Signetics

# NE/SA5211
## Transimpedance Amplifier (180MHz)

***Preliminary Specification***

**Linear Products**

## DESCRIPTION

The NE/SA5211 is a 28kΩ transimpedance, wide-band, low noise amplifier with differential outputs, particularly suitable for signal recovery in fiber optic receivers. The part is ideally suited for many other RF applications as a general purpose gain block.

## FEATURES

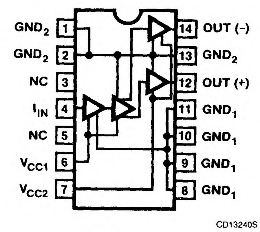
- **Extremely low noise: 1.8pA/√Hz**
- **Single 5V supply**
- **Large bandwidth: 180MHz**
- **Differential outputs**
- **Low input/output impedances**
- **High power supply rejection ratio**
- **28kΩ differential transresistance**

## APPLICATIONS

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

## PIN CONFIGURATION

**D Package**




**TOP VIEW**

## ORDERING INFORMATION

| DESCRIPTION | TEMPERATURE RANGE | ORDER CODE |
|---|---|---|
| 14-Pin Plastic SO | 0 to +70°C | NE5211D |
| 14-Pin Plastic SO | -40 to +85°C | SA5211D |

## ABSOLUTE MAXIMUM RATINGS

| SYMBOL | PARAMETER | RATING | | UNIT |
|---|---|---|---|---|
| | | NE5211 | SA5211 | |
| $V_{CC}$ | Power supply | 6 | 6 | V |
| $T_A$ | Operating ambient temperature range | 0 to +70 | -40 to +85 | °C |
| $T_J$ | Operating junction temperature range | -55 to +150 | -55 to +150 | °C |
| $T_{STG}$ | Storage temperature range | -65 to +150 | -65 to +150 | °C |
| $P_{D\ MAX}$ | Power dissipation, $T_A$ = 25°C (still-air)[1] | 1.0 | 1.0 | W |
| $I_{IN\ MAX}$ | Maximum input current[2] | 5 | 5 | mA |

**NOTES:**
1. Maximum dissipation is determined by the operating ambient temperature and the thermal resistance: $\theta_{JA}$ = 125°C/W
2. 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 | 4.5 to 5.5 | V |
| $T_A$ | Ambient temperature range | | |
| | NE Grade | 0 to +70 | °C |
| | SA Grade | −40 to +85 | °C |
| $T_J$ | Junction temperature range | | |
| | NE Grade | 0 to +90 | °C |
| | SA Grade | −40 to +105 | °C |

## DC ELECTRICAL CHARACTERISTICS

Min and Max limits apply over operating temperature at $V_{CC}$ = 5V, unless otherwise specified. Typical data apply at $V_{CC}$ = 5V and $T_A$ = 25°C.

| SYMBOL | PARAMETER | TEST CONDITIONS | NE5211 | | | SA5211 | | | UNIT |
|---|---|---|---|---|---|---|---|---|---|
| | | | Min | Typ | Max | Min | Typ | Max | |
| $V_{IN}$ | Input bias voltage | | 0.6 | 0.8 | 0.95 | 0.55 | 0.8 | 1.00 | V |
| $V_{O\pm}$ | Output bias voltage | | 2.8 | 3.4 | 3.7 | 2.7 | 3.4 | 3.7 | V |
| $V_{OS}$ | Output offset voltage | | | 0 | 120 | | 0 | 130 | mV |
| $I_{CC}$ | Supply current | | 21 | 24 | 30 | 20 | 26 | 31 | mA |
| $I_{OMAX}$ | Output sink/source current[1] | | 3 | 4 | | 3 | 4 | | mA |
| $I_{IN}$ | Input current (2% linearity) | Test Circuit 8, Procedure 2 | ±30 | ±40 | | ±20 | ±40 | | μA |
| $I_{IN\ MAX}$ | Maximum input current overload threshold | Test Circuit 8, Procedure 4 | ±40 | ±60 | | ±30 | ±60 | | μA |

**NOTE:**
1. Test condition: output quiescent voltage variation is less than 100mV for 3mA load current.

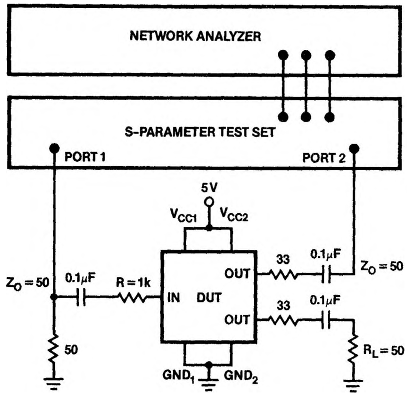
**AC ELECTRICAL CHARACTERISTICS** Typical data and Min and Max limits apply at $V_{CC}$ = 5V and $T_A$ = 25°C.

| SYMBOL | PARAMETER | TEST CONDITIONS | NE5211 | | | SA5211 | | | UNIT |
|---|---|---|---|---|---|---|---|---|---|
| | | | Min | Typ | Max | Min | Typ | Max | |
| $R_T$ | Transresistance (differential output) | DC tested<br>$R_L = \infty$<br>Test Circuit 8, Procedure 1 | 22 | 28 | 35 | 21 | 28 | 36 | kΩ |
| $R_O$ | Output resistance (differentialoutput) | DC tested | | 30 | | | 30 | | Ω |
| $R_T$ | Transresistance (single-ended output) | DC tested<br>$R_L = \infty$ | 11 | 14 | 17.5 | 10.5 | 14 | 18.0 | kΩ |
| $R_O$ | Output resistance (single-ended output) | DC tested | | 15 | | | 15 | | Ω |
| $f_{3dB}$ | Bandwidth (−3dB) | $T_A$ = 25°C<br>Test circuit 1 | | 180 | | | 180 | | MHz |
| $R_{IN}$ | Input resistance | | | 200 | | | 200 | | Ω |
| $C_{IN}$ | Input capacitance | | | 4 | | | 4 | | pF |
| $\Delta R/\Delta V$ | Transresistance power supply sensitivity | $V_{CC}$ = 5± 0.5V | | 3.7 | | | 3.7 | | %/V |
| $\Delta R/\Delta T$ | Transresistance ambient temperature sensitivity | $\Delta T_A = T_{A\ MAX} - T_{A\ MIN}$ | | 0.025 | | | 0.025 | | %/°C |
| $I_N$ | RMS noise current spectral density (referred to input) | Test Circuit 2<br>f = 10MHz<br>$T_A$ = 25°C | | 1.8 | | | 1.8 | | $pA/\sqrt{Hz}$ |
| $I_T$ | Integrated RMS noise current over the bandwidth (referred to input) | $T_A$ = 25°C<br>Test Circuit 2 | | | | | | | |
| | $C_S = 0$[1] | Δf = 50MHz<br>Δf = 100MHz<br>Δf = 200MHz | | 13<br>20<br>35 | | | 13<br>20<br>35 | | nA<br>nA<br>nA |
| | $C_S$ = 1pF | Δf = 50MHz<br>Δf = 100MHz<br>Δf = 200MHz | | 13<br>21<br>41 | | | 13<br>21<br>41 | | nA<br>nA<br>nA |
| PSRR | Power supply rejection ratio[2] ($V_{CC1} = V_{CC2}$) | DC tested, $\Delta V_{CC}$ = .01V<br>Equivalent AC<br>Test Circuit 3 | 26 | 32 | | 23 | 32 | | dB |
| PSRR | Power supply rejection ratio[2] ($V_{CC1}$) | DC tested, $\Delta V_{CC}$ = .01V<br>Equivalent AC<br>Test Circuit 4 | 26 | 32 | | 23 | 32 | | dB |
| PSRR | Power supply rejection ratio[2] ($V_{CC2}$) | DC tested, $\Delta V_{CC}$ = .01V<br>Equivalent AC<br>Test Circuit 5 | 45 | 65 | | 45 | 65 | | dB |
| PSRR | Power supply rejection ratio (ECL configuration)[2] | f = 0.1MHz<br>Test Circuit 6 | | 23 | | | 23 | | dB |
| $V_{OMAX}$ | Maximum differential output voltage swing | $R_L = \infty$<br>Test Circuit 8, Procedure 3 | 2.4 | 3.2 | | 1.7 | 3.2 | | $V_{P\text{-}P}$ |
| $V_{IN\ MAX}$ | Maximum input amplitude for output duty cycle of 50± 5%[3] | Test Circuit 7 | 160 | | | 160 | | | $mV_{P\text{-}P}$ |
| $t_R$ | Rise time for 50mV output signal[4] | Test Circuit 7 | | 0.8 | 1.2 | | 0.8 | 1.8 | ns |

**NOTES:**

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

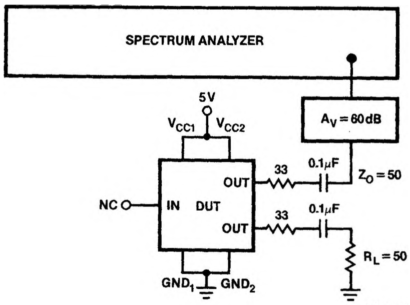
## TEST CIRCUITS




| SINGLE-ENDED | DIFFERENTIAL |
|---|---|
| $R_T \cong \frac{V_{OUT}}{V_{IN}}$ R = 2 × S21 × R | $R_T = \frac{V_{OUT}}{V_{IN}}$ R = 4 × S21 × R |
| $R_O \cong Z_O \left\|\frac{1+S22}{1-S22}\right\| - 33$ | $R_O = 2Z_O \left\|\frac{1+S22}{1-S22}\right\| - 66$ |

TC22561S

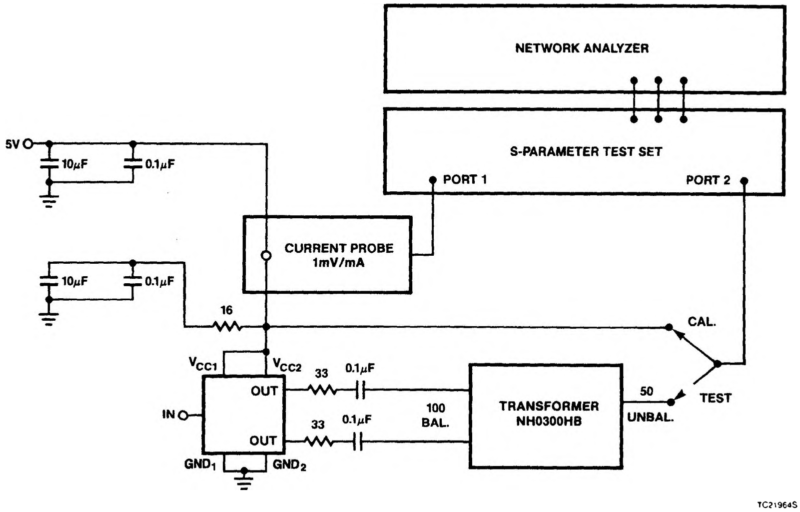
**Test Circuit 1**



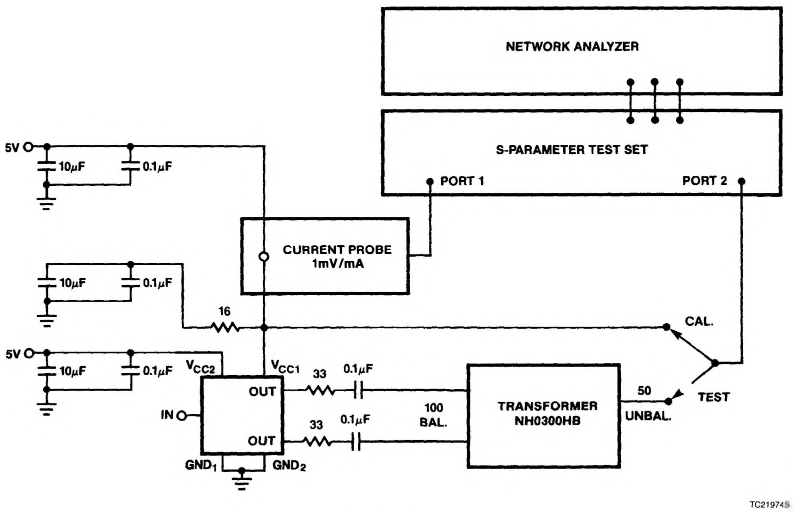

TC22551S

**Test Circuit 2**

**TEST CIRCUITS** (Continued)



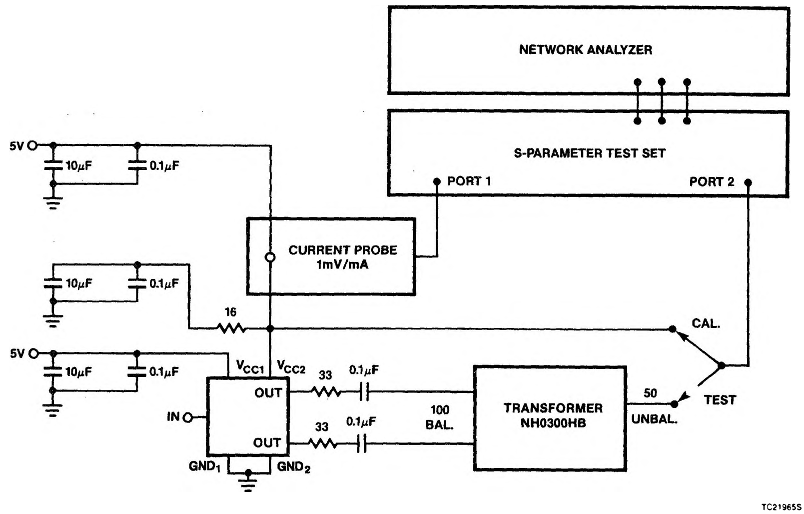

Test Circuit 3



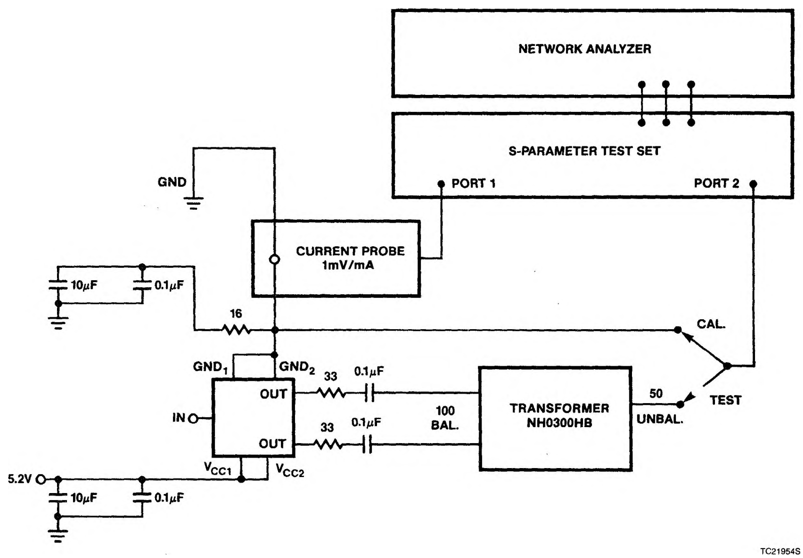

Test Circuit 4

# Transimpedance Amplifier (180MHz) NE/SA5211

## TEST CIRCUITS (Continued)

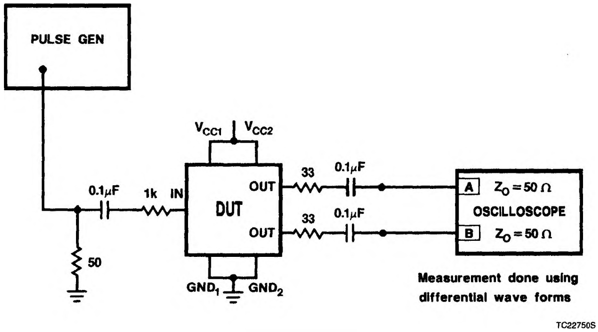



Test Circuit 5



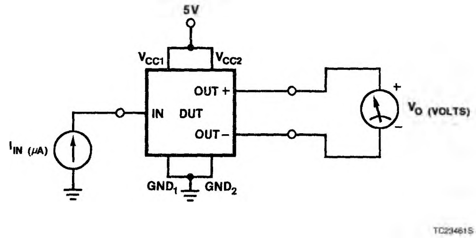

Test Circuit 6

**TEST CIRCUITS** (Continued)

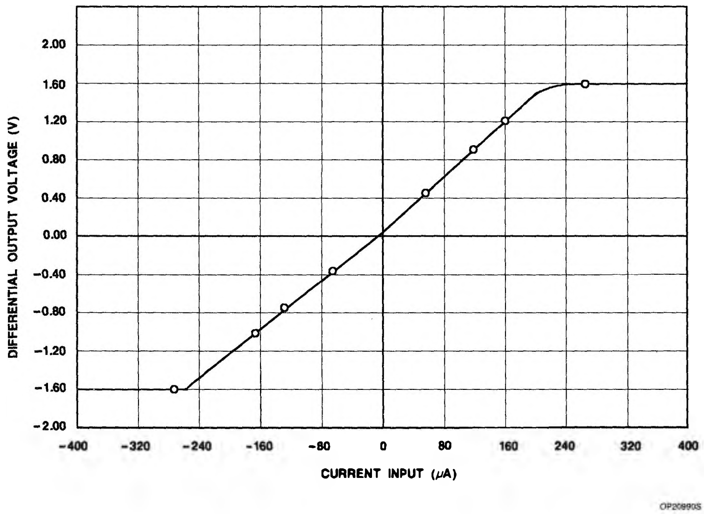



Test Circuit 7

## TEST CIRCUITS (Continued)




**Typical $V_O$ (Differential) vs $I_{IN}$**



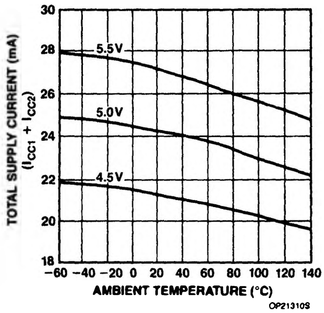

**NE5211 TEST CONDITIONS**

| | |
|---|---|
| Procedure 1 | $R_T$ measured at 15μA<br>$R_T = (V_{O1} - V_{O2})/(+15\mu A - (-15\mu A))$<br>Where: $V_{O1}$ Measured at $I_{IN} = +15\mu A$<br>$V_{O2}$ Measured at $I_{IN} = -15\mu A$ |
| Procedure 2 | Linearity = $1 - ABS((V_{OA} - V_{OB})/(V_{O3} - V_{O4}))$<br>Where: $V_{O3}$ Measured at $I_{IN} = +30\mu A$<br>$V_{O4}$ Measured at $I_{IN} = -30\mu A$<br>$V_{OA} = R_T * (+30\mu A) + V_{O8}$<br>$V_{OB} = R_T * (-30\mu A) + V_{O8}$ |
| Procedure 3 | $V_{OMAX} = V_{O7} - V_{O8}$<br>Where: $V_{O7}$ Measured at $I_{IN} = +65\mu A$<br>$V_{O8}$ Measured at $I_{IN} = -65\mu A$ |
| Procedure 4 | $I_{IN\ MAX}$ Test Pass Conditions:<br>$V_{O7} - V_{O5} > 50mV$ and $V_{O6} - V_{O5} > 50mV$<br>Where: $V_{O5}$ Measured at $I_{IN} = +40\mu A$<br>$V_{O6}$ Measured at $I_{IN} = -40\mu A$<br>$V_{O7}$ Measured at $I_{IN} = +65\mu A$<br>$V_{O8}$ Measured at $I_{IN} = -65\mu A$ |

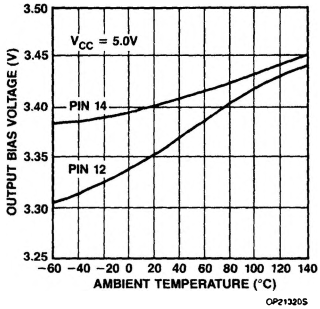
**Test Circuit 8**

## TYPICAL PERFORMANCE CHARACTERISTICS

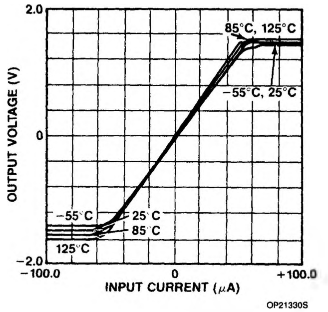
**NE5211 Supply Current vs Temperature**



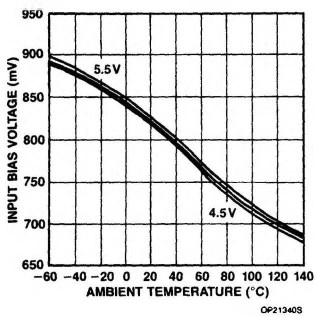

**NE5211 Output Bias Voltage vs Temperature**



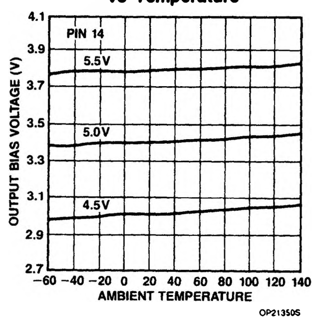

**Output Voltage vs Input Current**



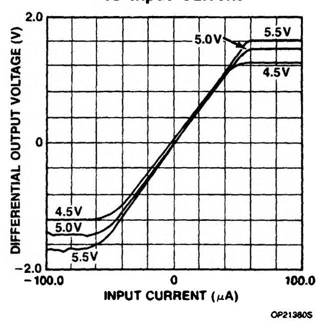

**NE5211 Input Bias Voltage vs Temperature**



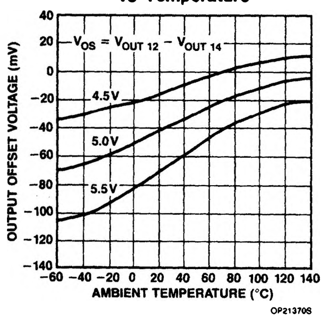

**NE5211 Output Bias Voltage vs Temperature**



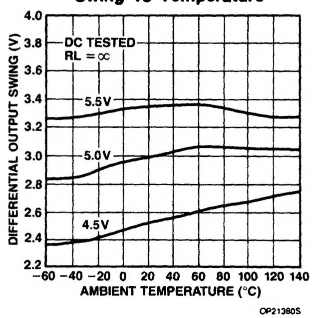

**Differential Output Voltage vs Input Current**



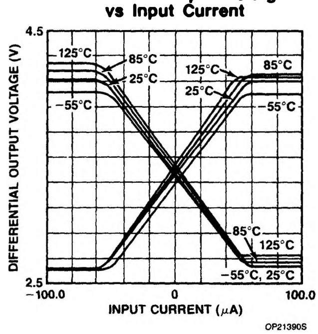

**NE5211 Output Offset Voltage vs Temperature**




**NE5211 Differential Output Swing vs Temperature**

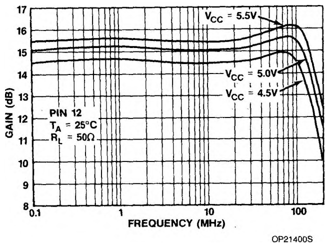



**Differential Output Voltage vs Input Current**

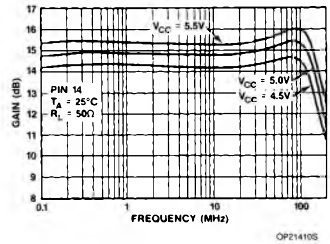


## TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

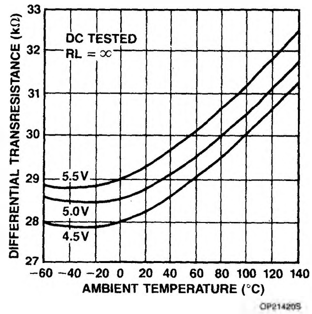
### Gain vs Frequency



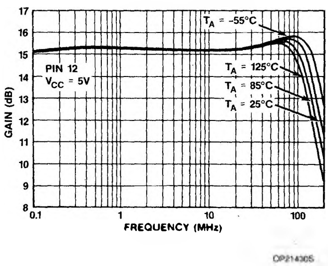

### Gain vs Frequency



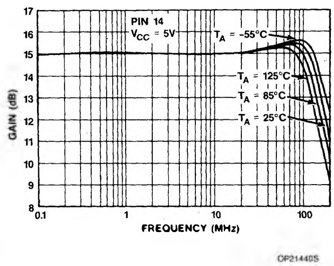

### NE5211 Differential Transresistance vs Temperature



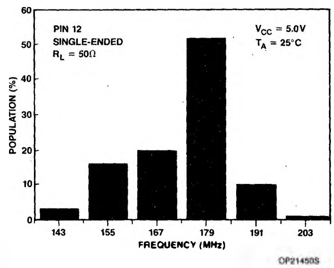

### Gain vs Frequency



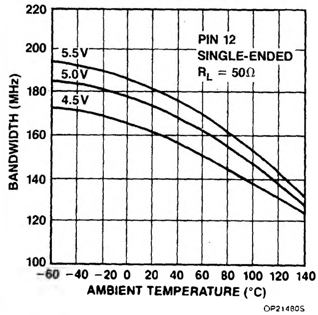

### Gain vs Frequency



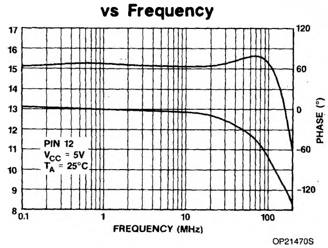

### NE5211 Typical Bandwidth Distribution (70 Parts from 3 Wafer Lots)



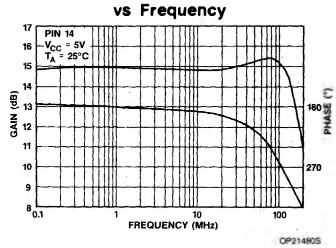

### NE5211 Bandwidth vs Temperature




### Gain and Phase Shift vs Frequency

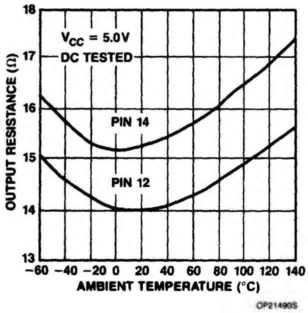



### Gain and Phase Shift vs Frequency

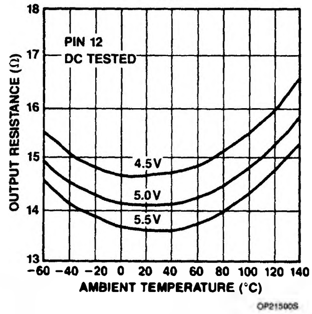


## TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

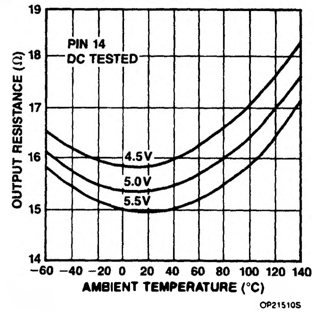
**NE5211 Output Resistance vs Temperature**



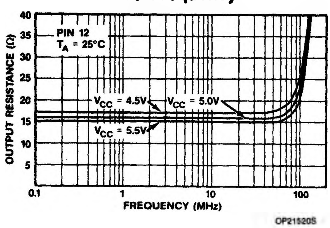

**NE5211 Output Resistance vs Temperature**



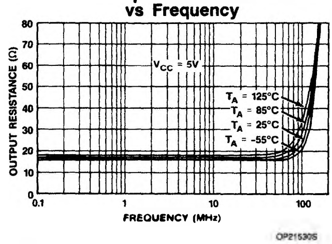

**NE5211 Output Resistance vs Temperature**



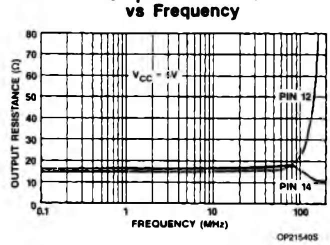

**Output Resistance vs Frequency**



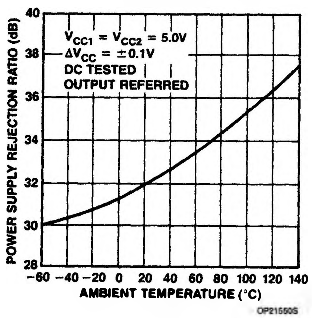

**Output Resistance vs Frequency**



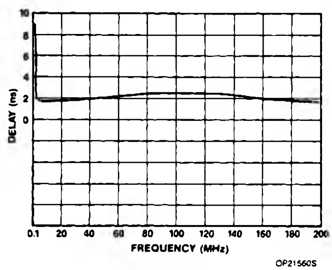

**Output Resistance vs Frequency**



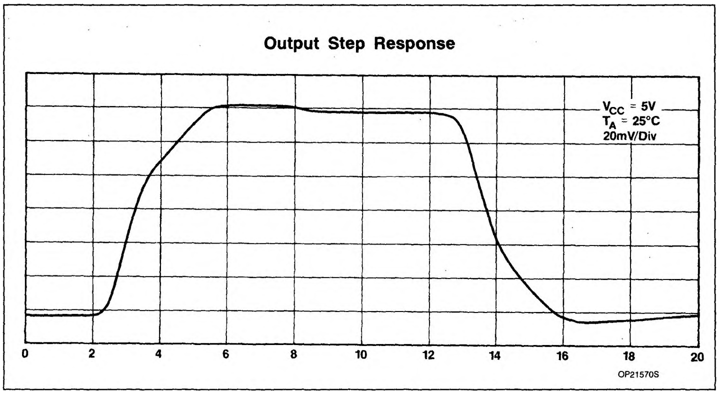

**NE5211 Power Supply Rejection Ratio vs Temperature**




**Group Delay vs Frequency**



## TYPICAL PERFORMANCE CHARACTERISTICS (Continued)




## THEORY OF OPERATION

Transimpedance amplifiers have been widely used as the preamplifier in fiber-optic receivers. The NE5211 is a wide bandwidth (typically 180MHz) transimpedance amplifier designed primarily for high sensitivity. The maximum input current before output stage clipping occurs at typically 50$\mu$A. The NE5211 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 wideband, 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 $Q_3$ is approximately the value of the feedback resistor, $R_F$ = 14.4k$\Omega$. 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

$$R_T = \frac{V_{OUT\ (diff)}}{I_{IN}} = 2R_F = 2(14.4k) = 28.8k\Omega.$$

The single-ended transresistance of the amplifier is typically 14.4k$\Omega$.

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, $A_{VOL}\approx 70$. The emitter follower $Q_2$ minimizes loading on $Q_1$. The transistor $Q_4$, resistor $R_7$, and $V_{B1}$ provide level shifting and interface with the $Q_{15}$–$Q_{16}$ differential pair of the second stage which is biased with an internal reference, $V_{B2}$. The differential outputs are derived from emitter followers $Q_{11}$–$Q_{12}$ which are biased by constant current sources. The collectors of $Q_{11}$–$Q_{12}$ are bonded to an external pin, $V_{CC2}$, in order to reduce the feedback to the input stage. The output impedance is about 17$\Omega$ single-ended. For ease of performance evaluation, a 33$\Omega$ resistor is used in series with each output to match to a 50$\Omega$ test system.

## 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 4pF, 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:

$$R_{IN} = \frac{V_{IN}}{I_{IN}} = \frac{R_F}{1 + A_{VOL}} = \frac{14.4k}{71} = 203\Omega.$$

More exact calculations would yield a value of 200$\Omega$.

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 $R_F$ = 14.4k$\Omega$, $R_{IN}$ = 200$\Omega$, $C_{IN}$ = 4pF:

$$f_{-3dB} = \frac{1}{2\pi\ 4pF\ 200} = 200MHz.$$

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, $R_{IN}$ = 200$\Omega$ then the total input capacitance, $C_{IN}$ = (1 + 4)pF which will lead to only a 20% 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 important. The input stage achieves a low input referred noise current (spectral density) of 1.8pA/$\sqrt{Hz}$. 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 41nA in a 200MHz bandwidth for $C_S$ = 1pF

## DYNAMIC RANGE

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

Electrical dynamic range, $D_E$, in a 200MHz bandwidth assuming $I_{INMAX}$ = 60$\mu$A and a wideband noise of $I_{EQ}$ = 41nA$_{RMS}$ for an external source capacitance of $C_S$ = 1pF.

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

$$= 20\ \log \frac{(60 \times 10^{-6})}{(\sqrt{2}\ 41 \times 10^{-9})}$$

# Transimpedance Amplifier (180MHz) NE/SA5211

$$= 20\log\frac{(60\mu A)}{(58nA)} = 60dB.$$

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

For a given wavelength λ;

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

Where h = Planck's Constant = $6.6 \times 10^{-34}$ Joule sec.
c = speed of light = $3 \times 10^{8}$ mt/sec
c/λ = optical frequency

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

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

Where η = quantum efficiency

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

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

where e = electron charge = $1.6 \times 10^{-19}$ Coulombs

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

I = P·R

Assuming a data rate of 400 Mbaud (Bandwidth, B = 200MHz), the noise parameter Z may be calculated as:[1]

$$Z = \frac{i_{eq}}{qB} = \frac{41 \times 10^{-9}}{(1.6 \times 10^{-19})(200 \times 10^{6})} = 1281$$

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:

$$P_{avMIN} = 12\frac{hc}{\lambda} B\ Z = 12\ 2.3 \times 10^{-19}$$

$$200 \times 10^{6}\ 1281 = 707nW = -31.5dBm,$$

where h is Planck's Constant, c is the speed of light, λ is the wavelength. The minimum input current to the NE5210, at this input power is:

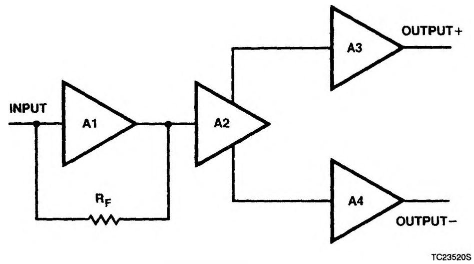



**Figure 1. NESA5211 Block Diagram**

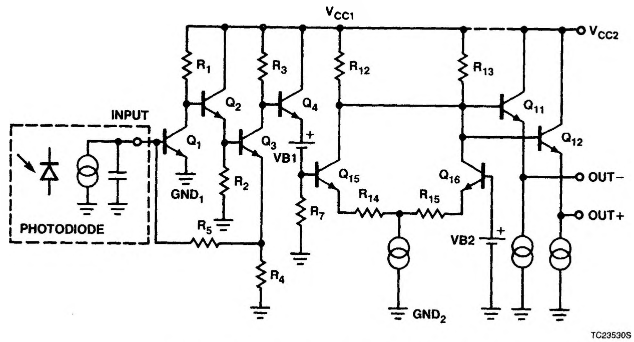



**Figure 2. Trans-Impedance Amplifier**

$$I_{avMIN} = qP_{avMIN}\frac{\lambda}{hc}$$

$$= \frac{707 \times 10^{-9} \times 1.6 \times 10^{-19}}{2.3 \times 10^{-19}}$$

$$= 492nA.$$

Choosing the maximum peak overload current of $I_{avMAX} = 60\mu A$, the maximum mean optical power is:

$$P_{avMAX} = \frac{hc\ I_{avMAX}}{\lambda\ q} = \frac{2.3 \times 10^{-19}}{1.6 \times 10^{-19}} 60 \times 10^{-6}$$

$$= 86mW \text{ or } -10.6dBm.$$

Thus the optical dynamic range, $D_o$ is:

$$D_o = P_{avMAX} - P_{avMIN} = -31.5 - (-10.6) = 20.8dB.$$

This represents the maximum limit attainable with the NE5211 operating at 200MHz bandwidth, with a half mark/half space digital transmission at 820nm wavelength.

## APPLICATION INFORMATION

Package parasitics, particularly ground lead inductances and parasitic capacitances, can significantly degrade the frequency response. Since the NE5211 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

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

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.

Figure 4 depicts a 50Mb/s TTL fiber-optic receiver using the BPF31, 850nm LED, the NE5211 and the NE5214 post amplifier. For more information on this circuit, please refer to Application Brief AB1432.

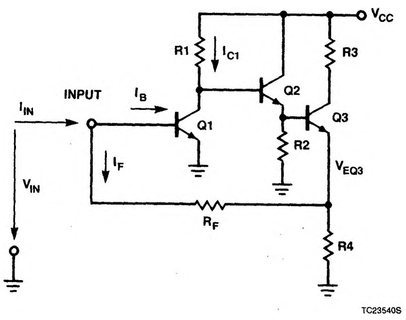



**Figure 3. Shunt-Series Input Stage**

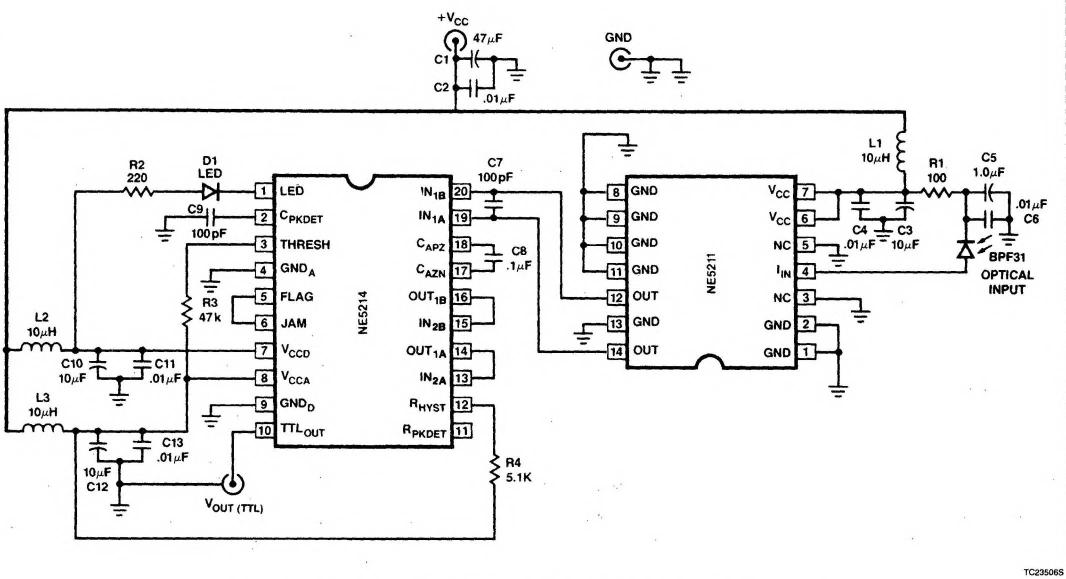



**Figure 4. A 50Mb/s TTL Fiber-Optic Receiver Using NE5210/NE5214**