2 Nanosecond, .1% Resolution Settling Time Measurement for Wideband Amplifiers

Quantifying Quick Quiescence

Jim Williams

INTRODUCTION

Instrumentation, waveform synthesis, data acquisition, feedback control systems and other application areas utilize wideband amplifiers. Current generation components (see box section, page 2, "A Precision Wideband Amplifier with 9ns Settling Time") feature good DC precision while maintaining high speed operation. Verifying precision operation at high speed is essential, and presents a high order measurement challenge.

SETTLING TIME DEFINED

Amplifier DC specifications are relatively easy to verify. Measurement techniques are well understood, albeit often tedious. AC specifications require more sophisticated approaches to produce reliable information. In particular, amplifier settling time is extraordinarily difficult to determine. Settling time is the elapsed time from input application until the output arrives at and remains within a specified error band around the final value. It is usually specified for a full-scale transition. Figure 1 shows that settling time has three distinct components. The *delay time* is small and almost entirely due to amplifier propagation delay. During this interval there is no output movement.
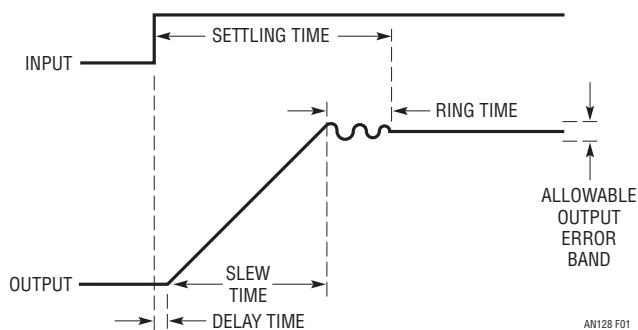
Figure 1. Settling Time Components Include Delay, Slew and Ring Times. Fast Amplifiers Reduce Slew Time, Although Longer Ring Time Usually Results. Delay Time Is Normally a Small Term

During *slew time* the amplifier moves at its highest possible speed towards the final value. *Ring time* defines the region where the amplifier recovers from slewing and ceases movement within some defined error band. There is normally a trade-off between slew and ring time. Fast slewing amplifiers generally have extended ring times, complicating amplifier choice and frequency compensation. Additionally, the architecture of very fast amplifiers usually dictates trade-offs which degrade DC error terms[1].

Measuring anything at any speed requires care. Dynamic measurement is particularly challenging. Reliable nanosecond region settling time measurement constitutes a high order difficulty problem requiring exceptional care in approach and experimental technique[2].

CONSIDERATIONS FOR MEASURING NANOSECOND REGION SETTLING TIME

Historically, settling time has been measured with circuits similar to that in Figure 2. The circuit uses the "false sum node" technique. The resistors and amplifier form a bridge type network. Assuming ideal resistors, the amplifier output will step to $-V_{IN}$ when the input is driven. During slew, the settle node is bounded by the diodes, limiting voltage excursion. When settling occurs, the oscilloscope probe voltage should be zero. Note that the resistor divider's attenuation means the probe's output will be one-half of the actual settled voltage.

In theory, this circuit allows settling to be observed to small amplitudes. In practice, it cannot be relied upon to produce useful measurements. Several flaws exist. The

Note 1. This issue is treated in detail in latter portions of the text. Also see Appendix B, "Practical Considerations for Amplifier Compensation".
Note 2. The approach used for settling time measurement and its description, while new, borrows from previous publications. See References 1-5, and Reference 9.

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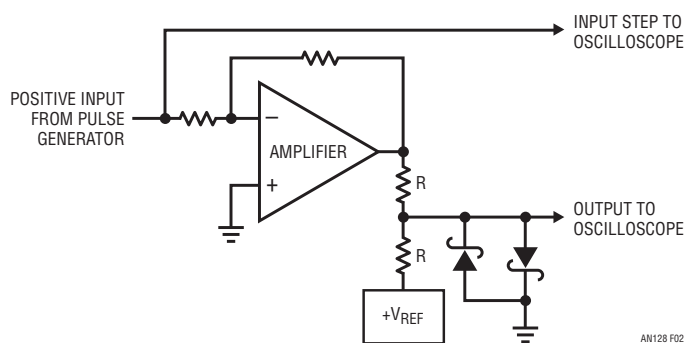
Figure 2. Popular Summing Scheme for Settling Time Measurement Provides Misleading Results. Pulse Generator Post-Transition Aberrations Appear at Output. Large Oscilloscope Overdrive Occurs. Displayed Information Is Meaningless

circuit requires the input pulse to have a flat top within the required measurement limits. Typically, settling within 5mV or less for a 5V step is of interest. No general purpose pulse generator is meant to hold output amplitude and noise within these limits. Generator output-caused aberrations appearing at the oscilloscope probe will be indistinguishable from

A PRECISION WIDEBAND AMPLIFIER WITH 9ns SETTLING TIME

Historically, wideband amplifiers provided speed, but sacrificed precision and, often, settling time. The LT1818 op amp does not require this compromise. It features low offset voltage and bias current with adequate gain for 0.1% accuracy. Settling time is 9ns to 0.1% for a 5V step. The output will drive a 100Ω load to ±3.75V with ±5V supplies, and up to 20pF capacitive loading is permissible at unity gain. The table below provides short form specifications.

LT1818 Short Form Specifications

CHARACTERISTIC	SPECIFICATION
Offset Voltage	0.2mV
Offset Voltage vs Temperature	10µV/°C
Bias Current	2µA
DC Gain	2500
Noise Voltage	6nV/$\sqrt{\text{Hz}}$
Output Current	70mA
Slew Rate	2500V/µs
Gain-Bandwidth	400MHz
Delay	1ns
Settling Time	9ns/0.1%
Supply Current	9mA

amplifier output movement, producing unreliable results. The oscilloscope connection also presents problems. As probe capacitance rises, AC loading of the resistor junction influences observed settling waveforms. 1x probes are not suitable because of their excessive input capacitance. A 10x probe's attenuation sacrifices oscilloscope gain and its 10pF capacitance still introduces significant lag at nanosecond speeds. An active 1x, 1pF FET probe largely alleviates the problem but a more serious issue remains.

The clamp diodes at the settle node are intended to reduce swing during amplifier slewing, preventing excessive oscilloscope overdrive. Unfortunately, oscilloscope overdrive recovery characteristics vary widely among different types and are not usually specified. The Schottky diodes' 400mV drop means the oscilloscope will undergo an unacceptable overload, bringing displayed results into question[3].

At 0.1% resolution (5mV at the amplifier output –2.5mV at the oscilloscope), the oscilloscope typically undergoes a 10x overdrive at 10mV/DIV, and the desired 2.5mV baseline is unattainable. At nanosecond speeds, the measurement becomes hopeless with this arrangement. There is clearly no chance of measurement integrity.

The preceding discussion indicates that measuring amplifier settling time requires an oscilloscope that is somehow immune to overdrive and a "flat top" pulse generator. These become the central issues in wideband amplifier settling time measurement.

The only oscilloscope technology that offers inherent overdrive immunity is the classical sampling 'scope[4]. Unfortunately, these instruments are no longer manufactured (although still available on the secondary market). It is possible, however, to construct a circuit that borrows the overload advantages of classical sampling 'scope technology. Additionally, the circuit can be endowed with features particularly suited for measuring nanosecond range settling time.

Note 3. For a discussion of oscilloscope overdrive considerations, see Appendix C, "Evaluating Oscilloscope Overdrive Performance".
Note 4. Classical sampling oscilloscopes should not be confused with modern era digital sampling 'scopes that have overdrive restrictions. See Appendix C, "Evaluating Oscilloscope Overload Performance" for comparisons of various type 'scopes with respect to overdrive. For detailed discussion of classical sampling 'scope operation, see References 23-26 and 29-31. Reference 24 is noteworthy; it is the most clearly written, concise explanation of classical sampling instruments the author is aware of—a 12-page jewel.

The flat-top pulse generator requirement can be avoided by switching current, rather than voltage. It is much easier to gate a quickly settling current into the amplifier's summing node than to control a voltage. This makes the input pulse generator's job easier, although it still must have a rise time of about 1 nanosecond to avoid measurement errors.

PRACTICAL NANOSECOND SETTLING TIME MEASUREMENT

Figure 3 is a conceptual diagram of a settling time measurement circuit. This figure shares attributes with Figure 2, although some new features appear. In this case, the oscilloscope is connected to the settle point by a switch. The switch state is determined by a delayed pulse generator, which is triggered from the input pulse. The delayed pulse generator's timing is arranged so that the switch does not close until settling is very nearly complete. In this way, the incoming waveform is sampled in time, as well as amplitude. The oscilloscope is never subjected to overdrive—no off-screen activity ever occurs.

A switch at the amplifier's summing junction is controlled by the input pulse. This switch gates current to the amplifier via a voltage-driven resistor. This eliminates the "flat-top" pulse generator requirement, although the switch must be fast and devoid of drive artifacts.

Figure 4 is a more complete representation of the settling time scheme. Figure 3's blocks appear in greater detail and some new refinements show up. The amplifier summing area is unchanged. Figure 3's delayed pulse generator has
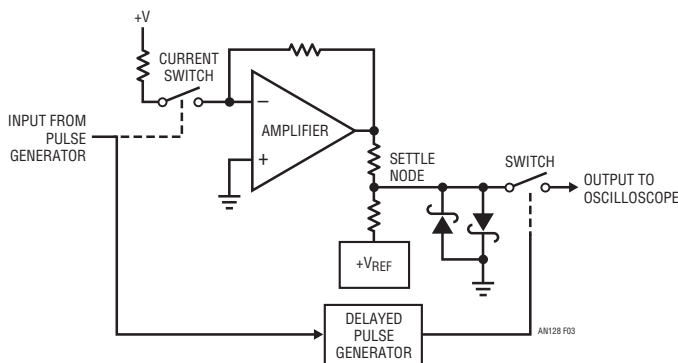
been split into two blocks; a delay and a pulse generator, both independently variable. The input step to the oscilloscope runs through a section that compensates for the propagation delay of the settling time measurement path. Similarly, another delay compensates sample gate pulse generator propagation delay. This delay causes the sample gate pulse generator to be driven with a phase-advanced version of the pulse which triggers the amplifier under test. This considerably improves minimum measurable settling time by making sample gate pulse generator propagation delay irrelevant.

The most striking new aspects of the diagram are the diode bridge switch and the multiplier. The diode bridge's balance, combined with matched, low capacitance Schottky diodes and high speed drive, yields clean switching. The bridge switches current into the amplifier's summing point very quickly, with settling inside a nanosecond . The diode clamp to ground prevents excessive bridge drive swings and ensures that non-ideal input pulse characteristics are nearly irrelevant.

Requirements for Figure 4's sample gate are stringent. It must faithfully pass wideband signal path information without introducing alien components, particularly those deriving from the switch command channel ("sample gate pulse")[5].

The sample gate multiplier functions as a wideband, high resolution, extremely low feedthrough switch. The great advantage of this approach is that the switch control channel can be maintained in-band; that is, its transition rate is held within the multipliers 250MHz bandpass. The multipliers wide bandwidth means the switch command transition is under control at all times. There are no out-of-band responses, greatly reducing feedthrough and parasitic artifacts.

Figure 3. Conceptual Arrangement Is Insensitive to Pulse Generator Aberrations and Eliminates Oscilloscope Overdrive. Input Switch Gates Current Step to Amplifier. Second Switch, Controlled by Delayed Pulse Generator, Prevents Oscilloscope from Monitoring Settle Node Until Settling Is Nearly Complete

Note 5. Conventional choices for the sample gate switch include FET's and the sampling diode bridge. FET parasitic gate to channel capacitances result in large gate drive originated feedthrough into the signal path. For almost all FETs, this feedthrough is many times larger than the signal to be observed, inducing overload and obviating the switches' purpose. The diode bridge is better; its small parasitic capacitances tend to cancel and the symmetrical, differential structure results in very low feedthrough. Practically, the bridge requires DC and AC trims and complex drive and support circuitry. LTC Application Note 74, "Component and Measurement Advances Ensure 16-bit DAC Settling Time" utilized such a sampling bridge and it is detailed in that text. See Reference 3. References 2, 9 and 11 describe a similar sampling bridge based approach.
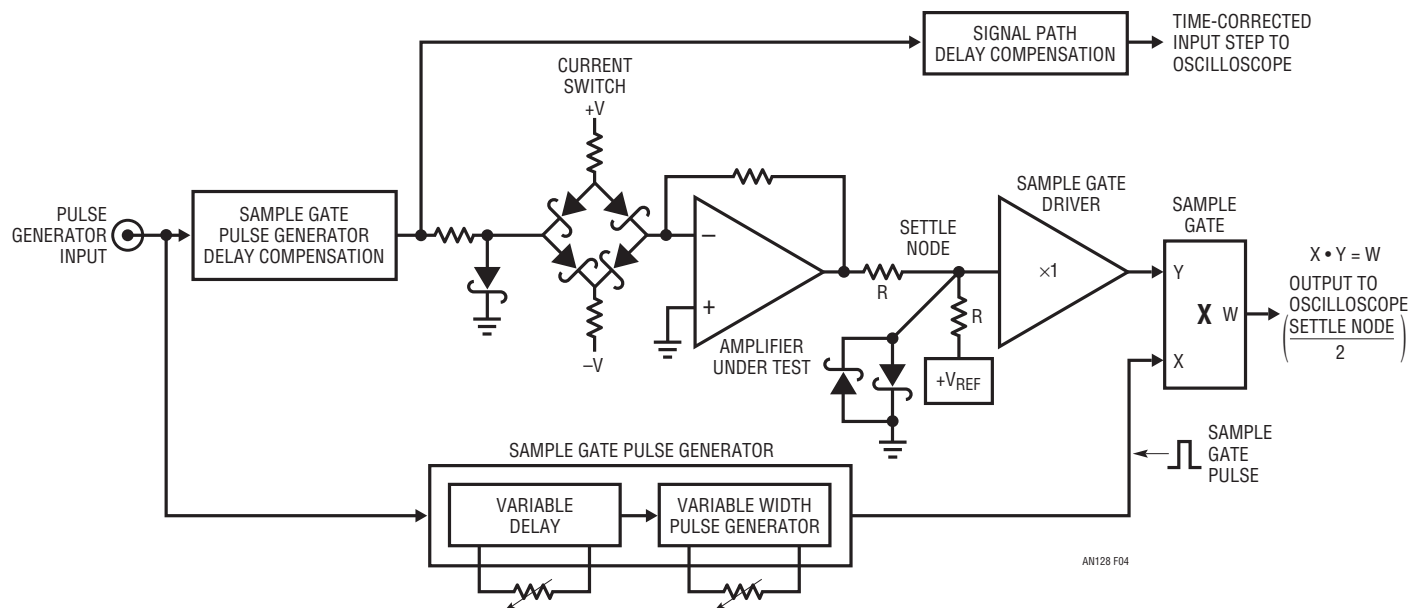
Figure 4. Block Diagram of Settling Time Measurement Scheme. Diode Bridge Cleanly Switches Input Current to Amplifier. Multiplier Based Sampling "Switch" Eliminates Signal Paths Pre-Settling Excursion, Preventing Oscilloscope Overdrive. Input Step Time Reference and Sample Gate Pulse Generator Are Compensated for Test Circuit Delays

DETAILED SETTLING TIME CIRCUITRY

Figure 5 is a detailed schematic of the settling time measurement circuitry. The input pulse switches the input bridge via a delay network ("A" inverters) and a driver stage ("C" inverters). The delay compensates the sample gate pulse generator's delayed response, ensuring that the sample gate pulse can occur immediately after the amplifier-under-tests' slew time ends. The delay range is chosen so that the sample gate pulse can be adjusted to occur *before* the amplifier slews. This capability is obviously unused in operation although it guarantees that the settling interval will always be capturable.

The "C" inverters form a non-inverting driver stage to switch the diode bridge. Various trims optimize driver output pulse shape, providing a clean, fast impulse to the diode bridge[6]. The high fidelity pulse, devoid of undamped components, prevents radiation and disruptive ground currents from degrading the measurement noise floor. The driver also activates the "B" inverters, which supply a time corrected input step to the oscilloscope.

The driver output pulse transitions through the 1N5712 diode clamp potential in under a nanosecond, causing essentially instantaneous diode bridge switching. The resultant cleanly settling current into the amplifier under tests' summing point causes proportionate amplifier output

movement. The negative bias current at the amplifiers summing point combined with the current step produces a +2.5V to −2.5V amplifier output transition. The amplifier's output is compared against a 5V supply derived reference via the summing resistors. The clamped "settle node" is unloaded by A1, which feeds the sample gate signal path information.

The comparator based sample gate pulse generator produces a delayed (controllable by the 20k potentiometer) pulse whose width (controllable by the 2k potentiometer) sets sample gate on-time. The Q1 stage forms the sample gate pulse into a fast rise, exceptionally clean event, furnishing high purity, calibrated amplitude, "on-off" switching instruction to the sample gate multiplier. If the sample gate pulse delay is set appropriately, the oscilloscope will not see any input until settling is nearly complete, eliminating overdrive. The sample window width is adjusted so that all remaining settling activity is observable. In this way, the oscilloscope's output is reliable and meaningful data may be taken.

Figure 6 shows circuit waveforms. Trace A is the time-corrected input pulse, Trace B the amplifier output, Trace C the

Note 6. To maintain text flow and focus, trimming procedures are not presented here. Detailed trimming information appears in Appendix A, "Measuring and Compensating Settling Circuit Delay and Trimming Procedures."

Figure 5. Detailed Schematic of Settling Time Measurement Circuitry Follows Block Diagram. Trimmed, Paralleled Logic Inverters Provide High Speed Drive to Current Switch Bridge. Additional Inverters Form Delay Compensation Networks for Signal Path and Sample Gate Pulse Generator. Transistor Stage Shapes Edges and Amplitude of Sample Gate Pulse Supplied to Multiplier. Multiplier, Functioning as Sample Gate, Passes Settling Time Signal when Sample Gate Pulse Is High

ALL TRIMMING PROCEDURES DETAILED IN APPENDIX A.
* = 1% METAL FILM RESISTOR
** = HSMS-2860, MATCHED 1 MILLIVOLT AT 10mA

⊿ = 1/6 74AHC04, LETTER; e.g. "A", "B", "C" DENOTES SEPARATE PACKAGE. GROUND ALL UNUSED INPUTS.

φ = FERRITE BEAD, FERRONICS #21-110J.

INPUT 50Ω TERMINATION MUST BE IN-LINE COAXIAL TYPE—DO NOT USE BOARD MOUNTED RESISTOR.
+5V AND –5V SUPPLIES DERIVED FROM BOARD MOUNTED REGULATORS. LT1175 = –5V, LT1761 = 5V. TRIM BOTH SUPPLIES—0.1%.
BYPASSING NOT SHOWN EXCEPT FOR AD835. BYPASS EACH IC AND INDICATED SUPPLY POINTS WITH 10μF SANYO OSCON PARALLELED WITH 0.1μF FILM CAPACITOR.

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sample gate pulse and Trace D the settling time output[7]. When the sample gate pulse goes high, the sample gate switches cleanly, and the last 20mV of slew are easily observed. Ring time is also clearly visible, and the amplifier settles nicely to final value. When the sample gate pulse goes low, the sample gate switches off with only 2mV of feedthrough. Note that there is no off-screen activity at any time—the oscilloscope is never subjected to overdrive.

Figure 7 expands vertical and horizontal scales so that settling detail is more visible[8]. Trace A is the time-corrected input pulse and Trace B the settling output. The last 50mV of slew are easily observed, and the amplifier settles inside 5mV (0.1%) in 9 nanoseconds when CF (see Figure 5) is optimized[9].

A = 5V/DIV

B = 5V/DIV

C = 5V/DIV

D = 10mV/DIV

HORIZ = 20ns/DIV AN128 F06
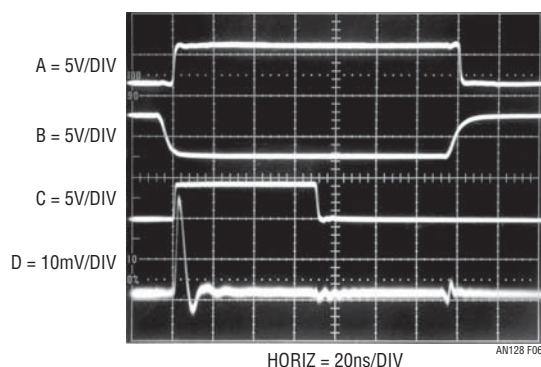
Figure 6. Settling Time Circuit Waveforms Include Time-Corrected Input Pulse (Trace A), Amplifier-Under-Test Output (Trace B), Sample Gate (Trace C) and Settling Time Output (Trace D). Sample Gate Window's Delay and Width Are Variable. Trace B Appears Time Skewed Relative to Time Corrected Trace A.

A = 2V/DIV

B = 10mV/DIV

HORIZ = 5ns/DIV AN128 F07
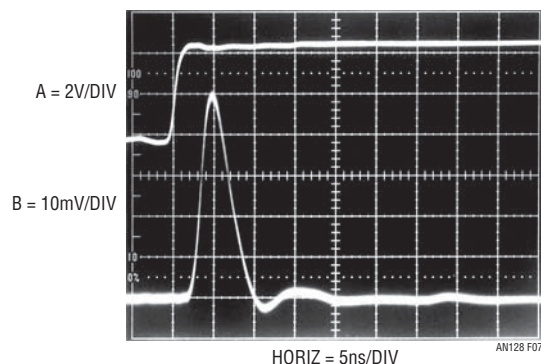
Figure 7. Expanded Vertical and Horizontal Scales Show 9ns Amplifier Settling within 5mV (Trace B). Trace A Is Time Corrected Input Step

Note 7. When interpreting waveform placement note that trace B appears time skewed relative to time corrected trace A. This accounts for trace B's falsely apparent movement before trace A's ascent.

USING THE SAMPLING-BASED SETTLING TIME CIRCUIT

In general, it is good practice to walk the sampling window "backwards" in time up to the last 50mV or so of amplifier slewing so that the onset of ring time is observable without encountering oscilloscope overdrive. The sampling based approach provides this capability and it is a very powerful measurement tool. Slower amplifiers may require extended delay and/or sampling window times, necessitating larger capacitor values in the delayed pulse generator timing networks.

VERIFYING RESULTS–ALTERNATE METHOD

The sampling-based settling time circuit appears to be a useful measurement solution. How can its results be tested to ensure confidence? A good way is to make the same measurement with an alternate method and see if results agree. It was stated earlier that classical sampling oscilloscopes were inherently immune to overdrive[10]. If this is so, why not utilize this feature and attempt settling time measurement directly at the clamped settle node? Figure 8 does this. Under these conditions, the sampling scope[11] is heavily overdriven, but is ostensibly immune to the insult. Figure 9 puts the sampling oscilloscope to the test. Trace A is the time corrected input pulse and Trace B the settle signal. Despite a brutal overdrive, the 'scope appears to respond cleanly, giving a very plausible settle signal presentation.

SUMMARY OF RESULTS AND MEASUREMENT LIMITS

The simplest way to summarize the different method's results is by visual comparison. Ideally, if both approaches represent good measurement technique and are properly constructed, results should be identical. If this is the case, the data produced by the two methods has a high probability of being valid. Examination of Figures 9 and 10 shows

Note 8. In this and all following photos, settling time is measured from the onset of the time-corrected input pulse. Additionally, settling signal amplitude is calibrated with respect to the amplifier, not the settle node. This eliminates ambiguity due to the settle node's resistance ratio.
Note 9. This section mentions amplifier frequency compensation within the context of sampling-based settling time measurement. As such, it is necessarily brief. Considerably more detail is available in Appendix B, "Practical Considerations for Amplifier Compensation."
Note 10. See Appendix C, "Evaluating Oscilloscope Overdrive Performance" for in-depth discussion.
Note 11. Tektronix type 661 with 4S1 vertical and 5T3 timing plug-ins.

LINEAR TECHNOLOGY

nearly identical settling times and highly similar settling waveform signatures. This kind of agreement provides a high degree of credibility to the measured results.

Close observation of settling time circuit operation indicates a noise floor/feedthrough imposed amplitude

resolution limit of 2mV. The time resolution limit is about 2 nanoseconds to 5mV settling. For details, see the section "Measurement Limits and Uncertainties", in Appendix A, "Measuring and Compensating Settling Circuit Delay and Trimming Procedures."

Figure 8. Settling Time Test Circuit Modifications Using Classical Tektronix 661/4S1/5T3 1GHz Sampling Oscilloscope. Sampling 'Scope's Inherent Overload Immunity Permits Large Off-Screen Excursions without Degrading Measurement Fidelity

Figure 9. Settling Time Measurement with Classical Sampling 'Scope. Oscilloscope's Overload Immunity Allows Accurate Measurement Despite Extreme Overdrive. 9ns Settling Time and Waveform Profile Are Consistent with Figure 7

Figure 10. Settling Time Measurement Using Figure 5's Circuit. T_{SETTLE} = 9ns. Results Correlate with Figure 9

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APPENDIX A

Measuring and Compensating Settling Circuit Delay and Trimming Procedures

The settling time circuit requires trimming to achieve quoted performance. The trims fall into four loosely defined categories including current switch bridge drive pulse shaping, circuit delays, sample gate pulse purity and sample gate feedthrough/DC adjustments[1].

Bridge Drive Trims

The current switch bridge drive is trimmed first. Disconnect all 5 bridge drive related trims and apply a 5V, 1MHz, 10 to 15 nanosecond wide pulse at the circuit input. The paralleled "C" inverter output viewed at the 43Ω back termination's undriven end should resemble Figure A1. Waveform edge times are fast but poorly controlled parasitic excursions risk corrupting the measurement noise floor and must be eliminated. Reconnect all 5 trims and adjust them according to their titles for Figure A2's much improved presentation. There is some interaction between the adjustments but it is limited and favorable results are easily attained. Figure A2's edge times are slightly slower than Figure A1's, but still pass through the 1N5712 clamp level in <1 nanosecond.

Delay Determination and Compensation

Circuit delay related trims come next. Before making these measurements and adjustments, probe/oscilloscope channel-to-channel time skewing must be corrected. Figure A3 shows 40 picosecond time skew error with both channel probes connected to a 100 picosecond rise time pulse source[2]. The error is corrected in Figure A4 by utilizing the oscilloscopes vertical amplifier variable delay feature (Tektronix 7A29, option 04, installed in a Tektronix 7104 mainframe). This correction permits high accuracy delay measurements to be made[3].

Note 1. The trims require considerable care in instrumentation selection as well as thoughtful wideband probing and oscilloscope measurement technique. See Appendixes D through H for tutorial guidance before proceeding.
Note 2. See Appendix H, "Verifying Rise Time and Delay Measurement Integrity" for fast pulse source recommendations.
Note 3. This assumes the oscilloscope time base has been verified for accuracy. For recommendations, see Appendix H, "Verifying Rise Time and Delay Measurement Integrity".

A = 0.5V/DIV

HORIZ = 2ns/DIV   AN128 FA01
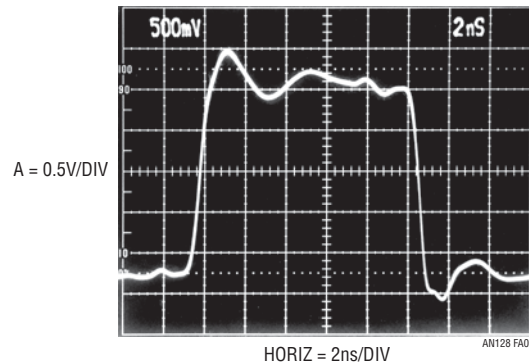
Figure A1. Untrimmed Current Switch Driver Response at 43Ω Back Termination Viewed in 1GHz, Real Time Bandwidth. Edge Times Are Fast, But Poorly Controlled. Undamped Waveform Artifacts Risk Corrupting Signal Path Noise Floor Via Radiation and Ground Current Disruption Induced Errors

A = 0.5V/DIV

HORIZ = 2ns/DIV   AN128 FA02
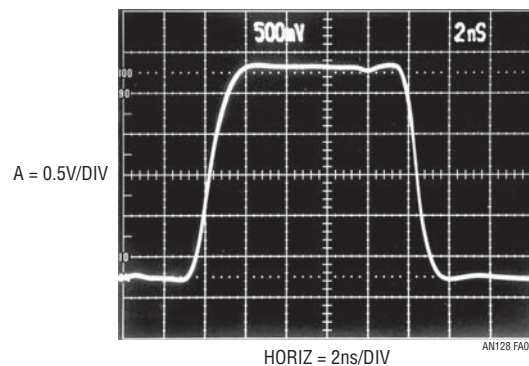
Figure A2. Trimmed Current Switch Driver Output at 43Ω Back Termination Passes Through 0.6V Diode Clamp Potential in <1 Nanosecond. AC Trims Promote Clean, Well Controlled Waveform

A = 0.1V/DIV
B = 0.1V/DIV

HORIZ = 200ps/DIV   AN128 FA03
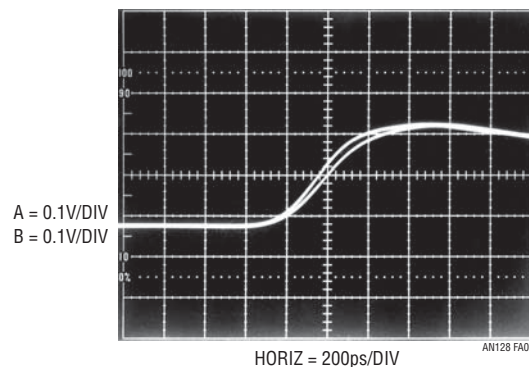
Figure A3. Probe-Oscilloscope Channel-to-Channel Timing Skew Measures 40 Picoseconds
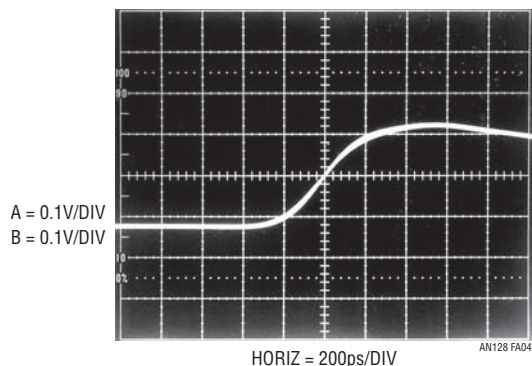
**Figure A4. Corrected Probe/Channel/Skew Shows Nearly Identical Time and Amplitude Response**

HORIZ = 200ps/DIV

A = 0.1V/DIV
B = 0.1V/DIV

AN128 FA04

Figure A5. Current Switch Driver (Trace A) to Amplifier-Under-Test Negative Input (Trace B) Delay Is 250 Picoseconds

HORIZ = 500ps/DIV

A = 5V/DIV
B = 20mA/DIV

AN128 FA05

Figure A6. Amplifier-Under-Test (Trace A) to Circuit Output (Trace B) Delay Measures 8.4 Nanoseconds. Multiplier X Input Held at 1V DC for This Test

HORIZ = 2ns/DIV

A = 2V/DIV
B = 0.2V/DIV

AN128 FA06

Figure A7. Multiplier Delay with X Input Held at 1V DC Measures 2ns

HORIZ = 2ns/DIV

A = 0.1V/DIV
B = 0.1V/DIV

AN128 FA07

The settling time circuit utilizes an adjustable delay network to time correct the input pulse for delays in the signal-processing path. Typically, these delays introduce errors approaching 10 nanoseconds, so an accurate correction is required. Setting the delay trim involves observing the network's input-output delay and adjusting for the appropriate time interval. Determining the "appropriate" time interval is somewhat more complex.

Referring to Figure 5, it is apparent that three delay measurements are of interest. The current switch driver to amplifier-under-test negative input, the amplifier-under-test output to circuit output and the sample gate multiplier delay. Figure A5 indicates 250 picoseconds delay from the current switch driver to the amplifier-under-test input. Figure A6 reveals 8.4 nanoseconds delay from the amplifier-under-test output to circuit output and Figure A7 shows sample gate multiplier delay of 2 nanoseconds. The measurements indicate a current switch driver-to-circuit output delay of 8.65 nanoseconds; the correction is implemented by adjusting the 1k trim in the "Signal Path Delay Compensation" network for that amount. Similarly, when the sampling 'scope is used, the relevant delays are Figure A5 plus A6 minus Figure A7, a total of 6.65 nanoseconds. This factor is adjusted into the signal path delay compensation network when the sampling 'scope-based measurement is taken.

The "Sample Gate Pulse Generator Path Delay Compensation" trim is less critical. The sole requirement is that it overlap the sample gate pulse generator's delay. Setting the 1k potentiometer in the "A" inverter chain to 15 nanoseconds satisfies this criteria, completing the delay related trims.
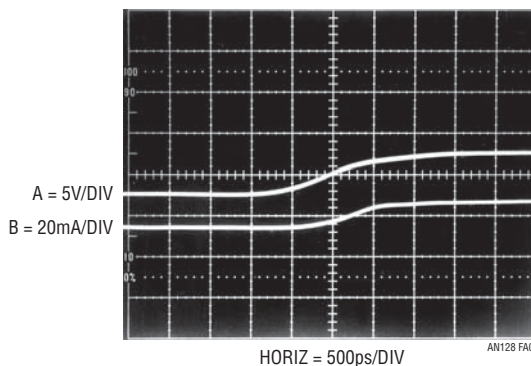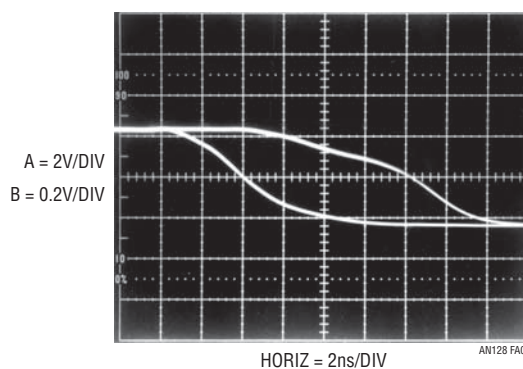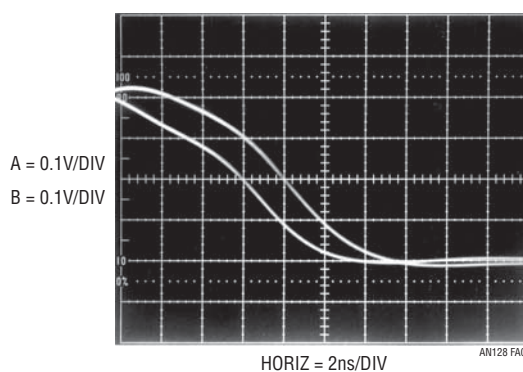
Sample Gate Pulse Purity Adjustment

The Q1 sample gate pulse edge shaping stage is adjusted for an optimized front corner, minimum rising edge time, pulse top smoothing and 1V amplitude with the indicated trims. The mildly interactive adjustments converge to
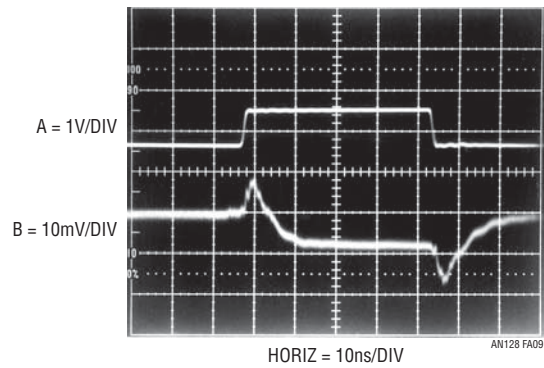
an128f
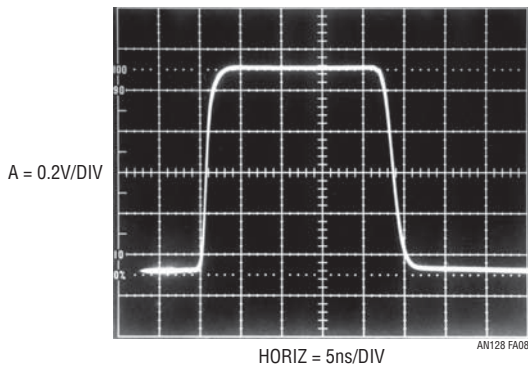
Figure A8. Sample Gate Pulse Characteristics, Controlled by Edge Shaping, Circuit Configuration and Transistor Choice, Are Kept Within Multiplier's 250MHz (T_{RISE} = 1.4ns) Bandwidth. Accurate, Low Feedthrough, Y Input Signal Path Switching Results

Figure A9. Settling Time Circuit's Output (Trace B) with Unadjusted Sample Gate Feedthrough and DC Offset. A1's Input Grounded for This Test. Excessive Switch Drive Feedthrough and Baseline Offset Are Present. Trace A Is Sample Gate Pulse

Figure A8's display, taken at the sample gate multiplier's X input. The pulse's 2 nanosecond rise time promotes rapid sample gate acquisition but remains within the multipliers 250MHz (t_{RISE} = 1.4ns) bandwidth, assuring freedom from out-of-band parasitic responses. The clean, 1V amplitude pulse top provides calibrated, consistent multiplier output devoid of aberrations which would masquerade as settling signal artifacts. Pulse fall time is irrelevant; it is not germaine to the measurement and its clean falling transition assures controlled multiplier turn-off, precluding off-screen excursions.

Sample Gate Path Optimization

The sample gate path adjustments are the final trims. First, put in 5V DC to the pulse generator input to lock the amplifier-under-test into its −2.5V output state. Adjust the "settle node zero" trim for zero volts within 1mV at A1's output. Next, restore the pulsed circuit input, disconnect the settle node from A1 and ground A1's input with a 750Ω resistor. Figure A9 is typical of the resultant untrimmed response. Ideally, the circuit output (trace B) should be static during sample gate (trace A) switching. The photo reveals errors; correction requires trimming DC offset and dynamic feedthrough related residue. The DC errors are eliminated by adjusting the "X" and "Y" offset trims for a continuous trace B baseline regardless of trace A's sample gate pulse state. Additionally, set the output offset adjustment for minimum multiplier baseline offset voltage. Sample gate gain is set to unity by shutting off the input pulse generator, applying 5V DC to C2's "+" input

and forcing 1.00V DC at the previously inserted 750Ω resistor. Under these conditions, adjust "scale factor" for 1.00V DC output. After completing this step, remove the DC bias voltages and the 750Ω resistor, reconnect the settle node and restore the pulsed input.

Feedthrough compensation is accomplished via feedthrough "time phase" and "amplitude" trims. These adjustments set timing and amplitude of the feedthrough correction applied at the multiplier "Z" input. Optimal adjustment results in Figure A10's presentation. This photograph shows the DC and feedthrough trims dramatic effect on Figure A9's pre-trim errors[4].
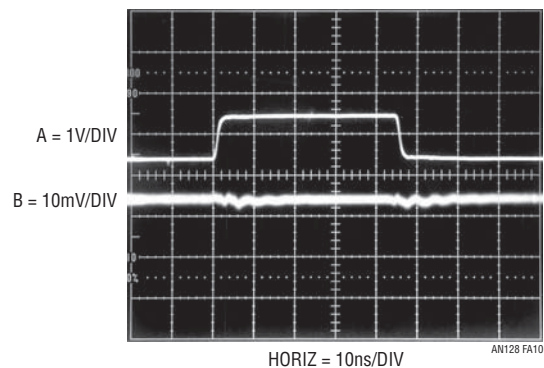
Figure A10. Settling Time Circuit's Output (Trace B) with Sample Gate Trimmed. As in Figure A9, A1's Input Is Grounded for This Test. Switch Drive Feedthrough and Baseline Offset Are Minimized. Trace A Is Sample Gate Pulse. Measurement Defines Circuit's 2mV Minimum Amplitude Resolution Limit

Note 4. The writer is not much for Hollywood's offerings, but does find drama in feedthrough trims.

Measurement Limits and Uncertainties

Figure A10's post trim response includes a flat baseline and greatly attenuated feedthrough. The measurement defines the circuit's minimum amplitude resolution at 2mV. In another test, A1's input is disconnected from the settle node and biased at 20mV DC via a 750Ω resistor to simulate an infinitely fast settling amplifier. Figure A11 shows circuit output (trace B) settling within 5mV in 2 nanoseconds, arriving inside the 2mV baseline noise limit in 3.6 nanoseconds. This data, taken with sample gate conduction beginning immediately after the time corrected input (trace A) rises, defines the circuit's minimum time resolution limit. Uncertainties in the quoted time and amplitude resolution limits are primarily due to delay compensation limitations, noise and residual feedthrough. Considering likely delay and measurement errors, a time uncertainty of ±500 picoseconds and a 2mV resolution limit

is probably realistic. Noise averaging would not improve the amplitude resolution limit because it is imposed by feedthrough residue, a coherent term.
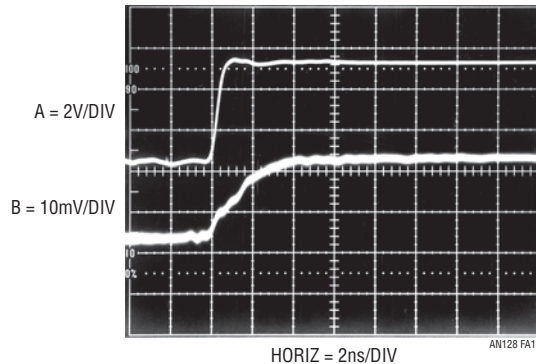
Figure A11. Circuit Response with 20mV DC Forced at A1's Input. Output (Trace B) Is within 5mV in 2ns, Arriving Inside 2mV Baseline Noise in 3.6ns. Measurement Defines Circuits Minimum Time Resolution Limit. Trace A Is Time Corrected Input Pulse

APPENDIX B

Practical Considerations for Amplifier Compensation

There are a number of practical considerations in compensating the amplifier to get fastest settling time. Our study begins by revisiting text Figure 1 (repeated here as Figure B1). Settling time components include delay, slew and ring times. Delay is due to propagation time through the amplifier and is a relatively small term. Slew time is set by the amplifier's maximum speed. Ring time defines the region where the amplifier recovers from slewing and ceases movement within some defined error band. Once an amplifier has been chosen, only ring time is readily adjustable. Because slew time is usually the dominant lag, it is tempting to select the fastest slewing amplifier

Figure B1. Settling Time Components Include Delay, Slew and Ring Times. For Given Components, Only Ring Time Is Readily Adjustable

available to obtain best settling. Unfortunately, fast slewing amplifiers usually have extended ring times, negating their brute force speed advantage. The penalty for raw speed is, invariably, prolonged ringing, which can only be damped with large compensation capacitors. Such compensation works, but results in protracted settling times. The key to good settling times is to choose an amplifier with the right balance of slew rate and recovery characteristics and compensate it properly. This is harder than it sounds because amplifier settling time cannot be predicted or extrapolated from any combination of data sheet specifications. It must be measured in the intended configuration. A number of terms combine to influence settling time. They include amplifier slew rate and AC dynamics, layout capacitance, source resistance and capacitance, and the compensation capacitor. These terms interact in a complex manner, making predictions hazardous[1]. If the parasitics are eliminated and replaced with a pure resistive source, amplifier settling time is still not readily predictable. The parasitic impedance terms just make a difficult problem more messy. The only real handle available to deal with all this is the feedback compensation capacitor, C_F. C_F's purpose is to roll off amplifier gain at the frequency that permits best dynamic response.


**Note 1.** Spice aficionados take notice.

Best settling results when the compensation capacitor is selected to functionally compensate for all the above terms. Figure B2 shows results for an optimally selected feedback capacitor. Trace A is the time-corrected input pulse and trace B the amplifier's settle signal. The amplifier comes cleanly out of slew (sample gate opens just after the second vertical division) and settles to 5mV in 9 nanoseconds. Waveform signature is tight and nearly critically damped.


A = 2V/DIV

B = 10mV/DIV

HORIZ = 5ns/DIV   AN128 FB02
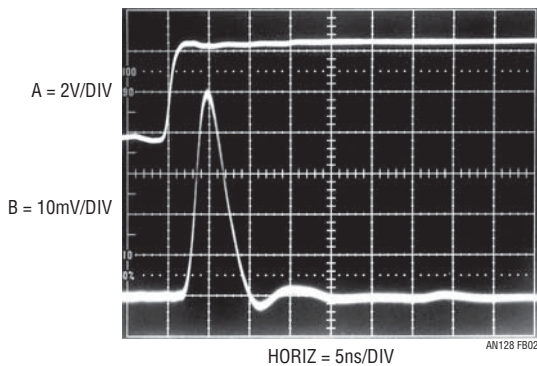
Figure B2. Optimized Compensation Capacitor Permits Tight Waveform Signature, Nearly Critically Damped Response and Fastest Settling Time. T$_{SETTLE}$ = 9ns. Trace A Is Time Corrected Input Step, Trace B, the Settle Signal

In Figure B3, the feedback capacitor is too large. Settling is smooth, although overdamped; a 13 nanosecond penalty results in 22 nanosecond settling. Figure B4 has no feedback capacitor, causing severely underdamped response with resultant excessive ring time excursions. Settling time goes out to 33 nanoseconds. B5 improves on B4 by restoring the feedback capacitor, but the value is too small, resulting in an underdamped response requiring 27 nanoseconds to settle. Note that Figures B3 to B5 require vertical scale reduction to capture non-optimal response.

When feedback capacitors are individually trimmed for optimal response, the source, stray, amplifier and compensation capacitor tolerances are irrelevant. If individual trimming is not used, these tolerances must be considered to determine the feedback capacitor's production value. Ring time is affected by stray and source capacitance and output loading, as well as the feedback capacitor's value. The relationship is nonlinear, although some guidelines are

A = 2V/DIV

B = 20mV/DIV

HORIZ = 5ns/DIV   AN128 FB03
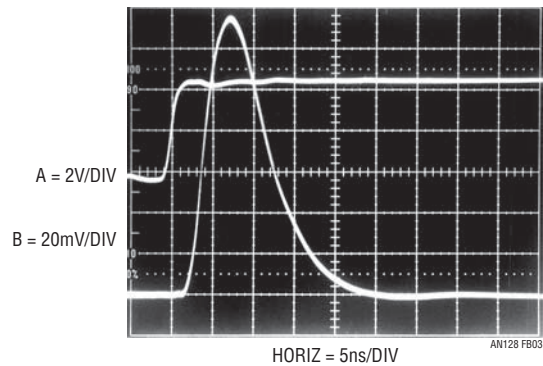
Figure B3. Overdamped Response Ensures Freedom from Ringing, Even with Component Variations in Production. Penalty Is Increased Settling Time. Note 2X Vertical Scale Change vs. Figure B2. T$_{SETTLE}$ = 22ns. Trace Assignments Same as Previous Figure

A = 2V/DIV

B = 50mV/DIV

HORIZ = 5ns/DIV   AN128 FB04
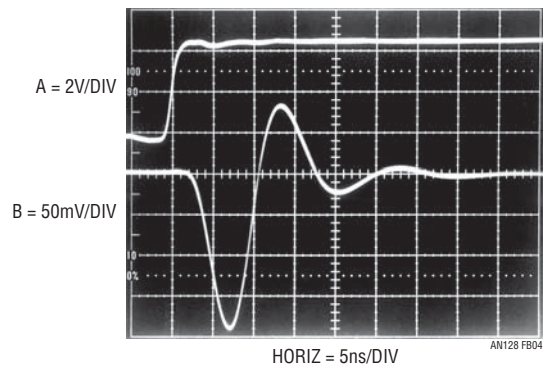
Figure B4. Severely Underdamped Response Due to No Feedback Capacitor. Note 5X Vertical Scale Change vs Figure B2. T$_{SETTLE}$ = 33ns. Trace Assignments as in Figure B2

A = 2V/DIV

B = 50mV/DIV

HORIZ = 5ns/DIV   AN128 FB05
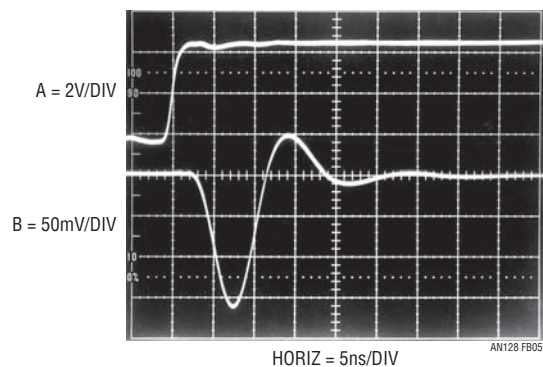
Figure B5. Underdamped Response Results from Undersized Capacitor. Component Tolerance Budgeting Will Prevent This Behavior. Note 5x Vertical Scale Change vs. Figure B2. T$_{SETTLE}$ = 27ns. Trace Assignments as in Figure B2

possible. The stray and source terms can vary by ±10% and the feedback capacitor is typically a ±5% component[2]. Additionally, amplifier slew rate has a significant tolerance, which is stated on the data sheet. To obtain a production feedback capacitor value, determine the optimum value by individual trimming with the production board layout (board layout parasitic capacitance counts too!). Then, factor in the worst-case percentage values for stray and source impedance terms, slew rate and feedback

capacitor tolerance. Add this information to the trimmed capacitors measured value to obtain the production value. This budgeting is perhaps unduly pessimistic (RMS error summing may be a defensible compromise), but will keep you out of trouble[3].

Note 2. This assumes a resistive source. If the source has substantial parasitic capacitance (photodiode, DAC, etc.), this number can easily enlarge to ±50%.
Note 3. The potential problems with RMS error summing become clear when sitting in an airliner that is landing in a snowstorm.

APPENDIX C

Evaluating Oscilloscope Overdrive Performance

The sampling based settling time circuit is heavily oriented towards preventing overdrive to the monitoring oscilloscope. This is done to avoid overdriving the oscilloscope. Oscilloscope recovery from overdrive is a grey area and almost never specified. How long must one wait after an overdrive before the display can be taken seriously? The answer to this question is quite complex. Factors involved include the degree of overdrive, its duty cycle, its magnitude in time and amplitude and other considerations. Oscilloscope response to overdrive varies widely between types and markedly different behavior can be observed in any individual instrument. For example, the recovery time for a 100x overload at 0.005V/DIV may be very different than at 0.1V/DIV. The recovery characteristic may also vary with waveform shape, DC content and repetition rate. With so many variables, it is clear that measurements involving oscilloscope overdrive must be approached with caution.

Why do most oscilloscopes have so much trouble recovering from overdrive? The answer to this question requires some study of the three basic oscilloscope types' vertical path. The types include analog (Figure C1A), digital (Figure C1B) and classical sampling (Figure C1C) oscilloscopes. Analog and digital 'scopes are susceptible to overdrive. The classical sampling 'scope is the only architecture that is inherently immune to overdrive.

An analog oscilloscope (Figure C1A) is a real time, continuous linear system[1]. The input is applied to an attenuator, which is unloaded by a wideband buffer. The vertical preamp provides gain, and drives the trigger pick-off, delay line and the vertical output amplifier. The attenuator and delay line

are passive elements and require little comment although they can display reactive behavior at speed and resolution extremes. The buffer, preamp and vertical output amplifier are complex linear gain blocks, each with dynamic operating range restrictions. Additionally, the operating point of each block may be set by inherent circuit balance, low frequency stabilization paths or both. When the input is overdriven, one or more of these stages may saturate, forcing internal nodes and components to abnormal operating points and temperatures. When the overload ceases, full recovery of the electronic and thermal time constants may require surprising lengths of time[2].

The digital sampling oscilloscope (Figure C1B) eliminates the vertical output amplifier, but has an attenuator buffer and amplifiers ahead of the A/D converter. Because of this, it is similarly susceptible to overdrive recovery problems.

The classical sampling oscilloscope is unique. Its nature of operation makes it inherently immune to overload. Figure C1C shows why. The sampling occurs before any gain is taken in the system. Unlike Figure C1B's digitally sampled 'scope, the input is fully passive to the sampling point. Additionally, the output is fed back to the sampling bridge, maintaining its operating point over a very wide range of inputs. The dynamic swing available to maintain the bridge output is large and easily accommodates a wide range of oscilloscope inputs. Because of all this, the amplifiers in this instrument do not see overload, even at 1000x overdrives, and there is no recovery problem. Additional immunity derives from the instrument's relatively slow

Note 1. Ergo, the Real Thing. Hopelessly bigoted residents of this locale mourn the passing of the analog 'scope era and frantically hoard every instrument they can find.
Note 2. Some discussion of input overdrive effects in analog oscilloscope circuitry is found in Reference 17.
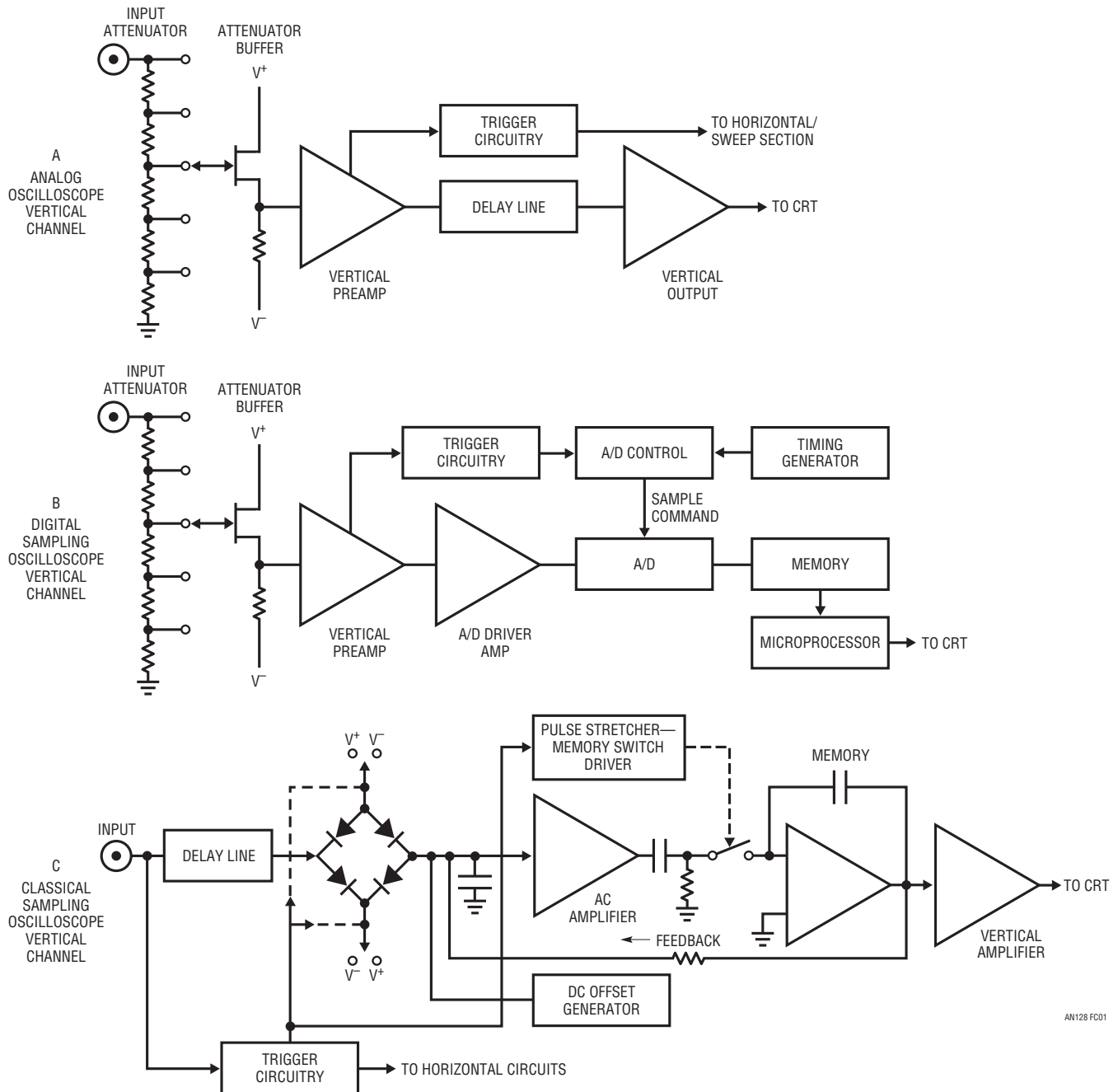
Figure C1. Simplified Vertical Channel Diagrams for Different Type Oscilloscopes. Only the Classical Sampling 'Scope (C) Has Inherent Overdrive Immunity. Offset Generator Allows Viewing Small Signals Riding On Large Excursions

sample rate—even if the amplifiers were overloaded, they would have plenty of time to recover between samples[3].

The designers of classical sampling 'scopes capitalized on the overdrive immunity by including variable DC offset generators to bias the feedback loop (see Figure C1C, lower right). This permits the user to offset a large input, so small amplitude activity on top of the signal can be accurately

observed. This is ideal for, among other things, settling time measurements. Unfortunately, classical sampling oscilloscopes are no longer manufactured, so if you have one, take care of it[4]!

Note 3. Additional information and detailed treatment of classical sampling oscilloscope operation appears in References 23-26 and 29-31.
Note 4. Modern variants of the classical architecture (e.g., Tektronix 11801B) may provide similar capability, although we have not tried them.

Although analog and digital oscilloscopes are susceptible to overdrive, many types can tolerate some degree of this abuse. The early portion of this appendix stressed that measurements involving oscilloscope overdrive must be approached with caution. Nevertheless, a simple test can indicate when the oscilloscope is being deleteriously affected by overdrive.

The waveform to be expanded is placed on the screen at a vertical sensitivity that eliminates all off-screen activity. Figure C2 shows the display. The lower right hand portion is to be expanded. Increasing the vertical sensitivity by a factor of two (Figure C3) drives the waveform off-screen, but the remaining display appears reasonable. Amplitude has doubled and waveshape is consistent with the original display. Looking carefully, it is possible to see small amplitude information presented as a dip in the waveform at about the third vertical division. Some small disturbances are also visible. This observed expansion of the original waveform is believable. In Figure C4, gain has been further increased, and all the features of Figure C3 are amplified

A = 1V/DIV
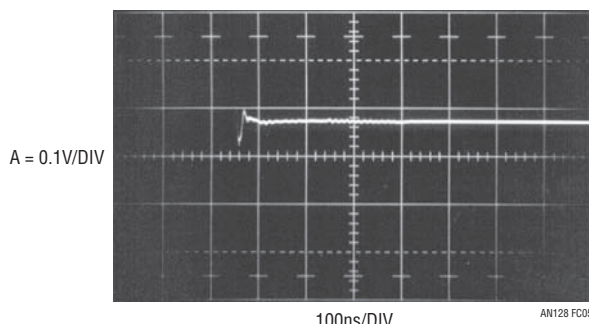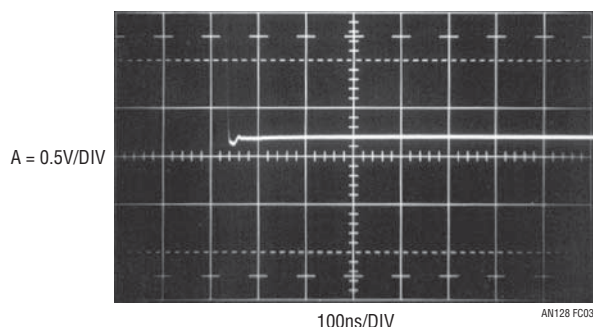
100ns/DIV   AN128 FC02

Figure C2

A = 0.1V/DIV

100ns/DIV   AN128 FC05

Figure C5

A = 0.5V/DIV

100ns/DIV   AN128 FC03

Figure C3

A = 0.1V/DIV

100ns/DIV   AN128 FC06

Figure C6

A = 0.2V/DIV

100ns/DIV   AN128 FC04

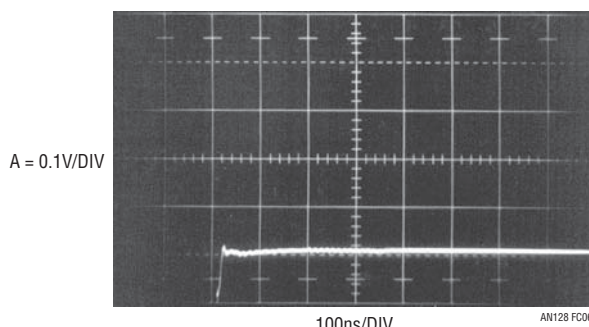**Figure C4**
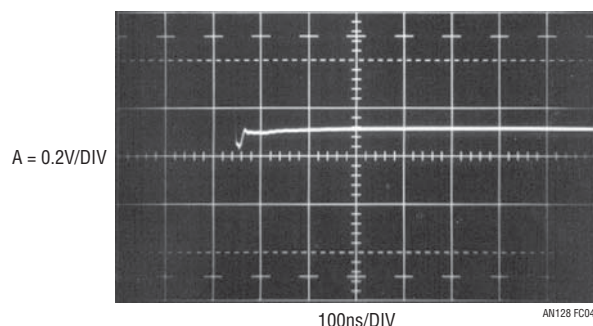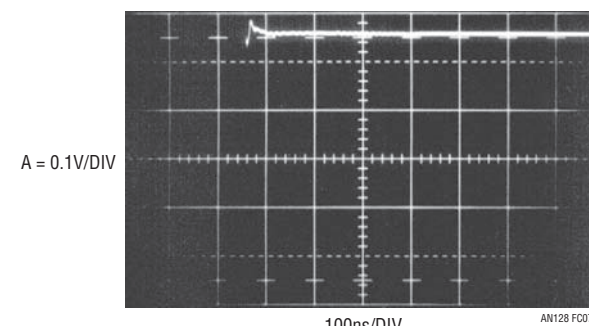

A = 0.1V/DIV

100ns/DIV   AN128 FC07
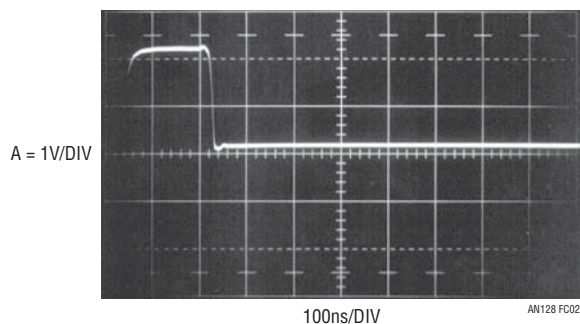
Figure C7

Figure C2 to C7. The Overdrive Limit Is Determined by Progressively Increasing Oscilloscope Gain and Watching for Waveform Aberrations

accordingly. The basic waveshape appears clearer and the dip and small disturbances are also easier to see. No new waveform characteristics are observed. Figure C5 brings some unpleasant surprises. This increase in gain causes definite distortion. The initial negative-going peak, although larger, has a different shape. Its bottom appears less broad than in Figure C4. Additionally, the peak's positive recovery is shaped slightly differently. A new rippling disturbance is visible in the center of the screen. This kind of change indicates that the oscilloscope is having trouble. A further test can confirm that this waveform is being influenced by overloading. In Figure C6 the gain remains the same

but the vertical position knob has been used to reposition the display at the screen's bottom[5]. This shifts the oscilloscope's DC operating point which, under normal circumstances, should not affect the displayed waveform. Instead, a marked shift in waveform amplitude and outline occurs. Repositioning the waveform to the screen's top produces a differently distorted waveform (Figure C7). It is obvious that for this particular waveform, accurate results cannot be obtained at this gain.

Note 5. Knobs (derived from Middle English, "knobbe", akin to Middle Low German, "knubbe"), cylindrically shaped, finger rotatable panel controls for controlling instrument functions, were utilized by the ancients.

APPENDIX D

ABOUT Z_0 PROBES

#### When to Roll Your Own and When to Pay the Money

Z_0 (e.g. low impedance) probes provide the most faithful high speed probing mechanism available for low source impedances. Their sub-picofarad input capacitance and near ideal transmission characteristic make them the first choice for high bandwidth oscilloscope measurement. Their deceptively simple operation invites "do-it-yourself" construction but numerous subtleties mandate difficulty for prospective constructors. Arcane parasitic effects introduce errors as speed increases beyond about 100MHz ($t_{RISE} = 3.5ns$). The selection and integration of probe materials and the probes physical incarnation require extreme care to obtain high fidelity at high speed. Additionally, the probe must include some form of adjustment to compensate small, residual parasitics. Finally, true coaxiality must be maintained when fixturing the probe at the measurement point, implying a high grade, readily disconnectable, coaxial connection capability.

Figure D1 shows that a Z_0 probe is basically a voltage divided input 50Ω transmission line. If R1 equals 450Ω, 10x attenuation and 500Ω input resistance result. R1 of 4950Ω causes a 100x attenuation with 5k input resistance. The 50Ω line theoretically constitutes a distortionless transmission environment. The apparent simplicity seemingly permits "do-it-yourself" construction but this section's remaining figures demonstrate a need for caution.

Figure D2 establishes a fidelity reference by measuring a clean 700ps rise time pulse using a 50Ω line terminated
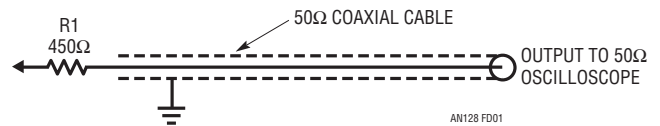
Figure D1. Conceptual 500Ω, "Z_0", 10x Oscilloscope Probe. If R1 = 4950Ω, 5k Input Resistance with 100x Signal Attenuation Results. Terminated Into 50Ω, Probe Theoretically Constitutes Distortionless Transmission Line. "Do-It-Yourself" Probes Suffer Uncompensated Parasitics, Causing Unfaithful Response Above ≈ 100MHz (t_{RISE} = 3.5ns)
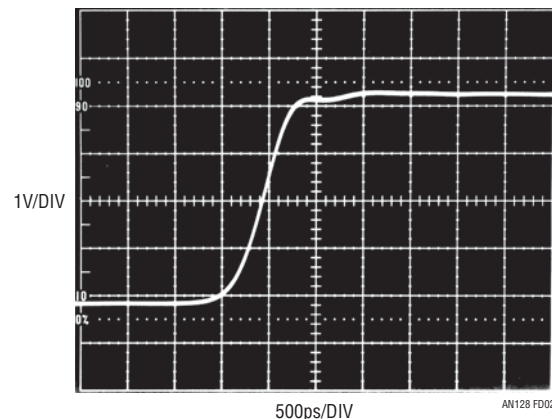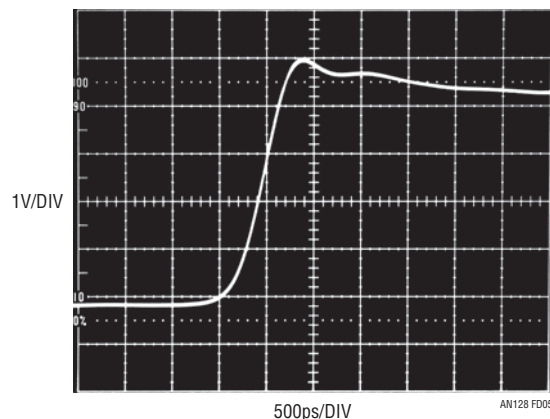
Figure D2. 700ps Rise Time Pulse Observed via 50Ω Line and 10x Coaxial Attenuator Has Good Pulse Edge Fidelity with Controlled Post-Transition Events

via a 10x coaxial attenuator—no probe is employed. The waveform is singularly clean and crisp with minimal edge and post-transition aberrations. Figure D3 depicts the same pulse with a commercially produced 10x Z_0 probe in use. The probe is faithful and there is barely discernible error in the presentation. Photos D4 and D5, taken with
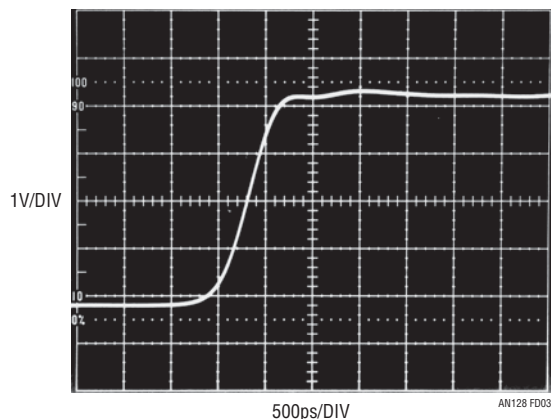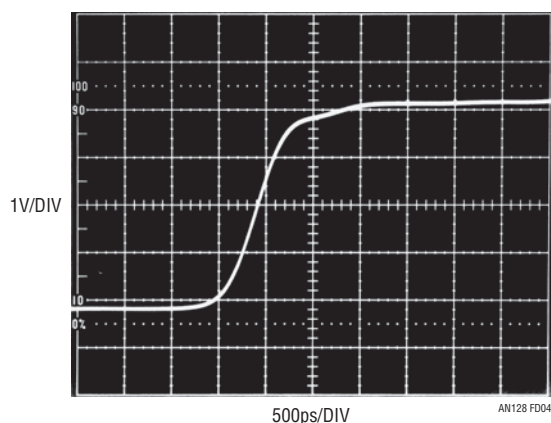
Figure D3. Figure D2's Pulse Viewed with Tektronix 10x, Z$_0$ 500Ω Probe (P-6056) Introduces Barely Discernible Error

Figure D4. "Do-It-Yourself" Z$_0$ Probe #1 Introduces Pulse Corner Rounding, Likely Due to Resistor/Cable Parasitic Terms or Incomplete Coaxiality. "Do-It-Yourself" Z$_0$ Probes Typically Manifest This Type of Error at Rise Times ≤2ns

Figure D5. "Do-It-Yourself" Z$_0$ Probe #2 Has Overshoot, Again Likely Due to Resistor/Cable Parasitic Terms or Incomplete Coaxiality. Lesson: At These Speeds, Don't "Do-It-Yourself"

two separately constructed "do-it-yourself" Z$_0$ probes, show errors. In D4, Probe #1 introduces pulse front corner rounding; Probe #2 in D5 causes pronounced corner peaking. In each case, some combination of resistor/cable parasitics and incomplete coaxiality are likely responsible for the errors. In general, "do-it-yourself" Z$_0$ probes cause these types of errors beyond about 100MHz (t$_{RISE}$ 3.5ns). At higher speeds, if waveform fidelity is critical, it's best to pay the money. For additional terror and wisdom along these lines, see pg. 2-4 → 2-8 of Reference 30 and Reference 35. Both are excellent, informative and, hopefully, sobering to still undaunted do-it-yourselfers.

APPENDIX E

Connections, Cables, Adapters, Attenuators, Probes and Picoseconds

Subnanosecond rise time signal paths must be considered as a transmission line. Connections, cables, adapters, attenuators and probes represent discontinuities in this transmission line, deleteriously affecting its ability to faithfully transmit desired signal. The degree of signal corruption contributed by a given element varies with its deviation from the transmission lines nominal impedance. The practical result of such introduced aberrations is degradation of pulse rise time, fidelity, or both. Accordingly, introduction of elements or connections to the signal path should be minimized and necessary connections and elements must be high grade components. Any form of connector, cable, attenuator or probe must be fully specified for high frequency use. Familiar BNC hardware becomes lossy at rise times much faster than 350ps. SMA components are preferred for the rise times described in the text. Additionally, cable should be 50Ω "hard line" or, at least, Teflon-based coaxial cable fully specified for high frequency operation. Optimal connection practice eliminates any cable by coupling the signal output *directly* to the measurement input.

Mixing signal path hardware types via adapters (e.g. BNC/SMA) should be avoided. Adapters introduce significant parasitics, resulting in reflections, rise time degradation, resonances and other degrading behavior. Similarly, oscilloscope connections should be made directly to the instrument's 50Ω inputs, avoiding probes. If probes must be used, their introduction to the signal path mandates attention to their connection mechanism and high frequency compensation. Passive Z_0 types, commercially available in 500Ω (10x) and 5kΩ (100x) impedances, have input capacitance below 1pf[1]. Any such probe must be carefully frequency compensated before use or misrepresented measurement will result. Inserting the probe into the signal path necessitates some form of signal pick-off which nominally does not influence signal transmission. In practice, some amount of disturbance must be tolerated and its effect on measurement results evaluated. High quality signal pick-offs always specify insertion loss, corruption factors and probe output scale factor.

The preceding emphasizes vigilance in designing and maintaining a signal path. Skepticism, tempered by enlightenment, is a useful tool when constructing a signal path and no amount of hope is as effective as preparation and directed experimentation.

Note 1. See Appendix D, "About Z_0 Probes".

APPENDIX F

Breadboarding, Layout and Connection Techniques

The measurement results presented in this publication required painstaking care in breadboarding, layout and connection techniques. Nanosecond domain, high resolution measurement does not tolerate cavalier laboratory attitude. The oscilloscope photographs presented, devoid of ringing, hops, spikes and similar aberrations, are the result of a careful breadboarding exercise. The breadboard required considerable experimentation before obtaining a noise/uncertainty floor worthy of the measurement.

Ohm's Law

It is worth considering that Ohm's law is a key to successful layout[1]. Consider that 10mA running through 1Ω generates 10mV—twice the measurement limit! Now, run that current at 1 nanosecond rise times ($\approx$350MHz) and the need for layout care becomes clear. A paramount concern is disposal of circuit ground return current and disposition of current in the ground plane. The impedance of the ground plane between any two points is *not* zero, particularly at nanosecond speeds. This is why the entry point and flow of "dirty" ground returns must be carefully placed within the grounding system.

A good example of the importance of grounding management involves delivering the input pulse to the breadboard. The pulse generator's 50Ω termination *must* be an in-line coaxial type. This ensures that pulse generator return current circulates in a tight local loop at the terminator, and does not mix into the signal plane. It is worth mentioning that, because of the nanosecond speeds involved, inductive parasitics may introduce more error than resistive terms. This often necessitates using flat wire braid for connections to minimize parasitic inductive and skin effect-based losses. Every ground return and signal connection in the entire circuit must be evaluated with these concerns in mind. A paranoiac mindset is quite useful.

Shielding

The most obvious way to handle radiation-induced errors is shielding. Determining where shields are required should come *after* considering what layout will minimize their necessity. Often, grounding requirements conflict with minimizing radiation effects, precluding maintaining distance between sensitive points. Shielding is usually an effective compromise in such situations.

A similar approach to ground path integrity should be pursued with radiation management. Consider what points are likely to radiate, and try to lay them out at a distance from sensitive nodes. When in doubt about odd effects, experiment with shield placement and note results, iterating towards favorable performance[2]. *Above all, never rely on filtering or measurement bandwidth limiting to "get rid of" undesired signals whose origin is not fully understood.* This is not only intellectually dishonest, but may produce wholly invalid measurement "results," even if they look pretty on the oscilloscope.

Connections

All signal connections to the breadboard must be coaxial. Ground wires used with oscilloscope probes are forbidden. A 1" ground lead used with a 'scope probe can easily generate large amounts of observed "noise" and seemingly inexplicable waveforms. Use coaxially mounting probe tip adapters[3]!

Note 1. I do not wax pedantic here. My guilt in this matter runs deep.
Note 2. After it works, you can figure out why.
**Note 3.** See Reference 34 for additional nagging along these lines.

APPENDIX G

How Much Bandwidth is Enough?

Accurate wideband oscilloscope measurements require bandwidth. A good question is just how much is needed. A classic guideline is that "end-to-end" measurement system rise time is equal to the root-sum-square of the system's individual component's rise times. The simplest case is two components; a signal source and an oscilloscope.

Figure G1's plot of $\sqrt{signal^2 + oscilloscope^2}$ rise time versus error is illuminating. The figure plots signal-to-oscilloscope rise time ratio versus observed rise time (rise time is bandwidth restated in the time domain, where:

$$\text{Rise Time (ns)} = \frac{350}{\text{Bandwidth (MHz)}})$$
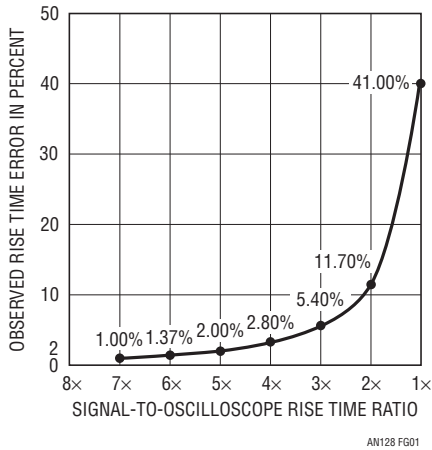
Figure G1. Oscilloscope Rise Time Effect on Rise Time Measurement Accuracy. Measurement Error Rises Rapidly as Signal-to-Oscilloscope Rise Time Ratio Approaches Unity. Data, Based on Root-Sum-Square Relationship, Does Not Include Passive Probe, Which Does Not Follow Root-Sum-Square Law

The curve shows that an oscilloscope 3 to 4 times faster than the input signal rise time is required for measurement accuracy inside about 5%. This is why trying to measure a 1ns rise time pulse with a 350MHz oscilloscope (t_{RISE} = 1ns) leads to erroneous conclusions. The curve indicates a monstrous 41% error. Note that this curve does not include the effects of passive probes or cables connecting the signal to the oscilloscope. Probes do not necessarily follow root-sum-square law and must be carefully chosen and applied for a given measurement. For details, see Appendix D. Figure G2, included for reference, gives 10 cardinal points of rise time/bandwidth equivalency between 1MHz and 5GHz.

RISE TIME	BANDWIDTH
70ps	5GHz
350ps	1GHz
700ps	500MHz
1ns	350MHz
2.33ns	150MHz
3.5ns	100MHz
7ns	50MHz
35ns	10MHz
70ns	5MHz
350ns	1MHz

Figure G2. Some Cardinal Points of Rise Time/Bandwidth Equivalency. Data Is Based on Text's Rise Time/Bandwidth Formula

APPENDIX H

### Verifying Rise Time and Delay Measurement Integrity

Any measurement requires the experimenter to insure measurement confidence. Some form of calibration check is always in order. High speed time domain measurement is particularly prone to error and various techniques can promote measurement integrity.

Figure H1's battery-powered 200MHz crystal oscillator produces 5ns markers, useful for verifying oscilloscope time base accuracy. A single 1.5V AA cell supplies the LTC3400 boost regulator, which produces 5V to run the oscillator. Oscillator output is delivered to the 50Ω load via a peaked attenuation network. This provides well defined 5ns markers (Figure H2) and prevents overdriving low level sampling oscilloscope inputs.

Once time base accuracy is confirmed it is necessary to check rise time. The lumped signal path rise time, including attenuators, connections, cables, probes, oscilloscope and anything else, should be included in this measurement. Such "end-to-end" rise time checking is an effective way to promote meaningful results. A guideline for insuring accuracy is to have 4x faster measurement path rise time than the rise time of interest. Thus, verifying the sample gate multipliers 250MHz (1.4ns risetime) bandwidth requires 1GHz (t_{RISE} = 350ps) oscilloscope bandwidth. Verifying the oscilloscope's 350 picosecond rise time, in turn, necessitates a 90 picosecond rise time step to ensure the 'scope is driven to its rise time limit. Figure H3 lists some very fast edge generators for rise time checking[1]. The Tektronix 284, specified at 70ps rise time, was used to check 'scope rise time. Figure H4 indicates 350ps rise time, promoting measurement confidence.

Note 1. This is a fairly exotic group, but equipment of this caliber really is necessary for rise time verification.
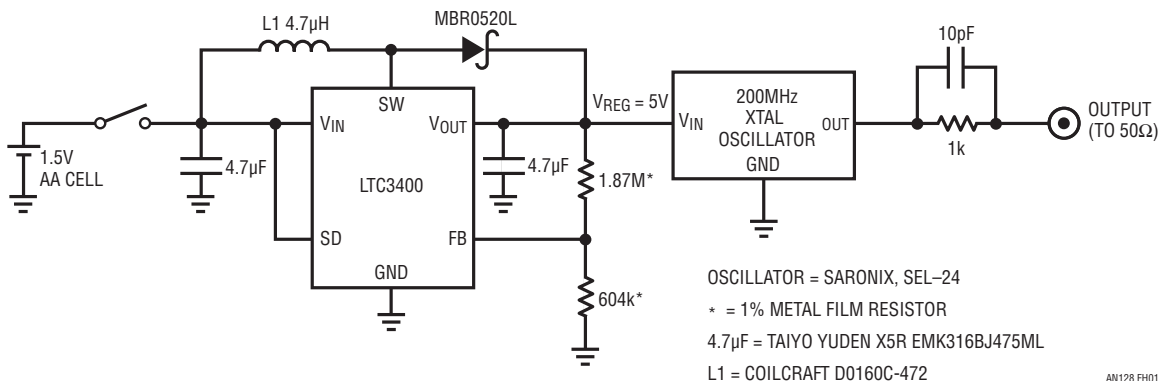

Figure H1. 1.5V Powered, 200MHz Crystal Oscillator Provides 5ns Time Markers. 1.5V to 5V Switching Regulator Powers Oscillator
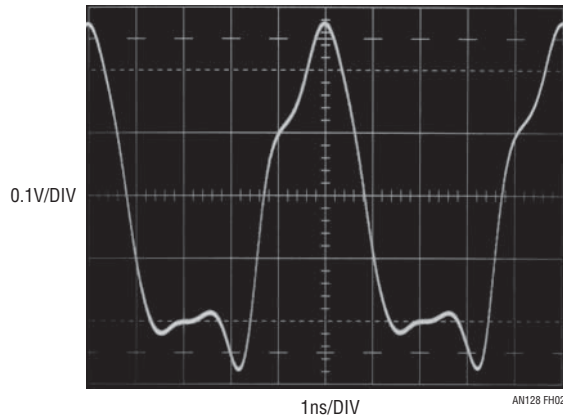
**Figure H2. Time Mark Generator Output Terminated into 50Ω.
Peaked Waveform Is Optimal for Verifying Time Base Calibration**

0.1V/DIV

1ns/DIV

AN128 FH02

MANUFACTURER	MODEL NUMBER	RISE TIME	AMPLITUDE	AVAILABILITY	COMMENTS
Avtech	AVP2S	40ps	0V to 2V	Current Production	Free Running or Triggered Operation, 0MHz to 1MHz
Hewlett-Packard	213B	100ps	≈175mV	Secondary Market	Free Running or Triggered Operation to 100kHz
Hewlett-Packard	1105A/1108A	60ps	≈200mV	Secondary Market	Free Running or Triggered Operation to 100kHz
Hewlett-Packard	1105A/1106A	20ps	≈200mV	Secondary Market	Free Running or Triggered Operation to 100kHz
Picosecond Pulse Labs	TD1110C/TD1107C	20ps	≈230mV	Current Production	Similar to Discontinued HP1105/1106/8A. See above.
Stanford Research Systems	DG535 OPT 04A	100ps	0.5V to 2V	Current Production	Must be Driven with Stand-alone Pulse Generator
Tektronix	284	70ps	≈200mV	Secondary Market	50kHz Repetition Rate. Pre-trigger 5ns, 75ns or 150ns Before Main Output. Calibrated 100MHz and 1GHz Sine Wave Auxiliary Outputs.
Tektronix	111	500ps	≈±10V	Secondary Market	10kHz to 100kHz Repetition Rate. Positive or Negative Outputs. 30ns to 250ns Pre-trigger Output. External Trigger Input. Pulse Width Set with Charge Lines
Tektronix	067-0513-00	30ps	≈400mV	Secondary Market	60ns Pre-trigger Output. 100kHz Repetition Rate
Tektronix	109	250ps	0V to ±55V	Secondary Market	≈600Hz Repetition Rate (High Pressure Hg Reed Relay Based). Positive or Negative Outputs. Pulse Width Set by Charge Lines

Figure H3. Picosecond Edge Generators Suitable for Rise Time Verification. Considerations Include Speeds, Features and Availability

50mV   200pS

VERT = 50mV/DIV

HORIZ = 200ps/DIV

AN128 FH04
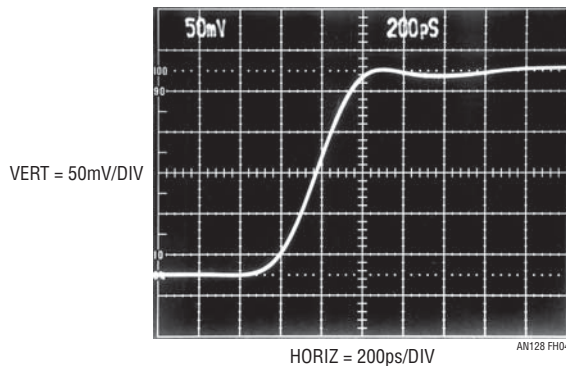
**Figure H4. 70 Picosecond Edge Drives Oscilloscope to its 350
Picosecond Rise Time Limit, Verifing 1GHz Bandwidth**

Contemplating Nanosecond Settling Time Measurement at Punta Cana, Dominican Republic, where this Application Note was written