

Designing an RFIC Mixer

Fitting a mixer within an RFIC requires optimized tradeoffs.

Stephen Long, University of California, Santa Barbara

Designing a mixer to work within an RFIC is not an easy task. There are many different performance specifications for mixers, which must be prioritized based upon the application. Some are intended for receive applications where the input signal level is not under your control and the maximum linearity under large signal drive conditions is often more critical than the noise figure. On the other hand, for transmit applications where you have control of the signal level, the design strategy shifts to trading off noise and intermodulation distortion (IMD) behavior to achieve the largest useable dynamic range.

In this article, a transmit mixer illustrates many of the tasks required to design a quality RFIC. The application is an upconversion mixer intended for a base-station transmitter power amplifier (figure 1). It is based on a Gilbert-cell MOSFET double-balanced differential mixer with an input baseband or IF signal centered at 200 MHz

and an output of 1.8 GHz. It uses 0.35- μm MOSFETs with a default device model parameter set. For your application, you would have to substitute your own verified model for the simulations to correspond to measured mixer characteristics.

After the intrinsic mixer performance was evaluated, the

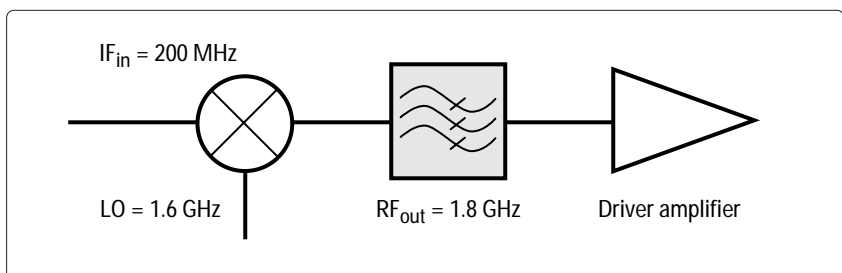


Figure 1. The upconversion mixer for the base-station transmitter power amplifier application

design was modified to improve conversion gain and image rejection by tuning the mixer output. Finally, a differential-to-single-ended converter was added to interface to an off-chip bandpass filter.

The mixer DesignGuide library was used as a starting point for the design exercise. The mixer DesignGuide is intended to enhance productivity of RF designers by providing an extensive collection of analysis tools that can be easily loaded into a project from a pulldown menu. Each pair of these analysis network and displays, which could take days to set up and verify, can be easily adapted for specific mixer circuit simulation requirements.

The initial design of the mixer was constructed as an Advanced Design System (ADS) subnetwork (figure 2). The mixer itself can be replaced with or modified to become your own design. You can declare any of the circuit parameters to be accessible outside of the subnetwork. In this example, V_{DD} , R_D (drain resistance), W_{csp} (current source control width), W_1 and W_2 (transconductance and switch MOSFET widths), R_S (source-degeneration resistance), and L_S (source-degeneration inductance) are all available for a parameter sweep.

Figure 3 shows a schematic of a MOSFET version of the Gilbert active double-balanced mixer. The lower FET differential pair serves as a transconductance amplifier. The upper FETs provide a fully balanced, phase-reversing current switch. A DC bias generator keeps the MOSFETs in their active region (not shown).

The large signal-handling capability of the mixer depends mainly upon the linearity of the transconductance amplifier,

and was measured by determining the maximum input voltage, V_{1dB} , (or power, P_{1dB} , in some cases) that causes a 1-dB compression in the conversion gain. The maximum linear input voltage range can be increased by increasing the source-degeneration resistors (R_S). Source inductance also can provide beneficial degeneration, but in this case there was a very low input IF frequency, 200 MHz. The inductance values required would be too large for RFIC implementation, so resistors

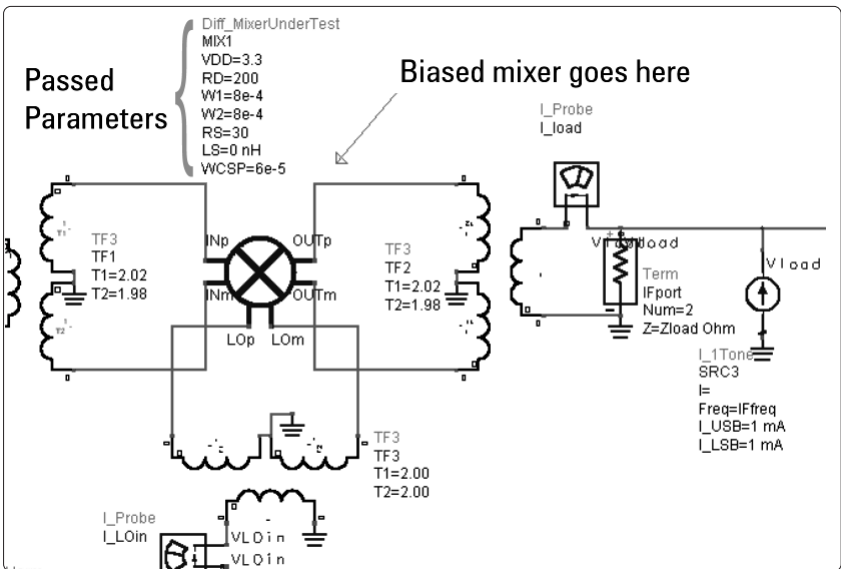
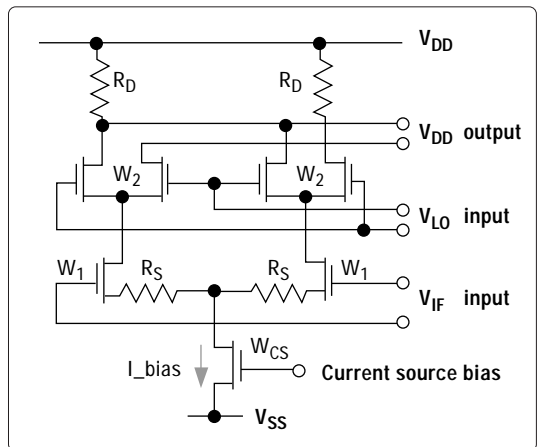


Figure 2. Differential mixer test bench example

Figure 3. The MOSFET Gilbert-cell active double-balanced mixer



must be used even though they add noise. The load resistors could also cause gain compression if the voltage swing at the drains is large enough to cause the output to clip under large signal drive conditions.

The double-balanced design rejects IF and LO feedthrough to the output if the output is taken differentially, because the LO component in the output is a common-mode signal and the RF output is differential.

Design Sequence

A mixer that is to be used for base-station transmit applications requires high linearity and low noise to minimize the amount of spurious power that is spread into the adjacent channel. This example mixer was optimized in the following sequence:

1) Determination of LO amplitude. The mixer commutating switch must be fully activated, as excess distortion can be produced with a weakly conducting or slowly activated switch. The conversion transducer gain and 1-dB gain compression input level were used to determine the sufficient LO voltage.

2) Evaluation of the influence of source and drain resistance on the 1-dB compression level, giving insight into the principal mechanisms that limit linearity.

3) Determination of how the added noise of the mixer affects the minimum signal level and thus limits dynamic range, so the tradeoff between noise, gain, and gain compression could be evaluated.

4) Evaluation of how the two-tone third-order IMD power and the noise figure affect the mixer dynamic range relative to the input voltage. As the designer has control over the input voltage for transmit applications, the optimum dynamic range—the mixer’s “sweet spot” for best performance—must be determined. If a fixed signal level is specified, the mixer must be designed to provide the best dynamic range at that signal level.

5) Finally, testing the mixer under a more realistic signal input, such as a CDMA source, to emulate a multicarrier environment. This is a more severe test than the two-tone IMD one, and is much more time consuming to simulate because a large number of symbols must be used for accurate results.

Once the basic resistively loaded Gilbert-cell mixer was characterized, two modifications were employed to improve performance. First, the mixer drain nodes were tuned with inductors and a capacitor for resonance at the output frequency. This improves conversion gain if inductors with reasonable Q_U can be fabricated. It also decreases the amplitude of the undesired output image because of its bandpass transfer function. The image must be removed anyway, and its presence can only degrade the distortion of the output stage by increasing the peak voltage present at its input.

The final stage was to convert the differential signal to a single-ended one. Because the output of the mixer must be filtered off-chip with a SAW filter before further amplification, a single-ended output is more efficient. It must have good common-mode rejection to suppress LO feedthrough, and good linearity so that it doesn’t degrade dynamic range.

Determining LO Voltage

The first step in designing this mixer was to determine a suitable LO voltage that provides a reasonable compromise between conversion gain and LO power and at the same time does not limit the 1-dB gain-compression input voltage. The MOSFETs forming the commutating switch (upper level) must be driven hard enough to present a low series resistance to the load. Many of the mixer analysis schematic and display templates available in the DesignGuide library include an LO power sweep capability. Also, an N-dB Gain Compression analysis could be used to evaluate the dependence of gain compression on LO drive.

These simulations showed that the input power at which gain compresses by 1 dB (P_{1dB}) does not closely depend on LO voltage, but conversion gain does depend somewhat on LO voltage (figure 4). The more gate voltage applied to the upper tier of MOSFETs, the lower their series resistance relative to the drain resistance and thus the higher the conversion gain. There was a conversion loss that became worse at the higher output RF frequency of 1.8 GHz, but this could be improved by tuning the RF output of the mixer.

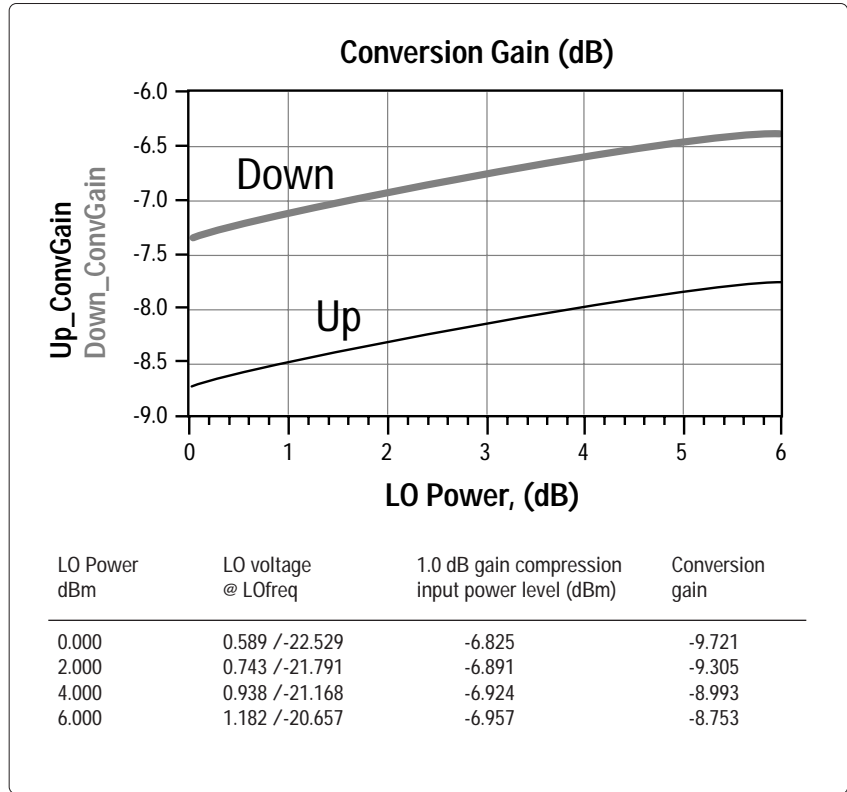


Figure 4. Simulations showing the dependence of the input power at which gain compresses by 1 dB and conversion gain upon LO voltage

Gain Compression Evaluation

Gain compression was evaluated using the N-dB Gain Compression Point analysis schematic. The 1-dB gain-compression input power and input voltage were found for swept parameters. For this example, the influence of R_S and R_D on V_{1dB} was determined. The R_S sweep used an R_D of 100 Ω , and the R_D sweep set used an R_S of 30 Ω . Conversion gain was measured at the 1-dB compressed level.

In an RFIC mixer where the input might not be matched to a source impedance, the input voltage is a more important metric of gain compression than the input available power (P_{1dB}) because available power assumes a conjugately matched source and load. Also, in a multi-signal environment, the peak input voltage can be quite large at the instant in time when all signals add in phase. It is this peak voltage that determines the distortion limits of the mixer. For example, two-tone IMD simulations predicted a

1-dB compression power that was 6 dB lower than predicted by single-tone simulations because the peak voltage was twice as high for the same power per tone.

It is also noteworthy that the conversion power gain depends inversely on R_D . In the simulation, the external load resistance was set to $2 R_D$ so that the output power (power absorbed in the load) was also the available output power, $P_{out} = V_{out}^2 / R_D$. The voltage gain would be expected to follow R_D/R_S , but increased less rapidly than anticipated, probably due to the output RC time-constant bandwidth limitations.

Determining Noise Tradeoffs

The next step was to evaluate how DC bias current (I_{bias}) and source resistance affect the mixer noise figure. The mixer single-sideband noise figure (SSB NF) was simulated as a function of DC bias current through the Gilbert cell (mixer core). The DC current was varied by sweeping the width of the PMOS current source (W_{csp}) and the mixer current mirror width (W_{cs}) using a parameter sweep.

SSB NF was appropriate because only one input frequency was applied to the mixer, but wideband noise at the image frequency and from LO harmonics was included in the signal-to-noise calculation. The simulation showed that the NF was reduced with increasing I_{bias} , but reached a point of diminishing returns. Thus, a width of $50 \mu m$ for the current source was selected as a compromise between power and noise.

The SSB NF was also found to be strongly dependent on the source resistance. This was expected because the thermal noise contributed by the resistor is directly in the input voltage loop of the differential pair. Thus, there needs to be a trade-off between V_{1dB} and NF to obtain the largest dynamic range of the mixer.

The dynamic range at low input signal power levels is limited by the carrier-to-noise ratio. The noise power for a minimum detectable signal ($S/N = 1$) depends on both NF and the noise bandwidth. This bandwidth is normally set by an external SAW filter between the

mixer and the driver amplifier. The filter is also required to reject the output difference ($F_{LO} - F_{in}$) image frequency at 1.4 GHz.

The conversion gain (a loss in this case) may also increase the noise figure because the drain-resistor thermal noise is input referred through the gain. Thus, a tuned output should be investigated to eliminate some of this noise.

Distortion Effects

At higher input signal levels, the dynamic range of the mixer is limited by distortion. The third-order IMD products are the most damaging because they show up in-band and cannot be rejected by the filter. A two-tone third-order IMD simulation with an RF power sweep was used to display the carrier-to-IMD power ratio. The IMD power present in the output increased at three times the rate of increase of input power. Thus, the difference between output power and IMD power shrinks with increasing input.

Dynamic Range versus Input Voltage

Determining the effect of input voltage on dynamic range required the output from two simulations: IMD RF power sweep and the SSB NF (figure 5). The dynamic range is controlled by the least of these two conditions:

$$DR = P_{out} \text{ (dBm)} - MDS \text{ (dBm)}$$

(noise-limited for low input levels)

$$DR = P_{out} \text{ (dBm)} - P_{IMD} \text{ (dBm)}$$

(distortion-limited for higher input levels)

R_s	DR (dB)	V_{in} (V) (differential)	NF (dB)
10	57.7	0.017	6.5
20	57.3	0.025	8
30	56.4	0.031	9.2
40	56.0	0.039	10.3

The dynamic-range peak depends on the noise bandwidth. For narrower bandwidths, the noise floor (MDS) drops and the peak dynamic range increases but shifts to lower differential input voltage. The 30-MHz noise bandwidth was chosen because of the base-station application. The transmitter should be capable of covering an entire frequency band.

Tuning Mixer Drain Nodes

The low conversion gain of the resistively loaded mixer caused higher noise due to the drain resistors. By resonating the output at 1.8 GHz, the conversion gain was increased and the gain at the image (1.4 GHz) was reduced. A comparison between the resistively loaded case and the tuned case showed an increase in conversion gain of about 3.5 dB.

You can perform an RF frequency sweep to find the resonant frequency of your design. From that, you can calculate how

much capacitance is contributed by the drain-to-substrate junction and absorb it into the resonator.

Gain Reduction Due to Inductor Q

On-chip inductor Q is limited by metal losses and substrate conduction in bulk silicon processes. An ordinary digital IC process produces low Q_U in spiral inductors. CMOS or BiCMOS RFIC processes can achieve higher Q inductors by using thicker dielectrics and thicker metal. Q values in the range of 5 to 15 are typical.

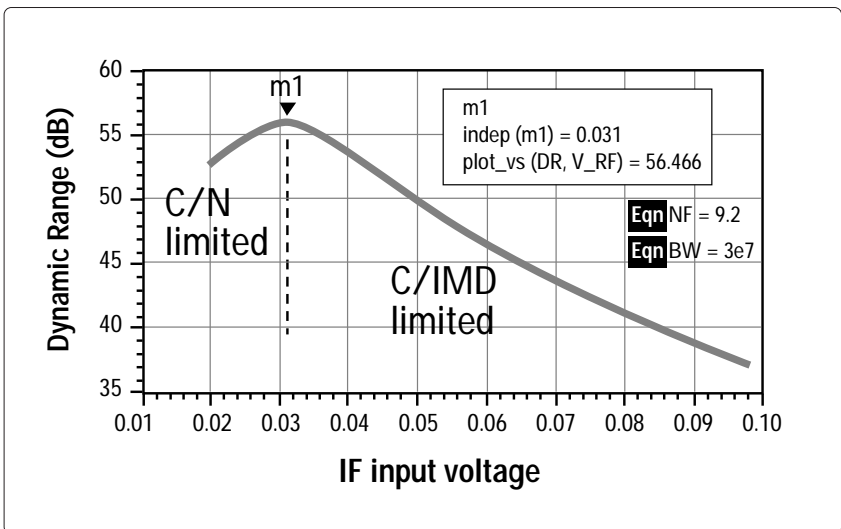


Figure 5. The effect of input voltage on dynamic range

Design Evaluation

For the initial design evaluation, it was easier to continue to measure the differential output so that tradeoffs and comparisons could be made between the differential tuned mixer and the mixer with output buffer. Once the design was complete, the mixer could then be evaluated in a single-ended configuration.

The SSB NF simulation was performed again with parameter sweeps for R_S and R_{ind} . Figure 7 shows that there is little noise sensitivity to R_{ind} ; however, it strongly affects the conversion gain. R_S affects both NF and conversion gain and also the carrier-to-IMD ratio versus IF input voltage. The mixer TOI/IMD simulation was performed again for an R_S of 10, 20, and 30 Ω . The dynamic range slowly improves for smaller R_S , but is very dependent on the noise bandwidth.

To speed up the process, a DesignGuide schematic intended for evaluation of single-ended mixers was copied from the menu and modified (figure 8).

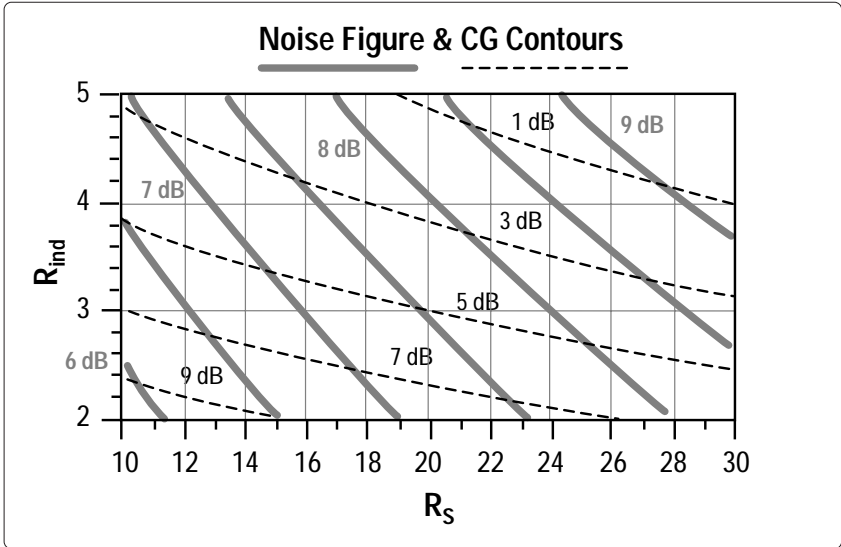


Figure 7. Noise-figure and conversion-gain contours

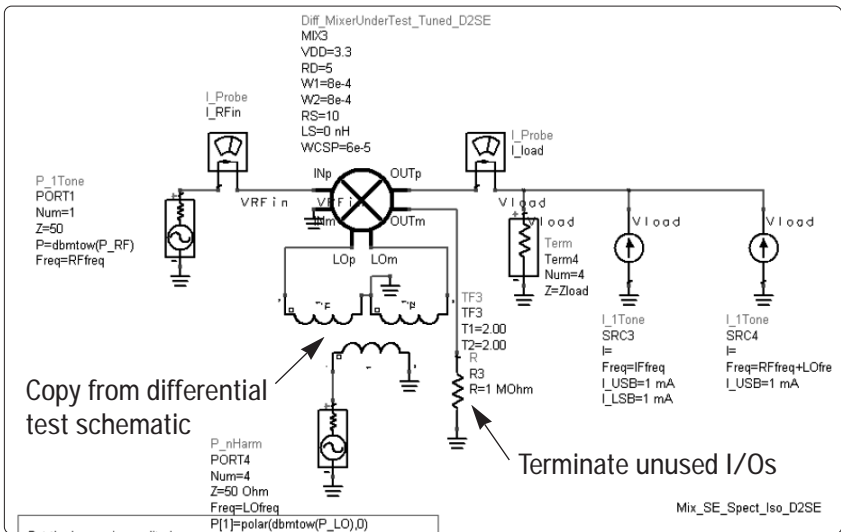


Figure 8. A single-ended mixer modified to evaluate the designed mixer

The tuned mixer with the D2SE output stage was then inserted from the component library. Unused inputs were terminated; the input was grounded and the output terminated in a large

resistance. The LO was differential. A transformer and source were copied from a differential test schematic and pasted into this schematic. An active LO single-ended-to-differential stage could also be designed and added to the mixer.

Again, NF and IMD-versus-RF power sweeps were performed for a range of R_S values from 10 to 30 Ω . These were combined to determine dynamic range (figure 9). An R_S of 10 Ω produced the best result: a peak dynamic range of 57.5 dB at an input voltage of 14 mV.

Simulation with a digital signal source such as CDMA is a more severe test of linearity. An IS-95 CDMA source with very good ACPR was used to drive the mixer input. When the input RF signal level was set to the optimum value for mixer dynamic range, relatively little spectral regrowth was observed.

The input of the mixer is badly mismatched, which may not be of much concern if the base-band and IF driver circuits are on the same chip with the upconversion mixer. In that case, the voltage levels are of greater interest. If the mixer is driven from off chip, the input impedance will be dominated by capacitive reactance. A

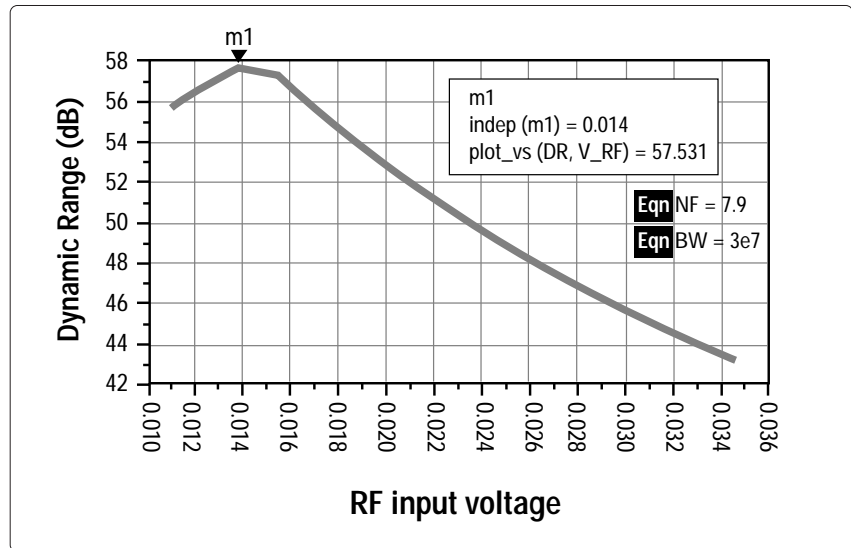


Figure 9. The dynamic range peaked at 57.5 dB.

matching network could increase the conversion gain significantly. The mixer DesignGuide contains several impedance-matching utilities that could be used to design and evaluate matching networks.

This study of optimizing an RFIC upconversion transmit mixer shows how design and analysis tools can help you evaluate the performance of your design and determine ways to improve it.

To get product information related
to this topic, visit our Web site at
www.agilent.com/find/eesof

Visit Agilent's library of application notes,
training courses, FAQs, tutorials, and more at
www.agilent.com/find/test

By internet, phone, or fax, get assistance
with all your test & measurement needs

Online assistance:
www.agilent.com/find/assist

Expanded from original article in Agilent
Measurement Solutions Volume 1. Issue 1

© Agilent Technologies, Inc. 2001
Printed in USA November 1, 2001
5988-4539EN



Agilent Technologies