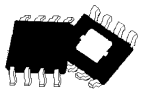
TS616

DUAL WIDE BAND OPERATIONAL AMPLIFIER WITH HIGH OUTPUT CURRENT

- **LOW NOISE :** 2.5nV/√Hz
- **HIGH OUTPUT CURRENT :** 420mA
- **VERY LOW HARMONIC AND INTERMODULATION DISTORTION**
- **HIGH SLEW RATE :** 420V/µs
- **-3dB BANDWIDTH :** 40MHz@gain=12dB on 25Ω load single ended.
- **20.7Vp-p DIFFERENTIAL OUTPUT SWING** on 50Ω load, 12V power supply
- **CURRENT FEEDBACK STRUCTURE**
- **5V to 12V POWER SUPPLY**
- **SPECIFIED FOR 20Ω and 50Ω DIFFERENTIAL LOAD**



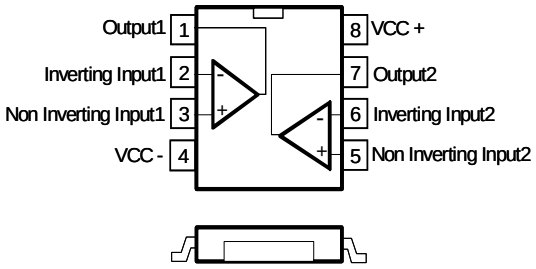
DW
SO8 Exposed-Pad
(Plastic Micro package)

ORDER CODE

Part Number	Temperature Range	Package
TS616IDW	-40, +85°C	DW
TS616IDWT	-40, +85°C	DW

DW = Small Outline Package with Exposed-Pad, **T** = Tape & Real

PIN CONNECTIONS (top view)



Cross Section View Showing Exposed-Pad
This pad can be connected to a (-Vcc) copper area on the PCB

DESCRIPTION

The TS616 is a dual operational amplifier featuring a high output current of 410mA. The drivers can be configured differentially for driving signals in telecommunication systems using multiple carriers. The TS616 is ideally suited for xDSL (High Speed Asymmetrical Digital Subscriber Line) applications. This circuit is capable of driving a 10Ω or 25Ω load at ±2.5V, 5V, ±6V or +12V power supply. The TS616 is able to reach a -3dB bandwidth of 40MHz on 25Ω load with a 12dB gain. This device is designed for high slew rates supporting low harmonic distortion and intermodulation.

APPLICATION

- Line driver for xDSL
- Multiple Video Line Driver

ABSOLUTE MAXIMUM RATINGS

Symbol	Parameter	Value	Unit
V_{CC}	Supply voltage [1]	±7	V
V_{id}	Differential Input Voltage [2]	±2	V
V_{in}	Input Voltage Range [3]	±6	V
T_{oper}	Operating Free Air Temperature Range	-40 to + 85	°C
T_{std}	Storage Temperature	-65 to +150	°C
T_j	Maximum Junction Temperature	150	°C
R_{thjc}	Thermal Resistance Junction to Case	16	°C/W
R_{thja}	Thermal Resistance Junction to Ambient Area	60	°C/W
$P_{max.}$	Maximum Power Dissipation (@Ta=25°C) for Tj=150°C	2	W
ESD only pins 1, 4, 7, 8	CDM : Charged Device Model	1.5	kV
	HBM : Human Body Model	2	kV
	MM : Machine Model	200	V
ESD only pins 2, 3, 5, 6	CDM : Charged Device Model	1.5	kV
	HBM : Human Body Model	2	kV
	MM : Machine Model	100	V
	Output Short Circuit	[4]	

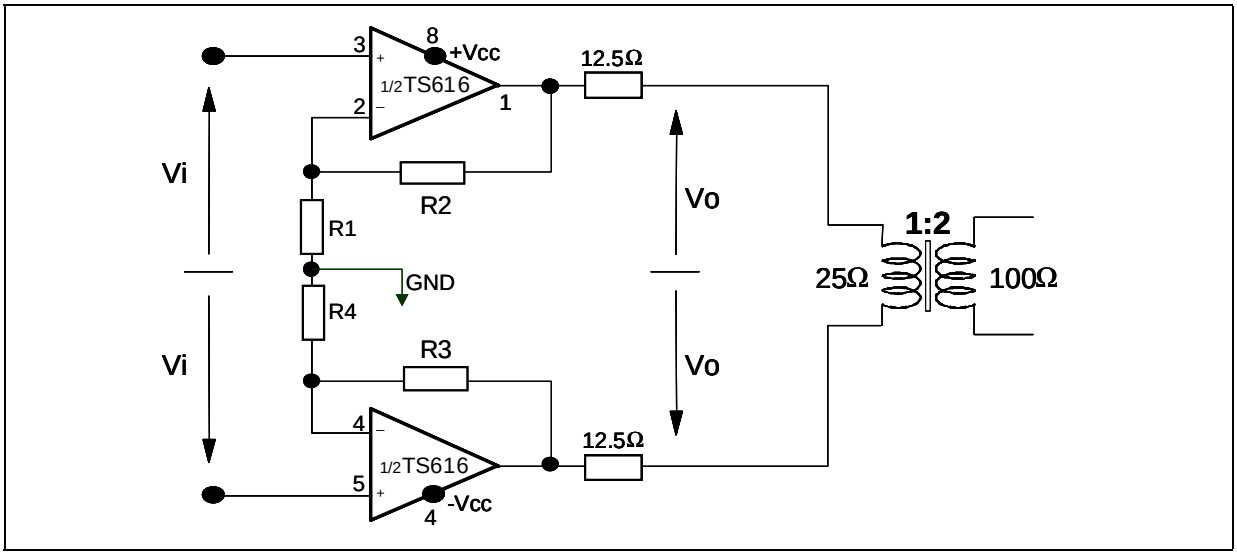
1. All voltage values, except differential voltage are with respect to network terminal.
2. Differential voltage are non-inverting input terminal with respect to the inverting input terminal.
3. The magnitude of input and output voltage must never exceed V_{CC} +0.3V.
4. An output current limitation protects the circuit from transient currents. Short-circuits can cause excessive heating. Destructive dissipation can result from short circuit on amplifiers.

OPERATING CONDITIONS

Symbol	Parameter	Value	Unit
V_{CC}	Power Supply Voltage	±2.5 to ±6	V
V_{icm}	Common Mode Input Voltage	$-V_{CC}$+1.5V to $+V_{CC}$-1.5V	V

TYPICAL APPLICATION:

Differential Line Driver for xDSL Applications



ELECTRICAL CHARACTERISTICS

V_{CC} = ±6Volts, R_{fb}=910Ω,T_{amb} = 25°C (unless otherwise specified)

Note: As described on page 24 (table 71), the TS616 requires a 620Ω feedback resistor for an optimized bandwidth with a gain of 12B for a 12V power supply. Nevertheless, due to production test constraints, the TS616 is tested with the same feedback resistor for 12V and 5V power supplies (910Ω).

Symbol	Parameter	Test Condition	Min.	Typ.	Max.	Unit
DC PERFORMANCE						
V_{io}	Input Offset Voltage	T_{amb}		1	3.5	mV
		$T_{min.} < T_{amb} < T_{max.}$		1.6		
ΔV_{io}	Differential Input Offset Voltage	T_{amb} = 25°C			2.5	mV
I_{ib+}	Positive Input Bias Current	T_{amb}		5	30	µA
		$T_{min.} < T_{amb} < T_{max.}$		7.2		
I_{ib-}	Negative Input Bias Current	T_{amb}		3	15	µA
		$T_{min.} < T_{amb} < T_{max.}$		3.1		
Z_{IN+}	Input(+) Impedance			82		kΩ
Z_{IN-}	Input(-) Impedance			54		Ω
C_{IN+}	Input(+) Capacitance			1		pF
CMR	Common Mode Rejection Ratio	ΔV_{ic} = ±4.5V	58	64		dB
	20 log ($\Delta V_{ic}/\Delta V_{io}$)	$T_{min.} < T_{amb} < T_{max.}$		62		
SVR	Supply Voltage Rejection Ratio	ΔV_{cc}=±2.5V to ±6V	72	81		dB
	20 log ($\Delta V_{cc}/\Delta V_{io}$)	$T_{min.} < T_{amb} < T_{max.}$		80		
I_{CC}	Total Supply Current per Operator	No load		13.5	17	mA
DYNAMIC PERFORMANCE and OUTPUT CHARACTERISTIC						
R_{OL}	Open Loop Transimpedance	V_{out} = 7Vp-p, R_L = 25Ω	5	13.5		MΩ
		$T_{min.} < T_{amb.} < T_{max.}$		5.7		
BW	-3dB Bandwidth	Small Signal V_{out}<20mVp A_V = 12dB, R_L = 25Ω	25	40		MHz
	Full Power Bandwidth	Large Signal V_{out}=3Vp A_V = 12dB, R_L = 25Ω		26		
	Gain Flatness @ 0.1dB	Small Signal V_{out}<20mVp A_V = 12dB, R_L = 25Ω		7		MHz
Tr	Rise Time	V_{out} = 6Vp-p, A_V = 12dB, R_L = 25Ω		10.6		ns
Tf	Fall Time	V_{out} = 6Vp-p, A_V = 12dB, R_L = 25Ω		12.2		ns
Ts	Settling Time	V_{out} = 6Vp-p, A_V = 12dB, R_L = 25Ω		50		ns
SR	Slew Rate	V_{out} = 6Vp-p, A_V = 12dB, R_L = 25Ω	330	420		V/µs
V_{OH}	High Level Output Voltage	R_L=25Ω Connected to GND	4.8	5.05		V
V_{OL}	Low Level Output Voltage	R_L=25Ω Connected to GND		-5.3	-5.1	V
I_{out}	Output Sink Current	V_{out} = -4Vp	-320	-490		mA
		$T_{min.} < T_{amb} < T_{max.}$		-395		
	Output Source Current	V_{out} = +4Vp	330	420		
		$T_{min.} < T_{amb} < T_{max.}$		370		

Note: As described on page 24 (table 71), the TS616 requires a 620Ω feedback resistor for an optimized bandwidth with a gain of 12B for a 12V power supply. Nevertheless, due to production test constraints, the TS616 is tested with the same feedback resistor for 12V and 5V power supplies (910Ω).

Symbol	Parameter	Test Condition	Min.	Typ.	Max.	Unit
NOISE AND DISTORTION						
eN	Equivalent Input Noise Voltage	F = 100kHz		2.5		nV/√Hz
iNp	Equivalent Input Noise Current (+)	F = 100kHz		15		pA/√Hz
iNn	Equivalent Input Noise Current (-)	F = 100kHz		21		pA/√Hz
HD2	2nd Harmonic Distortion (differential configuration)	V_{out} = 14Vp-p, A_V = 12dB F= 110kHz, R_L = 50Ω diff.		-87		dBc
HD3	3rd Harmonic Distortion (differential configuration)	V_{out} = 14Vp-p, A_V = 12dB F= 110kHz, R_L = 50Ω diff.		-83		dBc
IM2	2nd Order Intermodulation Product (differential configuration)	F1= 100kHz, F2 = 110kHz V_{out} = 16Vp-p, A_V = 12dB R_L = 50Ω diff.		-76		dBc
		F1= 370kHz, F2 = 400kHz V_{out} = 16Vp-p, A_V = 12dB R_L = 50Ω diff.		-75		
IM3	3rd Order Intermodulation Product (differential configuration)	F1 = 100kHz, F2 = 110kHz V_{out} = 16Vp-p, A_V = 12dB R_L = 50Ω diff.		-88		dBc
		F1 = 370kHz, F2 = 400kHz V_{out} = 16Vp-p, A_V = 12dB R_L = 50Ω diff.		-87		

ELECTRICAL CHARACTERISTICS

V_{CC} = ±2.5Volts, R_{fb}=910Ω,T_{amb} = 25°C (unless otherwise specified)

Symbol	Parameter	Test Condition	Min.	Typ.	Max.	Unit
DC PERFORMANCE						
V_{io}	Input Offset Voltage	T_{amb}		0.2	2.5	mV
		$T_{min.} < T_{amb} < T_{max.}$		1		
ΔV_{io}	Differential Input Offset Voltage	T_{amb} = 25°C			2.5	mV
I_{ib+}	Positive Input Bias Current	T_{amb}		4	30	µA
		$T_{min.} < T_{amb} < T_{max.}$		7		
I_{ib-}	Negative Input Bias Current	T_{amb}		1.1	11	µA
		$T_{min.} < T_{amb} < T_{max.}$		1.2		
Z_{IN+}	Input(+) Impedance			71		kΩ
Z_{IN-}	Input(-) Impedance			62		Ω
C_{IN+}	Input(+) Capacitance			1.5		pF
CMR	Common Mode Rejection Ratio	ΔV_{ic} = ±1V	55	61		dB
	20 log ($\Delta V_{ic}/\Delta V_{io}$)	$T_{min.} < T_{amb.} < T_{max.}$		60		
SVR	Supply Voltage Rejection Ratio	ΔV_{cc}=±2V to ±2.5V	63	79		dB
	20 log ($\Delta V_{cc}/\Delta V_{io}$)	$T_{min.} < T_{amb.} < T_{max.}$		78		
I_{CC}	Total Supply Current per Operator	No load		11.5	15	mA
DYNAMIC PERFORMANCE and OUTPUT CHARACTERISTICS						
R_{OL}	Open Loop Transimpedance	V_{out} = 2Vp-p, R_L = 10Ω	2	4.2		MΩ
		$T_{min.} < T_{amb.} < T_{max.}$		1.5		
BW	-3dB Bandwidth	Small Signal V_{out}<20mVp A_V = 12dB, R_L = 10Ω	20	28		MHz
	Full Power Bandwidth	Large Signal V_{out}= 1.4Vp A_V = 12dB, R_L = 10Ω		20		
	Gain Flatness @ 0.1dB	Small Signal V_{out}<20mVp A_V = 12dB, R_L = 10Ω		5.7		MHz
Tr	Rise Time	V_{out} = 2.8Vp-p, A_V = 12dB R_L = 10Ω		11		ns
Tf	Fall Time	V_{out} = 2.8Vp-p, A_V = 12dB R_L = 10Ω		11.5		ns
Ts	Settling Time	V_{out} = 2.2Vp-p, A_V = 12dB R_L = 10Ω		39		ns
SR	Slew Rate	V_{out} = 2.2Vp-p, A_V = 12dB R_L = 10Ω	100	130		V/µs
V_{OH}	High Level Output Voltage	R_L=10Ω Connected to GND	1.5	1.7		V
V_{OL}	Low Level Output Voltage	R_L=10Ω Connected to GND		-1.9	-1.7	V
I_{out}	Output Sink Current	V_{out} = -1.25Vp	-300	-400		mA
		$T_{min.} < T_{amb} < T_{max.}$		-360		
	Output Source Current	V_{out} = +1.25Vp	200	270		
		$T_{min.} < T_{amb} < T_{max.}$		240		

Symbol	Parameter	Test Condition	Min.	Typ.	Max.	Unit
NOISE AND DISTORTION						
eN	Equivalent Input Noise Voltage	F = 100kHz		2.5		nV/√Hz
iNp	Equivalent Input Noise Current (+)	F = 100kHz		15		pA/√Hz
iNn	Equivalent Input Noise Current (-)	F = 100kHz		21		pA/√Hz
HD2	2nd Harmonic Distortion (differential configuration)	V_{out} = 6Vp-p, A_V = 12dB F= 110kHz, R_L = 20Ω diff.		-97		dBc
HD3	3rd Harmonic Distortion (differential configuration)	V_{out} = 6Vp-p, A_V = 12dB F= 110kHz, R_L = 20Ω diff.		-98		dBc
IM2	2nd Order Intermodulation Product (differential configuration)	F1= 100kHz, F2 = 110kHz V_{out} = 6Vp-p, A_V = 12dB R_L = 20Ω diff.		-86		dBc
		F1= 370kHz, F2 = 400kHz V_{out} = 6Vp-p, A_V = 12dB R_L = 20Ω diff.		-88		
IM3	3rd Order Intermodulation Product (differential configuration)	F1 = 100kHz, F2 = 110kHz V_{out} = 6Vp-p, A_V = 12dB R_L = 20Ω diff.		-90		dBc
		F1 = 370kHz, F2 = 400kHz V_{out} = 6Vp-p, A_V = 12dB R_L = 20Ω diff.		-85		

Figure 1: Load Configuration
Load: R_L=25Ω, V_{CC}=±6V

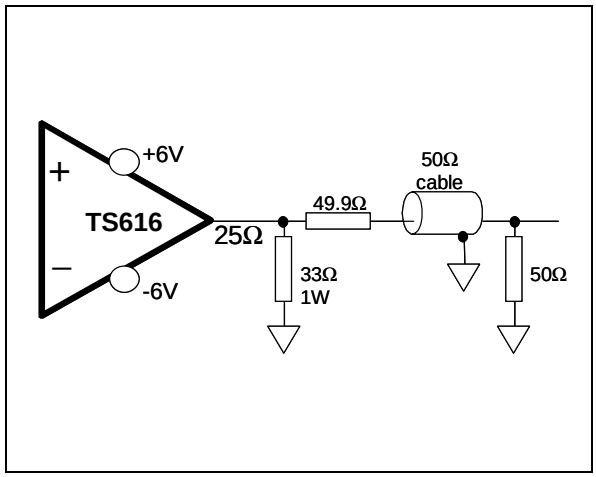


Figure 2: Closed Loop Gain vs. Frequency
A_V=+1

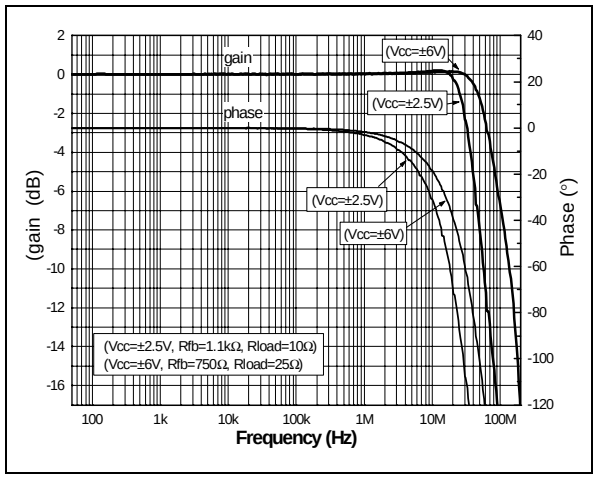


Figure 3: Closed Loop Gain vs. Frequency
A_V=+2

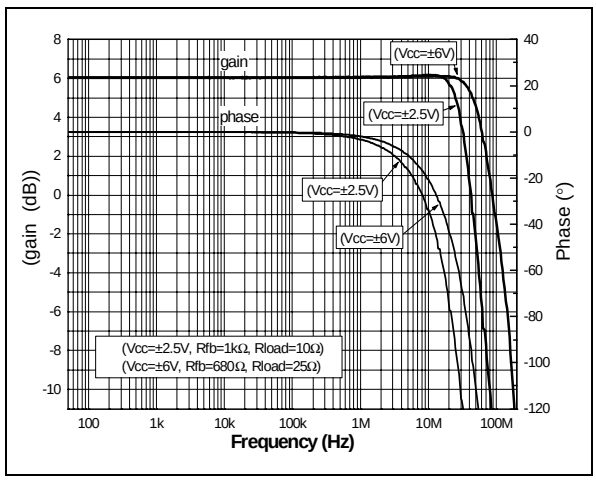


Figure 4: Load Configuration
Load: R_L=10Ω, V_{CC}=±2.5V

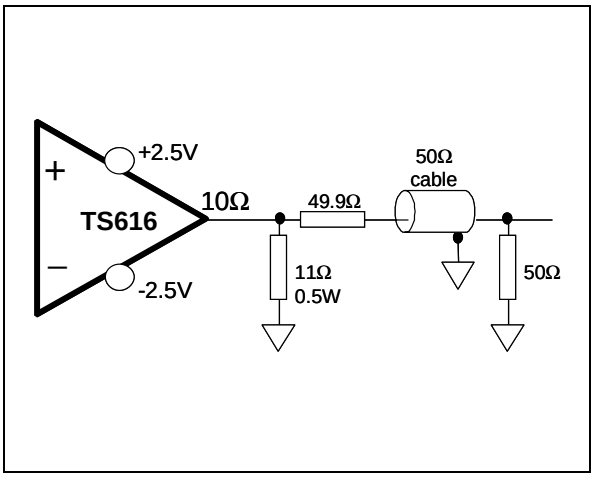


Figure 5: Closed Loop Gain vs. Frequency
A_V=-1

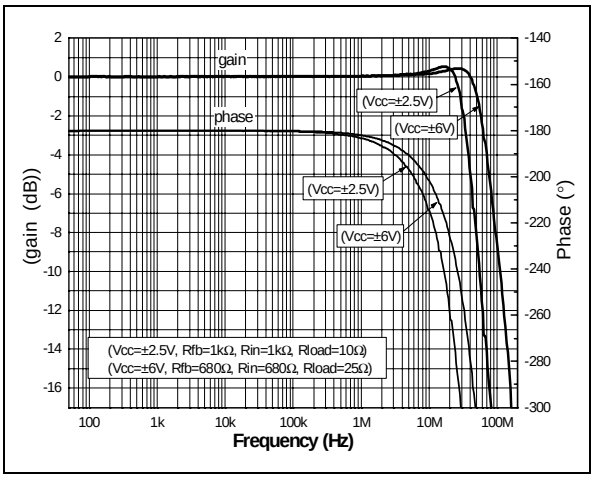


Figure 6: Closed Loop Gain vs. Frequency
A_V=-2

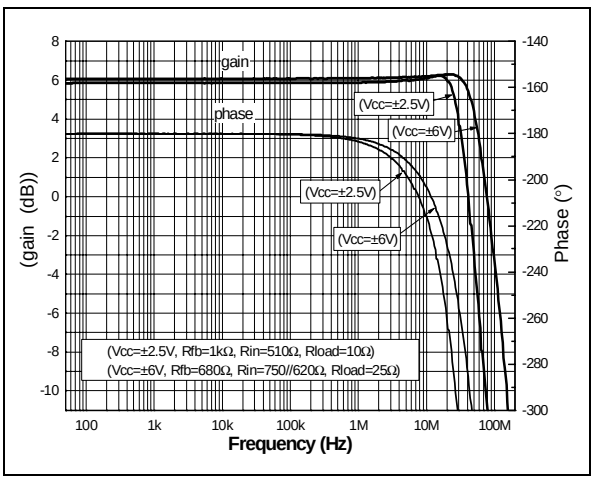


Figure 7: Closed Loop Gain vs. Frequency
A_V=+4

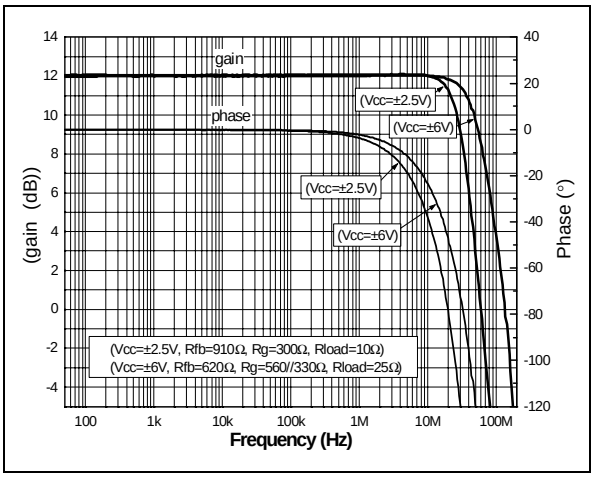


Figure 8: Closed Loop Gain vs. Frequency
A_V=+8

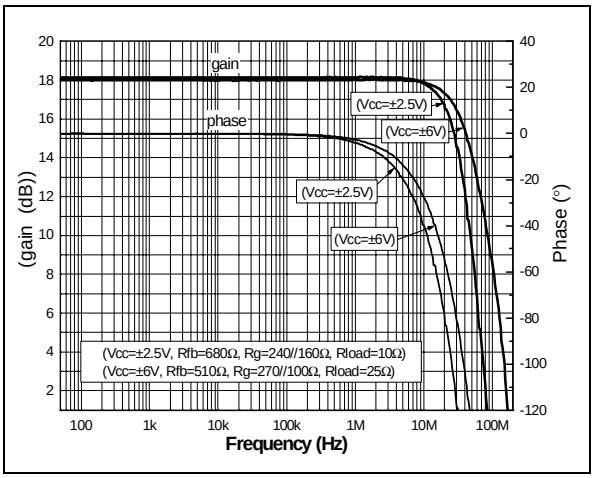


Figure 9: Bandwidth vs. Temperature
A_V=+4, R_{fb}=910Ω

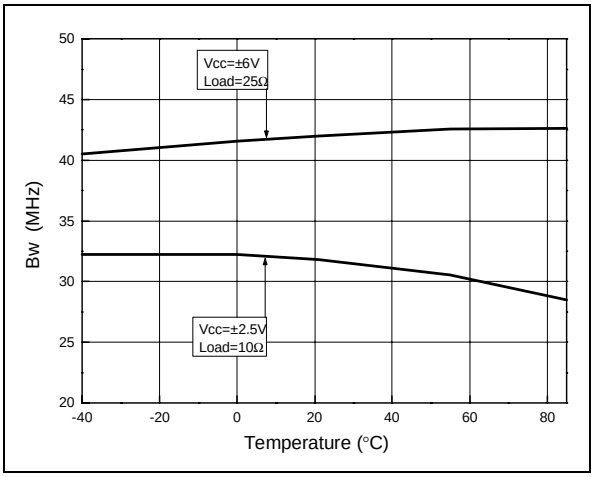


Figure 10: Closed Loop Gain vs. Frequency
A_V=-4

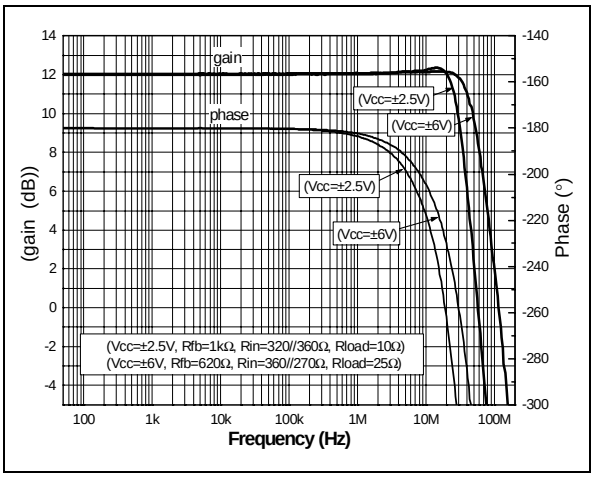


Figure 11: Closed Loop Gain vs. Frequency
A_V=-8

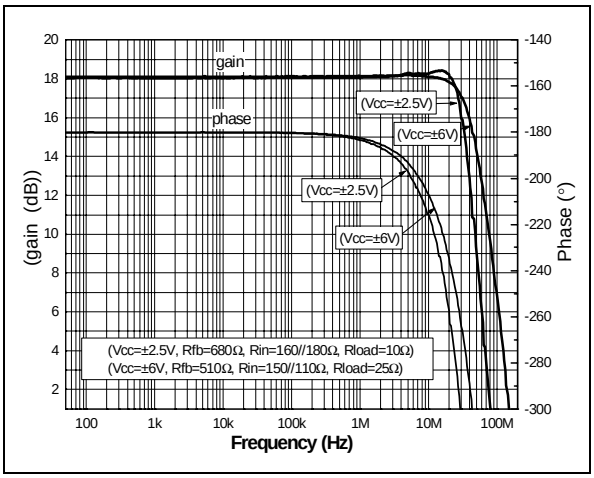


Figure 12: Positive Slew Rate
A_V=+4, R_{fb}=620Ω, V_{CC}=±6V, R_L=25Ω

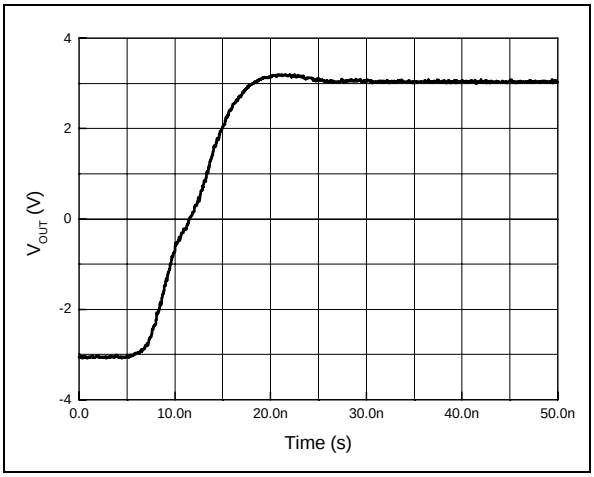


Figure 13: Positive Slew Rate
A_V=+4, R_{fb}=910Ω, V_{CC}=±2.5V, R_L=10Ω

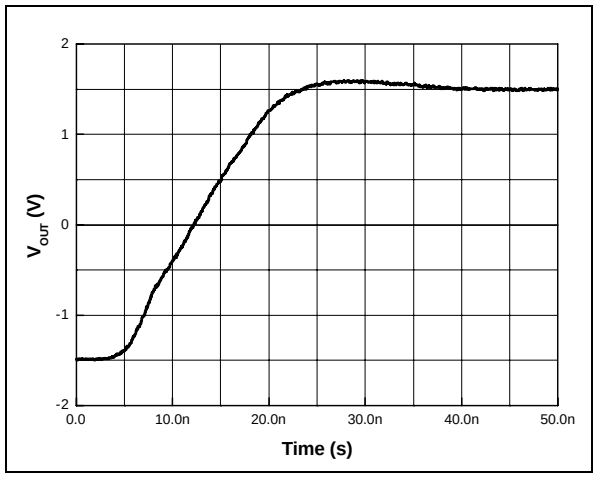


Figure 14: Negative Slew Rate
A_V=+4, R_{fb}=620Ω, V_{CC}=±6V, R_L=25Ω

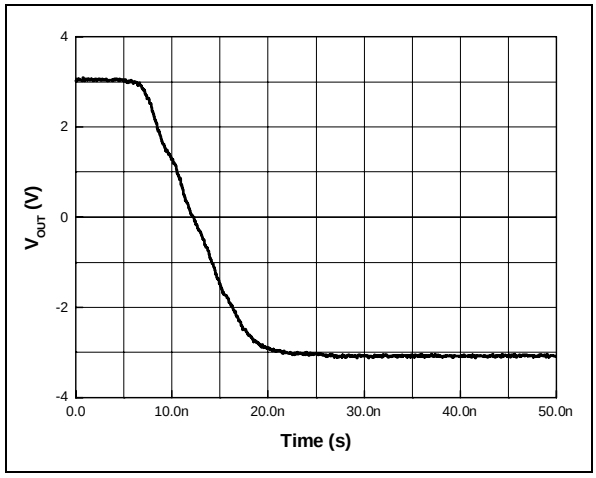


Figure 15: Negative Slew Rate
A_V=+4, R_{fb}=910Ω, V_{CC}=±2.5V, R_L=10Ω

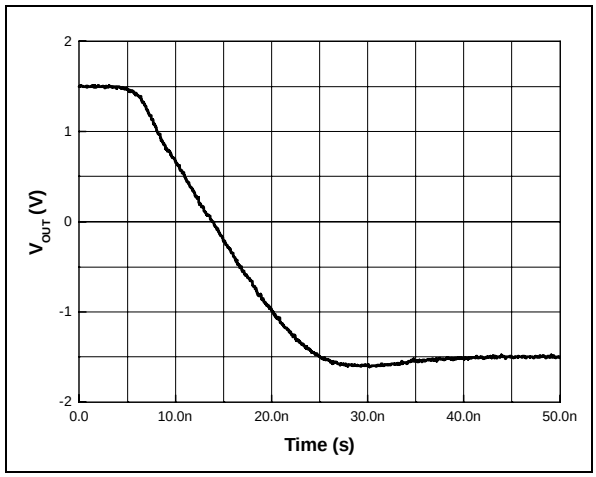


Figure 16: Positive Slew Rate
A_V= - 4, R_{fb}=620Ω, V_{CC}=±6V, R_L=25Ω

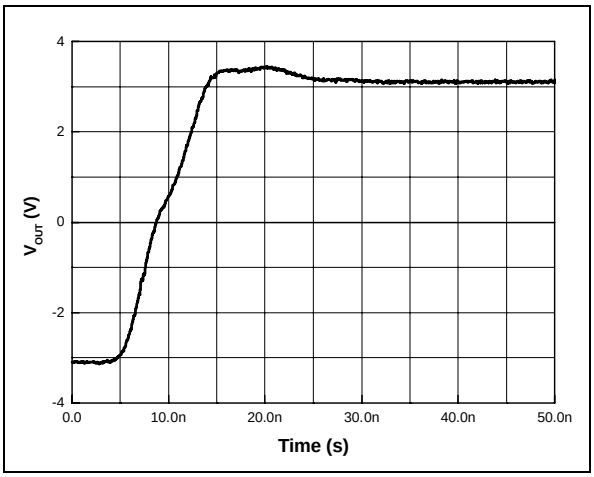


Figure 17: Positive Slew Rate
A_V= - 4, R_{fb}=910Ω, V_{CC}=±2.5V, R_L=10Ω

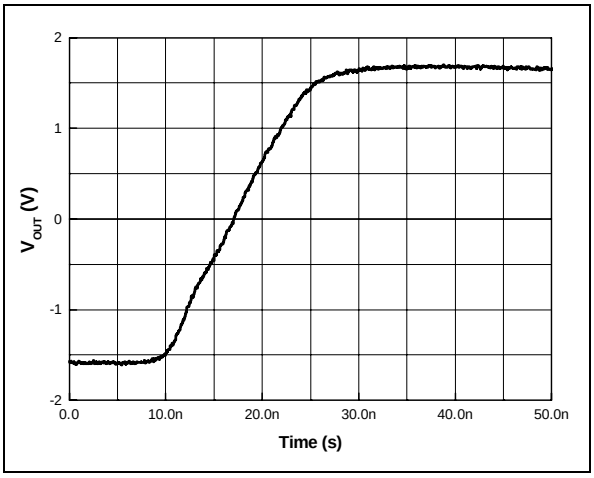


Figure 18: Negative Slew Rate
A_V= - 4, R_{fb}=620Ω, V_{CC}=±6V, R_L=25Ω

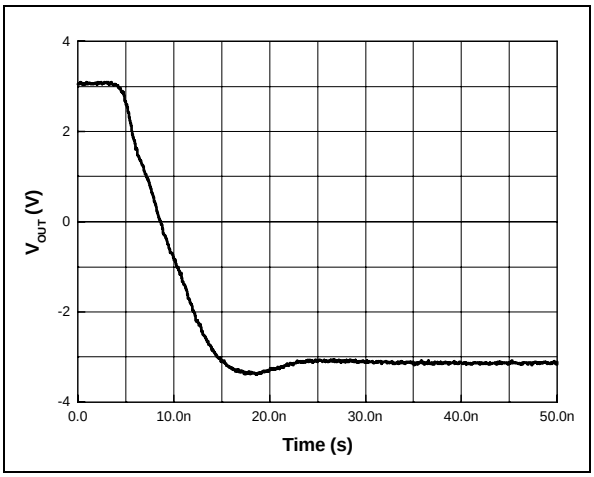


Figure 19: Negative Slew Rate
A_V= - 4, R_{fb}=910Ω, V_{CC}=±2.5V, R_L=10Ω

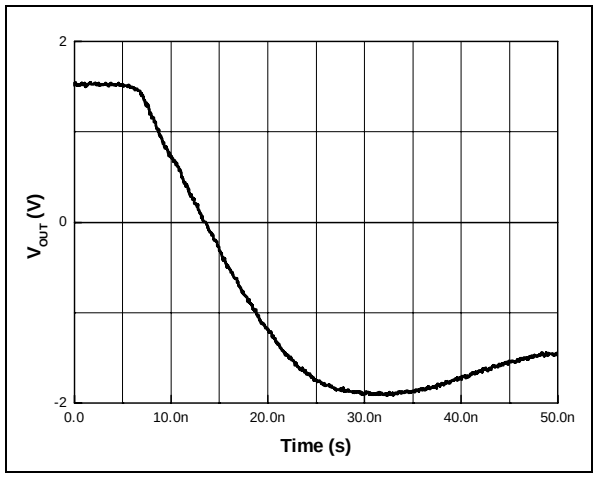


Figure 20: Slew Rate vs. Temperature
A_V=+4, R_{fb}=910Ω, V_{CC}=±2.5V, R_L=10Ω

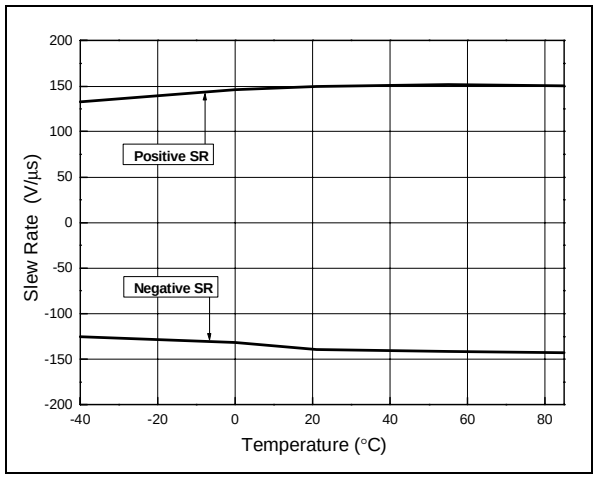


Figure 21: Slew Rate vs. Temperature
A_V=+4, R_{fb}=910Ω, V_{CC}=±6V, R_L=25Ω

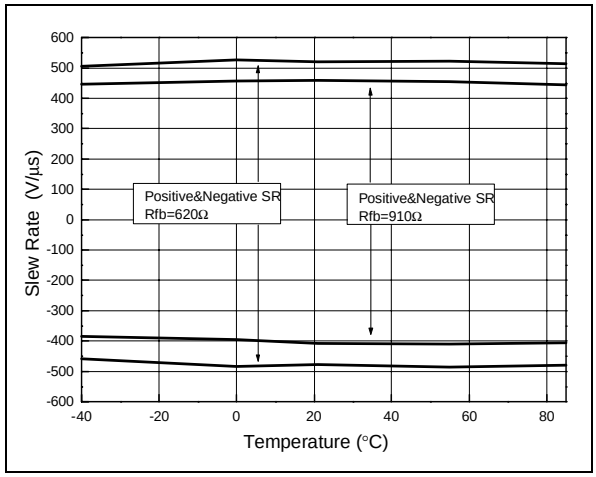


Figure 22: Input Voltage Noise Level
A_V=+92, R_{fb}=910Ω, Input+ connected to Gnd via 10Ω

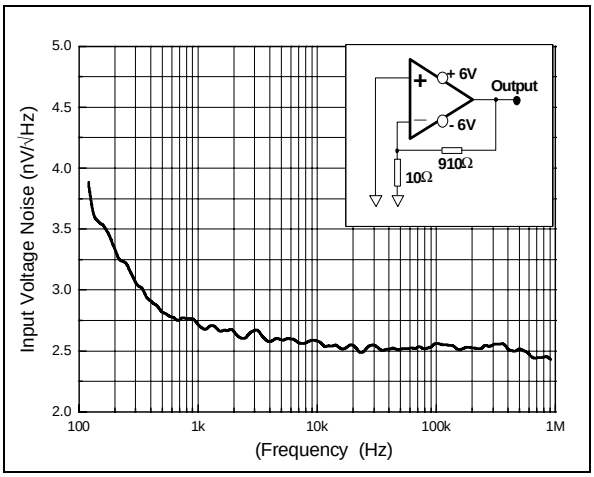


Figure 23: Transimpedance vs. Temperature
Open Loop

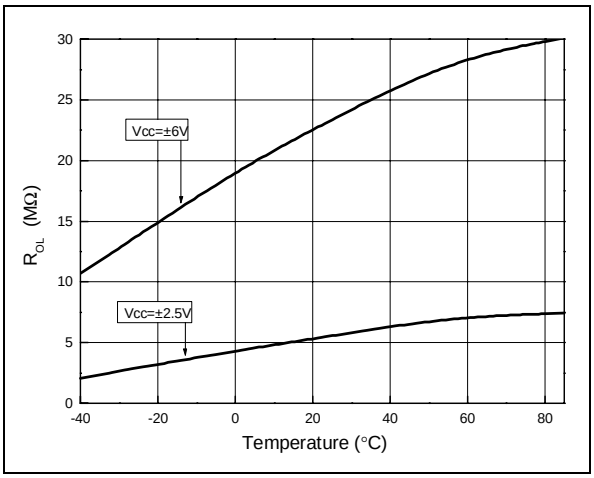


Figure 24: I_{CC} vs. Power Supply
Open loop, no load

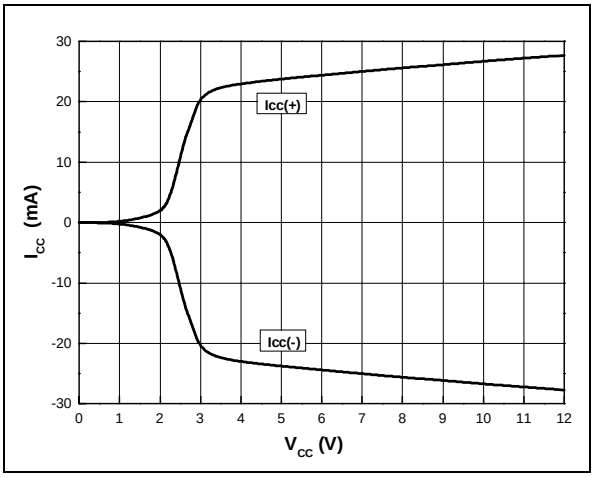


Figure 25: I_{ib} vs. Power Supply
Open loop, no load

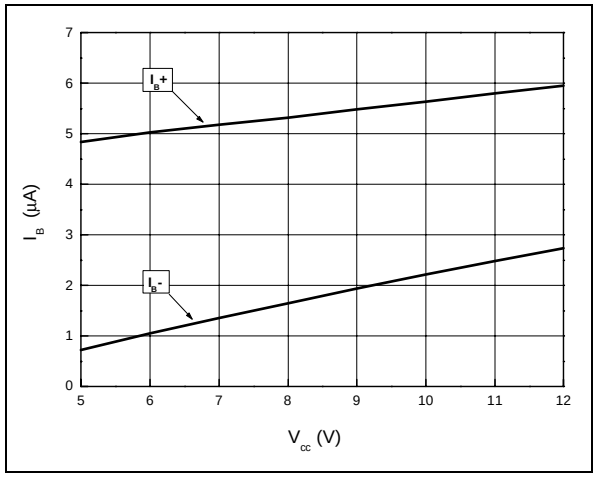


Figure 26: $I_{ib(-)}$ vs. Temperature
Open loop, no load

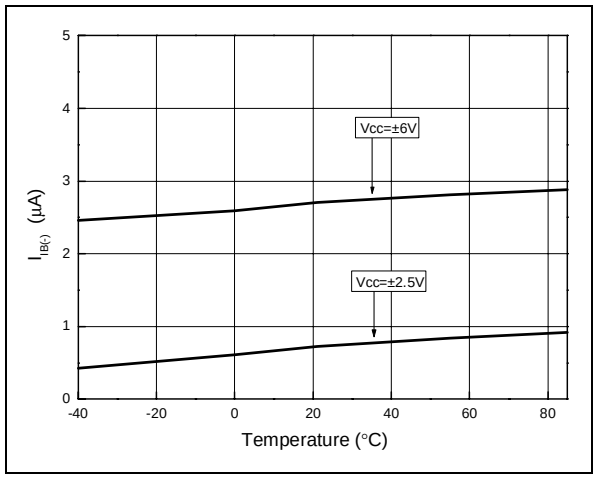


Figure 27: I_{cc} vs. Temperature
Open loop, no load

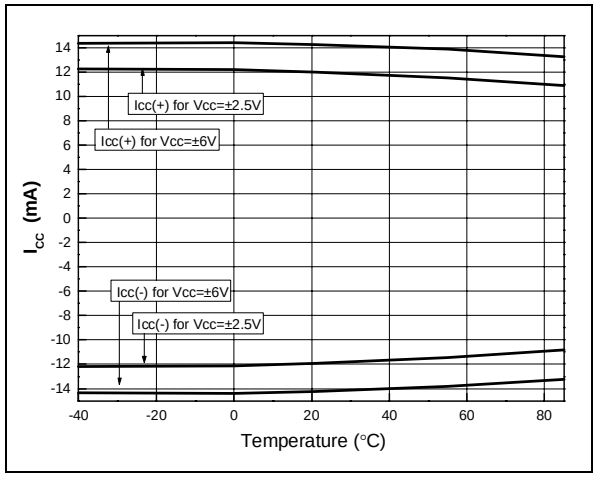


Figure 28: $I_{ib(+)}$ vs. Temperature
Open loop, no load

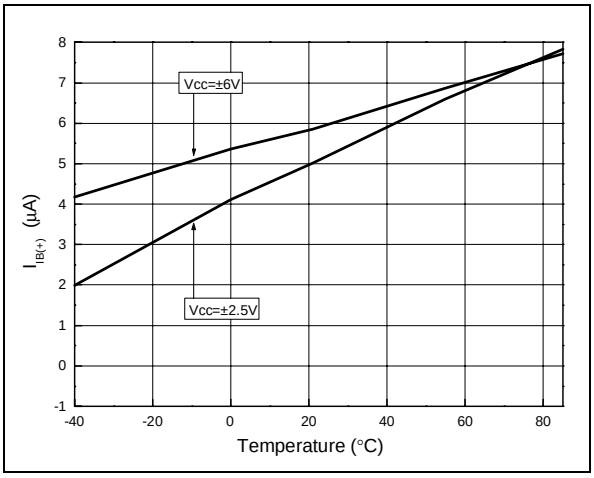


Figure 29: V_{oh} & V_{ol} vs. Power Supply
Open loop, R_L=25Ω

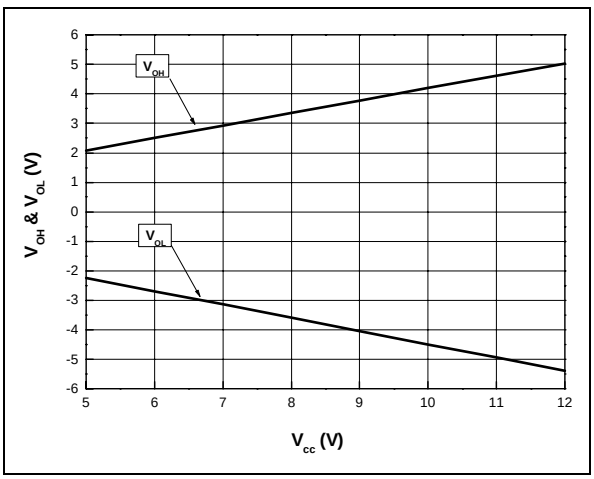


Figure 30: V_{oh} vs. Temperature
Open loop

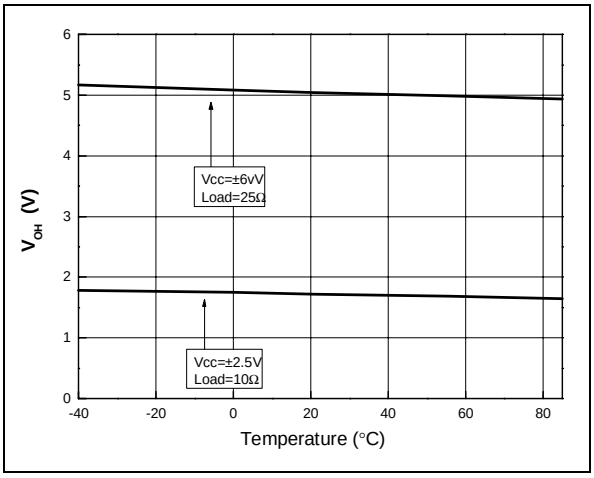


Figure 31: V_{ol} vs. Temperature
Open loop

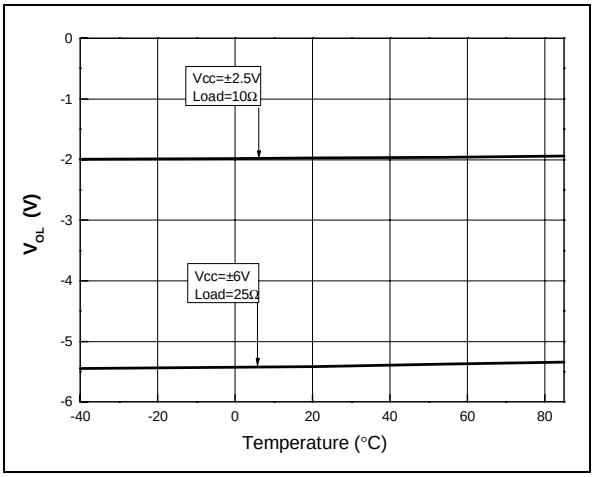


Figure 32: Differential V_{io} vs. Temperature
Open loop, no load

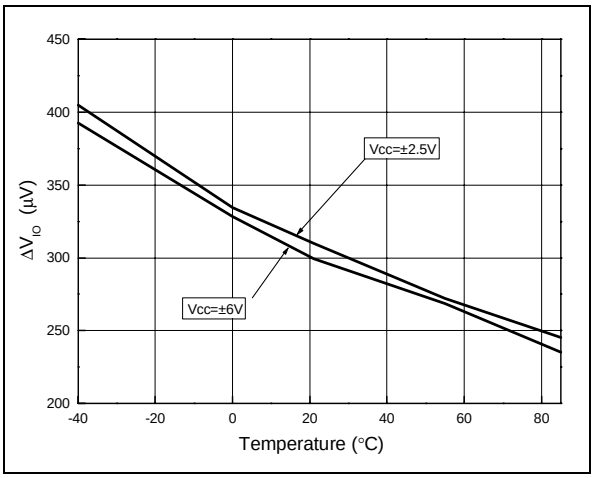


Figure 33: V_{io} vs. Temperature
Open loop, no load

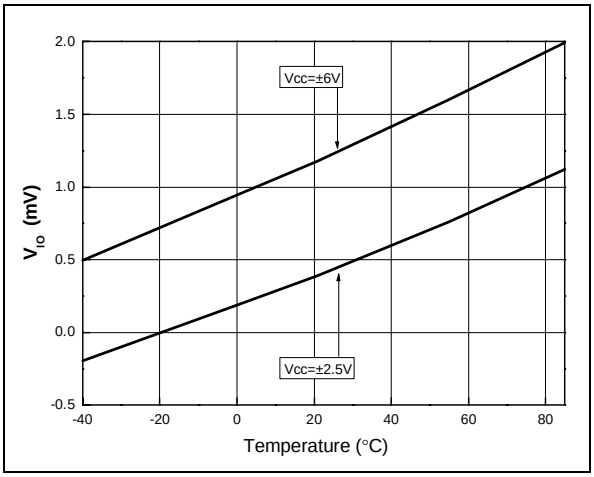


Figure 34: CMR vs. Temperature
Open loop, no load

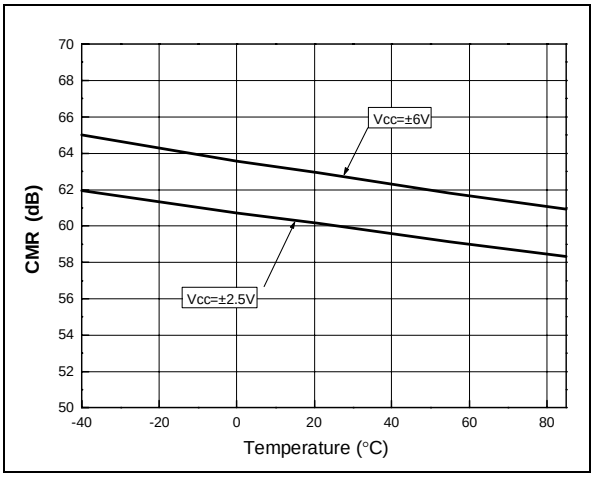


Figure 35: SVR vs. Temperature
Open loop, no load

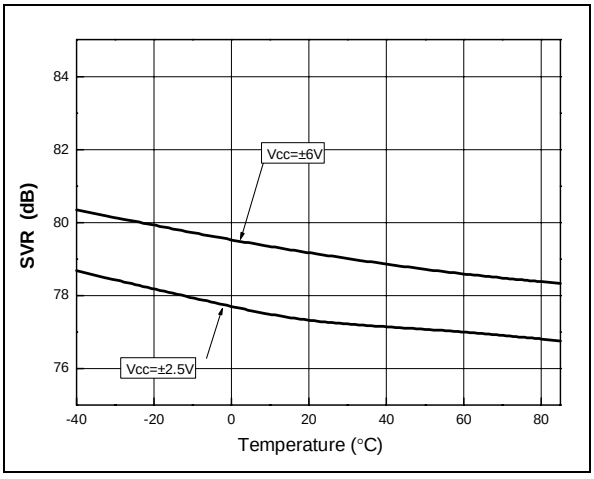


Figure 36: Iout vs. Temperature
Open loop, V_{CC}=±6V, R_L=10Ω

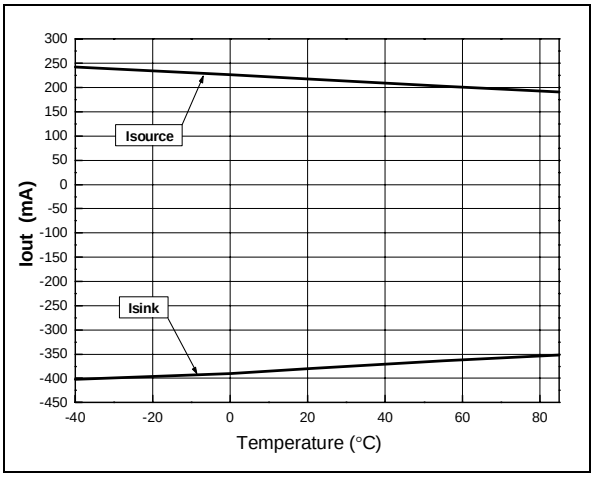


Figure 37: Iout vs. Temperature
Open loop, V_{CC}=±2.5V, R_L=25Ω

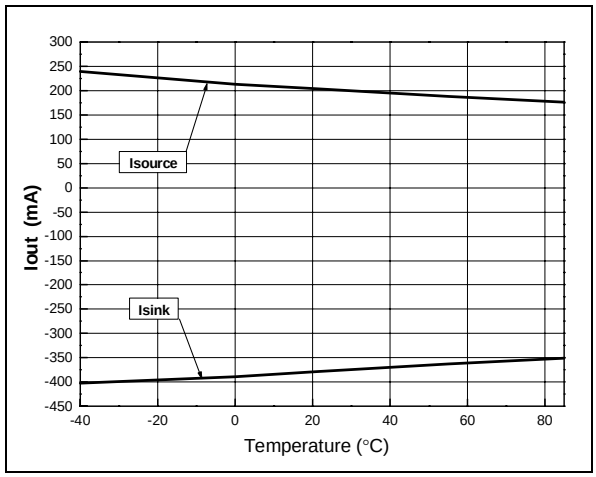


Figure 38: Maximum Output Amplitude vs. Load
A_V=+4, R_{fb}=620Ω, V_{CC}=±6V

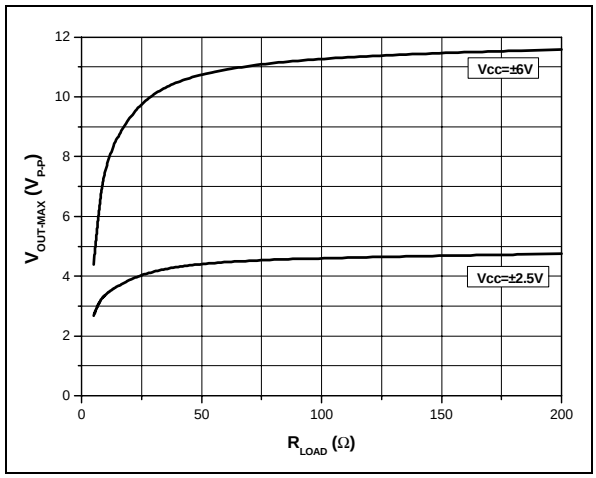


Figure 39: Isink vs. Output Amplitude.
V_{CC}=±2.5V, Open Loop, no Load

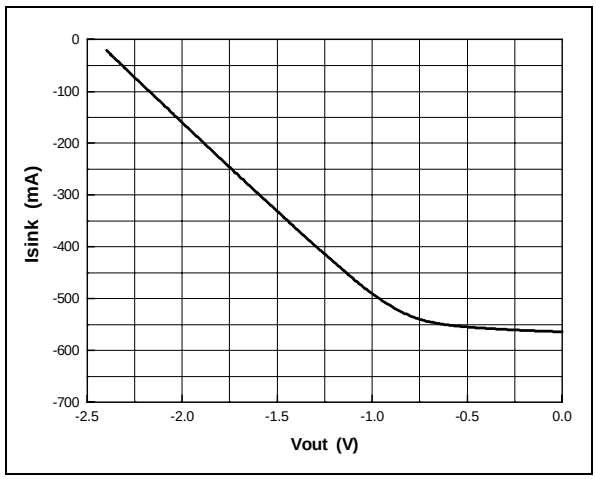


Figure 40: Isource vs. Output Amplitude.
V_{CC}=±2.5V, Open Loop, no Load

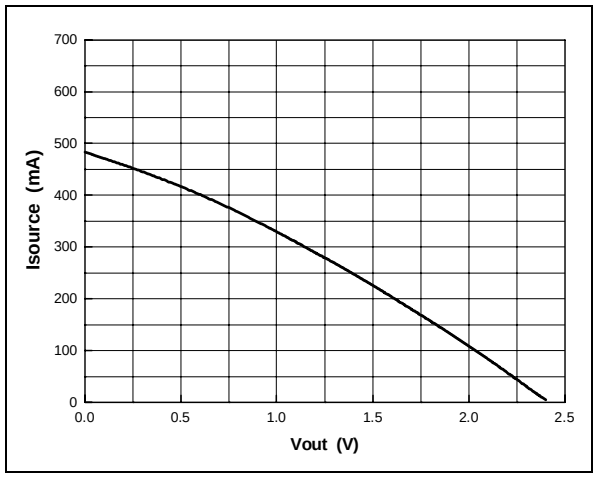


Figure 41: Isink vs. Output Amplitude
V_{CC}=±6V, Open Loop, no Load

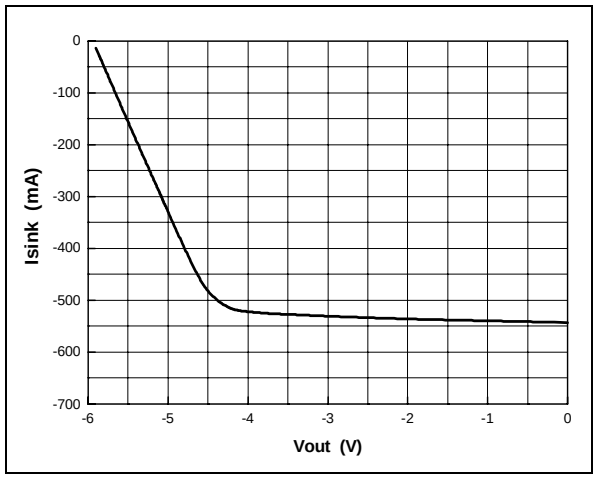


Figure 42: Isource vs. Output Amplitude
V_{CC}=±6V, Open Loop, no Load

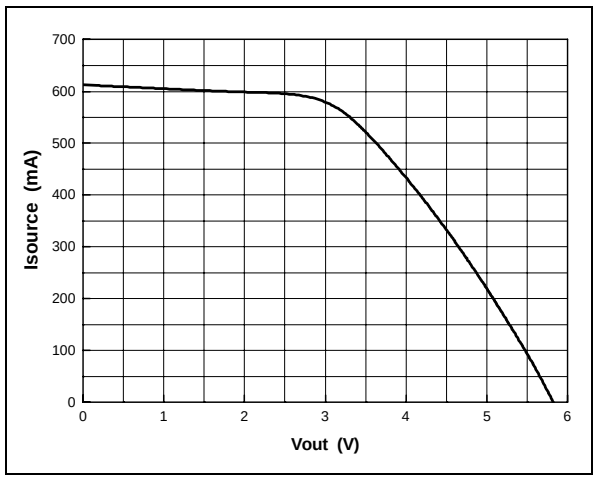
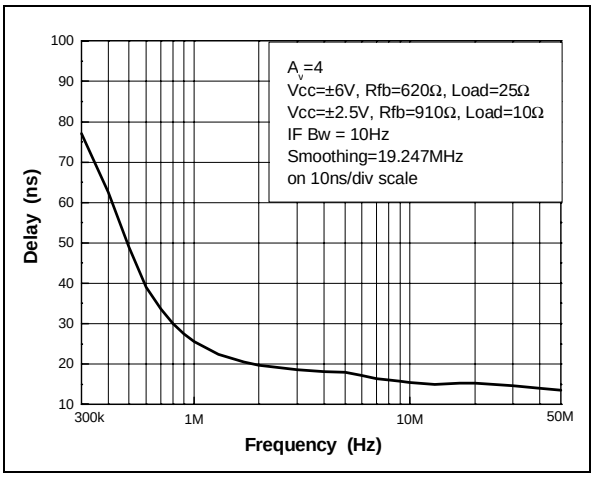


Figure 43: Group Delay
V_{CC}=±6V, V_{CC}=±2.5V



SAFE OPERATING AREA

Figure 44: Safe Operating Area

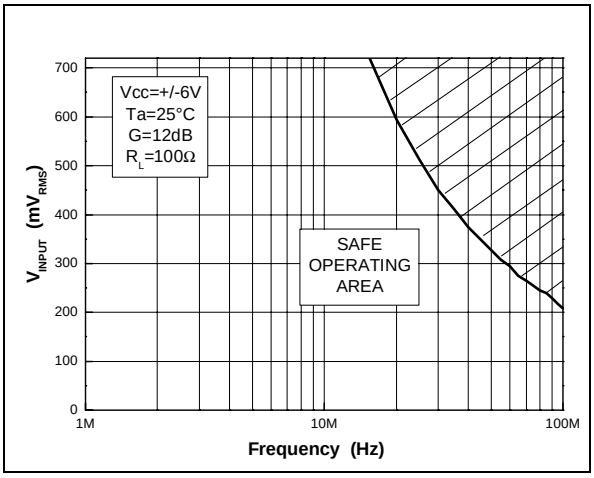


Figure 44 shows the safe operating condition. This curve shows the input level vs. the input frequency. It's necessary to consider this characteristic to guarantee the design. In the dash-lined zone, the consumption increases. Moreover, this increased consumption could do damage to the chip if the temperature increases.

INTERMODULATION DISTORTION PRODUCT

Non-ideal output of the amplifier can be described by the following series:

$$Vout = C_0 + C_1 V_{in} + C_2 V^2_{in} + \ldots C_n V^n_{in}$$

due to non-linearity in the input-output amplitude transfer, where the input is V_{in}=Asinωt, C_0 is the DC component, $C_1(V_{in})$ is the fundamental and C_n is the amplitude of the harmonics of the output signal V_{out}.

A one-frequency (one-tone) input signal contributes to harmonic distortion. A two-tone input signal contributes to harmonic distortion and intermodulation product.

The study of the intermodulation/distortionfor a two-tone input signal is the first step in characterizing the driving capability of multi-tone input signals.

In this case :

$$+ C_2(A\sin\omega_1 t + B\sin\omega_2 t)^2$$

$$\ldots + C_n(A\sin\omega_1 t + B\sin\omega_2 t)^n$$

$$V_{in} = A\sin\omega_1 t + B\sin\omega_2 t$$

$$V_{out} = C_0 + C_1(A\sin\omega_1 t + B\sin\omega_2 t)$$

and :

$$+ C_1(A\sin\omega_1 t + B\sin\omega_2 t)$$

$$-\frac{C_2}{2}\left(A^2\cos 2\omega_1 t + B^2\cos 2\omega_2 t\right)$$

$$+ 2C_2AB(\cos(\omega_1-\omega_2)t - \cos(\omega_1-\omega_2)t)$$

$$+\left(3\frac{C_3}{4}\right)$$

$$+\left(C_3A^3\sin 3\omega_1 t + B^3\sin 3\omega_2 t\right)$$

$$+\frac{3C_3A^2B}{2}\left(\sin(2\omega_1-\omega_2)t - \frac{1}{2}\sin(2\omega_1+\omega_2)t\right)$$

$$+\frac{3C_3A^2B}{2}\left(\sin(-\omega_1+2\omega_2)t - \frac{1}{2}\sin(\omega_1+2\omega_2)t\right)$$

$$\ldots + C_n(V_{in})^n$$

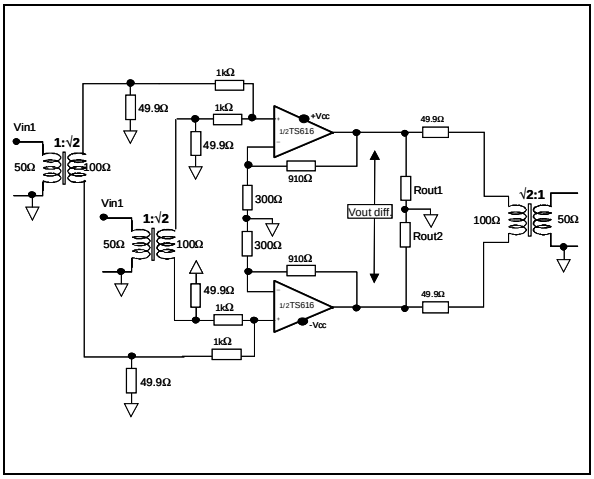
$$V_{out} = C_0 + C_2\left(\frac{A^2+B^2}{2}\right)$$

$$\left(A^3\sin\omega_1 t + B^3\sin\omega_2 t + 2A^2B\sin\omega_1 t + 2AB^2\sin\omega_2 t\right)$$

In this expression, we recognize the second order intermodulation IM2 by the frequencies (ω_1-ω_2) and (ω_1+ω_2) and the third order intermodulation IM3 by the frequencies ($2\omega_1$-ω_2), ($2\omega_1$+ω_2), ($-\omega_1$+$2\omega_2$) and (ω_1+$2\omega_2$).

The measurement of the intermodulation product of the driver is achieved by using the driver as a mixer by a summing amplifier configuration. In this way, the non-linearity problem of an external mixing device is avoided.

Figure 45: Non-inverting Summing Amplifier for Intermodulation measurements



The following graphs show the IM2 and the IM3 of the amplifier in different configurations. The two-tone input signal was generated by the multi-source generator Marconi 2026. Each tone has the same amplitude. The measurement was performed using a HP3585A spectrum analyzer.

Figure 46: Intermodulation vs. Output Amplitude
370kHz & 400kHz, A_V=+1.5, R_{fb}=1kΩ, R_L=14Ω diff.,V_{CC}=±2.5V

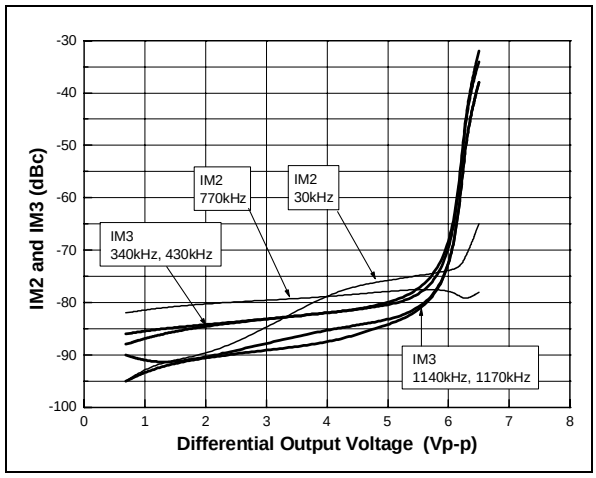


Figure 47: Intermodulation vs. Output Amplitude
370kHz & 400kHz, A_V=+1.5, R_{fb}=1kΩ, R_L=28Ω diff.,V_{CC}=±2.5V

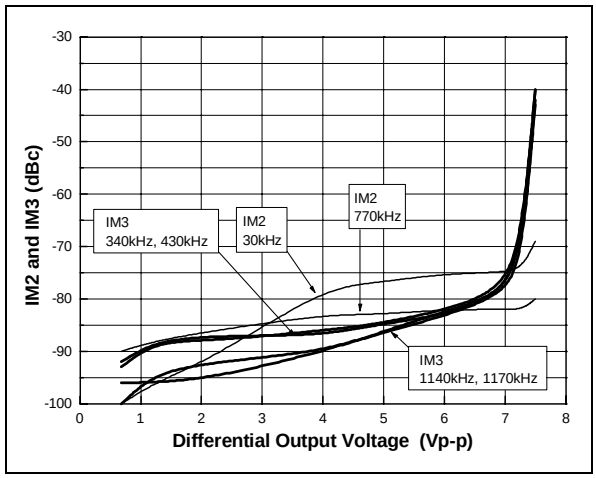


Figure 48: Intermodulation vs. Gain
370kHz & 400kHz, R_L=20Ω diff., Vout=6Vpp, V_{CC}=±2.5V

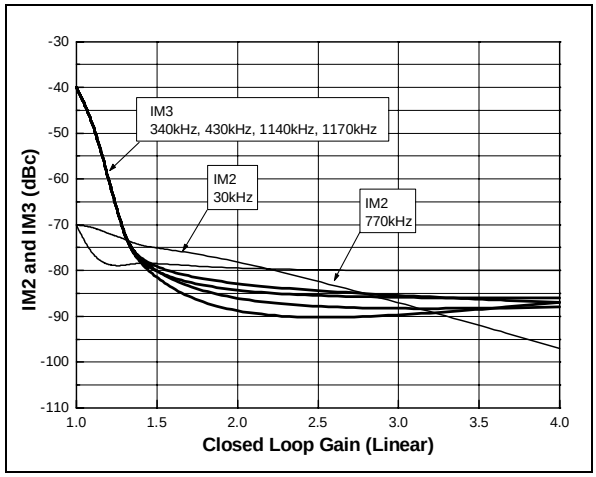


Figure 49: Intermodulation vs. Load
370kHz & 400kHz, A_V=+1.5, R_{fb}=1kΩ, Vout=6.5Vpp,V_{CC}=±2.5V

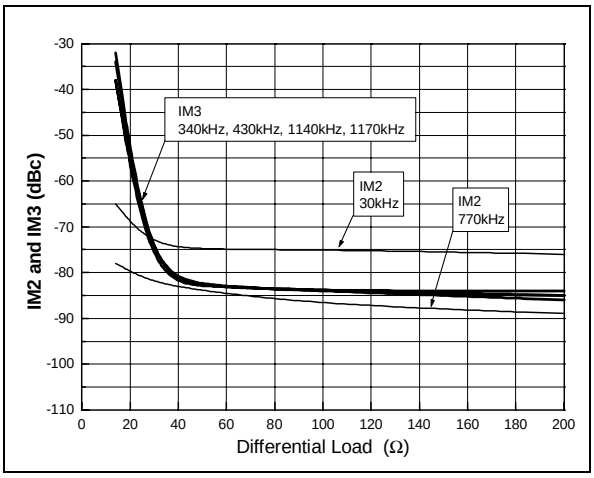


Figure 50: Intermodulation vs. Output Amplitude
100kHz & 110kHz, A_V=+4, R_{fb}=620Ω, R_L=200Ω diff.,V_{CC}=±6V

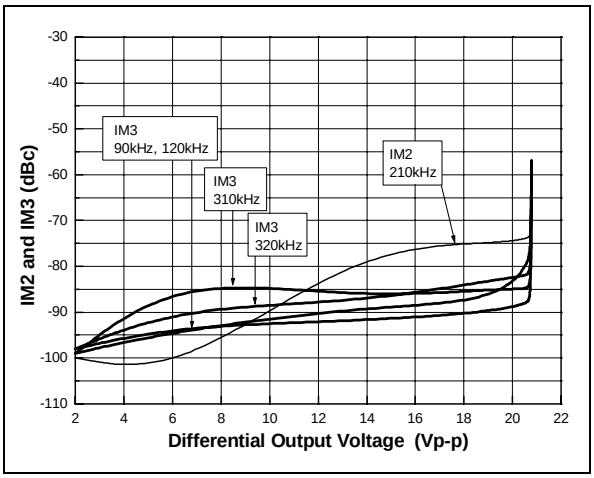


Figure 51: Intermodulation vs. Output Amplitude
100kHz & 110kHz, A_V=+4, R_{fb}=620Ω, R_L=50Ω diff., V_{CC}=±6V

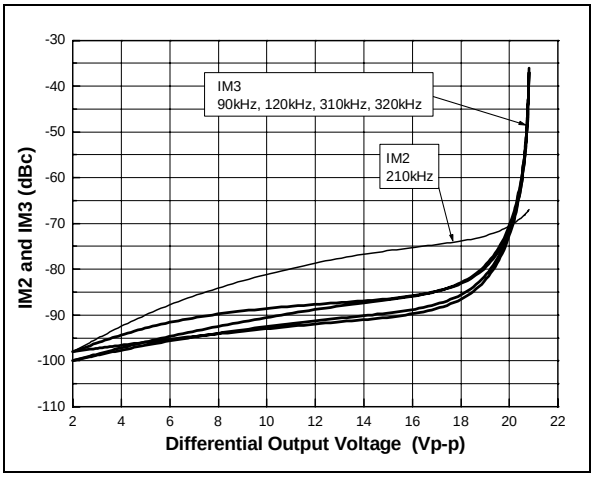


Figure 52: Intermodulation vs. Frequency Range
A_V=+4, R_{fb}=620Ω, R_L=50Ω diff., Vout=16Vpp, V_{CC}=±6V

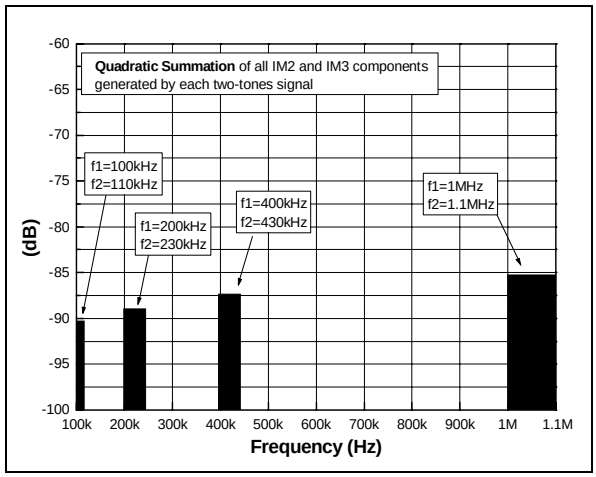


Figure 53: Intermodulation vs. Output Amplitude
370kHz & 400kHz, A_V=+4, R_{fb}=620Ω, R_L=200Ω diff.,V_{CC}=±6V

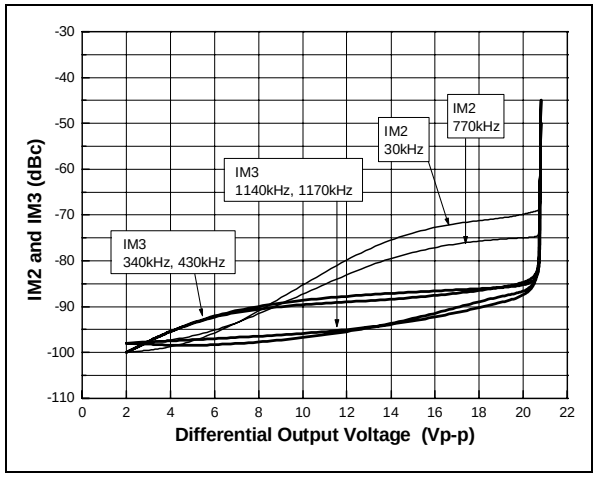
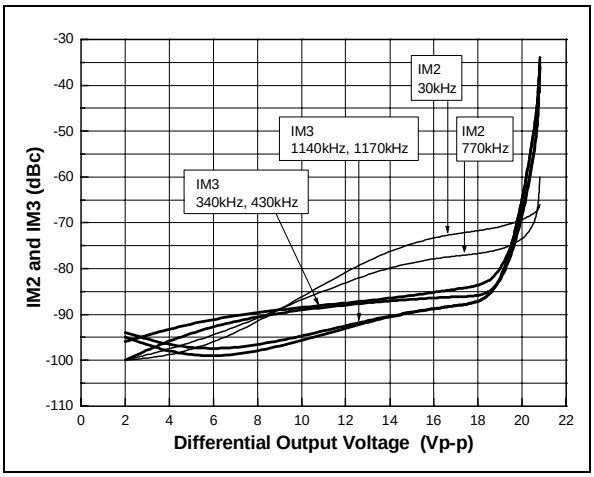


Figure 54: Intermodulation vs. Output Amplitude
370kHz & 400kHz, A_V=+4, R_{fb}=620Ω, R_L=50Ω diff., V_{CC}=±6V



PRINTED CIRCUIT BOARD LAYOUT CONSIDERATIONS

In the ADSL frequency range, printed circuit board parasites can affect the closed-loop performance.

The implementation of a proper ground plane on both sides of the PCB is mandatory to provide low inductance and low resistance common return. The most important factors affecting gain flatness and bandwidth are stray capacitances at the output and inverting input. To minimize these capacitances, the space between signal lines and ground plane should be increased. Feedback components connections must be as short as possible in order to decrease the associated inductance which affects high frequency gain errors. It is very important to choose the smallest possible external components, for example, surface mounted devices (SMD) in order to minimize the size of all DC and AC connections.

THERMAL INFORMATION

The TS616 is housed in an Exposed-Pad plastic package. As depicted in figure 55, this package uses a lead frame upon which the die is mounted. This lead frame is exposed as a thermal pad on the underside of the package. The thermal contact is direct with the dice. This thermal path provides excellent colling.

The thermal pad is electrically isolated from all pins in the package. It should be soldered to a copper area of the PCB underneath the package. Through these thermal paths within this copper area, heat can be conducted away from the package. In this case, the copper area should be connected to ($-V_{CC}$).

Figure 55: Exposed-Pad Package

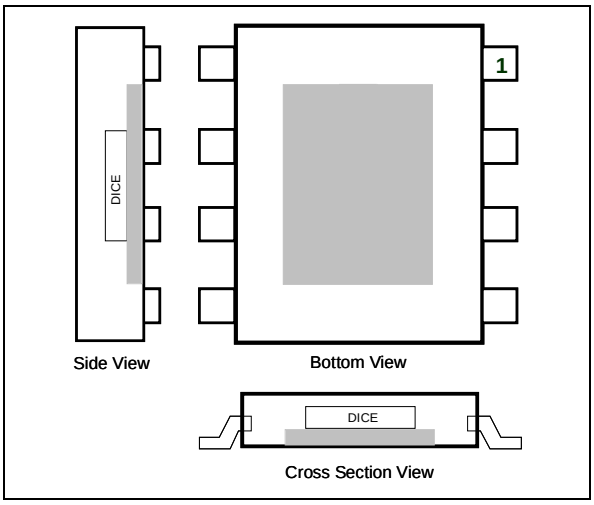


Figure 56: Evaluation Board

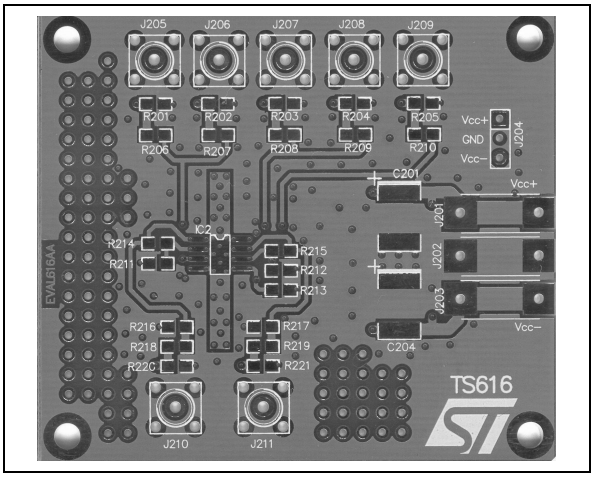


Figure 57: Schematic Diagram

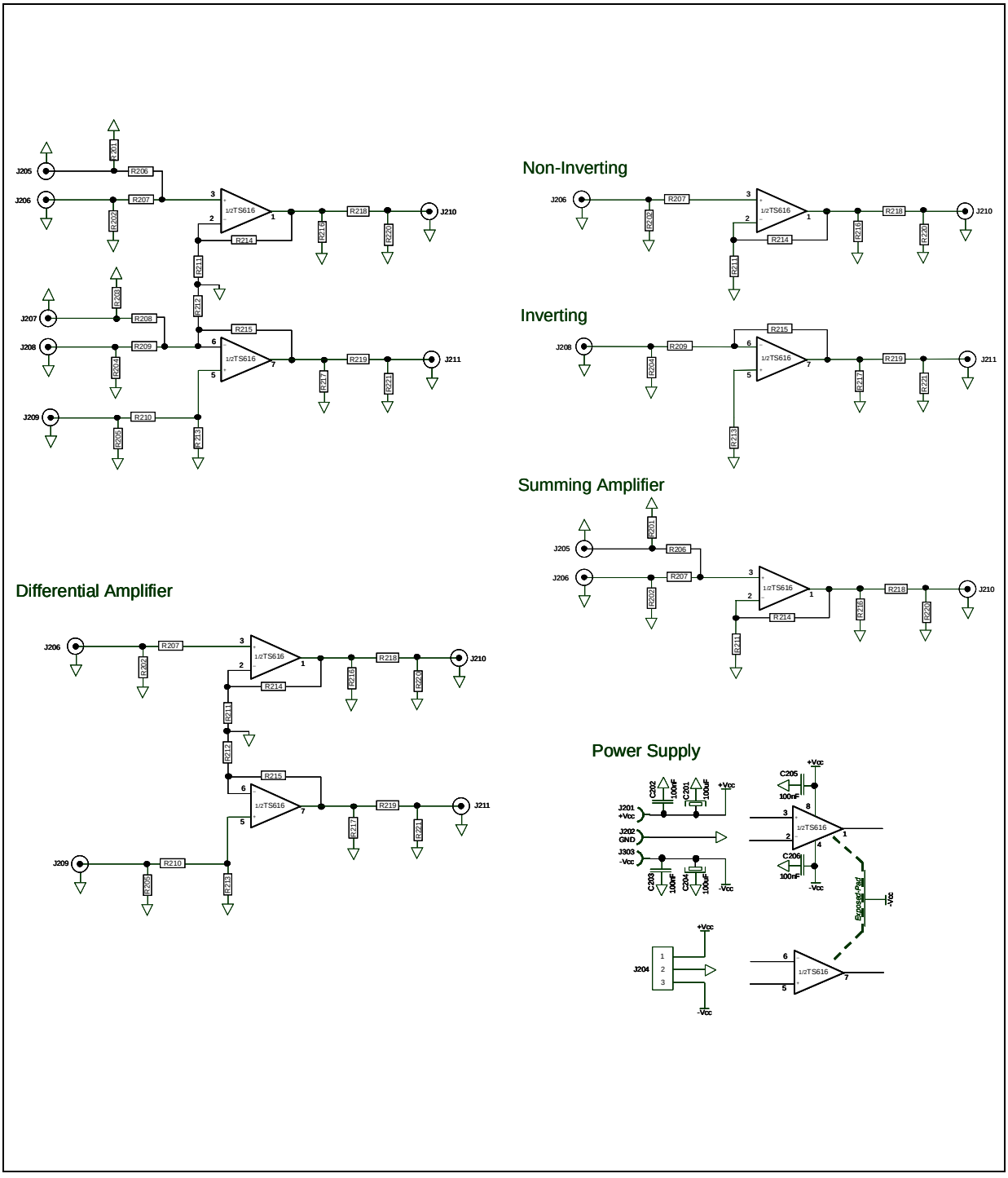


Figure 58: Component Locations - Top Side

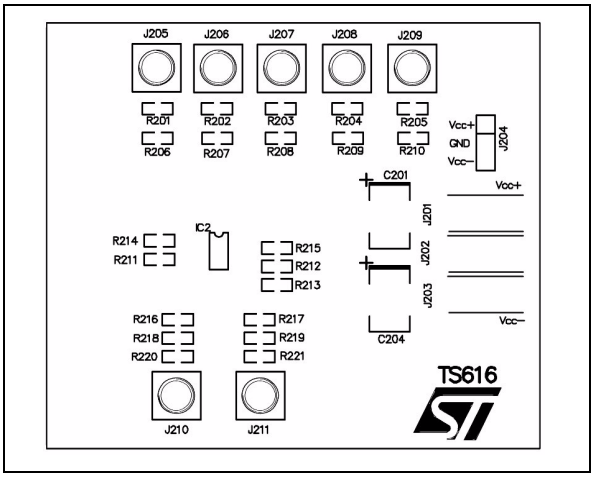


Figure 59: Component Locations - Bottom Side

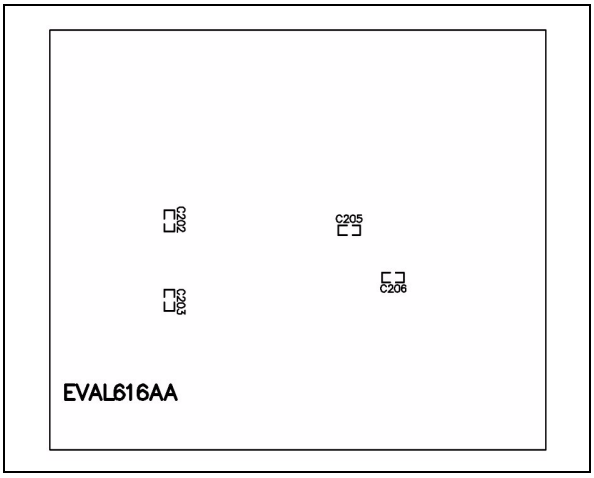


Figure 60: Top Side Board Layout

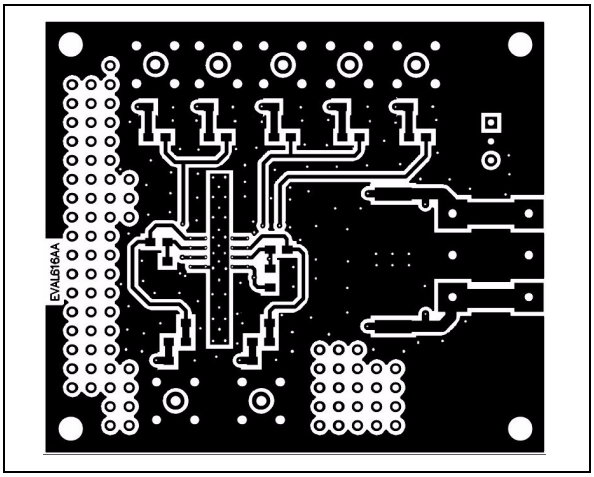
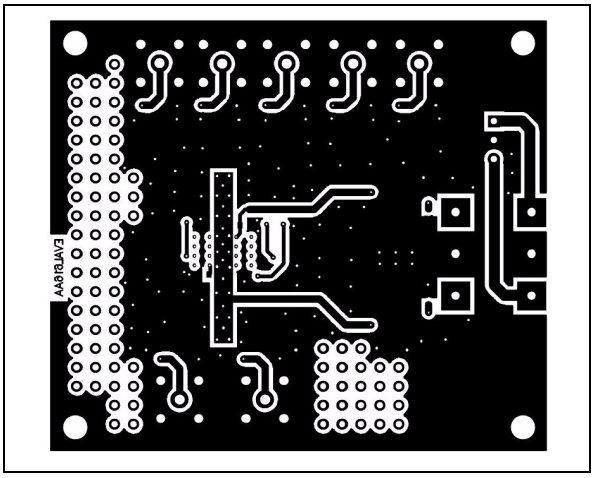
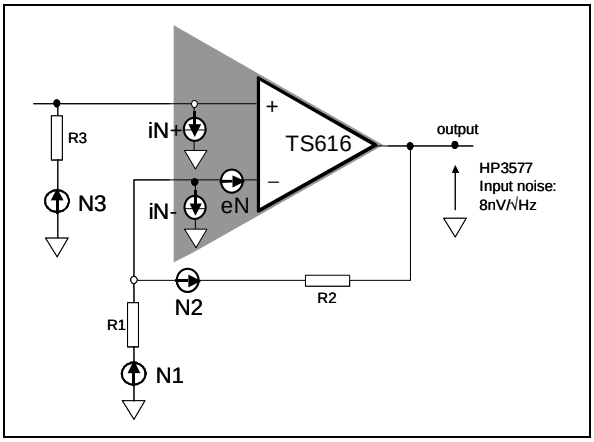


Figure 61: Bottom Side Board Layout



NOISE MEASUREMENT

Figure 62: Noise Model



eN : input voltage noise of the amplifier
iNn : negative input current noise of the amplifier
iNp : positive input current noise of the amplifier
The closed loop gain is :

$$A_V = g = 1 + \frac{R_{fb}}{R_g}$$

The six noise sources are :

$$V1 = eN \times \left(1 + \frac{R2}{R1}\right)$$

$$V2 = iNn \times R2$$

$$V3 = iNp \times R3 \times \left(1 + \frac{R2}{R1}\right)$$

$$V4 = -\frac{R2}{R1} \times \sqrt{4kTR1}$$

$$V5 = \sqrt{4kTR2}$$

$$V6 = \left(1 + \frac{R2}{R1}\right)\sqrt{4kTR3}$$

We assume that the thermal noise of a resistance R is:

$$\sqrt{4kTR\Delta F}$$

wher ΔF is the specified bandwidth.
On 1Hz bandwidth the thermal noise is reduced to

$$\sqrt{4kTR}$$

k is the Boltzmann's constant, equal to 1,374.10-23J/°K. T is the temperature (°K).

The output noise eNo is calculated using the Superposition Theorem. However it is not the simple sum of all noise sources. The square root of the sum of the square of each noise source.

$$eNo = \sqrt{V1^2 + V2^2 + V3^2 + V4^2 + V5^2 + V6^2}, (eq1)$$

$$eNo^2 = eN^2 \times g^2 + iNn^2 \times R2^2 + iNp^2 \times R3^2 \times g^2$$

$$.. + \left(\frac{R2}{R1}\right)^2 \times 4kTR1 + 4kTR2 + \left(1 + \frac{R2}{R1}\right)^2 \times 4kTR3, (eq2)$$

The input noise of the instrumentation must be extracted from the measured noise value. The real output noise value of the driver is:

$$eNo = \sqrt{(Measured)^2 - (instrumentation)^2}, (eq3)$$

The input noise is called the Equivalent Input Noise as it is not directly measured but it is evaluated from the measurement of the output divided by the closed loop gain (eNo/g).
After simplification of the fourth and the fifth term of (eq2) we obtain:

$$eNo^2 = eN^2 \times g^2 + iNn^2 \times R2^2 + iNp^2 \times R3^2 \times g^2$$

$$... + g \times 4kTR2 + \left(1 + \frac{R2}{R1}\right)^2 \times 4kTR3, (eq4)$$

Measurement of eN:
We assume a short-circuit on the non-inverting input (R3=0). (eq4) comes:

$$eNo = \sqrt{eN^2 \times g^2 + iNn^2 \times R2^2 + g \times 4kTR2}, (eq5)$$

In order to easily extract the value of eN, the resistance R2 will be chosen as low as possible. In the other hand, the gain must be large enough.
R1=10Ω, R2=910Ω, R3=0, Gain=92
Equivalent Input Noise: 2.57nV/√Hz
Input Voltage Noise: eN=2.5nV/√Hz

Measurement of iNn:
R3=0 and the output noise equation is still the (eq5). This time the gain must be decreased to decrease the thermal noise contribution.
R1=100Ω, R2=910Ω, R3=0, Gain=10.1
Equivalent Input Noise: 3.40nV/√Hz
Negative Input Current Noise: iNn =21pA/√Hz

Measurement of iNp:
To extract iNp from (eq3), a resistance R3 is connected to the non-inverting input. The value of R3 must be chosen in order to keep its thermal noise

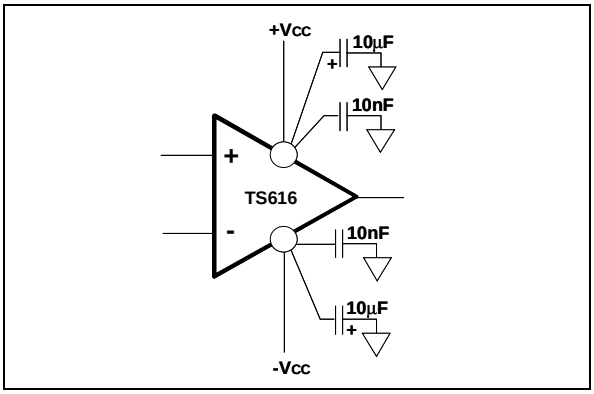
contribution as low as possible against the iNp contribution.
R1=100Ω, R2=910Ω, R3=100Ω, Gain=10.1
Equivalent Input Noise: 3.93nV/√Hz
Positive Input Current Noise: iNp=15pA/√Hz

Conditions: frequency=100kHz, V_{CC}=±2.5V
Instrumentation: Spectrum Analyzer HP3585A
(input noise of the HP3585A: 8nV/√Hz)

POWER SUPPLY BYPASSING

Correct power supply bypassing is very important for optimizing the performance in high frequency ranges. Bypass capacitors should be placed as close as possible to the IC pins to improve high frequency bypassing. A capacitor greater than 1µF is necessary to minimize the distortion. For a better quality bypassing a capacitor of 10nF can be added using the same implementation conditions. Bypass capacitors must be incorporated for both the negative and the positive supply.

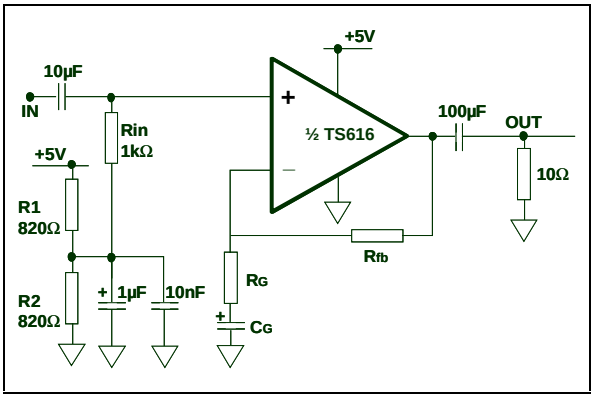
Figure 63: Circuit for Power Supply Bypassing



SINGLE POWER SUPPLY

The following figure shows the case of a 5V single power supply configuration

Figure 64: Circuit for +5V single supply



The TS616 operates with power supplies from 12V down to 5V. This can be achieved by either a dual power supplies of ±6V or ±2.5V or a single power supply of 12V or 5V referenced to the ground. In the case of asymmetrical supply, a new biasing is necessary to assume a positive output dynamic range between 0V and +V_{CC} supply rails. Considering the values of V_{OH} and V_{OL}, the amplifier will provide an output dynamic from +0.5V to 10.6V on 25Ω load for a 12V supply and from 0.45V to 3.8V on 10Ω load for a 5V supply.

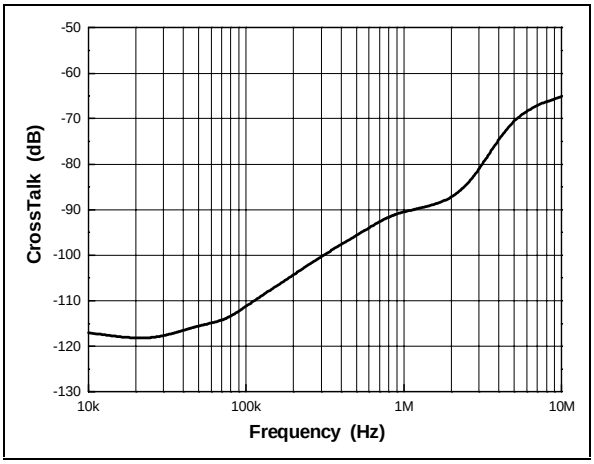
The amplifier must be biased with a mid-supply (nominally +V_{CC}/2), in order to maintain the DC component of the signal at this value. Several options are possible to provide this bias supply, such as a virtual ground using an operational amplifier or a two-resistance divider (which is the cheapest solution). A high resistance value is required to limit the current consumption. On the other hand, the current must be high enough to bias the non-inverting input of the amplifier. If we consider

this bias current (30µA max.) as the 1% of the current through the resistance divider to keep a stable mid-supply, two resistances of 2.2kΩ can be used in the case of a 12V power supply and two resistances of 820Ω can be used in the case of a 5V power supply.
The input provides a high pass filter with a break frequency below 10Hz which is necessary to remove the original 0 volt DC component of the input signal, and to fix it at $+V_{CC}/2$.

CHANNEL SEPARATION - CROSSTALK

The following figure shows the crosstalk from an amplifier to a second amplifier. This phenomenon, accentuated at high frequencies, is unavoidable and intrinsic to the circuit.
Nevertheless, the PCB layout also has an effect on the crosstalk level. Capacitive coupling between signal wires, distance between critical signal nodes and power supply bypassing are the most significant factors.

Figure 65: Crosstalk vs. Frequency
A_V=+4, R_{fb}=620Ω, V_{CC}=±6V, Vout=2Vp



CHOICE OF THE FEEDBACK CIRCUIT

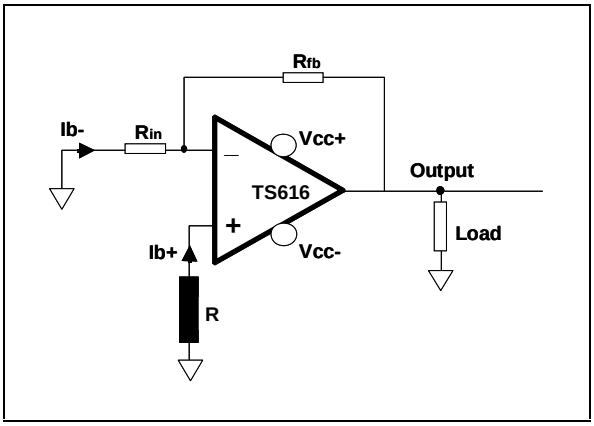
Table 1: Closed-Loop Gain - Feedback Components

V_{CC} (V)	Gain	R_{fb} (Ω)
±6	+1	750
	+2	680
	+4	620
	+8	510
	-1	680
	-2	680
	-4	620
	-8	510
±2.5	+1	1.1k
	+2	1k
	+4	910
	+8	680
	-1	1k
	-2	1k
	-4	910
	-8	680

INVERTING AMPLIFIER BIASING

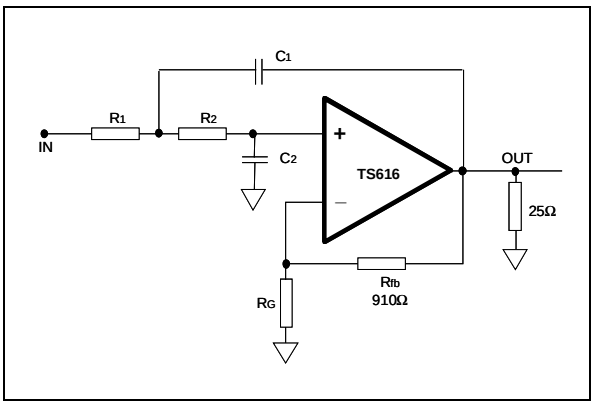
In this case a resistance R is necessary to achieve good input biasing, see Figure 66.
This resistance is calculated by assuming the negative and positive input bias current. The aim is to compensate for the offset bias current which could affect the input offset voltage and the output DC component. Assuming Ib-, Ib+, R_{in}, R_{fb} and a zero volt output, the resistance R comes: $R = R_{in} // R_{fb}$.

Figure 66: Compensation of the Input Bias Current



ACTIVE FILTERING

Figure 67: Low-Pass Active Filtering. Sallen-Key



The resistors R_{fb} and R_G give the gain of the filter as a classical non-inverting amplification configuration :

$$A_V = g = 1 + \frac{R_{fb}}{R_g}$$

Assume the following expression is the response of the system:

$$T_{j\omega} = \frac{Vout_{j\omega}}{Vin_{j\omega}} = \frac{g}{1 + 2\zeta\frac{j\omega}{\omega_c} + \frac{(j\omega)^2}{\omega_c^2}}$$

the cut-off frequency is not gain-dependent and it becomes:

$$\omega_c = \frac{1}{\sqrt{R1R2C1C2}}$$

The damping factor becomes:

$$\zeta = \frac{1}{2}\omega_c(C_1R_1 + C_1R_2 + C_2R_1 - C_1R_1g)$$

The higher the gain, the more sensitive the damping factor is. When the gain is higher than 1 it is preferable to use very stable resistors and capacitances values.

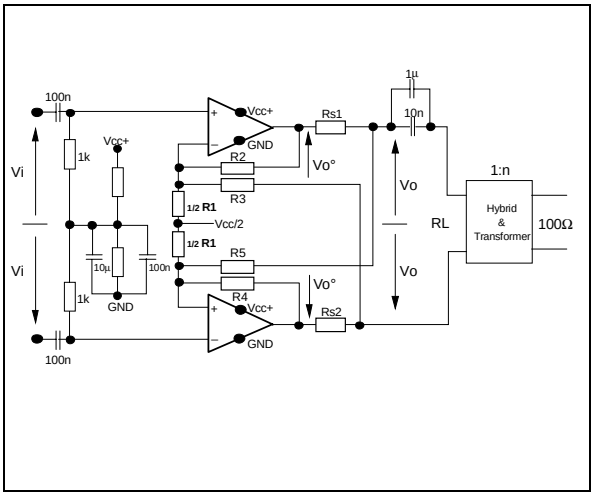
In the case of R1=R2:

$$\zeta = \frac{2C_2 - C_1\frac{R_{fb}}{R_g}}{2\sqrt{C_1C_2}}$$

INCREASING THE LINE LEVEL BY USING AN ACTIVE IMPEDANCE MATCHING

With passive matching, the output signal amplitude of the driver must be twice the amplitude on the load. To go beyond this limitation, active matching impedance can be used. With this technique, it is possible to keep good impedance matching with an amplitude on the load higher than half of the output driver amplitude. This concept is shown in Figure 68 for a differential line.

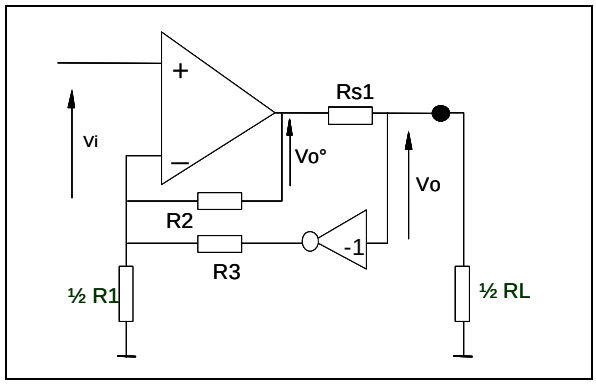
Figure 68: TS616 as a differential line driver with an active impedance matching



Component Calculation

Let us consider the equivalent circuit for a single-ended configuration as shown in Figure 69.

Figure 69: Single ended equivalent circuit



Let us consider the unloaded system. Assuming the currents through R1, R2 and R3 are respectively:

$$\frac{2Vi}{R1}, \frac{(Vi - Vo°)}{R2} \text{and} \frac{(Vi + Vo)}{R3}$$

as Vo° equals Vo without load, the gain in this case becomes :

$$G = \frac{Vo(noload)}{Vi} = \frac{1 + \frac{2R2}{R1} + \frac{R2}{R3}}{1 - \frac{R2}{R3}}$$

The gain, for the loaded system will be (eq1):

$$GL = \frac{Vo(withload)}{Vi} = \frac{1}{2}\frac{1 + \frac{2R2}{R1} + \frac{R2}{R3}}{1 - \frac{R2}{R3}}, (eq1)$$

As shown in Figure70, this system is an ideal generator with a synthesized impedance as the internal impedance of the system. From this, the output voltage becomes:

$$Vo = (ViG) - (RoIout), (eq2)$$

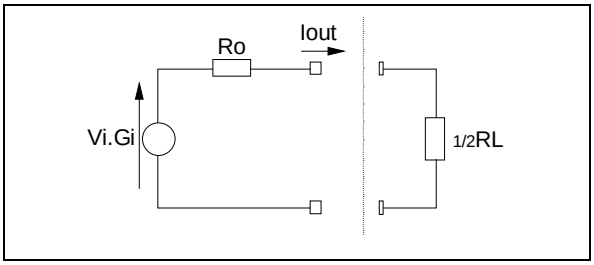
with Ro the synthesized impedance and Iout the output current. On the other hand Vo can be expressed as:

$$Vo = \frac{Vi\left(1 + \frac{2R2}{R1} + \frac{R2}{R3}\right)}{1 - \frac{R2}{R3}} - \frac{Rs1Iout}{1 - \frac{R2}{R3}}, (eq3)$$

By identification of both equations (eq2) and (eq3), the synthesized impedance is, with Rs1=Rs2=Rs:

$$Ro = \frac{Rs}{1 - \frac{R2}{R3}}, (eq4)$$

Figure 70: Equivalent schematic. Ro is the synthesized impedance.



Unlike the level Vo° required for a passive impedance, Vo° will be smaller than 2Vo in our case. Let us write Vo°=kVo with k the matching factor varying between 1 and 2. Assuming that the current through R3 is negligible, (eq4) becomes the following :

$$Ro = \frac{kVoRL}{RL + 2Rs1}$$

After choosing the k factor, Rs will equal to 1/2RL(k-1).

A good impedance matching assumes:

$$Ro = \frac{1}{2}RL, (eq5)$$

(eq4) and (eq5) give :

$$\frac{R2}{R3} = 1 - \frac{2Rs}{RL}, (eq6)$$

By fixing an arbitrary value of R2, (eq6) becomes :

$$R3 = \frac{R2}{1 - \frac{2Rs}{RL}}$$

Finally, the values of R2 and R3 allow us to extract R1 from (eq1), and it becomes:

$$R1 = \frac{2R2}{2\left(1-\frac{R2}{R3}\right)GL-1-\frac{R2}{R3}}, (eq7)$$

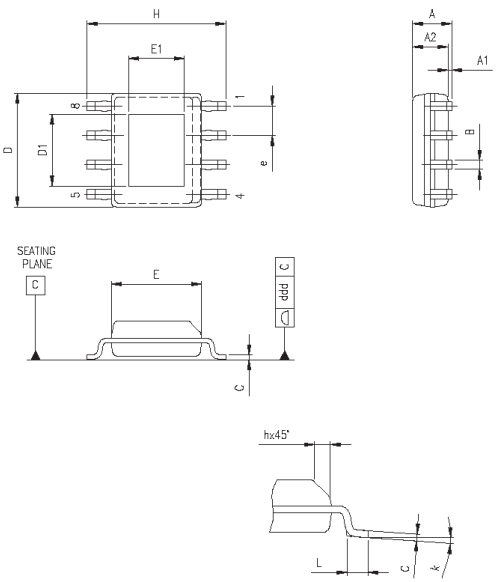
with GL the required gain.

Table 2 : Components Calculation for Impedance Matching Implementation

GL (gain for the loaded system)	GL is fixed for the application requirements GL=Vo/Vi=0.5(1+2R2/R1+R2/R3)/(1-R2/R3)
R1	2R2/[2(1-R2/R3)GL-1-R2/R3]
R2 (=R4)	Abritrary fixed
R3 (=R5)	R2/(1-Rs/0.5RL)
Rs	0.5RL(k-1)
Load viewed by each driver	kRL/2

PACKAGE MECHANICAL DATA

8 PINS - PLASTIC MICROPACKAGE (SO Exposed-Pad)



Dim.	Millimeters			Inches		
	Min.	Typ.	Max.	Min.	Typ.	Max.
A	1.350		1.750	0.053		0.069
A1	0.000		0.250	0.001		0.010
A2	1.100		1.650	0.043		0.065
B	0.330		0.510	0.013		0.020
C	0.190		0.250	0.007		0.010
D	4.800		5.000	0.189		0.197
D1		3.10			0.122	
E	3.800		4.000	0.150		0.157
E1		2.41			0.095	
e		1.270			0.050	
H	5.800		6.200	0.228		0.244
h	0.250		0.500	0.010		0.020
L	0.400		1.270	0.016		0.050
k	0d		8d	0d		8d
ddd			0.100			0.004

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