# APPLICATION NOTE

**AN1435**

# A family of wideband low noise transimpedance amplifiers

author

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## INTRODUCTION

Despite numerous advantages, the relatively high cost of fiber optic transmission prevented its wide-spread industrial acceptance. High bandwidth-distance products, a prerequisite for cost-effectiveness, could not be achieved with relatively inexpensive components. The latest technological advances on both transmitter and receiver sides, however, are about to change that.

## Transmitter

Starting at the transmitter side (Figure 1), the two major problems of the past were the lack of inexpensive, light emitting diode (LED) transmitters, capable of 10-20MHz modulation rates, and the compounded problem of cost and reliability of laser diodes, required for large channel capacity, single mode, long-distance systems.

In examining the present status of the fiber optic industry we observe, however, that new generations of LEDs, used in most short range, multi-mode transmitters, can achieve wide modulation bandwidths, enabling system designers to develop cost-effective systems. For example, commercially available 820-820nm AlGaAs surface emitting devices have significantly decreased in price and can be used up to and beyond 100MHz (200MBaud). InGaAsP LEDs can be used in the 1.3μm range. Their highly doped versions can be modulated up to bandwidths of several hundred MHz at the expense of lower output power.

InGaAsP laser diodes can go well beyond 1GHz. Their higher output power and an order of magnitude narrower spectral widths make these devices the ideal choice for long-range, very high data rate telecommunication systems.

## Receiver

The key to cost effectiveness at the receiver side is the ability to offer monolithic IC building blocks that can match those high transmitter data rates with bandwidth, large dynamic range and low noise. These kinds of IC building blocks weren't readily available in the past. Consequently, system designers had to choose between limiting data rates to below 20MBaud or using costly hybrid modules.

Philips Semiconductors solution to the problem is the introduction of a family of transimpedance amplifiers (TiA). These are the NE5210, NE5211 and NE5212.

Although the real meaning is different, "transresistance" and "transimpedance" are, in practice, used interchangeably. These names designate that these types of amplifiers are current-driven at their inputs and generate voltage at their outputs. The transmitter function is, therefore, a ratio of output voltage to input current with dimensions of ohms. Since the input is current driven, the input resistance must be low, which means low input voltage- swings, no capacitive charge/discharge currents and wide frequency response with a generous phase margin. Alternative approaches to the TIA, such as high input impedance FET preamplifiers with a shunt input resistor, tend to be more bandwidth limited. They exhibit integrating characteristics, and therefore must be equalized by a differentiating second stage to achieve broad frequency response. The integrating input stage, however, is prone to overload with signals that have high low-frequency content. If the amplifier overloads for any reason, the integrated waveform cannot be restored by differentiation and dynamic range suffers despite the low noise characteristics.

Since the transimpedance configuration does not have this problem, its superior dynamic range, inherently large bandwidth and compatibility with low cost IC technologies make it an attractive approach.

## Transimpedance Amplifier Family

The NE5210, NE5211 and NE5212 TIA is a low noise, wide band integrated circuit with single signal input and differential outputs, ideally suited for fiber optic receivers, both digital and analog, in addition to many other RF applications. Table 1 depicts the differences between the three amplifiers. As shown in Figure 2, a differential output configuration was chosen to achieve good power supply rejection ratio and to provide ease of interface with ECL type postamplifier circuitry. The input stage (A1) has a low noise shunt-series feedback configuration. The open loop gain of A1 ($R_F$ = infinite) is about 70; therefore, we can assume with good approximation an input stage transresistance equal to the value of $R_F$. Since the second stage differential amplifier (A2) and the output emitter-followers (A3 and A4) have a voltage gain of about two, the input to output transresistance is twice the value of $R_F$. The single-ended transresistance is half of this value.

Returning to the input stage (Figure 3), a simple analysis can be used to determine the performance of the TIA. The input resistance, $R_{IN}$, can be calculated as:

$$R_{IN} = \frac{V_{IN}}{I_{IN}} = \frac{R_F}{1 + A_{VOL}}$$

For the NE5210: $R_{IN} \cong \dfrac{3.6k}{1 + 70} = 60\Omega$

For the NE5211: $R_{IN} \cong \dfrac{14.4k}{1 + 70} = 200\Omega$

For the NE5212: $R_{IN} \cong \dfrac{7.2k}{1 + 70} = 110\Omega$

Typical input capacitance of the TIA, $C_{IN}$, are 7.5pF, 4pF and 10pF for NE5210, NE5211 and NE5212, respectively.

$$f_{-3dB} = \frac{1}{2\pi R_{IN}\, C_{IN}} \begin{aligned} &= 350MHz \ (NE5210) \\ &= 200MHz \ (NE5211) \\ &= 145MHz \ (NE5212) \end{aligned}$$

Thus, while neglecting driving source and stray capacitances, $R_{IN}$ and $C_{IN}$ will form the dominant hole of the entire amplifier: Although significantly wider bandwidths could have been achieved by a cascade input stage configuration, the present solution has the advantage of a very uniform, highly desensitized frequency response because the Miller-effect dominates over external photodiode and stray capacitances. Consequently, the NE5210, NE5211, NE5212 will be relatively insensitive to PIN photodiode source capacitance variations. Since the dominant pole of the amplifier is at the input node, PIN diode source capacitance will not degrade phase margin.

## Package Parasitics

Package parasitics, particularly ground-lead inductances, can significantly degrade frequency response. To minimize parasitics, multiple grounds are used in order to minimize ground wire-bond inductances.

Further bandwidth modifications can be achieved by a small capacitance between input and output or input and ground. Since each of the NE5210, NE5211 and NE5212 has differential outputs, both peaking and attenuating type frequency response shaping are possible.

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**Table 1.  Wideband Transimpedance Amplifier Family**

| PART | DIFFERENTIAL TRANSRESISTANCE k$\Omega$ (typ.) | BANDWIDTH -3dB (typ.) | INPUT NOISE CURRENT (typ.) | MAX. INPUT CURRENT |
|---|---|---|---|---|
| NE5212 | 14 | 140MHz | 2.5pA/$\sqrt{Hz}$ | ±120µA |
| NE5211 | 28 | 180MHz | 1.8pA/$\sqrt{Hz}$ | ±60µA |
| NE5210 | 7 | 280MHz | 3.5pA/$\sqrt{Hz}$ | ±240µA |

## Fighting Noise

Since most currently installed and planned fiber optic systems use non-coherent transmission and detect incident optical power, receiver noise performance becomes important.  The NE5210, NE5211 and NE5212 go a long way towards solving this problem.  Their input stage configurations achieve a respectably low input referred noise current spectral density of 3.5pA/$\sqrt{Hz}$ for the NE5210, 1.8pA/$\sqrt{Hz}$ for the NE5211 and 2.5pA/$\sqrt{Hz}$ for the NE5212, measured at 10MHz.  This low value is nearly flat over the entire bandwidth.  The transresistance configuration assures that the external high value bias resistors, often required for photodiode biasing, will not contribute to total system noise.  As shown in the following equation, the equivalent input $_{RMS}$ noise current is determined by the quiescent operating point of $Q_1$, the feedback resistor, $R_F$, and the bandwidth, $\Delta f^*$, however, it is not dependent on the internal Miller-capacitance.  The noise current equation is then

$$\overline{i_{eq2}} = 4kT\,\frac{\Delta f}{R_F}\ + 2q\,I_{BQ1}\,\Delta f$$
$$+\ 2q\,I_{CQ1}\,\frac{1}{g_m^2}\,\omega^2\,(C_S + C_{\pi 1})^2\,\Delta f$$
$$+\ 4kT\,r_{bQ1}\,\omega^2\,C_S^2\,\Delta f$$

The resulting integrated noise over 100MHz with $C_S = 1pF$ is
> 40nA for NE5210
> 21nA for NE5211
> 32nA for NE5212

## Testing the NE5212

*The remaining portion of this paper deals specifically with the NE5212 and is directly applicable to the NE5211 and the NE5210.*

Connecting the NE5212 in an actual fiber optic preamplifier configuration, dynamic range, transient response, noise and overload recovery tests are easily measured (Figure 4).  In order to replicate actual parasitic capacitances, effects of the photodiode bias network and circuit layout effects, the test circuit should closely resemble the real application conditions.  If the intention is to use the device in die form, then the actual hybrid circuit mounting techniques should be used while testing.

In the test circuit shown, an 850nm modulated laser light source feeds an HP-HFBR2202 PIN photodiode which is mounted in close proximity to the NE5212 input.  The RC filter in series with the photodiode eliminates possible disturbances from the power supply.  Both differential outputs are AC coupled through 33$\Omega$ resistors in order to match to the 50$\Omega$ test system.  In most applications these matching resistors are unnecessary.  Performance evaluation in the linear region, including amplitude and phase response and power supply rejection, can be accomplished by a network analyzer and S parameter test set (Figure 5).  The simple equations given in the figure for the calculation of transresistance, $R_T$, are accurate for R>>$R_{IN}$, where $R_{IN}$ is the input resistance of the NE5212.

## General Purpose RF Applications

Besides the main fiber optic receiver applications, many other interesting possibilities exist for the NE5212.  Simplicity and ease-of-use are the prevailing characteristics of this device.  For instance, amplifiers with 20dB gain can be built requiring only one external gain setting resistor (Figure 6).  The voltage gain of the differential configuration with no load at the outputs can be calculated as follows:

$$V_{OUT} = I_{IN} \times R_T = \frac{V_{IN}}{R_S + R + R_{IN}}\quad R_T\ \text{and}$$
$$A_V = \frac{V_{OUT}}{V_{IN}} = \frac{R_T}{R_S + R + R_{IN}}$$

where $R_S$ is the signal-source resistance, R is the external gain setting resistor and $R_{IN}$ is the input resistance of the NE5212. Substituting the actual values:

$$A_V = \frac{14000}{R_S + R + 110}$$

where all values are in ohms.  The graph of Figure 6 is an experimental verification of this formula in a single-ended, 50$\Omega$ system, using the test configuration of Figure 5.  Note the 6dB loss due to the single-ended configuration and another 6dB due to the 50$\Omega$ load.  As in all other RF applications, attention to power supply bypassing clean grounds and minimization of input stray capacitances are required for optimum performance.

Another useful application of the NE5212 is as a voltage controlled amplifier, using a DMOS FET device biased into the linear region (Figure 7).  An operational amplifier with supply-to-ground output swing and supply-to-ground input common mode range (such as the Philips Semiconductors NE5230) can provide adequate gate control voltage even with a single 5V power supply.  This type of circuit can have 25dB AGC range at 50MHz and 45dB at 10MHz with less than 1% harmonic content.  AGC range is determined by the ON resistance range of the FET and capacitive drain to source feedthrough.  If lowest RF feedthrough were required, the FET should be used in a shunt configuration rather than in a series.

Turning towards an entirely different area of application, where contrary to the NE5212's capabilities, poor phase margins are mandatory, a simple crystal oscillator with buffered output can be built using a minimum number of external components (Figure 8).  The feedback signal is taken from the non-inverting output, while the inverting output provides a low impedance (15$\Omega$) output drive.  The crystal operates in its series resonance mode.  Figure 9 shows a varactor tuned version with a large tuning range.  In Figure 10 the circuit has been optimized for stability at the expense of tuning range.

In RF amplifier applications it is often desirable to limit the amplifier bandwidth in order to minimize noise and RFI.  The 100-150MHz bandwidth of the NE5212 can be easily modified by connecting a capacitor to the input pin.  The device bandwidth then becomes
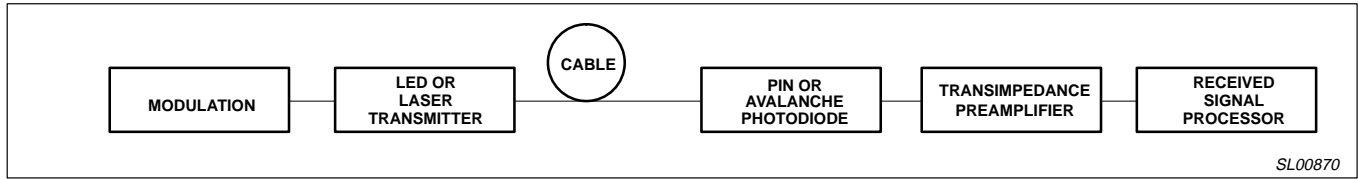
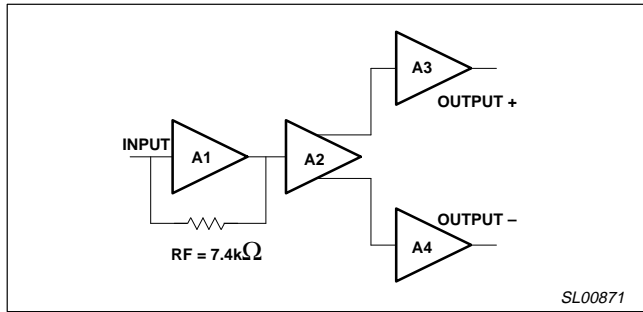**Figure 1.    Fiber Optic Transmission System Using Transresistance Preamplifier**



**Figure 2.    NE5212 Block Diagram Showing the Transresistance Preamplifier with the Internal Feedback Resistor and the Buffered Differential Outputs.**

$$f_{-3dB} = \frac{1}{2\pi \, R_{IN} \, (C_{IN} + C_{EXT})}$$

where $R_{IN}$ is the input resistance, $C_{IN}$ is the input capacitance as specified in the data sheet and $C_{EXT}$ is the external capacitance. For example, a $C_{EXT} = 33pF$ will reduce the amplifier bandwidth to 42MHz with a single pole roll-off. The penalty is an increase in noise current. The transfer curve is shown in Figure 11. Another way to limit the bandwidth is to connect a capacitor across the differential output. Single-ended to differential conversion is another useful application for the device. Impedance matching is easily accomplished by resistors connected in series with the outputs.



$$A_{VOL} \cong \frac{R_1 \parallel R_{IN\,(Q2)}}{v_T \, / \, I_{EQ1}} = 70 = \frac{v_{EQ3}}{v_{IN}}$$
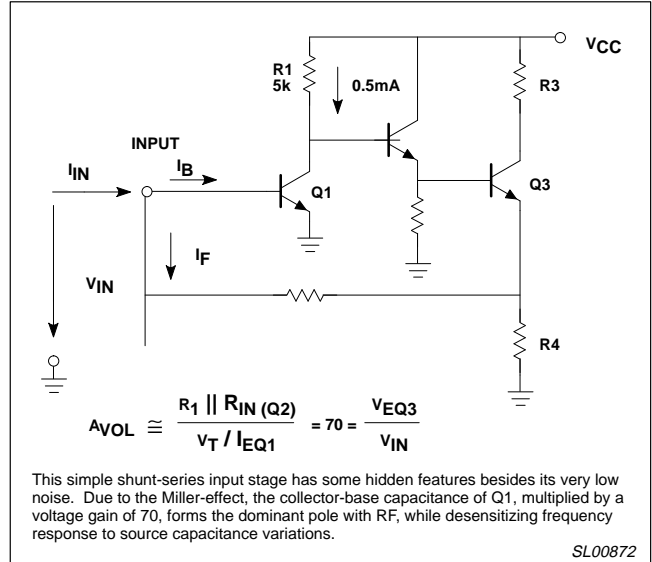
This simple shunt-series input stage has some hidden features besides its very low noise. Due to the Miller-effect, the collector-base capacitance of Q1, multiplied by a voltage gain of 70, forms the dominant pole with RF, while desensitizing frequency response to source capacitance variations.

**Figure 3.**

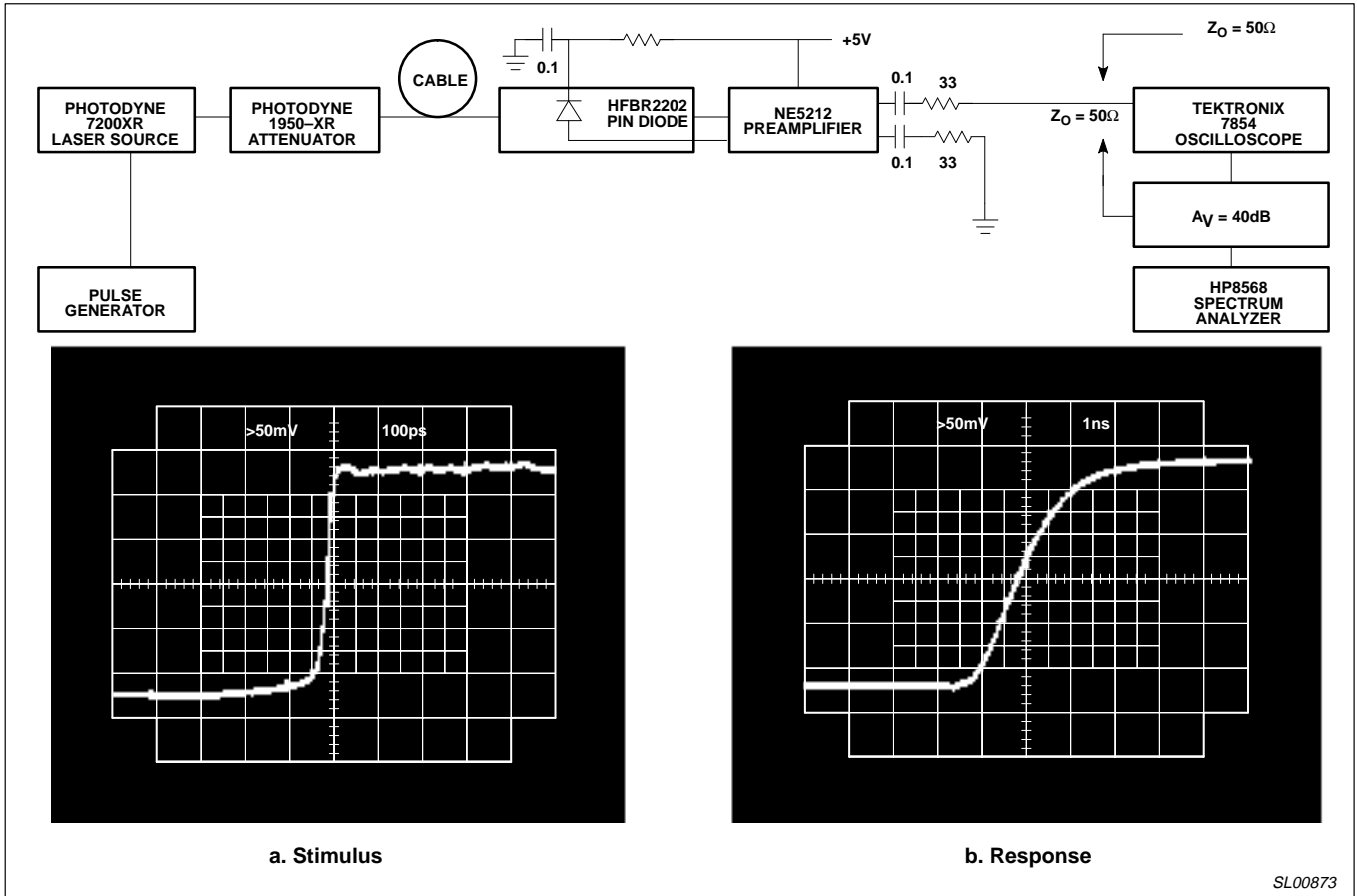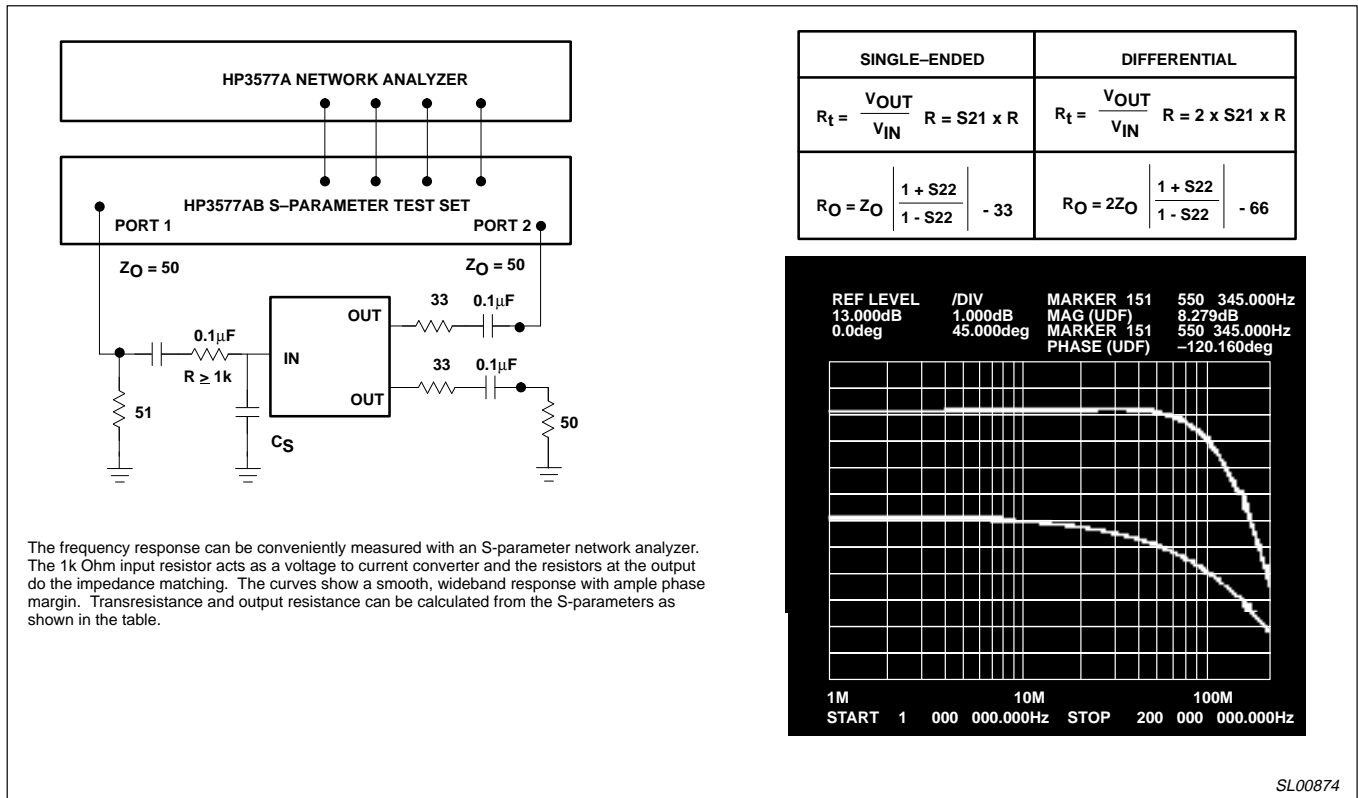# A family of wideband low noise transimpedance amplifiers

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**Figure 4.   Dynamic Range, Impulse Response and Noise Test Circuit**

| SINGLE–ENDED | | DIFFERENTIAL | |
|---|---|---|---|
| $R_t = \dfrac{V_{OUT}}{v_{IN}}$ | $R = S21 \times R$ | $R_t = \dfrac{V_{OUT}}{v_{IN}}$ | $R = 2 \times S21 \times R$ |
| $R_O = Z_O \left\lvert \dfrac{1 + S22}{1 - S22} \right\rvert - 33$ | | $R_O = 2Z_O \left\lvert \dfrac{1 + S22}{1 - S22} \right\rvert - 66$ | |

The frequency response can be conveniently measured with an S-parameter network analyzer. The 1k Ohm input resistor acts as a voltage to current converter and the resistors at the output do the impedance matching. The curves show a smooth, wideband response with ample phase margin. Transresistance and output resistance can be calculated from the S-parameters as shown in the table.

SL00874

**Figure 5.**



a. **Differential Amplifier**

b. **Non-Inverting Amplifier**

c. **Inverting Amplifier**

A 150MHz, 20dB amplifier requires only one external gain setting resistor. Differential or single-ended opeation with or without inversion is possible.

**Single-Ended Gain vs R**

SL00875

**Figure 6.**

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The variable gain RF amplifier has less than 1% distortion. 25dB range at 50MHz and 45dB at 10MHz, with a 0-1V control signal.

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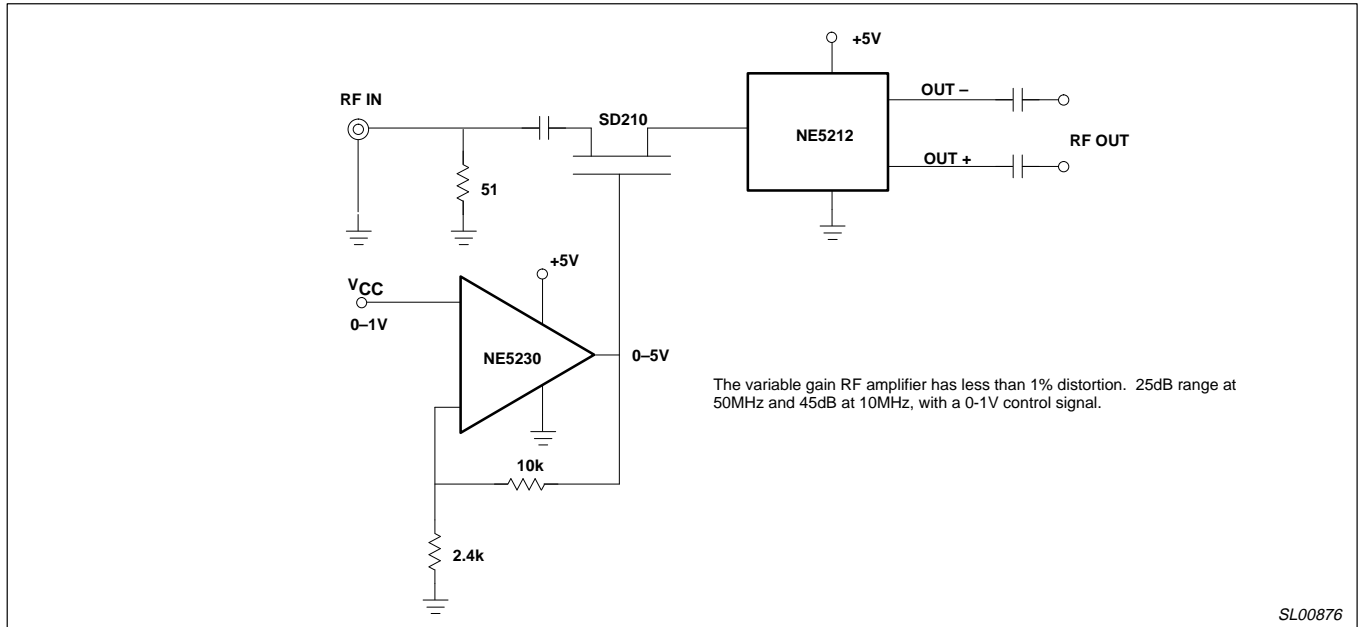
**Figure 7.**

SL00877
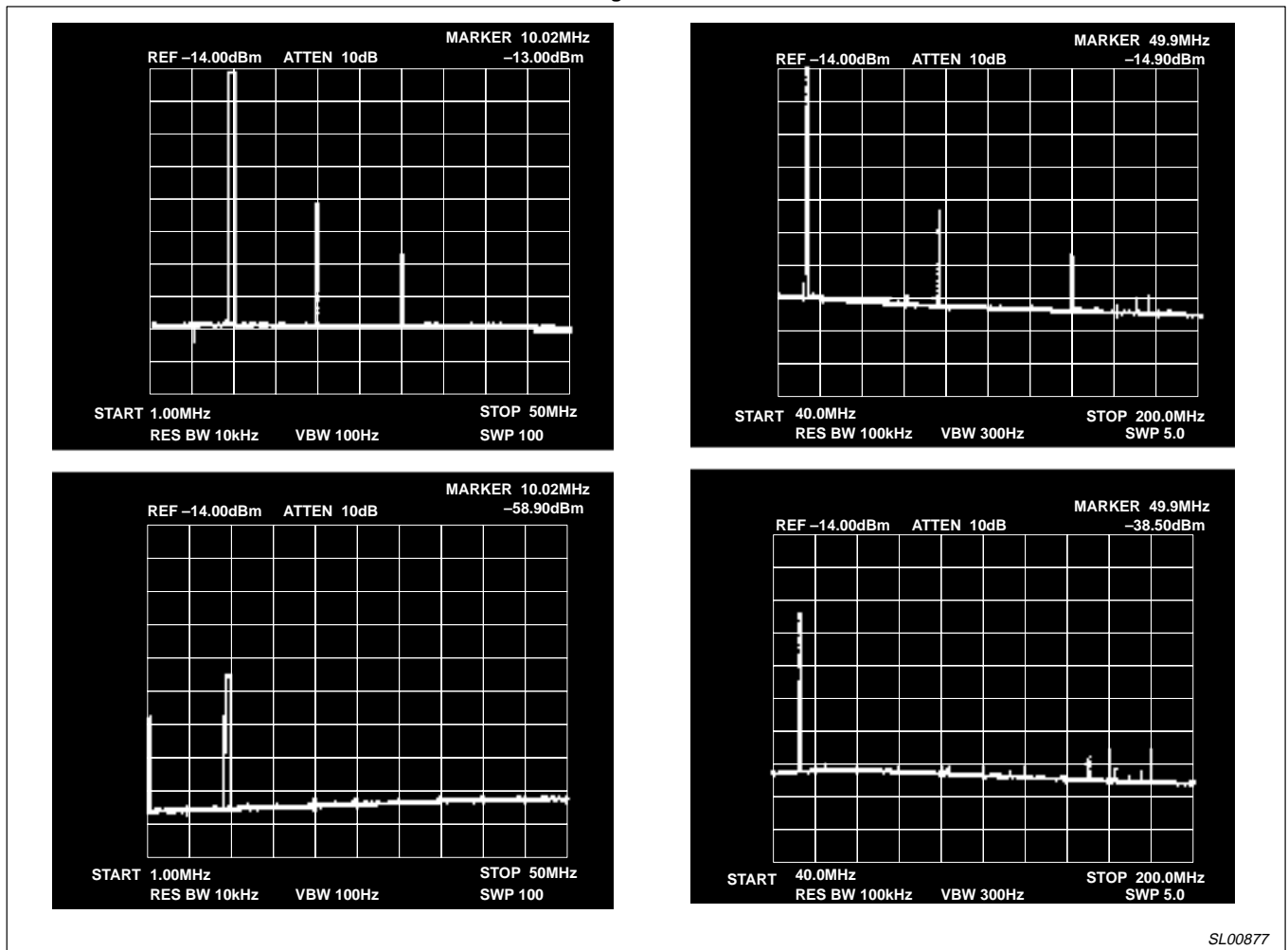
**Figure 8.    Performance Data of the RF Attenuator of Figure 7**

**FET resistance in its linear region at V$_{GS}$ = 0V (left) and at V$_{GS}$ = 5V (right).**

SL00878

**Figure 9.**



This 16MHz crystal oscillator has minimum component count and operates in the series resonant mode. Output provides independent 50 Ohm drive capability.
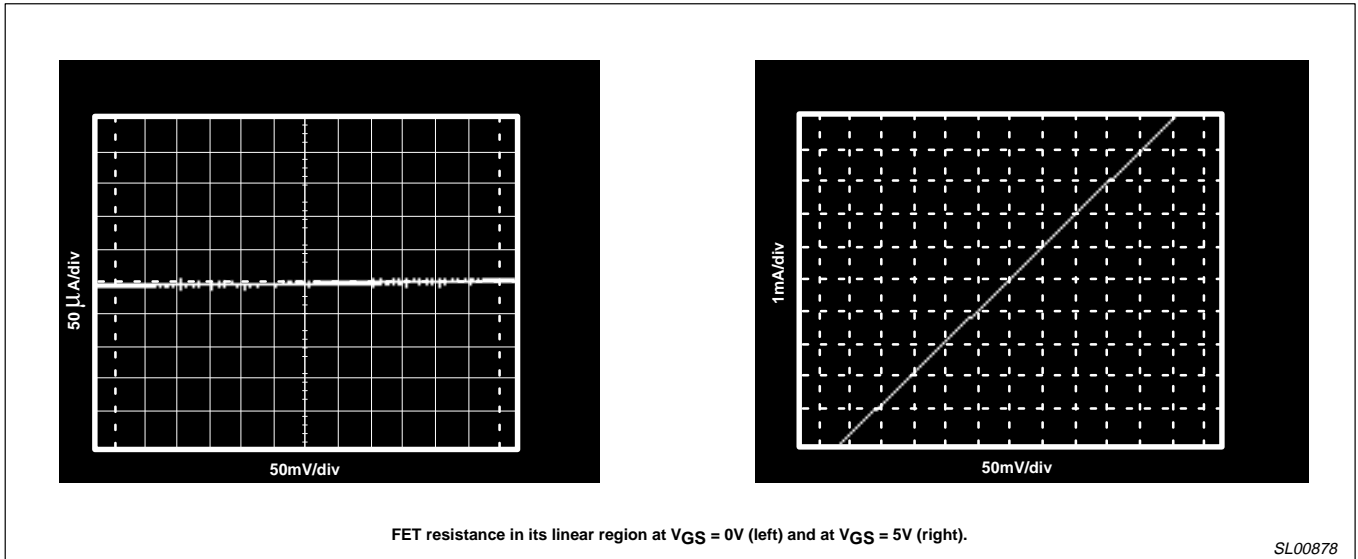
SL00879
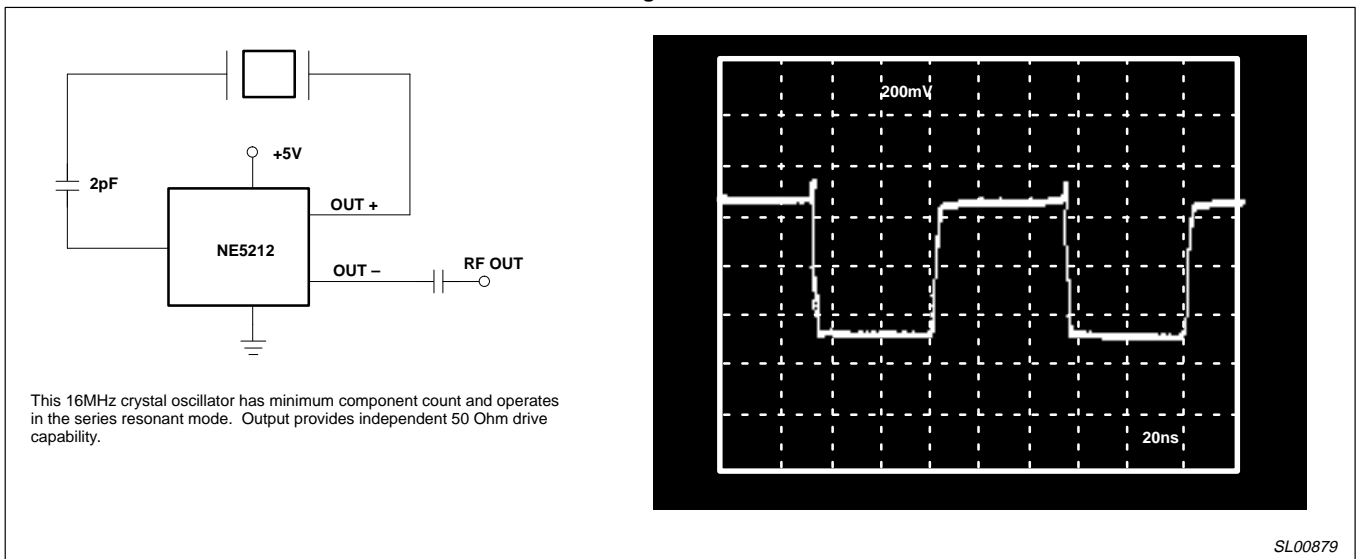
**Figure 10.**

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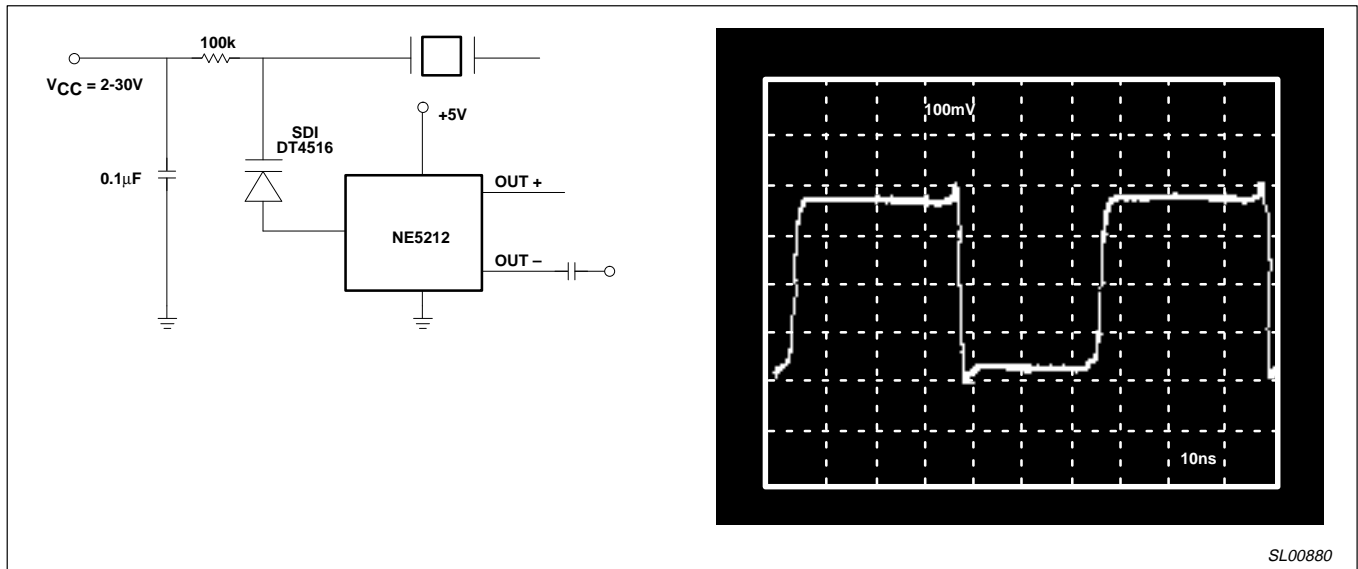
**Figure 11.   Varactor Tuned Crystal Oscillator with 0.28% Tuning Range**
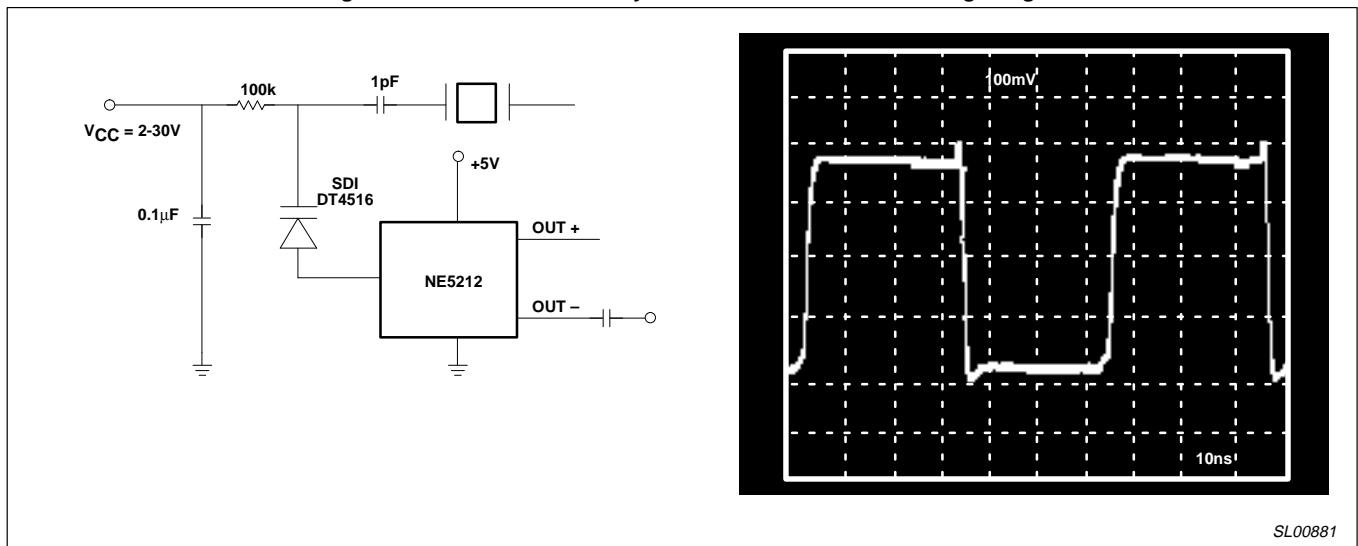


**Figure 12.   Varactor Tuned Oscillator with 1ppm/V Supply Sensitivity**

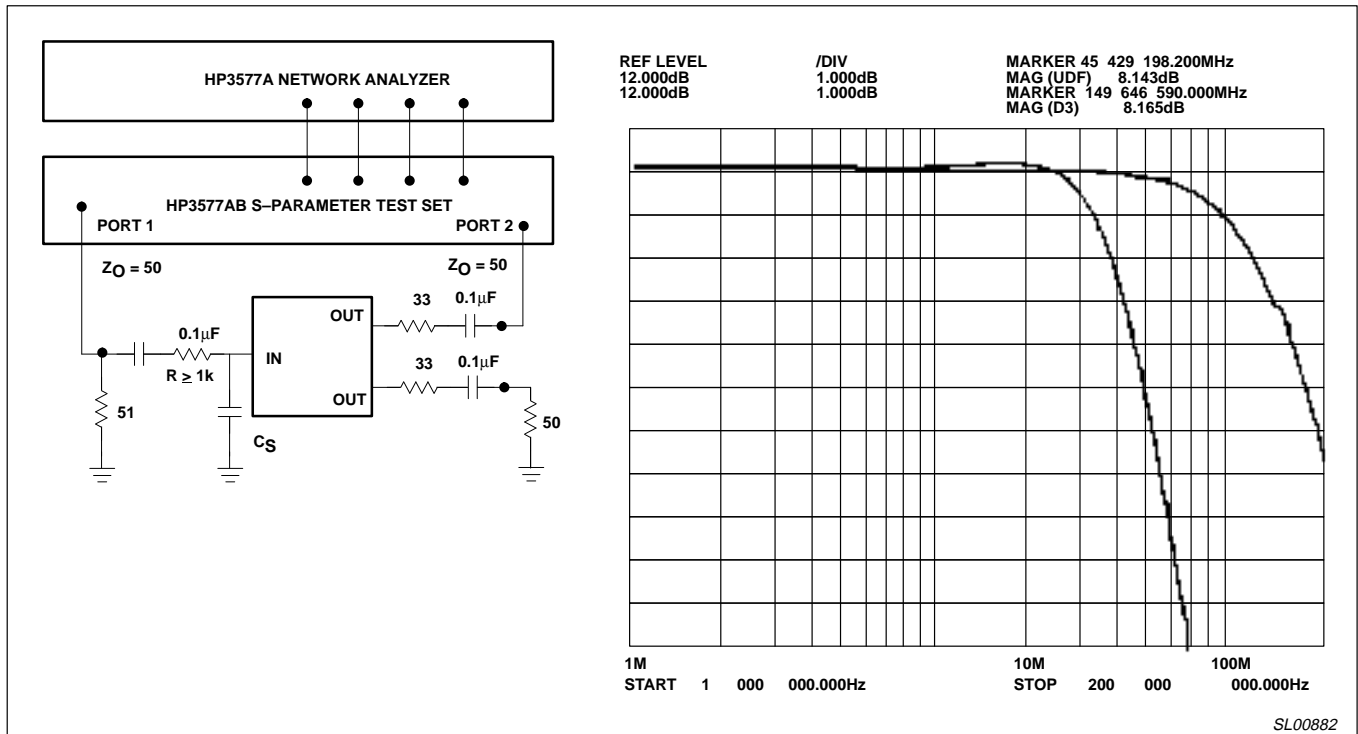# A family of wideband low noise transimpedance amplifiers

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**Figure 13.   Setting Bandwidth with Shunt Input Capacitor, C$_S$**