be represented by a single leakage inductance and series resistance.

The first assumption returns us to our original question of whether or not rectifier diode capacitance can cause transformer leakage inductance to ring. The second assumption is used to avoid the possibility that a particular transformer's six component model might uniquely tip the results one way or the other.

The model tests a 12V power supply using an EI split bobbin transformer (therefore having high leakage inductance and low stray capacitance), standard 1N4007 diodes in a centre-tapped rectifier (and therefore only one diode in series at any one time), into a 15,000µF reservoir capacitor, loaded by a regulator (which is a constant current load, and therefore provides no damping); **Figure 12**.

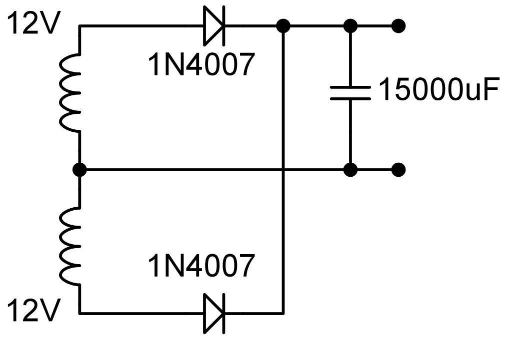

Figure 12 Simple power supply using full-wave rectifier.

Conversion into an LTspice model (**Figure 13**) requires the addition of parasitic components to the reservoir capacitor, such as 30mΩ equivalent series resistance, and 20nH series inductance. Because we are only interested in behaviour when the diode is off, we represent the diode as depletion capacitance and shunt resistance. We represent the transformer as a voltage source in series with leakage inductance and series resistance.

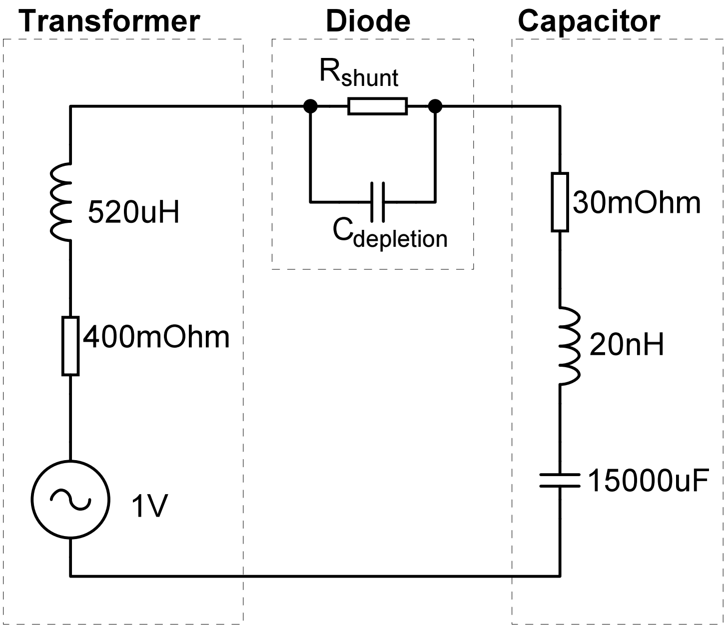

Figure 13 LTspice model of Figure 12.

We use a 1V interference source in order that the numerical results should directly represent attenuation, and perform six tests using different depletion capacitances and shunt resistances selected from the previously measurements at different diode voltages, **Figure 14**.

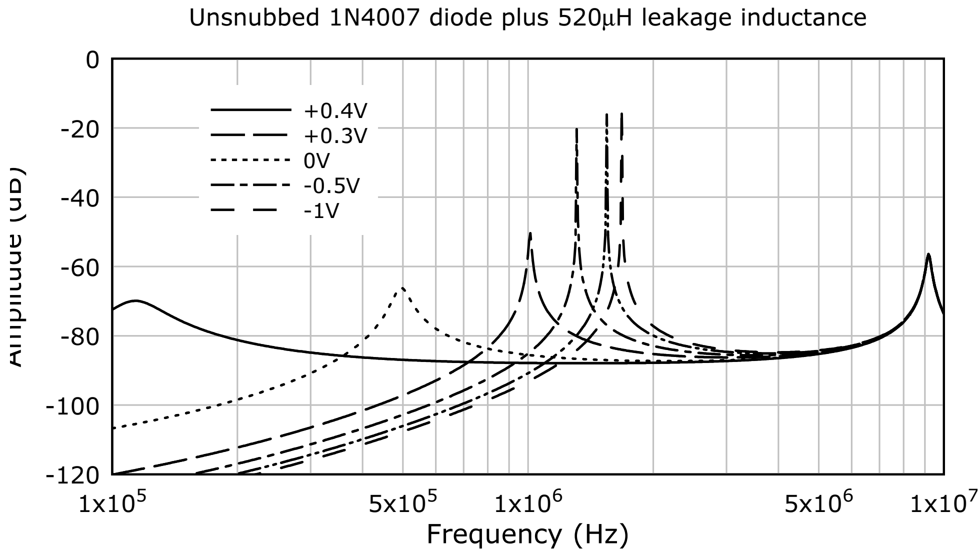

Figure 14 Modelled response of unsnubbed 1N4007 plus 520μH leakage inductance for various diode voltages.

Figure 14 is far less alarming than at first appears. The fact that LTspice *could* produce a frequency sweep without crashing answers the first of our questions at the beginning of this article – this transformer doesn't ring. Judging from this explicit investigation and my previous experience, it appears that a faulty transformer is required to provoke ringing.

However, what Figure 14 does show is that as the rectifier diode switches, it tunes a resonant circuit through the medium and long wave broadcast bands, and since mains wiring is an effective aerial at these frequencies, that can't be good, so we typically snub the 1N4007 diode with a 47n capacitor, **Figure 15**.

Figure 15 shows that a 47nF snubbing capacitance replaces the sweep through broadcast bands with a slightly attenuated but stationary peak at 32kHz. Worse, a low voltage transformer (<20V) will definitely be inductive at such a low frequency, so the peak *will* occur.

Rather than connecting an arbitrarily large capacitor across each diode, perhaps we could be a little more scientific and choose an optimum value? A good value might be five times depletion capacitance, but as 1N4007 capacitance changes viciously with voltage, what might work? Well, you have to start somewhere, and I had plenty of 1nF capacitors. Having chosen a capacitance that will

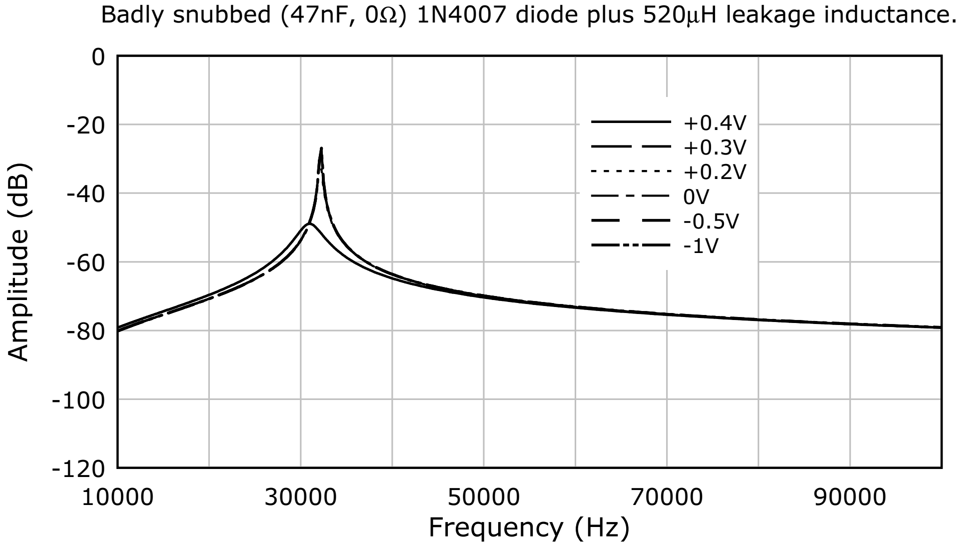

Figure 15 Typical capacitor-only snubbing.

swamp diode capacitance for most of its range, we can calculate the Q of the resulting resonant circuit using:

$$Q = \frac{1}{R}\sqrt{\frac{L}{C}}$$

Alternatively, we could state that we require Q = 0.707 (maximally flat response) and rearrange the equation:

$$R = \frac{1}{0.707}\sqrt{\frac{L}{C}} = \sqrt{\frac{2L}{C}} = \sqrt{\frac{2 \times 520 \times 10^{-6}}{1 \times 10^{-9}}} \approx 1k\Omega$$

The equation suggests that the resonance formed by the diode snubber capacitance and transformer leakage inductance could be damped by 1kΩ series resistance, so we add this to the model and the result of analysing this appears in **Figure 16**.

Note that not only does a damped smaller snubber capacitance leave the resonance further away from the audio band, but it reduces its amplitude by >40dB, rendering it entirely negligible. Note also that there is still a momentary (heavily damped) resonance at 100kHz just before the diode conducts, due to the very high 1N4007 capacitance just before conduction.

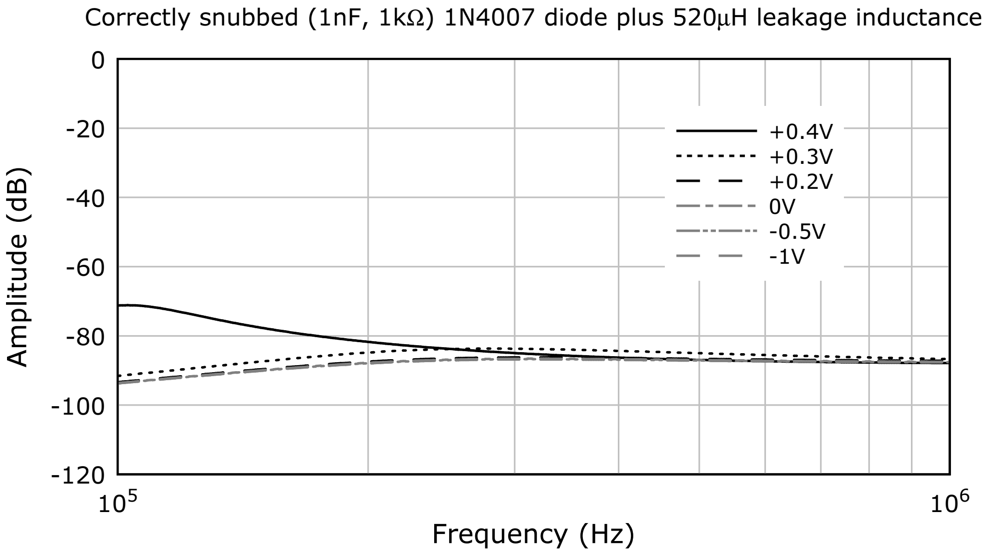

Figure 16 Correctly snubbed 1N4007 diode plus 520μH leakage inductance.

Schottky diodes

Schottky diodes are a popular (although expensive) substitution for the low voltage 1N4001. The argument runs that because conduction in a Schottky diode relies on electrons (rather than holes), there is no current overshoot to cause ringing. I substituted a 31DQ10 Schottky diode into the model, **Figure 17**.

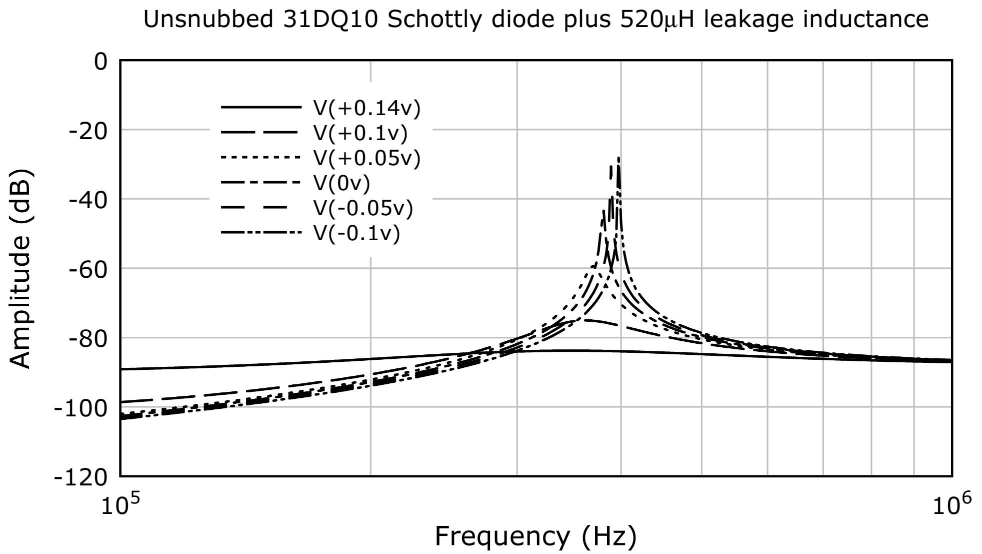

Figure 17 Unsnubbed Schottky plus 520μH leakage inductance.

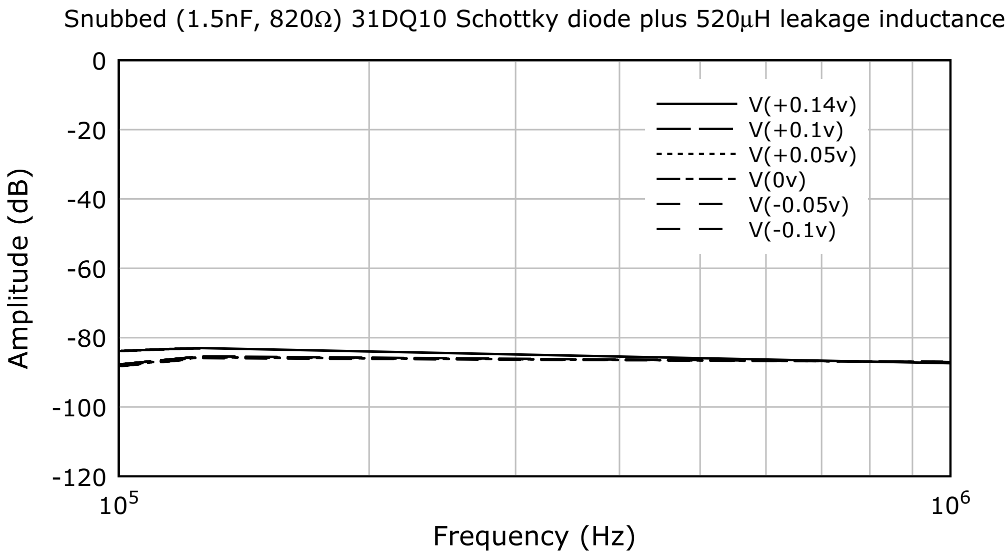

Figure 18 Snubbed 31DQ10 Schottky plus 520µH leakage inductance.

Note that the main difference between Figure 17 and Figure 14 is *not* reduced peak amplitude, but a reduction in the swept frequency range due to the Schottky diode's smaller variation of capacitance. Obviously, we could snub the Schottky, and if we assume 300pF diode capacitance (from Figure 10), we would need a 1.5nF snubber and 820Ω resistance, resulting in **Figure 18**.

If we compare Figure 18 (Schottky) with Figure 16 (1N4007), we have to look quite hard to see any difference, so provided that the diode is properly snubbed, the Schottky expense (and fragility) seems unjustified.

Higher voltages

The previous investigations focussed on low voltages, but what about voltages for valves? Rather than investigating 12V, we should investigate 250V. As mentioned in the leakage inductance section, higher voltage transformers have much higher leakage inductance, but they also have higher series resistance, and HT capacitors are smaller, perhaps 47µF, and so we obtain **Figure 19**.

As before, an unsnubbed 1N4007 rectifier sweeps across a range of broadcast frequencies but the amplitude is ≈60dB lower, necessitating graph rescaling (which is maintained for these higher voltage comparisons). Using a 1nF snubber,32mH leakage inductance requires 8k2 damping resistance.

Figure 20 shows that although a 1N4007 can be snubbed effectively, the necessarily large snubbing capacitance increases low frequency interference. Perhaps the (lower capacitance) STTA512F would be better?

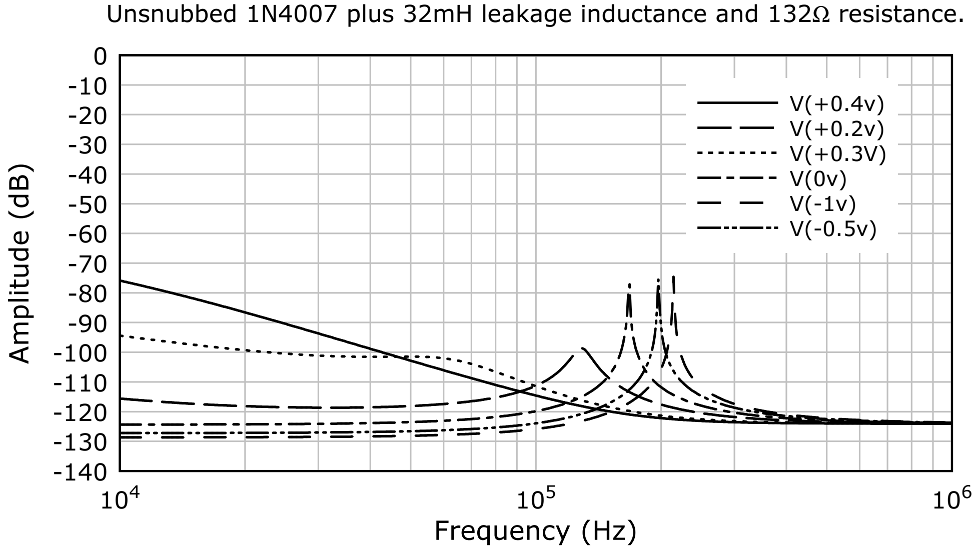

Figure 19 Unsnubbed 1N4007 behaviour in an HT supply. Note that the substantially increased leakage inductance lowers resonant frequencies by a factor of almost ten compared to Figure 14, but also that the greatly increased transformer series resistance reduces peak amplitude.

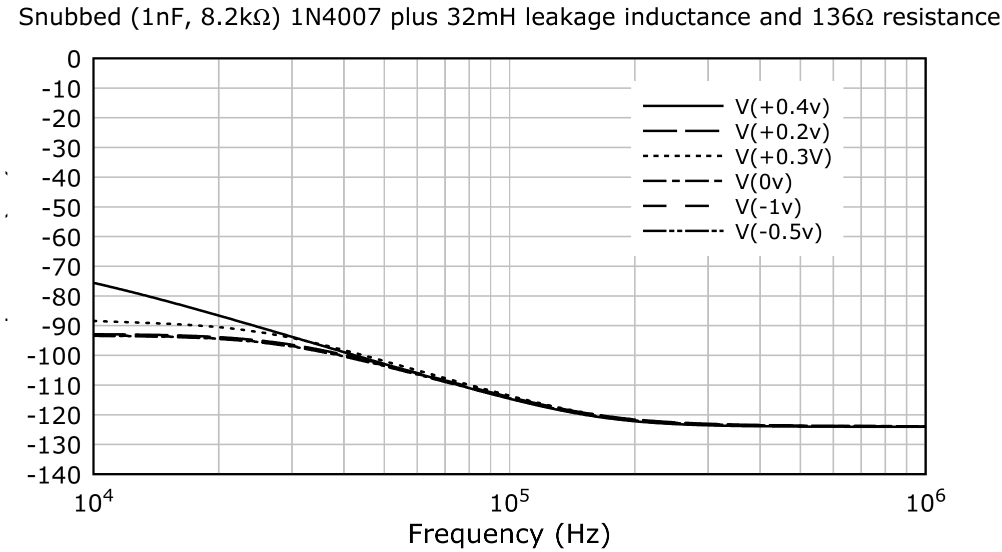

Figure 20 Snubbed 1N4007 as an HT rectifier.

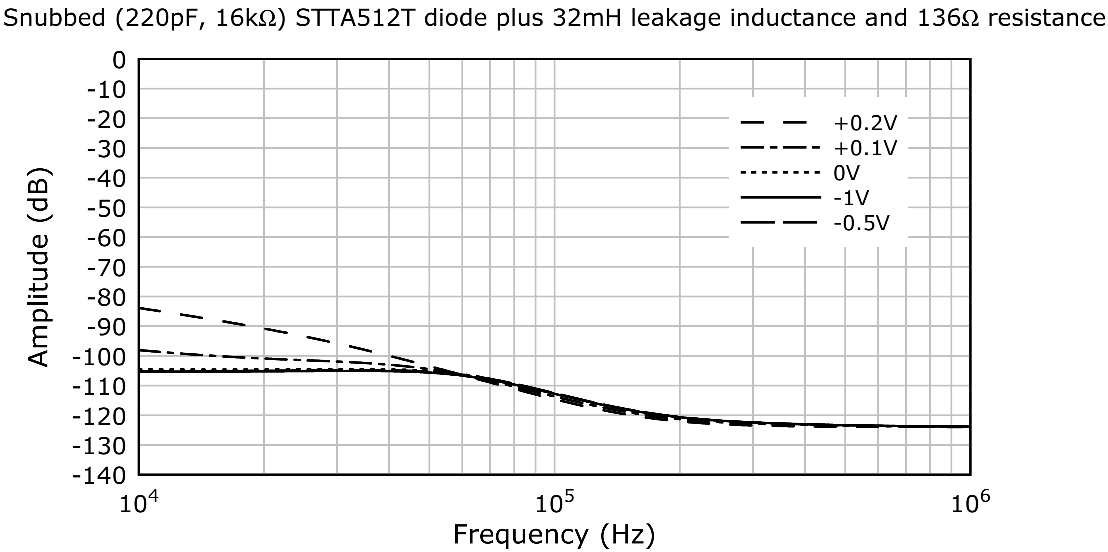

Figure 21 STTA512F as an HT rectifier.

Using an arbitrary (but useful) five times $C_{depletion}$ ratio, the reduced capacitance (<40pF) of the STTA512F permits a 220pF snubber and makes the combination >10dB quieter than the 1N4007 (**Figure 21**). Note also that the frequency of the damped resonance has risen from 30kHz to 60kHz, pushing it further away from the audio band.

For completeness, we should investigate a valve rectifier, such as the EZ81. C_{ak} is generally only specified for efficiency or damper diodes, so the EZ81 was measured to be 6.5pF (cold) and this seems to be typical for hard vacuum rectifiers. There's no reason to suppose that this capacitance will change either with voltage or with the heater on, so this single value was used. My nearest standard value of snubber capacitor was 33pF 1kV, and this would need a 47k damping resistor.

As **Figure 22** shows, a snubbed valve rectifier is very quiet, purely and simply because its capacitance is so low that it only needs a very small snubber capacitance.

The six component model

Analysis up until now has assumed a single leakage inductance and associated resistance exhibiting no resonance, but as secondary voltage increases, leakage inductance and stray capacitance increase, making it progressively more likely that the main transformer resonance will appear below 1MHz.

The EZ81 analysis was repeated using a 250V secondary that exhibited a resonance at 36kHz and needed a six component model, see **Figure 23**.

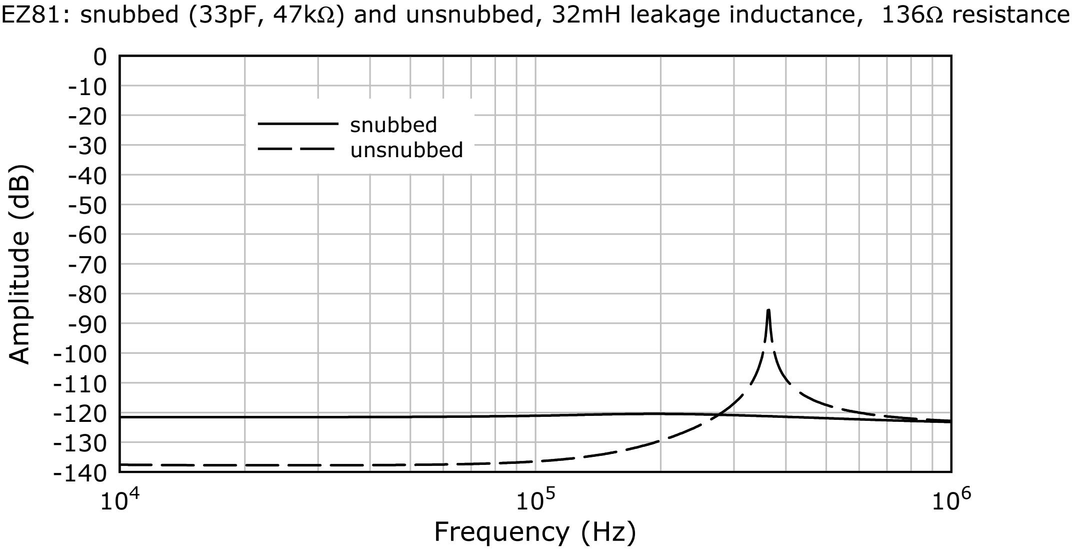

Figure 22 EZ81 valve rectifier, snubbed and unsnubbed. Adding a snubber eliminates the 362kHz peak at the expense of lifting broadband interference by 16dB.

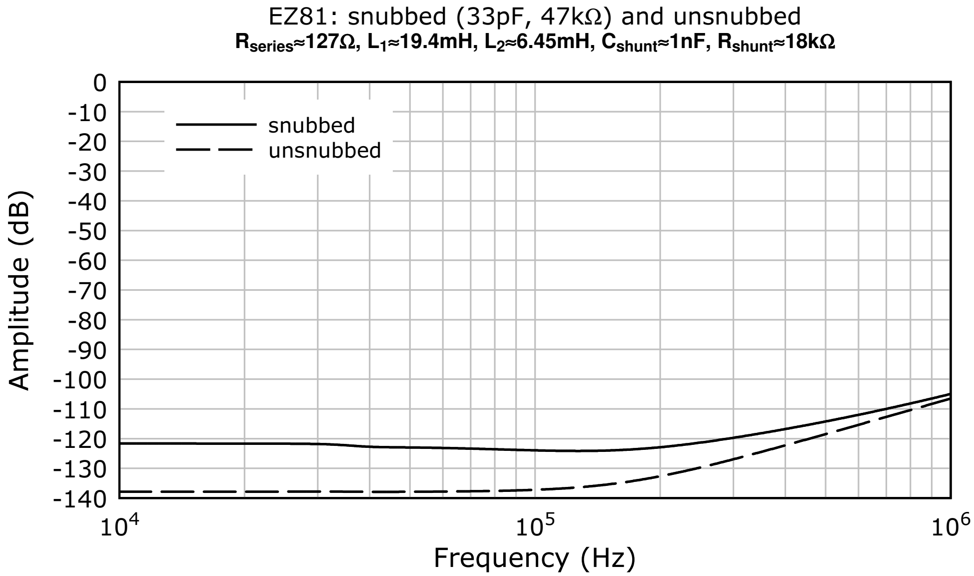

Figure 23 Note that in this instance, there is no peak, but the snubber still lifts broadband interference by 16dB.

The reason that Figure 23 shows no peak is as follows. Resonance between diode capacitance and transformer secondary can only occur at frequencies where secondary impedance is inductive. If diode capacitance is low, the theoretical resonant frequency combined with leakage inductance is above the transformer's own self-resonant frequency, the transformer appears capacitive, and swept resonance cannot occur.

Finally, the full six component model for the 12V EI split bobbin transformer first seen in Figure 7 was modelled with the results presented in **Figure 24**.

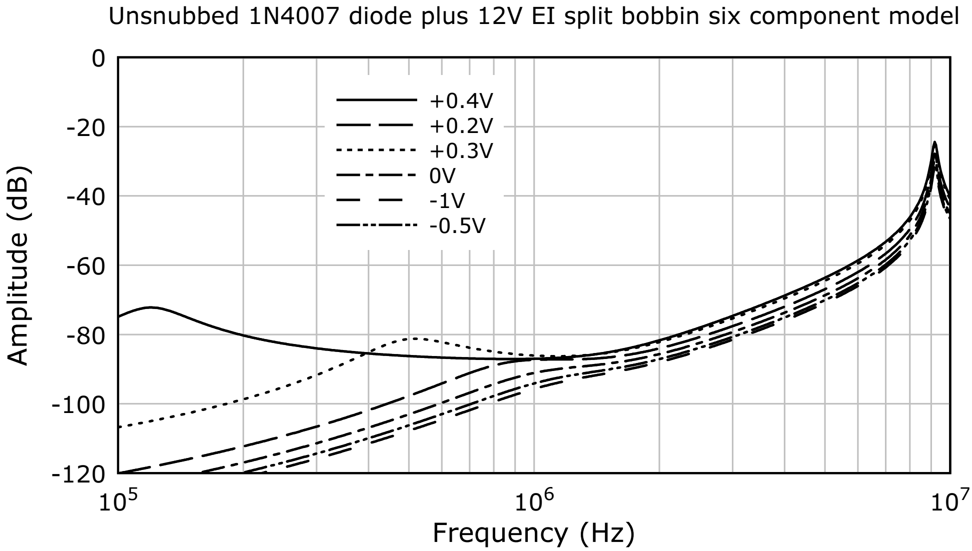

Figure 24 Note that despite the absence of an explicit snubber, there is only a hint of the swept resonance problem seen in Figure 14.

The transformer in Figure 24 is on the cusp of exhibiting swept frequency resonance and *might* benefit from a carefully damped snubber. Note that because a full six component transformer model was used, plus a diode model including changing depletion capacitance and leakage resistance, Figure 24 is expected to be a close approximation to reality below 2MHz.

Snubber position

The traditional position for the diode snubbers is across the diodes – they're treated as being switches acting upon an inductance, so for a bridge rectifier, we need four snubbers. But Figure 24 implies that all we really need to do is to make the transformer's leakage inductance look less inductive – by adding a parallel Zobel network. Perhaps we should change snubber position slightly and fit a single snubber/Zobel network across the transformer secondary?

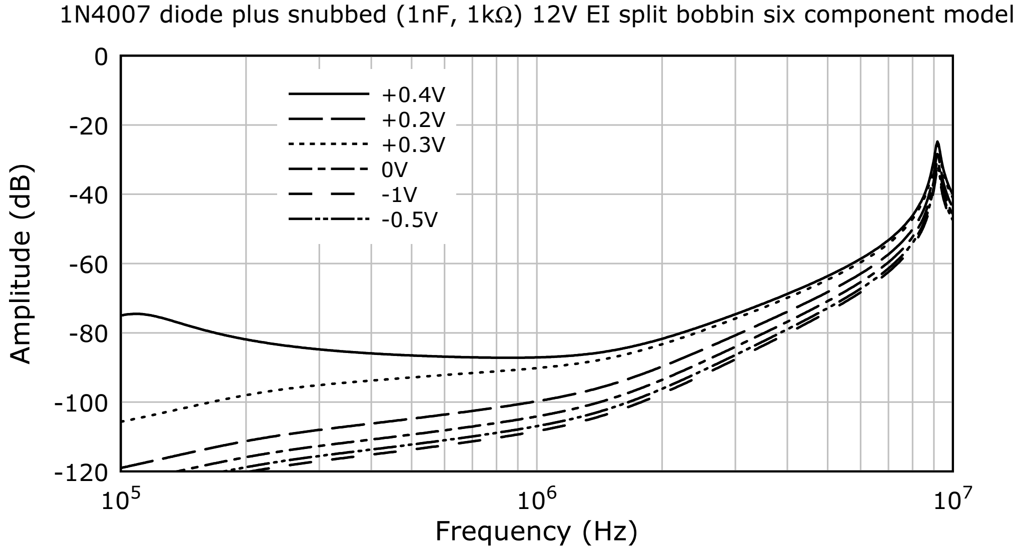

Figure 25 A single snubber directly across the transformer tames leakage inductance.

As hoped, **Figure 25** shows that the snubber works in its new position, with a saving of six components (assuming bridge rectifier); secondaries for centre-tapped rectifiers need two snubbers (one from each winding end to the centre tap). More significantly, it transpires that the new position makes it almost impossible to select snubber values that degrade performance, as witnessed by **Figure 26**.

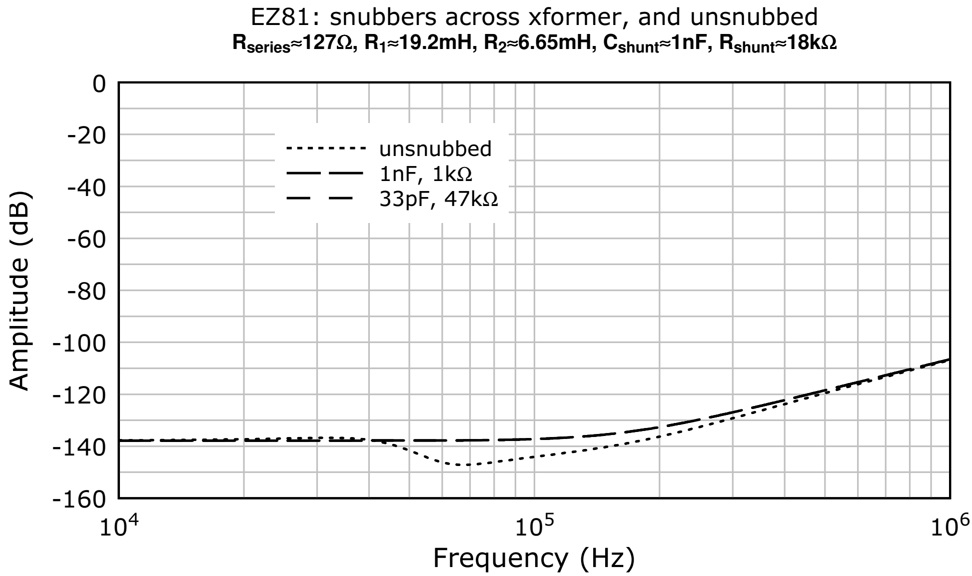

Figure 26 Once the snubber is across the transformer, its values become uncritical.

The EZ81 had previously been degraded 16dB by adding a snubber, but even applying wholly inappropriate snubber values (1nF, 1k) across either the simple LR or six component transformer secondary **improved** rather than degraded performance.

Conclusions

We have been chasing a demon that doesn't exist, then made matters worse by having the wrong parts in the wrong place – and too many of them. A single damped snubber/Zobel across each transformer secondary is all we need to suppress the swept resonance problem. Ideally, we would know diode capacitance, measure leakage inductance and calculate the correct resistance to set Q = 0.7, but even if we don't, 1nF and 1k will almost certainly do in both high and low voltage supplies.

References

[1] Wilson P R, "Effective Modelling of leakage Inductance for use in Circuit Simulation"
[2] Vishay Semiconductors, "1N4001 thru 1N4007 (General Purpose Plastic Rectifier)"