

FEATURES

Dual Matched PNP Transistor

Low Offset Voltage: 100 μ V Max

Low Noise: 1 nV/ $\sqrt{\text{Hz}}$ @ 1 kHz Max

High Gain: 100 Min

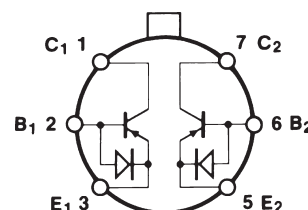
High Gain Bandwidth: 190 MHz Typ

Tight Gain Matching: 3% Max

Excellent Logarithmic Conformance: $r_{BE} \approx 0.3 \Omega$ typ

PIN CONNECTION

**TO-78
(H Suffix)**



GENERAL DESCRIPTION

The MAT03 dual monolithic PNP transistor offers excellent parametric matching and high frequency performance. Low noise characteristics (1 nV/ $\sqrt{\text{Hz}}$ max @ 1 kHz), high bandwidth (190 MHz typical), and low offset voltage (100 μ V max), makes the MAT03 an excellent choice for demanding preamplifier applications. Tight current gain matching (3% max mismatch) and high current gain (100 min), over a wide range of collector current, makes the MAT03 an excellent choice for current mirrors. A low value of bulk resistance (typically 0.3 Ω) also makes the MAT03 an ideal component for applications requiring accurate logarithmic conformance.

Each transistor is individually tested to data sheet specifications. Device performance is guaranteed at 25°C and over the extended industrial and military temperature ranges. To ensure the long-term stability of the matching parameters, internal protection diodes across the base-emitter junction clamp any reverse base-emitter junction potential. This prevents a base-emitter breakdown condition that can result in degradation of gain and matching performance due to excessive breakdown current.

REV. C

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MAT03—SPECIFICATIONS

ELECTRICAL CHARACTERISTICS (@ $T_A = 25^\circ\text{C}$, unless otherwise noted.)

Parameter	Symbol	Conditions	MAT03E			MAT03F			Unit
			Min	Typ	Max	Min	Typ	Max	
Current Gain ¹	h_{FE}	$V_{CB} = 0\text{ V}, -36\text{ V}$ $I_C = 1\text{ mA}$ $I_C = 100\text{ }\mu\text{A}$ $I_C = 10\text{ }\mu\text{A}$	100 90 80	165 150 120		80 70 60	165 150 120		
Current Gain Matching ²	Δh_{FE}	$I_C = 100\text{ }\mu\text{A}, V_{CB} = 0\text{ V}$		0.5	3		0.5	6	%
Offset Voltage ³	V_{OS}	$V_{CB} = 0\text{ V}, I_C = 100\text{ }\mu\text{A}$		40	100		40	200	μV
Offset Voltage Change vs. Collector Voltage	$\Delta V_{OS}/\Delta V_{CB}$	$I_C = 100\text{ }\mu\text{A}$ $V_{CB1} = 0\text{ V}$ $V_{CB2} = -36\text{ V}$		11 11	150 150		11 11	200 200	μV μV
Offset Voltage Change vs. Collector Current	$\Delta V_{OS}/\Delta I_C$	$V_{CB} = 0\text{ V}$ $I_{C1} = 10\text{ }\mu\text{A}, I_{C2} = 1\text{ mA}$		12 12	50 50		12 12	75 75	μV μV
Bulk Resistance	r_{BE}	$V_{CB} = 0\text{ V}$ $10\text{ }\mu\text{A} \leq I_C \leq 1\text{ mA}$		0.3 0.3	0.75 0.75		0.3 0.3	0.75 0.75	Ω Ω
Offset Current Collector-Base Leakage Current	I_{OS}	$I_C = 100\text{ }\mu\text{A}, V_{CB} = 0\text{ V}$		6	35		6	45	nA
Noise Voltage Density ⁴	I_{CB0} e_N	$V_{CB} = -36\text{ V} = V_{MAX}$ $I_C = 1\text{ mA}, V_{CB} = 0$ $f_o = 10\text{ Hz}$ $f_o = 100\text{ Hz}$ $f_o = 1\text{ kHz}$ $f_o = 10\text{ kHz}$		50 0.8 0.7 0.7 0.7	200		50 0.8 0.7 0.7 0.7	400	pA $\text{nV}/\sqrt{\text{Hz}}$ $\text{nV}/\sqrt{\text{Hz}}$ $\text{nV}/\sqrt{\text{Hz}}$ $\text{nV}/\sqrt{\text{Hz}}$
Collector Saturation Voltage	$V_{CE(SAT)}$	$I_C = 1\text{ mA}, I_B = 100\text{ }\mu\text{A}$		0.025	0.1		0.025	0.1	V

ELECTRICAL CHARACTERISTICS (@ $-40^\circ\text{C} \leq T_A \leq 85^\circ\text{C}$, unless otherwise noted.)

Parameter	Symbol	Conditions	MAT03E			MAT03F			Unit
			Min	Typ	Max	Min	Typ	Max	
Current Gain	h_{FE}	$V_{CB} = 0\text{ V}, -36\text{ V}$ $I_C = 1\text{ mA}$ $I_C = 100\text{ }\mu\text{A}$ $I_C = 10\text{ }\mu\text{A}$		70 60 50	120 105 90		60 50 40	120 105 90	
Offset Voltage	V_{OS}	$I_C = 100\text{ }\mu\text{A}, V_{CB} = 0\text{ V}$		30	135		30	265	μV
Offset Voltage Drift ⁵	TCV_{OS}	$I_C = 100\text{ }\mu\text{A}, V_{CB} = 0\text{ V}$		0.3	0.5		0.3	1.0	$\mu\text{V}/^\circ\text{C}$
Offset Current	I_{OS}	$I_C = 100\text{ }\mu\text{A}, V_{CB} = 0\text{ V}$		10	85		10	200	nA
Breakdown Voltage	BV_{CEO}			36			36		V

NOTES

¹Current gain is measured at collector-base voltages (V_{CB}) swept from 0 to V_{MAX} at indicated collector current. Typicals are measured at $V_{CB} = 0\text{ V}$.

²Current gain matching (Δh_{FE}) is defined as: $\Delta h_{FE} = \frac{100 (\Delta I_B) h_{FE}(\text{min})}{I_C}$.

³Offset voltage is defined as: $V_{OS} = V_{BE1} - V_{BE2}$, where V_{OS} is the differential voltage for $I_{C1} = I_{C2}$: $V_{OS} = V_{BE1} - V_{BE2} = \frac{KT}{q} \ln \left(\frac{I_{C1}}{I_{C2}} \right)$.

⁴Sample tested. Noise tested and specified as equivalent input voltage for each transistor.

⁵Guaranteed by V_{OS} test ($TCV_{OS} = V_{OS}/T$ for $V_{OS} \ll V_{BE}$) where $T = 298\text{ K}$ for $T_A = 25^\circ\text{C}$.

Specifications subject to change without notice.

ORDERING GUIDE

Model	V _{OS} max (T _A = +25°C)	Temperature Range	Package Option
MAT03EH	100 µV	–40°C to +85°C	TO-78
MAT03FH	200 µV	–40°C to +85°C	TO-78

ABSOLUTE MAXIMUM RATINGS¹

Collector-Base Voltage (BV _{CBO})	36 V
Collector-Emitter Voltage (BV _{CEO})	36 V
Collector-Collector Voltage (BV _{CC})	36 V
Emitter-Emitter Voltage (BV _{EE})	36 V
Collector Current (I _C)	20 mA
Emitter Current (I _E)	20 mA
Total Power Dissipation	
Ambient Temperature ≤ 70°C ²	500 mW
Operating Temperature Range	
MAT03E/F	–40°C to +85°C
Operating Junction Temperature	–55°C to +150°C
Storage Temperature	–65°C to +150°C
Lead Temperature (Soldering, 60 sec)	300°C
Junction Temperature	–65°C to +150°C

NOTES

¹Absolute maximum ratings apply to both DICE and packaged devices.

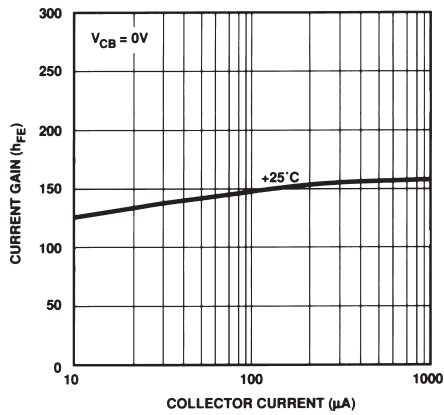
²Rating applies to TO-78 not using a heat sink and LCC; devices in free air only. For TO-78, derate linearly at 6.3 mW/°C above 70°C ambient temperature; for LCC, derate at 7.8 mW/°C.

CAUTION

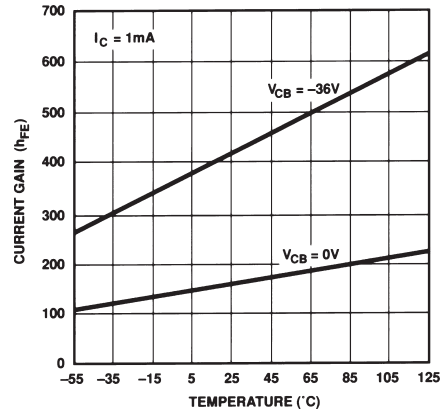
ESD (electrostatic discharge) sensitive device. Electrostatic charges as high as 4000 V readily accumulate on the human body and test equipment and can discharge without detection. Although the MAT03 features propriety ESD protection circuitry, permanent damage may occur on devices subjected to high-energy electrostatic discharges. Therefore, proper ESD precautions are recommended to avoid performance degradation or loss of functionality.



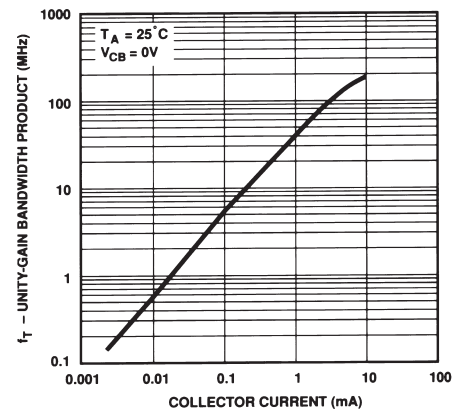
MAT03—Typical Performance Characteristics



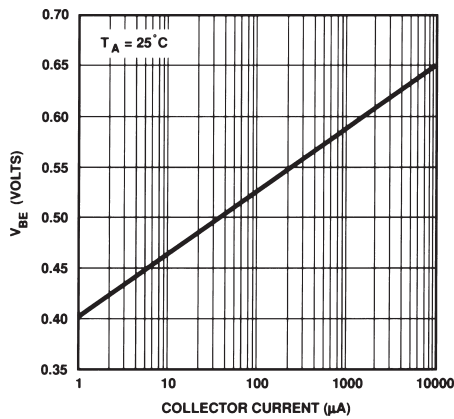
TPC 1. Current Gain vs. Collector Current



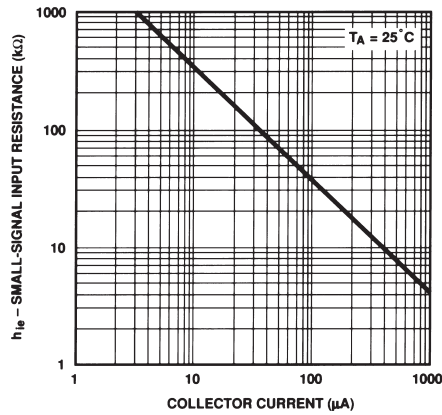
TPC 2. Current Gain vs. Temperature



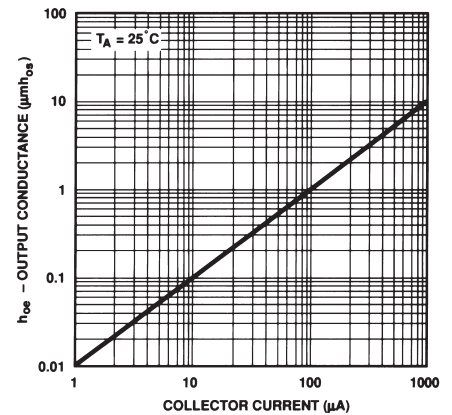
TPC 3. Gain Bandwidth vs. Collector Current



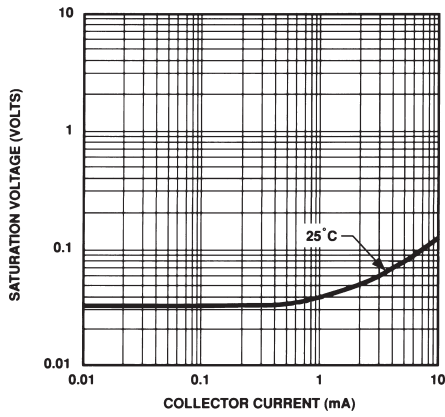
TPC 4. Base-Emitter Voltage vs. Collector Current



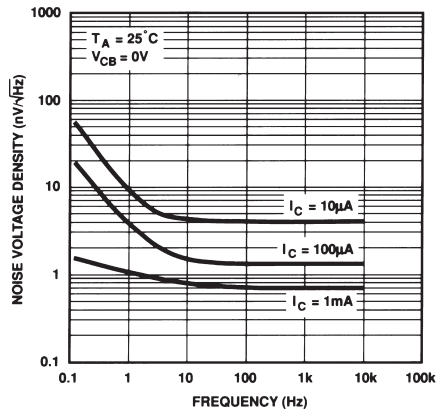
TPC 5. Small-Signal Input Resistance (h_{ie}) vs. Collector Current



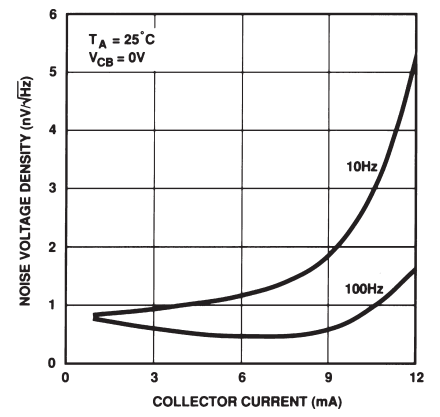
TPC 6. Small Signal Output Conductance (h_{oe}) vs. Collector Current



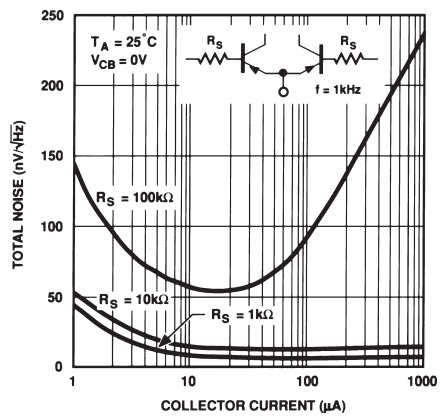
TPC 7. Saturation Voltage vs. Collector Current



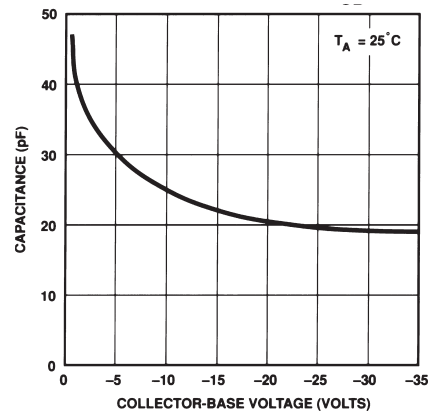
TPC 8. Noise Voltage Density vs. Frequency



TPC 9. Noise Voltage Density vs. Collector Current

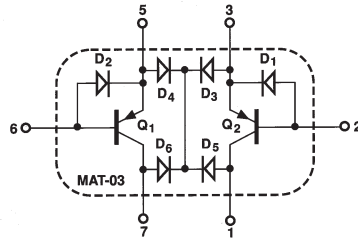


TPC 10. Total Noise vs. Collector Current



TPC 11. Collector-Base Capacitance vs. V_{CB}

MAT03



SPICE MODEL PARAMETERS

•SUBCKT MAT03 1 2 3 5 6 7
 Q1 1 2 3 PMAT
 Q2 7 6 5 PMAT
 D1 2 3 DMAT1
 D2 6 5 DMAT1
 D3 3 4 DMAT1
 D4 5 4 DMAT1
 D5 1 4 DMAT2
 D6 7 4 DMAT2
 •MODEL DMAT1 D IS = 7.2E-16 RS = 20
 •MODEL DMAT2 D IS = 1E-14 VJ = 0.6 CJ = 68P
 •MODEL PMAT PNP BF = 160 IS = 1.4E-13 VAF = 60 BR = 5 VAR = 7 RB = 16
 + PC = 12 RE = 0.35 CJE = 57P VJE = 0.7 MJE = 0.4 IF = 1.08E-9
 + TR = 3E-8 CJC = 40P VJC = 0.55 MJC = 0.5 CJS = 0 IKF = 160M)

SABER* MODEL PARAMETERS

template mat03 n1 n2 n3 n5 n6 n7
 electrical n1,n2,n3,n5,n6,n7
 {
 d..model dmat1 = (is = 7.2e-16,rs = 20)
 d..model dmat2 = (is = 1e-14,vj = 0.6,cjo = 68p)
 q..model pmat = (type = __ p,bf = 160,is = 1.4e-13,vaf = 60,br = 5,var = 7,
 rb = 16,rc = 12,re = 0.45,cje = 57p,vje = 0.7,mje = 0.4,tf = 1.08e-9,tr = 3e-8,
 cjc = 40p,vjc = 0.55,mjc = 0.5,cjs = 0,ikf = 160m)
 q.q1 n1 n2 n3 n4 = model = pmat
 q.q2 n7 n6 n5 n4 = model = pmat
 d.d1 n2 n3 = model = dmat1
 d.d2 n6 n5 = model = dmat1
 d.d3 n3 n4 = model = dmat1
 d.d4 n5 n4 = model = dmat1
 d.d5 n1 n4 = model = dmat2
 d.d6 n7 n4 = model = dmat2
 }

*SABER is a registered trademark of Analogly Inc.

Figure 1. SPICE or SABER Model

APPLICATIONS INFORMATION

MAT03 MODELS

The MAT03 model (Figure 1) includes parasitic diodes D₃ through D₆. D₁ and D₂ are internal protection diodes that prevent zenering of the base-emitter junctions.

The analysis programs, SPICE and SABER, are primarily used in evaluating the functional performance of systems. The models are provided only as an aid in using these simulation programs.

MAT03 NOISE MEASUREMENT

All resistive components (Johnson noise, $e_n^2 = 4kTBR$, or $e_n = 0.13\sqrt{R}$ nV/ \sqrt{Hz} , where R is in k Ω) and semiconductor junctions (shot noise, caused by current flowing through a junction, produces voltage noise in series impedances such as transistor-collector load resistors, $I_n = 0.566 \sqrt{I}$ pA/ \sqrt{Hz} where I is in μ A) contribute to the system input noise.

Figure 2 illustrates a technique for measuring the equivalent input noise voltage of the MAT03. 1 mA of stage current is used

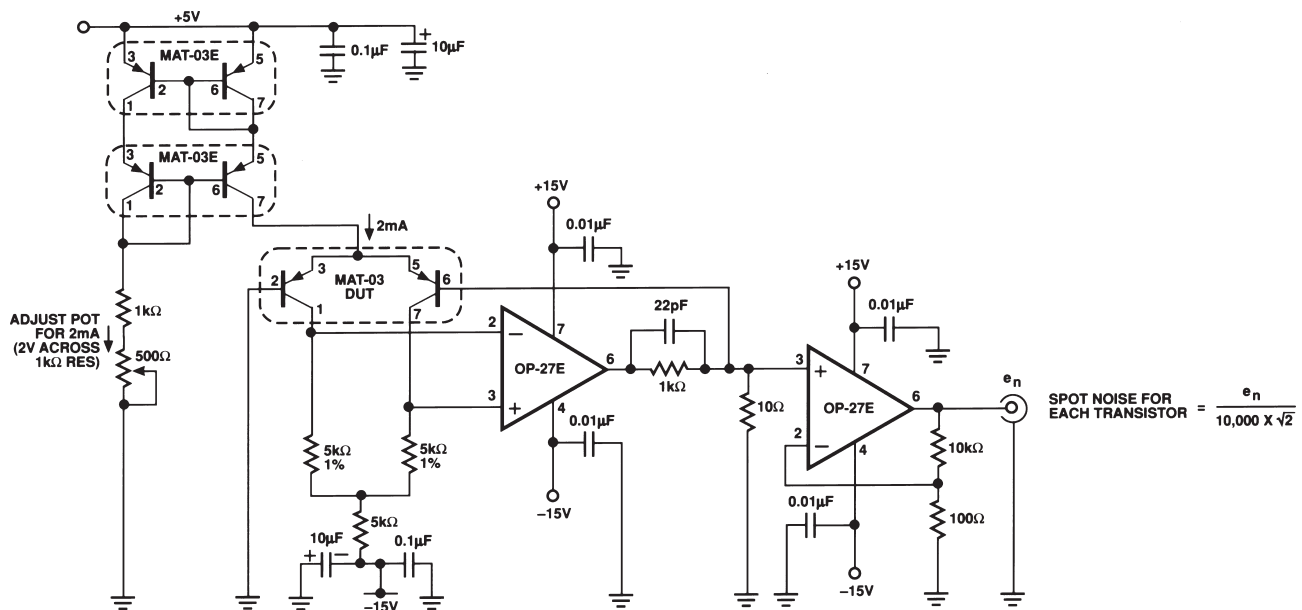


Figure 2. MAT03 Voltage Noise Measurement Circuit

to bias each side of the differential pair. The 5 k Ω collector resistors noise contribution is insignificant compared to the voltage noise of the MAT03. Since noise in the signal path is referred back to the input, this voltage noise is attenuated by the gain of the circuit. Consequently, the noise contribution of the collector load resistors is only 0.048 nV/ $\sqrt{\text{Hz}}$. This is considerably less than the typical 0.8 nV/ $\sqrt{\text{Hz}}$ input noise voltage of the MAT03 transistor.

The noise contribution of the OP27 gain stages is also negligible due to the gain in the signal path. The op amp stages amplify the input referred noise of the transistors to increase the signal strength to allow the noise spectral density ($e_{\text{in}} \times 10000$) to be measured with a spectrum analyzer. Since we assume equal noise contributions from each transistor in the MAT03, the output is divided by $\sqrt{2}$ to determine a single transistor's input noise.

Air currents cause small temperature changes that can appear as low frequency noise. To eliminate this noise source, the mea-

surement circuit must be thermally isolated. Effects of extraneous noise sources must also be eliminated by totally shielding the circuit.

SUPER LOW NOISE AMPLIFIER

The circuit in Figure 3a is a super low noise amplifier with equivalent input voltage noise of 0.32 nV/ $\sqrt{\text{Hz}}$. By paralleling three MAT03 matched pairs, a further reduction of amplifier noise is attained by a reduction of the base spreading resistance by a factor of 3, and consequently the noise by $\sqrt{3}$. Additionally, the shot noise contribution is reduced by maintaining a high collector current (2 mA/device) which reduces the dynamic emitter resistance and decreases voltage noise. The voltage noise is inversely proportional to the square root of the stage current, and current noise increases proportionally to the square root of the stage current. Accordingly, this amplifier capitalizes on voltage noise reduction techniques at the expense of increasing the current noise. However, high current noise is not usually important when dealing with low impedance sources.

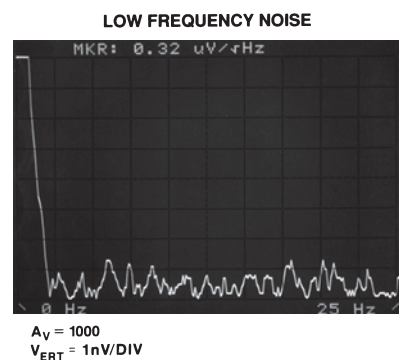
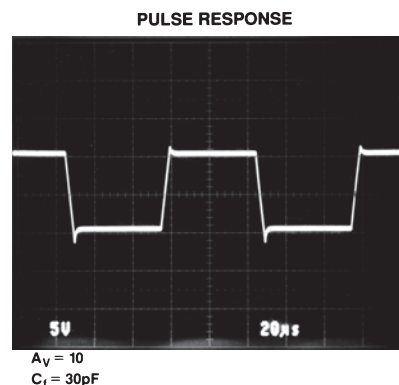
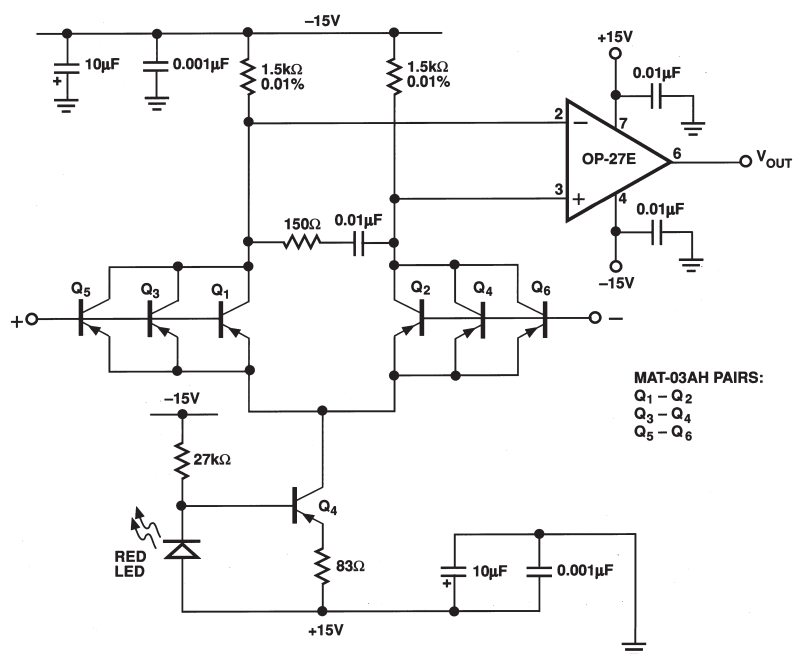


Figure 3a. Super Low Noise Amplifier

MAT03

This amplifier exhibits excellent full power ac performance, 0.08% THD into a 600 Ω load, making it suitable for exacting audio applications (see Figure 3b).

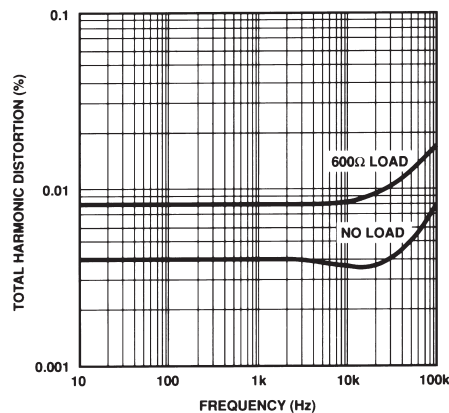


Figure 3b. Super Low Noise Amplifier—Total Harmonic Distortion

LOW NOISE MICROPHONE PREAMPLIFIER

Figure 4 shows a microphone preamplifier that consists of a MAT03 and a low noise op amp. The input stage operates at a relatively high quiescent current of 2 mA per side, which reduces the MAT03 transistor's voltage noise. The $1/f$ corner is less than 1 Hz. Total harmonic distortion is under 0.005% for a 10 V p-p signal from 20 Hz to 20 kHz. The preamp gain is 100, but can be modified by varying R_5 or R_6 ($V_{OUT}/V_{IN} = R_5/R_6 + 1$).

A total input stage emitter current of 4 mA is provided by Q_2 . The constant current in Q_2 is set by using the forward voltage of a GaAsP LED as a reference. The difference between this voltage

and the V_{BE} of a silicon transistor is predictable and constant (to a few percent) over a wide temperature range. The voltage difference, approximately 1 V, is dropped across the 250 Ω resistor which produces a temperature stabilized emitter current.

CURRENT SOURCES

A fundamental requirement for accurate current mirrors and active load stages is matched transistor components. Due to the excellent V_{BE} matching (the voltage difference between V_{BE} s required to equalize collector current) and gain matching, the MAT03 can be used to implement a variety of standard current mirrors that can source current into a load such as an amplifier stage. The advantages of current loads in amplifiers versus resistors is an increase of voltage gain due to higher impedances, larger signal range, and in many applications a wider signal bandwidth.

Figure 5 illustrates a cascode current mirror consisting of two MAT03 transistor pairs.

The cascode current source has a common base transistor in series with the output which causes an increase in output impedance of the current source since V_{CE} stays relatively constant. High frequency characteristics are improved due to a reduction of Miller capacitance. The small-signal output impedance can be determined by consulting “ h_{OF} vs. Collector Current” typical graph. Typical output impedance levels approach the performance of a perfect current source.

Considering a typical collector current of 100 μA , we have:

$$r_{OQ3} = \frac{1}{1.0 \mu MHOS} = 1 M\Omega$$

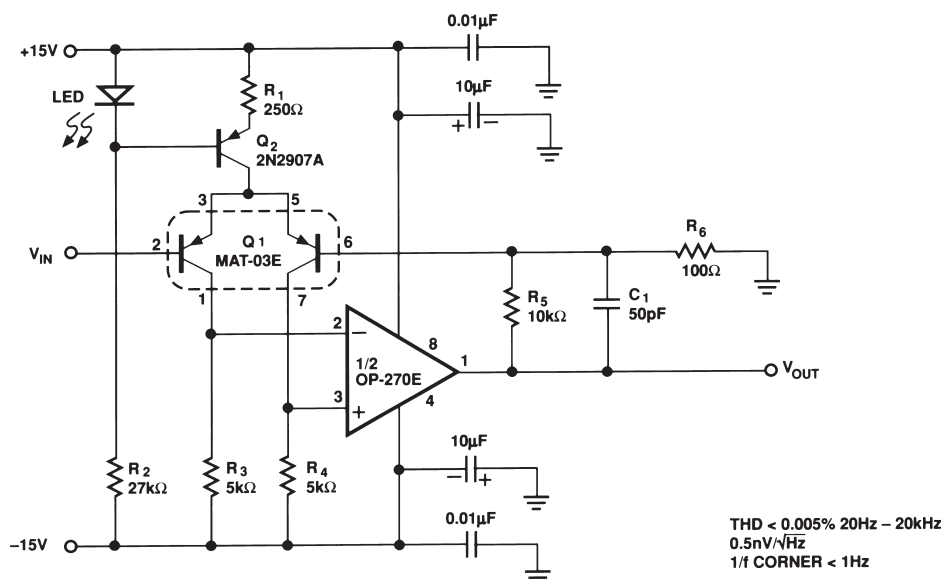


Figure 4. Low Noise Microphone Preamplifier

Q_2 and Q_3 are in series and operate at the same current levels so the total output impedance is:

$$R_O = h_{FE} r_{OQ3} @ (160)(1\text{ M}\Omega) = 160\text{ M}\Omega.$$

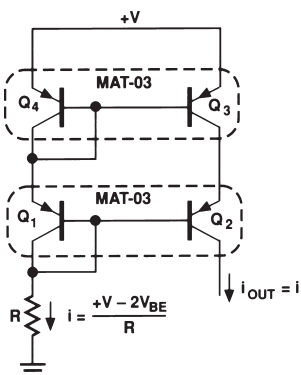


Figure 5. Cascode Current Source

CURRENT MATCHING

The objective of current source or mirror design is generation of currents that are either matched or must maintain a constant ratio. However, mismatch of base emitter voltages cause output current errors. Consider the example of Figure 5. If the resistors and transistors are equal and the collector voltages are the same, the collector currents will match precisely. Investigating the current matching errors resulting from a nonzero V_{OS} , we define ΔI_C as the current error between the two transistors.

Graph 6b describes the relationship of current matching errors versus offset voltage for a specified average current I_C . Note that since the relative error between the currents is exponentially proportional to the offset voltage, tight matching is required to design high accuracy current sources. For example, if the offset voltage is 5 mV at 100 μ A collector current, the current matching error would be 20%. Additionally, temperature effects such as offset drift (3 μ V/ $^{\circ}$ C per mV of V_{OS}) will degrade performance if Q_1 and Q_2 are not well matched.

DIGITALLY PROGRAMMABLE BIPOLAR CURRENT PUMP

The circuit of Figure 7 is a digitally programmable current pump. The current pump incorporates a DAC08, and a fast Wilson current source using the MAT03. Examining Figure 7, the DAC08 is set for 2 mA full-scale range so that bipolar current operation of ± 2 mA is achieved. The Wilson current mirror maintains linearity within the LSB range of the 8-bit DAC08 (± 2 mA/256 = 15.6 μ A resolution) as seen in Figure 8. A negative feedback path established by Q_2 regulates the collector current so that it matches the reference current programmed by the DAC08.

Collector-emitter voltages across both Q_1 and Q_3 are matched by D_1 , with Q_3 's collector-emitter voltage remaining constant, independent of the voltage across the current source output.

Since Q_2 buffers Q_3 , both transistors in the MAT03, Q_1 and Q_3 , maintain the same collector current. D_2 and D_3 form a Baker clamp which prevents Q_2 from turning off, thereby improving the switching speed of the current mirror. The feedback serves to increase the output impedance and improves accuracy by reducing the base-width modulation which occurs with varying collector-emitter voltages. Accuracy and linearity performance of the current pump is summarized in Figure 8.

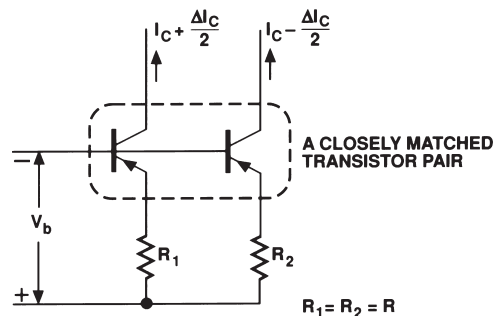


Figure 6a. Current Matching Circuit

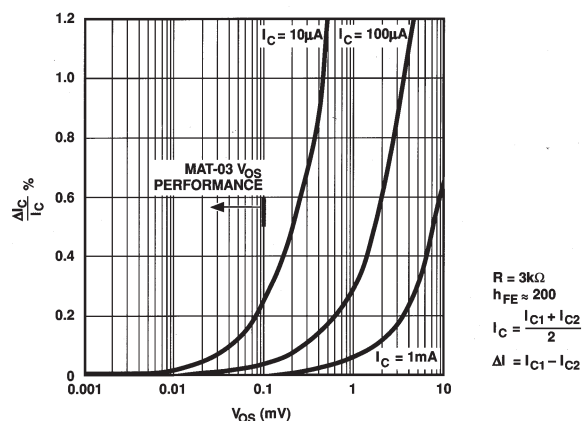


Figure 6b. Current Matching Accuracy % vs. Offset Voltage

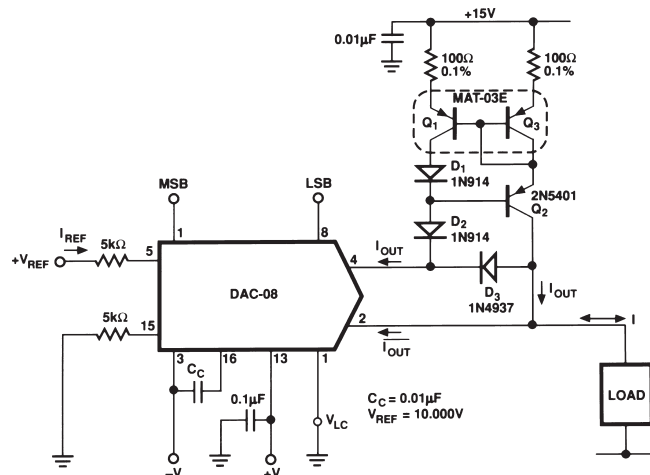


Figure 7. Digitally Programmable Bipolar Current Pump

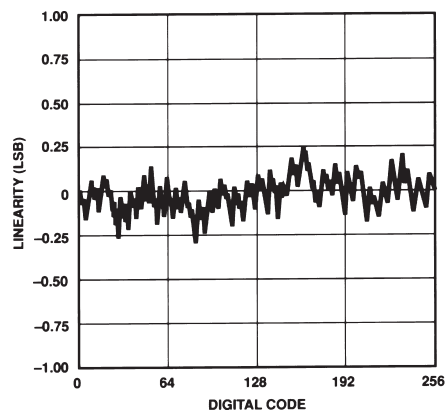


Figure 8. Digitally Programmable Current Pump—INL Error as Digital Code

The full-scale output of the DAC08, I_{OUT} , is a linear function of I_{REF}

$$I_{FR} = \frac{256}{256} \times I_{REF}, \text{ and } I_{OUT} + \overline{I_{OUT}} = I_{REF} \frac{256}{256}$$

The current mirror output is $I_{OUT} - \overline{I_{OUT}} = 1$, so that if

$$I_{REF} = 2 \text{ mA};$$

$$I = 2 I_{OUT} - 1.992 \text{ mA}$$

$$= 2 \left(\frac{\text{Input Code}}{256} \right) (2 \text{ mA}) - 1.992 \text{ mA}.$$

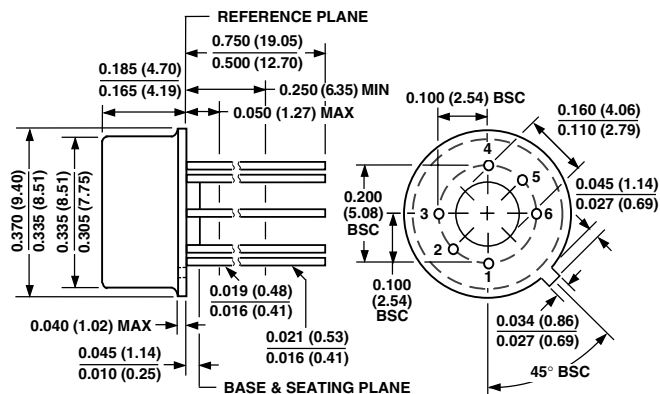
DIGITAL CURRENT PUMP CODING

	Digital Input B1 . . . B8	Output Current
FULL RANGE	1111 1111	I = 1.992 mA
HALF RANGE	1000 0000	I = 0.008 mA
ZERO SCALE	0000 0000	I = -1.992 mA

Dimensions shown in inches and (mm).

Dimensions shown in inches and (mm).

TO-78 Metal Can



Revision History

Location	Page
Data Sheet changed from REV. B to REV. C.	
Edits to ELECTRICAL CHARACTERISTICS	2
Deleted WAFER TEST LIMITS	3
Deleted DICE CHARACTERISTICS	3
Edits to ORDERING GUIDE	3
Edits to ABSOLUTE MAXIMUM RATINGS	3

