Deploying Current Transformers In Current And Voltage Monitoring

by Viktor Vogman, Power Conversion Consulting, Olympia, Wash.

Many power subsystems, containing switched-mode power supplies and dc-dc converters, have requirements for galvanic isolation of their loads and control circuits from the primary power source. Such isolation is required to conform to safety regulations and to prevent damage to the load (e.g. IT equipment) in the case of a primary-side active component failure. To meet this requirement high-frequency power transformers are used to electrically separate the primary side of the power conversion stages from the load. At the same time output voltage and input current monitoring used in these systems for control purposes must also include an isolating component in the control signal processing path.

Two of the most efficient methods of providing these features are use of isolated differential amplifiers for voltage sensing[1] and use of specialized current transformers (CTs) for current sensing. Such current sensors offer low power loss and high immunity to common-mode noise. However, besides accurate current monitoring, CTs can also be used as analog signal/voltage sensors, enabling the transfer of an analog voltage signal among two circuits that are not electrically coupled.[2]

Besides providing enhanced isolation through magnetic coupling and the other benefits mentioned above, CT-based voltage sensing has one more important advantage: it does not require the usage of separate converters supplying isolated Vcc to power up the isolating sensor input circuitry. This article discusses the implementation of techniques for adopting CTs in both current and voltage monitoring applications.

The article begins by discussing the principles of operation of a CT in its primary use in current monitoring. It then discusses CT design requirements for ac current monitoring in the case of symmetrical waveforms without a dc component and the limits imposed by core saturation. The discussion then addresses unidirectional current monitoring and how the sensor circuit can be modified for reconstruction of the dc component.

The second part of the article explores the use for CTs in voltage monitoring, presenting a circuit for this purpose and discussing its design requirements. This includes the option of implementing it with either a sample-and-hold circuit or a low-pass filter to produce the output voltage signal. Use of the circuit to sample constant- and time-varying waveforms is also discussed. Finally, guidance is provided on component selection for the CT-based voltage monitoring circuit and the choice of its operating parameters.

CT-Based Current Monitoring

The current transformer is considered a type of "instrument/signal" transformer that produces a varying current in its secondary winding which is proportional to the current flowing in its primary. There are many specialized types of current transformers operating at various frequencies (from tens of hertz to several megahertz) and monitoring currents up to thousands of amperes.

Similar to a voltage transformer, CT operation is based on the principle of Faraday's law of electromagnetic induction. In each of the transformers, a winding voltage ratio is directly proportional to the winding turns ratio and the current in the windings is inversely proportional to the voltage in the windings.

The major difference between current and voltage transformer types is in their utilization (application). While the primary winding of a voltage transformer is believed to be attached to a voltage source, the primary winding of a current transformer, like an ammeter, is connected in series with an external network that essentially functions as a current source, producing current through the CT primary winding (Fig. 1).

This current is determined by the external circuit operation/parameters (V_{in} and R_{LOAD} in Fig. 1) and practically does not depend on the CT characteristics. That's because, due to a commonly used large turns ratio, CT equivalent resistance referred to the primary side is negligible as compared to the external circuit resistance (R_{LOAD} in Fig. 1). To meet this requirement, the secondary CT winding must never be left open to prevent CT saturation and generation of excessive EMF voltage V_T across its secondary terminals. That is why in real applications, the CT is always loaded with a terminating resistor (R_T in Fig. 1) galvanically isolated from the main circuit.

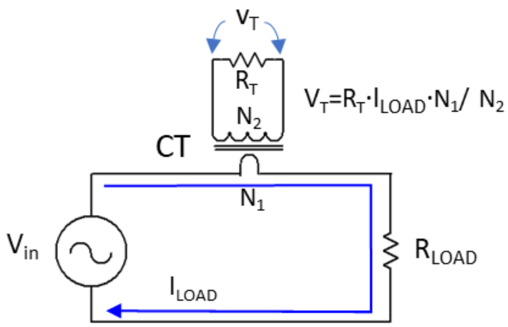


Fig. 1. Basic CT application for current monitoring. The CT primary winding current is determined by an external circuit impedance (resistor R_{LOAD}). The CT terminating resistance referred to the primary side $R_T N_1^2/N_2^2$ is negligible as compared to the external circuit resistance R_{LOAD}.

Using a large turns ratio in the CT provides high voltage gain that enables sensor operation with just a few millivolts of the voltage drop across the primary winding and a voltage output that does not need further amplification.[3] Thus, the major advantages of CT-based sensors are galvanic isolation and lower power dissipation as compared to shunt resistor sensors that require higher voltage drops to generate feasible signal levels.[4, 5]

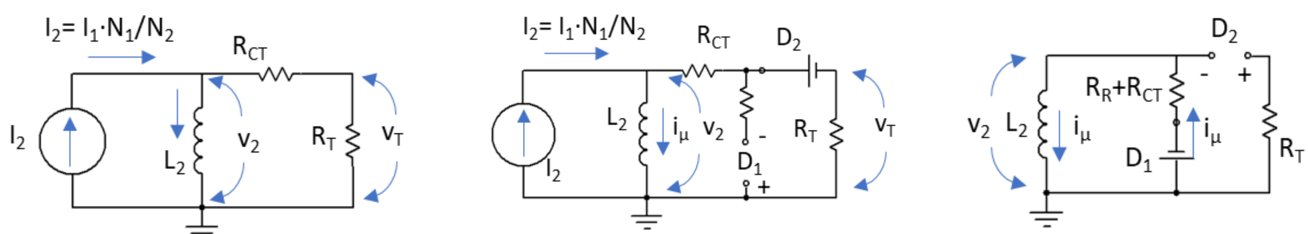
Next, let's examine a CT's basic operation and look at design considerations for its deployment in current sensing and voltage monitoring applications.

Current Sensing Applications

AC Current Monitoring

CTs are most commonly used in various ac current monitoring applications in which the current is driven in two directions and the current waveform is divided by a zero-line into two symmetrical halves. Ac waveforms provide automatic CT reset over each ac cycle. In such applications, CT-based current sensors can be placed, for example, on ac line current-carrying conductors, in series with a motor coil, or H-bridge power transformer primary winding.[4]

The Fig. 1 CT-based sensor equivalent circuit referred to the secondary is shown in Fig. 2a. This equivalent circuit is very basic and neglects stray inductances and core losses. However, it provides sufficient insight to understand the CT operation.



(a) *(b)* *(c)*

Fig. 2. Equivalent circuits of CT-based current sensors. These include the ac current monitoring case (a) and the unidirectional pulsing current monitoring case with a dc reconstruction network (b) and (c). In (b) and (c) the forward-biased diodes are represented by their potential barrier dc voltages.

As stated above, in this section we will be examining the case in which the monitored current $i_1(t)$ does not have a dc component and its waveform is symmetrical. When the current flows through the primary CT winding an equivalent voltage $v_T(t)$ generated at the burden terminals can be determined as follows:

$$v_T(t) = \frac{N_1}{N_2} \cdot i_1(t) R_T \qquad (1)$$

where R_T is the resistance of the terminating resistor attached to the secondary winding, and N_1 and N_2 are numbers of turns in the primary and the secondary windings, respectively.

Because the resistance of the secondary winding R_{CT} together with a burden resistance R_T represents a voltage divider, voltage $v_2(t)$ across equivalent magnetizing inductance L_2 can be determined as follows:

$$v_2(t) = v_1(t)\frac{N_2}{N_1} = \frac{v_T(t)(R_T + R_{CT})}{R_T} = \frac{N_1}{N_2} \cdot i_1(t)(R_T + R_{CT})$$

Defining the effective impedance of the equivalent shunt resistance $R_{1.eqv}$ acting at the primary side as a ratio of a primary winding voltage to the winding current, we find that

$$R_{1.eqv} = \frac{v_1(t)}{i_1(t)} = \left(\frac{N_1}{N_2}\right)^2 (R_T + R_{CT})$$

which is proportional to the CT turns ratio squared and that at $N_2 >> N_1$ it can be made very small which stipulates the high efficiency of a CT-based sensor.

The above equations are valid under the condition that the magnetizing current i_μ flowing through L_2 (Fig. 2a) is negligible as compared to the monitored current i_2. Defining magnetizing current magnitude $I_\mu$as an integral of the voltage applied to the inductance over one half of ac cycle T, we find:

$$I_\mu = \frac{1}{L_2}\int_0^{T/2} v_2(t)dt = \frac{1}{L_2}\int_0^{T/2} \frac{N_1}{N_2} i_1(t)(R_T + R_{CT})dt \qquad (2)$$

Because L_2 sinks a portion of the monitored current, it would be logical to define the current monitoring error as a ratio of the magnetizing current magnitude to the monitored current magnitude referred to secondary $\Delta = I_\mu / I_{m2}$. The smaller such error, the higher the monitoring accuracy.

This error needs to be set at the minimum current magnitude at which the required accuracy needs to be maintained. After expressing inductance L_2 from equation 2, we can determine the CT time constant that provides the required error value:

$$\tau_{CT} = \frac{L_2}{(R_T + R_{CT})} = \frac{1}{\Delta \cdot I_{m2}} \int_0^{T/2} \frac{N_1}{N_2} i_1(t)dt \qquad (3)$$

For example, for a sinusoidal waveform with magnitude I_{m2} $[i_2(t) = I_{m2} sin\omega t]$ we can find:

$$\tau_{CT.sin} = \frac{N_1}{\Delta N_2}\frac{T}{\pi}$$

For a square wave with the same magnitude:

$$\tau_{CT.sq} = \frac{N_1}{\Delta \cdot N_2}\frac{T}{2}$$

These time constants provide the required current monitoring accuracy under the condition that the CT core operates in the linear region of its magnetizing curve and inductance L_2 remains unchanged. To accommodate this requirement in a real application, some limitations on the current magnitude or/and terminating resistor value need to be considered.

CT Magnetizing Curve And Saturation

A current transformer, like any other magnetic component with given core dimensions, can properly operate only with a certain level of maximum flux density $B_{CT.max}$ in its core.[6] This flux density can be determined by the following equation:

$$B_{CT.max} = \frac{\int_0^{T/2} v_2(t)dt}{N_2 A_C} = \frac{\int_0^{T/2} \frac{N_1}{N_2} i_1(t)(R_T + R_{CT})dt}{N_2 A_C} \qquad (4)$$

where A_C is the core cross-sectional area.

This equation accounts for the startup condition, in which the core operating point is positioned at the origin ($B_{CT} = 0$). As long as the flux density $B_{CT.max}$ remains within a so-called unsaturated zone of the magnetizing curve, a change in flux will generate a signal on the secondary side, that with a given accuracy represents the primary current as described by equation 1.

When the primary current magnitude becomes so high that the core cannot handle any more flux, the CT core transitions into a saturation zone. In this zone, there is no flux change, which results in no secondary current flow. In this case, magnetizing inductance L_2 in the equivalent circuit in Fig. 2a essentially drops to zero and shorts the CT load. As a result, all of the monitored current is used as a magnetizing current and no current flows into the terminating resistor which makes the secondary signal equal zero. To prevent this from happening the following condition must be met:

$$B_{CT.max} < 0.5B_{sat}$$

where B_{sat} is the saturation flux density of the core material.

Meeting this condition practically guarantees that with dc offsets or load current surges not exceeding 30% to 50% of nominal, the CT operating point remains in the magnetizing curve linear region which automatically satisfies the requirement of providing the required current monitoring accuracy in normal operation.

This requirement is especially important for CT overcurrent protection applications. If the monitored current can exceed the projected level by greater than 50% there is a risk that the overcurrent protection will not sense this as an abnormal condition, so the protection trip point needs to be positioned as close to the nominal level as possible.

As can be seen from equation 4, another way of limiting the flux density is to reduce the terminating resistor value; however, in this case the provided margin will be associated with a reduction in signal magnitude.

Unidirectional Current Monitoring Specifics

Solid-state switches used in power conversion applications typically pass current flowing only in one direction. By definition switch currents are discontinuous and include a dc component creating a dc bias in the current transformer core (Fig. 3a). Because transformers operate only with ac voltages across any of their windings, the signal produced at the CT secondary winding terminals needs to be processed to accurately reconstruct the monitored current's dc content.

This is accomplished by adding extra components as shown in Fig. 3b and reflected in equivalent schematics in Fig. 2b and c. Circuits and timing diagrams in Fig. 3 further illustrate the dc bias impact and the details of such reconstruction.

The circuit shown in Fig. 3a represents the simplest CT sensor implementation and illustrates the usage of the basic current transformer sensor (Fig. 1) in the unidirectional application without reconstruction of the dc content on the secondary side. The corresponding output voltage waveform for a square-wave case is shown with a yellow trace in Fig. 3c.

In this configuration, the CT core needs to support the flux density $B_{CT.max}$ consisting of two components—the flux density B_{DC} corresponding to the magnetizing force, created by the primary average current (dc bias), and the ac flux density component $B_{\sim}$:

$$B_{CTmax} = B_{DC} + B_{\sim} = \frac{\mu\mu_0 N_1 I_{1m} D}{l_C} + \frac{V_{2.pk-pk}(1-D)DT}{N_2 A_C}$$

where μ is the magnetic permeability of the core material, $\mu_0 = 4\pi \times 10^{-7}$ H/m is the magnetic permeability of free space, l_C is the core magnetic path length, T is the switching cycle, D is the duty ratio, and $V_{2.pk-pk}$ is the peak-to-peak voltage generated at the burden terminals, which can be determined by equation 1:

$$V_{2.pk-pk} = \frac{N_1}{N_2} \cdot I_{1m} R_T$$

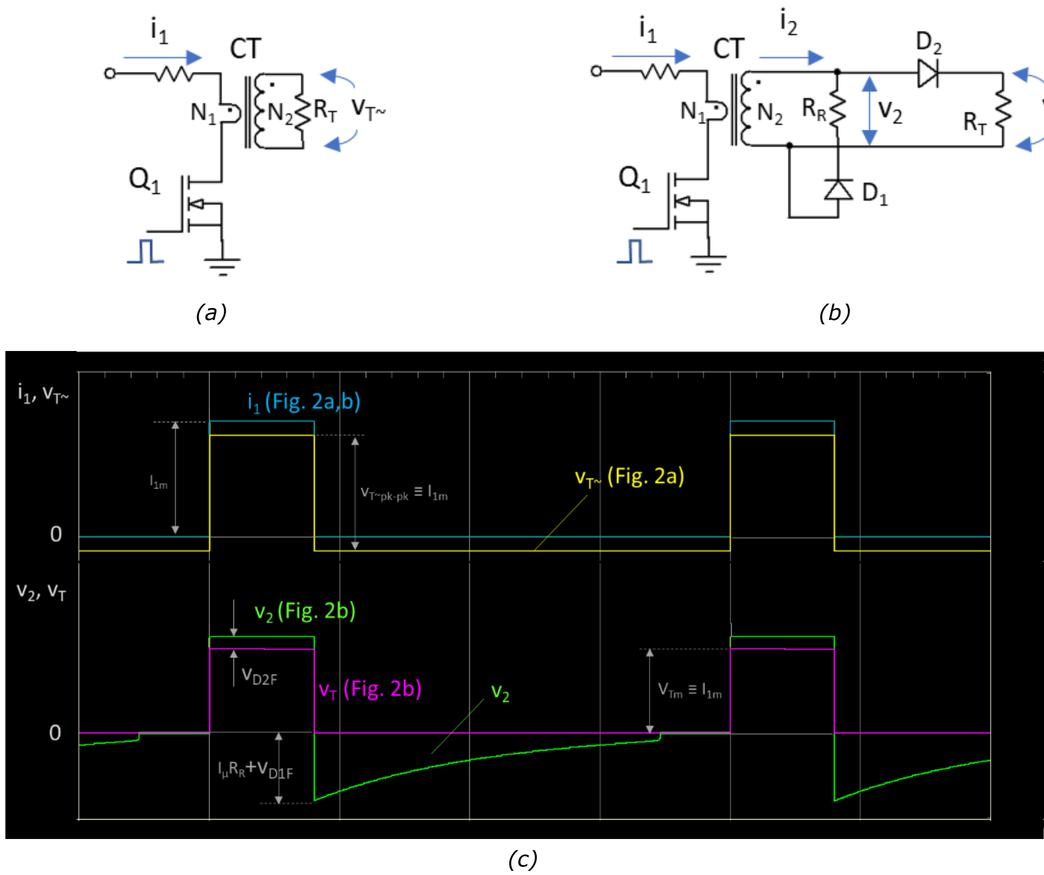


Fig. 3. Monitoring a unidirectional pulsing current with a basic (Fig. 1) CT sensor (a). Adding a few low-power components on the secondary side provides the transformer reset and reconstructs the dc component of the primary current (b). Voltage and current timing diagrams (c).

When due to a transient event the current magnitude or duty ratio instantly changes, the circuit shown in Fig. 3a will exhibit some latency in response to such change. This latency is associated with dc bias current change and is determined by L_2/R_T time constant, which is why such a CT application can be considered feasible only for qualitative current monitoring in which a steady-state cyclic current waveform shape is the subject of interest.

Reconstruction Of The DC Component

The circuit shown in Fig. 3b, repeated below in Fig 4a, can be used for accurate monitoring of unidirectional discontinuous (pulsing) current waveforms with zero minimum level. It senses current variations with no latency and contains three additional components (D_1, R_R and D_2) on the secondary side that are used for CT reset and reconstruction of the dc content of the original waveform.

Its equivalent circuits for conducting diodes D_2 and D_1 states are given in Fig. 2b and Fig. 2c, respectively (repeated below in Fig. 4c and d). The bottom diagrams in Fig. 3c (repeated below in Fig. 4b) illustrate the circuit operation. Rectifying diode D_2 detects the current magnitude at the burden terminals. Similar to the above case, the corresponding voltage magnitude can be determined using equation 1:

$$V_{Tm} = \frac{N_1}{N_2} \cdot I_{1m} R_T$$

If the ratio N_2/N_1 numerically equals R_T expressed in ohms, then the burden signal has a scale of 1 V per amp, which can be very convenient for lab measurement applications. The positive secondary voltage magnitude exceeds V_{Tm} by the rectifying diode forward voltage drop V_{D2F} (Fig. 4b) but since during the D_2 conduction time interval CT operates as a current transducer, equation 1 for voltage V_T across the terminating resistor remains valid.

Network D_1-R_R is used for the CT reset (de-magnetizing) during the dead time, i.e. when the monitored current is at the zero level. The R_R value needs to be selected so that the CT magnetizing current reaches zero over the dead time ($i_1(t) = 0$) interval $(1\text{-}D)T$:

$$R_R \geq \frac{3L_2}{(1-D)T}$$

In such an operating mode the CT is totally reset before the next current pulse (green trace waveform in Fig. 4b), the core magnetic flux is not impacted by the monitored current dc component, so the CT core type, the turns ratio, and the terminating resistance value can be selected based on equations 3 and 4 applicable to ac current monitoring CTs. The reset voltage magnitude $V_{m-} = I_\mu R_R + V_{D1F}$ (Fig. 4b) will be applied to the rectifying diode D_2 in the reverse direction (Fig. 4d) and represents the key parameter for its selection.

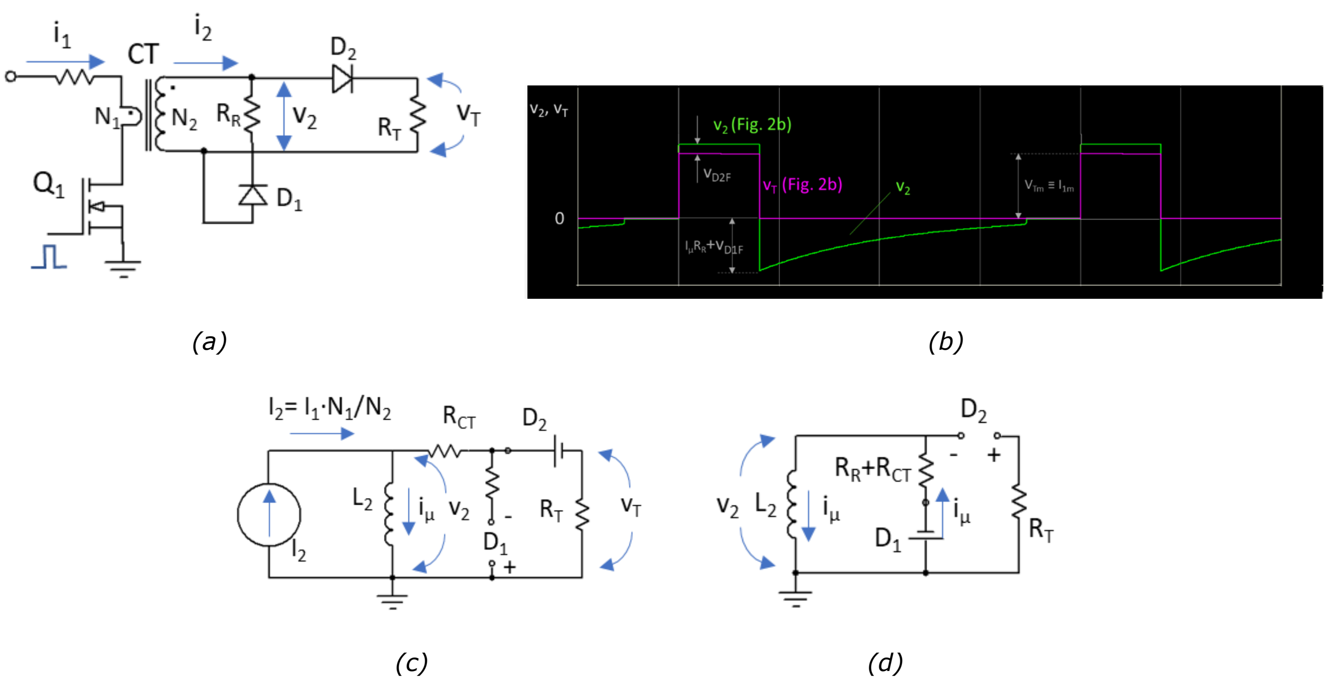


Fig. 4. CT sensor for monitoring a unidirectional pulsing current with reconstruction of the dc component (a). Voltage and current timing diagrams (b) for this sensor. Equivalent CT circuits during the pulse top (c) and dead time (d) intervals.

Waveforms With Non-Zero Minimum Level

It is important to reiterate, that because any transformer needs an alternating current for creating a changing magnetic flux in its core, only a varying component of a pulsing (peak-to-peak) current waveform can be captured by a CT-type sensor. For ac current waveforms or pulsing waveforms with a minimum level of zero such sensors provide current readings with a given accuracy.

Ac current waveforms with dc offsets or pulsing currents with minimum levels other than zero, such as a converter inductor current in CC mode, would be reflected on the CT secondary without their dc components. [4] That is why CT-based current sensors are not intended for use in such applications. It is also important to note, that using a more-sophisticated technique for dc component restoration and utilizing CT saturation mode provides an opportunity for dc current monitoring. This technique was discussed in detail in a previous article (see reference 7).

Using A CT For Voltage Monitoring

Besides providing accurate current monitoring, current transformers can also be used as analog signal/voltage sensors with galvanic isolation, enabling the transfer of such signals among two circuits that are not electrically coupled. As mentioned above, CT-based voltage sensors do not require the usage of converters supplying isolated Vcc to power up the input circuitry of isolated differential amplifiers.[1] Also similar to CT-based current sensors, they have low power loss and are not affected by common-mode noise which makes them an attractive option for isolated voltage sensing.

However, unlike the current transformer sensors examined above, CTs in most low-voltage monitoring applications do not need a large turns ratio because currents flowing through their primary and secondary windings are used just for "probing" the signal under test and are very small and comparable in magnitude. Now, let's take a closer look at this new application for CTs.

Basic Operation

A basic schematic diagram of the circuit utilizing a CT for voltage monitoring is shown in Fig. 5.

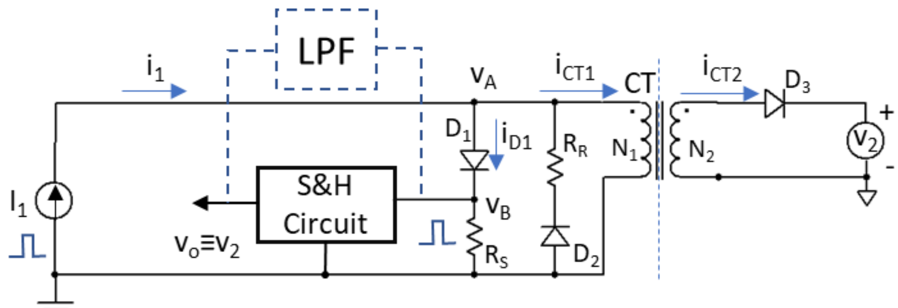


Fig. 5. Basic circuit for monitoring galvanically isolated voltage V_2 with a CT. Injecting "probing" pulses of current i_1 and sampling corresponding voltage samples (v_B) with a sample-and-hold circuit or a low-pass filter produces an analog signal V_o proportional to the sampled voltage.

In this network, a pulsing current created by the primary source I_1 gets injected into the CT-based monitoring circuit that samples a galvanically isolated voltage V_2 at the remote sense points. The injected current is split between two paths: D_1-Rs (i_{D1}) and the CT primary winding (i_{CT1}). The CT primary winding current, reflected to the secondary side as $i_{CT2} = i_{CT1}N_1/N_2$ (Fig. 5), flows through network D_3-V_2. An analog signal with a level proportional to the monitored voltage is detected by the sample-and-hold circuit synced with the injected current pulses.

The parameters for the CT-based voltage monitoring network can be determined by examining its step-by-step operation. Because D_3 is forward biased during the pulse top, the monitored voltage V_2 during this time interval is applied to the secondary CT winding via the conducting diode. Since injected currents are small, CT winding resistance can be neglected and the voltage across the D_1-Rs network during this time interval equals

$$v_A = V_{Am} = \frac{N_1}{N_2}(V_2 + V_{D3F})$$

Voltage magnitude V_{Bm} at the sample-and-hold circuit input will be less than the peak voltage V_{Am} in point A by the voltage drop V_{D1F} across diode D_1:

$$V_{Bm} = \frac{N_1}{N_2}(V_2 + V_{D3F}) - V_{D1F}$$

The sample-and-hold circuit output represents a dc voltage with a level proportional to V_2 and equal to V_{Bm}. If for noise immunity purposes the sample-and-hold circuit is replaced with a low-pass filter (dashed block in Fig. 5), then for square wave pulses having a flat top and produced at regular sampling time intervals T we get the following filter output:

$$V_o = \frac{1}{T}\int_0^T v_B(t)\,dt = V_{Bm} \cdot D_i = \left[\frac{N_1}{N_2}(V_2 + V_{D3F}) - V_{D1F}\right] \cdot D_i \qquad (5)$$

where D_i is the duty ratio of the injected pulses. Thus, with a flat pulse top, V_o also represents a linear function of V_2.

The implementation of current source I_1 is simplified if the housekeeping voltage available on the control side exceeds the V_2 level referred to the primary. In this case, the I_1 component can be generated with a source follower or a current source IC without employing magnetic components. The optimum transformer turns ratio can be determined as follows:

$$\frac{N_1}{N_2} = \frac{V_{cc} - V_\Delta}{V_2 + V_{D3F}} \tag{6}$$

where V_Δ is the margin required for reliable operation of the current-source stage which in practice can be selected in the range of 3 V to 5 V.

Based on this expression, since V_{cc} used by the control circuitry usually does not exceed 15 V, monitoring of higher voltage levels (V_2 > 12 V) will require a step-up transformer ($N_2 > N_1$). Meanwhile for many lower voltage applications N_1/N_2 can be selected as 1:1 for interchangeability purposes and simplified diode selection. A transformer with unity turns ratio also could be used for higher voltages monitoring when D3 is attached to a voltage divider.

Making currents through D_3 and D_1 equal: $I_{D1m} = I_{D3m} = I_{Dm}$ and high enough to operate above cut-in points of their V-I characteristics yields requirements for the injection current magnitude and resistor R_S value:

$$I_{1m} = I_{Dm}\left(1 + \frac{N_1}{N_2}\right); \quad R_S = \frac{V_{Bm}}{I_{Dm}}$$

As mentioned above, equation 5 is valid under the condition that voltage pulses in point B have flat tops. To meet this requirement the magnetizing current magnitude should always be lower than I_{Dm}, which determines the requirement for the minimum magnetizing inductance value:

$$L_{2.min} = \frac{(V_2 + V_{D3F}) D_i T}{I_{Dm}} \tag{7}$$

As follows from equation 5 describing the relationship between V_o and V_2, primary and secondary diode voltage drops are opposite in sign and under a certain condition ($V_{D3F}\, N_1/N_2 = V_{D1F}$) can cancel each other out. Such cancellation can be achieved by selecting identical diode types on primary and secondary sides.

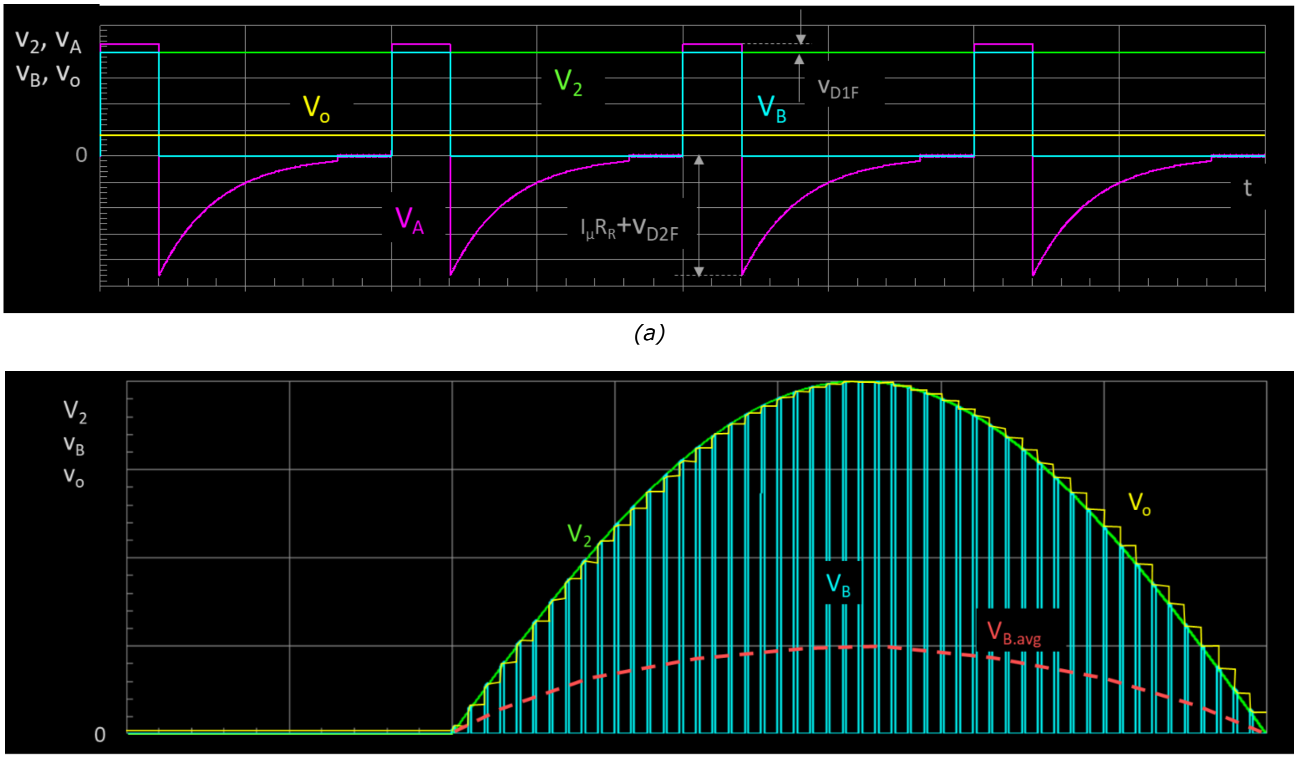
For example, for a 1:1 turns ratio and identical diodes on the primary and secondary sides ($V_{D3F} = V_{D1F}$), we get $V_{Bm} = V_2$. This means that at the low-pass filter output (equation 5) we get: $V_o = V_2 \cdot D_i$ independent of the diode characteristics.

Sampling Of Constant And Time-Varying Voltage Waveforms

Timing diagrams illustrating an isolated voltage sampling process of two different-shaped voltage waveforms are given in Fig. 6 with reference to Fig. 5 circuit designations.

Fig. 6a shows CT primary winding voltage V_A (magenta trace), rectified voltage V_B (blue trace) and output V_o (yellow trace) for a constant voltage V_2 sampling process. An analog signal proportional to the output voltage in this case is extracted with a low-pass filter. Fig. 6b shows secondary voltage V_2, rectified voltage V_B samples (blue trace), and their average level $V_{B.avg}$ for a positive sinusoidal waveform sampling case.

Since with widely varying V_2, the diode current magnitudes will not match most of the time, it would be recommended for a varying voltage sampling case to select diodes with the steepest V-I characteristics providing minimal forward voltage variations, and to have D_1 with a lower cut-in voltage. To provide the same accuracy in monitoring of a voltage varying over a wide range, passive resistor R_s in Fig. 5 needs to be replaced with an active current sink network synced with the I_1 source and capable of supporting the fixed sink current magnitude (I_{Dm}) at high and low V_A levels.



(a)

(b)

Fig. 6. Sampling of the isolated dc voltage (a), and positive portion of a sinusoidal waveform (b). Analog signal proportional to output voltage can be extracted with low pass filter (a) or with a sample- and-hold circuit (b).

To meet this requirement at V_2 levels close to zero the sink current network has to be attached to a negative voltage supply. Using an actual sample-and-hold circuit or LPF for extraction of the dynamic average from a sequence of pulses is associated with some latency (lag) Δt that increases at lower sampling frequencies. This latency can affect the control loop response time. The higher the sampling frequency, the higher the LPF cutoff frequency can be and therefore the lower latency that can be achieved.

Recommendations For Component And Operating Mode Parameter Selection

As shown in the above voltage sensing circuit analysis, the criteria used for selection of voltage monitoring circuit parameters are quite similar to those of the current monitoring CT application. The main difference is that in the voltage-monitoring case a designer has full flexibility in selecting CT operating frequency and current magnitude.

To achieve lower signal latency and reduce CT dimensions, the sampling frequency needs to be as high as possible. Maintaining a small magnetizing current magnitude providing small injected current magnitude variations is not as critical because as long as the diode current during the pulse top remains continuous, current variations do not noticeably impact the accuracy. This means that the magnetizing-to-injected current ratio Δ used in equation 3 could be set, for example, one order of magnitude higher than in the current monitoring case. This enables the designer to use a lower CT winding inductance as defined by equation 7.

The selection of flux density and reset components is very similar to the current monitoring CT application given above. The only difference is that a voltage magnitude across the winding in the voltage monitoring application does not depend on the load resistance but is determined by the V_2 level. For a flat-top pulse from equation 4 we find:

$$B_{CT.max} = \frac{(V_2 + V_{D3F})D_i T}{N_2 A_C}$$

To reduce the power used in the sampling process the injected current needs to be made as small as possible. This would also minimize its interference with the main power stage producing V_2. At the same time, similar to the current sensor case, the injected current magnitude needs to keep the diode operating points above their p-n junction potential barrier or cut-in voltage (about 0.7 V for silicon diodes). Practically, a few milliamps of current is good enough to make the diodes operate in their low dynamic resistance regions and can be considered as a reasonable tradeoff between minimizing the injection current magnitude and diode forward voltage variations.

The voltage monitoring accuracy can be further improved by replacing diodes D1 and D3 with synchronous rectifiers. Since such a replacement is associated with a part count increase and greater complexity it may be reasonable to consider packing the main voltage sensor components into an integrated circuit.

Summary

This article examined CT operation in current monitoring and voltage monitoring modes and determined the equations for computing basic parameters of CT-based current and voltage sensors. Application areas were defined for each of the examined CT-based current sensor topologies. As discussed above, the desired current monitoring accuracy and core saturation flux density must be taken into consideration when selecting parameters of a CT and a terminating resistor value.

Among other points discussed here, ac current waveforms having dc offsets or pulsing currents with minimum levels other than zero, such as a power converter's inductor current in continuous-conduction mode, would be reflected on the CT secondary side without their dc components. That is why CT-based current sensors are not intended for use in such applications.

However, usage of CT-based current sensors for unidirectional pulse-current monitoring with reconstruction of the dc component is capable of providing accurate current readings with no latency. In addition, this approach can be considered the best choice of current sensor for both steady-state and transient operating modes. Usage of a CT-based current sensor without reconstruction of the dc component can be considered feasible only for qualitative current monitoring in which a steady-state cyclic current waveform shape is the subject of interest.

Besides achieving enhanced isolation with magnetic coupling, immunity to common-mode noise, and low power dissipation, CT-based voltage sensing has one more important advantage: it does not require the use of separate converters supplying isolated Vcc to power up the isolating sensor input circuitry.

To provide high accuracy in monitoring of a voltage varying over a wide range, it is recommended to use an active current sink network synced with the I_1 source and capable of supporting the fixed sink current magnitude (I_{Dm}) at high and low V_A levels. In dc and varying unidirectional voltage monitoring applications selecting the highest possible sampling frequency provides the smallest size for the CT and the lowest analog signal latency.

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About The Author

Viktor Vogman currently works at [Power Conversion Consulting] as an analog design engineer, specializing in the design of various power test tools for ac and dc power delivery applications. Prior to this, he spent over 20 years at Intel, focused on hardware engineering and power delivery architectures. Viktor obtained an MS degree in Radio Communication, Television and Multimedia Technology and a PhD in Power Electronics from the Saint Petersburg University of Telecommunications, Russia. Vogman holds over 50 U.S. and foreign patents and has authored over 20 articles on various aspects of power delivery and analog design.

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