Avoiding Performance Pitfalls for the Pulse Response of High-Speed Op Amps

Part 1: Where the various estimates of rise time, slew rate, and full power bandwidth come from and why they might be misleading

by Michael Steffes

Time Domain Specifications in Wideband Amplifier Datasheets

As a starting point, it is critical to recognize that a dc-coupled signal path will probably require an op amp solution. Applications that are more frequency domain in nature can consider the large universe of RF and IF amplifiers available. Those devices are intended for ac-coupled, generally narrowband, applications where a step response plot would be pretty rare. Wideband op amps have improved to the point of providing an attractive option to these more classically RF-based designs. This is particularly true in the realm of dynamic range vs. quiescent power. If a true dc-coupled signal path is needed, a device that offers a frequency span all the way down to dc is the best option. Comparing the useable frequency span for any op amp vs. any RF or IF amplifier (on a log frequency scale) will typically show the op amp running out of steam on the high end, but extending an infinite number of decades down in frequency (to dc) vs. the RF or IF amplifiers. Much of the discussion here will apply to op amps of any speed range, with an increasing focus on high-speed devices in later sections. But the basic theoretical and industry practices discussion applies to both low speed, very high dc precision, devices as well as the higher speed (>100 MHz) dc-coupled op amps.

Way back in the mid to late 1980s, our Comlinear design and applications team was tasked with the job of coming up with characterization and specification techniques for amplifiers that were totally new and different from the classical VFB (voltage-feedback) devices. Building on a lot of earlier work among the op amp suppliers, we essentially tried to be consistent with industry best practices while highlighting the new characteristics offered by the CFB (current-feedback) topology. Fortunately, most of the design staff were former HP (now Agilent) test instrument designers, so our lab equipment and techniques were outstanding.

Most of the data sheet focus (then and now) was on the frequency domain characterization and that has rippled on to the other suppliers that came later to the high-speed op amp market. Small-signal frequency response vs. gain, frequency response vs. signal level, intermodulation distortion, noise, etc. Typically less than 20% of a wideband amplifier data sheet addresses time domain issues like pulse response, overdrive recovery, settling tails, etc. There is a sort of a hidden assumption in the data sheets that if you know the frequency response, you should be able to get to the pulse response. In retrospect, that might be a misplaced assumption with added confusion arising from occasionally different assumptions on some of these specifications between the different suppliers.

For example, one of the very common assumptions I used in reporting the rise time (10% → 90%) in a high-speed amplifier data sheet was that it will be equal to $0.35/F_{-3\text{ dB}}$. This does in fact pretty closely match what is normally measured for a (non slew-limited) pulse response. But this simple expression comes directly out of the analysis for a single-pole step response, when a wideband amplifier is almost never a single-pole response – so was this luck or what? (Part 2 will answer this in detail, but luck appears to be a strong contender). Other suppliers might nominally specify a very peaked frequency response and then report a measured or simulated rise time which, in that case, would be a much lower number (for the same bandwidth) than $0.35/F_{-3\text{ dB}}$. In other cases, you might see a *rise time* specified for what is clearly a slew-rate-limited pulse response.

More fundamentally, what is a designer trying to find out in looking at the time domain specifications of an amplifier? One of the really interesting things about op amp specifications (vs. data converters for instance), is that they are normally used to assess how much design margin over your signal characteristics the device offers. Whereas the principal specifications for an ADC (bits and speed) are used with the idea that the design will in fact use all of the bits and clock at (or near) the maximum clock rate, many of the amplifier specifications are written with the idea that the application should not approach these specified limits, or non-linear operation will result. For instance, it is rarely (probably never) the case that a designer specs in an op amp intending to operate it under full-power bandwidth conditions. There is an overriding assumption in many of the specifications that these are limits that are best avoided at all costs. Many times, the designer is most interested in the settling time to some accuracy, and some devices only specify that with no reported rise time specification. A correct understanding of settling time issues requires a good understanding of rise time and slew rate limits. This current set of articles focuses on those issues as a necessary, but not sufficient, condition for fast (and controlled) settling time while leaving the numerous details of a good settling time to a later time.

Op amp specifications are often used with the idea that the amplifier should impair the incoming waveform by an acceptably small amount while providing the desired gain, level shifting, filtering, or whatever specific operation desired. The most common test waveform for this in the time domain is a square wave input (since a true impulse response that would directly expose the amplifier's characteristic response is a little tough to generate). If the amplifier does not go into slew-limited operation, one of the most common approximations is that the output rise time (10% → 90%) will be the rss (Root Sum of Squares) of the input signal rise time and the amplifier's rise time:

$$t_R(output) = \sqrt{t_R^2(input) + t_R^2(amplifier)} \qquad \text{Eq 1}$$

For instance, if we want the output rise time to be reduced by no more than 1% of the original input signal, then this equation may be solved (Eq 2) for the maximum amplifier rise time that will meet this target. Restating this in terms of rise time ratios, where,

$\alpha \equiv \dfrac{t_R(output)}{t_R(input)}$ and $\beta \equiv \dfrac{t_R(amplifier)}{t_R(input)}$ then a required β from a target α may be resolved:

$$\beta = \sqrt{\alpha^2 - 1} \qquad \text{Eq 2}$$

Putting in $\alpha = 1.01$ will give a $\beta = 0.142$. If we assume for the moment that the amplifier rise time is linearly related to the $f_{-3\text{ dB}}$ bandwidth of the amplifier, then this suggests that the amplifier needs to be $1/\beta = 7.04$X the implied bandwidth of the input signal to degrade the output rise time by no more than 1% from the input rise time. This begs the question of why designers don't just specify amplifiers of vastly more bandwidth than the signal path requires. The immediate performance tradeoff limiting this approach is a much wider noise power bandwidth that will then degrade the output SNR. The more practical limit is that excessive speed for a particular requirement will always be more expensive in power and price.

Historical Perspective on Slew Rate

Op amp specifications evolved around the particular type of amplifier available in the earliest days of op amp development. This was most often a unity-gain stable, voltage feedback, design with what is sometimes called a *long-tailed pair* input stage. This input design was essentially a differential pair biased by a fixed-tail current source. Designs have moved far beyond this today and it has become quite a stretch to apply some of the specification concepts developed in the early 1970s to amplifiers being introduced in the new millennium (more on this in Part 4).

In reviewing the early op amp literature (late 1960s to early 1970s), you are staggered by the clarity and insight some of these early publications provided. I am reminded again that we are all standing on the shoulders of those who have come before us, and this is particularly true here. It is always necessary to select a finite subset of sources, and each author finds their preferred sources somewhat by chance. The ones cited here seem pretty good, but I am sure there are many others that are just as good (if not better) that I simply have not had time to search out. **The central topic in this current set of articles is to correctly anticipate the amplifier linear pulse response while considering any non-linear limits that might impair this expected response.** This issue was considered at the dawn of the op amp era in (at least) the following 3 sources:

1. *User's guide to Applying and Measuring Operational Amplifier Specifications*, by Ray Stata, Analogue Dialogue, Volume 1, Number 3 pages 3 & 4, 1967
2. *Burr-Brown Operational Amplifiers – Design and Applications* (book), by Jerald Graeme, Gene Tobey, and Dr. Lawrence Huelsman, 1971 pp 451 - 452
3. *Predicting Op Amp Slew Rate Limited Response*, by Marvin Vander Kooi, National Semiconductor application note LB-19, 1972

Several of these cite earlier sources, but since each of these were produced in the context of supplying op amp devices to the industry (as opposed to academic sources), they seem to have set the direction for later op amp data sheets. Each provides a snapshot of what the thinking was at the time and strongly influenced what has come later.

The first source, written by the founder of Analog Devices, was intended to codify what ADI meant by the numbers in their data sheets. Pulling Fig. 9 from this source shows a number of interesting features.

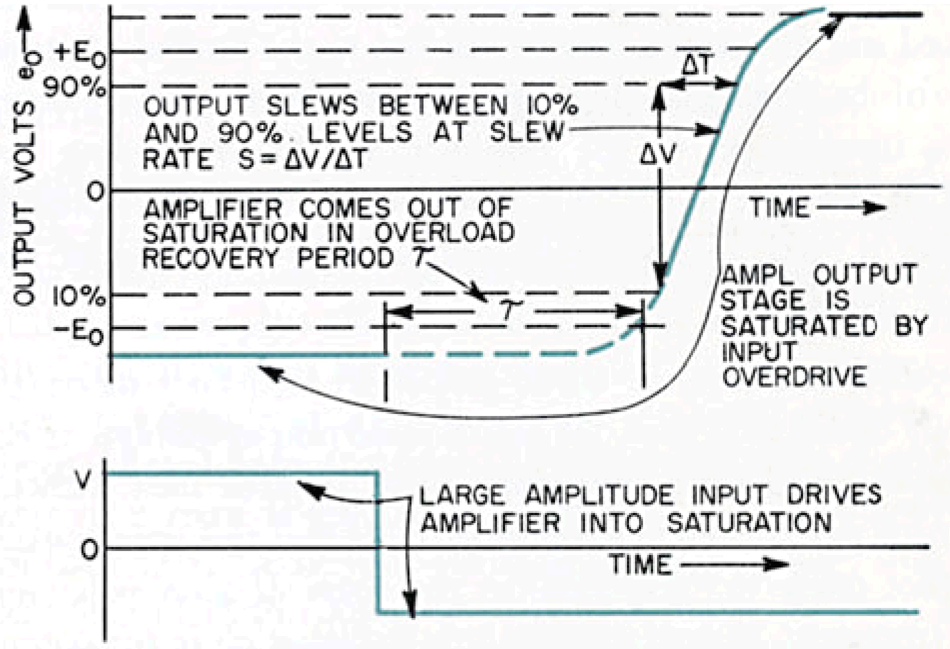

FIGURE 9. OVERLOAD RECOVERY AND SLEW RATE ILLUSTRATION

This original test was combined with an overdrive condition. The τ shown above was the delay time for the amplifier to come out of output saturation and then transition as fast as possible to the other clip limit. One of the interesting things here is the slew rate is defined over the 10% to 90% range – a concept picked up in numerous later sources (and possibly just repeated here from earlier work). Although not shown here, this 10% to 90% transition time has also come to be used as the most common rise time specification for a linear (non-slew limited) response. So the two concepts are linked even in this very early source. The text here also has a very good discussion of full-power bandwidth as it relates to slew rate and makes the point that physical slew limits are output stage related. In this early source, Mr. Stata emphasizes that the useable full power bandwidth is set by an acceptable level of signal distortion (not a true -3 dB bandwidth at this time). If the acceptable level is very small, it may be set by mechanisms other than the large signal slew rate defined in his Fig. 9 above. Both tests are performed in either the +1 V/V or -1 V/V gain condition. Other sources emphasize that slew rate mechanisms are input stage related. Eq 1 on page 3 (repeated here as Eq 3) reports the relationship of peak DVDs for a sinusoid that is actually used extensively to predict full-power bandwidth (but not in this source):

$$SR = 2 * \pi * f_{\max} * V_p \qquad \text{Eq 3}$$

where, Vp is the expected peak output excursion of a sinusoidal input signal at $f_{\max}$.

This equation arises from finding the peak dV/dt for a sine wave (at the zero crossing) and provides a way to use the measured SR along with Vp to estimate an f_{max}.

The second reference devotes Appendix B to op amp specification issues. The section on slew rate (pp 451 - 452 in my 1971 copy) emphasizes that the test is performed in a non-inverting unity-gain voltage-follower configuration as that will be the worst case condition. The text also implies a ± output swing to the clipping limits where the slew rate is the slope of the curve between the specified maximum linear output voltage range (not 10% to 90% as shown in the ADI figure, above) It does note asymmetrical slew rates (rising different than falling) are possible where the slower one would be reported. The text goes on to give the same Eq 3 above for the full power bandwidth. By this time, four years after Mr Stata's article, the SR in Eq 3 is the same SR measured by a rail-to-rail output swing fit. The text still uses an *acceptable level of distortion* at f_{max}, so the idea that f_{max} is a -3 dB frequency has not come in quite yet.

The third reference is a very succinct discussion of pulse response issues. In general, I find the early National application notes to be phenomenally useful. A summary excerpt from LB19 is repeated here.

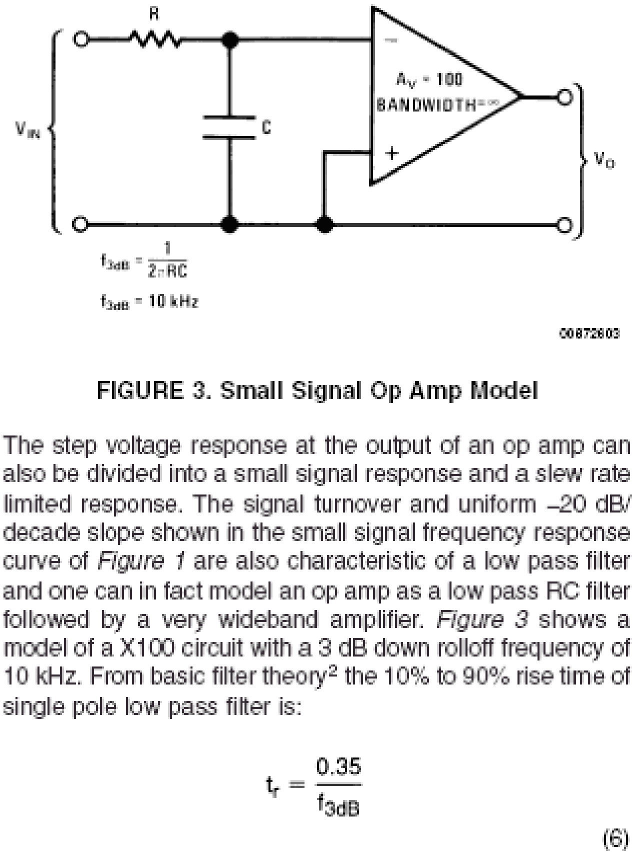

FIGURE 3. Small Signal Op Amp Model

The step voltage response at the output of an op amp can also be divided into a small signal response and a slew rate limited response. The signal turnover and uniform −20 dB/decade slope shown in the small signal frequency response curve of *Figure 1* are also characteristic of a low pass filter and one can in fact model an op amp as a low pass RC filter followed by a very wideband amplifier. *Figure 3* shows a model of a X100 circuit with a 3 dB down rolloff frequency of 10 kHz. From basic filter theory^2 the 10% to 90% rise time of single pole low pass filter is:

$$t_r = \frac{0.35}{f_{3dB}} \qquad (6)$$

What the author is saying here is that a slew limit occurs when the amplifier can no longer follow a single pole response having a rise time equal to $0.35/f_{3dB}$. As will be shown later, this equation is the 10% to 90% rise time of a single-pole response. He goes on to divide this t_r into the step size and compare it to an intrinsic slew rate limit as repeated here:

$$\frac{V_{STEP}\, f_{3db}}{0.35} \geq S_r$$

One slight error here is that while the rise time is implied to be the 10% to 90% level, the full V_{step} is used in the equation. That might be an intentional guardband (it will predict a signal dV/dt 20% higher than the actual), but to be consistent, this expression should have been what is shown as Eq 4 (where I have also substituted the more common $f_{-3\ dB}$ instead of the original f_{3dB} which today implies a +3 dB peaking in the response):

$$\text{Required } Sr = \frac{0.8 * V_{STEP} * f_{-3dB}}{0.35} \qquad \text{Eq 4}$$

It is interesting that this same error (interpreting the rise time as being for the full step size) shows up later in the most commonly used underdamped 2nd order response equations (more on that in part 2).

This author also discusses full power bandwidth issues in the context of the signal becoming distorted and repeats Eq 3 above.

Several basic ideas come out of these early documents:

1. The slew rate can be measured as some line fit to a non-linear output response to a square wave input over some span (usually the 10% to 90% range)
2. This same mechanism causes distortion at some frequency and output amplitude that requires that dV/dt at the zero crossing
3. The documents seem to imply that a single slew rate number (or mechanism) can be used under both conditions

Other sources emphasize that the non-linear limit to a large input step happens at the input stage for a unity gain follower application. This makes a lot of sense for these early amplifiers as a simple differential pair becomes fully switched to one side providing only the fixed-tail current as a charging current to the internal compensation cap. This suggests a reference to the first of several textbooks. Again, these just happen to be the ones I am most familiar with and I am sure there many others that address these issues in a similar fashion. The following text (and I happen to have a well used 1977 edition) covers the slew rate mechanism for a voltage feedback amplifier:

Analysis and Design of Analog Integrated Circuits Paul Gray and Robert Meyer, 1977, pp 541 - 551

In this simple case, they show the amplifier's slew rate to be:

$$\frac{dV}{dt} = \frac{2 * I_1}{C_{comp}}$$

where, I_1 = the transistor quiescent collector current on each side of the input differential pair when they are balanced.

The text continues with a very good discussion of increasing the slew rate by reducing the input stage dc gain – normally through input stage emitter degeneration. This allows the compensation capacitor to be reduced and still end up on the same open loop roll-off curve at high frequencies to satisfy stability issues. This particular approach, and many like it, pays the price of increased slew rate with higher input voltage noise. A great deal of IC op amp design revolves around trading off noise and full-power bandwidth with different topologies, technologies and quiescent power levels. This text also shows the same equation for large signal, slew limited, bandwidth that was shown earlier as Eq 3 but continues to consider it a frequency at which gross output distortion will occur.

As the literature and products progressed, the guardband to slew-limited operation for low distortion increased significantly, (ie operation far below the slew-limited condition is required for low distortion) and the full-power bandwidth number has come to mean a –3 dB estimate at (usually, but not always) the maximum available linear output swing.

Updated View of Pulse Response and Slew Rate Related Issues

Having the benefit of 35 some years of op amp introductions since these early definitional articles, it might be possible to refine their treatment. A good starting point would be a linear pulse response discussion. The first approximation to an op amp pulse response has always been a single-pole low-pass filter response where this single closed-loop pole is set by the amplifier type and model along with the external connections. It is almost never the case that a single-pole response is the actual frequency response in a high-speed application. But as a starting point, and for lower speed op amps, it is important to understand. For a voltage-feedback design, the closed-loop bandwidth is approximately the gain-bandwidth product divided by the noise gain (which is most often the non-inverting signal gain). That gain can always be set high enough to get the closed-loop amplifier to behave like a single-pole response. For a current-feedback design, the closed-loop bandwidth will be set by the amplifier chosen and the value of the feedback resistor. (see Reference 1 for a recent review of the internal workings and pros and cons of voltage-feedback vs. current-feedback amplifiers). Increasing the feedback resistor from the nominally-specified value (which is normally set to get a closed-loop 2nd order Butterworth frequency response) will move the small-signal frequency response towards a single pole.

When a step input is applied to a closed-loop op amp, it is normally preferable that the output response remain linear. If it does, the 10% → 90% transition time will remain a constant value with step size. **Linear response by definition has a constant rise time for any step size**. For a given amplifier frequency response, linear operation with increasing step size will be producing an increasing dV/dt at the output. This is for either the peak point slope or the more commonly used best line fit over the 10% to 90% range. If the step size is increased to the point that this peak (or average) dV/dt exceeds the amplifiers internal capability, the response will slip over into a slew-limited, non-linear, response. Going slew limited in the output response is essentially opening the feedback loop during the amplifier output transition to a final value. For a designer, slipping into slew-limited response can be detrimental for a couple of reasons. Often, this will result in more pulse overshoot than a linear analysis would predict. Secondly, once the loop opens up, it is almost certain that the final settling time will be significantly increased. So while every output step has some implied peak or average dV/dt, allowing this to exceed the maximum available dV/dt for that amplifier changes the pulse response details significantly.

The single-pole pulse response is covered in numerous text books. The one I like as an introductory text to a large number of op amp applications is:

Design with Operational Amplifiers and Analog Integrated Circuits (2nd Edition), Sergio Franco, 1998

The first-order pulse response and slew rate issues appear on pages 278 - 280. Given a Laplace frequency response described as a first-order low-pass pole:

$$H(s) = \frac{\omega_o}{s + \omega_o} \qquad \text{Eq 5}$$

where, $\omega_o = 2 * \pi * f_o$

We know in this case that the small signal frequency response has an $f_{-3dB} = f_o$

Applying a step input of amplitude A to this gives the familiar first-order pulse response:

$$V(t) = A * (1 - e^{\frac{-t}{\tau}}) \qquad \text{Eq 6}$$

where, τ = time constant = $\frac{1}{2 * \pi * f_o}$

An example plot of Eq 6 is shown as Fig. 1 where, in this case, f_o = 100 MHz (2π) and a 1 V step are shown.

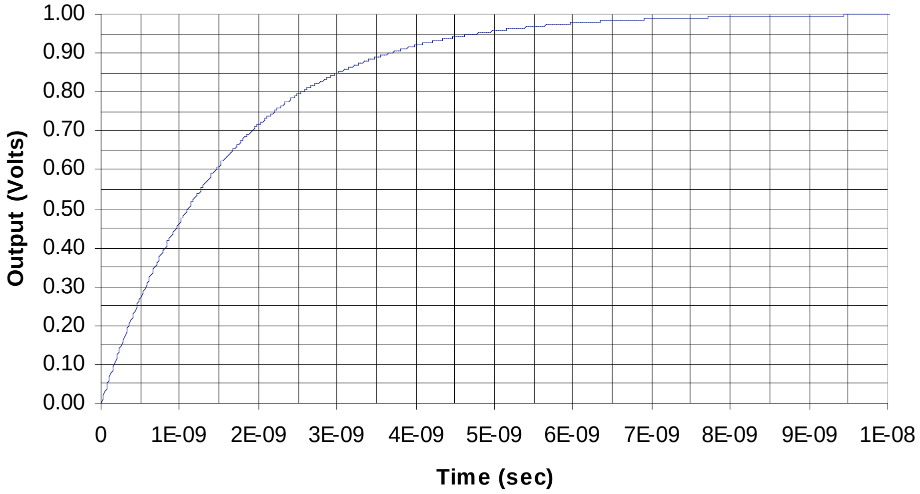

Fig. 1: Single-Pole Step Response

Several important concepts come from this simple equation. First the dV/dt may be solved by taking the first derivative of Eq 6 as shown in Eq 7:

$$\frac{dV}{dt} = \frac{A}{\tau} * e^{\frac{-t}{\tau}} \qquad \text{Eq 7}$$

By inspection, this point slope has a maximum value of A/τ at t = 0 and then decays to zero as $t \to \infty$ (as reported in Dr Franco's text).

It is also very easy to solve Eq 6 for specific points on the exponential waveform. Specifically, if we solve for a portion of the final value, A, by targeting a solution for t to get α*A, we will get Eq 8:

$$t = -\tau * \ln(1-\alpha) \qquad \text{Eq 8}$$

Then, if we solve for an $\alpha_1 = 0.1$ and $\alpha_2 = 0.9$ to give a t_1 and t_2, the 10% to 90% rise time can be solved as:

$$t_2 - t_1 = -\tau \ln(1-\alpha_2) + \tau \ln(1-\alpha_1) = \tau \ln\left(\frac{1-\alpha_1}{1-\alpha_2}\right) = 2.2\tau \qquad \text{Eq 9}$$

Recalling that $\tau = \frac{1}{2 * \pi * f_o}$ and that $f_{-3\ \text{dB}} = f_o$ for the single pole case, we get:

$$\text{Single pole rise time } t_r = t_2 - t_1 = \frac{2.2}{2 * \pi * f_{-3dB}} = \frac{0.35}{f_{-3dB}} \qquad \text{Eq 10}$$

This matches the rise time reported in Ref 3 from 1972, shown above. Even for this simple single-pole response, there is inconsistency in the literature on the required dV/dt for slew rate considerations. LB-19 reports a line fit to the 10% to 90% points (Eq 4) as the output dV/dt to use, while Dr Franco (along with most introductory texts) reports the t = 0 peak dV/dt = A/τ as the value to compare to the reported slew rate. Fig. 2 repeats Fig. 1 with these two possibilities superimposed as straight lines.

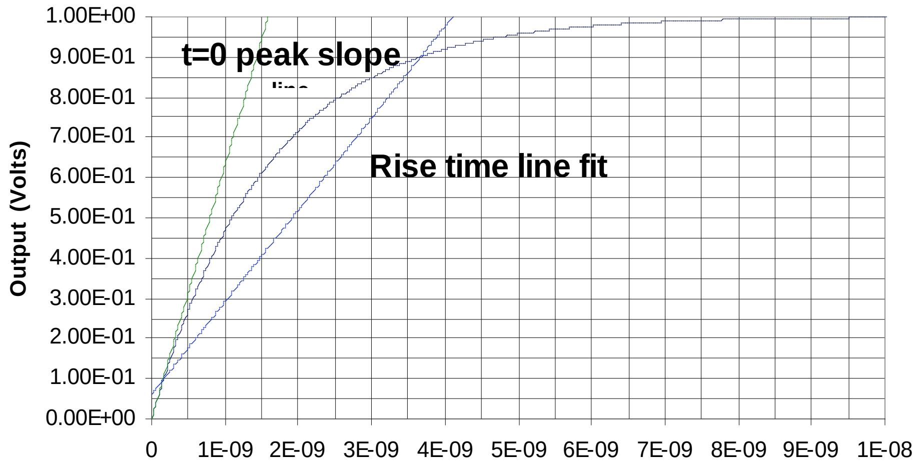

Fig. 2: Single-Pole Response With Slope Lines

Obviously this discrepancy comes from considering either the peak-point slope or the average slope between two points as the dV/dt of interest for slew rate considerations. This relatively simple issue is important to understand as it will recur in the 2nd-order analysis of Part 2. For this example (100 MHz single-pole response) the two slopes are:

t = 0 peak dV/dt = 1 V/1.59 ns = 628 V/µs
rise time dV/dt = 0.8 V/3.5 ns = 229 V/µs

Clearly a huge difference – so which one is right? Well, in this case, both should be interpreted as a simplified theoretical analysis since no real amplifier has a single-pole response. As will be shown in Part 2, the moment you add another pole, even a real pole at much higher frequencies than the dominant closed loop pole, the dV/dt at t = 0 drops to zero. So the simple equation shown in Dr Sergio's book (the t = 0 peak dV/dt = A/τ), while very common, gives a relatively high estimate of peak dV/dt to use in comparing to real amplifier slew rate limits. It probably should be interpreted as a very conservative estimate of required slew rate to support a desired output step size. As you move up in amplifier speed, this sort of conservativism (>2x in this example) might lead you to specifying an amplifier slew rate far higher (and much more expensive) than actually required. So the aim in Part 2 is to iterate closer to an actual dV/dt for real high-speed amplifiers to use in this analysis.

Practical Considerations in Reported Rise Times, Slew Rates, and Full Power Bandwidth

High-speed amplifier data sheets focus on the frequency response. The reported rise time is either calculated from the $f_{-3\text{ dB}}$ bandwidth and/or taken off a non-slew limited pulse response. Sometimes, the rise time specification is for a slew-limited response and then bears little resemblance to the linear analysis shown here. Hopefully, the specified rise time is close to what a pulse response plot shows under the same conditions. Unfortunately, the small-signal pulse response is often shown on such a long time scale that no real extraction of rise time is possible from the plots.

For a 2nd-order maximally-flat frequency response (Butterworth) it will turn out (in Part 2) that its 10% to 90% rise time is very nearly equal to $0.35/f_{-3\text{ dB}}$. For data sheets that report a -3 dB bandwidth for this type of frequency response, you would expect the specified rise time to be very close to this number if the specified step size does not go into slew limit. Occasionally, individual amplifier data sheets might use a different definition for rise time: sometimes a 20% to 80% span or a 25% to 75% span. It is easy to compute that rise time for a first-order response by letting $\alpha_1 = 0.2$ and $\alpha_2 = 0.8$ and then $\alpha_1 = 0.25$ and $\alpha_2 = 0.75$ in Eq. 9:

This gives $t_r = 1.4\tau = \dfrac{0.22}{f_{-3dB}}$ (a 20% to 80% single-pole rise time) Eq 11

Or, $t_r = 1.1\tau = \dfrac{0.175}{f_{-3dB}}$ (for a 25% to 75% single-pole rise time) Eq 12

If you see a rise time reported that seems much shorter than a $0.35/f_{-3\ dB}$ relationship would predict, try Eqs 11 or 12 to see if a narrower part of the output step is being used. The calculation above only applies to a single-pole response and the actual modification for a 2nd-order response will be developed in Part 2. The other possibility is that the small-signal frequency response is peaking quite a bit for the specified pulse response test. That will always give a faster rise time at the cost of more overshoot and ringing.

The overriding theme of this article is that a wideband amplifier will give you a linear output step response until the output dV/dt exceeds an internal limit – historically that limit has been called the slew rate. This number has normally been interpreted as a best line fit over a slew limited output pulse or, more recently, computed from the full-power bandwidth equation. Sometimes, it is reported for rising and falling edges separately and occasionally for different amplifier configurations (typically, +1 and -1 are pretty popular). In the very early days, this slewing rate also assumed that the amplifier was driven into its output swing limits. Over time, that idea has faded and it seems most data sheets are avoiding hitting the swing limits as part of this test. In general, for high-speed amplifiers, running the output into either swing limit is best avoided as very little of the data sheet applies under those conditions.

Given the close coupling between the very common 10% to 90% slope definition for slewing conditions, and the fact that linear rise time is also normally specified for that swing, it is not too surprising that the two concepts are sometimes confused. It is certainly possible to measure the average slope between 10% to 90% swing for a linear pulse response and call that the slew rate. When the amplifier is operating linearly, the rise time is constant, and an increasing output step size will give what appears to be an increasing slew rate if the linear slope is measured. The important distinction is that an output step operating linearly has an average dV/dt that is some portion of an available maximum slew rate for that specific device using the earliest definitions of slew rate available. As you increase the step size, the portion of the available maximum being used will increase until you hit the limit. Slew rate is non-linear phenomena while rise time is normally considered a linear phenomena.

One specific example of this specification convolution can be seen in the TI THS3110 data sheet. Numerous plots of slew rate vs. step size are shown where they all reach a moderately flat limit at higher step sizes. This flat limit is what is normally considered a slew rate limit while the rising portion of the curve is the fixed rise time of linear operation divided into an increasing step size. This might be just semantics, but this curve could also be misinterpreted as an amplifier with a low intrinsic slew rate but a very robust boost circuit that depends on step size. Fig. 3 shows an example from this data sheet. Where the curves go flat is the more widely recognized definition of slew rate.

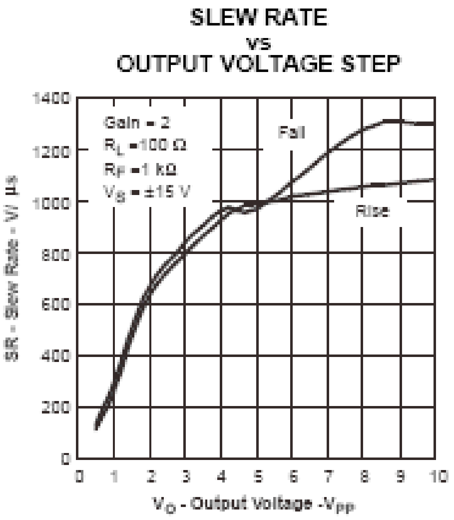

Fig. 3: THS3110 Slew Rate Vs. Output Step

As a final note, on getting to the real slew rate limits of an op amp, it is useful to know that the original full-power bandwidth equation has become pretty ambiguous as well. The earliest literature intended this bandwidth to indicate the onset of gross distortion (visible in the time domain, so >2% or about -36 dB THD). I believe the idea from the earliest literature was that, if you measure a rail to rail output swing (with a square wave input) for its transition slope and call that the slew rate, you can put that number into Eq 3 and get an estimate of the highest frequency of operation for a full-scale linear output swing. As will be shown in later parts, the maximum operating frequency for a targeted level of distortion is much lower than this equation would suggest when the desired output distortion moves into the -60 dBc (and lower) levels.

The earliest literature used *bandwidth* not as a -3 dB point on a full output swing swept sinusoidal test, but rather as the frequency where you knew the zero crossing was asking for a dV/dt that exceeded the maximum available slew rate. Over time, as full-power (or large-signal) frequency responses became commonly shown in high-speed data sheets, their -3 dB point came to be called the full-power bandwidth if the test condition was at or near the maximum available unclipped output voltage swing. If we took a full-power frequency response plot and tried to apply the original definition of *acceptable* level of distortion, we would first have to define that level today, then step back significantly from the measured $f_{-3\ dB}$ point to the useable bandwidth for distortion purposes.

While the FPBW vs. SR expression does not really give much information in the context that it was originally intended, it actually has been used in backing out an imputed slew rate for devices that don't show a classical slew limited response (eg current-feedback and some VFB designs). That expression, reported as Eq 3 earlier on and repeated here:

$$SR = 2 * \pi * f_{\max} * V_p \qquad \text{Eq 3}$$

has been used extensively to find a SR by applying a large-signal, swept-frequency output and looking for the $f_{-3\ dB}$ bandwidth. Normally, the FPBW (full power bandwidth) is defined as an output swing just below the clip limits and then swept up until a -3 dB point is found. The idea here is that the peak dV/dt for a sine wave occurs at the zero x-ing and once the output is demanding more dV/dt at the zero crossing than the amplifier has, the signal will distort and the amplitude decrease. The details of that distortion depend on the specific amplifier but another way to think about it is in the Fourier series domain where slew-limited operation is moving power into the harmonics while it is lowered in the fundamental (a network analyzer sweep of s_{21} for bandwidth is only measuring power in the fundamental frequency).

Anyway, when a FPBW or large signal sweep has reached a -3 dB point, it could be argued that the Vp for this equation needs to be reduced to 0.707 Vp where the original Vp was the lower frequency output swing before the onset of some slew-limiting mechanism. After all, the actual output power at the fundamental frequency has been reduced by -3 dB (and moved to the harmonics, so it's hard to say what the exact distorted waveform looks like without going over to the scope). Making that adjustment gives:

$$SR = 2 * \pi * f_{\max} * 0.707 * V_p \qquad \text{Eq 13}$$

These are very approximate relationships. But, in any case, making this adjustment does seem to match the measured data better on many amplifiers. For example, looking at the data sheet for the National LM6642 (using the single +5 V spec table), we see a nominal specified slew rate of 125 V/μs (coming probably from a pulse response measurement but it is hard to tell since the time scale is pretty course on those plots) and then a reported large-signal (but not full-power) 2 Vpp output bandwidth of 22 MHz. It is likely that number came from plugging into Eq 3 where, putting in the numbers gives an f_{max} = 20 MHz.

Since this original definition of f_{max} was intended as a course distortion point, it is hard to know what to do with it now. Plugging into Eq 13 gives 28 MHz and looking at the measured plots, it does in fact show a 2 Vpp output $f_{-3\,dB}$ point very close to 30 MHz. Since we can't really use Eq 3 in the sense that it was originally intended, Eq 13 will give better agreement between the plots of large-signal responses and reported large-signal bandwidth where, now, the idea of bandwidth is not the original *acceptable* distortion idea, but the actual $f_{-3\,dB}$ frequency. For a more detailed discussion of this issue, see Ref 2 where part of this web cast was devoted to mapping between FPBW and slew rate.

Conclusions

If you find that you have been a little confused about the relationships between rise time, slew rate and FPBW, you have good reason to be. The ideas have been mixed together in different ways by numerous credible sources over the years. Without a strong underpinning of what is really going on in the output time domain response, it would be very easy to come away with an inaccurate impression of how these terms relate. The starting point for me has always been a good understanding of a linear output response –

2^{nd}-order at minimum. Then a detailed understanding of what internal mechanisms might limit the dV/dt to include slew limiting ideas into a predicted output response.

The key points from this Part 1 discussion are:

1. A linear output response has both a point slope and average slope to its output dV/dt. Different authors use one or the other in talking about slew rate related issues
2. The first order dV/dt = A/τ (at t=0) is a maximum point slope analysis that should be interpreted as a very conservative, but simple, estimate of the output waveform's maximum required dV/dt. Actual linear outputs never reach this high a dV/dt for any 2^{nd} order response condition (shown in Part 2)
3. The mapping from slew rate to full-power bandwidth (FPBW) where this bandwidth is now a real $f_{-3\ dB}$ frequency for what we know will be a slew-limited condition, works better if the Vp is reduced to 0.707 Vp to be consistent with the –3 dB point idea

This first Part has been a pretty broad sweep of the extant literature for pulse response and large-signal swept frequency outputs. Any comments or questions (or references to additional sources that cover this topic) can be addressed to msteffes1@comcast.net. Part 2 in this series will step into a theoretical development of the point and average dV/dt for a 2^{nd}-order low pass system. Since this accurately represents most high-speed amplifiers, it should leave us better prepared to understand slew limiting mechanisms for the different types of high-speed amplifiers.

References

1. *Applications basics when applying wideband voltage and current feedback op amps*, Michael Steffes, Planet Analog, Dec. 2006
2. *Voltage Feedback or Current Feedback Op Amps in High-Speed Applications: Pros, Cons and Design Tricks*, TI web site, E-lab archives, Dec. 2005

About The Author

Since 1985 Michael Steffes has been deeply involved in the design, application and marketing of high-speed amplifiers. Starting as an IC designer at the original current feedback op amp supplier (Comlinear Corporation, acquired by National Semiconductor in 1994), he later managed the applications group writing most of the original data sheets and application notes for Comlinear wideband amplifiers. Taking the strategic marketing role at Burr-Brown Corporation (acquired by Texas Instruments in 2001), he set the product/process roadmap for the high-speed amplifier group while continuing to produce the product data sheets and numerous application articles. With the TI acquisition, Michael's role expanded to include managing product line rationalization and obsolescence issues, while contributing to (and presenting) worldwide technical seminars. After 22 years of handling high-speed amplifier issues 24/7, Michael is currently taking a break from full time employment.

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