RHF350

Rad-hard 550 MHz low noise operational amplifier

Preliminary data

Features

- Bandwidth: 550 MHz in unity gain
- Quiescent current: 4.1 mA
- Slew rate: 940 V/µs
- Input noise: 1.5 nV/√Hz
- Distortion: SFDR = -66 dBc (10 MHz, $1V_{pp}$)
- 2.8 V_{pp} minimum output swing on 100 Ω load for a +5 V supply
- 5 V power supply
- 300 krad MIL-STD-883 1019.7 ELDRS free compliant
- SEL immune at 125° C, LET up to 110 MEV.cm^2/mg
- SET characterized, LET up to 110 MEV.cm^2/mg

Applications

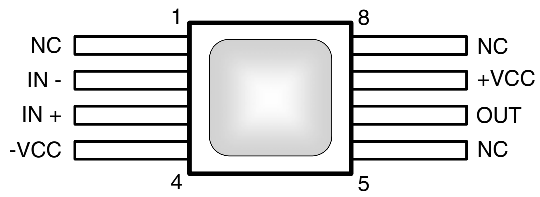
- Communication satellites
- Space data acquisition systems
- Aerospace instrumentation
- Nuclear and high energy physics
- Harsh radiation environments
- ADC drivers

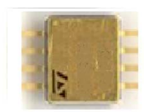
Description

The RHF350 is a current feedback operational amplifier that uses a very high speed complementary technology to provide a bandwidth of up to 410 MHz while drawing only 4.1 mA of quiescent current. With a slew rate of 940 V/µs and an output stage optimized for driving a standard 100 Ω load, this circuit is highly suitable for applications where speed and power-saving are the main requirements.

Pin connections
(top view)

Ceramic Flat-8





The RHF350 is a single operator available in the Flat-8 hermetic ceramic package, saving board room as well as providing excellent thermal and dynamic performance.

This is preliminary information on a new product now in development or undergoing evaluation. Details are subject to change without notice.

Contents

1 Absolute maximum ratings and operating conditions

Table 1. Absolute maximum ratings

Symbol	Parameter	Value	Unit
V_{CC}	Supply voltage(1)	6	V
V_{id}	Differential input voltage(2)	+/-0.5	V
V_{in}	Input voltage range(3)	+/-2.5	V
T_{oper}	Operating free air temperature range	-40 to + 85	°C
T_{stg}	Storage temperature	-65 to +150	°C
T_j	Maximum junction temperature	150	°C
R_{thja}	Flat-8 thermal resistance junction to ambient	50	°C/W
R_{thjc}	Flat-8 thermal resistance junction to case	30	°C/W
P_{max}	Flat-8 maximum power dissipation(4) (T_{amb} = 25° C) for T_j = 150° C	830	mW
ESD	HBM: human body model (5) pins 1, 4, 5, 6, 7 and 8 pins 2 and 3	 2 0.5	kV
	MM: machine model (6) pins 1, 4, 5, 6, 7 and 8 pins 2 and 3	 200 60	V
	CDM: charged device model(7) pins 1, 4, 5, 6, 7 and 8 pins 2 and 3	 1.5 1.5	kV
	Latch-up immunity	200	mA

1. All voltages values are measured with respect to the ground pin.
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} +0.3 V.
4. Short-circuits can cause excessive heating. Destructive dissipation can result from short-circuits on all amplifiers.
5. Human body model: a 100 pF capacitor is charged to the specified voltage, then discharged through a 1.5 kΩ resistor between two pins of the device. This is done for all couples of connected pin combinations while the other pins are floating.
6. This is a minimum value.
   Machine model: a 200 pF capacitor is charged to the specified voltage, then discharged directly between two pins of the device with no external series resistor (internal resistor < 5 Ω). This is done for all couples of connected pin combinations while the other pins are floating.
7. Charged device model: all pins and package are charged together to the specified voltage and then discharged directly to the ground through only one pin.

Table 2. Operating conditions

Symbol	Parameter	Value	Unit
V_{CC}	Supply voltage	4.5 to 5.5	V
V_{icm}	Common mode input voltage	$-V_{CC}$ +1.5 V to $+V_{CC}$ -1.5 V	V

2 Electrical characteristics

Table 3. Electrical characteristics for V_{CC} = ±2.5 V, T_{amb} = 25° C (unless otherwise specified) [1]

Symbol	Parameter	Test conditions	Min.	Typ.	Max.	Unit
DC performance						
V_{io}	Input offset voltage Offset voltage between both inputs	T_{amb}		0.8	4	mV
		$T_{min} < T_{amb} < T_{max}$	0.35		2.15	
I_{ib+}	Non-inverting input bias current DC current necessary to bias the input +	T_{amb}		12	35	µA
		$T_{min} < T_{amb} < T_{max}$	0.25		30	
I_{ib-}	Inverting input bias current DC current necessary to bias the input -	T_{amb}		1	20	µA
		$T_{min} < T_{amb} < T_{max}$	0.15		10	
CMR	Common mode rejection ratio 20 log ($\Delta V_{ic}/\Delta V_{io}$)	ΔV_{ic} = ±1 V	54	60		dB
		$T_{min} < T_{amb} < T_{max}$	57		63	
SVR	Supply voltage rejection ratio 20 log ($\Delta V_{CC}/\Delta V_{io}$)	ΔV_{CC} = +3.5 V to +5 V	68	81		dB
		$T_{min} < T_{amb} < T_{max}$	67		82	
PSR	Power supply rejection ratio 20 log ($\Delta V_{CC}/\Delta V_{out}$)	A_V = +1, ΔV_{CC} = ±100 mV at 1 kHz		51		dB
I_{CC}	Positive supply current DC consumption with no input signal	No load		4.1	4.9	mA
		$T_{min} < T_{amb} < T_{max}$	3.79		4.32	
Dynamic performance and output characteristics						
R_{OL}	Transimpedance Output voltage/input current gain in open loop of a CFA. For a VFA, the analog of this feature is the open-loop gain (A_{VD})	ΔV_{out} = ±1 V, R_L = 100 Ω	170	270		kΩ
		$T_{min} < T_{amb} < T_{max}$	236		325	kΩ
Bw	-3 dB bandwidth Frequency where the gain is 3 dB below the DC gain A_V[2]	Small signal V_{out} = 20 mV_{pp} A_V = +1, R_L = 100 Ω A_V = +2, R_L = 100 Ω A_V = +10, R_L = 100 Ω A_V = -2, R_L = 100 Ω	 250	 550 390 125 370		MHz
		$T_{min} < T_{amb} < T_{max}$	326			
	Gain flatness at 0.1 dB Band of frequency where the gain variation does not exceed 0.1 dB	Small signal V_{out} = 100 mV_p A_V = +1, R_L = 100 Ω		65		
SR	Slew rate Maximum output speed of sweep in large signal	V_{out} = 2 V_{pp}, A_V = +2, R_L = 100 Ω		940		V/µs
V_{OH}	High level output voltage	R_L = 100 Ω	1.44	1.56		V
		$T_{min} < T_{amb} < T_{max}$	1.46		1.67	

Table 3. Electrical characteristics for V_{CC} = ±2.5 V, T_{amb} = 25° C (unless otherwise specified) [1] (continued)

Symbol	Parameter	Test conditions	Min.	Typ.	Max.	Unit
V_{OL}	Low level output voltage	R_L = 100 Ω		-1.53	-1.44	V
		$T_{min} < T_{amb} < T_{max}$	-1.6		-1.5	
I_{out}	I_{sink} Short-circuit output current coming into the op-amp.[3]	Output to GND	135	205		mA
		$T_{min} < T_{amb} < T_{max}$	181		230	
	I_{source} Output current coming out from the op-amp.[4]	Output to GND	-140	-210		
		$T_{min} < T_{amb} < T_{max}$	179		273	
Noise and distortion						
eN	Equivalent input noise voltage[5]	F = 100 kHz		1.5		nV/√Hz
iN	Equivalent input noise current (+)[5]	F = 100 kHz		20		pA/√Hz
	Equivalent input noise current (-)[5]	F = 100 kHz		13		pA/√Hz
SFDR	Spurious free dynamic range The highest harmonic of the output spectrum when injecting a filtered sine wave	A_V = +1, V_{out} = 1V_{pp} F = 10 MHz F = 20 MHz F = 50 MHz F = 100 MHz		 -66 -57 -46 -42		dBc

1. $T_{min} < T_{amb} < T_{max}$: worst case of the parameter on a standard sample across the temperature range. The evaluation is done on 50 units in the SO-8 plastic package.
2. Gain bandwidth product criterion is not applicable for current feedback amplifiers.
3. See *Figure 7* for more details.
4. See *Figure 8* for more details.
5. See *Chapter 5 on page 14*.

Table 4. Closed-loop gain and feedback components

V_{CC} (V)	Gain	R_{fb} (Ω)
±2.5	+10	300
	-10	300
	+2	300
	-2	300
	+1	820
	-1	300

Figure 1. Frequency response, positive gain

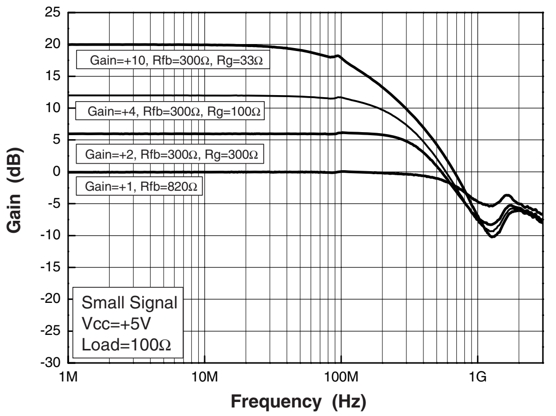


Figure 2. Flatness, gain = +1

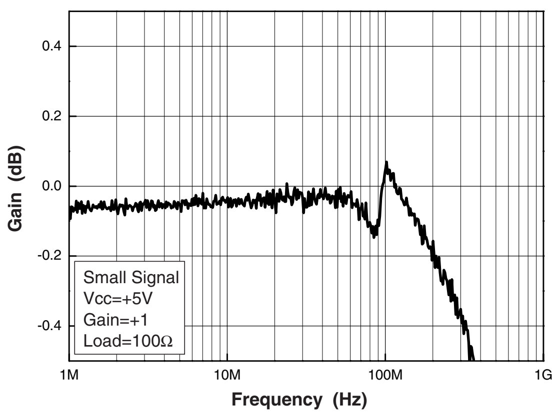


Figure 3. Flatness, gain = +2

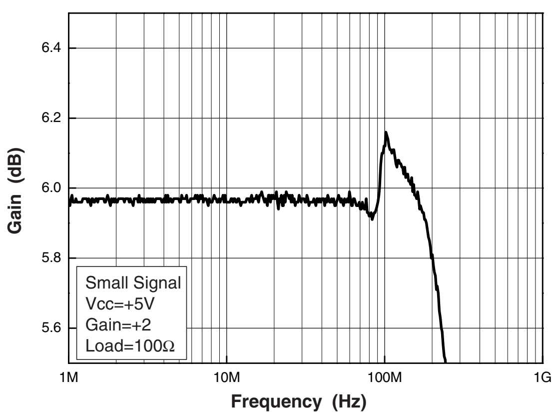


Figure 4. Flatness, gain = +4

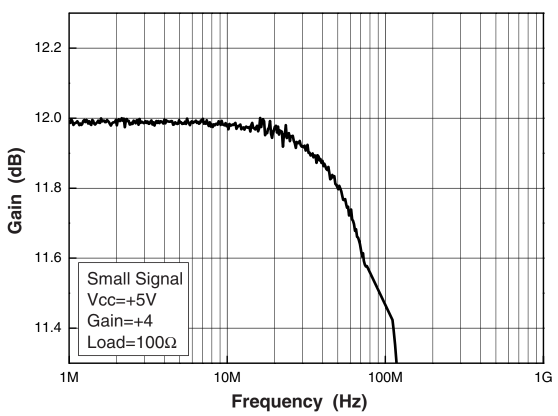


Figure 5. Flatness, gain = +10

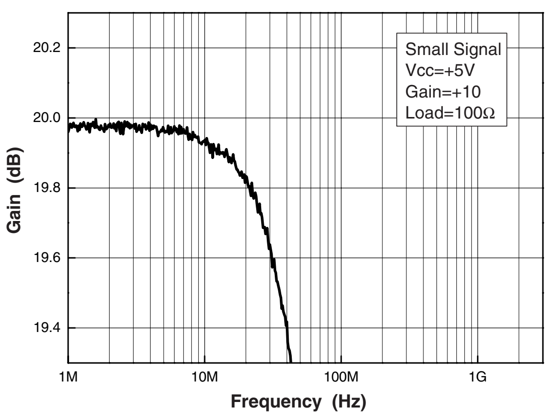


Figure 6. Slew rate

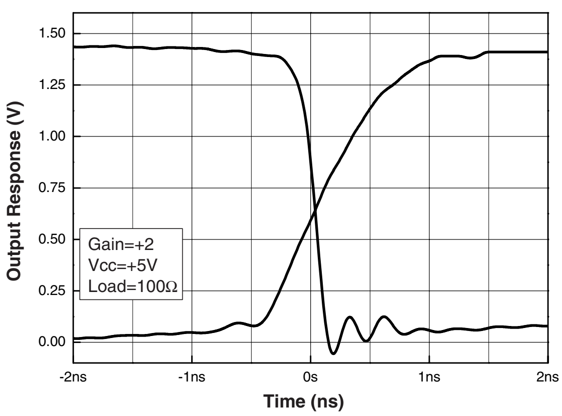


Figure 7. I_{sink}

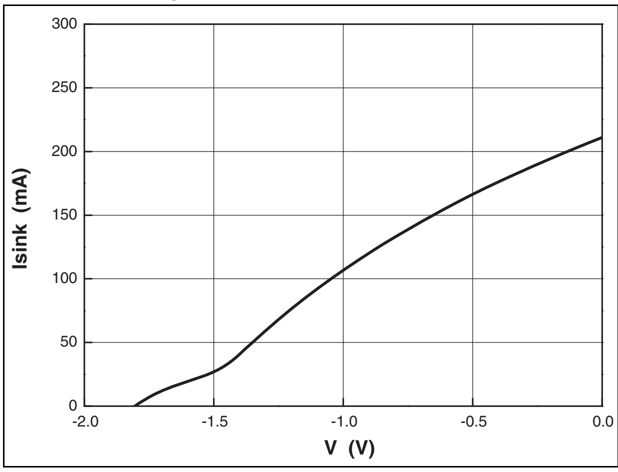


Figure 8. I_{source}

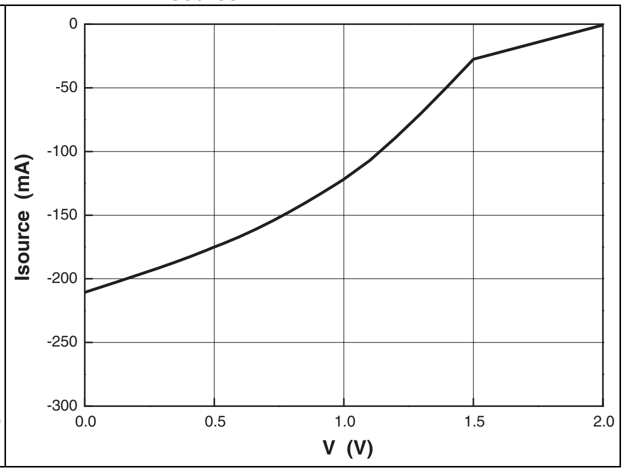


Figure 9. Input current noise vs. frequency

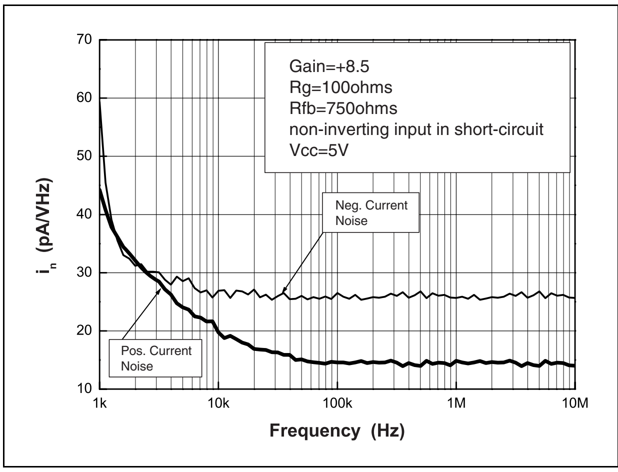


Figure 10. Input voltage noise vs. frequency

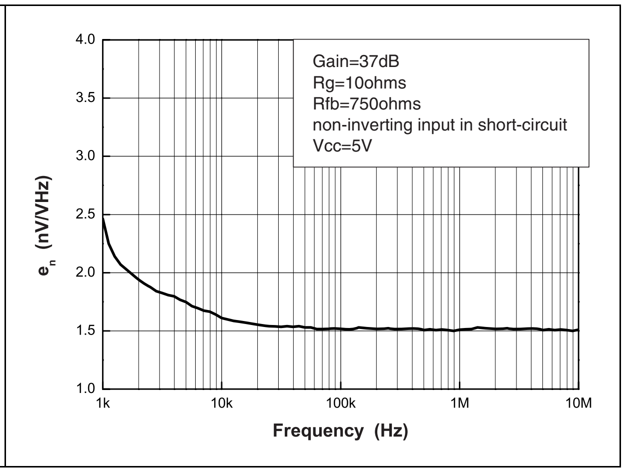


Figure 11. Quiescent current vs. V_{CC}

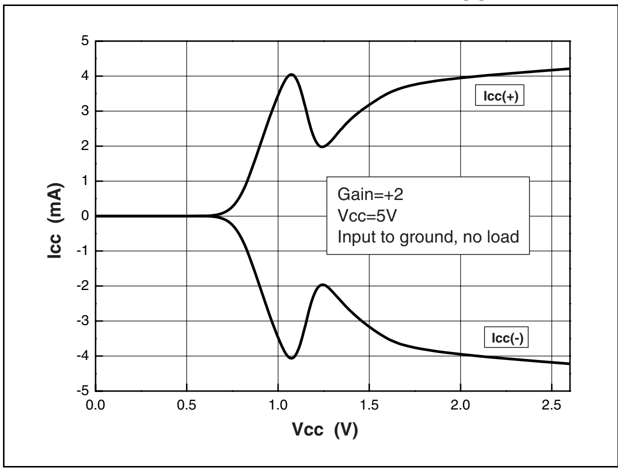


Figure 12. Noise figure

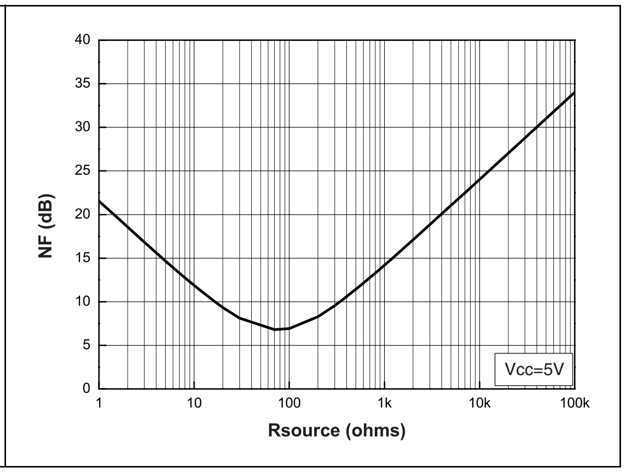


Figure 13. Distortion vs. output amplitude

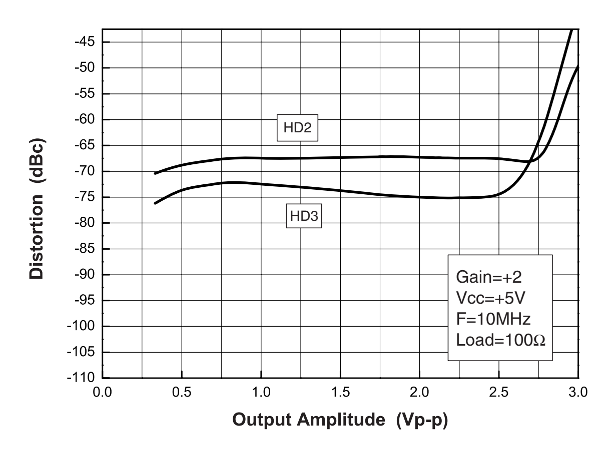


Figure 14. Output amplitude vs. load

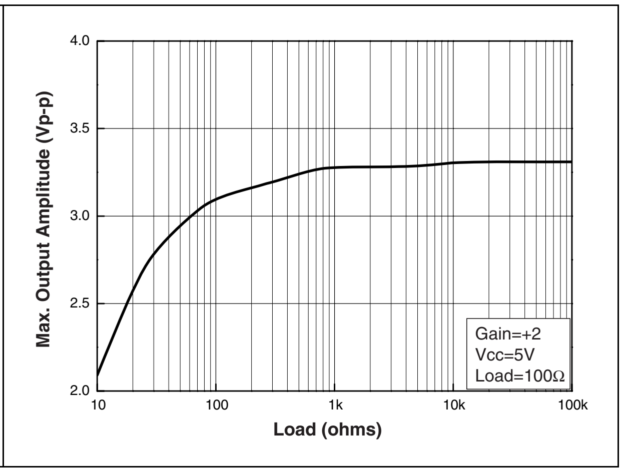


Figure 15. Reverse isolation vs. frequency

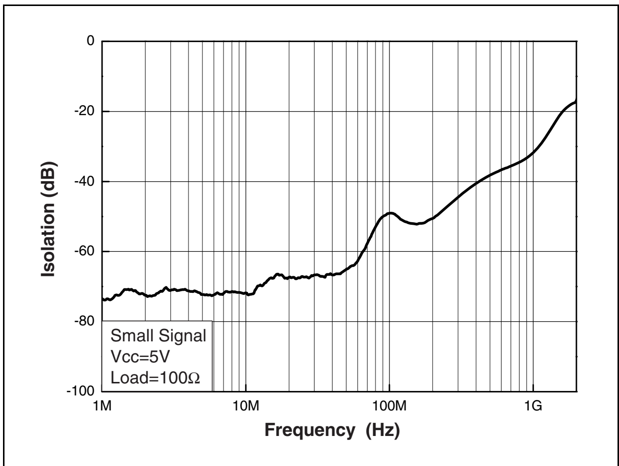


Figure 16. SVR vs. temperature

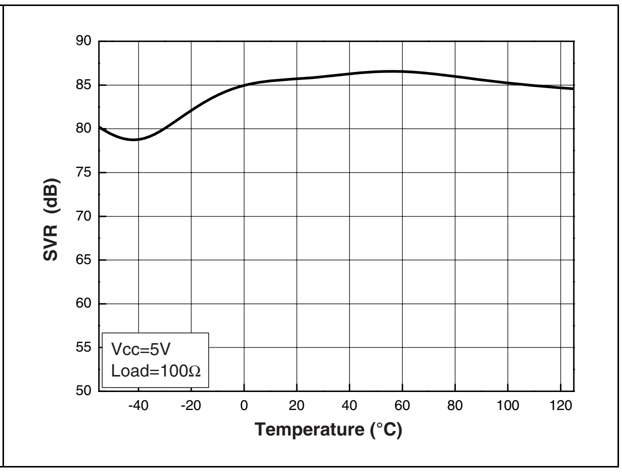


Figure 17. I_{out} vs. temperature

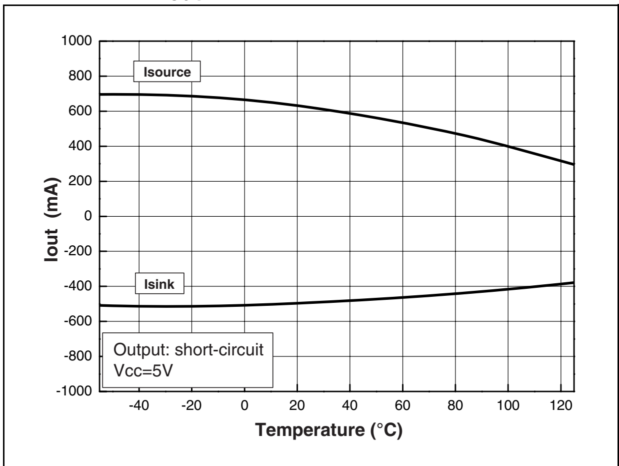


Figure 18. R_{OL} vs. temperature

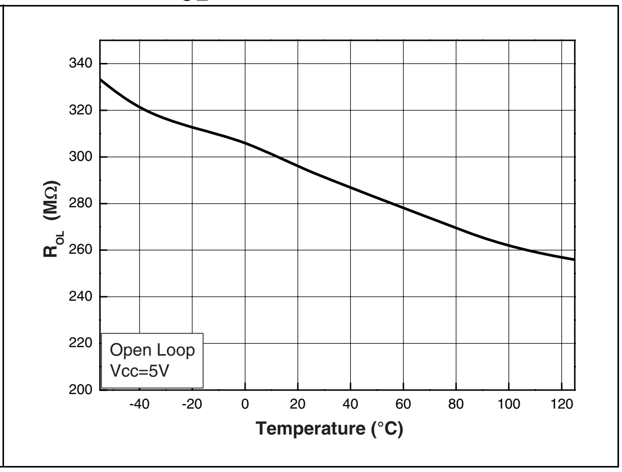


Figure 19. CMR vs. temperature

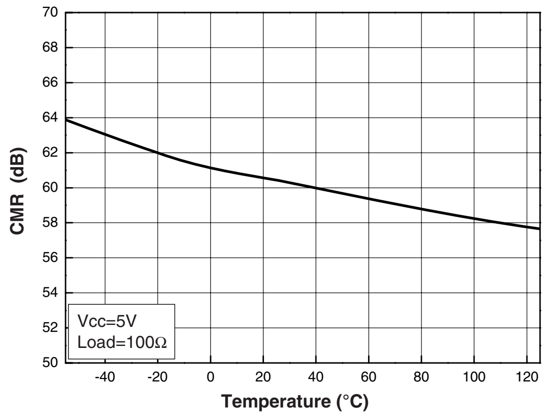


Figure 20. I_{bias} vs. temperature

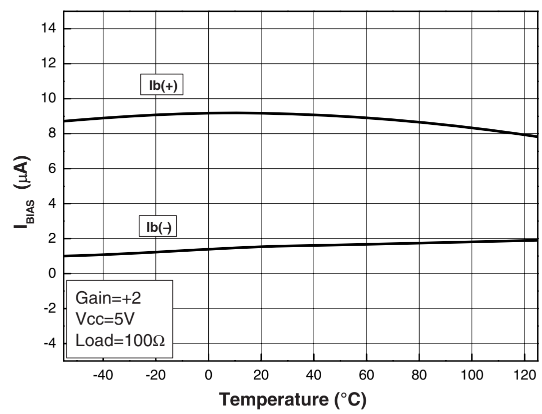


Figure 21. V_{io} vs. temperature

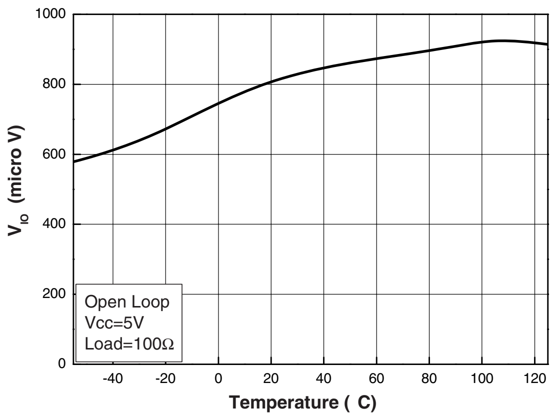


Figure 22. V_{OH} and V_{OL} vs. temperature

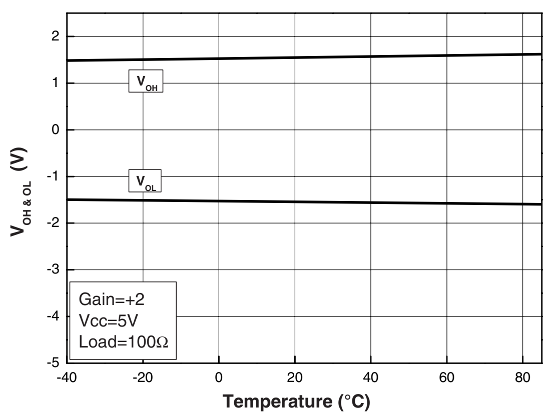
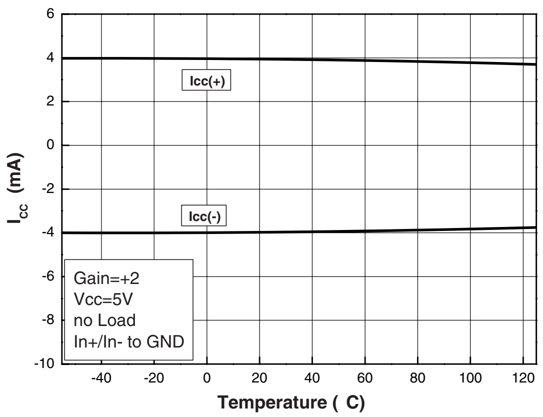


Figure 23. I_{CC} vs. temperature



3 Demonstration board schematics

Figure 24. Electrical schematics (inverting and non-inverting gain configuration)

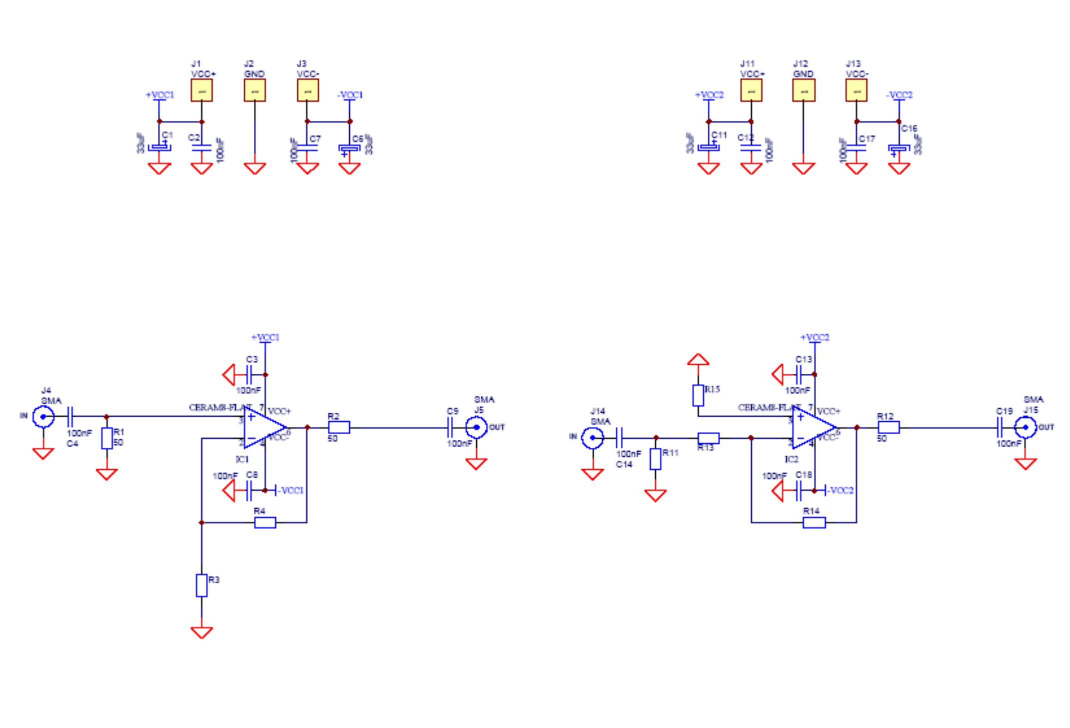


Figure 25. RHF3xx demonstration board

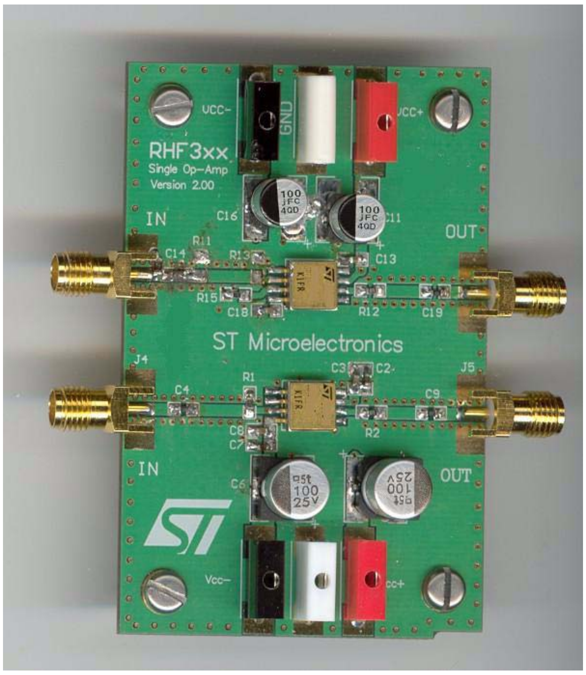


Figure 26. Top view layout

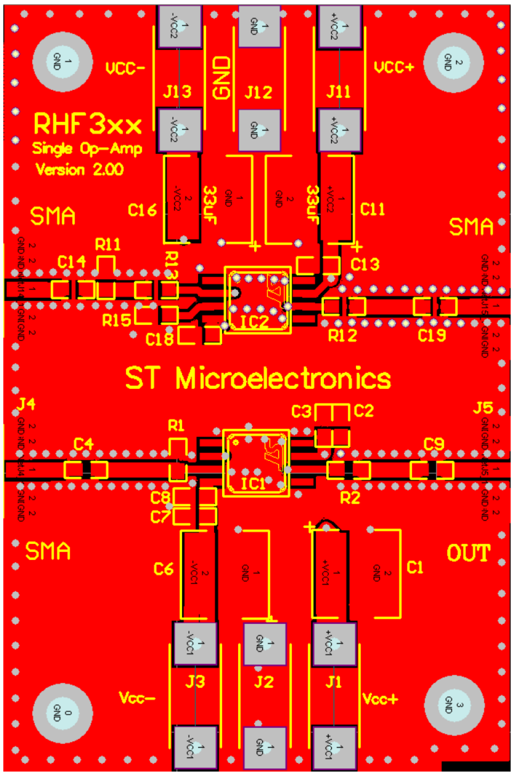
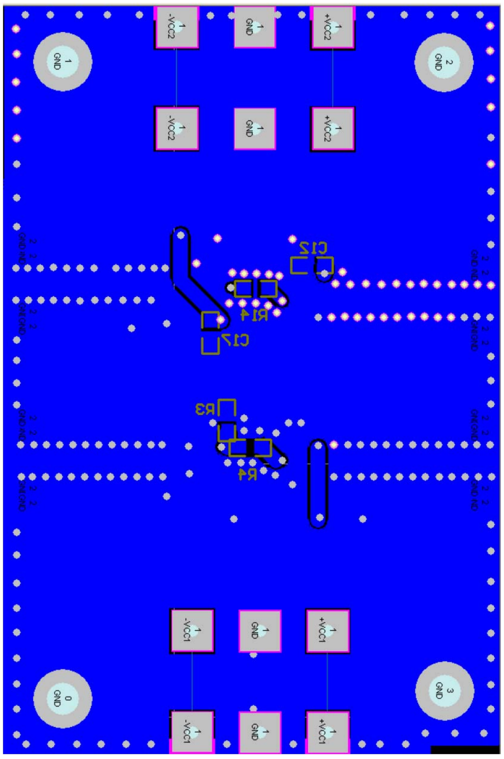


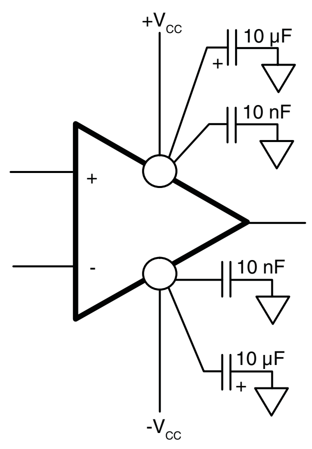
Figure 27. Bottom view layout



4 Power supply considerations

Correct power supply bypassing is very important to optimize performance in high-frequency ranges. The bypass capacitors should be placed as close as possible to the IC pins to improve high-frequency bypassing. A capacitor greater than 1 µF is necessary to minimize the distortion. For better quality bypassing, a 10 nF capacitor can be added. It should also be placed as close as possible to the IC pins. The bypass capacitors must be incorporated for both the negative and positive supply.

Figure 28. Circuit for power supply bypassing



4.1 Single power supply

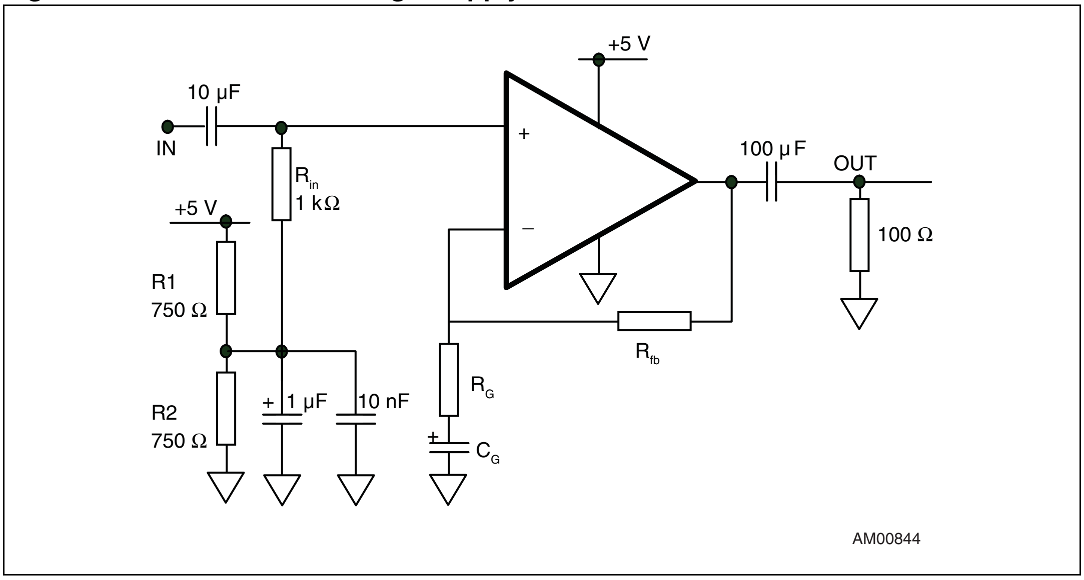
In the event that a single supply system is used, biasing is necessary to obtain a positive output dynamic range between 0 V and $+V_{CC}$ supply rails. Considering the values of V_{OH} and V_{OL}, the amplifier will provide an output swing from +0.9 V to +4.1 V on a 100 Ω load.

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 (35 µA maximum) as 1% of the current through the resistance divider, to keep a stable mid-supply, two resistances of 750 Ω can be used.

The input provides a high-pass filter with a break frequency below 10 Hz which is necessary to remove the original 0 V DC component of the input signal, and to set it at $+V_{CC}/2$.

Figure 29 on page 13 illustrates a 5 V single power supply configuration. A capacitor C_G is added in the gain network to ensure a unity gain in low frequencies to keep the right DC component at the output. C_G contributes to a high-pass filter with $R_{fb}//R_G$ and its value is calculated with regard to the cut-off frequency of this low-pass filter.

Figure 29. Circuit for +5 V single supply

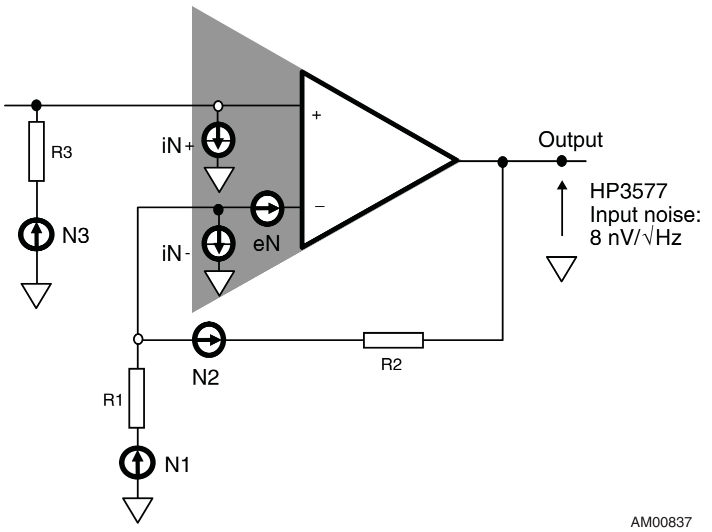


5 Noise measurements

The noise model is shown in *Figure 30*.

- eN: input voltage noise of the amplifier.
- iNn: negative input current noise of the amplifier.
- iNp: positive input current noise of the amplifier.

Figure 30. Noise model



The thermal noise of a resistance R is:

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

where ΔF is the specified bandwidth.

On a 1 Hz bandwidth the thermal noise is reduced to:

$$\sqrt{4kTR}$$

where k is the Boltzmann's constant, equal to 1,374.E(-23)J/°K. T is the temperature (°K).

The output noise eNo is calculated using the superposition theorem. However, eNo is not the simple sum of all noise sources, but rather the square root of the sum of the square of each noise source, as shown in *Equation 1*.

Equation 1

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

Equation 2

$$eNo^2 = eN^2 \times g^2 + iNn^2 \times R2^2 + iNp^2 \times R3^2 \times g^2 + \frac{R2^2}{R1} \times 4kTR1 + 4kTR2 + 1 + \frac{R2^2}{R1} \times 4kTR3$$

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

Equation 3

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

The input noise is called **equivalent input noise** because it is not directly measured but 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 *Equation 2* we obtain:

Equation 4

$$eNo^2 = eN^2 \times g^2 + iNn^2 \times R2^2 + iNp^2 \times R3^2 \times g^2 + g \times 4kTR2 + 1 + \frac{R2^2}{R1} \times 4kTR3$$

5.1 Measurement of the input voltage noise *eN*

If we assume a short-circuit on the non-inverting input (R3=0), from *Equation 4* we can derive:

Equation 5

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

In order to easily extract the value of eN, the resistance R2 will be chosen to be as low as possible. On the other hand, the gain must be large enough.

R3=0, gain: g=100

5.2 Measurement of the negative input current noise *iNn*

To measure the negative input current noise iNn, we set R3=0 and use *Equation 5*. This time, the gain must be lower in order to decrease the thermal noise contribution.

R3=0, gain: g=10

5.3 Measurement of the positive input current noise *iNp*

To extract iNp from *Equation 3*, 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.

R3=100 W, gain: g=10

6 Intermodulation distortion product

The non-ideal output of the amplifier can be described by the following series of equations.

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

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 to the intermodulation product.

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

In this case:

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

then:

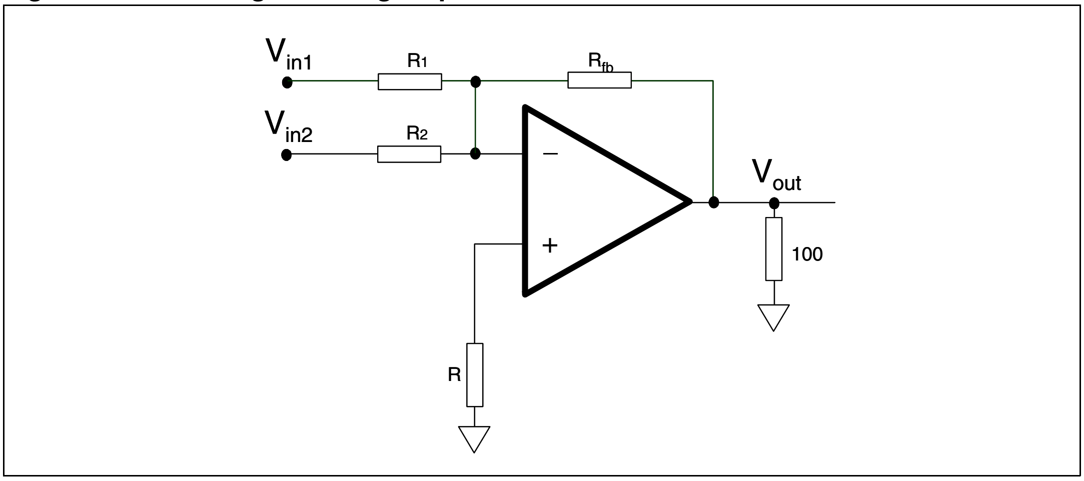
$$V_{out} = C_0 + C_1(A\sin\omega_1 t + A\sin\omega_2 t) + C_2(A\sin\omega_1 t + A\sin\omega_2 t)^2 \ldots + C_n(A\sin\omega_1 t + A\sin\omega_2 t)^n$$

From this expression, we can extract the distortion terms, and the intermodulation terms from a single sine wave.

- Second-order intermodulation terms IM2 by the frequencies $(\omega_1 - \omega_2)$ and $(\omega_1 + \omega_2)$ with an amplitude of $C2A^2$.
- Third-order intermodulation terms IM3 by the frequencies $(2\omega_1 - \omega_2)$, $(2\omega_1 + \omega_2)$, $(-\omega_1 + 2\omega_2)$ and $(\omega_1 + 2\omega_2)$ with an amplitude of $(3/4)C3A^3$.

The intermodulation product of the driver is measured by using the driver as a mixer in a summing amplifier configuration (*Figure 31*). In this way, the non-linearity problem of an external mixing device is avoided.

Figure 31. Inverting summing amplifier



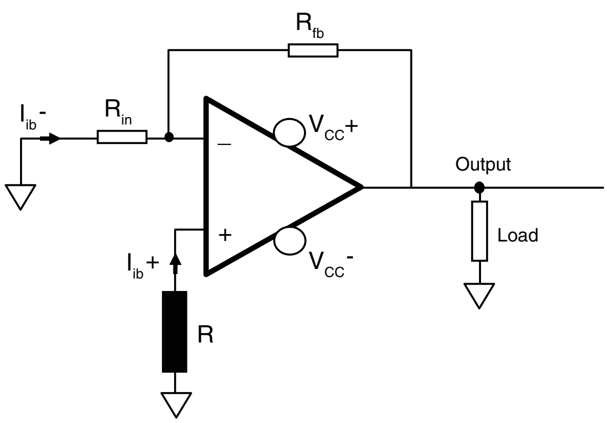
7 Inverting amplifier biasing

A resistance is necessary to achieve good input biasing, such as resistance R shown in *Figure 32*.

The value of this resistance is calculated from the negative and positive input bias current. The aim is to compensate for the offset bias current, which can affect the input offset voltage and the output DC component. Assuming I_{ib-}, I_{ib+}, R_{in}, R_{fb} and a 0 V output, the resistance R is:

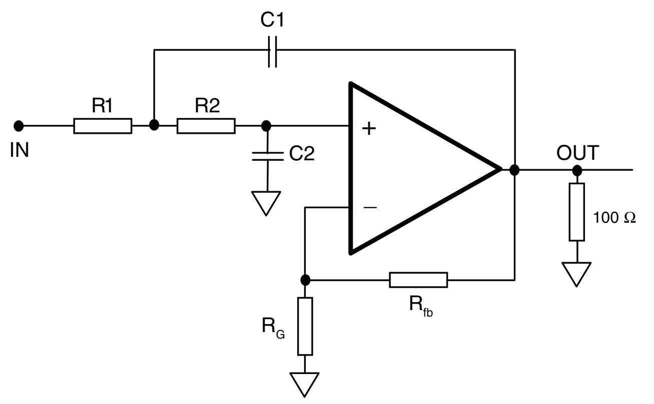
$$R = \frac{R_{in} \times R_{fb}}{R_{in} + R_{fb}}$$

Figure 32. Compensation of the input bias current



8 Active filtering

Figure 33. Low-pass active filtering, Sallen-Key



From the resistors R_{fb} and R_G we can directly calculate the gain of the filter in a classic non-inverting amplification configuration.

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

We 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 so becomes:

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

The damping factor is calculated by the following expression.

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

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 resistor and capacitor values. In the case of R1=R2=R:

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

Due to a limited selection of capacitor values in comparison with resistor values, we can set C1=C2=C, so that:

$$\zeta = \frac{2R_2 - R_1\frac{R_{fb}}{R_g}}{2\sqrt{R_1R_2}}$$

9 Package information

In order to meet environmental requirements, ST offers these devices in different grades of ECOPACK® packages, depending on their level of environmental compliance. ECOPACK® specifications, grade definitions and product status are available at: *www.st.com*. ECOPACK® is an ST trademark.

9.1 Ceramic Flat-8 package information

Figure 34. Ceramic Flat-8 package mechanical drawing

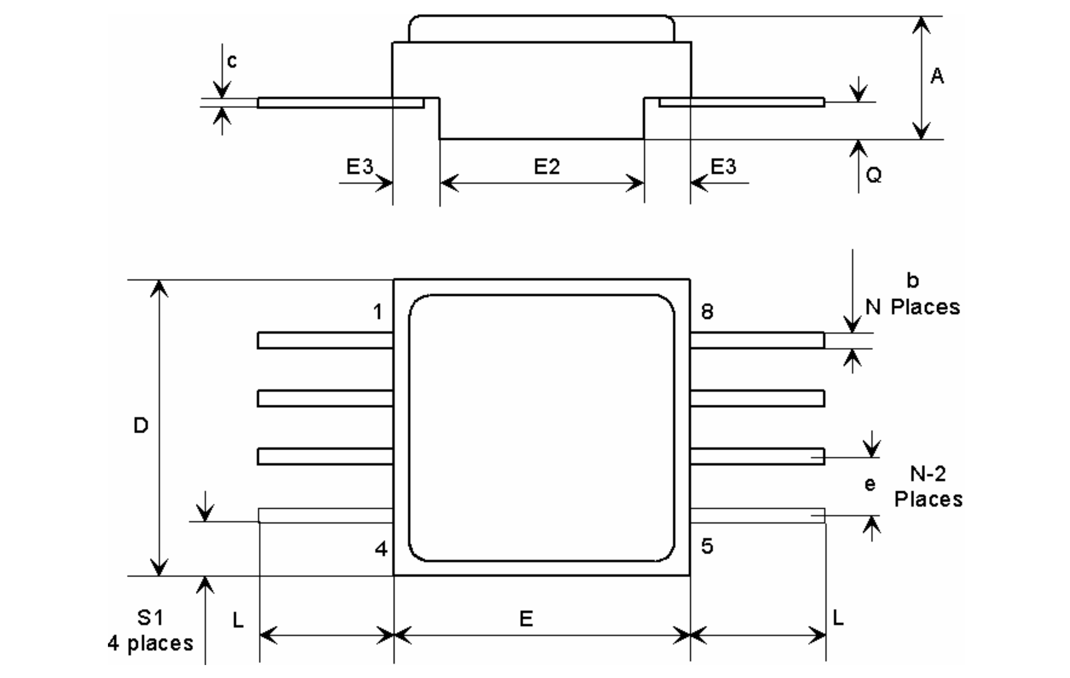


Table 5. Ceramic Flat-8 package mechanical data

Ref.	Dimensions					
	Millimeters			Inches		
	Min.	Typ.	Max.	Min.	Typ.	Max.
A	2.24	2.44	2.64	0.088	0.096	0.104
b	0.38	0.43	0.48	0.015	0.017	0.019
c	0.10	0.13	0.16	0.004	0.005	0.006
D	6.35	6.48	6.61	0.250	0.255	0.260
E	6.35	6.48	6.61	0.250	0.255	0.260
E2	4.32	4.45	4.58	0.170	0.175	0.180
E3	0.88	1.01	1.14	0.035	0.040	0.045
e		1.27			0.050	
L		3.00			0.118	
Q	0.66	0.79	0.92	0.026	0.031	0.092
S1	0.92	1.12	1.32	0.036	0.044	0.052
N	08			08		

10 Ordering information

Table 6. Order codes

Order code	Description	Temperature range	Package	Terminal finish	Marking
RHF350K-01V	Flight parts (QMLV)	-55°C to +125°C	Flat-8	Gold	TBD
RHF350K-02V	Flight parts (QMLV)	-55°C to +125°C	Flat-8	Solder	TBD
RHF350K1	Engineering samples	-55°C to +125°C	Flat-8	Gold	RHF350K1
RHF350K2	Engineering samples with 48-hrs burn-in	-55°C to +125°C	Flat-8	Gold	RHF350K2
RHF350DIE2V	Flight parts (QMLV)	-55°C to +125°C	Bare die	-	No marking

11 Revision history

Table 7. Document revision history

Date	Revision	Changes
20-May-2009	1	Initial release.