

NE/SA/SE5212 Transimpedance Amplifier (140MHz)

Product Specification

Linear Products

DESCRIPTION

The NE/SA/SE5212 is a $14k\Omega$ transimpedance, wideband, low noise differential output amplifier, particularly suitable for signal recovery in fiber optic receivers and in any other applications where very low signal levels obtained from high-impedance sources need to be amplified.

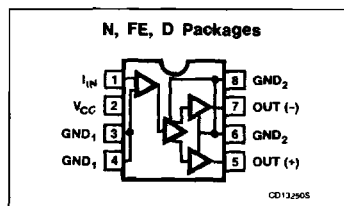
FEATURES

- Extremely low noise: $2.5pA/\sqrt{Hz}$
- Single 5V supply
- Large bandwidth: 140MHz
- Differential outputs
- Low input/output impedances
- High-power supply rejection ratio
- $14k\Omega$ differential transresistance

APPLICATIONS

- Fiber-optic receivers, analog and digital
- Current-to-voltage converters
- Wideband gain block
- Medical and scientific instrumentation
- Sensor preamplifiers
- Single-ended to differential conversion
- Low noise RF amplifiers
- RF signal processing

PIN CONFIGURATION



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ORDERING INFORMATION

DESCRIPTION	TEMPERATURE RANGE	ORDER CODE
8-Pin Plastic DIP	0 to +70°C	NE5212N
8-Pin Plastic SO	0 to +70°C	NE5212D
8-Pin Ceramic DIP	0 to +70°C	NE5212FE
8-Pin Plastic SO	-40°C to +85°C	SA5212D
8-Pin Plastic DIP	-40°C to +85°C	SA5212N
8-Pin Ceramic DIP	-40°C to +85°C	SA5212FE
8-Pin Plastic DIP	-55°C to +125°C	SE5212N
8-Pin Ceramic DIP	-55°C to +125°C	SE5212FE

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ABSOLUTE MAXIMUM RATINGS

SYMBOL	PARAMETER	RATING			UNIT
		NE5212	SA5212	SE5212	
V _{CC}	Power Supply	6	6	6	V
P _D MAX	Power dissipation, T _A = 25°C (still air) ¹				
	8-Pin Plastic DIP	1100	1100	1100	mW
	8-Pin Plastic SO	750	750	750	mW
	8-Pin Cerdip	750	750	750	mw
I _{IN} MAX	Maximum input current ²	5	5	5	mA
T _A	Operating ambient temperature range	0 to 70	-40 to 85	-55 to 125	°C
T _J	Operating junction	-55 to 150	-55 to 150	-55 to 150	°C
T _{STG}	Storage temperature range	-65 to 150	-65 to 150	-65 to 150	°C

NOTES:

- Maximum dissipation is determined by the operating ambient temperature³ and the thermal resistance:
8-Pin Plastic DIP: 110°C/W
8-Pin Plastic SO: 160°C/W
8-Pin Cerdip: 165°C/W
- The use of a pull-up resistor to V_{CC}, for the PIN diode, is recommended

RECOMMENDED OPERATING CONDITIONS

SYMBOL	PARAMETER	RATING	UNIT
V _{CC}	Supply voltage range	4.5 to 5.5	V
T _A	Ambient temperature ranges		
	NE Grade	0 to +70	°C
	SA Grade	-40 to +85	°C
	SE Grade	-55 to +125	°C
T _J	Junction temperature ranges		
	NE Grade	0 to +90	°C
	SA Grade	-40 to +105	°C
	SE Grade	-55 to +145	°C

DC ELECTRICAL CHARACTERISTICS Minimum and Maximum limits apply over operating temperature range at V_{CC} = 5V, unless otherwise specified. Typical data applies at V_{CC} = 5V and T_A = 25°C.

SYMBOL	PARAMETER	TEST CONDITIONS	NE5212			SA/SE5212			UNIT
			Min	Typ	Max	Min	Typ	Max	
V _{IN}	Input bias voltage		0.6	0.8	0.95	0.55	0.8	1.05	V
V _{O±}	Output bias voltage		2.8	3.3	3.7	2.5	3.3	3.8	V
V _{OS}	Output offset voltage				80			120	mV
I _{CC}	Supply current		21	26	32	20	26	33	mA
I _{OMAX}	Output sink/source current		3	4		3	4		mA
I _{IN}	Input current (2% linearity)	Test Circuit 6, Procedure 2	±60	±80		±40	±80		μA
I _N MAX	Maximum input current overload threshold	Test Circuit 6, Procedure 4	±80	±120		±60	±120		μA

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AC ELECTRICAL CHARACTERISTICS Minimum and Maximum limits apply over operating temperature range at $V_{CC} = 5V$, unless otherwise specified. Typical data applies at $V_{CC} = 5V$ and $T_A = 25^\circ C$.

SYMBOL	PARAMETER	TEST CONDITIONS	NE5212			SA/SE5212			UNIT
			Min	Typ	Max	Min	Typ	Max	
R_T	Transresistance (differential output)	DC tested, $R_L = \infty$ Test Circuit 6, Procedure 1	9.8	14	18.2	9.0	14	19	$k\Omega$
R_O	Output resistance (differential output)	DC tested	14	30	42	14	30	46	Ω
R_T	Transresistance (single-ended output)	DC tested, $R_L = \infty$	4.9	7	9.1	4.5	7	9.5	$k\Omega$
R_O	Output resistance (single-ended output)	DC tested	7	15	21	7	15	23	Ω
f_{3dB}	Bandwidth (-3dB)	Test Circuit 1 D package, $T_A = 25^\circ C$	100	140		100	140		MHz
		N, FE packages, $T_A = 25^\circ C$	100	120		100	120		MHz
R_{IN}	Input resistance		75	110	143	70	110	150	Ω
C_{IN}	Input capacitance			10	15		10	18	pF
$\Delta R/\Delta V$	Transresistance power supply sensitivity	$V_{CC} = 5 \pm 0.5V$		9.6			9.6		%/V
$\Delta R/\Delta T$	Transresistance ambient temperature sensitivity	D package $\Delta T_A = T_{A \text{ MAX}} - T_{A \text{ MIN}}$		0.05			0.05		%/°C
I_N	RMS noise current spectral density (referred to input)	Test Circuit 2 $f = 10MHz$ $T_A = 25^\circ C$		2.5			2.5		pA/\sqrt{Hz}
I_T	Integrated RMS noise current over the bandwidth (referred to input) $C_S = 0^1$	$T_A = 25^\circ C$ Test Circuit 2 $\Delta f = 50MHz$ $\Delta f = 100MHz$ $\Delta f = 200MHz$		20 27 40			20 27 40		nA nA nA
		$\Delta f = 50MHz$ $\Delta f = 100MHz$ $\Delta f = 200MHz$		22 32 52			22 32 52		nA nA nA
PSRR	Power supply rejection ratio ²	Any package DC tested $\Delta V_{CC} = 0.1V$ Equivalent AC Test Circuit 3	26	33		20	33		dB
PSRR	Power supply rejection ratio ² (ECL configuration)	Any package $f = 0.1MHz^1$ Test Circuit 4		23			23		dB
$V_{O \text{ MAX}}$	Maximum differential output voltage swing	$R_L = \infty$ Test Circuit 6, Procedure 3	2.4	3.2		1.7	3.2		$V_{p,p}$
$V_{IN \text{ MAX}}$	Maximum input amplitude for output duty cycle of $50 \pm 5\%$ ³	Test Circuit 5		325			325		$mV_{p,p}$
t_R	Rise time for 50mV output signal ⁴	Test Circuit 5		2.0			2.0		ns

NOTES:

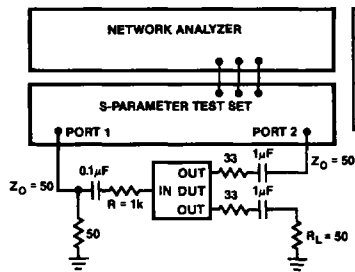
- Package parasitic capacitance amounts to about 0.2pF.
- PSRR is output referenced and is circuit board layout dependent at higher frequencies. For best performance use RF filter in V_{CC} line.
- Guaranteed by linearity and over load tests.
- t_R defined as 20 - 80% rise time. It is guaranteed by -3dB bandwidth test.

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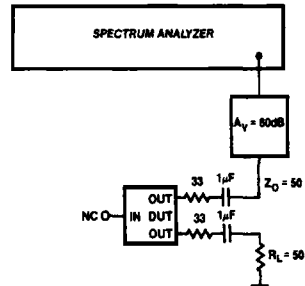
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TEST CIRCUITS

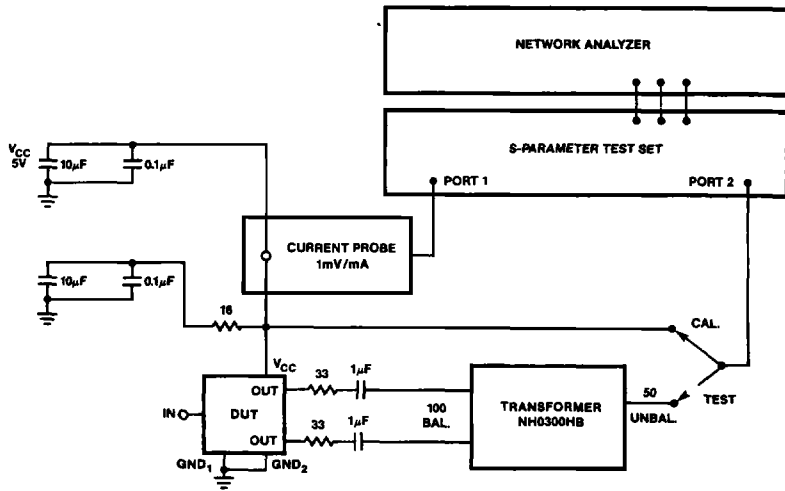


SINGLE-ENDED		DIFFERENTIAL	
$R_L = \frac{V_{OUT}}{V_{IN}}$	$R = 2 \times S_{21} \times R$	$R_L = \frac{V_{OUT}}{V_{IN}}$	$R = 4 \times S_{21} \times R$
$R_O = Z_O$	$\left \frac{1 + S_{22}}{1 - S_{22}} \right - 33$	$R_O = 2Z_O$	$\left \frac{1 + S_{22}}{1 - S_{22}} \right - 66$



Test Circuit 1

Test Circuit 2

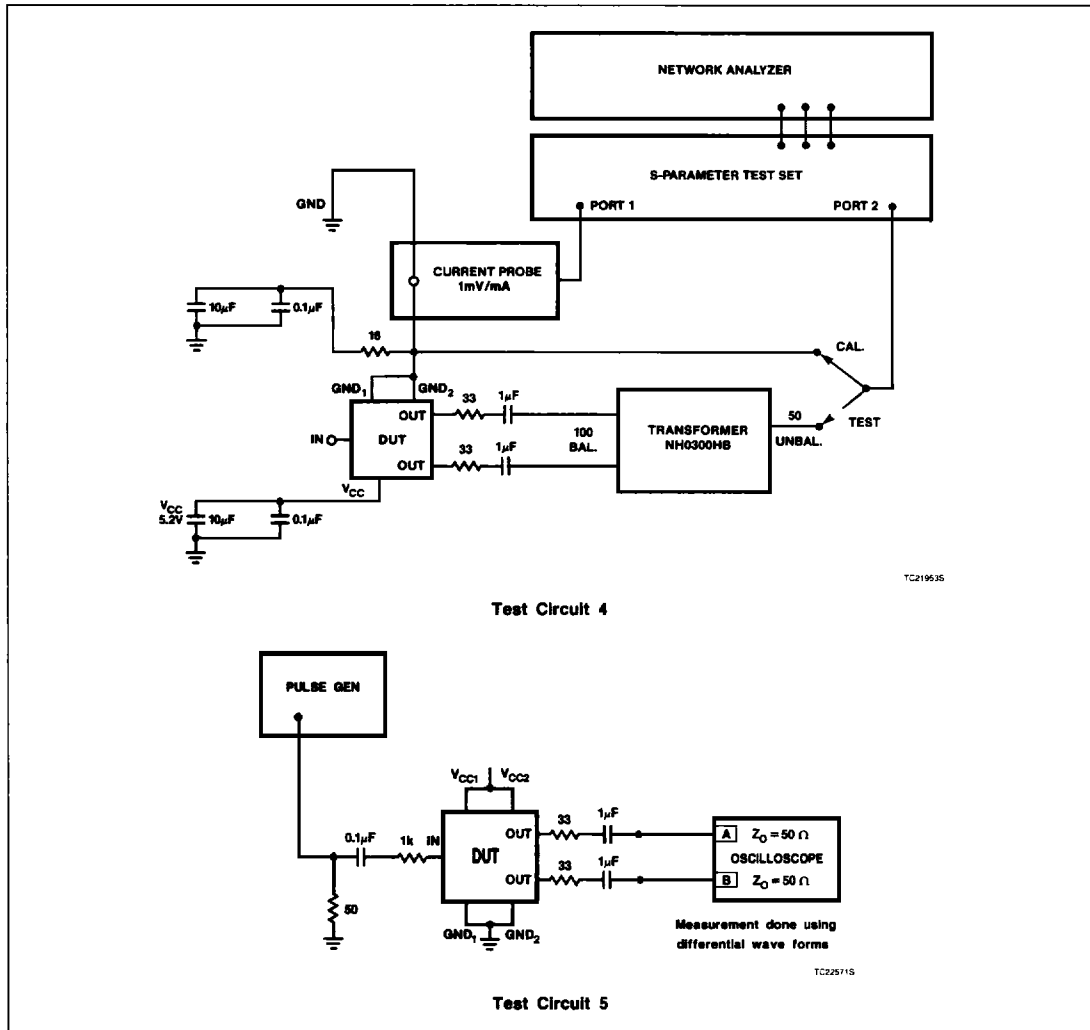


Test Circuit 3

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TEST CIRCUITS (Continued)

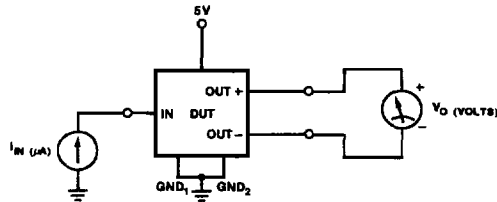


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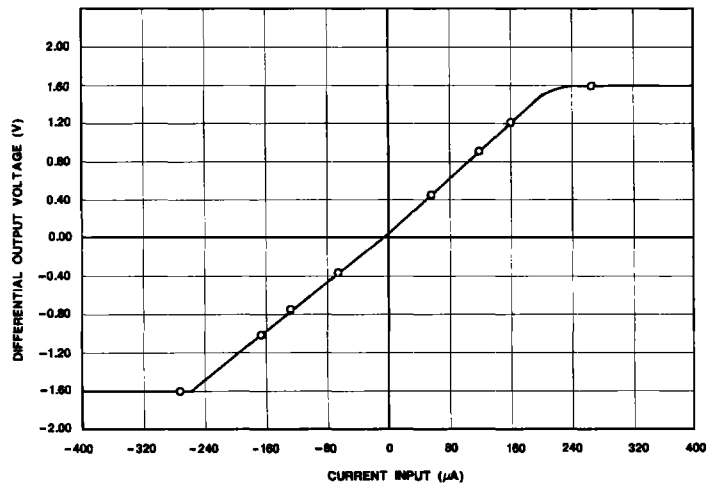
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TEST CIRCUITS (Continued)



TC234625

Typical V_O (Differential) vs I_{IN} 

OP20965

NE5212 TEST CONDITIONS

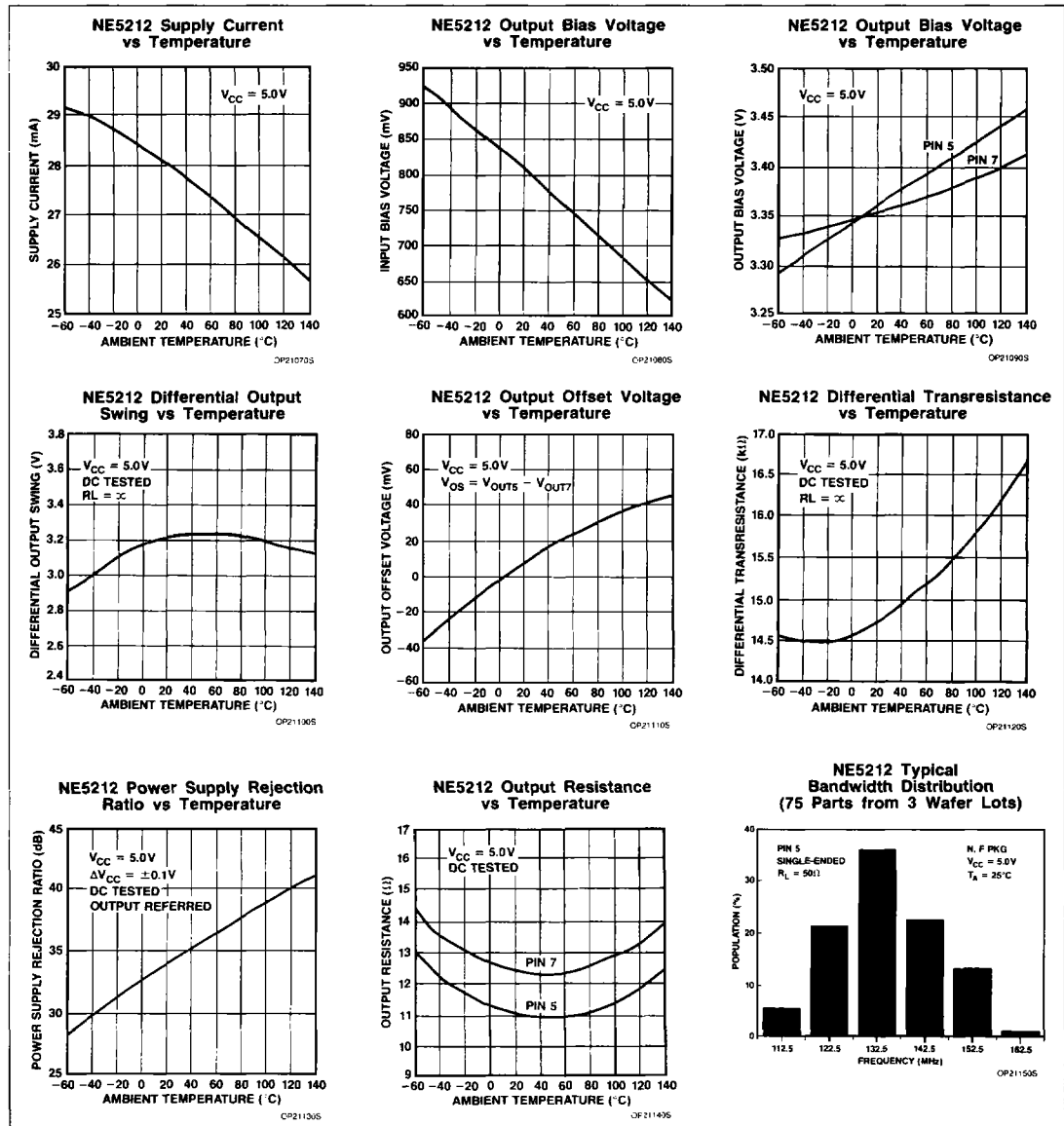
- Procedure 1 R_T measured at $30\mu A$
 $R_T = (V_{O1} - V_{O2}) / (+30\mu A - (-30\mu A))$
 Where: V_{O1} Measured at $I_{IN} = +30\mu A$
 V_{O2} Measured at $I_{IN} = -30\mu A$
- Procedure 2 Linearity = $1 - \text{ABS}((V_{O3} - V_{O5}) / (V_{O3} - V_{O4}))$
 Where: V_{O3} Measured at $I_{IN} = +60\mu A$
 V_{O4} Measured at $I_{IN} = -60\mu A$
 $V_{O3} = R_T * (+60\mu A) + V_{O8}$
 $V_{O5} = R_T * (-60\mu A) + V_{O8}$
- Procedure 3 $V_{OMAX} = V_{O7} - V_{O8}$
 Where: V_{O7} Measured at $I_{IN} = +130\mu A$
 V_{O8} Measured at $I_{IN} = -130\mu A$
- Procedure 4 I_{IN} MAX Test Pass Conditions:
 $V_{O7} - V_{O5} > 50mV$ and $V_{O6} - V_{O5} > 50mV$
 Where: V_{O5} Measured at $I_{IN} = +80\mu A$
 V_{O6} Measured at $I_{IN} = -80\mu A$
 V_{O7} Measured at $I_{IN} = +130\mu A$
 V_{O8} Measured at $I_{IN} = -130\mu A$

Test Circuit 6

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TYPICAL PERFORMANCE CHARACTERISTICS

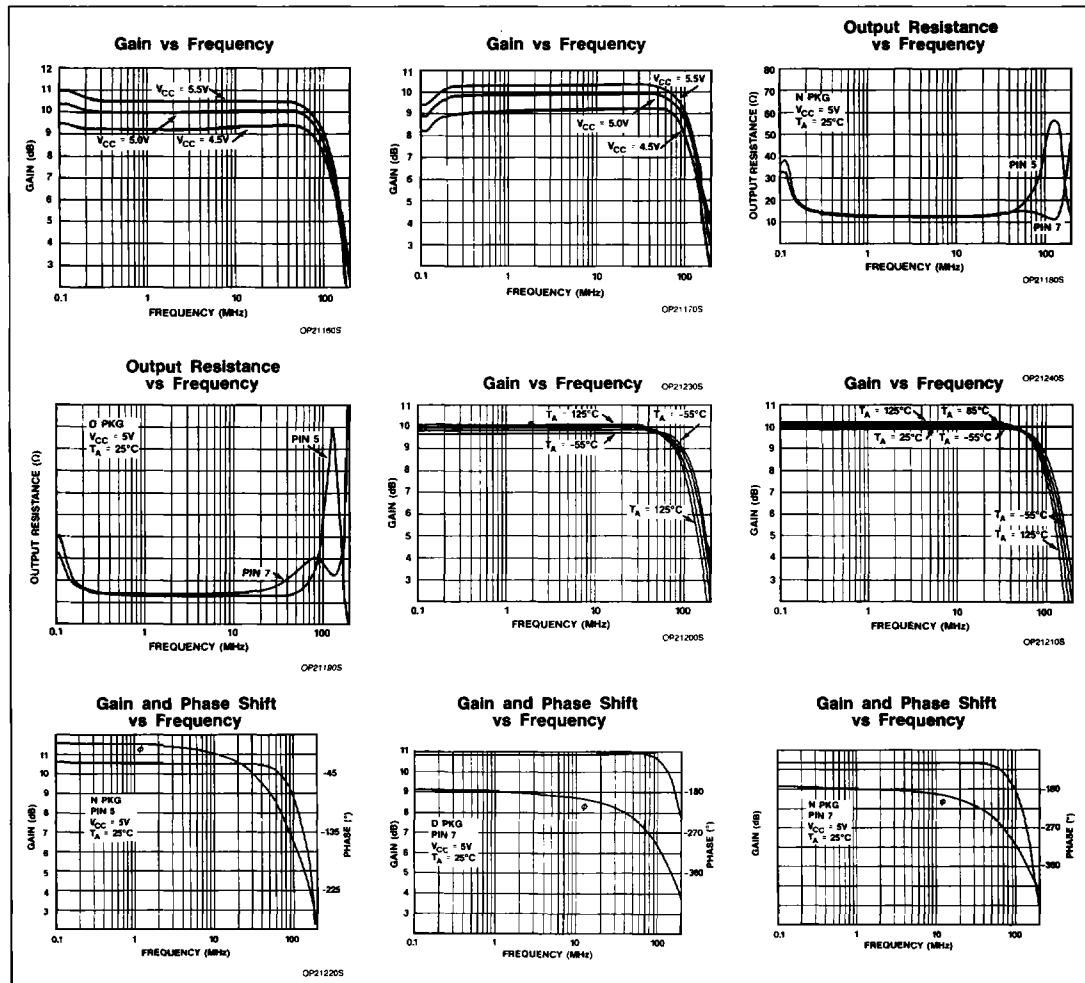


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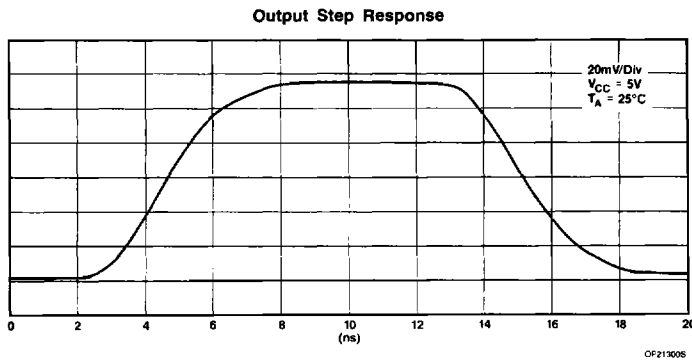
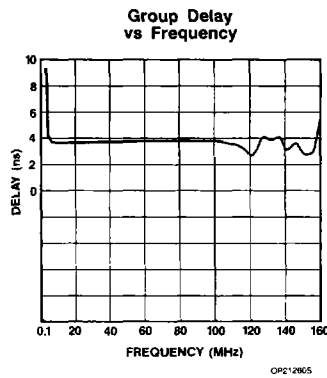
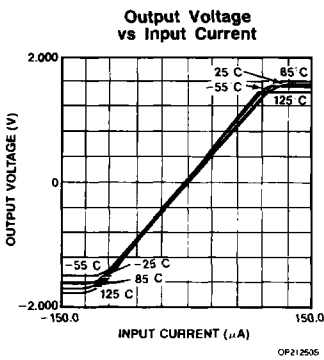
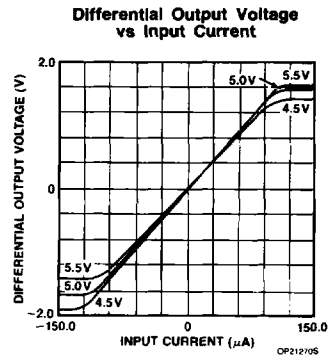
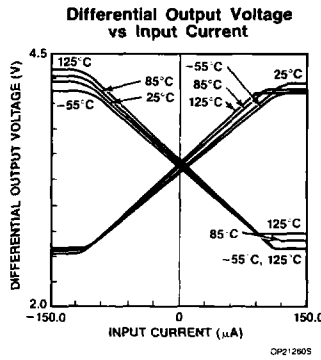
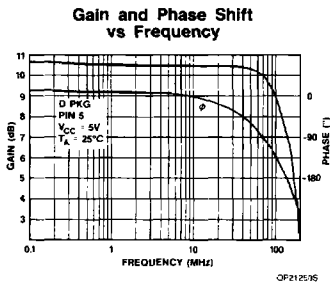
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)



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TYPICAL PERFORMANCE CHARACTERISTICS (Continued)



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THEORY OF OPERATION

Transimpedance amplifiers have been widely used as the preamplifier in fiber optic receivers. The NE5212 is a wide bandwidth (typically 130MHz) transimpedance amplifier designed primarily for high sensitivity. The maximum input current before output stage clipping occurs at typically 120µA. The NE5212 is a bipolar transimpedance amplifier which is current driven at the input and generates a differential voltage signal at the outputs. The forward transfer function is therefore a ratio of the differential output voltage to a given input current with the dimensions of ohms. The main feature of this amplifier is a wide-band, low-noise input stage which is desensitized to photodiode capacitance variations. When connected to a photodiode of a few picoFarads, the frequency response will not be degraded significantly. Except for the input stage, the entire signal path is differential to provide improved power-supply rejection and ease of interface to ECL type circuitry. A block diagram of the circuit is shown in Figure 1. The input stage (A1) employs shunt-series feedback to stabilize the current gain of the amplifier. The transresistance of the amplifier from the current source to the emitter of Q₃ is approximately the value of the feedback resistor, R_F = 7.2kΩ. The gain from the second stage (A2) and emitter followers (A3 and A4) is about two. Therefore, the differential transresistance of the entire amplifier, R_T is

$$R_T = \frac{V_{OUT(diff)}}{I_{IN}} = 2R_F = 2(7.2k) = 14.4k\Omega.$$

The single-ended transresistance of the amplifier is typically 7.2kΩ.

The simplified schematic in Figure 2 shows how an input current is converted to a differential output voltage. The amplifier has a single input for current which is referenced to Ground 1. An input current from a laser diode, for example, will be converted into a voltage by the feedback resistor R_F. The transistor Q₁ provides most of the open loop gain of the circuit, A_{VOL} ≈ 70. The emitter follower Q₂ minimizes loading on Q₁. The transistor Q₄, resistor R₇, and V_{B1} provide level shifting and interface with the Q₁₅-Q₁₆ differential pair of the second stage which is biased with an internal reference, V_{B2}. The differential outputs are derived from emitter followers Q₁₁-Q₁₂ which are biased by constant current sources. The collectors of Q₁₁-Q₁₂ are bonded to an external pin, V_{CC2}, in order to reduce the feedback to the input stage. The output impedance is about 17Ω single-ended. For ease of performance evaluation, a 33Ω resistor is used in series with each output to match to a 50Ω test system.

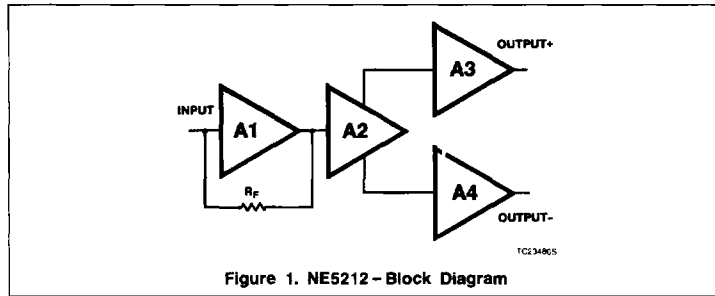


Figure 1. NE5212 - Block Diagram

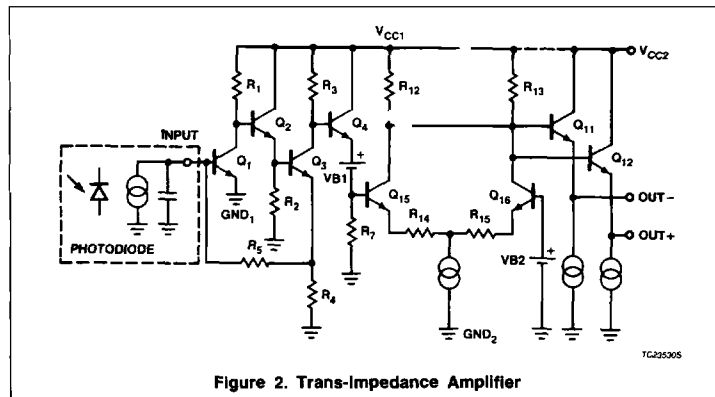


Figure 2. Trans-Impedance Amplifier

BANDWIDTH CALCULATIONS:

The input stage, shown in Figure 3, employs shunt-series feedback to stabilize the current gain of the amplifier. A simplified analysis can determine the performance of the amplifier. The equivalent input capacitance, C_{IN}, in parallel with the source, I_S, is approximately 10pF, assuming that C_S = 0 where C_S is the external source capacitance.

Since the input is driven by a current source the input must have a low input resistance. The input resistance, R_{IN}, is the ratio of the incremental input voltage, V_{IN}, to the corresponding input current, I_{IN} and can be calculated as:

$$R_{IN} = \frac{V_{IN}}{I_{IN}} = \frac{R_F}{1 + A_{VOL}} = \frac{7.2k}{70} = 103\Omega.$$

More exact calculations would yield a value of 110Ω.

Thus C_{IN} and R_{IN} will form the dominant pole of the entire amplifier;

$$f_{-3dB} = \frac{1}{2\pi R_{IN} C_{IN}}$$

Assuming typical values for R_F = 7.2kΩ, R_{IN} = 110Ω, C_{IN} = 10pF:

$$f_{-3dB} = \frac{1}{2\pi \cdot 110 \cdot 10 \times 10^{-12}} = 145\text{MHz}.$$

The operating point of Q₁ has been optimized for the lowest current noise without introducing a second dominant pole in the pass-band. All poles associated with subsequent stages have been kept at sufficiently high enough frequencies to yield an overall single pole response. Although wider bandwidths have been achieved by using a cascode input stage configuration, the present solution has the advantage of a very uniform, highly desensitized frequency response because the Miller effect dominates over the external photodiode and stray capacitances. For example, assuming a source capacitance of 1pF, input stage voltage gain of 70, R_{IN} = 110Ω then the total input capacitance, C_{IN} = (1 + 10) pF which will lead to only a 9% bandwidth reduction.

NOISE

Most of the currently installed fiber optic systems use non-coherent transmission and detect incident optical power. Therefore, receiver noise performance becomes very im-

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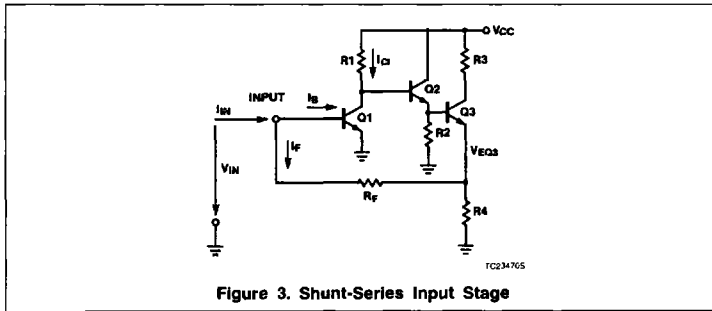


Figure 3. Shunt-Series Input Stage

portant. The input stage achieves a low input referred noise current (spectral density) of $2.5\text{pA}/\sqrt{\text{Hz}}$. The transresistance configuration assures that the external high value bias resistors often required for photodiode biasing will not contribute to the total noise system noise. The equivalent input RMS noise current is strongly determined by the quiescent current of Q_1 , the feedback resistor R_f , and the bandwidth; however, it is not dependent upon the internal Miller-capacitance. The measured wideband noise was 52nA in a 200MHz bandwidth for $C_S = 1\text{pF}$.

DYNAMIC RANGE:

The electrical dynamic range can be defined as the ratio of maximum input current to the peak noise current:

Electrical dynamic range, D_E , in a 200MHz bandwidth assuming $I_{INMAX} = 120\mu\text{A}$ and a wideband noise of $I_{EQ} = 52\text{nA}_{RMS}$ from an external source capacitance of $C_S = 1\text{pF}$.

$$D_E = \frac{(\text{Max. input current})}{(\text{Peak noise current})}$$

$$= 20 \log \frac{(120 \times 10^{-6})}{(\sqrt{2} \cdot 52 \times 10^{-9})}$$

$$= 20 \log \frac{(120\mu\text{A})}{(73\text{nA})} = 64\text{dB}$$

In order to calculate the optical dynamic range the incident optical power must be considered.

For a given wavelength λ ;

Energy of one photon = $\frac{hc}{\lambda}$ watt sec (Joule)

Where h = Planck's Constant = 6.6×10^{-34} Joule sec.

c = speed of light = 3×10^8 mt/sec

c/λ = optical frequency

$$\text{No. of incident photons/sec} = \frac{P}{hc \lambda}$$

where P = optical incident power

$$\text{No. of generated electrons/sec} = \eta \cdot \frac{P}{hc \lambda}$$

Where η = quantum efficiency

$$= \frac{\text{no. of generated electron hole pairs}}{\text{no. of incident photons}}$$

$$I = \eta \cdot \frac{P}{hc \lambda} \cdot e \text{ Amps (Coulombs/sec.)}$$

where e = electron charge = 1.6×10^{-19} Coulombs

$$\text{Responsivity } R = \frac{\eta e}{hc \lambda} \text{ Amp/watt}$$

$$I = P \cdot R$$

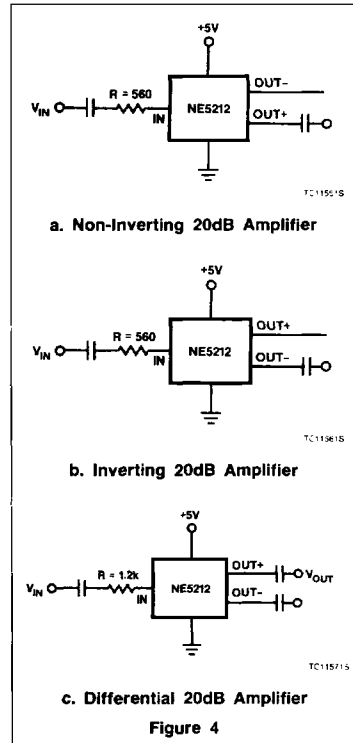
Assuming a data rate of 400Mbaud (Bandwidth, $B = 200\text{MHz}$), the noise parameter Z may be calculated as:

$$Z = \frac{I_{eq}}{qB} = \frac{52 \times 10^{-9}}{(1.6 \times 10^{-19})(200 \times 10^6)} = 1625$$

where Z is the ratio of RMS noise output to the peak response to a single hole-electron pair. Assuming 100% photodetector quantum efficiency, half mark/half space digital transmission, 850nm lightwave and using Gaussian approximation, the minimum required optical power to achieve 10^{-9} BER is:

$$P_{avMIN} = 12 \frac{hc}{\lambda} B Z = 12 \cdot 2.3 \times 10^{-19} \cdot 200 \times 10^6 \cdot 1625 = 897\text{nW} = -30.5\text{dBm}$$

where h is Planck's Constant, c is the speed of light, λ is the wavelength. The minimum



a. Non-Inverting 20dB Amplifier

b. Inverting 20dB Amplifier

c. Differential 20dB Amplifier

Figure 4

input current to the NE5212, at this input power is:

$$I_{avMIN} = q P_{avMIN} \frac{\lambda}{hc}$$

$$= \frac{897 \times 10^{-9} \times 1.6 \times 10^{-19}}{2.3 \times 10^{-19}}$$

$$= 624\text{nA}$$

Choosing the maximum peak overload current of $I_{avMAX} = 120\mu\text{A}$, the maximum mean optical power is:

$$P_{avMAX} = \frac{hc I_{avMAX}}{\lambda q} = \frac{2.3 \times 10^{-19}}{1.6 \times 10^{-19}} = 120 \times 10^{-6}$$

$$= 86\text{mW or } -7.6 \text{ dBm}$$

This the optical dynamic range, D_O is:

$$D_O = P_{avMAX} - P_{avMIN} = -30.5 - (-7.6) = 22.8\text{dB}$$

This represents the maximum limit attainable with the NE5212 operating at 200MHz bandwidth, with a half mark/half space digital transmission at 820nm wavelength.

¹ S. D. Personick, *Optical Fiber Transmission Systems*, Plenum Press, NY, 1981, Chapter 3.

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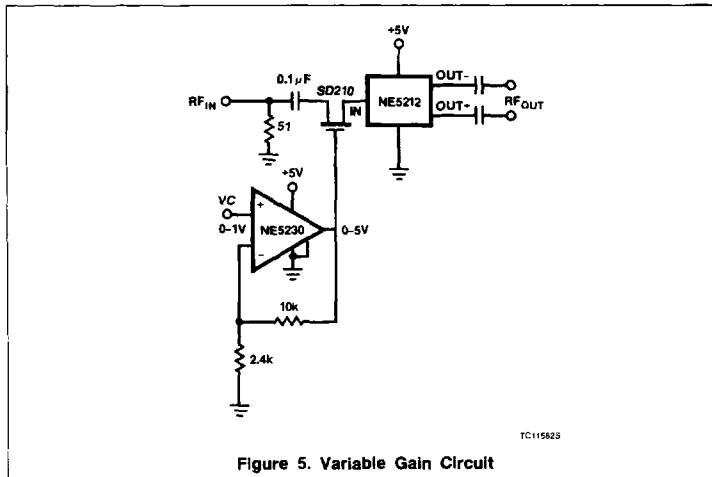


Figure 5. Variable Gain Circuit

APPLICATION INFORMATION

Package parasitics, particularly ground lead inductances and parasitic capacitances, can significantly degrade the frequency response. Since the NE5212 has differential outputs which can feed back signals to the input by parasitic package or board layout capacitances, both peaking and attenuating type frequency response shaping is possible. Constructing the board layout such that Ground 1 and Ground 2 have very low impedance paths have produced the best results. This was accomplished by adding a ground-plane stripe underneath the device connecting Ground 1, Pins 8-11, and Ground 2, Pins 1 and 2 on opposite ends of the SO14 package. This ground-plane stripe also provides isolation between the output return currents flowing to either V_{CC2} or Ground 2 and the input photodiode currents to flowing to Ground 1. Without this ground-plane stripe and with large lead inductances on the board, the part may be unstable and oscillate near 800MHz. The easiest way to realize that the part is not functioning normally is to measure the DC voltages at the outputs. If they are not close to their quiescent values of 3.3V (for a 5V supply), then the circuit may be oscillating. Input pin layout necessitates that the photodiode be physically very close to the input and Ground 1. Connecting Pins 3 and 5 to Ground 1 will tend to shield the input but it will also tend to increase the capacitance on the input and slightly reduce the bandwidth.

As with any high-frequency device, some precautions must be observed in order to enjoy reliable performance. The first of these is the use of a well-regulated power supply. The supply must be capable of providing

varying amounts of current without significantly changing the voltage level. Proper supply bypassing requires that a good quality 0.1µF high-frequency capacitor be inserted between V_{CC1} and V_{CC2} , preferably a chip capacitor, as close to the package pins as possible. Also, the parallel combination of 0.1µF capacitors with 10µF tantalum capacitors from each supply, V_{CC1} and V_{CC2} , to the ground plane should provide adequate decoupling. Some applications may require an RF choke in series with the power supply line. Separate analog and digital ground leads must be maintained and printed circuit board ground plane should be employed whenever possible.

BASIC CONFIGURATION

A trans resistance amplifier is a current-to-voltage converter. The forward transfer function then is defined as voltage out divided by current in, and is stated in ohms. The lower the source resistance, the higher the gain. The NE5212 has a differential transresistance of 14kΩ typically and a single-ended transresistance of 7kΩ typically. The device has two outputs: inverting and non-inverting. The output voltage in the differential output mode is twice that of the output voltage in the single-ended mode. Although the device can be used without coupling capacitors, more care is required to avoid upsetting the internal bias nodes of the device. Figure 4 shows some basic configurations.

VARIABLE GAIN

Figure 5 shows a variable gain circuit using the NE5212 and the NE5230 low voltage op

amp. This op amp is configured in a non-inverting gain of five. The output drives the gate of the SD210 DMOS FET. The series resistance of the FET changes with this output voltage which in turn changes the gain of the NE5212. This circuit has a distortion of less than 1% and a 25dB range, from -42.2dBm to -15.9dBm at 50MHz, and a 45dB range, from -60dBm to -14.9dBm at 10MHz with 0 to 1V of control voltage at V_C .

16MHz CRYSTAL OSCILLATOR

Figure 6 shows a 16MHz crystal oscillator operating in the series resonant mode using the NE5212. The non-inverting input is fed back to the input of the NE5212 in series with a 2pF capacitor. The output is taken from the inverting output.

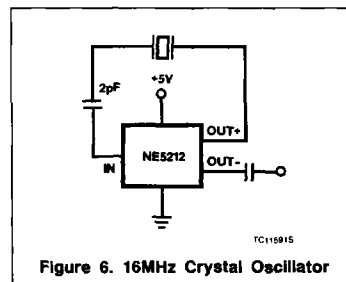


Figure 6. 16MHz Crystal Oscillator

DIGITAL FIBER OPTIC RECEIVER

Figures 7 and 8 show a fiber optic receiver using off-the-shelf components.

The receiver shown in Figure 7 uses the NE5212, the Signetics 10116 ECL line receiver, and Philips/Amperex BPF31 PIN diode. The circuit is a capacitor-coupled receiver and utilizes positive feedback in the last stage to provide the hysteresis. The amount of hysteresis can be tailored to the individual application by changing the values of the feedback resistors to maintain the desired balance between noise immunity and sensitivity. At room temperature, the circuit operates at 50Mbaud with a BER of 10E-10 and over the automotive temperature range at 40Mbaud with a BER of 10E-9. Higher speed experimental diodes have been used to operate this circuit at 220Mbaud with a BER of 10E-10.

Figure 8 depicts a TTL receiver using the NE5212 and the NE5214 fast amplifier system along with the Philips/Amperex PIN diode. The system shown is optimized for 50 Mb/s Non Return to Zero (NRZ) data. A link status indication is provided along with a jamming function when the input level is below a user-programmable threshold level.

Transimpedance Amplifier (140MHz)

NE/SA/SE5212

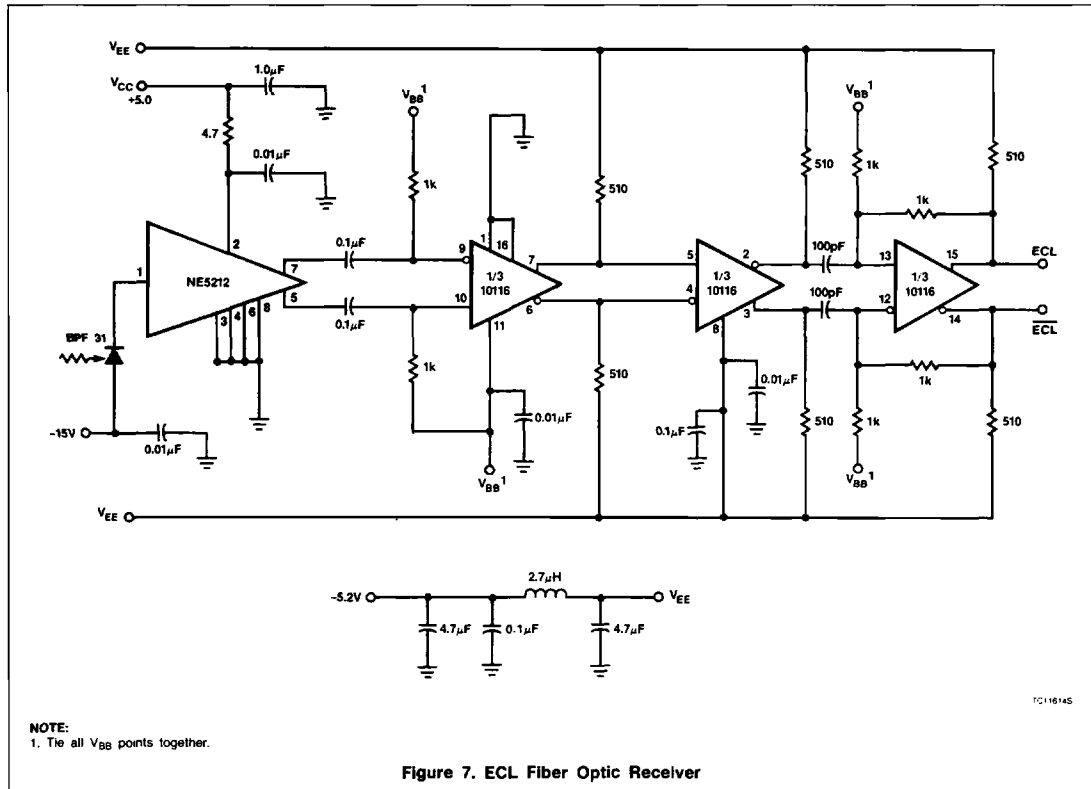


Figure 7. ECL Fiber Optic Receiver

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Transimpedance Amplifier (140MHz)

NE/SA/SE5212

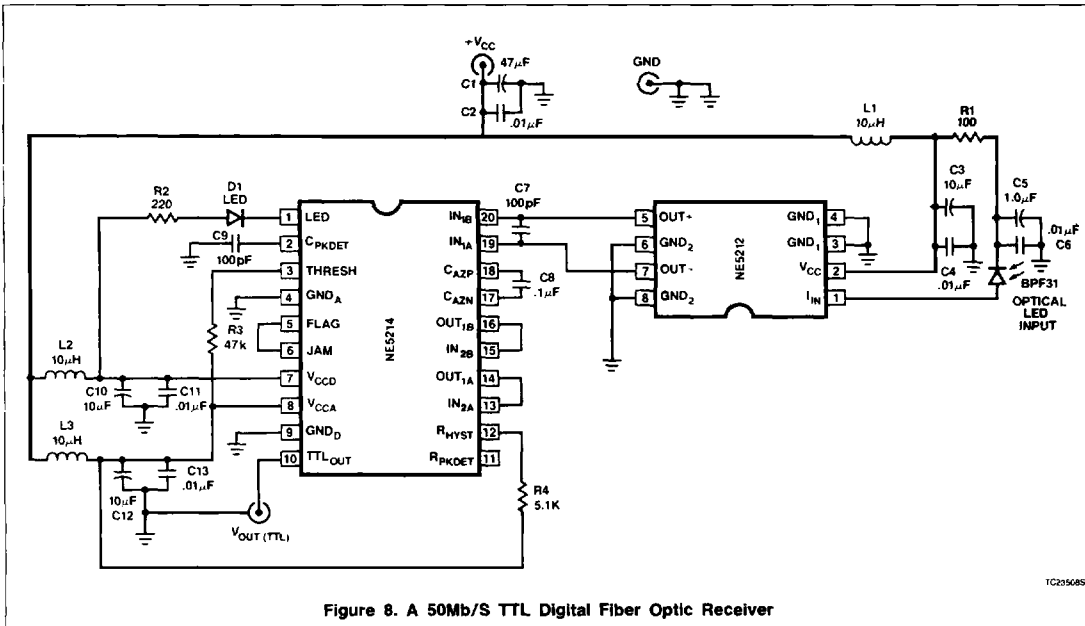


Figure 8. A 50Mb/S TTL Digital Fiber Optic Receiver