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APPLICATION NOTE

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Generating Multiple Isolated Bias Rails for IGBT Motor Drives with Flyback, SEPIC, and Ćuk Combination

by Bob Zwicker

INTRODUCTION

State-of-the-art motor drives use a 3-phase, insulated gate bipolar transistor (IGBT)-based inverter that is powered by a dc link voltage typically in the region of 400 V dc to 800 V dc. That high voltage rail can be derived directly from a 3-phase rectifier bridge filter combination or from a power factor corrected boost rectifier, which produces the high voltage rail from a 3-phase ac input (see Figure 1).

The IGBTs are the main power switches, which provide a (typically 10 kHz) pulse width modulated (PWM) output to each of the three motor phases. Induction and permanent magnet motors generally have high winding inductance, which integrates this PWM voltage into a low frequency winding current waveform that is approximately sinusoidal in shape. Whereas some IGBTs in smaller drives work well with unipolar (0 V to 15 V, for example) gate drive provided by a driver such as the ADuM4223, the usual requirement in larger systems is for bipolar gate drive levels (such as −7.5 V and +15 V) as driven by a suitable driver such as the ADuM4135. The negative turn-off level helps to avoid spurious turn-on of an IGBT, which can be induced by a rapid rise (high positive dV/dt) in the collector to emitter voltage (V_{CE}). This high dV/dt is commonly caused by the normal turn-on of the other device. (Turn-on of the upper device can induce undesired turn-on of the lower device or vice versa.) The six gate drivers need a power source to provide these +15 V and −7.5 V bias voltages.

In the example shown in Figure 1, two of the three motor phases have shunt resistors in series with the motor winding, across which are connected AD7403 isolated Σ-Δ modulators to measure the motor phase current. (The current is measured in only two phases because the third can be inferred.) These two Σ-Δ modulators are typically powered by 5 V.

The driver bias voltages for the three high-side (HS) IGBTs are each referenced to their respective motor phase, which means that the three high-side drivers (connected to the three motor phases) each have their own isolated bias power domains (HS-U, HS-V, and HS-W). In addition, the three low-side (LS) drivers are all referenced to the negative dc link, and therefore share one more bias power domain (LS). Table 1 lists the total requirement for bias power domains and included bias rails in a typical motor drive.

Table 1. Motor Inverter Power Supply Requirements

Inverter Circuit	Domains	Voltages (V)	Voltage Rails
Three Low-Side IGBTs	LS	+15, −7.5	2
Three High-Side IGBTs	HS-U, HS-V, HS-W	+15, −7.5	6
Two High-Side Σ-Δ Modulators	HS-V, HS-W	+5	2
Total	4		10

Whereas the calculation in Table 1 gives a total of 10 rails, the exact total can vary with the design of the motor drive and is not critical in the context of this application note. The exact number need not influence the techniques for providing these rails. It is the techniques that are the subject of this application note.

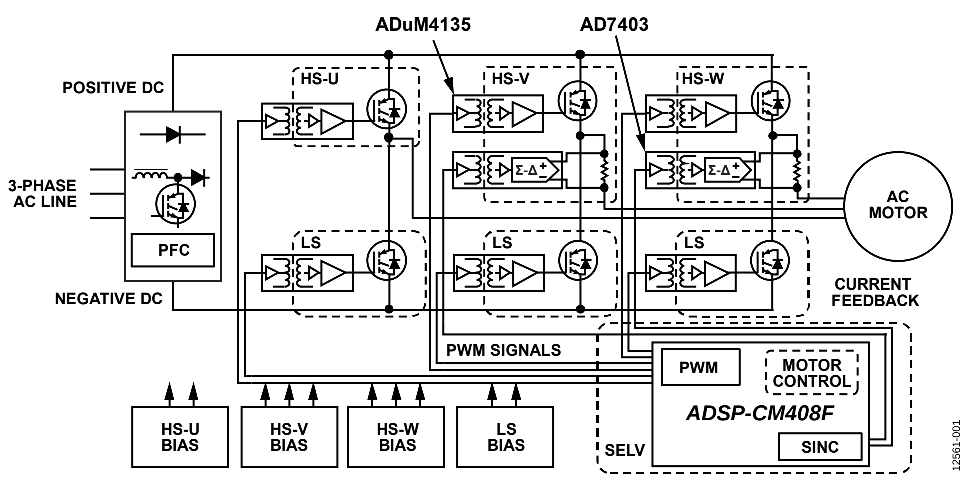

Figure 1. Block Diagram of a Typical Industrial Motor Drive

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REVISION HISTORY

11/15—Revision 0: Initial Version

BASIC CONSTRAINTS ON THE BIAS VOLTAGES

Any method for providing these bias rails must consider a few basic requirements.

ISOLATION

In mid to high end motor drives, the processor generally operates in the safety extra low voltage (SELV) power domain to optimize performance. Similar to the power that is supplied to the ports of common audio equipment or a PC, these voltage and current levels are low enough that they are not considered to be dangerous. Precautions against accidental human contact are not required. In this way, the processor human interface is easily accessible without the need for safety isolation. However, the IGBTs and motor phases typically operate with voltages of a few hundred or more volts, both relative to each other and relative to the SELV power domain. Therefore, the IGBT gates, the driver outputs, and the bias voltages that power them are all hazardous. Safety isolation is required between the IGBT gate voltage domains and the SELV power domain from which they are powered, and functional isolation is needed between the domains themselves. The bias power supply transformer needs an isolated output winding and at least two connection pins for each isolated power domain.

In addition to the absolute magnitude of the voltage, the common-mode slew rate (rate of change of voltage or dV/dt) of the motor phases must be considered. Figure 2 is an observation of the switching of an IGBT driven motor phase of a demonstration board. This measurement shows a slew rate of 11 V/ns. The bias voltage must ride on this common-mode voltage slew and must not be disturbed by it.

In Figure 2, Channel 1 is the emitter and Channel 2 is the gate on a high-side IGBT, which is turning on with positive load current flowing out of the emitter. Based on the Channel 1 cursor measurement, dV/dt is 11 V/ns.

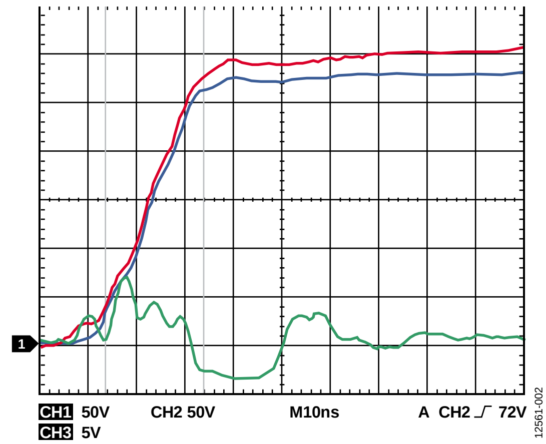

Figure 2. Motor Phase Voltage Waveform

DWELL

Depending on the motor drive algorithm, the motor phases may need to dwell in some state (such as high voltage output or low voltage output) for a relatively long period of time. In particular, some space vector modulation schemes can cause a motor phase to be switched high for milliseconds or longer. Some methods (such as bootstraps) of biasing the drivers are not compatible with these modulation schemes.

VOLTAGE REGULATION

Voltage regulation is one of the less demanding performance criteria for the isolated voltage converter. The output voltages to the gate drivers must stay within ±3% to ±5% over a load current range of approximately 10:1, which is relatively low precision. The 5 V output to the Σ-Δ modulator needs ±1% voltage regulation, which can be delivered by a low dropout (LDO) regulator such as the ADP7118, ADP7102, or ADP7104.

METHODS FOR GENERATING THE BIAS VOLTAGES

A resistor sourced charge pump is one of the cheaper methods of developing a bias voltage referenced to a motor phase. In the example in Figure 3, the charge pump generates a single positive rail. This may be adequate in a basic motor drive; however, its dissipative operation makes it very inefficient. The losses become unacceptable where more rails or higher current is required.

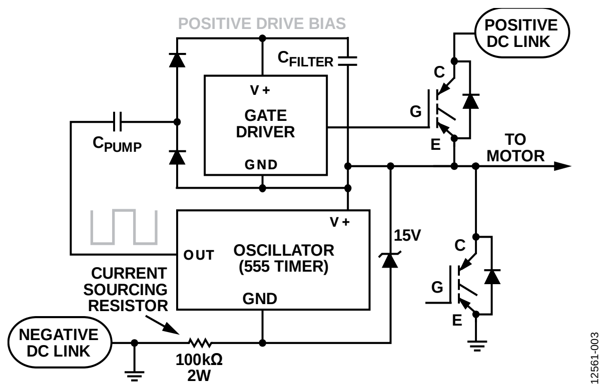

Figure 3. Resistor Sourced Charge Pump

Resistor sourced charge pumps are commonly used despite being inefficient.

The advantages of a resistor sourced charge pump are

- Multiple outputs from two transformer pins
- Low components cost
- Good load regulation
- Flexible voltage setpoints
- Avoids duty cycle limitations of the bootstrap

The disadvantages of a resistor sourced charge pump are

- Very low efficiency
- Low output current capability
- No power transfer when bottom IGBT is turned on

When only 15 V is needed, providing it for the low-side IGBTs is not a problem and those low-side IGBTs are never off for long periods, a bootstrap (as shown in Figure 4) may be the best way to power the high-side drivers, using the bias supply for the low-side drivers.

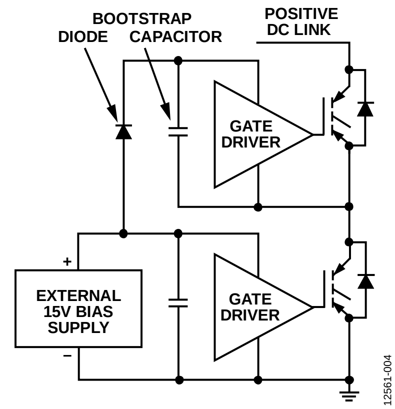

Figure 4. Bootstrap

Bootstraps are widely used, especially in buck dc-to-dc voltage conversion. They are highly recommended for applications where bootstraps can work satisfactorily.

The advantages of bootstraps are

- Avoid multiple transformer output windings
- Low cost
- Efficient

The disadvantages of bootstraps are

- Low-side IGBT must turn on frequently to recharge the bootstrap capacitor, which may be incompatible with space vector modulation
- Producing negative bias rails more difficult
- Charge pump aspect limits output current

Transformer-based techniques are the leading candidates for avoiding the limitations of the charge pump and bootstrap methods.

APPLYING FLYBACK CONVERTERS TO GENERATE MOTOR DRIVE BIAS

For low power isolated converters where the transformer can have low leakage inductance, the flyback topology is the most common and economical choice.

However, many commercial off-the-shelf flyback power supplies use optocouplers for providing feedback from the secondary side error amplifier to the primary side PWM controller. There are two issues with this approach.

The first issue is that secondary side voltage sensing is typically used with one main output of a converter. Secondary side voltage sensing can provide excellent voltage regulation (such as 1%) of this main sensed output. However, the typical motor drive example has 10 outputs in 4 isolated domains. When load current variations are applied to a main sensed output, the voltage regulation from the other, slaved outputs are usually adversely affected. This effect is commonly called cross regulation.

Secondary side voltage sensing can alternatively regulate a weighted combination of several outputs; however, these outputs normally must all be within one isolated domain. Variations in the combined loading of these sensed outputs likewise adversely affect the slaved outputs such as those in other isolation domains. Sensing several outputs in one domain does not appear to improve the cross regulation in the slaved or unsensed domains.

The second issue is that, due to capacitance and high gain at the base of the phototransistor, optocouplers tend to be adversely affected by high common-mode dV/dt. Figure 2 shows actual dV/dt at the IGBT emitter driving a motor phase; 11 V/ns is likely to interfere with proper operation of many optocouplers. The optocoupler can be replaced with a better device, such as the ADuM3190. The ADuM3190 replaces both the optocoupler and secondary side reference, which are typically used in an isolated power supply. It uses an integrated microelectronic transformer for coupling across the isolation barrier. It is not disturbed by the 11 V/ns common-mode slew rate.

With the many isolated outputs in a motor drive gate bias power converter, secondary side voltage sensing is not advantageous.

SENSING FLYBACK CONVERTER OUTPUT FROM THE PRIMARY SIDE

Another option is sensing the transformer output voltage from a winding on the primary side. A primary sensed flyback converter can provide simplicity and good output voltage regulation. In secondary side sensed converters, the main output is regulated tightly but the slaved (not sensed) outputs vary with the loading on the sensed output. In primary side sensed converters, the sensed output load is fixed and can be very low; therefore, there is no variation in the slaved outputs due to loading on the sensed output. Therefore, the worst-case voltage regulation from all of the outputs may be better with primary side voltage sensing. A ±3% (approximate) load regulation over a reasonably wide range of load currents is adequate for many purposes, including the typical motor control gate drive bias power requirements.

Individual output voltages are varied by changing the turns on that transformer winding. It is also possible to proportionally change all output voltages by changing the feedback network or by changing turns on the control winding.

Primary side voltage sensing also eliminates the secondary side reference and isolated feedback. It is usually a simpler and cheaper design with fewer components and a smaller printed circuit board (PCB) footprint. The technique has demonstrated good immunity to common-mode dV/dt across the isolation barrier.

Figure 5 shows the topology of an isolated, primary sensed, flyback voltage converter.

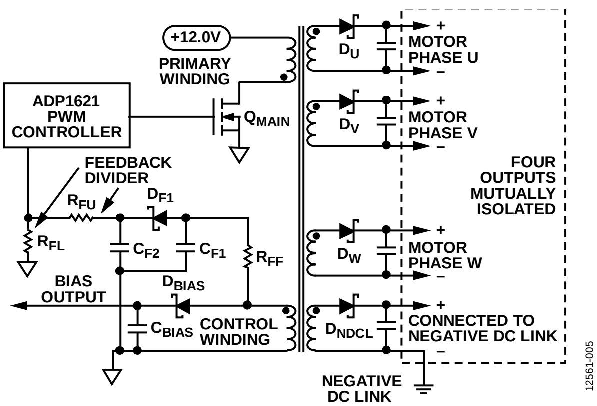

Figure 5. Simplified Schematic of Multiple Output, Isolated, Flyback Converter with Primary Side Voltage Sensing

When Q_{MAIN} turns off, the control winding delivers the same volts per turn as do the isolated outputs (on the right side of the transformer in Figure 5).

R_{FF} (100 Ω to 500 Ω) and C_{F1} (50 pF to 300 pF) form a low-pass filter that suppresses a leading edge voltage spike from the ac input waveform at the anode of D_{F1}. This spike coincides with Q_{MAIN} turnoff and is a function of the complex transformer leakage inductance. With the output of D_{F1} loaded as lightly as possible, this spike is rectified by D_{F1} and causes a significant degradation of the converter voltage regulation if it is not suppressed. The rectified output of D_{F1} is dc filtered by C_{F2}.

The cyclic charge in C_{F1} is a loss term; therefore, the value of C_{F1} must be minimized. With minimum C_{F1}, obtaining the necessary low-pass filter time constant requires maximizing the value of R_{FF}. (The optimal value of this time constant depends on the transformer but is typically in the 10 ns to 100 ns range.) However, the purpose of R_{FF} is to work with C_{FF} to form an ac low-pass filter only; the dc voltage drop in R_{FF} is an error term, which must be minimized. Minimizing this dc voltage drop requires choosing R_{FU} and R_{FL} to have the highest practical impedance that is consistent with the FB pin input bias current of the ADP1621 PWM controller IC. R_{FU} and R_{FL} form the feedback divider, which works with the control winding to set the output voltage.

D_{F1} must be a small signal (10 mA to 200 mA current rating) Schottky diode with adequate voltage rating.

One other consideration is the value of C_{F2}, which is used for dc filtering. Making C_{F2} too large moves an additional pole into the feedback loop passband and can cause instability. To avoid negatively impacting the feedback loop phase margin, C_{F2} must be as small as possible, consistent with its dc filtering task; 10 nF is a typical value.

The Figure 5 design also includes D_{BIAS} and C_{BIAS} (these are optional) to provide operating bias current to the controller IC for best efficiency when powering with a 12 V to 48 V input, and an efficient 5 V primary side bias rail is not available. Note that although feedback can be derived from D_{BIAS} so that it performs dual functions, separating D_{BIAS} and D_{F1} (with minimum loading on D_{F1}) provides the best voltage regulation.

COMBINING FLYBACK, SEPIC, AND ĆUK TOPOLOGIES

The following sections focus on output topologies for the flyback converter, developing a progression towards the SEPIC and Ćuk output circuits, which are the subject of this application note.

Figure 6 shows the most common method to produce two outputs from a flyback transformer. It is simple and efficient and offers independent output voltage setpoints based on the number of turns in each winding.

This method requires a transformer pin for each output, plus one for the common connection. This requirement is a disadvantage for producing a large number of outputs.

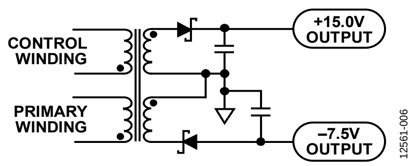

Figure 6. Straightforward Method to Produce Multiple Output Voltages from a Flyback Transformer

The advantages of this method are

- Good efficiency
- Low components cost/count
- Good load regulation
- Flexible voltage setpoints
- Easily produces negative or positive voltages

The disadvantage of this method is that it requires one transformer pin per output voltage, and one additional pin for the common point.

Figure 7 shows an approach with different tradeoffs. It uses only two transformer pins for an isolated output domain. It uses linear or dissipative means to accomplish rail splitting. It can regulate well; however, its application space is limited to low output current. It is the least efficient of the flyback output circuit architectures discussed in this application note but produces multiple motor drive bias voltages from one flyback output winding.

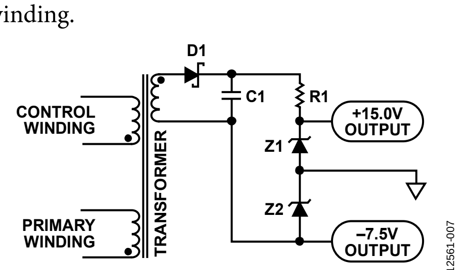

Figure 7. Dissipative Rail Splitting

The maximum required output current and the minimum Zener bias current must flow through D1 and R1 at all times.

The advantages of the dissipative rail splitting method are

- Multiple outputs from two transformer pins
- Low components cost
- Good load regulation
- Flexible voltage setpoints

The disadvantages of the dissipative rail splitting method are

- Low efficiency
- Low output current capability

Adding a coupling capacitor between output windings (as shown in Figure 8) can improve the voltage tracking between two dc outputs that have the same turns and produce the same voltage magnitude. The coupling capacitor effectively neutralizes the impact of the transformer leakage inductance on the output and therefore improves cross regulation. To illustrate how the added coupling can benefit flyback output regulation, Figure 9 shows experimental test results with a 36 W offline flyback converter using a PQ3230 core transformer.

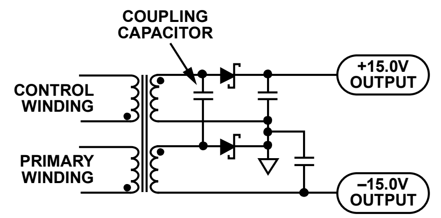

Figure 8. SEPIC and Ćuk Flyback Modification

The Figure 8 modification is based on Figure 6. Both windings have both ends connected to transformer pins, the negative rectifier was moved to the opposite end of the winding, and a coupling capacitor was added between the output windings. Both output windings must have the same turns.

The advantages of the Figure 8 modification are

- Good efficiency
- Low components cost
- Best cross regulation within a domain

The disadvantages of the Figure 8 modification are

- Two transformer pins required per output per domain
- Output voltage magnitudes must match each other

Table 2. Bias Supply Load Test Conditions

Load Combination	+12 V Output (A)	−12 V Output (A)
1	0.01	0.50
2	0.01	0.02
3	0.10	0.02
4	0.10	0.01
5	0.20	0.01
6	0.50	0.02
7	0.50	0.01
8	1.00	0.01
9	2.00	0.02
10	2.00	0.01
11	3.00	0.02
12	3.00	0.01

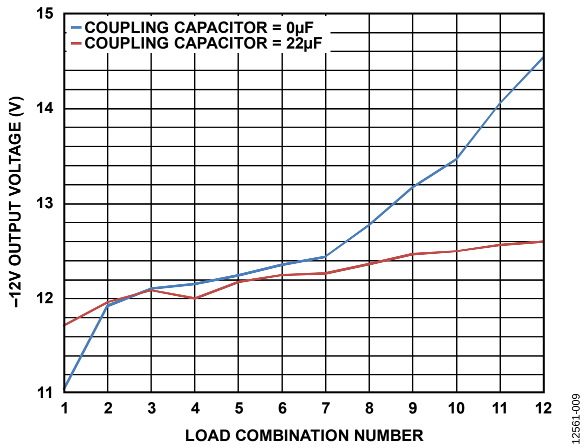

Figure 9. Test Data for the Dual Output Flyback Supply With and Without Coupling Capacitor Between Output Windings

Figure 9 shows test data made with a 36 W ± 12 V output flyback offline power supply. The output rectifier circuit architecture is similar to Figure 6; therefore, it operates similarly and has qualitative value in demonstrating the effect of the coupling capacitor. The output rectifiers are SS2PH10 for the 500 mA, −12 V output and SS5P10 for the 3 A, +12 V output. Outputs were measured using the same load current combinations, both with and without the coupling capacitor connected. In Table 2 and Figure 9, results were sequenced according to increasing negative output without the coupling capacitor. The feedback loop tightly regulated the +12 V output; therefore, its variation during the test was negligible.

With the capacitor omitted, the measured −12 V regulation band was −12.8 V ± 13.7%.

With the capacitor connected, the −12 V regulation band was −12.2 V ± 3.6%.

The next step in the design progression is to use one or more external (discrete or coupled) inductors to replace one or more transformer output windings, as shown in Figure 10.

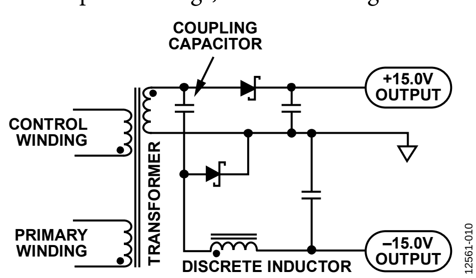

Figure 10. Dual Output Flyback Supply Using a Single Transformer Winding

The modification shown in Figure 10 is based on Figure 8, replacing one transformer winding with a discrete inductor. The −15 V output is a Ćuk output.

The advantages of the Figure 10 modification are

- Good efficiency
- Low components cost
- Improved load regulation
- Multiple outputs from one transformer winding with only two pins

The disadvantages of the Figure 10 modification are

- Additional discrete inductor needed
- Outputs must produce the same voltage magnitude

This modification provides similar performance to the circuit in Figure 6 yet reduces (to a total of two) the number of transformer pins required to make two outputs in one isolated domain.

Figure 11 shows a motor drive application with 7.5 V out of the transformer winding. The common point is connected as the most negative; therefore, the rectified transformer outputs are 7.5 V and 15 V. The 7.5 V feeds an LDO regulator to produce a 5 V rail, while the 15 V powers the unipolar gate driver.

Two outputs are not a limit. The circuit shown in Figure 12 uses a two-winding coupled inductor (Coilcraft LPD6235-473), which series connects three equal rectified outputs to produce −7.5 V, +7.5 V, and +15 V. This is the architecture used in the complete multiple output flyback converter design shown in Figure 18.

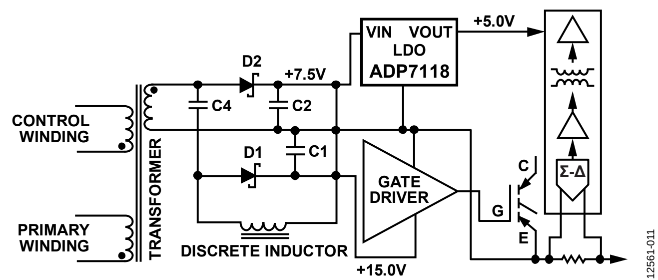

Figure 11. Dual Output Supply Supporting 15 V for the Isolated Gated Drive and 5 V for the Isolated Modulator

The Figure 11 design is similar to Figure 9, but produces 7.5 V and 15 V in a demonstration board. The 7.5 V powers a 5 V output LDO regulator to run the analog-to-digital converter (ADC).

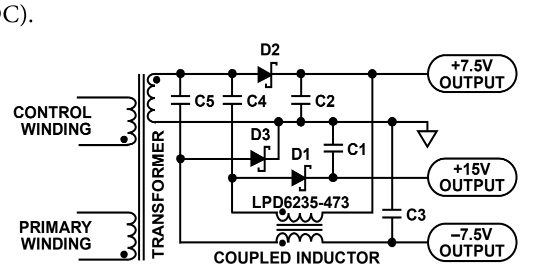

Figure 12. Triple Output Supply Supporting 15 V and 7.5 V for the Isolated Gated Drive and 7.5 V for the Isolated Modulator Circuit

The Figure 12 design is similar to Figure 10 but uses a coupled inductor to produce three output rails from one transformer winding. The +15 V power rail is for the gate turn-on drive; the +7.5 V rail powers the +5 V LDO regulator; and the −7.5 V rail is for the gate turn-off drive.

The load and cross regulation of the bias converter in Figure 12 was measured for the +15 V and −7.5 V outputs over load ranges of 5 mA to 200 mA. Figure 13 and Figure 14 show the calculated and graphical results that demonstrate that this topology is capable of maintaining good output voltage tolerance over at least a 40:1 load current range.

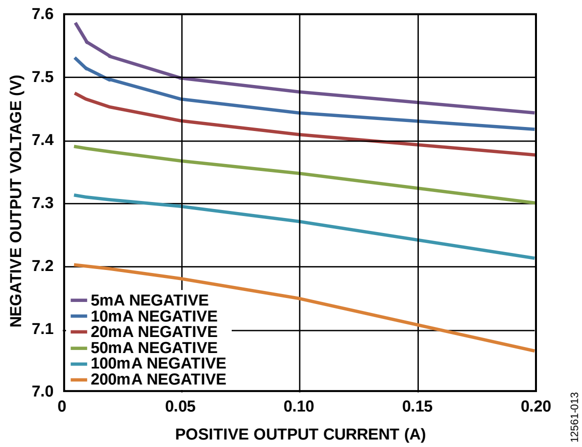

Figure 13. Measured Output Voltage Regulation of −7.5 V Output vs. Varying Load Current on −7.5 V and +15 V (Circuit Shown in Figure 12)

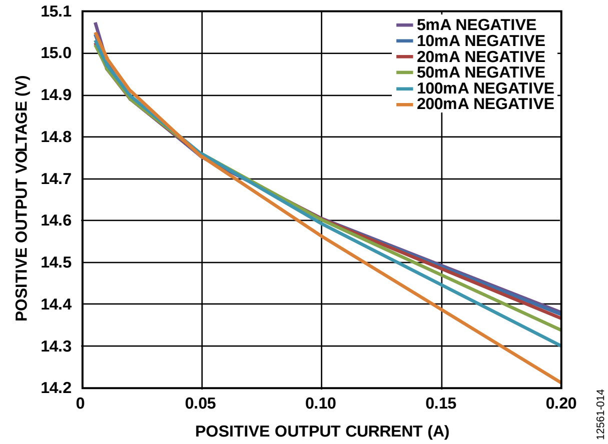

Figure 14. Measured Output Voltage Regulation of +15 V Output vs. Varying Load Current on −7.5 V and +15 V (Circuit Shown in Figure 12)

Table 3. Measured Load Regulation of Bias Converter shown in Figure 14

Output (V)	Tested Current Range (mA)	Center Point (V)	Tolerance (%)
+15	5 to 200	14.64	±3.0
−7.5	5 to 200	7.326	±3.6

EXPLANATION OF CIRCUIT THEORY AND TOPOLOGY COMPARISON

Flyback, SEPIC, and Ćuk converters are all buck or boost converters. They convert power by switching a winding across the input voltage to store energy in the magnetic core, then switching the same or other windings across the output(s) to deliver energy. (Buck, boost, and other topologies differ in important ways and are generally not as suitable for multiple outputs.) Because they all have the same basic operating mode, the voltages and duty cycles are all based on volt-second balance across the inductors and charge balance through the capacitors. Once any turns ratios are accounted for, the operating equations are the same. The basic blocks of varied buck or boost topologies can be combined in a variety of ways that are capable of producing proportional output voltages that track well over a wide current range.

The following are some relevant equations governing continuous conduction mode (CCM) and discontinuous conduction mode (DCM) operation of buck-boost converters.

The equation for CCM voltage conversion is

$$V_{OUT} = \frac{D \times V_{IN}}{1 - D}$$

where D is the duty cycle.

The equation for DCM total output power in watts is

$$P_{OUT} = \frac{(D \times V_{IN})^2}{2 \times L \times f}$$

where:
f is the frequency.
L is the total parallel inductance, in Henries.

The equation for DCM voltage conversion into a resistive load is

$$V_{OUT} = \frac{D \times V_{IN} \times R^{0.5}}{(2 \times L \times f)^{0.5}}$$

where R is the load resistance, in ohms.

Regarding inductance, in these SEPIC and Ćuk related designs, the coupling capacitors act as short circuits for switching frequency ac current. For the previous equations, consider that the transformer and output inductor(s) are in parallel. With coupled inductors (two or more windings with the same number of turns on one core), the inductance of either winding or both (or all) in parallel (connected in phase) is the same. This is usually the published inductance value of the coupled inductor.

When connecting multiple separate inductors or transformers in parallel (dc- or ac-coupled), as shown in Figure 10, Figure 11, or Figure 12, the effective inductance value is determined by the following equation for multiple inductors in parallel:

$$L_P = \frac{1}{(1/L_1) + (1/L_2) + ... + (1/L_n)}$$

In a flyback transformer, the core flux links all windings and produces the same volts per turn in all windings at any moment. This aspect allows the regulation of several outputs by monitoring the voltage produced by one of them. Leakage inductance is inductance that acts in series with one winding and is not shared with other windings. It decouples the windings and is usually minimized in the design of a flyback transformer. By comparison, coupled inductors can be designed for minimum leakage inductance, or they can be designed to have a specific leakage inductance. Some coupled inductors that are designed for minimum leakage inductance can work well as flyback transformers; however, others that are not designed for minimum leakage inductance do not work well.

The total inductance, L, of a winding is the sum of the mutual inductance, L_M, and the leakage inductance, L^σ:

$$L = L_M + L^\sigma$$

The mutual inductance of a winding, L_M, is the product of the total inductance, L, and the coupling coefficient, k:

$$L_M = L \times k$$

Transformers typically have safety isolation between primary and secondary windings, whereas coupled inductors typically do not; however, there are exceptions to both.

SEPIC and Ćuk converters can use coupled inductors with good or little magnetic coupling, or discrete inductors with no magnetic coupling. Energy transfer mainly or completely relies on the coupling capacitors. Voltage scaling between capacitor-coupled windings is not significantly affected by winding leakage inductance; however, for proper operation, coupled inductors must have some leakage inductance so as to maintain continuous or quasi-continuous current and allow the coupling capacitor to drive the ac voltage waveform. Low leakage inductance can increase capacitor size needed to obtain an LC resonant frequency well below the feedback loop unity gain crossover. The main advantage of using coupled rather than discrete inductors is economy in component cost and PCB area.

In the example in this application note, a flyback transformer provides isolation. The turns ratio is 1:1 (other ratios can also be used). Low leakage inductance between the primary and secondary windings is required as it is for any flyback. However, with SEPIC or Ćuk coupling linking multiple outputs in one isolated domain, the current waveforms and voltage cross regulation criteria within a domain become similar to SEPIC and Ćuk converters.

DESIGN CONSIDERATIONS FOR THE COMBINED FLYBACK, SEPIC, AND ĆUK CONVERTER

Design the flyback converter first. Determine the winding output voltage (7.5 V or 15 V, for example), and then refer the total output power to one output for total power calculations.

For example, if the current and power requirements are as shown in Table 4, the transformer and control can be designed as if for a single output: 1.5 W/7.5 V = 200 mA.

Table 4. Domain Power Requirements

Output Rail	Volts	Amps	Watts
V1	+7.5	+0.05	0.375
V2	+15	+0.06	0.9
V3	−7.5	−0.03	0.225
Total	Not applicable	Not applicable	1.5

Operating frequency is related to transformer, inductor, and ceramic capacitor size. Whereas higher operating frequency normally is associated with size reductions, frequencies over 200 kHz to 400 kHz are likely to increase loss and degrade voltage regulation due to transformer leakage inductance.

The minimum practical leakage inductance does not continue to scale inversely with transformer design frequency. The energy stored in the leakage inductance, $L \times I^2/2$, is usually wasted; because power is energy × frequency, the power loss through the leakage inductance scales with frequency.

Aside from power level and frequency, transformer pinout and safety spacings are additional factors that often determine a minimum transformer size. The PQ2625 was chosen for easy hand winding with the Rubadue multilayer Teflon safety insulated wire. The design operates at 200 kHz.

The following are notes concerning the design of the transformer and the power converter:

- For the waveform examples in Figure 13, Figure 14, and Table 3, the duty cycle in both of these figures is approximately 56%. Stable CCM operation with such larger duty cycles (approaching and exceeding 50%) requires more slope compensation. Slope compensation is the addition of a voltage ramp to the current ramp used by the current mode PWM controller IC. Duty cycles under 45% usually need less or no slope compensation, and lend themselves to easier control. Duty cycles in the range of 20% to 45% tend to be the easiest. With 12 V input, 7.5 V output, Schottky diodes, and a 1:1 turns ratios, the demonstration circuit runs at approximately a 40% duty cycle.
- In a size and power optimized flyback transformer with Schottky output rectifiers, continuous conduction mode (CCM) usually provides the best efficiency. Peak transformer flux density must be under 0.2 T to 0.22 T with peak load and minimum input voltage, to avoid core saturation when the transformer is hot. AC peak-to-peak flux density must be as high as possible but is limited by acceptable core loss; therefore, the design in this application note started with an ac peak-to-peak flux density limit of 0.05 T to 0.07 T at 200 kHz in Ferroxcube 3F3 ferrite.

In the example in this application note, using a PQ2625 core, minimum size is not a critical requirement. The transformer was designed for easy hand construction, low leakage inductance with reasonable core loss, and adequate creepage and clearance distances. The main tradeoff is that the core is significantly larger than needed for the power level. The primary, secondary, and output windings all have only four turns. With those few turns, the inductance that is needed for CCM cannot be obtained; therefore, the core assembly was gapped with a thick spacer (0.001 inches, or 0.025 mm), which was cut from polyester film. The resulting inductance was approximately 28 μH in all windings. The converter operates in discontinuous conduction mode (DCM) when it is loaded to normal limits. Note that the desired inductance can be achieved by increasing the turns; however, increasing the turns significantly increases the leakage inductance and is therefore a poor tradeoff.

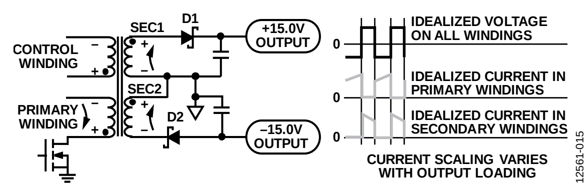

Figure 15. Dual Output Flyback Converter Transformer Winding Currents and Voltages

In a multiple output flyback converter, as shown in Figure 15, the total ampere-turns in the transformer can be continuous; however, the current in the individual windings must vary instantly to maintain waveform fidelity and voltage regulation. Low leakage inductance in the transformer windings is critical. This example has a transformer turns ratio of 1:1:1:1, idealized diodes, and 12 V dc input.

In a normal flyback, all of the windings share one magnetic core, and the core flux is proportional to the total ampere × turn product, scaled by the inverse of the reluctance of the magnetic path. When a separate inductor is used, the result is a separate (not shared) core. The coupling capacitor blocks any dc current flow between the transformer winding and the inductor, so that only ac current passes between the two magnetic components. The capacitor value is large enough that the ac current causes only a small ripple voltage across the capacitor. The capacitor acts as an ac short circuit, and its ripple voltage can be neglected in the simple analysis of circuit operation.

In the normal flyback example shown in Figure 15, the transformer output windings conduct no current when the transistor is on. This is not true of the transformer in the combined flyback and Ćuk converter (or any of the combined topologies) because the output winding needs to drive L1 and any other inductors through the coupling capacitors. The result is that the transformer output winding waveform includes components of both output diode current and inductor magnetization current. The primary winding current waveform looks like that of an ordinary flyback with the inductance being that of the parallel combination of all the magnetic structures. In the Figure 12 example, the transformer inductance measures approximately 28 μH. The Coilcraft LPD6235 coupled inductor similarly has 47 μH; therefore, the converter behaves similarly to a flyback with transformer inductance equal to the parallel equivalent value of 17.5 μH.

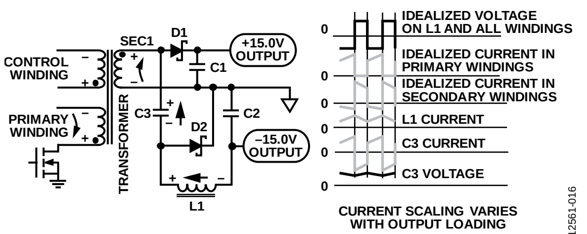

Figure 16. Dual Output Flyback Converter Transformer Winding Current and Voltages for the Single Secondary Topology

In this combined flyback-Ćuk converter, shown in Figure 16, low transformer leakage inductance is still critical to efficient energy transfer, but it has minimal impact on the cross regulation between outputs in one voltage domain. The critical ac component of Diode D2 current passes through Capacitor C3 and not Inductor L1. The critical stray inductance is that measured through the path of D1 and C1 and through the path of C3 and D2. Careful PCB layout must make this stray inductance much lower than the minimum leakage inductance, which can be obtained in a good transformer used in the normal flyback.

Choose the external inductor(s) based on inductance, core loss, dc resistance, and saturation current. Typically, if the peak-to-peak ripple current in an inductor approaches the saturation current rating at frequencies in the range from 50 kHz to 100 kHz, core losses are excessive. If possible, use core loss calculators offered by inductor manufacturers such as Coilcraft. More turns of smaller wire providing higher inductance in a given size decreases the ripple current and the core loss, but decreases the dc saturation current and increases the dc resistance.

In Figure 10, the dc current in L1 is the −15 V output current. In Figure 11, the dc inductor current is the +15 V output current.

In Figure 12, one side of the coupled inductor conducts the +15 V output current while the other side conducts the −7.5 V output current. To determine dc core excitation, these two magnitudes are added. It is not a common-mode choke.

Coilcraft and Cooper offer some small sized coupled inductors (such as the Coilcraft LPD6235) that tend to be single sourced. 12 mm square coupled inductors are made in interchangeable footprints by manufacturers including Pulse, Wurth, Cooper, and Coilcraft.

Coupling capacitor values must first be chosen so that the average charge is large relative to the cyclic charge (equal to I_{OUT}/switching frequency). Then, calculate the charge/capacitance to find the ripple voltage, which must not exceed a small percentage of dc volts; 5% is a good maximum. Note that ceramic capacitors lose a significant amount of capacitance with applied voltage and with time after soldering; therefore, be very conservative with ceramic capacitance ratings. (Murata offers online tools for graphing these coefficients.) This loss of capacitance with applied voltage is especially true for devices that have high C × V ratings in small packages. It is often convenient to use the same capacitor rating for output capacitors and for coupling capacitors.

Figure 17. Prototype Flyback Transformer with Bobbin Extender

Figure 17 shows the transformer using a PQ2625 3F3 with a bobbin extender, providing 10 mm creepage. This is the transformer used in the converter shown in Figure 18.

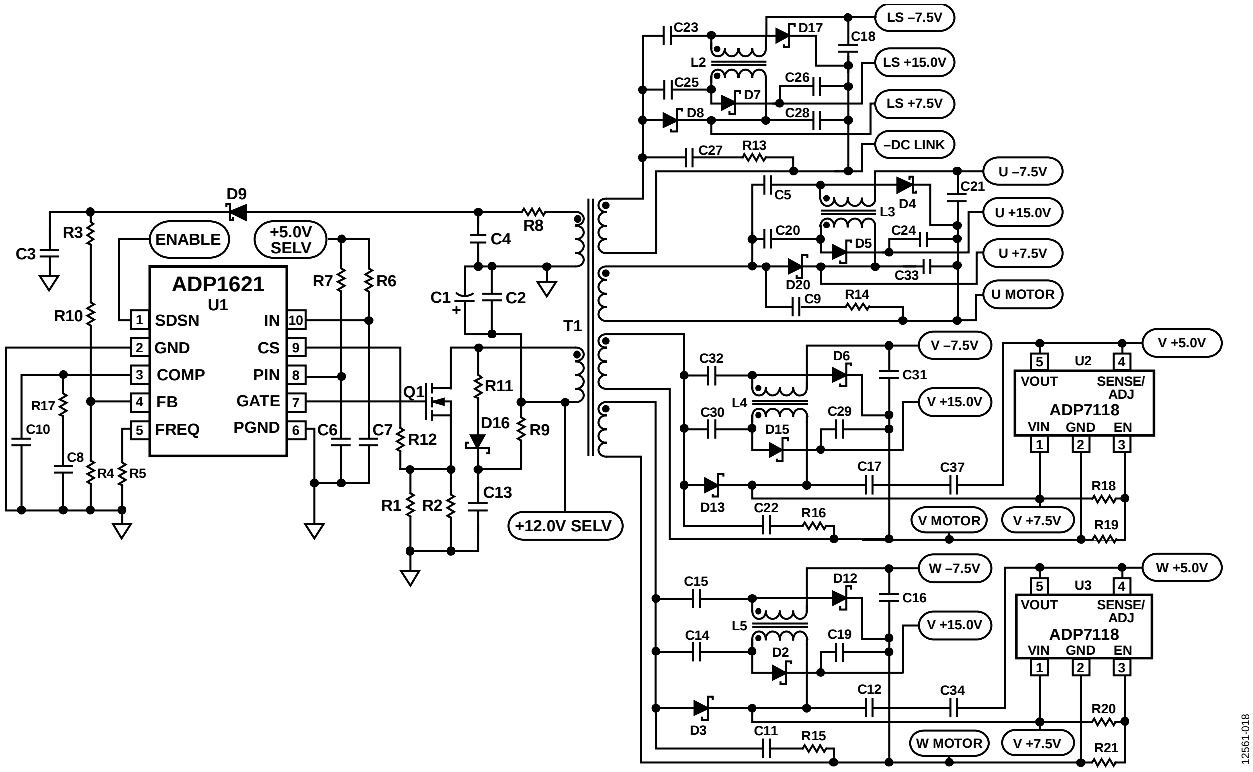

Figure 18. Schematic of the Complete Converter

Table 5. Bill of Materials for the Complete Converter

Item	Reference Designator	Value	Description
1	C1	100 µF	Nichicon UCL1C101MCL6GS
2	C2	1.00E-05	16 V, X5R, 1206
3	C3	10 nF	50 V, X7R, 0603
4	C4	100 pF	50 V, NP0, 0603
5	C5	2.2 µF	0805, X5R, 25 V
6	C6	1.0 µF	16 V, X5R, 0603
7	C7	1.0 µF	16 V, X5R, 0603
8	C8	10 nF	50 V, X7R, 0603
9	C9	100 pF	50 V, NP0, 0603
10	C10	Do not place	Do not place
11	C11	100 pF	50 V, NP0, 0603
12	C12	2.2 µF	0805, X5R, 25 V
13	C13	100 nF	50 V, X7R, 0603
14	C14	2.2 µF	0805, X5R, 25 V
15	C15	2.2 µF	0805, X5R, 25 V
16	C16	2.2 µF	0805, X5R, 25 V
17	C17	2.2 µF	0805, X5R, 25 V
18	C18	2.2 µF	0805, X5R, 25 V
19	C19	2.2 µF	0805, X5R, 25 V
20	C20	2.2 µF	0805, X5R, 25 V
21	C21	2.2 µF	0805, X5R, 25 V
22	C22	100 pF	50 V, NP0, 0603
23	C23	2.2 µF	0805, X5R, 25 V
24	C24	2.2 µF	0805, X5R, 25 V
25	C25	2.2 µF	0805, X5R, 25 V
26	C26	2.2 µF	0805, X5R, 25 V
27	C27	100 pF	50 V, NP0, 0603

Item	Reference Designator	Value	Description
28	C28	2.2 µF	0805, X5R, 25 V
29	C29	2.2 µF	0805, X5R, 25 V
30	C30	2.2 µF	0805, X5R, 25 V
31	C31	2.2 µF	0805, X5R, 25 V
32	C32	2.2 µF	0805, X5R, 25 V
33	C33	2.2 µF	0805, X5R, 25 V
34	C34	2.2 µF	0805, X5R, 25 V
35	C37	2.2 µF	0805, X5R, 25 V
36	D2	MBR0560	Micro Commercial
37	D3	MBR0560	Micro Commercial
38	D4	MBR0560	Micro Commercial
39	D5	MBR0560	Micro Commercial
40	D6	MBR0560	Micro Commercial
41	D7	MBR0560	Micro Commercial
42	D8	MBR0560	Micro Commercial
43	D9	LL101A	Vishay
44	D12	MBR0560	Micro Commercial
45	D13	MBR0560	Micro Commercial
46	D15	MBR0560	Micro Commercial
47	D16	LL103A	Vishay
48	D17	MBR0560	Micro Commercial
49	D20	MBR0560	Micro Commercial
50	L2	LPD6235-473	Coilcraft
51	L3	LPD6235-473	Coilcraft
52	L4	LPD6235-473	Coilcraft
53	L5	LPD6235-473	Coilcraft
54	Q1	IRLML0060	International rectifier
55	R1	Do not place	Do not place
56	R2	0.033 Ω, 5%	0805, Susumu
57	R3	499 kΩ, 1%	0603
58	R4	100 kΩ, 1%	0603
59	R5	100 kΩ, 1%	0603
60	R6	10 Ω, 1%	0603
61	R7	10 Ω, 1%	0603
62	R8	357 Ω, 1%	1206
63	R9	619 Ω, 1%	1206
64	R10	2.00E+04	1206
65	R11	8.2 Ω, 5%	1206
66	R12	200 Ω, 1%	0603
67	R13	Do not place	1206
68	R14	Do not place	1206
69	R15	Do not place	1206
70	R16	Do not place	1206
71	R17	4.99 kΩ, 1%	0603
72	R18	35.7 kΩ, 1%	0603
73	R19	10 kΩ, 1%	0603
74	R20	35.7 kΩ, 1%	0603
75	R21	10 kΩ, 1%	0603
76	T1	Transformer	Described in text
77	U1	ADP1621ARMZ	10-pin MSOP
78	U2	ADP7118AUJZ-5.0	5-pin TSOT
79	U3	ADP7118AUJZ-5.0	5-pin TSOT

CONCLUSION

This application note outlines different methods for generating isolated bias supplies for high-side and low-side gate drives and isolated current sense ICs in industrial motor drives. The advantages and limitations of methods such as charge pumps and bootstrap supplies are addressed, with the conclusion that transformer isolated topologies offer a distinct advantage in terms of efficiency, flexibility, and safety barriers. Flyback topologies are highly suited to the multiple output nature of these bias supplies; however, the standard flyback converter solution for gate driver bias supplies with multiple outputs or dissipative rail splitting suffers from the limitations of high transformer pin usage and poor efficiency, respectively. Moreover, the typical secondary sensing regulation approach suffers from poor cross regulation issues. Solutions are proposed to help mitigate these limitations: primary side sensing, which greatly improves overall cross regulation; the addition of secondary side coupling capacitance between the windings to further improve regulation; and the replacement of a transformer winding with either discrete or coupled inductors to reduce transformer pinout requirements. Results for cross regulation are demonstrated, and a full schematic and bill of materials for the coupled-inductor output version are provided. The discrete inductor version was also implemented on a full, 3-phase inverter platform running at a dc bus up to 800 V.

Figure 19 shows a photograph of the bias circuits on the EV-MCS-ISOINV-Z isolated inverter platform. This platform is available for order from the Analog Devices, Inc., website at www.analog.com/EVAL-ISO-INVERTER-MC.

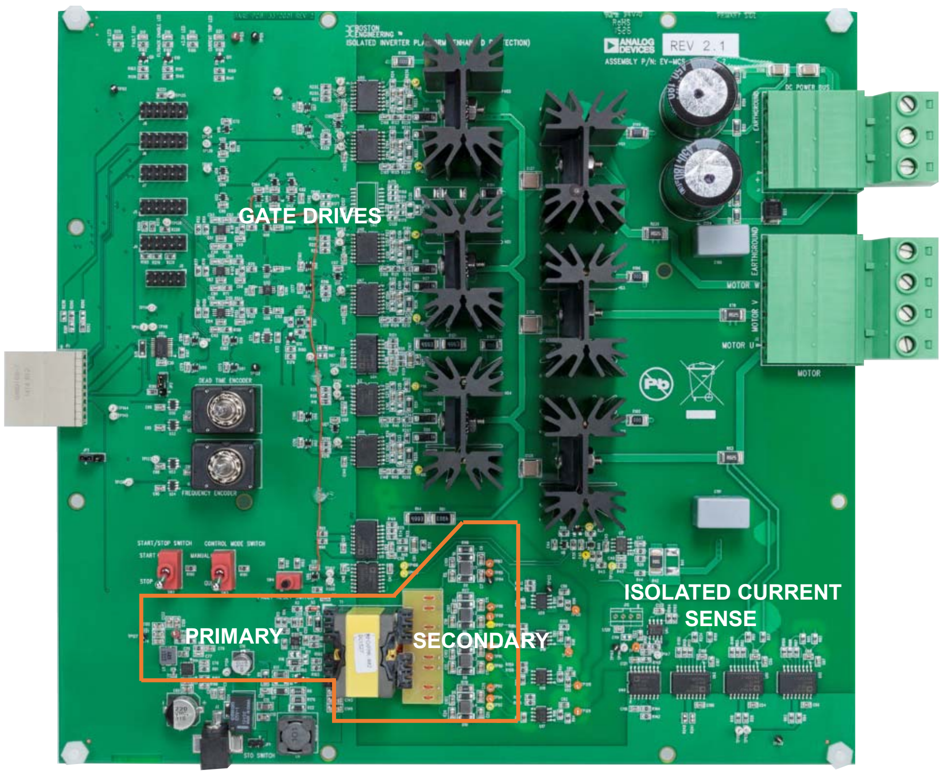

Figure 19. 3-Phase Motor Control Inverter with Bias Supplies

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