Design of Feedforward Active Ripple Filters for Power Converters

by

Mingjuan Zhu

Submitted to the Department of Electrical Engineering and Computer Science in partial fulfillment of the requirements for the degree of

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Abstract

An active ripple filter is an electronic circuit which cancels or suppresses the ripple current and EMI generated by the power stage of a power converter, thus reducing the passive filtration requirements. This thesis presents the design and evaluation of various current sensing and injection methods for feedforward active filter applications including a Rogowski-Coil sensor, a transformer sensor, a class A injector, and a current transformer-based injector. The advantages, tradeoffs, and limitation of each approach are discussed in detail. This thesis also presents the design and evaluation of a prototype feedforward active filter which employs a Rogowski-Coil current sensor and a class A current injector on a 42V/14V dc/dc power converter system. Quantitative comparisons between a hybrid passive/active filter and a purely passive filter having the same performance are presented. It is demonstrated that substantial improvements in filter mass and converter transient performance can be achieved using the proposed active ripple filtering method.

Thesis Supervisor: John Kassakian
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Chapter 1

Introduction

1.1 Motivation for Using Active Filtering Techniques

Switching power converters are often designed with filters at the input and output to suppress the ripple components which would otherwise appear at the power converter terminals. Filtration is necessary to maintain high quality input and output waveforms, and to prevent electromagnetic interference (EMI) with other systems. The specific filtration requirements vary depending on the application. In line interfaced applications, the ripple current conducted into the line is specified by organizations such as the Association of German Electrical Engineers (VDE) and the Federal Communications Commission (FCC) [1]. In many military applications, strict standards such as the MIL-STD-461B CEO3 [2] are imposed. In automotive applications, SAE standards, such as SAE J1113/41, apply [3]. For example, Figure 1.1 shows the Class 1 narrowband conducted disturbances limit of SAE J1113/41. In all of these standards, the tight ripple requirements necessitate the use of input and output filters which can account for an appreciable portion of the converter size, weight, and cost.

Traditionally, passive LC low-pass filters have been employed to achieve the desired level of ripple attenuation. Principles for designing such filters are well known [4]. Because of the tight ripple regulations, the cut off frequency must be set very low to achieve the necessary attenuation at the switching frequency and above. Implementing a low cutoff frequency results in bulky and expensive LC filter components which are detrimental to the transient performance of the system. Although higher order passive filters can be used to reduce the size of filter components while achieving the tight ripple
Figure 1.1: Limits for narrowband conducted disturbances on power input terminals (Class 1) SAE J1113/41 JUL 95. The ripple voltage is sensed with a quasi-peak detector across a 50Ω Line Impedance Stabilization Network (LISN).

attenuation requirements, they introduce a higher level of complexity, and can make the power circuit control more difficult.

One alternative approach that has been studied recently is the application of active EMI filters, in which active electronic circuits are used to achieve ripple attenuation [5,7,8,15,19]. In this technique, a passive filter with lower attenuation characteristics is used in conjunction with the active filter. The reduced passive filter is typically used to attenuate the ripple to an intermediate level from which the active filter provides the rest of the required attenuation. The advantages of using this hybrid passive/active filtering approach include reduction in total filter size, weight, and cost. In addition, since the passive filter element values are reduced, much better transient performance can be obtained.

1.2 Introduction to Active Filtering Techniques
Active filters can be classified in a number of ways. Voltage-ripple filters reduce the voltage ripple at a node, while current-ripple filters reduce the ripple current through a circuit
branch. Ripple reduction can be achieved through the use of feedforward or feedback techniques. Feedforward techniques operate by sensing the ripple component and injecting the opposite of that ripple, while feedback techniques suppress ripple through high-gain feedback control.

A simple illustration of the feedforward technique in a current-ripple active filter is shown in figure 1.2. The input port of the active filter is connected to the output of the power converter. The input port of the filter receives ripple voltage from the power converter which causes ripple current to flow from the input port to the output port. The current flowing through $Z_n$ is composed of a dc current $I_{dc}$ and a ripple current $i_{ac}$ as indicated in figure 1.2. The feedforward method senses the ripple current $i_{ac}$ flowing through $Z_n$, then uses a current driver to shunt this ripple current $i_{ac}$ to the ground so that output current $i_{OUT}$ is ripple free and is approximately equal to $I_{dc}$. Ideally, the feedforward gain, $\beta_F$, should be unity at the ripple frequencies for perfect ripple cancellation. This, however, is impossible to achieve due to the limited bandwidth and accuracy of the components involved.

![Figure 1.2: Feedforward active filter.](image-url)
The feedback technique senses the current $i_{\text{OUT}}$ flowing through $Z_q$, and forces the current $I_c$ shunted away from the quiet port to be much larger (by a factor $\beta_b$). This has the effect of amplifying the effective ripple frequency impedance $Z_q$ by the factor $\beta_b$, reducing the ripple current flowing to the output. Similar feedforward and feedback methods can be applied to ripple voltage filters, and hybrids of these types are also possible.

Active filters may be further classified by how the sensing and actuation (injection) functions are implemented [5]. In a current ripple filter, one must sense ripple current and inject a compensating current, and the manner in which these functions are accomplished has a dramatic impact on the performance of the active filter.

1.3 Thesis Objectives
The objectives of this thesis are to explore new and improved methods of implementing feedforward active ripple filters, to design a feedforward active filter based on these methods, and to investigate their effectiveness for ripple filtering in power converters. This thesis presents the design and evaluation of various current sensing and injection methods for feedforward active filter applications. In addition, a prototype active filter which employs
a Rogowski-Coil sensor is constructed and evaluated on a 230 W, 125 kHz buck converter with a 42 V nominal input voltage and a 14 V nominal output voltage.

Chapter 2 of this thesis presents the design and evaluation of a Rogowski-Coil current sensor. The Rogowski-Coil current sensor has never been applied in an active filtering application, yet its wide sensing bandwidth, insensitivity to parameter variation, and low cost core made it potentially very valuable in this application. Chapter 3 discusses the design of a current injector which uses a class A transistor circuit in conjunction with an op amp. Chapter 4 evaluates the prototype active filter with a Rogowski-Coil sensor and a class A injector on a dc/dc power converter. Chapter 4 also investigates the achievable reduction in passive filter elements through use of the active filter. Furthermore, an improvement of the transient performance through use of an active filter is also demonstrated. A phenomenon associated with the employment of feedforward active filters is that it affects the system dynamics. This phenomenon, which has not been previously discussed in the active filter literature, is investigated in Chapter 5. Chapter 6 studies the use of a current transformer sensor, and compares it to the Rogowski-Coil sensor under different current sensing conditions. Finally, Chapter 7 presents an alternative current injection method which makes use of a current transformer, and Chapter 8 draws conclusions and describes directions for the future work in this area.
Chapter 2

The Rogowski-Coil Current Sensor

2.1 Introduction

The Rogowski coil is a current sensor which is most commonly employed for sensing large-magnitude ac or transient currents. The coil consists of a uniform winding on a non-magnetic toroidal core (former) having a constant cross section (Fig. 2.1). When a current to be sensed, \( i_p \), is passed through the toroid, a voltage will be induced on the winding that is proportional to the time derivative of the primary current. This relationship is determined by Faraday’s law:

\[
\nu = \frac{\mu_0 N A_c}{l} \frac{di_p}{dt} = k \frac{di_p}{dt} \tag{2.1}
\]

Where \( \mu_0 \) is the permeability of the former (which is also the permeability of free space), \( N \) is the number of turns, \( A_c \) is the cross-sectional area of each turn and \( l \) is the mean circumference of the former.

Although the sensing method has been known since 1912 [6,10,11], the Rogowski-Coil sensor has not been previously employed in active ripple filtering. Because of its low sensing gain, it has been traditionally employed in applications where large, high frequency currents are being sensed. However, through proper amplification, it is possible to overcome the sensing gain limitation. The resulting sensor has a number of advantages for the active filtering application. For example, the Rogowski coil offers high accuracy and wide bandwidth, which are important in the design of a feedforward ripple filter. Because a non-magnetic core is used, the core does not have a saturation limitation and the size of the sensor does not change with the dc current level. Furthermore, the Rogowski-Coil sen-
Figure 2.1: A Rogowski-Coil. The sense coil is a uniform single layer winding on an air-core former.

The sensor is also simple and inexpensive to implement.

2.2 Designing the Rogowski Coil

To achieve accurate current sensing using the Rogowski coil, several factors need to be kept in mind. First, it is desirable to have evenly distributed coil turns with a constant cross sectional area. Also, the Rogowski coil can be either single layered or multi-layered. A single layer winding, however, gives a lower value of series self-inductance. A low series self-inductance is desirable for accurate current sensing at high frequencies, since a large value will limit the bandwidth of the sensor.

As shown in (2.1) the coil generates a voltage proportional to the derivative of the sensed current. An integrating amplifier stage needs to be used to generate a voltage proportional to the sensed current. One of the difficulties in implementing the Rogowski-Coil current sensor for many applications is the difficulty of designing an ideal integrator. For an ideal integrator op amp circuit, the integral of the constant offset voltage of the op amp will be a ramp, leading to an error that always grows with time which eventually will make the op amp go into saturation. For our application, this troublesome integrator design problem can be avoided since only the switching frequency ripple current needs to be
sensed. Therefore, only for the frequencies (ripple frequency and above) of interest do we need the amplifier stage to act as an integrator.

A challenge with implementing the Rogowski-Coil current sensor in the active filtering application is the low gain of the coil, where gain is defined as output voltage divided by the time derivative of the primary ripple current passing through the core. Given a maximum practical ratio of the toroid cross section to circumference \( (A_c/l) \), the only methods to increase the gain of the coil are to increase its size, to increase the number of turns on the coil, or to increase the number of times the sensed current is passed through the toroid.

The size of the coil and the number of times the sensed current is passed through it are limited by the desire to have a compact, inexpensive sensor. Increasing the number of turns on the coil increases the series self inductance, which limits the high frequency performance of the coil. To understand this, consider a Thevenin equivalent of the Rogowski coil consisting of an induced voltage source in series with the self inductance of the Rogowski coil, as shown in figure 2.2. The capacitor represents the input impedance of the amplifier stage that follows the coil. The resulting LC circuit is essentially a second order low pass filter where the undamped resonant frequency (also the cut-off frequency) is given by equation 2.2:

\[
\omega_n = \frac{1}{\sqrt{LC}}
\]  

(2.2)

If the series inductance is too large, then the cut-off frequency will be reduced, thus affecting the phase and eventually the magnitude of the sensed signal. Therefore, to achieve good high-frequency performance, there is a limit on the acceptable coil output inductance, and hence on the number of coil turns used. The resistor, \( R \), is a damping resistor that is required to keep the high-frequency impedance at the coil output node low (for con-
Figure 2.2: Rogowski coil impedance model including the amplifier input impedance.

verter noise rejection) and to prevent the coil inductance from resonating with the amplifier input. The selection of $R$ involves a trade-off between the rise time and overshoot (stability) of the system.

All these factors result in trade-offs among the size, gain, and bandwidth of the Rogowski-Coil sensor. For these reasons, the Rogowski coil usually is employed for sensing large currents. To apply it in this current ripple suppression application, a substantial amount of amplification is required with high accuracy and low phase shift at high frequencies. The amplifier design becomes a critical part of implementing the Rogowski-Coil current sensor.

After considering the design trade offs, a prototype Rogowski coil was constructed using a wooden toroid core. The inner diameter is 18 mm, the outer diameter is 32 mm, and the height is 13 mm. It was wound with 88 turns of 31 gauge wire, and the current to be sensed was passed through the coil 5 times, yielding a total calculated gain, $k$, of 0.64 V-μs/A. The coil with a self inductance of 14 μH was terminated with an amplifier input resistance of 1 kΩ, yielding acceptable bandwidth and damping. This prototype Rogowski coil achieves good sensing gain and bandwidth for ripple currents at the 100 mA level in the 100+ kHz range.
2.3 Amplifier Design for the Rogowski-Coil Sensor

2.3.1 Introduction
The amplifier circuit for the Rogowski-Coil current sensor serves two roles. The first role is to integrate the voltage sensed across the coil terminals so that the output voltage is directly proportional to the sensed current rather than its derivative. Second, the amplifier provides substantial gain in order to compensate for the low gain of the Rogowski coil.

The integrating amplifier frequency shapes (filters) the input signal with a gain that drops 20 dB/decade over the frequency range of interest. Furthermore, the amplifier must provide substantial gain (at the fundamental) and have very low deviations in phase over the whole frequency range of interest. This is difficult to achieve using a single stage op amp due to gain and bandwidth limitations. A solution to this problem is to distribute the gain among several stages. For this reason, a four stage amplifier circuit which includes three gain stages and one integrator stage is proposed.

2.3.2 Designing the Amplifier Stages
Figure 2.3 shows a complete schematic of the amplifier circuit. Sallen-Key op amp circuit structures [12] are utilized for all four amplifier stages. The parameter values for the capacitors and resistors are listed in Tables 2.1-2.4. The amplifier circuit includes one gain stage and two high-pass filter stages, each of which has a gain of 4.7 at the ripple frequency. Besides providing gain, the first op amp circuit also serves as a buffer stage. \( R_f \) is a damping resistor that is used to attenuate the resonance caused by the coil series inductance and the amplifier input impedance as discussed in previous section. The cut-off frequencies of the high pass filters should be set at least one decade below the sensed ripple current frequency to ensure enough room for the phase transition. Since this current sensor is designed for a power converter that has a fundamental switching frequency of 125 kHz,
Figure 2.3: Amplifier for the Rogowski-Coil current sensor. The circuit is based on an LT 1230 quad current feedback amplifier.

<table>
<thead>
<tr>
<th>$R_I$</th>
<th>$R_{a1}$</th>
<th>$R_{b1}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 kΩ</td>
<td>220 Ω</td>
<td>820 Ω</td>
</tr>
</tbody>
</table>

Table 2.1: Gain stage with a gain of 4.7.

<table>
<thead>
<tr>
<th>$R_2$</th>
<th>$R_{a2}$</th>
<th>$R_{b2}$</th>
<th>$C_1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 kΩ</td>
<td>220 Ω</td>
<td>820 Ω</td>
<td>0.2 μF</td>
</tr>
</tbody>
</table>

Table 2.2: First high-pass filter with a cut-off at 1 kHz and a gain of 4.7 at 125 kHz.

<table>
<thead>
<tr>
<th>$R_3$</th>
<th>$R_4$</th>
<th>$R_5$</th>
<th>$R_x$</th>
<th>$R_y$</th>
<th>$C_2$</th>
<th>$C_3$</th>
</tr>
</thead>
<tbody>
<tr>
<td>750 Ω</td>
<td>750 Ω</td>
<td>4.3 Ω</td>
<td>200 Ω</td>
<td>4.2 kΩ</td>
<td>0.2 μF</td>
<td>0.2 μF</td>
</tr>
</tbody>
</table>

Table 2.3: Integrator stage with double pole at 1 kHz and a gain of 0.16 at 125 kHz.

<table>
<thead>
<tr>
<th>$R_6$</th>
<th>$R_{a3}$</th>
<th>$R_{b3}$</th>
<th>$C_4$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 kΩ</td>
<td>220 Ω</td>
<td>820 Ω</td>
<td>0.2 μF</td>
</tr>
</tbody>
</table>

Table 2.4: Last high-pass filter with a cut-off at 1 kHz and a gain of 4.7 at 125 kHz.
the cut-off frequencies for these two high-pass filter are set to 1 kHz, two decades below the ripple current frequency. This attenuates the low (power) frequency components of the signal while ensuring that the phase transition will be completed at the switching frequency of the power converter.

The integrator stage is essentially a band pass filter that acts as a differentiator with a slope of 20 dB/dec below 1 kHz and functions as an integrator with a slope of -20 dB/dec above 1 kHz (and a gain of 0.16 at 100 kHz). This is acceptable in this application because we are only interested in sensing the ripple current of the converter at the frequencies of interest (125 kHz and above). The band pass filter shape suits this purpose well by providing an integrating function after the second cut-off frequency at 1 kHz, while attenuating low-frequency components. Moreover, the 1 kHz cut-off frequency is chosen so that it is two decades below the switching ripple frequency; this provides enough room for phase transition, while maintaining a sufficient gain at the fundamental switching frequency.

The arrangement of these four stages is not arbitrary. Rather, it is arranged to optimize the noise rejection and bandwidth performance of the sensor. The buffer-gain stage has to be placed first to keep the integrity of the sensed voltage signal from the coil. The other 3 stages can be arranged in three different ways: (high-pass, high-pass, integrator); (high-pass, integrator, high-pass); and (integrator, high-pass, high-pass). If the first arrangement is chosen, the bandwidth performance will be limited, due to limitations in the op amp output voltage. To understand this, recall that the Rogowski coil is essentially a differentiator, and thus its output voltage versus input current transfer function is a ramp in the frequency domain. This means that for a constant magnitude input signal, the output voltage increases as function of frequency. This voltage signal will be further amplified by the buffer-gain and the two high-pass-gain filter stages. As the input frequency increases, the voltage amplitude will surpass the output voltage range of the op amp, and the op amp will
saturate. To prevent the op amp from entering saturation, one of the other amplifier arrangement should be used. Another option is to move the integrator stage as far forward as possible. The integrator reduces the magnitude of the high frequency components, and compensating for the differentiating characteristic of the coil allows much higher frequency signals to be handled before the op amp saturates. However, this arrangement (integrator, high pass, high pass) also has drawbacks. Because the two high-pass gain filters are stacked next to each other, any high frequency noise injected in the first high-pass gain filter will be amplified as $4.7^2$. Therefore, to provide the best trade off between noise immunity and bandwidth, a better selection is to place the integrator after the first high-pass gain filter. This is precisely the amplifier arrangement shown in figure 2.3.

2.3.3 Selection of the Operational Amplifier
To achieve a high degree of ripple cancellation, accurate sensing of both the magnitude and the phase of the ripple current are necessary. This generally requires high bandwidth and high slew rate op amps. In the active filtering application, the desired sensing bandwidth can be up into the MHz range. Current feedback operational amplifiers [13] allow a better trade-off between gain and bandwidth than traditional voltage feedback op amps. In conventional voltage-feedback op amps, the ideal closed-loop gain $G$ enters the denominator of the closed-loop gain transfer function, and this makes the closed-loop bandwidth inversely related to the closed-loop gain of the op amp. In current feedback op amps, however, only the feedback resistor enters the denominator of the transfer function and affects the system pole locations. The ideal closed-loop gain $G$ is independent of the bandwidth, and therefore current feedback op amps offer very high gain-bandwidth product and high slew rate.
Although a current feedback op amp offers excellent bandwidth and slew rate, it is also very sensitive to the parasitics of the circuit layout. In most cases, careful PC-board layout is crucial to achieving good performance. Furthermore, many current feedback op amps are not well suited to driving capacitive loads. When driving a capacitive load, a resistor can be placed in series with the capacitive load to improve the dynamics. The components used should also have low parasitics and the op amp needs to be carefully bypassed using capacitors.

Slew rate (the maximum rate at which the output voltage can change) is also an important factor in choosing an op amp. The upper cut-off frequency from the manufacturers specification usually refers to an output voltage swing less than ± 1 V; when the output voltage swing is high, the upper cut-off frequency will be lowered. As a result, the op amp slew rate should be carefully chosen so that it is not a limiting factor in performance.

To achieve high bandwidth, a single LT1230 quad current feedback op amp IC was selected to implement the amplifier. This quad op amp has wide bandwidth (100 MHz) and high slew rate (1000 V/μs) at a low cost. The gain and two high pass filter stages, each have a passband gain of 4.7, and the integrator has a gain of 0.16 at 125 kHz. This results in an overall sensor midband gain of 6.6 V/A at the fundamental ripple current frequency (125 kHz).

2.4 Evaluation of the Current Sensor
The prototype sensor circuit is tested using an ac voltage function generator as shown in figure 2.4. The output terminal of the signal generator is connected to a resistor which is placed in the Rogowski coil to create a proportional ripple current. (The wire from the signal generator is wound 5 times on the Rogowski coil to provide the gain.) Figure 2.5 shows the scope measurement of the amplifier output voltage (v) as well as the input ripple
current \((i)\) at 125 kHz. Channel 4 displays the ripple current from the signal generator, and Channel 2 displays the ripple voltage at the output of the amplifier. The vertical scale is 50 mA/div for Channel 4 and 500 mV/div for Channel 2. The horizontal scale is 2.5 \(\mu\)s/div. This plot clearly shows that the output voltage tracks the input current waveform accurately with low phase shift at the desired fundamental frequency. The overall gain of this sensor is 6.6 V/A. From figure 2.5 we can also deduce gain of the Rogowski coil by dividing the overall gain of the sensor by the gain of the op amp circuit. The measured Rogowski coil gain \(k\) is 0.66 V-\(\mu\)s/A. The gain calculated using (2.1) is 0.64 V-\(\mu\)s/A, which is very close to the measured value.

The frequency performance of the current sensor was also tested using a network analyzer. Both the magnitude and the phase of the output voltage versus input current frequency are plotted in figure 2.6. The starting frequency is 10 kHz and the upper frequency is 5 MHz. From the plot we can conclude that the current sensor has a useful bandwidth from 20 kHz up to about 1 MHz. At 1 MHz the phase distortion becomes significant, and this phase shift will affect the filter performance. (For a sine wave, when the phase shift between the sensed signal to the injected signal is 5 degrees, a factor of 10 X ripple reduction can be achieved. However, if the phase shift increase to 10 degrees, only a factor of
Figure 2.5: Output waveform for the Rogowski-Coil sensor (top waveform as compared to the input ripple current measured by a current probe (bottom waveform).

Figure 2.6: Frequency response of the Rogowski-Coil sensor measured using Network analyzer. The top curve is the magnitude plot and the bottom curve is the phase plot. (10 dBmV/div, 10 deg/div)
5 X ripple reduction can be achieved.)

In this case, the upper cut-off frequency of this current sensor is limited by the low pass filter formed by the Rogowski coil inductance and the coupling capacitances before the amplifier stage. The damped resonant frequency of this low-pass filter is approximately 10 MHz, this means that the phase distortion of this low-pass filter begins to affect the phase of the sensor in the lower MHz range as shown in figure 2.6.

2.5 Conclusion
This chapter has demonstrated an effective ripple current sensor approach through use of a Rogowski coil. By using IC amplifiers, we are able to adapt this current sensing technique to the active filter application. The Rogowski-Coil current sensor can provide a useful bandwidth up into the MHz range. Its bandwidth is limited by the coil series inductance which forms a low-pass filter with the input impedance of the op amp. Reducing the number of turns on the coil can reduce the series self inductance, however, the gain of the coil is also reduced. The easiest way to increase the coil gain is to increase the number of turns of primary (to be sensed) current passed through the coil.

Overall, the Rogowski coil provides accurate and wide bandwidth current sensing. Because a non-magnetic core is used to wind the coil, the sensor is insensitive to parameter variations such as temperature. Furthermore, the gain of the Coil is independent of the dc current level of the sensed current, which makes it particularly advantageous in this application. A detailed study comparing the current transformer sensor and the Rogowski-Coil sensor is presented in Chapter 5.
Chapter 3

Current Injector

3.1 Introduction

The second major component of the feedforward active filter of figure 1.2 is the current injector. The current injector is designed to inject an exact copy of the sensed ripple current (with opposite polarity) at the active filter output node to cancel the ripple current. In order to achieve good ripple cancellation, the injector circuit must have high accuracy, wide bandwidth, and low phase shift, just as the current sensor must. To achieve this requirement, a dissipative linear injection circuit is usually used. Nevertheless, it should be pointed out that while the injected ripple currents may be large, the power injected into the output may actually be very small, since the ripple voltage seen at the injection point may be quite small. An ideal current injector would only inject an ac current into the output while only overcoming a small ac voltage ripple to do so, thus achieving low dissipation.

One possible approach to implementing a current injector is to use a linear amplifier along with a transformer and a coupling capacitor. The coupling capacitor which is in series with one of the windings blocks the dc voltage, while the transformer matches the low-voltage high-current output requirement to the high-voltage low-current capabilities of the linear amplifier. This method has been previously used in active filter applications [8,15]. It has the advantage of allowing high injection currents to be achieved at low power consumption, since only the ac current is being injected and only ac voltage is presented across the transformer. The drawbacks to this approach are the cost and volume of the transformer and coupling capacitor, and the performance limitations introduced by their parasitics. A more detailed study of the current transformer injector approach will be presented in Chapter 7.
In this chapter, an alternative injection approach using an inexpensive class-A injection circuit is presented. The advantage to this approach is its simplicity, accuracy, and high-bandwidth performance.

### 3.2 The class A Current Injector Circuit

A simplified schematic of the class A injector circuit is shown in figure 3.1. It is called a class A circuit since the output transistor conducts over the full cycle of a sine wave [14]. Because the output transistor conducts over the full cycle, it must carry a dc bias current greater than the peak ac injection current. The dc bias current times the dc injection point voltage represents a bias power loss that is substantially larger than the power loss due to the ac injection waveform. The class A injector is thus much less efficient than other possible structures, and it is the bias current power loss which fundamentally limits the maximum ac current injection.

Despite its inefficiency, the class A circuit does have the advantage of simplicity. To see how it operates, consider the schematic shown in figure 3.1. The voltage \( v_{ac} \) is the output voltage from the current sensor and is proportional to the ripple current to be injected. \( V_{bias} \) is a positive dc voltage proportional to the bias current that ensures the output transistor is always operating in the forward active region. By superposition, the non-inverting terminal has a voltage that is half of the sum of \( v_{ac} \) and \( V_{bias} \), that is:

\[
V_- = V_+ = \frac{v_{ac} + V_{bias}}{2}
\]  

(3.1)

For an ideal op amp, the inverting terminal \( V_- \) has the same voltage potential as the non-inverting terminal \( V_+ \), and the current flowing into the \( V_- \) terminal is zero. Therefore, by selecting the resistor \( R_c \), we can create a collector current that is proportional to \( v_{ac} \). This relation is shown in equation 3.2:
Figure 3.1: A simplified schematic of the class A injector circuit.

\[ i_c = \frac{\beta}{2(\beta + 1)R_e}(v_{ac} + V_{bias}) \]  \hspace{1cm} (3.2)

Therefore, for large values of $\beta$, we can generate an injection current proportional to $v_{ac}$.

Implementing a practical injector requires additional components. The complete schematic for the prototype injector circuit is shown in figure 3.2. Voltage $v_{ac}$ is the output waveform from the current sensor. $C_5$ is placed between $R_9$ and $v_{ac}$ to block dc (low frequency) components of $v_{ac}$. The bias voltage $V_{bias}$ is generated by using resistor $R_7$, a 1 kΩ potentiometer, and a zener diode $D_4$. The potentiometer allows tuning of the bias voltage, and $R_7$ provides a partial voltage drop down from the logic supply voltage $V_{p+}$. The purpose of the zener diode ($D_4$) is to limit the maximum value of $V_{bias}$, which in this case is 3.6 V. Capacitor $C_6$ helps attenuate high frequency noise in the bias voltage. Resistors $R_{10}$ and $R_{11}$ set the desired voltage to current gain to be the inverse of the current sensor gain.

The collector of the transistor is connected to the injection node of the filter via the parallel connection of a zener diode $D_5$ and ceramic capacitor $C_7$. These elements provide
**Figure 3.2:** Class A current injector circuit. $R_7=R_8=R_9=1\,\text{k}\Omega$, $R_{I0}=3.3\,\Omega$, $R_{I1}=50\,\Omega$, $C_5=C_6=C_7=0.1\,\mu\text{F}$, $D_1$ (1N4143), $D_2$ (BYV10-40), $D_3$ (1N4148TR), $D_4$ (BZX85C3V6), $D_5$ (2 BZX85C5V1 in series), LM6361 op amp, ZTX 649 transistor.

A low impedance voltage drop of 10.2 V which lowers the transistor dissipation while maintaining high bandwidth. An alternative connection that can be used is shown in figure 3.3. In this connection, bias current is drawn from the logic supply, while ac currents are injected into the output node. This can yield efficiency improvements when injecting current at a high-voltage node. In the case considered here, however, the two methods yield equal efficiency, since the logic supply voltage is generated from the output voltage via a linear regulator.

There are also several additional components used for circuit protection. Diodes $D_1$ and $D_2$ are put in series at the $V_+$ node to set the maximum commanded current. The purpose of $D_3$ is to limit a reverse bias voltage $V_{EB}$ to no more than 0.6 during transient operation in order to protect the transistor.
3.3 Conclusion

The class A current injector circuit has the advantage of extreme simplicity, and accurate, high-bandwidth current injection is easily achieved. The major drawback of the class A injector circuit is its low efficiency relative to other possible methods. The efficiency suffers because a dc bias current must be carried to keep the output transistor continuously operating in the forward active region, and this bias current results in significant losses. Nevertheless, when the dc voltage at the injection point is low, the injector can be directly coupled to the filter output without undue losses. For example, in the prototype system developed here, a 125 mA bias current drawn directly from the 14 V injection node causes only 1.75 W of dissipation, which is less than 1% of converter output power. Because of the simplicity of this approach, we have implemented a class A output injector in the prototype active filter.
Chapter 4

Evaluating the Feedforward Active Filter

4.1 Introduction
This chapter presents an experimental evaluation of the feedforward active filter with a Rogowski-Coil sensor and a class A injector. A prototype active filter of the design described in chapters 2 and 3 has been constructed on a pc-board (see Figure 4.1). We now describe an evaluation of this active filter on a prototype power converter. This evaluation has two goals. First, we want to test the performance of this active filter and determine how much ripple current it can cancel at the fundamental switching frequency of the power converter and at higher harmonic frequencies. Second, we want to determine how much this active filter can help to reduce the size, the weight, and the cost of the EMI filter as compared to a conventional filter. Finally, we want to establish how using an active filtering approach affects the converter transient performance.

Figure 4.1: Prototype active filter with a Rogowski-Coil sensor and a class A injector.
Having these goals in mind, a systematic evaluation procedure is outlined. The first step is to study the filter design with feedforward active elements. Specifically, the passive filter in the power converter needs to be designed to work with the feedforward active filter. This passive filter will be small in size since it only needs to attenuate the ripple current to a certain degree and leave the rest of the filtration job to the active filter. To avoid confusion in later discussion, we name this small-sized passive filter to be the SP (small passive) filter. After designing this filter, an EMI measurement of the power converter with just this SP filter is made. This measurement serves as a reference for later measurements. Then, the active filter elements are added to the power converter with this SP filter. The filter system with both the SP filter and the feedforward active filter is known as the hybrid passive/active filter. Another EMI measurement with this hybrid passive/active filter is then made. The results can be compared with the previous measurement in which only the SP filter is present. This comparison will show how effective the active filter is by displaying the additional current ripple reduction achieved.

The next goal is to find out how much this active filter can reduce the size, weight, and cost of the filter as compared to a conventional passive filter. To achieve this goal, another entirely passive filter that achieves the same ripple specification as the hybrid passive/active filter is designed and built. This passive filter is much larger in size because of the strict ripple reduction requirement. We name this big passive filter the BP filter to differentiate it from the SP filter. Again, an EMI measurement is performed to compare the filtration results of this BP filter with the hybrid passive/active filter. A size, weight, and cost comparison is then made between this BP filter and the hybrid passive/active filter.

The final goal is to establish how the use of active filtering affects the converter transient performance. To do this, the converter output voltage response to load step transients is measured with both the hybrid passive/active filter and the BP filter. A performance
**Figure 4.2:** Power converter model with hybrid passive/active filter.

<table>
<thead>
<tr>
<th>Filter Elements</th>
<th>Values</th>
<th>Parts</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1$</td>
<td>1.4 μH</td>
<td>Rm-10 ferrite core (3F3 material) with 3 turns of 10 gauge wire</td>
</tr>
<tr>
<td>$L_2$</td>
<td>51 μH</td>
<td>T37-52 powder core wound 44 turns with gauge 30 wire</td>
</tr>
<tr>
<td>$L_3$</td>
<td>9 μH</td>
<td>T157-40 powder core wound 15 turns with gauge 10 wire</td>
</tr>
<tr>
<td>$C_1$</td>
<td>20 μF</td>
<td>935C 1W20K film capacitor</td>
</tr>
<tr>
<td>$C_2$</td>
<td>5 μF</td>
<td>ECS-FIVE 105K (five 1 μF parallel connected)</td>
</tr>
<tr>
<td>$R$</td>
<td>0.56 Ω</td>
<td></td>
</tr>
<tr>
<td>$C_d$</td>
<td>47 μF</td>
<td>ECS-FIVE 106K (five 10 μF parallel connected)</td>
</tr>
<tr>
<td>$R_d$</td>
<td>3.3 Ω</td>
<td></td>
</tr>
</tbody>
</table>

**Table 4.1:** Converter and SP filter parameters.

comparison is then made using these responses.

**4.2 Design of the Hybrid Passive/Active EMI Filter**

The designed hybrid passive/active EMI filter will be applied to the output of a 230 W, 125 kHz buck converter with a 42 V nominal input voltage and a 14 V nominal output voltage.

A model of this single cell 42V/14V power converter with hybrid passive/active filters is
shown in Figure 4.2. The converter operates in discontinuous conduction mode with a buck inductance \( L_f \) of 1.4 \( \mu \text{F} \). Inductors and capacitors \( L_2, L_3, R, C_1, \) and \( C_2 \) are output filter elements. (Resistor, \( R_d \), and capacitor, \( C_d \), form a damping leg for system stability compensation and will be discussed in detail in Chapter 5.) Since the active filter is designed to sense and inject ripple currents on the order of 100 mA at the fundamental switching frequency (125 kHz), these passive filter elements are sized to allow 200 mA peak to peak ripple current \( i_{ac} \) in the sensed branch of the filter. This SP filter has a quality factor \( Q \) of 2.44 and the values for the filter elements are shown in table 4.1.

The current flowing through inductors \( L_2 \) and \( L_3 \) total to 200 mA of peak to peak ripple current and 17 A of dc current. Both inductors act as short circuits for dc current. Since inductor \( L_2 \) is in series with resistor \( R \) and hence represents a higher dc impedance path, most of the dc current will pass through \( L_3 \) instead. The 17 A dc current is quite large, so \( L_3 \) must be sized to meet the energy storage requirement. The energy storage requirement \( W_m \) is given by (4.1), where \( I \) is the peak branch current.

\[
W_m = \frac{1}{2} L I^2
\]  

(4.1)

For ripple frequencies, the current will split unevenly between \( L_2 \) and \( L_3 \). Since the ripple current has an insignificant amplitude compare to the dc current, the energy storage requirement for \( L_2 \) will be much smaller. As a result, \( L_3 \) is by far the dominant inductor in terms of size and cost.

From the energy calculation, a Micrometals powder core T157-40 is selected for winding \( L_3 \). The number of turns is determined using (4.2), where \( A_L \) is inductance rating, and percentage \( \mu \) is the operating point of initial permeability. This data can be found in the Micrometals catalog [16]. From (4.2), 15 turns of gauge 10 wire is used to wind this
The wire size is picked so that the current density will not exceed the limit 500A/cm² [17]. Also, the wire will need to fit into the window area of the core. For more information on inductor design, a detailed inductor design handbook is available. [18]

\[ N = \left[ \frac{desiredL(nH)}{(A_L)(percentage\mu)} \right] \] (4.2)

Similarly, the energy storage requirement for \( L_2 \) can be calculated by using (4.1). A much smaller core is required for \( L_2 \) since it only needs to handle 33 mA of the ac current. A T37-52 powder core from Micrometal is selected for making \( L_2 \). The core dimensions for both \( L_2 \) and \( L_3 \) are listed in table 4.2. From this table, we can see that the T37 core has a much smaller volume compared to the T157 core. \( L_2 \) requires 44 turns of gauge 30 wire. Because the winding resistance of the \( L_2 \) is 0.354 ohms, only 0.2 \( \Omega \) of additional resistance is needed to produce a net resistance of 0.56 \( \Omega \).

After constructing this SP filter, the next step is to perform EMI measurements both without and with the active filter elements and compare the results. The active filter is placed as indicated in figure 4.2. The sensor detects the current \( i_{ac} \) and shunts the ripple current to ground in parallel with \( C_2 \) and the load.

<table>
<thead>
<tr>
<th></th>
<th>T37</th>
<th>T157</th>
</tr>
</thead>
<tbody>
<tr>
<td>Out Diameter (in)</td>
<td>0.375</td>
<td>1.57</td>
</tr>
<tr>
<td>Inner Diameter (in)</td>
<td>0.205</td>
<td>0.95</td>
</tr>
<tr>
<td>Height (in)</td>
<td>0.128</td>
<td>0.57</td>
</tr>
</tbody>
</table>

Table 4.2: Powder Core Dimensions.

4.3 EMI Measurements of the SP and Hybrid Passive/Active EMI Filters

The EMI measurements are performed using standard techniques with Line Impedance Stabilization Networks (LISNs) at the input and output of the system under test. (see Figure 4.3). A simple model of the LISN is shown at the output of the converter. For ac fre-
quencies, the inductor of the LISN is approximately an open circuit and the capacitor of the LISN is approximately a short circuit. Therefore, most of the dc current is delivered to the load and most of the ac current flows through the 50 Ω resistor of the LISN. The ripple voltage across this 50 Ω resistor is the standard metric in conducted EMI specifications.

To analyze the ripple cancellation performance of the active filter, two sets of measurements are performed. The first set of measurements is taken by using an oscilloscope to measure the ripple voltage across the 50 Ω LISN resistor with both the SP and the hybrid passive/active filters. Figures 4.4 and 4.5 show the dramatic improvement in ripple attenuation achieved at the switching frequency by using the active filter. Figure 4.4 shows the ripple voltage with just the SP filter and Figure 4.5 shows the measurement with the active filter turned on. The vertical scale and horizontal scale for both plots are 20 mV/div and 5 μs/div. From these plots, we can see when only the SP filter is present, the ripple voltage across the LISN is about 50 mV peak to peak. However, figure 4.5 shows that this fundamental switching ripple is greatly attenuated to 10 mV when the active filter element is added.

The next set of measurements is performed using a spectrum analyzer, allowing the effects of the active filter to be evaluated over a wide frequency range. Figure 4.6 shows the plot of the LISN voltage spectrum both with and without active cancellation. The dashed-line curve is the voltage spectrum with only the SP filter and the solid-line curve is the voltage spectrum with the active filter turned on. From this plot we can again confirm that the active filter achieves a great ripple reduction at the fundamental switching frequency (125 kHz). Without the active filter, the ripple voltage at the 125 kHz is 84.11 dBμV. However, with the active filter, the ripple voltage is reduced to 53.3 dBμV, an attenuation of over 30 dB. At higher harmonic frequencies however, the active filter does not have a significant ripple attenuation. This is not because of a bandwidth limitation.
Figure 4.3: EMI measurement set up for testing the active filter. (The load is 1 Ω resistor).

Figure 4.4: Voltage across the LISN using only the SP filter. (20 mV/div)

Figure 4.5: Voltage ripple across the LISN using both the SP and the active filters. (20 mV/div)
Figure 4.6: Frequency spectrum measurement, SP filter and passive/active filter.

The sensor/injector pair has an effective bandwidth that exceeds 1 MHz but because the amplitude of the higher harmonics are down at the noise floor of the current sensor. As a result, the active filter cannot effectively attenuate these harmonics. Nevertheless, the active filter is advantageous because it substantially attenuates the fundamental, allowing the converter to meet a strict ripple specification (of 64 dBμV) across frequency with only small passive filter elements.

4.4 Design of the BP filter
The next evaluation procedure is to design a BP filter which meets the same flat ripple specification as the hybrid passive/active filter, and to compare the two filters. Through this procedure, we can compare and evaluate how much the hybrid passive/active filter approach can reduce the size, weight, and cost of the EMI filter.

For this BP filter design, the amount of ripple current flowing through the inductors is set to be 10 mA. The design approach is to keep $C_1$, $C_2$ the same, and adjust $L_2$, $L_3$, and $R$
so that the desired attenuation is achieved and the filter $Q$ is 2. The resulting element values are shown in Table 4.3.

<table>
<thead>
<tr>
<th>Filter element</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_2$</td>
<td>225 $\mu$H</td>
</tr>
<tr>
<td>$L_3$</td>
<td>576 $\mu$H</td>
</tr>
<tr>
<td>$C_1$</td>
<td>5 $\mu$F</td>
</tr>
<tr>
<td>$C_2$</td>
<td>20 $\mu$F</td>
</tr>
<tr>
<td>$R$</td>
<td>1.875 $\Omega$</td>
</tr>
</tbody>
</table>

**Table 4.3: Big passive filter parameter values.**

In this design, both capacitors are unchanged, but the inductors and the resistor values are updated. Again, the physical size of the inductors are determined by how much energy they need to store. Using (4.1), we can calculate the energy storage requirement for $L_3$, which is $32512$ $\mu$J. This energy storage requirement is over 10 times greater than the $L_3$ design for the SP filter. To have a fair comparison, the same core material (40) is selected. In this case, the core size increases to T400D. Similarly, inductor $L_2$ is designed by calculating how much peak energy it needs to store. In this case, the same core that was used for the SP filter is selected. The number of the turns for $L_2$ is 149, and gauge 32 wire is used to wind the inductor.

Here we measure the EMI performance of the BP filter, and compare the results to the hybrid passive/active filter measurement of the previous section (see Figure 4.7). At 125 kHz, the hybrid passive/active filter has a better attenuation than the BP filter by 10 dB. At higher frequencies however, the BP filter achieves better ripple attenuation. This is because the BP has to be over designed at high frequencies in order to achieve sufficient ripple attenuation at the fundamental frequency. Ultimately, both the hybrid/active filter and the BP filter meet the same flat ripple specification (64 dB$\mu$V) across frequency.
Figure 4.7: Ripple voltage spectrum ($R_L=1\Omega$), hybrid passive/active filter and BP filter.

Comparing the BP filter with the SP filter used in the hybrid passive/active filter, the only filter element size that has been altered is inductor $L_3$. $L_3$ is the largest inductor used in both the BP filter and the hybrid/active filter. Figure 4.8 shows a size comparison between these two inductors. The quarter on the left hand side of the plot serves as a size reference. Clearly from this picture we can see that $L_3$ used for the BP filter is significantly larger than $L_3$ used for the SP filter. Specifically, the core used for the BP filter inductor has dimension of (102mm/57.2mm/33mm) which is much larger than the core used for the hybrid passive/active filter (39.9mm/24.1mm/14.5mm). A direct consequence of this large inductor is that it is much heavier. To compare the mass of the BP filter and the hybrid passive/active filter, weights of both filters are measured. The results show that the hybrid passive/active filter has a total mass that is one fifth that of the BP filter! Thus the overall filter size and mass are substantially reduced using the hybrid passive/active filter approach.
Figure 4.8: Inductors size comparison.

4.5 Transient Performance with the Hybrid Passive/Active Filter
In the last section we showed that using the active filter can reduce the component value in the passive filter inductor (and the energy it stores) by a substantial amount. Intuition tells us that this reduction in inductance and energy storage will result in a better transient response for load changes. The filter cut-off frequency is inversely proportional to the square root of the inductance, so a smaller inductance means a faster settling time. In this section, the load step transient responses of the converter will be studied using both the BP filter and the hybrid passive/active filter. The measurement should confirm our expectation that the hybrid passive/active filter has a better transient performance with a faster settling time.
To set up the transient performance measurement, an electronic load is used to provide a step change in load resistance values. The lower bound of the load is 1.5 Ω and the upper bound is 3 Ω. The slew rate is specified in current: 5 A/μs. Figures 4.9 and 4.10 show the output voltage transient response with the BP filter and the hybrid passive/active filter respectively, when the load value steps from 3 Ω to 1.5 Ω. The vertical scale for both plots is 2 V/div and the horizontal scale is 100 μs/div. Clearly the hybrid filter has a much better transient performance, with only half of the voltage drop of the BP filter and a far shorter settling time. This transient measurement confirms our expectation and demonstrates the dynamic control advantages that can be obtained using the active filtering approach.

Figure 4.9: Output voltage transient response of the converter for a load step from 3 Ω to 1.5 Ω with the BP filter.
Figure 4.10: Output voltage transient response of the converter for a load step from 3 Ω to 1.5 Ω with the hybrid passive/active filter. The transient magnitude and duration are far smaller with this hybrid filter compared to the entirely passive BP filter.

4.6 Conclusion

The evaluation carried out in this chapter demonstrates that feedforward active cancellation can be used to achieve substantial ripple reduction in the EMI filter. In the example considered here, the fundamental component is reduced by over 30 dB. At higher frequencies, the harmonic amplitudes are small enough that the active filter cannot attenuate them, and at even higher frequencies the bandwidth of the active filter is exceeded and it has no effect. For EMI specifications that are flat across frequency, the hybrid passive/active filter approach is especially well-suited for ripple attenuation with a small filter. The moderate sized passive filter functions well at high frequencies while the active filter helps meet the low frequency requirement. In the example considered here, the mass of a hybrid active filter is only one fifth of that of an entirely passive filter meeting the same flat EMI specifi-
cation. However, in cases where the EMI limit becomes substantially tighter at higher frequencies, the hybrid passive/active filter becomes less advantageous. A larger passive filter will be necessary to filter out those harmonics.

Another advantage of using the hybrid passive/active filter is the improvement of the system transient performance. The introduction of an active filter reduces the passive filter element sizes, (in this case the major energy storage inductor is reduced), and the smaller passive elements result in better transient performance.
Chapter 5

Active Filter Effects on System Dynamics

5.1 Introduction
From a control point of view, use of a feedforward technique implies that information is sent forward; therefore, it should only affect the locations of zeros but not the poles of the system. However, this is only true if the variable that is fed forward is independent of the system state variables that are affected by the feedforward. Thus, depending on how it is used in a circuit, a “feedforward” active filter can substantially affect the filter dynamics.

Figures 5.1 and 5.2 show two feedforward active filters that sense and inject currents at different points. System A senses and injects the current \( i_x \) (a dependent variable) while System B senses and feeds forward the independent input current \( i_s \). We can verify that one of these filters will affect the system dynamics and the other will not by writing down their transfer functions. The transfer function \( \frac{I_y}{I_s} \) for System A is given by (5.1) and for System B is given by (5.2).

![Diagram of System A](image)

**Figure 5.1:** System A: feedforward of a dependent current.

\[
\frac{I_y}{I_s} = \frac{Z_1(1-K)}{Z_1 + Z_2 + (1-K)Z_3}
\]  

(5.1)
Figure 5.2: System B: sensing an independent input current.

\[
\frac{I_y}{I_s} = \frac{(1 - K)Z_1}{Z_1 + Z_2 + Z_3}
\]  

(5.2)

From the transfer functions, we see the expected results—the feedforward gain \( K \) in System A changes the pole and zero locations while in System B it only affects the zeros of the system and leaves the poles unchanged. We can also see why the System A pole locations are changed and the System B pole locations are not. Figures 5.3 and 5.4 are block diagram representations for the transfer functions in (5.1) and (5.2), respectively. It can be observed from these diagrams that the feedforward in System A changes the gain in the feedback loop, while feedforward in System B does not. Thus, the key to determining if the system dynamics will be affected is to find out whether the feedback loop information is altered; one should not be misled simply by the fact that information is fed forward.

Figure 5.3: Block diagram for System A.
Figure 5.4: Block diagram for System B.

In practical “feedforward” active filter designs, the active feedforward can substantially affect the filter dynamics. This fact is not often explicitly recognized in the literature on the subject, but can be an important issue in the filter design. In some cases, additional passive compensation (damping) may be required.

5.2 Feedforward Active Filter Effects on the System Dynamics.
In this section, we analyze how the feedforward active filter affects the power converter system dynamics. The power converter with hybrid passive/active filter described in Chapter 4 can be modeled as shown in figure 5.5. When $K$, the gain of the injection signal, is close to unity, the active filter effectively reduces the impedance of $C_f$ in parallel with the load. When $K$ is exactly one, the capacitor and load draw no current and are effectively shorted to ground. Obviously, the system dynamics will be changed due to this impedance reduction.

Using the model shown in Fig 5.5, we can derive the transfer function $I_x/I_s$ and observe this effect on the system dynamics. Figure 5.6 shows the bode plot of the model both with and without the active filter element in use. From these plots we can indeed see that there is 20 dB of peaking in the transfer function at 10 kHz for the hybrid filter system with the active element turned on. This is undesirable since it indicates poor damping. Also, any
noise in this frequency range will be amplified by a factor of 10. Essentially, the feedforward active filter makes the power converter system less stable by moving the poles of the system closer to the $j\omega$ axis. This was also observed experimentally; an oscillation was found to occur at about 10 kHz when the active filter element was added on to the passive filter to form a hybrid filter.

To compensate for the effects of the active filter, a damping leg consisting of a resistor ($R_d=3.3$ $\Omega$) in series with a capacitor ($C_d=47$ $\mu F$) is placed in parallel with $C_2$. This reduces the peaking to less than 10 dB, which is acceptable (see Figure 5.7). Because the damping leg can be implemented with inexpensive low-power, low-frequency components, it does not add significantly to the filter size and cost. This was found to be very effective experimentally.

![Diagram](image)

**Figure 5.5:** Model for the single cell power converter system with the hybrid/active filter.
Figure 5.6: Bode plot of the transfer function for the $I_a/I_s$ hybrid passive/active filter.

Figure 5.7: Bode plot of the transfer function $I_a/I_s$ for the hybrid/active filter with damping leg.
5.3 Conclusion

Although feedforward active filters have been previously applied in power converter systems [7,8,19], their potential effects on system dynamics has not received sufficient attention. This chapter addresses the fact that “feedforward” filters can significantly affect the system dynamics. In the case considered here, the feedforward filter makes the filter dynamics less stable. This is because the variables that are fed forward are often not independent inputs but dependent state variables. This issue is easily exposed through the use of block diagrams and transfer functions for the system.

While the effects of the active elements on filter damping can be undesirable, it should be pointed out that the problem can be easily addressed with simple compensation elements (like the RC damping leg circuit in this case). In short, the effects on the system dynamics due to the feedforward active filter is a phenomenon that designers need to consider, but, it does not overshadow the great benefits of the active filter.
Chapter 6

Current Transformer Sensor

6.1 Introduction
The current transformer is another reliable and commonly used technique for sensing power converter ripple current. It consists of a magnetic core with primary and secondary windings. When current $i_1$ is flowing through the primary winding it creates a magnetic flux in the core which induces the current $i_2$ to flow through the secondary winding. Figure 6.1 shows an ideal transformer model; the voltage and current relationships between the primary and secondary are given by:

\[
\frac{v_1}{v_2} = \frac{N_2}{N_1} \tag{6.1}
\]

\[
\frac{i_1}{i_2} = \frac{N_2}{N_1} \tag{6.2}
\]

To sense the (large) primary current $i_1$, one can use relation (6.2) to generate a (small) proportional secondary current, which is passed through a burden resistor to generate a sense voltage proportional to the primary current.

![Diagram of current transformer model](image)

**Figure 6.1:** Ideal current transformer model.
Figure 6.2: A more practical transformer model with leakage inductances ($L_{II}$, $L_{I2}$) and magnetizing inductance ($L_{\mu}$).

In practice, however, the above equations do not fully represent the voltage and current relationships between the primary and secondary side in a real transformer. First, not all the flux is linked from one winding to the other. There will be some flux leaking from the core to the surrounding air. Therefore, there are some inductances presented in the transformer due to the leakage flux. Also, the finite permeability of the core gives a finite magnetizing inductance in shunt with the ideal transformer. A more accurate transformer model including both leakage and magnetizing terms is shown in figure 6.2. The magnetizing inductance $L_{\mu}$ is a short circuit at dc and all the dc current will go through it. Therefore, the presence of this finite magnetizing inductance prevents the current transformer from working at dc. Additionally, some of the ac current will pass through this magnetizing inductance instead of through the “ideal” transformer so that not all of the ac current passing through the primary winding will get reflected to the secondary winding. We call this current magnetizing current, $i_{\mu}$, and it is desirable to minimize the ratio of this magnetizing current to the current passing through the primary side of the “ideal” transformer ($i_I$). The magnetizing inductance parasitic strongly affects the design and operation of the current transformer sensor.
Current transformers have been previously used for sensing ripple current in active ripple filtering applications [7]. The wide-bandwidth sensing and the isolation of the approach make it advantageous in many applications. Generally, the current transformer is an excellent sensor in applications where the dc component of the sensed current is small. In the cases where a large dc component is present, the transformer might not be an appropriate choice, because it must be sized so that the dc current component does not saturate the core. This contrasts with the Rogowski-Coil sensor, in which the core size is independent of the dc current level. Thus, in applications where the dc component is very large, one expects the Rogowski coil to be a better choice than the current transformer.

This chapter addresses the design of current transformers for active filter applications, and also presents a comparative study of the current transformer and the Rogowski coil. The purpose of the analysis is to determine when the transformer is a better current sensor choice and when the Rogowski-Coil current sensor is more appropriate.

6.2 Current Transformer Topology and Design Issues
Figure 6.3 shows the schematic for a current transformer sensor which includes a burden resistor and an amplifier in addition to the transformer. A more practical transformer model which contains both the leakage inductances ($L_{li}$, $L_{l2}$) and magnetizing inductance ($L_{\mu}$) is used. The primary current $i_1$, the current to be sensed, contains both ac and dc components. The transformer induces a secondary current $i_2$ which ideally should be proportional to the ac component of $i_1$. When the induced ripple current $i_2$ flows through $R$, the burden resistor, it creates a voltage drop $v_2$ across the burden resistor. This voltage is proportional to the primary ripple current and is further amplified through a gain stage.
Several issues need to be considered when designing this transformer current sensor. One issue is that the magnetizing inductance needs to be large enough so that the ripple current is sensed accurately with a low phase shift. To understand this, consider the impedance seen from the primary side of the transformer as shown in Figure 6.4, where secondary-side leakage is neglected for simplicity. $R'$ is the impedance of the burden resistor $R$ reflected to the primary side. This resistance is in parallel with the magnetizing inductance $L_\mu$. At the ac ripple frequency, the impedance of the magnetizing inductance $L_\mu$ should be much larger than $R'$ so that most of the ripple current will flow through $R'$. Specifically, (6.3) shows the transfer function from the ripple current to the sensed output voltage. This is a high pass filter response, and the pole should be set at least one decade below the rip-
ple frequency in order to achieve the desired gain and small phase shift at the ripple frequency. This constraint sets the minimum impedance ratio of the magnetizing inductance to the reflected burden resistance $R'$.

$$\frac{v_{out}}{i_1} = \frac{L_{\mu} s}{L_{\mu} s + R'} \left( \frac{N_1}{N_2} \right) RK$$

(6.3)

The magnetizing inductance has a direct influence on the energy storage requirement of the transformer and hence the size of the core. The relationship is described in (6.4), where $I_p$ is the peak current flowing through the transformer and $W_m$ is the required energy storage. The energy storage requirement is also proportional to the core volume. Equation 6.5 shows this relationship, where $B$ is the magnetic flux density, $V_c$ is the volume of the core, and $\mu$ is the permeability of the core.

$$W_m = \frac{1}{2} L_u I_p^2$$

(6.4)

$$W_m = \frac{B^2 V_c}{2 \mu}$$

(6.5)

Combining these equations, the required volume of the core can be expressed as

$$V_c = \frac{\mu L_u I_p^2}{B_{sat}^2}$$

(6.6)

From the above equations we can see that if $L_{\mu}$ is reduced, then the energy storage and the volume of the core will also be reduced. To reduce $L_{\mu}$, $R'$ needs to be relatively smaller in order to maintain the same impedance ratio. This requires a larger secondary to primary turns ratio. Therefore, large turn ratio reduces the energy storage requirement and the core
size of the transformer. On the other hand, it also reduces the magnitude of the induced current \(i_2\) and the voltage drop \(v_2\), thus requiring large amplification for the same gain. While a smaller \(R'\) value helps to reduce the size of the transformer core, there is also a constraint on the lower bound due to the presence of the leakage inductances. If \(R'\) is too small, then the voltage drop across the secondary-side leakage inductance becomes significant. This will introduce errors in the amplitude and the phase of the sensed signal. Ultimately, required amplification limits the minimum size of the core, along with other effects such as leakage inductance of the core.

### 6.3 Design of the Current Transformer

Here we carry out the current transformer design and use it to analyze when it is advantageous to use the current transformer and when it is advantageous to use the Rogowski coil.

As mentioned earlier, the key issue here is that the size of the transformer core increases as the square of the dc current level (assume the ac currents are much smaller than the dc currents), as seen in (6.6). It is desired to determine at what dc current level the size of the transformer core will exceed that of the Rogowski coil. For a fair comparison, the transformer is designed to sense the same 125 kHz ripple current as the Rogowski-Coil sensor is designed for. The magnitude of the ac ripple current is 100 mA. Designs were considered for dc current levels from 1 A to 50 A. The overall gain of the sensor is set to be 10 V/A, similar to the Rogowski-Coil sensor gain.

The design parameters are listed in Table 6.1. A primary to secondary turns ratio of 1:100 was chosen. The burden resistor is set to be 10 \(\Omega\); this gives a reflected resistance of 1 m\(\Omega\) on the primary side. The magnetizing inductance was chosen to be around 1 \(\mu\)H. This sets the cut off frequency at 159 Hz, almost two decades below the switching ripple frequency. Therefore, this design should give a unity gain and negligible phase shift at the switching frequency of 125 kHz.
After setting these design parameters, the next step is to choose the core material for the transformer. Ferrite is preferred for the core because it has higher permeability than many other materials, and yields low core losses. Material 3F3 is selected, it has a permeability of 1800*\(\mu\)o, a maximum magnetic flux density of 450 mT, and has low loss for frequencies up to 700 kHz. From these parameters, the volume of the core can be calculated using (6.6). Once the required core volume is determined then an appropriate core can be selected. Table 6.2 shows the parameters and selected cores for sensors designed to handle dc currents from 1 to 50 A.

<table>
<thead>
<tr>
<th>N1:N2</th>
<th>(R) ((\Omega))</th>
<th>(L_\mu) ((\mu)H)</th>
<th>(K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1:100</td>
<td>10</td>
<td>1</td>
<td>100</td>
</tr>
</tbody>
</table>

*Table 6.1: Transformer Design Parameters (a).*

<table>
<thead>
<tr>
<th>(I_{dc}) (A)</th>
<th>(i_{ac}) (mA)</th>
<th>(L_\mu) ((\mu)H)</th>
<th>Wm ((\mu)J)</th>
<th>Minimum Core Volume (mm(^3))</th>
<th>Ferrite Core (OD/ID/HT) (mm)</th>
<th>Actual (L_\mu) ((\mu)H)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>100</td>
<td>1</td>
<td>0.5</td>
<td>22.79</td>
<td>5.8/3.1/1.5</td>
<td>0.3276</td>
</tr>
<tr>
<td>10</td>
<td>100</td>
<td>1</td>
<td>50</td>
<td>2279</td>
<td>22/14/13</td>
<td>2.080</td>
</tr>
<tr>
<td>20</td>
<td>100</td>
<td>1</td>
<td>200</td>
<td>9118</td>
<td>39/20/13</td>
<td>3.014</td>
</tr>
<tr>
<td>50</td>
<td>100</td>
<td>1</td>
<td>1250</td>
<td>56990</td>
<td>102/66/15</td>
<td>2.314</td>
</tr>
</tbody>
</table>

*Table 6.2: Transformer Design Parameters (b).*

Table 6.2 shows that the transformer core size increases quadratically as the dc current component increased, in accordance with (6.6). For the dc current level of 1 A, the outer core diameter is only 0.6 cm. However, when the dc current level increases to 50 A, the core diameter increases to 10 cm. The core designed to handle 20 A of dc current is slightly larger than the Rogowski coil described in Chapter 2. Thus it may be concluded
that for dc currents of 20 A and larger, the Rogowski-Coil sensor will be smaller.

To implement the required gain of k=100, the same quad current-feedback op-amp is used as in the Rogowski-Coil sensor of Chapter 2. Only three stages on the quad opamp are utilized, and each stage has a gain of 4.7. Therefore, the overall gain is 105, slightly larger than original design. The complete prototype sensor circuit is shown in Figure 6.5.

6.4 Evaluation of the Current Transformer

To validate the current transformer design approach, and to allow direct comparison to the Rogowski coil method, the C.T. sensor for 20 A dc current was prototyped and tested. The current transformer was tested under similar conditions as the Rogowski coil. The ac ripple current was generated by using a signal generator. The dc current was created using a power supply. Figure 6.6 shows the output waveform of the transformer sensor as compared to the input ripple current (bottom waveform) at 125 kHz. The vertical scale for channel 2 (top waveform) is 500 mV/div, and this peak to peak voltage is about 1 V. The vertical scale for channel 3 (bottom waveform) is 50 mA/div and the peak to peak input ripple current is 100 mA. The horizontal scale is 2 μs/div and the period is 8 μs. This plot
Figure 6.6: Output waveform for transformer sensor (top waveform) as compared to input ripple current (bottom waveform) at 125 kHz.

demonstrates that the current transformer accurately senses current at the switching frequency.

A network analyzer was also used to measure the frequency response of this sensor. Figure 6.7 shows the magnitude and phase plots on a log scale. The starting frequency is 10 kHz and the stop frequency is 5 MHz. The vertical scale is 10 dB/div for the magnitude plot and 10 degrees/div for the phase plot. The reference level (0 dB) for the magnitude is the very top line and the reference level for the phase is the middle line (0 degrees). This plot demonstrates that the magnitude is very flat into the MHz range. The phase shift becomes significant at frequencies above 1 MHz. Overall, this sensor demonstrates similar bandwidth performance to the Rogowski-Coil sensor.
Figure 6.7: Current transformer sensor frequency response. (10 dB/div, 10 deg/div)

6.5 Conclusion

It has been shown that the size of a current transformer sensor for this application depends on the square of the dc current levels in the system. This is unlike a Rogowski-Coil sensor whose size is not dependent on the dc current levels. Based on the parameters used in our example, it is found that the C.T.-based sensor becomes larger than an equivalent Rogowski-Coil sensor for dc current levels of approximately 20 A and above. Furthermore, the cost is higher even at lower current levels because the Rogowski-Coil does not requires a magnetic core.

To validate these findings, a current transformer sensor based on these result was developed. The sensor was designed to handle dc current levels of 20 amps, and exhibited
similar size and complexity to the Rogowski-Coil sensor (though its mass was larger due to the magnetic core.) Its performance was comparable to that of the Rogowski-Coil sensor. In summary, for systems with low dc currents, the current transformer is a superior sensor, but at higher dc current levels, the Rogowski coil has clear advantages.
Chapter 7

Transformer-Based Current Injector

7.1 Introduction
In Chapter 3 we have introduced a current injector method which make use of a class A transistor stage in conjunction with an op amp. While this approach is simple, accurate, and wide-bandwidth, it could also result in unacceptable power dissipation if the required injection current is large due to the need for a dc bias current. An alternative current injector approach is the use of a linear amplifier (with a class B or AB output stage) which drives the injection current onto the output through a current transformer and a coupling capacitor [8]. The current transformer converts the relatively high-voltage low-current output of the linear amplifier to a low-voltage, high-current injection signal. The purpose of the capacitor is to block the dc voltage at the injection node, so that the transformer sees only the ac voltage.

A clear advantage of this approach is low power dissipation since only ac voltage and ac current are present at the injection point. Injection levels of as much as an amp have been achieved without undue dissipation using this method [8]. The drawbacks to this approach are the cost and volume of the current transformer and coupling capacitor, and the performance limitations introduced by their parasitics. This chapter presents the design and evaluation of a transformer-based injector circuit.

7.2 Design of the Transformer-Based Injector
A schematic of the current transformer based injector circuit is shown in Figure 7.1. The primary side of the transformer is connected to the op amp and the secondary side is connected to the injection point (the output of the power converter). The input voltage, \(v_{\text{inj}}\), is the output from the current sensor and is proportional to the desired injection current. The
The idea of this injector circuit is to create a low level current $i_p$ that is proportional to the input voltage $v_{inj}$, which causes a proportional current $i_s$ to be injected through the action of the transformer. For an ideal op amp, the $V_+$ terminal has zero voltage since $V_-$ is tied to ground, and the current flowing into the $V_-$ terminal is negligible. Therefore, a proportional current $i_p$ can be generated by simply connecting a resistor from $v_{inj}$ to $V_-$. In Figure 7.1, an additional capacitor $C_1$ is used to form a high-pass filter which blocks the low frequency content of input signal. As long as the cut off frequency of the filter is set far enough below the ripple current frequency, there is no phase distortion in the current $i_p$. By adjusting the resistor value and the turns ratio of the transformer, a high-level current $i_s$ of the proper magnitude is induced at the injection point.

There are design issues that constrain the scaling of the turns ratio and the resistor. The primary side of the transformer has a voltage limit which is determined by the voltage output swing of the op amp. The net impedance looking from primary side into the secondary side of the transformer should not be too large for a given $i_p$ in order to prevent the op amp
Figure 7.2: Impedance model looking from the primary side of the transformer into the secondary side.

from saturating. This impedance looking into the secondary side can be modeled as shown in figure 7.2, where $Z$ is the impedance looking from the secondary side of the transformer into the load, $L_{II}$ and $L_{I2}$ are the leakage inductance of the primary side and the secondary side respectively, and $L_\mu$ is the magnetizing inductance. Since both the reflected secondary leakage inductance and the reflected secondary impedance ($Z'$) increase quadratically with the turns ratio ($N_1/N_2$), there is a constraint on how large this turns ratio can be. If the turns ratio is too high, then there will be a significant voltage drop across the leakage inductance and the magnetizing current will increase; both of the results will affect the accuracy of the injection current. Furthermore, additional compensation might be required if these parasitics interfere with the op amp dynamics and cause stability problems.

Taking these issue into consideration, a prototype transformer injector circuit has been designed and constructed. Figure 7.3 shows the schematics for this prototype circuit. The impedance looking into the output of the power converter is modeled as a capacitor ($C_3$). There are several additional RC elements in this figure as compared to figure 7.1; these elements are to compensate the op amp. The objective of this design is to be able to inject ripple currents of up to 1 A on the secondary side. The turns ratio of the primary to the secondary is set to 1:25; this means the upper bound of the primary current is 40 mA.
Figure 7.3: Transformer-based current injector using an LM6361 op amp.

<table>
<thead>
<tr>
<th>Resistor</th>
<th>Capacitor</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R$</td>
<td>$C_1$</td>
<td>160 $\Omega$</td>
</tr>
<tr>
<td>$R_k$</td>
<td>$C_2$</td>
<td>3.9 k$\Omega$</td>
</tr>
<tr>
<td>$R_b$</td>
<td>$C_3$</td>
<td>100 $\Omega$</td>
</tr>
<tr>
<td>$R_k$</td>
<td>$C_k$</td>
<td>160 $\Omega$</td>
</tr>
</tbody>
</table>

Table 7.1: Parameter values for the transformer based injector circuit.

From this information, we can determine the values for $R$ and $C_1$. Assuming the Rogowski-Coil current sensor is used, then the maximum injection voltage is 6.6 V. By setting $R$ to be 164 $\Omega$ and $C_1$ to be 1 $\mu$F, we put the cut-off frequency of the high-pass filter at 1 kHz, two decades below the fundamental switching frequency of the converter, and the correct primary current is obtained.
The next task is to design the transformer. Two criteria are considered: first, the magnetizing inductance needs to be large enough so that the magnetizing current is negligible compare to the current flowing through $Z'$; second, the cross sectional area of the core has to be large enough to prevent the core from saturating. To understand the details involved in the transformer design a step by step procedure is presented.

**Step 1: Core selection**

For winding the transformer, a high permeability core is desired since it gives a high magnetizing inductance and low leakage inductances. Also, the core material needs to have a bandwidth large enough to cover the operation frequency. Considering these two criteria, a 3F3 ferrite core is selected. It has a permeability of $1800*\mu_0$ and a low-loss bandwidth up to 700 kHz.

**Step 2: Compute the magnetizing inductance**

The magnetizing inductance can be computed using (7.1), where $V_I$ is the voltage across the primary side of the transformer ($V_{out}$ to $V_i$), $i_m$ is the magnetizing current, and $f$ is the ripple current frequency. To ensure most of the current flows through $Z'$, the (primary side) magnetizing current $i_\mu$ is set to be a maximum of 400 $\mu$A, 100 times smaller than the total primary current. Given the voltage swing is 5 V and the switching frequency of the converter is 125 kHz, the required (primary side) magnetizing inductance is at least 0.016 H.

$$L_\mu = \frac{V_1}{i_m 2\pi f}$$  (7.1)

**Step 3: Determine the primary winding turns**

The number of turns is related to the magnetizing inductance by (7.2). After rewriting this equation, we can use (7.3) to calculate the number of turns required. We can choose
the specific values for \( l_c \) and \( A_c \), the effective circumference and cross section area of the core respectively; however, the ratio of obtainable \( A_c \) to \( l_c \) has a rather fixed range of values as can be determined from a core catalog. Therefore, an available \( A_c \) to \( l_c \) ratio value can be selected to calculate the number of turns required for the primary side. Since the primary to the secondary turn ratio is 25:1, the number of turns \( N_I \) needs to be rounded up so that the secondary winding is an integer value. In this case, \( N_I \) is rounded up to 50 to produce 2 turns on the secondary side.

\[
L_{\mu} = \frac{\mu N^2 A_c}{l_c} \quad (7.2)
\]

\[
N_I = \sqrt{\frac{L_{\mu} l_c}{\mu A_c}} \quad (7.3)
\]

**Step 4: Compute the cross section area required**

Equation 7.4 describes the constraint on the cross section area of the core \( A_c \) for preventing core saturation. \( K \) is a scaling factor which depends on the waveform of the ripple; for a sine wave, \( K \) is 4.44 [18]. \( B_{max} \) is the maximum magnetic flux density of the material and \( N_I \) is the number of turns on the primary side. (Note in other references, \( N_s \), the number of secondary winding turns might be used in this equation, since the secondary winding is defined as the one with more turns. It is essentially the same here, except that we have defined the primary side as the one with more turns. The reader should not be mislead by this difference.) The minimum cross sectional area of the core is calculated to be 4x10^{-3} cm^2; thus a very small core is sufficient for meeting the core saturation requirement. Therefore, the core size now is determined by the window area required for windings. In
the design considered here, the Phillips Ferrite core TX 22/14/13 3F3 is selected for winding the transformer.

\[ A_c > \frac{V_1}{KfN_1B_{max}} \]  \hspace{1cm} (7.4)

### 7.3 Evaluation of the Transformer Injector.

The transformer injector circuit is tested using a function generator to provide the ac \( v_{inj} \) signal. A scope measurement of this input voltage signal versus the injection current is shown in figure 7.4. Channel 3 displays the 125 kHz input voltage; the horizontal scale is 2 \( \mu \text{s/} \text{div} \) and the vertical scale is 500 mV/\text{div}. The peak-to-peak amplitude of the voltage is 0.66 V. Channel 4 shows the injection current. The vertical scale is 50 mA/\text{div}, resulting in 100 mA of peak to peak ripple current. This plot shows that the injection current has an accurate gain with a phase shift less than 4° at the fundamental switching frequency of the converter.

### 7.4 Conclusion

The transformer-based injector circuit has the main advantage of low power dissipation as compared to the class A injector circuit introduced in Chapter 3. For example, assume that the converter output voltage is composed of 14 V dc plus 0.2 V peak-peak ac. For the class A injector to cancel 1 A of peak to peak ripple current, it will have to draw greater than 0.5 A of dc bias current. The total power dissipation of the injector is the sum of the dc component and the ac component. The dc power dissipation is simply the product of the dc voltage and the dc current, resulting in 7 W of power loss in this case. The ac power dissipation is the product of the rms values of ac voltage and the ac current times the cosine of
the angle between them. Assuming the voltage and the current are both sinusoidal, this power

![Graph showing voltage and current waveforms.](image)

**Figure 7.4:** Scope measurement of the input voltage (Ch3) versus injection current (Ch4) at 125 kHz. The phase shift of the injection current is less than 4° at the fundamental.

loss is given by (7.5)

\[ P_{ac} = \frac{V_{pk}I_{pk}}{2} \cos \theta \]  

(7.5)

where \( V_{pk} \) and \( I_{pk} \) are the peak amplitude of the voltage and current respectively, and \( \theta \) is the angle between the voltage and the current. If there is no phase shift (worst case), then the ac power dissipation for the class A injector would be 0.025 W. Therefore, the total power dissipation for the class A injector is 7.025 W, over 3% of the converter output power. On the other hand, the transformer-based current injector only has ac power dissipation since it draws no dc current. Therefore, its total power dissipation is simply due to the ac component, 0.025 W in this case, about 0.1% of the converter output power. From
this example we can see that when a large ac current needs to be injected, the power loss for the class A injector is substantially larger than the transformer-based current injector, and can be impractical.

Despite its efficiency advantage at high injection current levels, the transformer-based current injector also has drawbacks. For example, the volume of the transformer increases and the core becomes more expensive as the injected current becomes large; therefore, there is a trade-off between the volume/cost of the core and the power loss of the system. Moreover, the transformer injector circuit is harder to implement since the parasitics (inductance especially) of the transformer affect the dynamics of the op amp and limit the bandwidth of the injector. Good compensation is necessary for accurate injection. Ultimately, a simpler injector circuit, such as the one considered in chapter 3, is less expensive and performs better for low injection currents, while the more complex design considered here is necessary if large currents are to be injected.
Chapter 8

Conclusion and Recommendation for Future Work

This thesis has demonstrated that the hybrid passive/active filtering approach has the advantage of reducing the size, weight, and cost of ripple filters for power converters by a substantial amount as compared to the conventional passive filtering approach. Because the use of an active filter reduces the size and energy storage of the passive filter, substantially improved transient performance can also be obtained. This thesis has also addressed the fact that “feedforward” filters can significantly affect the system dynamics since the ripple component being fed forward for cancellation is not an independent input. While this is a phenomenon that the designer needs to watch out for, it does not overshadow the great benefits of the active filter approach.

This thesis has also presented the innovative use of a Rogowski-Coil current sensor in the active filter application. The accuracy, wide bandwidth, and low cost of this sensor make it effective and advantageous in this application. It is demonstrated that for sensing currents with dc components of 20 A and higher, the Rogowski-Coil current sensor is significantly a better choice than the conventional current transformer sensor in terms of size, weight, and cost. In addition, a class A injector and a transformer injector are designed and evaluated in this thesis. The class A injector circuit is simpler, less expensive, and performs better for small injection currents, while the more complex transformer-based injector circuit is more efficient for injection of larger currents.
The active filtering approach for ripple current and ripple voltage cancellation is a very promising area that has only received very limited attention to date. Further investigation of feedforward active filtering techniques particularly for voltage-ripple cancellation, is definitely warranted. Another promising area is that of feedback active filtering, in which a high-gain feedback loop is used to suppress ripple. A more detailed study could be done which compares the advantages and limitations of these approaches. Also, hybrids of feedforward and feedback techniques could be investigated and compared to other approaches.
References


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