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CURRENT MEASUREMENT IN POWER ELECTRONIC AND MOTOR DRIVE  
APPLICATIONS – A COMPREHENSIVE STUDY

by

ASHABEN MEHUL PATEL

A THESIS

Presented to the Faculty of the Graduate School of the

UNIVERSITY OF MISSOURI-ROLLA

In Partial Fulfillment of the Requirements for the Degree

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

2007

Approved by

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## ABSTRACT

Current measurement has many applications in power electronics and motor drives. Current measurement is used for control, protection, monitoring, and power management purposes. Parameters such as low cost, accuracy, high current measurement, isolation needs, broad frequency bandwidth, linearity and stability with temperature variations, high immunity to  $dv/dt$ , low realization effort, fast response time, and compatibility with integration process are required to ensure high performance of current sensors. Various current sensing techniques based on different physical effects such as Faraday's induction law, Ohm's law, Lorentz force law, magneto-resistance effect, and magnetic saturation are studied in this thesis. Review and examination of these current measurement methods are presented.

The most common current sensing method is to insert a sensing resistor in the path of an unknown current. This method incurs significant power loss in a sense resistor at high output currents. Alternatives for accurate and lossless current measurement are presented in this thesis. Various current sensing techniques with self-tuning and self-calibration for accurate and continuous current measurement are also discussed. Isolation and large bandwidth from dc to several kilo-hertz or mega-hertz are the most difficult, but also most crucial characteristics of current measurement. Electromagnetic-based current sensing techniques, which are used to achieve these characteristics, are analyzed. Many applications require average current information for control purposes. Different average current sensing methods of measuring average current are also reviewed.

## ACKNOWLEDGEMENT

I owed thanks to many people who helped me through the Master's program at the University of Missouri-Rolla. My parent-in-laws deserve thanks for being so loving, supportive, and patient during all my years in graduate school. I am deeply indebted to my advisor, Dr. Mehdi Ferdowsi, whose stimulating suggestions and encouragement helped me throughout this research and the writing of this thesis. He provided expertise, inspiration, and invaluable support for this research, without which my studies could not have completed.

I would like to extend my sincere appreciation to Dr. Keith Corzine and Dr. Badrul Chowdhury for serving as my committee members and examining this thesis. The financial assistance provided to me in the form of a Graduate Research Assistantship through the Department of Electrical Engineering at the University of Missouri-Rolla is gratefully acknowledged.

Finally, I wish to thank my husband Mehul, who tricked me into the Master's program in the first place and without whom I could have never finished.

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# **1. INTRODUCTION**

To be able to control, protect, or monitor a system, certain information about the system is required. In electrical systems this information mainly consists of instantaneous, peak, or average values of the voltage and current signals. Hence, accuracy of current measurement plays an important role in the dynamic performance, efficiency, and safety of an electrical system. Knowing the current information about a system is also important for power management purposes. Applications of current sensing in motor control and power conversion are the main focal points of this thesis. Several detailed descriptions of such systems are provided in the following sub-sections.

## **1.1. MOTOR DRIVES**

Motor current is used as a control variable in motor drives. Vector control and direct torque control require current sensing for control purposes [1-8]. Speed sensorless and voltage sensorless control approaches require motor current measurement to provide accurate control with lower cost, noise, and complexity. Motor current measurement is also required to remove torque ripple in order to provide smooth torque [9, 10]. Modern motor drives are beginning to be digitally controlled and the control elements of such drives require accurate motor current feedback. Therefore, current sensing is of great importance to digitally controlled ac motor drives [11]. Current controlled pulse width modulation (PWM) inverters are used to serve as ac drives. These controllers are classified as hysteresis, ramp comparison, and predictive controllers, all of which require motor current information to determine inverter switching states [12]. Furthermore, information of stator parameters is important for several control schemes. Stator current

measurement is used for the estimation of these parameters [13]. Stator current information is also used for rotor velocity tracking control instead of measuring both rotor flux and velocity [14].

Motor current information is also required to detect rotor position and to check the motor's insulation condition in several applications. Starting from an unknown rotor position may create temporary reverse rotation or starting failure in permanent magnet motors, which is not acceptable in many applications. Rotor position can be detected by using phase current information [15]. Electrical insulation is the most critical component for operation of electrical motors. Stator insulation failure during motor operation can lead to motor failure, resulting in a costly outage. Insulation condition indicators such as capacitance and dissipation factor are calculated based on the measurement of the differential leakage currents of each phase winding [16].

Current sensing is also important for safety and efficient regulation of all the equipment using industrial motor drives, for example, a motor which is used to open and close elevator doors. An elevator door must react to an object or a person hindering closure. The obstacle detection in the elevator is performed by sensing motor currents. The obstacle creates higher torque, which causes higher currents in the motor. This current information is used to trigger the signal to open the door. Motor current measurement is also important for the elevator's smooth acceleration and stopping and for placing the cabin exactly level to its destination [17]. Another example is speed control in a forklift. The motor in the forklift is required to adjust the speed and the ascending and descending of the fork. For smooth and steady forklift operation, motor current information in the forklift is required to lift the fork with the same precision and

speed regardless of the amount of load it needs to transport. By measuring the motor current, the forklift adjusts the power required to perform the task.

## **1.2. POWER ELECTRONIC CONVERTERS**

A current sensor in a power electronic converter is used to measure the constantly varying current in order to provide feedback for control purposes. This leads to the accurate and smooth adjustment of converter operation. Current-mode control is implemented in converters to regulate the output voltage. Current-mode control approaches for converters are classified as peak, average, valley, hysteretic, PWM conductance, and output current feed-forward control. All current-mode control types require inductor waveform information to provide appropriate transient response [18-23].

Current sensing in converters is also required for mode hopping, current sharing, ripple cancellation, and integrated power electronic modules (IPEMs). In mode hopping, the sensed converter current is used to determine when to switch between continuous conduction mode and discontinuous conduction mode [24]. Current information is required for current sharing in parallel or multi-phase converters so that the stress among the paralleled dc–dc converters is balanced [25]. In some applications, an active ripple filter is required in power electronic converters to cancel the current ripple [26]. This active ripple filter requires information about converter current. IPEMs are used in vehicles to reduce cost, package size, and interconnect. Current sensing is the most important requirement for an IPEM [27, 28].

Converter current information is also important for power generation applications, for example, inverters used in gas co-generators. In gas co-generators, gas is used to produce steam to drive the turbine to produce electricity through an inverter. This inverter



is directly connected to the household electrical network. To ensure that the output of the inverter is perfectly matched with the electrical network, the output current of the inverter must be known. The grid needs to be supplied with continuous power at a controlled frequency and voltage [29]. Another example is wind generation. The electricity produced by on-shore or off-shore wind turbine generators needs to be safely fed into the grid. To do so, it is required to transform and optimize the generation of electricity from the wind turbine. Current measurement is required for every converter in wind energy turbines to enable optimal control and protection. Current sensors also continuously measure the current in the converters to position the turbine in relation to the wind [30].

### **1.3. PROTECTION AND MONITORING**

Protecting supply circuits against overloading conditions requires the system to sense its current delivery. The over current is a severe fault situation that can result in the failure of the device if appropriate action is not taken in time. The over current is caused by a short circuit, malfunction, or component failure. Current sensing is required for fault detection and its minimization [31, 32]. It enables emerging state-of-art switching supplies to improve transient response, efficiency, and compensation performance [33]. Current measurement is also required for ground fault detection on power lead for personnel protection. The safety lights on airport runways, towers, and chimneys require continuous current sensing.

Furthermore, current monitoring is required for process control, maintenance, and infrastructure management. Current consumption by industrial heaters, electroplating processes, smelting furnaces, electrolysis processes, electric dip coatings, etc, needs to be monitored for process control. A motor supplying a cutting tool drawing too much

current indicates that the cutting tool is blunt and needs to be sharpened or replaced. In air conditioning, ventilation, and heating applications, the contamination amount on an air filter is determined from the amount of current consumed by the fan's electric drive.

#### **1.4. BATTERY APPLICATIONS**

Power management in vehicles includes battery management, battery's state of health, and auto-disconnection of a battery; all of which require battery current information.

Battery current information is important in finding battery's state of charge (SoC) [34-36]. A battery's SoC is defined as its available capacity as a percentage of its rated capacity. Finding the SoC during fast charging in Ni-MH (nickel-metal hydride) battery prevents oxygen production under high charging currents. Only 10% oxygen generation will create 5-6 times more heat than normal operation. SoC information is also required to avoid the extreme operating conditions such as fully charged and fully discharged conditions.

#### **1.5. SELECTION CRITERIA**

There are many issues that need to be considered when choosing a current sensor for a particular application. The designer needs to think about all aspects of an application. Table 1.1 lists all the criteria which require special consideration while selecting a current sensor [37].

Table 1.1. Selection criteria for a current sensor

Parameters	Types	Criteria
Electrical	Current types	dc, ac, and complex
	Voltage	Primary working voltage, dielectric withstand voltage, compliance to the relevant standards, applicable standard for isolation, over voltage category, impulse withstand voltage, single or reinforced insulation
	Ranges	Nominal current, peak current, transient current, maximum peak value
	Required output signal	Output value at nominal or peak current, selection of desired or necessary output load impedance
	Measurement accuracy	Required accuracy at 25°C (ambient temperature), dc offset and non linearity of the output signal, global accuracy within the operating temperature range, the offset drift, and the gain drift
	Available power supply	Power supply voltage, maximum allowable current consumption
Environmental	Temperature	Extreme storage temperature, minimum and maximum working temperatures
	Presence of external field	Fields from transformers or inductors, magnetized materials in the area, other currents similar or greater amplitude, external current identical to the measured one
	Vibration and shock	Standards and levels to be considered
Dynamic	di/dt	Match between di/dt to be measured and the sensor's response and rise time, maximum possible di/dt overloads to be withstood by the sensor but not to be measured, maximum sensor recovery time after di/dt overload
	dv/dt	Maximum error allowed during dv/dt, maximum settling time of sensor after dv/dt disturbance
	Frequency range	Harmonic content, frequency range to be measured, fundamental operating frequency
Mechanical interfaces	External dimensions	Aperture location, connector location, required clearance distances, maximum specified dimensions
	Primary electrical connection	Printed circuit board (PCB) pins, busbar dimensions, aperture size and shape, other connections
	Secondary electrical connection	PCB pins, screw lugs, fasten tabs, connector, other connections
	Package fastening	PCB mount, aperture mount, rail mount, panel mount

## 1.6. LOCATIONS FOR SENSING CURRENT

Several possible current sensing locations in a power electronic converter are shown in Figure 1.1 [38]. The current sensing locations are selected based on the desired information and on which circuit branch it is available.

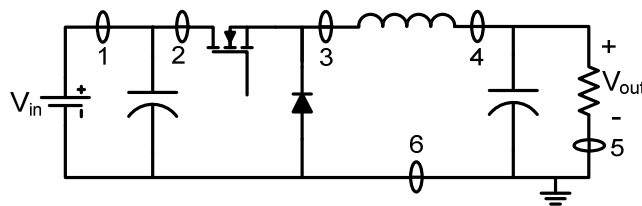


Figure 1.1. Possible current sensing locations in typical power switching stage

Sensing location 1 is good for sensing total input current (dc and ripple), which is required for power factor control, short circuit protection, and input power calculations. Sensing location 2 measures the switch current. It is a good location for peak current-mode control and overall short circuit protection. Sensing location 3 gives average current information. This location is not effective due to the wide voltage swing. Sensing location 4 gives information about both instantaneous inductor current and average load current, which are useful for average current-mode control. Sensing at location 5 is easier than all of the locations discussed above. This location yields true information about the output current. However, this information is not useful for short circuit protection. Grounding will affect the regulation at this location because the voltage across sensing element needs to be subtracted from the output voltage  $V_{out}$ . The problem with location 6 is the separation of load grounding from the input.

In motor drives systems, protection and control are implemented by measuring the current at different positions, as shown in Figure 1.2. Sensing location 1 protects the motor inverter by measuring the input current. Sensing location 2 protects the semiconductors by direct current measurement. Sensing location 3 effectively senses the motor current as a part of the speed control system. Sensing location 4 detects the dc link current and polarity to protect semiconductor devices.

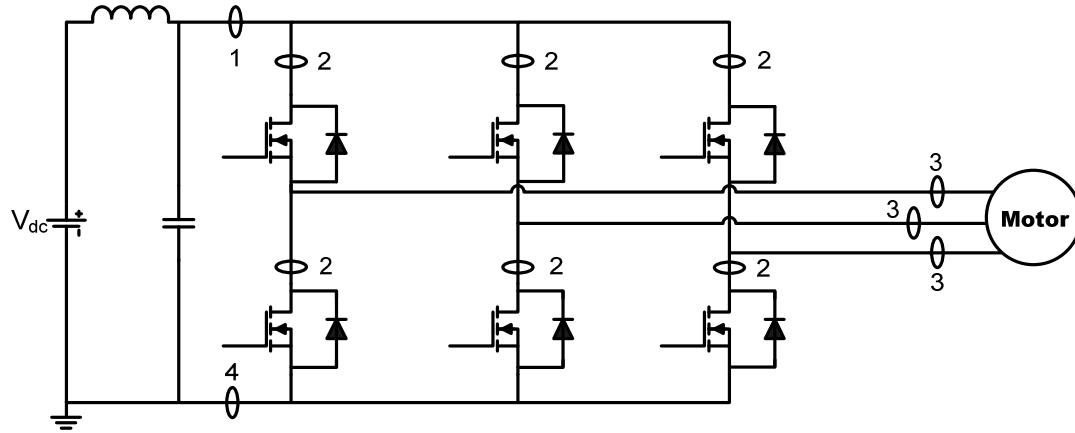


Figure 1.2. Possible current sensing locations in motor drives

## 1.7. THESIS ORGANIZATION

In this thesis, several current sensing techniques have been reviewed and evaluated which are different in accuracy, complexity, cost, operating range, linearity, current magnitude, bandwidth, electric isolation, switching noise sensitivity, and ac response. Different resistive-based current sensing methods are presented in Section 2. Section 3 discusses various electromagnetic-based current sensing techniques. Section 4 provides information about different current sensing techniques with self-tuning and/or self-calibration. Average current sensing techniques are discussed in Section 5. At the end of each Section, all of the current sensing techniques presented in the Section are compared. Finally, Section 6 presents an overall evaluation of current sensing techniques.

## 2. RESISTIVE-BASED CURRENT SENSING TECHNIQUES

According to Ohm's law, when current is flowing through a resistor, there is a voltage drop across it. Therefore, a resistor can be used to measure the current in a circuit and translate it into a voltage. This voltage signal is a representation of the current, which can be easily measured and monitored by control circuitry. The sense resistor, the resistor used for current measurement, must have low resistance to minimize its power consumption [39].

### 2.1. RESISTIVE-BASED CURRENT SENSING TECHNIQUE USING AN EXTERNALLY ADDED SENSE RESISTOR

A current detecting circuit of a dc-dc converter using an external sense resistor  $R_{sense}$  is shown in Figure 2.1.  $R_{sense}$  acts as a current to voltage converter.

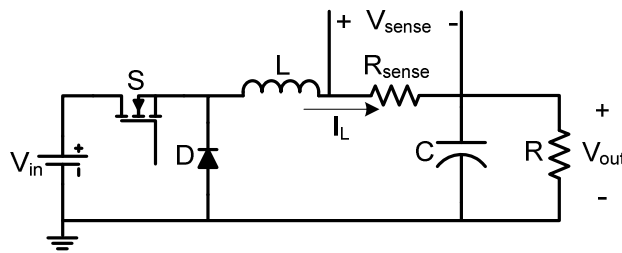


Figure 2.1. Current sensing using an externally added sense resistor

Discrete resistor and PCB trace are the most common types of  $R_{sense}$  [40-44]. If  $R_{sense}$  value has tight tolerance, this technique is accurate for low current values. When resistor  $R_{sense}$  is crossed by inductor current  $I_L$  in a buck converter, current  $I_L$  is sensed by

measuring the voltage ( $V_{sense}$ ) across sense resistor  $R_{sense}$ , as depicted in Figure 2.1. Current  $I_L$  is given by (2.1).

$$I_L = \frac{V_{sense}}{R_{sense}} \quad (2.1)$$

Due to its simplicity and accuracy, this method is used for power factor correction and over current protection. The criteria for the selection of  $R_{sense}$  are voltage drop, accuracy, efficiency and power dissipation, parasitic inductance, and cost. The drawback of this technique is the power loss incurred by resistor  $R_{sense}$ . Therefore, this method is inefficient for dc-dc converters. Additionally, it does not provide measurement isolation from transient voltage potentials on the load. A noise filter is required to reduce the noise in the signal output, which will affect the overall system bandwidth. This technique is not applicable to high performance dc-dc converters whose efficiency requirements are more than 85-90%.

If resistor  $R_{sense}$  is placed on the load side (see Figure 2.1), it only gives information about the output current. Input current measurement is required to achieve adaptive voltage positioning in voltage regulators. Input current sensing is achieved by placing  $R_{sense}$  on the input side. A modified metal oxide semiconductor (MOS) current sensing technique using a current mirror to overcome power loss incurred by sense resistor is presented in [45]. This method measures the current without requiring the entire output current to pass from the series sense resistor. This technique uses the microelectronic current mirroring concept. The current passing through the sense resistor is proportional to the output current and its magnitude is smaller.

## 2.2. RESISTIVE-BASED CURRENT SENSING TECHNIQUE USING THE INTERNAL RESISTANCE OF AN INDUCTOR

To avoid the use of sense resistor and to reduce the power loss it creates, an inductor is used for current sensing, as depicted in Figure 2.2 [38]. This method is appropriate for low voltage power converters. Resistor  $R_L$  in series with inductor  $L$  is the internal resistance or direct current resistance (DCR) of the inductor winding.

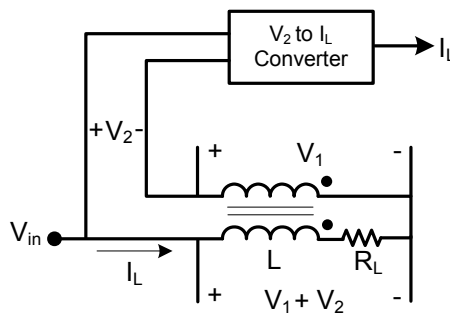


Figure 2.2. Current sensing using the internal resistance of an inductor

The voltage across the main inductor is made of two voltages  $V_1$  and  $V_2$ , which are given by (2.2).

$$V_L = V_1 + V_2 = L \frac{dI_L}{dt} + I_L R_L \quad (2.2)$$

An extra winding with an equal number of turns and minimum current loading is coupled with the main inductor, as shown in Figure 2.2. The voltage across the extra winding is  $V_1$  due to the equal number of turns. If the voltages of both windings are added, the resulting voltage is simply  $I_L R_L$  drop. The disadvantage of this technique is that the measured current is inaccurate because  $V_2$  which is the difference between two



large voltages is quite small. Inductor winding is built by copper wire, so, the temperature coefficient of copper's resistivity also applies.

### 2.3. MOSFET $R_{DS}$ -BASED CURRENT SENSING TECHNIQUE

A lossless MOSFET drain-source resistor-based current sensing method, which eliminates the need for resistor  $R_{sense}$  for current sensing, is depicted in Figure 2.3 [46-49]. The drain-source resistance ( $R_{DS}$ ) of the MOSFET is useful for current measurements. A MOSFET acts like a resistor when it is on, so it is possible to determine current  $I_L$  by measuring the drain-source voltage of the switch.

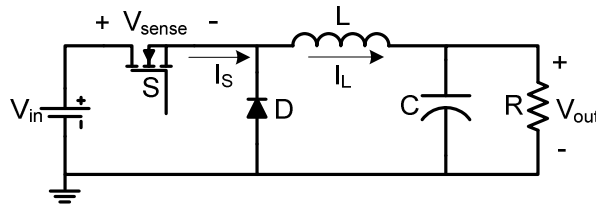


Figure 2.3. MOSFET  $R_{DS}$ -based current sensing

Current  $I_L$  can be determined by measuring switch current  $I_S$ . As switch S is on in a buck converter during the time interval  $0 < t < dT$  only, a sample-and-hold circuit is required to sense current  $I_L$ . Current  $I_S$  is determined by measuring the voltage across resistor  $R_{DS}$ . Current  $I_S$  is given by (2.3).

$$I_S = I_L = \frac{V_{sense}}{R_{DS}} \quad (2.3)$$

A MOSFET is in the triode region when it is conducting and voltage across it is low. MOSFET acts like a resistor in this case. The resistance  $R_{DS}$  is given by (2.4).

$$R_{DS} = \frac{L}{W\mu C_{OX}(V_{in} - V_T)} \quad (2.4)$$

Where  $\mu$  is the electron mobility,  $C_{ox}$  is the oxide capacitance, and  $V_T$  is the threshold voltage. The value of resistance  $R_{DS}$  is given in the MOSFET data sheet. The accuracy of this method depends on the tolerance of resistor  $R_{DS}$ . The datasheet gives only the maximum and typical values of resistor  $R_{DS}$ , not the minimum.

MOSFET  $R_{DS}$ -based current sensing is a very low cost method for current sensing. One of the uses of this method is in low-voltage, high-current point of load converter applications. However, the actual current threshold will vary with resistance  $R_{DS}$ . Resistance  $R_{DS}$  varies with temperature, gate drive voltage, and individual devices. Variations in the gate drive amplitude result in a poor current sensing accuracy in multiphase converters. Higher switching frequencies and higher input voltages make this method complicated.

#### **2.4. CURRENT-SENSING POWER MOSFET-BASED CURRENT SENSING TECHNIQUE**

Current-sensing power MOSFET technique is more accurate than the MOSFET  $R_{DS}$ -based current sensing technique [50-57]. A power MOSFET consists of a large number of parallel connected MOSFET cells. The gates, sources, and drains of all transistor cells are connected together. However, when power MOSFET is tuned on to sense the current, the entire amount of the sense current passes through sense resistor, so there is a significant and undesirable power loss. To avoid this power loss, few cells in

the power MOSFET are utilized to provide sensing FET (SENSEFET). Remaining transistor cells are used to provide the switching MOSFET. This new power MOSFET structure is known as a current-sensing power MOSFET. The current-sensing power MOSFET is a parallel connection of a SENSEFET ( $S_2$ ) and switching MOSFET ( $S_1$ ), as shown in Figure 2.4. Switch  $S_2$  has relatively fewer transistor cells to provide a small sensing signal proportional to the switch  $S_1$  current. A sense resistor is placed on the scaled down current to reduce the power dissipation; therefore, this technique is used to provide accurate and lossless current measurement. A current-sensing power MOSFET is symbolically represented in Figure 2.4.

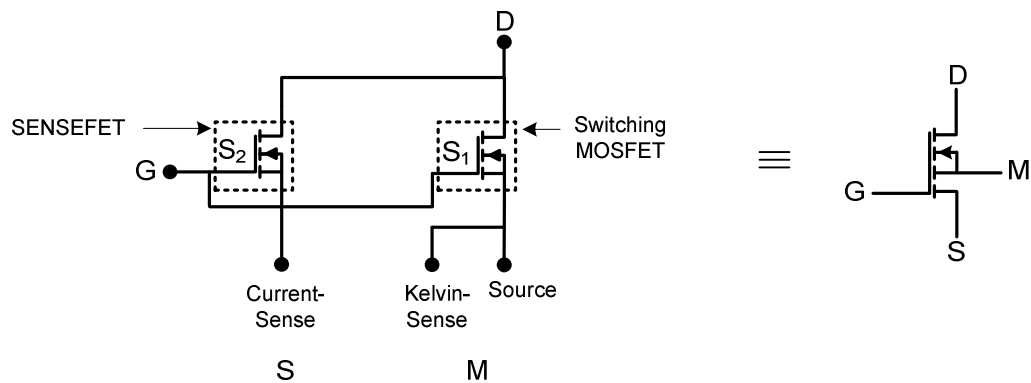


Figure 2.4. Current-sensing power MOSFET

The current-sensing power MOSFET is a five terminal device. Switches  $S_1$  and  $S_2$  have identical unit cell structures and reside on the same silicon substrate. Their gate terminals are connected to a common terminal G and their drain terminals are connected to a common terminal D. Source terminals of switch  $S_2$  are connected to a current sense or mirror terminal S and source terminals of switch  $S_1$  are connected to a main terminal

M which consists of Kelvin-sense terminal K and source terminal S. A Kelvin-sense terminal is shorted internally to the source terminal of switch  $S_1$  to bypass the packaging and interconnection parasitic resistance associated with switch  $S_1$ . It provides more accurate current sensing.

The most common practice of using  $S_1$  and  $S_2$  for current sensing in a buck converter is shown in Figure 2.5.  $N$  is a predetermined ratio of the transistor cells of switch  $S_1$  to switch  $S_2$ . For example,  $N$  may be in the order of 100 to 1000. If  $N$  increases, accuracy of the circuit decreases. All transistor cells of the current-sensing power MOSFET are similar; therefore, if the sources of switches  $S_1$  and  $S_2$  are virtually connected, switches  $S_1$  and  $S_2$  pass currents in the ratio  $N:1$ , and have resistances in ratio  $1:N$ .

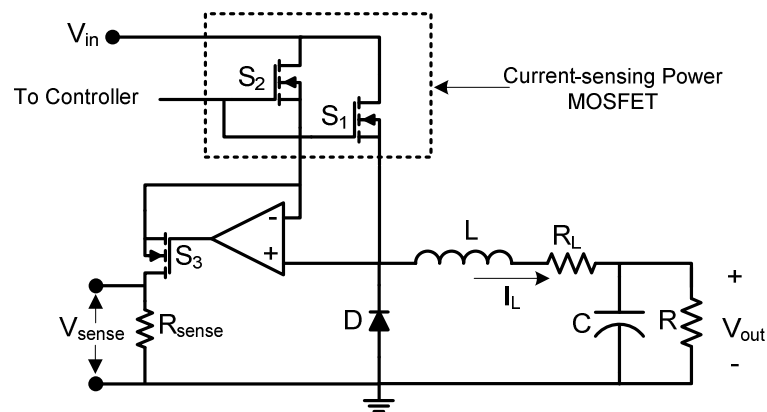


Figure 2.5. Current-sensing power MOSFET-based current sensing

An operational amplifier in Figure 2.5 is used to make the gate-to-source voltages of both switches equal. Since only a predetermined fraction of total current is passing

through  $R_{\text{sense}}$ , power dissipation in  $R_{\text{sense}}$  is low. As current passing through  $S_1$  is not continuous, a sample-and-hold circuit holds the sensed current when  $S_1$  is off. The sense voltage  $V_{\text{sense}}$  in this technique is given by (2.5).

$$V_{\text{sense}} = R_{\text{sense}} \left( \frac{I_L}{N} \right) \quad (2.5)$$

The use of current-sensing power MOSFET is impeded by its limited availability and high cost. This method is not applicable to high frequency systems because it introduces switching transients and noise to the sense signal. The accuracy of this method is limited because it works only when switches  $S_1$  and  $S_2$  are matched. Because this technique's current ratio is N:1, a low degree of coupling between  $S_1$  and  $S_2$  can induce a significant error and large spikes are injected in the sense signal during high di/dt periods.

## 2.5. PWM-VSI SWITCHING STRATEGIES

For the purpose of motor current sensing, the switching strategies of an inverter are discussed in this sub-section, assuming its load is a three phase star connected motor. However, the same reasoning is applied to single phase or poly-phase and star or delta connected synchronous or asynchronous motors.

PWM techniques are used in voltage source inverters (VSI) to control different types of motors. A three phase PWM-VSI driven ac motor is depicted in Figure 2.6. All three legs of an inverter are supplied by a dc voltage source. A large capacitor connected at the input terminals tends to make the input dc voltage constant and suppresses the harmonic fed back to the source. A control signal is provided through the gate of each transistor.

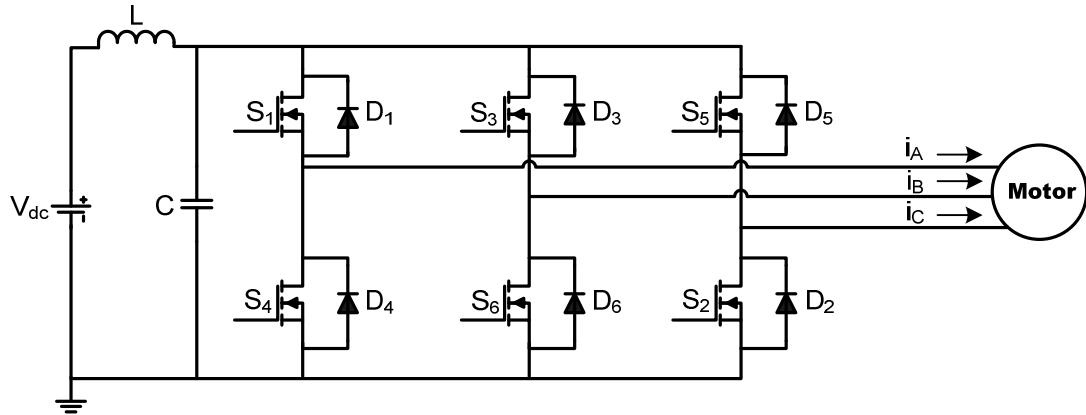


Figure 2.6. PWM-VSI driven ac motor

Switches in VSI are numbered in the sequence in which they are triggered. Three phase VSI works with either  $180^\circ$  or  $120^\circ$  conduction mode [58, 59].

If three phase VSI works with  $180^\circ$  conduction mode, switch pairs  $S_1, S_4$ ;  $S_3, S_6$ ; and  $S_5, S_2$  are turned on with a time interval of  $180^\circ$ , which means  $S_1$  conducts for the first  $180^\circ$  and  $S_4$  for the next  $180^\circ$  of a cycle, as shown in Figure 2.7.

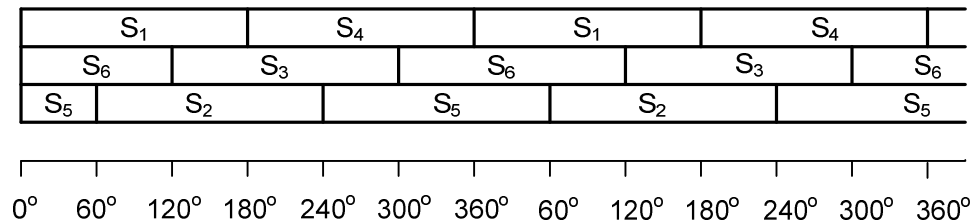


Figure 2.7. Switching sequence of a  $180^\circ$  VSI conduction

Switches in the upper group, i.e,  $S_1, S_3,$  and  $S_5,$  conduct at an interval of  $120^\circ$ , which implies that if  $S_1$  is fired at angle  $0^\circ$ , then  $S_3$  must be fired at angle  $120^\circ$  and  $S_5$  at

angle  $240^\circ$  [59]. Each arm of a three phase VSI has three possible states, so a total of twenty-seven combinations are possible. Only eight states that do not involve both transistors of any phase to be on at once are used. During remaining nineteen states the phase voltage depends on the conduction of freewheeling diodes. Eight different states of six switches are given in Table 2.1 for  $180^\circ$  VSI conduction. A simple logic circuit is required to identify states 1-6 of Table 2.1. Control signals are then fed to the inverter. However, control signals provided to the upper legs of inverter are opposite to the control signals provided to the lower legs.

Table 2.1. Different states of a  $180^\circ$  VSI conduction

State	S <sub>1</sub>	S <sub>2</sub>	S <sub>3</sub>	S <sub>4</sub>	S <sub>5</sub>	S <sub>6</sub>
0	off	on	off	on	off	on
1	on	off	off	off	on	on
2	on	on	off	off	off	on
3	on	on	on	off	off	off
4	off	on	on	on	off	off
5	off	off	on	on	on	off
6	off	off	off	on	on	on
7	on	off	on	off	on	off

If three phase VSI works with  $120^\circ$  conduction mode, each switch conducts for  $120^\circ$  of a cycle, as shown in Figure 2.8. Like  $180^\circ$  mode,  $120^\circ$  mode VSI also requires six steps, each of  $60^\circ$  duration, to complete one cycle of the output ac voltage. The six different possible states of six switches are given in Table 2.2 for  $120^\circ$  VSI conduction.

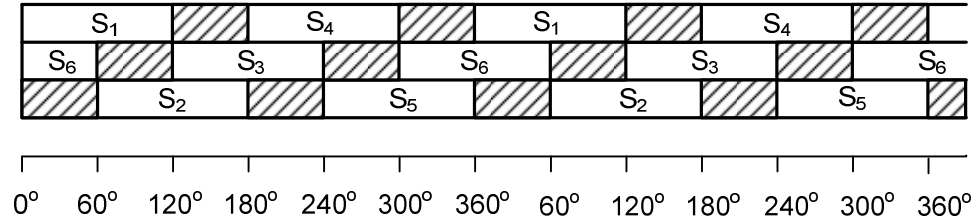


Figure 2.8. Switching sequence of a 120° VSI conduction

Table 2.2. Different states of a 120° VSI conduction

State	S <sub>1</sub>	S <sub>2</sub>	S <sub>3</sub>	S <sub>4</sub>	S <sub>5</sub>	S <sub>6</sub>
1	on	off	off	off	off	on
2	on	on	off	off	off	off
3	off	on	on	off	off	off
4	off	off	on	on	off	off
5	off	off	off	on	on	off
6	off	off	off	off	on	on

## 2.6. CURRENT SENSING TECHNIQUES FOR PWM-VSI DRIVEN AC MOTOR USING EXTERNALLY ADDED SENSE RESISTORS

Different current sensing strategies for measuring motor current using externally added sense resistors are described in this sub-section.

**2.6.1. An Externally Added Sense Resistor in Each Leg of the Inverter.** For certain applications, the current must be measured in individual legs of the motor controller inverter for various purposes. A VSI with current sensors in the three low side inverter switches to achieve closed-loop current regulation is shown in Figure 2.9. Sense resistors are used to sense the current flowing through each leg of the motor controller circuit [60-62]. Sense resistors are connected to the source pin of each low side transistor.



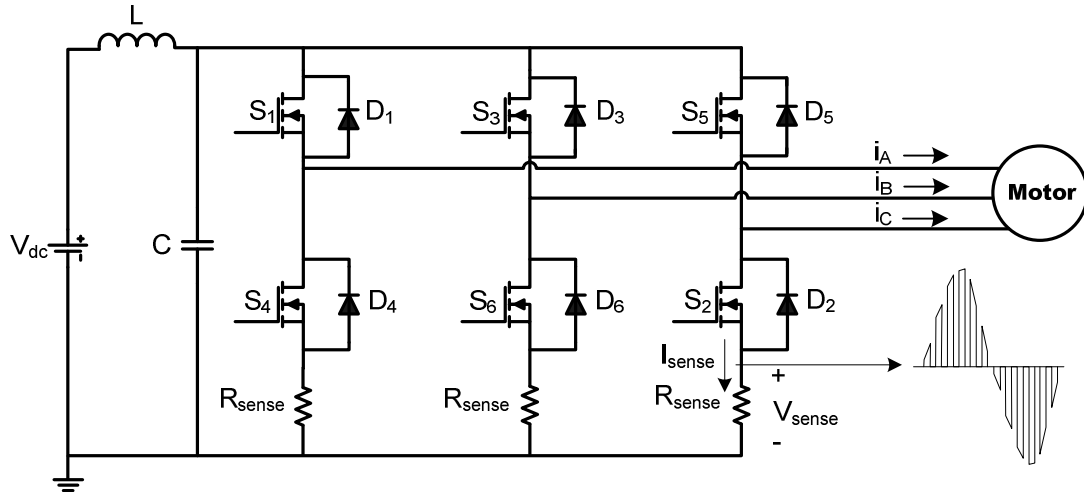


Figure 2.9. Motor current measurement using an externally added sense resistor in each leg of VSI

When VSI operates with 180° conduction, at any given instant of time all three motor phases carry currents. Therefore, when two lower leg switches are on, two of the three motor phase currents are determined by measuring the voltages developed across  $R_{\text{sense}}$  ( $V_{\text{sense}}$ ) to obtain complete motor line current information. For instance, if switches  $S_1$ ,  $S_6$ , and  $S_2$  are on, voltages developed across resistors  $R_{\text{sense}}$  are measured to determine currents  $i_B$  and  $i_C$ . Using the information about currents  $i_B$  and  $i_C$ , it is easy to calculate current  $i_A$ . However, the current measured by this technique is the half bridge current not the motor phase current, as shown in Figure 2.9. Sample-and-hold circuit is required to reconstruct the motor phase current.

In order to get a good sample of voltage  $V_{\text{sense}}$  for current measurement, the width of lower leg pulse must be long enough. However, if lower transistors are turned on for a high portion of PWM period, it will create noise problems for the entire sense output sampling circuit due to high  $dv/dt$  and  $di/dt$ .

**2.6.2. Externally Added Sense Resistors in Series with Motor Phases.** In this method, the sense resistors are connected in series with motor phases, as depicted in Figure 2.10 [63, 64]. The voltage drop across each resistor is measured to determine the current flowing through each phase. Also, at any given instant of time in this method, only two phase currents need to be measured to get full information about motor line currents.

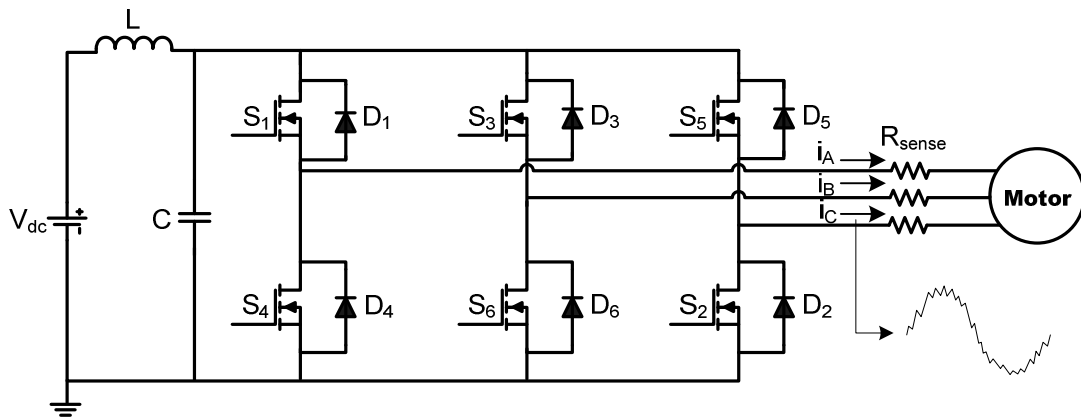


Figure 2.10. Motor current sensing using externally added line sense resistors

This method's major disadvantage is that the measured current has a common mode rejection problem because the small common mode signal with fast  $dv/dt$  rides on top of the measured current, as shown in Figure 2.10. A common method of solving the common mode rejection problem is to use an optically coupled isolated amplifier. However, if resistance is small when measuring higher currents, the inductive component of the impedance begins to dominate.

**2.6.3. An Externally Added Sense Resistor with DC Link.** Depending on the states of transistor switches and period of operation, the current through the motor phases can be measured or calculated by a single resistor, as shown in Figure 2.11. In the dc link sense resistor method, a single resistor is placed on the dc link between the dc power supply and the inverter to avoid level shifting during current measurement [65-78]. The dc link sense resistor minimizes the cost and power dissipation.

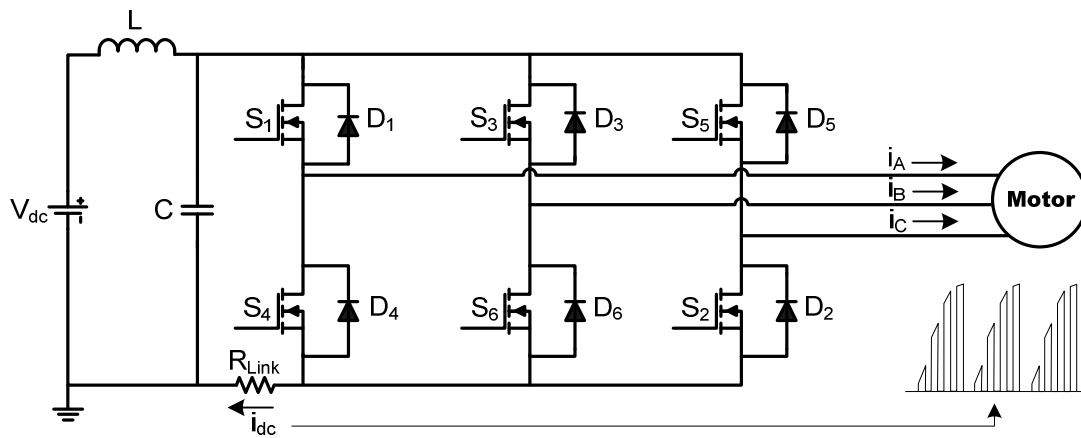


Figure 2.11. DC link current sensing using an externally added sense resistor

In ac drive systems, as mentioned earlier, if VSI works with  $180^\circ$  conduction, current  $I_{dc}$  is equal to one of the motor phase current or opposite to it for six different states, as depicted in Table 2.3.  $I_{dc}$  is zero when the upper three switches are on and the lower three switches are off or vice versa. If VSI works with  $120^\circ$  conduction, at any given instant the phase currents are expected to flow in only two phase windings (the third phase current is zero). Only one current measurement is necessary in the dc link and

other phase motor current can be reconstructed from dc link current information as depicted in Table 2.3.

It is also possible that the machine winding currents circulate within the VSI with some switching strategies, so there is no current available in the dc link for measurement. Furthermore, the measured current is the total bus current, not the phase current, as shown in Figure 2.11. The phase currents are reconstructed using measured dc link current and the knowledge of the switching states [78].

Table 2.3. DC link current for a PWM-VSI driven motor with 180° and 120° conduction

State	$i_{dc}$ for 180° Conduction	$i_{dc}$ for 120° Conduction
1	$-i_B$	$i_A = -i_B$
2	$i_A$	$i_A = -i_C$
3	$-i_C$	$i_B = -i_C$
4	$i_B$	$i_B = -i_A$
5	$-i_A$	$i_C = -i_A$
6	$i_C$	$i_C = -i_B$

No switch status information is needed in the inverter for current measurement if the sense resistor is connected to the positive rail of the VSI. However, the inverter circuit becomes more complicated to avoid reverse current flow into the sense resistor. The voltage information at two points of  $R_{sense}$  is required to find the voltage drop across  $R_{sense}$  in this method, whereas voltage measurement at only one point is required when  $R_{sense}$  is connected to the negative rail. It is also possible to connect the dc link sense resistor between the dc source and the capacitor, rather than between the capacitor and the inverter on the negative rail.

## 2.7. MOSFET $R_{DS}$ -BASED CURRENT SENSING FOR PWM-VSI DRIVEN AC MOTOR

A MOSFET switch has a linear resistance characteristic when it is on. Therefore, when the MOSFET switch is on in VSI, it is possible to get the motor current information from voltage drop observed across it. The circuit used to sense the single motor phase current by measuring the low side MOSFET switch ( $S_4$ ) voltage is depicted in Figure 2.12 [79]. The other two phases are exactly the same.

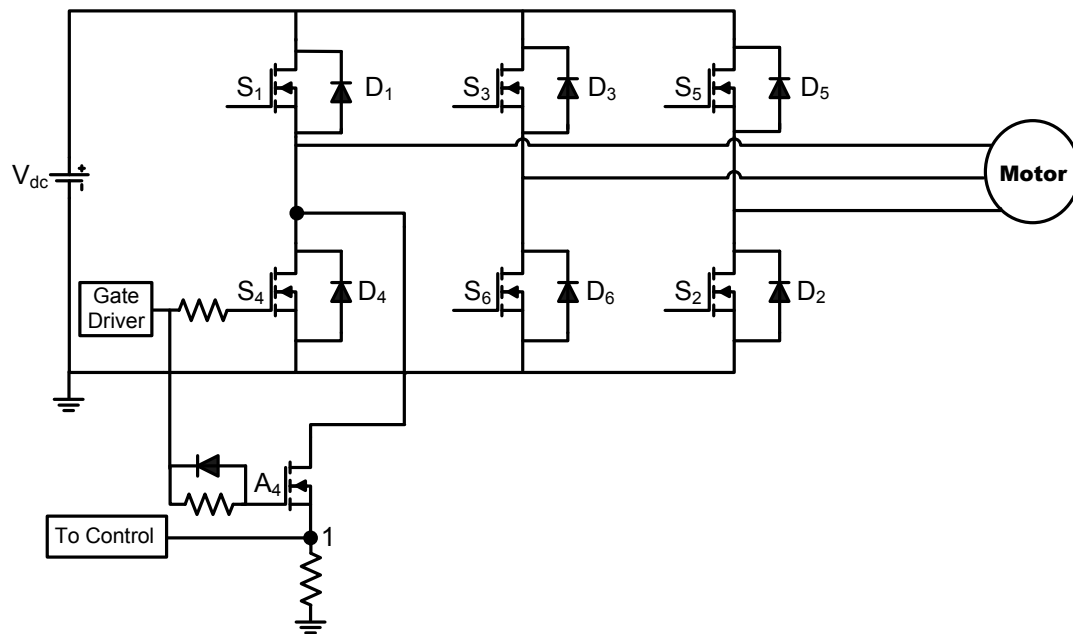


Figure 2.12. MOSFET  $R_{DS}$ -based current sensing technique for motor current detection

When low side MOSFET switch  $S_4$  is on, switch  $A_4$  also turns on, and the voltage at point 1 is the same as the voltage drop across switch  $S_4$ . When MOSFET switch  $S_4$  is off, switch  $A_4$  also turns off, and the voltage at point 1 is zero. If the voltage drop at point 1 and resistance  $R_{DS}$  of switch  $S_4$  are known, it is easy to derive the motor phase current

passing through switch  $S_4$ . The current information obtained from this technique is useful for current protection purposes; hence variation in  $R_{DS}$  is acceptable.

## 2.8. CURRENT-SENSING POWER MOSFET-BASED CURRENT SENSING TECHNIQUE FOR PWM-VSI DRIVEN AC MOTOR

A technique for phase current measurement in VSI using current-sensing power MOSFET is depicted in Figure 2.13 [80-88]. In this method each of the three lower leg MOSFET switches in VSI are replaced by current-sensing power MOSFETs. Current-sensing power MOSFET enables accurate and low cost current measurement. These current sensors are designed to accurately mirror the motor phase current.

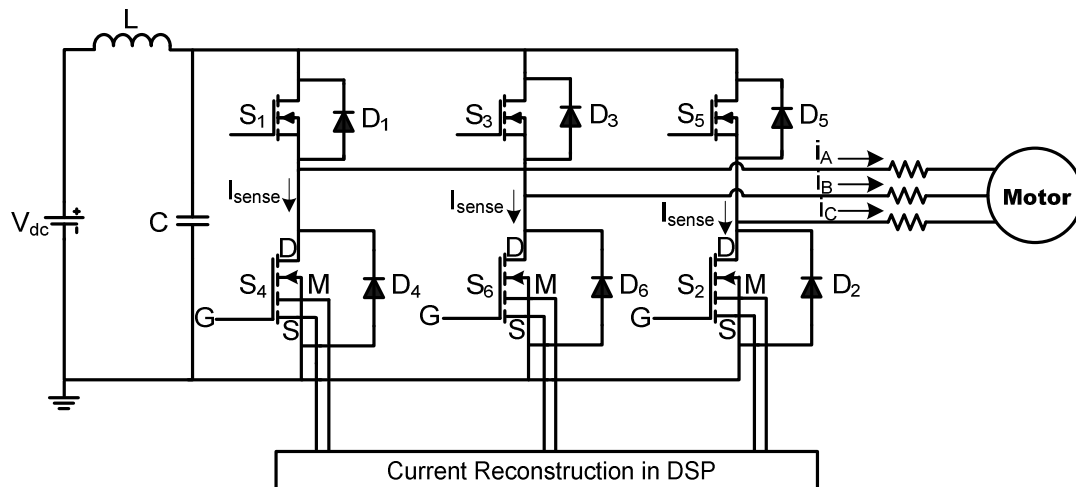


Figure 2.13. Current-sensing power MOSFET-based current sensing technique for motor current detection

A limitation of this technique is that its ability to measure the current only when the current flowing through it is in a forward direction, which is only possible when the polarity of the phase current is negative. Another limitation is that, at each particular

instant, two lower leg switches need to be on for complete current measurement. However, the current measurements taken using this technique are incomplete because the output phase current passes through the low side switch for only half of the fundamental cycle. Therefore, the alternating  $60^\circ$  intervals (three times during each fundamental frequency cycle) provide incomplete current information. A technique is depicted in [82] to reconstruct the three phase currents during the intervals when they are not measured.

It is also possible to connect the current sensors to the upper leg of VSI, but it requires high performance level shifting to transfer the sensed signal to a common reference. On the contrary, if the current sensors are in the lower leg of VSI, they can share the same reference terminal, simplifying the signal conditioning.

## **2.9. COMPARISON BETWEEN RESISTIVE-BASED CURRENT SENSING TECHNIQUES**

Table 2.4 summarizes the advantages and disadvantages of the presented resistive-based current sensing techniques.

Table 2.4. Comparative overview of resistive-based current sensing techniques

<b>Techniques</b>	<b>Advantages</b>	<b>Disadvantages</b>
Using an externally added resistor	Simple, accurate, low cost	Power loss incurred by sense resistor
Using the internal resistance of an inductor	Accurate, lossless	Not useful for high power applications
MOS current sensing	Lossless, accurate	Complicated circuit, accuracy depends on matching performance of the current mirror
Using the internal resistance of a MOSFET	Lossless, no additional sensing component required, low cost	Not accurate for high input voltage and high switching frequency, affected by temperature variations of $R_{DS}$ , discontinuous and noisy
Using current-sensing power MOSFET	Lossless, practical, accurate with respect to temperature variations, no additional sensing component required, low cost	Special MOSFET, introduce switching transients and noise at high frequency, accuracy depends on matching performance of the current mirror, switching noise, limited applications, discontinuous and noisy



### **3. ELECTROMAGNETIC-BASED CURRENT SENSING TECHNIQUES**

Resistive-based current sensors are acceptable where power loss, low bandwidth, noise, and non isolated measurement are acceptable. These sensors are not used in high power applications where isolation is required. Solution to these problems is electromagnetic-based current sensing techniques.

Based on Ampere's law, the current flowing through a conductor generates magnetic field. Hence, sensing the magnetic field surrounding a conductor provides information about its current. Electromagnetic-based techniques work based on this phenomenon while they provide galvanic isolation between the control and power stages, higher bandwidth, and lower power losses. The lower power dissipation of electromagnetic-based current sensors allow much higher signal level, significantly improves the signal-to-noise environment of the control system [39].

#### **3.1. CURRENT MEASUREMENT TECHNIQUES USING CURRENT TRANSFORMERS**

A current transformer (CT) is similar to a voltage transformer; except that the primary input is a current. There are four basic types of CTs [89-92]: ac current transformers (ACCTs), unidirectional current transformers (UCTs), dc current transformers (DCCTs), and fly-back type current transformers (FBCTs). ACCTs and UCTs are commonly used, DCCTs are used for high current applications, and FBCTs are used when current pulses are very short.

CTs are used with low range ammeters to measure currents in high voltage circuits. In addition to providing insulation from the high voltage side, CTs step down the

current in a known ratio. Their physical basis is the mutual induction between two circuits linked by a common magnetic flux. A CT consists of two inductive coils, which are electrically separated but magnetically linked through a path of low reluctance, as shown in Figure 3.1. If one coil is connected to an ac source, an alternating flux is set up in the core, most of which is linked with the other coil in which it produces mutually induced electromotive force (EMF) according to Faraday's law of electromagnetic induction. The first coil is called the primary coil, and second coil is called the secondary coil of the CT. If the secondary of the CT is closed, electric energy is magnetically transferred from primary to secondary.

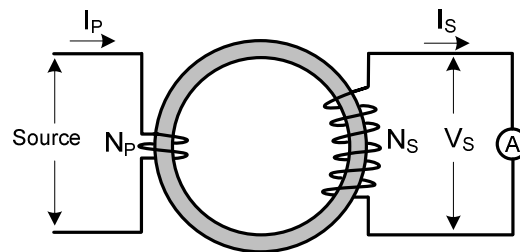


Figure 3.1. Current transformer

For an ideal transformer with no load, the induced secondary EMF is same as the secondary terminal voltage ( $V_s$ ). The relationship between the primary and secondary voltages, currents, and number of turns is given by (3.1).

$$\frac{V_P}{V_S} = \frac{I_S}{I_P} = \frac{N_P}{N_S} \quad (3.1)$$

Where  $V_P$  and  $V_S$  are primary and secondary terminal voltages,  $I_P$  and  $I_S$  are primary and secondary winding currents, and  $N_P$  and  $N_S$  are the number of primary and secondary turns, respectively.

For typical switching converter applications, a CT has single turn primary and multi-turn secondary. The basic CT schematic is given in Figure 3.2.

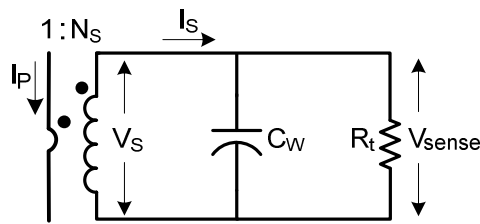


Figure 3.2. Basic circuit of single turn primary current transformer

The primary is formed by wire from which an unknown current is passing. The secondary has large number of turns and it is terminated by terminating or burden resistor ( $R_t$ ). If the number of secondary turns is too large, then there will be a significant inter-winding capacitance ( $C_W$ ). In addition to primary and secondary windings, capacitance  $C_W$  is also added in the model of a CT.

The relationship between currents  $I_P$  and  $I_S$  in this case is given by (3.2).

$$\frac{I_S}{I_P} = \frac{1}{N_S} \quad (3.2)$$

Capacitor  $C_w$  is not considered here since the measured frequency is low. The secondary output voltage ( $V_{sense}$ ) in a CT is proportional to the resistor  $R_t$  based on the current flowing through it. Voltage  $V_{sense}$  across the secondary is given by (3.3).

$$V_{sense} = I_S R_t = \frac{I_P}{N_S} R_t \quad (3.3)$$

One of the most commonly used CTs is known as clamp-on or clip-on type. It has a laminated core which is arranged in such a manner that it can be opened by pressing a switch, permitting the admission of the current carrying conductor. The current carrying conductor acts as a single turn primary, whereas secondary is connected across the standard ammeter.

CTs are used in control and limiting applications. The advantages of a CT include good signal-to-noise ratio (SNR), galvanic isolation between control circuit and the line being monitored, good common mode rejection, high bandwidth (50 or 60 Hz to 20 kHz), and low power loss in high-current applications. Galvanic isolation is required to protect the sensing device from high power signals and to reduce the power dissipated by the sensing resistor. The secondary side current can be made smaller by using a larger turns ratio. CTs are relatively simple to implement and are passive devices that do not require driving circuitry for operation. The major disadvantages of CTs are their higher cost, larger size, and non-integrality. An ACCT gets saturated if input is direct current. Yet, some methods are used to sense the switch dc current in switching power converter using CTs. It is due to the fact that switch currents in the converter are interrupting, allowing time for transformer reset.

**3.1.1. Using the Existing Inductor as the Primary of a Transformer.** This technique measures the current flowing through an inductor by using existing inductor of a converter as the primary of a transformer. Primary current  $I_L$  will generate magnetic field that is coupled into a secondary coil. As shown in Figure 3.3, in a buck converter, the inductor is used as the primary side of a transformer, and the secondary side develops voltage proportional to current  $I_L$  [42].

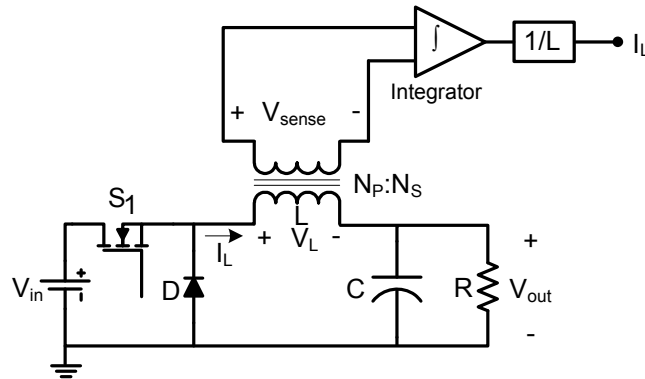


Figure 3.3. Current measurement using the existing inductor as the primary of a transformer

Inductor current in this technique is calculated by (3.4). For integration it is necessary to know the value of inductance  $L$ .

$$i_L = \frac{1}{L} \int v_{sense} dt \quad (3.4)$$

This technique measures the fraction of the high inductor current. The transformer only passes the ac part of current  $I_L$ , not the dc, and no information about average current is provided by technique. This technique is not appropriate for over current protection.

The presence of the dc current affects the low frequency performance of ACCT. In order to recover the dc component of the current signal during current measurement, the following techniques are used.

**3.1.2. AC Current Transformer with Sample-and-Hold Circuit.** The magnetic core of the ACCT needs to be optimized such that the ACCT is able to function with a primary current having a dc current component. One method of gathering full information about current  $I_L$  in the buck converter is depicted in Figure 3.4 [93]. Two ACCTs,  $CT_1$  and  $CT_2$ , are used to measure the two switch currents. Additional circuits are required to reconstruct the two switch current signals. Reconstructed current outputs of  $CT_1$  and  $CT_2$  are then added to obtain full information about current  $I_L$ . Two blocks with gain  $N_p/(N_s R_t)$  are required to adjust the different scaling used in the two CTs.

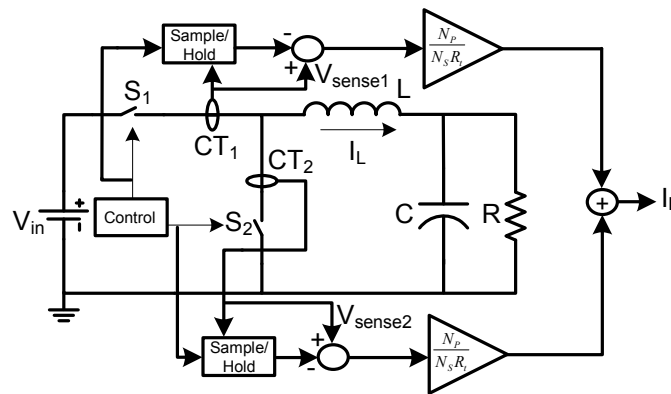


Figure 3.4. Inductor current estimation using two ac current transformers

Usually,  $CT_1$  and  $CT_2$  will filter off the dc component of the primary current, so they allow only the ac part on the secondary side. In order to obtain full information of current  $I_L$ , additional sample-and-hold circuits are required for switch current

reconstruction, as depicted in Figure 3.4. Current sensor waveforms for  $CT_1$  are shown in Figure 3.5.

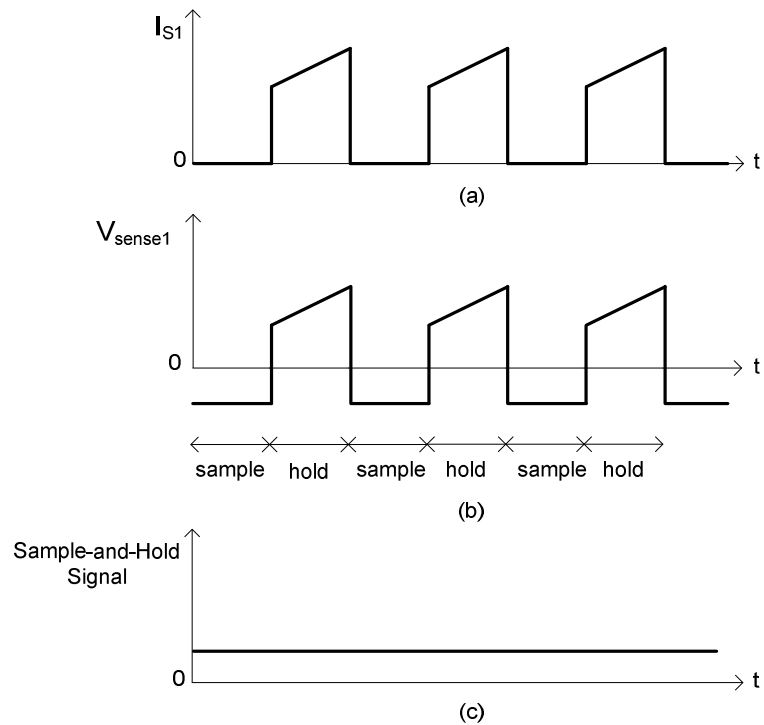


Figure 3.5. Current, voltage, and sample-and-hold circuit waveforms for  $CT_1$

Figure 3.5 (a) shows the ideal switch current ( $I_{S1}$ ) waveform for switch  $S_1$ . The dc component of current  $I_{S1}$  is filtered off in the secondary voltage ( $V_{sense1}$ ) of  $CT_1$  as shown in Figure 3.5 (b). However, no information is lost in this process. In order to obtain complete inductor current information in each cycle, sample the value of voltage  $V_{sense1}$  when switch  $S_1$  is turned off, hold that value when switch  $S_1$  is turned on, as shown in Figure 3.5 (c). The output of the sample-and-hold circuit is deducted from the voltage  $V_{sense1}$  to get a signal which is proportional to current  $I_{S1}$ . Current through switch  $S_2$

( $I_{\text{sense2}}$ ) is estimated in the same manner as with switch  $S_1$ . Two switch current signals are added to estimate  $I_L$ . High bandwidth sample-and-hold IC is required in this method.

There are some design requirements for the ACCT core in this method. For CT, if  $H$  is the magnetic field intensity and  $l$  is the magnetic mean path, then according to Ampere's law, the magneto-motive force (MMF) is given by (3.5).

$$N_P I_P + N_S I_S = Hl \quad (3.5)$$

If  $\phi$  is the total flux,  $B$  is flux density,  $A$  is the cross sectional area of the core,  $\mu_o$  is the vacuum permeability,  $\mu_r$  is the relative permeability, and  $R_e$  is the reluctance of the magnetic circuit, (3.5) can be rewritten as (3.6).

$$N_P I_P + N_S I_S = \frac{B}{\mu_o \mu_r} l = \frac{\phi}{\mu_o \mu_r A} l = R_e \phi \quad (3.6)$$

According to Faraday's law of electromagnetic induction, the secondary voltage is given by (3.7).

$$V_{\text{sense}} = I_S R_t = N_S \frac{d\phi}{dt} \quad (3.7)$$

From (3.6) and (3.7), ratio of currents  $I_S$  and  $I_P$  is given by (3.8).

$$\frac{I_S}{I_P} = \frac{N_P}{N_S} \left( 1 - \frac{1}{1 + ks} \right) \quad (3.8)$$



Where  $k$  is given by  $(N_s^2)/(R_t R_e)$ . Corner frequency ( $f_c$ ) of low pass filter is specified by (3.9).

$$f_c = \frac{2\pi}{k} = \frac{2\pi R_t R_e}{N_s^2} \quad (3.9)$$

The first design requirement is that the  $f_c$  needs to be much lower than the switching frequency ( $f_s$ ). The second design requirement is that the CT does not get saturated in its operation. More information about this requirement is discussed in detail in [3].

This technique has low power loss, fast response, and is accurate. It is also possible to obtain  $I_L$  by only measuring the high side switch  $S_1$  current. A suitable analog circuit is required to connect the maximum and minimum points of the switch  $S_1$  current to obtain full information about current  $I_L$ . The circuit required for that method is quite complicated.

**3.1.3. Current Measurement using AC and Unidirectional Current Transformers.** The magnetic flux in a typical ac transformer core alternates between positive and negative values, but not in the UCT. The UCT operates in a unipolar mode. For UCT, the core should have two properties; high permeability and low remanence.

The basic UCT circuit is shown in Figure 3.6 [94-97]. When a positive current pulse  $I_P$  flows in the primary of UCT, diode  $D$  conducts. Output voltage  $V_{\text{sense}}$  developed across resistor  $R_t$  is given by (3.3), and current  $I_S$  in this is given by (3.10).

$$I_S = \frac{I_P}{N_S} - I_M \quad (3.10)$$

Capacitor  $C_W$  is not considered here since frequency is low. Diode  $D$  will block the zero primary current and capacitor  $C$  will resonate with magnetizing inductor  $L_M$  to develop a negative half cycle sinusoidal pulse across secondary winding to reset the magnetizing current ( $I_M$ ). Current  $I_M$  has a small value due to this cycle by cycle reset. This provides a fast reset for the transformer core, and isolates the negative reset voltage from the control circuit. Capacitor  $C_f$  is a filtering capacitor required for suppressing high frequency noise.

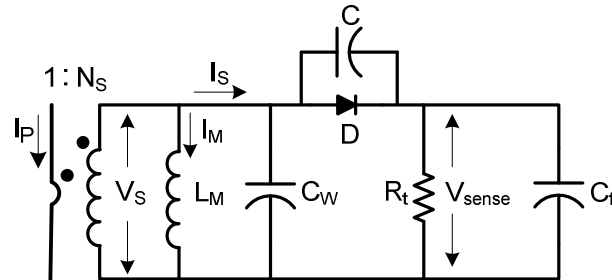


Figure 3.6. Basic circuit of a unidirectional current transformer

Waveforms for currents  $I_p$  and  $I_M$  and voltages  $V_S$  and  $V_{sense}$  of a UCT are given in Figure 3.7. During the pulse duration of current  $I_p$ , diode  $D$  conducts. The current  $I_s$  in  $R_t$  will be a transform of current  $I_p$ , and an analogue voltage of primary current will be developed across  $R_t$ . At the end of the conduction pulse, rapid UCT core recovery occurs because  $D$  blocks the reverse voltage across  $R_t$ . As a result, voltage  $V_S$  is large, enabling fast core restoration between pulses. For accurate current measurement by UCTs, inductance needs to be large to reduce current  $I_M$ . Current  $I_M$  increases during the pulse

duration. At the end of the conduction period, current  $I_M$  needs to be smaller than current  $I_P$ .

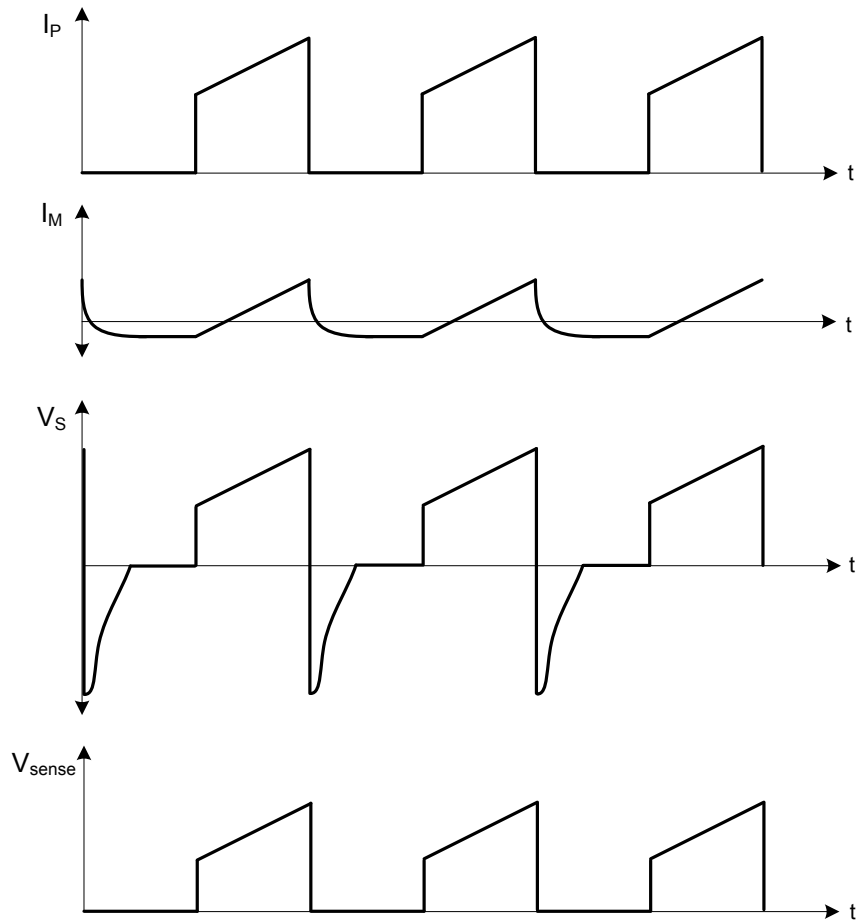


Figure 3.7. Unidirectional current transformer waveforms

A buck converter with UCTs and ACCT for current measurement is illustrated in Figure 3.8. The inductor current  $I_L$  and the output current  $I_O$  in any PWM converter are the combinations of the switch current  $I_S$ , the diode current  $I_D$ , and the capacitor current

$I_C$ . These currents are ac or zero return pulse, which can be measured by using ACCTs and UCTs.

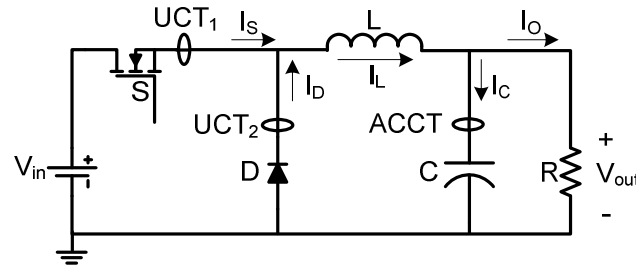


Figure 3.8. Current sensing using unidirectional and ac current transformers

In a buck converter  $I_L$  is given by (3.11).

$$I_L = I_S + I_D \quad (3.11)$$

UCT<sub>1</sub> and UCT<sub>2</sub> are used to measure  $I_S$  and  $I_D$ , as depicted in Figure 3.8. Current  $I_L$  is then derived from the information of the measured currents. Current  $I_C$  can be measured by using ACCT. Current  $I_O$  is derived from the information of currents  $I_L$  and  $I_C$ . In a buck converter, current  $I_O$  is given by (3.12).

$$I_O = I_L + I_C \quad (3.12)$$

This method is also useful for converters with isolation transformers. However, UCT has some limitations. There is not enough time for UCT to complete half cycle resonant reset at very high switching frequencies or very high or low duty cycles. The dc

magnetizing current in the transformer core will increase with a decrease in reset time. As a result, the dc level of the UCT output will drop.

**3.1.4. DC Current Transformer.** DCCT is used to accurately measure the dc current with low power loss. There are several different types of DCCT, one of which is shown in Figure 3.9 [98-100]. The secondary signal of DCCT is directly related to absolute value of primary current.

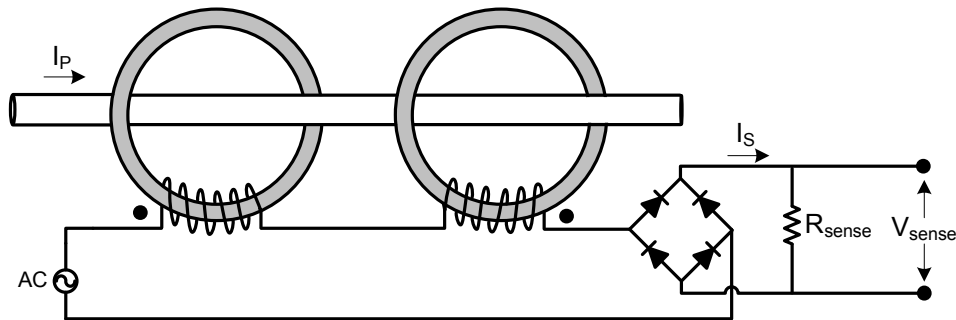


Figure 3.9. Current measurement using dc current transformer

DCCT has two ring cores, as depicted in Figure 3.9. AC supply is applied across their secondary to generate opposite polarity flux in each core, which will drive cores into and out of saturation. As each core comes out of saturation, the flux change induces a current  $I_s$  that is proportional to the current  $I_p$ . The rectifier is used to rectify the secondary signal. More information about this technique is given in [100].

### 3.2. CURRENT MEASUREMENT TECHNIQUES USING AIR CORE

The performance of a CT is often limited by the characteristics of its magnetic core material (hysteresis, non-linearity, losses, saturation, remanence (residual flux));

therefore, the design of an air core or coreless transformer is often considered. The challenge with air core current measurement techniques is to have enough measurement sensitivity and to be insensitive to external magnetic fields.

**3.2.1. Rogowski Coil Current Sensor.** The Rogowski Coil is a simple, inexpensive and accurate approach for current measurement. Structure of a Rogowski Coil is similar to a CT. However, instead of an iron core, Rogowski Coil is based on air or ironless bobbins with hundreds or thousand of turns, as shown in Figure 3.10. The Rogowski Coil has an air core, so it will never get saturated; therefore, its output of remains linear for high current measurement [101-106].

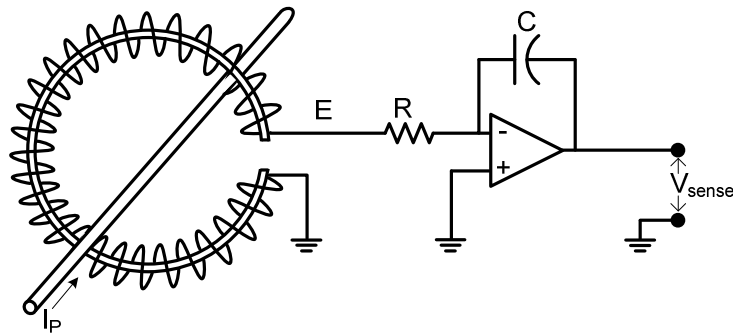


Figure 3.10. Rogowski Coil current sensor

The conductor from which the unknown current flows is surrounded by the Rogowski coil for current measurement, as shown in Figure 3.10. In order to place current carrying conductor inside, the Rogowski Coil can be opened without interrupting the circuit. The magnetic field produced by the current induces the voltage in the

secondary coil (E). Voltage E is proportional to the time derivative of current flowing through the conductor, which is given by (3.13).

$$E = M * \frac{dI_P}{dt} \quad (3.13)$$

Here,  $I_P$  is the unknown primary current and M is the mutual inductance of the circuit. M depends on the geometric parameters of the coil and is given by (3.14).

$$M = \frac{\mu_o AN_S}{l} \quad (3.14)$$

Where  $\mu_o$  is the permeability of free space, A is the cross sectional area of the coil,  $N_S$  is the total number of secondary winding turns, and  $l$  is the mean path length of the coil. Because the derivative of the direct current is zero, the Rogowski Coil current sensor can not measure dc currents. It is used to measure ac or pulsed dc current only.

The phase delayed secondary voltage is integrated to produce an output voltage ( $V_{sense}$ ) which is proportional to current  $I_P$ . If the Rogowski Coil is used to measure the current in a semiconductor switch, a simple resistor-capacitor integrator, as shown in Figure 3.10, can be used to reproduce current waveform as a voltage. Rogowski Coil terminals are connected in a special way to avoid the external field effects. The output voltage of the integrator circuit is given by (3.15), which is proportional to unknown current  $I_P$ .

$$V_{sense} = \frac{1}{RC} \int E * dt = \frac{1}{RC} \frac{\mu_o AN_S}{l} I_P \quad (3.15)$$

This technique is also used when approximate current is not known before application. It measures high energy current pulses with high frequency harmonic content because upper bandwidth can extend into the mega-hertz range. A higher bandwidth (100 kHz) allows the Rogowski Coil to measure switching transients in semiconductors. Other advantages include a wide dynamic range (from milli-amps to thousands of amps), easy calibration, linear output, and accurate response to transient currents, which makes it ideal for protecting systems and measuring current pulses. The Rogowski Coil sensor also has some disadvantages. Absolute value of the waveform cannot be reproduced using a Rogowski Coil. Also, voltage  $E$  is very small for hundreds of ampere current (rather than thousands) at typical power frequency, so it requires a high gain integrator. It is also difficult to satisfactorily connect a Rogowski Coil to the integrator to avoid ringing and reflections.

The Rogowski Coil works with a wound coil. The mechanically open structure of Rogowski Coil can create a slight gap in coil structure, which leads to errors (below 1% with a maximum of 2%) based on the position of the current carrying conductor in the aperture. This method's accuracy is affected by the external field due to the manufacturing tolerance of the wound coil. Planar Rogowski Coil current sensor is the solution to these problems.

**3.2.2. Planar Rogowski Coil Current Sensor.** A planar Rogowski Coil current sensor, used to measure ac or pulsed dc current, is an air core approach based on planar magnetic sensors, as shown in Figure 3.11[107-110]. It operates on the basic Rogowski coil principle in which a coil is magnetically coupled with the flux created by the current to be measured. In place of a continuous uniform coil used in the Rogowski Coil current



sensor, planar Rogowski Coil current sensor uses discrete coils placed at regular intervals on a fixed perimeter.

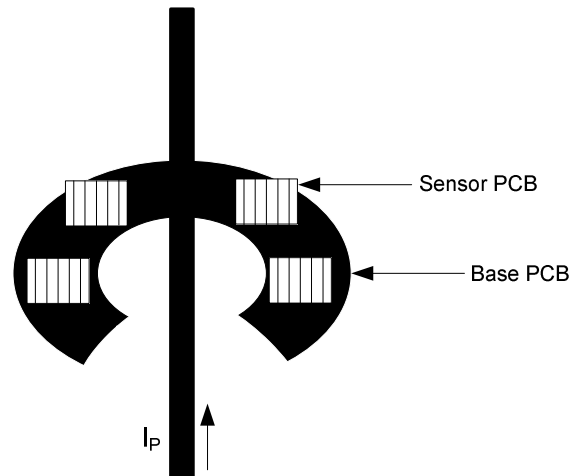


Figure 3.11. Planar Rogowski Coil current sensor

As shown in Figure 3.11, the planar Rogowski Coil current sensor consists of two parts; sensor PCB and base PCB. Each sensor PCB is made of two distinct air cored coils constructed on a multi layer PCB. Sensor PCBs, which are placed at right angle to the base PCB, are connected in a series to form inner and outer concentric loops, as shown in Figure 3.12. The great number of sensor PCBs gives lower sensitivity to conductor position, higher accuracy, and robustness to external field perturbations.

The current carrying conductor is placed inside the two loops without any electrical or mechanical interruption. Similar to a Rogowski Coil current sensor, the continuously changing magnetic flux due to current carrying conductor creates a voltage drop, which needs to be integrated. The integrator is kept very near to the current sensor

to reduce stray inductive pick up that can give rise to errors. The sense voltage  $V_{sense}$  in Figure 3.12 is given by (3.16).

$$V_{sense} = V_{inner} - \frac{V_{outer}}{\lambda} = V_{inner} \left( 1 - \frac{V_{outer}}{V_{inner} * \lambda} \right) \quad (3.16)$$

Where  $V_{inner}$  and  $V_{outer}$  are voltages induced in the inner and outer loop, respectively.

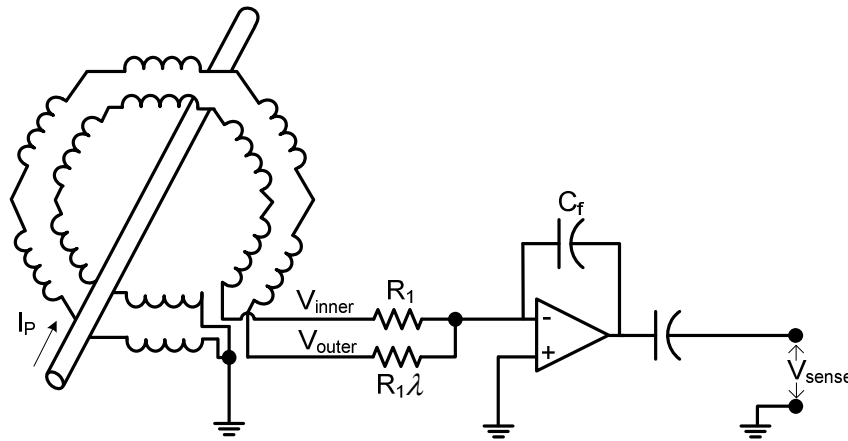


Figure 3.12. Planar Rogowski Coil current sensor with two concentric loops

In order to cancel the external field effect, this sensor is designed in such a way that ratio  $V_{outer}/V_{inner}$  is almost constant, irrespective of the position and magnitude of the external current source. Making  $\lambda = (V_{outer}/V_{inner})$  force the voltage  $V_{sense}$  in (3.16) zero, in the other words, an external current source will not produce any signal at output. To measure current of current carrying conductor which is placed inside the loops, the sensor is designed such that  $(\lambda/2) = (V_{outer}/V_{inner})$  irrespective of the position of current carrying

conductor. As a result, voltage  $V_{\text{sense}}$  in (3.16) is  $(V_{\text{inner}}/2)$ ; therefore, only half of the signal collected by inner loop is used for measurement.

This technique's advantages include its large bandwidth (10 Hz to 100 kHz), light weight, no magnetic saturation, low power consumption, low thermal losses, wide current measuring range, and high accuracy.

### 3.3. CURRENT MEASUREMENT TECHNIQUES USING HALL-EFFECT

Hall-effect sensors are used to measure ac, dc, and complex currents with electrical isolation. They are used to measure the current without interrupting the circuit. A Hall-effect sensor is small, provides noise-immune signal, and consumes little power [111-115]. A Hall-effect sensor works based on the Lorentz force, which acts on charges moving through a magnetic field. Figure 3.13 represents the Hall-effect principle. The Hall-effect principle states that when a magnetic field is applied to a conducting or semi-conducting material from which current is flowing, a voltage will be developed across the sides of the material.

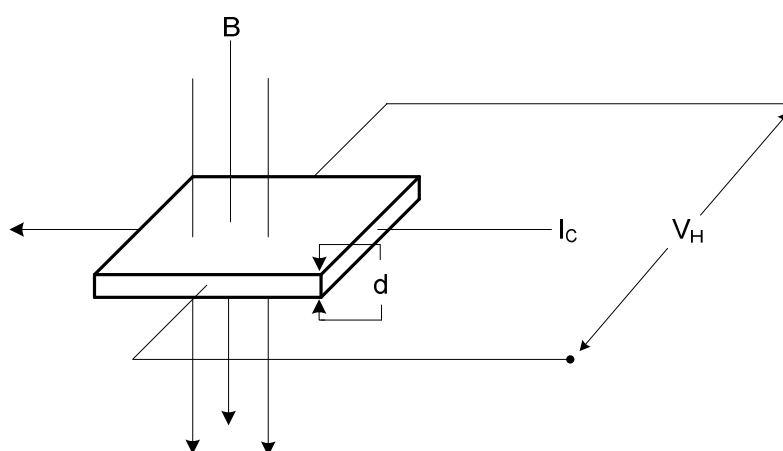


Figure 3.13. Hall-effect principle

If  $I_C$  is the control current passing through the Hall sensor,  $B$  is magnetic flux density created by unknown current carrying conductor,  $K$  is a constant of conducting material,  $d$  is the thickness of the sheet, and  $V_{OH}$  is the offset of the Hall sensor in the absence of the external field, the output Hall voltage  $V_H$  of Hall-effect sensor is given by (3.17).

$$V_H = \frac{K}{d} B I_C + V_{OH} \quad (3.17)$$

Here, product  $(K/d) \cdot I_C$  represents the sensitivity of the Hall-effect sensor. However, the sensitivity and voltage  $V_{OH}$  of a Hall-effect sensor are temperature dependent. As per (3.17), if constant current  $I_C$  is supplied to the Hall-effect sensor, then a voltage is produced that is proportional to flux density  $B$  in the core, which is proportional to unknown primary current  $I_p$ . Hall-effect sensors are also known as Hall generators.

Flux density  $B$  for a Hall-effect sensor is inversely proportional to the distance from the center of the conductor to the point of sensing, so usable flux density cannot be achieved at much greater distance from the conductor's center. Other disadvantages of Hall-effect sensors are their low sensitivity, which requires a concentrator, tricky mechanical positioning, limited linearity range, sensitivity to mechanical stresses and ambient temperature variations, limited maximum frequency range due to junction capacitance, and requirement of an isolated power supply. Different Hall-effect-based current measurement techniques have been proposed to overcome the abovementioned disadvantages. These techniques include open-loop Hall-effect sensing, closed-loop Hall-effect sensing, and combinations of open-loop or closed-loop Hall-effect sensing with a current transformer technique.

**3.3.1. Open-loop Hall-Effect Current Sensor.** An open-loop Hall-effect current sensor uses a high permeability magnetic core (as a field concentrator) with an air gap located around the conductor which carries current  $I_p$ , as shown in Figure 3.14 [116-118].

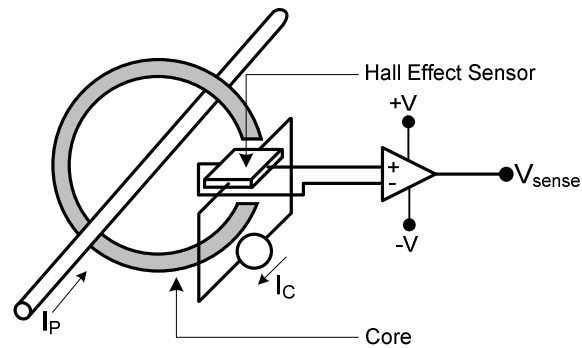


Figure 3.14. Open-loop Hall-effect current sensor with single turn primary

A linear Hall sensor is inserted into the air gap and provides voltage  $V_H$  proportional to flux density  $B$  produced by current  $I_p$ . Voltage  $V_H$  is amplified and output voltage is then read as a voltage which represents current  $I_p$  through a scaling factor. B-H loop for open-loop Hall-effect sensor is depicted in Figure 3.15. Voltage  $V_H$  is proportional to flux density  $B$ , and flux density  $B$  is proportional to current  $I_p$  within the linear region of hysteresis loop of the material used for the magnetic circuit, as shown in Figure 3.15. Voltage  $V_H$  needs to be amplified and compensated to remove the offset component.

Open-loop Hall-effect current sensors are able to measure ac, dc, and complex currents accurately. The benefits of an open-loop Hall-effect current sensor include

simple construction, low cost, low power consumption, low insertion losses, and small size for higher currents.

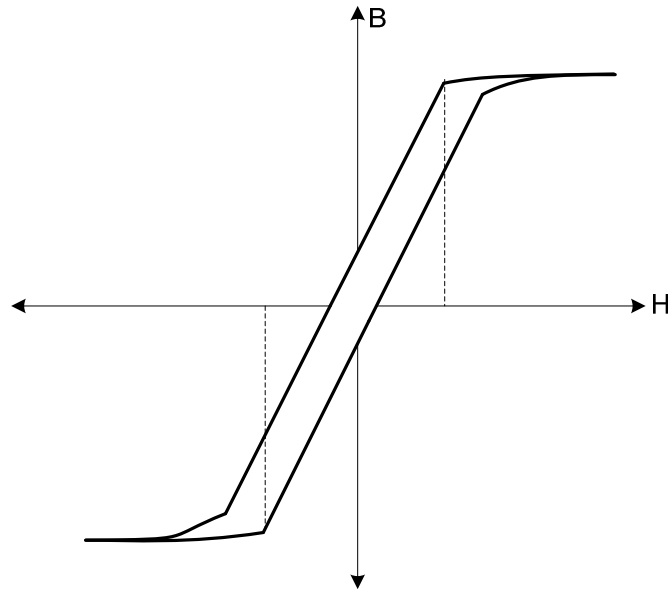


Figure 3.15. B-H loop for open-loop Hall-effect current sensor

However, disadvantages include magnetic core heating due to core losses at high frequency current measurement, narrow bandwidth (dc to 25 kHz), high offset and gain drift, limited range of linearity, and lower accuracy.

In order to get higher accuracy for lower current measurements ( $<50\text{A}$ ), the multiple turns of the conductor from which unknown current is flowing is wrapped around the field concentrating core as shown in Figure 3.16 [119]. The problem with this sensor is that the coil wrapped around the gaped core creates an inductor. Furthermore, core losses are increased by the square of the primary ampere turns in this case.

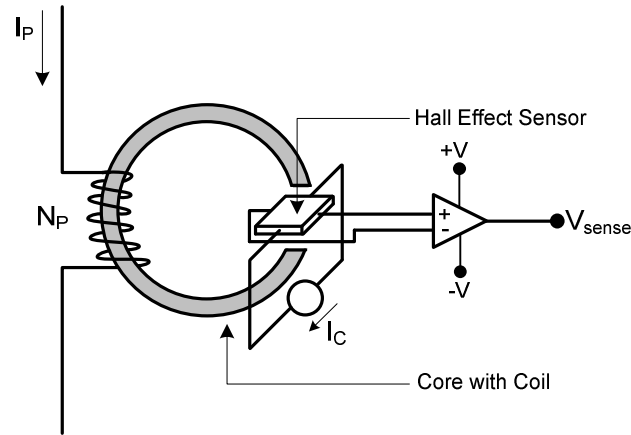


Figure 3.16. Open-loop Hall-effect current sensor with a multi-turn primary

**3.3.2. Closed-loop Hall-Effect Current Sensor.** Closed-loop Hall-effect current sensor improves the performance of open-loop Hall-effect current sensor by using a compensation circuit, as shown in Figure 3.17. In closed-loop Hall-effect sensor, a low current secondary winding is wrapped around the high permeability core to develop magnetic flux in opposition to the flux developed by current  $I_p$  [116]. The hall sensor is enclosed in an overall feedback loop, as depicted in Figure 3.17 [120-122]. A Hall sensor in the air gap produces a voltage  $V_H$  proportional to flux density in the core. Voltage  $V_H$  is then amplified by the operational amplifier and fed into a push-pull amplifier. Compensation current  $I_S$  is fed by the push-pull amplifier into the secondary coil to null the flux in the core.

The Hall sensor in the air gap is also used to detect zero flux. Therefore, the closed-loop Hall-effect sensors are also known as compensated or zero flux Hall-effect current sensors. Current  $I_S$  creates flux equal in amplitude, but opposite in direction, to the flux created by current  $I_p$ . Operating the core near zero flux eliminates dependence on the linearity of the core and Hall sensor and also reduces hysteresis errors.

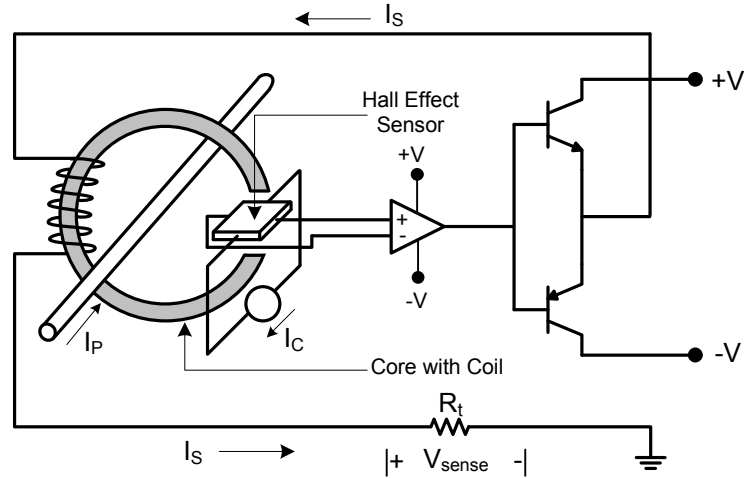


Figure 3.17. Closed-loop Hall-effect current sensor

When the magnetic flux is fully compensated, the ampere turns of two windings are identical, which is given by (3.18).

$$N_P I_P = N_S I_S \quad (3.18)$$

Therefore, current  $I_S$ , according to (3.18), is exact representation of unknown current  $I_P$ . The frequency response of the closed-loop Hall-effect current sensor is enhanced due to the zero flux. Current  $I_S$  can be converted to a voltage by placing a resistor  $R_t$  from the output of the secondary coil to ground, as shown in Figure 3.17. By selecting the proper value of resistor  $R_t$ , the voltage can be scaled for any application. A capacitor can be added in parallel to resistor  $R_t$  to filter high frequency components in the feedback signal.

Closed-loop Hall-effect current sensors are able to accurately measure ac, dc, and complex currents. Closed-loop Hall-effect current sensors provide many advantages, such



as high bandwidth (dc to 200 kHz), high accuracy and linearity, fast response time, low insertion losses, and low gain drift. However, their disadvantages include higher cost, high current consumption, larger dimensions, and limited output current due to the fact that closed-loop sensors can only drive a finite amount of secondary current. It is very difficult to design closed-loop Hall-effect sensors that will operate from low secondary voltage power supplies (less than or equal to 5 V) with primary currents more than 25 A. In closed-loop Hall-effect sensor, there is very limited voltage to drive a proper secondary winding and terminating resistance. Solution to this problem is combination of an open-loop Hall-effect current sensor with a current transformer.

**3.3.3. Combination of an Open-loop Hall-Effect Current Sensor with a Current Transformer.** This sensor's construction is similar to closed-loop Hall-effect current sensor with same magnetic circuit, Hall sensor, and secondary winding, as depicted in Figure 3.18 [116].

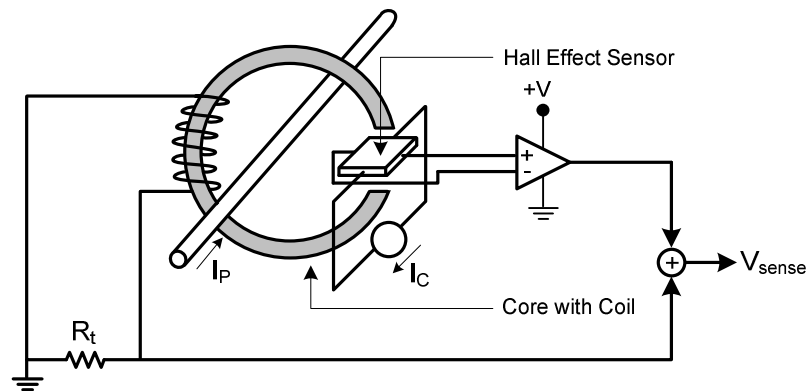


Figure 3.18. Combination of an open-loop Hall-effect current sensor and a current transformer

This technique combines the desirable properties of an open-loop Hall-effect current sensor and a CT [123-125]. It works as an open-loop Hall-effect sensor at low frequencies (2-10 kHz) with Hall generator providing signal proportional to current  $I_p$ . At high frequencies (10-100 kHz), this sensor works as a CT, where the output current is proportional to the ac current  $I_p$ . In this method, for flux measurement, a high performance application specific integrated circuit with temperature compensated Hall sensor is added in the gap of high permeability magnetic core, as depicted in Figure 3.18. It provides enhanced accuracy when this sensor is working as an open-loop Hall-effect sensor to measure dc and low frequency currents. Hall sensor and transformer signals are electronically added to form a common output signal. This sensor requires very low secondary power consumption because the power supply is not required to drive the secondary coil with compensation current  $I_s$

This technique provides broad bandwidth, as shown in Figure 3.19, because the open-loop Hall-effect sensor adds the dc response to the high frequency capability of the current transformer.

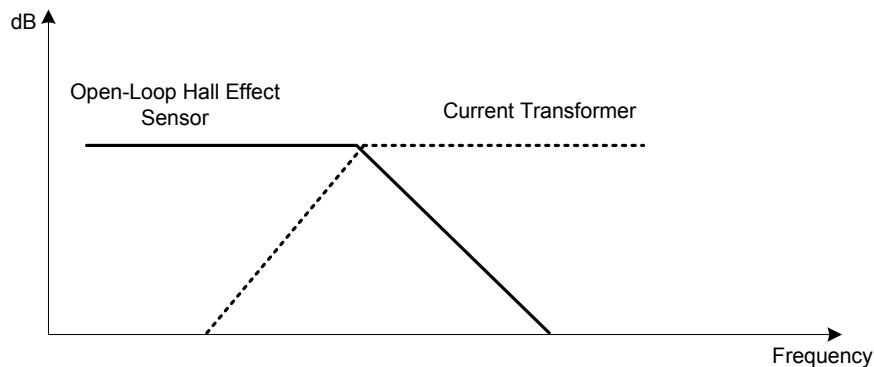


Figure 3.19. Broad bandwidth provided by combination of an open-loop Hall-effect current sensor and a current transformer

This technique is able to measure dc, ac, and complex currents with galvanic isolation and low insertion loss. Its construction is the same as that of a closed-loop Hall-effect current sensor; therefore, it does not provide any prize advantage. Other advantages include low power consumption, suitability for small secondary supply voltages, high bandwidth (dc to 100 kHz), and fast response time. Disadvantages include offset and gain drift, large core size at low frequency, need for a large secondary coil for high frequency detection, and moderate accuracy. It is also possible to combine a closed-loop Hall-effect current sensor with a CT. More information about this technique is given in [119].

### **3.4. SATURABLE INDUCTOR CURRENT SENSORS**

Saturable inductor current sensors work on the same measurement principle as Hall-effect-based current sensors: the magnetic field created by the primary current to be measured is detected by a specific sensing element [116, 126]. The design of the saturable inductor current sensor is similar to that of a closed-loop Hall-effect current sensor; the only difference is that this method uses the saturable inductor instead of the Hall-effect sensor in the air gap, as shown in Figure 3.20.

Saturable inductor current sensor is based on the detection of an inductance change. The saturable inductor is made of small and thin magnetic core wound with a coil around it, as depicted in Figure 3.21. The saturable inductor operates into its saturation region. It is designed in such a way that the external and internal flux density will affect its saturation level. Change in the saturation level of a saturable inductor will alter core's permeability and, consequently, its inductance  $L$ . The value of saturable inductance ( $L$ ) is high at low currents (based on the permeability of the core) and low at high currents (the core permeability becomes unity when saturated).

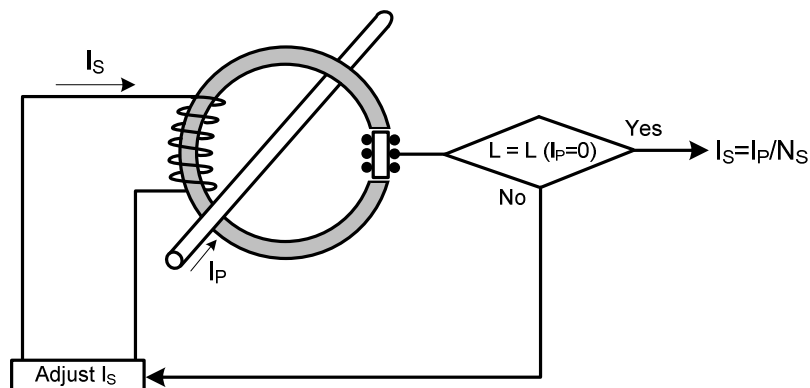


Figure 3.20. Saturable inductor current sensor

As shown in Figure 3.21, the saturable inductor current ( $I_{sic}$ ) passes through the coil wrapped around inductor, which produces the flux density  $B_{sic}$  in addition to the flux density  $B_P$  produced by current  $I_P$ .

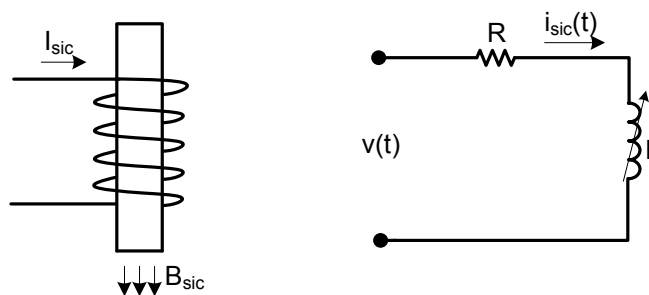


Figure 3.21. Saturable inductor and its equivalent circuit

When overall flux (due to  $I_P$  and  $I_{sic}$ ) is high, the core gets saturated and  $L$  drops. At lower flux, the value of  $L$  is high. If polarity of current  $I_{sic}$  gets change, the direction of the flux created by it will also change. If fluxes created by currents  $I_P$  and  $I_{sic}$  are opposite and have the same magnitude, they will create zero total flux. Saturable inductor is

designed to provide high  $L$  at zero total flux. As shown in Figure 3.20, the variations of  $L$  are detected by sensor electronics and compensated by using compensation current  $I_S$ , such that the total gap field is zero.

The current response of a saturable inductor to a voltage a voltage step when flux density due to current  $I_P$  and  $I_S$  ( $B_{P\&S}$ ) is zero and non-zero is illustrated in Figure 3.22.

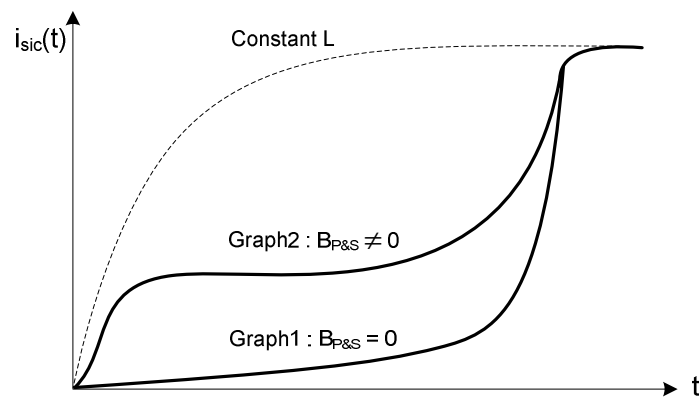


Figure 3.22. Current response of a saturable inductor to a voltage step

The graph for constant  $L$  shows the expected exponential current response. Graph-1 shows the current response of saturable inductor when  $B_{P\&S}$  is zero. For a small value of current  $I_{sic}$ , the inductor is non-saturated. When current  $I_{sic}$  exceeds the pre-defined level, the value of  $L$  drops, which results in fast current change. Graph-2 shows the current response of the saturable inductor when  $B_{P\&S}$  is non-zero. For small  $I_{sic}$  value,  $B_{P\&S}$  is predominant so  $L$  drops. When flux densities  $B_{sic}$  (due to current  $I_{sic}$ ) and  $B_{P\&S}$  are equal and opposite, the value of  $L$  is high, resulting in low current variation. As current  $I_{sic}$  continues to grow, the field caused by it becomes predominant. This leads the core

into opposite saturation, so  $L$  drops. The asymptotic level of current is defined by excitation voltage  $v(t)$  and coil resistance  $R$  (see Figure 3.21).

The current response of a saturable inductor when the square wave voltage is given as a supply in Figure 3.21 and flux density  $B_{P\&S}$  is zero and non-zero is illustrated in Figure 3.23.

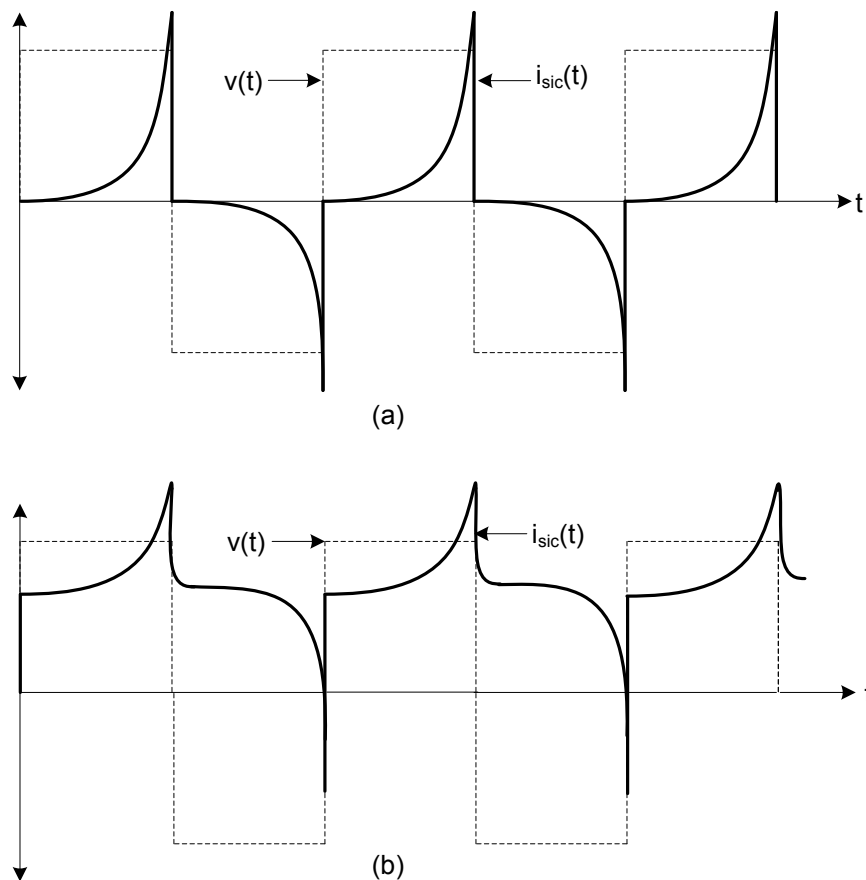


Figure 3.23. Voltage steps and current response of a saturable inductor; when (a)  $B_{P\&S} = 0$ , and (b)  $B_{P\&S} \neq 0$

If no primary current  $I_P$  is applied (or  $B_{P\&S}$  is zero), and the square wave voltage  $v(t)$  is given as a supply to saturable inductor in Figure 3.20, the behavior of current  $I_{sic}$  is

depicted in Figure 3.23(a). The shape of current  $I_{\text{sic}}$  is same as shown in Figure 3.22 for zero flux density  $B_{\text{p\&s}}$ . If the loop in Figure 3.20 is not closed and current  $I_{\text{s}}$  is not available, the behavior of current  $I_{\text{sic}}$ , when primary current  $I_{\text{p}}$  is present (or  $B_{\text{p\&s}}$  is non-zero), is given in Figure 3.23(b). The differences in two waveforms are the peak amplitude and dc current component. Different techniques used to sense the value of  $L$  in Figure 3.20 are measuring the dc current component, measuring the amplitude of the second harmonic, or measuring the duty cycle of the voltage of the Figure 3.23(b). Once the changes in  $L$  are detected, the closed-loop in Figure 3.20 is used to compensate the flux in the gap.

In this technique, high frequency performance is achieved by using two cores without air gaps. One of the two main cores is used to create a saturable inductor and the other is used to create a high frequency transformer effect. In another approach, three cores can be used without air gap. Two of the three cores are used to create saturable inductor, and the third core is used to create a high frequency transformer effect. More information about these techniques is given in [110] and [126]. It is difficult to compare different saturable inductor techniques. Advantages of saturable inductor sensors include high resolution, high accuracy, low offset and gain drift, and large bandwidth (up to 500 kHz). Drawbacks of saturable inductor technologies include limited bandwidth for simpler design, relatively high secondary power consumption, and risk of current or voltage noise injection into the primary conductor.

### **3.5. MAGNETO-RESISTIVE CURRENT SENSOR**

Every conducting material has some magneto-resistance. This magneto-resistance effect is large in pervalloids (Fe-Ni) and other ferromagnetic materials. A magneto-

resistor (MR) is a two terminal device which changes its resistance parabolically with applied magnetic field as shown in Figure 3.24. This variation of the resistance of MR due to the magnetic field is known as the anisotropic magneto-resistive (AMR) effect.

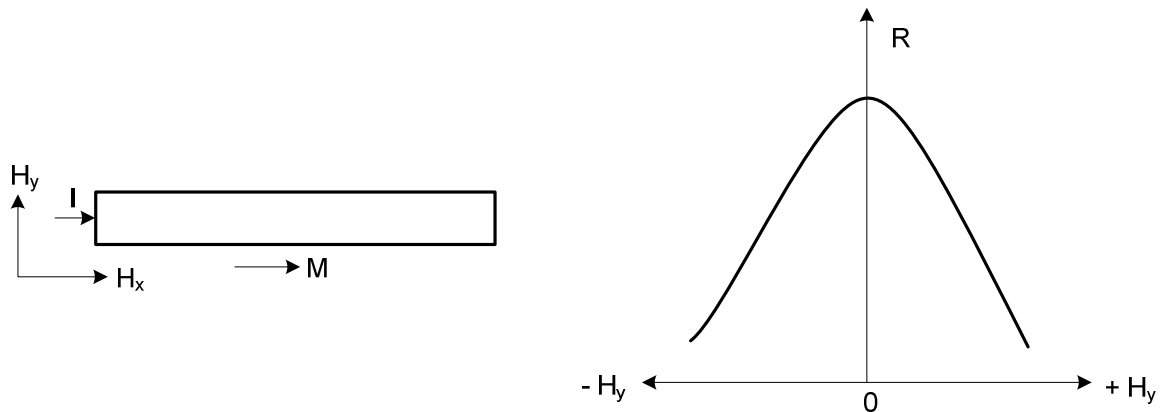


Figure 3.24. MR strip and resistance verses magnetic field for MR device

MR devices are more sensitive to magnetic field than are Hall-effect sensors. An MR device with a larger sensitivity is known as a giant magneto-resistor (GMR) [127]. An MR strip is depicted in Figure 3.24. Here  $H_y$  is the applied magnetic field,  $H_x$  is the internal or external (auxiliary) magnetic field, and  $M$  is the resultant magnetization due to  $H_y$  and  $H_x$ . Magnetic field  $H_y$  is coupled with an MR device and change the resistivity of the stripe, which is measured by the current  $I$  passing through the element. MR sensor sensitivity is controlled by controlling field  $H_x$ . When current  $I$  and magnetization  $M$  are parallel, the resistance of the strip is the greatest; when they are at a  $90^\circ$  angle to each other, it is the lowest [128-131].



An MR device cannot detect the direction of field  $H_y$  and has vanishing sensitivity for low fields. Other disadvantages include limited linear range, poor temperature characteristics, and a wide range of sensitivity from device to device. Non-linearity and non-directionality of an MR device are corrected by modifying the MR transfer curve, as shown in Figure 3.25.

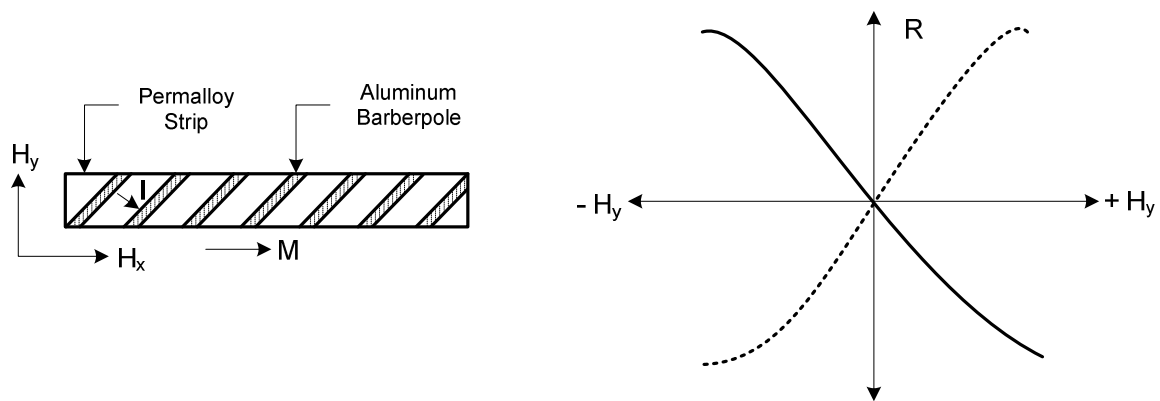


Figure 3.25. Barberpole MR device and resistance verses magnetic field for barberpole MR device

The most common way to modify the MR transfer curve is the use of barberpole configuration. Small structures of a highly conductive material such as aluminum are added to the MR stripe at an inclination of  $|45^\circ|$  to create the barberpole configurations. Barberpoles are tiny blocks of material sitting on top of the permalloy. One of such configuration with  $+45^\circ$  inclination of barberpoles is shown in Figure 3.25 ( $-45^\circ$  between current  $I$  and  $H_x$ -axis). Aluminum has a resistivity about five times lower than that of permalloy, so the barberpoles cause a change of operating point by changing the direction of current, as shown in Figure 3.25. With  $45^\circ$  barberpole biasing, if field  $H_y$  increases,

the resistance of barberpole MR device decreases, and if field  $H_Y$  decreases, the resistance of barberpole MR device increases, as shown by a solid line in the graph of Figure 3.25. The dashed line in the graph of Figure 3.25 shows the inverse behavior of barberpole MR device with  $-45^\circ$  barberpole biasing.

External magnetic fields can distort the current measurement of barberpole MR sensor. To avoid this distortion, barberpole MR devices can be configured into a half bridge or a full bridge arrangement. In the Wheatstone bridge configuration, shown in Figure 3.26, the arms of the bridge are made of four highly symmetrical MR devices.

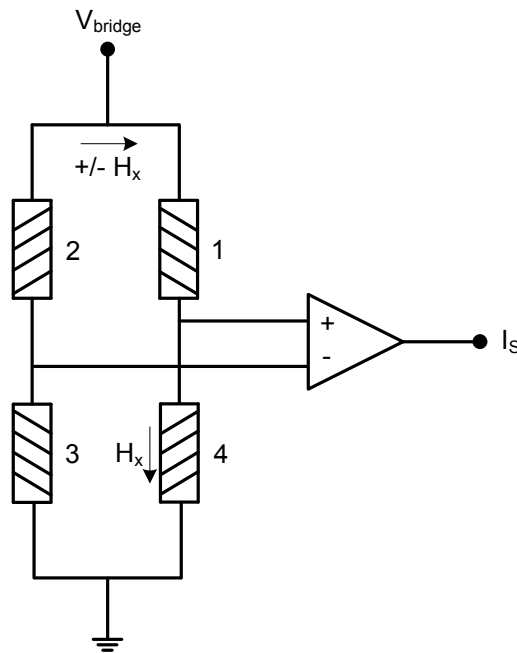


Figure 3.26. Four barberpole MR devices in a Wheatstone bridge configuration (1 and 3 with  $45^\circ$  and 2 and 4 with  $-45^\circ$  barberpole biasing)

As shown in Figure 3.26, in one arm of the bridge, the barberpoles are placed in opposite directions above the two MRs, such that, in the presence of a magnetic field, the

value of the first resistor increases and the value of the second decreases. The individual barberpole MR devices still sensitive to the temperature change; however, they all change at the same rate yielding a zero net drift at the output of the bridge.

Even with the barberpoles, the linearity of MR device is not very high; therefore, a compensation circuit is required, as in the closed-loop Hall-effect current sensor. An electrically isolated aluminum compensation conductor is integrated on the same substrate above the permalloy resistors as shown in Figure 3.27.

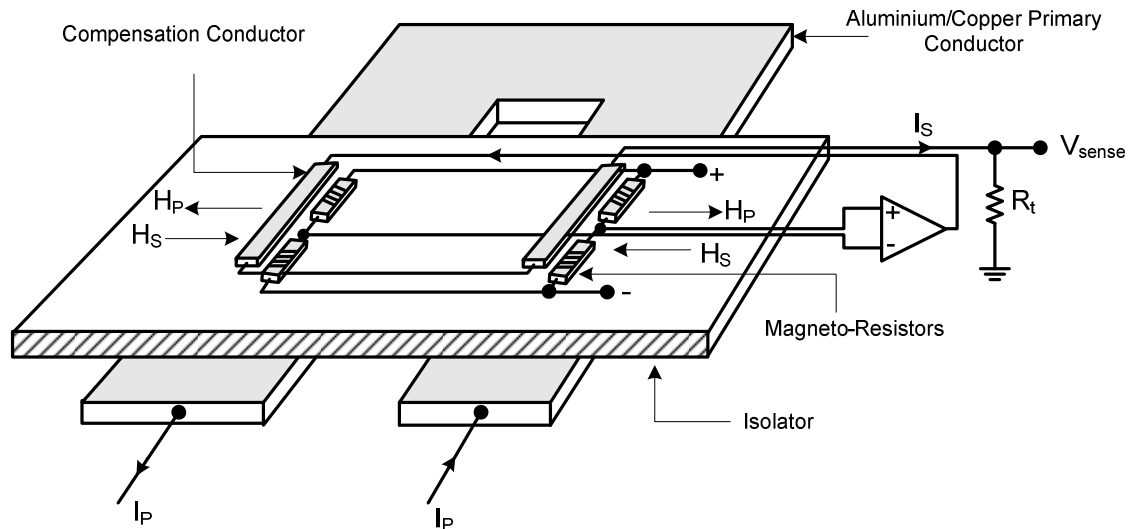


Figure 3.27. Barberpole MR sensors with compensation circuit

The output of the Wheatstone bridge is connected to the input of an operational amplifier, which generates the compensating current ( $I_s$ ). Current  $I_s$  then flows through the aluminum conductor to generate a magnetic field that exactly compensates the field created by current  $I_p$ . The bridge output voltage is usually close to zero; therefore, sensor nonlinearity is minimized. Current  $I_s$  is then measured through resistor  $R_t$ , which is a

exact representation of the current  $I_p$ . In order to have the same amplitude but opposite directions of the magnetic fields on two arms of the bridge, the primary current conductor under the substrate is U-shaped. More information about this technique is given in [131].

An MR current sensor is able to measure dc and ac currents accurately with galvanic isolation. Advantages of this technique include small size, wide bandwidth (dc to 100 kHz), and no remanence during overload. However, precise positioning of the barberpole MR device in the Wheatstone bridge is required with respect to the current carrying conductor. Barberpole MR devices have to be matched one another in sensitivity and temperature drifts, so they need to be placed close together to eliminate the effect of temperature gradient.

### 3.6. FIBER OPTIC CURRENT SENSOR

A fiber optics current sensor (FOCS) determines the current flow in an electrical conductor by measuring the magnetic field density within the vicinity of the conductor [132-136]. The operation principal is based on the Faraday Effect. When a polarized monochromatic light propagates parallel to a magnetic field, the polarization direction rotates, as shown in Figure 3.28.

The polarization angle is proportional to the magnetic field circulation on the optical path. The angular rotation experienced by the light passing through the sensor is given by (3.19).

$$\theta = VBl \quad (3.19)$$

Where  $V$  is the Verdet constant of the material used for sensor,  $B$  is magnetic flux density, and  $l$  is the length of FOCS exposed to the magnetic field. The polarization angle

also depends on the light wavelength and fiber material. FOCS measures the exact integral of the magnetic field along the closed-loop created by the fiber. A true current reading is obtained if FOCS completely encloses the conductor; otherwise the reading reflects the magnetic field intensity at the measurement point and has to be scaled accordingly.

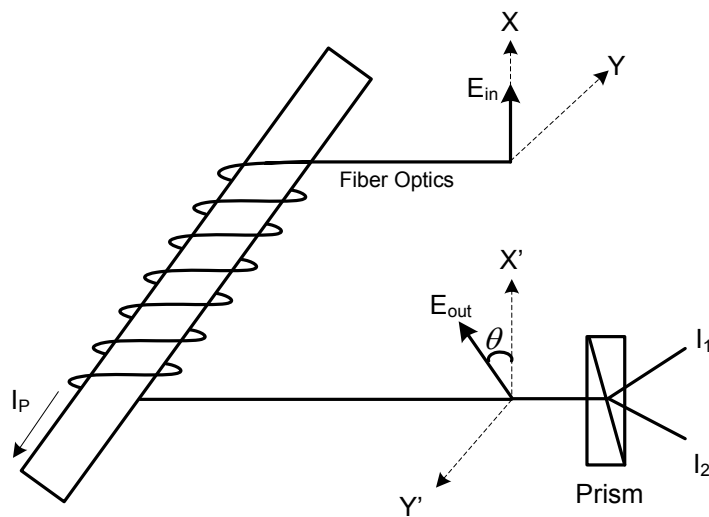


Figure 3.28. Polarimetric fiber optic current sensor

FOCSs are classified as intrinsic and extrinsic type. An intrinsic sensor uses fiber for current sensing, whereas an extrinsic sensor uses bulk optic. On the other hand, there are three different approaches of FOCS bulk, polarimetric, and interferometric. Bulk current sensor has high Verdet constant crystals, resulting in excellent sensitivity. Optical fiber has a lower Verdet constant, but many turns of the fiber around the conductor results in an improved sensitivity. The polarimetric sensor measures the rotation of a

linear polarization, while the interferometric sensor using a Sagnac interferometer measures the non-reciprocal phase shift.

FOCS offers many advantages over to conventional approaches. They can be used without any risk of discharge with the ground. Additionally, they offer a total protection against stray magnetic fields and have wide bandwidth that allows harmonic analysis of the measuring currents. They are very small; therefore, design of light and compact sensors is easy. Fiber optic sensor's installation is fast, simple, and can be done without interrupting the electric circuit to be measured. The built-in insulation of the optical fiber is an important characteristic for high voltage installations.

### 3.7. COMPARISION BETWEEN ELECTROMAGNETIC-BASED CURRENT SENSING TECHNIQUES

Table 3.1 summarizes the advantages and disadvantages of the presented electromagnetic-based current sensing techniques.

Table 3.1. Comparative overview of electromagnetic-based current sensing techniques

Techniques	Advantages	Disadvantages	Cost	Bandwidth	Measuring range
Using ACCT	Lossless, good SNR, good common mode rejection	Measures only ac current, core limitations, limited frequency range, not suitable for multiple-inductor converters	\$3 (10A)	0.1Hz-100MHz	mA-kA
Using ACCT with sample-and-hold	Accurate, low power loss, able to measure ac current having dc current component	Requirement of high bandwidth sample-and-hold ICs and two transformers, core limitations	-	-	-

Table 3.1. Comparative overview of electromagnetic-based current sensing techniques (Cont.)

Techniques	Advantages	Disadvantages	Cost	Bandwidth	Measuring range
Using UCT	Lossless, accurately measures ac current having dc current component	Not suitable for higher switching frequencies, core limitations	-	-	-
Using DCCT	DCCT can work with primary currents in either direction	Limited bandwidth (<100 kHz), significant output distortion, need of accurate transformers, core limitations	-	-	50-2000A
Using Rogowski Coil	Accurate, low weight, no dc current saturation, low sensitivity to parameter variations, ac and pulsed dc current measurements	External circuit is required to analyze output, open structure leads to measurement error, error is introduced by processing electronics	-	10Hz–100kHz	0-10kA
Using planar Rogowski Coil	Lossless, ac and pulsed dc current measurements accurate, light and small, insensitive to external field perturbations and conductor position inside the coil	Expensive, external circuit is required to analyze output, complicated circuit, high secondary power consumption	\$122 (400A)	10Hz–100kHz	0-10kA
Using open-loop Hall-effect current sensor	Low secondary power consumption, small size, low cost, ac, dc, and complex current measurements	Low sensitivity, temperature dependent output, linearity errors, prone to static drift, core limitations	\$43 (300A)	0-25kHz	0-15kA
Using closed-loop Hall-effect current sensor	Accurate, ac, dc, and complex current measurements, fast response time, wide bandwidth, low temperature drift	Compensation circuit is required, high secondary current consumption, expensive, bulky for low currents, core limitations	\$94 (300A)	0-200kHz	0-15kA

Table 3.1. Comparative overview of electromagnetic-based current sensing techniques (Cont.)

Techniques	Advantages	Disadvantages	Cost	Bandwidth	Measuring range
Using an open-loop Hall-effect current sensor with a CT	Unipolar supply, low power consumption, ac and dc, and complex current measurement, wide bandwidth	Linearity errors, low frequencies; offset and gain drift with temperature, moderate accuracy, core limitations	\$21 (100A)	0-100kHz	25-150A
Using saturable inductor current sensor with one core	High resolution, high accuracy, ac and dc current measurement	Limited bandwidth, high secondary power consumption, core limitations	-	0-200kHz	0-500A
Using saturable inductor current sensor with two cores	Good resolution, accurate, wide bandwidth, good immunity to surrounding magnetic fields, ac and dc current measurement	Risk of current or voltage noise injection into the primary conductor, high secondary power consumption, core limitations	\$636 (50A)	0-500kHz	0-150A
Using saturable inductor current sensor with three cores	High resolution, high accuracy, wide bandwidth, low noise on output signal, ac and dc current measurement	Limited operating temperature (10°-50° C), high secondary power consumption, large dimensions, costly, core limitations	\$992 (300A)	0-100kHz	0-700A
Using MR current sensor	Smaller volume and weight, no remanence, ac and dc current measurement, high sensitivity, noise immunity	Placement of barberpoles in Wheatstone bridge configuration, limited frequency response due to magnetic inertia of permalloy and skin effect of current carrying conductor	\$250 (<2A)	0-100kHz	0-50A
Using FOCS	Lossless, no electromagnetic interference, small sensing elements, ac and dc current measurement	Expensive, environmental sensitivity	-	10Hz-6kHz	up to 500kA



#### **4. CURRENT SENSING TECHNIQUES WITH SELF-TUNING AND/OR AUTO-CALIBRATION**

In most measurement techniques, the accuracy of the measurement depends on some assumed parameter e.g. resistor value in resistive-based approaches. This assumed value is subject to temperature variations and inaccuracies. Due to these non-ideal effects, conventional current measurement approaches are sensitive to temperature, tolerance of components, noise, and operating conditions. The solution is self-tuning approaches in which the assumption that is made on the value of a parameter will tune itself [39].

##### **4.1. MOSFET $R_{DS}$ -BASED CURRENT SENSING TECHNIQUE WITH REAL-TIME SELF CALIBRATION**

For accurate current measurement, it is necessary that circuit should somehow measure one of the current carrying elements in its path and sense the voltage across the same for accurate current measurement. One approach combines externally added sense resistor-based current sensing technique with another lossless method, such as MOSFET  $R_{DS}$ -based current sensing, filter-based current sensing, or observer current sensing for accurate current measurement. In Figure 4.1, externally added sense resistor-based current sensing method is combined with the MOSFET  $R_{DS}$ -based current sensing to effectively measure current  $I_L$  by determining the accurate value of resistance  $R_{DS}$ . Accuracy, along with losslessness, is achieved using an extra circuit which is a series connection of switch  $S_3$  and resistor  $R_{sense}$  [137-139]. This extra circuit is added into the circuit infrequently for calibration purpose; therefore, the current rating of switch  $S_3$  is small.

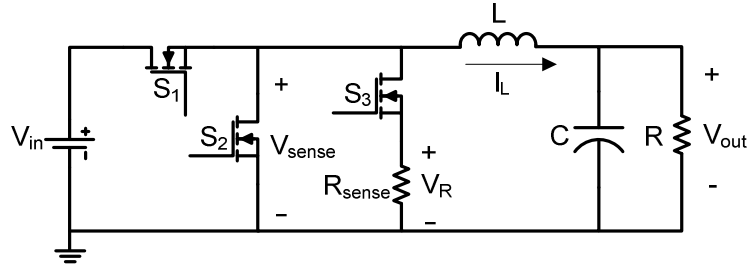


Figure 4.1. MOSFET  $R_{DS}$ -based current sensing technique with self-calibration

During a normal cycle, switch  $S_2$  is on for interval  $0 < t < (1-d)T$  and current  $I_L$  is specified by (4.1).

$$I_L = \frac{-V_{sense}}{R_{DS}} \quad (4.1)$$

Resistance  $R_{DS}$  is initially estimated from the MOSFET datasheet. To find calibrated value of resistance  $R_{DS}$  ( $R_{DS\_cali}$ ), main switch  $S_2$  is kept off, switch  $S_3$  is turned on, and the voltage across resistor  $R_{sense}$  ( $V_R$ ) is measured to find the value of current  $I_L$ , which is specified by (4.2).

$$I_L = \frac{-V_R}{R_{sense}} \quad (4.2)$$

If current  $I_L$  is not affected by the switching of  $S_3$ , from (4.1) and (4.2), the value of  $R_{DS\_cali}$  is derived as (4.3).

$$R_{DS\_cali} = \frac{R_{sense} V_{Sense}}{V_R} \quad (4.3)$$

The accuracy of (4.3) depends on the value of the  $R_{\text{sense}}$ . The combined resistance of switch  $S_3$  and resistor  $R_{\text{sense}}$  must be low enough, that the body diode of switch  $S_2$  is not tuned on when switch  $S_3$  is conducting. Another concern when finding the value of  $R_{\text{DS\_cali}}$  is that the calibration cycle is performed only when the circuit is operating in steady state. If these conditions are satisfied, then the value of (4.3) is used in place of resistance  $R_{\text{DS}}$  in (4.1) to accurately find current  $I_L$ .

The slope of current  $I_L$  will increase if the combined voltage of switch  $S_3$  and resistor  $R_{\text{sense}}$  is more than the voltage across switch  $S_2$  ( $V_{\text{sense}}$ ), introducing an error in current  $I_L$  ( $I_{L\_error}$ ), as depicted in Figure 4.2.

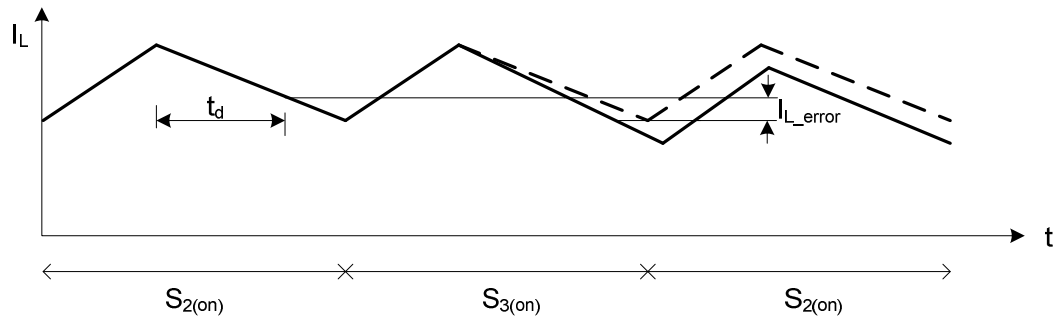


Figure 4.2. Error in the inductor current

In this situation, the value of error  $I_{L\_error}$  must be found in order to find the correct value of  $R_{\text{DS\_calib}}$ . However, this error can be large in converters having low output voltage and low inductance.  $I_{L\_error}$  can be measured if the values of inductor  $L$  and the time delay  $t_d$  for sampling are known. The error  $I_{L\_error}$  is calculated by using (4.4).

$$I_{L\_error} = t_d \left( \frac{V_{sense\_S_2(on)} - V_{sense\_S_3(on)}}{L} \right) \quad (4.4)$$

Once the value of  $I_{L\_error}$  is known, the value of  $R_{DS\_calib}$  is calculated using (4.5).

$$R_{DS\_calib} = R \frac{V_{sense\_S_2(on)}}{V_R} * \frac{1}{1 + \frac{I_{L\_error} R}{-V_R}} \quad (4.5)$$

The main difficulties with this method are the dependence on L and the need to find the values of voltages  $V_{sense}$  and  $V_R$  simultaneously. Once all of the above information is collected, then the value of (4.5) is used in (4.1) in place of  $R_{DS}$ , to find current  $I_L$ .

The error  $I_{L\_error}$  is canceled by adjusting the duty cycle. There are three steps to achieve accurate current  $I_L$  measurement, as depicted in Figure 4.3.

1. When switch  $S_3$  is on and switch  $S_2$  is off, voltage  $V_{sense\_S_3(on)}$  is sensed to find the slope of the inductor current.
2. When switch  $S_2$  is on and switch  $S_3$  is off, duty cycle  $d$  is increased by  $\Delta d$  for compensation. The value of  $\Delta d$ , which does not depend on the value of L, is calculated by using (4.6).

$$\Delta d = \frac{V_{sense\_S_2(on)} - V_{sense\_S_3(on)}}{V_{in}} \left( \frac{t_d}{T_S} + (1-d) \right) \quad (4.6)$$

This increased duty cycle compensates the error in inductor current.

3. When switch  $S_3$  is on and switch  $S_2$  is off, voltage  $V_R$  is measured and used in (4.3) to find the value of  $R_{DS\_calib}$ .

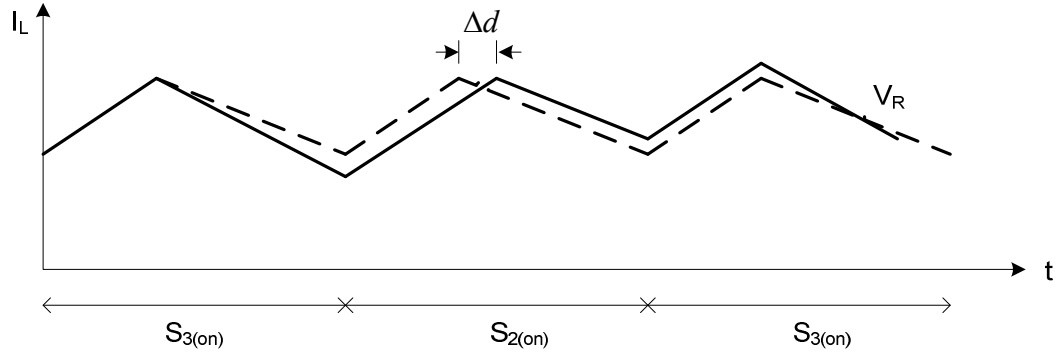
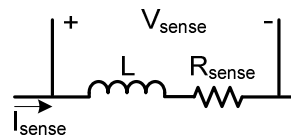


Figure 4.3. Duty cycle adjustment

#### 4.2. CURRENT SENSING TECHNIQUE USING THE INTERNAL RESISTANCE OF AN INDUCTOR AND THE FILTER

The basic idea for this technique is taken from the resistive-based current sensing using external sense resistor method of Section 2. At higher frequencies, the parasitic equivalent inductance of resistor  $R_{sense}$  appears. Hence, it is necessary to compensate for the parasitic inductance. The equivalent circuit of  $R_{sense}$  for this case is given in Figure 4.4.

Figure 4.4. Equivalent circuit of  $R_{sense}$ 

Voltage  $V_{sense}$  is given by (4.7).

$$V_{sense}(s) = (R_{sense} + Ls)I_{sense}(s) = K * I_{sense}(s) \quad (4.7)$$

At lower frequencies, voltage  $V_{\text{sense}}$  is proportional to current  $I_{\text{sense}}$ . At higher frequencies, gain  $K$  increases due to the parasitic inductance. A proper low pass filter, which can be active or passive, is required for compensating the gain  $K$ . Information about the use of an active filter for gain compensation is given in [140]. If a passive  $R_f C_f$  low pass filter is used, the voltage across the filter ( $V_{Cf}$ ) is then given by (4.8).

$$V_{Cf}(s) = \left( \frac{R_{\text{sense}} + Ls}{1 + R_f C_f s} \right) I_{\text{sense}}(s) \quad (4.8)$$

As mentioned in Section 2, inductor windings have DCR or internal resistance  $R_L$ . It is possible to use the DCR of inductor  $L$  as  $R_{\text{sense}}$  and inductor  $L$  itself as parasitic inductor in above case. Therefore, the filter-based current sensing technique uses the resistor  $R_L$  of inductor  $L$  and passive filter  $R_f C_f$ , as shown in Figure 4.5, for accurate current sensing. The total impedance of the  $R_f C_f$  filter is same as the total impedance of  $L$  and  $R_L$ . This technique is currently popular because of its accuracy, losslessness, and high bandwidth [141, 142]. Other advantages of this technique include; continuous current measurement, low cost, PCB space saving, and power efficiency improvement.

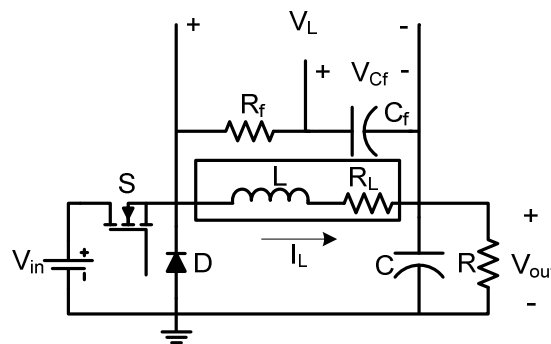


Figure 4.5. Filter-based current sensing technique

This technique detects the current  $I_L$  signal by sensing capacitor  $C_f$  voltage ( $V_{Cf}$ ). Voltage  $V_{Cf}$  is given by (4.9).

$$V_{Cf}(s) = I_L(s) \left( \frac{R_L + Ls}{1 + R_f C_f s} \right) \quad (4.9)$$

The parallel  $R_f C_f$  filter is designed in such a way that (4.10) and (4.11) are satisfied.

$$R_f C_f = \frac{L}{R_L} \quad (4.10)$$

$$\frac{L}{R_L} \gg T \quad (4.11)$$

Where  $\tau_L = L/R_L$  is an inductor time constant and  $\tau_C = R_f C_f$  is a low pass filter time constant. It is very difficult to satisfy (4.10). However, it is easy to satisfy (4.11) because the switching frequency is usually in the order of a few hundred kHz and resistance  $R_L$  is in mili-ohms.

If (4.10) and (4.11) are satisfied, the inductor current  $I_L$  is given by (4.12).

$$I_L = \frac{V_{Cf}}{R_L} \quad (4.12)$$

Figure 4.6 shows the response of voltage  $V_{Cf}$  and  $V_{RL}$  when time constants  $\tau_L$  and  $\tau_C$  match.

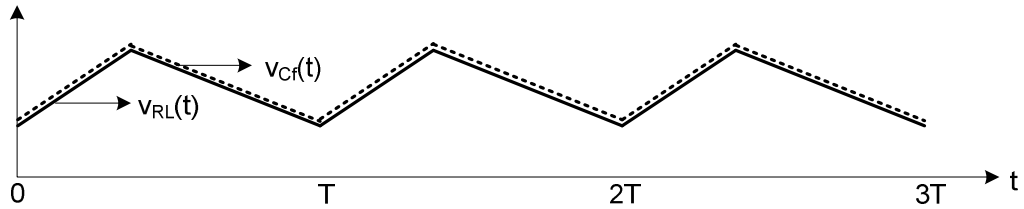


Figure 4.6. Voltage  $V_{Cf}$  and  $V_{RL}$ , when  $\tau_C = \tau_L$

It is possible that the time constants will not match; one time constant can be greater or less than the other due to the tolerance of components, temperature dependence of  $R_L$ , and change in inductance due to dc current bias. Figure 4.7 shows the response of voltage  $V_{Cf}$  in steady state. When switch S is on, current  $I_L$  increases at a rate of  $(V_{in} - V_{out})/L$  and capacitor  $C_f$  gets charged. When the freewheeling diode is on and switch S is off current  $I_L$  decreases at a rate of  $(-V_{out})/L$ , and capacitor  $C_f$  gets discharged. From charging and discharging rates, it is possible to find the slope of voltage  $V_{Cf}$ .

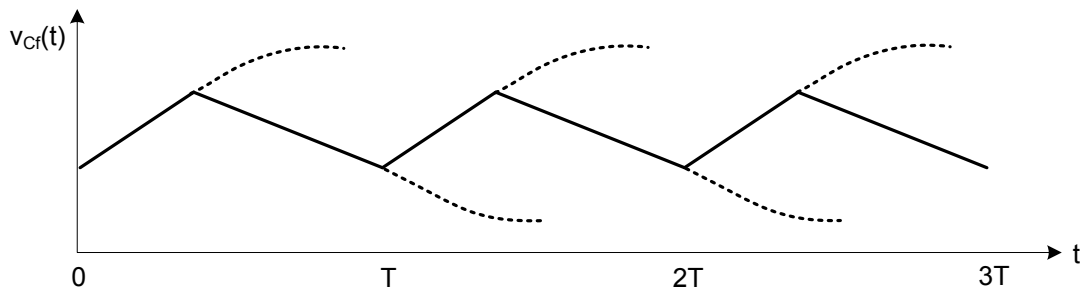


Figure 4.7. Voltage  $V_{Cf}$  in steady state

If time constants  $\tau_L$  and  $\tau_C$  are not matched, then peak to peak ripple of voltage  $V_{Cf}$  gets bigger or smaller than the peak-to-peak voltage ripple across resistor  $R_L$



generated by ac component of the current  $I_L$ . This will change the slew rate of voltage  $V_{Cf}$  of Figure 4.7.

**4.2.1. Filter-based Current Sensing with Temperature Compensation.** An error is introduced in filter-based current sensing due to the temperature dependency of resistor  $R_L$ . To minimize this error, resistor ( $R_{Cf}$ ) is added across the capacitor  $C_f$  for temperature compensation, as shown in Figure 4.8.  $R_{Cf}$  is the combination of resistors  $R_e$ ,  $R_g$ , and  $R_{NTC}$ .  $R_{NTC}$  is a negative temperature co-efficient resistor. Information about the values of  $R_e$ ,  $R_g$ , and  $R_{NTC}$  is given in [143]. Resistor  $R_{Cf}$  can also use for scaling down the detected current.

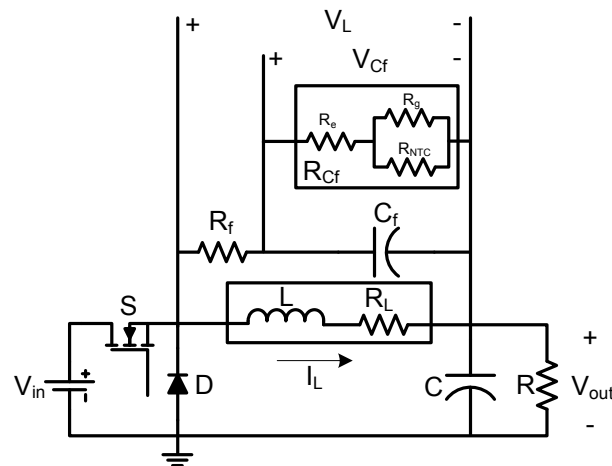


Figure 4.8. Filter-based current sensing with temperature compensated network

The addition of resistor  $R_{Cf}$  to the circuit changes the filter time constant. The new filter time constant is given by (4.13).

$$\tau_C = C_f \left( \frac{R_f R_{Cf}}{R_f + R_{Cf}} \right) \quad (4.13)$$

If  $R_{Cf}$  is added to the circuit and time constants  $\tau_L$  and  $\tau_C$  are matched, the voltage  $V_{Cf}$  is given by (4.14). Figure 4.9 shows the illustration of voltage  $V_{Cf}$  and  $V_{RL}$ . The dc voltage on the filter capacitor is level shifted by the factor  $R_{Cf} / (R_{Cf} + R_f)$  in the presence of resistor  $R_{Cf}$ , as shown in Figure 4.9.

$$V_{Cf} = I_L \left( \frac{R_L R_{Cf}}{R_f + R_{Cf}} \right) \quad (4.14)$$

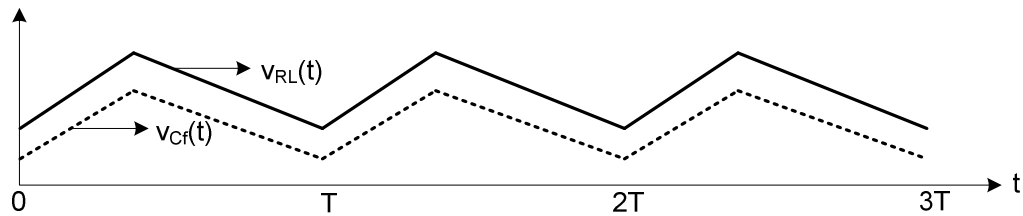


Figure 4.9. Voltage  $V_{Cf}$  and  $V_{RL}$ , when  $\tau_C = \tau_L$ , and resistor  $R_{Cf}$  is added into the circuit

**4.2.2. Filter-based Current Sensing with Accurate  $R_L$ .** If the filter-based current sensing technique depicted in Figure 4.5 is used for over current protection, and time constant  $\tau_L$  is greater than time constant  $\tau_C$ , the over current protection circuit can trip at a lower than desirable current. When time constant  $\tau_L$  is less than time constant  $\tau_C$ , the response is opposite. This is due to the fact that the value of  $R_L$  is not accurate; it is greater than the desired value. In order to deal with  $R_L$  parameter uncertainty, a modified

current sensing circuit for over current protection is shown in Figure 4.10 [145]. Design steps for this technique are given below.

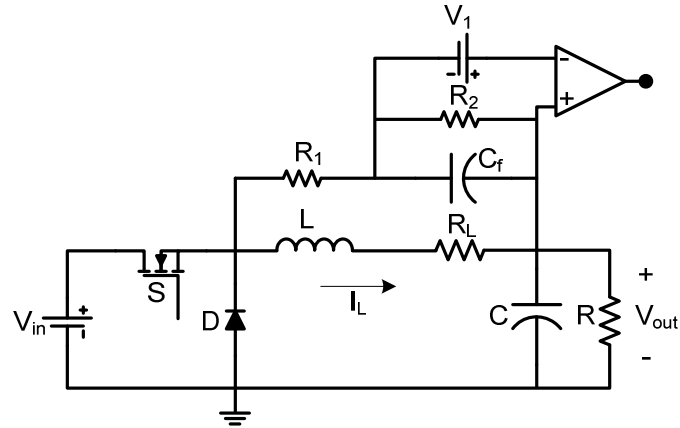


Figure 4.10. Filter-based current sensing with accurate  $R_L$

If  $V_1$  is the tripping voltage,  $I_1$  is the tripping current, current  $I_1$  is given by (4.15).

$$I_1 = \frac{V_1}{R_L} \quad (4.15)$$

In order to get actual value of  $R_L$ , a desired sense resistance  $R_{sense}$  can be calculated from (4.16).

$$R_{sense} = \frac{V_1}{I_1} \quad (4.16)$$

Once the value of  $R_{sense}$  is known, then the value of inductor  $L$  can be found from the value of  $R_L$ , which can be determined by (4.17).

$$R_L \geq R_{sense} \quad (4.17)$$

The maximum value of L at zero current and the minimum value of  $R_L$  at room temperature are measured to find value of time constant  $\tau_C$ , as depicted in (4.18).

$$R_{f \min} C_{f \min} > \frac{L_{\max}}{R_{L \min}} \quad (4.18)$$

If  $R_L$  is greater than  $R_{sense}$ , then the values of resistances  $R_1$  and  $R_2$  must be selected in such a way that (4.19) is satisfied.

$$\frac{R_2}{(R_1 + R_2)} = \frac{R_{sense}}{R_L} \quad (4.19)$$

$R_f$  of Figure 4.5 is equivalent to the resultant resistance of the parallel connection between resistors  $R_1$  and  $R_2$  in Figure 4.10.  $R_L$  of Figure 4.5 is equivalent to  $R_L * R_2 / (R_1 + R_2)$  in Figure 4.10.

**4.2.3. Filter-based Current Sensing with Self-tuning.** To overcome the time constants mismatching issue, a modified digital auto-tuning approach is presented in [146]. This technique automatically matches the time constants based on the measurement of the slope of voltage  $V_{Cf}$ , as depicted in Figure 4.11. The filter resistor  $R_f$  is variable.

A mismatch of time constants causes load transients, which create a large change in the value of voltage  $V_{Cf}$ . This change is sensed by a load transient detector. The controller measures the derivative of  $V_{Cf}$  signal and adjusts the value of variable resistor  $R_f$  to compensate for the time constant mismatch.

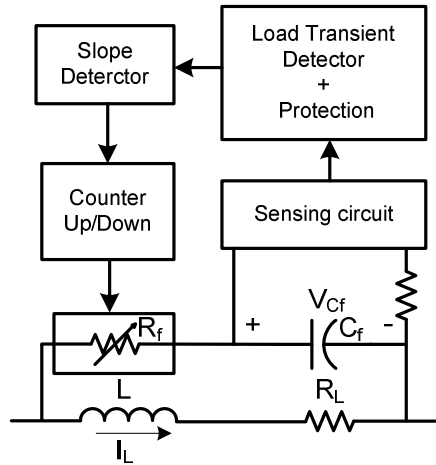


Figure 4.11. Digital auto-tuning circuit

**4.2.4. Combined-Sense Technique.** The sensed current in the filter-based current sensing has low magnitude and is susceptible to the interruption of noise, which is not acceptable in case of current-mode control [147]. In order to solve this problem, a new technique is introduced in [148] and shown in Figure 4.12. This technique provides better SNR of the sensed current.

Combined-sense technique includes additional switches  $S_3$ ,  $S_4$ , resistor  $R_d$ , and capacitor  $C_d$  in filter-based current sensing, as depicted in Figure 4.12.

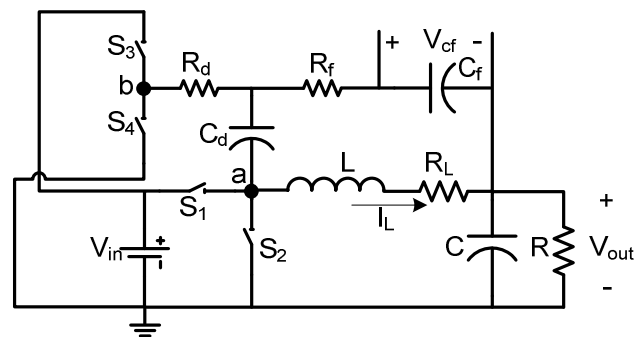


Figure 4.12. Combined-sense technique

Switches  $S_3$  and  $S_4$  are connected to node b. One end of their series connection is connected to  $V_{in}$  and the other is connected to the ground. Switches  $S_1$  and  $S_2$  are connected to node a.  $R_{DS}$  of switches  $S_1$ - $S_4$  are added into the filter according to the switching pattern. When switches  $S_1$  and  $S_3$  are on and  $S_2$  and  $S_4$  are off, the resistance of filter includes the  $R_{DS}$  of both switches  $S_1$  and  $S_3$ . Similarly, when switches  $S_2$  and  $S_4$  are on and  $S_1$  and  $S_3$  are off, the resistance of filter includes the  $R_{DS}$  of both switches  $S_2$  and  $S_4$ . In this way, the current can be sensed at any time during the cycle. The increased sense circuit resistance increases the voltage  $V_{Cf}$ , so the output signal is clean and less susceptible to noise. When the switches change their states from on to off or off to on, they temporarily fall into off state. The resistor  $R_d$  and capacitor  $C_d$  work as a low pass filter ( $R_d C_d$ ) during this dead time to block the high frequency transient signal from affecting the current sense signal. The  $R_d C_d$  filter is also synchronize the operation of nodes a and b.

Voltages at points a and b ( $V_a$  and  $V_b$ ) are depicted in Figure 4.13. Voltage  $V_a$  is a bit lower than voltage  $V_{in}$  and voltage  $V_b$  is approximately same as voltage  $V_{in}$ . because the rating of switches  $S_1$  and  $S_2$  are higher than the ratings of switches  $S_3$  and  $S_4$ . Filter  $R_d C_d$  averages the voltage difference for cleaner and exact current measurement.

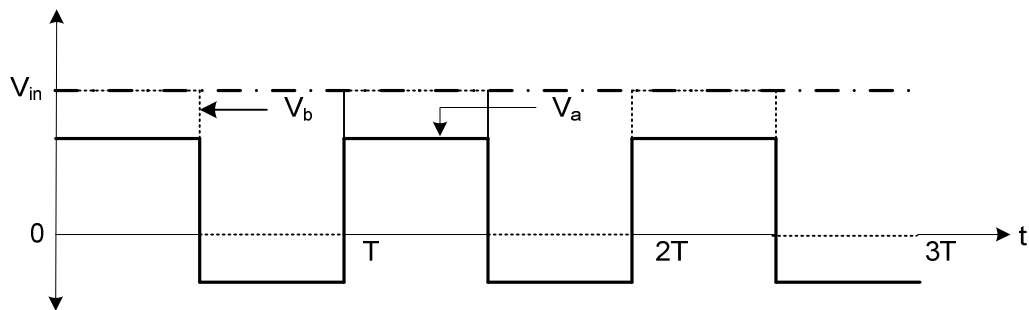


Figure 4.13. Voltages  $V_a$  and  $V_b$  in combined-sense technique

#### 4.2.5. Filter-based Current Sensing with Self-tuning and Self-calibration. To

achieve continuous current measurement with accuracy, losslessness, low switching noise, and high bandwidth, a modified technique is introduced in [149, 150] (see Figure 4.14).

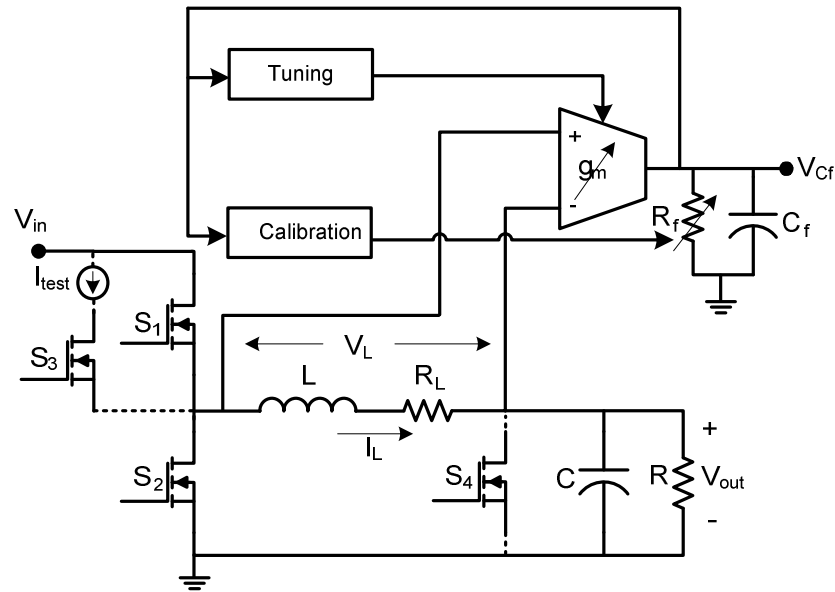


Figure 4.14. Filter-based technique with self-tuning and self-calibration

The sensed circuit is calibrated and tuned during the start-up and the inductor current is measured during normal operation, which ensures accuracy along with continuous current measurement benefits. The  $g_m$ - $C_f$  filter is a first order low pass filter. It is designed separately for each application as the inductor specifications are determined by the end user. The trans-conductance  $g_m$  and filter resistance  $R_f$  are variable. It is possible to change the frequency  $f_c$  by changing the value of  $R_f$  and to change the filter

gain by changing the value of  $g_m$ . The  $g_m$ - $C_f$  filter gives accurate current measurement after tuning and calibration.

By applying voltage  $V_L$  to a  $g_m$ - $C_f$  filter, whose voltage frequency response matches the current response of the inductor, one can write.

$$V_{C_f}(s) = g_m R_f \left( \frac{R_L + Ls}{1 + R_f C_f s} \right) I_L(s) \quad (4.20)$$

If  $R_f$  is adjusted to match the both time constants (equation (4.10)), current  $I_L$  is given by (4.21).

$$I_L = \left( \frac{V_{C_f}}{(g_m R_f) R_L} \right) \quad (4.21)$$

Here,  $g_m R_f R_L$  is current sensing gain, which can be adjusted to any value by changing  $R_f$  and/or  $g_m$ . Tuning and calibration are performed during the start-up to adjust the filter gain-bandwidth product and filter dc gain for accurate current measurement, as shown in Figure 4.15.

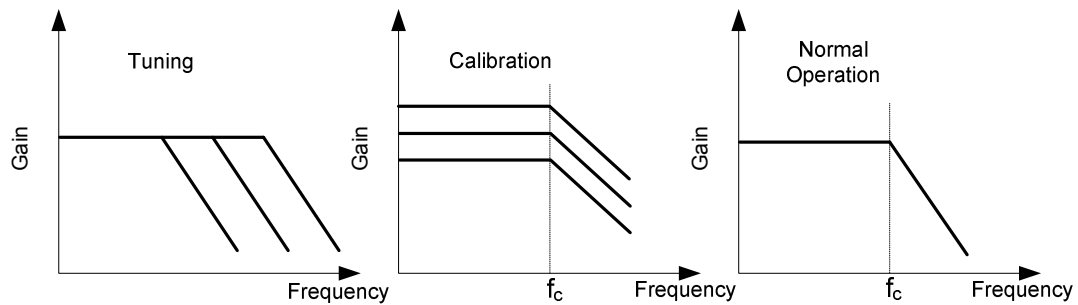


Figure 4.15. Tuning, calibration, and normal operation for filter-based current sensing technique



As tuning and calibration are performed during the start-up, switches  $S_1$  and  $S_2$  are kept disabled, so no power is lost during normal operation. This technique measures the inductor current during normal operation. Switches  $S_1$  and  $S_2$  are kept off and switches  $S_3$  and  $S_4$  are turned on during start-up for tuning purpose, as shown in Figure 4.14. A small ac test current ( $I_0/20$ ), triangular in shape, is injected into the circuit, as shown in Figure 4.16.

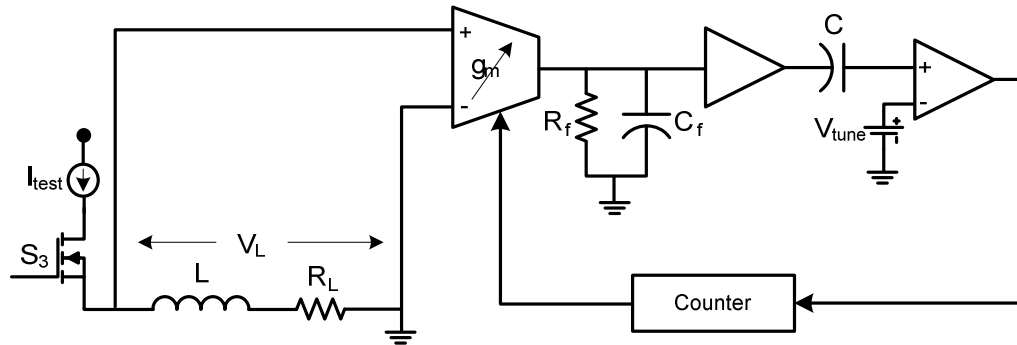


Figure 4.16. Tuning operation for filter-based current sensing technique

The test current's frequency is high enough that the voltage across inductor is dominated by inductance  $L$  and not by resistance  $R_L$  because inductor impedance is higher than resistance  $R_L$  at higher frequencies. The resulting ac voltage signal across inductor  $L$  is a square wave. The  $g_m$ - $C_f$  filter converts this square wave voltage  $V_L$  into the triangular wave voltage  $V_L$  because capacitor impedance is smaller than resistance  $R_f$  at higher frequency. Resistor  $R_f$  is kept at its minimum value to reduce the dc offset during the tuning process. The filter output is then buffered and the dc component is removed. The peak of the ac portion of voltage  $V_{Cf}$  matches with the tune voltage  $V_{tune}$

by adjusting  $g_m$ . Once both voltages are matched, the comparator stores their values and sends a stop signal to the counter.

The calibration phase immediately follows the tuning cycle. A small dc test current is supplied to the inductor for calibration, as shown in Figure 4.17. The value of filter resistance  $R_f$  is adjusted in the calibration process to match  $V_{Cf}$  with the calibration voltage  $V_{cali}$ .

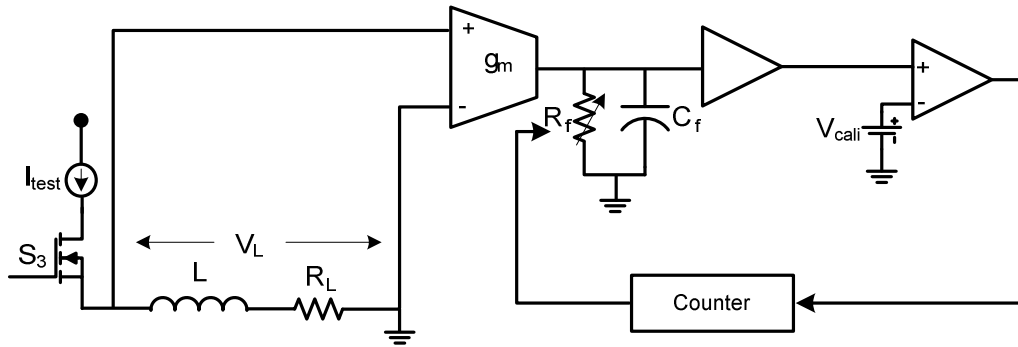


Figure 4.17. Calibration operation for filter-based current sensing technique

During tuning  $g_m$ - $C_f$  filter's cutoff frequency is matched with the inductor's cutoff frequency, so (4.10) is satisfied. During the calibration process, the current sensing gain  $g_m R_f R_L$  is adjusted to any value. If the value of the gain  $g_m R_f R_L$  is set to 1 ohm, then the value of  $V_{Cf}$  can directly provide information about inductor current, as shown in (4.22).

$$I_L = V_{Cf} \quad (4.22)$$

Once tuning and calibration parameters are properly set and stored, the converter is allowed to operate normally.

### 4.3. COMPARISON BETWEEN CURRENT SENSING TECHNIQUES WITH SELF-TUNING AND/OR AUTO-CALIBRATION

Table 4.1 summarizes the advantages and disadvantages of the presented current sensing techniques with self-tuning and/or auto-calibration.

Table 4.1. Comparative overview of current sensing techniques with self-tuning and/or auto-calibration

Techniques	Advantages	Disadvantages
Using MOSFET $R_{DS}$ -based current sensing with self-calibration	Lossless, accurate, combination of many lossless technique is also possible	Addition of an extra circuit, limited applications
Using filter-based current sensing	Lossless, continuous current measurement, low cost, low switching noise, high power efficiency	Low accuracy due to unknown L and $R_L$ , filter design, temperature dependence, tolerance of components, time constant mismatching, only for off-chip applications, change in inductance due to dc current bias
Using filter-based current sensing with self-tuning	Lossless, accurate, matched time constants, continuous current measurement	Complicated circuit, filter design
Using combined-sense technique	Lossless, accurate, continuous current measurement, low cost, improved SNR	Complicated circuit, filter design, unmatched $R_{DS}$ of main MOSFET switches, time constant mismatching
Using filter-based current sensing with self-tuning and self-calibration	Lossless, accurate, low switching noise, matched time constants, continuous current measurement, fully integrated	Complicated circuit, separate filter design for each application, accuracy of this technique depends on the tuning and calibration accuracy

## 5. AVERAGE CURRENT SENSING TECHNIQUES

It is hard to measure the instantaneous value of the current. In some applications, e.g. average current-mode control, there is no need to find the instantaneous value of current, only average value of current is enough. Different average current sensing techniques to calculate average inductor current are reviewed in this Section [39].

### 5.1. FILTER-BASED AVERAGE CURRENT SENSING

The simple, inexpensive, and efficient method of current sensing which measures the average value of the inductor current in a buck converter is depicted in Figure 5.1 [93, 151, 152,]. A simple  $R_f C_f$  filter is connected across low side switch  $S_2$  of the buck converter for average current measurement.

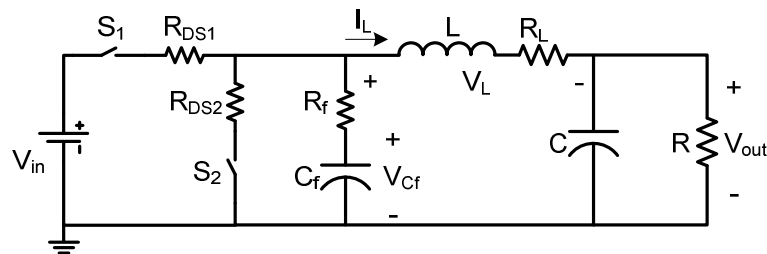


Figure 5.1. Filter-based average current sensing

Voltage  $V_L$  contains ac and dc parts. This technique measures the average voltage across the inductor, which is only a dc voltage across resistor  $R_L$ . Therefore, by measuring average inductor voltage, it is possible to measure average inductor current  $\langle I_L \rangle$ .

The value of  $R_f$  is much greater than the values of switch resistances  $R_{DS1}$  and  $R_{DS2}$ . Another requirement of this technique is given by (5.1).

$$\tau_c \gg \tau_L \quad (5.1)$$

Where  $\tau_c$  is a filter time constant and  $\tau_L$  is an inductor time constant. When switch  $S_1$  is on and  $S_2$  is off, capacitor  $C_f$  is charged. Similarly, when switch  $S_2$  is on and  $S_1$  off, capacitor  $C_f$  is discharged. Under steady state, the voltage across capacitor  $C_f$  will be equal to the average voltage across switch  $S_2$ . Since the average current through resistor  $R_f$  is zero, the average capacitor voltage ( $\langle V_{Cf} \rangle$ ) is given by (5.2).

$$\langle V_{Cf} \rangle = V_{out} + \langle I_L \rangle R_L \quad (5.2)$$

From (5.2), current  $\langle I_L \rangle$  can be derived as (5.3).

$$\langle I_L \rangle = \frac{\langle V_{Cf} \rangle - V_{out}}{R_L} \quad (5.3)$$

From (5.3), it is inferred that the average inductor current is only affected by  $R_L$ . The value of  $R_f$ ,  $C_f$ ,  $L$ , and  $R_{DS}$  have no effect on the current-sensing result. This technique is used in interleaved parallel dc-dc converters to balance the average load currents in various channels.

This technique is useful for low frequency measurements only. Its main drawback is that, for accurate current sensing, the value of  $R_L$  must be known. This technique provides information about average current only; it does not give any information about ac currents or transients.

## 5.2. CURRENT TRANSFORMER-BASED AVERAGE CURRENT SENSING

Transformers are used to measure the “switched” current, not the “average” current delivered to the load. However, in this sub-section, different current sensing techniques using current transformers for average inductor current measurement are described [38, 153].

**5.2.1. Average Current Measurement using Two Current Transformers.** This technique uses two current transformers for average inductor current measurement. In a boost converter, two current transformers are used to sense the switch currents to measure average inductor current, as shown in Figure 5.2.

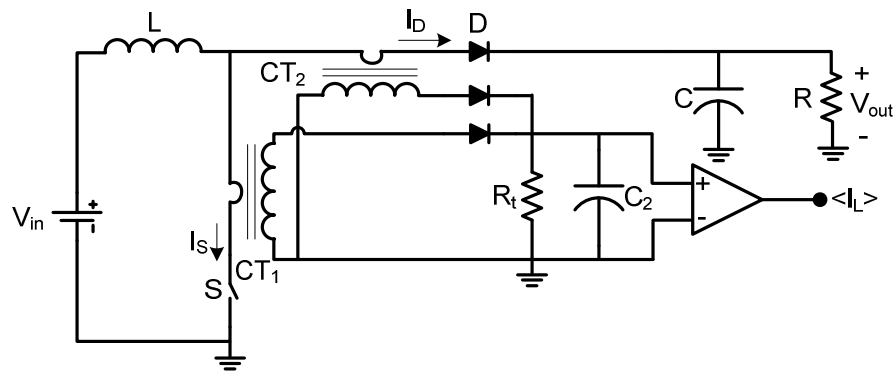


Figure 5.2. Average inductor current sensing using two current transformers

$CT_1$  measures switch current  $I_s$  and  $CT_2$  measures diode current  $I_D$ . The outputs of the two CTs are then added to get the average value of the inductor current. Output is accurate both in waveform and in dc value. As the two transformers do not have same the number of turns, this method does not provide very accurate information about average current.

**5.2.2. Average Current Measurement using One Current Transformer.** This technique uses only one current transformer for average current measurement. Forward converter with one current transformer for average inductor current measurement is depicted in Figure 5.3.

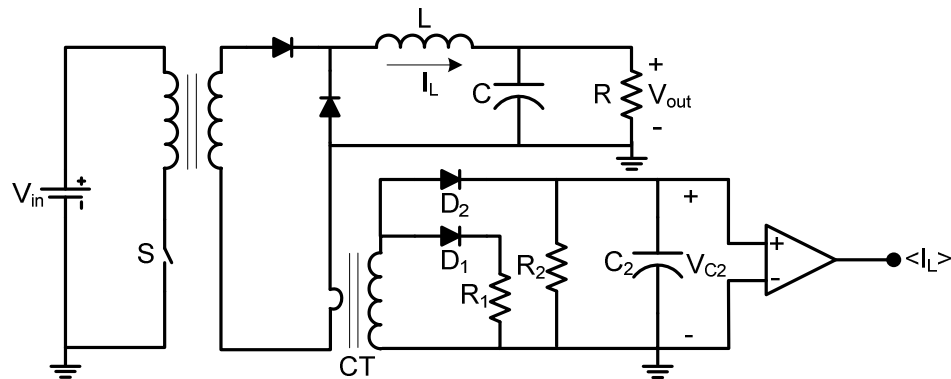


Figure 5.3. Average inductor current sensing using one current transformer

This technique replicates the inductor current waveform as capacitor voltage  $V_{C2}$  waveform. Voltage  $V_{C2}$  accurately follows the ac and dc excursion of inductor current. When switch  $S$  is on, the CT measures the inductor current. The output of CT is converted into the voltage using resistor  $R_1$ . This voltage is then used to charge capacitor  $C_2$ , which gives the rising portion of inductor current. When switch  $S$  is off, CT gets reset, and capacitor  $C_2$  is discharged through resistor  $R_2$ , which gives the down-slope of the inductor current. The waveform for voltages  $V_{R1}$  and  $V_{C2}$  are given in Figure 5.4 to enable better understanding of this technique.

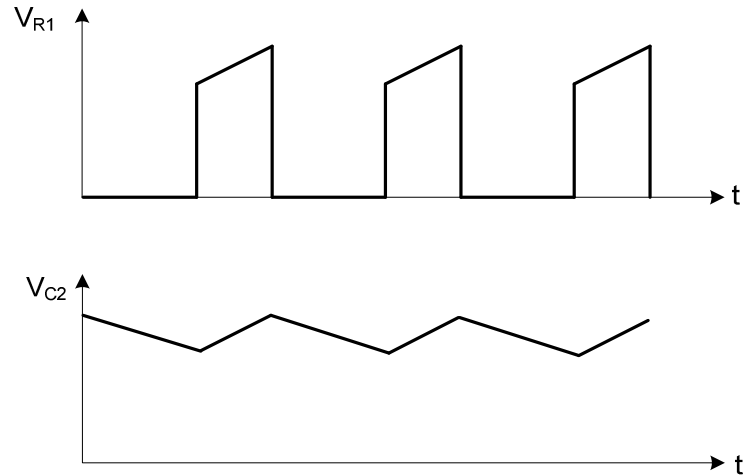


Figure 5.4. Voltage  $V_{R1}$  and  $V_{C2}$  for average current measurement

**5.2.3. Average Current Sensing by Adding the Second Winding on the Existing Inductor.** This technique mirrors the down-slope of the inductor current by using the second winding on the inductor, as depicted in Figure 5.5. The second winding generates the discharge current for capacitor  $C_2$  through push-pull amplifier. By using inductor as CT, isolation is achieved with both charge and discharge signals.

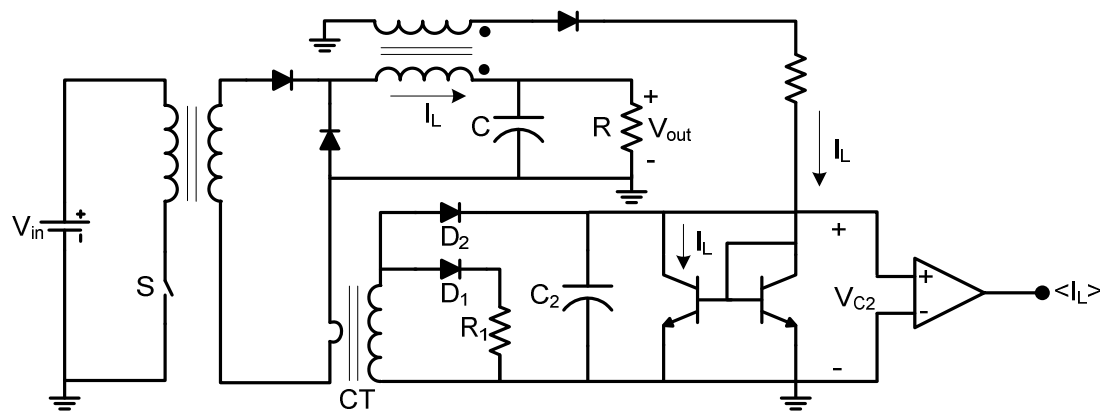


Figure 5.5. Average inductor current sensing by adding second winding on the existing inductor



In all of the above discussed techniques in this sub-section, the sensing circuits have their own grounds. Therefore, the sensing circuit can reside on either side of the isolation boundary.

### 5.3. COMPARISON BETWEEN AVERAGE CURRENT SENSING TECHNIQUES

Table 5.1 summarizes the advantages and disadvantages of the presented average current sensing techniques.

Table 5.1. Comparative overview of average current sensing techniques

Techniques	Advantages	Disadvantages
Using filter	Lossless, easily integrated on IC chip	Unknown $R_L$ , information of average inductor current only, sensing dependence on $R_f$ and $C_f$
Using two CTs	Simple, lossless	Requirement of two matched sensing transformers
Using one CT	Simple, lossless	Not very accurate, limited applications
Using inductor as a sense transformer	Simple, accurate, lossless	Additional winding expense, requirement of a high voltage approved isolation for off line converters

## 6. SUMMARY

Current measurement is a major contributor to the success in several applications. There are many applications for current sensing, including control, protection, monitoring, and power management. There are several issues related to current sensing, including ac or dc current measurement, complexity, linearity, sensitivity to switching noise, isolation requirements, accuracy, stability, robustness, bandwidth, transient response, cost, and acceptable power losses.

Current measurement is intrusive as there is a need to insert some type of sensor in the path of current flow for the current measurement. Most of the current measurement approaches can be categorized as resistive-based or electromagnetic-based techniques. In resistive-based current sensing techniques, the voltage drop across a sense resistor is measured to determine the current. These techniques have limited accuracy because the values of discrete elements of the circuit are unknown. Therefore, for accurate and continuous current measurement, different current sensing techniques with self-tuning and/or self-calibration are used. Electromagnetic-based current sensing techniques are used to measure current in high power applications where isolation is required. These techniques sense the magnetic field created by the current to be measured. Electromagnetic-based current sensing techniques are used to measure wide range of ac, dc, and complex currents with broad bandwidth.

Different current sensing strategies, such as resistive-based current sensing, electromagnetic-based current sensing, current sensing with self-tuning and/or auto-calibration, and average current sensing, are reviewed and evaluated in this thesis. Each

method has been introduced and its principle of operation is described. Different methods have been evaluated based on desirable characteristics. Several measurement techniques have been compared to each other, as well.

Applications of resistive-based and electromagnetic-based current sensing methods for low and high power circuits are listed in Table 6.1.

Table 6.1. Applications of resistive-based and electromagnetic-based current sensing techniques

Types of circuit	External $R_{sense}$	MOSFET- $R_{DS}$	Current sensing power MOSFET	Filter-based	CTs	Rogowski Coil	Hall-effect-based	Saturable inductor sensors	MR sensor	FOCS
Low power circuits										
• Battery charger					*		**			
• Voltage regulator modules			*	**						
• Electric vehicles							*		**	
• Power factor correction converters			*	**						
• Telecom applications					*					**
• Uninterruptible power supply				*			**			
• Control of power semiconductors						*		**		
High power circuits										
• Traction drive systems						**	*			
• Crusher and grinder motors						**	*			
• Converters for servo-motors					*		**			
• Current measurement for power analysis						**	*			
• Current measurement in busbar sets						**		*		
• High voltage distribution installations					*		**			
• Static converters for dc motor drives							**		*	

Resistive-based and electromagnetic-based current sensing techniques are suggested for particular application based on the criteria such as cost, linearity, sensor's space requirement, mass, operating range, isolation needs, bandwidth, response time, and type of current to be measured. In Table 6.1, '\*\*' indicates the best technique that can be used for particular application and '\*' indicates that one can use the particular technique for that application but it is not recommended. For example, in electric vehicle application, MR sensor is the best current measurement technique, however; one can also use Hall-effect-based sensors but they are not recommended due to their space requirement, weight, cost, and complexity.

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## VITA

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