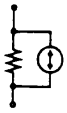
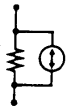
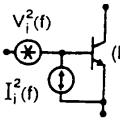
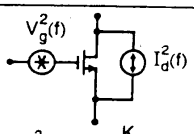
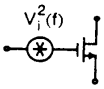
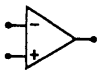
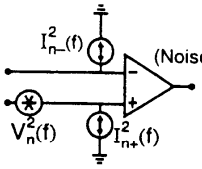
Curso 2003-2004
Tabla de ruido
hasta AN9-5 y
[illegible]
[illegible]

Element	Noise Models	
Resistor	R (Noiseless) $V_R^2(f) = 4kTR$	R (Noiseless)  $I_R^2(f) = \frac{4kT}{R}$
Diode (Forward biased)	$r_d = \frac{kT}{qI_D}$ (Noiseless) $V_d^2(f) = 2kTr_d$	$r_d = \frac{kT}{qI_D}$  $I_d^2(f) = 2qI_D$ (Noiseless)
BJT (Active region)	 $V_i^2(f)$ $I_i^2(f)$ (Noiseless)	$V_i^2(f) = 4kT\left(r_b + \frac{1}{2g_m}\right)$ $I_i^2(f) = 2q\left(I_B + \frac{KI_B}{f} + \frac{I_C}{\lvert\beta(f)\rvert^2}\right)$
MOSFET (Active region)	 $V_g^2(f)$ $I_d^2(f)$ $V_g^2(f) = \frac{K}{WLC_{ox}f}$ $I_d^2(f) = 4kT\left(\frac{2}{3}\right)g_m$	 $V_i^2(f)$ (Noiseless) $V_i^2(f) = 4kT\left(\frac{2}{3}\right)\frac{1}{g_m} + \frac{K}{WLC_{ox}f}$ Simplified model for low and moderate frequencies
Opamp 	 $I_{n-}^2(f)$ (Noiseless) $V_n^2(f)$ $I_{n+}^2(f)$	$V_n(f)$, $I_{n-}(f)$, $I_{n+}(f)$ — Values depend on opamp — Typically, all uncorrelated

Circuit elements and their noise models. Note that capacitors and inductors do not generate noise.

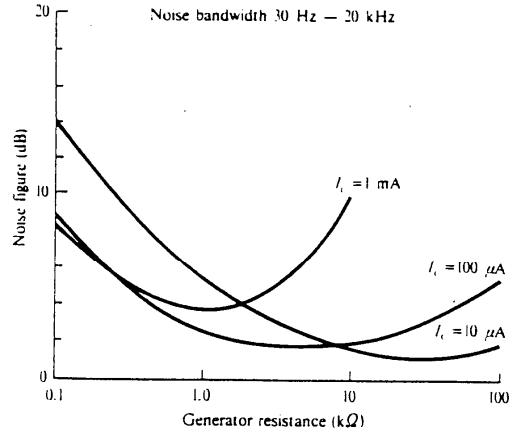
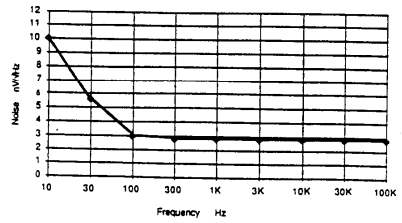


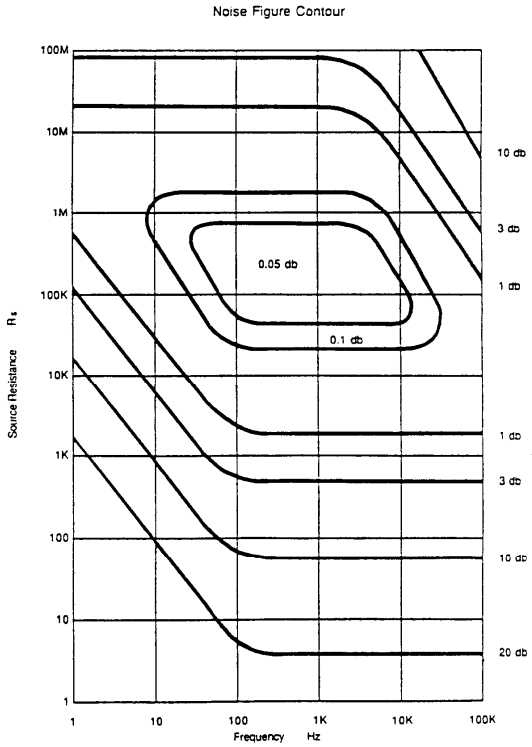
Fig. 5.9. Typical plot of noise figure against generator source resistance for a small silicon transistor such as BC109.

SR550 Preamplifier Specification Summary

Input Impedance	100 MΩ + 25 pF
Inputs	Single ended or differential (switch selectable)
Maximum Inputs	250 mV RMS for overload 100 Vdc, 10 Vac damage threshold
Noise	2.8 nV/√Hz at 1000 Hz (typical) 3.0 nV/√Hz at 100 Hz 10 nV/√Hz at 10 Hz
Common Mode	Range: 1 Volt peak Rejection: 110 dB at 100 Hz 110 dB at 1 kHz 90 dB at 10 kHz 70 dB at 100 kHz
Gain	1,2,5,10 Automatically set by SR510 or SR530 Lock-in depending on full scale sensitivity and dynamic reserve. Sensitivity ranges from 10 nV to 200 mV full scale (expand off)
Gain Accuracy	1% (2 Hz to 100 kHz)
Gain Stability	100 ppm/ C
Outputs	(A) single ended (600 Ω impedance) (B) shielded ground
Maximum Output	7 Volts peak to peak
Power	Supplied by SR510 or SR530 Lock-in via connector cable.
Mechanical	1.3" X 3.0" X 5.1", 1 lb.
Warranty	One year parts and labor on materials and workmanship.



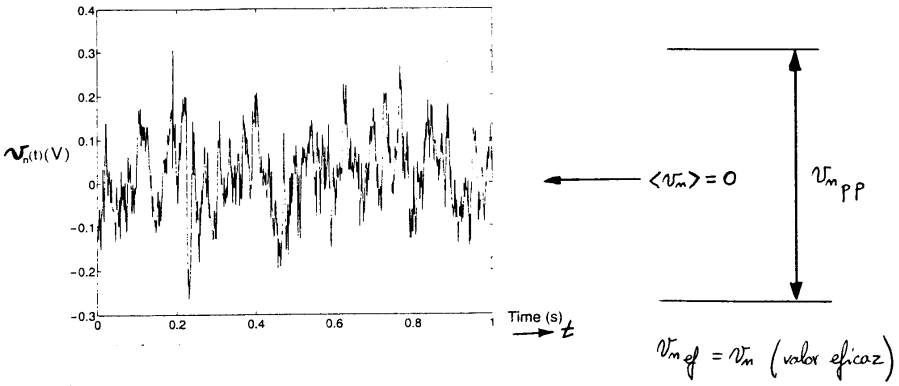
73



74

RUIDO (más de cerca)

DADO el 27-4-99 a las 9 horas. 14-12-00



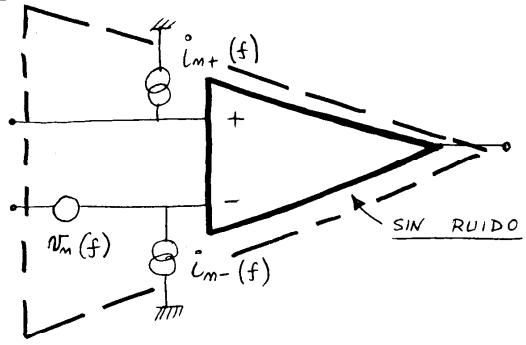
$v_{m_{pp}}$: tensión de ruido pico-pico

Debido a las propiedades estadísticas del ruido (el 99% de las veces que midamos o el 99% del tiempo, queda por debajo de 5 veces su valor eficaz) se puede decir que:

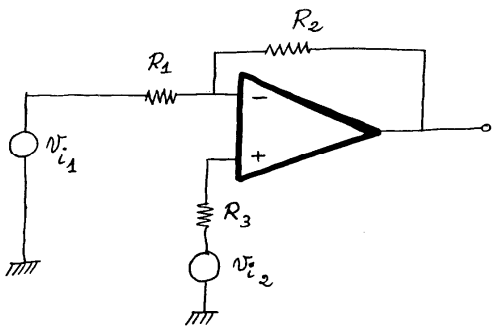
$v_{m_{pp}} \approx 5 \cdot v_{m_{ef}}$ (KEITLEY, 1972) INSTRUM. DE MEDIDA

$v_{m_{pp}} \approx 6 \cdot v_{m_{ef}}$ (LINEAR TECHNOLOGY, 1988) FABRIC. A.O. y C.I.

En un A.O. tenemos:



pero un A.O. suele estar realimentado negativamente (RN) por lo que hay que considerar más cosas....

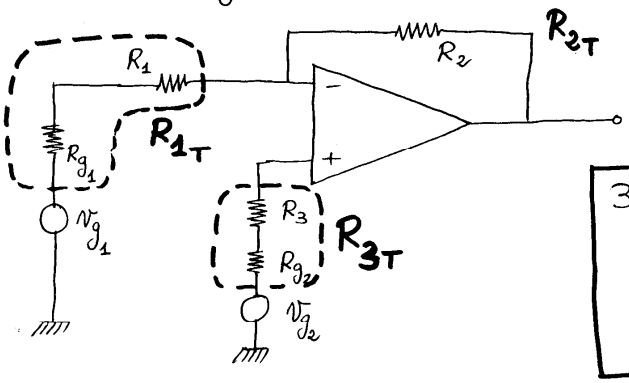


$$v_o = \underbrace{\frac{-R_2}{R_1} v_{i_1}}_{\text{INVERSOR}} + \underbrace{\left(1+\frac{R_2}{R_1}\right) v_{i_2}}_{\text{NO-INVERSOR}}$$

Si existe v_{i_1} y $v_{i_2}=0$ tenemos un amplificador inversor y si por el contrario $v_{i_1}=0$ y existe v_{i_2}, tenemos un amplif. no-inversor.

PARA EL RUIDO DA IGUAL, $\boxed{\text{EXISTEN } R_1, R_2, R_3 \text{ y el A.O.}}$

Hay que usar el "sentido común" al trabajar con fuentes de señal reales, no como v_{i_1} y v_{i_2}, sino con cierta R_g (equivalente Thèv. o Norton)



$\boxed{\begin{array}{c}\text{3 RESISTENCIAS} \\ R_{1T} \quad R_{2T} \quad y \quad R_{3T} \\ + \\ \text{A.O.}\end{array}}$

Nótese que v_{g_1} y R_{g_1} podían ser una sola resistencia R_{g_1} con su ruido termico asociado $v_{g_1} = \sqrt{4kTR_{g_1}(BW)}$ y análogamente v_{g_2} y R_{g_2}.

En este caso el ruido v_{g_1} aparece a la salida multiplicado por el factor $\frac{-R_{2T}}{R_1+R_{g_1}}$ que depende de R_{g_1} a su vez: $v_o = \sqrt{4kT(BW)\cdot R_{g_1}} \cdot \frac{-R_2}{R_1+R_{g_1}}$

por lo que la potencia de ruido de salida no es lineal con R_{g_1} (no es proporcional a R_{g_1})

Sin embargo, en el caso de $v_{g_2} = \sqrt{4KT(BW)\cdot R_{g_2}}$ las cosas son más "lineales":

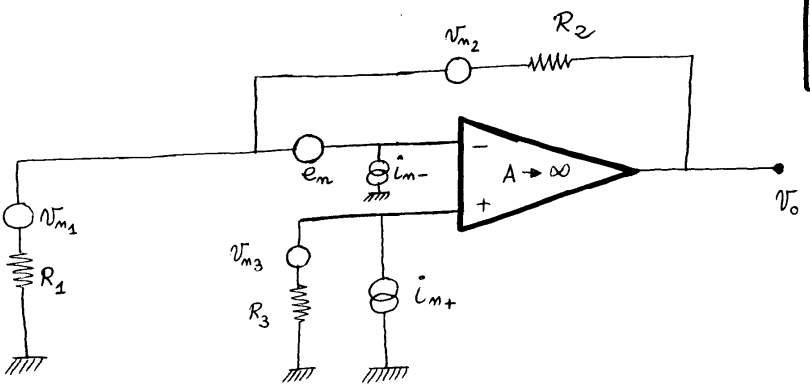
el ruido v_{g_2} aparece a la salida multiplicado por el factor: $\left(1 + \frac{R_{2T}}{R_{1T}}\right)$ que <u>NO</u> depende de la resistencia R_{g_2} que lo genera, por lo que podemos decir que el ruido v_{g_2} aparece a la salida amplificado por <u>el factor</u> $\left(1+\frac{R_{2T}}{R_{1T}}\right)$. De ahí el nombre de "<u>GANANCIA DE RUIDO</u>" dado a este factor.

La potencia de ruido a la salida, debida a R_{g_2}, sí es lineal o proporcional a R_{g_2} (que podría usarse como un sensor de temperatura absoluta poniendo $R_3 = 0$).

<u>CALCULO DE RUIDO EN A.O. CON R.N.</u>

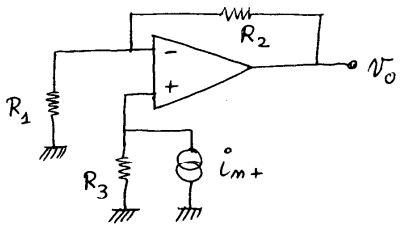
$R_{1T} = R_1$
$R_{2T} = R_2$
$R_{3T} = R_3$

MENOS LETRAS



Aplicamos superposición y además debido a $(RN + A \to \infty)$ las tensiones en las patillas (+) y (−) serán iguales.

<u>Empezamos por $i_{n+} \neq 0$</u> y todos los demás generadores anulados.

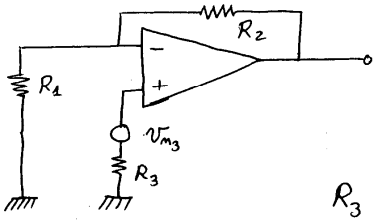


$$v_o = \underbrace{i_{m+} \cdot R_3 \cdot \left(1 + \frac{R_2}{R_1}\right)}$$

R_3 AYUDA A QUE i_{m+} genere ruido

MORALEJA: $R_3 = (R_1 // R_2)$ para compensar efectos de I_{BIAS} ⇒ **MAS RUIDO.**

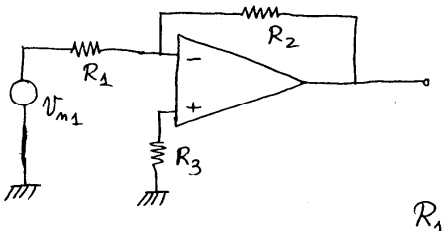
porque además, el ruido térmico que genera v_{n_3} también sale en v_o:



$$v_o = \underbrace{v_{n_3}\left(1 + \frac{R_2}{R_1}\right)}$$

R_3 genera ruido, que se amplifica.

Veamos ahora v_{n_1}:

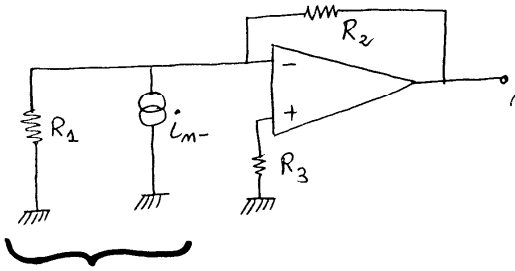


signo sobra en cálculos de ruido

$$v_o = \underbrace{\frac{-R_2}{R_1}\, v_{n_1}}$$

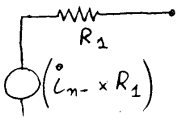
R_1 genera ruido, que se amplifica si $R_2 > R_1$.

Veamos ahora i_{n-}:



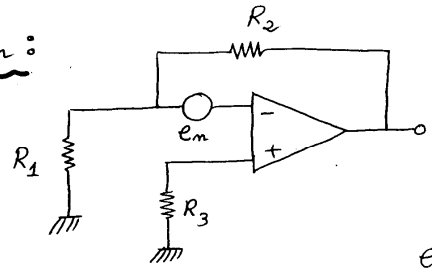
$$v_o = \frac{-R_2}{R_1} \cdot \left(i_{n-} \times R_1\right) = \underbrace{-R_2 \cdot i_{n-}}$$

R_2 convierte i_{n-} en ruido



Norton/Thèv o también: $v(+) = 0 = v(-)$ ⇒ i_{n-} debe circular toda por R_2 ⇒ $v = -R_2\, i_{n-}$

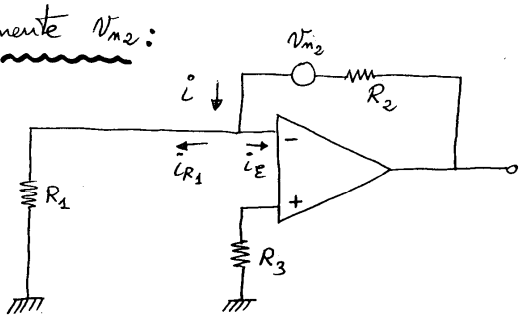
Efecto de e_n:



$$v_o = \underbrace{e_n\left(1+\frac{R_2}{R_1}\right)}$$

e_n se amplifica por la ganancia de ruido

Finalmente v_{n2}:

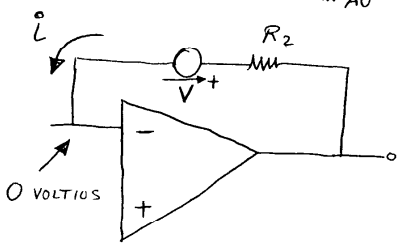


$$v_o = \underbrace{v_{n2}}$$

Ruido generado por R_2 directamente, sin amplificar ni atenuar.

Si $v(+) = 0$ y $v(-) = v(+) \Rightarrow v(-) = 0 \Rightarrow$

$i_{R_1} = 0$ y como $i_\varepsilon = \dfrac{v(-) - v(+)}{Z_{IN\,AO}} = \dfrac{0}{Z_{IN\,AO}} = 0 \Rightarrow i = 0$



$v_o = V$ para que $i = 0$

Por tanto, los **seis** generadores producen una salida de:

$$v_o = \left(1+\frac{R_2}{R_1}\right)\cdot\sqrt{e_n^2 + i_{n+}^2\cdot R_3^2 + \left[\frac{R_2\cdot R_1}{R_1+R_2}\, i_{n-}\right]^2 + \underbrace{v_{n1}^2\left(\frac{R_2}{R_1+R_2}\right)^2}_{a} + \underbrace{v_{n3}^2}_{b} + \underbrace{v_{n2}^2\left(\frac{R_1}{R_1+R_2}\right)^2}_{c}}$$

Ahora bien: $v_{n1}^2 = 4KT\,(BW)\cdot R_1$

$v_{n2}^2 = 4KT\,R_2$ ← 1 Hz (spot noise a cada frecuencia)

con lo que los términos debidos a v_{n_1}, v_{n_2} y v_{n_3} quedan:

$$a = 4KT\,\frac{R_1 R_2^2}{(R_1+R_2)^2} \qquad b = 4KT R_3 \quad y \quad c = 4KT\,\frac{R_2 R_1^2}{(R_1+R_2)^2}$$

Por tanto:

$$a+b+c = 4KT\left[R_3 + \frac{R_1\cdot R_2\cdot(R_2+R_1)}{(R_1+R_2)^2}\right] = 4KT\left[R_3 + \frac{R_1\cdot R_2}{R_1+R_2}\right]$$

$$= 4kT\left[R_3 + R_1 // R_2\right]$$

Si además suponemos $i_{n-} = i_{n+}$ (corrientes eficaces similares, nunca iguales pues no son la misma y son de ruido, no correladas...)

Tendremos: Por facilidad para recordar: EXACTO: $\left[R_3^2 + (R_1 // R_2)^2\right]$ OJO! $\left\{\text{Se usa } R_N^2 \text{ con } R_N = R_3 + R_1 // R_2\right.$

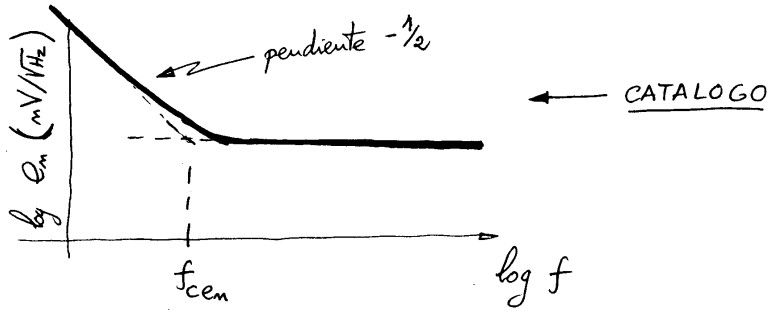
$$\boxed{v_o = \left(1+\frac{R_2}{R_1}\right)\sqrt{e_n^2 + i_n^2\left[R_3 + R_1//R_2\right]^{2?} + 4KT\left[R_3 + R_1//R_2\right]}} \quad (1)$$

Fórmula muy útil para hacer cálculos de ruido en etapas con A.O. a partir de las graficas de e_n e i_n del A.O. y que además deja bien claro que R_1, R_2 y R_3 deben ser valores bajos o al menos da un criterio para elegirlos.

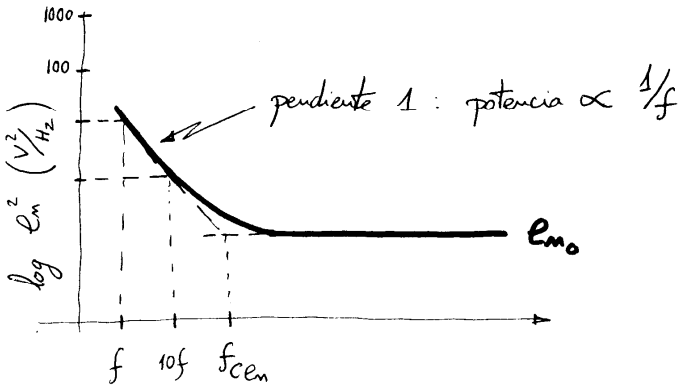
Tenemos tres tipos de contribución: la de e_n, la de i_n y una tercera que es plana con la frecuencia. Las otras dos presentan zona plana y zona $1/f$ con unas frecuencias de cado f_{ci_n} y f_{ce_n}.

Para el cálculo de ruido total a la salida hay que saber (o calcular) la frecuencia de corte superior del circuito y la frecuencia de corte inferior si la hay (puede haber acoplo en continua; en este caso f_{cin} y f_{cen} serán las mínimas que nos dé el fabricante o algo del orden del inverso del tiempo de toma de medidas).

Aportación debida a e_n:

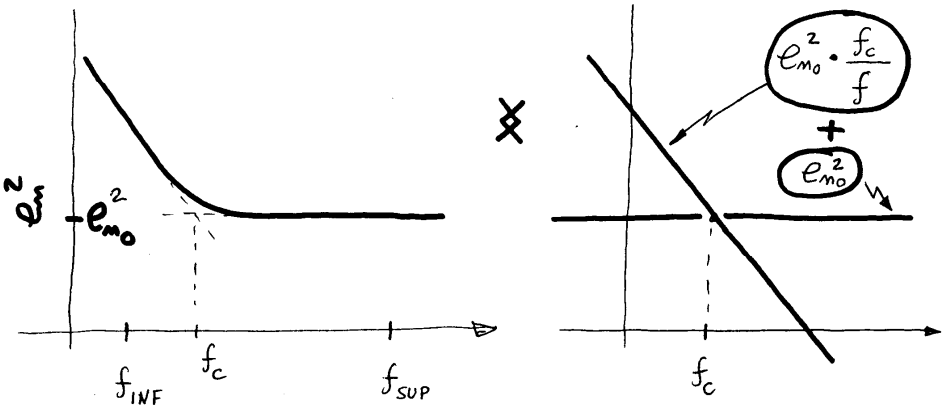


mejor "pensamos" en terminos de e_n^2 (potencia)



En la ecuación (1) aparecía e_n^2 dentro de la raíz y representaba la densidad de ruido a una frecuencia dada (spot noise).

Para calcular la aportación en un ancho de banda dado por f_{SUP} y f_{INF} (frec. de corte superior e inferior del sistema) tendremos:



$$V_o \text{ (Voltios eficaces)}\Big|_{\text{debidos a } e_n} = \left(1+\frac{R_2}{R_1}\right)\cdot\sqrt{\int_{f_{INF}}^{f_{SUP}}\left(e_{n_0}^2\cdot\frac{f_c}{f}+e_{n_0}^2\right)df}$$

$$= \left(1+\frac{R_2}{R_1}\right)\cdot e_{n_0}\sqrt{\left(f_{SUP}-f_{INF}\right)+f_c\,\ln\left(\frac{f_{SUP}}{f_{INF}}\right)}$$

De aquí que, salvo en el caso de f_c elevadas (A.O.'s con mucho ruido 1/f) el valor "exacto" de f_{INF} no sea muy importante

En audio: $f_{SUP} = 20.000\ Hz$,, $f_c = 50\,Hz$ (bastante alta, no es un "low-noise" A.O.)

si $f_{INF} = 20\,Hz \rightarrow \sqrt{19.980+345} = 142'6$

si $f_{INF} = 0'2\,Hz \rightarrow \sqrt{20000+610} = 143'6$

PMI

OP-27

LOW-NOISE PRECISION
OPERATIONAL AMPLIFIER

Precision Monolithics Inc.

FEATURES

- Low Noise ..... $80nV_{p-p}$ (0.1Hz to 10Hz)
  ..... $3nV/\sqrt{Hz}$
- Low Drift ..... $0.2\mu V/°C$
- High Speed ..... $2.8V/\mu s$ Slew Rate
  ..... 8MHz Gain Bandwidth
- Low V_{OS} ..... $10\mu V$
- Excellent CMRR ..... 126dB at V_{CM} of $\pm 11V$
- High Open-Loop Gain ..... 1.8 Million
- Fits 725, OP-07, OP-05, AD510, AD517, 5534A sockets
- Available in Die Form

ORDERING INFORMATION †

$T_A = +25°C$ V_{OS} MAX (mV)	PACKAGE TO-99	CERDIP 8-PIN	PLASTIC 8-PIN	LCC 20-CONTACT	OPERATING TEMPERATURE RANGE
25	OP27AJ*	OP27AZ*	–	–	MIL
25	OP27EJ	OP27EZ	OP27EP	–	IND/COM
60	OP27BJ*	OP27BZ*	–	OP27BR/883	MIL
60	OP27FJ	OP27FZ	OP27FP	–	IND/COM
100	OP27CJ	OP27CZ	–	–	MIL
100	OP27GJ	OP27GZ	OP27GP	–	XIND
100	–	–	OP27GS††	–	XIND

* For devices processed in total compliance to MIL-STD-883, add /883 after part number. Consult factory for 883 data sheet.

† Burn-in is available on commercial and industrial temperature range parts in CerDIP, plastic DIP, and TO-can packages. For ordering information, see 1990/91 Data Book, Section 2.

†† For availability and burn-in information on SO and PLCC packages, contact your local sales office.

GENERAL DESCRIPTION

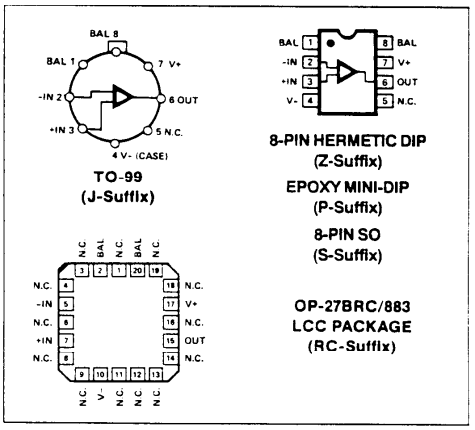
The OP-27 precision operational amplifier combines the low offset and drift of the OP-07 with both high-speed and low-noise. Offsets down to $25\mu V$ and drift of $0.6\mu V/°C$ maximum make the OP-27 ideal for precision instrumentation applications. Exceptionally low noise, $e_n = 3.5nV/\sqrt{Hz}$, at 10Hz, a low 1/f noise corner frequency of 2.7Hz, and high gain (1.8 million), allow accurate high-gain amplification of low-level signals. A gain-bandwidth product of 8MHz and a $2.8V/\mu sec$ slew rate provides excellent dynamic accuracy in high-speed data-acquisition systems.

A low input bias current of ±10nA is achieved by use of a bias-current-cancellation circuit. Over the military temperature range, this circuit typically holds I_B and I_{OS} to ±20nA and 15nA respectively.

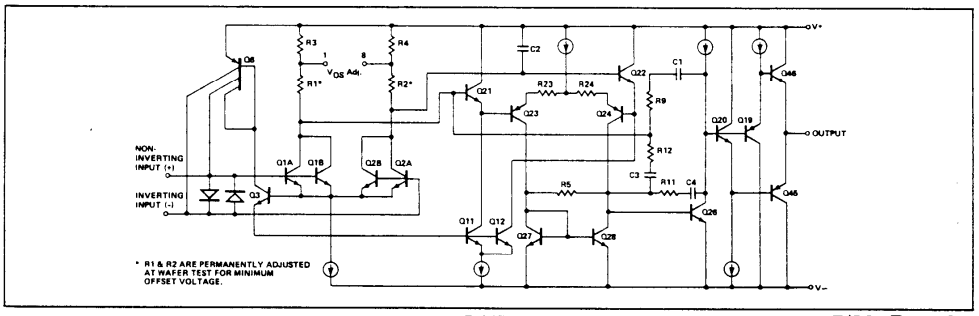
The output stage has good load driving capability. A guaranteed swing of ±10V into 600Ω and low output distortion make the OP-27 an excellent choice for professional audio applications.

PSRR and CMRR exceed 120dB. These characteristics, coupled with long-term drift of $0.2\mu V$/month, allow the circuit designer to achieve performance levels previously attained only by discrete designs.

PIN CONNECTIONS



SIMPLIFIED SCHEMATIC



Low cost, high-volume production of OP-27 is achieved by using an on-chip zener-zap trimming network. This reliable and stable offset trimming scheme has proved its effectiveness over many years of production history.

The OP-27 provides excellent performance in low-noise high-accuracy amplification of low-level signals. Applications include stable integrators, precision summing amplifiers, precision voltage-threshold detectors, comparators, and professional audio circuits such as tape-head and microphone preamplifiers.

The OP-27 is a direct replacement for 725, OP-06, OP-07 and OP-05 amplifiers; 741 types may be directly replaced by removing the 741's nulling potentiometer.

ABSOLUTE MAXIMUM RATINGS (Note 4)

Supply Voltage ..... ±22V
Input Voltage (Note 1) ..... ±22V
Output Short-Circuit Duration ..... Indefinite
Differential Input Voltage (Note 2) ..... ±0.7V
Differential Input Current (Note 2) ..... ±25mA
Storage Temperature Range ..... –65°C to +150°C
Operating Temperature Range
OP-27A, OP-27B, OP-27C (J, Z, RC)..... –55°C to +125°C
OP-27E, OP-27F (J, Z) ..... –25°C to +85°C
OP-27E, OP-27F (P) ..... 0°C to +70°C
OP-27G (P, S, J, Z) ..... –40°C to +85°C
Lead Temperature Range (Soldering, 60 sec) ..... 300°C
Junction Temperature ..... –65°C to +150°C

PACKAGE TYPE	Θ_{JA} (Note 3)	Θ_{JC}	UNITS
TO-99 (J)	150	18	°C/W
8-Pin Hermetic DIP (Z)	148	16	°C/W
8-Pin Plastic DIP (P)	103	43	°C/W
20-Contact LCC (RC)	98	38	°C/W
8-Pin SO (S)	158	43	°C/W

NOTES:
1. For supply voltages less than ±22V, the absolute maximum input voltage is equal to the supply voltage.
2. The OP-27's inputs are protected by back-to-back diodes. Current limiting resistors are not used in order to achieve low noise. If differential input voltage exceeds ±0.7V, the input current should be limited to 25mA.
3. Θ_{JA} is specified for worst case mounting conditions, i.e., Θ_{JA} is specified for device in socket for TO, CerDIP, P-DIP, and LCC packages; Θ_{JA} is specified for device soldered to printed circuit board for SO package.
4. Absolute maximum ratings apply to both DICE and packaged parts, unless otherwise noted.

ELECTRICAL CHARACTERISTICS at V_S = ±15V, T_A = 25°C, unless otherwise noted.

PARAMETER	SYMBOL	CONDITIONS	OP-27A/E MIN	OP-27A/E TYP	OP-27A/E MAX	OP-27B/F MIN	OP-27B/F TYP	OP-27B/F MAX	OP-27C/G MIN	OP-27C/G TYP	OP-27C/G MAX	UNITS
Input Offset Voltage	V_{OS}	Note 1	—	10	25	—	20	60	—	30	100	μV
Long-Term V_{OS} Stability	V_{OS}/Time	Notes 2, 3	—	0.2	1.0	—	0.3	1.5	—	0.4	2.0	μV/Mo
Input Offset Current	I_{OS}		—	7	35	—	9	50	—	12	75	nA
Input Bias Current	I_B		—	±10	±40	—	±12	±55	—	±15	±80	nA
Input Noise Voltage	e_{np-p}	0.1Hz to 10Hz Notes 3, 5	—	0.08	0.18	—	0.08	0.18	—	0.09	0.25	μVp-p
Input Noise Voltage Density	e_n	f_O = 10Hz (Note 3)	—	3.5	5.5	—	3.5	5.5	—	3.8	8.0	nV/√Hz
		f_O = 30Hz (Note 3)	—	3.1	4.5	—	3.1	4.5	—	3.3	5.6	
		f_O = 1000Hz (Note 3)	—	3.0	3.8	—	3.0	3.8	—	3.2	4.5	
Input Noise Current Density	i_n	f_O = 10Hz (Notes 3,6)	—	1.7	4.0	—	1.7	4.0	—	1.7	—	pA/√Hz
		f_O = 30Hz (Notes 3, 6)	—	1.0	2.3	—	1.0	2.3	—	1.0	—	
		f_O = 1000Hz (Notes 3, 6)	—	0.4	0.6	—	0.4	0.6	—	0.4	0.6	
Input Resistance — Differential-Mode	R_{IN}	Note 7	1.3	6	—	0.94	5	—	0.7	4	—	MΩ
Input Resistance — Common-Mode	R_{INCM}		—	3	—	—	2.5	—	—	2	—	GΩ
Input Voltage Range	IVR		±11.0	±12.3	—	±11.0	±12.3	—	±11.0	±12.3	—	V
Common-Mode Rejection Ratio	CMRR	V_{CM} = ±11V	114	126	—	106	123	—	100	120	—	dB
Power Supply Rejection Ratio	PSRR	V_S = ±4V to ±18V	—	1	10	—	1	10	—	2	20	μV/V
Large-Signal Voltage Gain	A_{VO}	R_L ≥ 2kΩ, V_O = ±10V	1000	1800	—	1000	1800	—	700	1500	—	V/mV
		R_L ≥ 600Ω, V_O = ±10V	800	1500	—	800	1500	—	600	1500	—	
Output Voltage Swing	V_O	R_L ≥ 2kΩ	±12.0	±13.8	—	±12.0	±13.8	—	±11.5	±13.5	—	V
		R_L ≥ 600Ω	±10.0	±11.5	—	±10.0	±11.5	—	±10.0	±11.5	—	
Slew Rate	SR	R_L ≥ 2kΩ (Note 4)	1.7	2.8	—	1.7	2.8	—	1.7	2.8	—	V/μs

ELECTRICAL CHARACTERISTICS at $V_S = \pm15V$, $T_A = 25°C$, unless otherwise noted. (Continued)

PARAMETER	SYMBOL	CONDITIONS	OP-27A/E MIN	OP-27A/E TYP	OP-27A/E MAX	OP-27B/F MIN	OP-27B/F TYP	OP-27B/F MAX	OP-27C/G MIN	OP-27C/G TYP	OP-27C/G MAX	UNITS
Gain Bandwidth Prod.	GBW	Note 4	5.0	8.0	—	5.0	8.0	—	5.0	8.0	—	MHz
Open-Loop Output Resistance	R_O	$V_O = 0$, $I_O = 0$	—	70	—	—	70	—	—	70	—	Ω
Power Consumption	P_d	V_O	—	90	140	—	90	140	—	100	170	mW
Offset Adjustment Range		$R_P = 10k\Omega$	—	±4.0	—	—	±4.0	—	—	±4.0	—	mV

NOTES:
1. Input offset voltage measurements are performed ~ 0.5 seconds after application of power. A/E grades guaranteed fully warmed-up.
2. Long-term input offset voltage stability refers to the average trend line of V_{OS} vs. Time over extended periods after the first 30 days of operation. Excluding the initial hour of operation, changes in V_{OS} during the first 30 days are typically 2.5μV — refer to typical performance curve.
3. Sample tested.
4. Guaranteed by design.
5. See test circuit and frequency response curve for 0.1Hz to 10Hz tester.
6. See test circuit for current noise measurement.
7. Guaranteed by input bias current.

ELECTRICAL CHARACTERISTICS for $V_S = \pm15V$, $-55°C \le T_A \le +125°C$, unless otherwise noted.

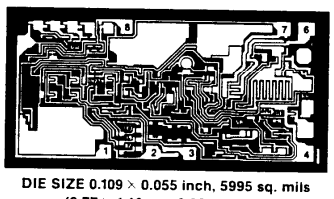
PARAMETER	SYMBOL	CONDITIONS	OP-27A MIN	OP-27A TYP	OP-27A MAX	OP-27B MIN	OP-27B TYP	OP-27B MAX	OP-27C MIN	OP-27C TYP	OP-27C MAX	UNITS
Input Offset Voltage	V_{OS}	Note 1	—	30	60	—	50	200	—	70	300	μV
Average Input Offset Drift	TCV_{OS} TCV_{OSn}	Note 2 Note 3	—	0.2	0.6	—	0.3	1.3	—	0.4	1.8	μV/°C
Input Offset Current	I_{OS}		—	15	50	—	22	85	—	30	135	nA
Input Bias Current	I_B		—	±20	±60	—	±28	±95	—	±35	±150	nA
Input Voltage Range	IVR		±10.3	±11.5	—	±10.3	±11.5	—	±10.2	±11.5	—	V
Common-Mode Rejection Ratio	CMRR	$V_{CM} = \pm10V$	108	122	—	100	119	—	94	116	—	dB
Power Supply Rejection Ratio	PSRR	$V_S = \pm4.5V$ to $\pm18V$	—	2	16	—	2	20	—	4	51	μV/V
Large-Signal Voltage Gain	A_{VO}	$R_L \ge 2k\Omega$, $V_O = \pm10V$	600	1200	—	500	1000	—	300	800	—	V/mV
Output Voltage Swing	V_O	$R_L \ge 2k\Omega$	±11.5	±13.5	—	±11.0	±13.2	—	±10.5	±13.0	—	V

ELECTRICAL CHARACTERISTICS at $V_S = \pm15V$, $-25°C \le T_A \le +85°C$ for OP-27J and OP-27Z, $0°C \le T_A \le +70°C$ for OP-27EP, FP and $-40°C \le T_A \le +85°C$ for OP-27GP, GS, unless otherwise noted.

PARAMETER	SYMBOL	CONDITIONS	OP-27E MIN	OP-27E TYP	OP-27E MAX	OP-27F MIN	OP-27F TYP	OP-27F MAX	OP-27G MIN	OP-27G TYP	OP-27G MAX	UNITS
Input Offset Voltage	V_{OS}		—	20	50	—	40	140	—	55	220	μV
Average Input Offset Drift	TCV_{OS} TCV_{OSn}	Note 2 Note 3	—	0.2	0.6	—	0.3	1.3	—	0.4	1.8	μV/°C
Input Offset Current	I_{OS}		—	10	50	—	14	85	—	20	135	nA
Input Bias Current	I_B		—	±14	±60	—	±18	±95	—	±25	±150	nA
Input Voltage Range	IVR		±10.5	±11.8	—	±10.5	±11.8	—	±10.5	±11.8	—	V
Common-Mode Rejection Ratio	CMRR	$V_{CM} = \pm10V$	110	124	—	102	121	—	96	118	—	dB
Power Supply Rejection Ratio	PSRR	$V_S = \pm4.5V$ to $\pm18V$	—	2	15	—	2	16	—	2	32	μV/V
Large-Signal Voltage Gain	A_{VO}	$R_L \ge 2k\Omega$, $V_O = \pm10V$	750	1500	—	700	1300	—	450	1000	—	V/mV
Output Voltage Swing	V_O	$R_L \ge 2k\Omega$	±11.7	±13.6	—	±11.4	±13.5	—	±11.0	±13.3	—	V

NOTES:
1. Input offset voltage measurements are performed by automated test equipment approximately 0.5 seconds after application of power. A/E grades guaranteed fully warmed-up.
2. The TCV_{OS} performance is within the specifications unnulled or when nulled with $R_P = 8k\Omega$ to $20k\Omega$. TCV_{OS} is 100% tested for A/E grades, sample tested for B/C/F/G grades.
3. Guaranteed by design.

DICE CHARACTERISTICS



DIE SIZE 0.109 × 0.055 inch, 5995 sq. mils
(2.77 × 1.40mm, 3.88 sq. mm)

1. NULL
2. (−) INPUT
3. (+) INPUT
4. V−
6. OUTPUT
7. V+
8. NULL

For additional DICE ordering information, refer to 1990/91 Data Book, Section 2.

WAFER TEST LIMITS at $V_S = \pm 15V$, $T_A = 25°C$ for OP-27N, OP-27G, and OP-27GR devices; $T_A = 125°C$ for OP-27NT and OP-27GT devices, unless otherwise noted.

PARAMETER	SYMBOL	CONDITIONS	OP-27NT LIMIT	OP-27N LIMIT	OP-27GT LIMIT	OP-27G LIMIT	OP-27GR LIMIT	UNITS
Input Offset Voltage	V_{OS}	(Note 1	60	35	200	60	100	μV MAX
Input Offset Current	I_{OS}		50	35	85	50	75	nA MAX
Input Bias Current	I_B		±60	±40	±95	±55	±80	nA MAX
Input Voltage Range	IVR		±10.3	±11	±10.3	±11	±11	V MIN
Common-Mode Rejection Ratio	CMRR	V_{CM} = IVR	108	114	100	106	100	dB MIN
Power Supply Rejection Ratio	PSRR	$V_S = \pm 4V$ to $\pm 18V$	—	10	—	10	20	μV/V MAX
Large-Signal Voltage Gain	A_{VO}	$R_L \geq 2k\Omega$, $V_O = \pm 10V$	600	1000	500	1000	700	V/mV MIN
		$R_L \geq 600\Omega$, $V_O = \pm 10V$	—	800	—	800	600	
Output Voltage Swing	V_O	$R_L \geq 2k\Omega$	±11.5	±12.0	±11.0	±12.0	±11.5	V MIN
		$R_L \geq 600\Omega$	—	±10.0	—	±10.0	±10.0	
Power Consumption	P_d	$V_O = 0$	—	140	—	140	170	mW MAX

NOTE:
Electrical tests are performed at wafer probe to the limits shown. Due to variations in assembly methods and normal yield loss, yield after packaging is not guaranteed for standard product dice. Consult factory to negotiate specifications based on dice lot qualification through sample lot assembly and testing.

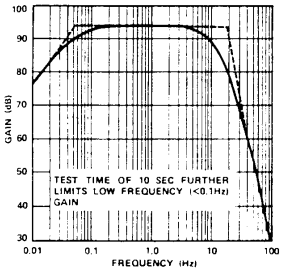
TYPICAL ELECTRICAL CHARACTERISTICS at $V_S = \pm 15V$, $T_A = +25°C$, unless otherwise noted.

PARAMETER	SYMBOL	CONDITIONS	OP-27N TYPICAL	OP-27G TYPICAL	OP-27GR TYPICAL	UNITS
Average Input Offset Voltage Drift	TCV_{OS} or TCV_{OSn}	Nulled or Unnulled R_P = 8kΩ to 20kΩ	0.2	0.3	0.4	μV/°C
Average Input Offset Current Drift	TCI_{OS}		80	130	180	pA/°C
Average Input Bias Current Drift	TCI_B		100	160	200	pA/°C
Input Noise Voltage Density	e_n	f_O = 10Hz	3.5	3.5	3.8	nV/√Hz
		f_O = 30Hz	3.1	3.1	3.3	
		f_O = 1000Hz	3.0	3.0	3.2	
Input Noise Current Density	i_n	f_O = 10Hz	1.7	1.7	1.7	pA/√Hz
		f_O = 30Hz	1.0	1.0	1.0	
		f_O = 1000Hz	0.4	0.4	0.4	
Input Noise Voltage	e_{np-p}	0.1Hz to 10Hz	0.08	0.08	0.09	μVp-p
Slew Rate	SR	$R_L \geq 2k\Omega$	2.8	2.8	2.8	V/μs
Gain Bandwidth Product	GBW		8	8	8	MHz

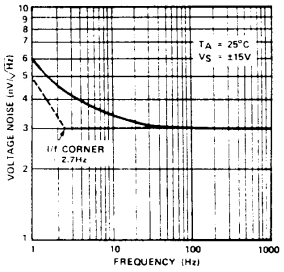
NOTE:
1. Input offset voltage measurements are performed by automated test equipment approximately 0.5 seconds after application of power.

TYPICAL PERFORMANCE CHARACTERISTICS

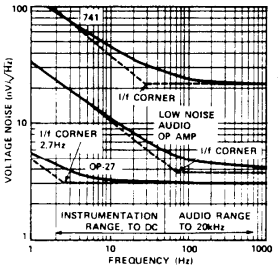
0.1Hz TO 10Hz$_{p-p}$ NOISE TESTER FREQUENCY RESPONSE



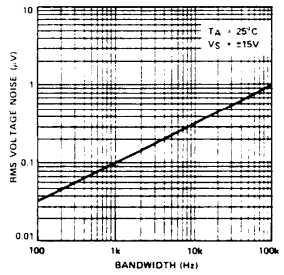
VOLTAGE NOISE DENSITY vs FREQUENCY



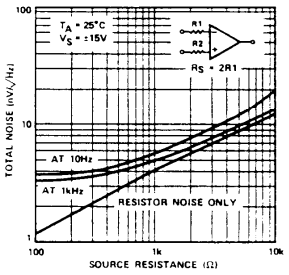
A COMPARISON OF OP AMP VOLTAGE NOISE SPECTRA



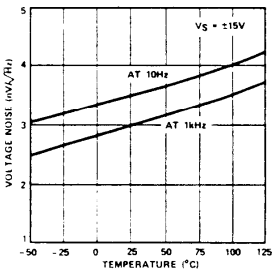
INPUT WIDEBAND VOLTAGE NOISE vs BANDWIDTH (0.1Hz TO FREQUENCY INDICATED)



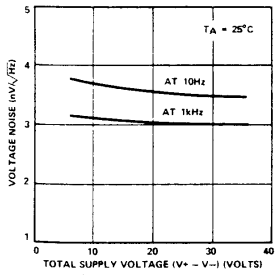
TOTAL NOISE vs SOURCE RESISTANCE



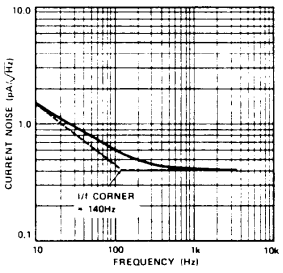
VOLTAGE NOISE DENSITY vs TEMPERATURE



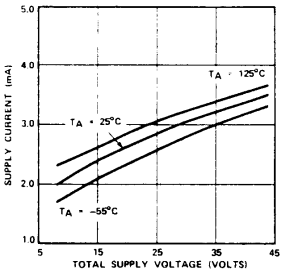
VOLTAGE NOISE DENSITY vs SUPPLY VOLTAGE



CURRENT NOISE DENSITY vs FREQUENCY

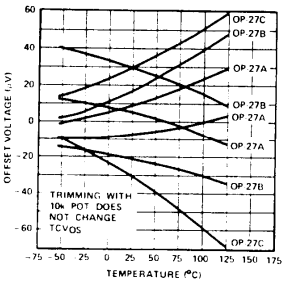


SUPPLY CURRENT vs SUPPLY VOLTAGE

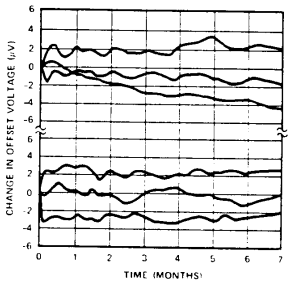


TYPICAL PERFORMANCE CHARACTERISTICS

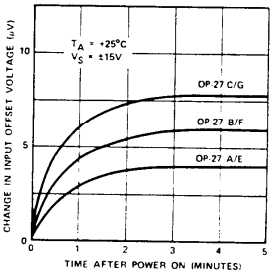
OFFSET VOLTAGE DRIFT OF EIGHT REPRESENTATIVE UNITS vs TEMPERATURE



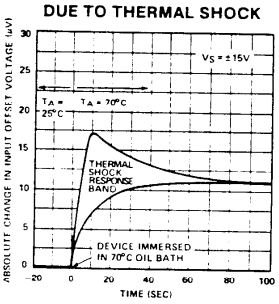
LONG-TERM OFFSET VOLTAGE DRIFT OF SIX REPRESENTATIVE UNITS



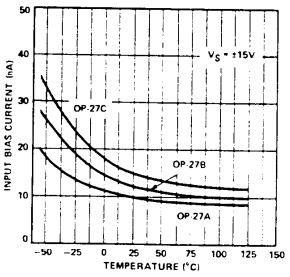
WARM-UP OFFSET VOLTAGE DRIFT



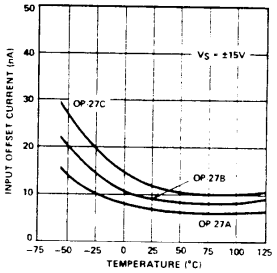
OFFSET VOLTAGE CHANGE DUE TO THERMAL SHOCK



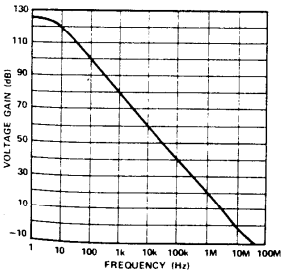
INPUT BIAS CURRENT vs TEMPERATURE



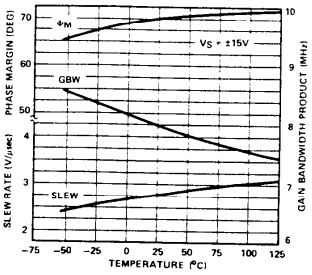
INPUT OFFSET CURRENT vs TEMPERATURE



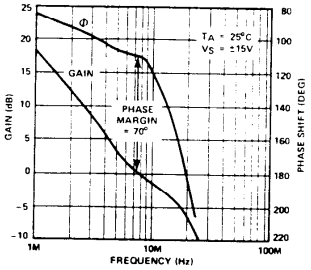
OPEN-LOOP GAIN vs FREQUENCY



SLEW RATE, GAIN-BANDWIDTH PRODUCT, PHASE MARGIN vs TEMPERATURE

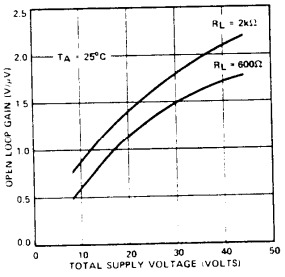


GAIN, PHASE SHIFT vs FREQUENCY

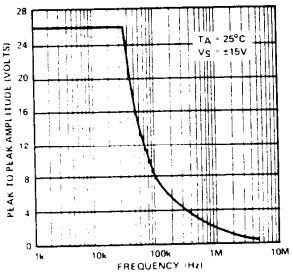


TYPICAL PERFORMANCE CHARACTERISTICS

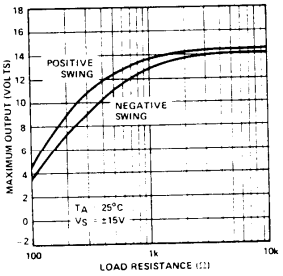
OPEN-LOOP VOLTAGE GAIN vs SUPPLY VOLTAGE



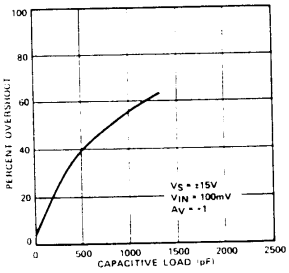
MAXIMUM OUTPUT SWING vs FREQUENCY



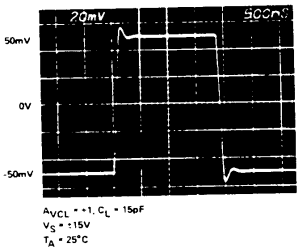
MAXIMUM OUTPUT VOLTAGE vs LOAD RESISTANCE



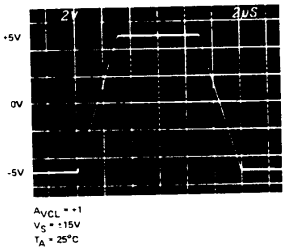
SMALL-SIGNAL OVERSHOOT vs CAPACITIVE LOAD



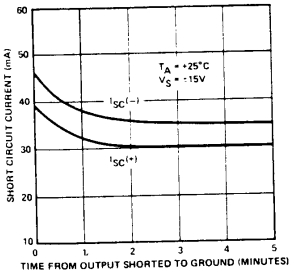
SMALL-SIGNAL TRANSIENT RESPONSE



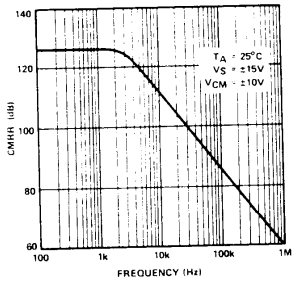
LARGE-SIGNAL TRANSIENT RESPONSE



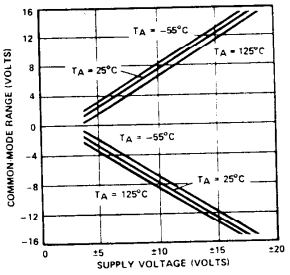
SHORT-CIRCUIT CURRENT vs TIME



CMRR vs FREQUENCY

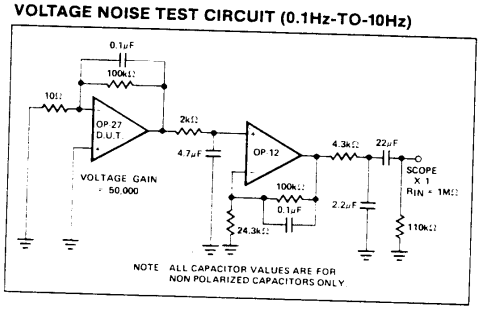


COMMON-MODE INPUT RANGE vs SUPPLY VOLTAGE

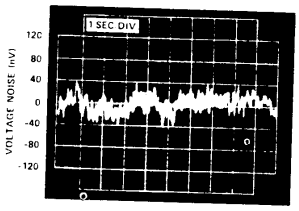


TYPICAL PERFORMANCE CHARACTERISTICS

VOLTAGE NOISE TEST CIRCUIT (0.1Hz-TO-10Hz)



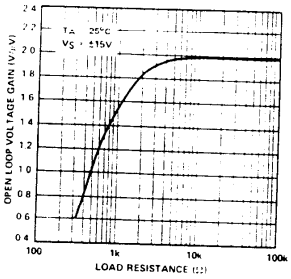
LOW-FREQUENCY NOISE



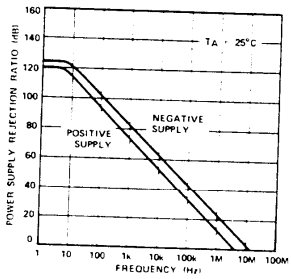
0.1Hz TO 10Hz PEAK-TO-PEAK NOISE

NOTE:
Observation time limited to 10 seconds.

OPEN-LOOP VOLTAGE GAIN vs LOAD RESISTANCE



PSRR vs FREQUENCY



APPLICATIONS INFORMATION

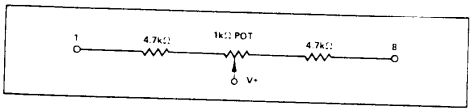
OP-27 Series units may be inserted directly into 725, OP-06, OP-07 and OP-05 sockets with or without removal of external compensation or nulling components. Additionally, the OP-27 may be fitted to unnulled 741-type sockets; however, if conventional 741 nulling circuitry is in use, it should be modified or removed to ensure correct OP-27 operation. OP-27 offset voltage may be nulled to zero (or other desired setting) using a potentiometer (see Offset Nulling Circuit).

The OP-27 provides stable operation with load capacitances of up to 2000pF and ±10V swings; larger capacitances should be decoupled with a 50Ω resistor inside the feedback loop. The OP-27 is unity-gain stable.

Thermoelectric voltages generated by dissimilar metals at the input terminal contacts can degrade the drift performance. Best operation will be obtained when both input contacts are maintained at the same temperature.

OFFSET VOLTAGE ADJUSTMENT

The input offset voltage of the OP-27 is trimmed at wafer level. However, if further adjustment of V_{OS} is necessary, a 10kΩ trim potentiometer may be used. TCV_{OS} is not degraded (see Offset Nulling Circuit). Other potentiometer values from 1kΩ to 1MΩ can be used with a slight degradation (0.1 to 0.2µV/°C) of TCV_{OS}. Trimming to a value other than zero creates a drift of approximately (V_{OS}/300) µV/°C. For example, the change in TCV_{OS} will be 0.33µV/°C if V_{OS} is adjusted to 100µV. The offset-voltage adjustment range with a 10kΩ potentiometer is ±4mV. If smaller adjustment range is required, the nulling sensitivity can be reduced by using a smaller pot in conjuction with fixed resistors. For example, the network below will have a ±280µV adjustment range.



NOISE MEASUREMENTS

To measure the 80nV peak-to-peak noise specification of the OP-27 in the 0.1Hz to 10Hz range, the following precautions must be observed:

(1) The device has to be warmed-up for at least five minutes. As shown in the warm-up drift curve, the offset voltage

typically changes 4μV due to increasing chip temperature after power-up. In the 10-second measurement interval, these temperature-induced effects can exceed tens-of-nanovolts.

(2) For similar reasons, the device has to be well-shielded from air currents. Shielding minimizes thermocouple effects.

(3) Sudden motion in the vicinity of the device can also "feed-through" to increase the observed noise.

(4) The test time to measure 0.1Hz-to-10Hz noise should not exceed 10 seconds. As shown in the noise-tester frequency-response curve, the 0.1Hz corner is defined by only one zero. The test time of 10 seconds acts as an additional zero to eliminate noise contributions from the frequency band below 0.1Hz.

(5) A noise-voltage-density test is recommended when measuring noise on a large number of units. A 10Hz noise-voltage-density measurement will correlate well with a 0.1Hz-to-10Hz peak-to-peak noise reading, since both results are determined by the white noise and the location of the 1/f corner frequency.

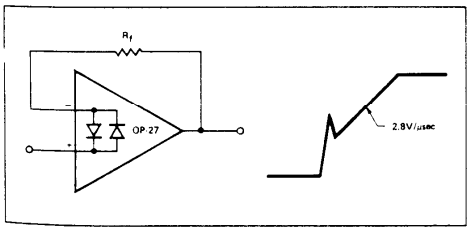
UNITY-GAIN BUFFER APPLICATIONS

When $R_f \leq 100\Omega$ and the input is driven with a fast, large signal pulse (>1V), the output waveform will look as shown in the pulsed operation diagram below.

During the fast feedthrough-like portion of the output, the input protection diodes effectively short the output to the input and a current, limited only by the output short-circuit protection, will be drawn by the signal generator. With $R_f \geq 500\Omega$, the output is capable of handling the current requirements ($I_L \leq 20mA$ at 10V); the amplifier will stay in its active mode and a smooth transition will occur.

When $R_f > 2k\Omega$, a pole will be created with R_f and the amplifier's input capacitance (8pF) that creates additional phase shift and reduces phase margin. A small capacitor (20 to 50pF) in parallel with R_f will eliminate this problem.

PULSED OPERATION



COMMENTS ON NOISE

The OP-27 is a very low-noise monolithic op amp. The outstanding input voltage noise characteristics of the OP-27 are achieved mainly by operating the input stage at a high quiescent current. The input bias and offset currents, which would normally increase, are held to reasonable values by the input-bias-current cancellation circuit. The OP-27A/E has I_B and I_{OS} of only ±40nA and 35nA respectively at 25°C. This is particularly important when the input has a high source-resistance. In addition, many audio amplifier designers prefer to use direct coupling. The high I_B, V_{OS}, TCV_{OS} of previous designs have made direct coupling difficult, if not impossible, to use.

Voltage noise is inversely proportional to the square-root of bias current, but current noise is proportional to the square-root of bias current. The OP-27's noise advantage disappears when high source-resistors are used. Figures 1, 2, and 3 compare OP-27 observed total noise with the noise performance of other devices in different circuit applications.

Total noise = $[(\text{Voltage noise})^2 + (\text{current noise} \times R_S)^2 + (\text{resistor noise})^2]^{1/2}$

Figure 1 shows noise-versus-source-resistance at 1000Hz. The same plot applies to wideband noise. To use this plot, just multiply the vertical scale by the square-root of the bandwidth.

NOISE vs SOURCE RESISTANCE (INCLUDING RESISTOR NOISE) AT 1000Hz.

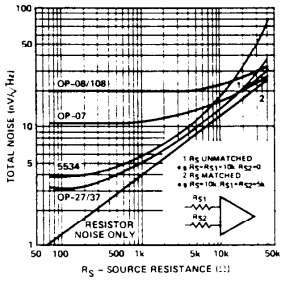


Figure 1

At $R_S < 1k\Omega$, the OP-27's low voltage noise is maintained. With $R_S > 1k\Omega$, total noise increases, but is dominated by the resistor noise rather than current or voltage noise. It is only beyond R_S of 20kΩ that current noise starts to dominate. The argument can be made that current noise is not important for applications with low-to-moderate source resistances. The crossover between the OP-27 and OP-07 and OP-08 noise occurs in the 15-to-40kΩ region.

Figure 2 shows the 0.1Hz-to-10Hz peak-to-peak noise. Here the picture is less favorable; resistor noise is negligible, current noise becomes important because it is inversely proportional to the square-root of frequency. The crossover with the OP-07 occurs in the 3-to-5kΩ range depending on whether balanced or unbalanced source resistors are used (at 3kΩ the I_B, I_{OS} error also can be three times the V_{OS} spec.).

PEAK-TO-PEAK NOISE (0.1 to 10Hz) vs SOURCE RESISTANCE (INCLUDES RESISTOR NOISE).

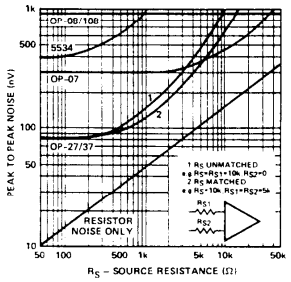


Figure 2

10Hz NOISE vs SOURCE RESISTANCE (INCLUDES RESISTOR NOISE).

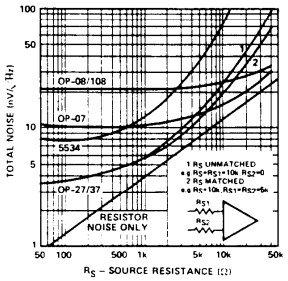


Figure 3

Therefore, for low-frequency applications, the OP-07 is better than the OP-27/37 when $R_S > 3k\Omega$. The only exception is when gain error is important. Figure 3 illustrates the 10Hz noise. As expected, the results are between the previous two figures.

For reference, typical source resistances of some signal sources are listed in Table 1.

Table 1

DEVICE	SOURCE IMPEDANCE	COMMENTS
Strain gauge	<500Ω	Typically used in low-frequency applications.
Magnetic tapehead	<1500Ω	Low i_B very important to reduce self-magnetization problems when direct coupling is used. OP-27 i_B can be neglected.
Magnetic phonograph cartridges	<1500Ω	Similar need for low i_B in direct coupled applications. OP-27 will not introduce any self-magnetization problem.
Linear variable differential transformer	<1500Ω	Used in rugged servo-feedback applications. Bandwidth of interest is 400Hz to 5kHz.

OPEN-LOOP GAIN

FREQUENCY AT:	OP-07	OP-27	OP-37
3Hz	100dB	124dB	125dB
10Hz	100dB	120dB	125dB
30Hz	90dB	110dB	124dB

For further information regarding noise calculations, see "Minimization of Noise in Op-Amp Applications", Application Note AN-15.

AUDIO APPLICATIONS

The following applications information has been abstracted from a PMI article in the 12/20/80 issue of Electronic Design magazine and updated.

Figure 4 is an example of a phono pre-amplifier circuit using the OP-27 for A_1; R_1-R_2-C_1-C_2 form a very accurate RIAA network with standard component values. The popular method to accomplish RIAA phono equalization is to employ frequency-dependent feedback around a high-quality gain block. Properly chosen, an RC network can provide the three necessary time constants of 3180, 318, and 75μs.[1]

For initial equalization accuracy and stability, precision metal-film resistors and film capacitors of polystyrene or polypropylene are recommended since they have low voltage coefficients, dissipation factors, and dielectric absorption.[4] (High-K ceramic capacitors should be avoided here, though low-K ceramics—such as NPO types, which have excellent dissipation factors, and somewhat lower dielectric absorption—can be considered for small values.)

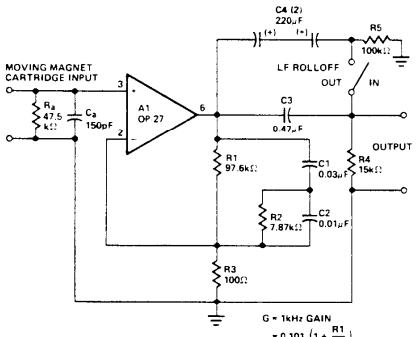


Figure 4

The OP-27 brings a 3.2nV/$\sqrt{Hz}$ voltage noise and 0.45 pA/$\sqrt{Hz}$ current noise to this circuit. To minimize noise from other sources, R_3 is set to a value of 100Ω, which generates a voltage noise of 1.3nV/$\sqrt{Hz}$. The noise increases the 3.2nV/$\sqrt{Hz}$ of the amplifier by only 0.7dB. With a 1kΩ source, the circuit noise measures 63dB below a 1mV reference level, unweighted, in a 20kHz noise bandwidth.

Gain (G) of the circuit at 1kHz can be calculated by the expression:

$$G = 0.101\left(1 + \frac{R_1}{R_3}\right)$$

For the values shown, the gain is just under 100 (or 40dB). Lower gains can be accommodated by increasing R_3, but gains higher than 40dB will show more equalization errors because of the 8MHz gain-bandwidth of the OP-27.

This circuit is capable of very low distortion over its entire range, generally below 0.01% at levels up to 7V rms. At 3V output levels, it will produce less than 0.03% total harmonic distortion at frequencies up to 20kHz.

Capacitor C_3 and resistor R_4 form a simple −6dB-per-octave rumble filter, with a corner at 22Hz. As an option, the switch-selected shunt capacitor C_4, a nonpolarized electrolytic, bypasses the low-frequency rolloff. Placing the rumble filter's high-pass action after the preamp has the desirable result of discriminating against the RIAA-amplified low-frequency noise components and pickup-produced low-frequency disturbances.

A preamplifier for NAB tape playback is similar to an RIAA phono preamp, though more gain is typically demanded, along with equalization requiring a heavy low-frequency boost. The circuit in Fig. 4 can be readily modified for tape use, as shown by Fig. 5.

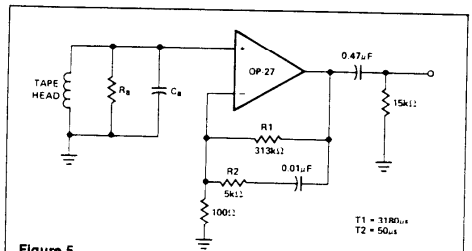


Figure 5

While the tape-equalization requirement has a flat high-frequency gain above 3kHz ($T_2 = 50\mu s$), the amplifier need not be stabilized for unity gain. The decompensated OP-37 provides a greater bandwidth and slew rate. For many applications, the idealized time constants shown may require trimming of R_1 and R_2 to optimize frequency response for nonideal tape-head performance and other factors.[5]

The network values of the configuration yield a 50dB gain at 1kHz, and the dc gain is greater than 70dB. Thus, the worst-case output offset is just over 500mV. A single 0.47μF output capacitor can block this level without affecting the dynamic range.

The tape head can be coupled directly to the amplifier input, since the worst-case bias current of 80nA with a 400mH, 100 μin. head (such as the PRB2H7K) will not be troublesome.

One potential tape-head problem is presented by amplifier bias-current transients which can magnetize a head. The OP-27 and OP-37 are free of bias-current transients upon power up or power down. However, it is always advantageous to control the speed of power supply rise and fall, to eliminate transients.

In addition, the dc resistance of the head should be carefully controlled, and preferably below 1kΩ. For this configuration, the bias-current-induced offset voltage can be greater than the 100μV maximum offset if the head resistance is not sufficiently controlled.

A simple, but effective, fixed-gain transformerless microphone preamp (Fig. 6) amplifies differential signals from low-impedance microphones by 50dB, and has an input impedance of 2kΩ. Because of the high working gain of the circuit, an OP-37 helps to preserve bandwidth, which will be 110kHz. As the OP-37 is a decompensated device (minimum stable gain of 5), a dummy resistor, R_p, may be necessary, if the microphone is to be unplugged. Otherwise the 100% feedback from the open input may cause the amplifier to oscillate.

Common-mode input-noise rejection will depend upon the match of the bridge-resistor ratios. Either close-tolerance (0.1%) types should be used, or R_4 should be trimmed for best CMRR. All resistors should be metal-film types for best stability and low noise.

Noise performance of this circuit is limited more by the input resistors R_1 and R_2 than by the op amp, as R_1 and R_2 each generate a 4nV/$\sqrt{Hz}$ noise, while the op amp generates a 3.2nV/$\sqrt{Hz}$ noise. The rms sum of these predominant noise sources will be about 6nV/$\sqrt{Hz}$, equivalent to 0.9μV in a 20kHz noise bandwidth, or nearly 61dB below a 1mV input signal. Measurements confirm this predicted performance.

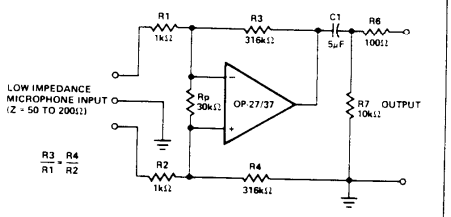


Figure 6

For applications demanding appreciably lower noise, a high-quality microphone-transformer-coupled preamp (Fig. 7) incorporates the internally-compensated OP-27. T_1 is a JE-115K-E 150Ω/15kΩ transformer which provides an optimum source resistance for the OP-27 device. The circuit has an overall gain of 40dB, the product of the transformer's voltage setup and the op amp's voltage gain.

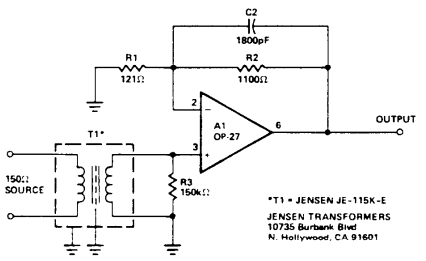


Figure 7

Gain may be trimmed to other levels, if desired, by adjusting R_2 or R_1. Because of the low offset voltage of the OP-27, the output offset of this circuit will be very low, 1.7mV or less, for a 40dB gain. The typical output blocking capacitor can be eliminated in such cases, but is desirable for higher gains to eliminate switching transients.

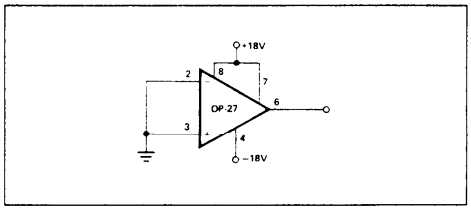
Capacitor C_2 and resistor R_2 form a 2μs time constant in this circuit, as recommended for optimum transient response by the transformer manufacturer. With C_2 in use, A_1 must have unity-gain stability. For situations where the 2μs time constant is not necessary, C_2 can be deleted, allowing the faster OP-37 to be employed.

Some comment on noise is appropriate to understand the capability of this circuit. A 150Ω resistor and R_1 and R_2 gain resistors connected to a noiseless amplifier will generate 220 nV of noise in a 20kHz bandwidth, or 73dB below a 1mV reference level. Any practical amplifier can only approach this noise level; it can never exceed it. With the OP-27 and T_1 specified, the additional noise degradation will be close to 3.6dB (or −69.5 referenced to 1mV).

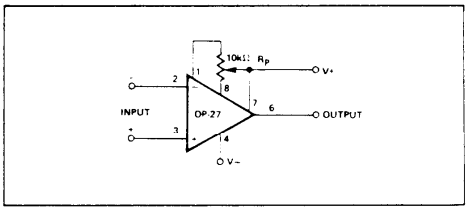
References

1. Lipshitz, S.P., "On RIAA Equalization Networks," *JAES*, Vol. 27, June 1979, p. 458-481.
2. Jung, W.G., *IC Op Amp Cookbook*, 2nd Ed., H.W. Sams and Company, 1980.
3. Jung, W.G., *Audio IC Op Amp Applications*, 2nd Ed., H.W. Sams and Company, 1978.
4. Jung, W.G., and Marsh, R.M., "Picking Capacitors," *Audio*, February & March, 1980.
5. Otala, M., "Feedback-Generated Phase Nonlinearity in Audio Amplifiers," London AES Convention, March 1980, preprint 1976.
6. Stout, D.F., and Kaufman, M., *Handbook of Operational Amplifier Circuit Design*, New York, McGraw Hill, 1976.

BURN-IN CIRCUIT



OFFSET NULLING CIRCUIT



FEATURES

- Low V_{OS} ..... 25μV Max
- Low V_{OS} Drift ..... 0.6μV/°C Max
- Ultra-Stable vs Time ..... 1.0μV/Month Max
- Low Noise ..... 0.6μV_{p-p} Max
- Wide Input Voltage Range ..... ±14V
- Wide Supply Voltage Range ..... ±3V to ±18V
- Fits 725, 108A/308A, 741, AD510 Sockets
- 125°C Temperature-Tested Dice

ORDERING INFORMATION †

T_A = +25°C V_{OS} MAX (μV)	PACKAGE TO-99	CERDIP 8-PIN	PLASTIC 8-PIN	LCC 20-CONTACT	OPERATING TEMPERATURE RANGE
25	OP07AJ*	OP07AZ*	—	—	MIL
75	OP07EJ	OP07EZ	OP07EP	—	COM
75	OP07J*	OP07Z*	—	OP07RC/883	MIL
150	OP07CJ	OP07CZ	OP07CP	—	XIND
150	—	—	OP07CS††	—	XIND
150	OP07DJ	—	OP07DP	—	XIND

* For devices processed in total compliance to MIL-STQ-883, add /883 after part number. Consult factory for 883 data sheet.

† Burn-in is available on commercial and industrial temperature range parts in CerDIP, plastic DIP, and TO-can packages. For ordering information, see PMI's Data Book, Section 2.

†† For availability and burn-in information on SO and PLCC packages, contact your local sales office.

GENERAL DESCRIPTION

The OP-07 has very low input offset voltage (25μV max for OP-07A) which is obtained by trimming at the wafer stage. These low offset voltages generally eliminate any need for external nulling. The OP-07 also features low input bias current (±2nA for OP-07A) and high open-loop gain (300V/mV for OP-07A). The low offsets and high open-loop gain make the OP-07 particularly useful for high-gain instrumentation applications.

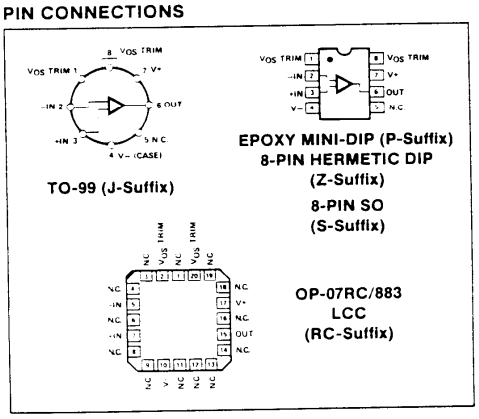
The wide input voltage range of ±13V minimum combined with high CMRR of 110dB (OP-07A) and high input impedace provides high accuracy in the noninverting circuit configuration. Excellent linearity and gain accuracy can be maintained even at high closed-loop gains.

Stability of offsets and gain with time or variations in temperture is excellent. The accuracy and stability of the OP-07, even at high gain, combined with the freedom from external nulling have made the OP-07 a new industry standard for instrumentation and military applications.

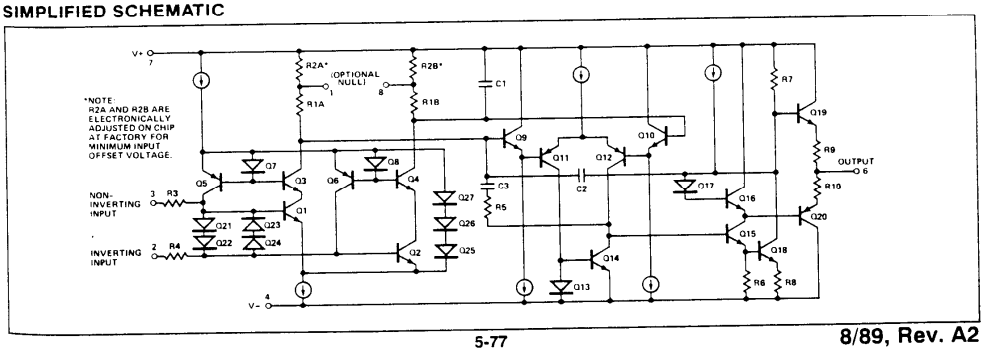
The OP-07 is available in five standard performance grades. The OP-07A and the OP-07 are specified for operation over the full military range of −55°C to +125°C; the OP-07E is specified for operation over the 0°C to +70°C range, and OP-07C and D over the −40°C to +85°C temperature range.

The OP-07 is available in hermetically-sealed TO-99 metal can or ceramic 8-pin Mini-DIP, and in epoxy 8-pin Mini-DIP. It is a direct replacement for 725, 108A, and OP-05 amplifiers; 741-types may be directly replaced by removing the 741's nulling potentiometer. The OP-207, a dual OP-07, is available for applications requiring close matching of two OP-07 amplifiers. For improved specifications, see the OP-77/OP-177.

PIN CONNECTIONS

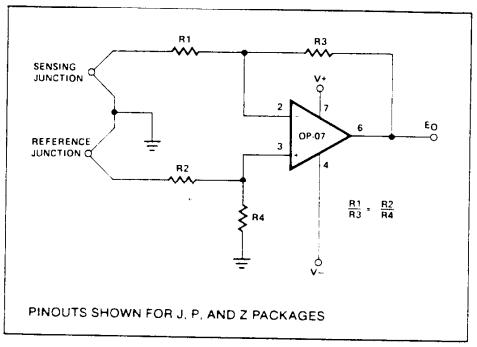


SIMPLIFIED SCHEMATIC



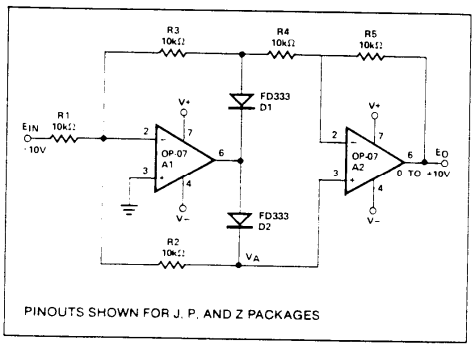
TYPICAL APPLICATIONS

HIGH-STABILITY THERMOCOUPLE AMPLIFIER



PINOUTS SHOWN FOR J, P, AND Z PACKAGES

PRECISION ABSOLUTE-VALUE CIRCUIT



PINOUTS SHOWN FOR J, P, AND Z PACKAGES

APPLICATIONS INFORMATION

OP-07 series units may be substituted directly into 725, 108A/308A* and OP-05 sockets with or without removal of external compensation or nulling components. Additionally, the OP-07 may be used in unnulled 741-type sockets. However, if conventional 741 nulling circuitry is in use, it should be modified or removed to enable proper OP-07 operation. OP-07 offset voltage may be nulled to zero through use of a potentiometer (see offset nulling circuit diagram).

The OP-07 provides stable operation with load capacitance of up to 500pF and ±10V swings; larger capacitances should be decoupled with a 50Ω decoupling resistor.

Stray thermoelectric voltages generated by dissimilar metals at the contacts to the input terminals can degrade drift performance. Therefore, best operation will be obtained when both input contacts are maintained at the same temperature, preferably close to the package temperature.

*TO-99 Package only

FEATURES

- Low Noise ..... 80nV p-p (0.1Hz to 10Hz)
  ..... 3nV/$\sqrt{Hz}$ at 1kHz
- Low Drift ..... 0.2μV/°C
- High Speed ..... 17V/μs Slew Rate
  ..... 63MHz Gain Bandwidth
- Low Input Offset Voltage 10μV
- Excellent CMRR ... 126dB (Common-Voltage of ±11V)
- High Open-Loop Gain ..... 1.8 Million
- Replaces 725, OP-05, OP-06, OP-07, AD510, AD517, SE5534 in Gains >5
- Available in Die Form

ORDERING INFORMATION †

T_A = +25°C V_{OS} MAX (mV)	PACKAGE TO-99	CERDIP 8-PIN	PLASTIC 8-PIN	LCC 20-CONTACT	OPERATING TEMPERATURE RANGE
25	OP37AJ*	OP37AZ*	–	–	MIL
25	OP37EJ	OP37EZ	OP37EP	–	IND/COM
60	OP37BJ*	OP37BZ*	–	OP37BRC/883	MIL
60	OP37FJ	OP37FZ	OP37FP	–	IND/COM
100	OP37CJ*	OP37CZ	–	–	MIL
100	OP37GJ	OP37GZ	OP37GP	–	XIND
100	–	–	OP37GS††	–	XIND

* For devices processed in total compliance to MIL-STD-883, add /883 after part number. Consult factory for 883 data sheet.

† Burn-in is available on commercial and industrial temperature range parts in CerDIP, plastic DIP, and TO-can packages. For ordering information, see 1990/91 Data Book, Section 2.

†† For availability and burn-in information on SO package, contact your local sales office.

GENERAL DESCRIPTION

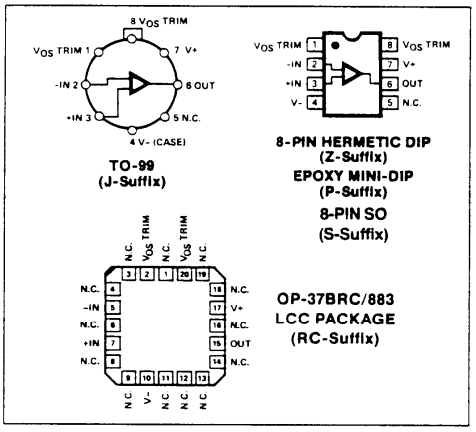
The OP-37 provides the same high performance as the OP-27, but the design is optimized for circuits with gains greater than five. This design change increases slew rate to 17V/μsec and gain-bandwidth product to 63MHz.

The OP-37 provides the low offset and drift of the OP-07 plus higher speed and lower noise. Offsets down to 25μV and drift of 0.6μV/°C maximum make the OP-37 ideal for precision instrumentation applications. Exceptionally low noise (e_n=3.5nV/$\sqrt{Hz}$ at 10Hz), a low 1/f noise corner frequency of 2.7Hz, and the high gain of 1.8 million, allow accurate high-gain amplification of low-level signals.

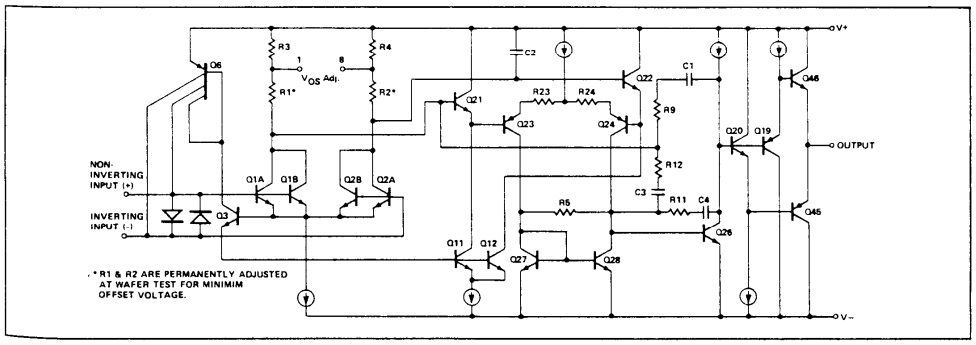
The low input bias current of ±10nA and offset current of 7nA are achieved by using a bias-current-cancellation circuit. Over the military temperature range this typically holds I_B and I_{OS} to ±20nA and 15nA respectively.

The output stage has good load driving capability. A guaranteed swing of ±10V into 600Ω and low output distortion make the OP-37 an excellent choice for professional audio applications.

PIN CONNECTIONS



SIMPLIFIED SCHEMATIC



LINEAR TECHNOLOGY

LT1028/LT1128

Ultra Low Noise Precision High Speed Op Amps

FEATURES

- Voltage Noise
  - 1.1nV/√Hz Max. at 1kHz
  - 0.85nV/√Hz Typ. at 1kHz
  - 1.0nV/√Hz Typ. at 10Hz
  - 35nV$_{P-P}$ Typ., 0.1Hz to 10Hz
- Voltage and Current Noise 100% Tested
- Gain-Bandwidth Product
  - LT1028: 50MHz Min.
  - LT1128: 13MHz Min.
- Slew Rate
  - LT1028: 11V/µs Min.
  - LT1128: 5V/µs Min.
- Offset Voltage: 40µV Max.
- Drift with Temperature: 0.8µV/°C Max.
- Voltage Gain: 7 Million Min.
- Available in 8-Pin SO Package

APPLICATIONS

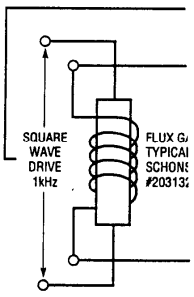
- Low Noise Frequency Synthesizers
- High Quality Audio
- Infrared Detectors
- Accelerometer and Gyro Amplifiers
- 350Ω Bridge Signal Conditioning
- Magnetic Search Coil Amplifiers
- Hydrophone Amplfiers

DESCRIPTION

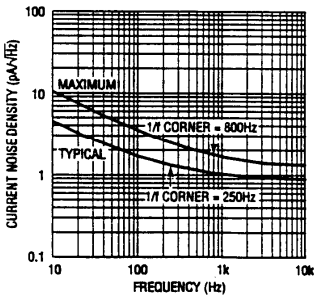
The LT1028(gain of –1 stable)/LT1128(gain of +1 stable) achieve a new standard of excellence in noise performance with 0.85nV/√Hz 1kHz noise, 1.0nV/√Hz 10Hz noise. This ultra low noise is combined with excellent high speed specifications (gain-bandwidth product is 75MHz for LT1028, 20MHz for LT1128), distortion-free output, and true precision parameters (0.1µV/°C drift, 10µV offset voltage, 30 million voltage gain). Although the LT1028/LT1128 input stage operates at nearly 1mA of collector current to achieve low voltage noise, input bias current is only 25nA.

The LT1028/LT1128's voltage noise is less than the noise of a 50Ω resistor. Therefore, even in very low source impedance transducer or audio amplifier applications, the LT1028/LT1128's contribution to total system noise will be negligible.

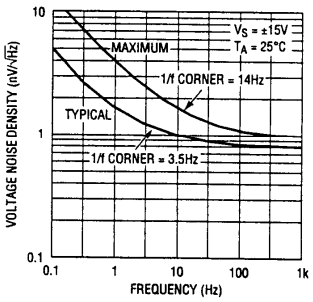
Flux Ga



Current Noise Spectrum



Voltage Noise vs Frequency



LINEAR TECHNOLOGY

LT1169

Dual Low Noise, Picoampere Bias Current, JFET Input Op Amp

FEATURES

- **Input Bias Current, Warmed Up: 20pA Max**
- **100% Tested Low Voltage Noise: 8nV/√Hz Max**
- S8 and N8 Package Standard Pinout
- Very Low Input Capacitance: 1.5pF
- Voltage Gain: 1.2 Million Min
- Offset Voltage: 2mV Max
- Input Resistance: $10^{13}\Omega$
- Gain-Bandwidth Product: 5.3MHz Typ
- Guaranteed Specifications with ±5V Supplies
- Guaranteed Matching Specifications

APPLICATIONS

- Photocurrent Amplifiers
- Hydrophone Amplifiers
- High Sensitivity Piezoelectric Accelerometers
- Low Voltage and Current Noise Instrumentation Amplifier Front Ends
- Two and Three Op Amp Instrumentation Amplifiers
- Active Filters

LT, LTC and LT are registered trademarks of Linear Technology Corporation.

DESCRIPTION

The LT1169 achieves a new standard of excellence in noise performance for a dual JFET op amp. For the first time low voltage noise (6nV/√Hz) is simultaneously offered with extremely low current noise (1fA/√Hz), providing the lowest total noise for high impedance transducer applications. Unlike most JFET op amps, the very low input bias current (5pA Typ) is maintained over the entire common mode range which results in an extremely high input resistance ($10^{13}\Omega$). When combined with a very low input capacitance (1.5pF) an extremely high input impedance results, making the LT1169 the first choice for amplifying low level signals from high impedance transducers. The low input capacitance also assures high gain linearity when buffering AC signals from high impedance transducers.

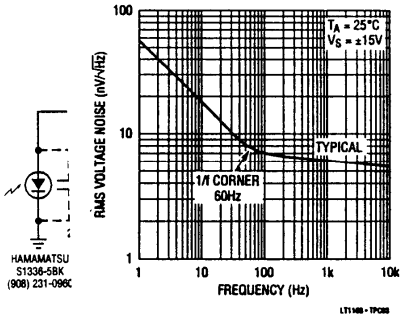
The LT1169 is unconditionally stable for gains of 1 or more, even with 1000pF capacitive loads. Other key features are 0.6mV V_{OS} and a voltage gain over 4 million. Each individual amplifier is 100% tested for voltage noise, slew rate (4.2V/µs), and gain-bandwidth product (5.3MHz).

The LT1169 is offered in the S8 and N8 packages.

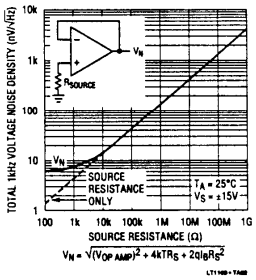
A full set of matching specifications are provided for precision instrumentation amplifier front ends. Specifications at ±5V supply operation are also provided. For an even lower voltage noise please see the LT1113 data sheet.

TYPIC

Voltage Noise vs Frequency

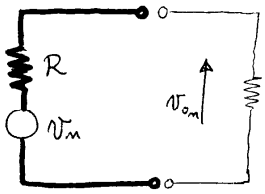


1kHz Output Voltage Noise Density vs Source Resistance



¿RUIDO EN CONDENSADORES? ¿Y EN BOBINAS?

Normalmente consideramos ruido térmico generado en una resistencia con densidad espectral de potencia plana con la frecuencia.

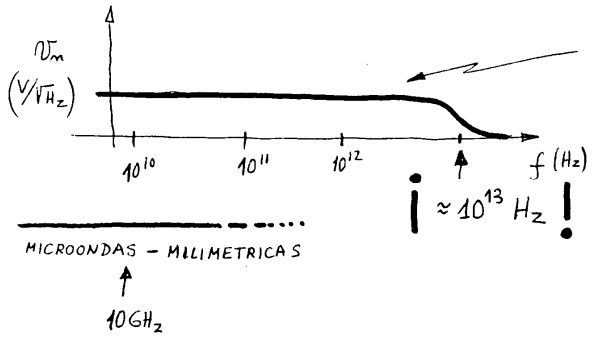


$R_L = R$ (máx. transfer. de potencia)

$$v_n^2 = 4KTR\cdot(BW) \quad \Rightarrow \quad v_{o_n}^2 = 4KT(BW)\cdot\frac{R\cdot R_L^2}{(R_L+R)^2} = \underline{\underline{KTR}}(BW)$$

Si $BW = 1Hz$ → dens. espectrales. (spot noise)

Si conectamos a la resistencia R un voltímetro con $R_L \to \infty$ y capaz de medir en cualquier frecuencia veríamos una tensión eficaz espectral de $\sqrt{4KTR}\,(V/\sqrt{Hz})$ constante con la frecuencia y ello ocurre hasta frecuencias cercanas al infrarrojo (luz) por lo que en anchos de banda "electrónicos" DC → 100 GHz (milimétricas) el ruido Johnson es plano o blanco. (Salvo que τ_d sea bajo.... hablar)



espectro plano en "electronica" BF
en " de Radiofrec
en " de microonda

¡ $\approx 10^{13}$ Hz ! (Termodinámica, Nyquist)

Ahora bien, ¿qué medirá nuestro voltímetro ideal ($R_L \to \infty$, $BW \to \infty$) al conectarse a esa resistencia?

Respuesta "teórica":

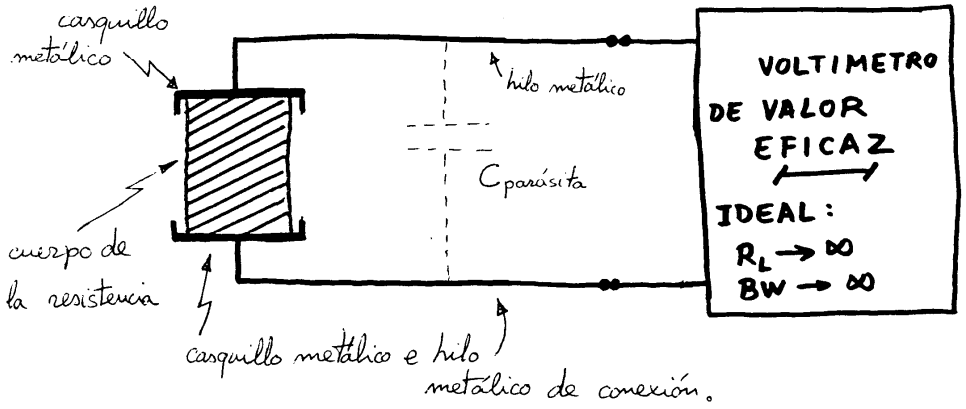
Medirá un valor eficaz de ruido: $v_{n_{ef}}(\text{voltios}) = \sqrt{4KTR \times (\sim 10^{13} Hz)}$;

algo del orden de: $\left.\begin{array}{l} K = 1'38 \cdot 10^{-23} J/K \\ T = 300 K \end{array}\right\} KT_{(300K)} = 4'14 \cdot 10^{-21} J$

Sea una $R = 1 k\Omega$

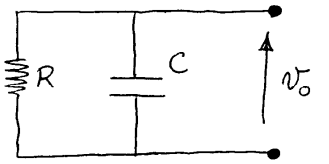
$$v_{n_{ef}} \approx \sqrt{4 \cdot 4'14 \cdot 10^{-21} \cdot 10^{3} \cdot 10^{13}} = 13 \text{ mV}_{\text{eficaces}}$$

Respuesta más real (habida cuenta cómo son las resistencias y las capacidades parásitas entre los hilos de conexion del voltímetro)



<u>C parásita</u>: Entre dos metales separados cierta distancia hay o aparece una capacidad...

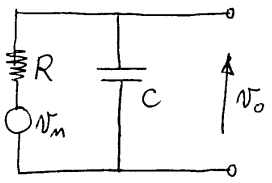
Por tanto <u>tenemos</u> (<u>sin poder evitarlo</u>) <u>un circuito RC</u>:



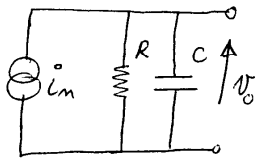
que será medida por el voltímetro ideal.

C puede ser $\leqslant 1 pF$ si somos cuidadosos

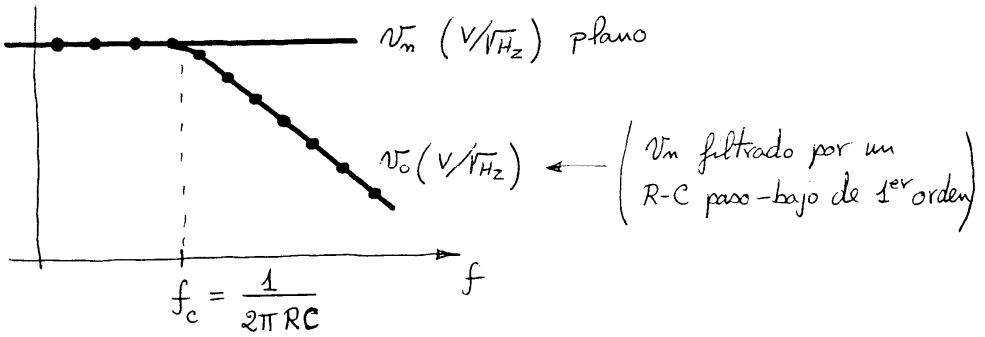
De cara al ruido podemos tomar este circuito:



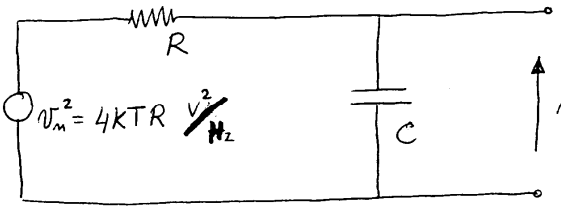
o su equivalente Norton:



Cualquiera de ellos dice lo mismo: v_o va a tener una forma no-plana debido al **FILTRADO** producido por C en altas frecuencias. En diagramas de Bode:



En otras palabras: el voltaje eficaz de ruido que midamos va a ser menor, ya que las componentes de "altas frecuencias" son atenuadas por el condensador. Calculemos $v_{n\,ef}$ en este caso.



$$v_o^2 = 4KTR \cdot \underbrace{\left| \frac{\frac{1}{j\omega C}}{\frac{1}{j\omega C} + R} \right|}_{V^2/Hz}$$

Tomamos el módulo de la tensión de salida a cada frecuencia, elevado al cuadrado, integramos y luego sacamos raíz...

$$v_{o\,ef} = \sqrt{\int_{f=0}^{f=\infty\ (\text{ó } 10^{13}\,Hz)} 4\,K\,T \cdot R \left(\frac{1}{1+\omega^2 R^2 C^2}\right) df}$$

El límite superior de integración (infinito por comodidad, ~10^{13} Hz si uno quiere ser más riguroso) ~~para~~ va a salir lo mismo a efectos de medida: $\text{arc tg}(\infty) = \pi/2$ y $\text{arc tg}(10000\ldots0) \approx \pi/2$

$$\int_0^\infty 4kTR \frac{1}{1+\left(f/f_c\right)^2} df = 4kTR \int_0^\infty \frac{1}{1+\left(f/f_c\right)^2} \cdot f_c \cdot d\left(f/f_c\right) =$$

$$= 4kTR f_c \ \text{arc tg}\left(f/f_c\right)\Big|_0^\infty = 4kTR \cdot \underbrace{\left(f_c \cdot \frac{\pi}{2}\right)}_{BW_m}$$

ANCHO DE BANDA DE RUIDO PARA EL R-C paso-bajo.

⊛ **1er resultado**: $BW_m = \frac{\pi}{2} f_c$ para un RC de 1er orden paso-bajo con $f_c = \frac{1}{2\pi RC}$

⊛ **2º resultado**:

$$v_{o\,ef} = \sqrt{4kTR\left(\frac{\pi}{2} f_c\right)} = \sqrt{4kTR \cdot \frac{\pi}{2} \cdot \frac{1}{2\pi RC}} = \sqrt{\frac{kT}{C}} \ \text{voltio}$$

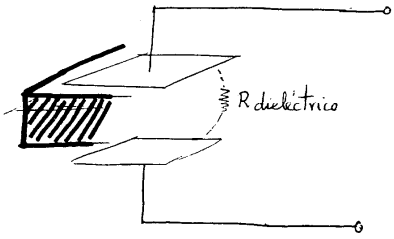
64 μV/cm 1 pF

"La tensión eficaz de ruido a la salida NO depende de R, sino de C"

¿Qué significa esto? Que sea cual sea R, C es quien determina

el valor eficaz del voltaje entre sus bornas: a mayor C menor fluctuación de tensión en C y viceversa.

Pero más aún: ¿Qué es un condensador?

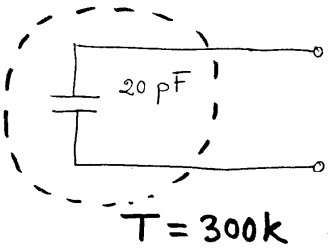


Dos placas metálicas con un dieléctrico entre ellas, que suele dar lugar a ciertas perdidas por conducción: cierta R aunque muy alta ---

Pero esto es nuestro circuito RC de una R muy alta...

Es decir: <u>un condensador</u> real va a presentar "el solito" una tensión de ruido térmico eficaz dada por: $\sqrt{\frac{kT}{C}}$ voltios

Por ejemplo: condensador de 20pF a Temp. ambiente

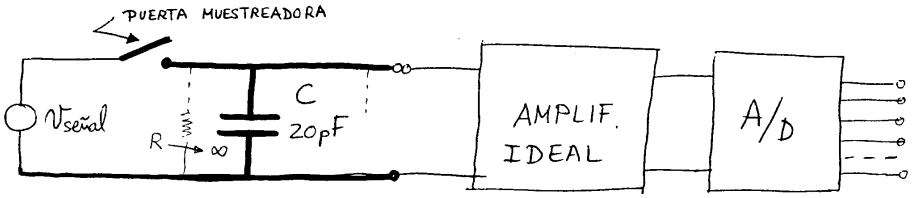


$$v_{0ef} = \sqrt{\frac{1'38 \cdot 10^{-23} \cdot 300}{20 \cdot 10^{-12}}} = 14'4\,\mu V_{ef}$$

14'4 microvoltios eficaces ⇒ bastantes más microvoltios pico-pico (y <u>no aplicamos</u> lo de 5 o 6 veces el valor eficaz porque <u>no es un ruido blanco</u> de espectro plano y densidad de probabilidad de amplitud gaussiana)

No obstante puede ser un factor 2 o 3 ----

En cualquier caso, pensemos en el circuito de muestreo-retención con un condensador de este valor: ⑥



Si $v_{señal} = 0$ voltios y se muestrea eso: 0 voltios. Al retener queda C a la entrada del amplificador. Pero C genera o tiene (por estar a T = 300K) 14'4 μV eficaces. Si el A/D es de 16 bits y el rango dinámico de conversión es de 1 voltio, el intervalo de cuantificación es $1V/2^{16} = 15\mu V \Rightarrow$ la salida no será 000---0 sino que tendrá un ruido de ± 1 bit (y a veces 2 o 3 según $v_{pico} = f(v_{eficaz\ ruido})$). ⇒ RESOLUCION $\equiv \sqrt{\frac{KT}{C}}$

Significado de ese $v_{o\,ef} = \sqrt{\frac{KT}{C}}$:

El condensador, en circuito abierto y a T = 300K, puede almacenar energía térmica (fluctuación de su voltaje). El valor eficaz de esa fluctuación es: $v_{o\,ef} = \sqrt{\frac{KT}{C}}$. Si fuera un término DC (el ruido que vemos es a.c. puro, su nivel DC es nulo) la energía almacenada en el condensador sería:

$$E = \frac{1}{2} C V^2 = \frac{1}{2} C \cdot \frac{KT}{C} = \frac{1}{2} KT \quad \text{Julios}$$

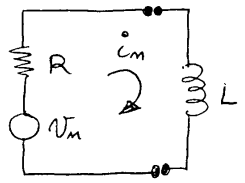
Por el significado de "valor eficaz" podemos decir que el condensador almacena $\frac{1}{2}KT$ Julios por estar en contacto térmico con el universo a

T = 300 **K**. Esto recuerda el fenómeno termodinámico de almacenamiento de $\frac{1}{2}kT$ unidades de energía por grado de libertad en un sistema a una temperatura T.

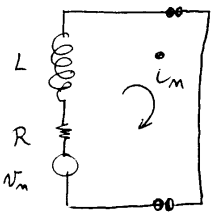
Si un <u>condensador en situación de almacenar energía</u> (es decir <u>con sus bornas NO-CORTOCIRCUITADAS</u>) almacena $\frac{kT}{2}$ Jul, una bobina debería hacer algo parecido.

Para <u>la bobina</u> la situación es dual (la energía se almacenará en forma de fluctuación de corriente) por lo que deberá estar <u>con sus bornas NO-ABIERTAS</u>. Así podrá almacenar energía como fluctuación de su campo magnético.

Partimos de la misma situación: resistencia a la que se le conecta una inductancia:



O realmente esto es una inductancia real (con cierta R serie) y con sus bornas cortocircuitadas. v_n es el ruido térmico debido a esa resistencia, que provocará cierta corriente de ruido i_n



$$i_n^2 = \frac{v_n^2}{|Z|^2} \quad \text{donde} \quad Z = R + j\omega L$$

$$i_n^2 = \frac{4kTR}{R^2+\omega^2L^2} = \frac{4kT}{R}\cdot\frac{1}{1+\frac{\omega^2}{\left(\frac{R}{L}\right)^2}} \quad \left(A^2/Hz\right)$$

La ecuación anterior representa la densidad espectral de corriente de ruido i_n debida a v_n. Para obtener un valor eficaz de corriente de ruido, integramos en frecuencia i_n^2 y luego sacamos la raíz cuadrada.

$$i_{n\,eficaz} = \sqrt{\int_0^\infty \frac{4kT}{R} \cdot \frac{1}{1+\left(\frac{f}{f_c}\right)^2}\, df} \qquad \text{con } f_c = \frac{1}{2\pi}\frac{R}{L}$$

integral similar a la del caso del condensador

$$\int_0^\infty \frac{4kT}{R} \frac{1}{1+\left(\frac{f}{f_c}\right)^2} \cdot f_c \cdot d\left(\frac{f}{f_c}\right) = \frac{4kT}{R}\left(f_c \operatorname{arctg}\left(\frac{f}{f_c}\right)\Big|_0^\infty\right) = \frac{4kT}{R}\cdot\left(f_c \cdot \frac{\pi}{2}\right.$$

cosa de esperar al ver el filtro paso-bajo de 1er orden que se aplica al término $4kTR$.

Por tanto:

$$i_{n\,eficaz} = \sqrt{\frac{4kT}{R} \cdot \frac{\pi}{2} \cdot \frac{R}{2\pi L}} = \sqrt{\frac{kT}{L}}$$

que como vemos, tampoco depende de R, sólo de L. ¿Que energía térmica almacenada supone?

$$E = \frac{1}{2} L\, i_{n\,ef.}^2 = \frac{1}{2} kT \quad \left(\frac{kT}{2} \text{ por grado de libertad...}\right)$$

Sea $L = 1\mu H$ a $T = 300K \Rightarrow i_{n\,ef}^2 = \sqrt{\frac{1'38\cdot 10^{-23}\cdot 300}{10^{-6} H}} = 64\ mA_{eficaces}$

$L = 100\mu H$ a $300K$ $\quad i_{n\,ef} = 64\ mA_{ef.}$! Extender al R-L-C

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