TA0105A

STEREO CLASS-T DIGITAL AUDIO AMPLIFIER DRIVER USING DIGITAL POWER PROCESSING (DPP™) TECHNOLOGY

Technical Information **Revision 2.2 – May 2005**

GENERAL DESCRIPTION

The TA0105A is a two-channel Amplifier Driver IC that uses Tripath's proprietary Digital Power Processing (DPP™) technology. Class-T amplifiers offer both the audio fidelity of Class-AB and the power efficiency of Class-D amplifiers.

The typical application for the TA0105A is direct drive (no output transformer) in 70V and 100V constant voltage amplifiers used for public address systems. The feedback and voltage range of the TA0105A can be configured externally unlike previous Tripath modules such as TA0104A.

APPLICATIONS

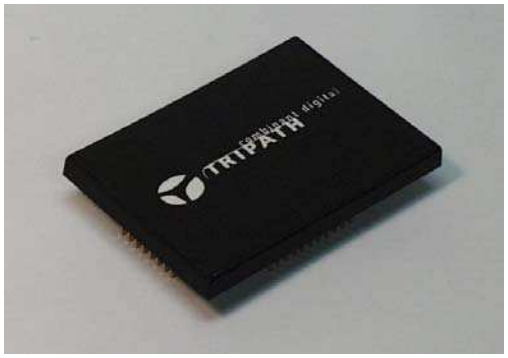
- Constant Voltage Amplifiers
- Distribution Amplifiers
- Pro-audio Amplifiers

BENEFITS

- Reduced system cost with smaller/less expensive power supply and heat sink
- Signal fidelity equal to high quality Class-AB amplifiers
- No output transformer is needed due to high supply voltage range
- High dynamic range compatible with digital media such as CD and DVD

FEATURES

- Class-T architecture
- Proprietary Digital Power Processing technology
- High Supply Voltage Range
- "Audiophile" Sound Quality
- High Efficiency
- Supports wide range of output power levels
- Output over-current protection
- Over- and under-voltage protection
- 38-pin Quad package



Absolute Maximum Ratings (Note 1)

SYMBOL	PARAMETER	Value	UNITS
VPP, VNN	Supply Voltage	+/-200	V
V5	Positive 5V Controller Voltage Voltage at Input Pins (pins 4-8, 10-11)	6 -0.3 to (V5+0.3)	V V
VN12	Voltage for FET drive	VNN+18	V
T_A	Operating Free-air Temperature Range	0º to 70º	C
T_J	Junction Temperature	150º	C
T_{STORE}	Storage Temperature Range	-40º to 150º	C
ESD_{HB}	ESD Susceptibility – Human Body Model (Note 2) All Pins	 2000	 V
ESD_{MM}	ESD Susceptibility – Machine Model (Note 3) All Pins	 200	 V

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. See the table below for Operating Conditions.
Note 2: Human body model, 100pF discharged through a 1.5KΩ resistor.
Note 3: Machine model, 220pF – 240pF discharged through all pins.

Operating Conditions (Note 4)

SYMBOL	PARAMETER	MIN.	TYP.	MAX.	UNITS
VPP, VNN	Supply Voltage (Note 4)	+/-125	+/-148	+/-185	V
V5	Positive 5V Controller Voltage	4.5	5	5.5	V
VN12	Voltage for FET drive (Volts about VNN)	10.8	12	13.2	V

Note 4: Recommended Operating Conditions indicate conditions for which the device is functional. The VPP and VNN supply limits are based on the internal OV/UV sensing resistor values. The supply voltage range can be lowered via external resistors. Please refer to the Application information section for a detailed discussion of changing the operating supply voltage range. See Electrical Characteristics for guaranteed specific performance limits.

Electrical Characteristics (Note 5)

T_A = 25 °C. See Application/Test Circuit on page 7. Unless otherwise noted, the supply voltage is VPP=|VNN|=148V. See note 8.

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNITS
I_q	Quiescent Current (No load, BBM0=1,BBM1=0, Mute = 0V)	VPP = +148V VNN = -148V V5 = 5V VN12 = 12V		35 40 45 120	100 100 80 250	mA mA mA mA
I_q	Quiescent Current (No load, BBM0=1,BBM1=0, Mute = 0V)	VPP = +106V VNN = -106V V5 = 5V VN12 = 12V		30 35 45 130	100 100 80 250	mA mA mA mA
I_{MUTE}	Mute Supply Current (No load, Mute = 5V)	VPP = +148V VNN = -148V V5 = 5V VN12 = 12V		1 1 20 1	 30 	mA mA mA mA
I_{Po}	Power Supply Current (Vo = 100Vrms, R_L = 50Ω)	VPP = +148V (Both Channels On) VNN = -148V (Both Channels On)		1.5 1.5	1.6 1.6	A A
I_{Po}	Power Supply Current (Vo = 70.7Vrms, R_L = 25Ω)	VPP = +106 (Both Channels On) VNN = -106 (Both Channels On)		2.1 2.1	2.22 2.22	A A
V_{IH}	High-level input voltage (MUTE)		3.5			V
V_{IL}	Low-level input voltage (MUTE)				1.0	V
V_{OH}	High-level output voltage (HMUTE)	RL = 10kohm	3.5			V
V_{OL}	Low-level output voltage (HMUTE)	RL = 10kohm			0.5	V
V_{OFFSET}	Output Offset Voltage	No Load, MUTE = Logic low, Measured without external trim circuit connected	-2.5		2.5	V
I_{OC}	Over Current Sense Voltage Threshold	Exceeding this threshold causes a latched mute condition	0.85	0.97	1.09	V
$V_{VPPSENSE}$	VPP Threshold Voltages	Over-voltage turn on (muted) Over-voltage restart (mute off) Under-voltage restart (mute off) Under-voltage turn on (muted)	193 185 80	227 216 111 101	250 125 118	V V V V
$V_{VNNSENSE}$	VNN Threshold Voltages	Over-voltage turn on (muted) Over-voltage restart (mute off) Under-voltage restart (mute off) Under-voltage turn on (muted)	-193 -185 -80	-221 -215 -110 -98	-250 -125 -118	V V V V
$V_{VPPSENSE}$	VPP Threshold Voltages (Externally shifted) (Note 6)	Over-voltage turn on (muted) Over-voltage restart (mute off) Under-voltage restart (mute off) Under-voltage turn on (muted)	147 140 60	167 159 82 74	185 95 85	V V V V
$V_{VNNSENSE}$	VNN Threshold Voltages (Externally shifted) (Note 6)	Over-voltage turn on (muted) Over-voltage restart (mute off) Under-voltage restart (mute off) Under-voltage turn on (muted)	-147 -140 -60	-166 -161 -82 -73	-185 -95 -85	V V V V

Note 5: Minimum and maximum limits are guaranteed but may not be 100% tested.

Note 6: These voltage values are calculated and not 100% tested. The voltages are based on 100% tested sense currents, an external “shift” resistor of 3.83MΩ from VLOW to VNN and another 3.83MΩ resistor from VHIGH to VPP, and the on board sense resistor values of 1.27MΩ for VNN, and 1.4MΩ for VPP. In addition, worse case resistor tolerances (+/-1%) were used to calculate the minimum and maximum values. Please refer to the Overvoltage and Undervoltage section of the Applications Information for a more detailed explanation.

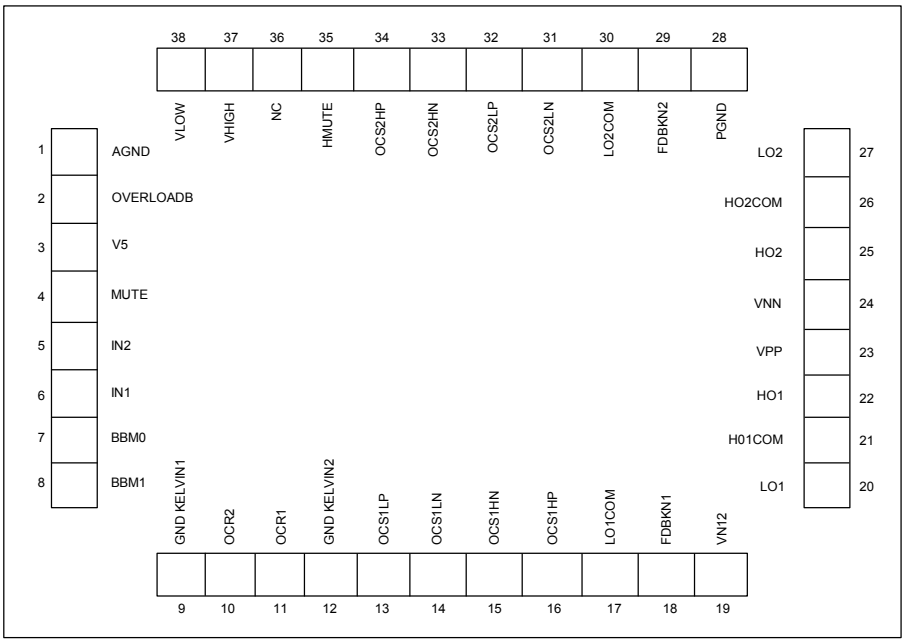
Electrical Characteristics (Notes 7 and 8)

T_A = 25 °C. Unless otherwise noted, the supply voltage is VPP=|VNN|=148V, the input frequency is 1kHz and the measurement bandwidth is 20kHz. See Application/Test Circuit on page 7.

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNITS
P_{OUT}	Output Power (Continuous Output/Channel)	VPP=|VNN|=148V THD+N = 0.1%, R_L = 25Ω THD+N = 1.0%, R_L = 25Ω VPP=|VNN|=106V THD+N = 0.1%, R_L = 12.5Ω THD+N = 1.0%, R_L = 12.5Ω		 300 400 300 400		 W W W W
THD + N	Total Harmonic Distortion Plus Noise	V_{OUT} = 50Vrms, f = 1kHz, R_L = 12.5Ω, VPP=|VNN|=106V		0.06	0.25	%
THD + N	Total Harmonic Distortion Plus Noise	V_{OUT} = 50Vrms, f = 7kHz, R_L = 12.5Ω, VPP=|VNN|=106V		0.2	1.0	%
THD + N	Total Harmonic Distortion Plus Noise	V_{OUT} = 70.7Vrms, f = 1kHz, R_L = 25.0Ω, VPP=|VNN|=148V		0.06	0.25	%
THD + N	Total Harmonic Distortion Plus Noise	V_{OUT} = 70.7Vrms, f = 7kHz, R_L = 25.0Ω, VPP=|VNN|=148V		0.2	1.0	%
IHF-IM	IHF Intermodulation Distortion	19kHz, 20kHz, 1:1 (IHF), R_L = 50Ω V_{OUT} = 25Vrms/Channel		0.02		%
SNR	Signal-to-Noise Ratio	A Weighted, R_L = 25Ω, P_{OUT} = 400W/Channel		103		dB
CS	Channel Separation	0dBr = 100W, R_L = 25Ω, f = 1kHz		100		dB
η	Power Efficiency	V_{OUT} = 100Vrms/Channel, R_L = 50Ω, VPP=|VNN|=148V	85	90		%
η	Power Efficiency	V_{OUT} = 70.7Vrms/Channel, R_L = 25Ω, VPP=|VNN|=106V	85	90		%
A_V	Amplifier Gain	P_{OUT} = 10W/Channel, R_L = 25Ω, Rin = 34.8kΩ, See Application / Test Circuit		40.5		V/V
A_{VERROR}	Channel to Channel Gain Error	P_{OUT} = 10W/Channel, R_L = 25Ω See Application / Test Circuit	-1		1	dB
e_{NOUT}	Output Noise Voltage	A-Weighted, input shorted, DC offset nulled to zero		700		μV

Note 7: Minimum and maximum limits are guaranteed but may not be 100% tested.
Note 8: Specific Components used:
Output MOSFETs (Q_O): ST Microelectronics STW20NM50FD
Output Diodes (D_O): International Rectifier MUR460
Feedback Resistors (R_{FB}): 39.2Kohm, 1W
Gate Diodes (D_G): General Semiconductor SS16

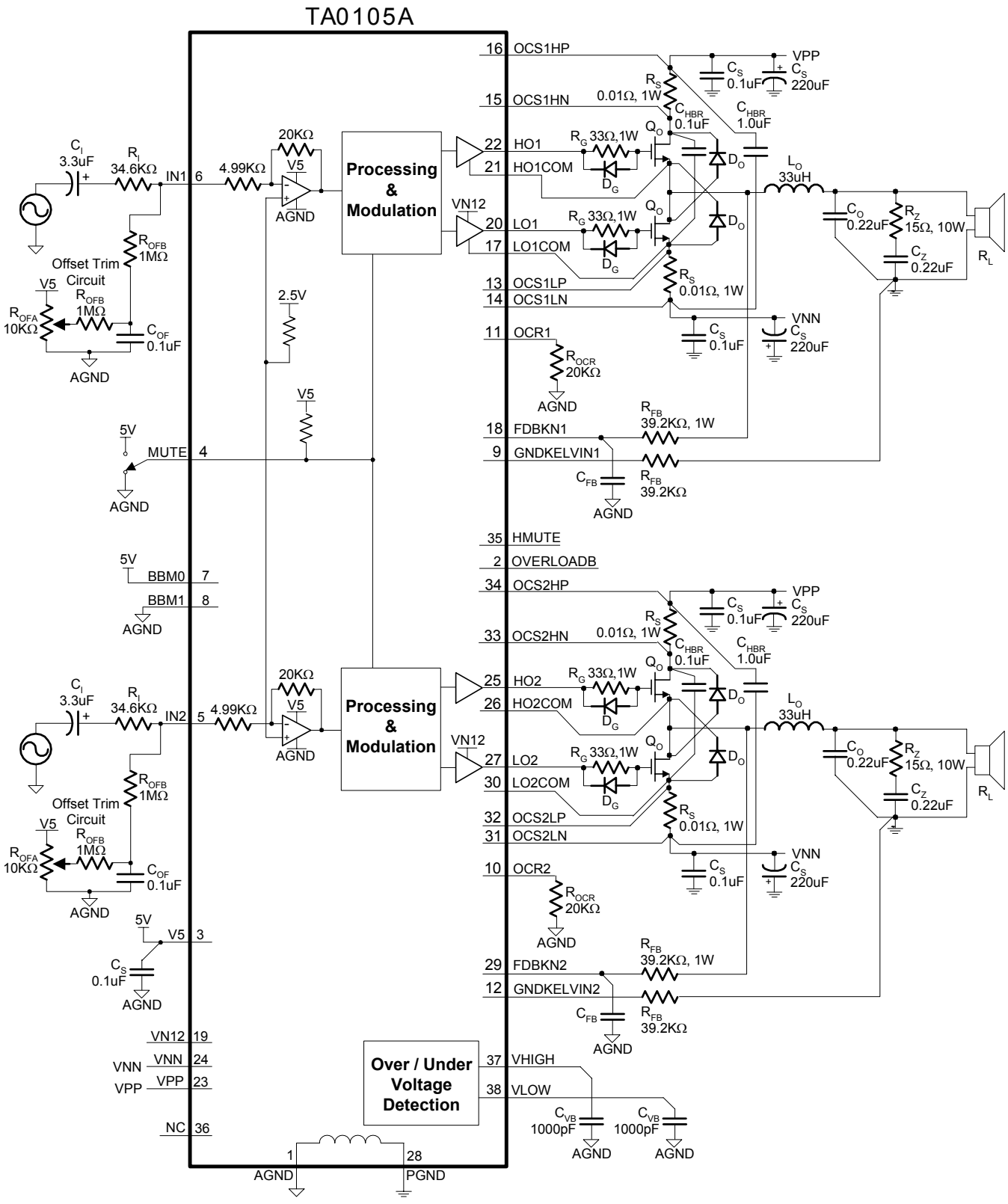
TA0105A Pinout



Pin Description

Pin	Function	Description
1	AGND	Analog ground. This should be the "star" point for all connections to analog ground.
2	OVERLOADB	Normally logic high. Logic low signals onset of clipping. Pin output impedance is approximately 100kΩ.
3	V5	5V power supply input.
4	MUTE	Logic input. A logic high puts the amplifier in mute mode. Ground pin if not used. Please refer to the section, Mute Control, in the Application Information.
5, 6	IN2, IN1	Audio inputs. (Channels 2 & 1)
7, 8	BBM0, BBM1	Break-before-make timing control to prevent shoot-through in the output MOSFETs.
9, 12	GNDKELVIN1, GNDKELVIN2	Output ground feedback (Channels 1 & 2)
10, 11	OCR2, OCR1	Over-current threshold adjustment (Channels 2 & 1)
13, 14	OCS1LP, OCS1LN	Over Current Sense inputs, Channel 1 low-side
15, 16	OCS1HN, OCS1HP	Over Current Sense inputs, Channel 1 high-side
17,30	LO1COM, LO2COM	Kelvin connection to source of low-side transistor (Channel 1 & 2)
18, 29	FDBKN1, FDBKN2	Switching feedback (Channels 1 & 2)
19	VN12	"Floating" supply input for the FET drive circuitry. This voltage must be stable and referenced to VNN.
20,27	LO1, LO2	Low side gate drive output (Channel 1 & 2)
21,26	HO1COM, HO2COM	Kelvin connection to source of high-side transistor (Channel 1 & 2)
22,25	HO1, HO2	High side gate drive output (Channel 1 & 2)
23	VPP	Positive supply voltage input. Connect to positive power supply. Used for power supply sensing.
24	VNN	Negative supply voltage input. Connect to positive power supply. Used for power supply sensing.
28	PGND	Power ground. This should be connected to the "star" point for the power (output) ground.
31, 32	OCS2LN, OCS2LP	Over Current Sense inputs, Channel 2 low-side
33, 34	OCS2HN, OCS2HP	Over Current Sense inputs, Channel 2 high-side
35	HMUTE	Logic Output. A logic high indicates both amplifiers are muted, due to the mute pin state, or a "fault" such as an overcurrent, undervoltage, or overvoltage condition.
36	NC	Do not connect.
37	VHIGH	Positive supply voltage sense input. This pin is biased at 2.5V nominally and left floating in typical applications. An external resistor may also be connected from VPP to VHIGH to lower the supply voltage operation range. See the Application Information for a detailed description on how to lower the supply voltage range.
38	VLOW	Negative supply voltage sense input. This pin is biased at 1.25V nominally and left floating in typical applications. An external resistor may also be connected from VNN to VLOW to lower the supply voltage operation range. See the Application Information for a detailed description on how to lower the supply voltage range.

Application/Test Circuit



NC - Not Connected (Must Be Left Floating)

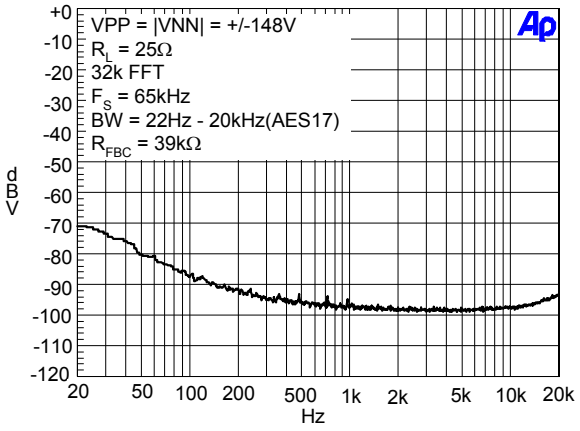
External Components Description (Refer to the Application/Test Circuit)

Components	Description
R_I	Inverting input resistance to provide AC gain in conjunction with R_F. This input is biased at the BIASCAP voltage (approximately 2.5VDC).
C_I	AC input coupling capacitor which, in conjunction with R_I, forms a highpass filter at $f_C = 1/(2\pi R_I C_I)$.
R_{FB}	Feedback resistor connected from either the half-bridge output to FDBKN1 (FDBKN2) or speaker ground to GNDKELVIN1 (GNDKELVIN2). The value of this depends on the supply voltage range and sets the TA0105A gain in conjunction with R_I. It should be noted that the feedback resistor from the half-bridge output must have a power rating of greater that $P_{DISS} = VPP^2/2R_{FB}$. Please see the Modulator Feedback Design paragraphs in the Application Information Section.
C_{FB}	Feedback delay capacitor that both lowers the idle switching frequency and filters very high frequency noise from the feedback signal, which improves amplifier performance. The value of C_{FB} should be offset between channel 1 and channel 2 so that the idle switching difference is greater than 40kHz. Please refer to the Application / Test Circuit.
R_{OFA}	Potentiometer used to manually trim the DC offset on the output of the TA0105A.
R_{OFB}	Resistor that limits the manual DC offset trim range and allows for more precise adjustment.
C_{OF}	Decoupling capacitor which low pass filters the offset trim voltage from noise and power supply fluctuations.
C_S	Supply decoupling for the power supply pins. For optimum performance, these components should be located close to the TA0105A and returned to their respective ground as shown in the Application/Test Circuit.
R_S	Over-current sense resistor. Please refer to the section, Setting the Over-current Threshold, in the Application Information for a discussion of how to choose the value of R_S to obtain a specific current limit trip point.
R_{OCR}	Over-current "trim" resistor, which, in conjunction with R_S, sets the current trip point. Please refer to the section, Setting the Over-current Threshold, in the Application Information for a discussion of how to calculate the value of R_{OCR}.
C_{HBR}	Supply decoupling for the high current Half-bridge supply pins. These components must be located as close to the output MOSFETs as possible to minimize output ringing which causes power supply overshoot. By reducing overshoot, these capacitors maximize both the TA0105A and output MOSFET reliability. These capacitors should have good high frequency performance including low ESR and low ESL. In addition, the capacitor rating must be twice the maximum VPP voltage.
Q_O	Output MOSFET. This is the main output switching device and to a large extent, sets the amplifier's limitations. This device must be a switching grade device with a good compromise between gate charge and on resistance while being able to withstand the full supply range. Please refer to the recommended devices in the Applications Information section.
D_O	Output diode, which is used to minimizes output overshoots/undershoots on the output node. These devices clamp the output to low impedance node formed by the close connection of CHBR. Note the connection shown in the Application/Test Circuit. The "drain to drain" diode protects the bottom side device from excessive BVDSS due to overshoots on the output node. The "source to source" diode protects the top side device from excessive BVDSS due to undershoots on the output node. This device must be an ultra fast rectifier capable of sustaining the entire supply range (VPP-VNN) and high peak currents.
R_G	Gate resistor, which is used to control the MOSFET rise/ fall times. This resistor serves to dampen the parasitics at the MOSFET gates, which, in turn, minimizes ringing and output overshoots. The typical power rating is 1 watt.
D_G	Gate diode, which is used to "speed-up" the turn off of the MOSFET. This minimizes cross conduction and idle VPP/VNN supply current. This device should be a switching grade type such as Schottky or ultra-fast rectifier.

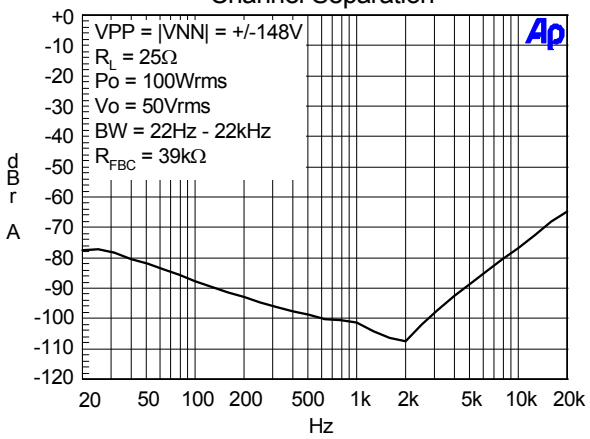
C_Z	Zobel capacitor, which in conjunction with R_Z, terminates the output filter at high frequencies. Use a high quality film capacitor capable of sustaining the ripple current caused by the switching outputs.
R_Z	Zobel resistor, which in conjunction with C_Z, terminates the output filter at high frequencies. The combination of R_Z and C_Z minimizes peaking of the output filter under both no load conditions or with real world loads, including loudspeakers which usually exhibit a rising impedance with increasing frequency. Depending on the program material, the power rating of R_Z may need to be adjusted. Typically 10 watts. If the system requires full power operation at 20kHz then the power rating for R_Z will likely need to be increased.
L_O	Output inductor, which in conjunction with C_O, demodulates (filters) the switching waveform into an audio signal. Forms a second order filter with a cutoff frequency of $f_C = 1/(2\pi\sqrt{L_O C_O})$ and a quality factor of $Q = R_L C_O / \sqrt{L_O C_O}$.
C_O	Output capacitor, which, in conjunction with L_O, demodulates (filters) the switching waveform into an audio signal. Forms a second order low-pass filter with a cutoff frequency of $f_C = 1/(2\pi\sqrt{L_O C_O})$ and a quality factor of $Q = R_L C_O / \sqrt{L_O C_O}$. Use a high quality film capacitor capable of sustaining the ripple current caused by the switching outputs.
C_{VB}	Supply decoupling for the power supply sensing pins. For optimum performance, these components should be located close to the TA0105A and returned to analog ground.

Typical Performance

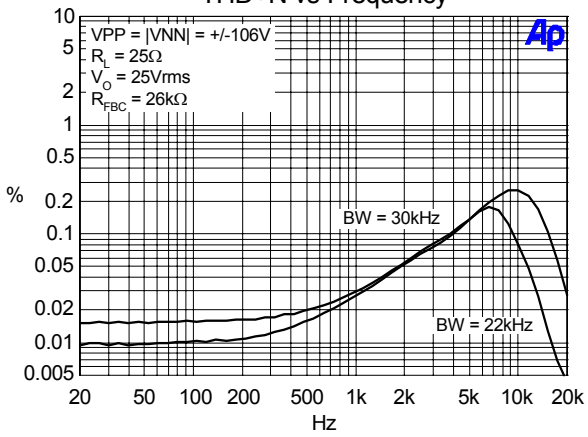
Noise Floor



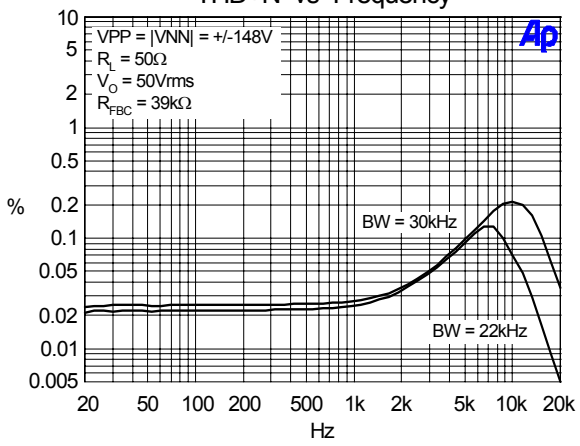
Channel Separation



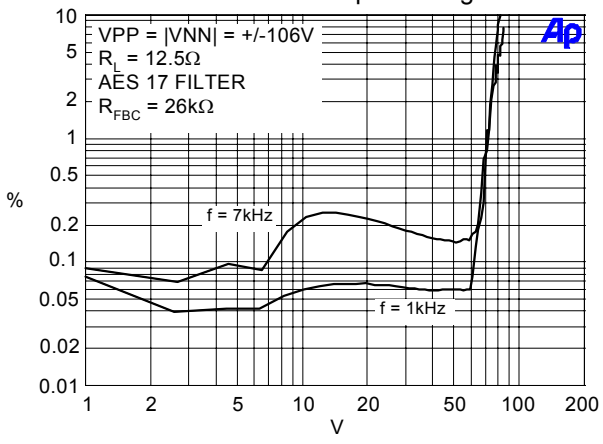
THD+N vs Frequency



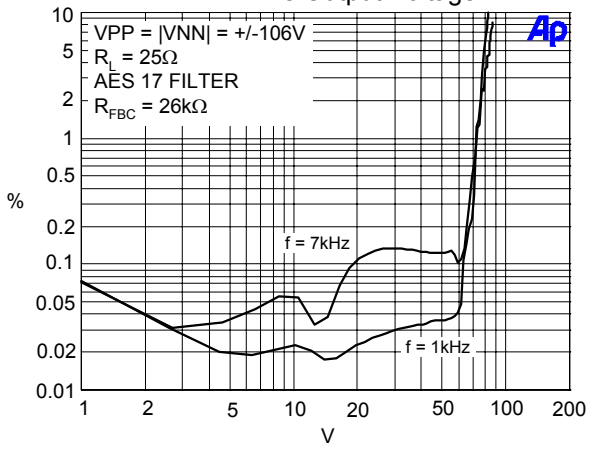
THD+N vs Frequency



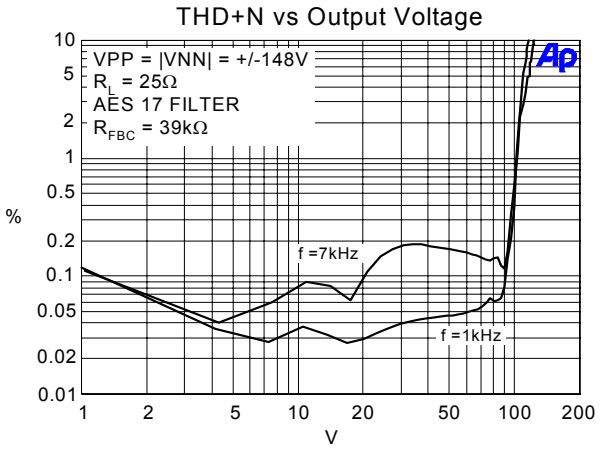
THD+N vs Output Voltage

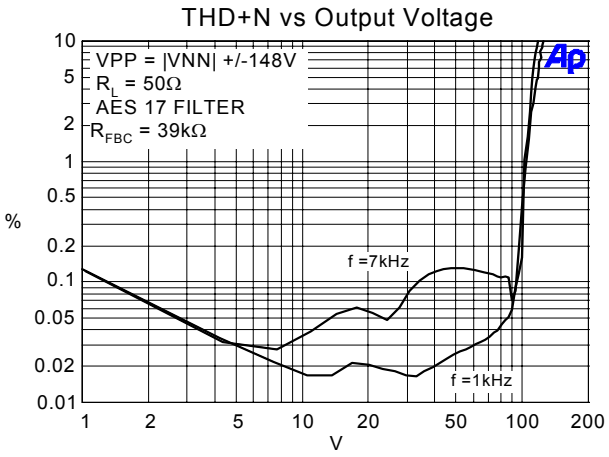


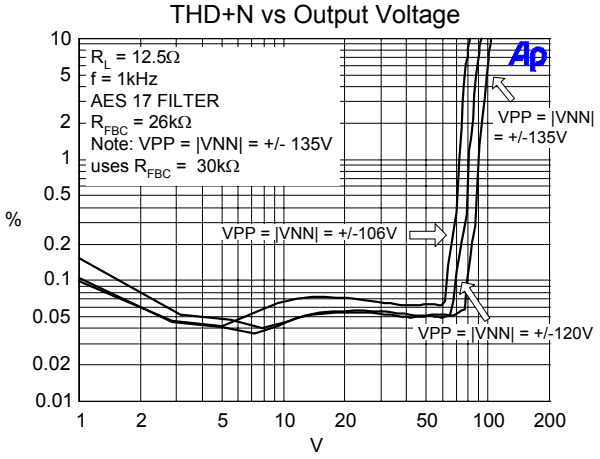
THD+N vs Output Voltage

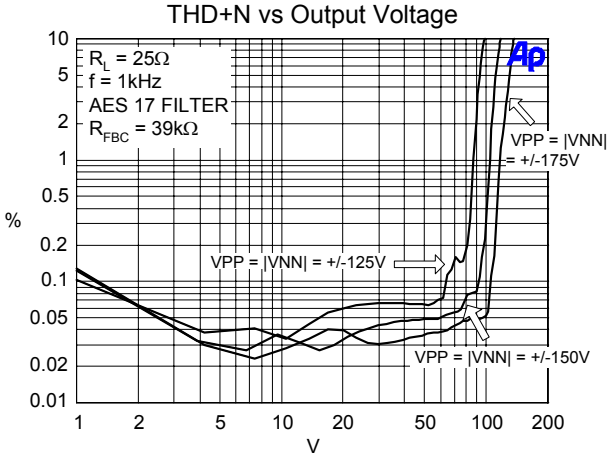


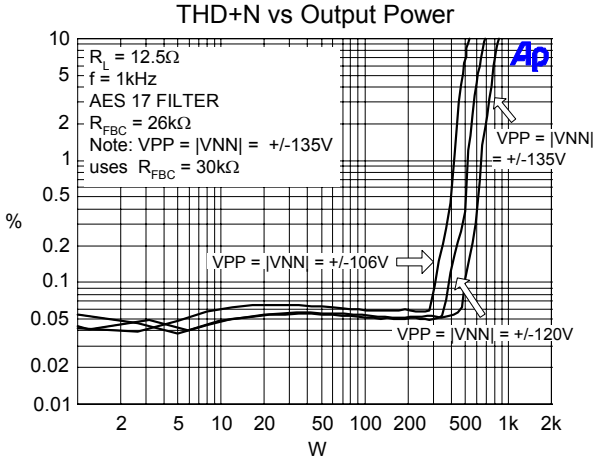
Typical Performance

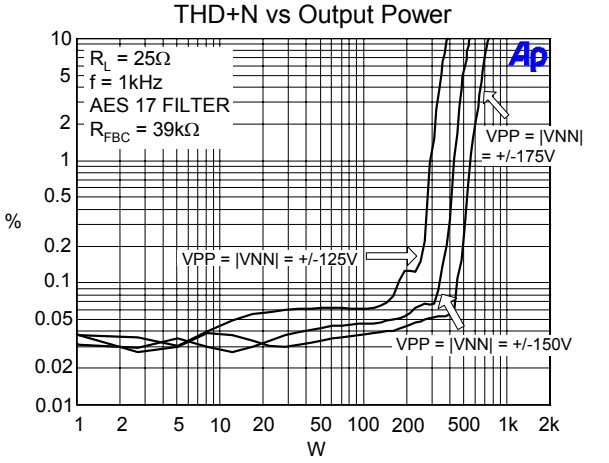




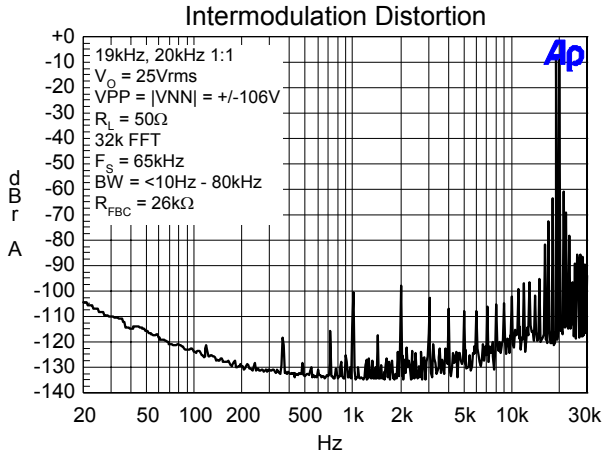


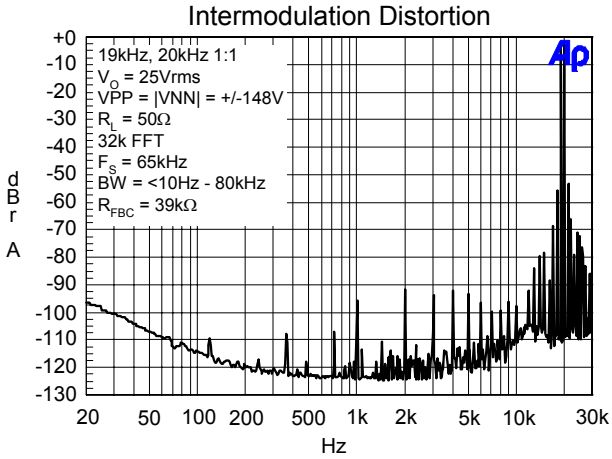


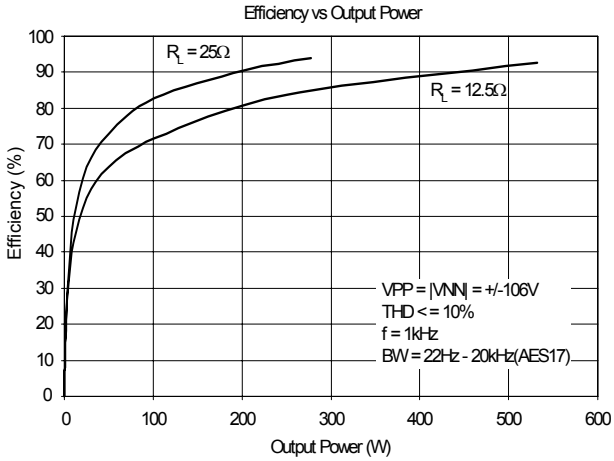


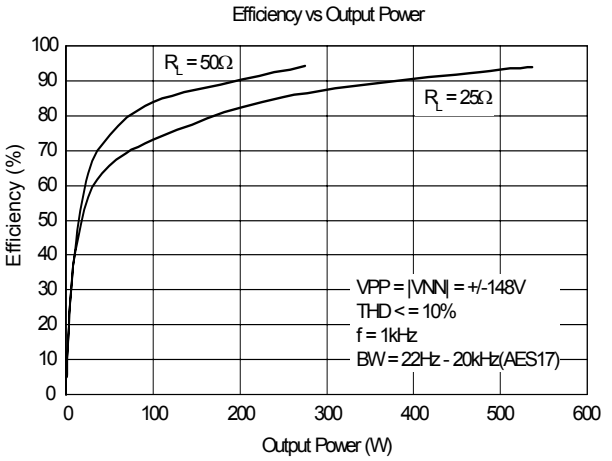


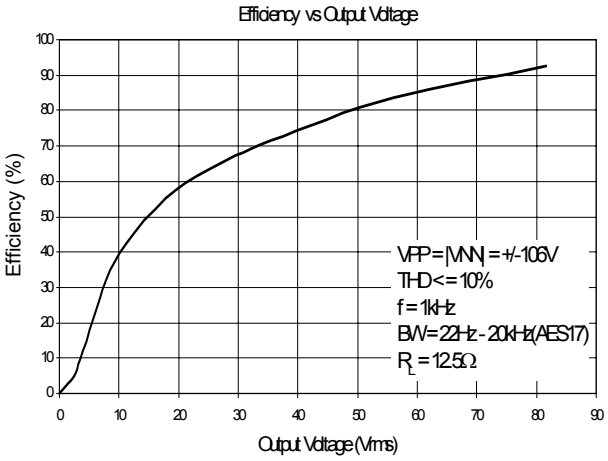
Typical Performance

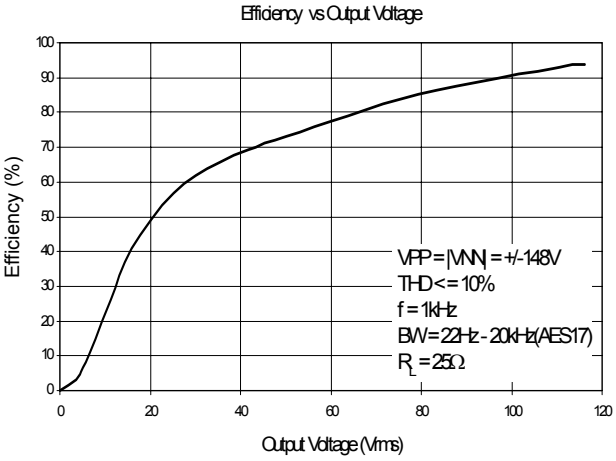












Application Information

Figure 1 is a simplified diagram of one channel (Channel 1) of a TA0105A amplifier to assist in understanding its operation.

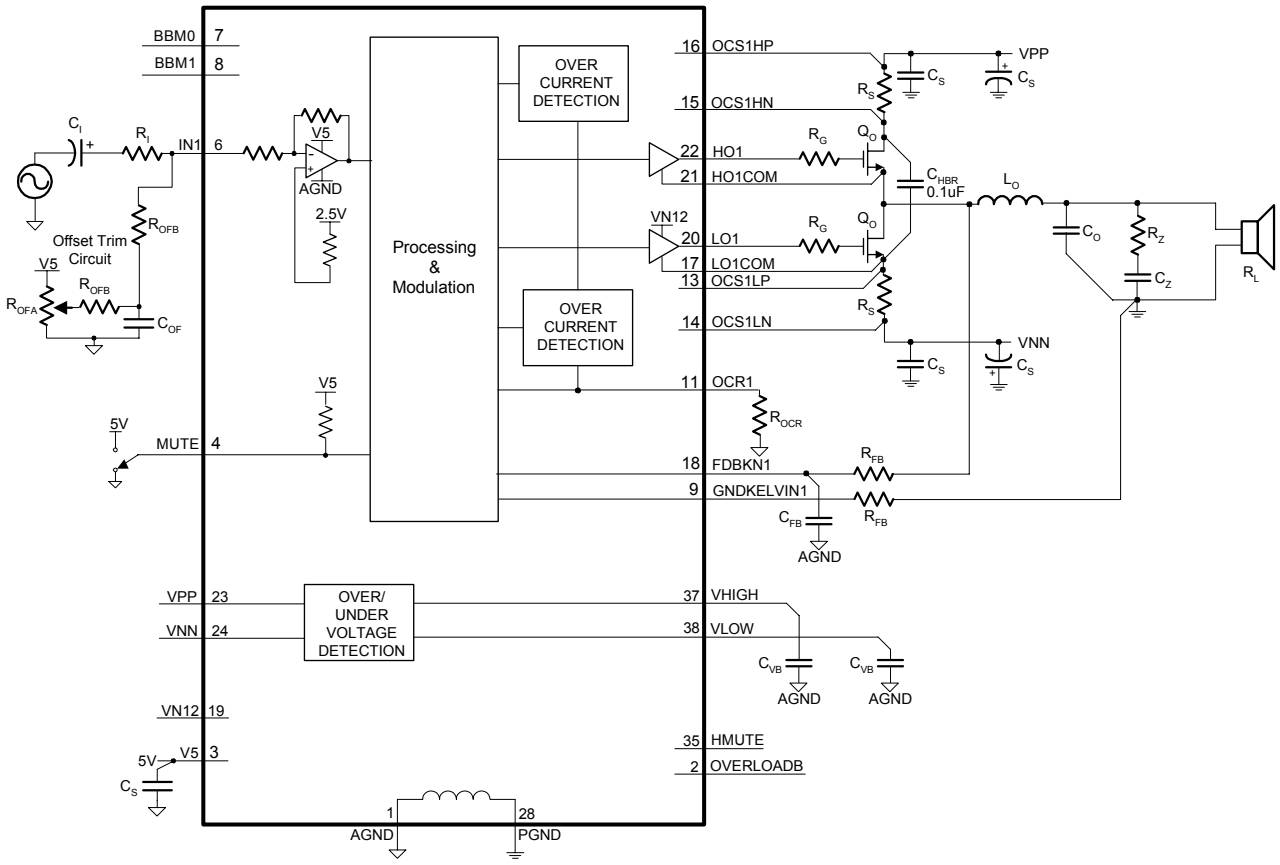


Figure 1: Simplified TA0105A Amplifier

TA0105A BASIC AMPLIFIER OPERATION

The audio input signal is fed to the processor internal to the TA0105A, where a switching pattern is generated. The average idle (no input) switching frequency is approximately 700kHz and can be adjusted by changing the C_{FB} value. The idle switching frequency must be maintained above 550kHz to ensure proper device operation. With an input signal, the pattern is spread spectrum and varies between approximately 200kHz and 1.5MHz depending on input signal level and frequency. Complementary copies of the switching pattern are level-shifted by the MOSFET drivers and output from the TA0105A where they drive the gates (HO1 and LO1) of external power MOSFETs that are connected as a half bridge. The output of the half bridge is a power-amplified version of the switching pattern that switches between VPP and VNN. This signal is then low-pass filtered to obtain an amplified reproduction of the audio input signal.

The processor portion of the TA0105A is operated from a 5-volt supply. In the generation of the switching patterns for the output MOSFETs, the processor inserts a “break-before-make” dead time between the turn-off of one transistor and the turn-on of the other in order to minimize shoot-through currents in the MOSFETs. The dead time can be programmed by setting the break-before-make control bits, BBM1 and BBM0. Feedback information from the output of the half-bridge is supplied to the processor via FBKOUT1. Additional feedback information to account for ground bounce is supplied via FBKGND1.

The MOSFET drivers in the TA0105A are operated from voltages obtained from VN12 and LO1COM for the low-side driver, and bootstrap voltage (internally generated) and HO1COM for the high-side driver. VN12 must be a regulated 12V above VNN.

N-Channel MOSFETs are used for both the top and bottom of the half bridge. The gate resistors, R_G, are used to control MOSFET slew rate and thereby minimize voltage overshoots. Though not shown, the gate diodes, D_G, reduce the MOSFET turn-off time, thus resucing cross conduction and idle supply current.

CIRCUIT BOARD LAYOUT

The TA0105A is a power (high current) amplifier that operates at relatively high switching frequencies. The output of the amplifier switches between VPP and VNN at high speeds while driving large currents. This high-frequency digital signal is passed through an LC low-pass filter to recover the amplified audio signal. Since the amplifier must drive the inductive LC output filter and speaker loads, the amplifier outputs can be pulled above the supply voltage and below ground by the energy in the output inductance. To avoid subjecting the TA0105A and external mosfets to potentially damaging voltage stress, it is critical to have a good printed circuit board layout. It is recommended that Tripath's layout and application circuit be used for all applications and only be deviated from after careful analysis of the effects of any changes. Please refer to the TA0105A evaluation board document, RB-TA0105A, available on the Tripath website, at www.tripath.com.

The following components are important to place near either their associated TA0105A or output MOSFET pins. The recommendations are ranked in order of layout importance, either for proper device operation or performance considerations.

- The impedance of the output node (the connection between the top side MOSFET source to bottom side MOSFET drain) must be minimized. Reducing the parasitic trace inductance is the most effective way of limiting output node ringing. A flat, bar conductor, in parallel with the PCB output node trace, is quite effective at minimizing the inductance thereby reducing output transients due to the switching architecture.

- The capacitors, $C_{HBR,}$ provide high frequency bypassing of the amplifier power supplies and will serve to reduce spikes and modulation of the power supply rails. Please note that both mosfet half-bridges must be decoupled separately. In addition, the voltage rating for C_{HBR} should be at least 400V as this capacitor is exposed to the full supply range, VPP-VNN.

- The output diodes, D_O, are used to minimize overshoots/undershoots on the output node. Please note that the proper connection of these is "Drain to Drain" and "Source to Source" as shown in the Application/Test Circuit. Improper routing of these diodes will render them useless due to PCB trace inductance.

- The gate resistors, R_G, should be located as close to the output MOSFET gates leads as possible. In addition, the trace length from the pins LOx/HOx to the gate resistor should be minimized. To reduce the loop area, a parallel trace from LOxCOM/HOxCOM should be routed directly to the respective MOSFET source lead.

- C_{FB} removes very high frequency components from the amplifier feedback signals and lowers the output switching frequency by delaying the feedback signals. In addition, the value of C_{FB} is different for channel 1 and channel 2 to keep the average switching frequency difference greater than 40kHz. This minimizes in-band audio noise. Locate these capacitors as close to their respective TA0105A pin as possible.

Some components are not sensitive to location but are very sensitive to layout and trace routing.

- The routing of the sense resistors, R_S, must be Kelvin connected. This implies a direct trace from the respective TA0105A pin to the sense resistor lead without interruption. If additional

connections are made to the TA0105A overcurrent sense pins or the traces, the overcurrent sense circuit may prematurely trigger.

- To maximize the damping factor and reduce distortion and noise, the modulator feedback connections should be routed directly to the pins of the output inductors. L_O. Please refer to the RB-TA0105AThis was done on the RB-TA0105A for additional information.

- The output filter capacitor, C_O, and zobel capacitor, C_Z, should be star connected with the load return. The output ground feedback signal should be taken from this star point.

- To minimize the possibility of any noise pickup, the trace lengths of IN1 and IN2 should be kept as short as possible. This is most easily accomplished by locating the input resistors, R_I as close to the TA0105A as possible. In addition, the offset trim resistor, R_{OFB}, which connects to either IN1,or IN2, should be located close to the TA0105A input section.

TA0105A GROUNDING

Proper grounding techniques are required to maximize TA0105A functionality and performance. Parametric parameters such as THD+N, Noise Floor and Crosstalk can be adversely affected if proper grounding techniques are not implemented on the PCB layout. The following discussion highlights some recommendations about grounding both with respect to the TA0105A as well as general “audio system” design rules.

The TA0105A is divided into two sections: the input section, which spans pins 1-12 and pins 35-38 and the output (high voltage) section, which spans pins 13 through pin 34. On the TA0105A evaluation board, the ground is also divided into distinct sections, one for the input and one for the output. To minimize ground loops and keep the audio noise floor as low as possible, the input and output ground should not be externally connected. They are already connected internally via a ferrite bead between pin 1 and pin 28. Additionally, any external input circuitry such as preamps, or active filters, should be referenced to pin 1.

For the power section, Tripath has traditionally used a “star” grounding scheme. Thus, the load ground returns and the power supply decoupling traces are routed separately back to the power supply. In addition, any type of shield or chassis connection would be connected directly to the ground star located at the power supply. These precautions will both minimize audible noise and enhance the crosstalk performance of the TA0105A.

The TA0105A incorporates a differential feedback system to minimize the effects of ground bounce and cancel out common mode ground noise. As such, the feedback from the output ground for each channel needs to be properly sensed. This can be accomplished by connecting the output ground “sensing” trace directly to the star formed by the output ground return, output capacitor, C_O, and the zobel capacitor, C_Z. Refer to the Application / Test Circuit for a schematic description.

TA0105A THERMAL MANAGEMENT

The bottom of the TA0105A module is a metal plate and serves as a heat sink for the internal MOSFET drivers. The temperature of this plate is directly related to the power dissipated in the output drivers. The power dissipated is broken up into two main areas, the VN12 power, and the power needed to charge the parasitic capacitances. These capacitances are internal to the MOSFET driver and the power to charge these comes from VPP and flows to VNN. Thus, as the supply voltage difference VPP-VNN increases, the amount of dissipation also increases.

Due to the increase in supply voltage, the TA0105A will run hotter than previous Tripath hybrids such as the TA0104A. Thus, depending on system airflow, and the actual power supply voltages, it may be necessary to attach an additional heat sink to the back plate or install a small fan to increase airflow directly around the hybrid. Of note, the back plate has a high impedance connection to VNN.

TA0105A AMPLIFIER GAIN

The gain of the TA0105A is the product of the input stage gain and the modulator gain. Please refer to the sections, Input Stage Design, and Modulator Feedback Design, for a complete explanation of how to determine the external component values.

$$A_{VTA0105A} = A_{VINPUTSTAGE} * A_{V\,MODULATOR}$$

$$A_{VTA0105A} \approx -\frac{20k\,\Omega}{4.99k\,\Omega + R_I}\left(\frac{(1.0k\,\Omega + R_{FB}) * 2.02}{1020} + 1\right)$$

For example, using a TA0105A with the following external components,

R_I = 34.8kΩ
R_{FB} = 39.2kΩ

$$A_{VTA0105A} \approx -\frac{20k\,\Omega}{39.79k\,\Omega}\left(\frac{40.2k\,\Omega * 2.02}{1020} + 1\right) = -40.52\ \frac{V}{V}$$

INPUT STAGE DESIGN

The TA0105A input stage is an inverting amplifier, with a maximum gain of 4. Figure 2 shows a typical application where the input stage is a constant gain inverting amplifier. The input stage gain should be set so that the maximum input signal level will drive the input stage output to 4Vpp. Please note that the input is biased between V5 and AGND. Thus, the polarity of C_I must be observed.

The gain of the input stage, above the low frequency high pass filter point, is that of a simple inverting amplifier:

$$A_{VINPUTSTAGE} = -\frac{20k\,\Omega}{4.99k\,\Omega + R_I}$$

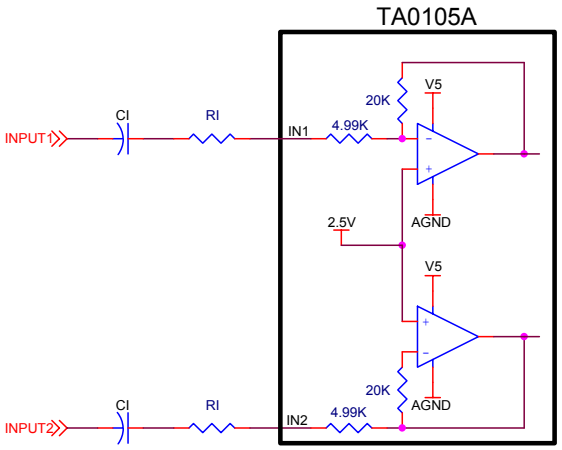


Figure 2: Input Stage

INPUT CAPACITOR SELECTION

C_I can be calculated once a value for R_{IN} has been determined. C_I and R_I determine the input low-frequency pole. Typically this pole is set at 10Hz. C_I is calculated according to:

$$C_I = 1 / (2\pi \times F_P \times R_I)$$

where: R_I = Input resistor value in ohms
F_P = Input low frequency pole (typically less than 10Hz)

MODULATOR FEEDBACK DESIGN

The modulator converts the signal from the input stage to the high-voltage output signal. The optimum gain of the modulator is determined from the maximum allowable feedback level for the modulator and maximum supply voltages for the power stage. Depending on the maximum supply voltage, the feedback ratio will need to be adjusted to maximize performance. The value of R_{FB}, in conjunction with resistors internal to the TA0105A hybrid, (see explanation below) define the gain of the modulator. Once these values are chosen, based on the maximum supply voltage, the gain of the modulator will be fixed even as the supply voltage fluctuates due to current draw.

For the best signal-to-noise ratio and lowest distortion, the maximum modulator feedback voltage should be approximately 4.5Vpp. This will keep the gain of the modulator as low as possible and still allow headroom so that the feedback signal does not clip the modulator feedback stage. It should be noted that the modulator works over basically a 2:1 supply voltage ratio with optimum performance around 3.5Vpp-4Vpp of feedback. Thus, the actual value of R_{FB} may need to be adjusted from the typical value (39.2kΩ) shown in the Application/Test Circuit to achieve maximum performance.

Figure 3 shows how the feedback from the output of the amplifier is returned to the input of the modulator. The input to the modulator (FDBKN1/GNDKELVIN1 for channel 1) can be viewed as inputs to an inverting differential amplifier. The internal 1kΩ and 1.02kΩ resistors bias the feedback signal to approximately 2.5V and R_{FB}, along with the internal series 1kΩ, scales the large output1 signal to down to approximately 4Vpp, depending on the supply voltage, VPP and VNN.

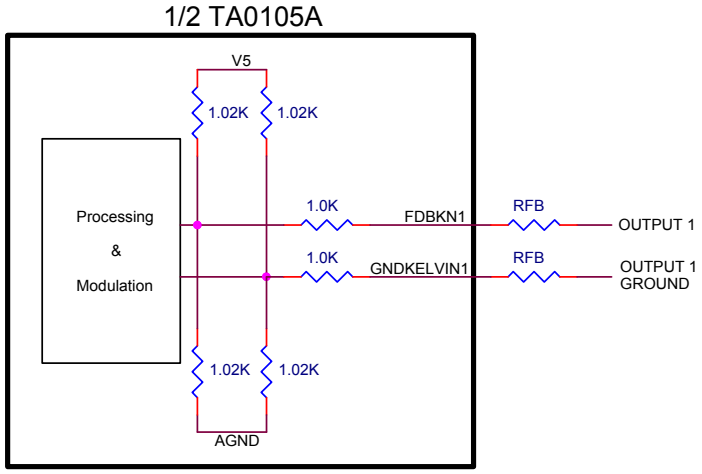


Figure 3: Modulator Feedback

The feedback resistors, R_{FB}, can be calculated using the following formula:

$$R_{FB} = \frac{1.0k\,\Omega * VPP}{4.5} - 1.0k\,\Omega$$

The above equation assumes that VPP=|VNN|.

The gain of the modulator can be calculated using the following formula:

$$A_{V\text{ - MODULATOR}} \approx \frac{(R_{FB} + 1.0k\,\Omega) * 2.02}{1020} + 1$$

For example, in a system with VPP_{MAX}=185V and VNN_{MAX}=-185V,

R_{FB} = 40.11kΩ, use 39.2kΩ, 1%

The resultant modulator gain is:

$$A_{V\text{ - MODULATOR}} \approx \frac{40.2k\,\Omega * 2.02}{1020} + 1 = 80.61V/V$$

MUTE

When a logic high signal is supplied to MUTE, both amplifier channels are muted (both high- and low-side transistors are turned off). When a logic level low is supplied to MUTE, both amplifiers are fully operational. There is a delay of approximately 200 milliseconds between the de-assertion of MUTE and the un-muting of the TA0105A. Please note that when the amplifier is in mute, the outputs are in a high impedance state and thus, the feedback resistors will set the output at approximately 2.5V without a load connected.

To ensure proper device operation, including minimization of turn on/off transients that can result in undesirable audio artifacts, Tripath recommends that the TA0105A device be muted prior to power up or power down of the 5V supply. The “sensing” of the V5 supply can be easily accomplished by using a “microcontroller supervisor” or equivalent to drive the TA0105A mute pin high when the V5 voltage is below 4.5V. This will ensure proper operation of the TA0105A input circuitry. A micro-controller supervisor such as the MCP101-450 from Microchip Corporation has been used by Tripath to implement clean power up/down operation.

HMUTE

The HMUTE pin is a 5V logic output that indicates various fault conditions within the device. These conditions include: over-current, overvoltage and undervoltage. The HMUTE output is capable of directly driving an LED through a series 2kΩ resistor.

TURN-ON & TURN-OFF NOISE

If turn-on or turn-off noise is present in a TA0105A amplifier, the cause is frequently due to other circuitry external to the TA0105A. While the TA0105A has circuitry to suppress turn-on and turn-off transients, the combination of the power supply and other audio circuitry with the TA0105A in a particular application may exhibit audible transients. In addition, a non-trimmed output offset will created an audible click on turn-on and turnoff. One solution that will completely eliminate turn-on and turn-off pops and clicks (assuming a nulled output offset) is to use a relay to connect/disconnect the amplifier from the speakers with the appropriate timing at power on/off. The relay can also be used to protect the speakers from a component failure (e.g. shorted output MOSFET). “DC protection” circuitry would need to be implemented external to the TA0105A detect such failures.

As stated in the Mute section above, a common cause of turn off pops can be attributed to the 5V supply collapsing while the other supply rails are still present. On power down, mute should be activated (pulled high) before the power supplies, especially the 5V, begin to collapse. A microcontroller supervisor, now available from multiple manufacturers, is a good way to insure proper control of the mute during power supply sequencing.

DC OFFSET

While the DC offset voltages that appear at the speaker terminals of a TA0105A amplifier are typically small, Tripath recommends that any offsets during operation be nulled out of the amplifier with a circuit like the one shown connected to IN1 and IN2 in the Application/Test Circuit. It should be noted that the DC voltage on the output of a TA0105A amplifier with no load in mute will not be zero. This offset does not need to be nulled. The output impedance of the amplifier in mute mode is approximately 40KΩ(R_FB + 1.0kΩ). This means that the DC voltage drops to essentially zero when a typical load is connected.

OVER-CURRENT PROTECTION

The TA0105A has over-current protection circuitry to protect itself and the output transistors from short-circuit conditions. The TA0105A measures the voltage across a resistor, R_S (via OCSxHP, OCSxHN, OCSxLP and OCSxLN) that is in series with each output MOSFET to detect an over-current condition. R_S and R_{OCR} are used to set the over-current threshold. The OCS pins must be Kelvin connected for proper operation. This implies connecting a trace directly from the resistor lead to the respective sense pin. No other current or power supply connections should be made to the OCS pins of the TA0105A. Doing so will result in false overcurrent events due to the IR losses of the PCB trace. See “Circuit Board Layout” in Application Information for additional details.

When the voltage across R_{OCR} becomes greater than V_{TOC} (typically 0.97) the TA0105A will shut off the output stages of its amplifiers. The occurrence of an over-current condition is latched in the TA0105A and can be cleared by toggling the MUTE input or cycling power.

SETTING OVER-CURRENT THRESHOLD

R_S and R_{OCR} determine the value of the over-current threshold, I_{OC}:

$$I_{OC} = 4990 \times (V_{TOC} – I_{BIAS} * (9100+R_{OCR}))/((9100+R_{OCR}) * R_S)$$

$$R_{OCR} = ((4990 \times V_{TOC})/(I_{OC} * R_S + 4990 * I_{BIAS})) - 9100$$

where:
R_S and R_{OCR} are in Ω
V_{TOC} = Over-current sense threshold voltage (See Electrical Characteristics Table)
= 0.97V typically
I_{BIAS} = 15uA

For example, to set an I_{OC} of 10A, R_{OCR} = 18.58KΩ (use 20KΩ, 1%) and R_S will be 10mΩ.

As high-wattage resistors are usually only available in a few low-resistance values (10mΩ, 25mΩ and 50mΩ), R_{OCR} can be used to adjust for a particular over-current threshold using one of these values for R_S.

It should be noted that the overcurrent trip level has a “duty cycle” dependence of roughly 2:1. This is due to the peak current detection (with some filtering) nature of the protection circuit implemented on the TA0105A. Thus, a current limit into a “short” will produce a peak current level roughly twice that of an over-current into a 12.5 (or higher) load. Most mosfets can withstand 3-4 times the rated continuous current for short durations (less than 100uS).

OVER- AND UNDER-VOLTAGE PROTECTION

The TA0105A senses the power rails through the VPP and VNN pins on the module. These voltages are converted to currents by internal resistor networks connected to VLOW and VHIGH. The over- and under-voltage limits are determined by the internal bias currents, the values of the resistors in the networks, along with process variations. If the supply voltage falls outside the upper and lower limits determined by the resistor networks, the TA0105A shuts off the output stages of the amplifiers. The removal of the over-voltage or under-voltage condition returns the TA0105A to normal operation. Please note that trip points specified in the Electrical Characteristics table are at 25°C and may change over temperature.

Once the supply comes back into the supply voltage operating range (as defined by the power supply sense resistors), the TA0105A will automatically be un-muted and will begin to amplify. There is a hysteresis range on both the VPP and VNN supplies. If the amplifier is powered up in the hysteresis band, the TA0105A will be muted. Thus, the usable supply range is the difference between the over-voltage restart and under-voltage restart points for both the VPP and VNN supplies. It should be noted that there is a timer of approximately 200mS with respect to the over and under voltage sensing circuit. Thus, the supply voltage must be outside of the user defined supply range for greater than 200mS for the TA0105A to be muted.

The overvoltage and undervoltage resistor values were chosen for the maximum supply range possible based on the internal hybrid components in conjunction with internal bias current settings. It is possible to lower the supply range via an external “parallel” resistor connected from VPP (pin23) to VHIGH (pin 37) and a second resistor connected from VNN (pin 24) to VLOW (pin 38). The delta between each of the trip points is a fixed ratio and not externally controllable. The current flowing into VHIGH controls the supply range for VPP while the current flowing out of VLOW controls the supply range for VNN.

The procedure for shifting the VPP range is as follows.

1) Choose the maximum VPP undervoltage turn on voltage point, $VPP_{UVTONMAX}$
2) Use the following equation to calculate the external parallel resistor, R_{VPP1}

$$R_{VPP1} = \frac{(VPP_{UVTONMAX} - 2.5V)}{80\,\mu A - (\frac{VPP_{UVTONMAX}}{1.4M\,\Omega})}$$

3) Use the following equation to calculate the resulting minimum VPP overvoltage restart point, $VPP_{OVRSTMIN}$

$$VPP_{OVRSTMIN} = \frac{2.5 * 1.4M\,\Omega + (1.4M\,\Omega * R_{VPP1} * 138\,\mu A)}{1.4M\,\Omega + R_{VPP1}}$$

4) Use the following equation to calculate the resulting maximum VPP undervoltage restart point, $VPP_{UVRSTMAX}$

$$VPP_{UVRSTMAX} = \frac{2.5 * 1.4M\,\Omega + (1.4M\,\Omega * R_{VPP1} * 86\,\mu A)}{1.4M\,\Omega + R_{VPP1}}$$

The usable (inside the hysteresis band) positive supply range is defined by $VPP_{OVRSTMIN}$ minus $VPP_{UVRSTMAX}$.

A similar procedure for shifting the VNN range is as follows.

1) Choose the maximum VNN undervoltage turn on voltage point, $VNN_{UVTONMAX}$.
2) Use the following equation to calculate the external parallel resistor, R_{VNN1}.

$$R_{VNN1} = \frac{(1.25V + |VNN_{UVTONMAX}|)}{87\,\mu A - \frac{|VNN_{UVTONMAX}|}{1.27M\,\Omega}}$$

3) Use the following equation to calculate the resulting minimum VNN overvoltage restart point, $VNN_{OVRSTMIN}$

$$VNN_{OVRSTMIN} = \frac{1.25 - R_{VNN1} * 152\,\mu A}{\frac{R_{VNN1}}{1.27M\,\Omega} + 1}$$

4) Use the following equation to calculate the resulting maximum VPP undervoltage restart point, $VNN_{UVRSTMAX}$

$$VNN_{UVRSTMAX} = \frac{1.25 - R_{VNN1} * 95\,\mu A}{\frac{R_{VNN1}}{1.27M\,\Omega} + 1}$$

The usable (inside the hysteresis band) negative supply range is defined by $VNN_{OVRSTMIN}$ minus $VNN_{UVRSTMAX}$.

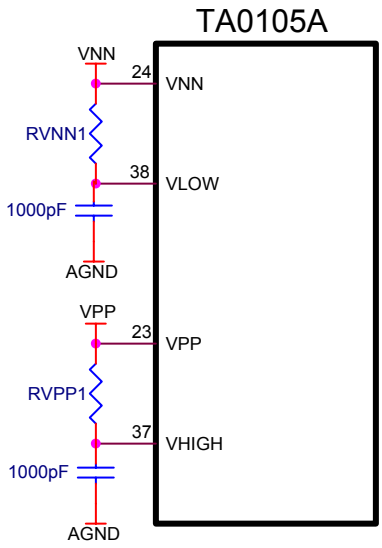


Figure 4: External Overvoltage and Undervoltage Shift

VN12 SUPPLY

VN12 is an additional supply voltage required by the TA0105A. VN12 must be 12 volts more positive than the nominal VNN. VN12 must track VNN. Generating the VN12 supply requires some care.

The proper way to generate the voltage for VN12 is to use a 12V-postive supply voltage referenced to the VNN supply. Figure 5 shows the correct way to power the TA0105A:

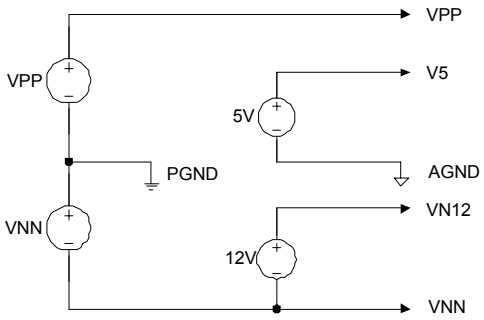


Figure 5: Proper Power Supply Connection

One apparent method to generate the VN12 supply voltage is to use a negative IC regulator to drop PGND down to 12V (relative to VNN). This method will not work since negative regulators only sink current into the regulator output and will not be capable of sourcing the current required by VN12. Furthermore, problems will arise since VN12 will not track movements in VNN.

A common approach is to use an additional secondary on the power transformer to generate an isolated, say 15VAC voltage. This AC voltage is then full bridge rectified and filtered to produce a DC input voltage for a LM7812 or similar. The "ground" of the LM7812 is then connected to VNN and thus VN12 will be properly referenced. Please refer to Figure 6.

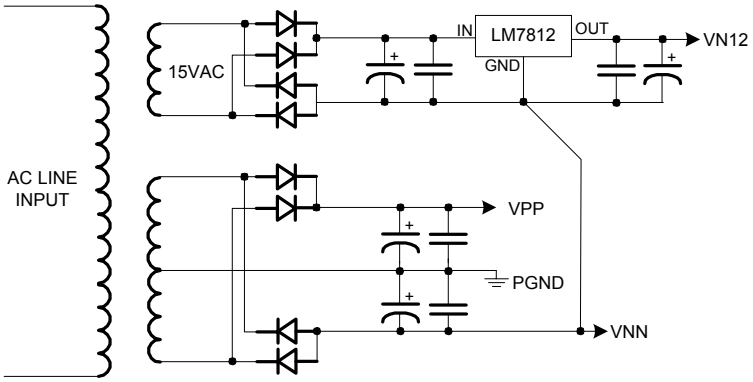


Figure 6: Proper VN12 Supply Generation

OUTPUT TRANSISTOR SELECTION

The key parameters to consider when selecting what MOSFET to use with the TA0105A are drain-source breakdown voltage (BVdss), gate charge (Qg), and on-resistance ($R_{DS(ON)}$).

The BVdss rating of the MOSFET needs to be selected to accommodate the voltage swing between V_{SPOS} and V_{SNEG} as well as any voltage peaks caused by voltage ringing due to switching transients. With a 'good' circuit board layout, a BVdss that is 25% higher than the VPP and VNN voltage swing is a reasonable starting point. The BVdss rating should be verified by measuring the actual voltages

experienced by the MOSFET in the final circuit. Thus, for TA0105A "typical" applications a mosfet with 500V rating is required.

Ideally a low Qg (total gate charge) and low $R_{DS(ON)}$ are desired for the best amplifier performance. Unfortunately, these are conflicting requirements since $R_{DS(ON)}$ is inversely proportional to Qg for a typical MOSFET. The design trade-off is one of cost versus performance. A lower $R_{DS(ON)}$ means lower $I^2R_{DS(ON)}$ losses but the associated higher Qg translates into higher switching losses (losses = Qg x 12 x 1.2MHz). A lower $R_{DS(ON)}$ also means a larger silicon die and higher cost. A higher $R_{DS(ON)}$ means lower cost and lower switching losses but higher I^2R_{DSON} losses.

The following table lists BVdss, Qg and $R_{DS(ON)}$ for MOSFETs that Tripath has used with the TA0105A.

Part Number	Manufacturer	BV_{DSS} (V)	I_D (A)	Q_g (nC)	$R_{DS(on)}$ (Ω)	P_D (W)	Package
STW20NM50FD	ST Microelectronics	500	20	38	0.22	214	TO247
STW20NM50	ST Microelectronics	500	20	40	0.22	214	TO247
STW18NB40	ST Microelectronics	400	18.4	60	0.19	190	TO247

GATE RESISTOR / GATE DIODE SELECTION

The gate resistors, R_G, are used to control MOSFET switching rise/fall times and thereby minimize voltage overshoots. They also dissipate a portion of the power resulting from moving the gate charge each time the MOSFET is switched. If R_G is too small, excessive heat can be generated in the driver. Large gate resistors lead to slower MOSFET switching, which requires a larger break-before-make (BBM) delay.

In addition, it is strongly recommended to use a schottky or ultra-fast PN junction diode in parallel with the gate resistor as shown in the Application/Test Schematic. This diode serves to "speed up" the turn-off of the output devices further reducing cross conduction and minimizing output stage idle current.

A typical gate resistor value for the mosfets recommended above is 33ohms. This resistor value assumes the use of a 1A 40V(or greater) schottky diode such as an IRF 11DQ04 or General Semiconductor SS16. Ultra fast recovery diodes will also work adequately for the gate diode, D_G.

BREAK-BEFORE-MAKE (BBM) TIMING CONTROL

The half-bridge power MOSFETs require a deadtime between when one transistor is turned off and the other is turned on (break-before-make) in order to minimize shoot through currents. BBM0 and BBM1 are logic inputs (connected to logic high or pulled down to logic low) that control the break-before-make timing of the output transistors according to the following table.

BBM1	BBM0	Delay
0	0	145 ns
0	1	105 ns
1	0	65 ns
1	1	25 ns

Table 1: BBM Delay

The tradeoff involved in making this setting is that as the delay is reduced, distortion levels improve but shoot-through and power dissipation increase. The actual amount of BBM required is dependent upon components such as MOSFET type and gate resistor value as well as circuit board layout. The BBM value selected should be verified in the actual application circuit board. It should also be verified under maximum temperature and power conditions since shoot-through in the output MOSFETs can increase under these conditions, possibly requiring a higher BBM setting than at room temperature.

OUTPUT FILTER DESIGN

One advantage of Tripath amplifiers over PWM solutions is the ability to use higher-cutoff-frequency filters. This means load-dependent peaking/droop in the 20kHz audio band potentially caused by the filter can be made negligible. Furthermore, speakers are not purely resistive loads and the impedance they present changes over frequency and from speaker model to speaker model.

Tripath recommends designing the filter as a 2nd order, LC filter. Tripath has obtained good results with L_O = 33uH and C_O = 0.22uF (resonant frequency of 59kHz). The filter capacitor must be able of sustain the ripple current caused by the high frequency switching. Thus, a high quality film capacitor is strongly recommended.

The typical application of the TA0105A is driving "high impedance" loads from 12.5 ohms and above. This dictates the use of a larger value output inductor, L_O, as compared to other Tripath amplifiers to minimize in band output filter peaking and match better to the intended load impedance.

There is a compromise between inductor value and amplifier efficiency. Tripath amplifiers count on the inductor current making "free" transitions. Take the case where the inductor current is flowing out towards the load. This is the case where there is a positive going output waveform. When the top side device turns off, the output voltage will "flip" to keep the inductor current in the same direction. If the entire transition of the output voltage (from VPP to VNN) occurs before the bottom side device is enhanced, then the transition is free. This has a positive effect on amplifier efficiency. If the bottom side device turns on before the transition is completed then power is wasted and the amplifier efficiency suffers. The output transition time is directly proportional to the inductor value and the supply voltage. Thus, larger values of inductance (for a given fet output capacitance) will result in longer transition times and decreased efficiency for a fixed supply rail. The value of L_O, 33uH, recommended above was chosen as a reasonable compromise between efficiency and load "damping." An upper bound on L_O without totally sacrificing efficiency, is 47uH for typical TA0105A supply voltages and the STW20NM50FD fets. Above this value, the designer should fully characterize the amplifier efficiency before settling on the inductor value. The peaking exhibited by a lightly loaded LC filter can be equalized out (to some degree) by an input RC filter located before the input coupling capacitor, C_I. This will result in a flatter magnitude response over a wider range of output loads. In addition, it will provide additional protection (beyond that provided by the zobel network) against high frequency signals that can cause the output filter to resonate.

The core material of the output filter inductor has an effect on the distortion levels produced by a TA0105A amplifier. Tripath recommends low-mu type-2 iron powder cores because of their low loss and high linearity (available from Micrometals, www.micrometals.com). The specific core used on the RB-TA0105A was a T106-2 wound with 49 turns of 18AWG wire.

Tripath also recommends that an RC damper be used after the LC low-pass filter. No-load operation of a TA0105A amplifier can create significant peaking in the LC filter, which produces strong resonant currents that can overheat the output MOSFETs and other components. The RC dampens the peaking and prevents problems. Tripath has obtained good results with R_D = 15Ω and C_D = 0.22uF. The zobel resistor must be able dissipate the power of the LC resonance as well as the remainder of high frequency energy that passes through the LC filter. A typical power rating for this resistor is 10W. The zobel resistor power capability will need to increased if the application requires full power at 20kHz. The zobel capacitor must be able to sustain the ripple current caused by the high frequency switching. Thus, a high quality film capacitor is recommended.

LOW-FREQUENCY POWER SUPPLY PUMPING

A potentially troublesome phenomenon in single-ended switching amplifiers is power supply pumping. This phenomenon is caused by current from the output filter inductor flowing into the power supply output filter capacitors in the opposite direction as a DC load would drain current from them. Under certain conditions (usually low-frequency input signals), this current can cause the supply voltage to "pump" (increase in magnitude) and eventually cause over-voltage/under-voltage shut down. Moreover, since

over/under-voltage are not “latched” shutdowns, the effect would be an amplifier that oscillates between on and off states. If a DC offset on the order of 0.3V is allowed to develop on the output of the amplifier (see “DC Offset Adjust”), the supplies can be boosted to the point where the amplifier’s over-voltage protection triggers.

One solution to the pumping issue it to use large power supply capacitors to absorb the pumped supply current without significant voltage boost. The low-frequency pole used at the input to the amplifier determines the value of the capacitor required. This works for AC signals only.

A no-cost solution to the pumping problem uses the fact that music has low frequency information that is correlated in both channels (it is in phase). This information can be used to eliminate boost by putting the two channels of a TA0105A amplifier out of phase with each other. This works because each channel is pumping out of phase with the other, and the net effect is a cancellation of pumping currents in the power supply. The phase of the audio signals needs to be corrected by connecting one of the speakers in the opposite polarity as the other channel.

PERFORMANCE MEASUREMENTS OF A TA0105A AMPLIFIER

Tripath amplifiers operate by modulating the input signal with a high-frequency switching pattern. This signal is sent through a low-pass filter (external to the TA0105A) that demodulates it to recover an amplified version of the audio input. The frequency of the switching pattern is spread spectrum and typically varies between 200kHz and 1.5MHz, which is well above the 20Hz – 22kHz audio band. The pattern itself does not alter or distort the audio input signal but it does introduce some inaudible noise components.

The measurements of certain performance parameters, particularly those that have anything to do with noise, like THD+N, are significantly affected by the design of the low-pass filter used on the output of the TA0105A and also the bandwidth setting of the measurement instrument used. Unless the filter has a very sharp roll-off just past the audio band or the bandwidth of the measurement instrument ends there, some of the inaudible noise components introduced by the Tripath amplifier switching pattern will get integrated into the measurement, degrading it.

Tripath amplifiers do not require large multi-pole filters to achieve excellent performance in listening tests, usually a more critical factor than performance measurements. Though using a multi-pole filter may remove high-frequency noise and improve THD+N type measurements (when they are made with wide-bandwidth measuring equipment), these same filters can increase distortion due to inductor non-linearity. Multi-pole filters require relatively large inductors, and inductor non-linearity increases with inductor value.

EMULATING A TA0104A USING A TA0105A MODULE

This following information is provided as a legacy set of instructions. Tripath has recently released a product named the TDA2500. The over-current circuit in the TDA2500 is much closer to that in the TA0104A, as compared to the TA0105A. Thus, for new designs, when trying to replace a TA0104A (or TA0102A or TA0103A), please use the TDA2500. A data sheet for the TDA2500 is available at www.tripath.com.

The TA0105A and TA0104A are structurally very similar employing the same block diagram. The TA0105A gain and Overvoltage/Undervoltage range is roughly double that of the TA0104A. The voltage rating on the TA0105A hybrid components are 200V, thus operating at lower voltages does not cause any problem assuming that the external, user selectable, components are properly chosen.

For ease of use, the “voltage shifting” components are external to theTA0105A, allowing the user to choose the voltage range, depending on the specific application. A common application is emulating a TA0104A with its associated gain and voltage range. Below is a list of instructions along with diagrams of the modifications needed to implement a “TA0104A” design. It should be noted that if some intermediate range is needed, that the feedback and overvoltage/undervoltage resistors can be adjusted based on the equations given in previous sections of the Application Information.

- Change the feedback resistors, R_{FB}, to 18.7K, 1/4W. This requires a total of four resistors (2 per channel) as both the FDBKNx and GNDKELVINx nodes need to have the series resistors inserted. This scales the amplifier feedback properly for the TA0104A supply range and results in the same gain as a standard TA0104A design. The resulting gain equation is as follows. The modifications needed for channel 1 are shown in Figure 7.

$$A_V \approx -\frac{20k\,\Omega}{4.99k\,\Omega + R_I}\left(\frac{19.7k\ * 2.02}{1020} + 1\right)$$

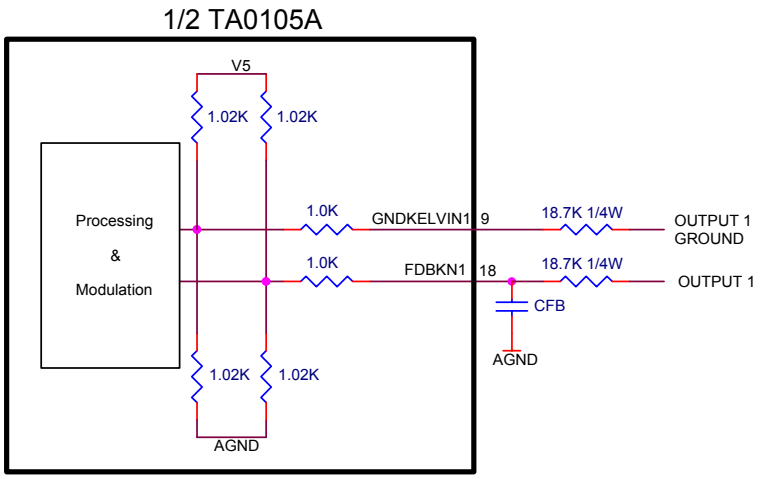


Figure 7: Feedback Structure for TA0104A Emulation

- Add the resistor dividers to both VLOW and VHIGH as shown in Figure 8. These resistors lower the supply range of the TA0105A to roughly +/-59V to +/-93V, with a maximum undervoltage turn on voltage of +/-55V, assuming worse case tolerances. It should be noted that the TA0104A voltage specification of +/-55V to +/-92V were the undervoltage and overvoltage turn on points, not the inner hysteresis band. The “hot side” of the VNN and VPP resistors should be connected to pin 24 and pin 23, respectively. Surface mount types can be used (1/8W is fine) though the resistors need to 1% tolerance. Please note that the recommended resistor values are slightly different than those used in the TA0104A. This was done intentionally to produce a symmetrical supply range for VPP and VNN. The overvoltage and undervoltage values used on the TA0104A resulted in a slightly asymmetrical voltage supply range that is clearly undesirable. Using 4 external resistors (as opposed to two shown earlier in this data sheet) results in the most symmetrical supply range possible.

-

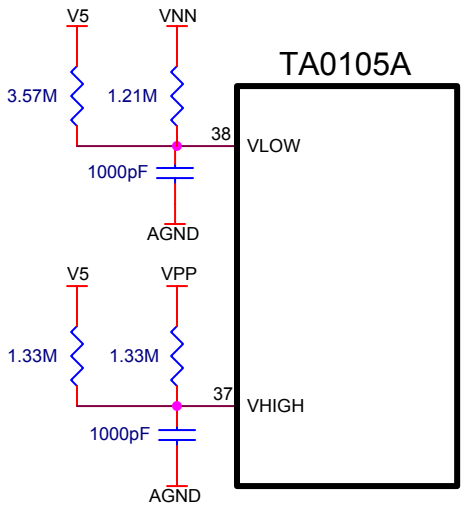


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Figure 8: Voltage Supply Sensing Structure for TA0104A Emulation

- Add the 1000pF capacitors (C_{VB}). These capacitors stabilize the sensing circuit resulting in repeatable voltage trip points. Please note that the return point is to analog ground (pin 1 on the TA0105A).

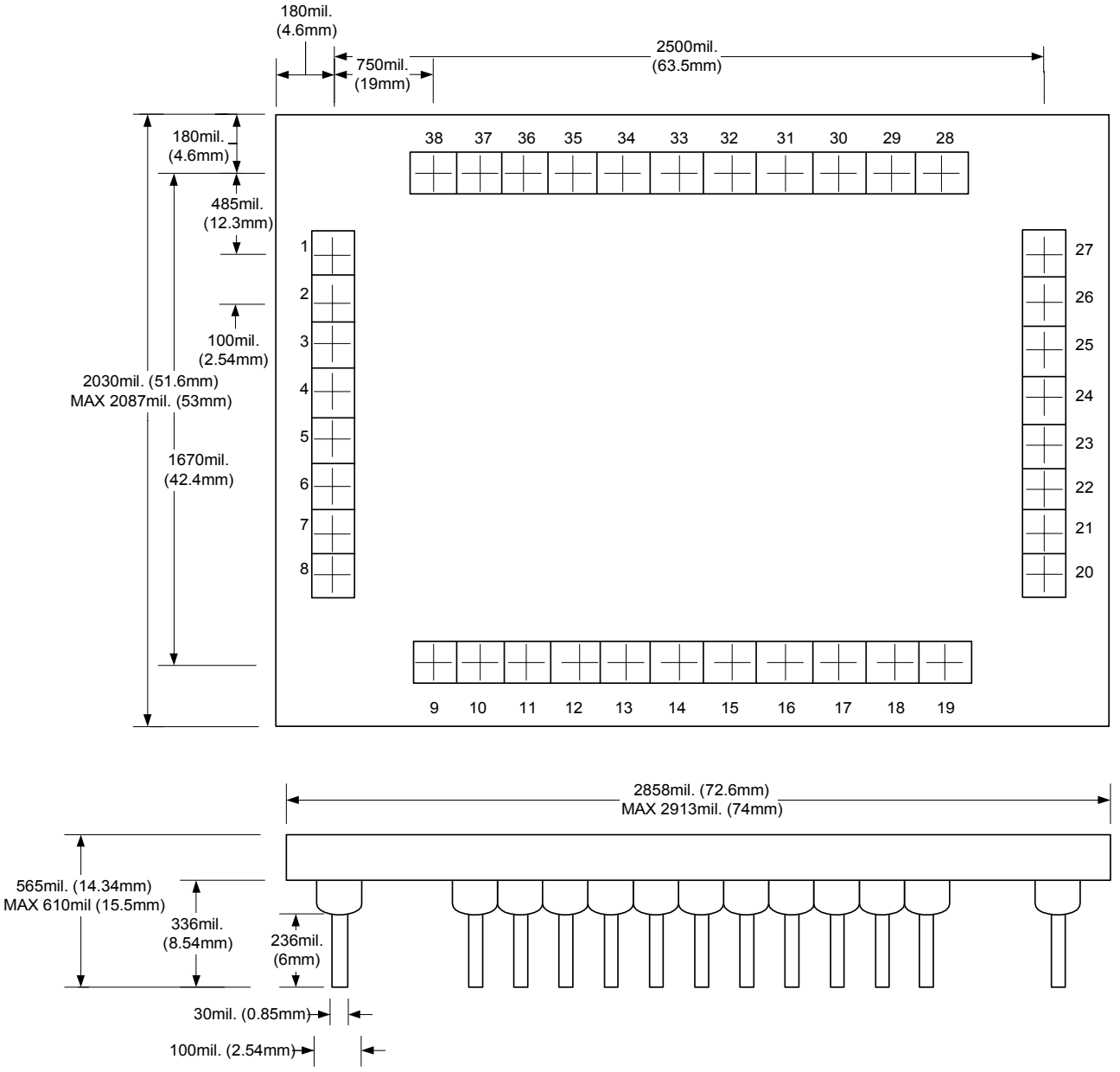
- The values of C_{FB} should be reevaluated. It is likely that the value for channel 2 will need to be increased as compared to the previous TA0104A design due to slightly different internal compensation. For best performance make sure that the difference between the two channels idle switching frequency is greater than 40kHz. In addition, make sure that the idle switching frequency of both channels is maintained above 575kHz.

- Other components such as output filter values, MOSFET type, gate resistor values, etc. should remain unchanged from the TA0104A design. Typical output filter components are 11uH, 0.22uF along with appropriate zobel compensation (15ohm/5W and 0.22uF). Typical MOSFET choice is the STW34NB20 or similar along with 5.6ohm gate resistors. The Application/Test Circuit provided earlier in this data sheet is intended for +/-185V (maximum), “high impedance” operation, not for driving low impedance loads like those typically used in TA0104A applications.

- It is highly recommended that the supply bypassing (C_{HBR}) and diode (D_O) clamping structure shown in the Application/Test circuit is utilized for new designs. This structure has been shown to minimize output node transients during high current events and will result in a more robust design.

Package Information

38 Pin Quad Module



Phyco Socket: 4150-1 x 8SF1 8 position header female
4150-1 x 1SF1 11 position header female

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For more Technical Information, please visit us @ www.tripath.com/data.htm