A Tutorial on Applying Op Amps to RF Applications

National Semiconductor
OA-11
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With operating frequencies exceeding 300MHz, National's line of monolithic and hybrid current feedback operational amplifiers have become an attractive option for the RF (and IF) design engineer. Typical operational amplifier specifications do not, however, include many of the common specifications familiar to RF engineers. To help the designer exploit the many advantages these amplifiers can offer, this application note will define the RF specifications of most interest to designers, detail what determines each of these particular performance characteristics for National's current feedback op amps, and, where possible, discuss performance optimization techniques. To apply op amps to RF applications, questions in three general areas must be addressed:

1. Setting the op amp's operating conditions
2. Small signal AC performance in an RF context
3. Typical limits to RF amplifier dynamic range applied to op amps

Wherever possible, tested performance using the CLC404 will be used to demonstrate performance. The CLC404 is a ±5V power supply monolithic amplifier intended for use over a voltage gain range of ±1 to ±10. At its optimum gain of +6, the CLC404 offers a DC to 175MHz frequency range while delivering 12dBm power into a 50Ω load while dissipating only 110mΩ quiescent power. National offers a wide range of additional monolithic op amps, as well as higher supply voltage (and hence higher power output) hybrid amplifiers. The best amplifier for a particular application will depend upon the desired gain, power output, frequency range and dynamic range.

Operation of National's Current Feedback Op Amps

The current feedback op amp, developed by National Semiconductor Corporation, provides a very wideband, DC coupled op amp that has the distinct advantage of being relatively gain-bandwidth independent. As with all op amps using a closed loop negative feedback structure, the frequency response for the National op amps is set by the loop gain characteristics. The key development of the National amplifiers is to de-couple the signal gain from the loop gain part of the transfer function.

This de-coupling allows the desired signal gain to be changed without radically impacting the frequency response. If compared to voltage feedback amplifiers, which are constrained to a gain-bandwidth product operation, the current feedback topology offers truly impressive equivalent gain-bandwidth products (e.g. the CLC401 at a gain of 20 yields a flat response with a -3dB bandwidth of 150MHz. To match this, a voltage feedback op amp would require 20*150MHz = 3GHz gain bandwidth product). Please refer to National application note OA-13 for a description of the current feedback op amp topology and transfer function.

One of the big changes in going from a classical RF amplifier to using an op amp is the exceptional flexibility offered by the op amps. The designer is now charged with setting up the proper operating conditions for the op amp, defining the gain, and determining the I/O impedances with external components. Op amps allow the designer the option of running either a non-inverting or an inverting gain path. For RF applications, the 180° phase shift provided by the inverting mode is often incidental. There are, however, advantages and disadvantages to each mode, depending on the desired performance, and both will be considered at each stage in this development.

Most of this discussion on applying op amps to RF applications applies to any type of op amp. The unique advantages of the current feedback topology are its higher frequency capabilities and its intrinsically low distortion at low operating currents. If not specifically stated as being unique to the current feedback topology, the items considered here apply equally as well to a voltage feedback op amp.

As a starting point for describing op amps for RF applications, it is useful to summarize some of the standard operating assumptions for typical RF amplifiers. Although there are certainly exceptions to the typical conditions shown here, RF amplifiers generally have:

1. AC coupled input and output. A DC voltage generally has little meaning in RF applications.
2. Input and output impedances nominally set to 50Ω (AC) over the frequency range of operation. This is seldom a physical 50Ω resister, but rather a combination of active element I/O impedances along with passive matching networks.
3. Fixed signal gain operations over a certain band of frequencies. Any particular RF amp is purchased to provide a particular gain and is not user adjustable. A two decade range of operating frequencies seems typical.
4. Single power supply operation. Since both input and output are AC coupled, bipolar power supplies, balanced around ground, are not needed. The DC bias point is maintained internally with minimal user adjustment possible.

Operation of National's Current Feedback Op Amps (Continued)

Figure 1 shows a typical RF amplifier connection, while *Figure 2* and *Figure 3* show an ideal op amp, either current or voltage feedback, connected for non-inverting and inverting gains, respectively.

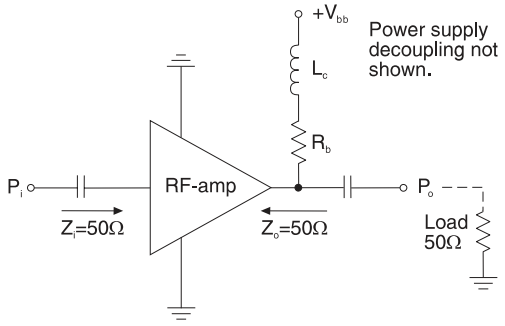

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FIGURE 1. Typical RF Amplifier Connection

For the RF amplifier, both input and output are AC coupled, while a single power supply biases the part through R_b. L_c chokes off the AC output signal from seeing the power supply as a load. The RF amplifier signal gain is specified with the output driving a 50Ω load and is defined as 10*log (power gain)

The two ideal op amp circuits assume that the source is coming from a ground referenced, zero impedance voltage source while their outputs are intended to act as ideal (zero ohm output impedance) voltage sources to a ground referenced load. The non-inverting configuration ideally presents an infinite input impedance, a zero ohm output impedance, and a voltage gain, as shown in *Figure 2*, from the plus input to the output pin.

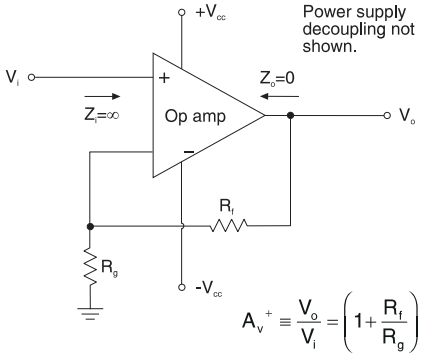

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FIGURE 2. Ideal Non-Inverting Op Amp

The ideal inverting op amp differs in several respects from the non-inverting. The output voltage is ideally 180˚ out of phase from the input, which accounts for the signal inversion. The op amp's (-) input ideally presents a virtual ground, while drawing minimal current, for either voltage or current feedback op amps. This leaves R_g as the ideal input impedance seen by the source, while the voltage gain from the input of R_g to the output is simply $-R_f/R_g$. This signal inversion is usually of no consequence in an RF application, and most of this discussion will deal only with the magnitude of the inverting gain.

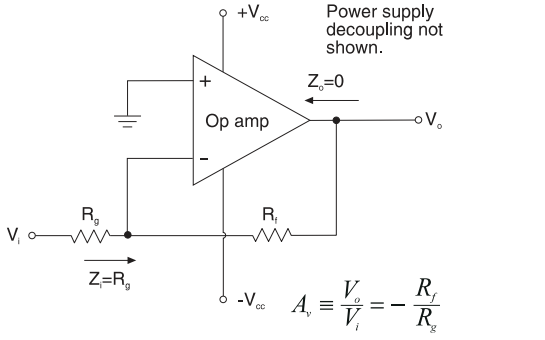

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FIGURE 3. Ideal Inverting Op Amp

When using op amps as RF amplifiers, we must first satisfy the I/O impedance matching requirements, recast the gain from a voltage gain to a power gain (in dB), and possibly configure for operation from a single power supply. *Figure 4* and *Figure 5* show the op amps of *Figure 2* and *Figure 3* set up to provide I/O impedance matching with the resulting power gain equations, but still using bipolar supplies. The bipolar power supplies allow operation to be maintained all the way down to DC. Single supply operation is possible and will be considered next

For the non-inverting case, setting Z_i = 50Ω simply requires a 50Ω termination resistor to ground on the non-inverting input, R_T . Getting Z_o = 50Ω simply requires a series 50Ω resistor in the output, R_O.

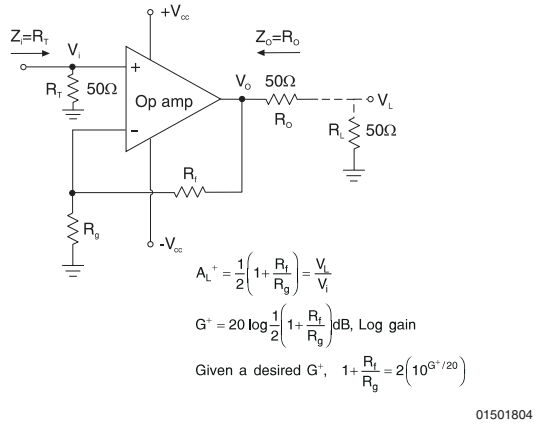

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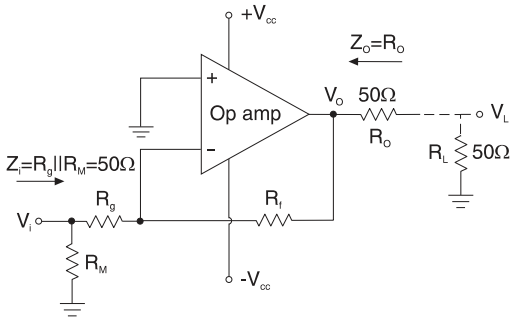
FIGURE 4. Non-Inverting Op Amp Configured for RF Application

For the inverting mode of op amp operation, the (+) input is ground referenced, while the signal channel input impedance becomes the parallel combination of $R_g$and R_M . As OA-13 describes, the current feedback topology depends on the value of the feedback resistor to determine the frequency

Operation of National's Current Feedback Op Amps (Continued)

response. With each particular op amp calling out a particular optimum R_f , R_g can then be used to set the gain and R_M, along with R_g, will set the input impedance. Setting R_g to yield the desired gain and then setting R_M to satisfy $Z_i = 50\Omega$ will work until the required $R_g < 50\Omega$. Having fixed R_f to satisfy the amplifier's stability requirements, going to higher and higher inverting gains will eventually yield R_g's $< 50\Omega$. Non-inverting operation should be used if this limitation is reached. R_f can, however, be increased beyond the recommended value for a current feedback op amp in order to allow an $R_g = 50$ at higher gains, but only at the expense of decreasing bandwidth.



$$A_L^- = -\frac{1}{2}\frac{R_f}{R_g} = \frac{V_L}{V_I}; \text{ neglecting the signal inversion}$$

$$G^- = 20\log\left(\frac{1}{2}\frac{R_f}{R_g}\right) \text{dB, Log gain}$$

$$\text{Given a desired } G^-, \frac{R_f}{R_g} = 2\,(10^{G-/20})$$

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FIGURE 5. Inverting Op Amp Configured for RF Application

Note that for both topologies the gain to the matched load has been cut in half (-6dB), from the earlier ideal case, through the voltage divider action of $R_O = R_L$. It is a simple, but critical, conversion from any description of output voltage swing to and from a power (in dBm) defined at the load. *Figure 6* shows these conversions for a purely sinusoidal signal. Basically, for whatever initial description of voltage swing given, we need to convert that into an RMS voltage, square it and divide by the load ($R_L = 50\Omega$ normally) to get the absolute power in watts. This is then divided by 0.001 to reference that power to 1mW and 10*log of that expression is taken to yield the power in dBm.

$$P_o = 10\log\frac{\left(\frac{V_{L_{pp}}}{2\sqrt{2}}\right)^2}{50\,\Omega\,(1\text{mW})} = 20\log\left(8(50\,\Omega)(0.001)\right)$$
$$= 20\log V_{L_{pp}} + 4\,\text{dBm}$$

$$\text{Conversely, for a given } P_o \text{ (in dBm)}$$

$$V_{L_{pp}} = 10^{(P_0 - 4)/20}$$

Peak – Peak voltage swing at load

$$V_{0_{pp}} = 2^{*}10^{(P_0 - 4)/20}$$

Peak – Peak voltage swing at output pin

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FIGURE 6. Converting Between Voltage Swings and Power

Every op amp has a specified maximum output voltage swing that is generally shown as a peak excursion from ground. This type of specification, for balanced bipolar power supplies, is really inferring how close the output may come to the supply voltages before non-linear limiting occurs. For AC coupled RF applications, it is always best to hold the output pin DC level centered between the two supply pins in order to provide the maximum output V_{pp} . Application note OA-15 discusses in more detail input and output voltage range considerations.

Most of National's op amps do not require a ground reference for proper operation and can be easily operated from a single supply. Generally, all that is required is to keep the DC voltage on the (+) input and the output pin centered between the voltages appearing on the two supply pins. For a single supply operation (with one supply pin held at ground) this translates into the (+) input and V_O being held at V_{cc} /2. For those amplifiers requiring a ground pin, that pin should also be driven with a low source impedance voltage midway between the supply pins.

There are many possible implementations of single power supply op amp operation. *Figure 7* and*Figure 8* show two simple ways to operate non-inverting and inverting op amps as AC coupled RF amplifiers using a single power supply.

In the non-inverting case, the input termination is still DC coupled, while the (+) input bias is set by the two R_b 's to yield V_{cc} /2. R_b should be large enough to limit excessive quiescent current in the bias path, but not so large as to

Operation of National's Current Feedback Op Amps (Continued)

generate excessive DC errors due to the amplifier's input bias current. The gain setting resistor, R_g , is also AC coupled to limit the DC gain to 1. Hence, the (+) input DC bias voltage also appears at the output pin. The output should be AC coupled in both circuits to limit the DC current that would be required if a grounded load were driven.

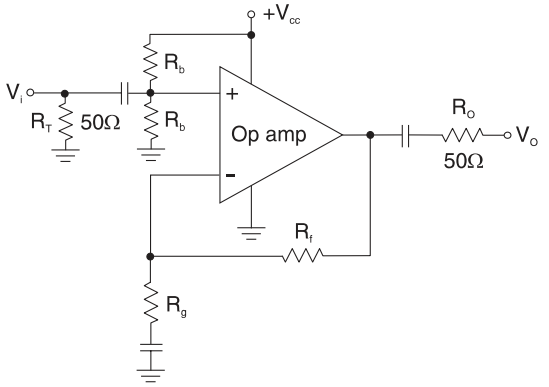

FIGURE 7. Single Supply, Non-Inverting Op Amp Operation

Single Supply, Non-Inverting Op Amp Operation For the single supply inverting amplifier of *Figure 8*, we still require the midpoint reference to be brought in on the (+) input. A de-coupling capacitor on that node is also suggested to decrease the AC source impedance for the non-inverting input noise current. The gain for this non-inverting input reference voltage is again AC coupled to yield a unity DC gain to get $V_{cc}/2$ at the output pin. The inverting input impedance goes from R_M at DC to 50Ω at higher frequencies. R_M, as well as R_T in *Figure 7*, could also be AC coupled to avoid DC loading on the source.

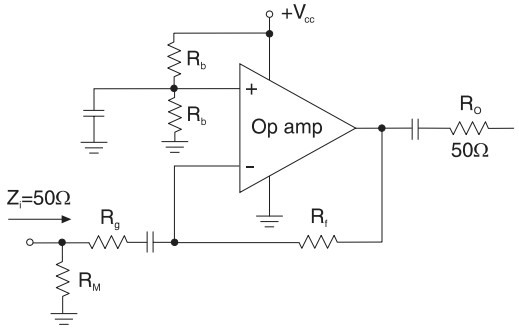

FIGURE 8. Single Supply, Inverting Op Amp Operation

For both of these single supply circuits, we have given up the DC coupling for the signal path. The low frequency limits to operation will now be set by the AC coupling capacitors, along with impedances in each part of the circuit. All of the subsequent discussions assume balanced bipolar supplies, but apply equally as well to single supply operation.

Small Signal AC Performance Characteristics

All of the typical small signal AC parameters specified for RF amplifiers are derived from the S-parameters (reference 1). These are:

Scattering Parameters	RF Amplifier Specification
S_{11} Input reflection	Input VSWR
S_{22} Output reflection	Output VSWR
S_{21} Forward transmission	Amplifier gain and bandwidth
S_{12} Reverse transmission	Reverse isolation

These frequency dependent specifications are measured using a network analyzer and an S-parameter test set. A full 2-port calibration should be performed prior to any device measurements. The HP8753A, used for the measurements reported here, incorporates full 12 term error correction in its 2-port calibration. This basically normalizes all measurement errors due to imperfections in the cabling and test hardware (reference 2).

Figure 9 and *Figure 10* show the two configurations for the CLC404 used in demonstrating the small signal AC performance parameters listed above. In each case, the S-parameter test set places the device into a 50Ω input and output environment. Both configurations achieve a voltage gain of 6 to the output pin and 3 to the 50Ω load. This yields a gain of 20*log(3) = 9.54dB measured by the network analyzer. Recall that one of the advantages to using op amps in RF applications is the exceptional flexibility in setting the gain. A wide range of gains could have been selected for the test circuits of *Figure 9*. and *Figure 10*. ±6 was selected to allow easy comparisons to the CLC404's data sheet specifications, which are all defined at a gain of +6.

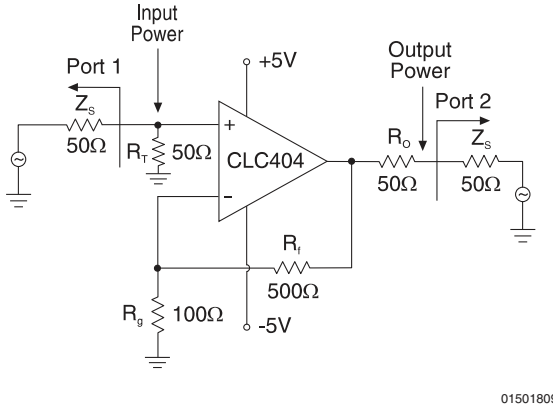

FIGURE 9. Non-Inverting Amplifier S-parameter Test Circuit

For the inverting gain configuration, R_M along with R_g sets the input impedance to 50Ω. An R_T of 50Ω is retained on the

Small Signal AC Performance Characteristics (Continued)

non-inverting input to limit the possibility of self-oscillation in the non-inverting input transistors (See application note OA-15).

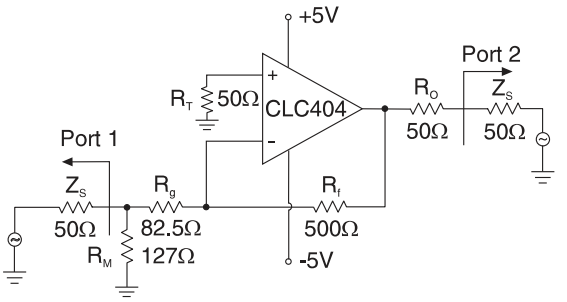

FIGURE 10. Inverting Amplifier S-parameter Test Circuit

Input/Output VSWR

The Voltage Standing Wave Ratio (VSWR) is a measure of how well the input and output impedances are matched to the source impedance. (It is assumed throughout that the transmission line characteristic impedance is also equal to the source impedance of both ports-50Ω in this case). It is desirable that the input and output impedances be as closely matched as possible to the source for maximum power transfer and minimum reflections.

$$\text{VSWR} = \frac{Z_I}{Z_s} \text{ or } \frac{Z_S}{Z_I} \quad \text{whichever} > 1$$

$Z_I \rightarrow$ amplifier input or output impedance

$Z_S \rightarrow$ test system source impedance

$$\text{Return loss} = 20\log\left(\frac{\text{VSWR}+1}{\text{VSWR}-1}\right) = 10\log(S_{11})^2 \text{ input}$$

$$\text{or } 10\log(S_{22})^2 \text{ output}$$

Ideal VSWR = 1

Typically, VSWR = 1.5, for RF-amps over their operating frequency range

Measuring the input VSWR is simply a matter of measuring the ratio of the reflected power vs. incident power on Port 1 of *Figure 9* and *Figure 10* (S_{11}). A perfect match will reflect no power. Output VSWR is measured similarly at Port 2 (S_{22}).

As described earlier, an op amp's input and output impedances are determined by external components selected by the designer. For this reason, I/O VSWR is never shown on an op amp's data sheet. Excellent VSWR can, nevertheless, be achieved using the components shown in *Figure 4* and *Figure 5*.

An op amp's gain polarity has minimal effect on the output VSWR. At low frequencies, R_O by itself will determine the output VSWR. Setting this resistor to 50Ω will yield excellent output VSWR to reasonably high frequencies. As the test frequency increases, however, the op amp's output impedance will begin to increase as the loop gain rolls off (reference 3, page 237). This inductive characteristic can be partially compensated by a small shunt capacitance across R_O. *Figure 11* shows this, for either gain polarity, along with tested output VSWR with and without this shunt capacitance. The value of this capacitance will depend on the amplifier and, to some extent, on the gain setting, and was determined empirically for this test by using a small adjustable cap (5-20pF) directly across R_O.

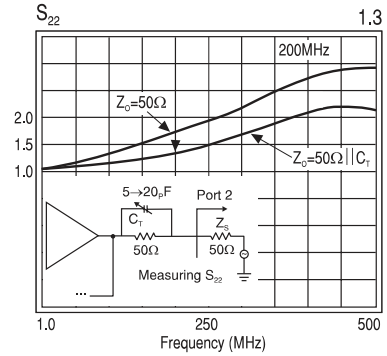

FIGURE 11. Measuring and Tuning CLC404 Output VSWR

The marker at 200MHz indicates an output VSWR of 1.3:1 when CT is tuned optimally. Tuning CT also extends the frequency response (S_{21}) lightly and will be left in place for the remainder of the tests.

The input impedance match of the non-inverting topology *Figure 9* is principally set by R_T . As the frequency increases, the input capacitance of the op amp will eventually degrade the input VSWR. This effect is so negligible over the expected operating frequency range, however, that no tuning is required.

The input impedance match of the inverting topology *Figure 10* is, at low frequencies, set by the parallel combination of R_g and R_M . This holds very well as long as the amplifier's inverting input acts like a low impedance over frequency. For current feedback amplifiers, the inverting input is actually a driven, low impedance buffer. It's impedance will, however, increase with frequency. A voltage feedback amplifier's apparent inverting input impedance will also increase with frequency as its loop gain rolls off. In the voltage feedback case, the increase in inverting input impedance will be seen at a lower frequency than for a current feedback amplifier and will depend strongly on the amplifier gain setting.

Input/Output VSWR (Continued)

Figure 12 shows the tested input VSWR for the two gain polarities of *Figure 9* and *Figure 10*. In this case, we are measuring S_{11} and allowing the HP8753A to convert the measurement and display VSWR directly.

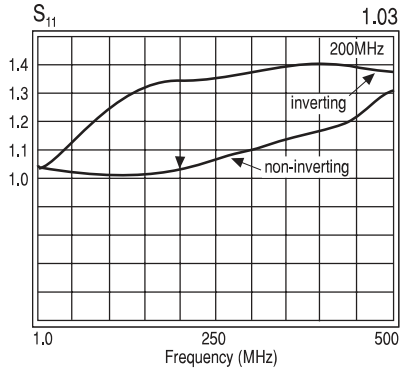

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FIGURE 12. CLC404 Input VSWR

Note carefully the change in scale for the input VSWR vs. the output VSWR plot. The marker on the non-inverting test trace shows an exceptional input VSWR of 1.03:1 at 200MHz, while the inverting, though higher, remains under 1.4:1 through this range.

Forward Gain and Bandwidth

Typical RF amplifier specifications show a fixed gain, as defined in *Figure 1*, with a specified frequency range for 0.5dB gain flatness, along with -3dB cutoff frequencies. For the designer using a current feedback op amp, a wide range of possible gains are easily obtainable. With the CLC404's specified voltage gain range of ±1 to ±10, and including the additional 6dB loss from the output to the load, -6dB to 14dB gains may be achieved using the CLC404. Higher gains can be achieved with this, or any other current feedback amplifier, with some sacrifice in bandwidth (see application note OA-13). For example, the CLC401, specified over a ±7 to ±50 voltage gain range, translates into an 11dB to 28dB gain range for RF applications.

The forward gain over frequency (commonly called the frequency response and measured as S_{21}) always appear in the National data sheets over a range of gains. Small signal -3dB bandwidth and gain flatness are also guaranteed at a particular gain for each amplifier. Rarely does a voltage feedback op amp show the S_{21} characteristics, since it so strongly depends upon the gain setting. Rather, these amplifiers show an open loop gain and phase plot and leave it to the designer to predict closed loop gain and phase. The frequency response plots for the National op amps are normalized to show each gain coming in at the same grid on the plot for easier comparisons of frequency response shape over a wide range of gains. Another advantage of the excellent loop gain control of the current feedback topology is exceptional forward gain phase linearity. This phase is also shown on the frequency response plot. A maximum deviation from linear phase is guaranteed at a particular gain setting in the data sheet specifications.

The part to part variation in the frequency response is minimal for the hybrid amplifiers from National, with more variation seen for the monolithic op amps. As application note OA-13 describes, the current feedback topology allows an easy, resistive trim for the frequency response shape that has no impact on the forward gain. This frequency response flatness trim has the same effect for either non-inverting or inverting topologies. *Figure 13* shows this adjustment added to the circuit of *Figure 9*, along with the measured S_{21} with and without this trim. As OA-13 describes, this resistive trim inside the feedback loop has the effect of adjusting the loop gain, and hence the frequency response, without adjusting the signal gain, which would still be set by only R_f and R_g . This particular test achieved a flatness of ±.1dB from DC to 110MHz at a gain of 9.54dB for the non-inverting test circuit shown (with identical results for an inverting configuration).

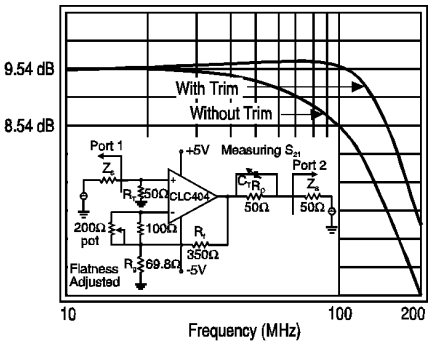

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FIGURE 13. Measuring and Adjusting the Frequency Response S_{21}

Note that the values for R_f and R_g have been reduced from those used in the circuit of *Figure 9*, although their ratio and hence the gain, have remained the same. With the adjustment pot set to zero ohms, this lower R_f value ensures that the frequency response will be peaked for any particular CLC404 used in the circuit. Then, by increasing the resistance into the inverting input, the amplifier can be compensated and S_{21} adjusted to the excellent flatness shown above.

The part to part variation in frequency response becomes more pronounced as the desired operating frequencies and signal gains increase. Operation of the CLC404 through 50MHz at 9.54dB gain would, for example, have minimal variation relative to operation through 100MHz and 14dB gain. For ±.1dB flatness, and considering the rapid degradation in distortion performance at higher frequencies, 100MHz is probably a reasonable upper limit for the operation of National op amps (available at the time of publication) in RF (or IF) applications. Higher frequency operation can be achieved if the degraded flatness and distortion characteristics are acceptable to the application. New product introductions can be expected to extend this operating frequency.

Reverse Isolation

This small signal AC characteristic is a measure of how much signal injected into the output port makes it back into the input source. The magnitude of S_{12} is the measure of

Reverse Isolation (Continued)

reverse isolation. National's current feedback op amps exhibit excellent reverse isolation relative to most RF amplifiers. This results from both the output and inverting input being driven, low impedance, nodes. To the extent that the output of the op amp and its inverting input both present very low impedances over wide frequency ranges, significant signal attenuation can be expected in taking a signal voltage applied to the output matching resistor and tracing it back to either an inverting or non-inverting input signal. Slightly more attenuation can be expected for the non-inverting vs. inverting configurations, since the signal must also get from the inverting to non-inverting pin in the non-inverting case.

The circuit of *Figure 13*, along with the inverting circuit of *Figure 14* were used to measure the reverse isolation for both gain polarities, as shown in *Figure 15*. Although reverse isolation is generally specified as a positive number, this is simply the negative of the log gain in going backwards through the amplifier. Hence, the plot of *Figure 14* shows a rising "gain" that would be interpreted as a decreasing reverse isolation as we go to higher frequencies. As *Figure 15* shows, isolations in excess of 30dB are easily obtainable through frequencies far higher than the operating frequency range, with very high isolations observed at low frequencies.

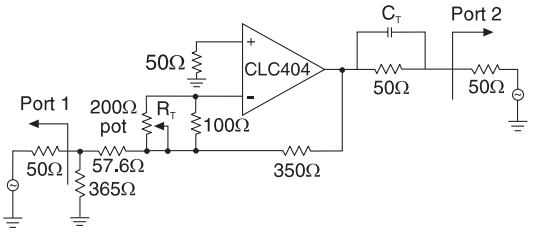


FIGURE 14. Inverting Reverse Isolation Test Circuit

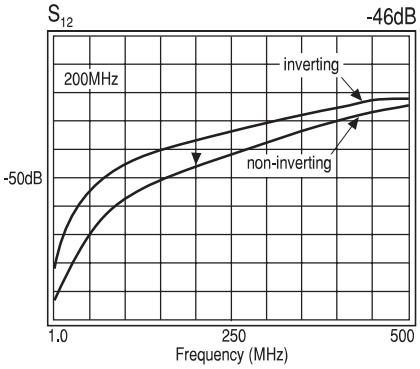


FIGURE 15. Reverse Gain for the Circuits of Figures 13 and 14

Dynamic Range Limiting Characteristics

The final area of concern in applying op amps to RF applications are the limits to dynamic range familiar to RF amplifier users. These are generally limited to:

-1dB Compression Point

2-Tone, 3rd Order, Intermodulation Intercept

Noise Figure

The -1dB Compression Point is a measure of the maximum output power capability of the amplifier. The 2-Tone Intercept allows the prediction of spurious signals caused by amplifier non-linearities when two input signals closely spaced in frequency are applied to the input. The noise figure is a measure of how much noise is added by the amplifier and will set a limit to the minimum detectable signal.

Dynamic Range Limiting Characteristics (Continued)

Although each of these can certainly be measured for any particular op amp configuration, their interpretation for op amps may vary from RF amplifiers depending on the op amp being used and the specification. Each of these will be described generally and developed, and/or measured, for the CLC404 with any anomalies in interpretation noted.

-1dB Compression Point

Briefly stated, this is the expected output power, at a fixed input frequency, where the amplifier's actual output power is 1dBm less than expected. As *Figure 16* shows, it can also be interpreted as the ideal output power at which the actual amplifier gain has been reduced by 1dB from its value at lower output powers. With both the X and Y axis of *Figure 16* a dBm scale, the output power vs. the input power will have a slope of 1. If we shift the X-axis by the amplifier's low power gain (a 20dB gain was used arbitrarily in *Figure 16*), the amplifier's input to output transfer would ideally be a unity slope line through the origin.

An additional interpretation of *Figure 16* is that beyond the -1dB compression point the output power remains fixed as the input power is increased. If S_{21} were measured at a fixed frequency, with a swept input power, we would get a horizontal line, showing the low power gain, that eventually transitions to a -1 slope line as the output power becomes fixed while the input power continues to increase.

The -1dB compression power is commonly used as a maximum output power limit when computing an amplifier's dynamic range. Standard AC coupled RF amplifiers show a relatively constant -1dB compression power over their operating frequency range.

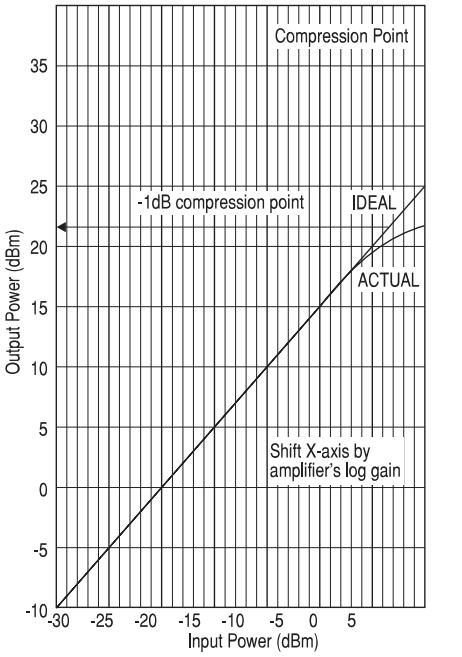


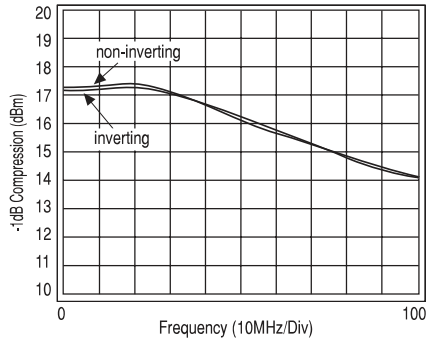
FIGURE 16. Illustration of -1dB Compression

For an operational amplifier, the maximum output power depends strongly on the input frequency. The two op amp specifications that serve a similar purpose to -1dB compression are output voltage range and slew rate. At low frequen-

-1dB Compression Point (Continued)

cies, increasing the power of a fixed frequency input will eventually drive the output “into the rails” - a saturation limit typically some number of diode drops below the supply voltages. In addition, as the input frequency increases, all op amps will reach a limit on how fast the output can transition. This is typically specified as a slew rate indicating the maximum dV/dT at the output pin voltage. Half this slew rate is available at the matched load when an output series matching resistor is used. For a sinusoidal signal, the maximum slew rate occurs at the 0 crossing. This maximum dV/dT is simply the peak voltage exursion times the radian frequency. Given a slew rate in Volts/sec (SR) and a frequency, the maximum peak amplitude before slew limited operation is experienced is predicted to be SR/(2*π*frequency). However, this peak amplitude, which can be converted to a dBm power at the load using the expressions developed earlier, does not relate directly to the measured -1dB compression.

Figure 17 shows the measured -1dB compression powers vs. frequency for the CLC404 in the circuits of *Figure 13* and *Figure 14*. Since the maximum output power is principally a function of the output stage, there is very little difference between the non-inverting and inverting -1dB compression points. For the National amplifiers that show a higher inverting slew rate than non-inverting (e.g. CLC400), a higher -1dB compression power at higher frequencies in inverting configurations would be expected. The low frequency value, however, should be similar between polarities, since it is determined by the maximum output voltage swing (principally set by the power supply voltages and the headroom requirements in the output stage).

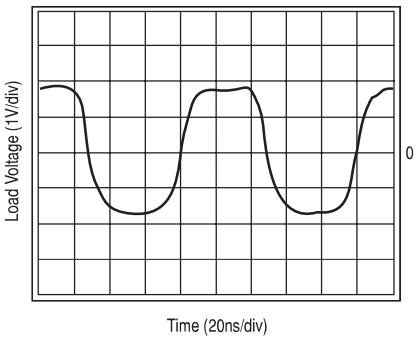


01501818

FIGURE 17. 1dB Compression for the CLC404

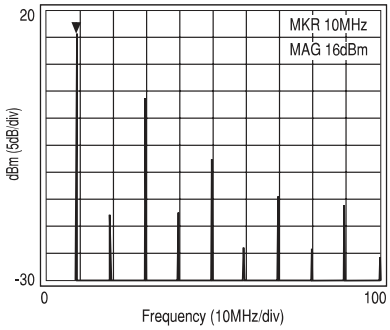
Although *Figure 17* shows the -1dB compression as defined in *Figure 16*, it is also very useful to look at the output waveforms and spectrums to gain an understanding of what is setting the measured -1dB compression power.

Figure 18 and *Figure 19* show the time waveform and the spectrum at the load for the input power that yields the -1dB compression point for the CLC404 operating at 10MHz.



01501819

FIGURE 18. Output Waveform at 10MHz - 1dB Compression



01501820

FIGURE 19. Output Spectrum at 10MHz - 1dB Compression

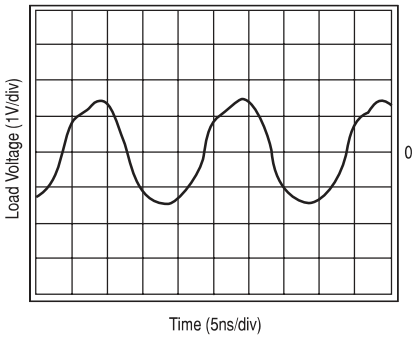
At this low frequency, we are clearly running into an output voltage swing limitation. With a 17.3dBm -1dB compression (as shown at 10MHz in *Figure 17*), we would expect the fundamental amplitude in the spectrum to be at 16.3dBm. The observed 16dBm in the spectrum of *Figure 19* is a reasonable match to this expected fundamental power. It is, however, incorrect to directly convert this fundamental power at -1dB compression into a sinusoid and expect that the amplifier can deliver a sinusoid of this amplitude. For the 16.3dBm fundamental power predicted by the -1dB compression measurement, we might expect that the output is delivering an approximately $4V_{pp}$ sinusoidal swing at the load, or ±4V swing at the output pin. Although this would

-1dB Compression Point (Continued)

exceed the maximum output swing specification for the CLC404 operating at ±5 volt supplies, this amplitude of sinusoid is in fact available if a zero loss filter is used to pass on only the fundamental harmonic.

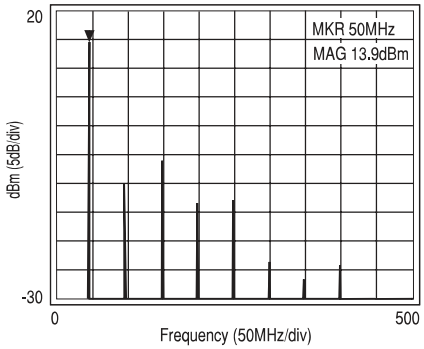
Notice that a considerable portion of the output power has been spread into the odd order harmonics. This is typical of the square wave output observed in the time domain trace of *Figure 18*. The fundamental (10MHz) power can be related to the output time waveform amplitude through the Fourier series expansion of the output waveform. If the output were a perfect square wave, under conditions of output voltage limited operation, a peak square wave amplitude of A would generate a fundamental frequency amplitude of $4*A/\pi$. Going from the measured peak amplitude of the output time waveform, the anticipated -1dB compression would be calculated as the power in a sinusoid $4/\pi$ times the square wave amplitude +1dBm. Doing this for the measured ±1.8V swing of *Figure 18* would predict 15.1dBm (peak-peak square wave amplitude converted to dBm) + 2.1dBm ($20*\log(4/\pi)$) + 1dBm (reported -1dBm output power is 1dBm higher than measured power) = 18.3dBm. This is 1dBm higher than measured. This can be explained by the less than perfect square wave shape shown in the time waveform of *Figure 18*. This less than perfect square wave will yield a coefficient for the fundamental term in the Fourier expansion that is actually less than the predicted $A*4/\pi$.

As the operating frequency increases, the slew limit for the op amp will eventually restrict the achievable output swing to something less than the output voltage swing limit of the amplifier. This can be observed in *Figure 11* at approximately 30MHz for the CLC404. Again, it is instructive to look at the time waveform and resulting spectrum when operating at an input power that yielded a -1dB compression in measured gain at the these higher frequencies. *Figure 20* and *Figure 21* show this for the non-inverting circuit (*Figure 13*) operating at the input power necessary to produce the measured -1dB compression with a 50MHz sinusoidal input signal (From *Figure 17*, this input power would be 16.3dBm - 9.54 (gain) = 6.8dBm)



01501821

FIGURE 20. Measured Output Waveform at 50MHz -1dB Compression



01501822

FIGURE 21. Measured Output Spectrum at 50MHz -1dB Compression

The measured -1dB compression power under slew limited conditions is dependent on the amount of power in the fundamental frequency generated by the time waveform shown in *Figure 20*. Although we can say that the -1dB compression must be related to the amplifier's slew rate, it would be very difficult to relate the slew rate to the waveform shape and then, through the Fourier series, to the fundamental power and hence -1dB compression. The exact distribution of power into the fundamental and harmonics is changing over frequency. All that can really be said is that at these higher frequency -1dB compressions, a significantly distorted waveform with a peak to peak excursion less than that seen at lower frequencies is being generated.

At low frequencies, the -1dB compression power can be predicted approximately using the analysis shown earlier by assuming a square wave output set by the output voltage swing limits shown in the op amp data sheet. Remember that the output voltage range specified in the data sheet is twice what can be delivered through the 6dB loss taken from the matching resistor to the load. It is not, however, possible to easily predict the higher frequency -1dB compression from the slew rate specification. As will become apparent in the next section, it is also not possible to relate the -1dB compression to the third order intercept. Typical RF amplifiers will show a 3rd order intercept 10dBm higher than the -1dB compression point. National op amps, if they show an intercept characteristic, have an intercept considerably higher than what would be predicted by adding 10dBm to the -1dB compression.

2-Tone, 3rd Order Intermodulation Intercept

This specification is directed at predicting the 3rd order intermodulation distortion powers for any combination of two closely spaced (in frequency) input signals. Any amplifier can be modeled to have a polynomial approximation to its transfer function from input to output. When two input signal frequencies are present, the 3rd order term of this polynomial approximation will give rise to distortion terms at frequencies that can be very near the input signal frequencies.

2-Tone, 3rd Order Intermodulation Intercept (Continued)

These closely spaced distortions are considerably more troublesome to narrowband IF channels than the simple harmonic distortion terms that appear in integer increments away from the input signal frequency.

Appendix B expands all of the spurious frequency locations and distortion coefficients for two input signals at frequencies of f_O-Δf and f_O+Δf when passed through a 5th order polynomial. With this simple definition of equal deviations from a center frequency (an average frequency), all of the spurious frequency locations become very simple algebraic expressions of f_O and Δf. Using this approach to defining the test frequency locations also allows a clear illustration of the symmetric clusters of spurious terms around integer multiples of f_O. From appendix B, the 3rd order inter-modulation terms fall at $f_O \pm \Delta Df_O$. With an input signal defined as V_i = Acos(2Δ(f_O-Δf)t) + Bcos(2Δ(f_O +Δf)t), and an input to output voltage gain transfer function of $V_O = K_0 + K_1 V_i + K_2 V_i^2 + K_3 V_i^3$ (ignoring the higher order terms for now), a lower 3rd order spurious term at f_O - 3Δf with an amplitude of (3/4)*K_3 *A*B_2 and an upper spurious at f_O + 3Δf with an amplitude of (3/4)*K_3 *A_2 *B will result.

If equal amplitude signals were applied to the input (A = B), and if these were increased in an equal fashion, the two spurious amplitudes would increase in a cubic fashion. In dBm terms, if the two input, and hence output, powers were increased by 1dBm, this model predicts that the two output third order spurious powers will increase by 3dBm. It is interesting to note the effect of adjusting just one of the input frequency power's. Changing the lower test frequency power by 1dBm will change the lower spurious by 1dBm and the upper spurious by 2dBm. Conversely, changing the upper test frequency power by 1dBm will change the lower spurious by 2dBm and the upper by 1dBm. The dependence of the 3rd order spurious power to output test frequency power (assuming equal powers for each test frequency) is shown graphically in *Figure 22*.

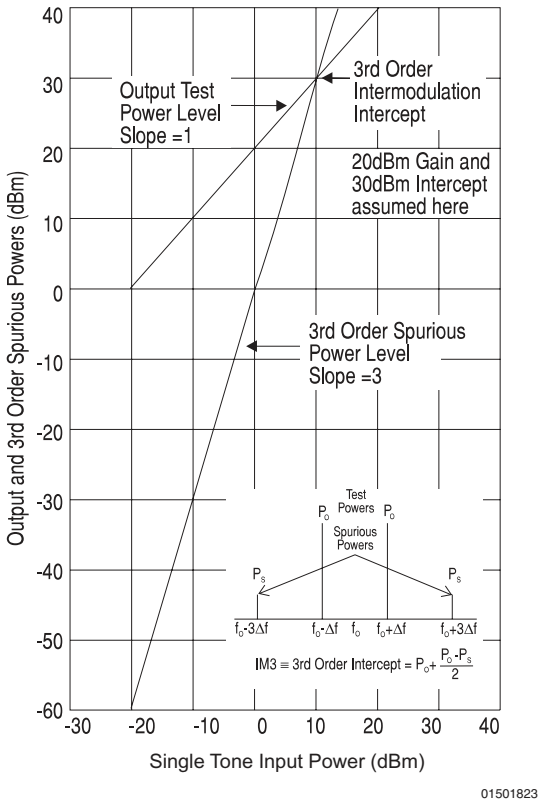


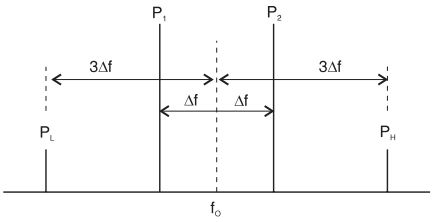
FIGURE 22. Output and 3rd Order Spurious Power vs. Input Power

As shown in *Figure 22*, the 3rd order spurious powers, increasing at a 3X rate vs. the input power, will, at some output power, "intercept" the desired output powers that are increasing at a 1X rate vs. the input power. Another way of

2-Tone, 3rd Order Intermodulation Intercept (Continued)

saying this is that there is a 2X closure rate between the desired output powers and the undesired 3rd order intermodulation spurious powers. The graph of *Figure 22* was arbitrarily set up for an amplifier gain of 20dB (the x-axis has been shifted to yield 0dBm output power for -20dBm input power) and for a 30dBm 3rd order intercept. No actual amplifier will be able to reach the intercept point from an output power standpoint since this intercept typically exceeds the -1dB compression power by at least 10dBm. The intercept is intended as a mathematical construct to allow the prediction of the spurious power level for a given output signal power. For an amplifier that shows a 3rd order intermodulation intercept characteristic, a single measurement of output powers and spurious levels are sufficient to solve for the intercept point as shown by the equation in *Figure 22*.

Figure 22 assumes equal output power levels for the two desired output signals. A more general approach, with unequal test power levels, shows that, from one set of measurements, two estimates for the 3rd order intercept can be made. *Figure 23* steps through this analysis and concludes with the predictive equations for each of the two 3rd order spurious levels. The graphical representation shown in *Figure 23* is modeling what would be observed for a spectrum analyzer measurement of the test and spurious powers.



$$1^{st}\ \text{estimate of intercept } IM3_1 = P_1 + \frac{P_2 - P_L}{2}$$

$$2^{nd}\ \text{estimate of intercept } IM3_2 = P_2 + \frac{P1 - P_H}{2}$$

$$\text{Averaging these } IM3_{AVG} = \frac{P_1 + P_2}{2} + \frac{P_2 - P_L}{4} + \frac{P_1 - P_H}{4}$$

$$\text{If } P_1 = P_2 = P_T \text{ and } P_L = P_H, \text{ define } \Delta dB_c = P_O - P_L = P_O - P_H$$

$$IM3 = P_O + \frac{\Delta dB_c}{2}$$

Solving the two intercept estimates for P_L and P_H

$$P_L = P_2 - 2\,(IM3 - P_1)$$

$$P_H = P_1 - 2\,(IM3 - P_2)$$

$$\text{or, for } P_1 = P_2 = P_O$$

$$P_L = P_H = 2\left(\frac{3}{2}P_O - IM3\right)$$

01501824

FIGURE 23. 3rd Order Intermodulation intercept Calculations

Typical RF amplifiers closely approximate this 3rd order intermodulation spurious model with an intercept that is relatively constant over the specified operating frequency range of the amplifier. Op amps, however, show significant deviations from this simple model. The principal difference is that all op amps will show a strongly frequency dependent 2-tone, 3rd order spurious performance. The observed intermodulation spurious levels will be a function of the intrinsic distortion in the forward path of the amplifier corrected by whatever loop gain the amplifier has at that frequency. All op amps show a loop gain that decreases with frequency. Hence, the 3rd order spurious levels will, in general, increase with frequency for fixed output test powers (reference 4 discusses this loop gain dependence in detail).

An additional concern is at what point in the circuit to define the 3rd order intercept. In order to make direct comparisons to RF amplifiers, National defines the 3rd order intercept to be at a 50Ω load when driving from a 50Ω output impedance. Some of the earlier National data sheets (e.g. CLC220, CLC221) defined the intercept for a voltage swing at the output pin converted into a power (as if it were driving 50Ω) while actually applying that output swing to the series 50Ω into a 50Ω load. This has the effect of defining an intercept that is 6dBm higher than what is actually available at the load. This can be seen from the equation shown above for IM3. Recall that IM3, for equal test power levels for the two test frequencies, is simply the test power level +1/2 the difference between the test power levels and the spurious power levels. This difference does not change in going from the output pin to the matched load. However, the output voltage swing will drop by 6dB and, since the output pin power was erroneously defined as being a particular voltage swing across a 50Ω load (when it in fact sees a 100Ω load), this will translate into a 6dBm drop in the test power level to the matched 50Ω load. Therefore, the usable intercept at the matched load is 6dBm lower than specified in those earlier data sheets that call for an output power calculation at the output pin.

Given that the test power level is being defined at the matched load, it is important to consider the amplifier limitations on the maximum power and frequency of test. For a two tone test of equal powers and closely spaced frequencies, the available peak to peak voltage swing for each test frequency at the load is 1/4 the peak to peak output voltage available at the amplifier's output pin while the available slew rate for each test tone can be estimated as 1/8 the amplifier's specified slew rate. For a 2-tone test signal being generated at a matched load, twice the peak to peak swing is being generated in the envelope (and twice the slew rate). Going back through the matching resistor to the output pin will double this swing and slew rate again. In addition, empirical testing has revealed that an overall maximum slew rate at the output pin that is 1/2 the specified op amp slew rate will show low spurious performance. As the slew rate of the output pin waveform exceeds this limit, additional nonlinearities come into play rapidly increasing the 3rd order spurious powers.

Using the circuit of *Figure 13* and the typical specifications for the CLC404, the maximum test power level at the load for each test tone, from an output swing standpoint, would be (1/4)*6V_{pp} = 1.5V_{pp} . This translates into a maximum test power level for each tone of approximately 8dBm. At this maximum output swing, the available slew rate of (1/8)*2000Vμmsec = 250V/μsec will limit the frequency of operation to less than (SR/(2*π*V_{pp}/2) = 250E6/(2*π*.75V) = 53MHz. As the test or operation powers decrease, this upper frequency limit set by the slew rate limit will increase. For example, dropping the power 6dB to 2dBm will push this limit out to 106MHz.

Although some of the National current feedback amplifiers (e.g. CLC400, CLC401, CLC560) show a good approximation to the 3rd order intercept model, the CLC404, used in

2-Tone, 3rd Order Intermodulation Intercept (Continued)

the example circuits shown thus far, shows a spurious power vs. test power characteristic that deviates significantly from the simple model of *Figure 22*. *Figure 24* shows the difference between the test and spurious powers plotted as a function of single tone test power at the load. Note that the independent variable axis is output power; not the input power shown in *Figure 22*. Ideally, this would, at each frequency, yield a straight line with a slope of +2 (instead of the +3 slope shown in *Figure 22*). A similar plot for the CLC401, which more closely approximates this ideal, is shown in *Figure 25*. If an op amp closely approximated the 3rd order intercept model, a single measurement at one operating power would be adequate to predict the intercept at that frequency.

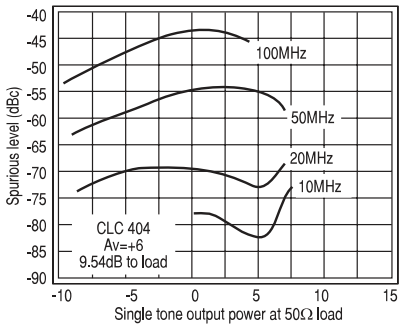


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FIGURE 24. Measured 3rd Order Spurious for the CLC404

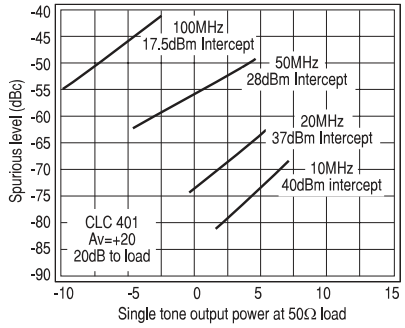

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FIGURE 25. Measured 3rd Order Spurious for the CLC401

The 3rd order spurious plot for the CLC404 is clearly showing some additional mechanism that is holding the spurious levels down as the output power level moves above 0dBm. At lower power levels, it appears that the spurious characteristic is moving towards a linear slope of 2 as predicted by the simple intercept model. Looking again at the 5th order expansion of the 2-tone coefficients shown in appendix B, an additional 5th order term contributes to the spurious powers observed at the 3rd order intermodulation frequencies. Normally, it would be expected that the K5 coefficient is so much lower than the K3 value that this 5th order contribution can be neglected. However, in the case of the CLC404, the K3 coefficient is so low as to make this second term significant at higher operating powers. Note that the contribution of this 5th order term increases as the 5th power of the two test powers vs. the more slowly increasing 3rd order term. It can be theorized that the 5th order coefficient is of opposite sign to the 3rd order coefficient. Then, as the test powers increase to the level that this 5th order term becomes significant in magnitude vs. the 3rd order, the spurious levels actually decrease for increasing output power.

The projected intercept at very low power levels can still be used to predict the spurious free dynamic range. In *Figure 24*, the intercept at low output powers may be estimated for a particular frequency as the output power minus 1/2 the y-axis value. However, it should be realized that wideband op amps like the CLC404 actually provide better spurious performance at high powers than would be predicted by this low power intercept model.

The 3rd order intercept performance is typically very similar between inverting and non-inverting topologies. As discussed in reference 4, anything that changes the loop gain of the op amp will have an effect on the 3rd order spurious performance. Increasing loop gain, either by going to low feedback resistor values for current feedback op amps or low signal gains for voltage feedback op amps, will decrease the spurious powers. In both cases, however, increasing the loop gain by changing the external operating point is constrained by closed loop stability considerations. 3rd order distortions and intermodulations can be further reduced by operating any op amp at higher quiescent currents (if possible) and/or driving the output into a higher impedance load for those situations not requiring a 50Ω matched impedance environment.

Noise Figure

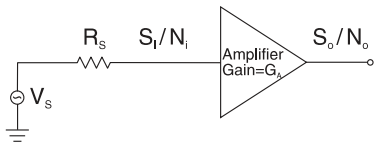
Unlike the compression point and 3rd order intermodulation intercept, the noise figure for an op amp is always usable in the same way that it is for an RF amp. It is important to remember that, like compression and intercept, a noise figure is generally developed at a particular frequency and may change over frequency. Normally, however, a single value can be used above the op amp's 1/f noise corner frequency (see application note OA-12 for an additional noise discussion and its appendices for 1/f noise corner discussion and tabulated op amp input noise terms for National op amps).

The noise figure can be accurately calculated from the equivalent input noise terms for an op amp and the resistor values used to achieve the desired gain and input impedance. Unlike an RF amplifier with a fixed gain and noise figure, an op amp's noise figure will be strongly dependent on the gain setting. We can, however, easily predict the noise figure with the equations developed here.

A very general development for an op amp's non-inverting noise figure will be performed in order to allow easy comparison to noise figure expressions found in earlier National data sheets. The inverting op amp's noise figure will, however, proceed with the assumption normally used – that the input impedance is to be matched to the source impedance.

Noise Figure (Continued)

An idealized schematic illustrating the definition of noise figure is shown in *Figure 26*.



$$\text{Noise Figure} = NF = 10\log\left[\frac{S_i / N_i}{S_o / N_o}\right] = 10\log\left[\frac{S_i\, N_o}{S_o\, N_i}\right]$$

01501827

FIGURE 26. Noise Figure Definition

All of the input and output noise and signal terms in the equation for noise figure (NF) are considered to be powers. Ni is the noise power delivered by the source resistor to the input of the amplifier. All other noise sources are considered to be part of the amplifier and contribute to the noise power, N_o , seen at the output.

Looking at the two parts of the NF expression (inside the log function) yields:

$S_i / S_o \rightarrow$ Inverse of the power gain provided by the amplifier

$N_o / N_i \rightarrow$ Total output noise power, including the contribution of R_S , divided by the noise power at the input due to R_S

To simplify this, consider Na as the noise power added by the amplifier (reflected to its input port):

$S_i / S_o \rightarrow$ 1/G

$N_o / N_i \rightarrow$ $G^*(N_i + N_a)/N_i$ (where $G^*(N_i + N_a) = N_o$)

Substituting these two expressions into the NF expression:

$$NF = 10\log\left[\frac{1}{G}\left(\frac{G[N_i + N_a]}{N_i}\right)\right] = 10\log\left(1 + \frac{N_a}{N_i}\right) \quad (1)$$

The noise figure expression has simplified to depend only on the ratio of the noise power added by the amplifier at its input (considering the source resistor to be in place but noiseless in getting Na) to the noise power delivered by the source resistor (considering all amplifier elements to be in place but noiseless in getting N_i). Generally, the definition for NF also constrains the input impedance for the amplifier to be conjugate matched to the source resistor (this yields $N_i = k_T$ with this constraint). We will, however, relax this constraint initially to allow comparison to the NF expressions found in National's earlier datasheets.

The NF of *Equation (1)* is specified in terms of a power ratio. The individual noise terms for the op amp are, however, expressed as spot noise voltages or currents (Spot means in a 1Hz bandwidth, as opposed to integrated over some noise power bandwidth. See OA-12). Combining separately contributing noise sources is a matter of adding noise powers. This can be done by converting all current noises to a voltage through the appropriate impedance, then summing all of the squared noise voltage terms. Any impedance (normally needed to define a power) or noise power bandwidth (used to convert from spot to integrated noise) will normalize out since we are developing the ratio of two powers at the same point in the circuit. Getting to the total spot noise power is then simply a matter of summing all the relevant squared noise voltages.

Figure 27 below shows an op amp in the non-inverting configuration with all of the individual resistor and amplifier input noise terms detailed.

Where;

e_n = op amp input voltage noise

in_i = op amp non-inverting input current noise

i_i = op amp inverting input current noise

$R_T \rightarrow$ input terminating resistor

e_T = R_T's voltage noise

e_f = R_f's voltage noise

i_g = R_g's current noise

e_S = R_S's voltage noise

01501856

FIGURE 27. Non-Inverting Op Amp Noise Figure Analysis Circuit

Recall that the noise of a resistor (Johnson Noise) can be defined as either a spot current or voltage noise. For a resister of value R, these two possible expressions are:

$$\text{voltage noise,}\quad e_R = \sqrt{4kTR}$$

$$\text{current noise,}\quad i_R = \sqrt{\frac{4kT}{R}}$$

where;

k $\rightarrow$ Boltzman's constant

k = 1.38E-23 Joules/˚Kelvin

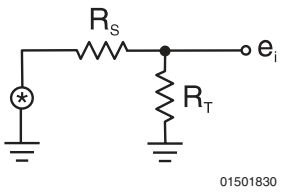
T $\rightarrow$ ˚Kelvin = 290˚ in this analysis

4kT = 16E-21 Joules at T = 290˚K

The 3 amplifier noise terms are available for most of National's amplifiers in appendix 2 of OA-12. If the spot noise Figure below the 1/f noise corner is of interest, appendix I of OA-12 also shows how to approximate the low frequency spot noise from the high frequency flat band value and the 1/f noise corner frequency.

Using the circuit of *Figure 27*, the NF expression can be developed by generating an expression for N_i and N_a . N_i is the noise power delivered by the source resistor noise to the input of the amplifier. This analysis simply proceeds by considering the noise voltages as sources in normal linear circuit analysis, but eventually squaring the resulting noise voltage delivered to RT from es. *Figure 28* shows the equivalent circuit and the resulting N_i . This is considering the amplifier to have an infinite non-inverting input impedance, with all other noise sources neglected for now (superposition of noise voltage contributions are used throughout this analysis).

Noise Figure (Continued)



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$$e_i = \frac{R_T}{R_S + R_T}\sqrt{4kTR_S}$$

$$N_i = e_i^2 = \frac{R_T{}^2\, 4kTR_S}{(R_s \div R_T)^2}$$

$$\text{define} \quad R_p = R_s \,||\, R_T = \frac{R_S R_T}{R_S + R_T}$$

$$N_i = \frac{4kTR_pR_T}{R_S + R_T} = \frac{4kTR_p}{1 + R_T / R_S}$$

01501831

FIGURE 28. Input Noise Power calculation

To get an expression for N_a , all other noise voltages and currents are referred to the non-inverting input and summed as voltages squared. For the noise terms on the inverting side of the amplifier, it is best to find each term's gain to the output voltage, then reflect back to the non-inverting input by dividing by the non-inverting voltage gain of the amplifier. At this point, since we are dealing with linear voltage gains, define this gain as $A_v + = 1 + R_f / R_g$. Table 1 tabulates each individual voltage and current noise and its "gain" to the input of *Figure 27*. Note that all current noise terms have an impedance in their gain expression to yield all voltage noise terms at the input of the amplifier.

TABLE 1. Noise Terms Contributing to N_a for the Non-inverting Op Amp Configuration

Noise Source	Value	Voltage Gain to Input
Non-inverting input voltage noise	e_n	1
Non-inverting input current noise	in_i	$R_n \,\|\|\, R_T \rightarrow (R_P)$
Inverting input current noise	i_i	$R_f/A_v{}^+$
Input terminating resistor voltage noise	$\sqrt{4kT/R_g}$	R_S/R_S+R_T
Gain setting resistor current noise	$\sqrt{4kT/R_g}$	$R_f/A_v{}^+$
Feedback resistor voltage noise	$\sqrt{4kT/R_g}$	$1/A_v{}^+$

One point of possible confusion is that, although we are trying to develop the total noise power at the input of the op amp, what relation does this have to the input voltage noise term that already appears in the op amp model, en. As described in OA-12, the noise model for an op amp attempts to lump all the internal noise sources of the actual amplifier into an equivalent input noise voltage at the non-inverting input and two input noise currents. The intent is to provide a means of predicting the noise performance over a wide range of external operating conditions. The en shown in the analysis model of *Figure 27* is associated only with the internal characteristics of the op amp itself. The total amplifier output noise includes this and contributions from all of the other noise sources shown there. Having gotten to an expression for the total output noise voltage, an equivalent input noise voltage may be derived by simply dividing by the voltage gain of the op amp. This step of input referring each noise source is performed for each term in Table 1.

To form an expression for N_a, we need only to sum the squared product of each noise source and its associated gain, as shown in Table 1.

$$N_a = e_n{}^2 + \left(i_{ni}R_p\right)^2 = \left(i_i \frac{R_f}{A_v{}^+}\right)^2 + \frac{4kTR_T R_S{}^2}{(R_S + R_T)^2} + \frac{4kTR_f{}^2}{R_g(A_v{}^+)^2} + \frac{4kTR_f}{(A_v{}^+)^2}$$

This will simplify to:

$$N_a = e_n{}^2 + (i_{ni}R_p)^2 + \left(i_i \frac{R_f}{A_v{}^+}\right)^2 + \frac{4kTR_p}{1 + R_T / R_S} + \frac{4kTR_f}{{}^+A_v}$$

op amp noise terms + input terminating resistor noise term + combined feedback and gain setting resistor noise terms

(2)

Noise Figure (Continued)

An expression for the non-inverting noise Figure (N+) may now be derived by substituting the equation in *Figure 28* and *Equation (2)* back into *Equation (1)*.

$$NF^+ = 10\log\left[1 + \frac{e_n^2 + (i_{ni}R_p)^2 + \left(i_i \frac{R_i}{A_v^+}\right)^2 + \frac{4kTR_p}{1+R_T/R_S} + \frac{4kTR_f}{A_v^+}}{\frac{4kTR_p}{1+R_S/R_T}}\right]$$

This will further simplify to:

$$NF = 10\log\left[1 + \frac{1+R_s/R_T}{1+R_T/R_S} + \frac{(1+R_s/R_T)R_p^2}{4kTR_p}\left(\left(\frac{e_{ni}}{R_p}\right)^2 + i_{ni}^2 + \left(\frac{i_iR_f}{R_pA_v^+}\right)^2 + \frac{4kTR_f}{R_p^2A_v^+}\right)\right] \quad (3)$$

Simplifying two of these terms:

$$\frac{1+\frac{R_S}{R_T}}{1+\frac{R_T}{R_S}} = \frac{\frac{R_T+R_S}{R_T}}{\frac{R_S+R_T}{R_S}} = \frac{R_S}{R_T}$$

$$\text{Recall that } R_p = \frac{R_SR_T}{R_S+R_T}$$

$$\text{then } \frac{\left(1+\frac{R_S}{R_T}\right)R_p^2}{4kTR_p} = \frac{\frac{R_T+R_S}{R_T}\frac{R_SR_T}{R_S+R_T}}{4kT} = \frac{R_S}{4kT}$$

Putting these simplifications back into Equation 4 yields:

$$NF^+ = 10\log\left[1 + \frac{R_S}{R_T} + \frac{R_S}{4kT}\left(\left(\frac{e_{ni}}{R_p}\right)^2 + i_{ni}^2 + \left(\frac{i_iR_f}{R_pA_v^+}\right)^2 + \frac{4kTR_f}{A_v^+R_p^2}\right)\right] \quad (4)$$

This expression for the non-inverting noise Figure closely matches the equation shown in the CLC205 and CLC206 data sheets. *Equation (4)* differs only in some of the variable names and in the addition of a term due to the R_f and R_g noise, which the CLC205 and CLC206 equations neglected.

If we were to let $R_T \rightarrow \infty$, driving the signal directly into the non-inverting input with no input termination and neglect any noise contribution from i_{ni}, R_f and R_g, *Equation (4)* will reduce to:

$$NF^+_{est} \approx 10\log\left[1 + \frac{e_{ni}^2 + \left(i_iR_f/A_v^+\right)^2}{4kTR_S}\right] \quad (5)$$

This expression for NF matches that appearing in several of the National hybrid amplifier data sheets (e.g. CLC200, CLC201, CLC103, CLC203, CLC220 and CLC221), where the Δf term has been replaced by a 1 in *Equation (5)* to consider only spot noise figure.

Equation (4) above is the most general expression for an op amp's non-inverting spot noise figure, considering an arbitrary input termination resister R_T and all possible contributing terms (even though some may prove negligible). The simplified *Equation (5)* assumes no input terminating resister and neglects any noise contribution of the op amps non-inverting input bias current noise and the feedback and gain setting resister noises. The expression found in the CLC205 and CLC206 data sheets include an arbitrary R_T , but still neglected the noise contribution of R_f and R_g.

Having labored through all of this to clarify where some of the earlier noise figure expressions published by National came from, we can now step to the most useful form of the noise figure expression where $R_S = R_T$. Doing this in *Equation (4)* yields:

$$NF^+ = 10\log\left[2 + \frac{e_{ni}^2 + \left(i_{ni}\frac{R_S}{2}\right)^2 + \left(\frac{i_iR_f}{A_v^+}\right)^2 + \frac{4kTR_f}{A_v^+}}{kTR_S}\right]$$

With $R_S = R_T$ (6)

This is probably the most useful formulation of the noise figure for the non-inverting op amp. The "2" arises from the signal attenuation we take in getting from the source to the input by using an external, noisy, resister matched to R_S (e.g. R_T). Note that the noise figure will decrease as the signal gain is increased due to the two numerator terms showing an A_v + in their denominators. Also note that for current feedback amplifiers, the feedback resister R_f is fixed to satisfy the amplifier's loop gain phase margin requirements (application note OA-13 discusses relaxing this requirement somewhat). Hence, the latter two terms in *Equation (6)*'s numerator do indeed decrease with increasing gain. If R_f were not particularly constrained in value, as with voltage feedback amplifiers, the R_f/A_v + term appearing in the last two terms of *Equation (6)* would probably make more sense if replaced by an $R_f||R_g$ term.

Inverting Op Amp Noise Figure

In this case, the discussion will be simplified by constraining the input impedance of the op amp to be equal to R_S. *Figure 29* shows the circuit for analysis with all of the contributing noise sources:

RT has been retained on the non-inverting input, along with its noise voltage source, for complete generality. R_g's noise now appears as a voltage source instead of the current noise term used in the non-inverting analysis. Again, developing the noise figure expression for the inverting amplifier configuration is simply a matter of resolving N_a and N_i and placing these expressions into *Equation (1)*. Knowing that the input impedance is matched to R_S, 1/2 of the noise voltage attributed to R_S will be delivered to the input port of the amplifier. This yields a (voltage) 2 at the input.

Inverting Op Amp Noise Figure

(Continued)

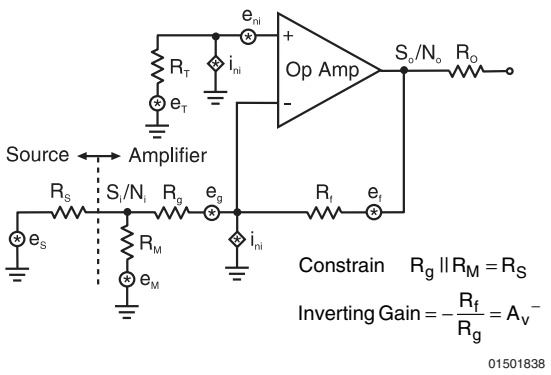

FIGURE 29. Inverting Op Amp Noise Figure Analysis

$$e_i^2 = \left(\frac{1}{2}e_S\right)^2 = \frac{1}{4}4kTR_S = kTR_S \quad (7)$$

Table II shows each individual noise terms, except e_s, with each term's "gain" to the inverting input. The noise terms on the non-inverting input have a gain of A_t to the inverting input. This represents the non-inverting gain to the output divided by the inverting gain back to the inverting input. The two resistor noise terms for R_M and R_g are taken to have a voltage gain to the inverting input defined simply by the resistor divider networks and simplified with the constraint on R_M that it is to be set to yield $R_g||R_M = R_S$. It is, perhaps, easiest to confirm the gain equations for R_g 's and R_M 's noise by computing the current those voltages generate into R_g , taking this current to the output by multiplying by R_f and then reflecting back to the inverting input by dividing by A_v - = R_f /R_g . Doing this and then substituting in for R_M , as shown in Table II, will (with some manipulation) yield the simple gain expressions found in Table II. The inverting noise current and R_f noise voltage are taken to the output then reflected back to the inverting input by dividing by the inverting gain.

TABLE 2. Noise Terms Contributing to N_a for the Inverting Op Amp Configuration

Noise Source	Value	Voltage Gain to Input
Non-inverting input voltage noise	e_n	A_T
Non-inverting input current noise	in_i	R_TA_T
Non-inverting input source resistor noise	$\sqrt{4kT/R_T}$	A_T
Inverting input current noise	i_i	$\frac{R_f}{A_v^-}$
Inverting input impedance matching resistor noise	$\sqrt{4kT/R_M}$	$\frac{1}{2}\left[1-\frac{R_f}{A_v^-}\right]$
Gain setting resistor	$\sqrt{4kT/R_g}$	$1-\frac{1}{2}\frac{R_s}{R_g}$
Feedback resistor voltage noise	$\sqrt{4kT/R_f}$	$\frac{1}{A_v^-}$

where:

$$A_v^- = \frac{R_f}{R_g} \text{ and } A_T = \frac{1 + A_v^- \left(1 - \frac{1R_S}{2R_g}\right)}{A_v^-}$$

$$\text{For } A_v^- > \sqrt{\frac{2R_f}{R_S}}, A_{T<1}$$

$$R_M = \frac{R_g R_S}{R_g - R_S} \text{ to get } R_M \| R_g = R_S$$

The noise terms on the non-inverting side of the op amp have a gain of A_T to the inverting input. As A_v^- increases, this gain drops to < 1, which contributes to the lower noise figure achievable using the inverting amplifier configuration. Again, an expression for a noise (voltage) 2 at the input may be obtained by taking the sum of the squared product of each noise source and associated gain shown in Table II.

$$e_a^2 (e_n A_T)^2 + (i_m R_T A_T)^2 + 4kTR_T A_T^2 + \left(\frac{i_i R_f}{A_v}\right)^2 +$$

$$4kTR_m \left(\frac{1}{2}\left(1 - \frac{R_S}{R_g}\right)\right)^2 + 4kTR_g \left(1 - \frac{1R_S}{2R_g}\right)^2 + \frac{4kTR_f}{(A_v^-)^2}$$

Combining the two noise powers attributed to the input matching network will allow considerable simplification in the final inverting noise figure expression. Substituting in for R_M with the expression shown as part of Table II and expanding the squared gain expressions.

$$\frac{4kTR_g R_S}{R_g - R_S}\left[\frac{1}{4}\frac{(R_g - R_S)^2}{R_g^2}\right] + 4kTR_g \frac{1}{4}\left(2 - \frac{R_S}{R_g}\right)^2 =$$

$$kTR_S \left(1 - \frac{R_S}{R_g}\right) + kTR_S \left[4\frac{R_g}{R_S} - 4 + \frac{R_S}{R_g}\right] = kT\left(4R_g - 3R_S\right)$$

Putting this back into the inverting e_a^2 expression and grouping the non-inverting input noise terms together yields:

$$e_a^2 = \left(e_n^2 + (i_{ni} R_T)^2 + 4kTR_T\right) A_T^2 + \left(\frac{i_i R_f}{A_v^-}\right)^2 + kTR_S \left(4\frac{R_g}{R_S} - 3\right) + \frac{4kTR_f}{\left(A_v^-\right)^2} \quad (8)$$

Putting the expressions for inverting e_a^2 and e_i^2 (*Equation (8)* and *Equation (9)*) back into the noise figure expression (Equation 1), and recognizing that dividing each term by R_S will yield N_a and N_i respectively, shows that the kTR_s term that arose in e_a^2 from the R_M and R_g resistor noises will collapse to a simple term, not including kTR_s (very reminiscent of the "2" appearing in the NF+ expression). One difference is that this part of the expression includes the contribution of both R_M and R_g, while the non-inverting equation kept the R_g noise as part of the equivalent input noise. This arises since R_g is now constrained by the input impedance matching requirement and can, therefore, be taken into this simplified form. The inverting noise figure NF- is then:

$$NF^- = 10\log\left(2\left(2\frac{R_g}{R_S} - 1\right) + \frac{(e_n^2 + (i_{ni} R_T)^2 + 4kTR_T) A_T^2 + \left(\frac{i_i R_f}{A_v^-}\right)^2 + \frac{4kTR_f}{(A_v^-)^2}}{kTR_S}\right) \quad (9)$$

To compare the non-inverting noise figure expression (*Equation (7)*) to the inverting expressions (*Equation (9)*), note that noise terms on the non-inverting input side have a gain of 1 for the non-inverting configuration but a gain of A_t for the inverting. Also note that the term associated with the feedback resistor noise is divided by simply A_v^+ in the non-inverting case. This arises because it also includes the R_g noise in the non-inverting expression. However, it is divided by $(A_v^-)2$ in the inverting case. This arises from the R_g noise term being considered part of the input termination. In this case, R_g and R_M's noise, appears in $2*(2*R_g/R_s - 1)$ as part of the noise figure expression. Note that this collapses to simply equal 2 when $R_g = R_s$ similar to the NF+ case.

At low inverting gain, the non-inverting input noise terms have a larger impact for the inverting configuration than for an equivalent non-inverting gain, yielding a higher noise figure. As A_v^- increases, however, the non-inverting noise terms will be attenuated in going to the inverting signal input reference point, yielding a lower inverting noise figure than for an equivalent gain non-inverting configuration.

Figure 30 compares the noise figures over gain for the non-inverting vs. inverting configurations using the CLC404. For this comparison, R_f is assumed fixed at 500Ω, R_T for the non-inverting case = 50Ω; but is set to 25Ω for the inverting case. With these constraints, R_g and R_M will be set by the desired gain and the requirement that $R_g \| R_M = R_S$ in the inverting mode. R_S is assumed = 50Ω throughout. The inverting noise figure plot simply stops at the point where R_g = 50Ω, since higher gains are not possible (with a fixed R_f) while retaining the input impedance matching requirement. Also note that log gains are shown on the x-axis to the matched load, while the voltage gains used in the noise figure calculations are the linear voltage gains to the output pin.

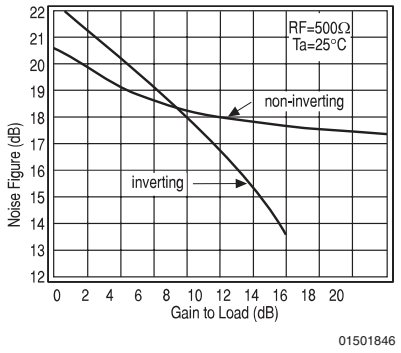

FIGURE 30. Noise Figure vs. Gain For the CLC404

Dynamic Range Calculation

Having developed the 3 limits to dynamic range commonly used in describing RF amplifiers as they apply to op amps, it is now possible to combine them into a single dynamic range number. The usable dynamic range is typically described in terms of the difference between the minimum detectable signal at the amplifier output and either the -1dB compression or the output power that would bring the 3rd order spurious up to this minimally detectable level. As described in reference 5 (page 175) the minimally detectable signal at the output of an amplifier is:

$Po_{MDS} = kTG_A(NF)BX$

Dynamic Range Calculation

(Continued)

where:

kT →	Noise power delivered to matched input in dBm
GA →	Power gain in dB
NF →	Noise Figure
B →	Noise bandwidth
X →	Additional margin above the noise floor for detectability; typically 3dBm

Note that the kTGA(NF) term solves to yield the spot output noise power. Substituting for NF and recalling that k_T = noise power delivered to the input matching resistor = N_i.

$$kTG_A(NF) = N_i G_A \left[\frac{S_i/N_i}{S_0/N_0}\right] = N_0 G_A \frac{S_i}{S_0} = N_0$$

Adding 10*log(B) will then show the integrated noise floor at the output of the amplifier. It is important to remember that this bandwidth need not be the bandwidth of the amplifier itself. It is advantageous to bandlimit the response as narrowly as possible at some point after the amplifier immediately prior to the desired signal extraction. It is this later bandlimiting bandwidth that would be used in the equation for determining the minimum detectable signal.

As an example, consider the CLC404 circuit used throughout this discussion with a post-filter to yield a bandwidth of interest from 10MHz to 20MHz. The non-inverting, gain of 9.54dB topology would yield a minimum detectable power level at the output:

$$Po_{MDS} = 10\log(kT) + 10\log(G_A) + NF + 10\log(10MHz) + 3db = -174dBm + 9.54dB + 18dB + 70dB + 3dB = -73.5dBm$$

Where the noise figure was read off of *Figure 30* and

$$10\log\left(\frac{kT}{.001}\right) = 10\log\left(1000(4E-21)\right) = -174dBm.$$

Having determined the minimally detectable signal, at the output, a maximum output signal set by some constraint will determine the dynamic range. Typically, a simple dynamic range specification uses the -1dB compression power as the maximum output power. From the earlier discussion on -1dB compression, we know that the actual output power at the fundamental frequency is -1dBm less than the reported -1dB compression point, and, that the true achievable sinusoidal power is approximately 2.1dBm less than this due to increase in power showing in the fundamental when the output is approaching a square wave. With these considerations, it would seem more realistic to use a maximum output power 3dBm less than the measured -1dB compression power. Going to *Figure 17* and subtracting 3dBm from the measured -1dB compression at the maximum operating frequency will yield a maximum usable output power at the matched, load of 14.5dBm. Subtracting the minimum detectable signal at the output from this shows a 14.5 – (-73.5) = 88dB dynamic range.

An alternative approach is to define a spurious free dynamic range. This approach sets the maximum output power to yield a 3rd order spurious level equal to the minimum detectable signal. At this point, the amplifier generated spurious is just equalling what can be detected from a noise floor consideration. From *Figure 23*, the 3rd order spurious levels are:

$$P_S = 2\left(\frac{3}{2}P_0 - IM3\right)$$

Setting this equal to the minimum detectable signal and solving for P_o –

$$P_S = 2\left(\frac{3}{2}P_o - IM3\right) = -174dBm + 10\log G_A + 10\log(B) + 3dB$$

$$P_{o\,max} = \frac{1}{3}\left[-174dBm + 10\log G_A + NF + 10\log(B) + 3dB + 2(IM3)\right]$$

Putting in the previously develop minimum detectable signal and pulling the 20MHz intercept from *Figure 24* shows a maximum spurious free output power of –

$$P_{0_{max}} = \frac{1}{3}\left[-73.5 + 2(28)\right] = -5.8\,dBM$$

and a spurious free dynamic range of:

DRf = -5.8dBm – (-73.5dBm) = 67.7dB

An additional check on this spurious free dynamic range is to recall that the total output power for the equal power 2-tone condition (that will generate the spurious level at the minimum detectable signal) is actually a voltage envelope that is twice the individual signals, or, 6dBm higher in power. This would imply a 0dBm total output power when the spurious is just rising above the noise floor. This is well below the dynamic range set by the -1dB compression limit.

Another way to interpret the 3rd order spurious plot of *Figure 24* is to compute the absolute spurious power level as the output power is swept higher and simply compare that to the minimum detectable signal power at the output. The absolute spurious power can be derived from the data of *Figure 24* as simply the x-axis value minus the y-axis value, (P_O -(P_O - P_S)) = P_S . When the measured spurious free range is in a region of -1 slope (2→4dBm on the 20MHz line), P_S is remaining constant as the output power increases. Using the actual measured data, as opposed to an intercept, can become a more appropriate way to compare the spurious power to the noise floor when the spurious level begins to level out and become constant (for a part like the CLC404) just below the minimum detectable signal.

For example, if the minimum detectable signal were actually -68dBm in the example considered earlier, the 3rd order spurious would equal this level at 4dBm output power instead of the -4dBm level that would be predicted from the equation used above with the low power estimate of a 28dBm intercept. Increasing the amplifier gain by 6dB would raise the output noise floor to -67.5 to take advantage of this improved spurious performance at higher output powers. Doing this would actually yield a 72dB spurious free dynamic range vs. the 68dB calculated earlier.

The primary determinants to dynamic range are noise power bandwidth, noise figure, -1dB compression, and 2- tone, 3rd order spurious. To maximize dynamic range, the following steps may be taken –

Dynamic Range Calculation

(Continued)

1. Limit the noise power bandwidth after the amplifier as much as possible.
2. To reduce noise figure, run the amplifier at as high a gain as possible consistent with bandwidth limitations and/or use the amplifier at high inverting gains. Alternatively, using a transformer coupled non-inverting amplifier configuration as described in National application note OA-14 can typically reduce the noise figure to the 6dB level.
3. If the -1dB compression limits are inadequate, use a higher supply voltage amplifier (such the hybrid amplifiers offered by National), or, increase the power supply voltage above the recommended value. National's low voltage monolithic amplifier's specify a maximum voltage across the supply pins of 14 volts. Increasing the supplies from ±5V to ±6V (or using a single 12V supply and the single supply circuits described earlier) will typically increase the maximum usable output power by 2dBm.
4. Be sure to consider the actual spurious performance if an intercept characteristic is not followed. Increasing the supply current (if possible) or increasing the load impedance can dramatically drop the 3rd order distortion terms. Recall that the feedback network remains as an upper limit on the output loading. Reference 4 describes an additional technique of loop gain shaping that can be used to further improve the distortion performance.

Conclusions

High speed current feedback amplifier's can offer considerable performance advantages when used in IF and RF applications. The flexible gain and I/O impedance capability can be used to the designers benefit in tailoring the amplifier to the specific requirement. Last minute gain changes can be accommodated with resistor value changes as opposed to requiring a new amplifier. Exceptional I/O VSWR and reverse isolation are easily attainable using wideband op amps. Although somewhat different, the dynamic range can be calculated, or measured, and compared between op amps and more typical RF amplifiers.

One of the most significant advantages of wideband current feedback amplifiers is the low 3rd order spurious level for their relatively low quiescent power dissipation. Most of National's monolithic amplifier's dissipate less than 150mW while delivering in excess of 40dBm intercepts below 10MHz. The primary drawback to the closed loop op amps are their rapid rolloff in distortion performance as the loop gain decreases at higher frequencies. Another area for improvement are the relatively high noise figures using standard op amp topologies. Using an input transformer can reduce the overall noise figure to around 6dB (see National application note OA-14). Additional external circuit techniques, along with a new low noise op amp (the CLC425), show a potential for noise figures as low as 2dB. For IF and RF applications below 100MHz, and particularly below 50MHz (when a high spurious free dynamic range is required), a wideband op amp solution can probably offer significant performance, power dissipation, and price advantages over more typical fixed gain amplifiers. Appendix A summarizes the comparison between RF amplifiers and wideband op amps.

References:

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5. Microwave Transistor Amplifiers: Analysis and Design, Guillermo Gonzalez; Prentice Hall, Inc., 1984.

The circuits included in this application note have been tested with National Semiconductor parts that may have been obsoleted and/or replaced with newer products. Please refer to the CLC to LMH conversion table to find the appropriate replacement part for the obsolete device.

Note: The circuits included in this application note have been tested with National Semiconductor parts that may have been obsoleted and/or replaced with newer products. Please refer to the CLC to LMH conversion table to find the appropriate replacement part for the obsolete device.

Appendix A
Amplifier Comparison Table

Parameter	RF Amplifiers	National Semiconductor Op Amps
Gain	Almost always fixed gain	Easily adjustable over wide range
Bandwidth	Limited capacity < 1MHz. Can be very high frequency. Generally, 2 decade range .	DC capability. Upper limit around 100MHz to match RF amp flatness specs
I/O VSWR	Typically 1.5:1	Can be tuned to much better match through 100MHz than RF amps
Reverse Isolation	20 to 30dB considered good. Not too frequency dependent.	Much better isolation possible degrades at high frequencies. Better non-inverting than inverting.
Noise Figure	Can be very low 2 to 5dB typical	Varies with gain setting. Higher gains better but bottoming out about 12dB for typical op amp. Circuit can be improved to < 5dB

Appendix A
Amplifier Comparison Table (Continued)

Parameter	RF Amplifiers	National Semiconductor Op Amps
3rd Order	Moderate levels, needs high I_{cc} for excellent numbers. Relatively frequency independent.	Very good intercepts for quiescent power. Strong frequency dependent-degrading rapidly at Can be improved at low frequencies. See App. Note OA-22.
–1dB Compression	Good levels for voltage supplies. Relatively frequency independent. -1dB + 3rd intercept are order related.	Requires more head room for available output power. Drops rapidly with frequency due to slew rate -1dB and 3rd order not related.
Supply Current	Usually single polarity supply. High quiescent current vs. PoCapability.	Bi-polar supplies. Almost all can be run single supply. Much lower quiescent currents for PoCapability.

Appendix B
Harmonic & Intermodulation Terms for a 5th Order Polynomial Transfer Function

For an input signal that is two sinusoidal signals

$V_i = A\cos 2\pi f_1 t + B\cos 2\pi f_1 t$

$f_1 = f_o - \Delta f$

$f_2 = f_o - \Delta f$

Processed through a 5th order polynomial transfer function

$V_0 = K_0 + K_1 V_i + K_2 V_i^2 + K_3 V_i^3 + K_4 V_i^4 + K_5 V_i^5$

yields the following frequencies and coefficients:

Frequency terms and coefficients in order of ascending frequency

Frequency	Coefficient	
DC	$K_O + K_2 (A^2/2 + B^2/2) + K_4 (3/8 * A^4 + 3/8 * B^4 + 3/2 * A^2B^2)$	
$2\Delta f$	$(K^2 + AB/2) + 2K_4 (A_3B + AB^3) + 5/4 * K_5 (A^4B + AB^4)$	2nd Order Intermod
f_0	$K_5 * 10/16 * A^3B^2$	5th Order Intermod
$f_0 - 3\Delta f$	$K_3 * 3/4 * AB^2 + 30/16 * K_5A^2B^3$	3rd Order Intermod
$f_0 - \Delta f$ (=f_1)	$K_1 + K_3 (3/4 * A_3 + 3/2 * A^2B) + K_5 (5/8 * A^5 + 30/8 * A^3B^2 + 15/8 * AB^4)$	Lower Test Tone f_1
f_0	No coefficient	
$f_0 - \Delta f$ (=f_2)	$K_1 + K_3 (3/4 * B_3 + 3/2 * AB^2) + K_5 (5/8 * B^5 + 30/8 * A^2B^3 + 15/8 * A^4B)$	Upper Test Tone f_2
$f_0 + 3\Delta f$	$K_3 * 3/4 * A^2B + 30/16 * K_5A^3B^2$	3rd Order Intermod
$f_0 + 5\Delta f$	$K_5 * 10/16 * A^2B^3$	5th Order Intermod
$2f_0 + 4\Delta f$	$2K_4A^3B$	4th Order Intermod
$2f_0 + 2\Delta f$ (=$2f_1$)	$K_2 * A_2/2 + K_4 * A^4/2 + 3/2 * K_4A^2B^2$	2nd Harmonic for f_1
$2f_0$	$K_2 * AB/2 + 2K_4 (A^3B + AB^3) + K^5 * 5/4 * (A^4B + AB^4)$	2nd Order Intermod
$2f_0 + 2\Delta f$ (=$2f_2$)	$K_2 * B^2/2 + K_4 * B^4/2 + 3/2K_4A^2B_2$	2nd Harmonic for f_2
$2f_0 + 4\Delta f$	$2K_4AB^3$	4th Order Intermod
$3f_0 - 5\Delta f$	$5/16 * K_5A^4B$	5th Order Intermod
$3f_0 - 3\Delta f$ (=$3f_1$)	$K3/4 * A^3 + K_5/16 * A^5$	3rd Harmonic for f_1
$3f_0 - \Delta f$	$3/4 * K_3AB^2 + 30/16 * K_5A^2B^3$	Higher 3rd Order Intermod
$3f_0$	No Coefficient	
$3f_0 - \Delta f$	$3/4 * K_3A^2B + 30/16 * K_5A^3B^2$	Higher 3rd Order Intermod
$3f_0 - 3\Delta f$ (=$3f_2$)	$K3/4 * B^3 + K_5/16 * B^5$	3rd Harmonic for f_2
$3f_0 - 5\Delta f$	$5/16 * K_5AB^4$	5th Order Intermod
$4f_0 - 4\Delta f$ (=$4f_1$)	$1/8 * K_4A_4$	4th Harmonic for f_1
$4f_0 - 2\Delta f$	$2K_4A^3B$	4th Order Intermod
$4f_0$	No Coefficient	

Appendix B
Harmonic & Intermodulation Terms for a 5th Order Polynomial Transfer Function (Continued)

$4f_0 - 2\Delta f$	$2K_4AB^3$	4th Order Intermod
$4f_0 - 4\Delta f (=4f_2)$	$1/8 * K_4B^4$	4th Harmonic for f_2
$5f_0 - 5\Delta f (=5f_1)$	$1/16 * K_5A^5$	5th Harmonic for f_1
$5f_0 - 3\Delta f$	$5/16 * K_5A^4B$	5th Order Intermod
$5f_0$	No Coefficient	
$5f_0 - 3\Delta f$	$5/16 * K_5AB^4$	5th Order Intermod
$5f_0 - 5\Delta f (=5f_2)$	$1/16 * K_5B^5$	5th Harmonic for f_2

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