MAKING SENSE OF CURRENT SENSE TRANSFORMERS

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ABSTRACT

Current sense transformers are commonly used for monitoring, control and protection and can be as simple as a single winding on a toroidal core to sense the current in a wire passing through the core. This article describes some pitfalls to avoid and techniques to employ to achieve best performance. A high frequency current transformer is used as an example but the principles apply equally at low frequencies.

ARTICLE

Units – throughout this article, units in the equations are: microseconds, amps, volts, teslas, square millimetres, farads, ohms, henries,

A current sense transformer (CT) operates in exactly the same way as any magnetic transformer and follows the basic rules that voltage and current are transformed in the ratio of the primary and secondary winding turns, load impedance is transformed in the ratio of the square of the turns and the volt-second product applied to any winding must average zero over each switching cycle.

Shown in a schematic, a CT will look like Figure 1 as a simple two-winding ideal transformer with infinite secondary inductance, zero winding resistances, perfect coupling and no parasitics. As such, if terminated with a pure resistance, it would have a response from the slowest varying DC to any high frequency at any current with no losses or distortion. A real CT shown in Figure 2 has a finite value of magnetising or winding inductance L*m* and the parasitic components of leakage inductance L*lk*, winding resistances R*p* and R*s*, coupling capacitance C*c*, winding self capacitances C*p* and C*s* and core loss R*loss*. R*f* is the secondary impedance reflected to the primary. The core can also magnetically saturate. These practical characteristics, interacting with the external circuit, limit the performance of the CT which can be summarised by its *accuracy* , $\frac{V_b}{I_t}$ or ability to reproduce a perfectly scaled representation in voltage, V*b*, of the primary current I*t* when terminated with a resistance R*b*.

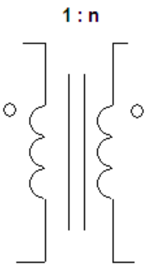


Figure 1. An 'ideal' CT

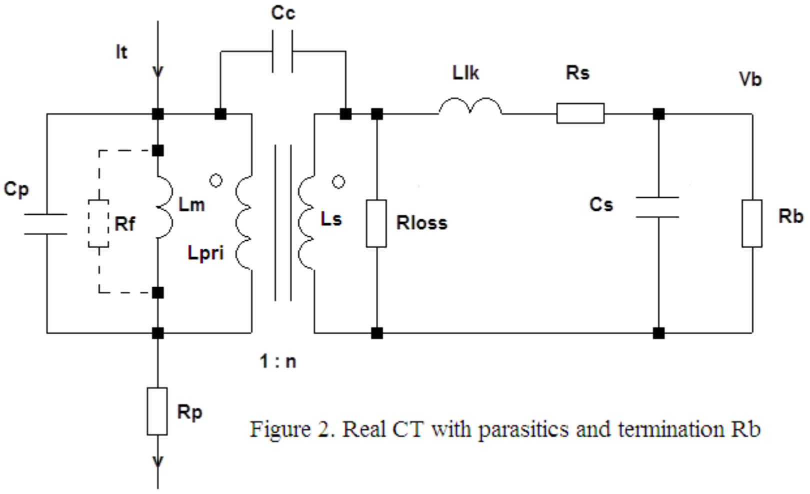


Figure 2. Real CT with parasitics and termination Rb

Low frequency accuracy is defined by the magnetising inductance L*m*, secondary winding resistance R*s* and burden resistor R*b*. This is because R*s* in series with R*b* reflects to the primary of the transformer in the square of the turns ratio *n* and appears in parallel with L*m* as:

$$R_f = \frac{R_b + R_s}{n^2}$$

The sensed current I*t* therefore divides between R*f* and L*m* with only the partial current through R*f* reflecting to current through R*b*, indicating a value less than expected. As frequency reduces, so does the impedance of L*m*, increasing the error. Current through L*m* does not appear in the secondary of the transformer.

So how much current is ‘lost’ through L*m*? If the measured current is a sine wave then the current splits in the ratio of the complex impedance R*f*, to the impedance of L*m* = ωL*m*. Manipulation of the relationships gives the following result for accuracy, $\frac{V_b}{I_t}$

$$\frac{Vb}{It} = \frac{R_b}{n}\left[\left(1+\frac{L_{lk}}{n^2L_m}+\frac{R_s}{R_l}+\frac{R_b}{R_l}+\frac{\omega^2C_sR_bL_{lk}}{R_l}+\frac{C_sR_bR_s}{L_mn^2}\right)+j\left(\omega R_bC_s+\frac{\omega R_bC_sL_{lk}}{n^2L_m}+\frac{\omega C_sR_bR_s}{R_l}+\frac{\omega L_{lk}}{R_l}-\frac{R_s}{\omega L_mn^2}-\frac{R_b}{\omega L_mn^2}\right)\right]^{-1} \quad \text{Eq. 1}$$

Inspection of the equation shows that accuracy tends to $\frac{R_b}{n}$, the ideal result, at mid frequencies where $\omega L_m n^2 >> R_s$ or R_b and if C_s, L_{lk} are small and R_{loss} high. At lower frequencies, the terms $\frac{R_s}{\omega L_m n^2}$ and $\frac{R_b}{\omega L_m n^2}$ begin to dominate and accuracy suffers. However, this can be counteracted by reducing R*b*. This is shown graphically in Figure 3 using the Murata Power Solutions 53200C 200:1 CT as an example which has L*s* = 8 mH, C*s* = 30 pF, L*lk* = 50 µH and R*s* = 34 ohms. Accuracy is plotted for burden resistors R*b* of 50, 100 and 200 ohms with the Y axis volts per amp scale normalised to unity. The improvement in low frequency bandwidth with decreasing R*b* can be clearly seen. The phase advance from sensed current to indicated voltage is also plotted in Figure 4.

A low frequency bandwidth limit is also where the core begins to magnetically saturate. A version of Faraday's law can be used to derive flux density B from voltage, core cross sectional area A*e* and winding turns n:

$$B = \frac{V_{rms}.10^6}{4.44\,F\,A_e\,n} \quad \text{Eq. 2}$$

Substituting in our equation for accuracy, rearranging and ignoring high frequency terms, the maximum rms current before saturation occurs can be expressed as:

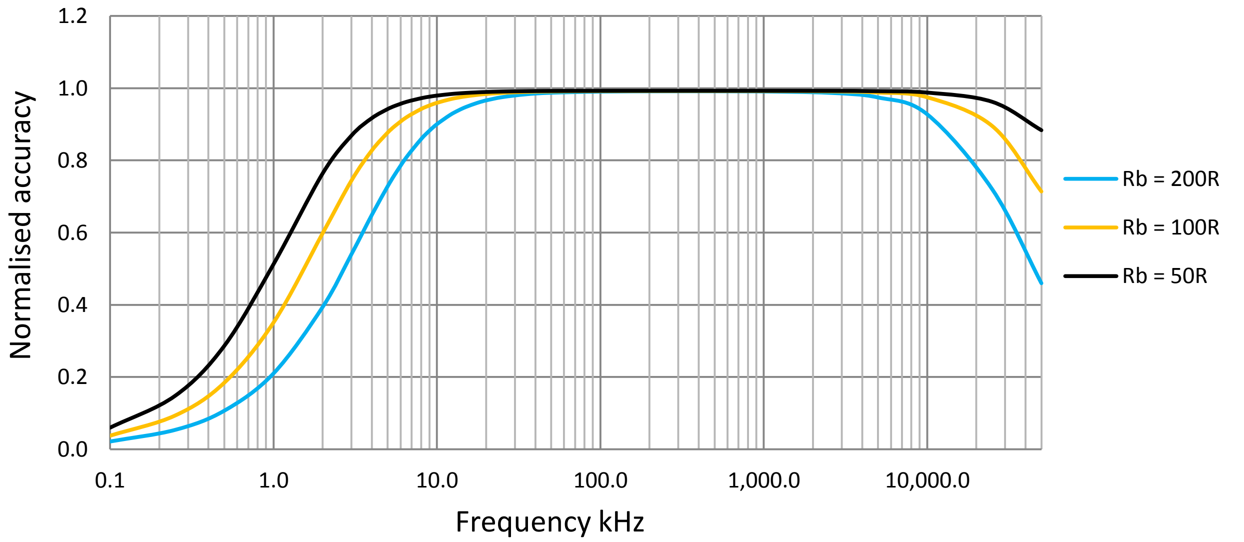
$$I_{tmax} = \frac{B_{sat}R_b\omega A_e n_{sec}}{\sqrt{2}Acc(R_b+R_s)10^6} \quad \text{Eq. 3}$$

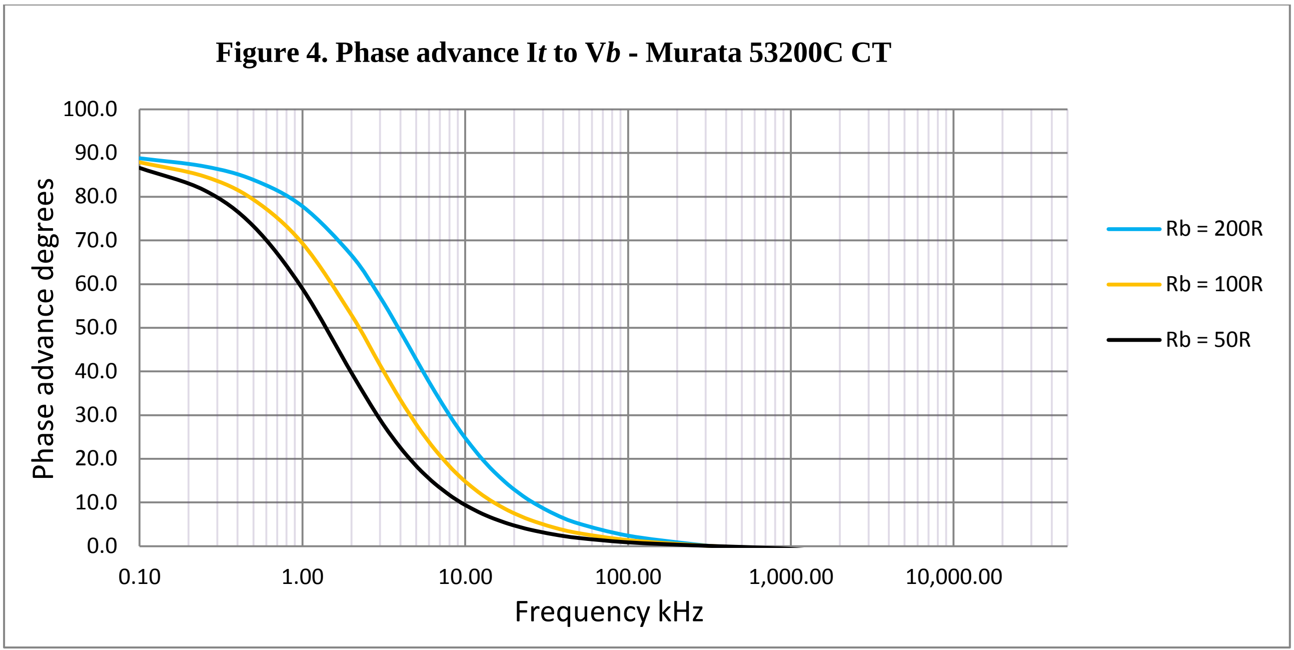
Where Bsat is the maximum flux density for the material used, typically 0.4 tesla and Acc is the accuracy computed in Equation 1 for the frequency in question, $F = \frac{\omega}{2\Pi}$. This is plotted in Figure 5 showing that saturation occurs well below the rated bandwidth but again improves with reducing R*b*. The maximum current of 10 A within the rated bandwidth of this product is a thermal limit. B*sat* does reduce with temperature, typically by about 20% from 25 Celsius to 100 Celsius which should be allowed for. Note that the sensed current itself does not directly cause saturation as it ideally transforms to burden resistor current rather than magnetising the core. It is the ‘diverted’ current into the magnetising inductance which causes saturation.

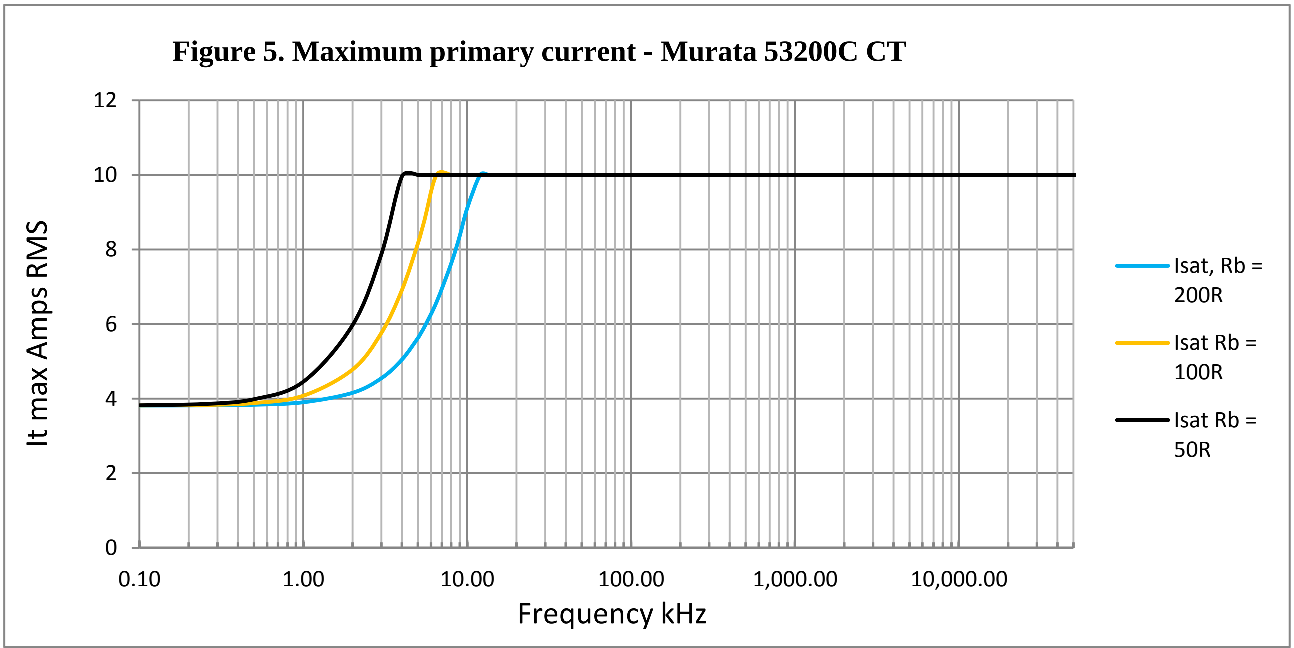
At higher frequencies, C*s* and the decreasing effective value of R*loss* start to shunt away current from R*b* and the accuracy falls. C*s* can be minimised by making the secondary winding a single layer which is easier with fewer turns, arguing for selection of CTs with lower turns ratios, perhaps 50:1 for best high frequency performance. Core losses depend on material and increase exponentially with frequency at constant flux density but are mitigated by the fact that flux density decreases linearly with frequency, everything else being equal. It can be expected that the permeability of the ferrite falls off with increasing frequency reducing inductance but this has minimal effect. Lower values of R*b* also improve the high frequency bandwidth as can be seen in Figure 3. It can be seen that the rated high frequency bandwidth of this part, 500 kHz, is very conservative with a useful response up to several megahertz especially with a lower burden resistor value.

It is clear that a low value of burden resistor directly improves low and high ends of bandwidth so what is the disadvantage of reducing R*b*? Simply less voltage is developed across it for the same sensed current perhaps making the circuit more prone to noise pickup. In some situations, CT reset time can be affected adversely, see later.

Figure 3. Bandwidth - Murata 53200C CT





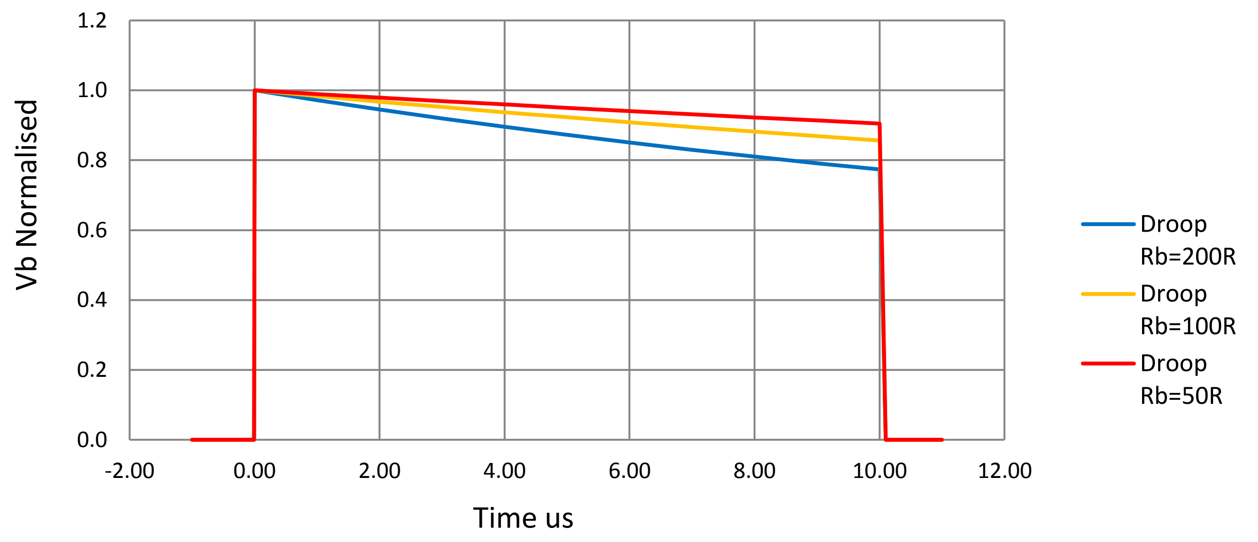


If the primary current is a unipolar or bipolar square wave, the 'lost' current into the magnetising inductance increases from zero with time according to I*m* = Et/L*m* where E is the decreasing voltage dropped across the corresponding decreasing current through R*f* with time, t. The effect is therefore not linear and V*b* follows the relationship:

$$V_b = \frac{I_t R_b L_{sec}}{n(t_{on}(R_b+R_s)+L_{sec})} \qquad \text{Eq.4}$$

The practical effect of this is to show an exponential 'droop' on the top of the voltage waveform V*b*. From the relationship it can be seen that reducing R*b* or increasing L*sec* lessens the amount of droop in a given time. Using the Murata 53200C CT again as an example, with a square wave sensed current of 10 μs duration, its recommended value for R*b* = 200 ohms gives 22.6% droop while R*b* = 50 ohms gives just 9.5%, shown diagrammatically in Figure 6.

Figure 6. 'Droop', Murata 53200C CT



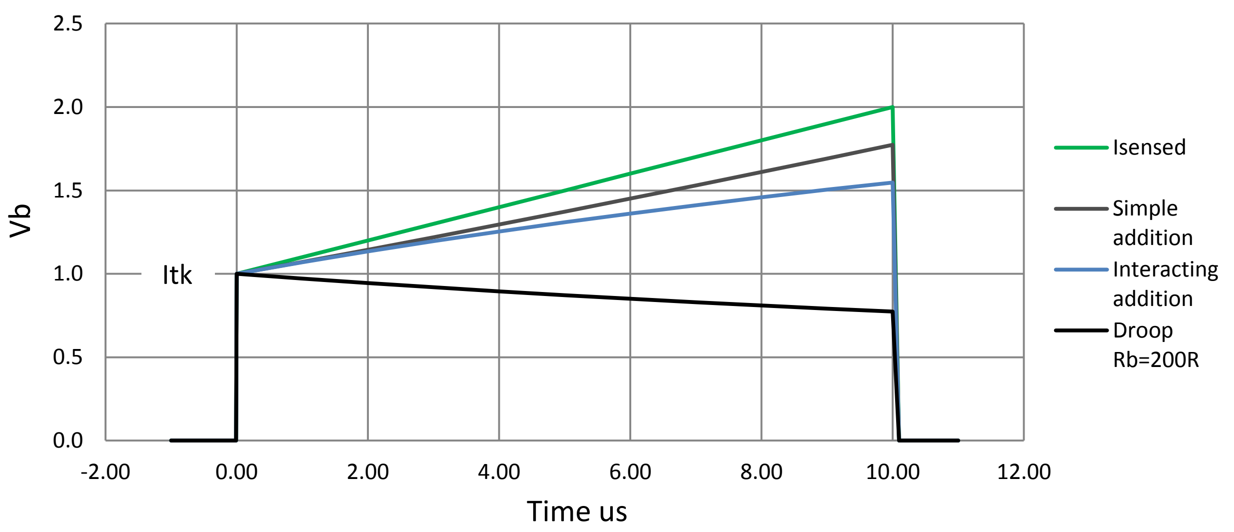
A more common sensed current waveform has a rising, approximately linear current superimposed on the square wave as shown in Figure 7. The rising current could be for example magnetising current in a flyback converter transformer or reflected output storage inductor current in a forward converter transformer. When the droop effect is added to these waveforms, it could be that the voltage across R*b* still shows a net fall representing a problem if the circuit is intended to detect overcurrent above a certain level. Also, in current mode control of switched-mode converters, droop must be limited to give a net positive slope for V*b*, necessary to initiate switch turn-off. Note that the amount of droop is in turn affected by the positive slope of the sensed current itself with the accuracy of the voltage V*b* now given by

$$V_b = \frac{(I_{tk}+\frac{di}{dt}.t_{on})R_b L_{sec}}{n(t_{on}(R_b+R_s)+L_{sec})} \qquad \text{Eq.5}$$

Where Itk is the starting value of the rising current with slope $\frac{di}{dt}$ (A/µs). Figure 7 shows an example of the 53200CT with a 1 A step current rising linearly to 2 A over 10 µs. The plot 'Droop, Rb = 200R' is that which would occur with a constant 1 A sensed current.

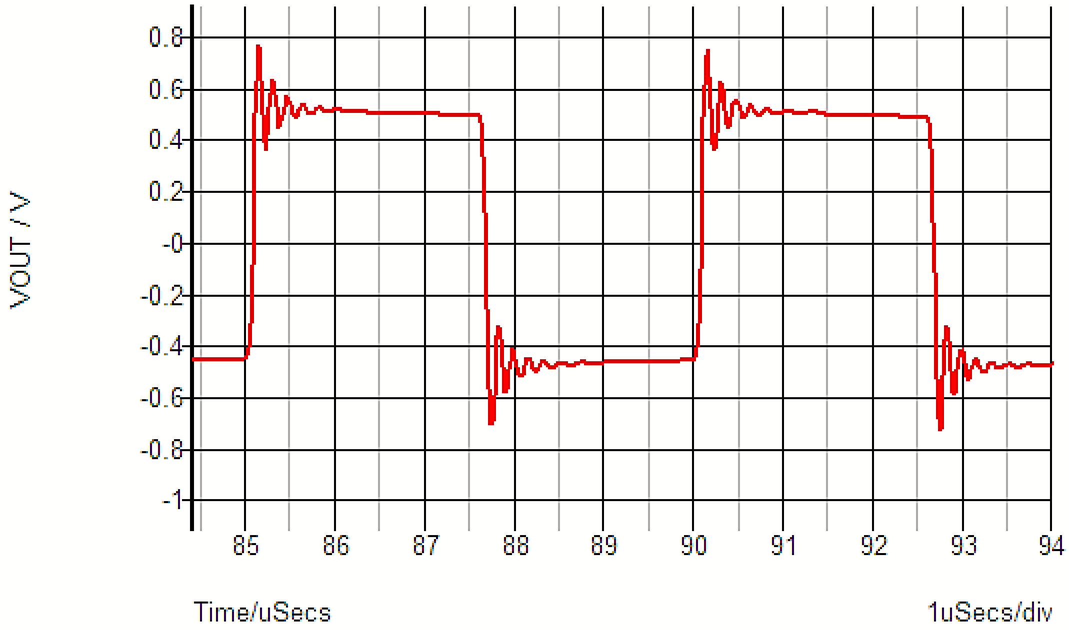
Applying Equation 5 shows that the resultant slope of the Vb waveform is actually lower that that given by simply adding the droop and rising sensed current calculated separately, requiring perhaps less droop to be tolerated to give a net positive slope for Vb if the $\frac{di}{dt}$ of the sensed current were less. Remember the droop is actually exponential so the slope of the waveform at Vb tends to a constant, higher value over time.

Figure 7 - Droop and Sensed current interacting - Murata 53200C CT



A more realistic waveform for a pulsed current with rising peak value is shown in Figure 8 with skew and ringing to the rising and falling edges.

Figure 8. Ringing and skew on edges



As expected from the analysis of the effect of R*b* on bandwidth, reducing R*b* reduces the edge skew somewhat. The inherent values of C*s* and L*s* have an effect although ringing is damped by R*b* and core losses. Skew caused by L*s* is the time it takes for the current through it, I*s*, to reach the settling value, nominally $\frac{It}{n}$. This in turn is driven by the voltage across L*s* during the skew time according to:

$$\frac{dIs}{dt} = \frac{E}{Ls}$$

E is driven from the primary voltage reduced by the turns ratio, varying and not very determinate, depending on the source impedance of the external driving circuit and its compliance voltage.

Coupling capacitance C*c*, rings with L*s* if the primary and secondary of the transformer share the same ground or if the grounds have significant mutual capacitance, which is commonplace. If this represents a problem in the circuit, toroidal current sense transformers with a single primary turn such as the Murata Power Solutions 5600C series can have lower values of C*c* than the bobbin-wound types such as their 5300C series.

With pulsed square wave currents, low end bandwidth, effectively the maximum pulse width allowed T*onmax*, is dependent on what droop or inaccuracy can be tolerated in the V*b* waveform and ultimately saturation of the transformer core given by:

$$T_{onmax} = \frac{B_{max}A_e}{V_p} \qquad \text{Eq. 6}$$

Primary voltage, V*p* depends on the value of R*b* chosen and sensed current I*t* and reduces as I*t* diverts increasingly into magnetising current with time. Including this effect T*onmax* for a square wave sensed current is given by:

$$T_{onmax} = \frac{B_{max}n^2A_en_{pri}L_{sec}}{I_tL_{sec}(R_b+R_s)-B_{max}n^2A_en_{pri}(R_b+R_s)} \qquad \text{Eq. 7}$$

A sensed current with rising value $\frac{di}{dt}$ from an initial value of I*tk* yields the quadratic:

$$T_{onmax}^2L_{sec}(R_b+R_{s)}\frac{di}{dt} + T_{onmax}\left(L_{sec}(R_b+R_s)I_{tk} - B_{max}n^2A_en_{pri}(R_b+R_s)\right) - B_{max}n^2A_en_{pri}L_{sec} = 0$$

Eq. 8

Using the example of the 53200C CT, for a sensed current of 10 A, T*onmax* calculates to be 23 µs, 40 µs and 63.5 µs with R*b* = 200, 100 and 50 ohms respectively. If the current has a rising value $\frac{di}{dt}$ of 0.1 A/µs from the initial value of 10 A, from Equation 8, T*onmax* reduces to 21.1 µs.

Even if a CT is used within its rated bandwidth and current, consideration must be given to the duty cycle of unipolar sensed current. Flux in any transformer must be allowed to return to zero or 'reset' between successive pulses otherwise residual flux adds on successive cycles and can ultimately cause 'staircase saturation' of the core. CTs fed with symmetrical sine waves or bipolar pulsed currents do not suffer from the problem as flux is actively driven to equal positive and negative values each switching cycle. The CT primary will often be coupled through a capacitor to guarantee that volt-second equality is always preserved and there is no net DC through the sense winding.

In the circuit of Figure 2, when the sensed current goes to zero, the energy in the core, $\frac{1}{2}L_mI_m^2$ generated by the magnetising current causes both winding voltages to reverse and the resistive path on the secondary winding provides a route for current to continue to flow and energy to dissipate in R*b*, R*s* and in core losses. The secondary current starts at value $(-)\frac{I_m}{n}$ and rises exponentially to zero with a time to a final current given by:

$$t_{final} = \frac{L_S}{R_S+R_b} ln\left(\frac{i_{initial}}{i_{final}}\right) \qquad \text{Eq. 9}$$

The reset time can be lengthy, for example for the 53200 CT with a burden resistor of 200 ohms, the time for the reset current to fall to 10% of its initial value is 78 µs. This time is so long that the effects of C*s*, R*loss* and L*lk* are negligible. This is an example of where reducing the value of the burden resistor does not produce a beneficial effect.

In practice unipolar current waveforms are sensed with the circuit of Figure 9. Here a diode blocks the reset current and R*r* can be a high value giving fast reset as long as there is also no reset current path around the primary of the transformer. The practical limit is that the secondary reset voltage increases in proportion and must be kept within the reverse breakdown limit of D1.

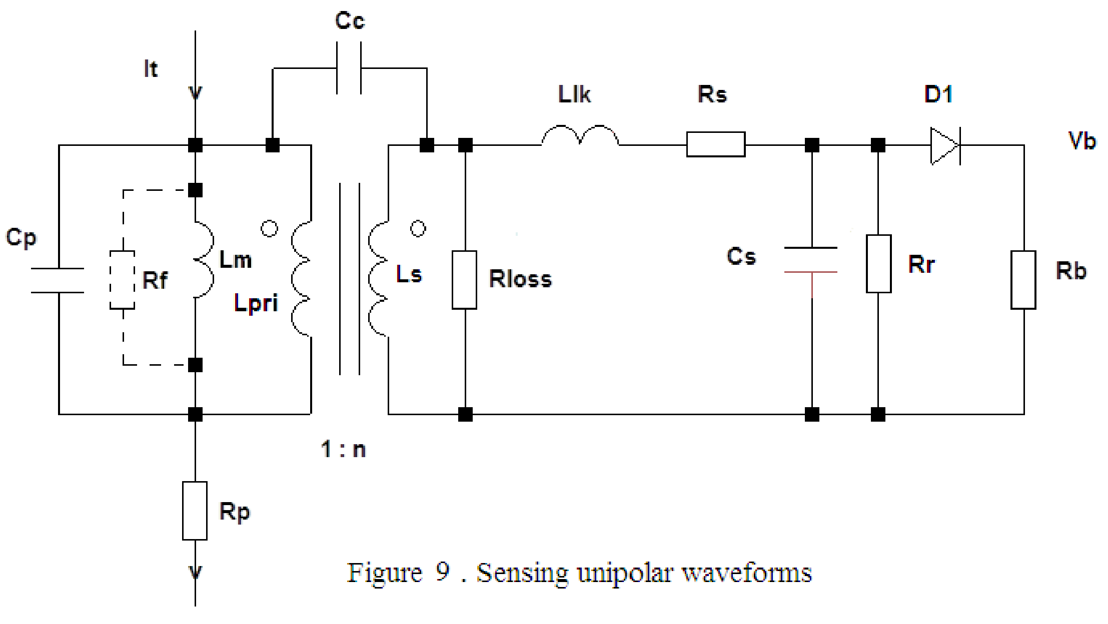


Figure 9 . Sensing unipolar waveforms

In real circuits, the secondary winding capacitance sinks current and limits the speed of reset and also limits the voltage during reset. If all of the magnetising energy from the transformer resonantly transfers to the capacitor, the maximum possible peak voltage V*r* is:

$$V_r = \frac{i_m}{n}\sqrt{\frac{L_S}{C_S}} \qquad \text{Eq. 10}$$

For our example CT, with Ls = 8 mH, n = 200, Cs = 30 pF and for a primary magnetising current of 2 A peak, the secondary reset voltage Vr would be 163 V.

It would be tempting to leave off Rr altogether and just ensure that D1 could stand 163 V reset plus any cathode voltage for minimum reset time. Remember however that the reverse voltage also appears on the primary winding by transformer action, in this case, $(-)\ \frac{163}{n} = -0.815\ V$

If for example, the CT were in the emitter of a bipolar transistor or low threshold MOSFET, held off by 0V on its base or gate, this voltage could turn on the transistor, possibly with disastrous consequences.

The effect is exacerbated by the fact that D1 adds to Rb voltage drop reflecting to the primary causing additional magnetising current and hence reset voltage according to Equation 10, compared with the bipolar sensing arrangement of Figure 2. A Schottky diode is often preferred for D1 for this reason but it itself has limited reverse voltage capability. A fast recovery diode should anyway be used as the recovery current acts to delay reset. As an example, Figure 10 shows the simulated resonant reset achieved with the circuit of Figure 9 with Rr = 10k ohms and with 1N4148 and 1N4003 diodes showing the effect of the long recovery of the 1N4003. The plot is the voltage on the secondary of the CT with a 10 V, 5 µs positive pulse followed by reset. Equation 9 gives the time for reset current to fall to a specified value with (Rb + Rs) replaced with
(Rr + Rs) so now with D1 fitted and Rr = 10k, under the same conditions, reset reduces from 78 µs to 1.8 µs.

Other reset schemes are possible including forcing reset current from an external supply to achieve duty cycles of greater than 95%.

In summary, commercially available current sense transformers are characterised under specific conditions. Being aware of the effects of the termination impedance and associated components can enable the designer to optimise his performance for best accuracy over the widest frequency or pulse width range often exceeding the headline specifications of the part.

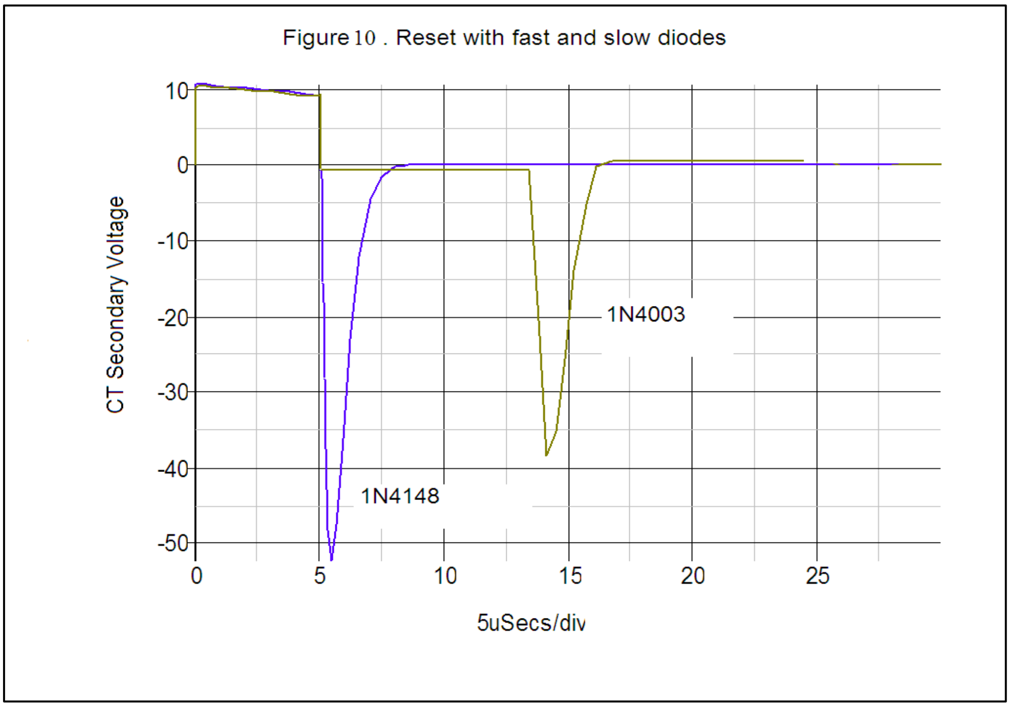


Figure 11. Murata Power Solutions 5300 series current sense transformer

http://www.murata-ps.com/data/magnetics/kmp_5300c.pdf