HIGH CURRENT VOLTAGE REGULATORS

$2.00

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369 Whisman Road, Mountain View, California 94043

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Introduction

The Fairchild Hybrid Division was established to fulfill the high-volume, high-quality, low-cost requirements of a growing number of companies turning to hybrid technology. A broad range of precision hybrid voltage regulators is available off-the-shelf as well as automotive ignition systems. For custom hybrid programs we offer full design capability with rapid prototyping and translation to volume production.

No other hybrid manufacturer in the world can match Fairchild's total capabilities. Since Fairchild is one of the world's leading semiconductor manufacturers, there is no need to depend on outside suppliers for delivery, reliability or quality of semiconductor components. Our years of experience in developing and qualifying sources of passive components for the automotive hybrid market assure dependable performance in the finished product. Facilities in Northern California and Hong Kong can turn out more reliable hybrid products in a day than most hybrid suppliers can produce in a week.

Fairchild is equipped to produce hybrids for any company in any business that uses hybrid products in large quantities. The markets served include automotive, consumer electronics, computer, telecommunications, industrial controls, aerospace and military.

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Table of Contents

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Chapter 1 Capabilities

1

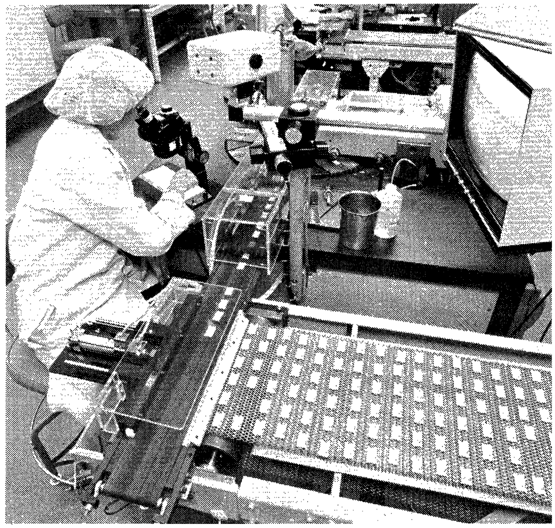
Design

Hybrids offer all the advantages of other semiconductor components—small size, high reliability and low cost in high volume—while providing fully tested functions that readily interface with the user's overall system. In addition they offer unique advantages such as ability to mix technologies not achievable monolithically, ability to operate in extremely hostile environments, and improved ratios of reliability to complexity, while retaining their size advantage over the discrete approach. A broad range of process techniques has been developed from which to select the manufacturing method best suited to a specific objective. This variety of techniques provides for wide design flexibility with minimal process constraints. Also, since the semiconductors used in Fairchild hybrids are almost exclusively supplied internally, parameters can be selected and stringently controlled for compatibility with the overall circuit design.

New process development and materials research takes place on a continuous basis to ensure that our methods and materials are selected from the best alternatives available. Internal IC design compatibility exists for linear and other bipolar technologies, power transistors are also designed within the division. Design and production of custom products follow a thorough routine. Once a user's input is submitted, whether in the form of functional specifications, a circuit diagram or breadboard, a comprehensive cost analysis and circuit evaluation is made. Only when Fairchild is satisfied with the circuit/cost analysis is the paper design submitted. Upon approval, a breadboard is produced for user evaluation. A complete detailed layout is then constructed for use in producing preliminary parts for customer approval. When the customer gives the go-ahead, volume production begins. In either instance, standard or custom, a Fairchild hybrid subsystem, fabricated using mixed technologies, is fully tested and delivered on time, in volume quantities.

Production

Fairchild excels in high volume production of hybrid devices. Facilities include: complete thick film production; rubylith, photo reduction and screen manufacturing; active and passive laser trim; wafer sort and scribe; assembly and packaging; testing and quality control.



All materials are pre-tested and qualified before admission to the thick film process; pastes are subjected, on a lot basis, to stringent incoming tests using computer controlled equipment. The flatness and surface finish of the alumina substrates are rigidly controlled. Printing onto these substrates is accomplished using automatic magazine fed machines having a high degree of stability and print alignment. The substrates are then transferred by automatic collation equipment to a drying and firing furnace, in which the environment is moisture and oil-free to ensure the control required for maintenance of tight resistor temperature coefficient of resistance (TCR) distributions. Furnace zones are microprocessor controlled, and interfaced to a computer system capable of providing check profiles on demand. Post-firing offload is accomplished using automatic equipment eliminating handling damage. Base conductor and dielectric layers are individually printed and fired; subsequent resistor prints are dried between applications and finally co-fired to the desired pre-trim values. Process monitoring, again using computer controlled equipment, constantly verifies visual, electrical and dry-print thickness measurements, assuring high-yield low-cost production. Tailoring the substrate resistors to final value is achieved using active and passive laser trimming systems with carousel feeds and closed-circuit TV monitoring systems. Extensive computer control at this step provides flexibility and accuracy.

Total thick-film production capability encompasses gold and palladium-silver conductor systems, resistor prints in the range 1Ω to 10MΩ, high quality pinhole-free dielectric systems and multilayer techniques. Fairchild's thick film capability is complemented by wafer sort and scribing facilities employing diamond and saw techniques. Wafers are supplied to this area from the company's integrated circuit and discrete fabrication areas. Linear devices and power transistors can be supplied from the division's internal capability. The proximity of all facilities ensures rapid resolution of any technical or scheduling problems.

Hybrid production utilizes all the standard manufacturing methods plus a number of proprietary processes designed to meet exacting customer requirements. For example, an exclusive flip-chip solder reflow process has been developed to eliminate bonding steps in large-volume custom applications, while simultaneously providing an extemely rugged micro-interconnect capable of withstanding wide temperature excursions and the most demanding corrosion and vibration environments. For applications involving the use of LSI chips with large area and I/O counts, a versatile interconnect scheme using the latest tape automated bonding techniques is currently under development. Because of its potential for computer controlled assembly when combined with automatic pick and place equipment, tape carrier appears to be the best solution for high volume production of reliable hybrids at minimum cost.

Many other advanced techniques in hybrid technology have been developed over the past decade to meet a variety of customer objectives. This trend will continue. Fairchild is committed to producing the highest quality hybrids possible to endure the most stringent environmental conditions in both commercial and military applications.

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Chapter 2 Reliability

What Is Reliability

Reliability is defined as the behavior of a component, a machine, or a system, as a function of time. Statistically, it is also expressed as the probability that the item will perform a required function under established conditions for a given period of time.

This *Time* is a variable covering minutes, hours, months, or years. Some of the equipment in oil well logging operations must have, on the average, a useful life of only five minutes. At the other extreme, the telephone companies want their equipment to last a minimum of 25 years. Missiles are subjected to thousands of "g" (gravity) forces and the components must not lose their monitoring, telemetering, or sensing capabilities during the first critical seconds of their flight. Power supplies are constantly being thermally cycled as they are turned on and off. Voltage regulators undergo similar stresses, and both must provide thousands of hours of flawless operation.

Some Reliability Terms

Initial/Infant Failure

A failure occurring during the early stages of operation, the failure rate during infancy is higher than during long term operation.

Infant failures are caused by weaknesses not removed by the numerous inspection operations, if detectable at all.

Random Failure

A failure which occurs sometimes between the infant mortality and the wear-out periods; the failure rate during this period is generally constant.

Wear-Out Failure

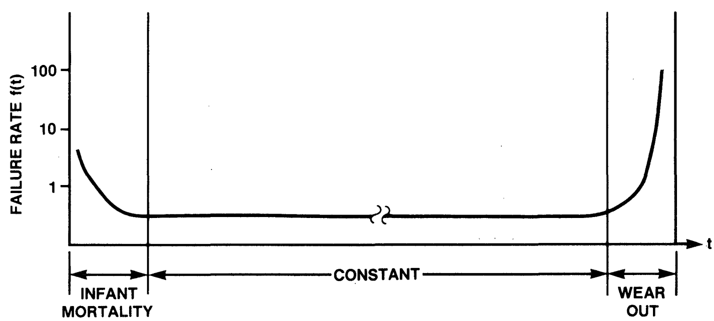
This failure occurs as a result of degradation, physical or chemical.

- Repeated thermal cycling will cause electrical discontinuities between conductors having different coefficients of thermal expansion.
- Voltage incursions may cause shorting across a dielectric.
- Excessive current densities may introduce metal migration, causing shorts.
- Continuous vibrations may cause loss of contact or create loose conducting particles and subsequent shorting.
- Moisture is known to degrade components because of chemical reactions resulting in parameter changes.

Figure 1 suggests that, to obtain the lowest failure rate a removal of the infant mortality weaknesses is required. Traditionally, this removal was done, as a rule, on military products and for certain nonmilitary users requiring the highest reliability.

More recently, the growing complexities of various electronic systems, of multi-million dollar computers, some of which contain as many as a quarter million integrated circuits, have created demands by the commercial users for a reliability level similar to that of the military, the difference being only in the temperature range of operation.

Fig. 1 A Diagrammatic Representation of Failure Patterns



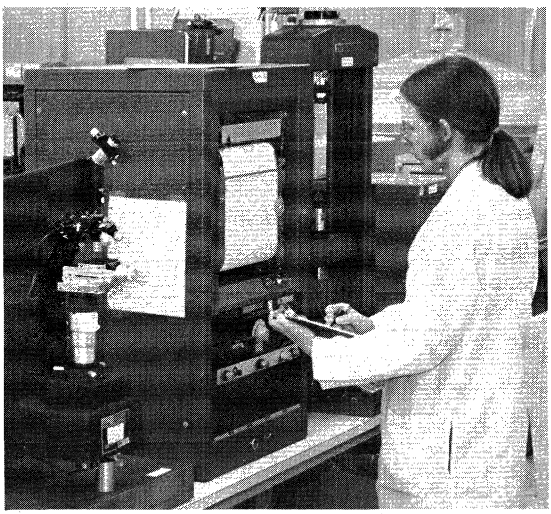
How Is Reliability Obtained?

Design

High Current voltage regulators are comprised of three major parts: Active components, substrate and package.

A linear integrated circuit (LIC) and a power transistor provide all the drives and controls: A start circuit, voltage regulation, current sourcing, short-circuit protection, thermal shutdown and amplification.

The high current densities typical of high power devices and especially of power transistors, integrated on a chip or discrete, require uniform current distribution, especially along the emitter contacts, to prevent current hogging, hot spots and excessive heating. This joule heating can be sufficient to reach the aluminum-silicon eutectic temperature, melt the silicon and short the emitter and collector. This effect is minimized by large geometries that decrease the current densities and by diffusion and concentration profiles that insure better current distribution. Long term aluminum migration, a concern wherever large current densities exist, is also eliminated by the proper "sizing" of the conduction lines. This concept of safe margins and of conservative design rules applies also to all components procured from outside vendors.



Materials

The substrate provides a base for the thick film resistors, the connections for attaching the active components, and a means of heat dissipation.

The package connects to the equipment for both testing and usage and gives an additional path for heat dissipation.

In *Table 1* the major component parts are listed with tests normally performed for conformance, and, for reliability.

Table 1 Test and Control of Major Component Parts

Piece Part	Test/Control
Substrate	Mechanical dimensions Mechanical strength Thermal conduction properties Electrical insulation Chemical composition Thermal cycling
Package	Mechanical dimensions and characteristics Solderability of leads Electrical insulation Lead seal check Lead strength
Solder Preform	Mechanical dimensions Chemical composition
Die	Visual inspection
Capacitor	Mechanical dimensions Electrical properties Electrical stress on a sample

Process

Control of Wafer Fabrication Operations

Wafer fabrication is a very complex and disciplined operation and listing all the numerous control points and monitoring operations is beyond the scope of this Reliability section.

Reliability, quality and yield are major concerns for any wafer fab operation, and by extension, of any manufacturing operation. Only a few items will be mentioned here:

- Wafer purity
- Analysis of chemicals
- Dust particle concentrations
- Temperature/humidity
- Analysis of dopant sources
- Gas flows into furnaces
- Furnace profiles
- C/V plots for checking ionic drift
- Equipment calibration
- Exposure
- Development
- Etching
- Cleaning, etc. . . .

Control of Assembly Operations
Completed wafers are electrically probed, and good dice are identified for assembly.

Every assembly operation is critical, and every effort is made to guarantee long term life of the product.

A typical assembly flow, shown in *Table 2,* shows both the operation and its control equivalent.

Environmental Testing
An additional level of reliability can be obtained by performing environmental and electrical tests along the flow referenced in MIL-STD 883 (Test Methods & Procedures for Microelectronics), Method 5008 (Test Procedures for Hybrids and Multichip Microcircuits). This "hi-rel" flow performs the following:

- Storage. Isolates product not capable, for mechanical reasons, of storage at 150°C for 24 hours.
- Temperature cycle. Eliminates product exhibiting mechanical damages that would cause functional failures.
- Constant acceleration. Eliminates structural and mechanical weaknesses:
  - — Poor wire-to-die bonding
  - — Poor substrate-to-package attach
  - — Poor die-to-substrate attach
- Seal. Prevents the components, active and passive, from being influenced by outside factors of the working environment, mostly humidity.
- Burn-In. Screens or eliminates all marginal devices, those with inherent defects resulting from manufacturing, aberrations which cause time and stress dependent failures. In the absence of burn-in, these defective units would result in infant/early mortality failures (see paragraph on "Some Reliability Terms").
- Electrical testing at temperature extremes. Removes all units not meeting functional and parametric criteria.

Standard and hi-rel flows are compared on *Table 3.*

Table 2 A Typical Assembly Operation

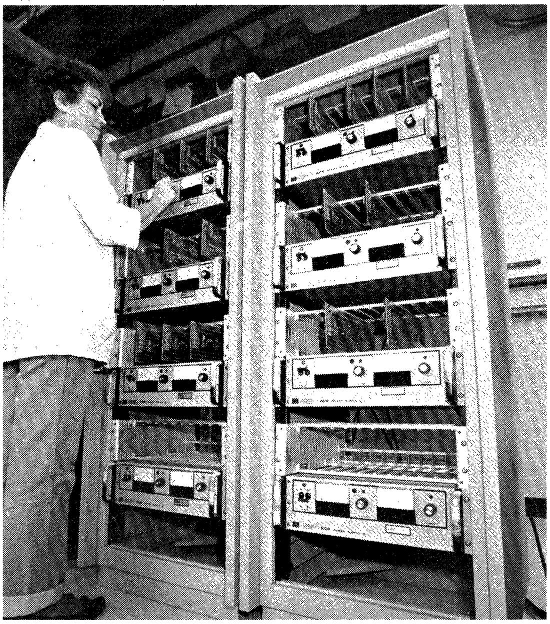
Operation	Control
Scribing (or sawing)	Maintenance
Separation of good & bad die	Visual inspection, QC sample, conformance inspection
Die attach	Functional check for adherence and wetting; 100% X-Ray; 100%visual check Push Test
Substrate attach	Functional check for adherence and wetting; 100% X-Ray; 100% visual check Push Test
Wire bonding	Incoming wiretest Pull strength Visual check
Optical check (preseal)	QC sample conformance inspection
Package seal	Hermeticity check for fine & gross leak
100% Optical check (post-seal)	QC sample conformance inspection

Table 3 A Comparison of Hi Rel and Standard Flows

Fairchild Unique Level B

	Hi Rel	Standard
O	Package Seal	Package Seal
O	Post Seal Visual Inspection	Post Seal Visual Inspection
Q	Post Seal Sample Inspection	Post Seal Sample Inspection
O	Bake — 150°C/24 hrs	—
O	Temperature Cycling -65°C to 150°C — 10X	—
O	Constant Acceleration 10KG — Y1 Axis	—
O	Seal — Fine Gross	—
Q	—	Seal — Fine Gross
O	Electrical Test	Electrical Test
O	Burn In — T_J = 150°C Max Time = 160 Hrs. Min.	—
O	Electrical Test — Post Burn In • 25°C • −55°C • 125°C	—
Q	Quality Conformance 1. Electrical • 25°C • −55°C • 125°C 2. Visual/Mechanical 3. Group B — Re: Mil Std 883 As Applicable 4. Group C — Re: Mil Std 883 As Applicable	Quality Conformance 1. Electrical • 25°C • 0°C •100°C 2. Visual/Mechanical — —

O = 100% Operation
Q = Quality Conformance Inspection



How Is Reliability Tested And Maintained?
No product will be put on the market unless it meets the stringent reliability requirements determined by the procurement agency or by the factory. These requirements can be in terms of FITs (Failures In Time) or percent per thousand hours, quantities that give a mathematical limit to the failure rates resulting from a given stress. They can also be in terms of time, e.g., time to 10%, 20%, or 50% failure of a given sample for a given test.

MIL-STD 883, previously mentioned, lists both the tests and the frequency of these tests performed to maintain qualification—or suitability for sale—of a given product. New products, new processes, new design, new materials, are all "qualified" along the lines originally established by MIL-STD-883.

Two major series of tests designed for periodic monitoring or for original qualification and are listed in *Table 4*, together with a brief description and the respective LTPD (Lot Tolerance Percent Defective) that gives sample size and allowed failures.

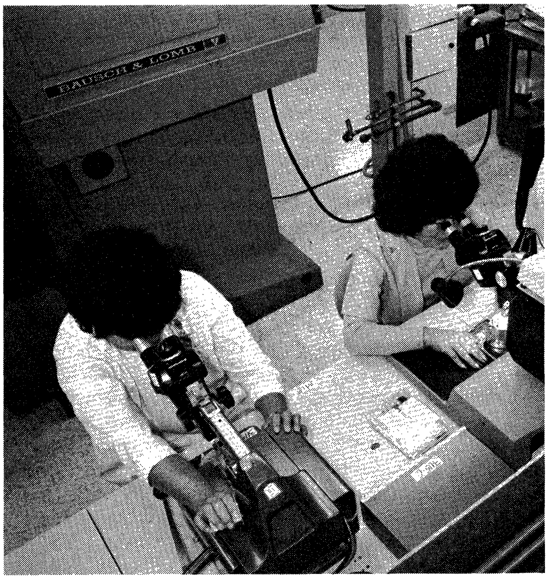
Table 4 Qualification and Monitor Testing

A. Die Related Tests — Group C

Test	Description	LTPD
Temperature Cycling	−65°C to 150°C	15
Constant Acceleration	10 kg along Y1 axis	
Seal Fine	Helium or Krypton	
Seal Gross	Fluorocarbon/bubble	
Electrical Test		
Operating Life	Static and/or dynamic	10

B. Package Related Tests — Group D

• Lead Integrity	Lead bending	15
Seal Fine	See Group C above	
Seal Gross	See Group C above	
• Thermal Shock	−55°C to 125°C 15X	15
Temperature Cycling	−65°C to 150°C 100X	
Moisture Resistance	Variable temp/humidity l0x	
Seal Fine	See Group C above	
Seal Gross	See Group C above	
Visual Inspection		
• Mechanical Shock	3000 g .3 ms	15
Constant Acceleration	1kg Y1 axis	
Seal Fine	See Group C above	
Seal Gross	See Group C above	
Visual Examination		
• Salt Atmosphere	Salt Atmosphere at 35°C	15
Seal Fine	See Group C above	
Seal Gross	See Group C above	
Visual Examination		



Conclusion

The most critical area in any electrical/electronic system is the power supply. If the power supply fails, the system goes down. Power supply failure may result in loss of critical data or damage to other system components.

To the equipment user, this means idle labor hours and unexpected replacement, repair and service costs.

To the equipment manufacturer, it can mean customer dissatisfaction and excessive warranty and rework costs.

The power supply is critical to any system and the heart of the power supply is the voltage regulator. The Fairchild Hybrid Division recognizes the importance of quality and reliability to our customer . . . and to his customer. Quality and reliability standards are established before the product is designed and are rigidly adhered to throughout the production flow.

High Current Voltage Regulators are presently shipped to a guaranteed AQL of 0.1%, with an actual return rate far less. Fairchild shares its customers' concern for quality and reliability and will continue to improve its products to insure their equipment achieves optimum performance.

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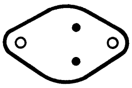
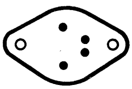
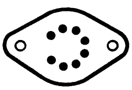
Chapter 3 Cross Reference Guide and Ordering Information

Cross Reference Guide

Voltage Regulator	Output Capability	Fairchild Device	Cross Reference Lambda	National	Silicon General
Fixed-Positive	5 V, 3 A	SH323	LAS1405	LM323	SG323
	5 V, 5 A	78H05	LAS1405, 1905	Not Available	Not Available
	5 V, 5 A	78H05A	LAS1405, 1905	Not Available	Not Available
	5 V, 8 A	78P05	LAS3905	Not Available	Not Available
	5 V, 10 A	78P05	Not Available	Not Available	Not Available
	12 V, 5 A	78H12A	LAS1412, 1912	Not Available	Not Available
Adjustable-Positive	5 To 24 V, 3 A	78HGA	LAS14U	LM350	SG350
	5 To 24 V, 5 A	78HGA	LAS19U	LM338	Not Available
Adjustable-Negative	−2 To −24 V, 5 A	79HG	LAS18U	Not Available	Not Available
Adjustable-Switching (Step Down) 	3 To 30 V, 5 A	SH1605	Not Available	LH1605	Not Available

Ordering Information
Fairchild High Current Voltage Regulators may be ordered using a simplified purchasing code.

```
XXXXX     S    C— Temperature Range Code
  |       |
  |       L———— Package Code
  |
  L——————————— Device Type (5 to 8 Digits)
```

Temperature Range Code
Operating Junction Temperature
C = Commercial
0°C to +150°C
(unless otherwise specified)
V = Industrial (SH 223 only)
−25°C to +150°C
M = Military
−55°C to +150°C

Package Code
S = Steel TO −3 Package
2−Lead
4−Lead } Refer To Chapter 4
8−Lead

Device Type (5 to 8 Digits)

SH323	3 A, 5 V Fixed Regulator
78H05A	5 A, 5 V Fixed Regulator
SH1605	5 A Switching Regulator

Examples
(a) SH 323 SC
This number code indicates a 3 amp, 5 volt fixed regulator packaged in a steel, 2−lead TO −3 with an operating junction temperature range of −25°C TO +150°C

(b) 78 HG ASM
This number code indicates a 5 amp, adjustable regulator with guaranteed maximum dropout voltage limits, packaged in a steel, 4−lead TO −3 with an operating junction temperature range of −55°C to +150°C.

Unique Level B Processing.
To meet the need for improved reliability in the military market, high current voltage regulators are available with special processing. Devices ordered to this program are subject to 100% screening as outlined in chapter 2. Devices may be ordered by simply adding the letters "QB" to the end of the ordering code.

Example
(a) 79 HG SM QB
This number code indicates a 5 amp, adjustable negative voltage regulator, packaged in a steel, 4−lead TO −3 with an operating junction temperature range of −55°C TO +150°C and screened to the Fairchild unique level B program as outlined in Chapter 2.

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μA78H05 • μA78H05A 5-Volt 5-Amp Voltage Regulators

Hybrid Products

4

Description

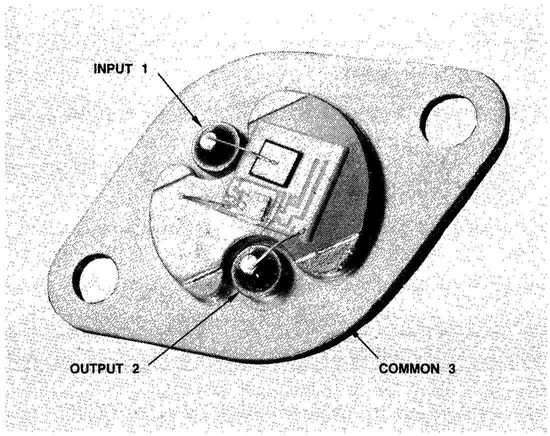
The μA78H05 and μA78H05A are hybrid regulators with 5.0 V fixed outputs and 5.0 A output capabilities. They have the inherent characteristics of the monolithic 3-terminal regulators, i.e., full thermal overload, short-circuit and safe-area protection. All devices are packaged in hermetically sealed TO-3s providing 50 W power dissipation. If the safe operating area is exceeded, the device shuts down rather than failing or damaging other system components (Note 1). This feature eliminates costly output circuitry and overly conservative heat sinks typical of high-current regulators built from discrete components.

- **5.0 A OUTPUT CURRENT**
- **INTERNAL CURRENT AND THERMAL OVERLOAD PROTECTION**
- **INTERNAL SHORT CIRCUIT PROTECTION**
- **LOW DROPOUT VOLTAGE (TYPICALLY 2.3 V @ 5.0 A)**
- **50 W POWER DISSIPATION**
- **STEEL TO-3 PACKAGE**
- **ALL PIN-FOR-PIN COMPATIBLE WITH THE SH323**

Note

1. These voltage regulators offer output transistor safe-area protection. However, to maintain full protection, the devices must be operated within the maximum input-to-output voltage differential ratings, as listed on this data sheet under "Absolute Maximum Ratings." For applications violating these limits, devices will not be fully protected.

Connection Diagram
TO-3 Metal Package

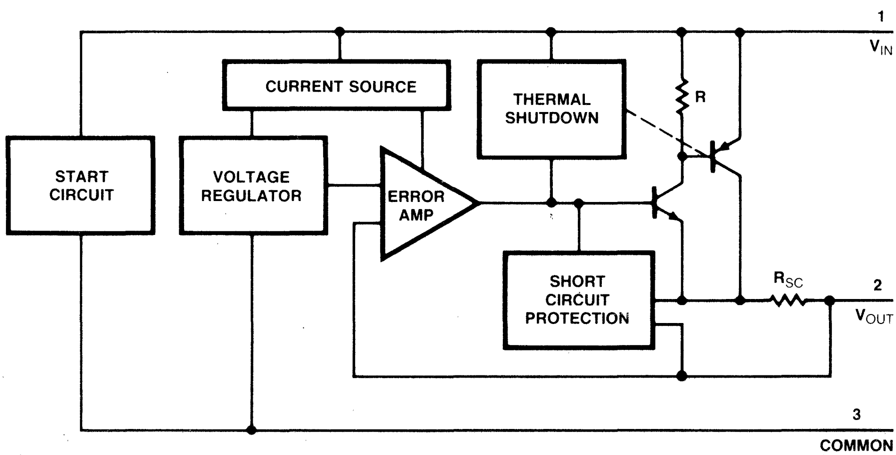


(Top View)

Order Information

Type	Package	Code	Part No.
μA7805	Metal	GN	μA78H05SC
μA7805A	Metal	GN	μA78H05ASC
μA7805	Metal	GN	μA78H05SM
μA7805A	Metal	GN	μA78H05ASM

Block Diagram



Absolute Maximum Ratings

Input Voltage	40 V
Input-to-Output Voltage Differential, Output Short Circuited	35 V
Internal Power Dissipation	50 W @ 25°C Case
Operating Junction Temperature	150°C
Military Temperature Range	
μA78H05SM	−55°C to +150°C
μA78H05ASM	−55°C to +150°C
Commercial Temperature Range	
μA78H05SC	0°C to +150°C
μA78H05ASC	0°C to +150°C
Storage Temperature Range	−55°C to +150°C
Pin Temperature (Soldering, 60 s)	300°C

μA78H05 • μA78H05A
Electrical Characteristics T_J = 25°C, V_{IN} = 10 V, I_{OUT} = 2.0 A unless otherwise specified.

Symbol	Characteristic	Condition		Limits Min	Typ	Max	Unit
V_{OUT}	Output Voltage	I_{OUT} = 2.0 A		4.85	5.0	5.25	V
ΔV_{OUT}	Line Regulation (Note 2)	V_{IN} = 8.5 to 25 V (μA78H05)			10	50	mV
		V_{IN} = 7.5 to 25 V (μA78H05A)			10	50	mV
ΔV_{OUT}	Load Regulation (Note 2)	10 mA ≤ I_{OUT} ≤ 5.0 A			10	50	mV
I_Q	Quiescent Current	I_{OUT} = 0			3.0	10	mA
RR	Ripple Rejection	I_{OUT} = 1.0 A, f = 120 Hz, 5.0 $V_{pk\text{-}pk}$		60			dB
V_n	Output Noise	10 Hz ≤ f ≤ 100 kHz			40		μV_{RMS}
V_{DD}	Dropout Voltage (Note 3)	μA78H05	I_{OUT} = 5.0 A		2.3		V
			I_{OUT} = 3.0 A		2.0		V
		μA78H05A	I_{OUT} = 5.0 A		2.3	2.5	V
			I_{OUT} = 3.0 A		2.0	2.3	V
I_{OS}	Short-Circuit Current Limit				7.0	12.0	A_{pk}

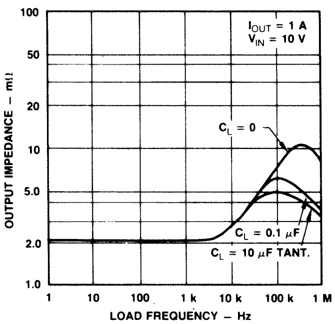
Notes

2. Load and line regulation are specified at constant junction temperature. Pulse testing is required with a pulse width ≤ 1 ms and a duty cycle of ≤ 5%. Full Kelvin connection methods must be used to measure these parameters.

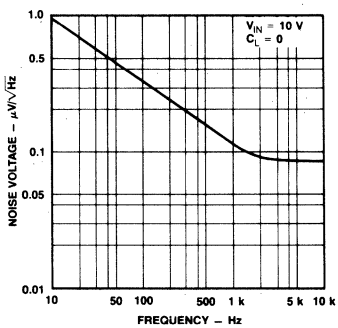
3. Dropout Voltage is the input-output voltage differential that causes the output voltage to decrease by 5% of its initial value.

Typical Performance Curves

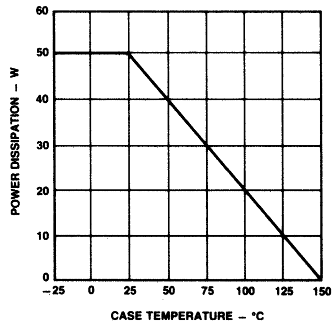
Output Impedance



Output Noise Voltage

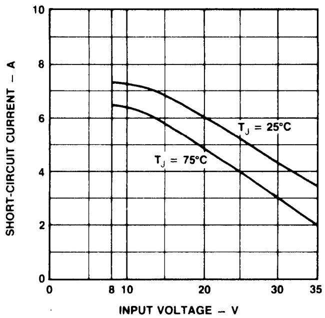


Maximum Power Dissipation

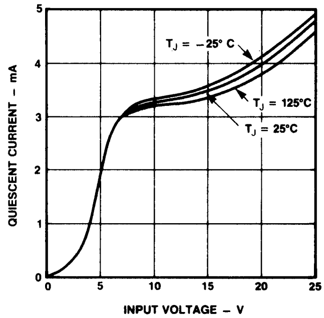


Typical Performance Curves (Cont.)

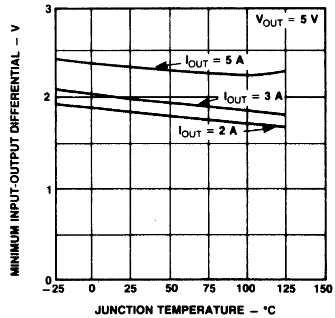
Short Circuit Current



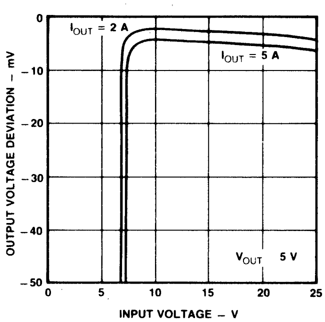
Quiescent Current



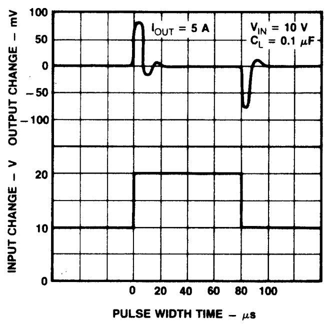
Dropout Voltage



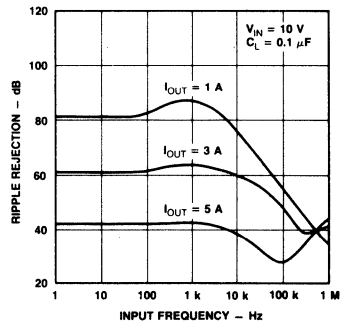
Line Regulation



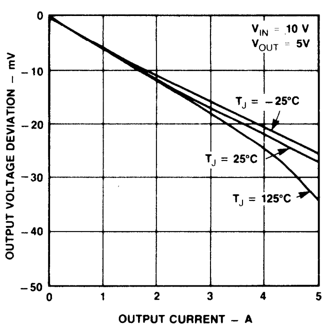
Line Transient Response



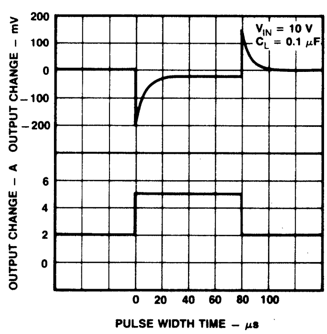
Ripple Rejection



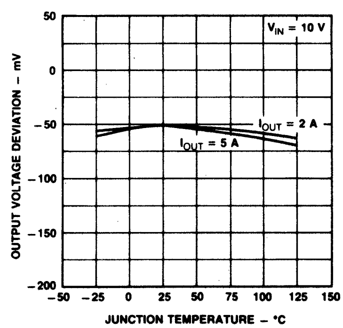
Load Regulation



Load Transient Response

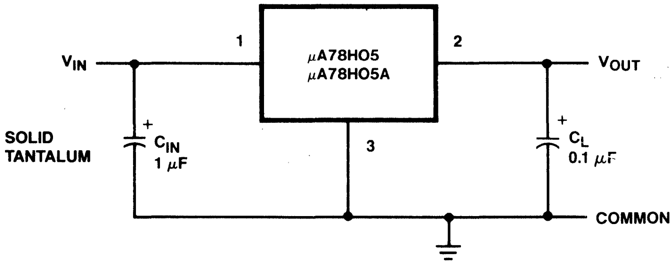


Output Voltage Deviation vs Junction Temperature



Test Circuit

Fixed Output Voltage



Design Considerations

These devices have thermal-overload protection from excessive power and internal short-circuit protection which limits the circuit's maximum current. Thus, the devices are protected from overload abnormalities. Although the internal power dissipation is limited, the junction temperature must be kept below the maximum specified temperature (150°C). It is recommended by the manufacturer that the maximum junction temperature be kept as low as possible for increased reliability. To calculate the maximum junction temperature or heat sink required, the following thermal resistance values should be used:

Package	Typ θ_{JC}	Max θ_{JC}
TO-3	1.8	2.5

$$P_{D(max)} = \frac{T_{J(max)} - T_A}{\theta_{JC} + \theta_{CA}}$$

$$\theta_{CA} = \theta_{CS} + \theta_{SA}$$

Solving for T_J:

$$T_J = T_A + P_D (\theta_{JC} + \theta_{CA})$$

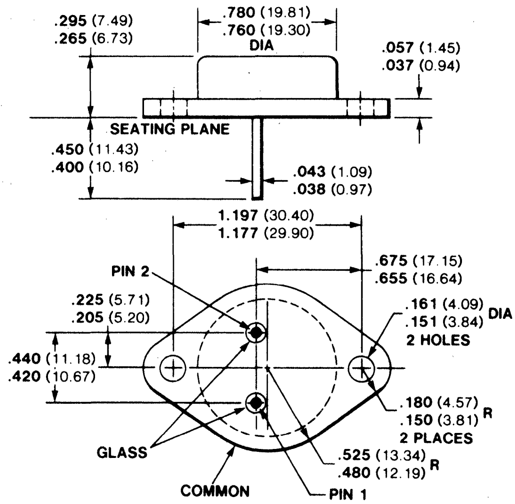
Where:

T_J = Junction Temperature
T_A = Ambient Temperature
P_D = Power Dissipation
θ_{JC} = Junction-to-case thermal resistance
θ_{CA} = Case-to-ambient thermal resistance
θ_{CS} = Case-to-heat sink thermal resistance
θ_{SA} = Heat sink-to-ambient thermal resistance

The devices are designed to operate without external compensation components. However, the amount of external filtering of these voltage regulators depends upon the circuit layout. If in a specific application the regulator is more than four inches from the filter capacitor, a 1 μF solid tantalum capacitor should be used at the input. A 0.1 μF capacitor should be used at the output to reduce transients created by fast switching loads, as seen in the basic test circuit. These filter capacitors must be located as close to the regulator as possible.

Caution: Permanent damage can result from forcing the output voltage higher than the input voltage. A protection diode from output to input should be used if this condition exists.

Package Outline
(S Package — Steel)



Notes

All dimensions in inches bold and millimeters (parentheses)
Pins are solder-dipped alloy 52

FAIRCHILD

A Schlumberger Company

μA78H12A 5-Amp Voltage Regulator

Hybrid Products

Description

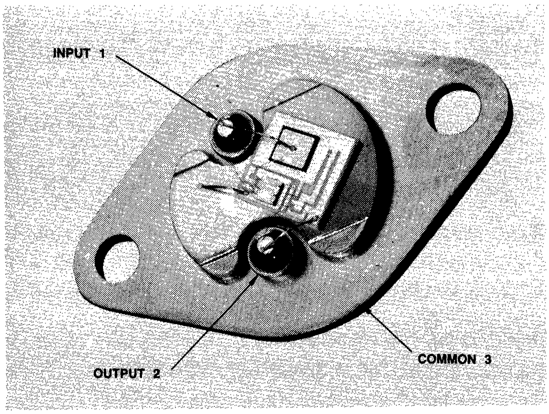
The μA78H12A is a hybrid regulator with 12.0 V fixed output and 5.0 A output capability. It has the inherent characteristics of the monolithic 3-terminal regulators; i.e., full thermal overload, short-circuit and safe-area protection. All devices are packaged in hermetically sealed TO-3s providing 50 W power dissipation. If the safe operating area is exceeded, the device shuts down, rather than failing or damaging other system components (Note 1). This feature eliminates costly output circuitry and overly conservative heat sinks typical of high-current regulators built from discrete components.

- 5.0 A OUTPUT CURRENT
- INTERNAL CURRENT AND THERMAL OVERLOAD PROTECTION
- INTERNAL SHORT CIRCUIT PROTECTION
- LOW DROPOUT VOLTAGE (TYPICALLY 2.3 V @ 5.0 A)
- 50 W POWER DISSIPATION
- STEEL TO-3 PACKAGE

Note

1. This voltage regulator offers output transistor safe-area protection. However, to maintain full protection, the device must be operated within the maximum input-to-output voltage differential ratings, as listed on this data sheet under "Absolute Maximum Ratings." For applications violating these limits, device will not be fully protected.

Connection Diagram
TO-3 Metal Package



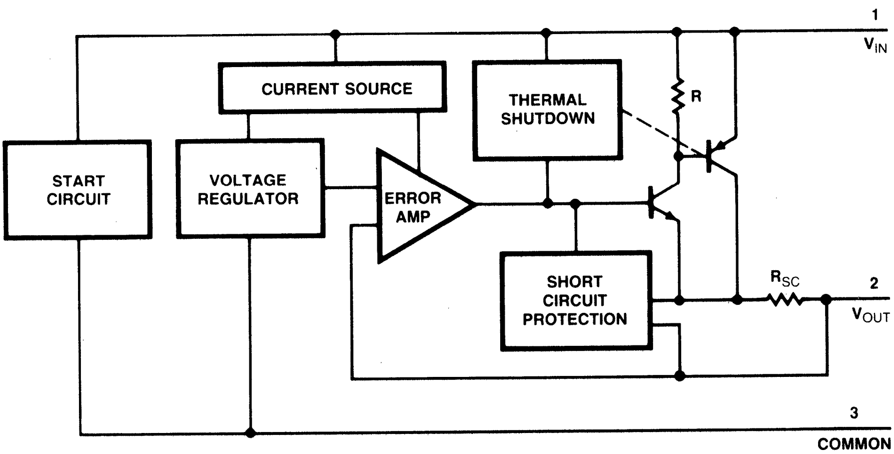
(Top View)

Order Information

Type	Package	Code	Part No.
μA78H12A	Metal	GN	μA78H12ASC
μA78H12A	Metal	GN	μA78H12ASM

4

Block Diagram



Absolute Maximum Ratings

Input Voltage	40 V
Input-to-Output Voltage Differential, Output Short-Circuited	35 V
Internal Power Dissipation	50 W @ 25°C Case
Operating Junction Temperature	150°C
Military Temperature Range μA78H12ASM	−55°C to +150°C
Commercial Temperature Range μA78H12ASC	0°C to +150°C
Storage Temperature Range	−55°C to +150°C
Pin Temperature (Soldering, 60 s)	300°C

μA7812A

Electrical Characteristics T_J = 25°C, V_{IN} = 19 V, I_{OUT} = 2.0 A unless otherwise specified

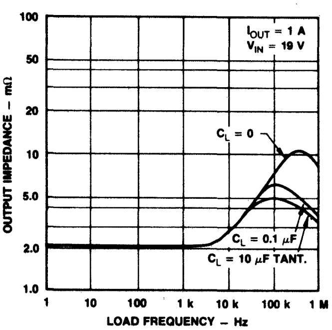
Symbol	Characteristic	Condition	Limits Min	Typ	Max	Unit
V_{OUT}	Output Voltage	I_{OUT} = 2.0 A	11.5	12	12.5	V
ΔV_{OUT}	Line Regulation (Note 2)	V_{IN} = 16 to 25 V		20	120	mV
ΔV_{OUT}	Load Regulation (Note 2)	10 mA $\leq I_{OUT} \leq$ 5.0 A		20	120	mV
I_Q	Quiescent Current	I_{OUT} = 0, V_{IN} = 17 V		3.7	10	mA
RR	Ripple Rejection	I_{OUT} = 1.0 A, f = 120 Hz, 5.0 $V_{pk\text{-}pk}$	60			dB
V_n	Output Noise	10 Hz $\leq$ f $\leq$ 100 kHz, V_{IN} = 17 V		75		V_{RMS}
V_{DD}	Dropout Voltage (Note 3)	I_{OUT} = 5.0 A		2.3	2.5	V
		I_{OUT} = 3.0 A		2.0	2.3	V
I_{OS}	Short-Circuit Current Limit			7.0	12.0	A_{pk}

Notes

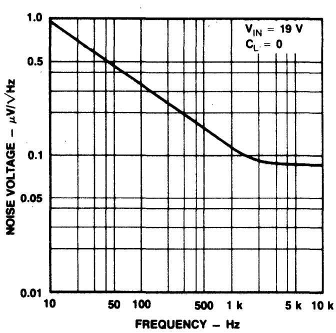
2. Load and line regulation are specified at constant junction temperature. Pulse testing is required with a pulse width $\leq$ 1 ms and a duty cycle $\leq$ 5%. Full Kelvin connection methods must be used to measure these parameters.
3. Dropout Voltage is the input-to-output voltage differential that causes the output voltage to decrease by 5% of its initial value.

Typical Performance Curves

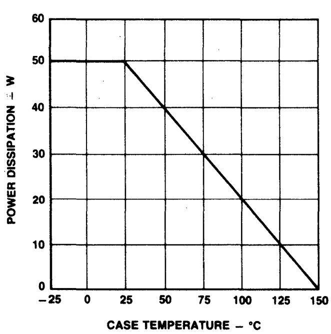
Output Impedance



Output Noise Voltage

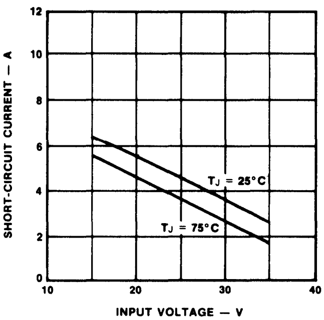


Maximum Power Dissipation

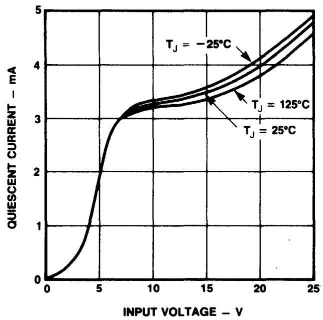


Typical Performance Curves (Cont.)

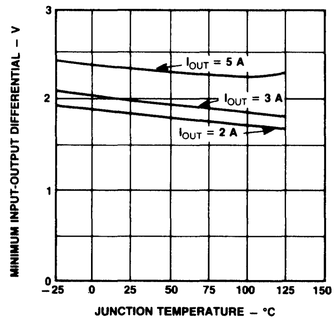
Short Circuit Current



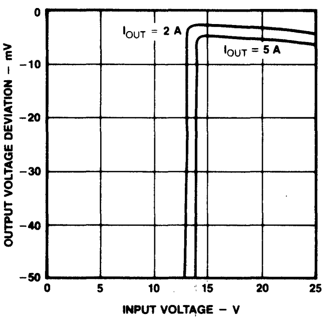
Quiescent Current



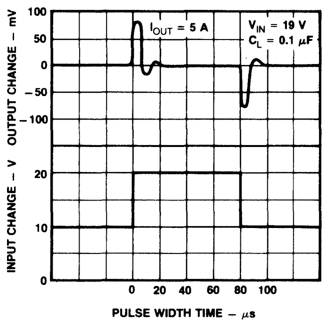
Dropout Voltage



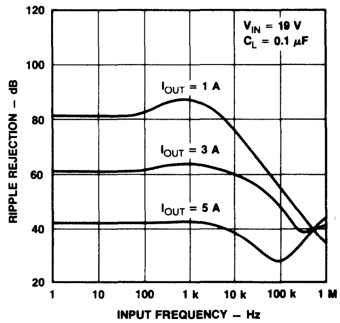
Line Regulation



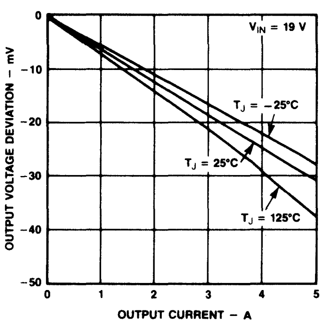
Line Transient Response



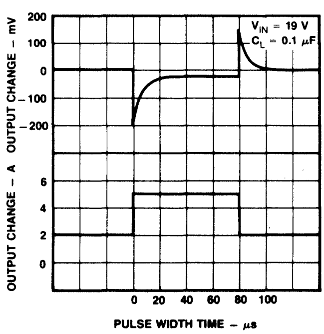
Ripple Rejection



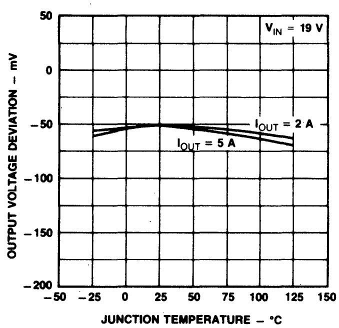
Load Regulation



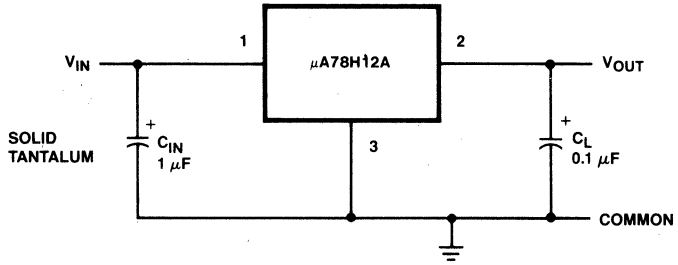
Load Transient Response



Output Voltage Deviation vs Junction Temperature



Basic Test Circuit



Design Considerations

This device has thermal-overload protection from excessive power and internal short-circuit protection which limits the circuit's maximum current. Thus, the device is protected from overload abnormalities. Although the internal power dissipation is limited, the junction temperature must be kept below the maximum specified temperature (150°C). It is recommended by the manufacturer that the maximum junction temperature be kept as low as possible for increased reliability. To calculate the maximum junction temperature or heat sink required, the following thermal resistance values should be used:

Package	Typ θ_{JC}	Max θ_{JC}
TO-3	1.8	2.5

$$P_{D(max)} = \frac{T_{J(max)} - T_A}{\theta_{JC} + \theta_{CA}}$$

$$\theta_{CA} = \theta_{CS} + \theta_{SA}$$

Solving for T_J:

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

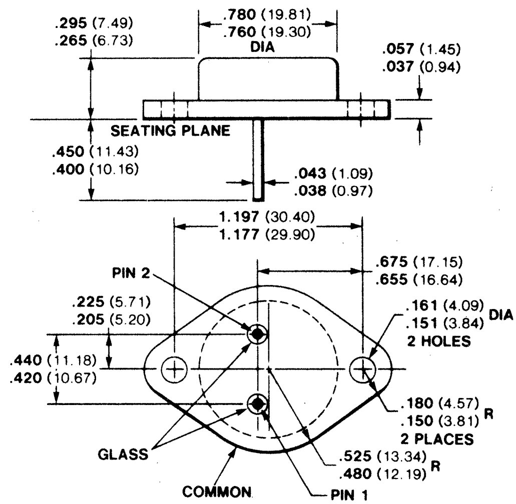
Where:

T_J = Junction Temperature
T_A = Ambient Temperature
P_D = Power Dissipation
θ_{JC} = Junction-to-case thermal resistance
θ_{CA} = Case-to-ambient thermal resistance
θ_{CS} = Case-to-heat sink thermal resistance
θ_{SA} = Heat sink-to-ambient thermal resistance

The devices are designed to operate without external compensation components. However, the amount of external filtering of these voltage regulators depends upon the circuit layout. If in a specific application the regulator is more than four inches from the filter capacitor, a 1 μF solid tantalum capacitor should be used at the input. A 0.1 μF capacitor should be used at the output to reduce transients created by fast switching loads, as seen in the basic test circuit. These filter capacitors must be located as close to the regulator as possible.

Caution: Permanent damage can result from forcing the output voltage higher than the input voltage. A protection diode from output to input should be used if this condition exists.

Package Outline
(S Package — Steel)



Notes

All dimensions in inches bold and millimeters (parentheses)
Pins are solder-dipped alloy 52

FAIRCHILD

A Schlumberger Company

μA78HGA
Positive Adjustable 5-Amp Voltage Regulator

Hybrid Products

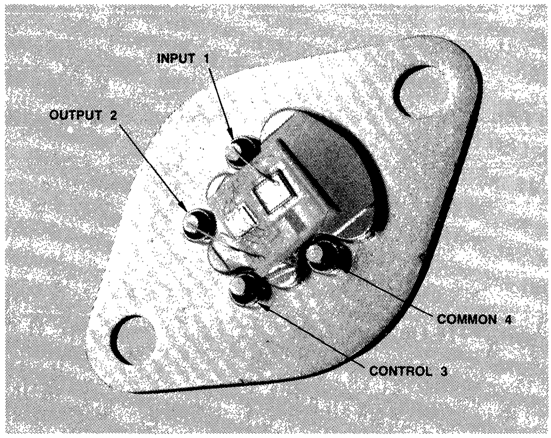
Description

The μA78HGA is an adjustable 4-terminal positive voltage regulator capable of supplying in excess of 5.0 A over a 5.0 V to 24 V output range. Only two external resistors are required to set the output voltage.

The μA78HGA is packaged in a hermetically sealed TO-3, providing 50 W power dissipation. The regulator consists of a monolithic chip driving a discrete series-pass element. A beryllium-oxide substrate is used in conjunction with an isothermal layout to optimize the thermal characteristics of each device and still maintain electrical isolation between the various chips. This unique circuit design limits the maximum junction temperature of the power output transistor to provide full automatic thermal overload protection. If the safe operating area is ever exceeded (Note 1), the device simply shuts down rather than failing or damaging other system components. This feature eliminates the need to design costly regulators built from discrete components.

- 5.0 A OUTPUT CURRENT
- INTERNAL CURRENT AND THERMAL LIMITING
- INTERNAL SHORT CIRCUIT CURRENT LIMIT
- LOW DROPOUT VOLTAGE (TYPICALLY 2.3 V @ 5.0 A)
- 50 W POWER DISSIPATION
- ELECTRICALLY NEUTRAL CASE
- STEEL TO-3 PACKAGE
- ALL PIN-FOR-PIN COMPATIBLE WITH μA78HG

Connection Diagram
TO-3 Metal Package



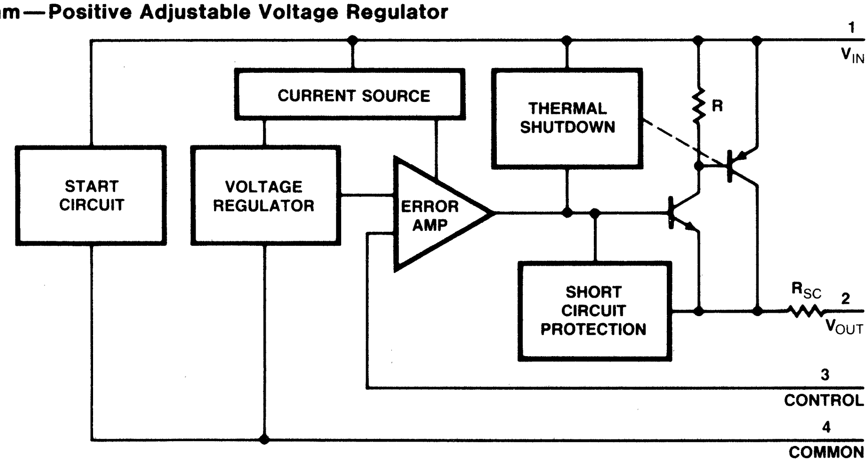
(Top View)

Order Information

Type	Package	Code	Part No.
μA78HGA	Metal	JA	μA78HGASC
μA78HGA	Metal	JA	μA78HGASM

4

Block Diagram—Positive Adjustable Voltage Regulator



Notes on following pages.

Absolute Maximum Ratings

Input Voltage	40 V
Internal Power Dissipation	50 W @ 25°C Case
Maximum Input-to-Output Voltage	
Differential Output Short Circuit	35 V
Operating Junction Temperature	150°C
Military Temperature Range μA78HGASM	−55°C to +150°C
Commercial Temperature Range μA78HGASC	0°C to +150°C
Storage Temperature Range	−55°C to +150°C
Pin Temperature (Soldering, 60 s)	300°C

Electrical Characteristics T_J = 25°C, V_{IN} = 10 V, I_{OUT} = 2.0 A unless otherwise specified

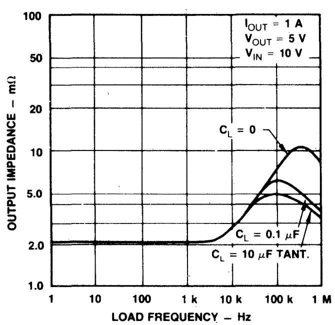
			Limits			
Symbol	**Characteristic**	**Condition (Note 3)**	**Min**	**Typ**	**Max**	**Unit**
V_{OUT}	Output Voltage (Note 4)	I_{OUT} = 2.0 A, V_{IN} = V_{OUT} + 3.5 V	5.0		24	V
ΔV_{OUT}	Line Regulation (Note 2)	V_{IN} = 7.5 to 25 V		0.2%	1%	V
ΔV_{OUT}	Load Regulation (Note 2)	10 mA ≤ I_{OUT} ≤ 5.0 A		0.2%	1%	V
I_Q	Quiescent Current	I_{OUT} = 0		3.4	10	mA
RR	Ripple Rejection	I_{OUT} = 1.0 A, f = 210 Hz, 5.0 V_{pk-pk}	60			dB
V_n	Output Noise	10 Hz ≤ f ≤ 100 kHz, V_{IN} = V_{OUT} + 5.0 V		50		μV_{RMS}
V_{DD}	Dropout Voltage (Note 5)	I_{OUT} = 5.0 A		2.3	2.5	V
		I_{OUT} = 3.0 A		2.0	2.3	V
I_{OS}	Short-Circuit Current Limit	V_{IN} = 15 V		7.0	12.0	A_{pk}
V_C	Control Pin Voltage		4.85	5.0	5.25	V

Notes

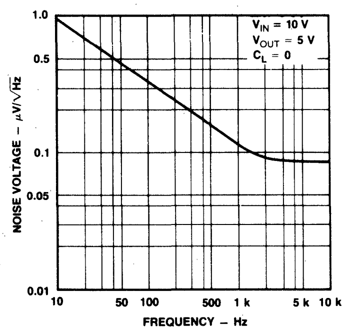
1. This voltage regulator offers output transistor safe-area protection. However, to maintain full protection, the device must be operated within the maximum input-to-output voltage differential rating listed on the data sheet under "Absolute Maximum Ratings." For applications violating these limits, device will not be fully protected.
2. Load and line regulation are specified at constant junction temperature. Pulse testing is required with a pulse width ≤ 1 ms and a duty cycle ≤ 5%. Full Kelvin connection methods must be used to measure these parameters.
3. The performance characteristics of the adjustable series (μA78HGA) is specified for V_{OUT} = 5.0 V, unless otherwise noted.
4. V_{OUT} is defined as $V_{OUT} = \frac{R1 + R2}{R2} (V_{CONT})$ where R1 and R2 are defined in the Basic Test Circuit diagram.
5. Dropout Voltage is the input-output voltage differential that causes the output voltage to decrease by 5% of its initial value.

Typical Performance Curves

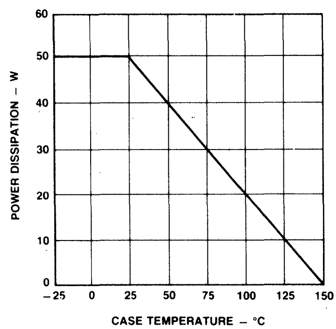
Output Impedance



Output Noise Voltage

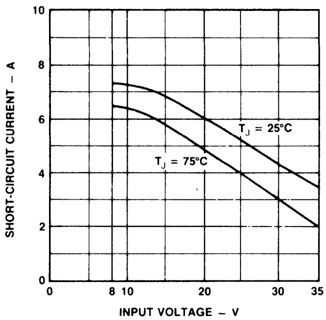


Maximum Power Dissipation

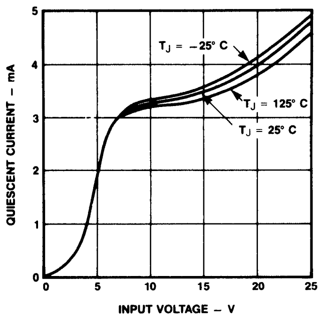


Typical Performance Curves (Cont.)

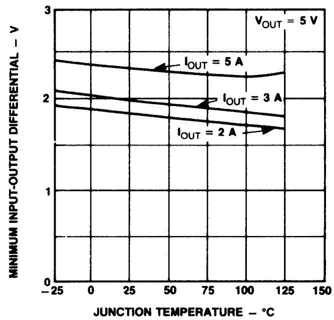
Short Circuit Current



Quiescent Current



Dropout Voltage



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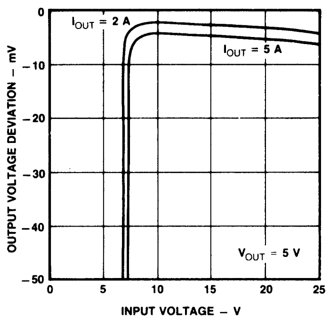
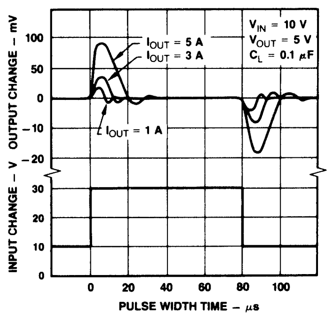
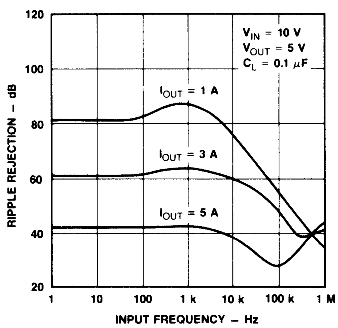
Line Regulation



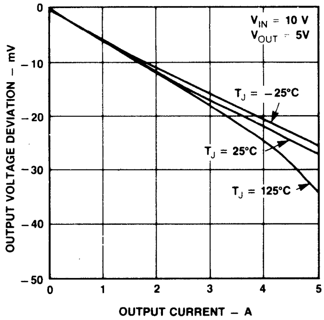
Line Transient Response



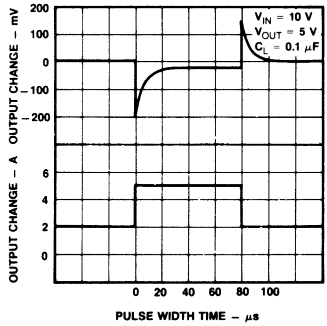
Ripple Rejection



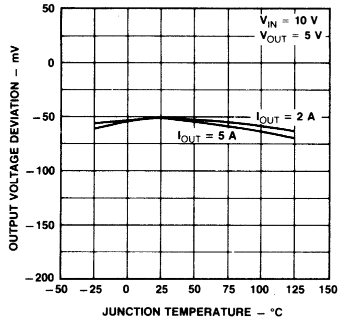
Load Regulation



Load Transient Response

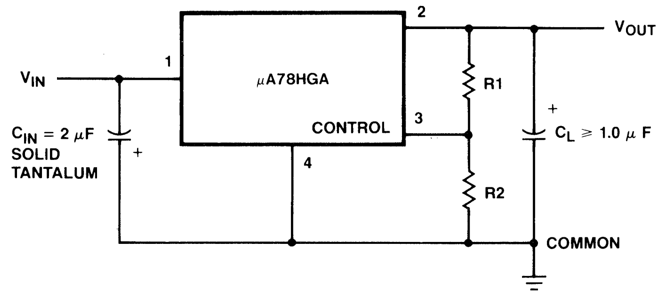


Output Voltage Deviation vs Junction Temperature



Test Circuit

Adjustable Output Voltage



Design Considerations

This device has thermal-overload protection from excessive power and internal short-circuit protection which limits the circuit's maximum current. Thus, the device is protected from overload abnormalities. Although the internal power dissipation is limited, the junction temperature must be kept below the maximum specified temperature (150°C). It is recommended by the manufacturer that the maximum junction temperature be kept as low as possible for increased reliability. To calculate the maximum junction temperature or heat sink required, the following thermal resistance values should be used:

Package	Typ θ_{JC}	Max θ_{JC}
TO-3	1.8	2.5

$$P_{D(MAX)} = \frac{T_{J(max)} - T_A}{\theta_{JC} + \theta_{CA}}$$

$$\theta_{CA} = \theta_{CS} + \theta_{SA}$$

Solving for T_J:

$$T_J = T_A + P_D\,(\theta_{JC} + \theta_{CA})$$

Where:

T_J = Junction Temperature
T_A = Ambient Temperature
P_D = Power Dissipation
θ_{JC} = Junction-to-case thermal resistance
θ_{CA} = Case-to-ambient thermal resistance
θ_{SA} = Heat sink-to-ambient thermal resistance
θ_{CS} = Case-to-heat sink thermal resistance

This device is designed to operate without external compensation components. However, the amount of external filtering of this voltage regulator depends upon the circuit layout. If in a specific application the regulator is more than four inches from the filter capacitor, a 1 μF solid tantalum capacitor should be used at the input. A 0.1 μF capacitor should be used at the output to reduce transients created by fast switching loads, as seen in the basic test circuit. These filter capacitors must be located as close to the regulator as possible.

Caution: Permanent damage can result from forcing the output voltage higher than the input voltage. A protection diode from output to input should be used if this condition exists.

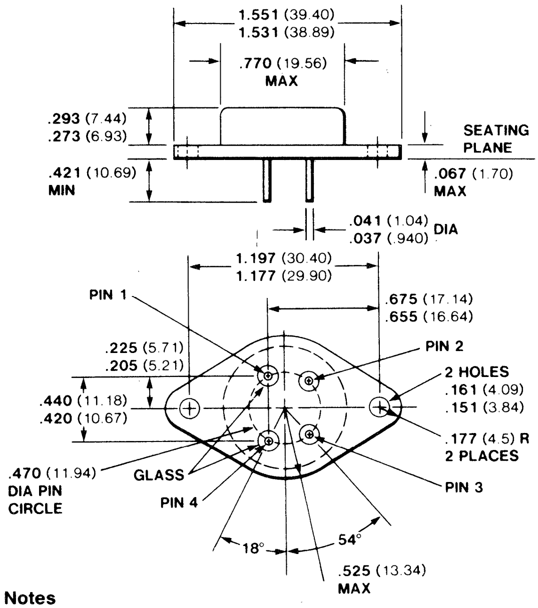
Voltage Output

The device has an adjustable output voltage from 5.0 V to 24 V which can be programmed by the external resistor network (potentiometer or two fixed resistors) using the relationship

$$V_{OUT} = V_{CONTROL}\left(\frac{R1 + R2}{R2}\right)$$

Example: If R1 = 0 Ω and R2 = 5 kΩ, then
V_{OUT} = 5 V nominal.
Or, if R1 = 10 kΩ and R2 = 5 kΩ, then
V_{OUT} = 15 V.

Package Outline
(S Package — Steel)



Notes

All dimensions in inches bold and millimeters (parentheses)

FAIRCHILD

A Schlumberger Company

μA78P05 5-Volt 10-Amp Voltage Regulator

Hybrid Products

Description

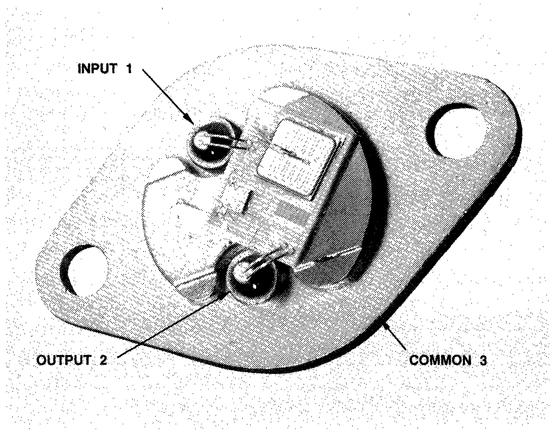
The μA78P05 3-terminal positive 5 V regulator, consisting of a monolithic control chip driving a series-pass transistor, is capable of delivering 10 A. This hybrid device is virtually blow-out proof and contains all the protection features inherent in monolithic regulators such as internal short-circuit current limiting, thermal overload and safe-area protection. If the safe-operating area is exceeded, the device shuts down rather than failing or damaging other system components (Note 1). This feature eliminates costly output circuitry and overly conservative heat sinks typical of high-current regulators built with discrete components. The μA78P05 is packaged in a hermetically sealed TO-3 providing 70 W power dissipation.

- **10 A OUTPUT CURRENT**
- **INTERNAL THERMAL OVERLOAD PROTECTION**
- **INTERNAL SHORT CIRCUIT CURRENT LIMIT**
- **LOW DROPOUT VOLTAGE (TYPICALLY 2.3 V @ 10 A)**
- **70 W POWER DISSIPATION**
- **PIN-FOR-PIN COMPATIBLE WITH THE μA78H05, μA78H05A AND SH323**
- **STEEL TO-3 PACKAGE**

Note

1. This voltage regulator offers output transistor safe-area protection. However, to maintain full protection, the device must be operated within the maximum input-to-output voltage differential ratings as listed on this data sheet under "Absolute Maximum Ratings." For applications violating these limits, device will not be fully protected.

Connection Diagram
TO-3 Metal Package

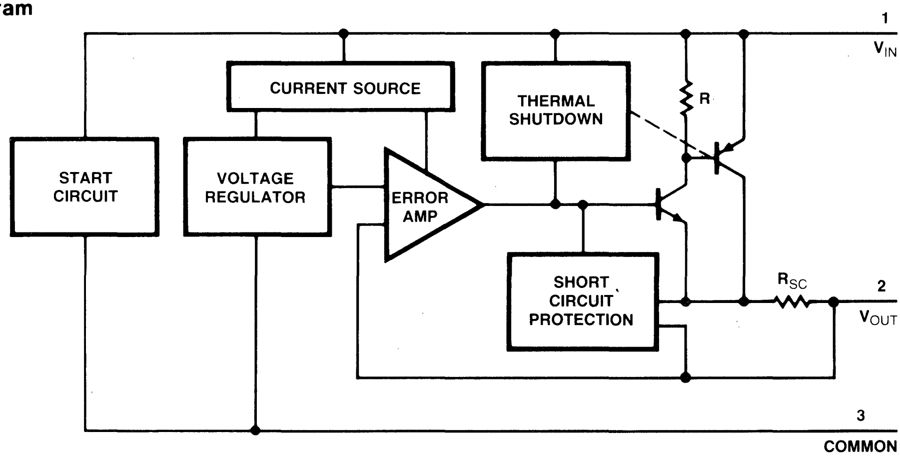


(Top View)

Order Information

Type	Package	Code	Part No.
μA78P05	Metal	6N	μA78P05SC
μA78P05	Metal	6N	μA78P05SM

Block Diagram



4

Absolute Maximum Ratings

Input Voltage	40 V
Input-to-Output Voltage Differential, Output Short-Circuited	35 V
Internal Power Dissipation	70 W @ 25°C Case
Operating Junction Temperature	150°C
Military Temperature Range μA78P05SM	−55°C to +150°C
Commercial Temperature Range μA78P05SC	0°C to +150°C
Storage Temperature Range	−55°C to +150°C
Pin Temperature (Soldering, 60 s)	300°C

μA78P05

Electrical Characteristics T_J = 25°C, V_{IN} = 10 V, I_{OUT} = 2.0 A unless otherwise specified

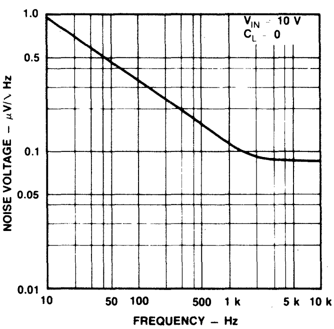
Symbol	Characteristic	Condition	Limits Min	Limits Typ	Limits Max	Unit
V_{OUT}	Output Voltage	I_{OUT} = 2.0 A	4.85	5.0	5.25	V
ΔV_{OUT}	Line Regulation (Note 2)	V_{IN} = 8 to 25 V		10	50	mV
ΔV_{OUT}	Load Regulation (Note 2)	10 mA $\leq I_{OUT} \leq$ 5 A		25	40	mV
ΔV_{OUT}	Load Regulation (Note 2)	10 mA $\leq I_{OUT} \leq$ 10 A		50	75	mV
I_Q	Quiescent Current	I_{OUT} = 0		3.4	10	mA
RR	Ripple Rejection	I_{OUT} = 1.0 A, f = 120 Hz, 5.0 $V_{pk\text{-}pk}$	60			dB
V_n	Output Noise	10 Hz $\leq$ f $\leq$ 100 kHz		40		μV_{RMS}
V_{DD}	Dropout Voltage (Note 3)	I_{OUT} = 5.0 A		2.0	2.3	V
		I_{OUT} = 10 A		2.5	3.0	V
I_{OS}	Short-Circuit Current Limit			14		A_{pk}

Notes

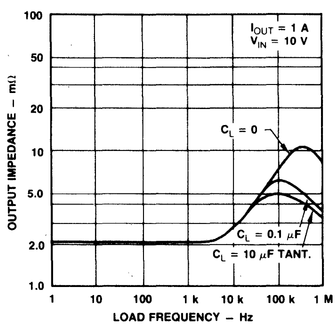
2. Load and line regulation are specified at constant junction temperature. Pulse testing is required with a pulse width ≤ 1 ms and a duty cycle ≤ 5%. Full Kelvin connection methods must be used to measure these parameters.
3. Dropout Voltage is the input-output voltage differential that causes the output voltage to decrease by 5% of its initial value.

Typical Performance Curves

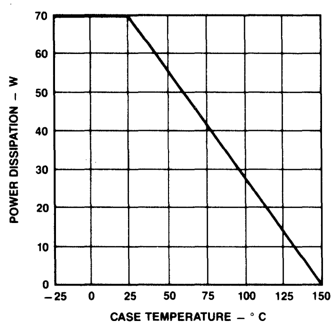
Output Noise Voltage



Output Impedance

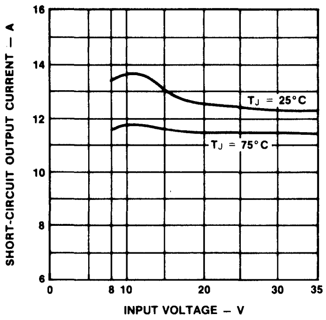


Maximum Power Dissipation

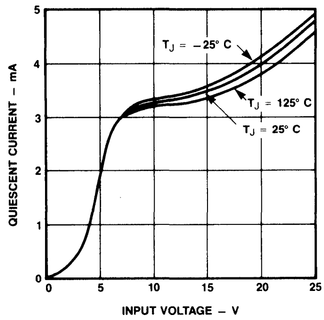


Typical Performance Curves (Cont.)

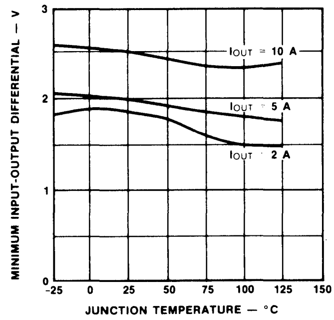
Short Circuit Current



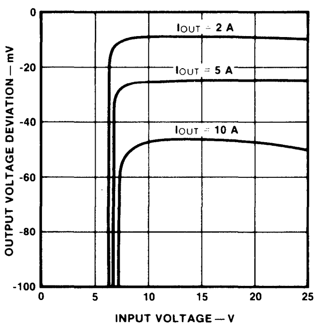
Quiescent Current



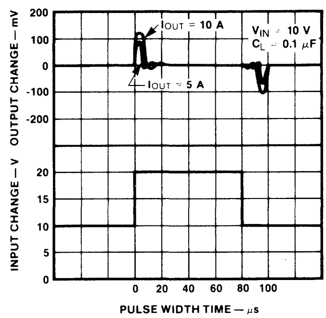
Dropout Voltage



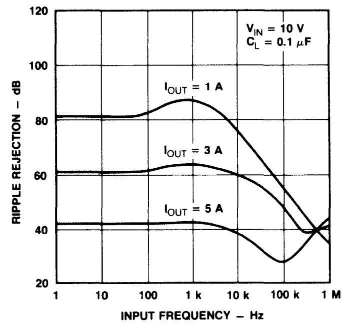
Line Regulation



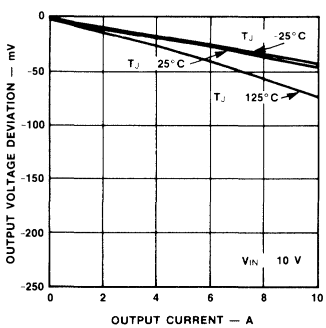
Line Transient Response



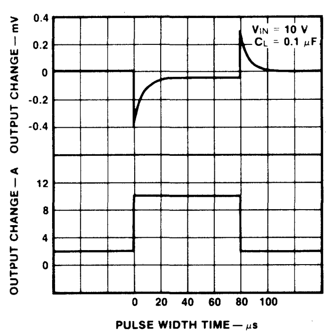
Ripple Rejection



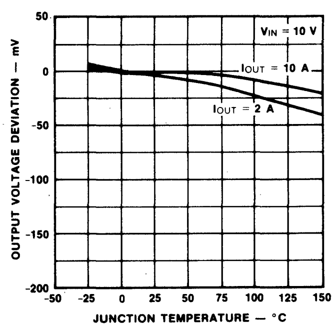
Load Regulation



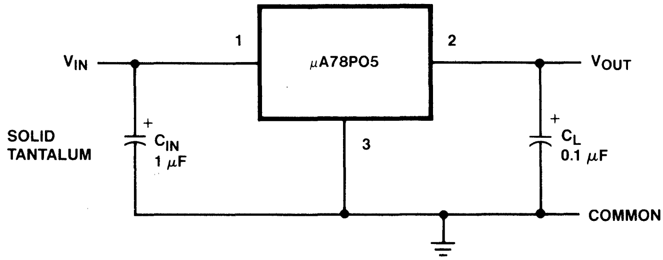
Load Transient Response



Output Voltage Deviation vs Junction Temperature



Basic Test Circuit



Design Considerations

This device has thermal-overload protection from excessive power and internal short-circuit protection which limits the circuit's maximum current. Thus, the devices are protected from overload abnormalities. Although the internal power dissipation is limited, the junction temperature must be kept below the maximum specified temperature (150°C). It is recommended by the manufacturer that the maximum junction temperature be kept as low as possible for increased reliability. To calculate the maximum junction temperature or heat sink required, the following thermal resistance values should be used:

Package	Typ θ_{JC}	Max θ_{JC}
TO-3	1.5	1.8

$$P_{D(max)} = \frac{T_{J(max)} - T_A}{\theta_{JC} + \theta_{CA}}$$

$$\theta_{CA} = \theta_{CS} + \theta_{SA}$$

Solving for T_J:

$$T_J = T_A + P_D\,(\theta_{JC} + \theta_{CA})$$

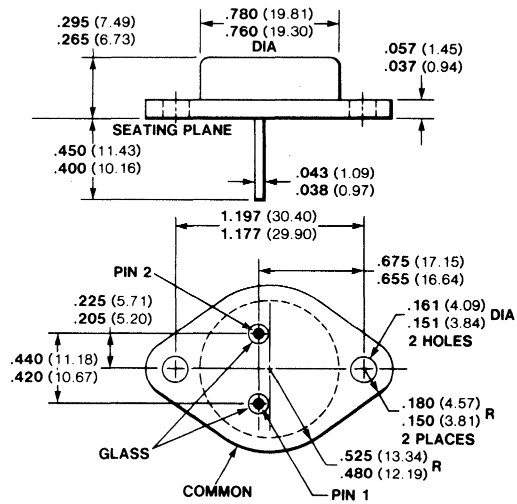
Where:

T_J = Junction Temperature
T_A = Ambient Temperature
P_D = Power Dissipation
θ_{JC} = Junction-to-case thermal resistance
θ_{CA} = Case-to-ambient thermal resistance
θ_{CS} = Case-to-heat sink thermal resistance
θ_{SA} = Heat sink-to-ambient thermal resistance

The μA78P05 is designed to operate without external compensation components. However, the amount of external filtering of this voltage regulator depends upon the circuit layout. If in a specific application the regulator is more than four inches from the filter capacitor, a 1 μF solid tantalum capacitor should be used at the input. A 0.1 μF capacitor should be used at the output to reduce transients created by fast switching loads, as seen in the basic test circuit. These filter capacitors must be located as close to the regulator as possible.

Caution: Permanent damage can result from forcing the output voltage higher than the input voltage. A protection diode from output to input should be used if this condition exists.

Package Outline
(S Package — Steel)



Notes
All dimensions in inches bold and millimeters (parentheses)
Pins are solder-dipped alloy 52

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μA79HG
5 A Negative Adjustable Voltage Regulator

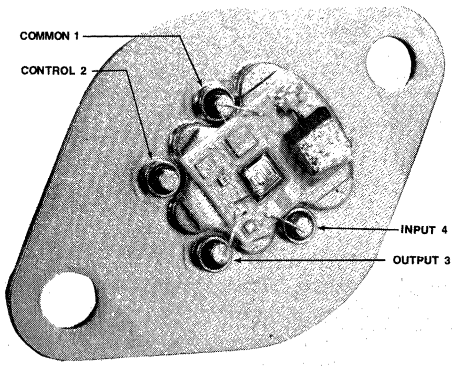
Hybrid Products

Description

The μA79HG is an adjustable 4-terminal negative voltage regulator capable of supplying in excess of −5 A over a −24 V to −2.11 V output range. The μA79HG hybrid voltage regulator has been designed with all the inherent characteristics of the monolithic 4-terminal regulator; i.e., full thermal overload and short circuit protection. The μA79HG is packaged in a hermetically-sealed 4-pin TO-3 package providing 50 W power dissipation. The regulator consists of a monolithic chip driving a discrete-series pass element and short circuit detection transistors.

- −5.0 A OUTPUT CURRENT
- INTERNAL CURRENT AND THERMAL OVERLOAD PROTECTION
- INTERNAL SHORT CIRCUIT CURRENT LIMIT
- LOW DROP-OUT VOLTAGE (TYPICALLY 2.2 V @ 5.0 A)
- 50 W POWER DISSIPATION
- ELECTRICALLY NEUTRAL CASE
- STEEL TO-3 CASE

Connection Diagram
4-Pin Metal Package

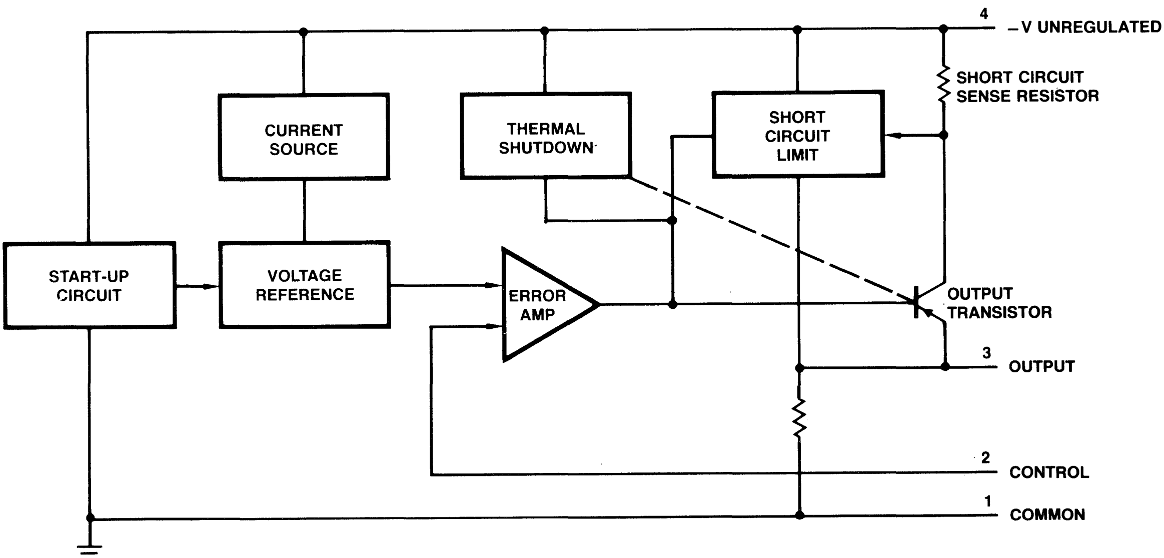


(Top View)

Order Information

Type	Package	Code	Part No.
μA79HG	Metal	JA	μA79HGSC
μA79HG	Metal	JA	μA79HGSM

Block Diagram



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Absolute Maximum Ratings

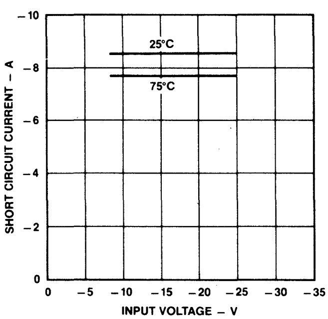
Input Voltage	−40 V
Internal Power Dissipation	50 W @ 25°C Case
Maximum Input-to-Output Voltage Differential	−35 V
Operating Junction Temperature Range	0°C to +150°C
Storage Temperature Range	−55°C to +150°C
Pin Temperature (Soldering, 60 s)	300°C

μA79HG
Electrical Characteristics $T_J = 25°C$, $V_{IN} = -10$ V and $I_{OUT} = -2.0$ A unless otherwise specified.

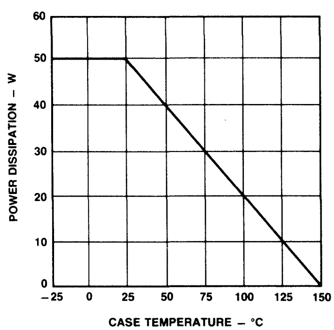
Characteristic	Limits Min	Typ	Max	Unit	Condition
Input Voltage Range	−40		−7.0	V	
Nominal Output Voltage Range	−24		−2.11	V	$V_{IN} = V_{OUT} -5$ V
Output Voltage Tolerance			4	%(V_{OUT})	$-40\text{ V} \leq V_{IN} \leq -7\text{ V}$
Line Regulation		0.4	1.0	%(V_{OUT})	$-40\text{ V} \leq V_{IN} \leq -7\text{ V}$
Load Regulation		0.7	1.0	%(V_{OUT})	$V_{IN} = V_{OUT} -10$ V, $-10\text{ mA} \leq I_{OUT} \leq -5.0\text{ A}$
Control Pin Current			3.0	μA	
Quiescent Current			−5.0	mA	$V_{IN} = -10$ V
Ripple Rejection		50		dB	$-18\text{ V} \leq V_{IN} \leq -8.5\text{ V}$, $V_{OUT} = -5$ V, f = 120 Hz
Output Noise Voltage		200		μV	$10\text{ Hz} \leq f \leq 100\text{ kHz}$, $V_{OUT} = -5.0$ V
Dropout Voltage		2.2		V	$I_{OUT} = -5$ A
Short Circuit Current Limit		−8	−12	A	$V_{IN} = -15$ V
Control Pin Voltage (Reference)	−2.35		−2.11	V	$V_{IN} = -10$ V

Typical Performance Curves

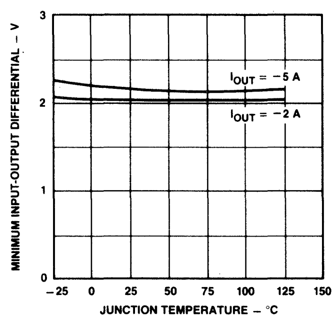
Short Circuit Current



Quiescent Current

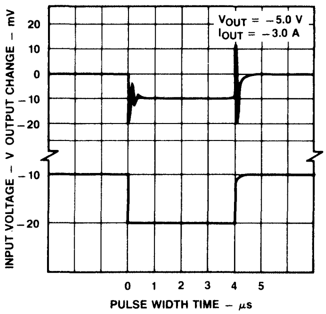


Dropout Voltage

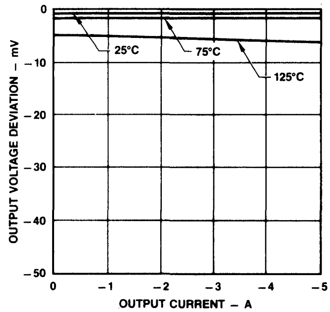


Typical Performance Curves (Cont.)

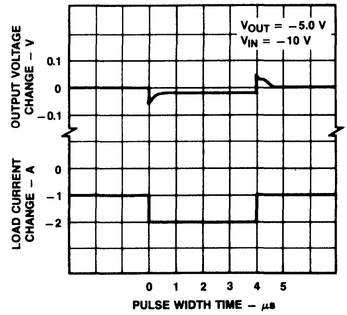
Line Transient Response



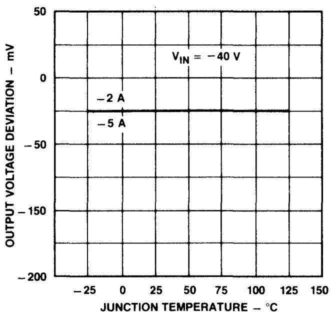
Load Regulation



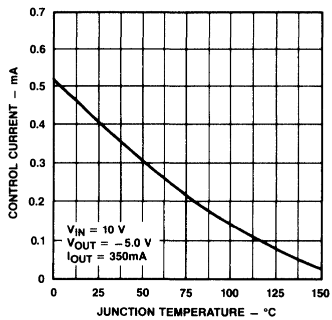
Load Transient Response



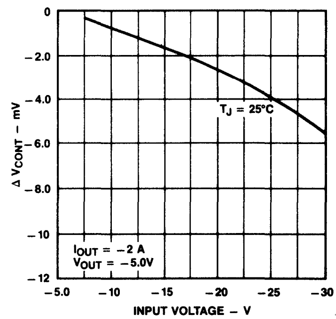
Output Voltage Deviation vs Junction Temperature



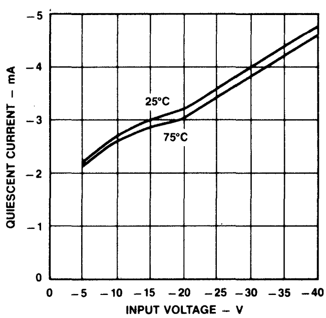
Control Current vs Temperature



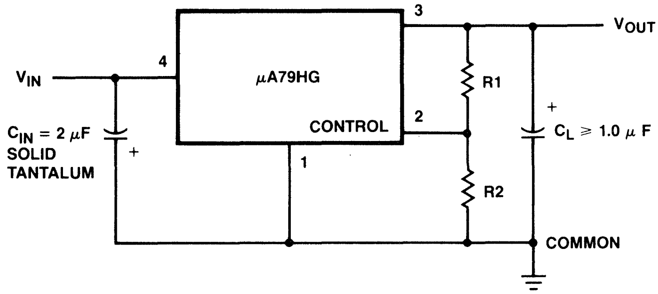
Differential Control Voltage vs Input Voltage



Maximum Power Dissipation



Basic Test Circuit, Adjustable Output Voltage



$$V_{OUT} = V_{CONT}\left(\frac{R1 + R2}{R2}\right)$$

Design Considerations

This device has thermal overload protection from excessive power and internal short circuit protection which limits the circuit's maximum current. Thus, the device is protected from overload abnormalities. Although the internal power dissipation is limited, the junction temperature must be kept below the maximum specified temperature (150°C). It is recommended by the manufacturer that the maximum junction temperature be kept as low as possible for increased reliability. To calculate the maximum junction temperature or heat sink required, the following thermal resistance values should be used.

Package	Typ	Max
	θ_{JC}	θ_{JC}
TO-3	1.8	2.5

$$P_{D(MAX)} = \frac{T_{J(MAX)} - T_A}{\theta_{JC} + \theta_{CA}}$$

$$\theta_{CA} = \theta_{CS} + \theta_{SA}$$

Solving for T_J:

$$T_J = T_A + P_D (\theta_{JC} + \theta_{CA})$$

Where:

T_J = Junction Temperature
T_A = Ambient Temperature
P_D = Power Dissipation
θ_{JC} = Junction-to-case thermal resistance
θ_{CA} = Case-to-ambient thermal resistance
θ_{CS} = Case-to-heat sink thermal resistance
θ_{SA} = Heat sink-to-ambient thermal resistance

The device is designed to operate without external compensation components. However, the amount of external filtering of these voltage regulators depends upon the circuit layout. If in a specific application the regulator is more than four inches from the filter capacitor, a 2 μF solid tantalum capacitor should be used at the input. A 1 μF capacitor should be used at the output to reduce transients created by fast switching loads, as seen in the basic test circuit. These filter capacitors must be located as close to the regulator as possible.

Caution: Permanent damage can result from forcing the output voltage higher than the input voltage. A protection diode from output to input should be used if this condition exists.

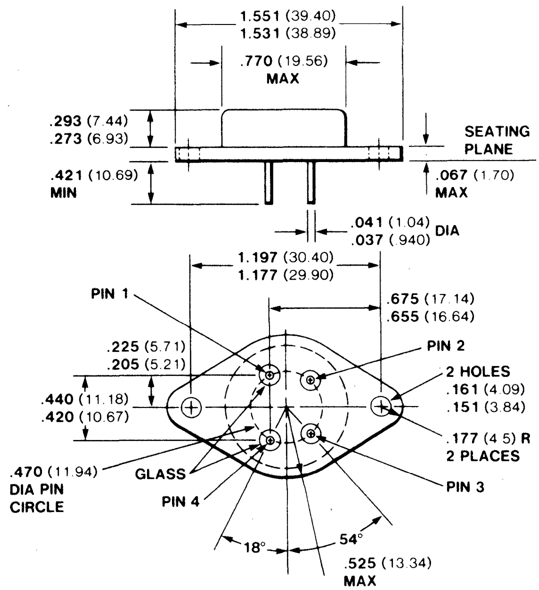
Voltage Output

The device has an adjustable output voltage from −2.11 to −24 V which can be programmed by the external resistor network (potentiometer or two fixed resistors) using the relationship:

$$V_{OUT} = V_{CONTROL}\left(\frac{R1 + R2}{R2}\right)$$

Example: If R1 = 0 Ω and R2 = 5 kΩ, then V_{OUT} = −2.11 V nominal.
Or, if R1 = 12.8 kΩ and R2 = 2.1 kΩ then V_{OUT} = −15 V.

Package Outline
(S Package — Steel)



Notes

All dimensions in inches bold and millimeters (parentheses)
Pins are solder-dipped alloy 52

FAIRCHILD

A Schlumberger Company

SH323 • SH223 • SH123
3 A, 5 V
Voltage Regulator

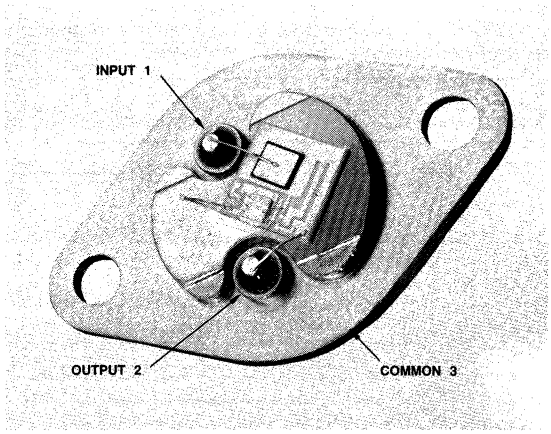
Hybrid Products

Description

The SH323 is a hybrid regulator with 5.0 V fixed output and 3.0 A output capability. It has the inherent characteristics of the monolithic 3-terminal regulators, i.e., full thermal overload, short circuit and safe area protection. All devices are packaged in hermetically sealed TO-3s providing 50 W power dissipation. If the safe operating area is exceeded, the device shuts down rather than failing or damaging other system components (Note 1). This feature eliminates costly output circuitry and overly conservative heat sinks typical of high-current regulators built from discrete components.

- 3.0 A OUTPUT CURRENT
- INTERNAL CURRENT AND THERMAL OVERLOAD PROTECTION
- INTERNAL SHORT CIRCUIT PROTECTION
- LOW DROPOUT VOLTAGE (TYPICALLY 2.0 V @ 3.0 A)
- 50 W POWER DISSIPATION
- STEEL TO-3 PACKAGE
- ALL PIN-FOR-PIN COMPATIBLE WITH THE LM323, SG323

Connection Diagram
2-Pin Metal Package



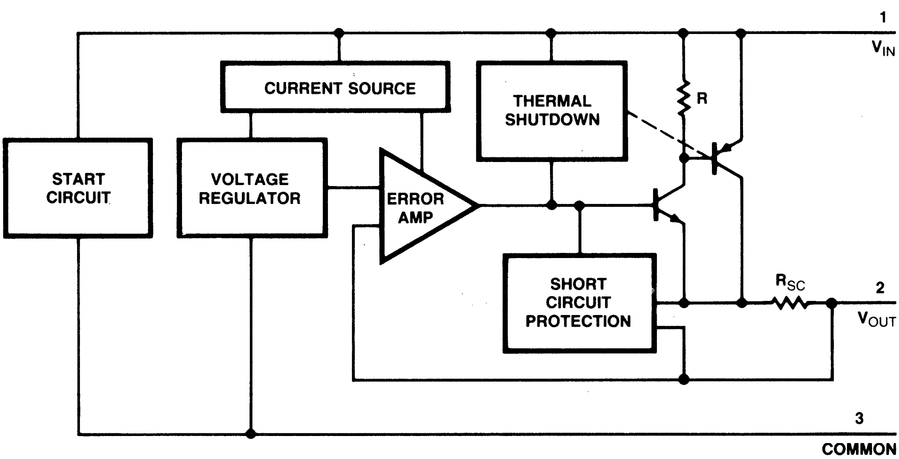
(Top View)

Order Information

Type	Package	Code	Part No.
SH323	Metal	GN	SH323SC
SH223	Metal	GN	SH223SV
SH123	Metal	GN	SH123SM

4

Block Diagram



Absolute Maximum Ratings

Input Voltage	40 V
Input-to-Output Voltage Differential	
Output Short Circuited	35 V
Internal Power Dissipation	50 W @ 25°C Case
Operating Junction Temperature	150°C
Industrial Temperature Range SH223SV	−25°C to +150°C
Military Temperature Range SH123SM	−55°C to +150°C
Commercial Temperature Range SH323SC	0°C to +150°C
Storage Temperature Range	−55°C to +150°C
Pin Temperature (Soldering, 60 s)	300°C

Electrical Characteristics $T_J = 25°C$, $V_{IN} = 10$ V, $I_{OUT} = 2.0$ A unless otherwise specified.

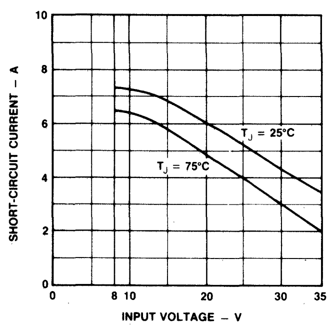
Symbol	Characteristic	Limits Min	Typ	Max	Unit	Condition
V_{OUT}	Output Voltage	4.85	5.0	5.25	V	$I_{OUT} = 2.0$ A
ΔV_{OUT}	Line Regulation (Note 2)		10	25	mV	$V_{IN} = 7.5$ to 25 V
ΔV_{OUT}	Load Regulation (Note 2)		10	50	mV	10 mA $\leq I_{OUT} \leq$ 3.0 A
I_Q	Quiescent Current		3.0	10	mA	$I_{OUT} = 0$
RR	Ripple Rejection	60			dB	$I_{OUT} = 1.0$ A, f = 120 Hz, 5.0 $V_{pk\text{-}pk}$
V_n	Output Noise		40		μV_{RMS}	10 Hz $\leq$ f $\leq$ 100 kHz, $V_{IN} = 10$ V
V_{DD}	Dropout Voltage (Note 3)		2.0	2.3	V	$I_{OUT} = 3$ A
I_{OS}	Short Circuit Current Limit		7.0	12.0	A_{pk}	$V_{IN} = 10$ V

Notes

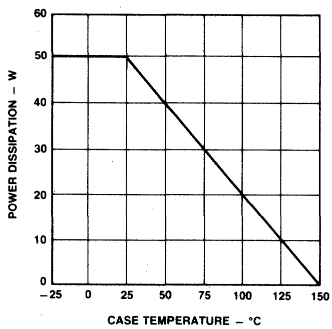
1. This voltage regulator offers output transistor safe area protection. However, to maintain full protection, the device must be operated within the maximum input-to-output voltage differential ratings, as listed on this data sheet under "Absolute Maximum Ratings." For applications violating these limits, device will not be fully protected.
2. Load and line regulation are specified at constant junction temperature. Pulse testing is required with a pulse width ≤ 1 ms and a duty cycle ≤ 5%. Full Kelvin connection methods must be used to measure these parameters.
3. Dropout Voltage is the input-output voltage differential that causes the output voltage to decrease by 5% of its initial value.

Typical Performance Curves

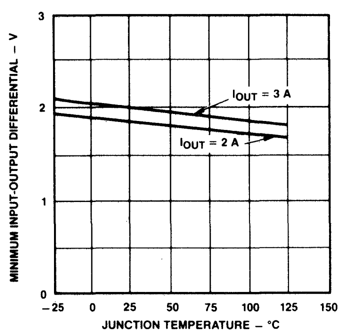
Short Circuit Current



Maximum Power Dissipation

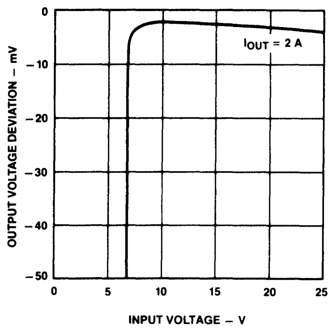


Dropout Voltage

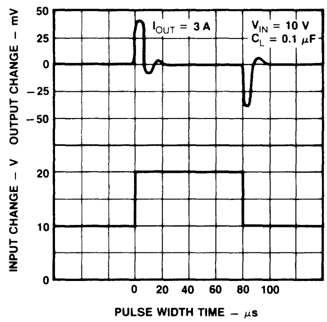


Typical Performance Curves (Cont.)

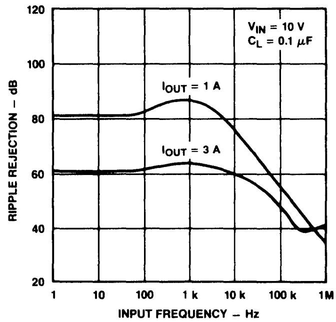
Line Regulation



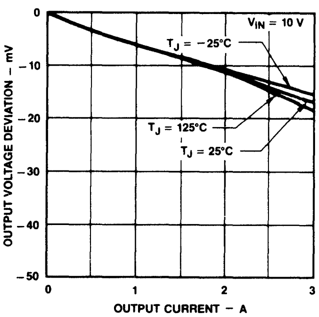
Line Transient Response



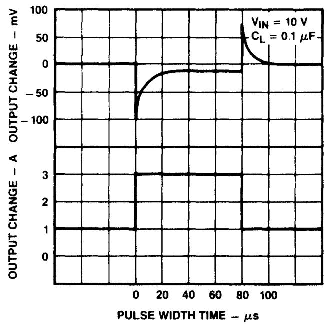
Ripple Rejection



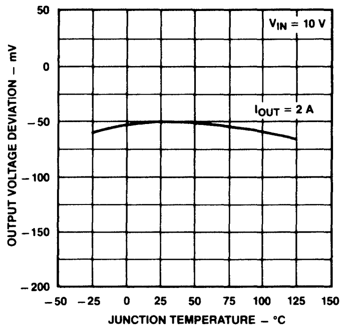
Load Regulation



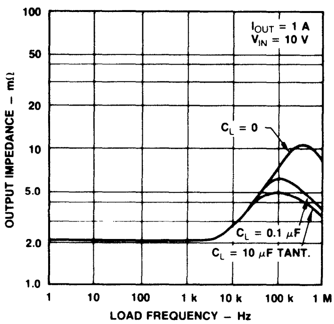
Load Transient Response



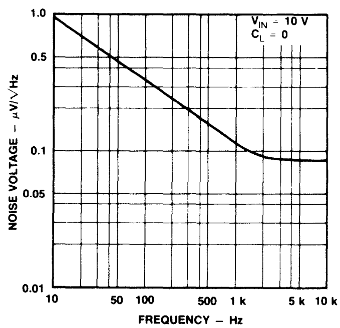
V_{OUT} vs Junction Temperature



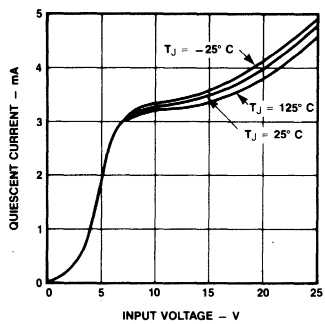
Output Impedance



Output Noise Voltage

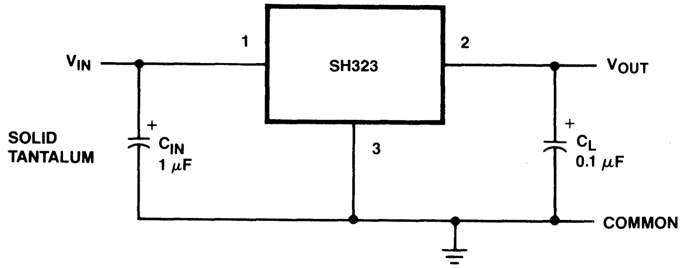


Quiescent Current



Test Circuit

Fixed Output Voltage



Design Considerations

This device has thermal overload protection from excessive power and internal short circuit protection which limits the circuit's maximum current. Thus, the device is protected from overload abnormalities. Although the internal power dissipation is limited, the junction temperature must be kept below the maximum specified temperature (150°C). It is recommended by the manufacturer that the maximum junction temperature be kept as low as possible for increased reliability. To calculate the maximum junction temperature or heat sink required, the following thermal resistance values should be used.

Package	Typ θ_{JC}	Max θ_{JC}
TO-3	1.8	2.5

$$P_{D(MAX)} = \frac{T_{J(MAX)} - T_A}{\theta_{JC} + \theta_{CA}}$$

$$\theta_{CA} = \theta_{CS} + \theta_{SA}$$

Solving for T_J:

$$T_J = T_A + P_D\,(\theta_{JC} + \theta_{CA})$$

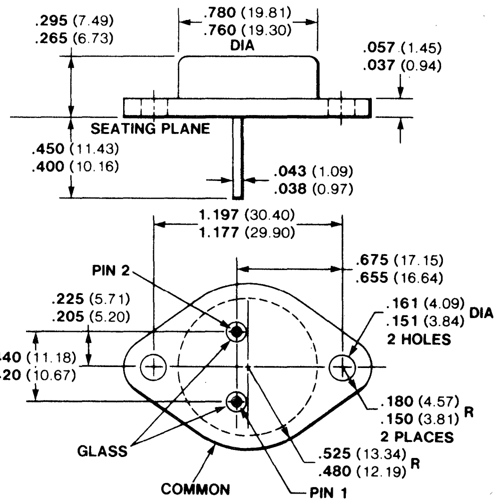
Where:

T_J = Junction Temperature
T_A = Ambient Temperature
P_D = Power Dissipation
θ_{JC} = Junction-to-case thermal resistance
θ_{CA} = Case-to-ambient thermal resistance
θ_{CS} = Case-to-heat sink thermal resistance
θ_{SA} = Heat sink-to-ambient thermal resistance

The device is designed to operated without external compensation components. However, the amount of external filtering of this voltage regulator depends upon the circuit layout. If in a specific application the regulator is more than four inches from the filter capacitor, a 1 μF solid tantalum capacitor should be used at the input. A 0.1 μF capacitor should be used at the output to reduce transients created by fast switching loads, as seen in the basic test circuit. These filter capacitors must be located as close to the regulator as possible.

Caution: Permanent damage can result from forcing the output voltage higher than the input voltage. A protection diode from output to input should be used if this condition exists.

Package Outline
(S Package — Steel)



Notes

All dimensions in inches bold and millimeters (parentheses)
Pins are solder-dipped alloy 52

FAIRCHILD

A Schlumberger Company

SH1605
5-Amp, High-Efficiency Switching Regulator

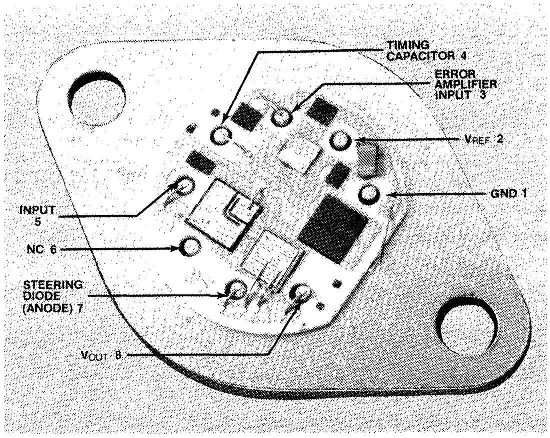
Hybrid Products

Description

The SH1605 is a hybrid switching regulator with high output current capabilities. It incorporates a temperature-compensated voltage reference, a duty-cycle controllable oscillator, error amplifier, high current-high voltage output switch, and a power diode. The SH1605 can supply 5 A of regulated output current over a wide range of output voltage.

- **STEP-DOWN SWITCHING REGULATOR**
- **OUTPUT ADJUSTABLE FROM 3 TO 30 V**
- **5 A OUTPUT CURRENT**
- **HIGH EFFICIENCY**
- **FREQUENCY UP TO 100 KHz**
- **UP TO 150 W OUTPUT POWER**
- **STANDARD 8-PIN, TO-3 PACKAGE**

Connection Diagram
8-Pin TO-3 Type

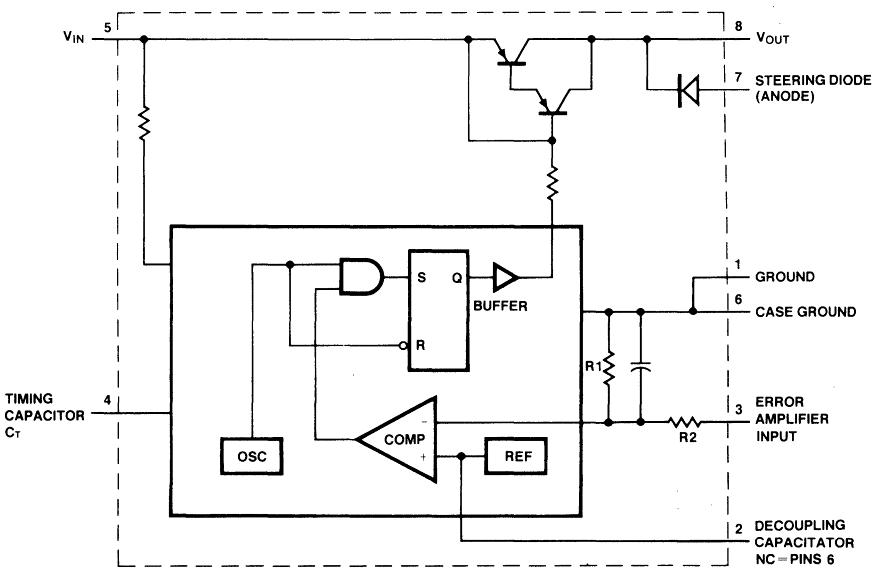


(Bottom View)

Order Information

Output Voltage	Temperature Range	Part Number
3 V To 30 V	0°C to +70°C	SH1605SC
3 V To 30 V	−55°C to +150°C	SH1605SM

Block Diagram



4

Absolute Maximum Ratings

$V_{in} - V_{out}$ (Min)	5 V	Storage Temperature	−65°C to +150°C
Input Voltage	35 V Max	Internal Power Dissipation	20 W
Output Current	6 A	Duty Cycle	20% to 80%
Operating Temperature (T_J)	150°C	Steering Diode Reverse Voltage	60 V
Operating Temperature (T_A)		Steering Diode Forward Current	6 A
SH1605SC	0°C to +70°C		
SH1605SM	−55°C to +125°C		

Electrical Characteristics: $T_C = 25°C$, $T_{IN} = 15$ V, $V_{OUT} = 10$ V unless otherwise specified.

Symbol	Characteristics	Conditions	SH1605SC/SH1605SM Min	Typ	Max	Units
V_{OUT}	Output Voltage	$V_{IN} \geq V_O + 5$ V, Io = 2 A	3.0		30.0	V
V_S	Switch Saturation	$I_{OUT} = 5.0$ A, $I_{OUT} = 2.0$ A		1.5 1.0	2.0 1.2	V V
V_F	Diode On Voltage	$I_{OUT} = 5.0$ A, $I_{OUT} = 2.0$ A		2.2 1.6	2.8 2.0	V V
V_{CC}	Supply Voltage		10		35	V
I_{RD}	Diode Reverse Current	$V_{RD} = 25$ V		2.0		μA
I_O	Quiescent Current	$I_{OUT} = 0.2$ A		30		mA

Reference and Oscillator Section

Symbol	Characteristics	Conditions	Min	Typ	Max	Units
XY_{REF}	Voltage on Pin 3			2.5		V
$\triangle V_3/T$	V_3 Temperature Coefficient			150		ppm/°C
XI_C	Charging Current-Pin 4	$V_{IN} = 10$ V	20	25	50	μA
		$V_{IN} = 35$ V	20		70	
$\triangle V_c$	Voltage Swing-Pin 4			0.5		V
I_D	Discharging Current - Pin 4	$V_{IN} = 10$ V	150	225	250	μA
		$V_{IN} = 35$ V	150		350	

Switching Characteristics (See Test Circuit)

Symbol	Characteristics	Conditions	Min	Typ	Max	Units
t_r	Voltage Rise Time	$I_{OUT} = 2.0$ A $I_{OUT} = 5.0$ A		700 1.8		ns μs
t_f	Voltage Fall Time	$I_{OUT} = 2.0$ A $I_{OUT} = 5.0$ A		700 900		ns ns
t_s	Storage Time	$I_{OUT} = 5.0$ A		2.6		μs
t_d	Delay Time	$I_{OUT} = 2.0$ A		2.5		μs

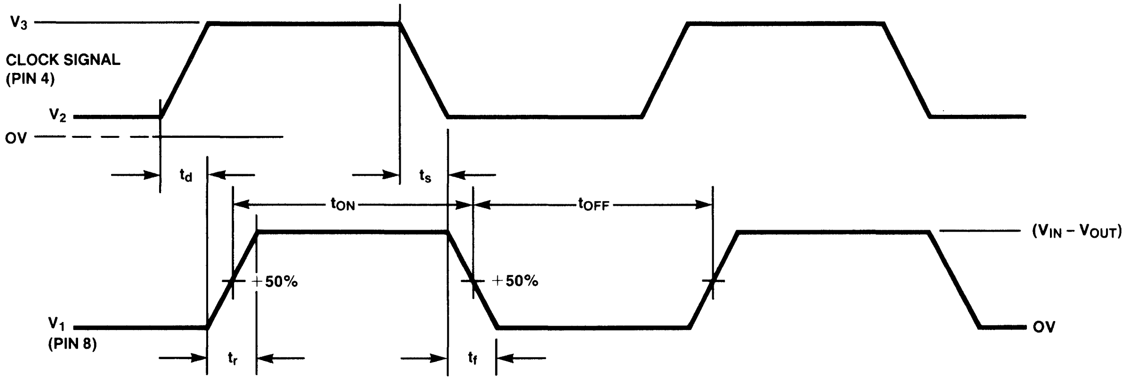
Thermal Characteristics

Symbol	Characteristics	Conditions	Min	Typ	Max	Units
P_D	Power Dissipation	$I_{OUT} = 5.0$ A $V_{OUT} = 10$ V		16		W
η	Efficiency	$I_{OUT} = 10$ V $V_{OUT} = 5$ A		75		%
$\theta_{J\text{-}C}$	Thermal Resistance			4.5		°C/W

Notes

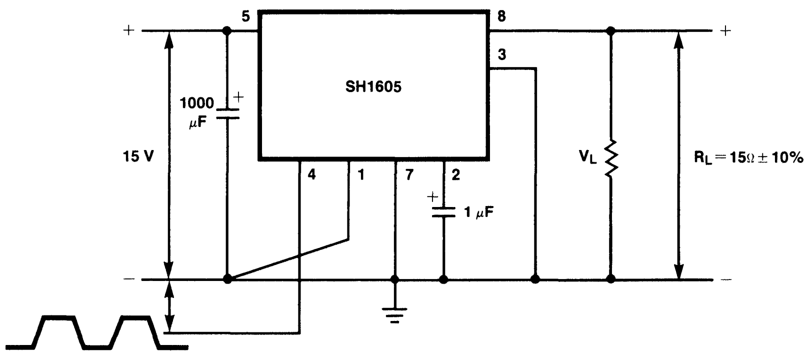
1. Typical is 30°C/W for natural convection cooling.
2. For heatsinking requirements see power derating curve.
3. V_{OUT} refers to the output voltage range of a switching supply the output LC filter as shown in the typical application circuit.

Switching Waveforms

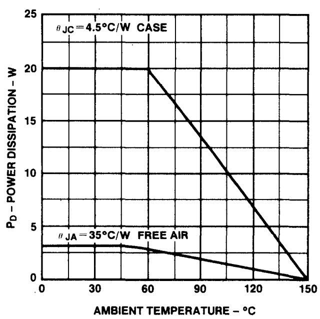


WHERE $V_2 \leq 0.2$ V
$4.0 \text{ V} \geq V_3 \geq 2.0$ V

Switching Characteristics Test Circuit



Power Derating Curve



Design Equations

$$\text{Efficiency } (\eta) = \frac{P_{OUT} \times 100}{P_{IN}}$$

$$\text{Transistor DC Losses } (P_T) = I_{OUT} \times V_s \frac{t_{ON}}{t_{ON} + t_{OFF}}$$

$$\text{Diode DC Losses } (P_D) = I_{OUT} \times V_F \frac{t_{OFF}}{t_{ON} + t_{OFF}}$$

$$\text{Drive Circuit Losses } (D_L) = \frac{V_{IN}^2}{300} \times \frac{t_{ON}}{t_{ON} + t_{OFF}}$$

Switching Losses Transistor

$$(P_S) = V_{IN} \times I_{OUT} \frac{t_r + t_f}{2\,(t_{ON} + t_{OFF})}$$

$$\text{Transistor Duty Cycle} = \frac{t_{ON}}{t_{ON} + t_{OFF}} = \frac{V_{OUT}}{V_{IN}}$$

$$\text{Diode Duty Cycle} = \frac{t_{OFF}}{t_{ON} + t_{OFF}} = 1 - \frac{V_{OUT}}{V_{IN}}$$

Power Inductor

$(P_L) = I_{OUT}^2 \times R_L$ (Winding Resistance)

Efficiency (η) =

$$\frac{V_{OUT}\, I_{OUT}}{V_{OUT}\, I_{OUT} + P_T + P_D + D_L + P_S + P_L} \times 100$$

Where:
- P_{OUT} = Output Power Dissipation
- P_{IN} = Input Power Dissipation
- I_{OUT} = Output Current
- V_S = Darlington Switching Saturation Voltage
- t_{on} = Regulator "On" Time
- t_{off} = Regulator "Off" Time
- V_F = Steering Diode Forward Voltage Drop
- V_{IN} = Input Voltage
- t_r = Regulator Switching Rise Time
- t_f = Regulator Switching Fall Time
- V_{OUT} = Output Voltage
- R_L = Inductor Winding Resistance

V_{OUT} SET RESISTANCE = R_S

$$= \left[\frac{V_{OUT}\,(R_1 + R_2)}{V_{REF}}\right] - \left[R_1 + R_2\right]$$

$$= \left[\frac{2 \times 10^3\, V_{OUT}}{V_{REF}} - 2 \times 10^3\right]$$

$= 8 \times 10^2\, V_{OUT} - 2 \times 10^3$ Typical

Where: Internal Resistors = $R_1 = R_2$
$= 1 \times 10^3\, \Omega$
Reference Voltage On Pin 3 = V_{REF}
= 2.5 V Typical

$$\text{Inductance} = L_1 = \left(\frac{V_{in(nom)} - V_{OUT}}{\Delta I_1}\right) \times t_{ON}$$

Where: Change in Inductor Current = ΔI_1
$= 2 \times I_{OUT(Min)}$
Minimum Continuous Output Current
$= I_{OUT\,(Min)}$
On Time = $t_{ON} > (td + ts)$
t_{ON} is determined by the design and depends upon the desired frequency of operation under constant load conditions where frequency = $1/(t_{on} + t_{off})$. Off Time, t_{off}, is determined by the ratio of input voltage and output voltage where

$$t_{OFF} = t_{ON} \times \left(\frac{V_{IN}}{V_{OUT}} - 1\right)$$

Delay Time = td = 2.5 μs Typical
Storage Time = ts = 2.6μs Typical
Nominal Input Voltage = $V_{IN(NOM)}$
Output Voltage = V_{OUT}

$$\text{Timing Capacitance } (C_T) = \frac{t_{ON} \times I_c}{\Delta V_c}$$

Where: Charging Current on Pin 4 = I_c
= 25 μA Typical
Voltage Swing on Pin 4 = ΔV_c
= 0.5 V Typical

$$\text{Frequency} = F = \frac{1}{\dfrac{C_T\, \Delta V_C}{I_c} + \dfrac{\Delta I_1\, L_1}{V_{OUT} + V_F}}$$

Where: Steering Diode Forward Voltage Drop
$= V_F$
= 2.2 V @ 5 A Typical
(From Elect. Char.)
= 1.6V @ 2 A Typical
(From Elect. Char.)

Minimum Output Capacitance = $C_{OUT\,(MIN)}$

$$= \frac{\Delta I_1}{(8 \times F_{(MIN)} \times V_{RIPPLE(MAX)})}$$

Where: Minimum Expected Frequency = $F_{(MIN)}$

$$F_{(MIN)} = \frac{1}{\dfrac{C_T\, \Delta V_C}{I_c} + \dfrac{\Delta I_1\,(MAX)\, L}{V_{OUT} + V_F}}$$

Maximum Change in Inductor Current

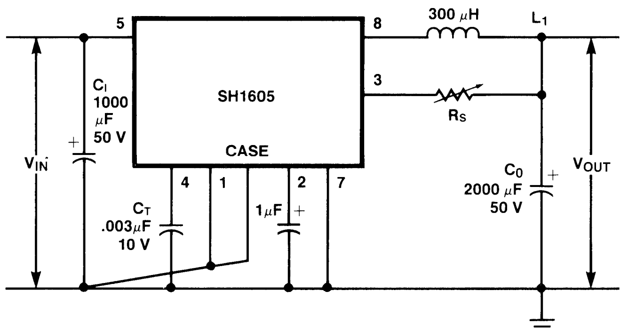
$$= \Delta I_{1(MAX)} = \left(\frac{V_{IN(MAX)} - V_{OUT}}{L_1}\right) \times t_{ON}$$

Maximum Expected Input Voltage
$= V_{IN(MAX)}$
Maximum Expected Ripple Voltage
$= V_{RIPPLE(MAX)}$

Effective Series Resistance of C_{ONT} = ESR

$$= \frac{V_{RIPPLE(MAX)}}{\Delta I_{1(MAX)}}$$

Typical Application



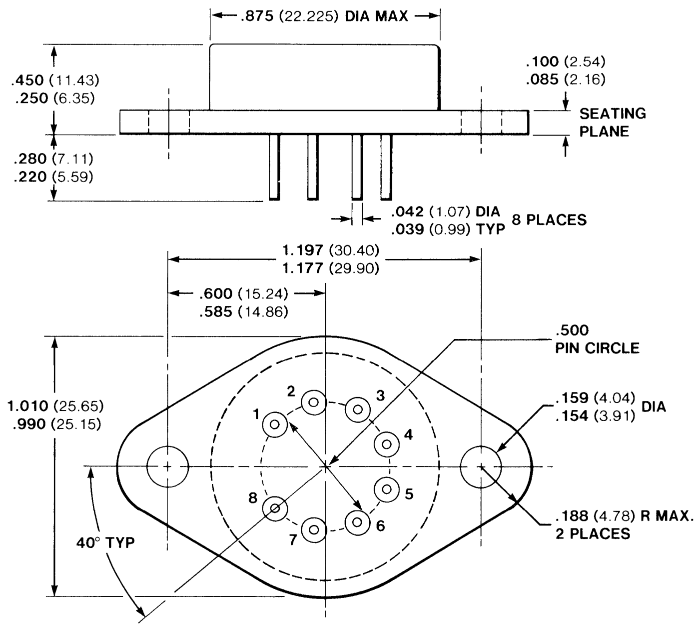
R_S = V_{OUT} Set Resistor

$$R_S = \frac{2 \times 10^3 (V_{OUT} - 2.5)}{2.5}$$

V_{IN} = 12 − 18 V
V_{OUT} = 5 V
I_{OUT} = 5 A (Max)
I_{OUT} = 1 A (Min)

η = 70%
Load Reg. = 50 mV
Line Reg. = 50 mV
Ripple = 100 mV

Package Outline
(S Package — Steel)



Notes
All dimensions in inches bold and millimeters (parentheses)
Pins are solder-dipped alloy 52

Following is a partial list of sockets and heat dissipators for use with the SH1605. Fairchild assumes no responsibility for their quality or availability.

8-Lead TO-3 Hardware

Sockets	Heat Sinks	Mica Washers
Robinson Nugent 0002011	Thermalloy 2266B (35°C/W)	Keystone 4858
Azimuth 6028 (test socket)	IERC LAIC 3B4CB	
Augat 8112 - AG6	IERC HP1-TO3-33CB (7°C/W)	
	AAVID 5791B	

AAVID
ENGINEERING
30 Cook Court
Laconia, New Hampshire 03246

Azimuth Electronics
2377 S. El Camino Real
San Clemente, CA 92672

Augat
P.O. Box 779
Attleboro, MA 02703

IERC
135 W. Magnolia Blvd.
Burbank, CA 91502

Keystone Electronics Corp
49 Bleecker St.
New York, N.Y. 10012

ROBINSON
NUGENT INC.
800 E. 8th St.
New Albany, IN 47150

Thermalloy
P.O. Box 34829
Dallas, Texas 75234

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High Current Voltage Regulator Applications

This application note is to assist the user in designing power supplies and on card regulation systems using Fairchild's family of series pass High Current Voltage Regulators.

Selecting the Correct High Current Voltage Regulator

The regulator selection guide *(Table 1)* provides a concise table of key regulator specifications by device number. Select the device that provides the desired output voltage and current, then proceed as follows.

Determine the required input voltage (V_{IN}).

$V_{IN(max)} > V_{IN} > V_{OUT(max)} + V_{DD(max)} + \triangle V_L + V_{R(pk)}$

where

$V_{IN(max)}$ = Maximum allowable input voltage

V_{IN} = Regulator input voltage under load

$V_{OUT(Max)}$ = Maximum output voltage of regulator

$V_{DD(max)}$ = Maximum dropout voltage

$\triangle V_L$ = Maximum line voltage change

$V_{R(pk)}$ = Peak ripple voltage

Also determine $T_{A(max)}$ = Maximum ambient temperature and select $T_J < T_{J(max)}$ from the data sheet and see the application note titled "Thermal Considerations" for heat sink requirements.

Design Precautions

When designing and laying out a regulator circuit, follow these guidelines to save time, money and simplify design.

- Keep all ground leads as short as possible. Use ground conductors sufficiently large enough to handle rated currents to reduce unwanted voltage drops across leads, and to minimize heating effects and lead inductance.

High Current Voltage Regulator Selection Guide

Device	Function	Input Voltage Max (V)	Output Voltage Range (V)	Output Current Max (A)	Line Regulation (%)	Load Regulation (%)	Quiescent Current (MA)	Ripple Rejection Min (dB)	Dropout Voltage (V)	Thermal Resistance Max (°C/W)		Package
										θ_{JC}	θ_{JA}	
SH323	Fixed Positive	40	4.85 5.25	3	0.2	0.2	3	60	2	2.5	38	2-Pin TO-3
78H05A	Fixed Positive	40	4.85 5.25	5	0.2	0.2	3	60	2.3	2.5	38	2-Pin TO-3
78H12A	Fixed Positive	40	11.5 12.5	5	0.2	0.2	3.7	60	2.3	2.5	38	2-Pin TO-3
78HGA	Adjustable Positive	40	5 24	5	0.2	0.2	3.4	60	2.3	2.5	38	4-Pin TO-3
79HG	Adjustable Negative	−40	−2.11 24	−5	0.4	0.7	−5	60	−2.2	2.5	38	4-Pin TO-3
78P05	Fixed Positive	40	4.85 5.25	10	0.2	1.0	3.4	60	2.5	1.8	38	2-Pin TO-3

- Use single-point grounding at the regulator common terminal whenever possible to prevent circulating currents or ground loops.
- When using the adjustable multi-terminal regulators, especially at high output current levels, derive the feedback sense voltage from across the load rather than from across the regulator to improve circuit performance.
- High Current Voltage Regulators are particularly attractive because of the small number of external components required. It is good practice, however to use bypass capacitors at all times. Input bypass capacitors (1 μF for positive positive regulators and 2 μF for negative regulators) are especially critical if the regulator is located any appreciable distance from the power supply filter. Output bypass capacitors (0.1 μF for positive regulators and 1.0 μF for negative regulators) are also required to improve transient response. These bypass capacitors should be mylar, ceramic or tantalum with good high frequency characteristics. If more than one bypass capacitor source or more than one type is used, stability should be checked on each source or type. Stable operation with one capacitor from one vendor may not necessarily result in stable operation with a capacitor of the same type from a second vendor, since the characteristics of the capacitors may vary.

  Regulator output impedance is in the order of 100 mΩ or less and increases as a function of frequency above 10 kHz due to the gain rolloff of the error amplifier. A tantalum electrolytic bypass capacitor connected to the regulator output will maintain low impedance for frequencies up to 1 MHz. A ceramic capacitor should be placed in parallel with the tantalum capacitor for driving fast switching loads to compensate for the rising impedance of the electrolytic capacitor above 1 MHz. If switching loads are distributed over a large area, additional ceramic bypass capacitors should be located at the loads. Very large-value output bypass capacitors should not be used unless adequate measures are taken to prevent the output from rising above the input, or to avoid discharging the bypass capacitor through the series-pass transistor of the regulator if the input is accidentally grounded. A reverse-biased diode connected from input to output is normally sufficient to achieve this protection.
- Internal protection circuits are provided in all High Current Voltage Regulators to improve reliability and make these regulators immune to certain types of overloads. These on-chip components protect the regulators against short-circuit conditions (current limit), excessive input-output voltage differential conditions (safe-area limit) and excessive junction temperatures (thermal limit). The protection circuits protect the device against abuse and fault conditions that may be encountered *occasionally*. Continuous operation of the device under fault conditions such as a short or in a thermal shutdown mode is *not* a recommended procedure.

  Proper attention must be paid to the safe-area protection network when these regulators are operating with excessive input voltage or excessive input-output differential-voltage conditions. In addition to reducing the available output current with high input-output differential conditions, the safe-area protection network may, under certain conditions, cause the device to latch-up if the output is shorted to ground. This situation is aggravated as the input voltage, load current or the operating junction temperature is increased. This mode of operation does not damage the device but power (input voltage or load current) must be interrupted momentarily for the device to recover from the latched condition.

  Precautions must also be taken to avoid regulator operation beyond its absolute maximum ratings. Switching transients exceeding the maximum input voltage rating of a regulator, for instance, can destroy a regulator. These transients, which occur especially if the regulator input voltage is switched instantaneously rather than ramped by the natural smoothing provided by the ac line and the filter capacitors, are usually hard to track and normally caused by lead inductance and fast switching currents. Good quality bypass capacitors that have low series resistance cause the inrush current to increase further, thereby causing a higher magnitude transient at the input of the regulator. In such cases, a lower quality and cheaper bypass capacitor may be the answer.

  Because of their output stage configurations, positive regulators source current and negative regulators sink current. These restrictions should be kept in mind and, under no circumstance, should a regulator output terminal be allowed to go more than a few volts higher than the regulated output of the regulator. The power should be turned off before removing or inserting a regulator into a test socket. However, if it is necessary to insert a regulator into a "live" socket, care must be taken to ensure that the common terminal connection is made prior to, or simultaneously with, the input terminal connection. In the absence of the common terminal connection, the output voltage of the regulator is 1 or 2 V below the input voltage. This type of fault condition can cause an excessive output voltage which may adversely affect the circuits supplied by the regulator. If the common terminal is quickly connected, the regulator can be destroyed. Also, damage to the regulator may result from the discharging of the bypass capacitor through the output and common terminals.

- The thermal properties and limitations of voltage regulators are extremely important in circuit design. Whether or not a heat sink is required should be determined before the circuit is laid out. See the application note entitled "Thermal Considerations."

Applications
A few of the most popular High Current Voltage Regulator Applications are illustrated in this section. These illustrations include both basic applications and some applications more exotic to extend the capabilities of the regulator.

Basic High Current Voltage Regulator Configurations
Figure 1 shows the basic connection diagram for fixed positive high current voltage regulators including the SH323, SH223, SH123, μA78H05, μA78H05A, μA78H12A and the μA78P05. The user may refer to *Table 1* or the individual data sheets to determine which regulator satisfies his system needs.

Adjustable regulators are ideal for applications that require non-standard output voltages. Output voltages are determined by the following equation:

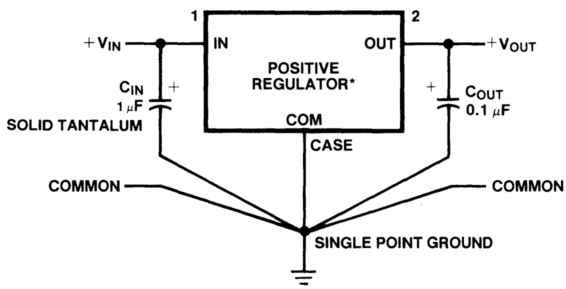
$$V_{OUT} = \left(\frac{R_1 + R_2}{R_2}\right) V_{CONTROL}$$

Where: R_1 and R_2 are set resistors as shown in *Figures 2* and *3*.

$V_{CONTROL} = 5\ V_{(NOMINAL)}$ for the μA78HGA
$= -2.23\ V_{(NOMINAL)}$ for the μA79HG

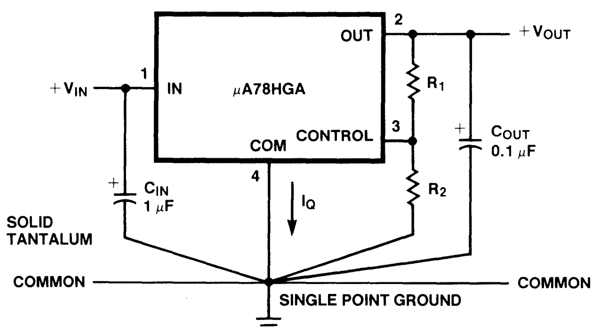
Output voltage can be set anywhere between +5 V to +24 V for the μA78HGA and −2.11 V to −24 V for the μA79HG. A trimpot may also be substituted for R_1 and R_2 to allow for either full range adjustments or output voltage trimming.

Fig. 1 Basic Fixed Positive High Current Voltage Regulator with Bypass Capacitors



* Device Type SH323, SH223, SH123, μA78H05, μA78H05A, μA78H12A, or μA78P05 depending upon desired system parameters. See *Table 1.*

Fig. 2 A Basic Positive Adjustable High Current Voltage Regulator

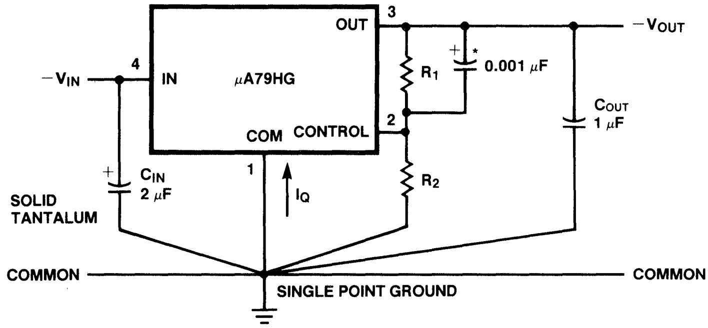


Notes

$$V_{OUT} = \frac{R1 + R2}{R2} V_{CONTROL}$$

$V_{CONTROL}$ Nominal = 5 V
Recommended R2 current ≈ 1mA

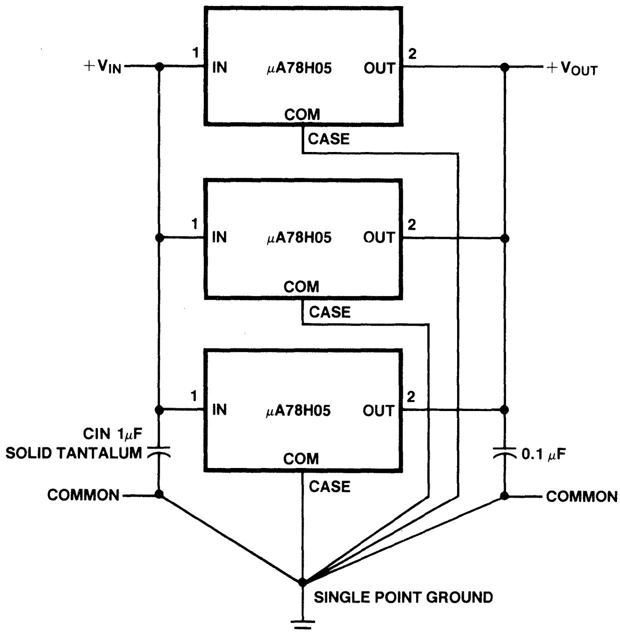
Fig. 3 A Basic Negative Adjustable High Current Voltage Regulator



*May be necessary with long leads

$$V_{OUT} = \left(\frac{R1 + R2}{R2} \right) V_{CONTROL} \qquad V_{CONTROL}\ \text{Nominal} = -2.23\ \text{V}$$

Fig. 4 Parallel Operation of Regulators For Very High Current



Parallel Regulators
To obtain even higher output current, several regulators in parallel may be used. Regulation of the overall system can be improved if the individual devices are matched for output voltages as shown in *Figure 4.* If the outputs are not matched, it is likely that the output current will not be shared between the regulators and, as a result, some of the regulators will operate at or near the current limit while others are at their quiescent no-load levels.

Excessive Input/Output Differential
When a regulator is operating with a large input-output differential, the addition of a series resistor with the input extends the operating range of the device by sharing the power dissipation, see *Figure 5.* The value of the series resistor R1 must be low enough so that, under worst-case conditions, (lowest supply voltage, highest output voltage, and highest load) the device remains in regulation. R1 can be calculated as follows.

$$R1 = \frac{V_{S(min)} - V_{OUT(Max)} - V_{DD(max)}}{I_{OUT(max)} + I_{Q(max)}}$$

where

$V_{s(min)}$ is the minimum supply voltage
$V_{DD(max)}$ is the maximum dropout voltage
$I_{Q(max)}$ is the maximum quiescent current
$V_{OUT(max)}$ is the maximum output voltage

Maximum regulator dissipation, however, occurs with highest supply voltage and highest load current.

$$P_{D(max)} = [V_{IN(max)} - V_{OUT(min)}]\, I_{OUT(max)}$$

where

$$V_{IN(max)} = V_{S(max)} - [I_{OUT(max)} + I_{Q(max)}]\, R1$$

For a constant load, the regulator input voltage varies by the same amount as the supply voltage and consequently the line regulation of the device remains essentially the same.

For load regulation, assuming constant supply voltage, the combined effects of the change at the input due to the voltage change across R1 must be taken into consideration. In this configuration, as the load is increased, the regulator input voltage decreases and the net result, in most cases, is a slight degradation in the performance of the regulator since these two effects are additive.

The load regulation can therefore be calculated as follows.

Load regulation at constant V_S = load regulation at constant V_{IN} + line regulation

Example: Assume a supply range of 25 to 35 V used with a μA78HG12A regulator delivering an output current of 1 to 3 Amps.

From the data sheet: $V_{OUT(min)} = 11.5$ V

$V_{OUT(max)} = 12.5$ V

$V_{DD(max)} = 2.5$ V

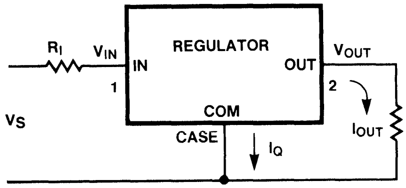
$I_{Q(max)} = 10$ mA

$$R1 = \frac{25 - 12.5 - 2.5}{3 + .01} = 3.3\ \Omega$$

With this value of R1 and a load varying from 1 to 3A, the input voltage to the regulator varies,

$$\triangle V_{IN} = \triangle I_{OUT}\, R1 = 6.6V$$

Fig. 5 Reducing Power Dissipation in a Regulator with Dropping Resistor R1



The effect of the 6.6 V change at the regulator input under worst case conditions can be determined from a ratio of data sheet parameters:

$$\frac{\text{Line Regulation (Max)}}{\triangle V_{IN}\ \text{(For Line Regulation Test)}} \times \triangle V_{IN} =$$

$$\frac{120\ \text{mV}}{25\ \text{V} - 16\ \text{V}} \times 6.6\ \text{V} = 88\ \text{mV}$$

The effect is 88 mV additional change at the output terminal.

The inclusion of the 33 Ω reduces the maximum power dissipation of the regulator as shown below.

From $P_{D(max)} = (35 - 11.5) \times 3 =$ 70.5 W (without R1).

To $P_{D(max)} = (35 - 3.3 \times 3 - 11.5) \times 3 =$ 40.8 W (with R1)

Note that the power dissipation is shared between the regulator and R1.

Although bypass capacitors are not shown in *Figure 5*, it is recommended that they be incorporated in the design as illustrated in *Figure 1*.

Input Voltage > $V_{IN(max)}$

When a regulator is used with supply voltages greater than the rated regulator maximum input voltage, the circuit shown in *Figure 6* can be used. This circuit essentially provides a constant voltage to the regulator with supply voltage variations. The choice of Zener diode voltage is dictated by $V_{IN(min)}$ of the regulator and $V_{BE(max)}$ of Q1.

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Fig 6 Regulator Input Circuit for Input Voltage Source Greater than $V_{IN(max)}$

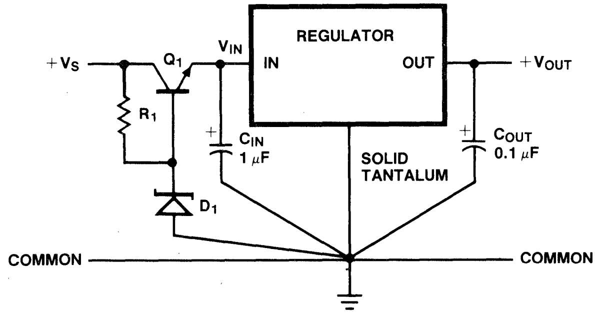


Fig. 7 High Output Voltage Regulator, No Short-Circuit Protection

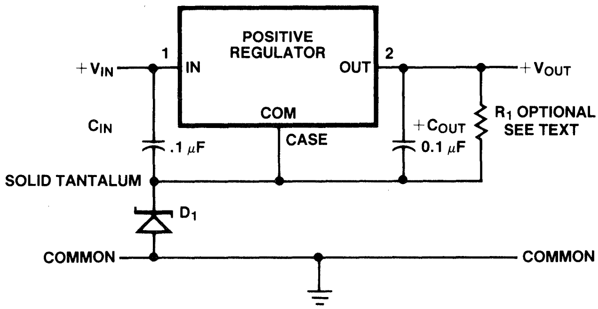
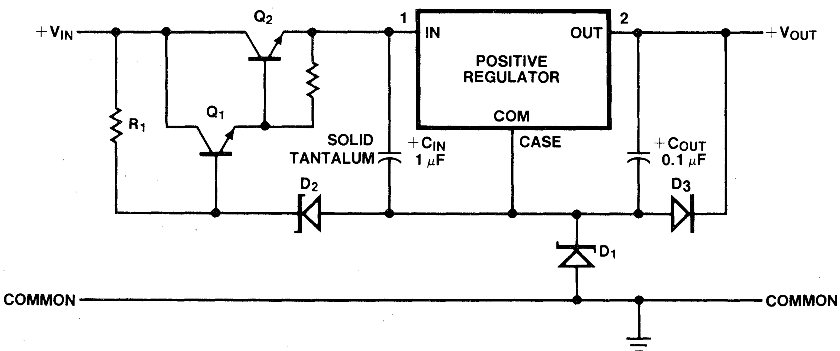


Fig 8 High Output Voltage Regulator with Short-Circuit Protection



High Output Voltages

Figure 7 shows a simple circuit that can be used to obtain an output voltage greater than the standard fixed voltages available. The quiescent current biases Zener diode D1 and the regulator common terminal rides on the pedestal established by D1. If the Zener must be operated at currents greater than the quiescent current level of the regulator, then R1 can be used to increase the Zener current. If, on the other hand, lower Zener current is satisfactory, R1 can be placed in parallel with D1 to shunt some of the current. *Caution:* this circuit configuration cannot utilize the thermal shutdown or short-circuit protection features of the regulator if the input voltage exceeds the maximum input voltage rating of the regulator.

Figure 8 can be used to take advantage of the protective features of the regulator. Here too, the regulator common terminal operates on the pedestal established by Zener diode D1. Zener diode D2 and the Darlington configuration of Q1, Q2 reduces the regulator input voltage to a safe value. The Darlington configuration prevents loading of Zener diode D2, and thus maintains a high level of regulation. Diode D3 protects the regulator against accidental shorts by clamping the common terminal of the regulator to a diode drop above the shorted output.

Remote Shutdown

Electronic shutdown is used in some applications where, under certain conditions, the removal of power from the load is desired. The 3-terminal regulator circuit of *Figure 9* has a remote shutdown feature. Under normal conditions, Q2 is on and provides the base current of Q1.

Q1 acts as a switch and is either in saturation, when the signal to the base of Q2 is high, or is off when the signal to the base of Q2 is low. It must have a current rating equal to the load current. Turn-off time is dependent on C2 and the load current; the higher the load current, the faster the turn-off time.

Constant Current Regulator

Any regulator can be used as a constant-current regulator as shown in *Figure 10.* The current I_{OUT} which dictates the regulator type to be used is determined by this equation.

$$I_{OUT} = \frac{V_{OUT}}{R1} + I_Q$$

where V_{OUT} is the regulator output voltage and I_Q is the quiescent current.

The input voltage V_{IN} must be high enough to accommodate the dropout voltage at the low end, but must not exceed the maximum input voltage rating at the high end.

Positive and Negative Adjustable Regulators

The concepts used above for positive fixed regulators can easily be extended to the μA78HGA, positive adjustable regulator, by simply including the R_1, R_2 resistor network shown in *Figure 2.*

Also, since negative voltage regulators are complements of the positive voltage regulators, almost all the positive regulator applications can be converted into negative versions by appropriate changes in the polarity of the input voltages. If external active components such as series-pass transistors are used, they should be the complements of those used in the positive-regulator application, *i.e.,* npn transistors replaced by pnp and vice versa. Finally, these concepts can be extended to the μA79HG, negative adjustable regulator, by simply including the R_1, R_2 Resistor Network shown in *Figure 3.*

5

Fig. 9 Remote Shutdown

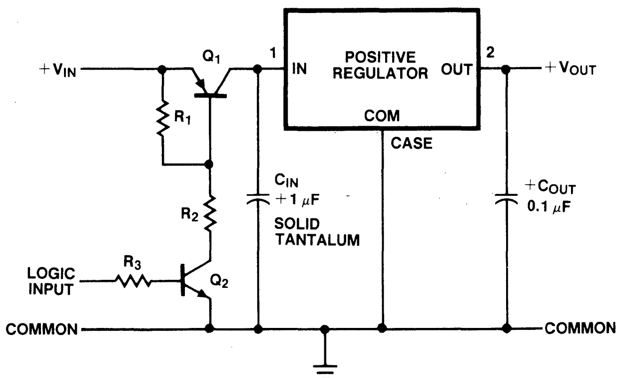
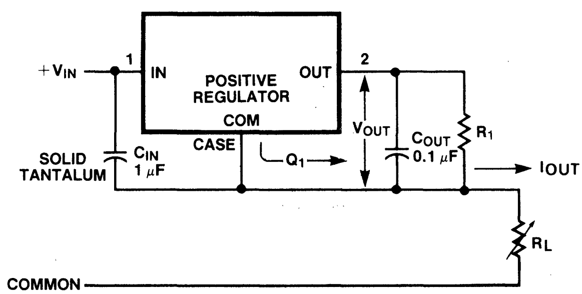


Fig. 10 Constant Current Regulator (Positive Output)



$$I_{OUT} = \frac{V_{OUT}}{R_1} + I_Q$$

Dual Regulators

Dual regulators, or dual power supplies, are normally used for applications requiring two output voltages of opposite polarities that do not necessarily have equal magnitudes, for example, +12 V, −5 V. However, the word dual can also imply two supplies of the same polarity but of different magnitudes, such as +5 V, + 12 V. With dual tracking, not only are the output voltages of different polarities, but one output voltage always follows the other one, *i.e.*, an increase in the positive voltage results in a decrease in the negative output voltage.

Dual Supplies

The simplest dual-polarity high current supply can be obtained by using a positive and a negative adjustable regulator with a center-tapped transformer as shown in *Figure 11.* The same type of dual supply can be achieved with two positive (or two negative) adjustable regulators if a transformer with two isolated windings is used as shown in *Figure 12.*

The reverse-biased diodes connected across the outputs of the dual regulator circuits are not necessary if the loads are referenced to ground. If the loads are tied between the two outputs, however, a latch-up may occur at the instant power is turned on, especially if one regulator input voltage rises faster than the second one. The diodes, that ensure proper start-up of the regulators by preventing a parasitic action from taking place when power is turned on, should have a current rating equal to half the load current.

Fig. 11 Dual Supply using a Center Tapped Transformer with a Positive and a Negative Adjustable High Current Voltage Regulator

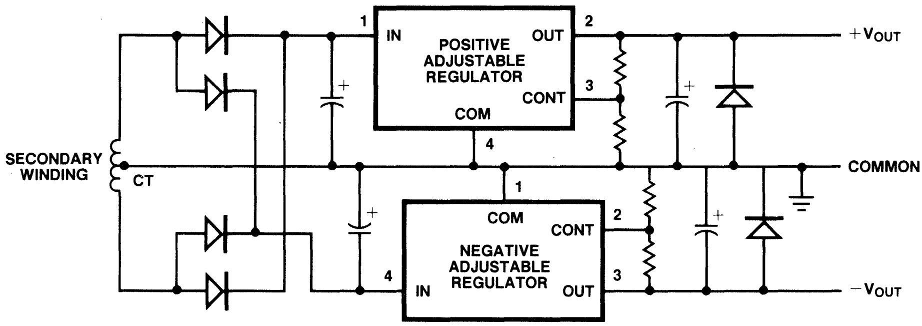
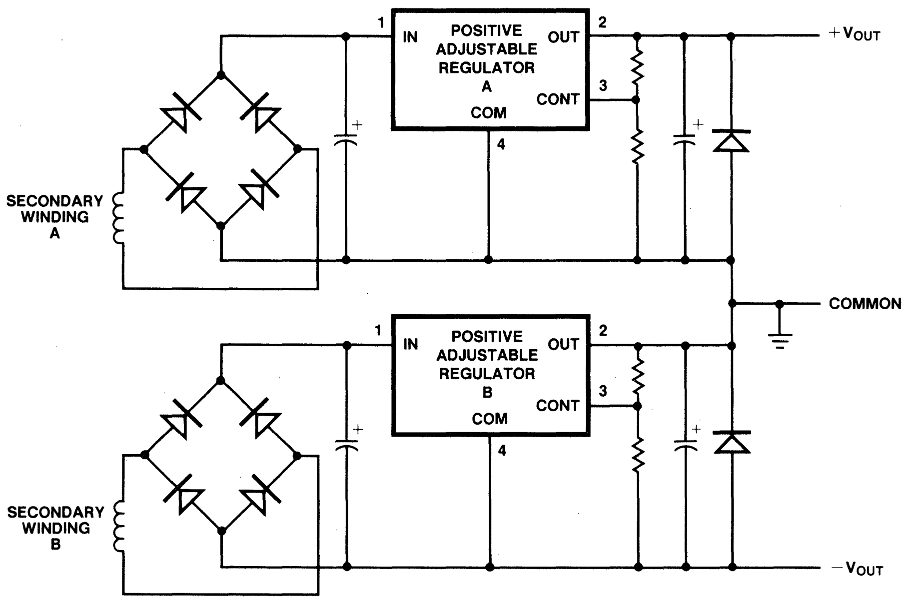


Fig. 12 Dual Supply using a Transformer with Two Windings and Two Positive Adjustable High Current Voltage Regulators



Absolute Maximum Ratings

$V_{in} - V_{out}$ (Min)	5 V	Storage Temperature	−65°C to +150°C
Input Voltage	35 V Max	Internal Power Dissipation	20 W
Output Current	6 A	Duty Cycle	20% to 80%
Operating Temperature (T_J)	150°C	Steering Diode Reverse Voltage	60 V
Operating Temperature (T_A)		Steering Diode Forward Current	6 A
SH1605SC	0°C to +70°C		
SH1605SM	−55°C to +125°C		

Electrical Characteristics: T_C = 25°C, T_{IN} = 15 V, V_{OUT} = 10 V unless otherwise specified.

Symbol	Characteristics	Conditions	SH1605SC/SH1605SM Min	Typ	Max	Units
V_{OUT}	Output Voltage	$V_{IN} \geq V_O + 5$ V, Io = 2 A	3.0		30.0	V
V_S	Switch Saturation	I_{OUT} = 5.0 A, I_{OUT} = 2.0 A		1.5 1.0	2.0 1.2	V V
V_F	Diode On Voltage	I_{OUT} = 5.0 A, I_{OUT} = 2.0 A		2.2 1.6	2.8 2.0	V V
V_{CC}	Supply Voltage		10		35	V
I_{RD}	Diode Reverse Current	V_{RD} = 25 V		2.0		μA
I_O	Quiescent Current	I_{OUT} = 0.2 A		30		mA

Reference and Oscillator Section

Symbol	Characteristics	Conditions	Min	Typ	Max	Units
XY_{REF}	Voltage on Pin 3			2.5		V
$\triangle V_3/T$	V_3 Temperature Coefficient			150		ppm/°C
XI_C	Charging Current-Pin 4	V_{IN} = 10 V	20	25	50	μA
		V_{IN} = 35 V	20		70	
$\triangle V_c$	Voltage Swing-Pin 4			0.5		V
I_D	Discharging Current - Pin 4	V_{IN} = 10 V	150	225	250	μA
		V_{IN} = 35 V	150		350	

Switching Characteristics (See Test Circuit)

Symbol	Characteristics	Conditions	Min	Typ	Max	Units
t_r	Voltage Rise Time	I_{OUT} = 2.0 A I_{OUT} = 5.0 A		700 1.8		ns μs
t_f	Voltage Fall Time	I_{OUT} = 2.0 A I_{OUT} = 5.0 A		700 900		ns ns
t_s	Storage Time	I_{OUT} = 5.0 A		2.6		μs
t_d	Delay Time	I_{OUT} = 2.0 A		2.5		μs

Thermal Characteristics

P_D	Power Dissipation	I_{OUT} = 5.0 A V_{OUT} = 10 V		16		W
η	Efficiency	I_{OUT} = 10 V V_{OUT} = 5 A		75		%
θ_{J-C}	Thermal Resistance			4.5		°C/W

Notes

1. Typical is 30°C/W for natural convection cooling.
2. For heatsinking requirements see power derating curve.
3. V_{OUT} refers to the output voltage range of a switching supply the output LC filter as shown in the typical application circuit.

FAIRCHILD

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SH1605
5-Amp, High-Efficiency Switching Regulator

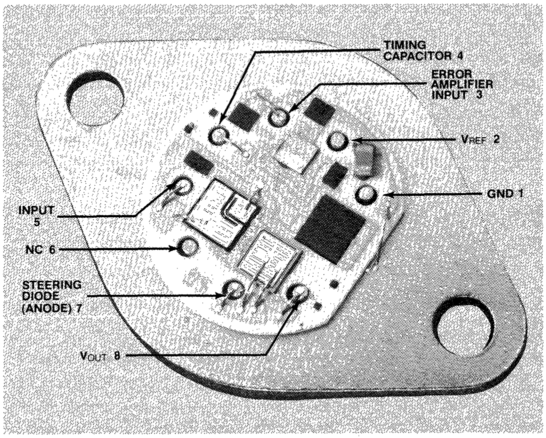
Hybrid Products

Description

The SH1605 is a hybrid switching regulator with high output current capabilities. It incorporates a temperature-compensated voltage reference, a duty-cycle controllable oscillator, error amplifier, high current-high voltage output switch, and a power diode. The SH1605 can supply 5 A of regulated output current over a wide range of output voltage.

- STEP-DOWN SWITCHING REGULATOR
- OUTPUT ADJUSTABLE FROM 3 TO 30 V
- 5 A OUTPUT CURRENT
- HIGH EFFICIENCY
- FREQUENCY UP TO 100 KHz
- UP TO 150 W OUTPUT POWER
- STANDARD 8-PIN, TO-3 PACKAGE

Connection Diagram
8-Pin TO-3 Type



(Bottom View)

Order Information

Output Voltage	Temperature Range	Part Number
3 V To 30 V	0°C to +70°C	SH1605SC
3 V To 30 V	−55°C to +150°C	SH1605SM

Block Diagram

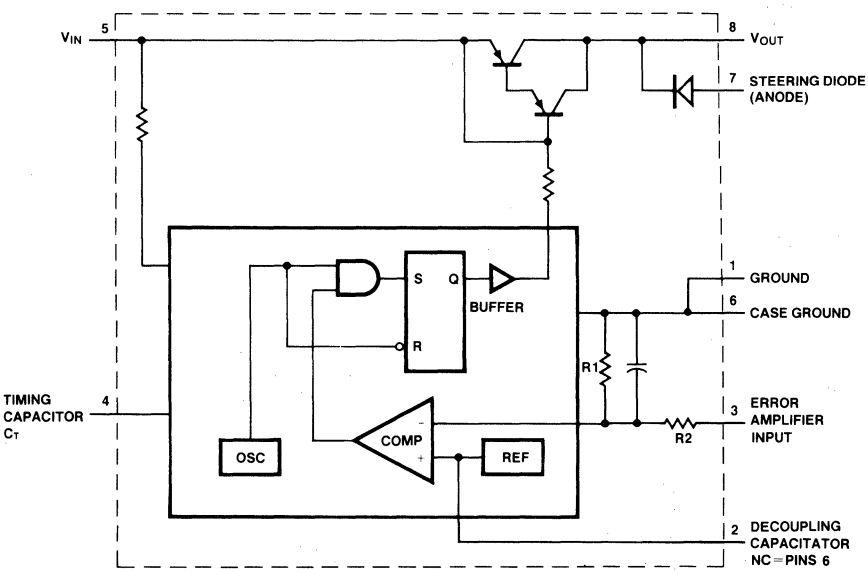


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Fig. 13 A Dual Positive Supply with a +5 V and +10 V Output

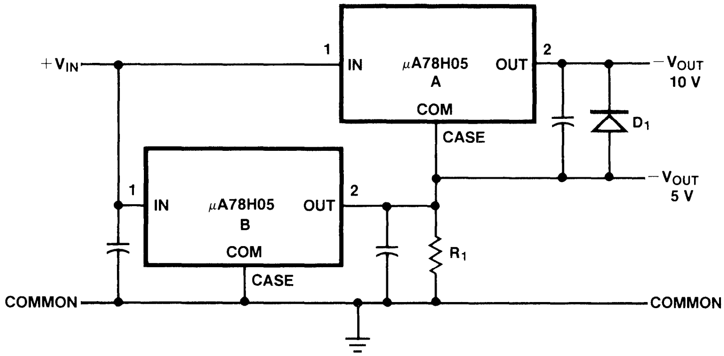
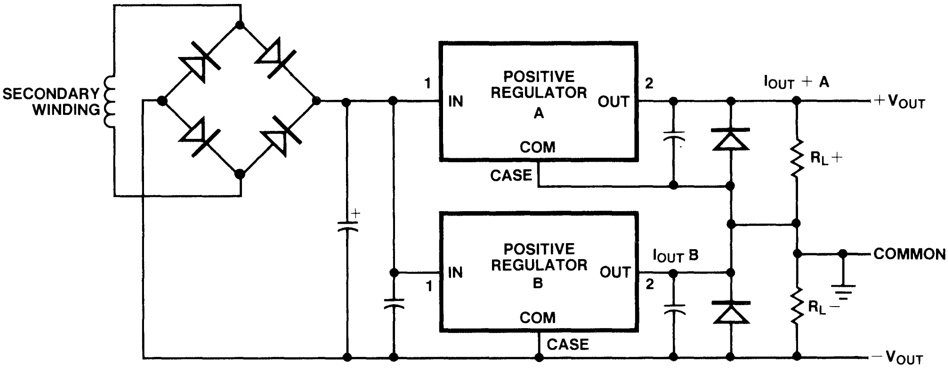


Fig. 14 A Dual Polarity Supply from a Single Transformer Winding



$R_L+ > R_L-$

Figure 13 shows a single positive-polarity dual supply with +5 and +10 V output. It uses two 5 V μA78H05 regulators operating from a single positive voltage source. The +10 V output is achieved by connecting the common terminal of the top regulator A to the output of the bottom regulator B. Diode D1 ensures proper start-up of the top regulator and prevents a latch-up that may occur under a heavy load condition on regulator A. Resistor R1 provides a path for the quiescent current of regulator A and can be eliminated if regulator B has a minimum load current greater than the quiescent current of regulator A.

The concept of *Figure 13* can be used to achieve a dual-polarity output from a floating single supply as shown in *Figure 14*. This circuit is restricted in that $R_L+ > R_L-$, since all of the current provided by the positive regulator A must flow through R_L-.

Adjustble Dual Tracking Regulators

For applications requiring adjustble tracking outputs, the circuit of *Figure 15* can be used. Tracking is accomplished by connecting a common resistor between the control terminals of the two 4-terminal adjustable regulators. Because of the internal feedback of the 4-terminal regulators, a constant voltage is developed across the resistor string R1, R2 and R3. Variations at one of the output nodes are reflected at the control nodes causing corresponding variations at the opposite output node. Note that tracking between the two outputs is not one to one but rather depends on the absolute value of the two references and the feedback resistors R1, R2, and R3. The output voltages are determined by

$$V_{OUT}^{+} = V_{REF}^{+} + \frac{R1}{R2}(V_{REF}^{+} - V_{REF}^{-})$$

$$V_{OUT}^{-} = V_{REF} - \frac{R4}{R3}(V_{REF}^{+} - V_{REF}^{-})$$

Tracking between the outputs can be improved by adding a μA741 and modifying the circuit as shown in *Figure 16*. This circuit yields an adjustable true dual-tracking regulator with internal short-circuit protection, safe-area limiting, thermal overload protection, and is capable of a 5 A maximum output current. The outputs of the regulators are independently adjustable by potentiometers R1 and

R2. With the component values shown, the output voltage of the positive μA78HGA can be varied from 5 to 24 V, and the negative μA79H6 can be varied from −2.11 to −24 V.

This circuit has a positive and a negative regulator and an operational amplifier used as a comparator. Tracking is accomplished by connecting the two regulator common terminals to the output of the μA741 that provides a potential on which the common terminals of the regulators float. The summed regulator outputs, V_O+ and V_O-, are then compared to the power supply common.

The positive and negative regulator outputs track as follows: any change in the positive regulator output causes an opposite change on the common terminals and also on the negative regulator output. For example, a decrease in the positive regulator output voltage causes a like change in the amplitude of the negative regulator output. Since each regulator has a reference, no slaving exists between the outputs and, as a result, tracking is true and independent of polarity.

Proper care must be taken to insure that the maximum supply voltage rating of the μA741 is not exceeded when the regulators are operating with high input voltage sources.

Fig. 15 Adjustble Dual Tracking Regulator

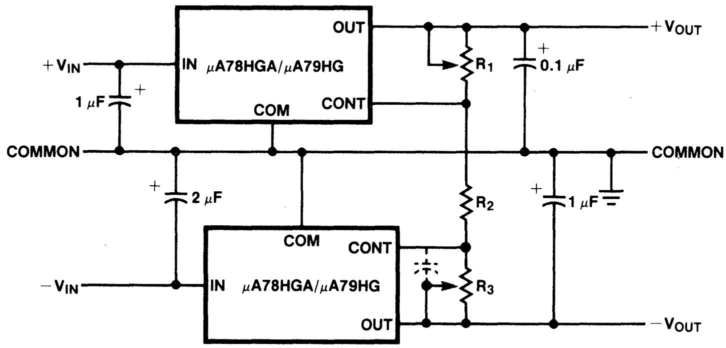
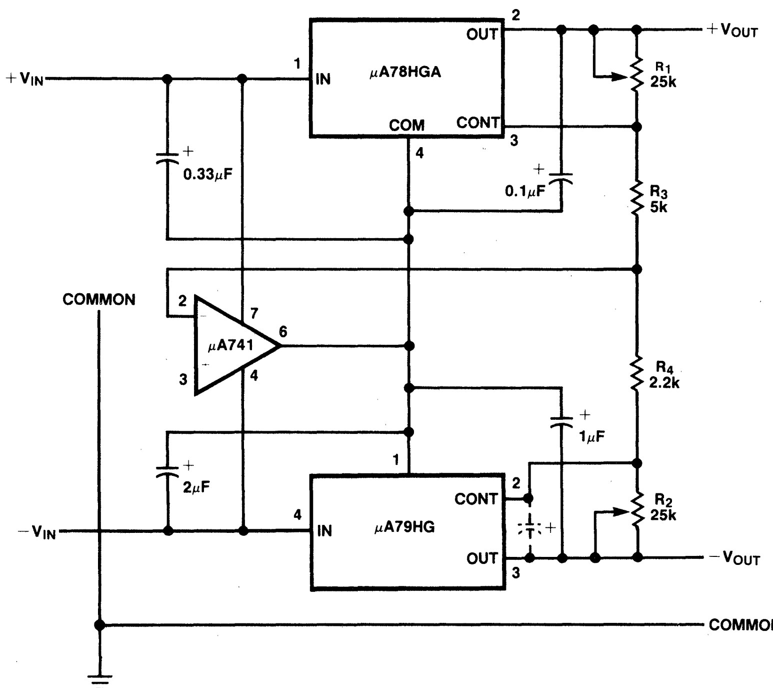


Fig. 16 Independently Adjustable True Dual Tracking Power Supply



Miscellaneous Applications
This section consists of a set of illustrations showing a variety of additional high current voltage regulator applications.

Fig. 17 Negative Output Voltage Circuit

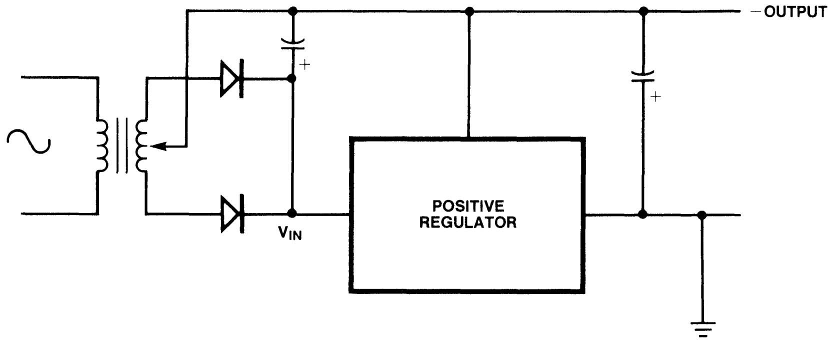
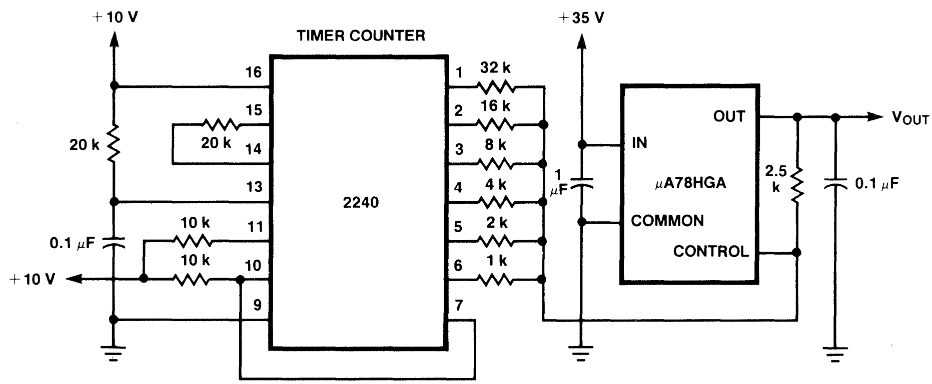


Fig. 18 Programmable Supply



Output Waveform

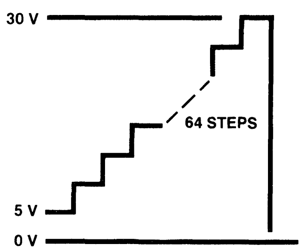
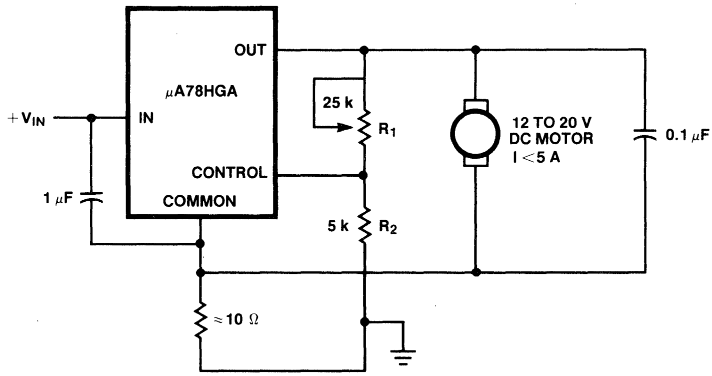
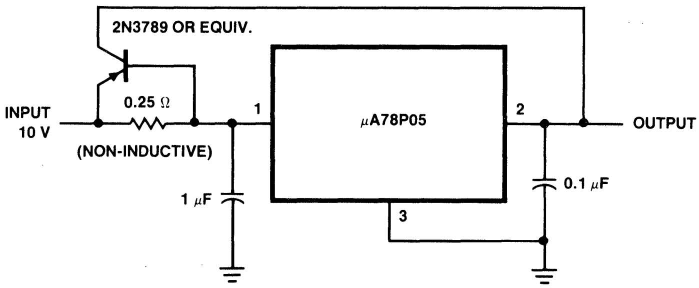


Fig. 19 Motor Speed Control



Use flyback diode across motor if necessary

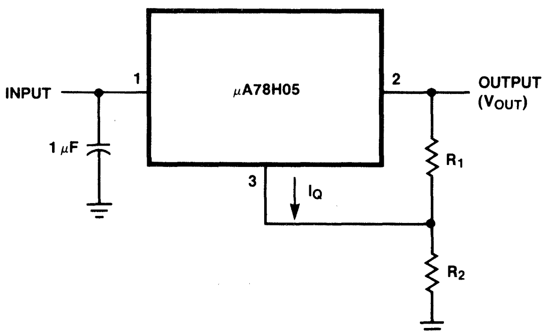
Fig. 20 15+ Amp Regulator



Notes

a. No current limit in effect
b. At 10 A out 9.0 A passes 4.0 A
c. For 10 A output change, V_{OUT} changes approx. 80 mV

Fig. 21 Increased Output Voltage



$$V_{OUT} = V_{RI}\left(1 + \frac{R2}{R1}\right) + I_Q R2$$

Fig. 22 Signal Driver/Modulator

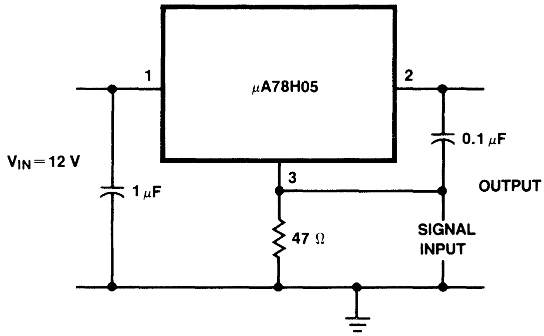
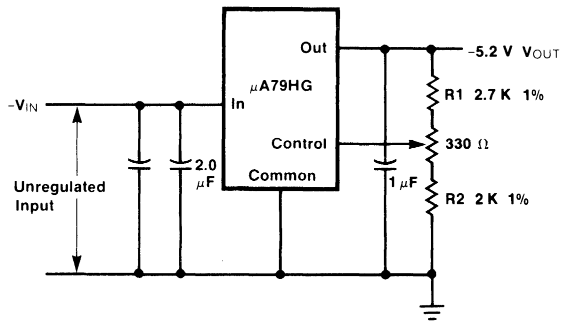
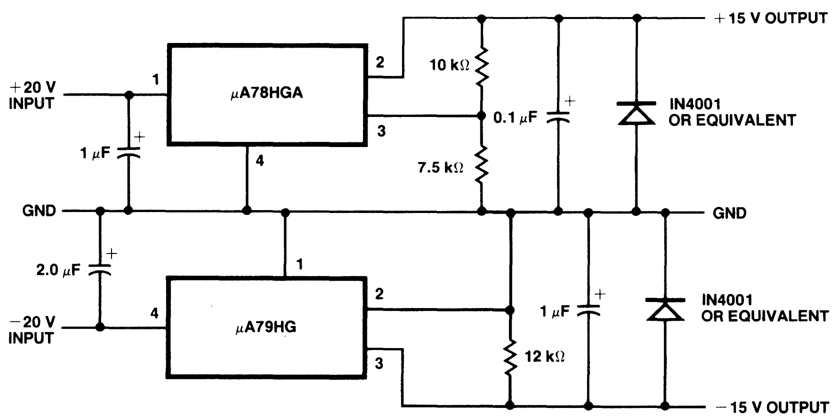


Fig. 23 High-Current ECL Regulator Using μA79HG



*Solid Tantalum Close to Device

Fig. 24 Operational Amplifier Supply (±15 V @ 5 A)



5

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Understanding the Switching Regulator

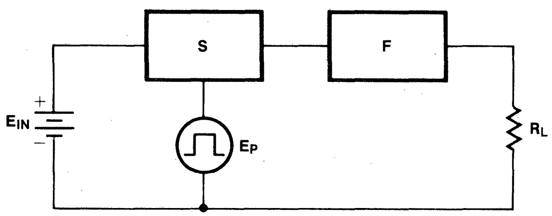
A basic switching regulator is composed of four major components: a voltage source E_{in}, a switch S1, a pulse generator E_p, and a filter F1. The block diagram in *Figure 1* shows the interconnection between these elements. The voltage source may be any dc supply needing conversion and/or regulation — a battery, an unregulated rectified and filtered supply, or even a regulated voltage to be converted into some other required voltage. The requirements for a voltage source are:

- It must be capable of supplying the required output power plus the losses associated with the switching regulator.

$$P_{in} = \frac{P_{out}}{\eta}$$

- The input voltage must be sufficiently high to overcome any IR drops and meet the minimum requirements of the system.
- The input voltage must be large enough to supply sufficient dynamic range for line and load variations.
- In a modern computer power supply, the input voltage may be required to store energy for a specified amount of time during brownouts or power failures.

Fig. 1 Basic Switching Regulator



Switch S is typically a transistor or thyristor connected as a power switch. The switch is inherently efficient because it is operated in the saturated mode. The pulse generator alternately turns S on and off. The pulse is generally an asymmetrical square wave varying in either frequency (frequency modulation) or pulse width (pulse with modulation). Theoretical analysis and formulas generally apply to both frequency or pulse width modulated systems. The frequency f_1 of the pulse generator is usually in the tens of kilohertz to keep components small and switching inaudible. Filter F serves as an averaging filter, converting the pulse from S into a dc voltage. Assuming no losses, the power in equals the power out:

$$E_{in}I_{in} = E_{out}I_{out}$$

This switching mechanism allows a conversion similar to transformers, thus the switching regulator has often been referred to as a dc transformer. The relationship of input and output voltage is a function of duty cycle. The duty cycle is the ratio of the on-time (t_{on}) to the period ($t_p = t_{on} + t_{off} = 1/f$). Thus, the duty cycle

$$\sigma = \frac{t_{on}}{t_{on} + t_{off}}, \quad E_{out} = \sigma E_{in} = \left(\frac{t_{on}}{t_{on} + t_{off}}\right) E_{in}$$

With ($t_{on} + t_{off} = t_p$) constant, the output voltage E_{out} is directly proportional to the on-time t_{on}. Thus, varying t_{on} varies the output voltage (i.e., pulse width modulation).

With t_{on} held constant, the output voltage, E_{out}, is inversely proportional to the period, $t_p = t_{on} + t_{off}$, or directly proportional to the frequency, $f = 1/(t_{on} + t_{off})$.

These techniques allow efficient generation of low voltages from high voltages in a stepdown regulator. Operating from voltages much too high for linear conversion affords a wide dynamic range and high energy input storage for brownouts and missing cycles.

The Filter

The filter or integrating network is of major importance in the proper design for the switching regulator. The filter basically has three forms:

RC filter   RL filter   RLC filter

While all these filters are used in switching regulators, the RLC filter is most often used in series switching regulators. A brief analysis of the RC and RL filters gives the foresight needed for understanding the RLC filter design.

The RC Filter

A simple switching regulator employing an RC filter is shown in *Figure 2.* When a switch Q1 closes, the instantaneous current in capacitor C1 is very large and limited only by the series resistance R_s and the ESR* of the capacitor. This instantaneous current can be found from Kirchhoff's Voltage Law, using Laplace transforms.

$$\frac{E}{S} = I_s R + \frac{I_s}{CS}; \qquad R = R_S + ESR^*$$

$$\frac{E}{S} = I_S\left(R + \frac{1}{CS}\right)$$

$$^*ESR = \text{Effective Series Resistance} = \frac{X_c}{Q}$$

$$I_S = \frac{E}{R\left(S + \frac{1}{RC}\right)}; \qquad a = \frac{-1}{RC}$$

$$\frac{1}{s - a} = e^{at}$$

$$I_S = \frac{E}{R}\, e^{\frac{-t}{RC}}$$

The resultant formula is the familiar equation for finding the current in an RC circuit. It can easily be seen that at $t = 0+$, the current is limited by R only. When switch Q1 is open, the voltage across C1 starts to decay in accordance with the formula:

$$e_c = E\left[1 - e^{\frac{-t}{RC}}\right] \qquad \text{(See } \textit{Figure 2}\text{)}$$

In order to maintain the voltage on C1 (i.e., E_{out}) relatively constant, it is necessary to make the charge time constant much shorter than the load time constant

$$RC1 << R_L C1$$

As R becomes smaller, the averaged square wave approaches a dc source. However, as R becomes smaller, the peak current I_c becomes larger. These peak currents are very high and impractical to switch reliably. As R is increased to limit the current, it becomes noticeably lossey and dissipates excessive power.

The RL Filter

A simple switching regulator employing an RL filter is shown in *Figure 3.* As switch Q1 closes, the voltage across the inductor is the full power supply voltage E. The current supplied to the load at $t = 0+$ is approximately equal to zero and exponentially

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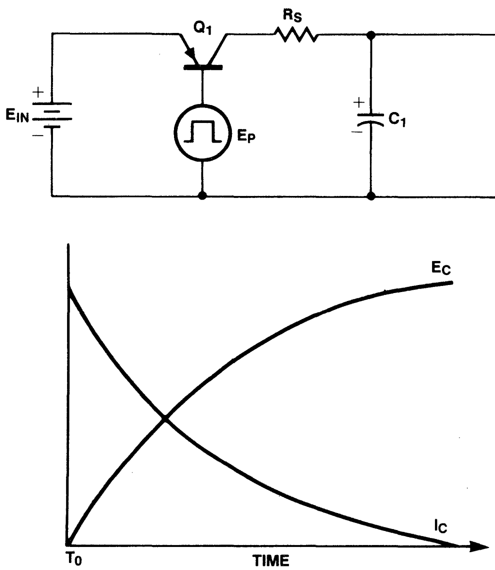
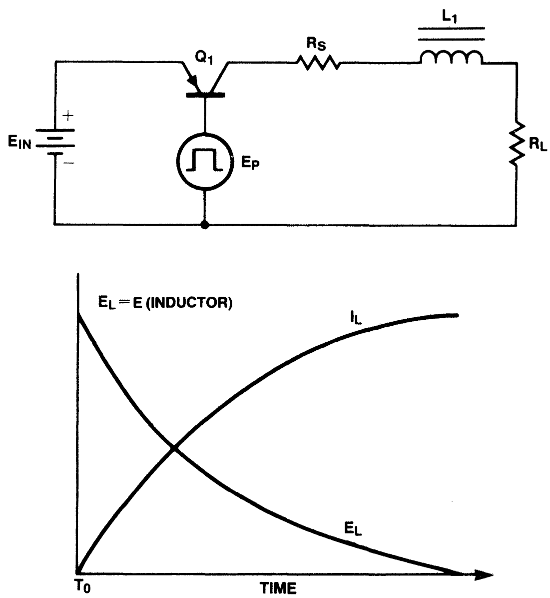


Fig. 3 Simple Switching Regulator with RL Filter



increases as shown in the curve in *Figure 3.* In a similar fashion, the instantaneous current can be found using Laplace transforms and Kirchhoff's Voltage Law:

$$\frac{E}{S} = I_S R + I_S L S; \qquad R = R_S + R_{inductor}$$

$$\text{Yielding: } i_L = \frac{E}{R}\left[1 - e^{\frac{-Rt}{L}}\right]$$

The resultant formula is the familiar equation for finding the current in an RL circuit. It can easily be seen that at $t = 0+$, the current is zero. Thus the time constant L/R must be smaller than the load time constant to average the square wave into a dc source.

$$\frac{L1}{R} << \frac{L1}{R_L}$$

While the inductor does overcome the large peak current phenomenon of the RC filter, there are three important disadvantages associated with the RL filter.

- Since the current cannot change instantaneously in an inductor, a sudden change in the load (R_L) will cause an abrupt change in the output voltage. This phenomenon is the limiting factor that determines the transient response of a switching regulator.
- The energy stored in an inductor is determined from the equation

$$e = \frac{1}{2} LI^2.$$

Since the energy changes with the square of the current, the inductor must be very large to provide constant current flow when the load current is small.

- The disruption of current in Q1 (shutting Q1 off) causes the magnetic field associated with L1 to collapse and induce a potential in accordance with Lenz' law:

$$e_L = -L \frac{di}{dt}$$

This negative voltage places a very large voltage across transistor switch Q1 and will probably result in its destruction.

$$V_{CE(off)} = E_{in} + |e_L|$$

The RLC Filter

Combining the RC and RL filters gives all of the advantages of both, with few of the disadvantages. *Figure 4* depicts the RLC filter in the simple switching regulator. The inductor L1 is used to limit the peak currents associated with the charging of capacitor C1. This current will be highest during the initial turn-on of the power supply with all initial conditions set to zero. This circuit is shown in *Figure 5.* The peak current then is again derived by Laplace transforms from Kirchhoff's Voltage Law.

$$\frac{E}{S} = I_S R_S + I_S L S + I_S R_L // \frac{1}{CS}$$

$$\frac{E}{S} = I_s \left[R_s + L_s + \frac{\frac{R_L}{CS}}{R_L + \frac{1}{CS}} \right]$$

$$I_S = \left[\frac{E}{S}\right] \times \left[\frac{R_L CS + 1}{R_L LCS^2 + (R_S R_L C + L) S + R_S + R_L}\right]$$

$$I_S = \left[\frac{E}{S}\right]\left[\frac{RLC}{R_L LC}\right] \times \left[\frac{S + \frac{1}{R_L C}}{S^2 + S\frac{(R_s R_L C + L)}{R_L LC} + \frac{R_s R_L}{R_L LC}}\right]$$

$$I_S = \left[\frac{E}{L}\right]\left[\frac{I}{S}\right] \times \left[\frac{S + \frac{1}{R_L C}}{S^2 + S\frac{(R_s R_L C + L)}{R_L LC} + \frac{R_s R_L}{R_L LC}}\right]$$

Resulting in the form:

$$f_{(s)} = \frac{S + d}{S(S - a)(S - b)}$$

Yielding the general equation: $I = Ae^{at} + Be^{bt} + K$

$$A = \frac{a + d}{a(a - b)}$$

$$B = \frac{b + d}{b(b - a)}$$

$$K = \frac{d}{ab}$$

This formula is used in the section analyzing a typical switching regulator. A typical curve for I_s is shown in *Figure 6.*

Under light load conditions, the capacitor C1 supplies the necessary current to the load in accordance with the equation $e = \frac{1}{2} CE^2$. Under heavy load conditions, the energy stored in L1 supplies the current in accordance with the equation $e = \frac{1}{2} LI^2$. The energy stored in the magnetic field that resulted in the negative induced voltage can now be applied as an advantage. As shown in *Figure 4*, diode D1 steers the current developed by the collapsing magnetic field and charges capacitor C1 during the off-time; D1 also acts as a clamp and limits the negative potential to one diode drop. Diode D1 is called a steering diode, commutating diode or free wheeling diode. This circuit not only protects switch Q1, but also uses the energy stored in the magnetic field to charge capacitor C1; thus, $LI^2 = CE^2$.

Optimization of the RLC filter requires examination of the current loop equation for the RLC filter. *Figure 5* depicts the RLC circuit used for the analysis.

$$I = Ae^{s_1 t} - Ae^{s_2 t}; \quad A = \frac{V_o}{2L}$$

$$S_1 = \frac{-R}{2L} + \sqrt{\frac{R^2}{4L^2} - \frac{1}{LC}}$$

$$S_2 = \frac{-R}{2L} + \sqrt{\frac{R^2}{4L^2} - \frac{1}{LC}}$$

Fig. 4 Basic Switching Regulator with RLC Filter

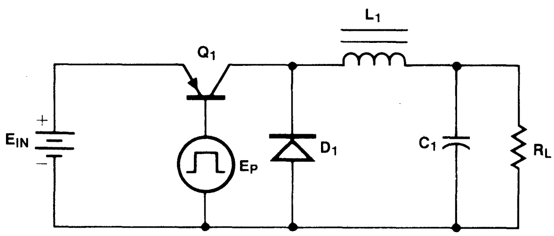


Fig. 5 Equivalent Circuit (Turn-On)

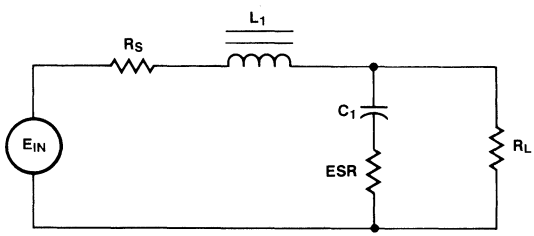


Fig. 6 Turn-On Peak Current

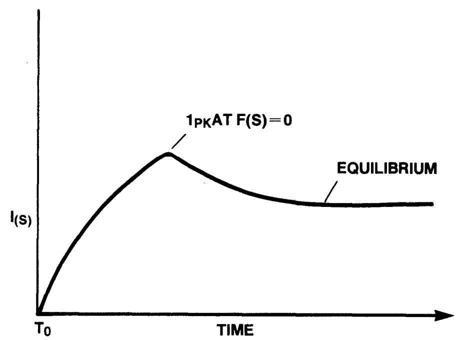
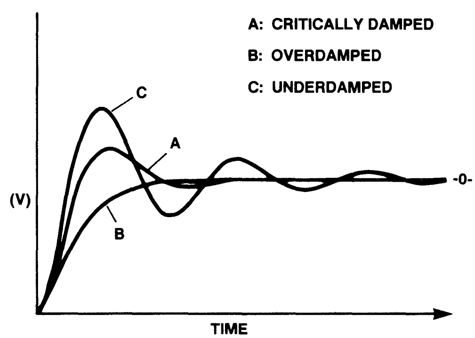


Fig. 7 Frequency Response Curve of an RLC Filter



Examining the roots of the equation shows that three special conditions exist:

Underdamped case:

$$\frac{R^2}{4L^2} < \frac{1}{LC}; \quad \sqrt{\frac{L}{C}} < 0.5\ R$$

The solution is complex and is exhibited as an oscillatory condition. This condition is undesirable due to the associated losses, i.e., energy in the ringing, and the RFI produced. See *Figure 7.*

Overdamped case:

$$\frac{R^2}{4L^2} > \frac{1}{LC}; \quad \sqrt{\frac{L}{C}} > 0.5\ R$$

The solution for these roots is real. This condition is also undesirable in the extreme case due to its associated losses. See *Figure 7.*

Critically damped case:

$$\frac{R^2}{4L^2} = \frac{1}{LC}; \quad \sqrt{\frac{L}{C}} = 0.5\ R$$

The equation has a real solution and is the most desirable case since losses are at a minimum. However, since this is not a practical case state, the circuit is operated in a slightly overdamped condition. See *Figure 7.*

During the off-condition of switch Q1, the circuit becomes the dual of *Figure 5.* This too has three similar conditions:

Underdamped case $\sqrt{\frac{C}{L}} > \frac{0.5}{R}$

Overdamped case $\sqrt{\frac{C}{L}} < \frac{0.5}{R}$

Critically damped case $\sqrt{\frac{C}{L}} = \frac{0.5}{R}$

The waveforms for the dual circuit are the same as those in *Figure 7.* Thus, to insure that for both on and off conditions of switch Q1, both circuits are slightly overdamped.

$$\sqrt{\frac{L}{C}} < 0.5\ R_s \text{ (on-condition)}$$

$$\sqrt{\frac{C}{L}} < 0.5\ R_L \text{ (off-condition)}$$

In the on-condition, R_s is a fixed quantity and should be made small to minimize IR losses. Thus, a snubber network may be required to dampen any oscillations associated with a small R_s.

In the off-condition, R_L is variable, with the worst case occurring during light loads. This can be alleviated with a minimum R_L or a snubber network as used in the on-condition. As aforementioned, meeting these inequalities enhances both the RFI characteristics and the possibility of parasitic oscillations.

Overshoot and Undershoot
When the load is abruptly changed (i.e., load current), the output voltage changes accordingly. This is called overshoot for decreasing loads and undershoot for increasing loads. Expressions for overshoot and undershoot can be derived from the two equations:

$$-e_L = L\,\frac{di}{dt} \qquad dt = \frac{L\,di}{e_L}$$

where

$e_L = E_{in} - E_{out}$ for increasing loads
$e_L = E_{out}$ for decreasing loads
$di = i_c =$ change in load current $= \triangle I$
$t =$ transient time
$dv =$ overshoot/undershoot voltage

$$i_C = C\,\frac{dv}{dt}, \quad i_C = di$$

$$dt = C\,\frac{dv}{di}$$

$$\frac{Ldi}{e_L} = C\,\frac{dv}{di}$$

$$dv = \frac{Ldi^2}{C\,e_L}$$

$$\triangle E_{out} = \frac{L\,\triangle I^2}{C\,(E_{in} - E_{out})} \text{ for increasing loads}$$

$$\triangle E_{out} = \frac{L\,\triangle I^2}{C\,E_{out}} \qquad \text{for decreasing loads}$$

Transient Response

The transient response, as mentioned earlier, is limited by the size of inductor L1. This transient response time t_R is the time necessary before the system can compensate for an abrupt change in the load, assuming zero loop response. Transient response time can be found from the equation:

$$-e_L = L\frac{di}{dt}$$

$$t_R = \frac{2L \triangle I}{E_{in} - E_{out}} \quad \text{for increasing loads}$$

$$t_R = \frac{2L \triangle I}{E_{out}} \quad \text{for decreasing loads}$$

The Inductor

The inductor is perhaps the least understood of the switching regulator components and yet one of the most important. There are seven major areas with tradeoffs to be considered.

1. Energy storage for the regulator
2. Peak current limiting in Q1
3. Output ripple
4. Transient response
5. Overshoot
6. Size and cost limits
7. Radiated electric and magnetic fields

As inductance is increased, items 1 through 3 are enhanced. Item 1, the energy ($e = \frac{1}{2}$ $L1^2$) is directly proportional to the inductance. Item 2, the peak current,

$$\left[I_{pk} = \left(\frac{E_{in} - E_{out}}{2L} \right) t_{on} \right]$$

is inversely proportional to the inductance. Item 3, the ripple voltage

$$\left[E_{RIPPLE} = \frac{E_{in} - E_{out}}{4\pi^2 f^2 LC} \right]$$

is also inversely proportional to the inductance.

However, as the increase in inductance enhances operation of items 1 through 3, it is detrimental to items 4 through 6. In item 4, the transient response

$$\left[t_R = \frac{2 \triangle I}{E_{in} - E_{out}} \right]$$

is directly proportional to the inductance. Item 5, the circuit overshoot

$$\left[\triangle E_{out} = \frac{L \triangle I^2}{C (E_{in} - E_{out})} \right]$$

is directly proportional to the inductance. Item 6, the size and cost are directly effected by the inductance as well as a host of other factors.

Item 7, the effect of the inductor on radiated electrical and magnetic noise is a function of geometry, frequency, size and cost. It becomes apparent that selecting the inductor requires careful consideration of the aforementioned tradeoffs. Applications of these tradeoffs are considered in the analysis of a typical switching regulator.

Inductor design has many philosophies associated with it. Size constraints are radiated electrical and magnetic fields may dictate a powder toroid or pot core; however, in most applications (computer and peripherals), the decision is left to the design philosophy. The three most common techniques employed in the industry are:

- Powdered permalloy toroids
- Ferrite EI, U and toroid cores
- Silicon steel EI butt stacks

The first technique, the powdered permalloy toroid, yields perhaps the most stable and predictable inductor. Powdered permalloy toroids have low leakage inductance, high permeability and low core losses. The major disadvantage is the cost of manufacturing and mounting toroid inductors.

The ferrite EI, U and toroid cores exhibit low losses. The ferrite toroid has low leakage inductance but is as expensive to manufacture as its powdered permalloy counterpart. All ferrite cores have low permeability, poor high temperature performance and the expense in mounting. The silicon steel EI butt stack offers one of the best tradeoffs in low voltage switching regulators. The silicon steel laminations exhibit high permeability, high flux densities, ease of construction and mounting. Core losses, while higher than the powdered permalloy and ferrite cores, are usually insignificant at low voltage levels. The silicon steel lamination is a common material in most magnetic houses and often can be found on the shelf.

Inductor Design

Combining Faraday's Law and Lenz' Law yields:

$$E = N \frac{d\phi}{dt} \times 10^{-8} = L \frac{di}{dt}$$

$$\int N \frac{d\phi}{dt} \times 10^{-8}$$

$$= \int L \frac{di}{dt} = LI = BAeN \times 10^{-8}$$

Multiplying both sides by $\frac{1}{2}$ yields:

$$\frac{1}{2} L I^2 = \frac{B\,Ae\,NI \times 10^{-8}}{2}$$

which is the energy stored in the inductor. Integrating Faraday's Law $E = N \frac{d\phi}{dt}$ yields the formula for ac flux in the core.

$$B_{ac} = \frac{3.49\,E \times 10^6}{f\,A_C\,N}\ \text{Gauss}$$

The dc flux is found from the equation:

$$B_{dc} = \frac{0.6\,NI_{dc}}{l_g}\ \text{Gauss}$$

The magnetizing from Ampere's Law is found from the equation:

$$H_{dc} = 0.4\,NI_{dc}$$

The inductance is found from the equation:

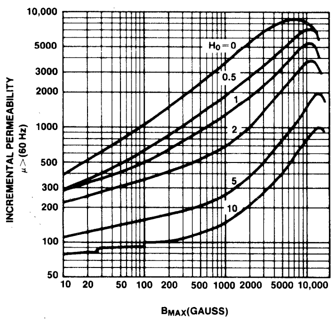
$$L = \frac{3.19 N^2 A_c \times 10^{-8}}{l_g + \frac{L_C}{\Delta\mu}}$$

Incremental permeability can be found from manufacturers' data sheets as shown in *Figure 8.* Linearity of the inductor can be enhanced by making l_g large.

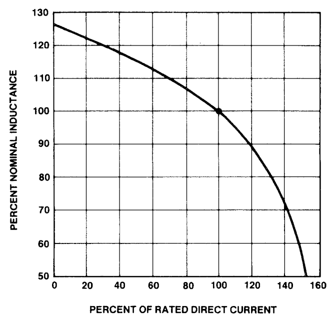
Units:
A_C = cross sectional area (in.^{2})
B_{ac} = ac flux (Gauss)
B_{dc} = dc flux (Gauss)
f = frequency (Hz)
H_{dc} = magnetizing force (Oersteds)
l_c = mean length of core (in.)
l_g = gap (in.)
N = number of turns
$\Delta\mu$ = incremental permeability

There are several off-the-shelf inductors manufactured by Sprague called Soft Inductors. The Soft Inductor is designed specifically for switching regulators, with a special variable reluctance gap.

Figure 8



a. Incremental permeability curve for AISI grade M-22 laminations where H_o is the dc magnetizing force in core.



b. Effect of dc in a typical filter choke. Inductance drops linearly until rated dc is flowing through coil, then drops rapidly as core saturates. The linear portion of the curve has less slope for inductors that have larger air gaps.

The Output Capacitor

Selection of the output capacitor also requires care. Consideration must be given to both the ESL and the ESR. Very often the ESR contributes more to ripple and noise than its reactance does. Desirable characteristics can be achieved by carefully paralleling three or four different types of capacitors such as tantalum, electrolytic and ceramic capacitors. Capacitors especially developed for switching regulators are now available in a multitude of ranges, sizes and types, with low ESR and low ESL at the switching frequencies. A curve of the 4-terminal UFT capacitor (manufactured by Cornell Dubilier) compared to a conventional electrolytic is shown in *Figure 9.* The curve plots impedance versus frequency. The UFT capacitor remains quite flat beyond 1 MHz. The UPT capacitor, also manufactured by Cornell Dubilier, is designed for switching regulators and gives one of the best performance/cost tradeoffs available. A simplified equivalent circuit is shown in *Figure 10.*

Figure 9

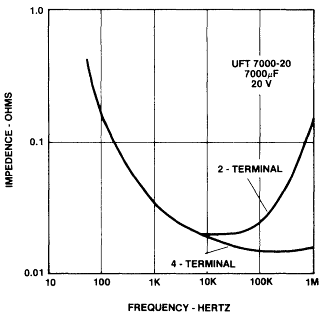
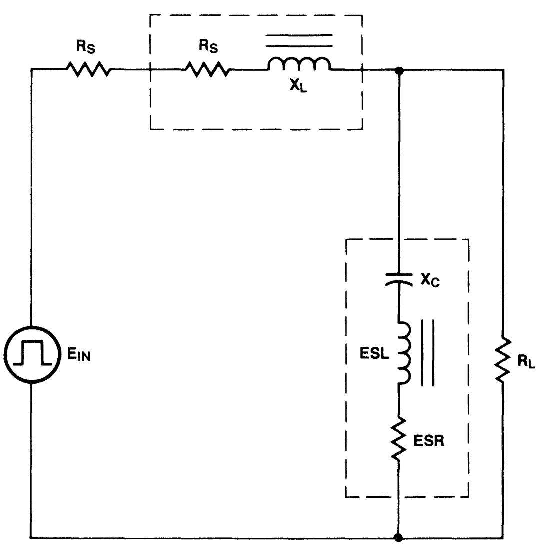


Fig. 10 Simplified Equivalent Circuit of RLC Filter



$$Z1 = Rs + rs + jX_L$$

$$Z2 = ESR + jX_{ESL} - jX_c$$

$$Z3 = R_Lj_o$$

$$\gamma = \frac{Z2//Z3}{Z1 + Z2//Z3}$$

$$\gamma \approx \frac{ESR + j(X_{ESL} - X_C)}{(R_S + r_s + ESR) + j(X_2 + X_{ESL} - X_C)}$$

$$E_{RIPPLE} =$$

$$E_{in}\left[\frac{ESR + j(X_{ESL} - X_C)}{(R_s + r_s + ESR) + j(X_C + X_{ESL} - X_C)}\right]$$

$$E_{RIPPLE} = \gamma\, E_{in}$$

The formula: $C1 = \dfrac{E_{in} - E_{out}}{4\,\pi^2 f^2 L E_{RIPPLE}}$

is a good approximation for finding the minimum capacitance; however, the preceding formula must be used to accurately determine the ripple voltage.

Closed Loop

In order for the switching regulator to maintain an output voltage relatively constant, some feedback mechanism must be employed. *Figure 11* shows a typical feedback system.

- K3 represents the power switch, filter and all the associated losses.
- K2 represents the transfer function for the pulse generator.
- K1 is the open loop gain of the error amplifier.
- β is the attenuation factor usually determined by a simple voltage divider.
- Σ is a summing network that produces an error voltage $\triangle e$ from the difference between the reference voltage E_{REF} and the feedback voltage E_{fb}.

$$\triangle \epsilon = E_{REF} - \triangle E_{fb}$$

The total loop gain will determine the percentage regulation of the switching regulator.

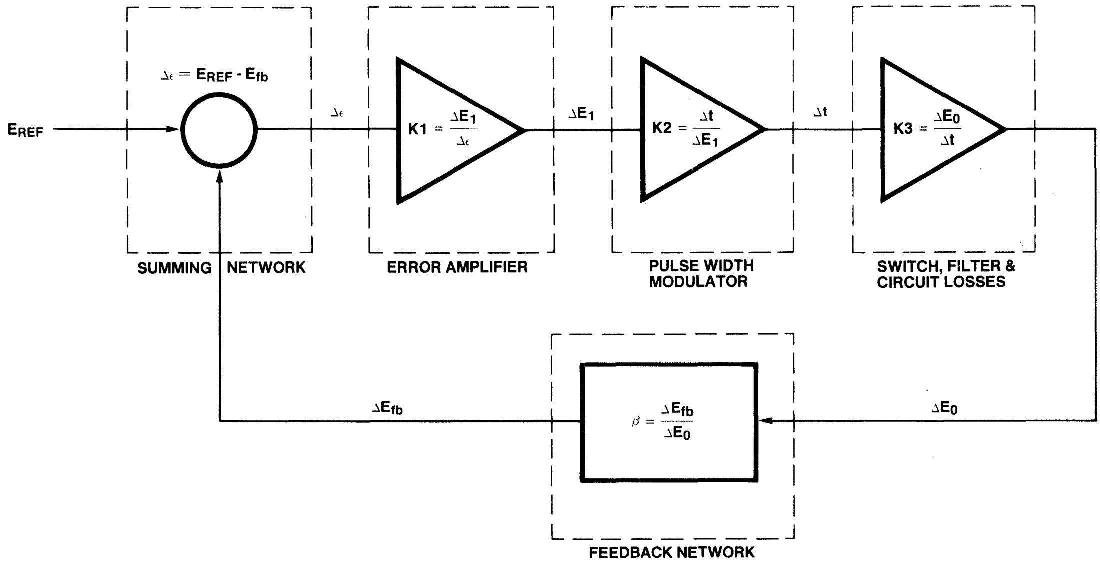
$$A = \frac{\triangle E_{out}}{\triangle \epsilon} = \%\ \text{regulation}$$

The error amplifier gain then is defined by:

$$K1 = \frac{A - 1}{\beta K2,\ K3}$$

While this model is an approximation, it yields relatively close results.

Fig. 11 Typical Switching Regulator Feedback Loop



A Switching Regulator Using the SH1605

The SH1605 is a hybrid integrated circuit, designed specifically to be used as a major building block in high-current, step-down switching regulator systems. It contains a temperature-compensated voltage reference, comparator, oscillator, high-current Darlington and high-power steering diode. This device is capable of supplying up to 5 A continuous current; its package dissipation capability is 20 W maximum. This circuit provides excellent performance, with efficiencies up to 85%, for applications requiring high power densities and large operating currents.

Switching Regulator Theory

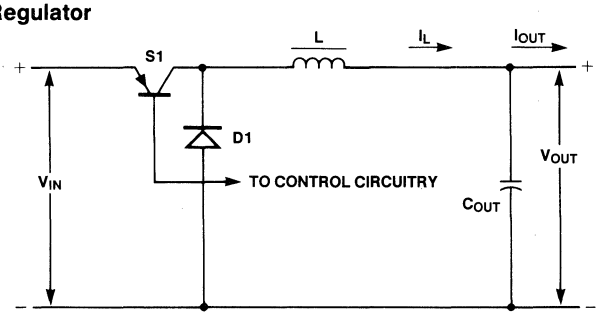
Figure 12 shows the basic switching regulator configuration. This circuit provides an output voltage V_{OUT}, related to the input voltage V_{IN} by the duty cycle of switch S1. Thus:

$$V_{OUT} = V_{IN}\left(\frac{t_{on}}{t_{on} + t_{off}}\right)$$

Therefore, a switching regulator maintains a constant output voltage against variations in input by appropriately modifying the system duty cycle. The basic switching regulator operates as follows. Control transistor S1 switches on when first energized, thereby applying a voltage approximately equal to the input across L1 and C_0. This causes current I_1 to increase linearly with time, supplying current to the load while storing energy in L1. Diode D1 insures that, when S1 switches off, current continues to flow to the load thereby achieving a continuous load-current flow.

At equilibrium, the average current through L1 is equal to the load current. The rate of current change

Fig. 12 Basic Switching Regulator



through the inductor, $\triangle I_1$, during the on and off period is defined by *Equations 1* and *2* below.

$$\triangle I_1 = \left(\frac{V_{IN} - V_{OUT}}{L1}\right) t_{on} \qquad (1)$$

$$\triangle I_1 = \left(\frac{V_{OUT}}{L1}\right) t_{off} \qquad (2)$$

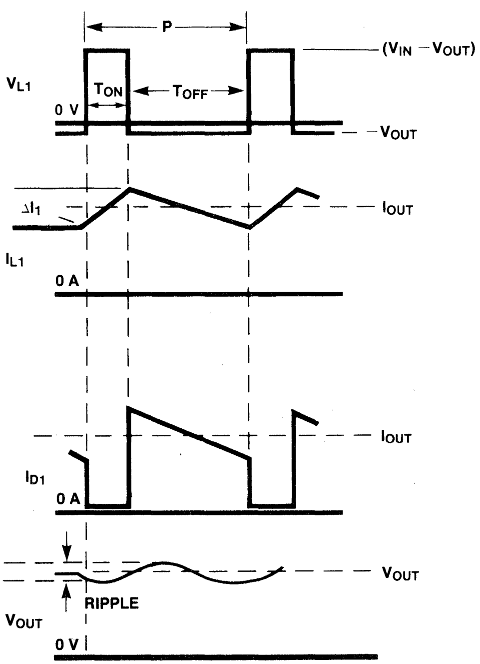
Since, in a conventional switching regulator the excursions $I_{1(on)}$ and $1_{1(off)}$ are equal, *Equations 1* and *2* can be written as follows:

$$\frac{V_{IN}}{V_{OUT}} - 1 = \frac{t_{off}}{t_{on}} \qquad (3)$$

Equation 3 shows the natural tendency for the on-to-off time ratios to remain proportional to the input-output voltage differential. Voltage regulation can be achieved when this information is properly fed back to the switch.

Since the duty cycle is dependent only upon the magnitude of the input-to-output voltage differential, it follows that variations of output voltage with load should be minimal. Basic switching regulator waveforms are shown in *Figure 13.*

Fig. 13 Typical Switching Regulator Waveforms



SH1605 Theory of Operation

The SH1605 simplified block diagram is shown in *Figure 14.* Circuit operation is as follows. When power is first applied, the output voltage V_{OUT} is low, thus forcing the comparator output into a HIGH state. As a result, the oscillator freely toggles the output switch on and off at a rate determined by the charge and discharge rate of the timing capacitor C_T. This is a temporary condition that continues until V_{OUT} has exceeded the reference voltage level times the factor set by R_S, R2 and L_{RS}. The output voltage can be expressed as follows.

$$V_{OUT} = V_{REF} \frac{(R_S + R2 + Rs)}{(R1 + R2)} \qquad (4)$$

Since the value of R1 and R2 (1 kΩ each) inside the SH1605 is established, R_S can be determined as follows.

$$Rs = \frac{(2 \times 10^3)(V_{OUT} - 2.5)}{2.5} \text{ for Rs in ohms} \qquad (5)$$

Equilibrium is reached at the completion of the on cycle when the comparator input has exceeded the reference level. When the comparator output goes LOW, the oscillator output is disabled and Q1 switches off. V_{OUT} then begins to fall at a rate determined by the ratio of the output voltage to the inductor value.

$$\frac{\triangle I_1}{t_{off}} = \frac{V_{OUT} + V_{D1}}{L1} \qquad (6)$$

Whenever V_{OUT} falls to the level specified in *Equation 4*, the comparator changes state and the output switches on. It remains in this state until the voltage across C_T reaches a positive threshold level. The rate of C_T charge is determined by the size of the timing capacitor and the magnitude of the constant current source inside the oscillator. Charging current is typically 25 μA and discharging current is 225 μA. From the equation describing on and off time duration, the frequency of oscillation can be deduced:

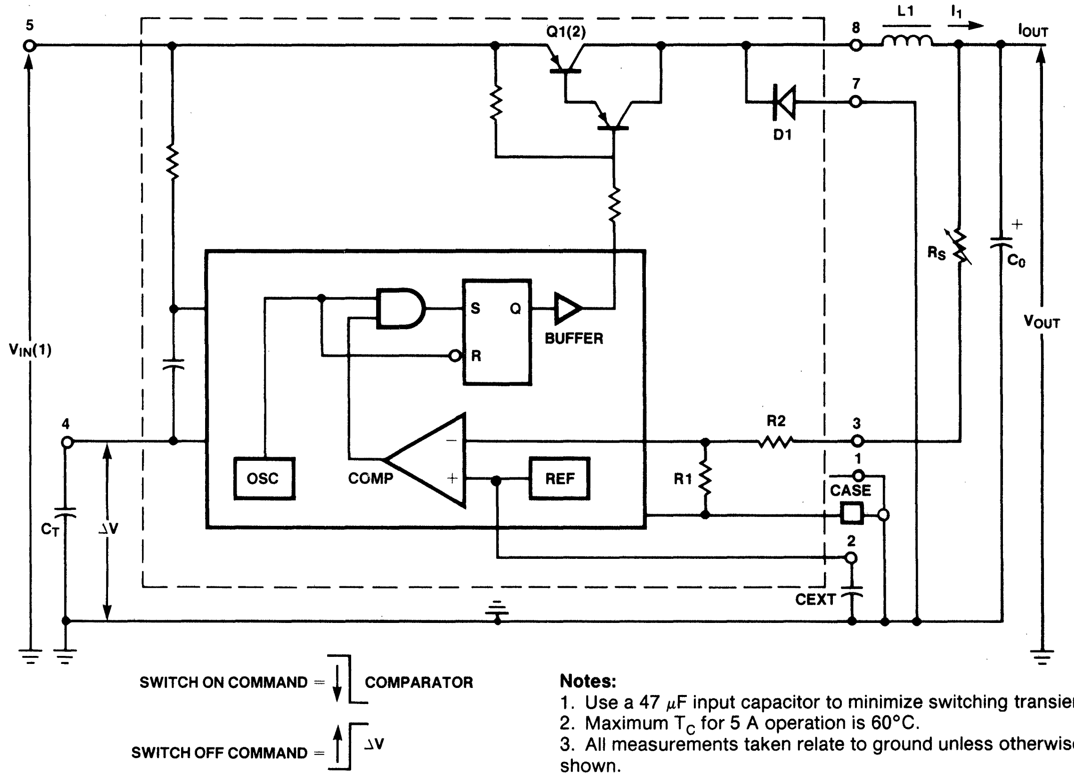
$$t_{on} = \frac{C_T \triangle v}{I_C}$$

$$t_{off} = \frac{\triangle I_1 L I}{V_{OUT} + V_F} \qquad (8)$$

where:

$L1$ = Filter Inductance
$\triangle I_1$ = Change in Inductor Current
V_{OUT} = Output Voltage
C_T = Timing capacitor
I_C = Oscillator charging current
$\triangle v$ = 0.5 V Typical
V_F = Steering Diode Forward Voltage Drop

Fig. 14 SH1605 Block Diagram



Notes:
1. Use a 47 μF input capacitor to minimize switching transients.
2. Maximum T_C for 5 A operation is 60°C.
3. All measurements taken relate to ground unless otherwise shown.

$$\text{Nominal Frequency} = \frac{1}{\dfrac{C_T \Delta V}{I_C} + \dfrac{\Delta I_{1(nom)} L1}{V_{OUT} + V_F}} \quad (9)$$

For improved system efficiency, the operating period should always be many times longer than the device transition times. A trade off must be sought between inductor size and efficiency when selecting the frequency of operation.

Design Example

Figure 15 is a typical design of a step-down switching regulator using the SH1605.

Nominal Design Objectives

$V_{OUT} = +5$ V	Line Regulation = 2%
$I_{OUT(max)} = 5.0$ A	Load Regulation = 2%
$I_{OUT(min)} = 1.0$ A	Ripple (max) = 0.1 $V_{pk\text{-}pk}$
$V_{IN} = 12$ to 18 V	Efficiency = 70%

First, R_S is calculated from *Equation 5:*

$$R_S = \frac{(2 \times 10^3)(V_{OUT} - 2.5)}{2.5} = 2\ \text{k}\Omega$$

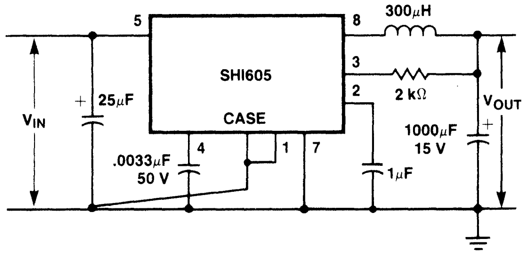
Since the required $I_{OUT(min)}$ is 1 A to maintain continuous operation, the peak-to-peak current excursion must be equal to 2 A or less, i.e.,

$$\Delta I_1 = 2\ I_{OUT(min)}$$

To calculate the value of the inductor keeping the efficiency/component-size tradeoff in mind use *Equation 1.* For this example $t_{on} = 60\ \mu$s is selected. t_{on} is determined by the designer and depends upon the desired frequency of operation under expected constant load conditions where frequency = $1/(t_{on} + t_{off})$. t_{on} must always be greater than ts + td = 5.1μs, typically, from the SH1605 data sheet. Off time, t_{off}, is determined by the ratio of input voltage to output voltage where

$$t_{off} = t_{on} \times \left(\frac{V_{IN}}{V_{OUT}} - 1\right)$$

Fig. 15 Design Example



Circuit Performance
V_{IN} = 12-18 V
V_{OUT} = 5.06 V

Load Reg. = 50 mV (1 A $\leq I_{OUT} \leq$ 5 A)
Line Reg. = 50 mV (12 V $\leq V_{IN} \leq$ 18 V)

Note
In this example the SH1605 must be mounted on a heat sink with a maximum thermal resistance of $\phi_{CA} > 4°$ C/W.

(*Equation 3*). Thus with a known ratio of V_{IN}/V_{OUT} the designer is offered a trade-off between frequency of operation, efficiency and component size.

From *Equation 1:*

$$L1 = \left(\frac{V_{IN(nom)} - V_{OUT}}{\triangle I_1}\right) t_{on}$$

$$= \frac{10}{2}(6 \times 10^{-5}) = 300\ \mu H$$

where $V_{IN(nom)} = 15$ V, $t_{on} = 60\ \mu s$

$\triangle I_1 = 2$ A

One very important element in achieving the optimum performance in a switching regulator is to insure the inductor is kept below the specified saturation limits.

Since the timing capacitor controls the 60 μs on time, C_T can be determied using *Equation 7:*

$$C_T = \frac{(t_{on})(I_C)}{\triangle v} = \frac{(6 \times 10^{-5})(2.5 \times 10^{-5})}{5 \times 10^{-1}} = 3000 \text{ pF}$$

where $I_C = 25\ \mu A$ nominal per data sheet.

The final step is to determine the requirements for the output capacitor C_O to obtain the desired value of ripple voltage. Consideration must be given to the absolute value of C_O as well as the internal effective series resistance (ESR). Since the capacitor size is inversely proportional to the operating frequency, the lowest frequency of operation must be calculated. Minimum operating frequency can be determined by using $\triangle I_{1(max)}$ vs $\triangle I_{1(nom)}$ in *Equation 9.*

$$\text{Minimum Frequency} = \frac{1}{\dfrac{C_T \triangle v}{I_c} + \dfrac{\triangle I_{1(max)} L_1}{V_{out} + V_F}}$$

$$= \frac{1}{1.7 \times 10^{-4}} = 5.9 \text{ kHz}$$

$$\text{Where: } \triangle I_{1(max)} = \left(\frac{V_{IN(max)} - V_{OUT}}{L_1}\right) \times t_{on}$$

$$= \left(\frac{18 - 5}{3 \times 10^{-4}}\right) \times 6 \times 10^{-5}$$

$$= 2.6 \text{ A}$$

From *Equation 1*

The output capacitor can now be determined as follows:

$$C_{O(min)} = \frac{\triangle I_1}{(8\, f_{(min)}\, V_{ripple(max)})}$$

$$= \frac{2}{(8 \times 5.9 \times 10^3) \times (1 \times 10^{-1})}$$

$$= 423 \mu F$$

The maximum acceptable ESR is therefore

$$ESR(max) = \frac{V_{ripple(max)}}{\triangle 1_{1(max)}}$$

$$= \frac{1 \times 10^{-1}}{2.6} = 0.038\Omega$$

5

Normally, the minimum capacitance value should be increased considerably if a low ESR capacitor is not used.

As a final step for minimizing switching transients at the device input, a low ESR capacitor must be used for decoupling purposes between the input terminal and ground.

Conclusion

The SH1605 is a highly versatile building block for high current, step-down switching regulator systems. However, to attain optimum performance and reliability the following guidelines should be followed:

- Keep operating period long, relative to the device switching times, for optimum efficiency
- Insure that the inductor stays out of saturation and minimize the series resistance.
- Use high quality capacitors for input and output to minimize ripple and noise.

Designer's Note

As an aid in designing with the SH1605, 5 Amp Switching Regulator, the following is a review of several characteristics of the device which should be recognized and understood by the designer.

Short Circuit Current Limit

Space limitations and the already high packing density attained in the SH1605 prevent the inclusion of a short circuit current limit in the product. For those occasions where short circuit protection is required (i.e., prototype designs and lab testing), a schematic for an external protection network is shown in *Figure 16.*

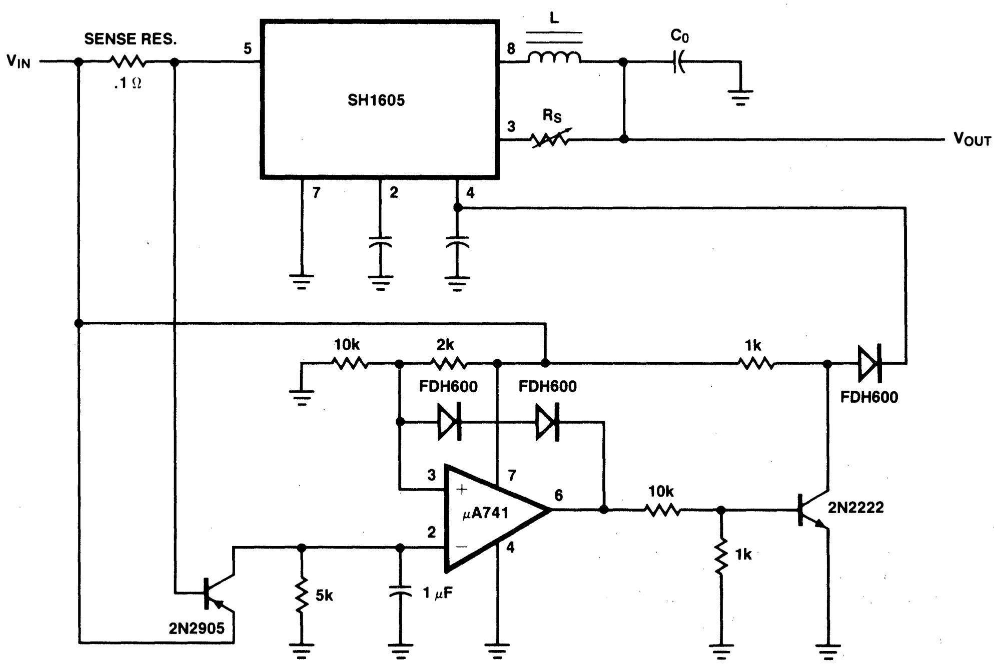
Heatsink Designs

While heatsinking is generally not a problem with the SH1605 due to its high efficiency, mounting of the package can have a dramatic effect on θ_{JA}. Cutting a large hole or curved slot around all eight leads leaves only the package fringes for heat transfer. A θ_{CS} thermal resistance of 4.0 to 4.5°C/W will result from this type of mounting. A θ_{CS} thermal resistance of 0.3 to 0.4°C/W can be obtained using a hole configuration similar to Thermalloy pattern 15 or IERC pattern LAIC, UP or HP and a good thermal conducting compound.

8 Pin TO-3 Sockets

Sockets are a definite convenience when prototyping, testing and even sometimes for small volume production runs. Standard sockets are commercially available from a number of manufacturers. For a partial list of suppliers refer to the SH1605 data sheet.

Figure 16. Switching Regulator with Short Circuit Protection



Grounding

Switching power supplies are by nature more susceptible to grounding problems than linear power supplies because of larger ripple currents. It is generally recommended that a ground plane be used. An ideal connection diagram to minimize grounding problems is shown in *Figure 17.*

A common problem encountered with the SH1605 is excessive noise, or ripple, which is almost always generated by improper grounding. Care must be taken in the design and layout of the breadboard to eliminate any possible ground loops. This is accomplished by observing very standard layout procedures. The following diagram explicitly illustrates where the ground connections must be made to avoid potential problems.

Pin 7, which is the anode of the steering diode and which carries up to 5 A of ripple, must be tied to "input ground" . . . not the case and not "output ground". An incorrect connection here accounts for at least 80% of the field problems. To further improve system performance, the negative sides of both the timing capacitor and the decoupling capacitor should be tied together at the case with a single lead going to "output ground" and the negative side of the output filter capacitor should be connected directly to "input ground." Note that there are two distinct grounding points in the system. "Input ground" is defined as the connection point between the negative side of the input filter capacitor and the incoming ground line. "Output ground" is a ground point as close to the load as possible. The input and output ground points are connected but distinctly separate thus minimizing system ground loops and their effect on output voltage regulation.

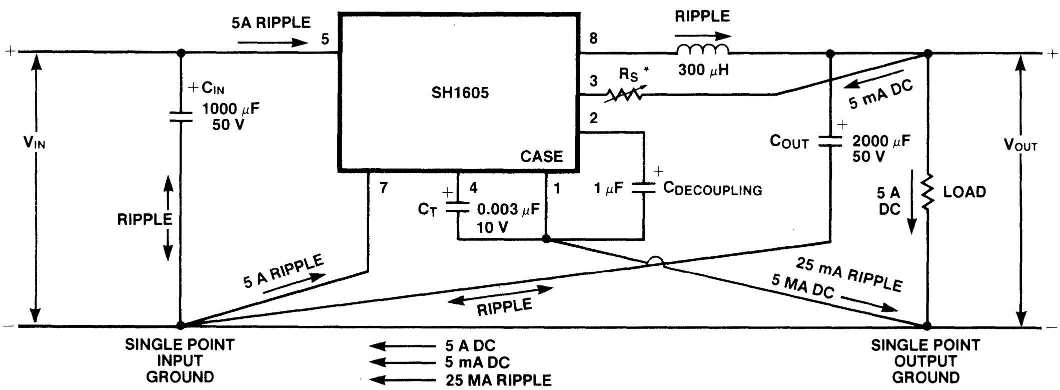
Frequency of Operation

The SH1605 is a frequency modulated switcher. Thus, frequency will vary somewhat during operation depending upon power demand. When frequency is designed to fall mostly within audio ranges, users may find the continuously varying tone an annoyance. It is, therefore, recommended that users either provide for sound insulation or design for frequencies outside the normal human audio range.

Although the SH1605 is capable of operating across a broad range of frequencies, it is recommended that the user design his system to operate between 20KHz and 30KHz for optimum efficiencies and performance.

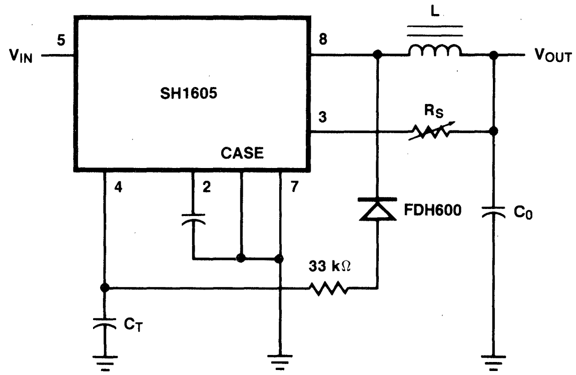
For convenience, a circuit for frequency locking is shown in *Figure 18.*

Figure 17. An Ideal Connection Diagram



*Metal Film Resistor or Temp. Coef. < 100ppm/°C

Figure 18. SH1605 Frequency Locking Network



Notes

1. Diode FDH600 in series with 33KΩ resistor are the frequency locking network which facilitates measurement and minimizes noise.
2. As input to output voltage ratio is increased, the operating frequency (fo), will decrease according to the expression shown below:

$$F(o) = \left(\frac{Vo}{Vin}\right)\left(\frac{1}{Ton}\right) \text{ where } Ton = \frac{C^T \triangle V}{I^C}$$

Power Supply Design

A power supply normally operates from an ac line. This ac input voltage must be converted to unregulated dc by some form of rectifier/filter combination and then to regulated dc using a voltage regulator. This chapter discusses the performance characteristics of the most common forms of rectifier/filter combinations and provides appropriate design equations for any output voltage and current.

Single Phase, Half Wave Rectifier
Figure 1 is a half wave rectifier and capacitor filter. Without the capacitor, peak current is

$$I_M = \frac{V_M}{R_S + R_L}$$

on the positive half cycle (or forward conduction cycle) of the input voltage. Some additional electrical characteristics follow.

$$I_{rms} = \frac{I_M}{\sqrt{2}} \qquad I_O = \frac{IM}{\pi} \qquad \gamma = 1.21$$

$$P_O = \frac{1}{\pi^2}\left(\frac{V_M{}^2 R_L}{(R_S+R_L)^2}\right)$$

$$\eta R = \frac{40.6}{\left(1 + \frac{R_S}{R_L}\right)} \%$$

Note that for a resistive load, the maximum ripple factor is 121% which, under most circumstances, requires filtering. When the capacitor is added across the load resistor, the ripple is reduced proportionate to the R_LC product *(Figure 2).*

One possible problem with any capacitive filter is the high peak current drawn due to the diode back-bias present throughout most of the input cycle. This is a result of the voltage stored across the filter capacitor. The rectifier conducts only during that short period of time when the input voltage exceeds the capacitor voltage by one diode drop. During conduction, the rectifier must supply the capacitor with sufficient energy to hold the ripple within specification until the next conduction cycle. *Figure 3* is a plot of the I_M/I_O ratio versus the R_LC product with the R_S/R_L ratio as a variable. Notice that the surge-to-dc ratio of current increases as a function of both increasing capacitor value and of a reduced source-to-load impedance ratio.

Definition of Terms

Parameter	Definition
V_M	peak input voltage
V_O	dc output voltage
V_{pk}	transformer peak voltage
V_S	ac input voltage
F	form factor of the load current: I_{rms}/I_O
I_{ac}	effective value of all alternating components of load current, *i.e.*, the current reading on an ac meter
I_M	peak current through each rectifier
I_O	average value of the load current, the reading on a dc meter
I_{rms}	effective value of the total load current $\sqrt{1_{ac}{}^2 + I_O{}^2}$
P_{in}	ac input power
P_O	dc output power
R_L	load resistance
R_S	total series resistance, or the source resistance plus any added resistance plus the diode series resistance
γ	ripple factor in all charts normalized as 100% equal to 1, $\gamma = (F^2-1) = \left[\left(\frac{I_{rms}}{I_O}\right)^2 - 1\right]^{1/2}$
ηR	rectification efficiency, $\frac{P_O}{P_{in}} \times 100\%$
ω	$2\pi f$ where f = line frequency

5

Fig. 1 Half-Wave Rectifier Circuit with Capacitive Filtering

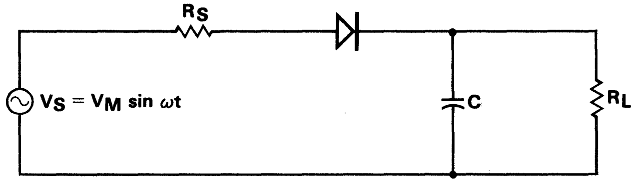


Fig. 2 Ripple Factor vs $\omega R_L C$

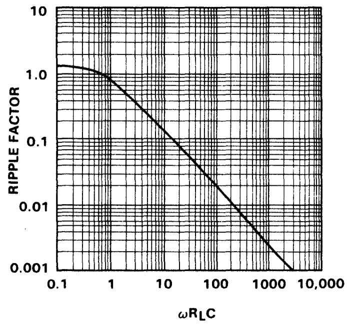


Fig. 3 I_M/I_O vs $\omega R_L C$

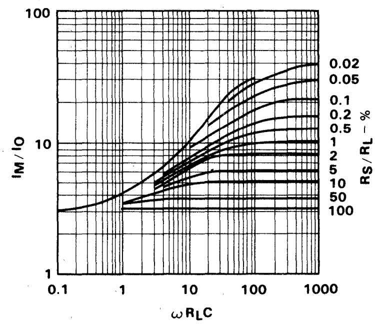


Fig. 4 DC-to-Peak Ratio

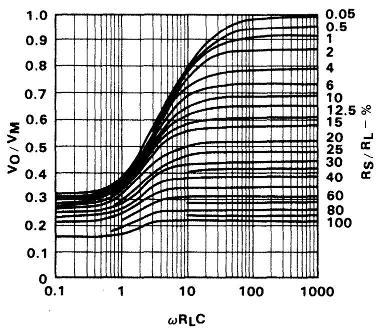
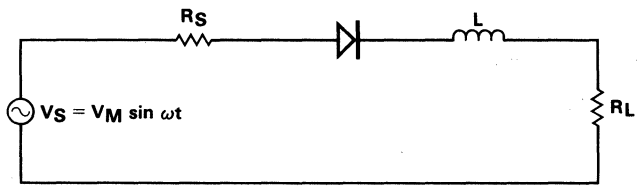


Fig. 5 Half-Wave Rectifier



When the ripple factor, load impedance, and ω are known, the required capacitance can be determined from *Figure 2*. Because of the high turn-on surge, an external series limiting resistor is normally needed. *Figure 4* is a plot of the dc-to-peak voltage ratio with the filter product as the X axis and the source/load impedance ratio as the third parameter. Note that the dc output-to-peak input voltage ratio approaches unity as the filter factor goes up and also as the source-to-load impedance ratio decreases. Because of the relatively large value of the filter capacitor required for a given ripple factor, the use of the half-wave capacitor filter is usually limited to low current applications.

Half Wave Rectifier With Series Inductive Filter

Figure 5 is a half-wave rectifier with series inductive filtering. The inductor, in series with the load, prevents any rapid changes in the current flow and thus reduces the ripple factor by acting as an energy storage device. When the current flow is above the average current required, energy is stored in the inductor, and when the current is below the average, the stored energy is released. *Figure 6* is the plot of ripple factor versus filter product for the inductor input filter. Because of the energy storage available with an inductor, the peak current through the rectifier is little more than the average current. However, the peak inverse voltage PIV seen by the rectifier is simply V_M, the peak input voltage. *Figure 7* is a plot of V_O/V_M ratio as a function of the inductor filter product.

Fig. 6 Ripple Factor vs Filter Product

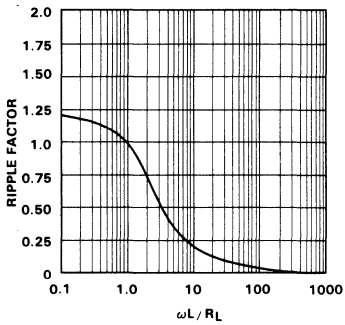
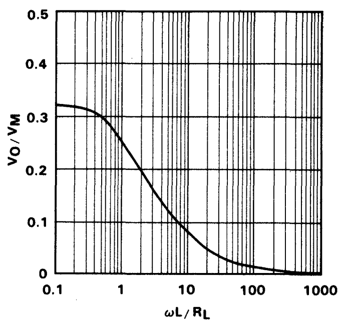


Fig. 7 V_O/V_M Ratio



Single-Phase Full-Wave Rectifier

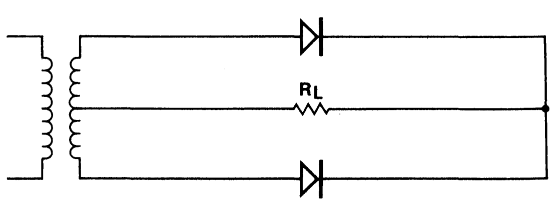
Figure 8, a basic full wave rectifier, has the following electrical characteristics.

$$I_{rms} = \frac{I_M}{\sqrt{2}} \qquad I_O = \frac{2I_M}{\pi} \qquad \gamma = 0.48$$

$$P_O = \left(\frac{2}{\pi}\right)^2 \frac{V_M{}^2 R_L}{(R_S + R_L)^2}$$

$$\eta R = \frac{81.2}{\left(1 + \frac{R_S}{R_L}\right)} \%$$

Fig. 8 Basic Full-Wave Rectifier



There are two interesting features. Efficiency has doubled, as can be expected when doubling the number of rectifiers. In addition, the ripple factor has decreased from 121% to 48% in comparison with the half-wave circuit. Even with ripple reduction, a 48% factor is normally too high to be useful and must be filtered. *Figure 9* is the filter product plot for both capacitive and inductive filters, assuming $R_S << R_L$. High peak currrents are always associated with capacitive filters and *Figure 10* plots the ratio of peak-to-dc current as a function of the filter product. The relationship between the filter product, the R_S/R_L ratio and the dc output-to-peak input voltage is given in *Figure 11* for the capacitive input filter. Load regulation may also be determined from *Figure 11* by using the high and low limits for R_L.

Fig. 9 Filter Product

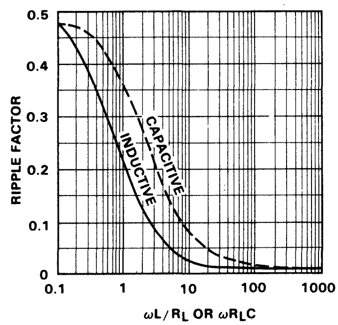


Fig. 10 Peak-to-DC Ratio

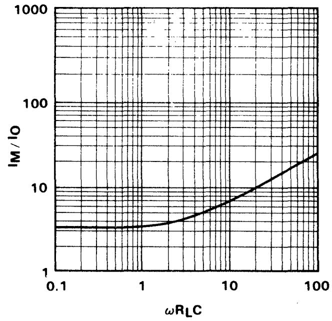
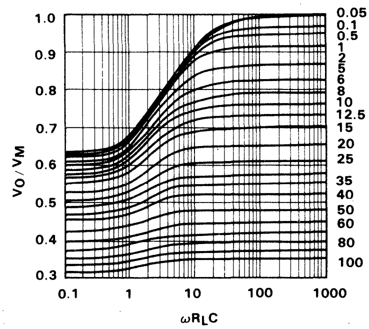


Fig. 11 Load Regulation



Design Example

For a full-wave circuit with the following requirements,

$V_O = 20$ V

$I_O = 1$ A

$\gamma < 0.1$

with $R_S = 1\ \Omega$

proceed with the following steps.

Step 1 Find the filter product from *Figure 9*

for $\gamma < 0.1 (\omega R_L C = 10)$

Step 2 Calculate R_L

$$R_L = \frac{20}{1} = 20\ \Omega$$

Step 3 Calculate C

$$C = \frac{10}{\omega R_L} = \frac{10}{120 \times 20\pi}$$

$$= \frac{1}{240\pi} = 1300 \mu F$$

Step 4 Calculate $\frac{R_S}{R_L}$

$$\frac{R_S}{R_L} = \frac{1}{20} = 5\%$$

Step 5 Find the transformer peak input voltage from the following.

V_{pk} = diode forward voltage. One diode forward-voltage drop for a center-tapped full-wave input, two diode forward-voltage drops for a full-wave bridge

$$+ \frac{V_O}{V_O/V_M}$$

using the filter values from *Figure 11.*

$V_{pk} = 0.7 + \frac{.20}{0.82}$ (intersection of

$5\% \frac{R_S}{R_L}$ and $\omega R_L C = 10$ from *Figure 11.*)

$V_{pk} = 0.7 + 25.3 =$ 26 V peak or
52 V pk-pk or
18.6 V_{rms}

Step 6 Check peak diode current from *Figure 10.* For this example at a filter product of 10, the peak current is seven times the dc current, or 7 A.

LC Section Filter

The LC section filter is one method of reducing ripple levels without the need for single, large value filter omponents. The basic circuit is shown in *Figure 12.* As a general rule, the capacitive reactance should always be less than 10% of the load resistance at the second harmonic of the incoming frequency. All the succeeding information is based upon this ratio. The ripple factor for an L-section filter has the form:

$$\gamma = \frac{0.47}{4\omega^2 LC - 1}$$

or, if n L-section filters are cascaded, then the ripple factor is:

$$\gamma = \frac{0.47}{(4\omega^2 L1C1 - 1)(4\omega^2 L2C2 - 1) - (4\omega^2 L_n C_n - 1)}$$

Fig. 12 LC Filter

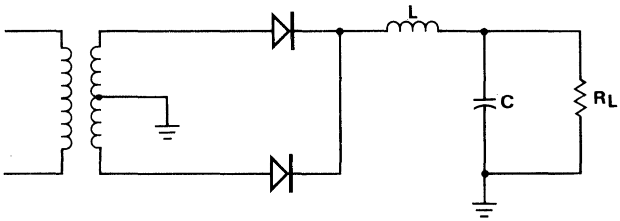


Figure 13 is a plot of the filter factor versus the ω^2LC product. The one additional requirement is continuous current flow through the inductance. This says, in effect, that there is a critical inductor size. To assure this continuous current flow, a bleeder resistor R_K must be used at the filter output. The critical value of inductance is

$$L_C = \frac{R_S + R_{eff}}{3\omega}$$

where $R_{eff} = \dfrac{R_K R_{L(max)}}{R_K + R_{L(max)}}$ and $R_K = \dfrac{V_O}{I_K}$

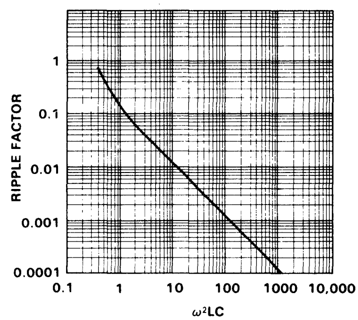
Bleeder current I_K may be assumed to be 10% of minimum load current or, if this is not a practical value, then some reasonable minimum bleeder current is selected. Once the critical inductance is found, then the capacitor value may be determined by the following steps: Set $L = 2\ L_C$. Determine ω^2LC from *Figure 13* for the required ripple factor. Solve for C from $\omega^2LC = X$, where X is the product from *Figure 13.*

The peak rectifier currents depend upon the size of the inductor selected such that if $L = L_C$ then $I_M = 2\ I_L$ and if $L = 2\ L_C$ then $I_M = 1.5\ I_L$. The transformer secondary voltage is given by

$$V_{rms} = 1.11 \left[V_O + R_S\left(I_{L(max)}\ I_K\right) \right]$$

and the minimum PIV for the rectifier is $1.57\ V_{O(max)}$ for a full-wave bridge rectifier.

Fig. 13 Filter Factor vs ω^2LC



For minimum power dissipation, R_K should be as large as possible. In some cases, since the value of critical inductance is proportional to the value of the bleeder resistor, the selection of a high value results in an inductance too large to be practical. In this case, a swinging choke or a choke whose inductance decreases with increasing current flow is needed.

5

Design Example
Full wave, single-section, choke input filter design,

$V_O = 50\text{ V}$ $I_K = 100\text{ mA}$ $\gamma = 1\%$

$I_O = 1\text{ A}$ $R_S = 10\,\Omega$

Step 1 Calculate R_K

$$R_K = \frac{V_O}{I_K} = \frac{50\text{ V}}{100\text{ mA}} = 500\,\Omega$$

Step 2 Calculate R_{eff}

$$R_{eff} = \frac{R_K R_{L(max)}}{R_K + R_{L(max)}} = 500\,\Omega\ (R_{L(max)} = \infty)$$

Step 3 Calculate L_C

$$L_C = \frac{R_{eff} + R_S}{3\omega} = \frac{500 + 10}{1130}$$

$$= \frac{510}{1130} \cong 0.5\text{ H}$$

Step 4 Calculate C

$\gamma = 0.01$

then, $\omega^2 L_C C = 12$ from *Figure 13*

$$C = \frac{12}{\omega^2 L_C} = \frac{12}{(120\pi)^2 0.5} = \frac{2}{142 \times 10^3 \times 0.5}$$

$$= 0.169 \times 10^{-3} = 169\,\mu\text{F}$$

Step 5 Calculate I_M

Since $L = L_C$

then $I_M = 2\,(I_O + I_K)$

$I_M = 2 \times 1.1 = 2.2\text{ A}$

Step 6 Calculate voltage drop both at no load and full load

V_D no load $= I_K\,(R_S) = 0.1 \times 10 = 1\text{ V}$

V_D full load $= (I_O + I_K)\,R_S = 1.1 \times 10 = 11\text{ V}$

Step 7 Calculate transformer minimum rms voltages

$V_{rms} = 1.11\,[V_O + R_S\,(I_{O(max)} + I_K)]$

$V_{rms} = 1.11\,(50 + 10 \times 1.1)$

$V_{rms} = 1.11\,(61)$

$V_{rms} = 67.5\ V_{rms}$

Step 8 Calculate maximum output voltage

$$V_{O(max)} = \frac{V_{rms}}{1.11} - I_K R_S$$

$$V_{O(max)} = \frac{67.5}{1.11} - 0.1 \times 10 = 61 - 1 = 60\text{ Vdc}$$

Step 9 Calculate PIV rating required

$PIV = (1.57)\,V_{O(max)}$ (See Table 4-1)

$PIV = 1.57 \times 60 = 94\text{ V}$

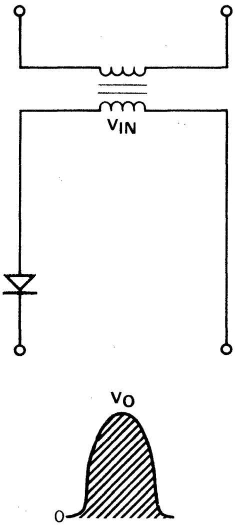
Transformer ratios are determined from *Table 1.*

Table 1. Electrical Reference Table and Rectifier Circuit Wave Shapes

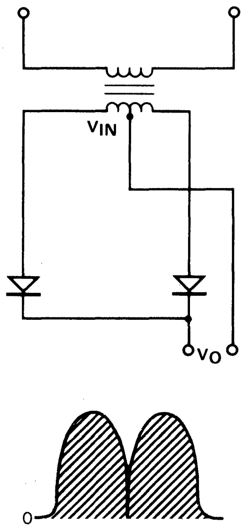
Characteristic	Load	Single Phase Half Wave (See 1-A)	Single Phase Full Wave Center-Tap (See 1-B)	Single Phase Full Wave Bridge (See 1-C)	Three Phase Star (Half-Wave) (See 1-D)	Three Phase Full Wave Bridge (See 1-E)	Six-Phase Star (Three Phase Diametric) (See 1-F)	Three Phase Double Wave With Interphase Transformer (See 1-G)
R M S Input Voltage Per Transformer Leg (V_1)	Resistive & Inductive	2.22 V_O	1.11 V_O	1.11 V_O	0.855 V_O	0.428 V_O	0.741 V_O	0.855 V_O
	Capacitive	0.707 V_O	0.707 V_O	0.707 V_O	0.707 V_O	0.408 V_O	0.707 V_O	0.707 V_O
Peak Inverse Voltage Per Rectifier (P & V)	R & L	3.14 V_O	3.14 V_O	1.57 V_O	2.09 V_O	1.05 V_O	2.09 V_O	2.09 V_O
	C	2.00 V_O	2.00 V_O	1.00 V_O	2.00 V_O	1.00 V_O	2.00 V_O	2.00 V_O
Average Current Through Through Rectifier I_F	R.L. & C	1.00 I_O	0.50 I_O	0.50 I_O	0.333 I_O	0.333 I_O	0.167 I_O	0.167 I_O
Current Through Rectifier I_M	R	3.14 I_O	1.57 I_O	1.57 I_O	1.21 I_O	1.05 I_O	1.05 I_O	0.525 I_O
	L		1.00 I_O	1.00 I_O	1.00 I_O	1.00 I_O	1.00 I_O	0.500 I_O
	C	Depends on Size of Capacitor						
Transformer Total Secondary PA	Sine Wave	3.49 P_O	1.75 P_O	1.23 P_O	1.50 P_O	1.05 P_O	1.81 P_O	1.49 P_O
	Sq. Wave	3.14 P_O	1.57 P_O	1.11 P_O	1.48 P_O	1.05 P_O	1.81 P_O	1.48 P_O
Transformer Total Primary PA	Sine Wave	3.49 P_O	1.23 P_O	1.23 P_O	1.23 P_O	1.05 P_O	1.28 P_O	1.06 P_O
	Sq. Wave	3.14 P_O	1.11 P_O	1.11 P_O	1.21 P_O	1.05 P_O	1.28 P_O	1.05 P_O
% Ripple	Sine Wave Resistive Load	121%	47%	47%	17%	4%	4%	4%
Lowest Ripple Frequency	—	1 F_I	2 F_I	2 F_I	3 F_I	6 F_I	6 F_I	6 F_I
Conversion Efficiency	—	40.6%	81.2%	81.2%	97%	99.5%	99.5%	99.5%

Table 1. Electrical Reference Table and Rectifier Circuit Wave Shapes (Cont.)

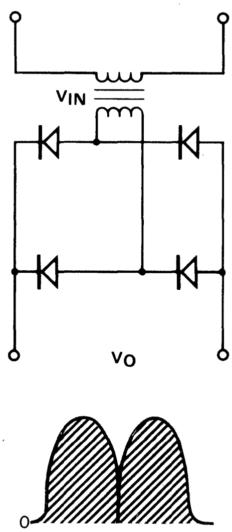
SINGLE PHASE HALF WAVE (1-A)



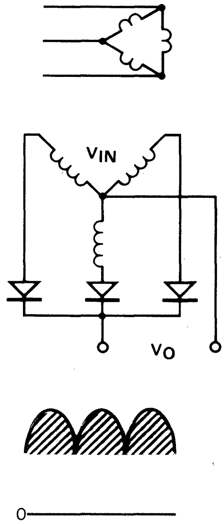
SINGLE PHASE FULL WAVE CENTER TAP (1-B)



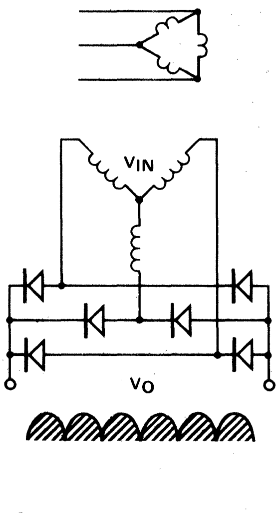
SINGLE PHASE FULL WAVE BRIDGE (1-C)



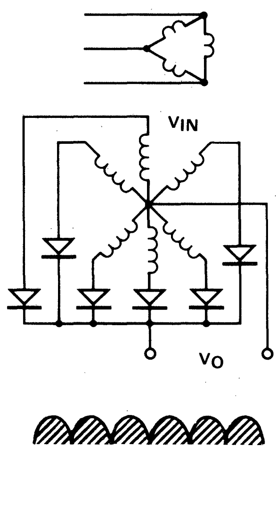
THREE PHASE STAR (HALF WAVE) (1-D)



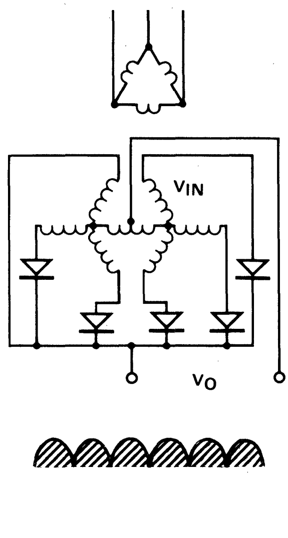
THREE PHASE FULL WAVE BRIDGE (1-E)



SIX PHASE STAR (THREE PHASE DIAMETRIC) (1-F)



THREE PHASE DOUBLE WAVE WITH INTERPHASE TRANSFORMER (1-G)



Swinging Choke LC Section Filter

When designing a swinging choke section filter, the inductance required at the minimum and maximum output currents can be determined as follows.

1. Find L_C (critical inductance)

$$L_C = \frac{R_S + R_{eff}}{3\omega}$$

where, as before:

$$R_{eff} = \frac{R_K R_{L(max)}}{R_K + R_{L(max)}}$$

2. Find L2 (inductance at maximum load current)

$$L2 = \frac{R_S + R_{eff2}}{3\omega}$$

where:

$$R_{eff2} = \frac{R_{L(min)} R_K}{R_{L(min)} + R_K}$$

When L_C has been determined, then the capacitor value may be calculated as before. The condition $\omega^2 L_C \leq 1/4$ should be avoided due to possible filter instabilities.

Capacitive Input Filter Characteristics

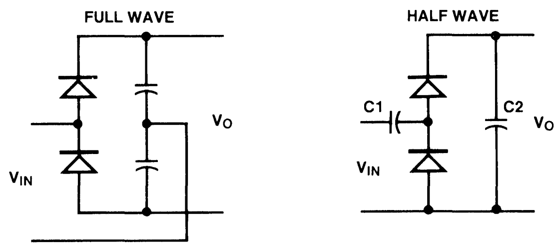
$R_S/R_{L(min)} = 0.02 \qquad \omega C R_{L(min)} = 12$

When the voltage and current levels are known, *Table 2* can be used to select the optimum configuration and determine transformer and rectifier characteristics.

Voltage Doublers

Increased dc output voltage from a transformer winding can be obtained using a voltage multiplier circuit. However, this method requires additional components, *i.e.,* two filter capacitors, and reduces the output current. A full-wave doubler and a half-wave doubler are shown in *Figure 14.* The half-wave doubler is generally preferred since it has a common input and output terminal. In operation, C2 is charged on one half cycle; on the second half cycle, C1 is charged thereby summing the voltages across each capacitor. This provides a doubling effect since the output voltage is approximately twice the input voltage.

Fig. 14. Voltage Doublers

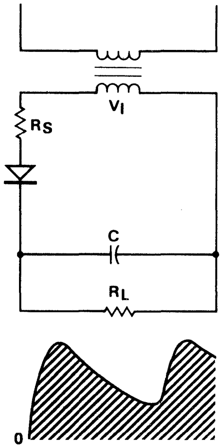
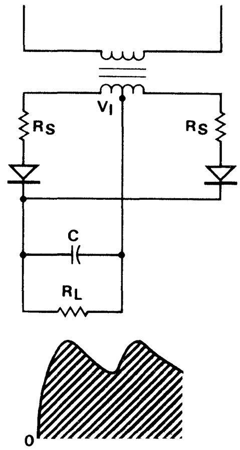


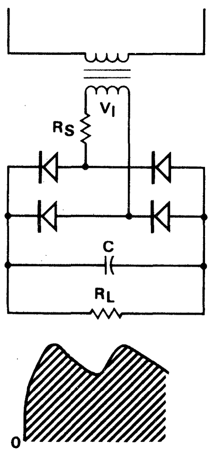
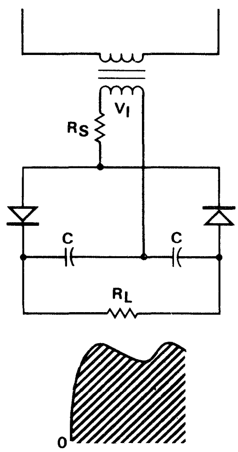
5

Table 2. Capacitive Input Filter Characteristics and Rectifier Circuit Wave Shapes

Characteristic	Single Phase Half Wave (See 2-A)	Single Phase Full Wave Center-Tap (See 2-B)	Single Phase Full Wave Bridge (See 2-C)	Single Phase Full Wave Voltage Doubler (See 2-D)
V_1	0.910 V_O	0.825 V_O	0.805 V_O	0.552 V_O
PIV	2.56 V_O	2.34 V_O	1.14 V_O	1.56 V_O
Ripple	0.12 V_O	.06 V_O	.06 V_O	.09 V_O
I_M/Rect.	7.80 I_O	4.75 I_O	4.75 I_O	3.00 I_O
I_{RMS}/Rect.	2.50 I_O	1.33 I_O	1.33 I_O	1.10 I_O
SEC VA	2.35 P_O	2.16 P_O	2.16 P_O	1.22 P_O
PRI VA	2.35 P_O	3.05 P_O	2.16 P_O	1.72 P_O

Table 2. Capacitive Input Filter Characteristics and Rectifier Circuit Wave Shapes (Cont.)

SINGLE PHASE HALF WAVE (2-A)	SINGLE PHASE FULL WAVE CENTER TAP (2-B)
	

SINGLE PHASE FULL WAVE BRIDGE (2-C)	SINGLE PHASE FULL WAVE VOLTAGE DOUBLER (2-D)
	

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1. Schade, O.H., "Analysis of Rectifier Operation," Proc IRE, July 1943, Vol. 31 #7, pp. 341-361.
2. Ryder, John D., "Electronic Engineering Principles," Prentice Hall, Inc. 1947, pp. 94-125.
3. Martin, Thomas L., Jr., "Electronic Circuits," Prentice Hall, Inc., 1955, pp. 506-541.
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5. Gray, Truman, S., "Applied Electronics," John Wiley and Sons, Inc., 1954 pp. 250-277.

Thermal Considerations

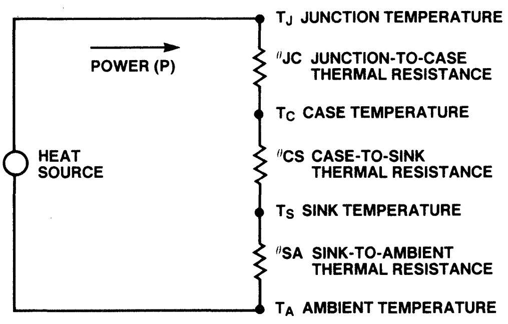
To realize the full capabilities of the High Current Voltage Regulator, sufficient attention must be paid to proper heat removal. For efficient thermal management, the user must rely on important parameters supplied by the manufacturer, such as junction-to-case and junction-to-ambient thermal resistance and maximum operating junction temperature. The device temperature depends on the power dissipation level, the means for removing the heat generated by this power dissipation and the temperature of the body (heat sink) to which this heat is removed.

Figure 1 shows a simplified equivalent circuit for a typical semiconductor device in equilibrium. The power dissipation, which is analogous to current flow in electrical terms, is caused by a heat source similar to a voltage source. Temperature is analogous to voltage potential and thermal resistance to ohmic resistance. Extending the analogy of Ohm's law to

$$\theta_{JA(tot)} = \theta_{JC} + \theta_{CS} + \theta_{SA} = \frac{T_J - T_A}{P_D}$$

Thermal resistance, then, is the rise in the temperature of a package above some reference level per unit of power dissipation in that package, usually expressed in degrees in centigrade per watt. The reference temperature may be ambient or it may be the temperature of a heat sink to which the package is connected. There are several factors that affect thermal resistance including die size, the size of the heat source on the die (or substrate), die-attach material and thickness, substrate material and thickness, and package material, construction and thickness.

Fig. 1 Simplified Thermal Circuit



Thermal Evaluation of Regulators

To measure thermal resistance, the difference between the junction temperature and the chosen reference temperature, case, sink or ambient, must be determined. Ambient or sink temperature measurement is straightforward. For case-temperature measurement, the device should have a sufficiently large heat sink and the power level should be close to the specified rating of the package-die combination. The case temperature can be measured by an infrared microradiometer or by using a thermocouple soldered to a point in the center of the case heat-sink interface as close to the die as practical.

Measurement of the junction temperature, unfortunately, is not as simple and involves some calibrations. There are several methods available for junction-temperature measurement; the one most commonly used is described here.

Thermal Shutdown Method

With this method, the thermal shutdown temperature of each device is used as the thermometer in determining the thermal resistance. The device is first heated externally, with as little internal power dissipation as practical, until it reaches thermal shutdown. Then, with the device mounted on a heat sink, the regulator is powered externally until it reaches thermal shutdown again. In some cases, the ambient of the device and its heat sink may have to be elevated sufficiently to force the regulator into shutdown. The thermal resistance of the device can then be calculated by using

$$\theta_{JC} = \frac{T_J - T_C}{P_D}$$

where θ_{JC} is the junction-to-case thermal resistance
T_J is the measured thermal shutdown temperature
T_C is the measured case temperature
P_D is the power dissipated to force the device into shutdown and is equal to

$$(V_{IN} - V_{OUT})\ I_{OUT} + V_{IN}\ I_Q$$

I_Q is the quiescent current of the device and can be neglected for low thermal resistance packages such as the TO-3

Heat Sink Requirements

When is a heat sink necessary, and what type of a heat sink should one use? The answers to these questions depend on reliability and cost requirements. Heat sinking is necessary to keep the operating junction temperature T_J of the regulator below the specified maximum value. Since semiconductor reliability improves as operating junction temperature is lowered, a reliability/cost compromise is usually made in the device design.

Thermal characteristics of voltage-regulator chips and packages determine that some form of heat sinking is mandatory whenever the power dissipation exceeds 3.2 W for the high current voltage regulator TO−3 package at 25°C ambient or lower power levels at ambients above 25°C.

To choose or design a heat sink, the designer must determine the following regulator parameters.

$P_{D(max)}$ — Maximum power dissipation: $(V_{IN} - V_{OUT})\, I_{OUT} + V_{IN}\, I_Q$

$T_{A(max)}$ — Maximum ambient temperature the regulator will encounter during operation.

$T_{J(max)}$ — Maximum operating junction temperature, specified by the manufacturer.

θ_{JC}, θ_{JA} — Junction-to-case and junction-to-ambient thermal resistance values, also specified by the regulator manufacturer. (θ_{JA} = 38°C/W max. θ_{JC} = 2.50°C/W max).

θ_{CS} — Case-to-heat-sink thermal resistance which for large packages, can range from about 0.2°C/W to about 1°C/W depending on the quality of the contact between the package and the heat sink.

θ_{SA} — Heat-sink-to-ambient thermal resistance, specified by heat-sink manufacturer.

Maximum permissible dissipation without a heat sink is determined by

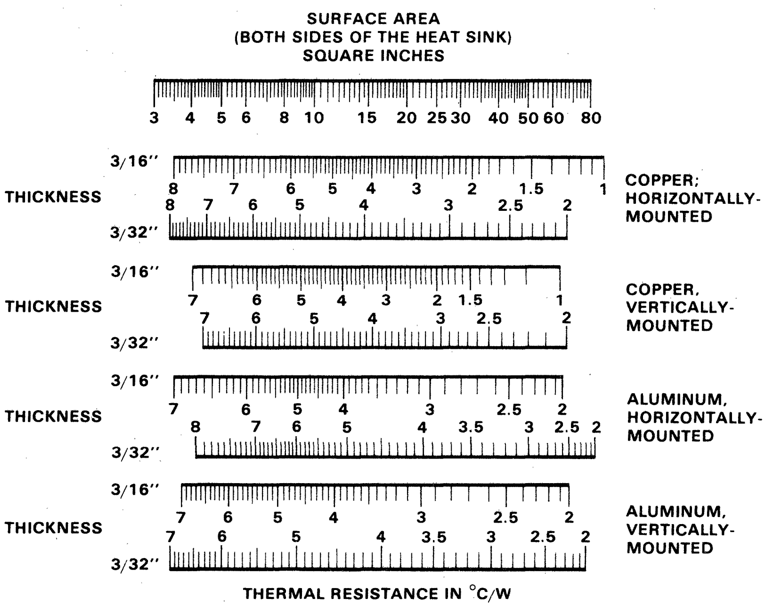
$$P_{D(max)} = \frac{T_{J(max)} - T_{A(max)}}{\theta_{JA}}$$

If the device dissipation P_D exceeds this figure, a heat sink is necessary. The total required thermal resistance may then be calculated.

$$\theta_{JA(tot)} = \theta_{JC} + \theta_{CS} = \theta_{SA} = \frac{T_{J(max)} - T_{A(max)}}{P_D}$$

Case-to-sink and sink-to-ambient thermal resistance information on commercially available heat sinks is normally provided by the heat sink manufacturer. A summary of some commercially available heat sinks is shown in *Table 1.* However, if a chassis or other conventional surface is used as a heat sink, *Figure 2* can be used as a guide to estimate the required surface area.

Fig. 2 Heat Sink Material Selection Guide



To determine either area required or thermal resistance of a given area, draw a vertical line between the top (or area) line down to the material of interest.

How to Choose a Heat Sink — Example
Determine the heat sink required for a regulator which has the following system requirements:

Operating ambient temperature range: 0°C−40°C
Maximum junction temperature: 125°C
Maximum output current: 3 A
Maximum input to output differential: 5 V

For this example assume the μA78HGA, 5 Amp Positive Adjustable High Current Voltage Regulator has been selected.

θ_{JC} = 2.5°C/W maximum (from data sheet)

$$\theta_{JA(tot)} = \theta_{JC} + \theta_{CS} + \theta_{SA} = \frac{T_J - T_A}{P_D}$$

$$\theta_{CS} + \theta_{SA} = \frac{125 - 40}{3 \times 5} - 2.5 = 3.16°C/W$$

Assuming θ_{CS} = .16°C/W then θ_{SA} = 3°C/W

This thermal resistance value can be achieved by using either 22 square inches of 3/16 inch thick vertically mounted aluminum (*Figure 2*) or a commercial heat sink (*Table 1*).

Tips for Better Regulator Heat Sinking
Avoid placing heat-dissipating components such as power resistors next to regulators.

Keep lead lengths to a minimum and use the largest possible area of the printed board traces or mounting hardware to provide a heat dissipation path for the regulator.

Be sure the heat sink surface is flat and free from ridges or high spots. Check the regulator package for burrs or peened-over corners. Regardless of the smoothness and flatness of the package and heat-sink contact, air pockets between them are unavoidable unless a lubricant is used. Therefore, for good thermal conduction, use a thin layer of thermal lubricant such as Dow Corning DC-340, General Electric 662 or Thermacote by Thermalloy.

If the regulator is mounted on a heat sink with fins, the most efficient heat transfer takes place when the fin is in a vertical plane, as this type of mounting forces the heat transfer from fin to air in a combination of radiation and convection.

If it is necessary to bend any of the regulator leads, handle them carefully to avoid straining the package. Furthermore, lead bending should be restricted since repeated bending will fatigue and eventually break the leads.

Table 1
Heat Sink Selection Guide
This list is only representative. No attempt has been made to provide a complete list of all heat sink manufacturers. All values are typical as given by manufacturer or as determined from characteristic curves supplied by manufacturer.

θ_{SA} Approx. (°C/W)	Manufacturer and Type
0.4 (9″ length)	Thermalloy (Extruded) 6590 Series
0.4 - 0.5 (6″ length)	Thermalloy (Extruded) 6660, 6560 Series
0.56 - 3.0	Wakefield 400 Series
0.6 (7.5″ length)	Thermalloy (Extruded) 6470 Series
0.7 - 1.2 (5 - 5.5″ length)	Thermalloy (Extruded) 6423, 6443, 6441, 6450 Series
1.0 - 5.4 (3″ length)	Thermalloy (Extruded) 6427, 6500, 6123, 6401, 6403, 6421, 6463, 6176, 6129, 6141, 6169, 6135, 6442 Series
1.9	IERC E2 Series (Extruded)
2.1	IERC E1, E3 Series (Extruded)
2.3 - 4.7	Wakefield 600 Series
4.2	IERC HP3 Series
4.5	Staver V3-5-2
4.8 - 7.5	Thermalloy 6001 Sries
5 - 6	IERC HP3 Series
5 - 10	Thermalloy 6013 Series
5.6	Staver V3-3-2
5.9 - 10	Wakefield 680 Series
6	Wakefield 390 Series
6.4	Staver V3-7-224
6.5 - 7.5	IERC Up SEries
8	Staver V1-5
8.1	Staver V3-5
8.8	Staver V3-7-96
9.5	Staver V3-3
9.5 - 10.5	IERC LA Series
9.8 - 13.9	Wakefield 630 Series
10	Staver V1-3
11	Thermalloy 6103, 6117 Series

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