CURRENT TRANSFORMERS

Current transformers (CT) are magnetic devices, which assist in the measurement of current. Different applications necessitate a wide variety of designs. Current transformers used in the ac power field, along with voltage or "potential" transformers, are generally known as instrument transformers, and are discussed only briefly here. Other types of CT discussed in this article are:

- high-frequency CTs;
- pulse current transformers;
- ac + dc current-sensing transformers;
- precision dc current-sensing transformers;
- Rogowski coils.

Classical ac CTs are used to change an ac current to a more convenient level for measurement, while isolating the ammeter from the current carrying conductor. Current transformers used in electrical power applications typically operate at frequencies of 50 or 60 Hz, but may range from 25 Hz to 400 Hz or so. Measured current levels may be less than an amp to thousands of amperes or more, and transformation accuracies may range from 0.1% or better for critical measurements, to several percent or more for general indication or overload detection.

Higher frequency ac current transformers are used in a number of other applications, including RF transmitters, induction heaters and electronic power converters to control output, monitor operation, or diagnose problems. Frequencies may range from hundreds of hertz to hundreds of megahertz, while current levels are usually lower at milliamperes to hundreds of amperes. Accuracy requirements are often not as stringent, often in the 1 to 10% range, although errors down to 0.1% are obtainable at frequencies up to the low megahertz band. When very wide bandwidths are required, active electronic circuits can be used to extend the low-frequency limit by several orders of magnitude, while improving accuracy at midband frequencies.

Ac transformers can be adapted to measuring pulse currents, finding use in electronic power circuits and pulse energy systems. They can operate with a small net dc current, as long as the current is zero for a sufficient interval between pulses.

Conventional ac current transformers cannot measure dc current, nor can they usually tolerate significant dc current without magnetic saturation of the core. The CT can be adapted to measure dc as well as ac current, with modifications and the addition of electronic circuitry, albeit at some potential sacrifice in available accuracy. These devices are sometimes termed current transducers, to distinguish them from "ac only" CTs. Some oscilloscope current probes also use this approach to provide broadband sensing from tens of megahertz down to dc.

A close relative of the current transformer is the Rogowski coil, which physically resembles a current transformer but (usually) without a magnetic core. The open circuit output voltage is proportional to the time derivative of the measured current, rather than the current itself. Lacking a core, the device cannot be saturated by large ac or dc currents. As such, it may find use where high currents are involved, or at very high frequencies, where magnetic cores become a hindrance to accurate measurement.

HIGH-FREQUENCY CURRENT TRANSFORMERS

High-frequency current transformers (HFCT) are similar in design and construction to an instrument CT, principally differing in the common use of ferrite cores, which perform as well or better than nickel–iron cores above about 100 kHz. Minimization of capacitive loading and parasitic coupling also become much more important at high frequencies.

The maximum initial permeability of ferrite is about 10 k, but this value holds to low flux densities and up to at least 100 kHz. This permeability is superior to that of conventional silicon steel at very low flux densities for any frequency. The permeability of nickel–iron can reach values of 100 k, but begins to fall at a few kilohertz to similar values for ferrite above 100 kHz. However, ferrite often finds use at frequencies well below 100 kHz, due to a significantly lower cost than nickel–iron.

Unlike ac line frequency CTs, HFCTs are often used over a very broad band of frequencies, typically three to six decades of frequency. HFCTs are often used to reduce the current to be measured, as are ac line frequency CTs, but not necessarily for identical reasons. The load or burden on a HFCT is usually a resistance (instead of an ammeter), to convert the current to a corresponding voltage, which is easier to measure at high frequencies. The shunt resistance required to measure currents above a few amperes is typically much less than 1 Ω, in order to limit power dissipation. The parasitic inductance of current shunts becomes problematic at high frequencies, as described in the article titled CURRENT SHUNTS.

The shunt inductance problem can be greatly alleviated with current transformers, even with a moderate turns ratio. A CT with a ten-turn secondary and a one-turn primary decreases the primary current and increases the voltage by ten times on the secondary. The secondary side current shunt now has 100 times the resistance required on the primary for the same power dissipation. The useable frequency increases by the same factor for a given shunt inductance, and the higher voltage is usually easier to measure accurately. For HFCTs, the secondary side current shunt is usually known as the CT *load resistance*, and this term will be used hereinafter.

Many high-frequency CTs use a single-layer secondary winding on a toroidal core with a single-turn primary, as shown in Fig. 1. Often the primary is composed of the conductor carrying the current to be measured.

In some instances the temporary installation of a current transformer is desired when it is not feasible to break the circuit carrying the current to install the transformer. Clamp-on current transformers are made for these applications, with a split core for placing around conductors without having to open the primary circuit. A hinged version of a clamp-on transformer is shown diagrammatically in Fig. 2. Other configurations may use a split toroidal core with a hinge, or a portion of the core may be made to slide into place to complete the magnetic circuit. The inevitable air gap between core

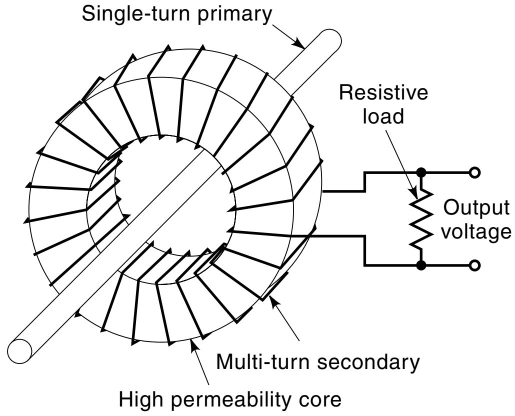

Figure 1. A high-frequency ac current transformer is typically used with a resistive load or "shunt" to measure current.

pieces tends to reduce the accuracy and raise the minimum useable frequency of clamp-on current transformers.

Current transformers will algebraically sum the currents in two or more primary windings. This ability can be quite useful and sensitive in comparing two nearly identical currents flowing in opposite directions, as the CT responds only to the current difference. This allows a comparison of two currents to within parts per million in precision laboratory test equipment, or the detection of ground fault currents of a few milliamperes out of 10 or more amperes in the humble household ground fault interrupter (GFI) breaker.

HFCT Error Sources

An equivalent circuit for a typical transformer is shown in Fig. 3, including the usual parasitic "components," in addition to the ideal transformer. The relative importance of the various parasitic effects are quite different for current transformers, when compared with other types of transformers.

Primary Resistance and Leakage Inductance. The primary winding resistance R_p and transformer leakage inductance

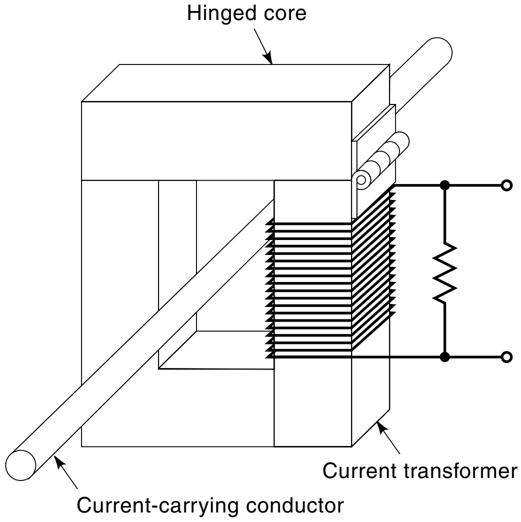

Figure 2. A clamp-on current transformer allows current measurement without breaking the circuit.

L' have no direct impact on current transformation accuracy, as they are in series with the primary input current I_p. Their only effect is to increase the insertion impedance of the transformer, which may have an influence on the primary current if the insertion impedance is not insignificant, compared with the impedance of the rest of the circuit.

Core Magnetization Currents. The core magnetizing inductance L_m and parallel loss resistance R_m have been shown on the secondary side of the ideal transformer, instead of the more conventional primary side placement. This helps clarify the sources of error, as the secondary current Is is still an "ideal" transformation of the primary current. The ideal secondary current now divides among four parallel paths:

1. the core magnetizing inductance L_m;
2. the core's parallel loss resistance R_m;
3. the parasitic secondary loading capacitance C_s; and
4. the output load resistance R_l.

The currents in L_m, R_m, and C_s bypass current from R_l, and are direct sources of error in the CT. The magnetizing inductance is principally of concern at lower frequencies, below about 100 kHz for a ferrite core permeability of 10 k, and at proportionally higher frequencies if lower permeabilities are used. As long as core flux densities remain below saturation, the minimum operating frequency f_1 is defined by:

$$f_1 = \frac{1}{[2\pi L_m/(R_l + R_s)]} \tag{1}$$

which is the corner frequency below which the current in L_m exceeds that in the load resistance R_l. The transformation error at this frequency is large, about 30%, and the phase displacement is 45°. If an accurate transformation is desired, the transformer has to be designed for a much lower corner frequency. Lowering the corner frequency requires increasing the core area/magnetic path length ratio, increasing the number of primary and secondary turns, and/or using a higher permeability core if one is available.

A second limitation to the minimum CT operating frequency may be core saturation, which occurs when the maximum secondary voltage experienced is too high for the number of turns used. Full saturation occurs at about 0.3 to 0.4 T in ferrites, but peak flux must be kept below 0.1 to 0.2 T (depending on material and operating temperature) for operation in the high permeability "linear" region of the core B–H loop. The number of turns required with a sinuosidal voltage is given by:

$$N = \frac{U}{4.44BAf} \tag{2}$$

where N is the number of turns; U is the rms voltage across L_m, in V; B is the allowable peak flux density, in T; A is the cross section of the core, on m^2; and f is the operating frequency, in Hz. If core flux is too high for the required secondary voltage and operating frequency, the flux can be reduced by increasing the core area and/or the number of turns, while keeping the desired primary/secondary turns ratio. A higher core permeability will not lower the frequency of core saturation for the same saturation flux density.

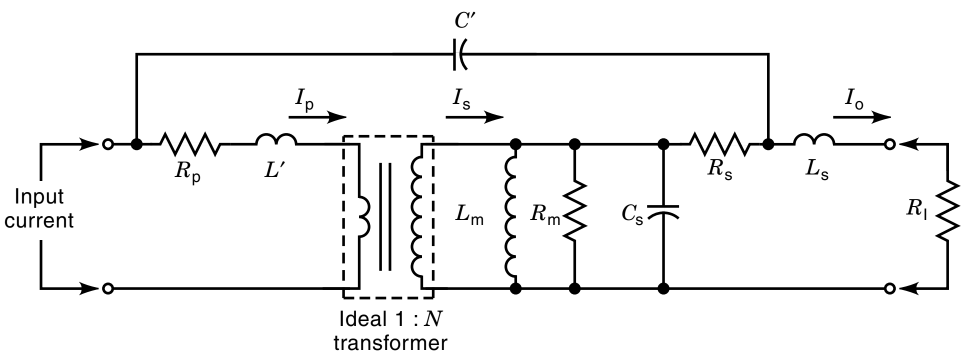

Figure 3. An equivalent circuit of a high-frequency current transformer is useful for identifying sources of error.

The parallel loss resistance of ferrites is roughly constant at high frequencies, so R_m represents a fairly consistent "broadband" error current. This error can also be reduced by increasing the core area/magnetic path length ratio or increasing the number of turns, but a higher or lower permeability core will not significantly affect R_m.

Secondary Capacitance Loading. The upper frequency limit of a CT usually occurs when the impedance of the secondary parasitic capacitance loading C_s becomes less than that of the load R_l. Secondary capacitance is largely composed of winding layer-to-layer capacitance, seconary winding to core capacitance, and turn-to-turn capacitance, in that approximate order. Single-layer windings are often used to eliminate large layer-to-layer capacitances, although some space between the winding ends on a toroidal core must also be provided to reduce end-to-end capacitance. Extra insulation between the winding and a toroidal core is also often useful in reducing secondary capacitance. Turn-to-turn capacitance is usually insignificant in all but the highest frequency CTs with a few secondary turns, when spaced turns or wire with a thicker insulation may be beneficial.

The number of secondary turns tends to have only a moderate effect on secondary capacitance, which may increase or decrease, depending on other winding geometric factors.

Secondary Resistance and Inductance. As long as the secondary winding resistance R_s and the impedance of the secondary output inductance L_s are less than the load resistance R_l, the voltage drop on the output shunt largely determines the voltage on L_m, R_m, C_s, and their respective error currents. Only when the impedance of R_s and L_s become comparable to or larger than R_l do they have a significant or major effect on the currents in L_m and R_m and, hence, on the current transformation error. Increasing the number of turns decreases the magnetizing current errors, but the secondary resistance will increase at least as the square of the number of turns, and as the cube of the turns if the secondary winding is kept to a single layer. A point of diminishing returns is eventually reached when $R_s > R_l$, where more turns begin to degrade accuracy. If higher accuracy is still required, a larger core may be required.

Output inductance is produced by any secondary current flux, which is not coupled to the primary. It is due partly to secondary lead inductance, and partly to "air core" inductance of the secondary winding. The impedance of L_s is usually much less than R_s, and is usually not a significant problem, except at the highest frequencies with a large number of secondary turns.

Primary to Secondary Capacitance. The primary to secondary capacitance coupling C' is usually minimized by having a single-turn primary centrally located in a toroidal CT, with significant insulation thickness on the primary. As such, C' usually has an insignificant contribution to secondary loading capacitance C_s. However, significant error currents can flow through C', if the CT primary and secondary windings are at a high ac potential with respect to each other, which usually occurs when the CT is placed on a primary conductor at a large high-frequency voltage. The worst case occurs with square or pulse voltage waveforms, as the displacement current is

$$I_c = C'(dV/dt) \tag{3}$$

where I_c is the instantaneous capacitance displacement current, in A; C' is the primary to secondary capacitance, in F; and dV/dt is the time derivative of the voltage, in V/s.

These capacitance currents can be surprisingly high with "square-wave" voltages. For example, a 600 V p-p pulse or square wave with a 30 ns rise or fall time will produce a 200 mA pulse of current through only 10 pF of capacitance.

If large primary ac voltages exist, the best solution may be to try to move the CT to a conductor carrying the same current, but at a lower ac potential. If this is not practical and the problem is not too severe, increasing the primary–secondary spacing and going to an insulation with a lower dielectric constant (ideally "air") may suffice.

Otherwise, an electrostatic shield is required around the secondary winding, or at least between the primary and secondary winding. This shield is preferably tied to a "ground" point in the primary circuitry, although a connection to the ground side of the secondary may also serve. Superior isolation is achieved when two shields are used, with the one closer to the primary "grounded" to the primary circuit and the secondary shield tied to the "ground" side of the secondary.

It must be noted that the term "ground," as used here, does not refer to an "earth" connection external to the circuit, but to a low ac potential conductor of the respective circuit. This is often a dc bus for the circuit, and may be known as the *ac quiet* or *RF low* conductor.

Extending HFCT Bandwidth Accuracy with Active Circuits

It has been shown that core magnetizing currents represent the fundamental limit to CT low- and medium-frequency

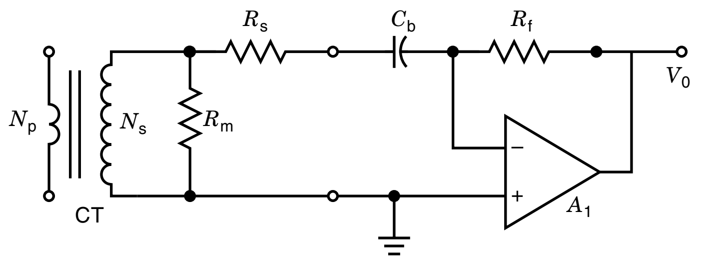

Figure 4. An operational amplifier (op amp) can be used to create a "virtual ground" on the secondary of a CT, improving accuracy and lowering the minimum usable frequency.

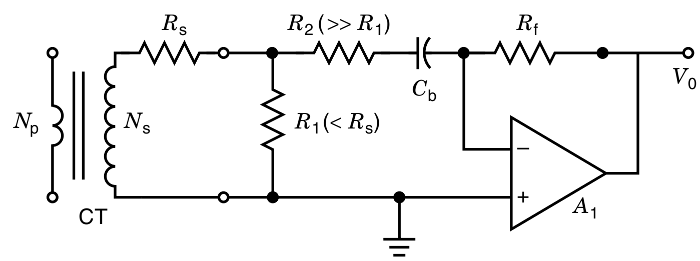

Figure 5. The operational amplifier current required to create a "virtual ground" on the secondary of a CT is reduced by the ratio of $R_1/(R_1 + R_2)$.

bandwidth and accuracy. There are two ways that active electronic circuitry can be used to reduce these errors: the "virtual ground" load, and flux cancellation through feedback.

CT Virtual Ground. The CT load resistance is usually much larger than the secondary winding resistance, so the secondary current accuracy will improve as the load resistance is decreased to zero, dropping the voltage on the core's L_m and R_m to a minimum limited by R_s (and L_s, if significant). Unfortunately, the error in reading the shunt voltage tends to increase with lower voltages, due to system noise and other effects. One way to overcome this tradeoff is to use feedback to force the secondary output voltage to zero (creating a virtual ground on the secondary winding), while measuring the current required to achieve the virtual ground.

The simplest virtual ground load circuit uses an operational amplifier (op amp) A_1, as shown in Fig. 4. The "ground" side of the CT secondary is connected to the positive (noninverting) input of A_1, while the other side is connected to the negative (inverting) input. A feedback resistor R_f is connected between the A_1 output and the inverting input. The high gain of A_1 forces the voltage on the negative input to be essentially equal to the "zero voltage" on the positive input. The current required to do this flows through R_f, so the output voltage is, essentially,

$$V_o = I_s \times R_f \tag{4}$$

where V_o is the output voltage, in V; I_s is the secondary current, in A; and R_f represents feedback resistance, in Ω. In effect, the feedback resistor R_f becomes the CT load resistor, but the CT does not "see" this voltage on the secondary.

A blocking capacitor C_b may be required in series with the CT secondary, to avoid gain multiplication of the A_1 input offset voltage V_{os}. Without C_b, the dc output of A_1 would be V_{os} times R_f/R_s, while, with C_b, the dc output is simply equal to V_{os}. C_b must be a relatively large capacitor, to avoid limiting the low-frequency bandwidth; specifically,

$$(C_b R_s) > L_m/R_s \tag{5}$$

where C_b is the dc blocking capacity, in F; R_s signifies secondary winding resistance, in Ω; and L_m is the secondary winding magnetizing inductance, in H.

The equivalent series resistance (ESR) of C_b should be less than R_s, and C_b should have an rms current rating equal to the full secondary current. A low-voltage electrolytic capacitor will usually suffice, which need not be a nonpolar type as the dc voltage will only be the input offset voltage of A_1, which is usually less than 5 to 20 mV.

A drawback of this simple but effective circuit is that A_1 must be able to source the full CT secondary current through R_f, while the current current capability of most op amps is limited to 10 to 100 mA. The feedback resistance R_f will usually be much less than the CT magnetizing resistance R_m, so a unity gain stable op amp is required, and the output HF bandwidth is limited by the gain–bandwidth of A_1.

This limitation can be partially circumvented by the circuit of Fig. 5, where the op amp current is reduced by the current division ratio of $R_1/(R_1 + R_2)$. R_1 must be less than R_s, to avoid raising the L_F end of the bandwidth. The op amp A_1 need now only be stable at a gain of $R_f/(R_1 + R_2)$, which can increase the HF bandwidth (the actual voltage gain is R_f/R_2). An alternative circuit, with a noninverting gain of $(R_f + R_2)/R_2$, is shown in Fig. 6.

Flux Cancellation with Feedback. The low-frequency bandwidth of a HFCT can be extended even further, with circuits that sense and cancel the CT core flux through feedback techniques (1). A basic circuit that illustrates this principle is shown conceptually in Fig. 7. The HFCT now has a tertiary winding with turns N_t, which drive the inputs of high-gain amplifier A_1, while the output of A_1 drives a current through the secondary winding and load resistor R_l. The high gain of A_1 forces a secondary current to flow, which almost exactly cancels the primary current, in order to hold the voltage on the tertiary winding to nearly zero. With essentially no cur-

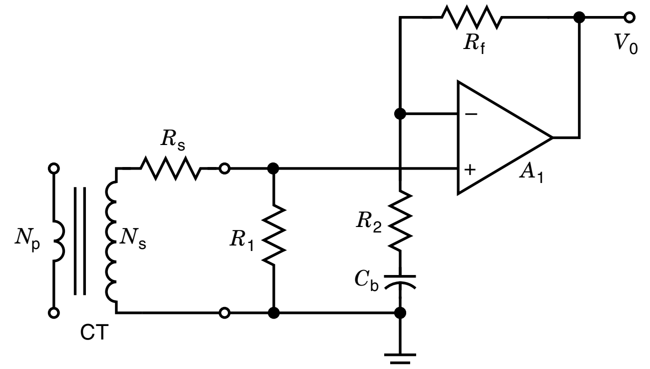

Figure 6. A noninverting buffer amplifier can be used with a CT shunt or load resistance of $R_1 < R_s$, to extend the low-frequency bandwidth.

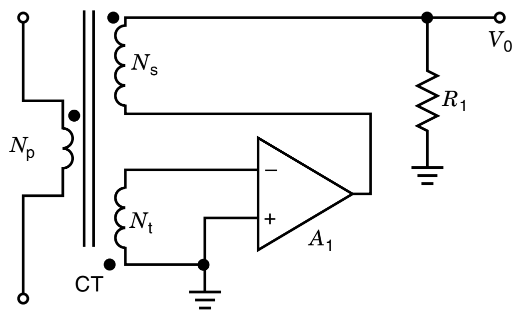

Figure 7. Using feedback to cancel the core flux of a CT extends the low-frequency bandwidth by orders of magnitude and enhances midband accuracy.

rent in the tertiary winding, and thus no $I \times R$ voltage drop, the core flux and magnetizing current errors are also forced to be nearly zero.

Whereas the circuits of Figs. 4, 5, or 6 may lower the LF end of the bandwidth by a factor of ten or more, flux cancellation can lower the minimum frequency by a factor of 1,000 or more. Both approaches also improve midband accuracy by similar factors and, in the case of flux cancellation, it becomes difficult to measure the actual midband CT error.

The circuit of Fig. 7 is not a practical circuit as shown. A dc feedback path must be provided around A_1, and stability of the circuit must be ensured at both the lower and upper unity gain crossover frequencies. With "zero" magnetizing current the LF bandwidth could extend arbitrarily close to dc. In principle, the finite open loop gain of the op amp limits the minimum frequency, but, in practice, the maximum gain must be controlled, in order to achieve a stable unity gain crossover at the low-frequency end.

Ac current transformers tend to be 10 to 100 times larger than power transformers of the same "power" or volt–ampere rating, but active flux canceling allows the size of CTs to be reduced dramatically, particularly at lower frequencies. A practical limitation is that the op amp must again be able to supply the full secondary current.

A major advantage of the circuit of Fig. 7 is that the HF bandwith is not necessarily limited by the bandwidth of A_1. If the output of A_1 has a low impedance, and the input can tolerate the tertiary voltage, the circuit reverts to normal CT operation at frequencies greater than the bandwidth of A_1.

PULSE CURRENT TRANSFORMERS

Pulse current transformers (PCT) may resemble ac current transformers, but the performance parameters of interest are somewhat different. The principle specifications for an ac CT are frequency range, maximum current, maximum burden or load resistance, and accuracy. Critical ratings for a pulse transformer are maximum peak current, rise time, pulse top droop, and the maximum current-time product for a pulse.

Low Duty Cycle PCTs

An ac current transformer with a resistive load can be operated with unipolar pulse currents, if the duty cycle is sufficiently low (often less than 1%) and other conditions are met:

1. The volt–seconds on the CT windings during the pulse must not cause the core flux to reach saturation.
2. The primary current must be zero between current pulses.
3. The interval between pulses must be long enough to allow the core to reset with the resistive load on the secondary.

Residual flux B_r in the core may become a significant portion of the saturation flux with some core materials, which reduces the pulse amp-second product. Occasionally a very small core "air gap" is introduced to reduce B_r. This also allows the PCT to operate with a slightly greater amount of dc current, but at the expense of a similar drop in magnetizing inductance, which increases pulse droop. For applications with a significant dc current, a "dc compensation" winding must usually be added, to cancel the dc flux in the core.

This type of pulse transformer is usually also rated for high-frequency ac CT use. They are usually supplied with an internal resistive load to produce an output voltage proportional to current, and are typically terminated with a 50 Ω source impedance for driving coaxial cables (2) and/or 50 Ω load impedances. Clamp-on versions are also produced, which can operate over broad frequency ranges (3).

High Duty Cycle PCTs

In some applications, like switched mode power converters, it is desirable to operate a PCT with a relatively high duty cycle. This can be readily accomplished for unipolar pulses with the addition of a rectifying diode D_1 between the CT and the resistive load, as shown in Fig. 8.

During the current pulse, D_1 conducts to connect the CT to the resistive load. After the current pulse, D_1 conduction stops and allows the magnetizing inductance of the CT to "kick back" and generate a reverse voltage, which can reset the core very quickly if the kick back voltage is much higher than the forward voltage during the pulse. Pulse duty cycles of >90% can be accommodated with this simple modification to a CT.

A reset voltage clamp is also shown in Fig. 8, which prevents reverse breakdown of D_1. The clamp typically consists of a Zener or avalanche diode Z_1 and a second diode D_2, which blocks forward conduction of Z_1 during the pulse. Schottky diodes may be used for D_1 at high frequencies, partly to minimize additional secondary voltage when the CT resistive load voltage is about a volt or less, and partly for the high switching speed of Schottky diodes. The current rating of the Schottky diode should not greatly exceed that of the actual CT secondary current, as the high diode capacitance will tend to

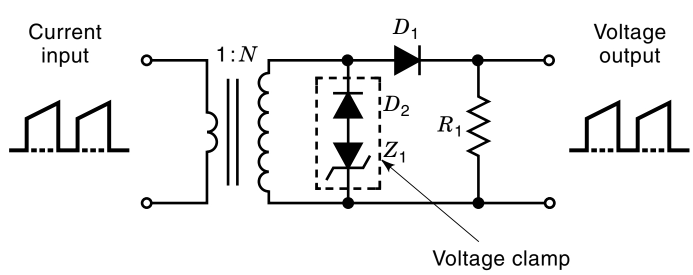

Figure 8. A CT secondary side diode allows unipolar pulse currents to be monitored.

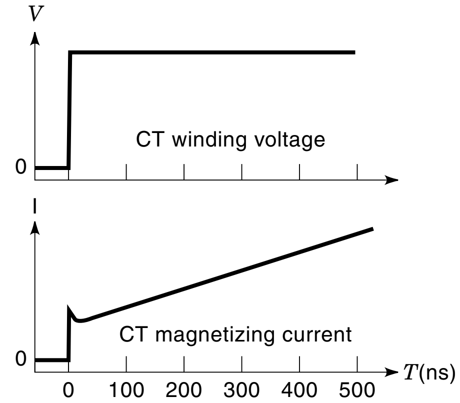

Figure 9. The pulse magnetization current of a ferrite has an initial step in current due to core losses. The effect is shown for a permeability of 10,000, where the core loss current exceeds the inductive magnetization current for about 200 to 300 ns.

slow down the rise of the reset voltage. p–n junction diodes for D_1 (and D_2) should be of the ultra-fast epitaxial variety, with reverse recovery times of 20 to 40 ns.

The voltage clamp is not always required for short current pulses, particularly for those of less than 1 to 10 μs duration. The maximum possible reset voltage on a PCT is limited by the peak magnetizing current in L_m (in Fig. 3), flowing through the core loss resistance R_m.

This core loss limit to the reset voltage also tends to limit the maximum pulse duty cycle with short pulses. However, an incomplete core reset after one pulse will leave a residual magnetizing current in L_m, which adds to the reset current after the next pulse and increases the reset voltage. The residual magnetizing current may build up to a stable value before core saturation, which could be acceptable if the increase in magnetizing error current can be tolerated. Alternatives are to use a lower permeability core with more magnetizing current, or to provide a small core reset bias current.

The core loss resistance also creates an unexpectedly high magnetizing current for short pulses, as illustrated in Fig. 9 for a typical 10,000 permeability ferrite. For a pulse duration of 100 ns, the magnetizing current is four times higher than would be calculated from the low-frequency permeability of the ferrite alone. Under these conditions, the open circuit CT kick back voltage can be no higher than $\frac{1}{3}$ of the forward voltage during the pulse.

The high duty cycle PCT is not limited to unipolar pulse currents. The antiparallel diodes D_1 and D_2 of Fig. 10 allow sensing of bipolar current pulses, while still allowing for re-

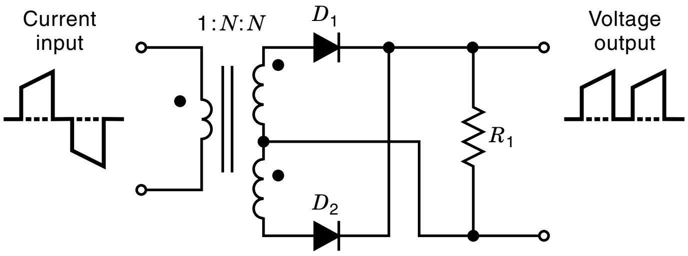

Figure 11. A center taped current transformer secondary can be used to rectify bipolar current pulses to unipolar output voltage pulses.

setting of residual core magnetizing currents during the zero current interval between pulses. Alternatively, the bipolar input current pulses can be rectified to unipolar output voltage pulses with the circuit of Fig. 11.

The secondary volt–second mismatch between the two polarities of input pulse cannot be too great, or the core will not be able to reset between pulses with the reset voltage limited to a diode drop. More reset voltage is available with p–n diodes than Schottky diodes, and several diodes could be used in series to allow greater reset voltages. A drawback is that the diode voltage also adds to the load voltage and magnetizing current errors.

It must always be borne in mind that PCT core resetting requires that the input current be zero for a finite interval. A pulse current waveform like that shown in Fig. 12 cannot be accurately sensed.

DC + AC CURRENT TRANSFORMERS

Dc current-sensing capability can be added to an ac current transformer by placing a Hall effect element in a small air gap in the CT core (4), as sketched in Fig. 13. A Hall effect element is typically a small semiconductor chip, with current input and voltage output electrodes on opposite edges, which is placed in a magnetic field with the flux normal to the surface. The output Hall effect voltage is proportional to the input current I_e and the flux density.

An amplifier A_1 senses the Hall voltage produced by a current in the primary, and produces a current in a secondary winding to cancel the core flux produced by the primary winding. Operation is similar to the flux canceling circuit of Fig. 5 in most regards, except that the flux-sensing tertiary winding is replaced with the Hall effect element. Since the Hall effect responds to dc as well as ac flux, the frequency response ex-

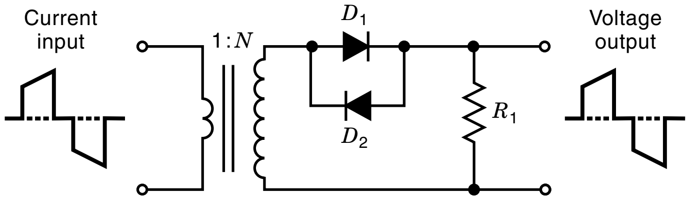

Figure 10. Antiparallel CT secondary diodes allow bipolar pulse currents to be monitored.

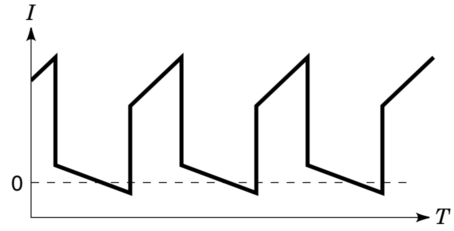

Figure 12. A pulse current waveform with no finite I = 0 interval, such as that shown here, cannot be monitored with a pulse current transformer.

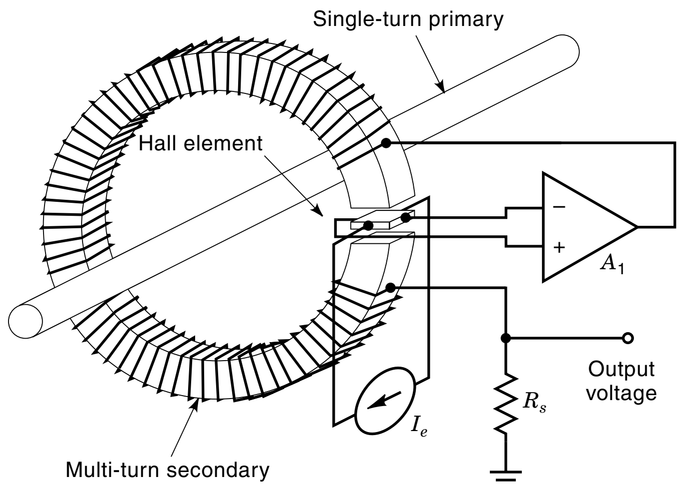

Figure 13. A Hall effect element is placed in an air gap in the core of a flux-cancellation CT circuit, to extend the low-frequency response to dc.

tends to dc. The necessary gap in the core of the CT, however, degrades the performance of the device. In practice, the accuracy is limited to a few tenths of a percent.

As in the flux-canceling circuit of Fig. 5, the bandwidth can exceed that of the amplifier, as the circuit can revert to conventional ac CT operation at high frequencies, which may range from tens of kilohertz to tens of megahertz.

Other designs eliminate the flux canceling winding in Fig. 13, and just amplify the output of the Hall effect element directly for current indication. The advantages are lower cost and power consumption, at the expense of reduced bandwidth and accuracy.

PRECISION DC CURRENT TRANSFORMERS

Although the conventional CT cannot measure dc current directly, a number of ingenious techniques have been developed (4–7) which use magnetic transformers to sense and measure dc current, with accuracies of parts per million in some cases. Typically, these approaches use two to four magnetic cores, one or two of which are driven to the verge of saturation or beyond by an ac source. A winding with the dc current to be measured links all cores, as does a dc compensation or canceling winding. Any dc amp-turn imbalance between the measured and compensation windings causes an asymmetrical response in the cores with an ac flux, which is sensed by a feedback and control circuit which adjusts the compensation current to remove the imbalance. The measured current is known from the compensation current and the turns ratios of the windings. Some of these circuits are suitable for measuring ac as well as dc current. A detailed description of these specialized approaches is beyond the scope of this overview article.

AIR-CORED CURRENT TRANSFORMERS (ROGOWSKI COILS)

The construction of classical Rogowski coils (8) is similar to that of ac current transformers, but with a nonmagnetic or "air" core. The "secondary" winding is usually in the shape of a toroid, with the return conductor brought back inside the winding to minimize spurious response to ambient magnetic fields, as shown in Fig. 14.

The output of the Rogowski coil is proportional to the rate of change of the enclosed magnetic field, or to the dI/dt of a conductor through the middle of the toroid. The response of a Rogowski coil is theoretically uniform to any current inside the hole of the toroid, and zero to any current outside the toroid. In practice, the response is usually slightly nonuniform to internal currents, with a slight spurious response to external currents, depending on the uniformity of the coil winding. These nonideal responses are typically within a few percent for well-constructed coils.

The output voltage may be used directly to monitor the amplitude of sinusoidal currents of a constant frequency, but the voltage is usually integrated to produce a flat response over a range of frequencies (9,10). If phase shifts in the integrator are well controlled, the time domain fidelity of the waveform can be within one percent over a broad range of frequencies (11).

Rogowski coils have several advantages over conventional CTs. Although they cannot measure dc current, their performance is not affected even by very large dc currents. They cannot be saturated by large ac currents, and actually become increasingly advantageous as ac current increases. Conventional CTs require secondary turns to increase as measured currents rise to keep the output current constant, while Rogowski coils require fewer turns at higher currents, to produce a constant output voltage. Thus Rogowski coils can be much smaller than CTs at high current levels.

Magnetic Cores for Rogowski Coils

Rogowski coils tend to become less useful at low currents and low frequencies, where an excessive number of turns becomes required to produce a useful output voltage. This can be somewhat overcome by using a magnetic core, which increases the output in proportion to the permeability. However, the effective permeability must be stable and well defined, and insensitive to frequency, temperature, and dc and ac flux density over the operating conditions. This stability does not occur with solid ferromagnetic cores, but can be achieved with a core (such as ferrite) with one or more discrete air gaps. Un-

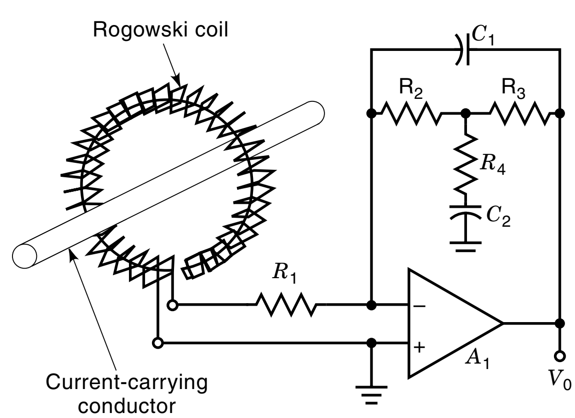

Figure 14. The output of a Rogowski coil is proportional to the derivative of the sensed current, and requires an integrator to reconstruct the primary current waveform.

fortunately, the discrete gaps will increase sensitivity to external fields and primary conductor placement, unless a large number of equally spaced air gaps are used.

The best alternative is to use a "distributed gap" material. The most suitable types are molypermalloy powder (MPP) cores, which are available in permeabilities of 14 to 300. Achievable stabilities vs. flux range from better than $\pm$ 1% at $\mu = 14$ to about $\pm$ 4% at $\mu = 300$, although the initial standard tolerance of $\pm$ 8% on permeability usually requires calibration for reasonable accuracy. The -3 dB permeability rolloff frequency ranges from about 600 kHz for $\mu = 300$ to about 20 MHz for $\mu = 14$.

Powdered iron cores are not recommended, in general, because they typically have relatively low permeabilities and exhibit large changes in permeability with ac flux density. Possible exceptions are the unannealed carbonyl iron powder cores, which have stable permeabilities of about ten or less, but are useable to hundreds of megahertz.

Note that the output voltage of a CT (for a given load resistance) is proportional to N_p/N_s, while the output voltage of a Rogowski coil is proportional to $N_p \times N_s$. The sensitivity to low currents with either device can be increased by increasing the number of primary turns.

Design of Rogowski Coils

If the secondary inductance of the Rogowski coil is known, the output voltage for a sinusoidal primary current is

$$V_o = 6.28 f L_s I_p (N_p/N_s) \tag{6}$$

where V_o = output voltage, in V; f = frequency, in Hz; L_s = secondary inductance, in H; I_p = primary current, in A; N_p = number of primary turns (usually $N_p = 1$); and N_s = number of secondary turns.

For an arbitrary current waveform

$$V_o = L_s (N_p/N_s)(dI_p/dt) \tag{7}$$

where V_o represents the instantaneous output voltage, in V and dI_p/dt is the time derivative of primary current, in A/s. The inductance of a toroidal winding (or any cored inductor) is

$$L = (1.257 \times 10^{-6})(N^2 A_e \mu)/l_e \tag{8}$$

where L = inductance, in H; N = number of turns; A_e = magnetic core area, or winding area for air core, in m^2; μ = relative core permeability ($\mu = 1$ for air); and l_e = magnetic path length, in m.

Replacing L in Eqs. (6) and (7) with the formula in Eq. (8),

$$V_o = (7.896 \times 10^{-6})(f I_p N_s N_p A_e \mu)/l_e \tag{9}$$

for sinusoidal currents, and

$$V_o = (1.257 \times 10^{-6} N_s N_p A_e \mu)(dI_p/dt)/l_e \tag{10}$$

for an arbitrary primary current waveform.

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