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Development of a Ripple Rejection Controller for DC Commutated Motors

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DEVELOPMENT OF A RIPPLE REJECTION CONTROLLER FOR DC COMMUTATED MOTORS

by

Spencer M. Atkin

A report submitted in partial fulfillment
of the requirements for the degree

of

MASTER OF SCIENCE

in

Mechanical Engineering

Approved:

Dr. R. Rees Fullmer Major Professor	Dr. Thomas Fronk Committee Member

Dr. Don Cripps
Committee Member

UTAH STATE UNIVERSITY
Logan, Utah
2012

ABSTRACT

Development of a Ripple Rejection Controller for DC Commutated Motors

by

Spencer M. Atkin, Master of Science

Utah State University, 2012

Major Professor: Dr. R. Rees Fullmer
Department: Mechanical Engineering

The requirement of modern systems to be quiet and run smoothly increases production costs. The manufacture and purchase of high quality motors that meet these demands becomes increasingly costly. By using the ever increasing computer power available in micro-controllers, at the same costs, it is possible to use current sensing to develop control adjustments which decrease power ripples caused by the commutators of basic DC commutated motors. These ripples if left unmitigated propagate through to torque ripples which then increase the acoustic noise level present.

(28 pages)

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INTRODUCTION

The Automotive industry has been pushing the smooth and quiet ride of cars to new standards. As these standards have developed and new levels of quality have been set it has become apparent that new standards for quiet operation are needed for some parts that previously had negligible impacts on the overall sound level. Muffling and other methods have been used to reduce the acoustic levels associated with automobiles which historically have been caused by the engine. When the power window was introduced to automobiles the acoustic background noise was great enough that the sound caused by the electric motor was magnitudes less and therefore, negligible. Since the Automotive industry is one of the largest and most competitive they have had the resources and drive to produce quiet designs in which the sound produced by the electric motor becomes significant. As stated by Dr.-Ing. Wolfgang Schulter, these noises can be largely contributed to the “current ripple phenomena” which is present in DC commutated motors (6). A typical commutated motor works by use of brushes which carry current to the commutation surfaces which in turn power the motor. As the motor rotates the brushes touch either one or two contacts. The repetition of this process causes the current ripple which is seen on such motors. These current ripples cause torque fluctuations which in turn cause changing forces and impacts creating acoustic noise sources. Since cars already need to employ passive systems to reduce exterior, non-controllable, sounds, only lower frequency noises need to be considered. Higher frequency noises are easily damped out by enclosing the motors in absorbent materials which unfortunately have little effect on lower frequencies.

Thesis: The acoustic noise levels of the motor can be reduced by sensing the current draw and controlling the input voltage without adding hardware nontypical to the system. In short the only feedback data available to the controller is the current draw.

LITERATURE REVIEW

The literature in the field of acoustic noise reduction of commutated motors has been sparse due to the fact that normally when smooth operation is required, brushless motors have been used despite the higher cost. There has been some work done in the field of motor control based off ripple counting. These methods are focused on determining motor position and are considered unreliable (Nottelmann).

In 2011 Schulter developed a model of a DC commutated motor (Schulter). Dr. Schulter's model shows that the majority of the acoustic noise produced by this type of motor can be attributed to this ripple effect and that the sensing of the ripple with sufficient computational power can be used to reduce the ripple's amplitude and the subsequent sound.

PHYSICS MODEL

DC motors are closely modeled by a linear model, unlike the true motor the linear model has no position dependence. The equation for the current draw can be written as (Schulter,2):

$$I_a(s) = \frac{V_m(s)}{(1+sT_a)R_a}\left[1 - \frac{1}{1+s\frac{2\pi J R_a}{c_1 c_2 \Phi^2}+s^2\frac{2\pi J T_a R_a}{c_1 c_2 \Phi^2}}\right] \qquad \text{Equation (1)}$$

Table (1) Equation terms

Ra	Armature impedance
La	Armature inductance
Φ	Magnetic Flux
C1,C2	Motor constants
J	Mass Moment of Inertia
Ta	La/Ra

This can be simplified in the steady state as s goes to infinity producing the following equation.

$$I_a = \frac{V_m}{R_a} \qquad \text{Equation (2)}$$

As stated by Schulter the non-linear aspects of the commutation process are what cause the ripple in question. Figure 1 depicts a typical commutator and Figures 2 (a) and (b) show how the resistance path changes as the commutator rotates. A typical commutated motor works by use of brushes which carry current to the commutation surfaces which in turn power the coils that create magnetic fields to produce torque on the shaft of the motor. As this torque rotates the motor the brushes come in contact with a different set of commutation surfaces. The majority of the time each brush touches two commutation surfaces however during every transition it touches only one contact for a brief instant. During this transition the current path changes and the internal resistance and inductance change. This in turn alters the current draw by the motor. This is depicted in Figure (2). Experimental results are shown in the Experimental Test Setup section to compare the model to the physical system.

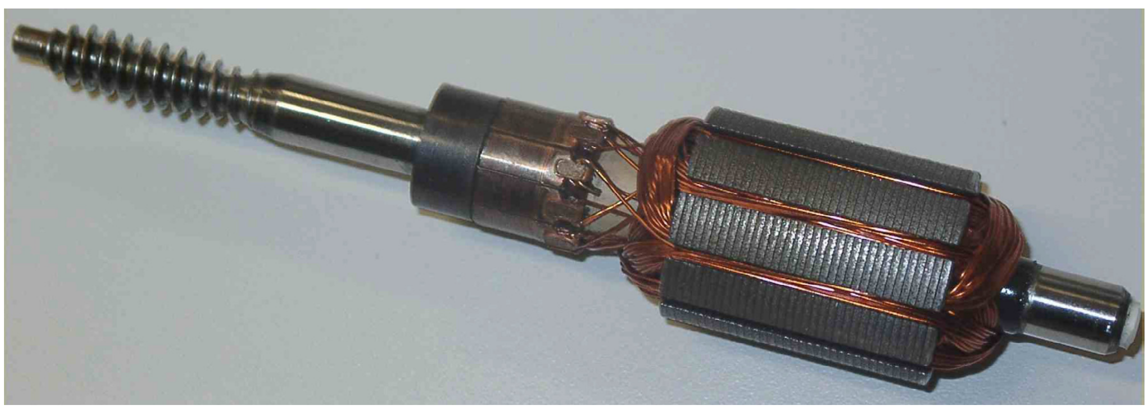


Figure 1: Typical DC-motor rotor (reproduced from Schulter with permission)

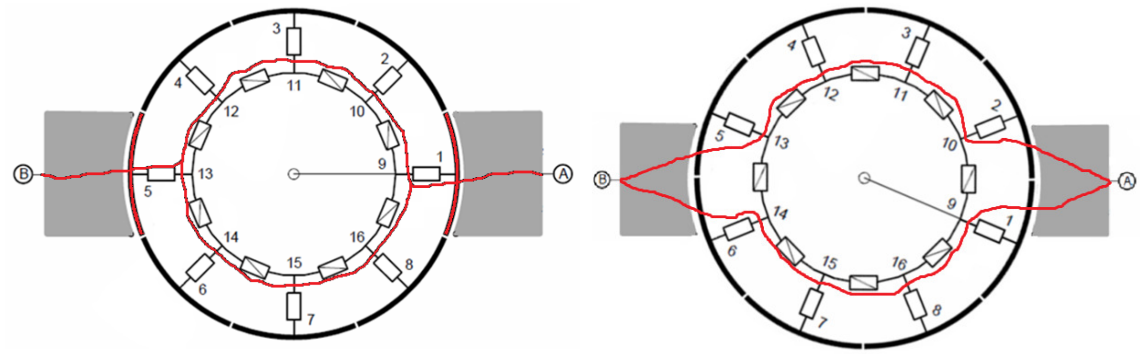


Figure (2) (a) Commutator without overlap (b) Commutator with overlap (adapted from Schulter with permission)

Figure (2) shows on a typical commutator how the alignment causes overlap with the resistive and inductive path it produces depicted as sketched lines. The open boxes represent resistive elements while the slashed boxes represent inductive and resistive elements such as windings.

SIMULATION MODEL

For the initial design and understanding of the system, the Simulink model adapted from the model developed by Prof. Schulter is used. This model (Figure 3) incorporates the nonlinearities of the commutator which cause these current ripples.

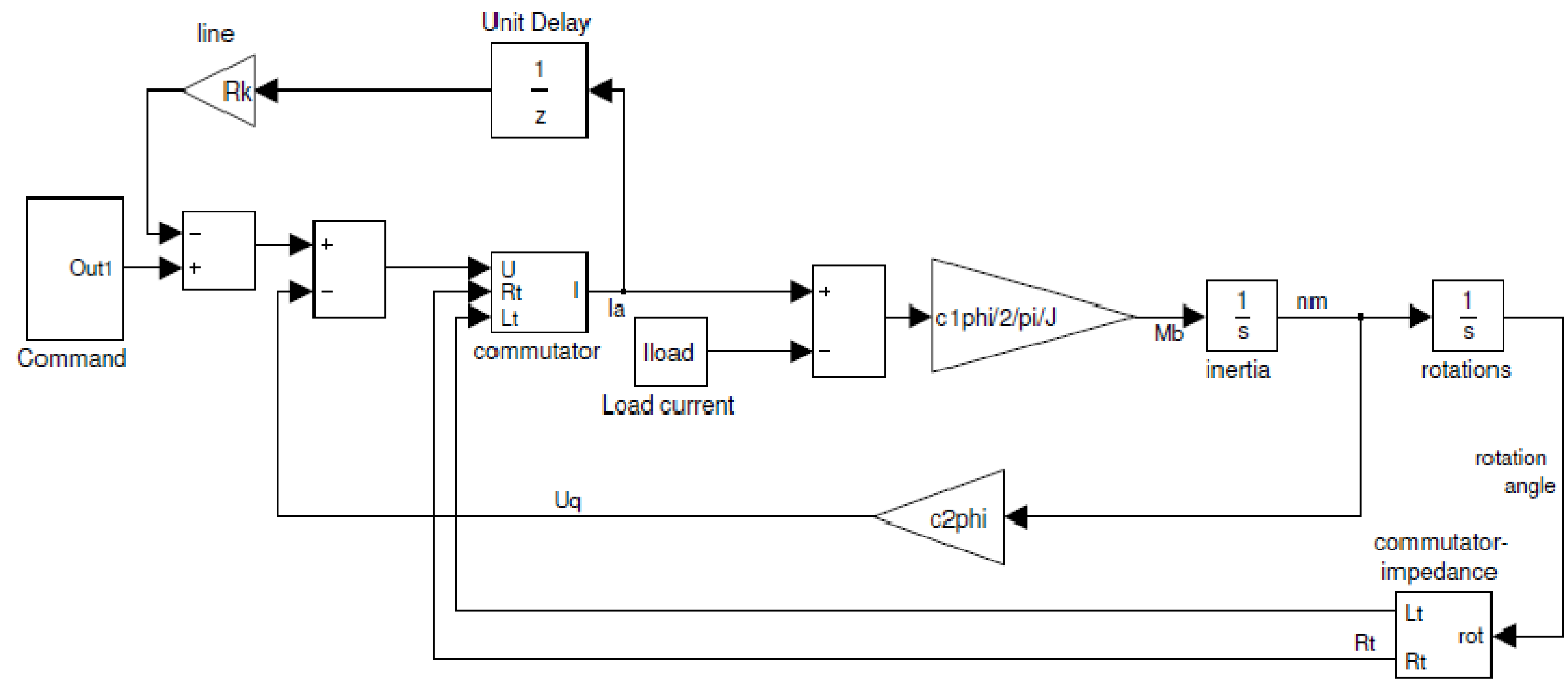


Figure (3) Model of motor including effects of commutation

The importance aspects of this model are contained in the commutator. The model outputs its rotational position from which it determines the respective resistance and inductance. This is done by determining whether the brushes would be in contact with one or two surfaces. The Impedance and Inductance are then fed into the commutator model itself where Equation (1) as derived by Schulter is applied. The equation for the commutator is:

$$I(t) = \frac{V(t) - \dot{I}(t) * L(\theta)}{R(\theta)} \qquad \text{Equation (3)}$$

Where the current, its derivative and the Voltage are functions of time, while the Inductance and the Impedance are functions of motor position. Here it is assumed that under effective control, where current is constant, inductance is irrelevant as $\dot{I}(t)$ is zero so inductance has no effect on the current. And the equation becomes:

$$I(t) = \frac{V(t)}{R(\theta)} \qquad \text{Equation (4)}$$

EXPERIMENTAL TEST SETUP

The Objective of the testing is to first determine the similarity of the proposed simulation model to a physical motor. Firstly it is necessary to run the motor without any control so that this system may be compared to the free running simulation. The testing is used to apply the controller models. The results from the controlled tests can then be compared to check that the physical system reacts similar to the simulation.

The test rig employs Matlab\Simulink real time toolbox to program and load the controller, which is discussed in the next section, to run in real time on a 32-bit Dell Optiplex 790 with an Intel i3-2100 3.10GHz processor and 2.00 GB RAM. The operating system is Windows 7 Professional. The data acquisition system used is a National Instrument PCI-6251 with a BNC 2120 connector block. The PCI-6251 is a 16-bit, 1MS/s (Multichannel) Data Acquisition system (Appendix B). The combined limitations of interfacing between Matlab the computer and the Data Acquisition system restrict the amount of data throughput, allowing only 0.1MS/s. When performing the averaging tasks the system can only achieve a closed loop throughput of 67kS/s. These rates are sufficient to demonstrate the control system.

The power system is built around a Burr-Brown OPA549 high power amplifier rated for 8A continuous current (Appendix B). It is wired to provide a gain of -2 so that voltages above the 10 V limit of the data acquisition system can be achieved. The system is protected by two high voltage power Schottky rectifiers (Appendix B) to prevent excessive voltages from motor EMF damaging the power amplifier. The amplifier draws it power from two Black and Decker 18V NiCd batteries in order to provide sufficient current, as no other power source with sufficient current was available. Figure (4) shows the wiring diagram with voltage potentials. This figure shows how the shunt resistor was placed on the grounding side of the motor in order to keep all leads connected to the data acquisition system within the -10V to 10V limits.

Figure (5) shows measured current data compared simulated current date. The tests consisted of starting the motors from rest with a step voltage command. The fidelity found between the data sets demonstrates the reliability of the model. This is also demonstrated more heavily in the Controller Design section. The results clearly show sizing differences. The motor is not loaded in the exact same manner and is rotating slower than the simulation, thus creating larger ripples since they have more time to settle. However, the fundamentals of the models are the same, showing that the system works for different size motors in the same manner.

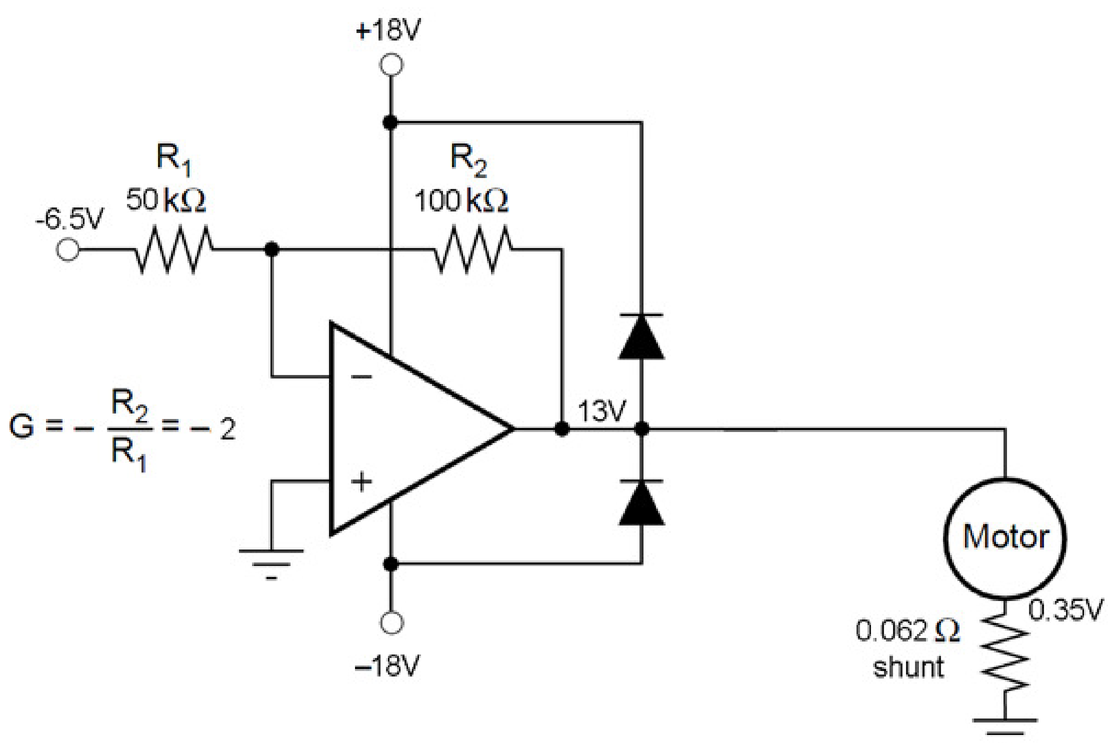


Figure (4) Power system wiring diagram

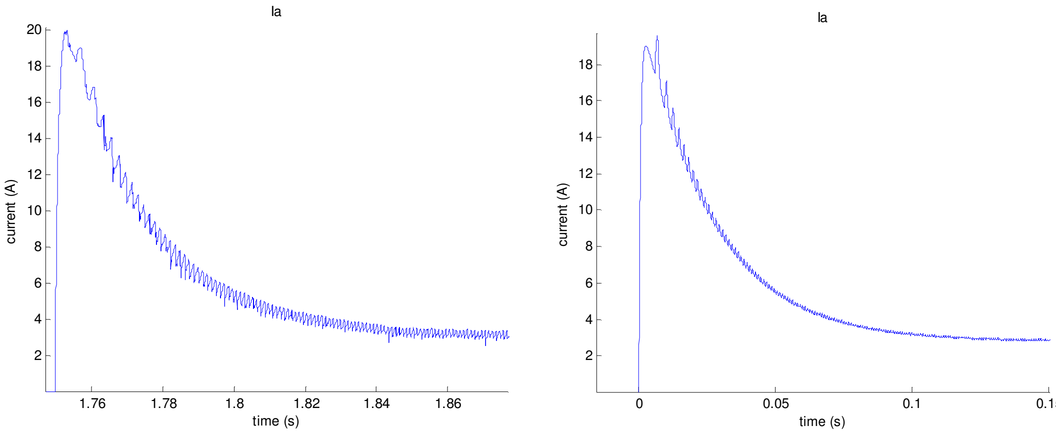


Figure (5) Measured Current data (left) next to simulated data (Right)

The easiest and most effective method of control would be to know the motor's angular position and determine the resistance from that. That would require an extra sensor and the intent of this design is to improve performance based on increasing computer power without additional hardware. The only available data for the controller is the voltage the motor sees which can be measured directly and the current which can be found by measuring the voltage across a low impedance shunt resistor in the motor line. Therefore, the model is rearranged to insert the controller where it receives an on or off signal and has a current feedback to control from (Figure 7). A simplified diagram of the overall system is depicted in figure (6).

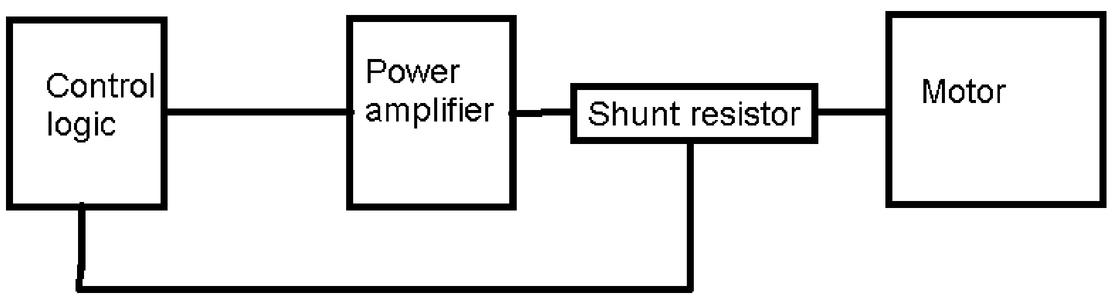


Figure (6) Simplified system diagram

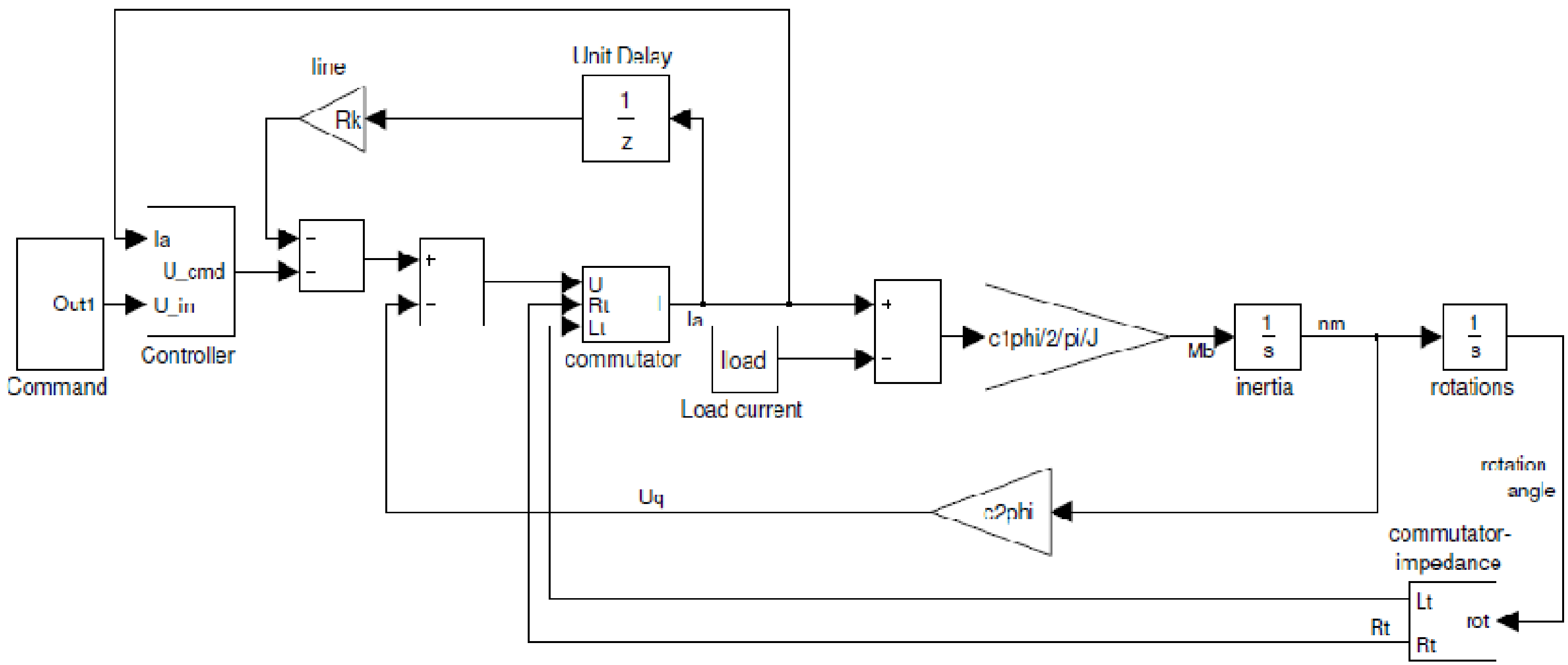


Figure (7) Motor model with controller in place using current feedback

Most control tasks involve a target and hitting that target. This task requires the torque as controlled by the current is regulated to reduce low frequency ripples. Since the load is unknown and variable, the control logic becomes resisting change. Mathematically this means suppressing $\dot{I}(t)$ toward zero while allowing sufficient slew to react to varying conditions.

Given the peak-style ripple in the current a derivative controller is the simplest method to sense the change in contacts. Difficulties caused by noise to a derivative will be addressed later. While the change in resistance is a step wise function, the change in $\dot{I}(t)$ will be extremely large or infinite. To mitigate the effects of a derivative controller and to prevent irrational control commands, a saturation limit is placed in the control line. To determine appropriate limits for the control refer to equation (4).

$$I(t) = \frac{V(t)}{R(\theta)} \qquad \text{Equation (4)}$$

Under steady state conditions with a constant drive voltage given sufficient settling time this equation holds true. Writing this out for both contact possibilities yields:

$$I_1(t) = \frac{V_1(t)}{R_1} \qquad I_2(t) = \frac{V_2(t)}{R_2} \qquad \text{Equation (5), (6)}$$

Assuming the ripple is stabilized through control

$$I_1(t) = I_2(t) \qquad \text{Equation (7)}$$

Therefore

$$\frac{V_1(t)}{R_1} = \frac{V_2(t)}{R_2} \qquad \text{Equation (8)}$$

Rearranging this

$$\frac{V_1(t)}{V_2(t)} = \frac{R_1}{R_2} \qquad \text{Equation (9)}$$

This shows that the appropriate ratio of command voltages is the ratio between the respective resistances.

In order to even out the distribution this will be divided above and below the initial command voltage. This will produce the following controller design (Figure (8)). There is a unit delay on the output to simulate the delay experienced when controlling a real system.

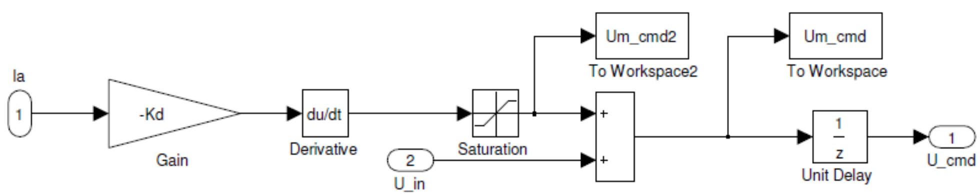


Figure (8) Simple Derivative Controller

When controlling a noiseless system in simulation the results are reasonable given appropriate controller settings.

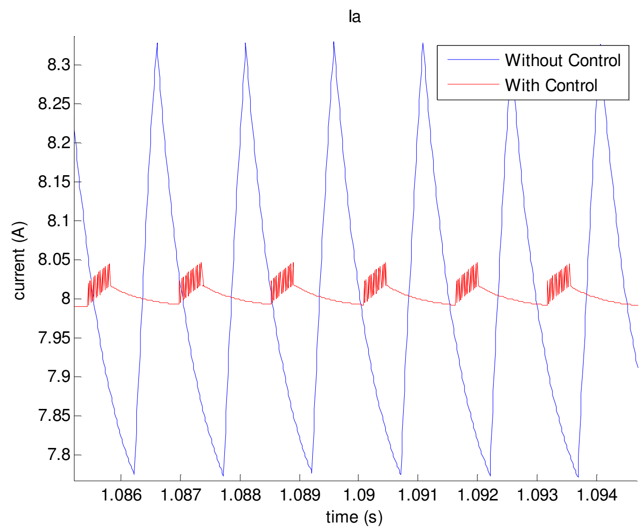


Figure (9) Close up of derivative control of a noiseless system, factor of 12 reduction on peak to peak ripple.

The use of this controller with simulated systems where significant levels of measurement noise is present is not recommendable as the derivative is highly affected by noise.

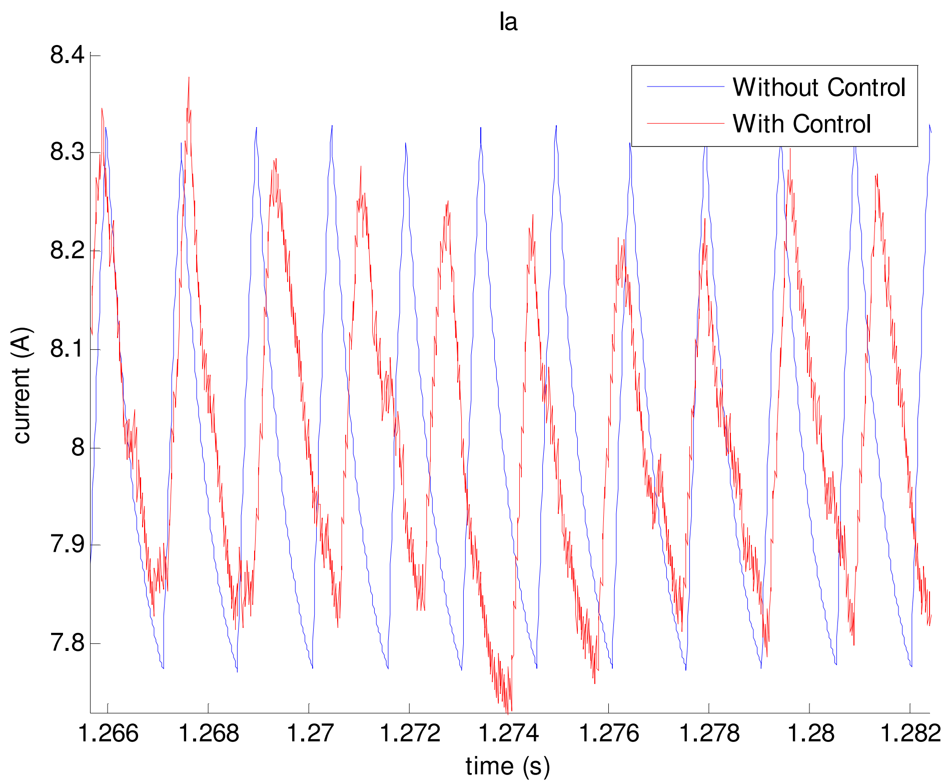


Figure (10) Close up of derivative control of system with noise

It is clear that a simplistic derivative controller is unsuitable for realistic applications where some level of noise is present. Another method for control would be using applications to create a scenario similar to typical control designs with a target and a system to drive it there. The uncontrolled motor system adjusts to load changes by adjusting current draw. The controller needs to force the current ripples down while still allowing the overall system to adjust the current draw to match the load. This can be achieved by employing sliding averages. A larger sliding average can be used to determine the target voltage. This method of digital filtering allows relatively fast response times with minimal oscillations. The classical filters proved to be ineffective or unstable for this application. Using more points the target can be set relatively firmly with little variation. Larger numbers of points can be used at the cost of speed as it makes the system “sluggish” in response to load variations. By using fewer points the reaction time can be improved since the overall update of the system is shorter. However, this causes the target to be less damped. It then oscillates at lower amplitudes and lags the true signal. To be noted is the fact that as the controller reduces the ripple the average of the ripples reduces as well and the effects of the fewer points average can be minimal. By forcing the sample frequency higher, enough samples can be taken to mitigate some of the effects of noise with digital filtering. This is accomplished by taking a very small number of samples and averaging them. The result lags slightly behind the true signal. However, given sufficient sampling this is minimal and the reduction of noise can be very effective. The resulting controller can be seen in figure (11) where, for the model, the signal comes in with noise. It is then sent to both the large and the small averages where the latest set of data is stored and averaged. The averages are then fed into the controller where the small average is subtracted from the large, producing the “error”. This is then multiplied by a gain and saturated at the 1.2 V to -1.2 V limit to avoid extreme controls. This value is then added to the initial command voltage and then, for the model, it goes through a unit delay to simulate the connection time with hardware.

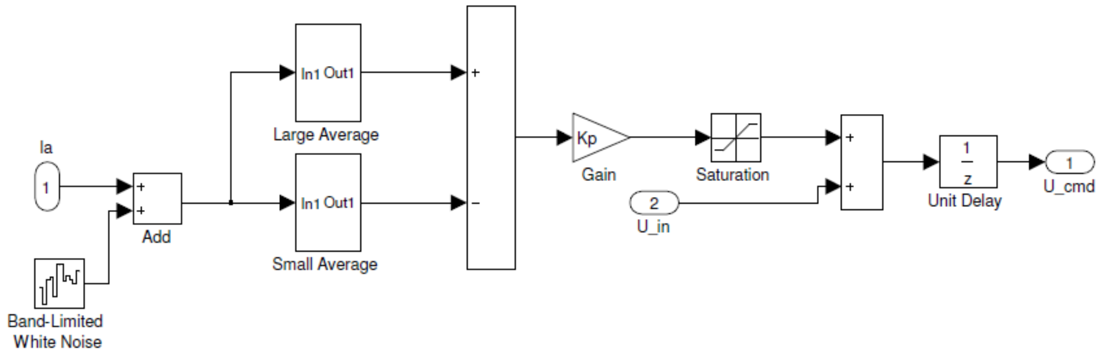


Figure (11) Sliding average controller with noise and unit delay

This can be mathematically represented with the following equation where superscripts refer to time steps:

$$V_{cmd}^{2} = \left(\left(Avg_{large}(I_a) - Avg_{small}(I_a)\right) K_p + V_{in}\right)^{1} \quad \text{Equation(10)}$$

Applying this controller with a sample time of 15 microseconds and a large average of 60 points with a small average of 3 points the simulation results can be seen in figures (12-14). Figure (13) shows the start up peak where many of the larger peaks are drastically reduced. Figure (14) shows the steady state conditions where higher frequency noise is present; however the overall peak to peak magnitude of the ripples has been reduced to less than a third of the free running magnitude. Table (2) provides a comparison of the methods used to this point.

Table (2) Comparison of simulations

Value\Simulation	Without control	Derivative w/o noise	Derivative with noise	Average with noise
RMS mAmps	162.2	15.2	149.2	49.0
RMS-RMS(w/o control) mAmps	0	-146.0	-12.0	-112.3
PSD mdBA/Hz 1-2kHz	7678.3	7665.4	7676.3	7666.3
PSD-PSD(w/o control) mdBA/Hz 1-2kHz	0	-12.6	-2.0	-11.7
PSD mdBA/Hz 1-20kHz	7679.5	7666.4	7677.5	7667.5
PSD-PSD(w/o control) mdBA/Hz 1-20kHz	0	-12.9	-1.8	-11.8

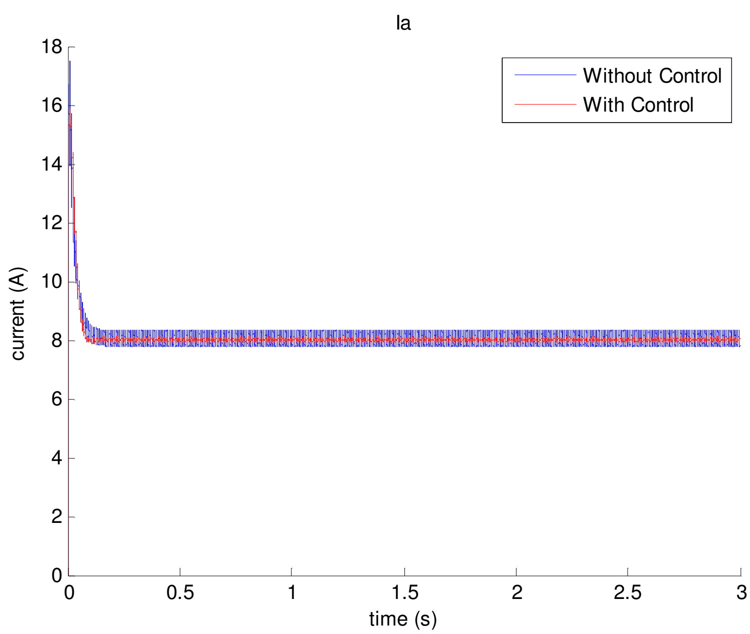


Figure (12) Averaged simulation model run

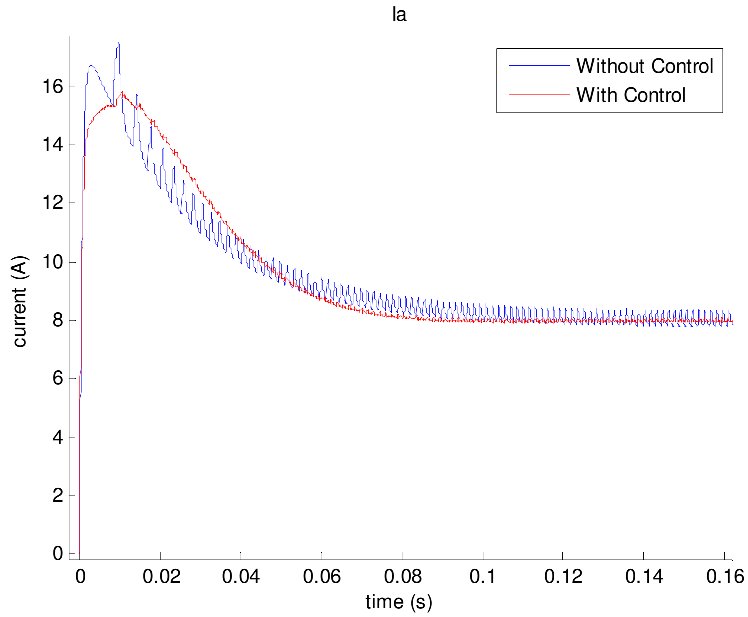


Figure (13) Averaged simulation model Start up from rest with 8 amp load

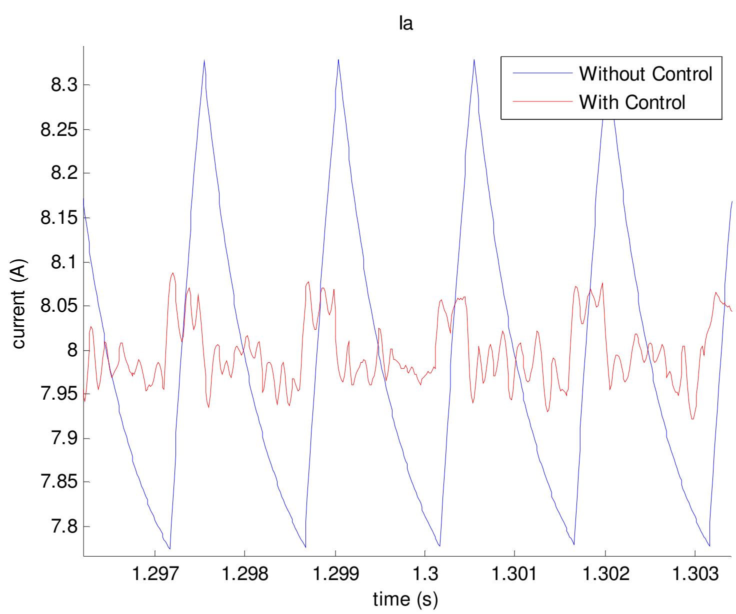


Figure (14) Averaged noise free simulation model steady state

EXPERIMENTAL TESTING

The final control program has a clock based switching system by which the controller can be turned on and off. It also pulls the controller gain from a data file which is initiated in a setup program to facilitate testing multiple gains within the same run for comparison purposes.

Table (3) Comparison of 2.5 sec test

Value\Simulation	w/o control	Gain of 2.5	Gain of 8
RMS mAmps	205.8	188.1	108.9
RMS-RMS(w/o control) mAmps	0	-17.7	-96.9
PSD mdBA/Hz 1-2kHz	950.4	1019.8	390.2
PSD-PSD(w/o control) mdBA/Hz 1-2kHz	0	69.3	-560.3
PSD mdBA/Hz 1-20kHz	952.2	1021.6	391.0
PSD-PSD(w/o control) mdBA/Hz 1-20kHz	0	69.4	-561.2

Table (4) Comparison of 0.5 sec test runs

Value\Simulation	W/o control	Gain of 7	Gain of 7.5	Gain of 8	Gain of 9
RMS mAmps	161.8	48.0	34.3	38.8	38.8
RMS-RMS(w/o control) mAmps	0	-106.9	-131.6	-123.0	-117.6
PSD mdBA/Hz 1-2kHz	23107.2	10046.8	9451.2	9619.6	9783.9
PSD-PSD(w/o control) mdBA/Hz 1-2kHz	0	-14329.8	-13041.5	-13487.6	-13802.3
PSD mdB/Hz 1-20kHz	23110.1	10047.8	9452.0	9620.7	9785.0
PSD-PSD(w/o control) mdBA/Hz 1-20kHz	0	-14332.0	-13043.5	-13489.4	-13804.1

These data sets demonstrate how the controller decreases the amplitude of the current ripple significantly. The higher gain settings produce larger spikes in irregular noise. Using this controller it is prudent to select a gain of approximately 7.5 as it produces well reduced lower amplitudes without the introduction of significantly more powerful higher amplitudes or an excessive number of irregular spikes. Using the assumption that the acoustic output of the motor is directly related to current ripple reductions in sound power can be expected (Schulter,6).

In order to test the Hypothesis of the sound reduction six acoustic samples were taken of the motor in operation. The samples were taken with a common computer microphone and compared to the uncontrolled motor’s sound recording. The PSD of the controlled recordings showed minor reductions of sound. For frequencies between 0 and 2000 Hz the PSD was reduced between 6% and 8% while for the 0 to 5000 Hz spectrum the reduction of PSD was between 2% and 5%. These results show that the control of the current flow will affect the sound output. However, since the reduction seen is minimal to the point of being unnoticeable to the human ear it would be advisable to research the correlation between current ripple and acoustic noise more thoroughly.

RESULTS

From the above background and testing it is clear that the control of ripple in commutated motors is indeed feasible given the increasing quality and speed of modern computer hardware. The current results display a significant amount of higher frequency ripples. However, the PSD of the ripples is reduced along with the RMS values. The effect of this control reduces the acoustic impact of these motors, if only minimally. By proving the capability of this type of control the grounds have been set to refine and improve these motors. The noise free simulations demonstrate that given sufficient noise filtering on the current measurement, it is possible to reduce the ripples of current on commutated motors down to negligible levels.

SUGGESTIONS FOR CONTINUED WORK

There was some research done during the project which suggested that by measuring the back EMF voltages and taking them into account with the controller better results could be achieved. It should be approached to see if such an application could be realized and improve on the performance.

The next stage of controller would be a predictive controller. While the challenges associated with such a control would be intriguing they are beyond the scope of this project. The nonlinearities of the system make such a controller difficult. The nonlinear aspects do provide fairly definitive points however which may be able to define the system well.

It would be useful to set up a test system so that acoustic sound and torque measurements could be directly correlated to the current measurements from the controller.

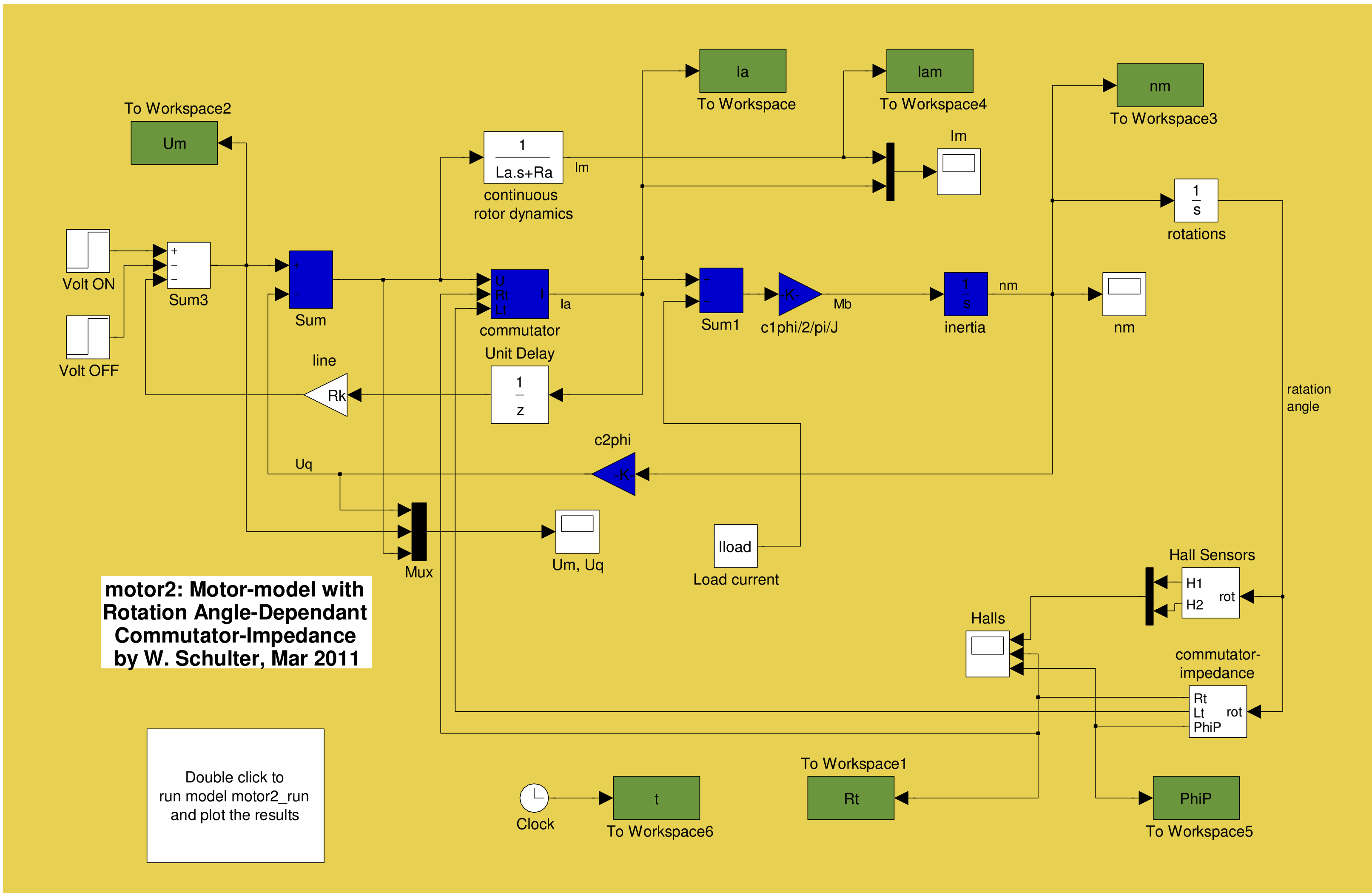
REFERENCES

[1] Schulter, Wolfgang, *DC Motor Modell,* 2011

[2] Nottelmann, Jan, *Sensor-less Rotation Counting in Brush Commutated DC motors*, IDEAdvance Ltd, 2010

Appendices

Appendix A: Simulink Model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Volt ON
Volt OFF
Sum3
Sum
U
Rt
Lt
I
Ia
commutator
Sum1
c1phi/2/pi/J
Mb
inertia
nm
nm
Unit Delay
1
z
line
Rk
c2phi
-K-
Uq
ratation
angle
Mux
Um, Uq
Iload
Load current
Hall Sensors
H1
H2
rot
Halls
commutator-
impedance
Rt
Lt
rot
PhiP
motor2: Motor-model with
Rotation Angle-Dependant
Commutator-Impedance
by W. Schulter, Mar 2011
Double click to
run model motor2_run
and plot the results
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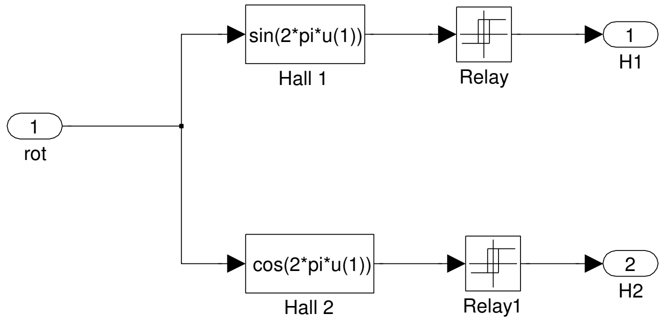
To Workspace6
To Workspace1
Rt
PhiP
To Workspace5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rot
cos(2*pi*u(1))
Hall 2
Relay1
2
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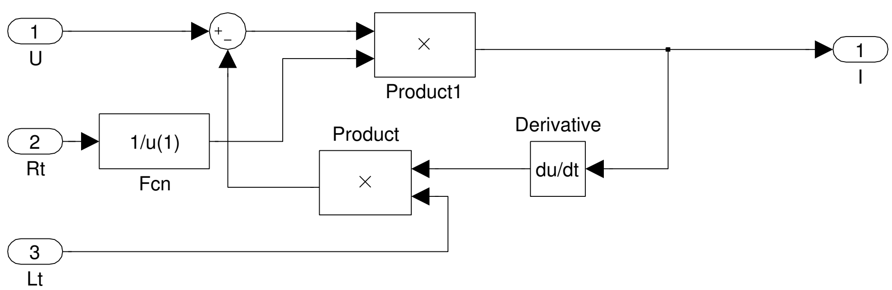
1
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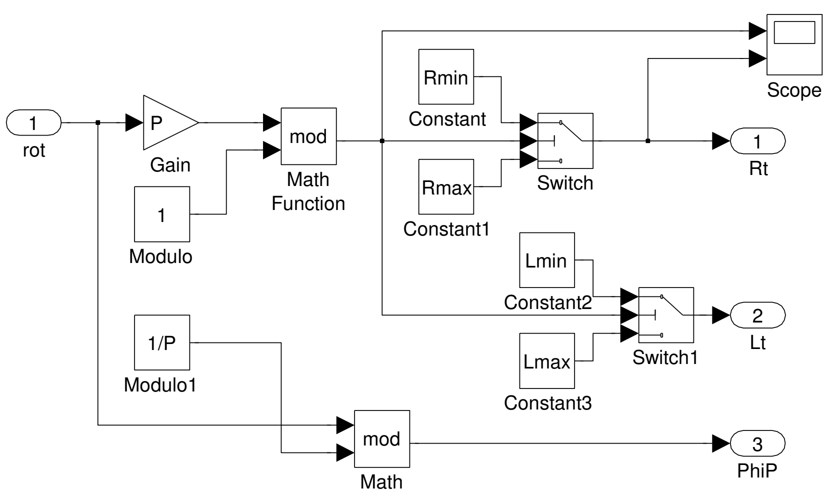
I(t) = (U - I'(t)*L)/R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Rmax
Constant1
Switch
Lmin
Constant2
Lmax
Constant3
Switch1
Scope
1
Rt
2
Lt
3
PhiP

APPENDIX B: SPECIFICATION SHEETS

Technical Sales
United States
(866) 531-6285
info@ni.com

NI PCI-6251

16-Bit, 1 MS/s (Multichannel), 1.25 MS/s (1-Channel), 16 Analog Inputs

- Two 16-bit analog outputs (2.8 MS/s); 24 digital I/O; 32-bit counters
- NIST-traceable calibration certificate and more than 70 signal conditioning options
- Correlated DIO (8 clocked lines, 10 MHz); analog and digital triggering
- NI-MCal calibration technology for increased measurement accuracy
- Get improved measurement accuracy, resolution, and sensitivity by choosing high-accuracy M Series.
- NI-DAQmx driver software and NI LabVIEW SignalExpress LE interactive data-logging software

Overview

The National Instruments PCI-6251 is a high-speed multifunction M Series data acquisition (DAQ) board optimized for superior accuracy at fast sampling rates. For increased measurement accuracy, consider the high-accuracy M Series devices with an 18-bit analog-to-digital converter providing a 4X resolution increase.

High-speed M Series devices incorporate advanced features such as the NI-STC 2 system controller, NI-PGIA 2 programmable amplifier, and NI-MCal calibration technology to increase performance and accuracy. High-speed M Series devices have an onboard NI-PGIA 2 amplifier designed for fast settling time at high scanning rates, ensuring 16-bit accuracy even when measuring all channels at maximum speeds. To learn more about M Series technologies, device specifications, and information on recommended cables and accessories, please refer to the data sheet and specifications.

Driver Software

M Series devices work with multiple operating systems using three driver software options including NI-DAQmx, NI-DAQmx Base, and the Measurement Hardware DDK. Browse the information in the Resources tab to learn more about driver software or download a driver. M Series devices are not compatible with the Traditional NI-DAQ (Legacy) driver.

Application Software

With NI LabVIEW, you can create custom data acquisition applications with the ease of graphical programming and power of more than 500 analysis functions and advanced programming tools. LabVIEW Full and Professional Development Systems include LabVIEW SignalExpress for interactive data logging. M Series data acquisition devices are compatible with the following versions (or later) of NI application software – LabVIEW 7.x, LabWindows™/CVI 7.x, or Measurement Studio 7.x; LabVIEW SignalExpress 1.x; or LabVIEW with the LabVIEW Real-Time Module 7.1. M Series data acquisition devices are also compatible with Visual Studio .NET, C/C++, and Visual Basic 6.

Specifications

Specifications Documents

- Specifications
- Data Sheet

Specifications Summary

General	
Product Name	PCI-6251
Product Family	Multifunction Data Acquisition
Form Factor	PCI
Part Number	779070-01
Operating System/Target	Real-Time , Linux , Mac OS , Windows
LabVIEW RT Support	Yes
DAQ Product Family	M Series
Measurement Type	Digital , Voltage , Frequency , Quadrature encoder
RoHS Compliant	Yes
Analog Input	
Channels	16 , 8
Single-Ended Channels	16
Differential Channels	8
Resolution	16 bits
Sample Rate	1.25 MS/s
Max Voltage	10 V
Maximum Voltage Range	-10 V , 10 V
Maximum Voltage Range Accuracy	1920 µV
Maximum Voltage Range Sensitivity	112 µV
Minimum Voltage Range	-100 mV , 100 mV
Minimum Voltage Range Accuracy	52 µV
Minimum Voltage Range Sensitivity	6 µV
Number of Ranges	7
Simultaneous Sampling	No
On-Board Memory	4095 samples
Analog Output	
Channels	2
Resolution	16 bits
Max Voltage	10 V

Maximum Voltage Range	-10 V , 10 V
Maximum Voltage Range Accuracy	2080 µV
Minimum Voltage Range	-5 V , 5 V
Minimum Voltage Range Accuracy	1045 µV
Update Rate	2.86 MS/s
Current Drive Single	5 mA
Digital I/O	
Bidirectional Channels	24
Input-Only Channels	0
Output-Only Channels	0
Number of Channels	24 , 0
Timing	Software , Hardware
Max Clock Rate	10 MHz
Logic Levels	TTL
Input Current Flow	Sinking , Sourcing
Output Current Flow	Sinking , Sourcing
Programmable Input Filters	Yes
Supports Programmable Power-Up States?	Yes
Current Drive Single	24 mA
Current Drive All	448 mA
Watchdog Timer	No
Supports Handshaking I/O?	No
Supports Pattern I/O?	Yes
Maximum Input Range	0 V , 5 V
Maximum Output Range	0 V , 5 V
Counter/Timers	
Counters	2
Number of DMA Channels	2
Buffered Operations	Yes
Debouncing/Glitch Removal	Yes
GPS Synchronization	No

Maximum Range	0 V , 5 V
Max Source Frequency	80 MHz
Pulse Generation	Yes
Resolution	32 bits
Timebase Stability	50 ppm
Logic Levels	TTL
Physical Specifications	
Length	15.5 cm
Width	9.7 cm
I/O Connector	68-pin VHDCI female
Timing/Triggering/Synchronization	
Triggering	Digital , Analog
Synchronization Bus (RTSI)	Yes

Pricing

NI PCI-6251 Complete Package

Each NI PCI-6251 requires:

1 Connector Block	1 Cable	**NI PCI-6251**	Software

Roll over icons above to learn why you need each item in the package.

NI PCI-6251 and Accessories

Hardware Subtotal: $ 1,537

NI PCI-6251 - 779070-01	Qty	$ 1,079 each
Required Accessories		
Connector Block - Screw Terminal **SCB-68** - 776844-01	Qty	$ 319 each
Cable - Shielded **SHC68-68-EPM Cable (2m)** - 192061-02 Select length: 1m 2m 5m 10m 0.5m	Qty	$ 139 each

Change Accessories
Learn about all compatible accessories and get recommendations based on your application needs.

Optional Accessories Show

Software

Software Subtotal: $ 0
Note : You should only purchase this device without software if you already own compatible application software. [View Full List]

LabVIEW Full Development System for Windows (English) , 1 year SSP - 776670-35	Qty	$ 2,699 each
LabVIEW Application Builder for Windows - 776675-35	Qty	$ 999 each

Change Software
Learn about all compatible software and get recommendations based on your needs.

Estimated Shipping Days: 1 - 3
Package Price: $ 1,537

Add to Cart >>

Guarantee your price for 30 days (learn more) Instant Quote

Save for Later

Since your device does not have direct signal connectivity, you need a connector block to act as an interface between your sensors/signals and your device. A connector block provides easy access to the inputs and outputs of your hardware.
You need a cable to transmit signals between your device and your signal connections on your connector block.
You need software to interface with your hardware and to collect, analyze, present, and store your measurements. This board is compatible with a variety of programming languages, including LabVIEW, C/C++, Visual Basic, and .NET. LabVIEW provides the easiest integration with all of your NI hardware and is recommended to maximize your hardware investment.
You have selected **United States** as the country where you will use the product(s) (change).

Services

Extended Warranties

National Instruments designs and manufactures all products to minimize failures, however unexpected failures can still occur. Extended warranties provide a fixed economical price at the time of system purchase, covering any repair costs for up to three years. In addition, they offer the following benefits:

- Significant cost savings compared to individual repair incidents
- Fault location, diagnostics, and repair by NI any time the system product fails
- All parts and labor costs covered as well as any adjustments needed to restore the hardware to manufacturing specifications

For more information about your warranty options:

- Learn More About Warranty Services [http://www.ni.com/services/warranty.htm]
- Talk to an Expert About Extended Warranties [javascript:openCallMeWindowCTA(document.referrer,%20'US')]
- View Warranty Repair Policies [http://www.ni.com/services/warranty_repair_policies.htm]

Calibration

NI recognizes the need to maintain properly calibrated devices for high-accuracy measurements. NI provides manual calibration procedures, services to recalibrate your products, and automated calibration software to calibrate many NI measurement products.

- Learn More About Calibration Services [http://www.ni.com/services/calibration.htm]

Training

NI training is the fastest, most certain route to productivity with NI tools and successful application development.

- Learn More About NI Training and Certification [http://www.ni.com/training/]
- Find a Course Near You and View Schedules [http://sine.ni.com/apps/utf8/nisv.custed]

Repair Services

Return your registered product under warranty at no additional labor and parts cost. NI offers fault location, diagnostics, and repair any time the system fails as well as any adjustments needed to restore the hardware to manufacturing specifications.

- Learn More About Repair Services [http://www.ni.com/services/warranty.htm]
- Contact NI to obtain a Return Material Authorization (RMA) form and shipping instructions. [http://sine.ni.com/apps/utf8/nicc.call_me]
- View your RMA support request status online. [http://www.ni.com/support/servicereq/]
- Register your product [http://www.ni.com/register] .

Technical Support

ni.com/support [http://www.ni.com/support/]

Resources

Additional Product Information

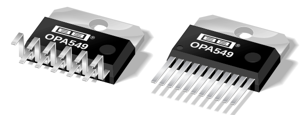
- Manuals (6)
- Dimensional Drawings
- Product Certifications

Related Information

- Obtain superior accuracy in your measurements with 18-bit, high-accuracy M Series
- Upgrade to S Series with simultaneous sampling
- Learn about LabVIEW Graphical Programming
- Learn about Data Acquisition Driver Software
- Learn about academic pricing and product kits
- Obtain OEM pricing information
- Download NI drivers
- Read More About NI Software Maintenance and Support
- View an introductory video on getting started with NI data acquisition

Burr-Brown Products
from Texas Instruments

OPA549



SBOS093E – MARCH 1999 – REVISED OCTOBER 2005

High-Voltage, High-Current OPERATIONAL AMPLIFIER

FEATURES

- **HIGH OUTPUT CURRENT:**
  8A Continuous
  10A Peak
- **WIDE POWER-SUPPLY RANGE:**
  Single Supply: +8V to +60V
  Dual Supply: ±4V to ±30V
- **WIDE OUTPUT VOLTAGE SWING**
- **FULLY PROTECTED:**
  Thermal Shutdown
  Adjustable Current Limit
- **OUTPUT DISABLE CONTROL**
- **THERMAL SHUTDOWN INDICATOR**
- **HIGH SLEW RATE: 9V/µs**
- **CONTROL REFERENCE PIN**
- **11-LEAD POWER PACKAGE**

APPLICATIONS

- **VALVE, ACTUATOR DRIVERS**
- **SYNCHRO, SERVO DRIVERS**
- **POWER SUPPLIES**
- **TEST EQUIPMENT**
- **TRANSDUCER EXCITATION**
- **AUDIO POWER AMPLIFIERS**

DESCRIPTION

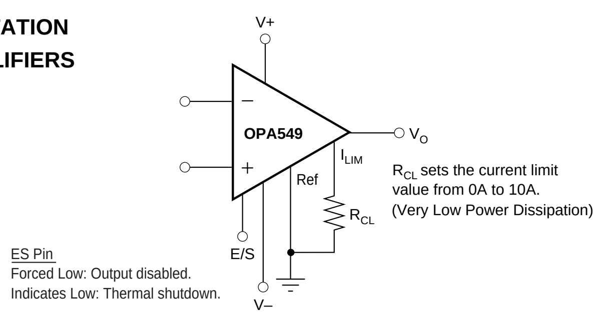
The OPA549 is a low-cost, high-voltage/high-current operational amplifier ideal for driving a wide variety of loads. This laser-trimmed monolithic integrated circuit provides excellent low-level signal accuracy and high output voltage and current.

The OPA549 operates from either single or dual supplies for design flexibility. The input common-mode range extends below the negative supply.

The OPA549 is internally protected against over-temperature conditions and current overloads. In addition, the OPA549 provides an accurate, user-selected current limit. Unlike other designs which use a “power” resistor in series with the output current path, the OPA549 senses the load indirectly. This allows the current limit to be adjusted from 0A to 10A with a resistor/potentiometer, or controlled digitally with a voltage-out or current-out Digital-to-Analog Converter (DAC).

The Enable/Status (E/S) pin provides two functions. It can be monitored to determine if the device is in thermal shutdown, and it can be forced low to disable the output stage and effectively disconnect the load.

The OPA549 is available in an 11-lead power package. Its copper tab allows easy mounting to a heat sink for excellent thermal performance. Operation is specified over the extended industrial temperature range, –40°C to +85°C.



ABSOLUTE MAXIMUM RATINGS(1)

Output Current	See SOA Curve (Figure 6)
Supply Voltage, V+ to V–	60V
Input Voltage Range	(V–) – 0.5V to (V+) + 0.5V
Input Shutdown Voltage	Ref – 0.5 to V+
Operating Temperature	–40°C to +125°C
Storage Temperature	–55°C to +125°C
Junction Temperature	150°C
Lead Temperature (soldering, 10s)	300°C
ESD Capability (Human Body Model)	2000V

NOTE: (1) Stresses above these ratings may cause permanent damage. Exposure to absolute maximum conditions for extended periods may degrade device reliability.

PACKAGE/ORDERING INFORMATION

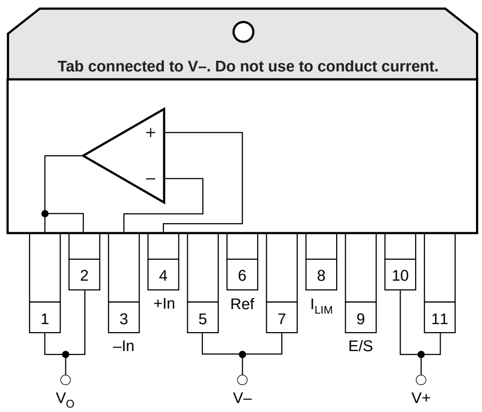
For the most current package and ordering information, see the Package Option Addendum at the end of this datasheet or see the TI website at www.ti.com.

ELECTROSTATIC DISCHARGE SENSITIVITY

This integrated circuit can be damaged by ESD. Texas Instruments recommends that all integrated circuits be handled with appropriate precautions. Failure to observe proper handling and installation procedures can cause damage.

ESD damage can range from subtle performance degradation to complete device failure. Precision integrated circuits may be more susceptible to damage because very small parametric changes could cause the device not to meet its published specifications.

CONNECTION DIAGRAM



Connect both pins 1 and 2 to output.
Connect both pins 5 and 7 to V–.
Connect both pins 10 and 11 to V+.

ELECTRICAL CHARACTERISTICS

Boldface limits apply over the specified temperature range, $\mathbf{T_A = -40°C\ to\ +85°C}$.

At T_{CASE} = +25°C, V_S = ±30V, Ref = 0V, and E/S pin open, unless otherwise noted.

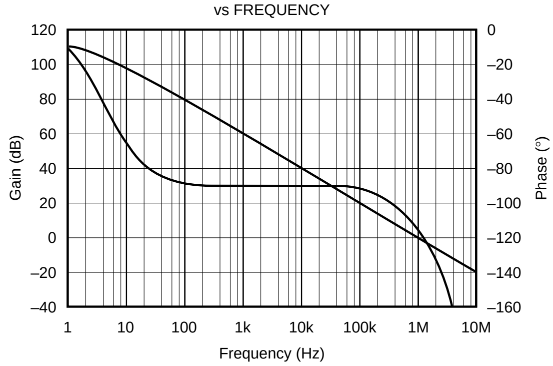
			OPA549T, S			
PARAMETER		**CONDITION**	**MIN**	**TYP**	**MAX**	**UNITS**
OFFSET VOLTAGE	V_{OS}					
Input Offset Voltage		V_{CM} = 0V, I_O = 0		±1	±5	mV
vs Temperature	**dV_{OS}/dT**	**T_{CASE} = –40°C to +85°C**		**±20**		**μV/°C**
vs Power Supply	PSRR	V_S = ±4V to ±30V, Ref = V–		25	100	μV/V
INPUT BIAS CURRENT[1]						
Input Bias Current[2]	I_B	V_{CM} = 0V		–100	–500	nA
vs Temperature		**T_{CASE} = –40°C to +85°C**		**±0.5**		**nA/°C**
Input Offset Current	I_{OS}	V_{CM} = 0V		±5	±50	nA
NOISE						
Input Voltage Noise Density	e_n	f = 1kHz		70		$nV/\sqrt{Hz}$
Current Noise Density	i_n	f = 1kHz		1		$pA/\sqrt{Hz}$
INPUT VOLTAGE RANGE						
Common-Mode Voltage Range: Positive	V_{CM}	Linear Operation	(V+) – 3	(V+) – 2.3		V
Negative	V_{CM}	Linear Operation	(V–) – 0.1	(V–) – 0.2		V
Common-Mode Rejection Ratio	CMRR	V_{CM} = (V–) – 0.1V to (V+) – 3V	80	95		dB
INPUT IMPEDANCE						
Differential				10^7 || 6		Ω || pF
Common-Mode				10^9 || 4		Ω || pF
OPEN-LOOP GAIN						
Open-Loop Voltage Gain	A_{OL}	V_O = ±25V, R_L = 1kΩ	100	110		dB
		V_O = ±25V, R_L = 4Ω		100		dB
FREQUENCY RESPONSE						
Gain Bandwidth Product	GBW			0.9		MHz
Slew Rate	SR	G = 1, 50Vp-p Step, R_L = 4Ω		9		V/μs
Full-Power Bandwidth				See Typical Curve		
Settling Time: ±0.1%		G = –10, 50V Step		20		μs
Total Harmonic Distortion + Noise[3]	THD+N	f = 1kHz, R_L = 4Ω, G = +3, Power = 25W		0.015		%
OUTPUT						
Voltage Output, Positive		I_O = 2A	(V+) – 3.2	(V+) – 2.7		V
Negative		I_O = –2A	(V–) + 1.7	(V–) + 1.4		V
Positive		I_O = 8A	(V+) – 4.8	(V+) – 4.3		V
Negative		I_O = –8A	(V–) + 4.6	(V–) + 3.9		V
Negative		R_L = 8Ω to V–	(V–) + 0.3	(V–) + 0.1		V
Maximum Continuous Current Output: dc[4]			±8			A
ac[4]		Waveform Cannot Exceed 10A peak	8			A rms
Output Current Limit						
Current Limit Range				0 to ±10		A
Current Limit Equation				I_{LIM} = 15800 • 4.75V/(7500Ω + R_{CL})		A
Current Limit Tolerance[1]		R_{CL} = 7.5kΩ (I_{LIM} = ±5A), R_L = 4Ω		±200	±500	mA
Capacitive Load Drive (Stable Operation)	C_{LOAD}			See Typical Curve		
Output Disabled						
Leakage Current		Output Disabled, V_O = 0V	–2000	±200	+2000	μA
Output Capacitance		Output Disabled		750		pF
OUTPUT ENABLE/STATUS (E/S) PIN						
Shutdown Input Mode						
$V_{E/S}$ High (output enabled)		E/S Pin Open or Forced High	(Ref) + 2.4			V
$V_{E/S}$ Low (output disabled)		E/S Pin Forced Low			(Ref) + 0.8	V
$I_{E/S}$ High (output enabled)		E/S Pin Indicates High		–50		μA
$I_{E/S}$ Low (output disabled)		E/S Pin Indicates Low		–55		μA
Output Disable Time				1		μs
Output Enable Time				3		μs
Thermal Shutdown Status Output						
Normal Operation		Sourcing 20μA	(Ref) + 2.4	(Ref) + 3.5		V
Thermally Shutdown		Sinking 5μA, T_J > 160°C		(Ref) + 0.2	(Ref) + 0.8	V
Junction Temperature, Shutdown				+160		°C
Reset from Shutdown				+140		°C
Ref (Reference Pin for Control Signals)						
Voltage Range			V–		(V+) – 8	V
Current[2]				–3.5		mA
POWER SUPPLY						
Specified Voltage	V_S			±30		V
Operating Voltage Range, (V+) – (V–)			8		60	V
Quiescent Current	I_Q	I_{LIM} Connected to Ref I_O = 0		±26	±35	mA
Quiescent Current in Shutdown Mode		I_{LIM} Connected to Ref		±6		mA
TEMPERATURE RANGE						
Specified Range			–40		+85	°C
Operating Range			–40		+125	°C
Storage Range			–55		+125	°C
Thermal Resistance, θ_{JC}				1.4		°C/W
Thermal Resistance, θ_{JA}		No Heat Sink		30		°C/W

NOTES: (1) High-speed test at T_J = +25°C. (2) Positive conventional current is defined as flowing into the terminal. (3) See "Total Harmonic Distortion + Noise vs Frequency" in the Typical Characteristics section for additional power levels. (4) See "Safe Operating Area" (SOA) in the Typical Characteristics section.

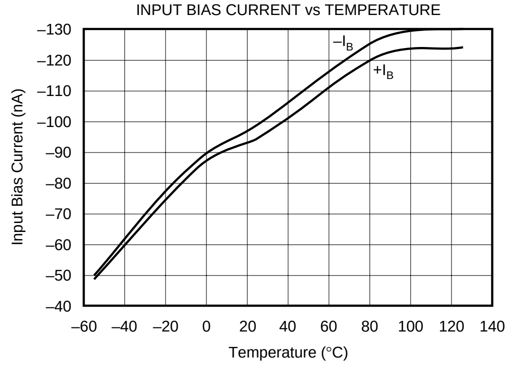
TYPICAL CHARACTERISTICS

At T_{CASE} = +25°C, V_S = ±30V, and E/S pin open, unless otherwise noted.

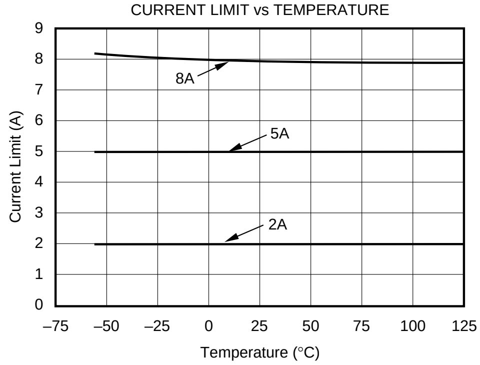
OPEN-LOOP GAIN AND PHASE
vs FREQUENCY



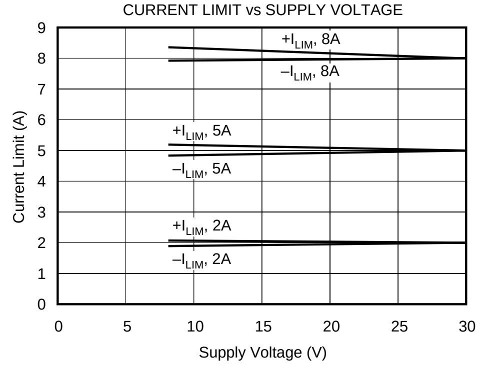
INPUT BIAS CURRENT vs TEMPERATURE



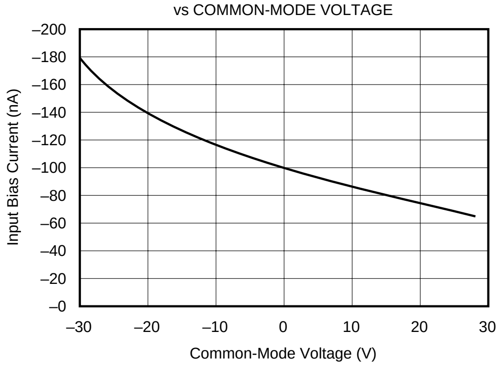
CURRENT LIMIT vs TEMPERATURE



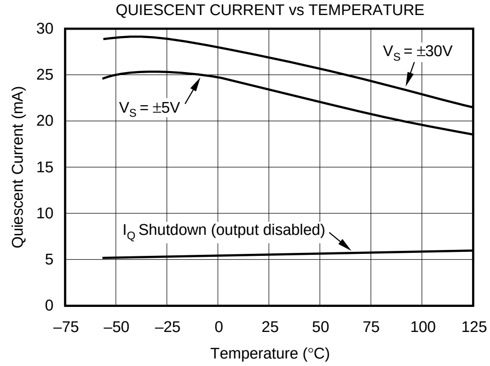
CURRENT LIMIT vs SUPPLY VOLTAGE



INPUT BIAS CURRENT
vs COMMON-MODE VOLTAGE



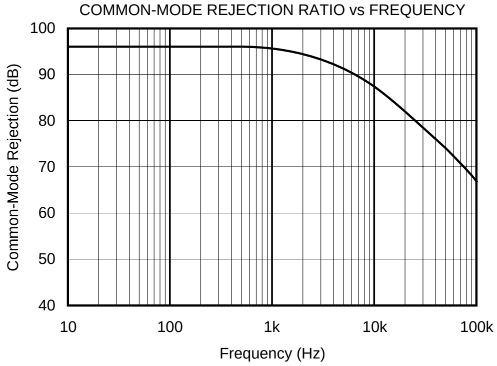
QUIESCENT CURRENT vs TEMPERATURE



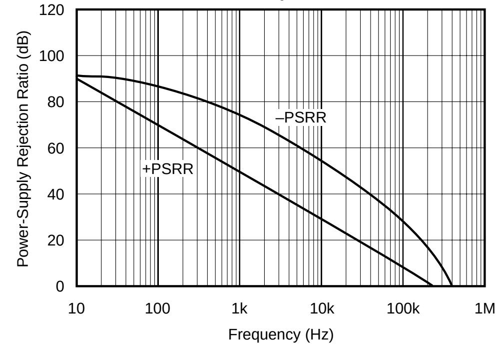
TYPICAL CHARACTERISTICS (Cont.)

At T_{CASE} = +25°C, V_S = ±30V, and E/S pin open, unless otherwise noted.

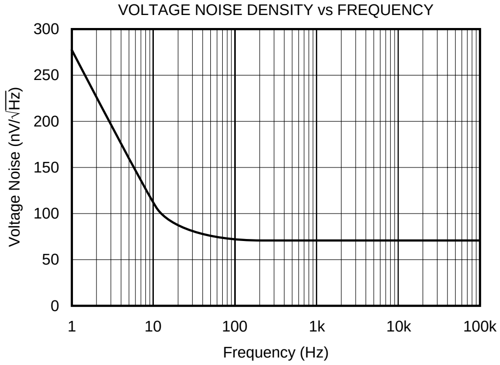
COMMON-MODE REJECTION RATIO vs FREQUENCY



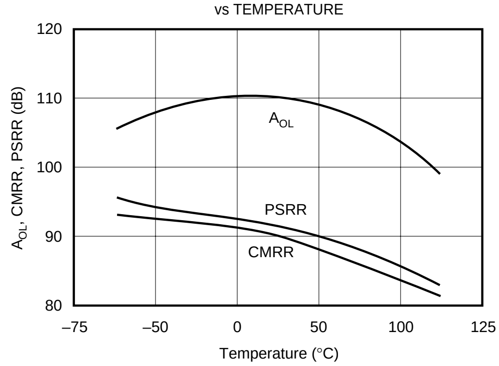
POWER-SUPPLY REJECTION RATIO
vs FREQUENCY



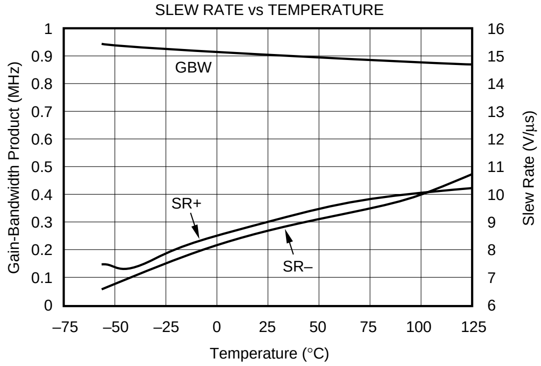
VOLTAGE NOISE DENSITY vs FREQUENCY



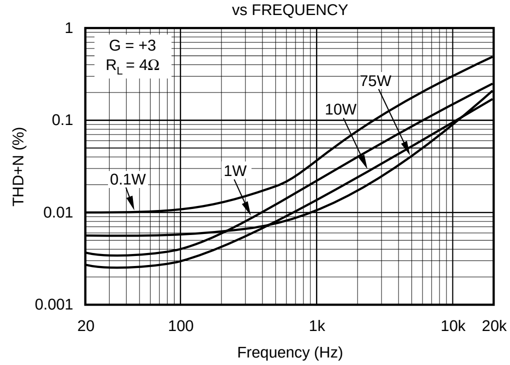
OPEN-LOOP GAIN, COMMON-MODE REJECTION RATIO,
AND POWER-SUPPLY REJECTION RATIO
vs TEMPERATURE



GAIN-BANDWIDTH PRODUCT AND
SLEW RATE vs TEMPERATURE



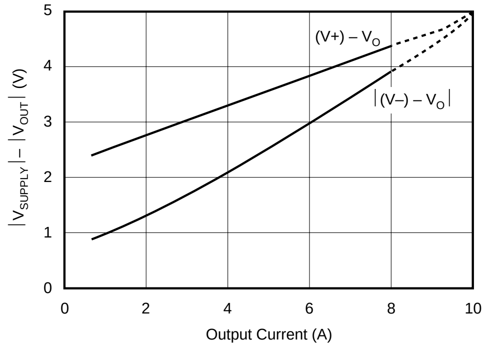
TOTAL HARMONIC DISTORTION + NOISE
vs FREQUENCY



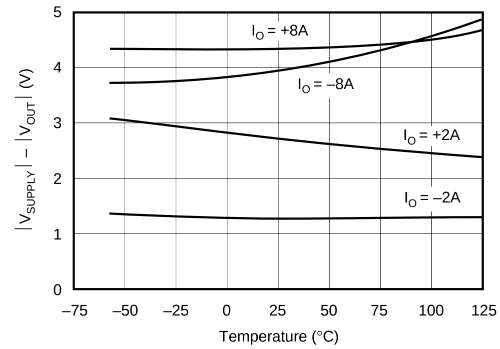
TYPICAL CHARACTERISTICS (Cont.)

At T_{CASE} = +25°C, V_S = ±30V, and E/S pin open, unless otherwise noted.

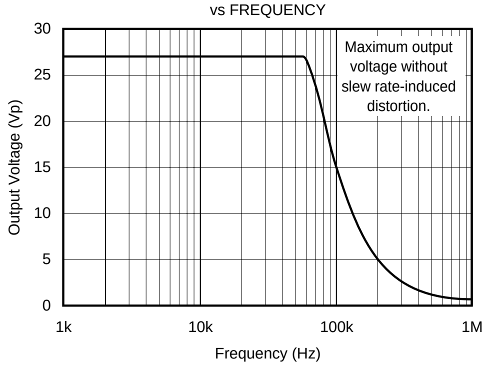
OUTPUT VOLTAGE SWING vs OUTPUT CURRENT



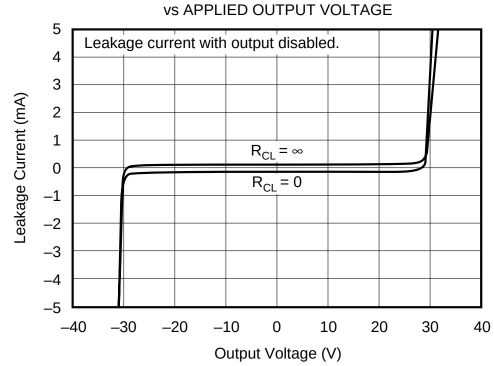
OUTPUT VOLTAGE SWING vs TEMPERATURE



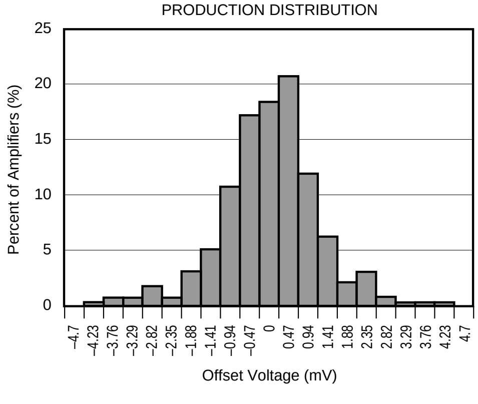
MAXIMUM OUTPUT VOLTAGE SWING
vs FREQUENCY



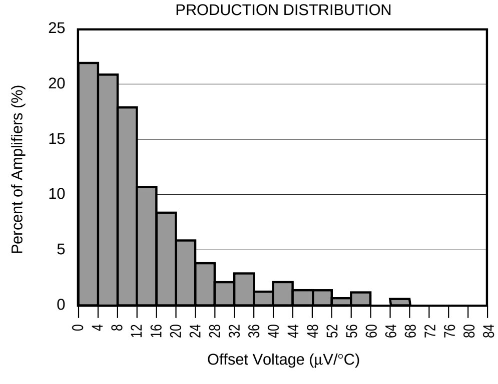
OUTPUT LEAKAGE CURRENT
vs APPLIED OUTPUT VOLTAGE



OFFSET VOLTAGE
PRODUCTION DISTRIBUTION



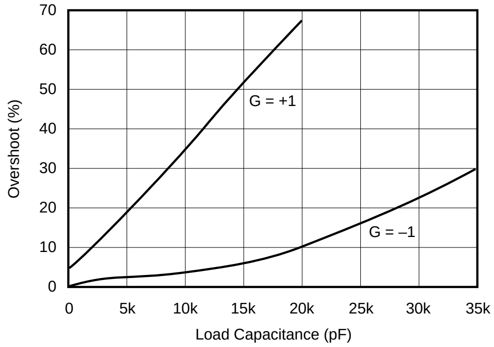
OFFSET VOLTAGE DRIFT
PRODUCTION DISTRIBUTION



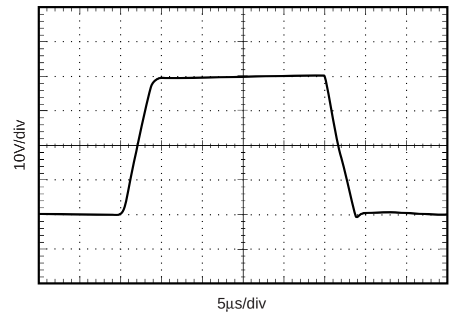
TYPICAL CHARACTERISTICS (Cont.)

At T_{CASE} = +25°C, V_S = ±30V, and E/S pin open, unless otherwise noted.

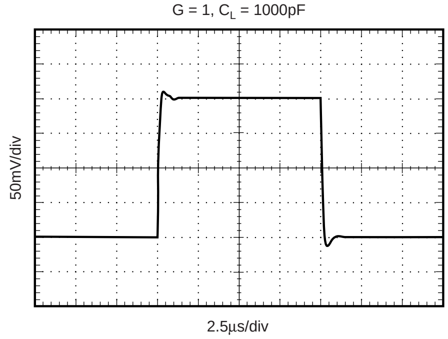
SMALL-SIGNAL OVERSHOOT vs LOAD CAPACITANCE



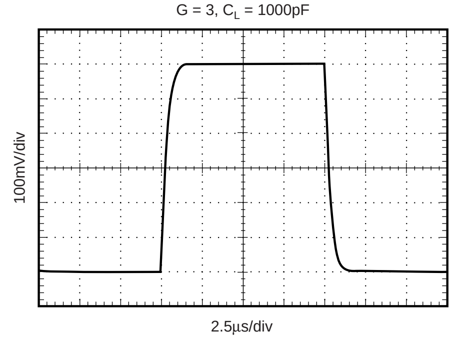
LARGE-SIGNAL STEP RESPONSE G = 3, C_L = 1000pF



SMALL-SIGNAL STEP RESPONSE G = 1, C_L = 1000pF



SMALL-SIGNAL STEP RESPONSE G = 3, C_L = 1000pF

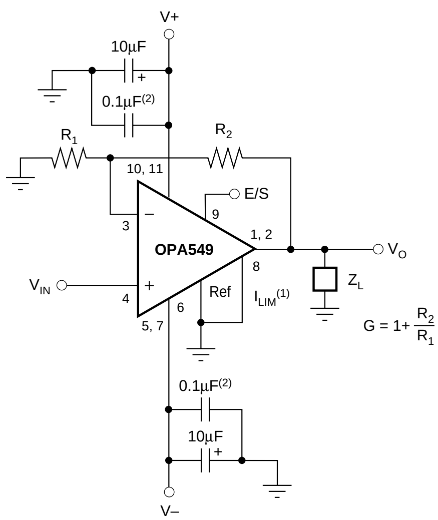


APPLICATIONS INFORMATION

Figure 1 shows the OPA549 connected as a basic noninverting amplifier. The OPA549 can be used in virtually any op amp configuration.

Power-supply terminals should be bypassed with low series impedance capacitors. The technique shown in Figure 1, using a ceramic and tantalum type in parallel, is recommended. Power-supply wiring should have low series impedance.

Be sure to connect *both* output pins (pins 1 and 2).



NOTES: (1) I_{LIM} connected to Ref gives the maximum current limit, 10A (peak). (2) Connect capacitors directly to package power-supply pins.

FIGURE 1. Basic Circuit Connections.

POWER SUPPLIES

The OPA549 operates from single (+8V to +60V) or dual (±4V to ±30V) supplies with excellent performance. Most behavior remains unchanged throughout the full operating voltage range. Parameters that vary significantly with operating voltage are shown in the Typical Characteristics. Some applications do not require equal positive and negative output voltage swing. Power-supply voltages do not need to be equal. The OPA549 can operate with as little as 8V between the supplies and with up to 60V between the supplies. For example, the positive supply could be set to 55V with the negative supply at –5V. **Be sure to connect both V– pins (pins 5 and 7) to the negative power supply, and both V+ pins (pins 10 and 11) to the positive power supply. Package tab is internally connected to V–; however, do not use the tab to conduct current.**

CONTROL REFERENCE (Ref) PIN

The OPA549 features a reference (Ref) pin to which the I_{LIM} and the E/S pin are referred. Ref simply provides a reference point accessible to the user that can be set to V–, ground, or any reference of the user's choice. **Ref cannot be set below the negative supply or above (V+) – 8V. If the minimum V_S is used, Ref must be set at V–.**

ADJUSTABLE CURRENT LIMIT

The OPA549's accurate, user-defined current limit can be set from 0A to 10A by controlling the input to the I_{LIM} pin. Unlike other designs, which use a power resistor in series with the output current path, the OPA549 senses the load indirectly. This allows the current limit to be set with a 0µA to 633µA control signal. In contrast, other designs require a limiting resistor to handle the full output current (up to 10A in this case).

Although the design of the OPA549 allows output currents up to 10A, it is not recommended that the device be operated continuously at that level. The highest rated continuous current capability is 8A. Continuously running the OPA549 at output currents greater than 8A will degrade long-term reliability.

Operation of the OPA549 with current limit less than 1A results in reduced current limit accuracy. Applications requiring lower output current may be better suited to the OPA547 or OPA548.

Resistor-Controlled Current Limit

See Figure 2a for a simplified schematic of the internal circuitry used to set the current limit. Leaving the I_{LIM} pin open programs the output current to zero, while connecting I_{LIM} directly to Ref programs the maximum output current limit, typically 10A.

With the OPA549, the simplest method for adjusting the current limit uses a resistor or potentiometer connected between the I_{LIM} pin and Ref according to Equation 1:

$$R_{CL} = \frac{75\text{kV}}{I_{LIM}} - 7.5\text{k}\Omega \qquad (1)$$

Refer to Figure 2 for commonly used values.

Digitally-Controlled Current Limit

The low-level control signal (0µA to 633µA) also allows the current limit to be digitally controlled by setting either a current (I_{SET}) or voltage (V_{SET}). The output current I_{LIM} can be adjusted by varying I_{SET} according to Equation 2:

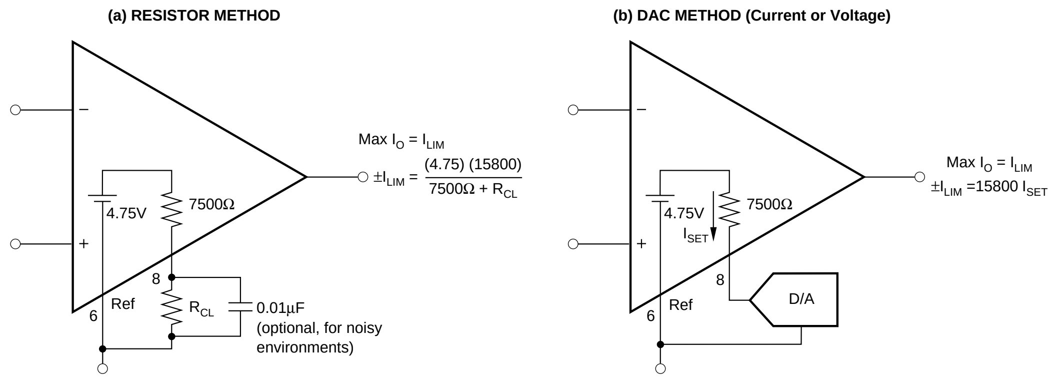
$$I_{SET} = I_{LIM}/15800 \qquad (2)$$

Figure 2b demonstrates a circuit configuration implementing this feature.

The output current I_{LIM} can be adjusted by varying V_{SET} according to Equation 3:

$$V_{SET} = (\text{Ref}) + 4.75\text{V} - (7500\text{W})(I_{LIM})/15800 \qquad (3)$$

Figure 11 demonstrates a circuit configuration implementing this feature.



$$R_{CL} = \frac{15800\ (4.75V)}{I_{LIM}} - 7500\Omega$$

$$= \frac{75k\Omega}{I_{LIM}} - 7.5k\Omega$$

$I_{SET} = I_{LIM}/15800$

$V_{SET} = (Ref) + 4.75V - (7500\Omega)\ (I_{LIM})/15800$

OPA549 CURRENT LIMIT: 0A to 10A

DESIRED CURRENT LIMIT	RESISTOR[1] (R_{CL})	CURRENT (I_{SET})	VOLTAGE (V_{SET})
0A[2]	I_{LIM} Open	0µA	(Ref) + 4.75V
2.5A	22.6kΩ	158µA	(Ref) + 3.56V
3A	17.4kΩ	190µA	(Ref) + 3.33V
4A	11.3kΩ	253µA	(Ref) + 2.85V
5A	7.5kΩ	316µA	(Ref) + 2.38V
6A	4.99kΩ	380µA	(Ref) + 1.90V
7A	3.24kΩ	443µA	(Ref) + 1.43V
8A	1.87kΩ	506µA	(Ref) + 0.95V
9A	845Ω	570µA	(Ref) + 0.48V
10A	I_{LIM} Connected to Ref	633µA	(Ref)

NOTES: (1) Resistors are nearest standard 1% values. (2) Offset in the current limit circuitry may introduce approximately ±0.25A variation at low current limit values.

FIGURE 2. Adjustable Current Limit.

ENABLE/STATUS (E/S) PIN

The Enable/Status Pin provides two unique functions: 1) output disable by forcing the pin low, and 2) thermal shutdown indication by monitoring the voltage level at the pin. Either or both of these functions can be utilized in an application. For normal operation (output enabled), the E/S pin can be left open or driven high (at least 2.4V above Ref). A small value capacitor connected between the E/S pin and C_{REF} may be required for noisy applications.

Output Disable

To disable the output, the E/S pin is pulled to a logic low (no greater than 0.8V above Ref). Typically the output is shut down in 1µs. To return the output to an enabled state, the E/S pin should be disconnected (open) or pulled to at least 2.4V above Ref. It should be noted that driving the E/S pin high (output enabled) *does not defeat internal thermal shutdown*; however, it does prevent the user from monitoring the thermal shutdown status. Figure 3 shows an example implementing this function.

This function not only conserves power during idle periods (quiescent current drops to approximately 6mA) but also allows multiplexing in multi-channel applications. See Figure 12 for two OPA549s in a switched amplifier configuration. The on/off state of the two amplifiers is controlled by the voltage on the E/S pin. Under these conditions, the disabled device will behave like a 750pF load. Slewing faster than 3V/µs will cause leakage current to rapidly increase in devices that are disabled, and will contribute additional load. At high temperature (125°C), the slewing threshold drops to approximately 2V/µs. Input signals must be limited to avoid excessive slewing in multiplexed applications.

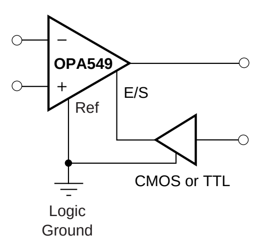


FIGURE 3. Output Disable.

Thermal Shutdown Status

The OPA549 has thermal shutdown circuitry that protects the amplifier from damage. The thermal protection circuitry disables the output when the junction temperature reaches approximately 160°C and allows the device to cool. When the junction temperature cools to approximately 140°C, the output circuitry is automatically re-enabled. Depending on load and signal conditions, the thermal protection circuit may cycle on and off. The E/S pin can be monitored to determine if the device is in shutdown. During normal operation, the voltage on the E/S pin is typically 3.5V above Ref. Once shutdown has occurred, this voltage drops to approximately 200mV above Ref. Figure 4 shows an example implementing this function.

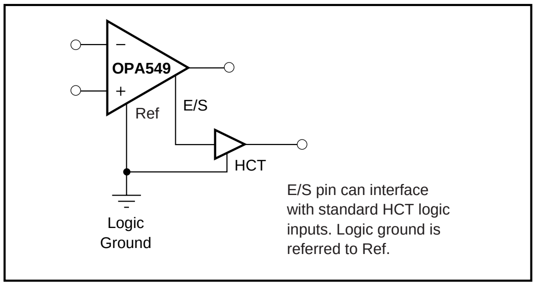


FIGURE 4. Thermal Shutdown Status.

External logic circuitry or an LED can be used to indicate if the output has been thermally shutdown, see Figure 10.

Output Disable and Thermal Shutdown Status

As mentioned earlier, the OPA549's output can be disabled and the disable status can be monitored simultaneously. Figure 5 provides an example of interfacing to the E/S pin.

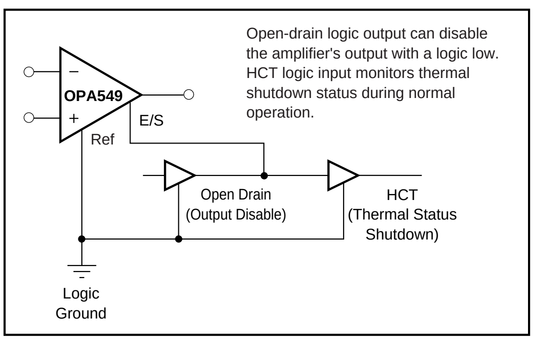


FIGURE 5. Output Disable and Thermal Shutdown Status.

SAFE OPERATING AREA

Stress on the output transistors is determined both by the output current and by the output voltage across the conducting output transistor, $V_S - V_O$. The power dissipated by the output transistor is equal to the product of the output current and the voltage across the conducting transistor, $V_S - V_O$. The Safe Operating Area (SOA curve, Figure 6) shows the permissible range of voltage and current.

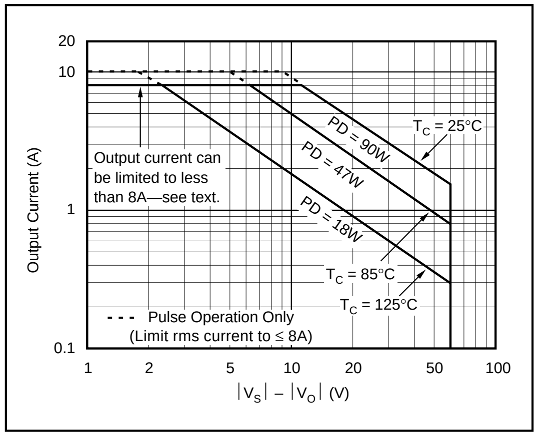


FIGURE 6. Safe Operating Area.

The safe output current decreases as $V_S - V_O$ increases. Output short circuits are a very demanding case for SOA. A short circuit to ground forces the full power-supply voltage (V+ or V–) across the conducting transistor. Increasing the case temperature reduces the safe output current that can be tolerated without activating the thermal shutdown circuit of the OPA549. For further insight on SOA, consult Application Report SBOA022 at the Texas Instruments web site (www.ti.com).

POWER DISSIPATION

Power dissipation depends on power supply, signal, and load conditions. For dc signals, power dissipation is equal to the product of output current times the voltage across the conducting output transistor. Power dissipation can be minimized by using the lowest possible power-supply voltage necessary to assure the required output voltage swing.

For resistive loads, the maximum power dissipation occurs at a dc output voltage of one-half the power-supply voltage. Dissipation with ac signals is lower. Application Bulletin SBOA022 explains how to calculate or measure power dissipation with unusual signals and loads.

THERMAL PROTECTION

Power dissipated in the OPA549 will cause the junction temperature to rise. Internal thermal shutdown circuitry shuts down the output when the die temperature reaches approximately 160°C and resets when the die has cooled to 140°C. Depending on load and signal conditions, the thermal protection circuit may cycle on and off. This limits the dissipation of the amplifier but may have an undesirable effect on the load.

Any tendency to activate the thermal protection circuit indicates excessive power dissipation or an inadequate heat sink. For reliable operation, junction temperature should be limited to 125°C maximum. To estimate the margin of safety in a complete design (including heat sink) increase the ambient temperature until the thermal protection is triggered.

Use worst-case load and signal conditions. For good reliability, thermal protection should trigger more than 35°C above the maximum expected ambient condition of your application. This produces a junction temperature of 125°C at the maximum expected ambient condition.

The internal protection circuitry of the OPA549 was designed to protect against overload conditions. It was not intended to replace proper heat sinking. Continuously running the OPA549 into thermal shutdown will degrade reliability.

AMPLIFIER MOUNTING AND HEAT SINKING

Most applications require a heat sink to assure that the maximum operating junction temperature (125°C) is not exceeded. In addition, the junction temperature should be kept as low as possible for increased reliability. Junction temperature can be determined according to the Equations:

$$T_J = T_A + P_D\theta_{JA} \quad (4)$$

where

$$\theta_{JA} = \theta_{JC} + \theta_{CH} + \theta_{HA} \quad (5)$$

- T_J = Junction Temperature (°C)
- T_A = Ambient Temperature (°C)
- P_D = Power Dissipated (W)
- θ_{JC} = Junction-to-Case Thermal Resistance (°C/W)
- θ_{CH} = Case-to-Heat Sink Thermal Resistance (°C/W)
- θ_{HA} = Heat Sink-to-Ambient Thermal Resistance (°C/W)
- θ_{JA} = Junction-to-Air Thermal Resistance (°C/W)

Figure 7 shows maximum power dissipation versus ambient temperature with and without the use of a heat sink. Using a heat sink significantly increases the maximum power dissipation at a given ambient temperature, as shown in Figure 7.

The challenge in selecting the heat sink required lies in determining the power dissipated by the OPA549. For dc output, power dissipation is simply the load current times the voltage developed across the conducting output transistor, $P_D = I_L (V_S - V_O)$. Other loads are not as simple. Consult the SBOA022 Application Report for further insight on calculating power dissipation. Once power dissipation for an application is known, the proper heat sink can be selected.

Heat Sink Selection Example—An 11-lead power ZIP package is dissipating 10 Watts. The maximum expected ambient temperature is 40°C. Find the proper heat sink to keep the junction temperature below 125°C (150°C minus 25°C safety margin).

Combining Equations (4) and (5) gives:

$$T_J = T_A + P_D (\theta_{JC} + \theta_{CH} + \theta_{HA}) \quad (6)$$

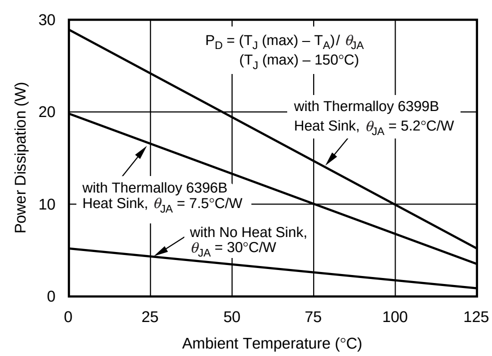
T_J, T_A, and P_D are given. θ_{JC} is provided in the Specifications Table, 1.4°C/W (dc). θ_{CH} can be obtained from the heat sink manufacturer. Its value depends on heat sink size, area, and material used. Semiconductor package type, mounting screw torque, insulating material used (if any), and thermal joint compound used (if any) also affect θ_{CH}. A typical θ_{CH} for a mounted 11-lead power ZIP package is 0.5°C/W. Now we can solve for θ_{HA}:

$$\theta_{HA} = [(T_J - T_A)/P_D] - \theta_{JC} - \theta_{CH}$$

$$\theta_{HA} = [(125°C - 40°C)/10W] - 1.4°C/W - 0.5°C/W$$

$$\theta_{HA} = 6.6°C/W$$

To maintain junction temperature below 125°C, the heat sink selected must have a θ_{HA} less than 6.6°C/W. In other words, the heat sink temperature rise above ambient must be less than 66°C (6.6°C/W • 10W). For example, at 10W Thermalloy model number 6396B has a heat sink temperature rise of 56°C (θ_{HA} = 56°C/10W = 5.6°C/W), which is below the required 66°C required in this example. Thermalloy model number 6399B has a sink temperature rise of 33°C (θ_{HA} = 33°C/10W = 3.3°C/W), which is also below the required 66°C required in this example. Figure 7 shows power dissipation versus ambient temperature for a 11-lead power ZIP package with the Thermalloy 6396B and 6399B heat sinks.



Thermalloy 6399B	θ_{HA}	= 5.6°C/W
assume	θ_{CH}	= 0.5°C/W
OPA549	θ_{JC}	= 1.4°C/W
	θ_{JA}	= 7.5°C/W

Thermalloy 6396B	θ_{HA}	= 3.3°C/W
assume	θ_{CH}	= 0.5°C/W
OPA549	θ_{JC}	= 1.4°C/W
	θ_{JA}	= 5.2°C/W

FIGURE 7. Maximum Power Dissipation vs Ambient Temperature.

Another variable to consider is natural convection versus forced convection air flow. Forced-air cooling by a small fan can lower θ_{CA} ($\theta_{CH} + \theta_{HA}$) dramatically. Some heat sink manufacturers provide thermal data for both of these cases. Heat sink performance is generally specified under idealized conditions that may be difficult to achieve in an actual application. For additional information on determining heat sink requirements, consult Application Report SBOA021.

As mentioned earlier, once a heat sink has been selected, the complete design should be tested under worst-case load and signal conditions to ensure proper thermal protection. Any tendency to activate the thermal protection circuitry may indicate inadequate heat sinking.

The tab of the 11-lead power ZIP package is electrically connected to the negative supply, V–. It may be desirable to isolate the tab of the 11-lead power ZIP package from its mounting surface with a mica (or other film) insulator. For lowest overall thermal resistance, it is best to isolate the entire heat sink/OPA549 structure from the mounting surface rather than to use an insulator between the semiconductor and heat sink.

OUTPUT STAGE COMPENSATION

The complex load impedances common in power op amp applications can cause output stage instability. For normal operation, output compensation circuitry is typically not required. However, for difficult loads or if the OPA549 is intended to be driven into current limit, an R/C network may be required. Figure 8 shows an output R/C compensation (snubber) network which generally provides excellent stability.

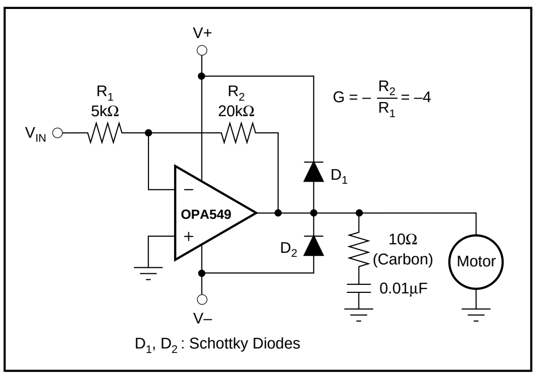


FIGURE 8. Motor Drive Circuit.

A snubber circuit may also enhance stability when driving large capacitive loads (> 1000pF) or inductive loads (motors, loads separated from the amplifier by long cables). Typically, 3Ω to 10Ω resistors in series with 0.01μF to 0.1μF capacitors is adequate. Some variations in circuit values may be required with certain loads.

OUTPUT PROTECTION

Reactive and EMF-generating loads can return load current to the amplifier, causing the output voltage to exceed the power-supply voltage. This damaging condition can be avoided with clamp diodes from the output terminal to the power supplies, as shown in Figure 8. Schottky rectifier diodes with a 8A or greater continuous rating are recommended.

VOLTAGE SOURCE APPLICATION

Figure 9 illustrates how to use the OPA549 to provide an accurate voltage source with only three external resistors. First, the current limit resistor, R_{CL}, is chosen according to the desired output current. The resulting voltage at the I_{LIM} pin is constant and stable over temperature. This voltage, V_{CL}, is connected to the noninverting input of the op amp and used as a voltage reference, thus eliminating the need for an external reference. The feedback resistors are selected to gain V_{CL} to the desired output voltage level.

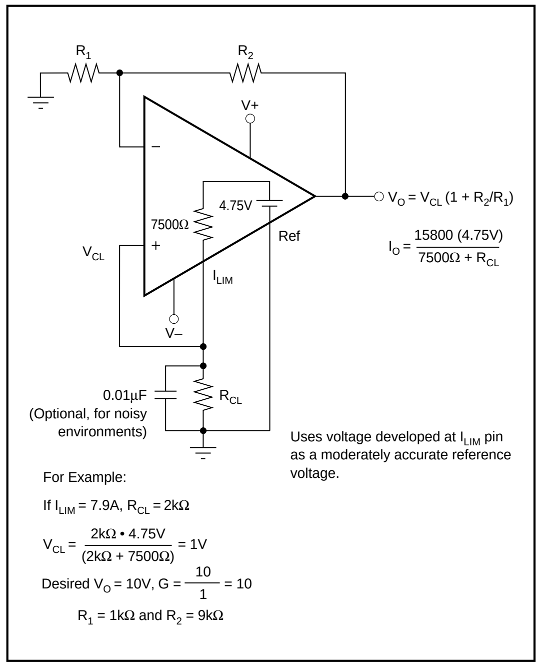


FIGURE 9. Voltage Source.

PROGRAMMABLE POWER SUPPLY

A programmable source/sink power supply can easily be built using the OPA549. Both the output voltage and output current are user-controlled. See Figure 10 for a circuit using potentiometers to adjust the output voltage and current while Figure 11 uses DACs. An LED connected to the E/S pin through a logic gate indicates if the OPA549 is in thermal shutdown.

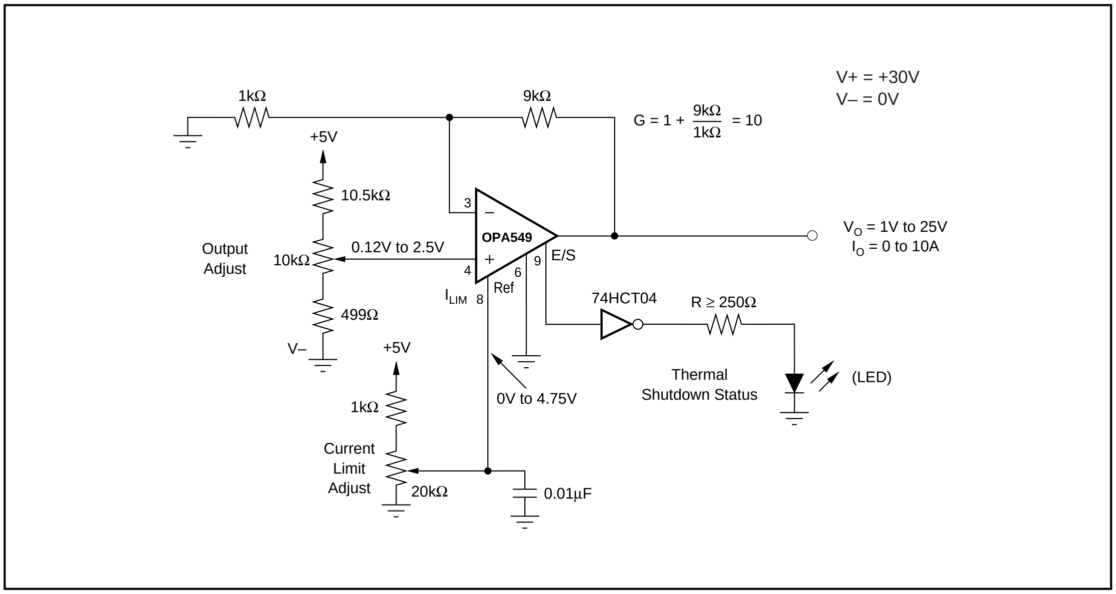


FIGURE 10. Resistor-Controlled Programmable Power Supply.

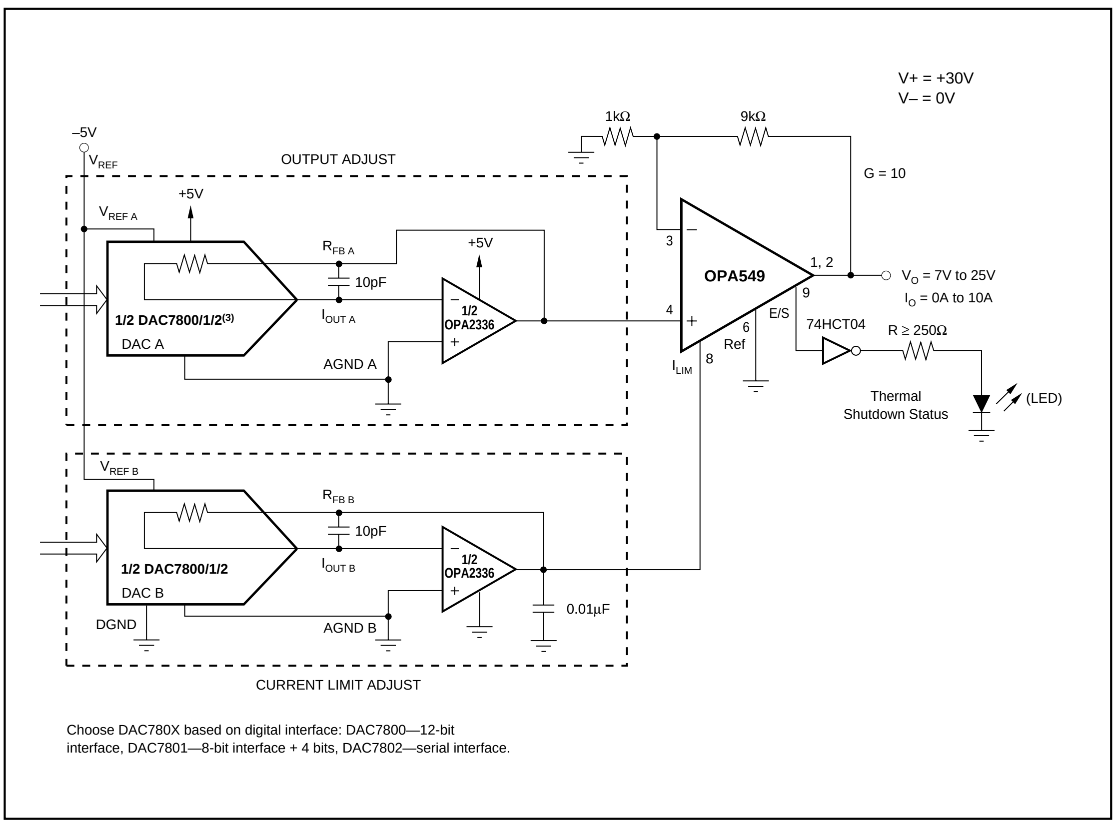


FIGURE 11. Digitally-Controlled Programmable Power Supply.

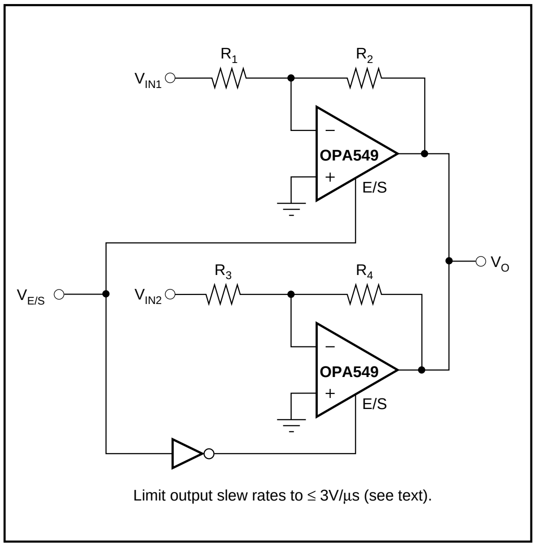


FIGURE 12. Switched Amplifier.

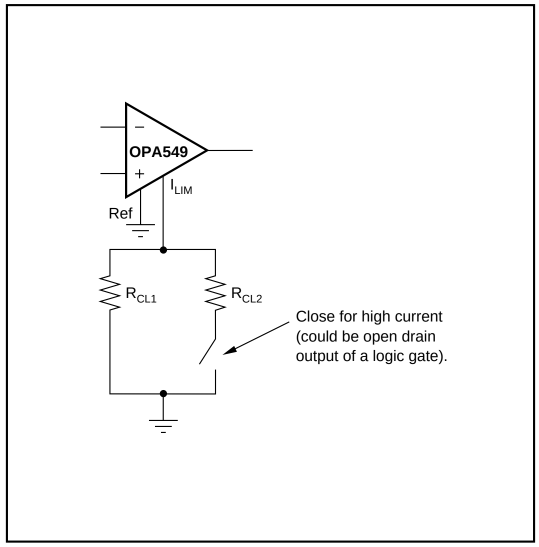


FIGURE 13. Multiple Current Limit Values.

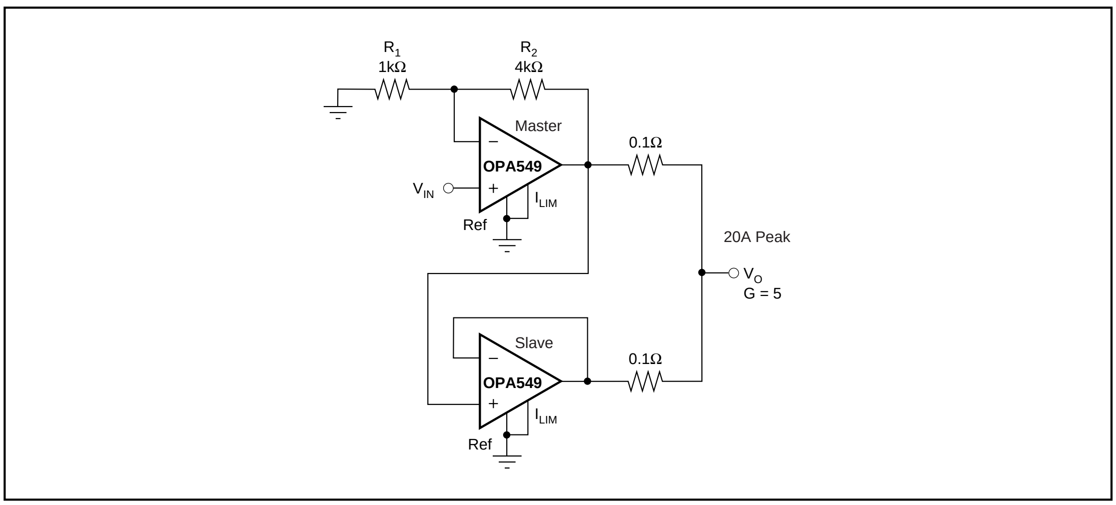


FIGURE 14. Parallel Output for Increased Output Current.

PACKAGING INFORMATION

Orderable Device	Status (1)	Package Type	Package Drawing	Pins	Package Qty	Eco Plan (2)	Lead/ Ball Finish	MSL Peak Temp (3)	Samples (Requires Login)
OPA549S	ACTIVE	Power Package	KVC	11	25	Green (RoHS & no Sb/Br)	CU SN	N / A for Pkg Type	
OPA549SG3	ACTIVE	Power Package	KVC	11	25	Green (RoHS & no Sb/Br)	CU SN	N / A for Pkg Type	
OPA549T	ACTIVE	TO-220	KV	11	25	Green (RoHS & no Sb/Br)	CU SN	N / A for Pkg Type	
OPA549TG3	ACTIVE	TO-220	KV	11	25	Green (RoHS & no Sb/Br)	CU SN	N / A for Pkg Type	

(1) The marketing status values are defined as follows:
ACTIVE: Product device recommended for new designs.
LIFEBUY: TI has announced that the device will be discontinued, and a lifetime-buy period is in effect.
NRND: Not recommended for new designs. Device is in production to support existing customers, but TI does not recommend using this part in a new design.
PREVIEW: Device has been announced but is not in production. Samples may or may not be available.
OBSOLETE: TI has discontinued the production of the device.

(2) Eco Plan - The planned eco-friendly classification: Pb-Free (RoHS), Pb-Free (RoHS Exempt), or Green (RoHS & no Sb/Br) - please check http://www.ti.com/productcontent for the latest availability information and additional product content details.
TBD: The Pb-Free/Green conversion plan has not been defined.
Pb-Free (RoHS): TI's terms "Lead-Free" or "Pb-Free" mean semiconductor products that are compatible with the current RoHS requirements for all 6 substances, including the requirement that lead not exceed 0.1% by weight in homogeneous materials. Where designed to be soldered at high temperatures, TI Pb-Free products are suitable for use in specified lead-free processes.
Pb-Free (RoHS Exempt): This component has a RoHS exemption for either 1) lead-based flip-chip solder bumps used between the die and package, or 2) lead-based die adhesive used between the die and leadframe. The component is otherwise considered Pb-Free (RoHS compatible) as defined above.
Green (RoHS & no Sb/Br): TI defines "Green" to mean Pb-Free (RoHS compatible), and free of Bromine (Br) and Antimony (Sb) based flame retardants (Br or Sb do not exceed 0.1% by weight in homogeneous material)

(3) MSL, Peak Temp. -- The Moisture Sensitivity Level rating according to the JEDEC industry standard classifications, and peak solder temperature.

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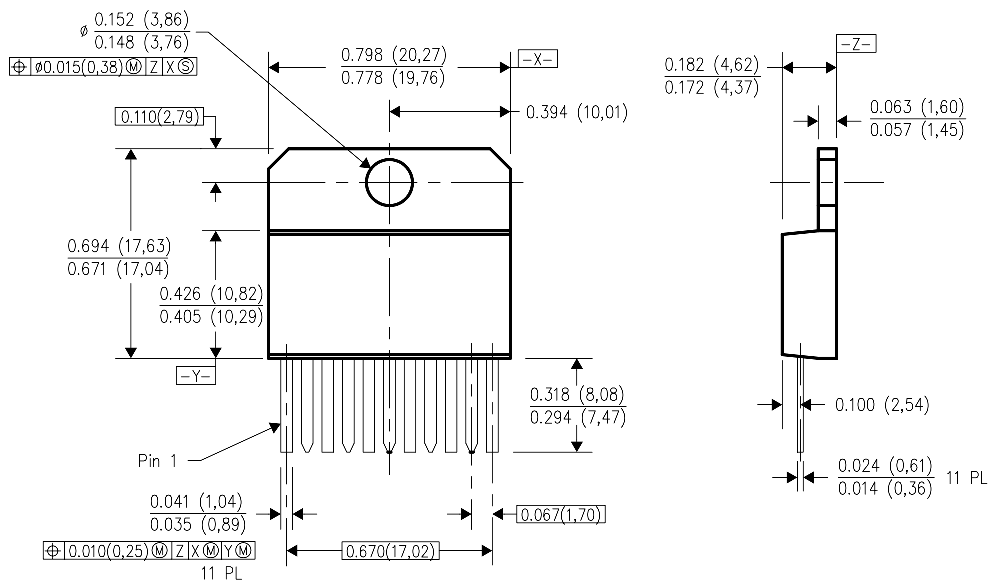
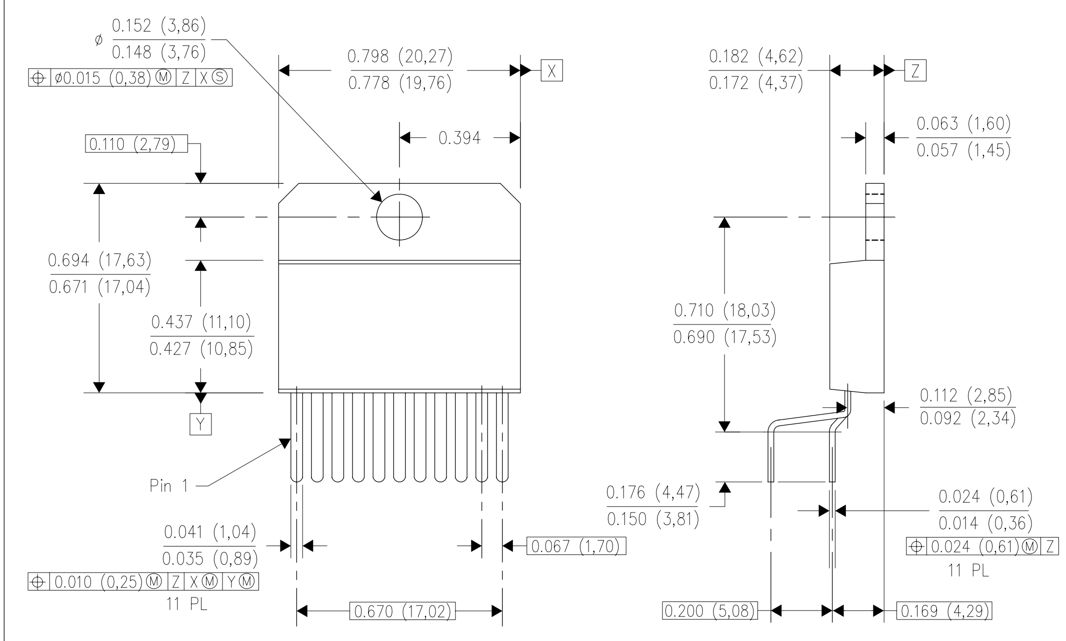
KVC (R–PSFM–T11) PLASTIC FLANGE–MOUNT



NOTES:
A. All linear dimensions are in inches (millimeters).
B. This drawing is subject to change without notice.
C. Controlling dimension in inches.
D. Falls within JEDEC MO–48–AA. Reference for body dimensions only (excluding lead forming dimensions).

KV (R–PZFM–T11) PLASTIC FLANGE–MOUNT



NOTES:
A. All linear dimensions are in inches (millimeters).
B. This drawing is subject to change without notice.
C. Controlling dimension: inch.
D. All lead dimensions apply before solder dip.
E. Falls within JEDEC MO–48–AA.

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DLP® Products	www.dlp.com	Consumer Electronics	www.ti.com/consumer-apps
DSP	dsp.ti.com	Energy and Lighting	www.ti.com/energy
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Logic	logic.ti.com	Security	www.ti.com/security
Power Mgmt	power.ti.com	Space, Avionics and Defense	www.ti.com/space-avionics-defense
Microcontrollers	microcontroller.ti.com	Video and Imaging	www.ti.com/video
RFID	www.ti-rfid.com		
OMAP Mobile Processors	www.ti.com/omap		
Wireless Connectivity	www.ti.com/wirelessconnectivity		
	TI E2E Community Home Page		e2e.ti.com

Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265

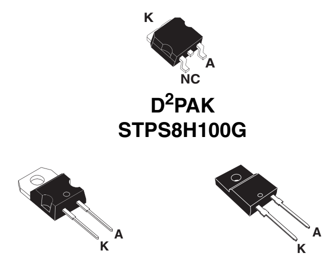
![ST logo]

STPS8H100

High voltage power Schottky rectifier

Main product characteristics

$I_{F(AV)}$	8 A
V_{RRM}	100 V
T_j	175° C
V_F(max)	0.58 V



D^2PAK
STPS8H100G

TO-220AC
STPS8H100D

TO-220FPAC
STPS8H100FP

Features and benefits

- Negligible switching losses
- High junction temperature capability
- Low leakage current
- Good trade off between leakage current and forward voltage drop
- Insulated package:
  - TO-220FPAC
    Insulating voltage = 2000 V DC
    Typical package capacitance = 12 pF
- Avalanche capability specified

Description

Schottky barrier rectifier designed for high frequency compact Switched Mode Power Supplies such as adaptators and on board DC/DC converters.

Order Codes

Part Number	Marking
STPS8H100D	STPS8H100D
STPS8H100G	STPS8H100G
STPS8H100G-TR	STPS8H100G
STPS8H100FP	STPS8H100FP

Table 1. Absolute ratings (limiting values)

Symbol	Parameter			Value	Unit
V_{RRM}	Repetitive peak reverse voltage			100	V
$I_{F(RMS)}$	RMS forward voltage			30	A
$I_{F(AV)}$	Average forward current δ = 0.5	TO-220AC, D^2PAK	T_C = 165° C	8	A
		DO-15	T_C = 150° C		
I_{FSM}	Surge non repetitive forward current		t_p = 10 ms sinusoidal	250	A
P_{ARM}	Repetitive peak avalanche power		t_p = 1 µs T_j = 25° C	10800	W
T_{stg}	Storage temperature range			-65 to + 175	° C
T_j	Maximum operating junction temperature			175	° C

1 Characteristics

Table 2. Thermal resistance

Symbol	Parameter		Value	Unit
$R_{th(j-c)}$	Junction to case	TO-220AC, D^2PAK	1.6	° C/W
		TO-220FPAC	4	

Table 3. Static electrical characteristics (per diode)

Symbol	Parameter	Tests conditions		Min.	Typ	Max.	Unit
I_R (1)	Reverse leakage current	T_j = 25° C	$V_R = V_{RRM}$			4.5	µA
		T_j = 125° C			2	6.0	mA
V_F (2)	Forward voltage drop	T_j = 25° C	I_F = 8 A			0.71	V
		T_j = 125° C			0.56	0.58	
		T_j = 25° C	I_F = 10 A			0.77	
		T_j = 125° C			0.59	0.64	
		T_j = 25° C	I_F = 16 A			0.81	
		T_j = 125° C			0.65	0.68	

1. t_p = 5 ms, δ < 2%
2. t_p = 380 µs, δ < 2%

To evaluate the conduction losses use the following equation:
$P = 0.48 \times I_{F(AV)} + 0.0125\ I_F{}^2{}_{(RMS)}$

Figure 1. Average forward power dissipation versus average forward current

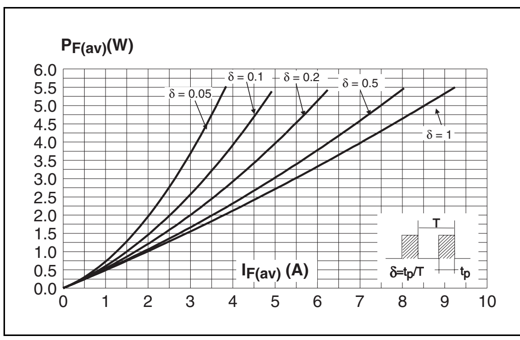


Figure 2. Normalized avalanche power derating versus pulse duration

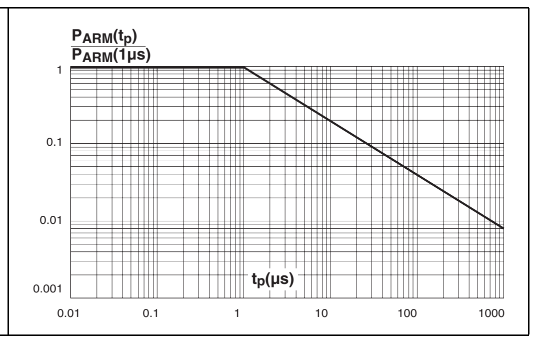


Figure 3. **Normalized avalanche power derating versus junction temperature**

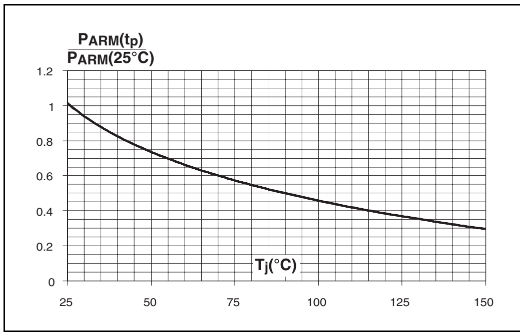


Figure 4. **Average forward current versus ambient temperature, δ = 0.5, (TO-220AC, D^2PAK)**

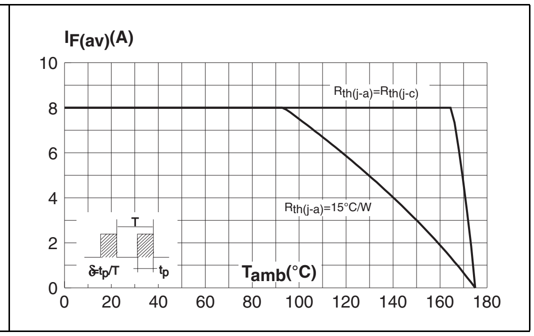


Figure 5. **Average forward current versus ambient temperature, δ = 0.5, (TO-220FPAC)**

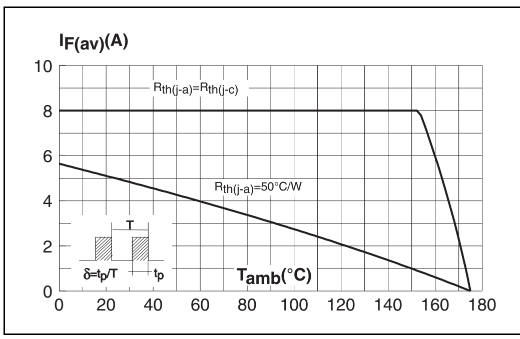


Figure 6. **Non repetitive surge peak forward current versus overload duration - maximum values, per diode (TO-220AC, D^2PAK)**

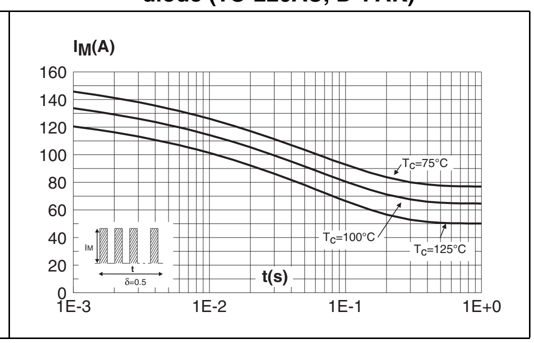


Figure 7. **Non repetitive surge peak forward current versus overload duration - maximum values (TO-220FPAC)**

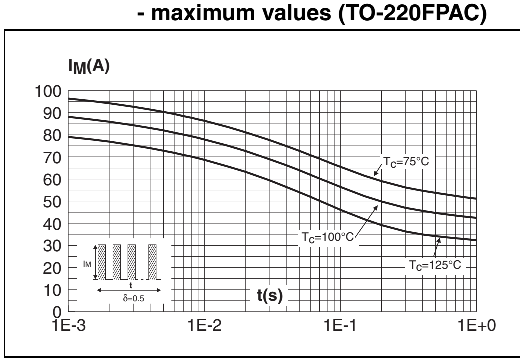


Figure 8. **Relative variation of thermal impedance junction to case versus pulse duration (TO-220AC, D^2PAK)**

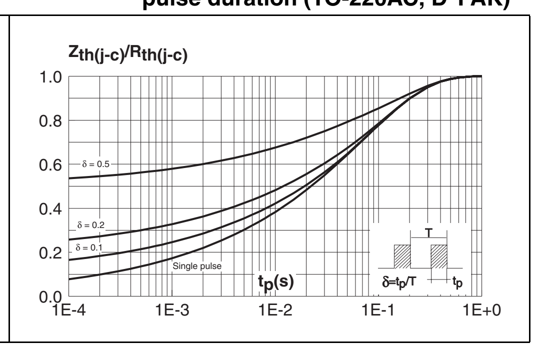


Figure 9. Relative variation of thermal impedance junction to case versus pulse duration (TO-220FPAC)

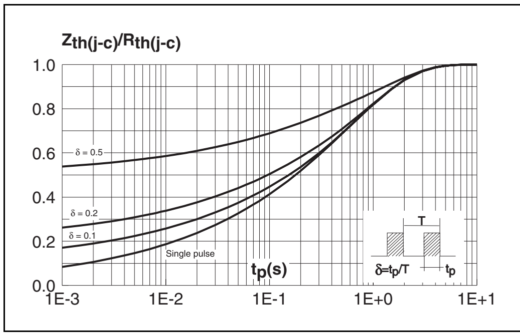


Figure 10. Reverse leakage current versus reverse voltage applied (typical values)

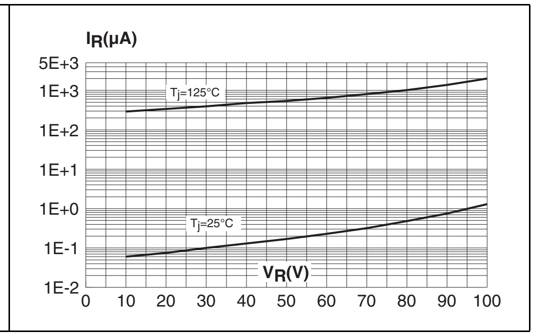


Figure 11. Junction capacitance versus reverse voltage applied (typical values)

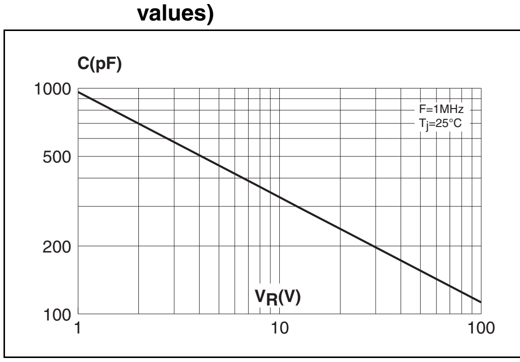


Figure 12. Forward voltage drop versus forward current (maximum values)

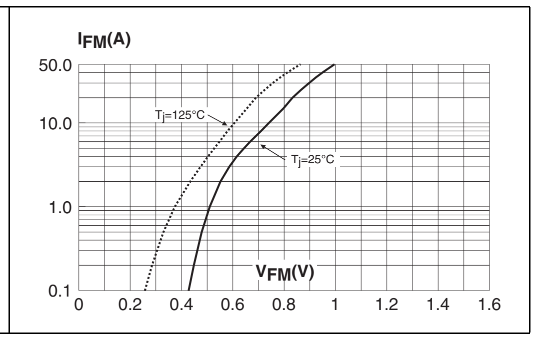
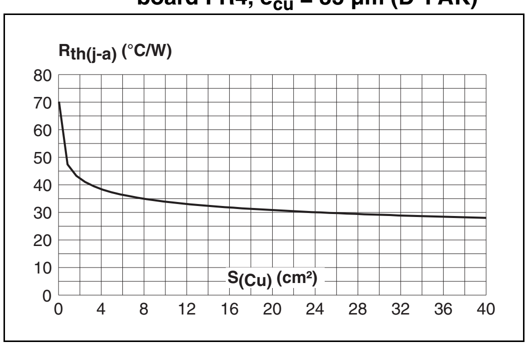


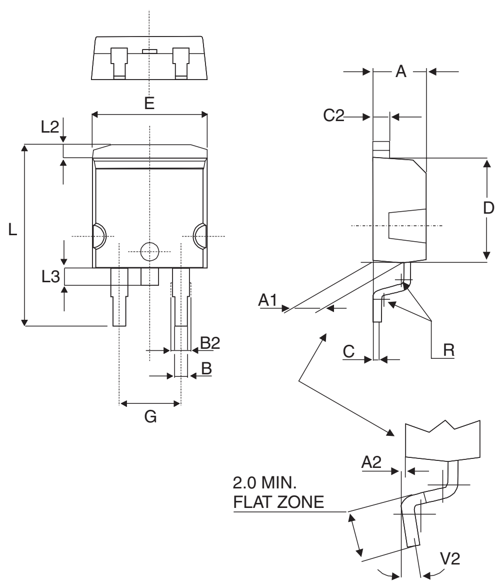
Figure 13. Thermal resistance junction to ambient versus copper surface under tab - Epoxy printed circuit board FR4, e_{cu} = 35 µm (D^2PAK)



2 Package information

Epoxy meets UL94, V0.

Table 4. D²PAK Dimensions



REF.	Dimensions			
	Millimeters		Inches	
	Min.	Max.	Min.	Max.
A	4.40	4.60	0.173	0.181
A1	2.49	2.69	0.098	0.106
A2	0.03	0.23	0.001	0.009
B	0.70	0.93	0.027	0.037
B2	1.14	1.70	0.045	0.067
C	0.45	0.60	0.017	0.024
C2	1.23	1.36	0.048	0.054
D	8.95	9.35	0.352	0.368
E	10.00	10.40	0.393	0.409
G	4.88	5.28	0.192	0.208
L	15.00	15.85	0.590	0.624
L2	1.27	1.40	0.050	0.055
L3	1.40	1.75	0.055	0.069
M	2.40	3.20	0.094	0.126
R	0.40 typ.		0.016 typ.	
V2	0°	8°	0°	8°

Figure 14. D²PAK footprint dimensions (in mm)

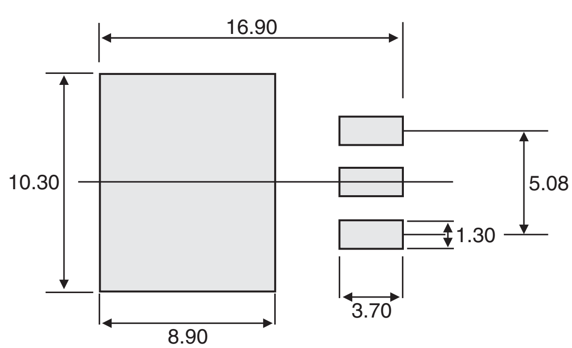
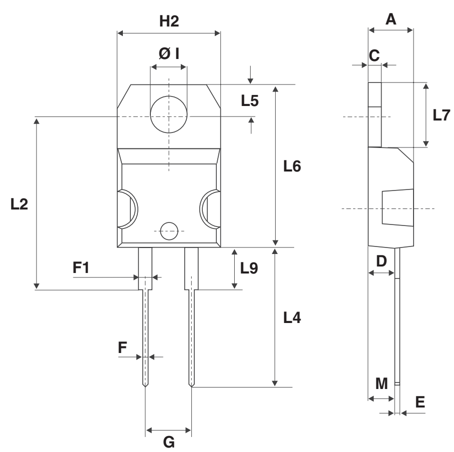
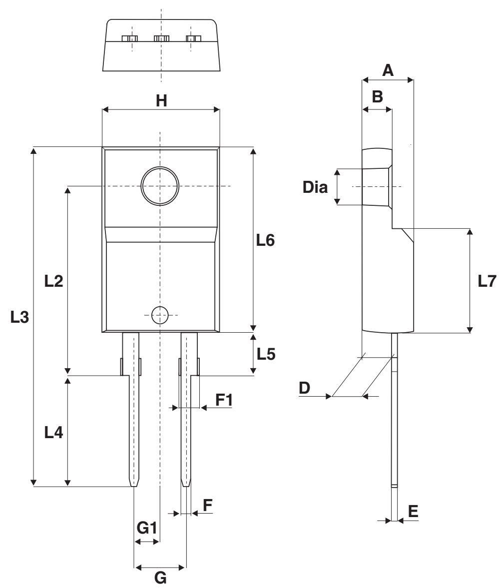


Table 5. TO-220AC Dimensions



REF.	Dimensions			
	Millimeters		Inches	
	Min.	Max.	Min.	Max.
A	4.40	4.60	0.173	0.181
C	1.23	1.32	0.048	0.051
D	2.40	2.72	0.094	0.107
E	0.49	0.70	0.019	0.027
F	0.61	0.88	0.024	0.034
F1	1.14	1.70	0.044	0.066
G	4.95	5.15	0.194	0.202
H2	10.00	10.40	0.393	0.409
L2	16.40 typ.		0.645 typ.	
L4	13.00	14.00	0.511	0.551
L5	2.65	2.95	0.104	0.116
L6	15.25	15.75	0.600	0.620
L7	6.20	6.60	0.244	0.259
L9	3.50	3.93	0.137	0.154
M	2.6 typ.		0.102 typ.	
Diam. I	3.75	3.85	0.147	0.151

Table 6. TO-220FPAC Dimensions



REF.	Dimensions			
	Millimeters		Inches	
	Min.	Max.	Min.	Max.
A	4.4	4.6	0.173	0.181
B	2.5	2.7	0.098	0.106
D	2.5	2.75	0.098	0.108
E	0.45	0.70	0.018	0.027
F	0.75	1	0.030	0.039
F1	1.15	1.70	0.045	0.067
G	4.95	5.20	0.195	0.205
G1	2.4	2.7	0.094	0.106
H	10	10.4	0.393	0.409
L2	16 Typ.		0.63 Typ.	
L3	28.6	30.6	1.126	1.205
L4	9.8	10.6	0.386	0.417
L5	2.9	3.6	0.114	0.142
L6	15.9	16.4	0.626	0.646
L7	9.00	9.30	0.354	0.366
Dia.	3.00	3.20	0.118	0.126

In order to meet environmental requirements, ST offers these devices in ECOPACK® packages. These packages have a lead-free second level interconnect. The category of second level interconnect is marked on the package and on the inner box label, in compliance with JEDEC Standard JESD97. The maximum ratings related to soldering conditions are also marked on the inner box label. ECOPACK is an ST trademark. ECOPACK specifications are available at: www.st.com.

3 Ordering information

Ordering type	Marking	Package	Weight	Base qty	Delivery mode
STPS8H100D	STPS8H100D	TO-220AC	1.86 g	50	Tube
STPS8H100FP	STPS8H100FP	TO-220FPAC	1.9 g	50	Tube
STPS8H100G	STPS8H100G	D^2PAK	1.48 g	50	Tube
STPS8H100G-TR	STPS8H100G	D^2PAK	1.48 g	500	Tape and reel

4 Revision history

Date	Revision	Description of Changes
Jul-2003	6D	Last update.
1-June-2006	10	Reformatted to current standard. Added ECOPACK statement. Changed nF to pF in Figure 11. Revision number set to 10 to align with on-line versioning.

APPENDIX C: PERMISSION FROM SCHULTER

June, 27 2012

Spencer Atkin
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208-339-5738
s.atkin@aggiemail.usu.edu

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- Abbildung 1.3: 8-nutiger Kommutator

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