参考資料

Design Guide: TIDA-060019

高速、ローサイド電流センスのリファレンス・デザイン

概要

このリファレンス・デザインでは、スイッチング周波数が高いローサイド電流検出アプリケーションにおける広ゲイン帯域幅積 (GBWP) アンプの利点について紹介します。このリファレンス・デザインでは、1mΩ のローサイド・シャント抵抗を監視するために 3 種類のアンプ (OPA365-Q1、OPA836、OPA863、OPA607-Q1) を選択できます。応用例には、モーター用シングル・シャント・ローサイド・センシング、大電流および高速モーター、高周波スイッチング電源、オンボード・チャージャ (OBC)、フォルト検出ハードウェアが含まれます。実装された高速アンプは、広いダイナミック・レンジで電流を監視し、±50A を超える過渡を高い精度で検出するために 33V/V のゲインを持つように構成されています。この構成では、各アンプは大信号ステップ応答を 0.5µs 以内に収束させ、テキサス・インスツルメンツのほとんどのマイコンでモーター制御に使われている 2MSPS、12 ビット内部 ADC のアクイジション期間 (250ns (標準値)) を正確に駆動できます。これらの優れた性能が、低帯域幅デバイスと同様のコストと消費電力で利用できます。

リソース

TIDA-060019	デザイン・フォルダ
OPA2836-Q1、OPA607-Q1、OPA365-Q1	プロダクト・フォルダ

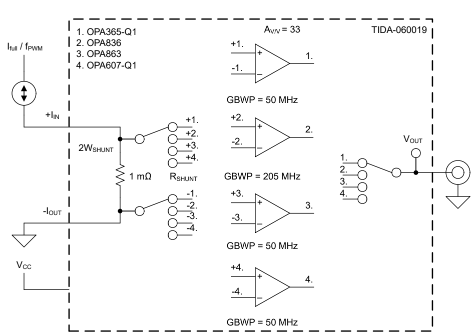
Ask our TI E2E™ support experts

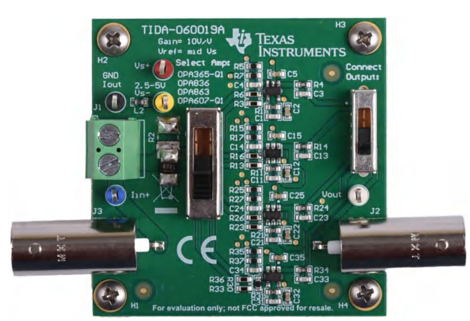
特長

- 4 つのアンプおよび / または回路を比較可能
- 必要に応じてシャント抵抗とゲイン抵抗を構成可能
- アンプの GBWP：
  - OPA365-Q1 = 50MHz
  - OPA836 (OPA2836-Q1 と同じ) = 205MHz
  - OPA863 = 50MHz
  - OPA607-Q1 = 50MHz
- 1% セトリング・タイム：0.5µs 以下
- 2W、1mΩ のシャント抵抗 (2512 パッケージ)
- 2MSPS、12 ビット、3.3V ADC に対応する小信号セトリング・タイム：250ns 以下、0.5LSB
- 総合ノイズ：250µV_{RMS} 未満
- 既存のアプリケーションに接続して迅速な検証が可能

アプリケーション

- 車載用 HVAC (エアコン) コンプレッサ・モジュール
- バッテリ管理システム (BMS)
- コードレス電動工具
- DC/DC コンバータ
- DC 入力 BLDC モーター・ドライブ
- ドライブ・ライン・コンポーネント
- ドローンのプロペラ ESC
- 電動自転車 / 電動アシスト自転車
- 電気駆動
- eTurbo チャージャ
- インバータおよびモーター制御
- オンボード・チャージャ (OBC) およびワイヤレス・チャージャ
- ポンプ





English Document: TIDUEH2

1 System Description

A low-side current shunt is a simple, low cost way to measure wideband current signals up to ± 100 A, with a greater degree of accuracy and dynamic range when compared with magnetic and in-phase current sensing.

In the current shunt circuit, the amplifier's gain ($A_{V/V}$) is required to minimize shunt power (W_{shunt}) and amplify the shunt's voltage signal up to the full ADC input range. A sufficient gain allows a minimized shunt resistance (R_{shunt}) and ensures that a maximized signal-to-noise ratio is being sent to the analog-to-digital converter (ADC). The amplifier's bandwidth is required to measure a short minimum pulse width so that a high pulse-width-modulation frequency (f_{PWM}) can be used to control phase current. This design guide will compare three high gain-bandwidth product devices and review the necessary considerations for fast and accurate current sensing.

Magnetic solutions like Rogowski coils and current transformers are optimized to measure alternating currents within a frequency range, but their bandwidth and dynamic range are constrained by the derivative relationship between magnetic field strength and current frequency. DC currents cannot be measured and can saturate the core of a current transformer. Alternatively, in-phase shunt monitors require an amplifier with an input common-mode voltage range up to and above the maximum system voltage. The large input common-mode step that occurs during switching injects noise into the measurement. This method can add to the complexity and cost of the overall design. A low-side shunt can measure a wide range of current over a wide bandwidth and minimize design complexity and cost.

1.1 Key System Specifications

表 1-1. Key System Specifications

PARAMETER	SPECIFICATIONS
Transient Measurement Range	± 50 A (Vcc = 3.3 V, Gain = 33 V/V, R_{sh} = 1mΩ)
Continuous Shunt Current	≤ 45 A_{rms} (2 W, 1 mΩ) ≤ 63 A_{rms} (2 W, 0.5 mΩ)
Minimum Pulse Width: Gain = 100 V/V	OPA365-Q1: 2 µs
	OPA2836-Q1: 0.5 µs
	OPA863: 2 µs
	OPA607-Q1: 1 µs
Output range (3.3 V supply)	OPA365-Q1: 0.010 V to 3.29 V
	OPA2836-Q1: 0.15 V to 3.1 V
	OPA863: 0.14 V to 3.16 V
	OPA607-Q1: 0.012 V to 3.292 V

2 System Overview

One concern of low-side current sense circuits is that large and/or very fast currents will cause the local ground voltage to change beneath the shunt due to ground plane impedance. Therefore, a difference-amplifier configuration (図 2-1 and 図 2-2) is needed to measure the differential voltage directly across the shunt resistor.

Non-inverting gain and inverting gain give similar results, and either gain polarity can be used to satisfy hardware system requirements.

To measure a bidirectional current with a single positive supply, the output voltage for 0-A must be offset to mid-supply, by providing a reference voltage (図 2-1, V_{REF}) to the amplifier. A convenient way to ensure this reference voltage is quiet and referenced to the same analog ground as the ADC, is to place a voltage divider close to the microcontroller's analog voltage reference inputs, and to use another amplifier (図 2-1, IOP1) to buffer this voltage. The reference amplifier can supply this low noise, low impedance, mid-supply reference voltage to all three low-side shunt monitor amplifiers.

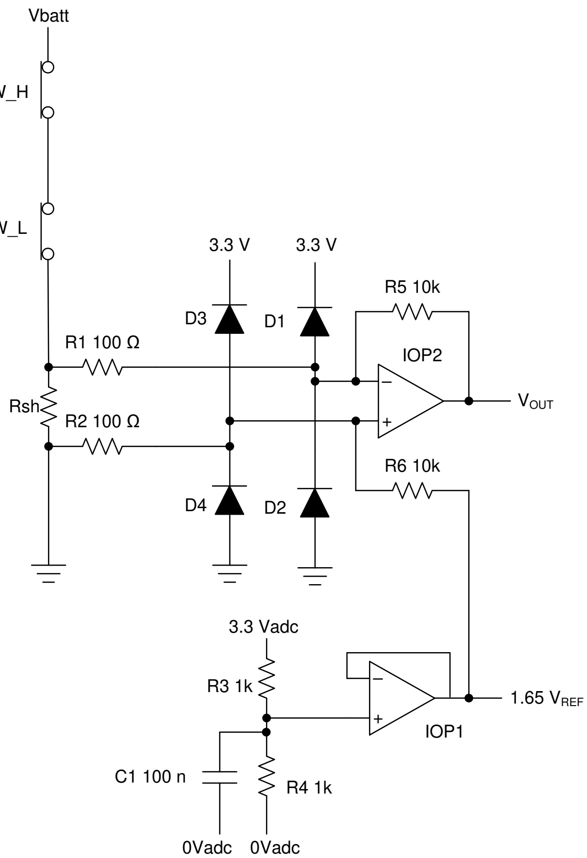


図 2-1. Difference Amplifier with Analog Ground Reference Voltage and Input Protection Diodes

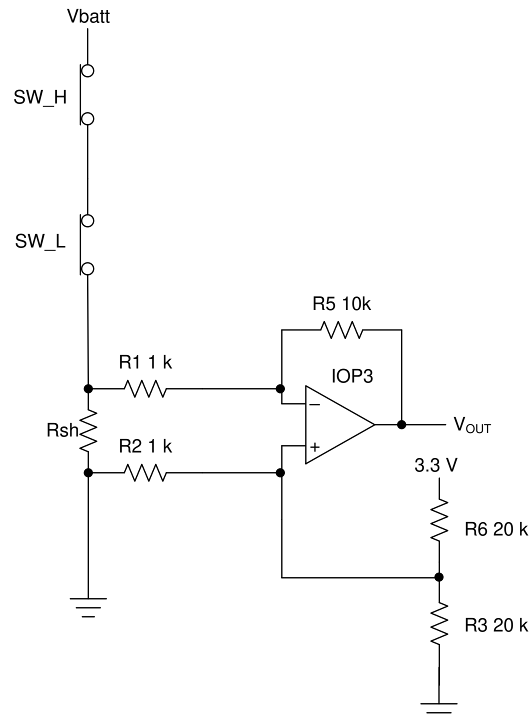


図 2-2. Difference Amplifier with Local Thévinin Equivalent Reference Voltage

In low cost, lower-current applications, the voltage-reference amplifier and 10kΩ series resistor (図 2-1, R6) can be replaced with a Thévenin-equivalent resistor divider, for each amplifier, as has been done in this reference design. The drawback of using an unbuffered voltage divider is that when the amplifier is next to the shunt, the ground bounce caused by the large shunt current might be included in the reference voltage, and thereby offset the measurement sent to the ADC.

2.1 Block Diagram

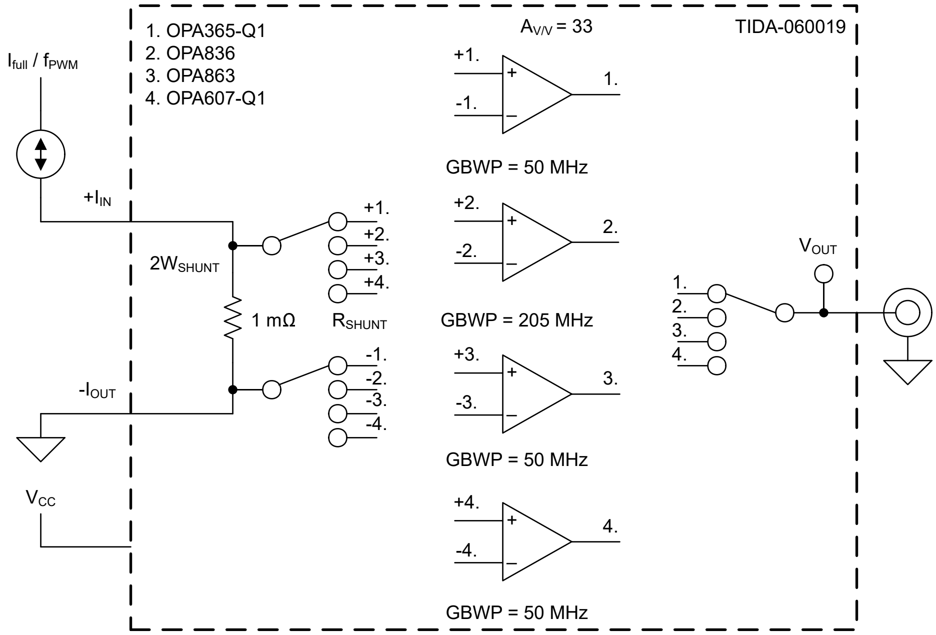


図 2-3. Block Diagram

2.2 Design Considerations

The Shunt Current: I_{full}

The first step in designing a current shunt monitor is determining the maximum current that the shunt resistor will need to conduct. For this design discussion, we will use an example application of a battery-operated motor drive circuit to determine the design of the shunt measurement. In motor drives, the maximum current is the inrush current that occurs at start up. Since inrush is a transient condition from a cold start, the shunt power rating does not have to include the peak inrush current, although the peak may be an important max input value for software-controlled startup or hardware overcurrent interruption based on the low-side shunt. The shunt power derating for inrush current will depend on the motor torque constant, load, and target acceleration. A good rule of thumb is that a motor will draw six times its rated current during start up. Each of the three shunt resistors should therefore be rated to conduct double the full load current. In this example, the full rated load current is 20 A and the shunt resistor current rating is 45 A_{rms} per phase.

The Shunt Power: W_{shunt}

The next step is to determine the shunt power rating based on the available heatsinking. The shunt resistor is a power component and should be given the same thermal consideration as a power switch. The shunt power rating often assumes ideal heatsinking conditions which should be verified in custom hardware. Although a 2512 resistor package is rated for 2 W, the true thermal limit is determined by the PCB heatsinking ability. If unable to effectively sink 2 W of heat, 170 ºC is approaching the melting point of solder, and the resistor solder joint may melt well below its 2 W rating. Temperature, which increases resistance, will also affect measurement accuracy.

An easy way to increase the shunt power rating, space permitting, is to use parallel shunt resistors, since the negative thermal feedback will keep the current evenly distributed, just like for parallel switches. In this example, a single resistor is used, but use of two stacked resistors per phase was successfully tested. This shunt has a <1 % thermal variation up to 170 ºC and a 2 W power rating. With a fast-settling high-speed amplifier to drive the ADC, use of 0.1 % gain-setting resistors can keep this thermal limit as the accuracy floor of the system.

Some have tried to replace the shunt resistors entirely with the Rdson of the low-side MOSFETs. While this is possible in theory, the resistance will vary much more strongly across individual devices than for a 1% shunt resistor, and greater attention to individual device detail will be necessary. Also, this requires the full battery voltage to be applied between the inputs of the difference amplifier when the switch is open, resulting in a much higher amplifier input current. Using a high-speed amplifier can keep a shunt resistor's power sufficiently low for all but the most thermally sensitive applications.

The Shunt Value: R_{shunt}

Once the maximum allowable shunt power has been determined, the next step is picking the right shunt resistor value. To minimize shunt power, this means determining the minimum possible resistance. Theoretically, this is based on the shunt inductance, since fast current edge rates will produce a higher gain from the shunt inductance compared to the DC resistance. A larger shunt resistance can maintain consistent impedance over a higher frequency range. The shunt in this example is the CRE2512-FZ-R001E-2. This 1-mΩ resistor has < 5 nH inductance, suitable for frequencies ≤ 300 kHz.

In practice, a larger minimum resistance is often required due to the amplifier's limited gain at the bandwidth required to measure the minimum pulse width. However, using a high-speed amplifier allows a larger gain and minimizes shunt resistance.

The Amplifier Gain: $A_{V/V}$

The amplifier gain should be chosen to match the maximum software-controlled current to the full ADC input range with some headroom. In battery-based applications, the maximum current is typically around the same maximum current expected from the battery. Meeting the full ADC input range increases the signal-to-noise ratio and increases the resolution of the data used by the motor controller. In this example, a gain of 67-V/V was chosen to convert a ± 20-A (24.6-A ceiling) signal to a ±1.65-V signal for a 3.3-V ADC with a 1-mΩ shunt resistor. A 12-bit ADC results in a measurement resolution of 12 mA.

While it is often not necessary to accurately measure the peak inrush current, the series gain resistors should be large enough to protect the amplifiers' inputs during the large shunt voltage transients. For extremely high inrush shunt voltages, diodes can be placed from each amplifier's input to each supply rail to increase the allowable input current, but this will add leakage current and increase the amplifier’s offset. Scaling up to larger resistors is often sufficient.

The Switching Frequency: f_{PWM}

The motor's phase current frequency in Hz = (RPM / 60 seconds) × (Stator Poles). In a motor controller, reconstructing this sine wave with a higher PWM frequency can reduce ripple and the corresponding vibration noise and mechanical wear on the motor windings, especially at top speeds. To minimize distortion of the phase current, a rule of thumb is that the f_{PWM} should be 60 times the maximum expected phase current frequency. In this example, a 600 RPM, 50 stator pole (23 rotor magnet pole pairs) motor is spun with a 500 Hz phase current frequency. A good switching frequency for this motor is f_{PWM} = 60 × 500 Hz = 30 kHz. Switching frequencies of 15 kHz, 30 kHz, 45 kHz, and 60 kHz are compared in 図 2-4, 図 2-5, 図 2-6, and 図 2-7 for this motor at top speed with 10 A of field weakening current. The OPA2836-Q1 GBWP supports all of these frequencies, and the phase current of the motor controller more accurately reproduces a sinusoid as the f_{PWM} increases.

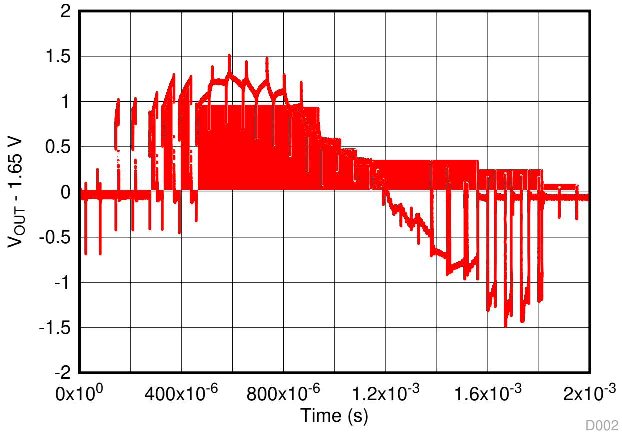


図 2-4. f_{PWM}= 15 kHz, OPA2836 Gain = 67 V/V, R_{shunt} = 1 mΩ

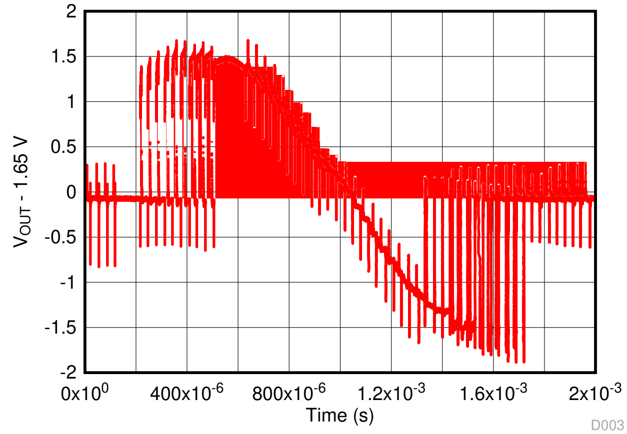


図 2-5. f_{PWM} = 30 kHz, OPA2836 Gain = 67 V/V, R_{shunt} = 1 mΩ

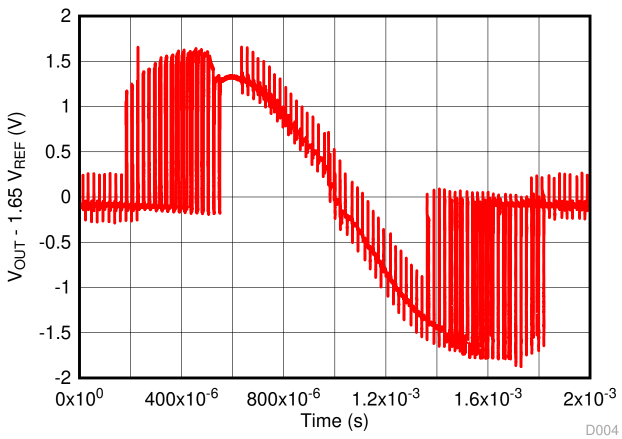


図 2-6. f_{PWM}= 45 kHz, OPA2836 Gain = 67 V/V, R_{shunt} = 1 mΩ

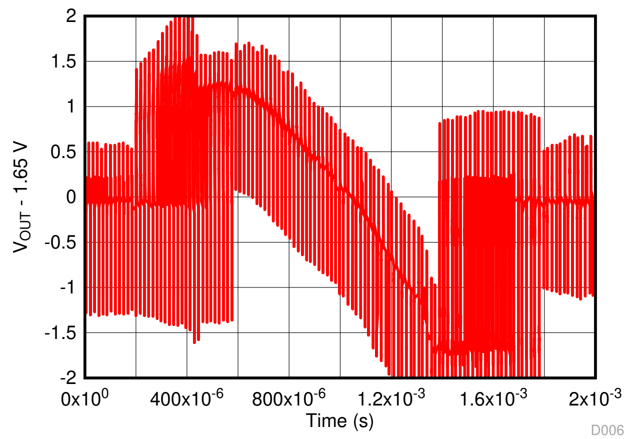


図 2-7. f_{PWM}= 60 kHz, OPA2836 Gain = 67 V/V, R_{shunt} = 1 mΩ

A high f_{PWM} is typically required for high RPM motors and motors with a high stator-pole count due to their high voltage and low inductance. TI offers excellent, high-ampere gate drivers to reduce switching time and switching losses for high-current switches, and in this example a < 100 ns switch settling time is achieved with a 1 us dead time (0.5 us for 60 kHz PWM). Since the shunt is in series with the low-side switch, it produces three types of signals during a cycle: 0 V when the switch is open, pure current when the switch is closed, and PWM current when the switch is modulated at the PWM frequency. The most challenging measurement occurs during the transition from open switch to PWM region, where the minimum pulse-width occurs (D_{min}). The motor will still work with an amplifier GBWP below the minimum pulse width requirement, being able to power through a fraction of missed measurements, but this will begin to degrade the accuracy of the motor control and limit the top speed that can be achieved.

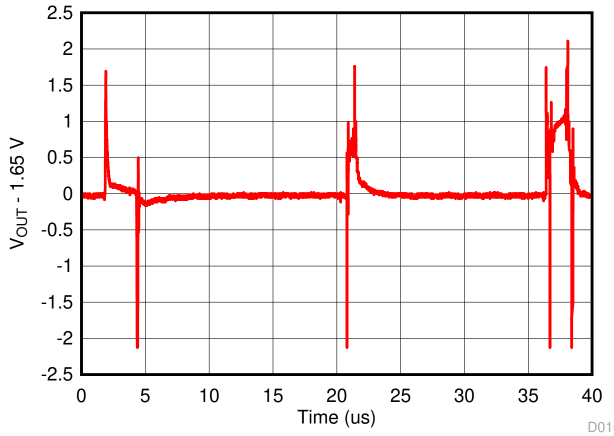


図 2-8. < 1 us Minimum Pulse Width using OPA2836-Q1, Gain = 67 V/V, 60 kHz PWM, 5% Duty Cycle

At top speed, the minimum pulse measurement is its most important, being a larger fraction of the total measurements. With a f_{PWM} = 60 × the electrical frequency, there will be 15 PWM periods between 0 % and 100% duty cycle . This implies a minimum duty cycle of 1/15 = 6.7%. Designing for a 5% duty cycle will accommodate slightly slower speeds and simplify calculation.

The Amplifier Gain Bandwidth Product: GBWP

The amplifier's gain-bandwidth product should exceed (switching frequency) × (Amplifier gain) / (Minimum Duty Cycle Ratio) as shown in 式 1.

$$\mathrm{GBWP} > \frac{f_{PWM} \times A_{V/V}}{\mathrm{Min}(D)} \quad (1)$$

Additional bandwidth will improve settling time and increase the number of accurate data points available in a minimum width pulse. In this example, the maximum f_{PWM} is 60 kHz, the minimum duty cycle is ~5%, and the amplifier gain is 67 V/V. This requires an amplifier with a gain-bandwidth product of at least 120 MHz. The OPA2836-Q1 satisfies this requirement with a large signal gain-bandwidth product of 120 MHz. Additionally, the OPA2836-Q1 features a low noise of 4.6 nV/rtHz to maximize the SNR of the system.

With these design considerations, a simplified equation can determine the required gain-bandwidth product for a current shunt amplifier. A f_{PWM} = 60 × electrical frequency, a three-phase system, a shunt power rated for 6 × full load current at inrush, and a 1.65 × amplifier gain derating that leaves 65% measurement headroom from a mid-supply referenced 3.3-V ADC input is assumed.

$$A_{V/V} = 1.65V_{adcin} / (I_{full} \times R_{shunt} \times 1.65 \text{ headroom derating}) \quad (2)$$

$$R_{shunt} \le W_{shunt} / (6_{inrush}/3_{phases} \times I_{full})^2 \quad (3)$$

$$A_{V/V} \ge 4 \times I_{full} / W_{shunt} \quad (4)$$

$$\mathrm{GBWP} \ge (f_{PWM}) \times (A_{V/V}) / (D_{min}) \quad (5)$$

$$\mathrm{GBWP} \ge (60 \times \mathrm{RPM}/60\mathrm{s} \times \text{Stator Poles}) \times (4 \times I_{full} / W_{shunt}) / (0.05) \quad (6)$$

$$\mathrm{GBWP} \ge (\mathrm{RPM} \times \text{Stator Poles}) \times I_{full} / W_{shunt} \times 80 \quad (7)$$

$$\mathrm{GBWP} \ge (f_{PWM}) \times I_{full} / W_{shunt} \times 80 \quad (8)$$

These are estimates and should be verified based on design needs, but they provided a good starting point in picking the right amplifier. Extra amplifier gain-bandwidth is beneficial to further reduce R_{shunt} to the next available standard value.

eScooter

For RPM = 600, Stator Poles = 50, I_{full} = 20 A and W_{shunt} = 2 W, the amplifier's gain-bandwidth product should be at least

$$R_{shunt} \leq 1.25\ m\Omega,\ A_{V/V} \geq 40\ V/V \tag{9}$$

$$GBWP \geq (600 \times 50) \times 20 / 2 \times 80 = \mathbf{24\ MHz} \tag{10}$$

eBike

To produce more power using a similar motor, the voltage can be increased from 42 V to 70 V. The RPM is now (70 / 42) × 600 = 1000, and the new motor current is 25 A. The shunt resistance must be reduced, the f_{PWM} increased, and the amplifier's gain-bandwidth product should now be

$$R_{shunt} \leq 0.8\ m\Omega,\ A_{V/V} \geq 50\ V/V \tag{11}$$

$$GBWP \geq (1000 \times 50) \times 25 / 2 \times 80 = \mathbf{50\ MHz} \tag{12}$$

Propeller ESC

For a 5 kW rated propeller motor, there are 12 stator poles (7 rotor magnet pole pairs). The target RPM is 8000, and the full motor current is 45 A. The W_{shunt} = 3W. The amplifier's gain-bandwidth product should now be

$$R_{shunt} \leq 0.37\ m\Omega,\ A_{V/V} \geq 60\ V/V \tag{13}$$

$$GBWP \geq (8000 \times 12) \times 45 / 3 \times 80 = \mathbf{115.2\ MHz} \tag{14}$$

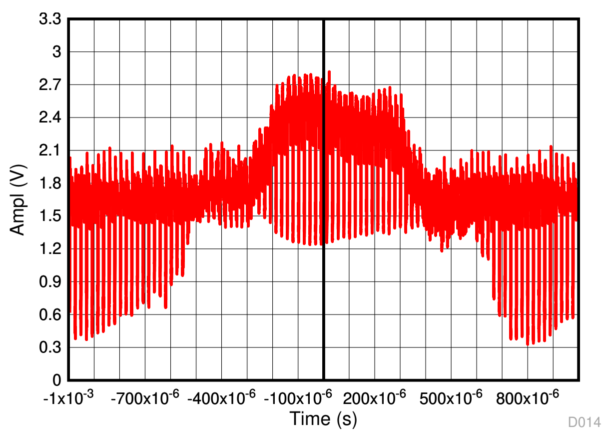


図 2-9. OPA2836 Gain = 67 V/V, 45 kHz PWM, R_{shunt} = 1/3 mΩ, 5kRPM

Inrush Current

Before a brushless DC motor starts to rotate, the inverter circuit forms a synchronous buck converter with a shorted output, with the motor windings acting as the inductor (図 2-10). This creates a high starting current condition known as inrush.

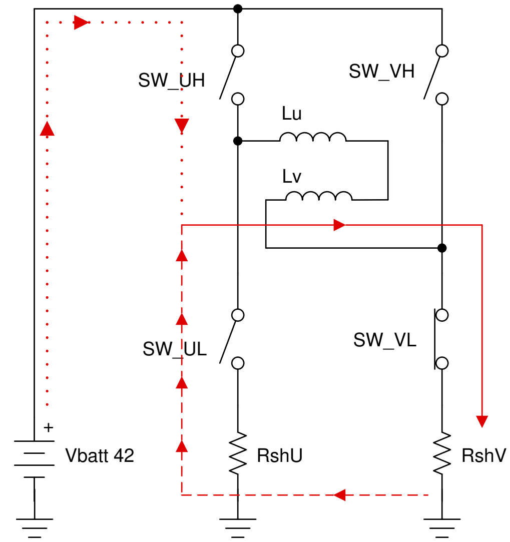


図 2-10. Synchronous ¼ Buck Converter with V_{in}=42 V and V_{out} = Short

In this example, Hall-Effect sensors have given the initial fixed rotor position for sensor-controlled start up. In phase V, the high-side switch (SW_VH) is permanently open and the low-side switch (SW_VL) is permanently closed. In phase U, the high-side switch (SW_UH) is pulsed with a fixed duty cycle of 25% and the low-side switch (SW_UL) is pulsed with a complimentary 75% duty cycle. During the 25% cycle, the full voltage of the battery quickly ramps up the current through the series U and V stator pole windings. During the 75% cycle, this winding current circulates in a loop (through ground and backwards through RshU and SW_UL) at a constant, incrementally increasing value. The windings, switches, and shunt resistors produce only a small opposing voltage due to series resistance. Since this high current circulates in a loop between the motor and the switches, a battery fuse cannot detect or protect the motor inverter from the high current in this stalled condition. The current through the phase V low-side shunt (RshV) here is four times higher than the current from the battery, which only supplies current during 25% of the cycle.

The large current forms a large magnetic field in the windings, and if the magnetic force exceeds the inertia of the motor, the rotor begins to accelerate. The U and V windings acquire an increasing voltage based on the rotor speed to oppose the current, the Hall-Effect sensor outputs update based on the new rotor position, and SW_VL opens, extinguishing the V winding current into the high voltage of the battery and allowing RshV to cool down. Typically, the motor accelerates quickly, and the winding voltage increases towards 25% of the battery voltage. This voltage is detected by the microcontroller, which switches to field-oriented control and limits the motor current to ≤ 20 A for the rest of operation (図 2-11).

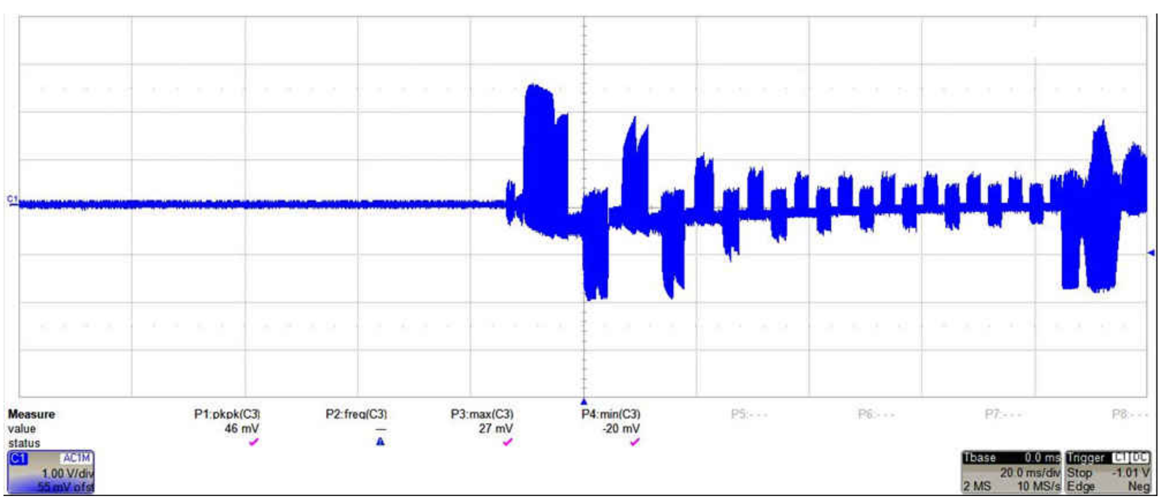


図 2-11. >100-A Peak Measured Inrush Current: V_s = 5 V, V_{ref} = 1.65 V, OPA2836-Q1 $A_{V/V}$ = 67, R_{shunt} = 0.5 mΩ,

However, for heavy loads, high attempted acceleration, or a mismatched motor, the winding voltage will not increase quickly enough to stabilize the winding current. Although this current will be shared between the three shunt resistors, a high and sustained three phase current load will still cause the shunt resistors to overheat and fail. Low-side shunt monitors accurately measure inrush current and can be used in soft-starting and power-factor-correction techniques to sufficiently reduce this transient value with varying degrees of complexity.

2.3 Highlighted Products

2.3.1 OPA365-Q1 (50 MHz)

OPA365-Q1 is the industry standard for shunt monitoring applications. It has 50 MHz GBW, with a 2.2 V to 5.5 V operating single-supply range. It has 4.5 nV/√Hz of broadband noise and 100 µV of offset with 1 uV/ºC of drift.

2.3.2 OPA607-Q1 (50 MHz)

The OPA607-Q1 has 50 MHz of GBW and 3.8 nV/√Hz of noise. OPA607 has 100 µV typical offset and 0.3 µV/ºC of drift. OPA607 is optimized for lowest cost for this application and for gains > 7 V/V. OPA607 is not suitable for lower gains.

2.3.3 OPA836 (same as OPA2836-Q1) (205 MHz)

The OPA836 is a bipolar operational amplifier with 110 MHz of large signal GBW (205 MHz small signal). It has 4.6 nV/√Hz of broadband noise, a 65 µV typical offset and 1 µV/ºC of drift. The 560 V/µs slew rate can be applied in hard-wired, comparator-based current interruption schemes to interrupt faults faster than the sampling rate of the ADC. The OPA2836-Q1 is a dual amplifier in an automotive-qualified 8-pin VSSOP package.

2.3.4 OPA863 (50 MHz)

The OPA863 is a bipolar operational amplifier with 50-MHz GBW. The device has 5.9 nV/√Hz of broadband noise, a 400-µV typical offset, and 1 µV/ºC of drift. The OPA863 is optimized for a faster overdrive recovery time from the negative rail.

3 Hardware, Software, Testing Requirements, and Test Results

3.1 Hardware Requirements

- Function Generator with BNC cable (or current path if using on-board shunt)
- Oscilloscope with BNC cable
- DC bench power supply with hook connector cables
- 0603 gain resistors (Rg) for each gain tested. The feedback resistance (Rf) is 1 kΩ for <20V/V gain and 10 kΩ for high gain. To account for the function generator's 50 Ω source impedance, the IN+ gain resistor should be 50 Ω less than the IN- gain resistor.
  – **Gain, Rg:** 1 V/V, 1 kΩ; 5 V/V, 200 Ω; 10 V/V, 100 Ω; 20 V/V, 200 Ω; 200 V/V, 50 Ω

3.2 Test Setup

1. Adjust or remove the feedback and gain components to provide reduced or additional gain and filtering. When using a feedback capacitor, be aware that an integrator amplifier circuit can be created unintentionally. The following measurements use no external capacitors (See the OPA607 low gain stabilization, documented in the *Decompensated Amplifier Stabilization Circuit* design to use the OPA607 for < 5 V/V gain).
2. 図 3-1 shows the red test point supplies V_{S+} and the yellow test point supplies V_{S-} to all four amplifiers. The black test point is connected to chassis ground, and is shorted to the yellow test point V_{S-} by a ferrite bead L2. L2 can be opened to enable a bipolar supply for use with ground-referenced test equipment.
3. The total supply should be no higher than 5 V to avoid damaging the amplifiers. To avoid damage to a typical target ADC, this test uses 3.3 V_S. The default amplifier output reference is mid-supply; for zero current, the amplifier has an output of 1.65 V.
4. Set the oscilloscope input to high-impedance mode and attach the BNC output (labeled J2) to the oscilloscope. Alternatively, the test point V_o can be used with a hook-probe lead.
5. Attach the function generator to the J3 BNC input.
6. Set both the input slide switch and the output slide switch to the same incremental position to enable the corresponding amplifier. When both switches are in position 1, the OPA365-Q1 is enabled.
7. Power-on the system and use the oscilloscope to compare the rise times, noise, THD, and overdrive recovery time of the four different amplifier devices and or circuit configurations.

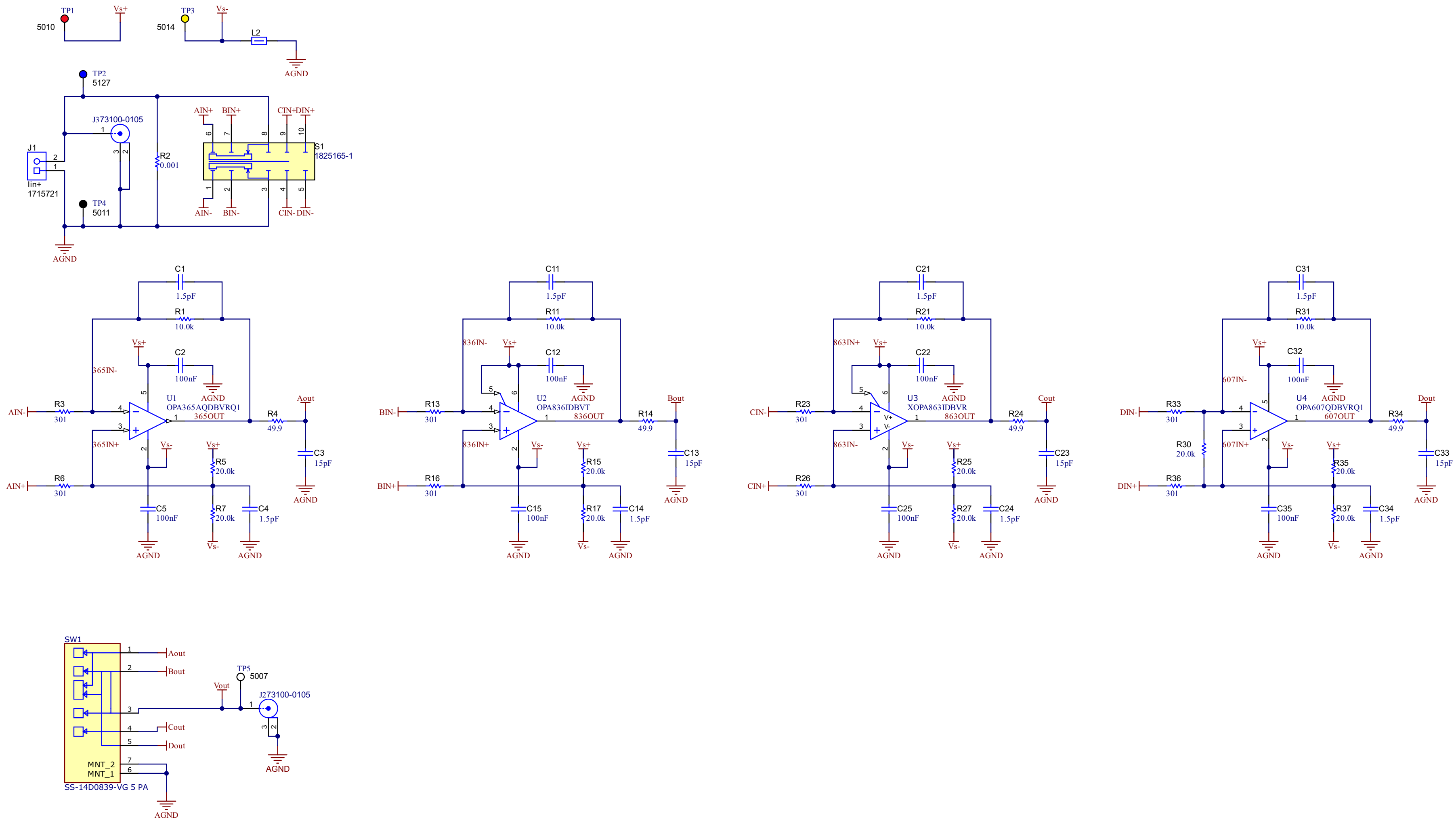


図 3-1. TIDA-060019A Schematic

3.3 Test Results

These test results show how accurately each amplifier can settle within the minimum pulse width for several gain increments. For more information pertaining to 図 3-2, see the *Decompensated Amplifier Stabilization Circuit* design.

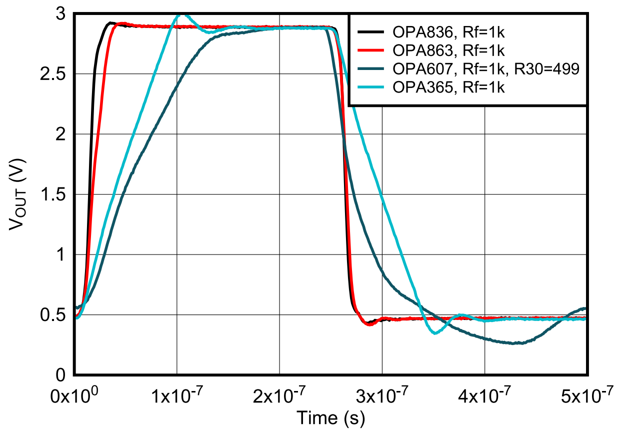


図 3-2. 1 V/V Gain, 250-ns Pulse

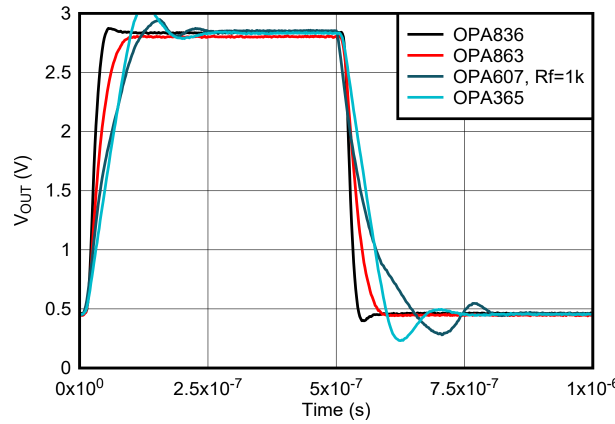


図 3-3. 5 V/V Gain, 500-ns Pulse

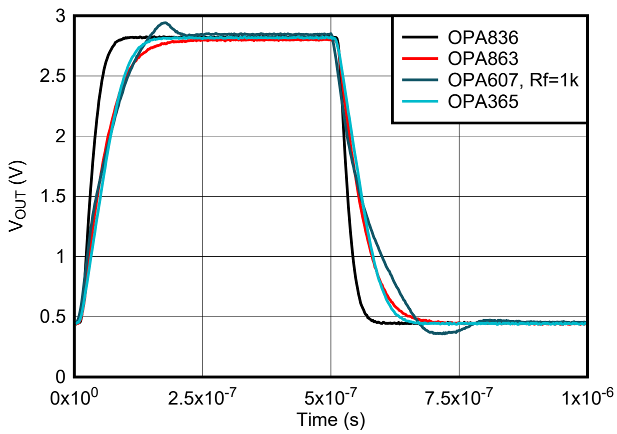


図 3-4. 10 V/V Gain, 500-ns Pulse

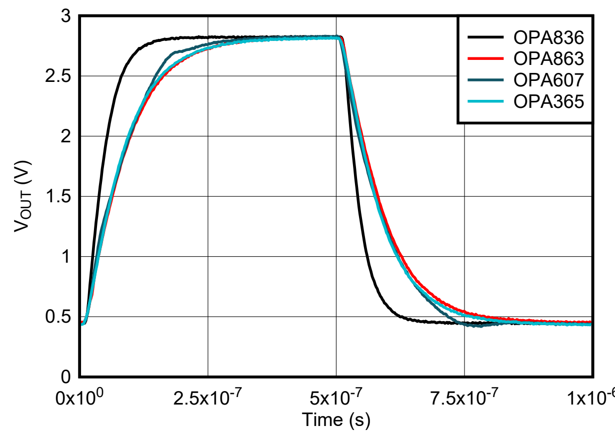


図 3-5. 20 V/V Gain, 500-ns Pulse

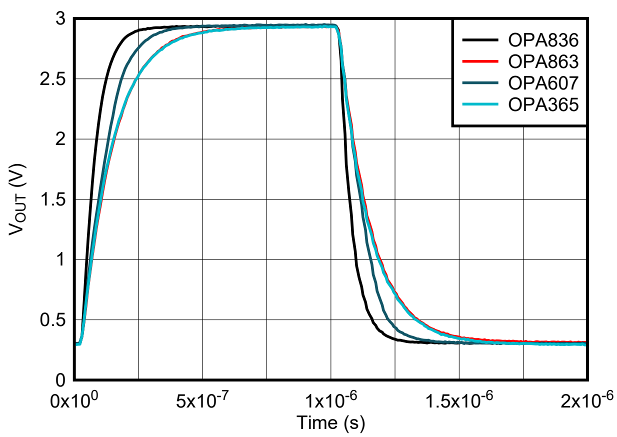


図 3-6. 33 V/V Gain, 1µ ns Pulse

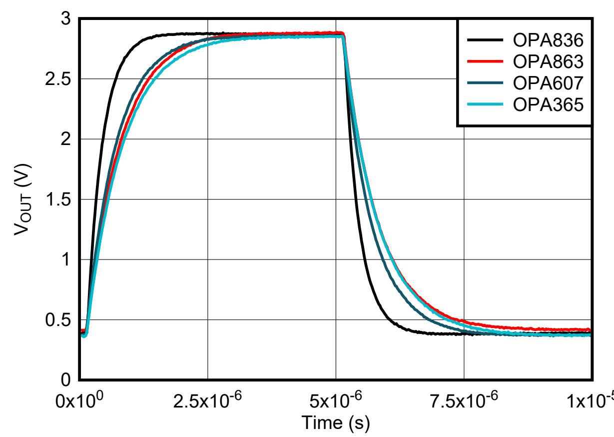


図 3-7. 200 V/V Gain, 5µ ns Pulse

English Document: TIDUEH2

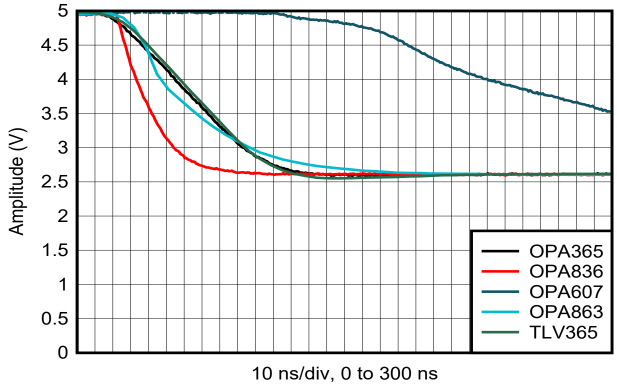


図 3-8. +1.5 V_{OUT} Overdrive Recovery

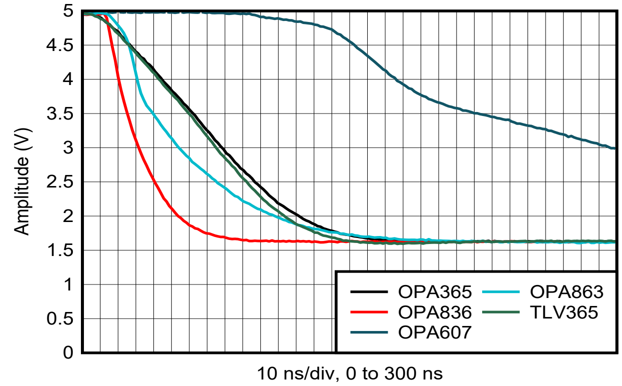


図 3-9. +0.5 V_{OUT} Overdrive Recovery

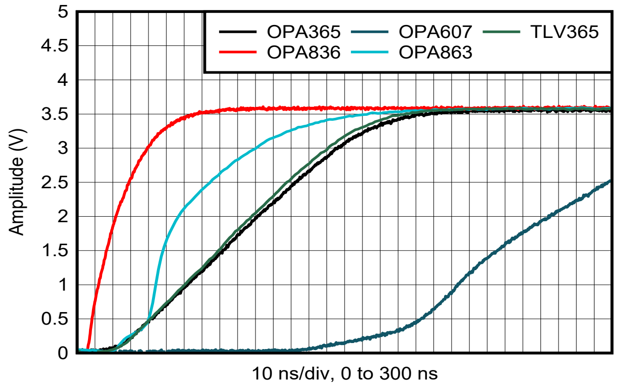


図 3-10. –0.5 V_{OUT} Overdrive Recovery

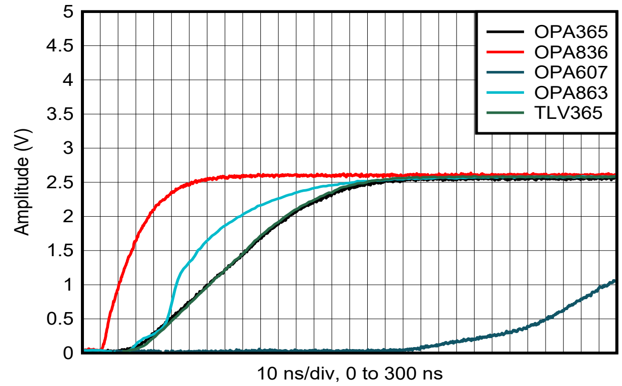


図 3-11. –1.5 V_{OUT} Overdrive Recovery

4 Design and Documentation Support

4.1 Design Files

4.1.1 Schematics

To download the schematics, see the design files at TIDA-060019.

4.1.2 BOM

To download the bill of materials (BOM), see the design files at TIDA-060019.

4.2 Documentation Support

1. Texas Instruments, *Low Voltage, 50A Sensorless FOC Reference Design for PM or BLDC Motors* reference design
2. Texas Instruments, *DRV8301-69M-KIT* evaluation kit
3. Texas Instruments, *OPA607 Low Gain Stabilization* circuit design

4.3 サポート・リソース

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5 About the Author

Sean V. Cashin has been an applications engineer in High-Speed Amplifiers for 5 years. He received his B.S. EE in 2017 from the University of Illinois at Urbana-Champaign.

6 Revision History

資料番号末尾の英字は改訂を表しています。その改訂履歴は英語版に準じています。

Changes from Revision B (October 2021) to Revision C (September 2022)	Page
• Added OPA863 to device comparison in セクション 2.3.4	11
• Updated セクション 3.3 graphs for transient response and added overdrive recovery	13

Changes from Revision A (April 2021) to Revision B (October 2021)	Page
• 出版タイトルを更新	1
• Updated block diagram image	4
• Updated *OPA2836 Gain = 67 V/V, 45 kHz PWM,* R_{shunt} *= 1/3 mΩ, 5kRPM* image	4

Changes from Revision * (December 2020) to Revision A (April 2021)	Page
• 「概要」を更新	1
• OPA607 を OPA607-Q1 に変更	1
• 特長を追加	1
• 「アプリケーション」を追加	1
• Updated Design Considerations to include switching frequency comparison	4
• Updated Test Results images to include setting time comparison	13
• Added Low Voltage, 50A Sensorless FOC Reference Design for PM or BLDC Motors	15

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