# Voltage feedback vs. current feedback amplifiers: Advantages and limitations

Xavier Ramus

February 26th, 2009

# Agenda

- Voltage Feedback Amplifier
  - Analytical Models & Architecture
    - Simple Model
    - Advanced Model
    - Input Architecture and trade-offs
  - Limitations
- Current Feedback Amplifier
  - Analytical Models & Architecture
    - Simple Model
    - Advanced Model
    - Input Architecture and trade-offs
  - Limitations
- Side-by-side comparison
- Applications:
  - Transmit Function
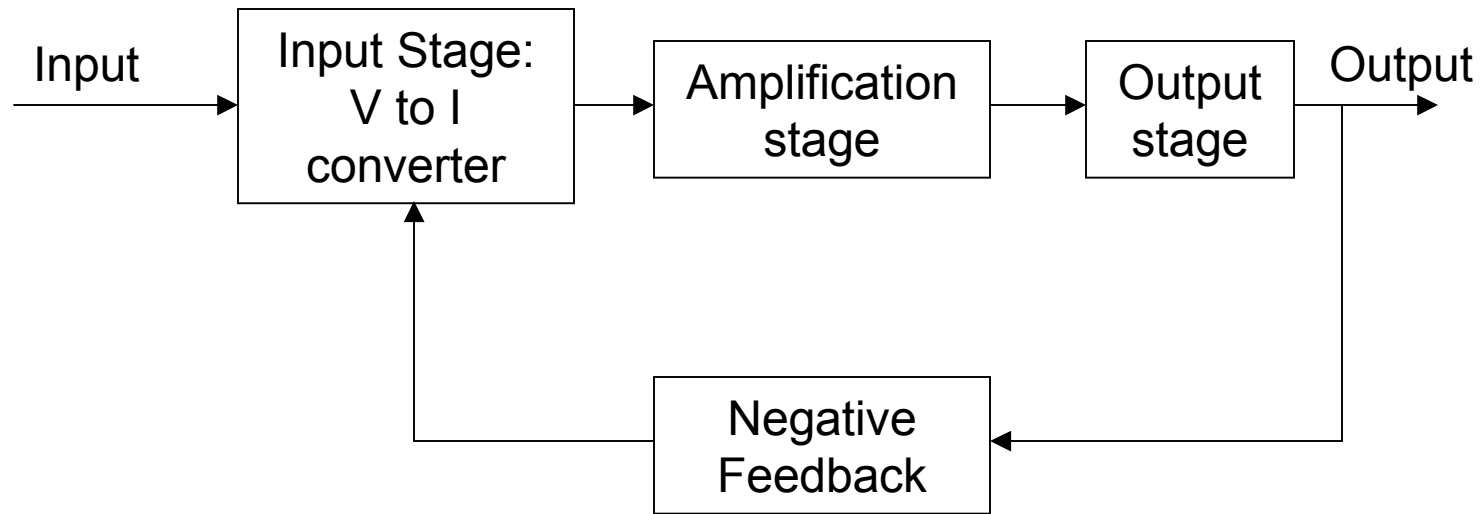  - Receive function
  - Filtering

# Overview

# Introduction

- Commercially available amplifiers are built on state of the art semiconductor processes –

    - These have great density, and good transistor parameters, but poorly controlled absolute specifications

- Open loop amplifiers are available, but they typically have poor power efficiency and more performance variation than closed loop devices.

    - The uA733 is an example of an open loop amplifier.

- More details on that device later, for now let's have a look at an operational amplifier structure

# Operational Amplifier structure

- This structure is valid for any operational amplifier
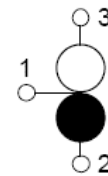- The implementation of the VI converter makes an operational amplifier of voltage mode or current mode.

Input → **Input Stage: V to I converter** → **Amplification stage** → **Output stage** → Output

**Negative Feedback**

# Typical Voltage Feedback Input Stage Architecture

- Note that both input are high impedance
- Note that the maximum current that this type of input stage can deliver to the following stage is limited by the total bias current.  In practice, this limits the maximum achievable slew rate given a quiescent current.  A voltage feedback amplifier with this architecture operating on 20mA may achieve 1000V/us.
- Specification typically better in VFB:
  - Input offset voltage
  - Matched input bias current
  - High power supply rejection ratio
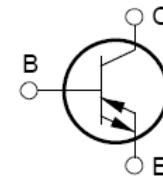  - Good common mode rejection ratio
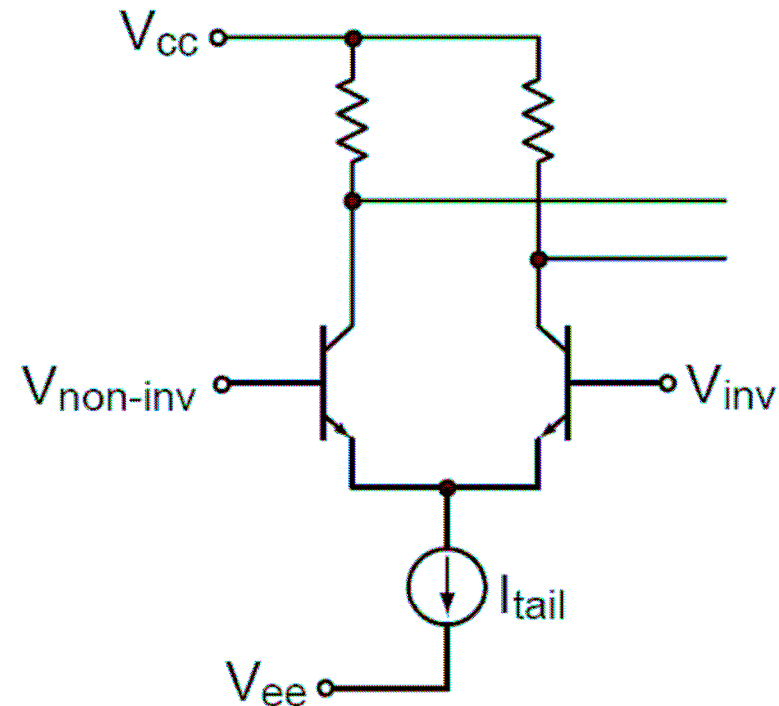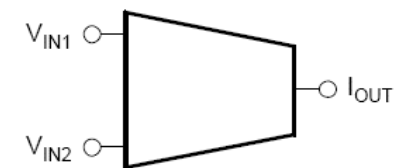
# Operational Transconductance Amplifier

- An operational transconductance amplifier is a voltage - controlled current source or in other term a V to I converter.

- This is an open-loop amplifier and can be represented as shown here.
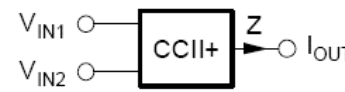
- One such device is the OPA861

# OPA861

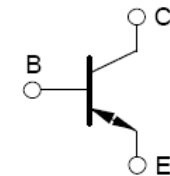## Wide Bandwidth Operational Transconductance Amplifier

### Features

- Wide Bandwidth OTA (>80Mhz)
- 900V/usec Slew Rate
- 2.4nV/√Hz Input Noise Voltage
- Very Flexible Circuit Building Block
- Ideal Complementary Transistor Function
- Tuneable Transconductance
- Externally settable supply current

### Benefits

- Simple to use Ideal Transistor
- Very high speed flexible circuit element
- Externally adjustable transconductance
- High I/O voltage range (+/-4.2V on +/-5V supply)
- Low Cost Filter Design Element

### Applications

- DC Restore Circuits
- NIC Filters
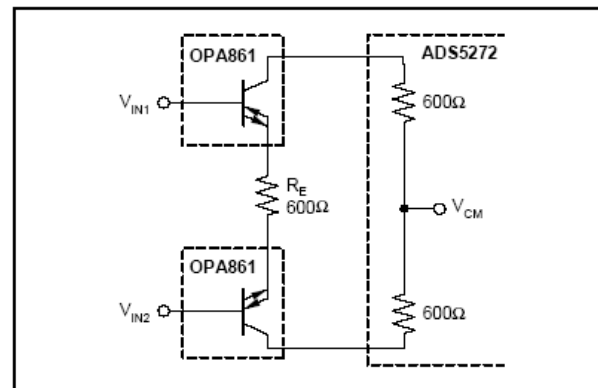- High CMRR ADC Driver
- Capacitive Load Driver

1ku / $0.95

EVM



Figure 45. High CMRR, Moderate Precision, Differential I/O ADC Driver
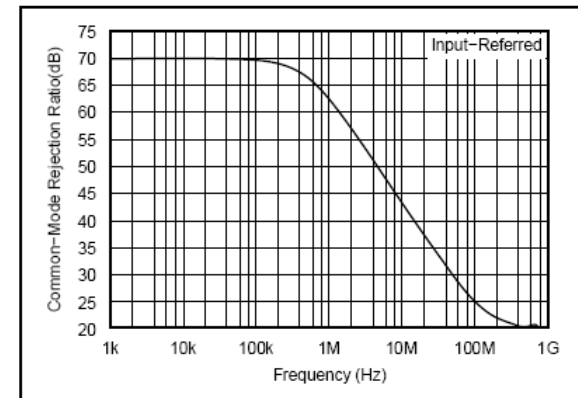


Figure 47. CMRR of the ADC Driver

This 70dB CMRR at 100kHz compares to 25dB for the INA331

TEXAS INSTRUMENTS

# Operational Transconductance Amplifier
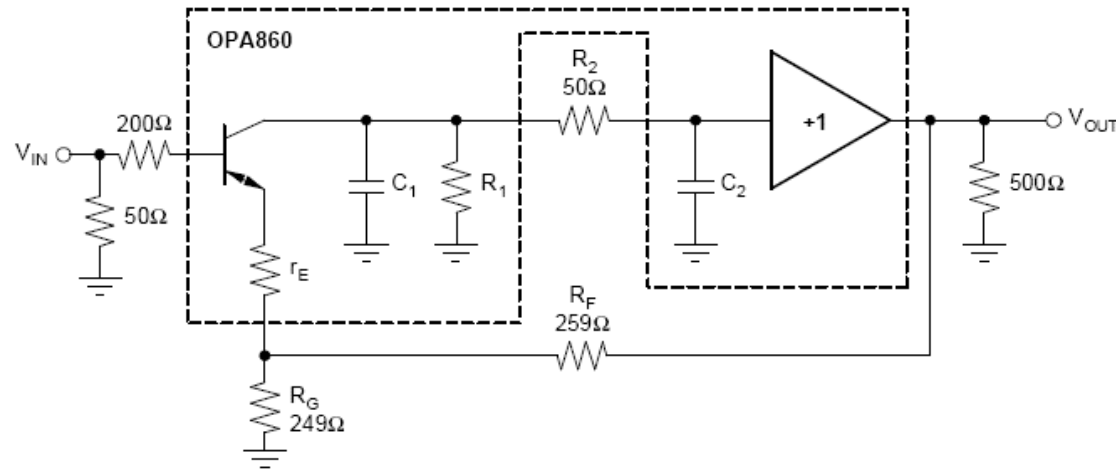


- Combining a transconductance amplifier with a buffer and adding some negative feedback gives the architecture of a current feedback amplifier.
- An older device, the OPA860 shows how this architecture was initially introduced in an integrated circuit.  Both devices were separated to allow a large variety of application circuits from the same package.

TEXAS INSTRUMENTS

# Current Feedback Amplifier
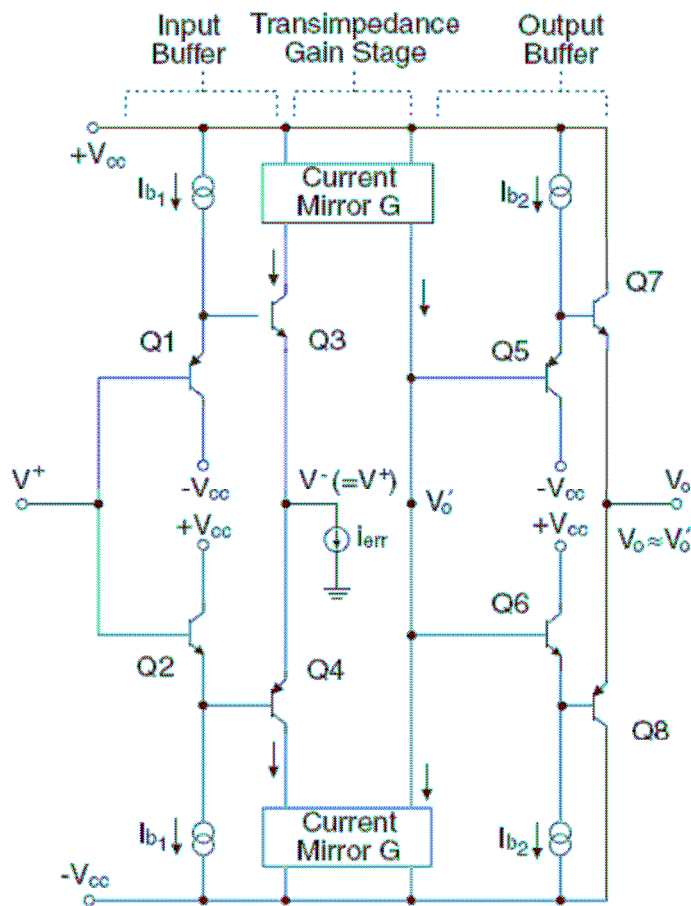


- A simplified schematic of a CFB is constituted of:
  - An input buffer
  - A transimpedance stage
  - An output buffer
- $V_O'$ is the compensation node that in conjunction with a feedback resistor selected during design provides the optimum bandwidth while guaranteeing unconditional stability.
- The feedback resistor is the compensation element for a CFB. Reducing its value will decrease the phase margin directly and possibly impacting stability depending on external components. Increasing its value will increase the phase margin, possibly compensating for a capacitive load zero.

# Current Feedback Input Stage Architecture
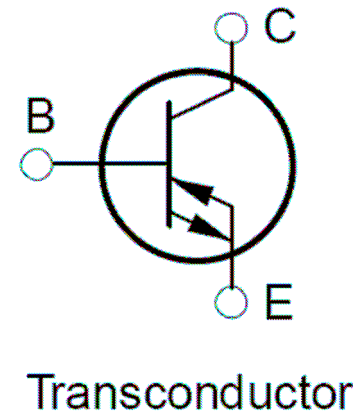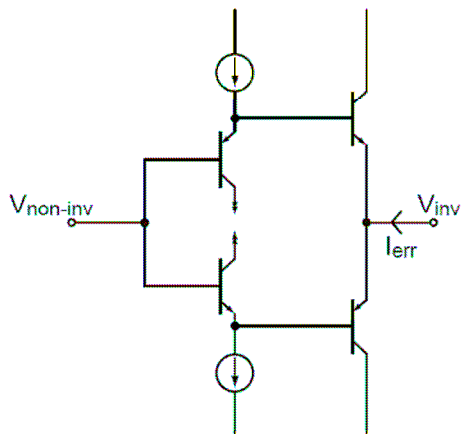


Transconductor

- The B-Input is the non-inverting input of a CFB and the E-input/output is the inverting input.
- The transconductor is nothing else but a bi-directional transistor.
- From a direct comparison with transistor, it comes that the B-input is high impedance and the E-input/output is low impedance
- Note the maximum current flowing out of C-output is not limited by the biasing of the circuit.  Hence CFB can routinely achieve 2000V/us slew rate on 5mA.  (2x the SR of a VFB on ¼ of the $I_Q$).
- The DC specification of a CFB inherent to the input stage architecture:
  - Non-zero input offset voltage
  - Unmatched Input bias current

TEXAS INSTRUMENTS

# Modern Voltage Feedback Amplifier

- Modern VFB use the same architecture as the CFB in order to reduce the power consumption while maintaining a large slew rate.

- This architecture eliminates the dependency of the SR on the $I_Q$ although at the cost of more circuitry. This architecture will achieve almost as good a SR as an equivalent $I_Q$ CFB.

- This approach will ensure a good matching between both inverting and non-inverting input of the amplifier restoring the typical VFB good DC performance.

$I_{OUT}$ to next stage

+IN

R

-IN
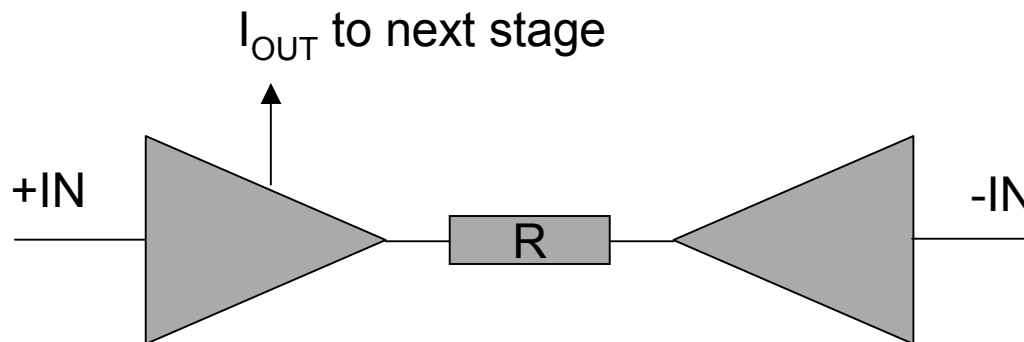
# Architecture Implications

- This difference in architecture generates changes in the following amplifier specifications
  - Stability concerns
  - Bandwidth & Gain Bandwidth Product
  - Open-loop gain
  - Noise
  - Input impedance
  - Slew Rate
  - DC precision
  - Distortion

- First we will have a closer look on modeling these architecture.

# Voltage Feedback Amplifiers

# Simplified VFB Analysis



$$\frac{V_o}{V_i} = \frac{-\dfrac{R_2}{R_1}}{1 + \dfrac{1 + \dfrac{R_2}{R_1}}{A(s)}}$$

$Z^+$

$V_1$  +

$Z_o$

$A(s)$  $V_o$

$V_2$  −

$Z^-$
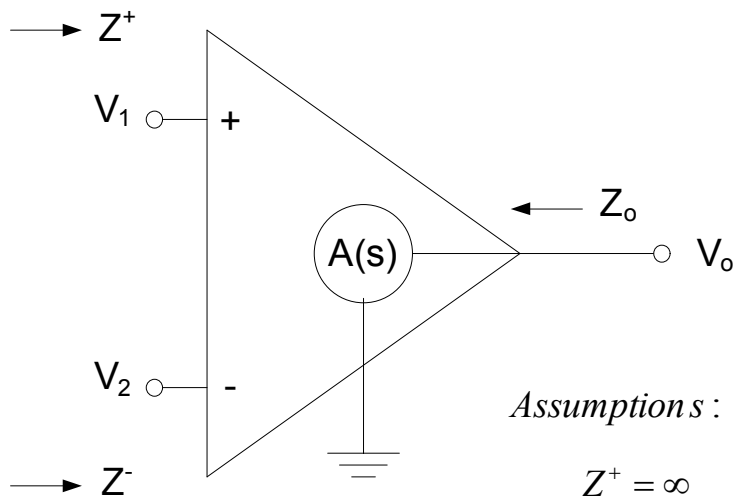
*Assumptions* :

$Z^+ = \infty$

$Z^- = \infty$

$Z_o = 0$

$V_o = A(s)[V_1 - V_2]$

$R_2$

$R_1$

$V_i$

−

VFB

+

$V_o$

TEXAS INSTRUMENTS
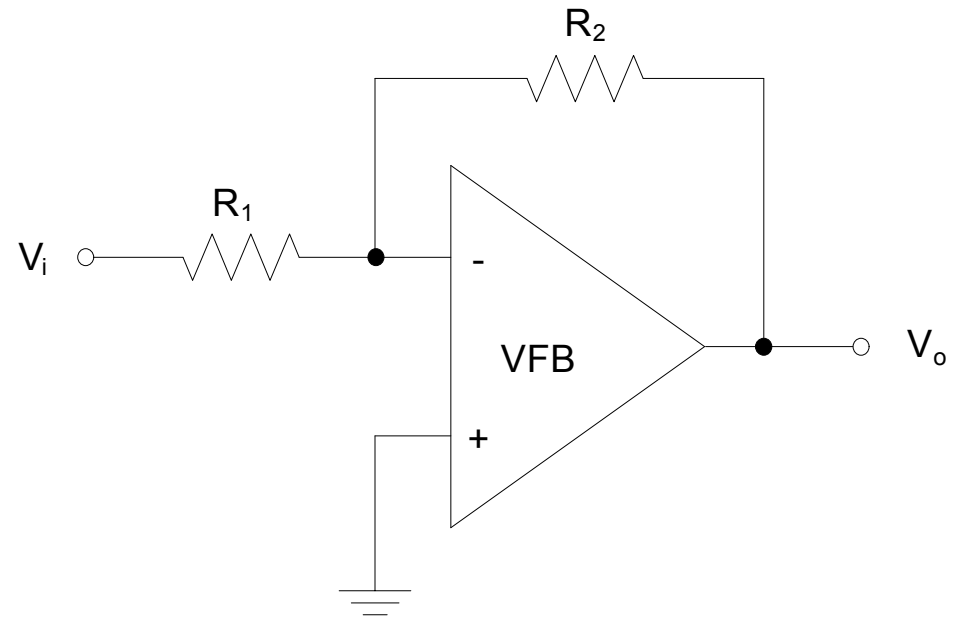
# Simplified VFB Loop Gain Analysis



20 Log (A$_{(S)}$)

Log Gains

Loop Gain

$$20\log\left(1+\frac{R_f}{R_g}\right)$$

Loop Gain x-over

Loop
Bandwidth

Log(f)

0°

$\angle(LG)$

Phase
Margin

-180°

TEXAS
INSTRUMENTS
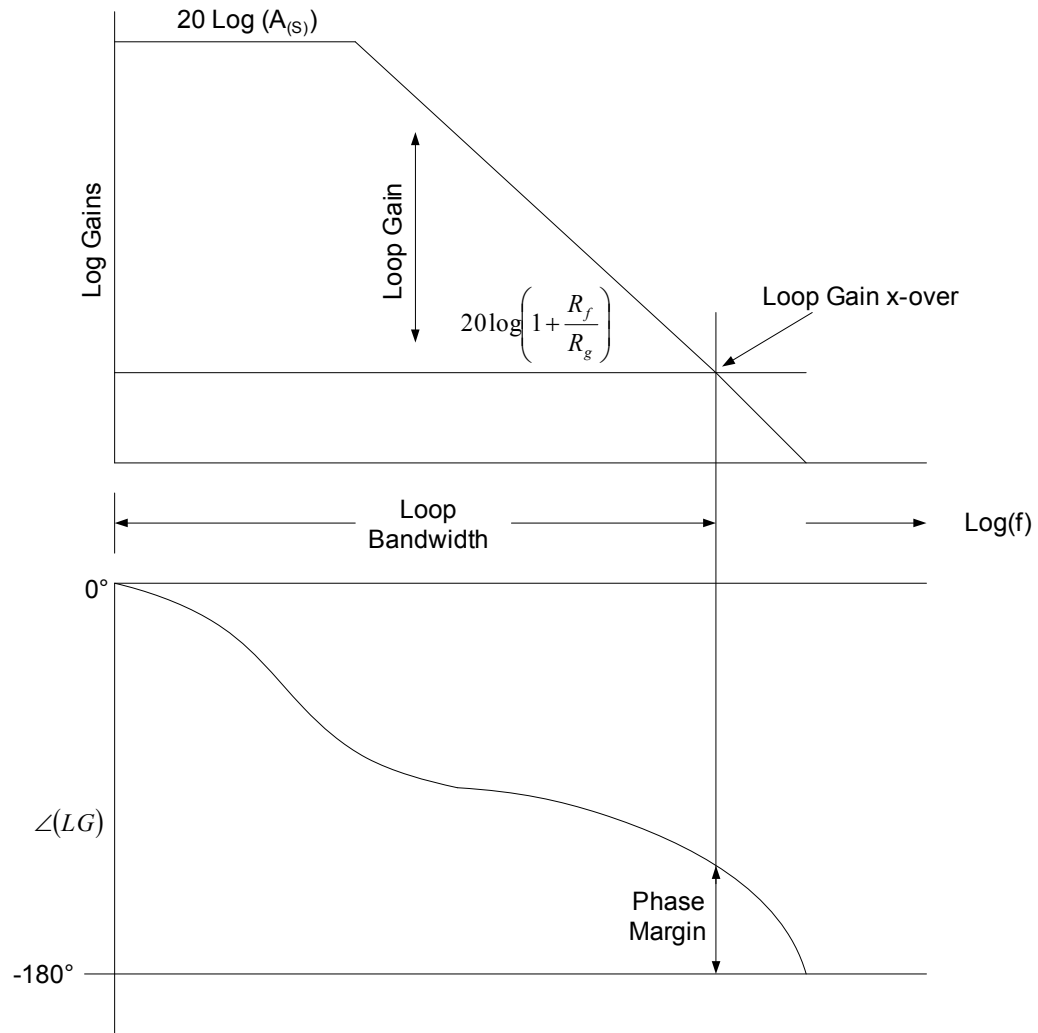
# Advanced Model

- The simplified model is sufficient in most cases to be able to predict VFB behavior.

- The advanced model adds input and output impedance, including parasitic.

- This model will becomes critical when looking for real world stability in transimpedance and capacitive load driving applications.

- The compensation mechanism of the VFB is provided by the noise gain. At low frequency, it is set by $1 + R_F/R_G$ and at high frequency it is set by $1 + C_G/C_F$.
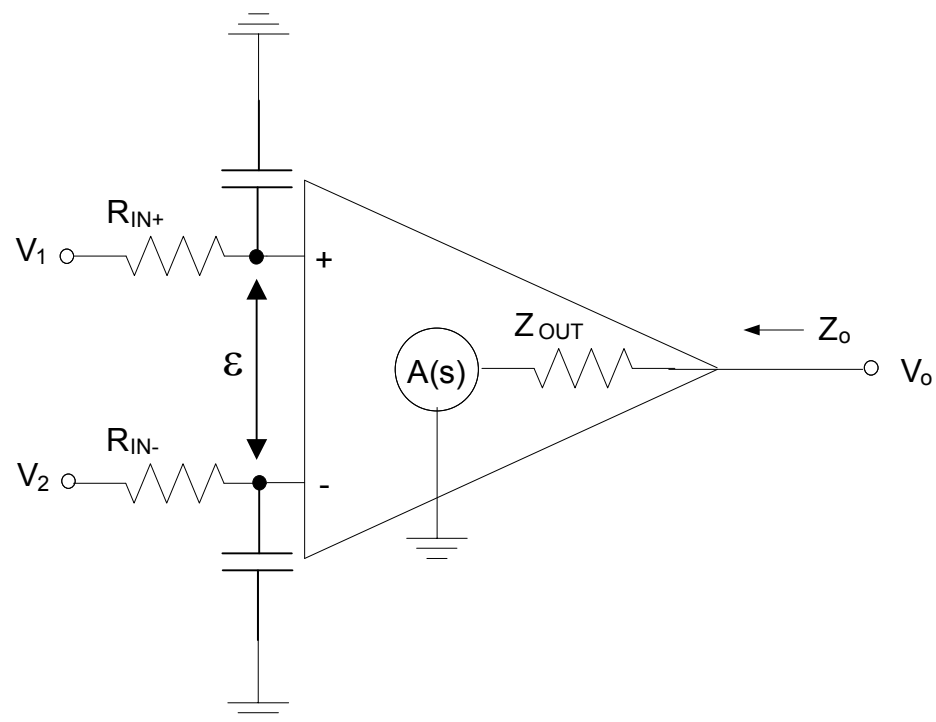
# Gain Bandwidth Product

- From the simplified model, you can easily see that the achievable bandwidth is directly related to the noise gain $1+R_F/R_G$. As the gain increase, the bandwidth decreases. In practice, for high speed amplifier, this is valid for large gain (>10V/V) only as peaking in the frequency response as well as package parasitic will influence the actual low gain bandwidth.
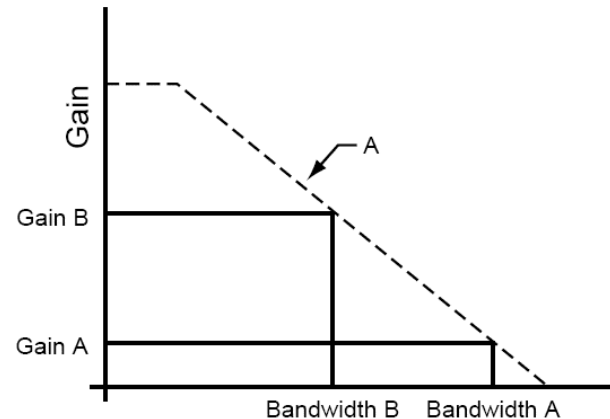
$$\frac{V_{OUT}}{V_{IN}} = \frac{-\frac{R_F}{R_G}}{1+\frac{1+\frac{R_F}{R_G}}{A(s)}}$$

TEXAS INSTRUMENTS

# Gain Bandwidth Product

- Note that to compare VFB using the GBWP figure of merit, the spec table may not always provide the information and you will have to look at the open loop gain and phase plot in the curve section.

- Here you have the OPA820 $A_{OL}$ plot showing a ~3MHz bandwidth at 40dB gain. This translates to ~300MHz GBWP. The specification table provides a 280MHz number. Note that the unity gain bandwidth show a exceptional 800MHz -3dB bandwidth.



OPEN-LOOP GAIN AND PHASE
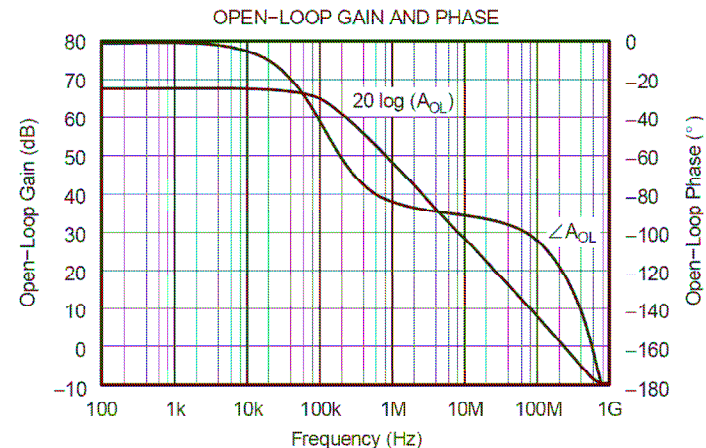
| | | | OPA820ID, IDBV | | | | | |
|---|---|---|---|---|---|---|---|---|
| | | TYP | MIN/MAX OVER TEMPERATURE | | | | | TEST |
| PARAMETER | CONDITIONS | +25°C | +25°C(1) | 0°C to 70°C(2) | −40°C to +85°C(2) | UNITS | MIN/ MAX | LEVEL (3) |
| AC PERFORMANCE | | | | | | | | |
| Small-Signal Bandwidth | G = +1, $V_O$ = 0.1$V_{PP}$, $R_F$ = 0Ω | 800 | | | | MHz | typ | C |
| | G = +2, $V_O$ = 0.1$V_{PP}$ | 240 | 170 | 160 | 155 | MHz | min | B |
| | G = +10, $V_O$ = 0.1$V_{PP}$ | 30 | 23 | 21 | 20 | MHz | min | B |
| Gain-Bandwidth Product | G ≥ 20 | 280 | 220 | 204 | 200 | MHz | min | B |

# Current Feedback Amplifiers

# *Simplified CFB Analysis*
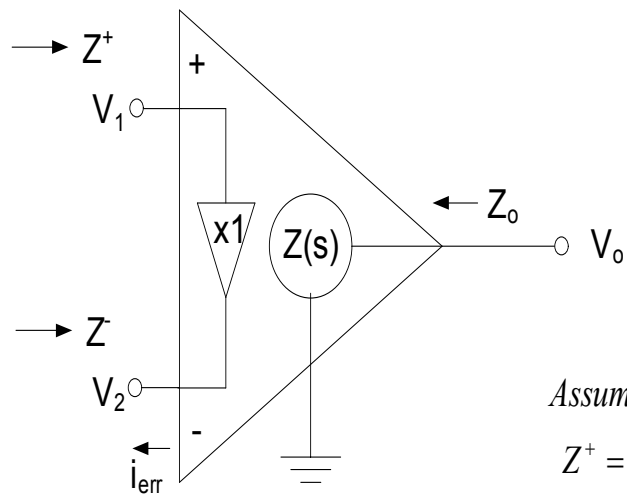


Assumptions:

$$Z^+ = \infty$$

$$Z^- = 0$$

$$Z_o = 0$$

$$V_o = Z(s)i_{err}$$

$i_{err}$ is the error <u>current</u>
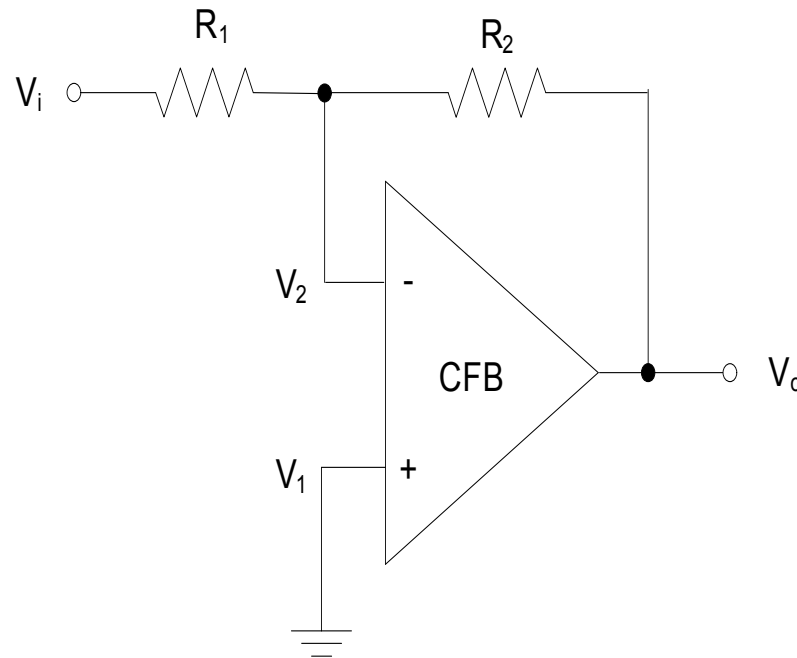
$$\frac{V_o}{V_i} = \frac{-\dfrac{R_2}{R_1}}{1 + \dfrac{R_2}{Z(s)}}$$

TEXAS INSTRUMENTS

# Simplified CFB Loop Gain Analysis
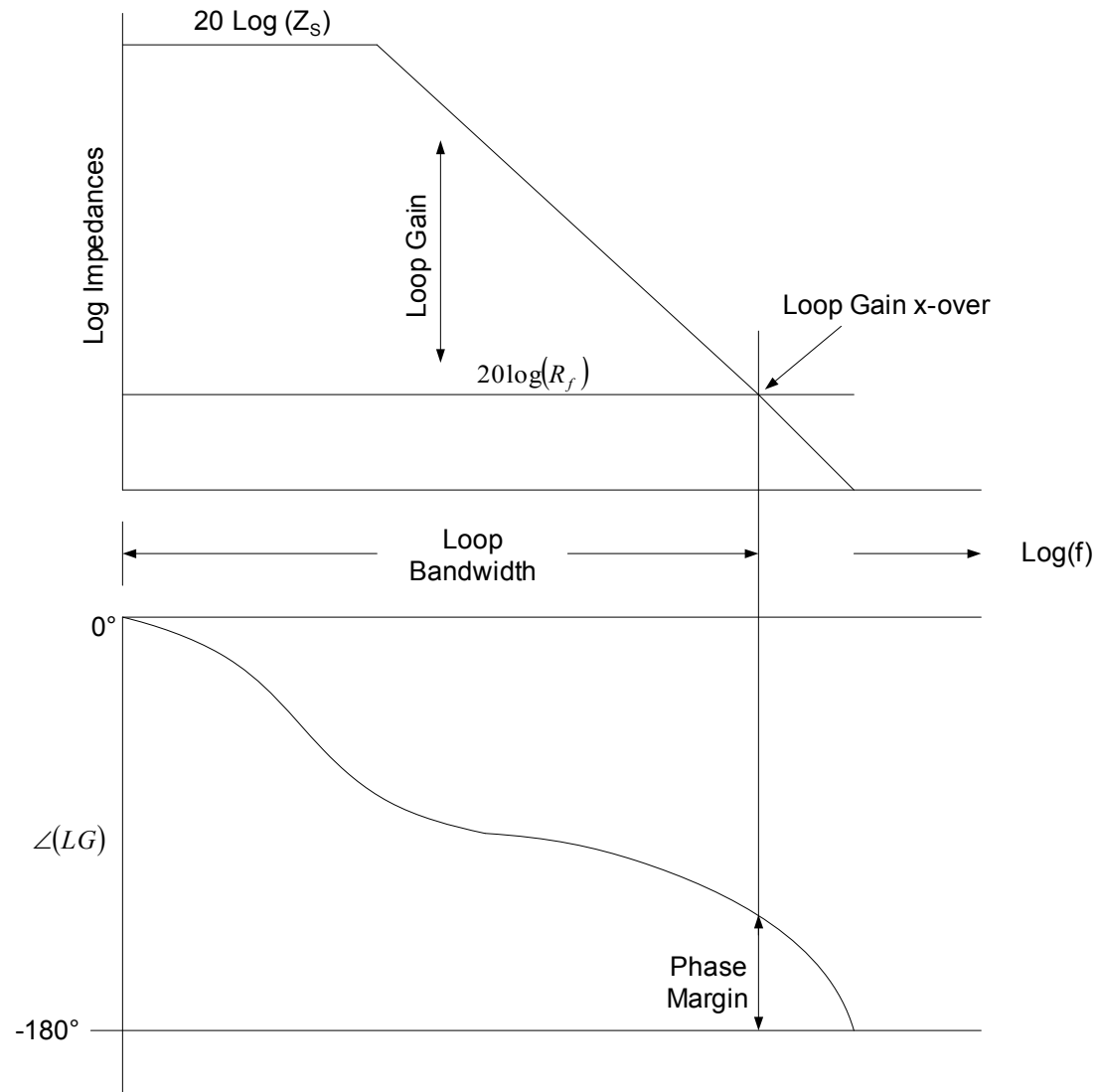
# Gain Bandwidth Independence

- CFB are gain bandwidth independent
- The reason for that can be found in the gain equation, here the inverting configuration.
  - The -3dB bandwidth is reached when Z(s) = $R_F$
  - This is independent of the value of $R_G$. If the gain increased is achieved by lowering $R_G$ instead of increasing $R_F$ then the bandwidth is independent of the gain.

$$\frac{V_{OUT}}{V_{IN}} = \frac{-\dfrac{R_F}{R_G}}{1 + \dfrac{R_F}{Z(s)}}$$

# Advanced Model

- For CFB, the simplified model is only a tool to help understand the architecture and has little use beyond that.

- The advanced model adds:
  - Inverting input resistor

- This adds a term in the denominator that is very useful.



$$\frac{V_o}{V^+} = \frac{1 + \dfrac{R_f}{R_g}}{1 + \dfrac{R_f + R_i\left(1 + \dfrac{R_f}{R_g}\right)}{Z(s)}}$$

Where $1 + \dfrac{R_f}{R_g}$ desired non $-$ inverting gain
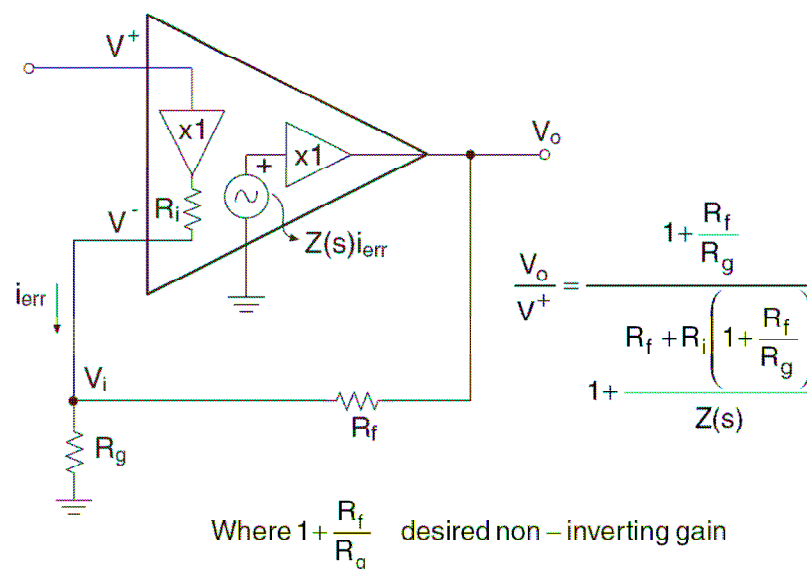
# Gain Bandwidth Independence

- Using the advanced model, we find out that this is not completely true as there is a 2nd order dependency of the frequency response to the noise gain.

- $R_{IN-}$ can vary from a few ohms to $100\Omega$ in CFB.

- One of the device with the lowest inverting input resistance is the OPA683 that operates on <1mA and achieve an equivalent GBWP of 3.5GHz. Note that the OPA684 on 1.7mA has a 7GHz Equivalent GBWP.

- A comparable GBWP VFB is the OPA847 with 3.6GHz but operates on 18mA.

$$\frac{V_{OUT}}{V_{IN}} = \frac{-\dfrac{R_F}{R_G}}{1 + \dfrac{R_F + R_{IN-} \cdot NG}{Z(s)}}$$

*with*

$$NG = 1 + \frac{R_F}{R_G}$$

OPA683 BANDWIDTH vs GAIN



Normalized Gain (dB) vs Frequency (MHz), $R_F = 1.2k\Omega$, curves for G = 2, G = 5, G = 10, G = 20, G = 50, G = 100
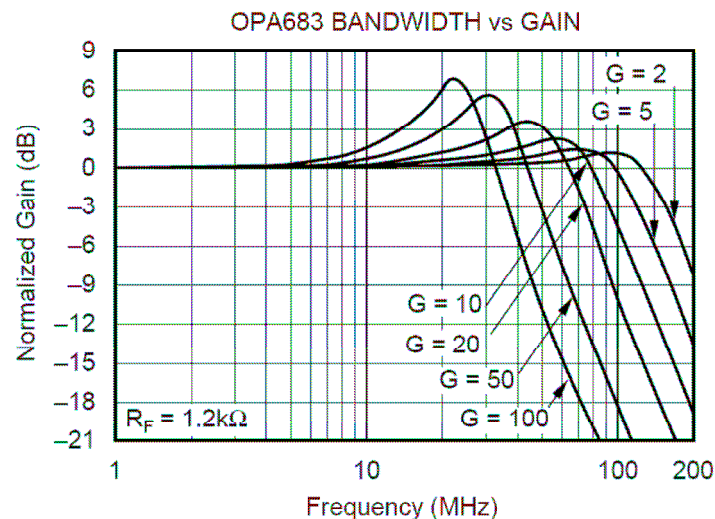
# Summary so far

# *Comparing Voltage and Current Feedback Op Amps*

- Classical Advantages of Voltage Feedback Op Amps

    – Typically can deliver better DC accuracy

        • This is most applicable to pulse oriented signal requirements - typically, DC precision is less important in AC coupled (communications) channels

    – Can be the lowest overall equivalent input noise

        • Best noise (< 1.2nV/$\sqrt{Hz}$) comes at the price of high quiescent current and non-unity gain stability.

    – Typically internally compensated. Note that some external compensation VFB exist.

    – The highest accuracy, lower noise devices also have a typical architecture, limiting the maximum achievable slew rate.

    – Low noise Transimpedance application are ideal target application

# *Comparing Voltage and Current Feedback Op Amps*

- Classical Advantages of Current Feedback Op Amps

    – Essentially unlimited slew rate - gives very high full power bandwidth

        • Most data sheet slew rate numbers are either limited by the input stage buffer or are actually reporting bandwidth limited rise time by mistake

    – Nearly  gain bandwidth independent

        • Most useful aspect of this is intrinsic low gain stability with very high closed loop BW

    – Most CFB also provide a large output current drive capability.

    – Application such as adder and high gain application are ideal target application

# Table Comparison: DC & AC specification

| | VFB | CFB |
|---|---|---|
| Input Offset Voltage (max at 25°C) | ±0.6mV to ±4mV | ±2.5mV to ±7mV |
| Non Inverting Input Bias Current (max at 25°C) | ±5pA to ±10uA | +25uA to +40uA |
| Input Offset Current (max at 25°C) | ±5pA to ±1uA | |
| Inverting Input Bias Current (max at 25°C) | | ±25uA to ±50uA |

| | Traditional VFB | Modern VFB | CFB |
|---|---|---|---|
| Gain Bandwidth Product | 3600MHz (18mA) | 500MHz (5mA) | N/A |
| Slew Rate | 800V/us (18mA) | 1800V/us (5mA) | 3500V/us (12mA) |

# Open-Loop Gain and Stability

# *Loop Gain Review*



OPEN-LOOP GAIN AND PHASE

$$20 \cdot \log_{10}\left(1 + \frac{R_F}{R_G}\right)$$

**OPA690**

OPEN-LOOP TRANSIMPEDANCE GAIN/PHASE

$$20 \cdot \log_{10}\left(R_F + R_{IN-} \cdot \left(1 + \frac{R_F}{R_G}\right)\right)$$
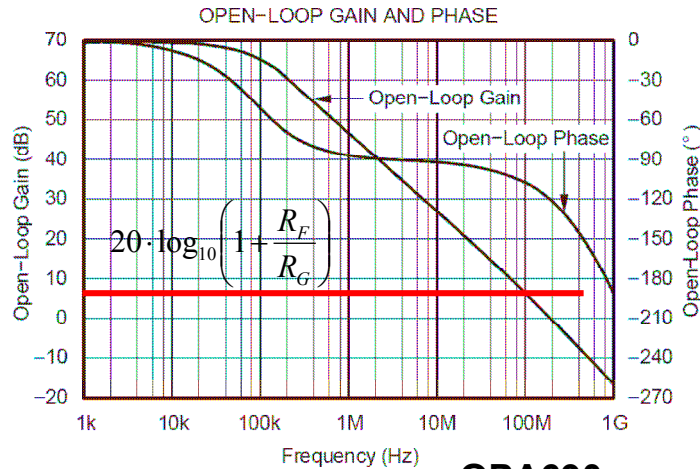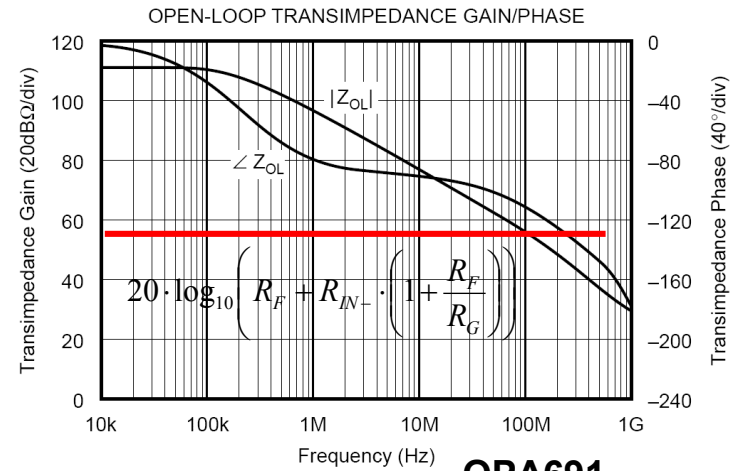
**OPA691**

- For Voltage Feedback op amps, the loop gain varies directly with the signal gain for simple external circuits. Changing the gain, changes the frequency response directly.

- For Current Feedback op amps, the loop gain is set by the feedback impedance allowing an independent setting for the signal gain. The feedback resistor becomes the frequency response compensation.

- Both amplifiers are showing the feedback path compensation for a gain of 2V/V.  This corresponds to a noise gain of 6dB for the OPA690 and to a compensation element of 466Ω for the OPA691.  Notice that you have ~70° phase margin for the VFB and ~65° phase margin for the CFB.

# Loop Gain is Everything in Op Amps

- Op Amp suppliers are essentially selling a device that does impedance transformation (high input Z to low output Z) and a whole lot of open loop gain.

- The customer then closes the loop to get a more controlled voltage gain, but also gets a huge improvement in precision (both DC and AC) due to the high open loop gain.

- For high frequency parts, the DC open loop gain is a secondary issue and it is really the one pole roll-off curve that is of interest and where the magnitude of the open loop gain equals the inverse of the feedback ratio. (Loop Gain x-over).

- While the closed loop response is what is normally observed and reported, hiding inside this is a loop gain over frequency that is critical for distortion and stability analysis.

# Comparing Voltage and Current Feedback Op Amps

- Two parts on the same process, at the same quiescent power, will have pretty similar open loop gain curves for VFB and CFB devices – Compare the OPA690 (VFB) and the OPA691(CFB) below.
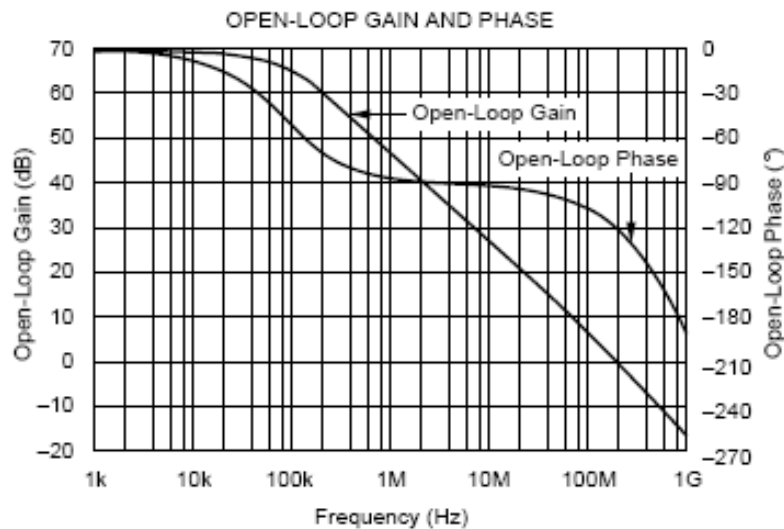


**OPA690 Voltage Feedback (VFB)**

**Dominant Pole at 80kHz**

Gain of 2 (6dB) Loop Gain at 20Mhz is 14dB



**OPA691 Current Feedback (CFB)**

**Dominant Pole at 200kHz**

Gain of 2, Rf = 402ohms, Loop gain at 20Mhz is 16dB

**The loop gain profile is just slightly higher over frequency for the CFB version due to the higher dominant pole location**

TEXAS INSTRUMENTS

# *Minimum Stable Gain for VFB Op Amps*

- **Classical Tradeoff's in selecting Voltage Feedback (VFB) Op Amps**

  - **Minimum Stable gain (primary IC circuit design variable) influences several key parameters.**

    - **Useable gain range**

    - **As minimum stable gain increases, input noise goes down and slew rate goes up.**

    - **Notice the 2nd pole at 800MHz before the unity gain cross-over on the OPA847.**

  - **Wideband, low gain, operation has been very difficult for VFB amplifiers. Newer parts, like the OPA690, use a high transconductance input stage that gives very high slew rate in a unity gain stable device – at the cost of higher input noise voltage**



**OPA847**



**OPA690**

**TEXAS INSTRUMENTS**
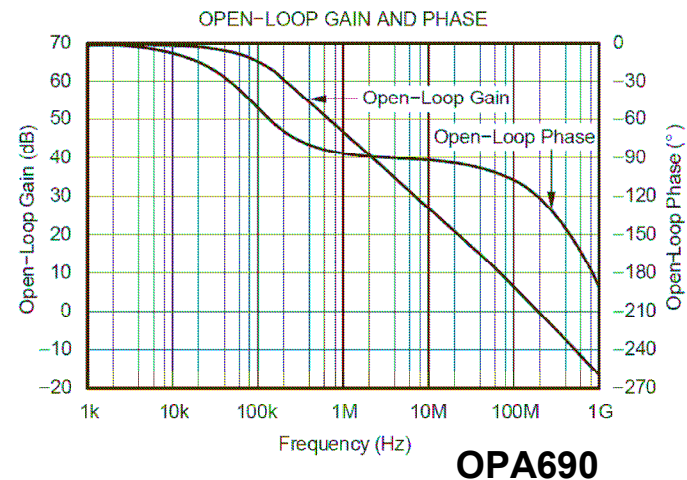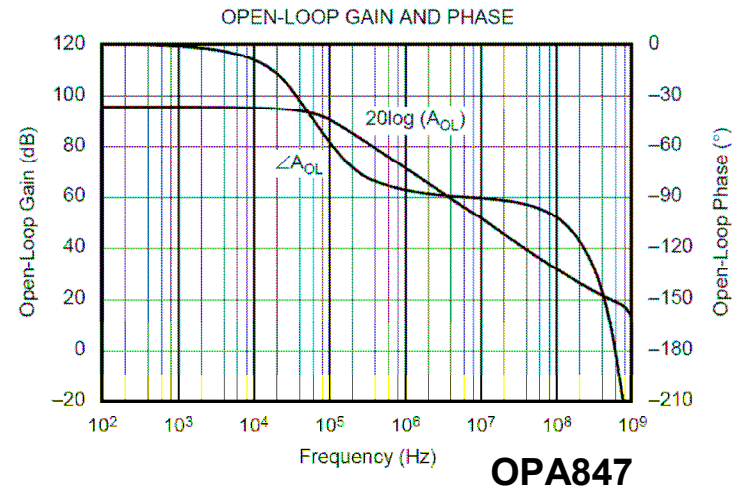
# *Selecting Current Feedback Op Amps*

- Classical Tradeoff's in selecting  Current Feedback (CFB) Op Amps

  – Although input voltage noise can be low, inverting input current noise is always much higher than VFB equivalents

    • This limits the usability of the CFB for receivers.  Most time a VFB will offer better dynamic range.

  – Feedback element is constrained in its impedance range since it is the compensation element

    • This limits the usability of the CFB for transimpedance.  A CFB can be used in transimpedance applications but mostly to low gain as the feedback resistor is the compensation element.  Too little and you have oscillation, to much and the bandwidth becomes quickly limited.

  – Input bias currents are large and unmatched - limits achievable DC accuracy

    • Time domain application and pulse oriented application generally uses VFB.

# Typical Signs of Instability

- These signs of instability are valid for both VFB and CFB

  - Time Domain, or Pulse Response

    - Overshoot and/or sustained ringing.

  - Frequency Domain

    - Higher apparent noise than you would expect

    - Sharp spike in the frequency response

  - DC

    - Elevated case temperature

    - Higher Output Offset Voltage

    - Higher supply current than expected.

# "Before you can fix it, you have to find it" issue.

- Once a sustained oscillation is present, it shows up everywhere – so probing on a board is almost useless – actually worse than useless due to a form of the
  – "Heisenberg uncertainty principal"
  – One form of this stated that improving the accuracy in time of a measurement, you would have to lose accuracy in position – or in other words, the act of measuring something will change it.
- In trying to measure an oscillation, it is extremely likely that the measurement means will change the effect you are trying to see. For instance, a part that is almost, but not quite, oscillating may well go into oscillation if you probe with a 10X (10pF) probe. This is a common issue where
  – Test points are brought out from inside the high speed path for probing
  – In circuit board testing can put strange parasitic inside the loop of the DUT.

# "Before you can fix it, you have to find it" issue.

- At higher frequencies, it is much more useful to scan the board with a home-made loop antennae into a spectrum analyzer than to probe directly on the board. At lower frequencies (<1Mhz) probing might be ok.

- Even worse, a 100ohm measurement path might stabilize a loop that once removed will break back into oscillations.

- So – it is extremely critical to hunt for these issues in as non-invasive a way as possible to retain the loop issues inherent to the board (and not part of the measurement circuit)

TEXAS INSTRUMENTS

# Isolating the oscillation using a loop antennae

- Assuming you are pretty sure you have a sustained oscillation on a board – how to isolate where it is coming from?
  - First you have to get a reliable detection of the oscillation that you know is not changing the circuit – a loop antennae is how we do that.
- Essentially you scan the board with the loop antennae stepping through smaller frequency bands on the spectrum analyzer to get a low noise floor. (I usually go in 100MHz center frequency steps with a 100Mhz span)
- If you see something that looks promising, turn the board power off and see if it goes away – if it doesn't, you are picking up something else in the ambient (FM stations show up pretty well with this technique)

# Isolating the oscillation using a loop antennae

- Here we show a loop antennae (20 turns of magnetic wire with about ¾" diameter) sensing the LO output of the HP8568B spectrum analyzer at 20Mhz.

- This is just an example but essentially 10-20 turns of magnetic wire connected into a female BNC connector is what we use.



TEXAS INSTRUMENTS
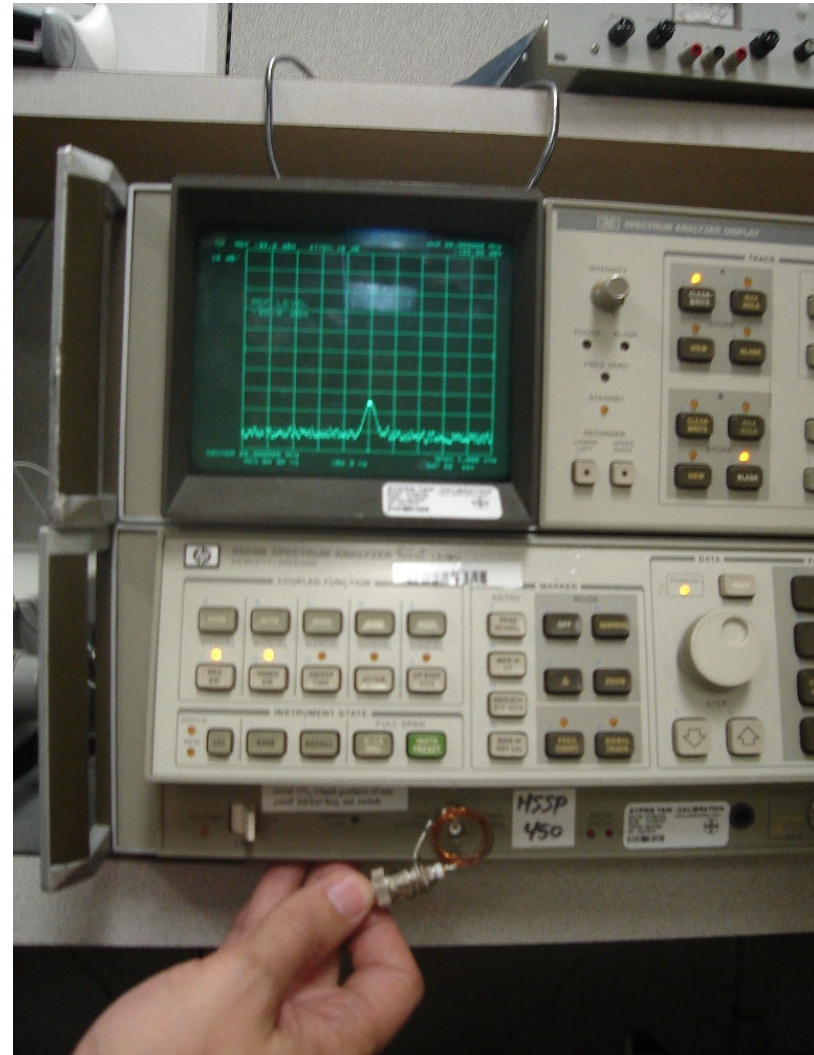
# Isolating the oscillation using a loop antennae

- Here – you are just looking for narrowband power at some frequency – remember, if the oscillation is a Rail-Rail square wave phenomena (common for supply related issues) then you will see a square wave spectrum (fundamental then a bunch of odd harmonics from that). It is only the fundamental that is of interest and those other harmonics are not really added oscillation frequencies just harmonics of a non-sinusoidal oscillation. The amplitude of the power is not too important – just that it exists and can be attributed to the board.

- Once you are sure that a solid spur is being picked up that is not part of the normal board operation or ambient pickup, how do you isolate it?

- A first order effort can be made moving the loop antennae around on the board for maximum amplitude.

- Then, we need to isolate the feedback loop associated with the instability.

# Isolating the oscillation using a loop antennae

- Commonly, designers will touch or probe with a small cap. lead in the circuit to look for changes – key point –
  - almost anywhere you touch will change the amplitude – you know you are in the loop when it changes the frequency (and it will also probably change the amplitude – but ignore that).
- For lower frequency oscillations (<1Mhz), you will have to probe somewhere in the circuit since the loop antennae won't pick this up.
- Since the oscillation will show up everywhere once its established, try to probe somewhere that the probe capacitance will not change the circuit – then touch suspect points and look for a frequency change in the oscillation.

# Common types of oscillations

- We see 3 general types of oscillations most commonly in high speed circuits.
  - power supply related issues
  - self oscillation in the I/O transistors
  - and, most commonly, overall feedback loop oscillations (sometimes called a loop gain oscillation)
- This is certainly not a comprehensive list and there can certainly be other types of phenomena, but these seem to be the most common.
- Any instability discussion needs to start with "what is the frequency of oscillation?" – and be careful to ignore harmonics if it is a non-sinusoidal oscillation – just the fundamental frequency.

# Fundamental requirements to have an oscillation

- Haven't talked yet about loop phase margin explicitly, but that is the question hiding behind every consideration of circuit instability.
- Essentially, you must have a source of gain in the circuit and a power supply (passive elements by themselves can't oscillate) – and, some feedback loop inside the system must show >180° phase shift around the loop at some frequency while the system still has a loop gain >1 at that frequency.
- IC amplifiers have many internal feedback loops. Part of the designers job is to verify those are all stable internally over process and temperature. Most difficulties occur in getting signals into and out of the devices along with the overall feedback loop for a negative feedback device.

# Clues from the Frequency of Oscillation

- If the frequency of oscillation is somewhere around what we would expect the overall loop gain x-over to be, then we probably have a loop gain oscillation. For instance, using a 200Mhz unity gain stable voltage feedback amplifier with a cap across the feedback resistor, and seeing an oscillation at 200Mhz would suggest an overall loop gain oscillation issue.

- If the frequency of oscillation is << loop gain x-over, this is most often a power supply issue or an interaction (often through the power supply) of different stages in the design.

- If the frequency of oscillation is >> loop gain x-over, then this is usually a self oscillation in an I/O transistor stage.

# Stability in Current Feedback

- Never place a capacitance in parallel with the feedback resistance.
  - The compensation resistor is placed in the feedback path.
- Limit the capacitance on the inverting node of the amplifier.
  - The inverting node is a buffer.
- Limit of isolate the capacitance on the output pin.



CLCXXX
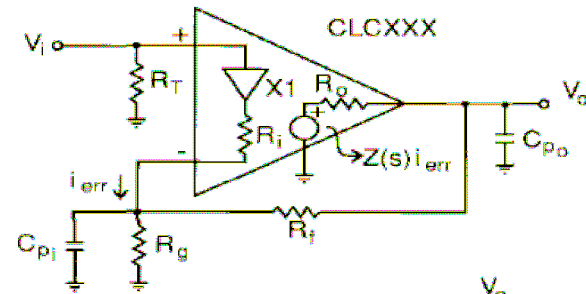
$$\text{ideal } A_v = \frac{V_o}{V_i} = 1 + \frac{R_f}{R_g}$$

Most critical parasitic capacitances
$C_{po} \rightarrow$ output capacitance
$C_{pi} \rightarrow$ inverting input capacitance



RECOMMENDED $R_S$ vs CAPACITIVE LOAD
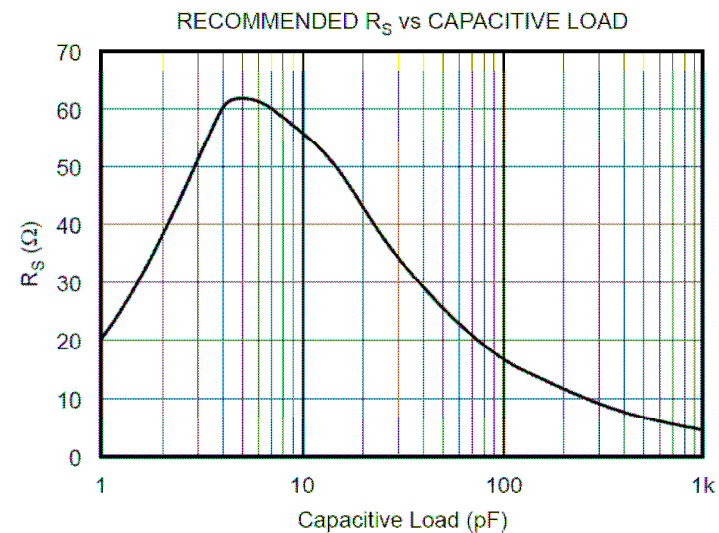
# Noise

# Input and Output Noise Calculations

- Noise can be a very confusing issue. Some points to keep in mind.

    - The only noise that can be measured is at the output of the amplifier.

    - Input referred noise is simply the output noise divided by the gain back to the input that you care about - could be the non-inverting input, inverting input, or the input of a prior stage.

    - Output noise power is made up of the sum of numerous noise contributors. Often, one or two of these are clearly dominant and swamp out all others. This leads to simplified noise equations that drop out terms - leading to much confusion. General equations should include a fairly complete model even if some terms are often (but not necessarily always) negligible.
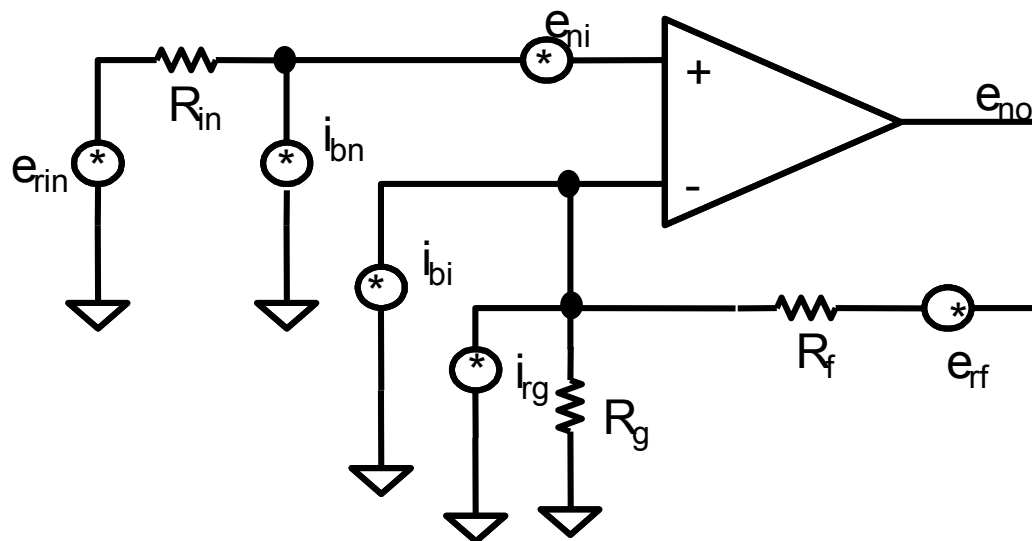
# Noise Analysis for any Op Amp

$$e_{no}^2 = \left[(e_{rin}NG)^2 + (i_{bn}R_{in}NG)^2 + \left((e_{ni}NG)^2 + (i_{bi}R_f)^2 + (i_{rg}R_f)^2 + e_{rf}^2\right]\right.$$

$$e_{no}^2 = \left[e_{ni}^2 + (i_{bn}R_{in})^2 + 4kTR_{in}\right]NG^2 + \left[(i_{bi}R_f)^2 + (4kTR_f)NG\right]$$

4kT=16.4E-21 J        at        T = 298 $^O$ C

$$NG = 1 + \frac{R_f}{R_g}$$

See TI App. Note SBOA066
"Noise Analysis for High Speed
Op Amps" for more detail.
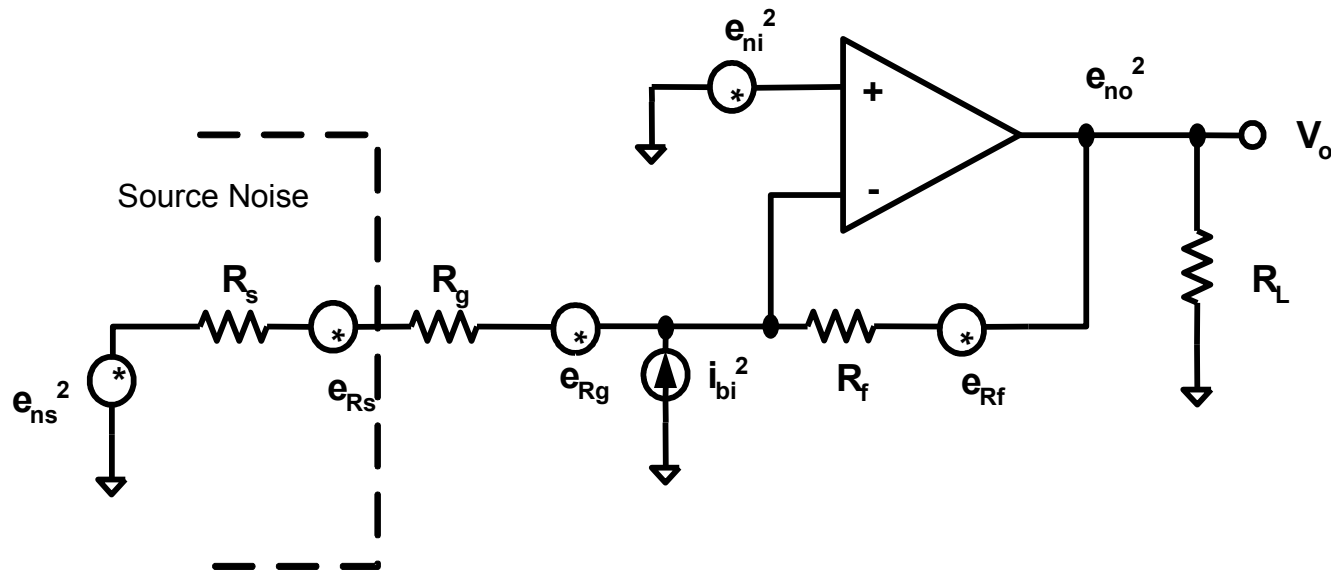
# Non-Inverting Input Referred Total Noise

- Dividing the total output noise by the non-inverting gain will mathematically develop an input noise that, if this term were placed at the input of a noiseless amplifier of the same gain - you would get the same total output noise.

$$e_{no} = \sqrt{e_{ni}^2 + (i_{bn}R_s)^2 + 4kTR_s + \left(\frac{i_{bi}R_f}{NG}\right)^2 + \frac{4kTR_f}{NG}}$$

Where NG = 1 + Rf/Rg = Noise Gain

This shows that, as gain increases, the non-inverting input referred noise approaches just those terms present at the non-inverting input. Conversely, at low gains, the apparent input noise can be dominated by the terms at inverting input - this is commonly the case for current feedback op amps.

# Inverting Amplifier Noise Model with Input Matching



$$e_{no} = \sqrt{e_{ni}^2 \cdot NG^2 + 4kT\left(\left(R_s + R_g\right)\|R_f\right) + \left(i_{bi}R_f\right)^2 + e_{ns}^2\left(NG - 1\right)^2}$$
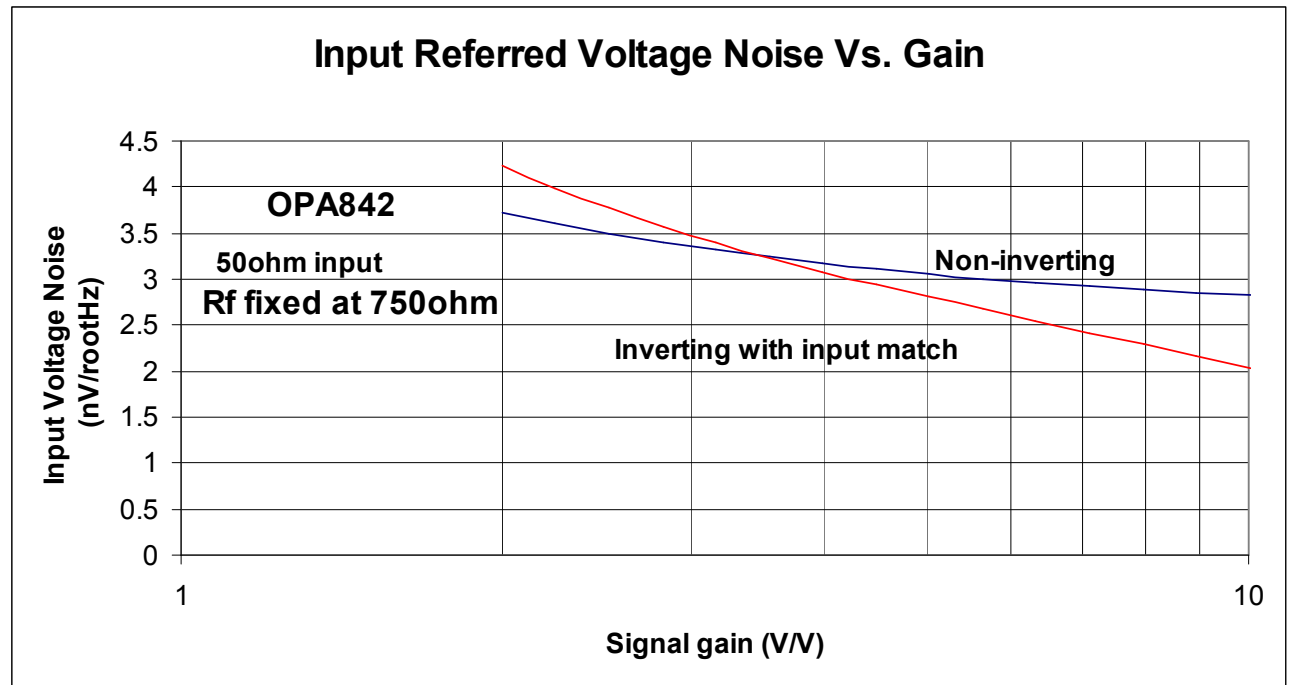
$$NG = 1 + \frac{R_f}{R_s + R_g}$$

$$4kT = 16.0 \times 1E - 21J \quad \text{For T} = 290\text{K}$$

# Inverting Input Referred Total Noise

- Dividing the total output noise by the inverting gain will mathematically develop an input noise that, if this term were placed at the input of a noiseless amplifier of the same gain - you would get the same total output noise.

- This is particularly useful for low input voltage noise parts when Rg = Rs. Total input referred noise in this case can be very low. (The VFB OPA842 and OPA847 and CFB OPA695 are good examples).

# Input Referred Voltage Feedback Noise vs. Gain Setting

•The Inverting mode becomes lower noise when the equivalent gain from the $e_n$ term at the non-inverting input to the inverting input signal point becomes <1. This applies when a finite source impedance is matched to the input impedance.

**Input Referred Voltage Noise Vs. Gain**

OPA842

50ohm input

Rf fixed at 750ohm

Non-inverting

Inverting with input match

Input Voltage Noise (nV/rootHz)

4.5
4
3.5
3
2.5
2
1.5
1
0.5
0

Signal gain (V/V)

1                                    10

# Input Referred Current Feedback Noise vs. Gain Setting

• CFB Amps at high gains can provide very low input referred noise

• Low Noise CFB amps in the Inverting mode will have even lower noise than non-inverting configuration

•Big increase at the lower gains is the effect of the higher inverting bias current noise times the feedback R.

**Input Referred Noise vs Gain**

**Non-Inverting & Inverting CFB OPA695**



Non-Inverting Gain

Inverting Gain

Noise (nV/rt Hz)

Gain V/V
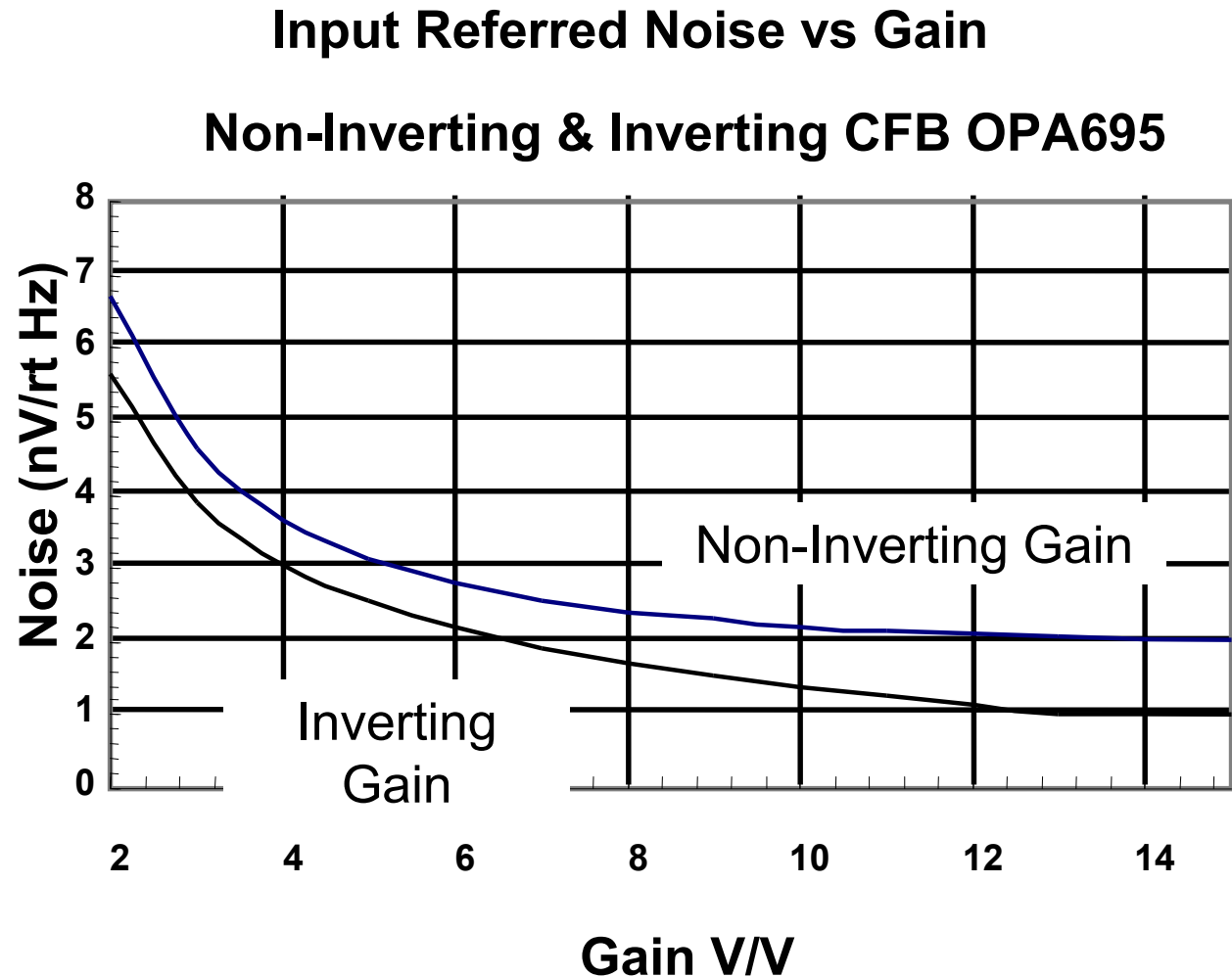
The Feedback R is adjusted here to get maximum flat bandwidth at each gain.

Texas Instruments

# Distortion

# *Understanding Noise & Distortion Tradeoff's Between Op Amp Types and Application Topologies*

- Loop Gain and other contributors to linearity

- Differential circuits and why

- Distortion dependence on external conditions

  - Voltage Feedback, Current Feedback

  - Noise models

  - Differences between amplifier types

- Example solutions and conclusions

# *Distortion Issues*

- At lower frequencies, the lowest distortion will be given by voltage feedback amplifiers

  - We believe this is due to a linearity floor set by the error sensing point in the CFB topology. The CFB inverting input linearity sets a floor to distortion much higher than the best VFB designs. This is normally a 2nd harmonic term.

  - CFB will, however, give relatively constant distortion vs. Gain setting and hold better numbers to higher frequencies due to considerably more slew rate margin.

# *Theoretical Determinants of Harmonic Distortion*

- An Ideal amplifier would take an input spectrum and pass it on to the output with the same gain for each Fourier component and no added power in the spectrum.

    - We have not quite achieved that ideal, hence new amplifiers and techniques moving closer to this are still being introduced.

- Output spectral purity has many levels of consideration – the better you aspire to, the more of these levels you will have to consider.

- The first level is that, for a high open loop gain type of part, the closed loop output linearity will be the open loop linearity intrinsic to the output stage corrected by the loop gain at the fundamental frequency.

    - Low loop gain devices, like most RF amplifiers, achieve high linearity by making the signal power a very small part of the quiescent power. Hence you will see >80dBc SFDR type devices to very high frequencies using > 1.5W quiescent power

# Distortion Analysis using Negative Feedback with Distortion modeled only as an Output Stage Distortion
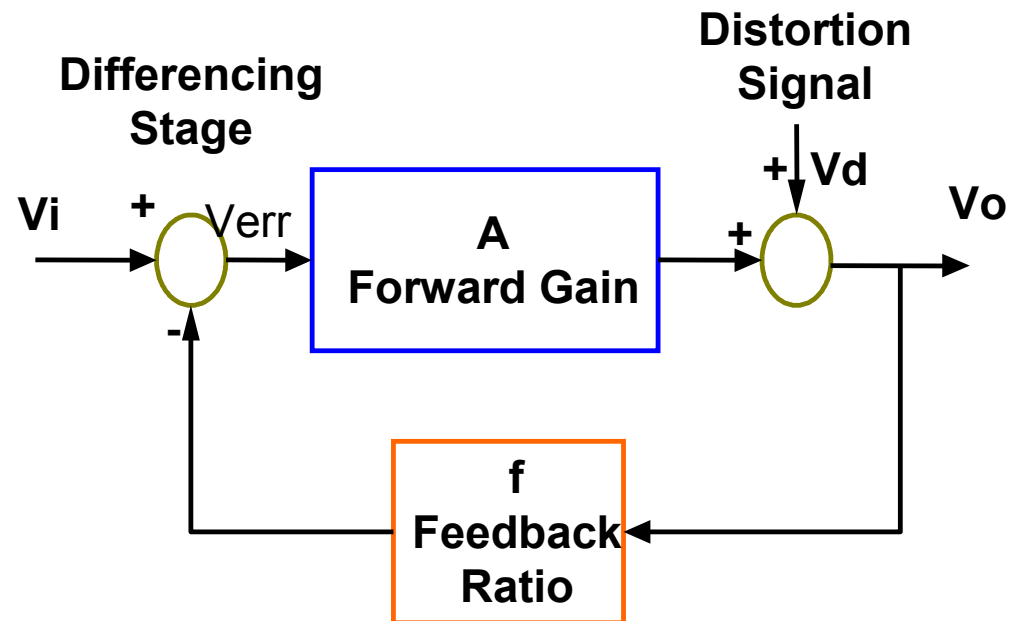
$$V_o = A \cdot V_{err} + V_d$$

$$V_{err} = V_i - f \cdot V_o$$

$$V_o = A \cdot V_i - A \cdot f \cdot V_o + V_d$$

$$(1 + A \cdot f) V_o = A \cdot V_i + V_d$$

$$V_o = A \frac{V_i}{(1 + A \cdot f)} + \frac{V_d}{(1 + A \cdot f)}$$

**Differencing Stage**

**Distortion Signal**

**Vi**

**+** Verr

**-**

**A Forward Gain**

**+** Vd

**+**

**Vo**

**f Feedback Ratio**

where Af ≡ Loop Gain.   Output stage non-linearities are corrected by loop gain.

# *Paths to Improved Distortion Suggested by the Control Theory Model.*

- At a first level, output linearity is the open loop distortion of the output stage, corrected by the loop gain. So, improving either of these will improve distortion.

- One key conclusion from the Loop Gain comparison between VFB and CFB is that the CFB holds a more constant loop gain over signal gain (Gain Bandwidth Independence). This should hold more constant distortion to higher gains than VFB.

- Comparing those plots for the VFB OPA690 and CFB OPA691 -

**OPA690, VFB, HD linear with log gain**

**OPA691,CFB, HD more constant over gain**
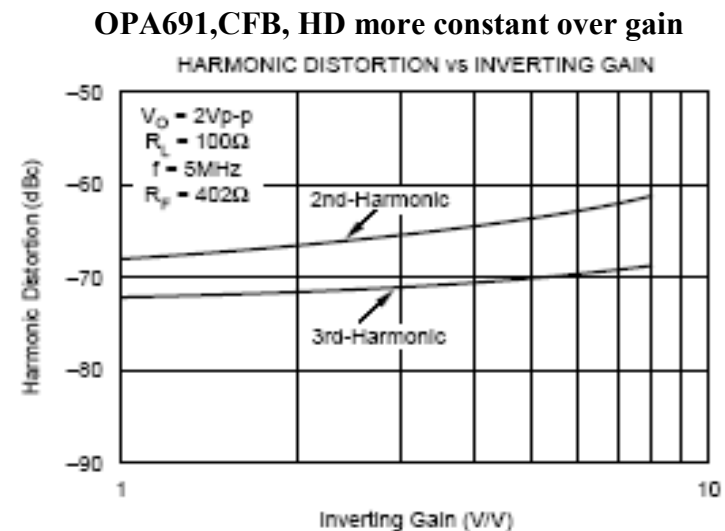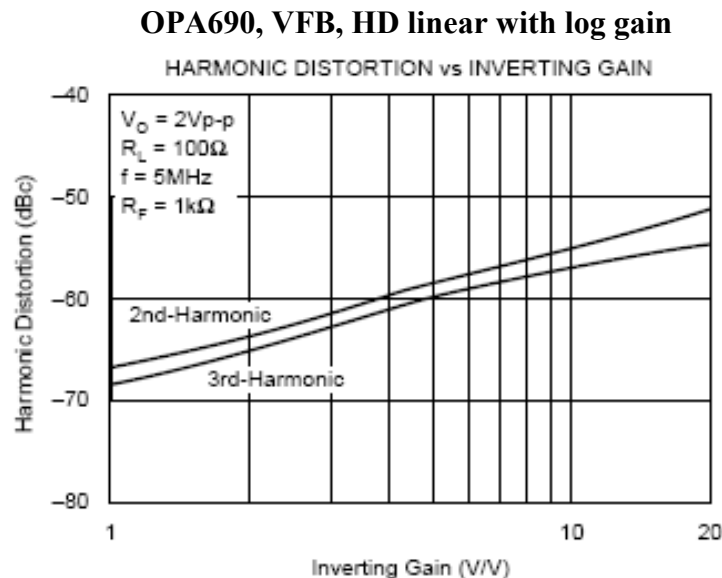
# *Continued Improvement in SFDR??*

- The 2$^{nd}$ Harmonic typically does not follow this theory exactly. There are other, external, effects that come into play on the even order terms for a single ended amplifier.

- Even order distortion can be visualized as ½ cycle imbalance on a sine wave. Odd order distortion can be visualized as curvature through zero on a sine wave or a very balanced deviation on each ½ cycle.

- Anything that will take a purely balanced output sine wave and introduce perturbation on one ½ cycle but not the other, will be generating even order distortion terms.

- Suspects include –

  - Mutual coupling in the negative supply pin to the non-inverting input.

  - Slightly imbalanced ground return currents getting into the input signal paths.

  - Imbalanced supply decoupling impedance.

- One of the best ways to eliminate this issue is to run the signal path differentially – but exactly why does that work??

# Why is it that a Differential Configuration Suppresses the 2nd harmonic??

Differential even order harmonic cancellation
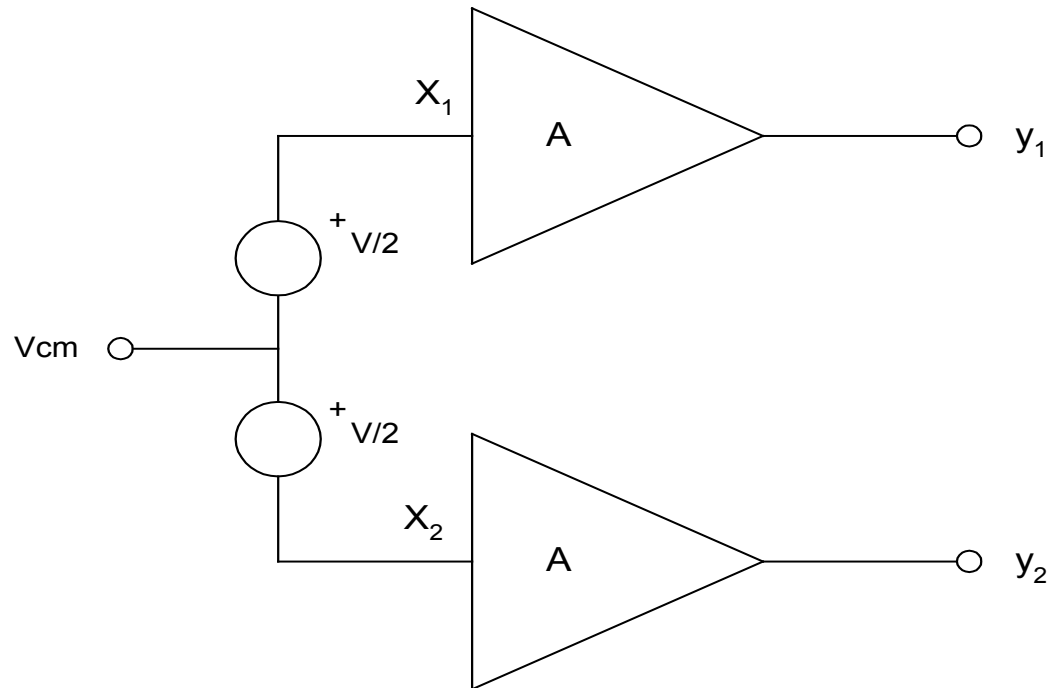


Let both gain elements A have the same polynomial approximation to a transfer function

# Why is it that Differential configurations suppress the 2nd harmonic??

$$y = A_o + A_1 X + A_2 X^2 + A_3 X^3$$

$$X_1 = V/2$$

&

$$X_2 = -V/2$$

$$y_1 = A_o + A_1\left(V/2\right) + A_2\left(V/2\right)^2 + A_3\left(V/2\right)^3$$

$$y_2 = A_o - A_1\left(V/2\right) + A_2\left(V/2\right)^2 - A_3\left(V/2\right)^3$$

*then*

$$(y_1 - y_2) = 0 + A_1 V + 2 A_3\left(V/2\right)^3$$

$$= A_1 V + \frac{A_3}{4} V^3$$

- Substituting in the two halves of differential input signal, getting to each output signal, then taking the difference - shows we are theoretically only left with the desired linear signal and the 3rd order term.  Even if the A2 coefficient is not exactly matched between the two amplifiers, it is their difference that ends up being the gain for this 2nd order non-linearity at the output. We also see a reduction in the 3rd order coefficient - arising from only applying 1/2 of the input through each channel.

**TEXAS INSTRUMENTS**

# Single Ended Even order Terms become Odds in the Differential Configuration

- In the time domain, this effect can be seen by producing a clipped waveform for the two outputs, then taking the difference. The individual outputs would have a very high even order harmonic content, while the differential signal will still be distorted, but will give rise to only odd harmonics since the clipping is now symmetric on each 1/2 cycle of the sinusoid.



**Single ended to Differential Distortion**

Differential Output

Output Voltages vs Time (sec)
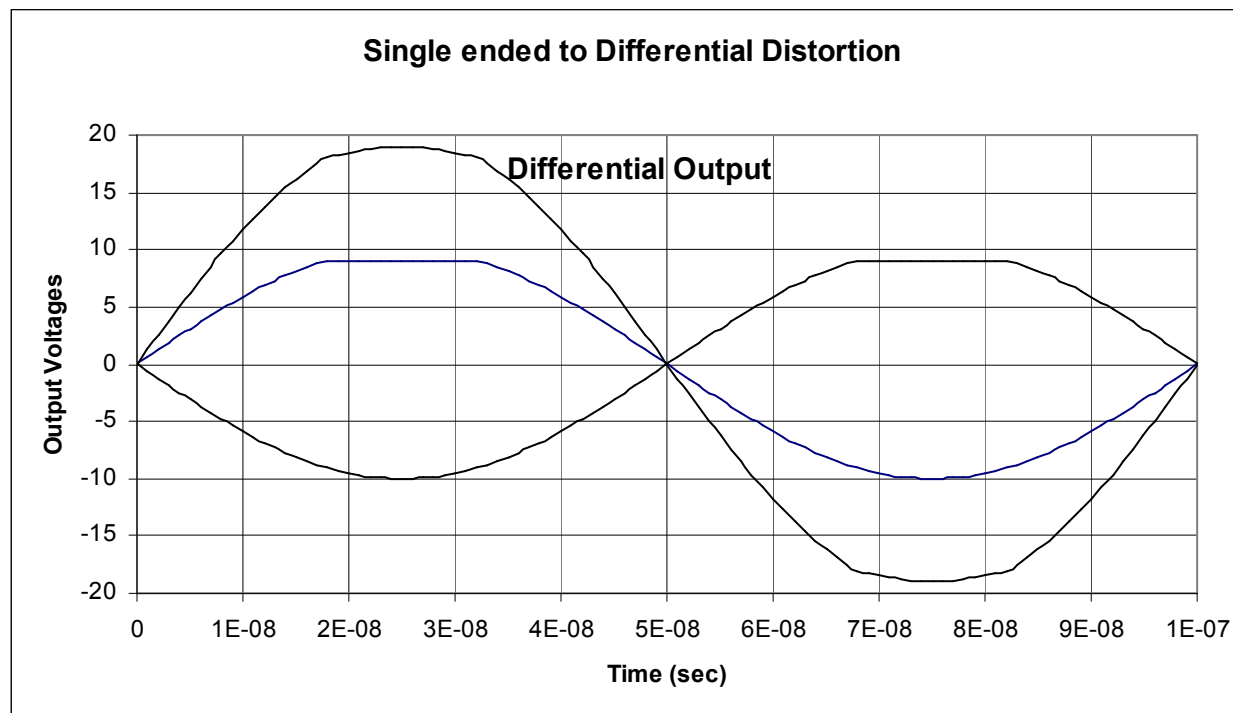
# Single Ended vs. Differential SFDR

- To illustrate the power of differential designs in suppressing HD2, the plots on the next slide show the HD2 and HD3 for a low noise, low distortion VFB dual amplifier in both single ended and differential configurations. The test conditions give the same loop gain, but the differential test had a 35ohm load to each output while the single ended was a 100ohm – which raised the HD3 quite a bit.

- The single ended performance is HD2 dominated, while running the same part differentially, pushes the 2nd down to be on the order of the HD3 number.

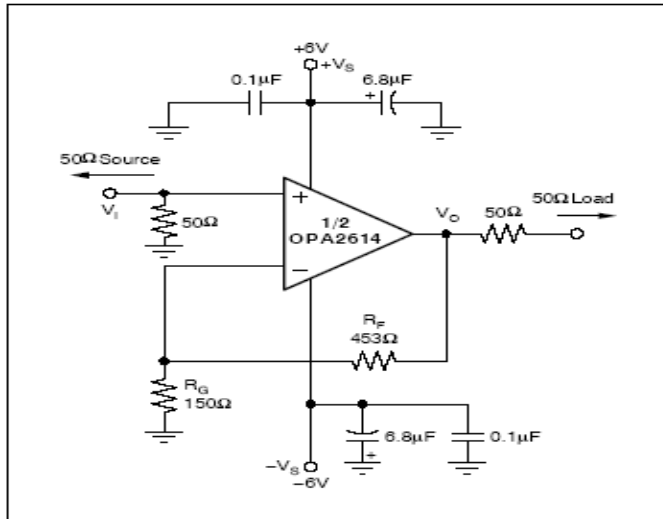# Single Ended vs. Differential SFDR



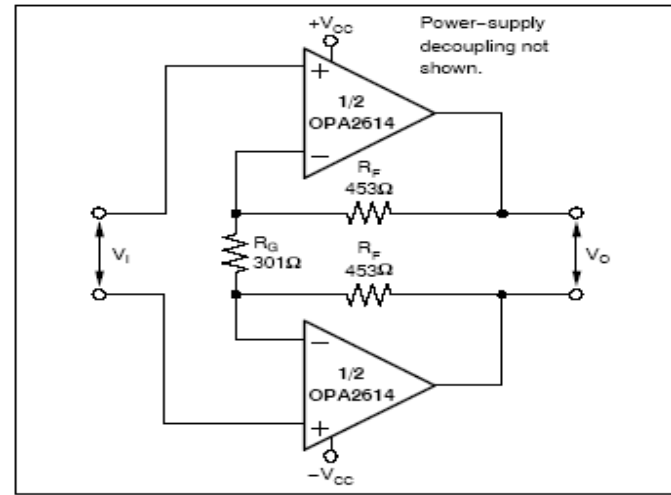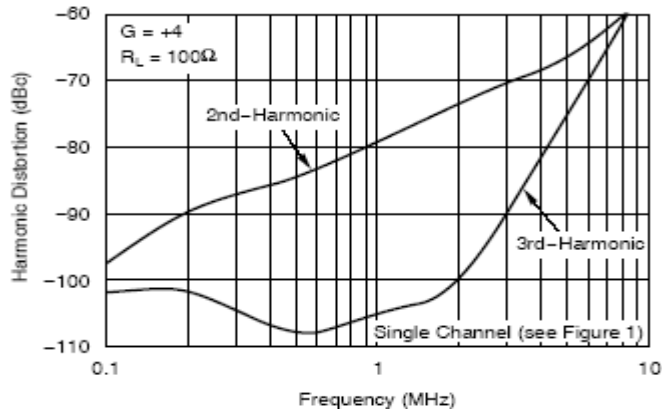Figure 1. DC-Coupled, G = +4, Bipolar Supply, Specification and Test Circuit



Figure 5. Noninverting Differential I/O Amplifier



*TEXAS INSTRUMENTS*
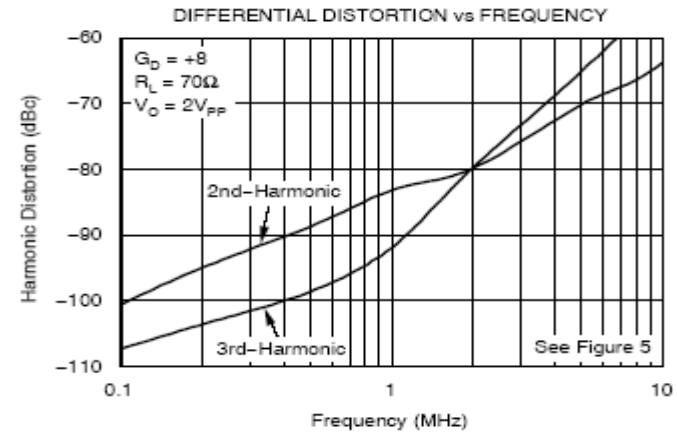
# *Key Elements to Understanding and Improving Distortion*

- External conditions that will influence distortion

    - Required Output Voltage and Current as a portion of the quiescent power and design of the output stage

        - This is including loading and supply voltage effects as well.

        - Adding a higher standing current in the output stage will often lower distortion with no effect on noise. This Class A current can pick up about 10dB on the 3$^{rd}$.

    - Loop gain – use a VFB designed for the desired gain setting or, at higher gains use a CFB device.

    - Frequency – since loop gain changes with frequency, a fixed output stage non-linearity will give a changing distortion over frequency.

    - Layout and Supply Decoupling

        - This is covered in detail in TI – app. Note SBAA113

- To improve distortion, we can perhaps shape the loop gain over frequency to get enhanced low frequency distortion while holding a stable response.

# *Summary Suggestions to get the Lowest Noise and Distortion*

- Differential signal paths allow much lower distortion versus quiescent power than single ended signal paths.

    - Making the last stage of gain before the converter a differential path will be moving in the direction of better SFDR. If the signal path is intrinsically single ended, make the conversion to differential at a lower power (voltage) level then use the last stage interface to get the remaining gain in a differential structure.

- At higher gains, where an input match is desired, the inverting topology can offer lower input referred voltage noise – this arises due the effective attenuation of the amplifiers voltage noise term when referred to the inverting input.

- At lower frequencies, VFB will probably get to the lowest distortion vs. Iq. At higher frequencies, the CFB  has been used, but emerging high slew rate FDA's are also doing very well in this application.

# Applications:

– Receive function

# New Compensation Technique for Non-Unity Stable Voltage Feedback Op Amps
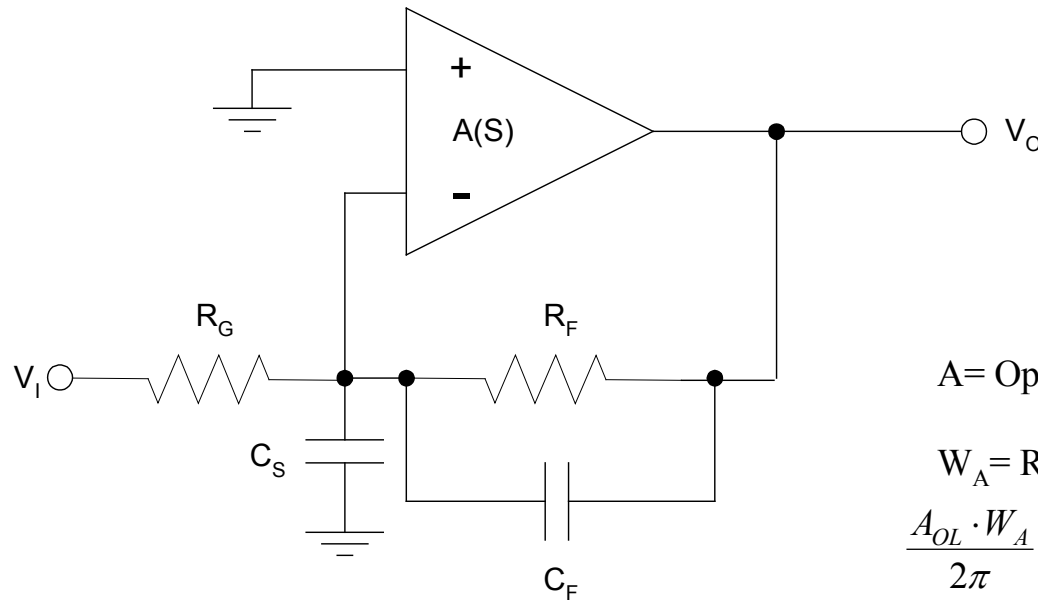


$$A(s) = \frac{A_{OL} \cdot W_A}{s + W_A}$$

A= Open Loop Gain

$W_A$= Radian Dominant Pole

$$\frac{A_{OL} \cdot W_A}{2\pi} = \text{Gain Bandwidth Product (GBW) Hz}$$

This is an inverting mode circuit that will shape the noise gain to transition from the resistive divider at low frequencies to a capacitive divider at loop gain x-over. It has proven very effective at improving low frequency distortion where very non-unity gain stable VFB devices can be applied at low gains with excellent results.

This circuit and its performance is developed in detail in "Unique Compensation Technique Tames High Bandwidth Voltage Feedback Op Amps", EDN, August 1st, 1997, pp133-150

# Bode Analysis for the New Compensation Circuit
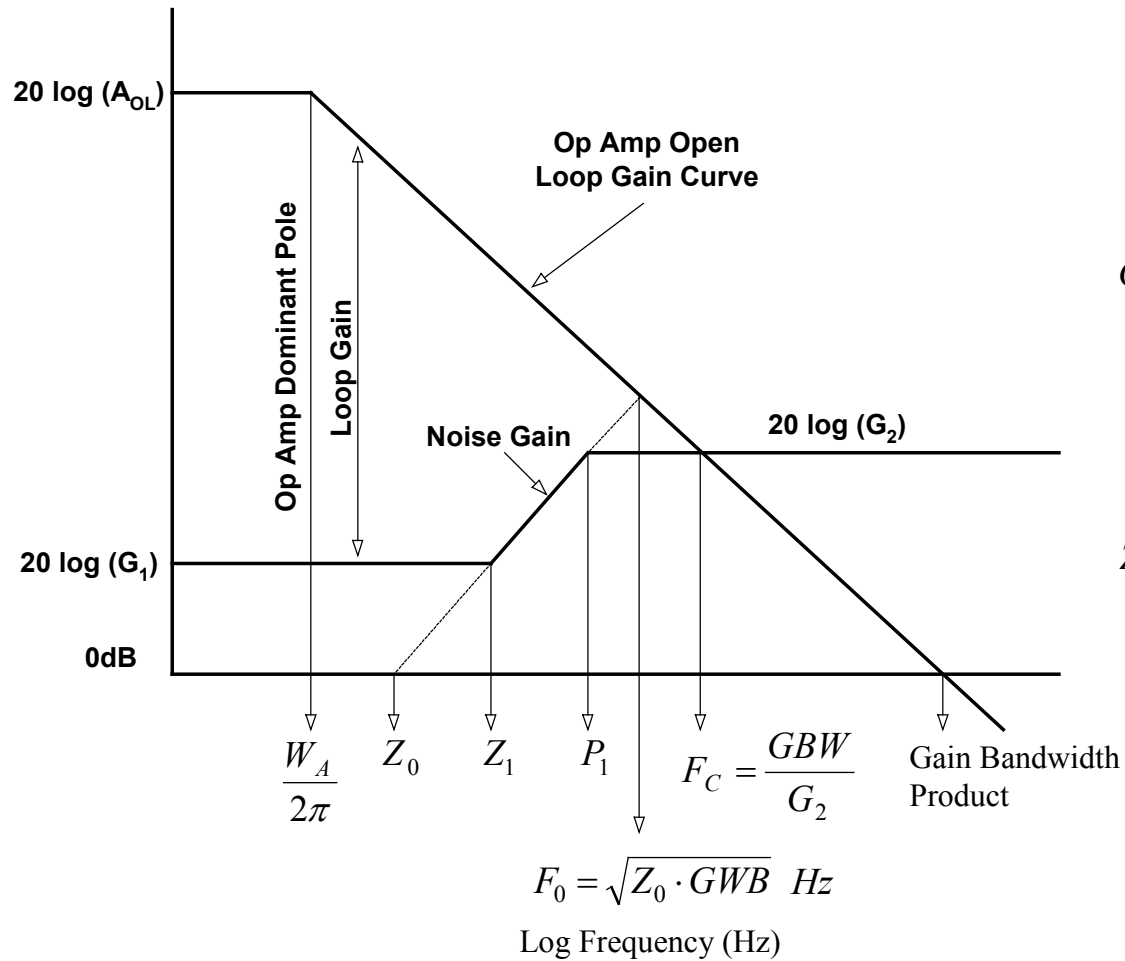

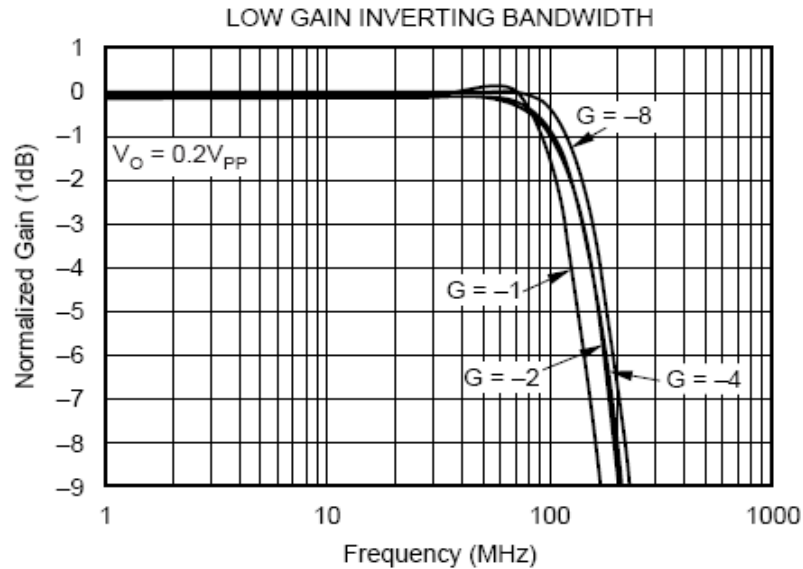
$$G_1 = 1 + \frac{R_F}{R_G}$$

$$G_2 = 1 + \frac{C_S}{C_F}$$

$$Z_0 = \frac{1}{R_F(C_F + C_S)2\pi}$$

$$Z_1 = \frac{1}{(R_F \parallel R_G)(C_F + C_S)2\pi} = G_1 \cdot Z_0$$

$$P_1 = \frac{1}{R_F \cdot C_F \cdot 2\pi} = G_2 \cdot Z_0$$

Graph labels:

- 20 log ($A_{OL}$)
- Op Amp Open Loop Gain Curve
- Op Amp Dominant Pole
- Loop Gain
- Noise Gain
- 20 log ($G_2$)
- 20 log ($G_1$)
- 0dB
- $\frac{W_A}{2\pi}$
- $Z_0$
- $Z_1$
- $P_1$
- $F_C = \frac{GBW}{G_2}$
- Gain Bandwidth Product
- $F_0 = \sqrt{Z_0 \cdot GWB} \ Hz$
- Log Frequency (Hz)

# Application of Inverting External Compensation to 3.9GHz GBP VFB Op Amp

LOW GAIN INVERTING BANDWIDTH



For each of these plots the Rg was changed to step the gain up and Cs, Cf adjusted to get a Butterworth response. Each of these shows about 140Mhz BW but hiding inside the response is exceptional loop gain below Z1 in the Bode plot. This gives extremely low distortion for this circuit below 10Mhz.

Without these comp. caps, the OPA847 is specified as minimum stable gain of 14V/V.

There is an analogous technique for CFB devices
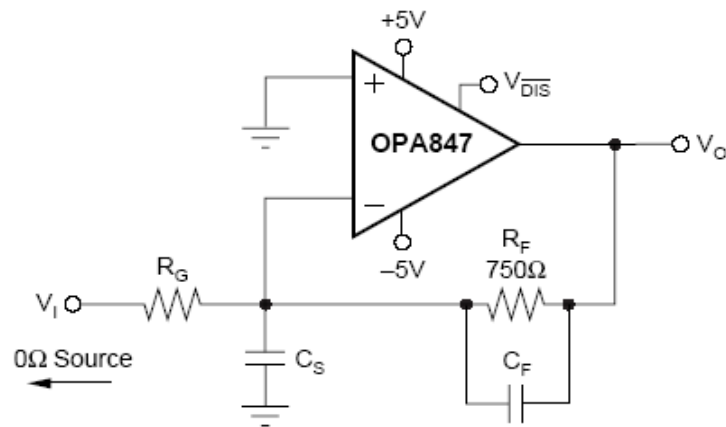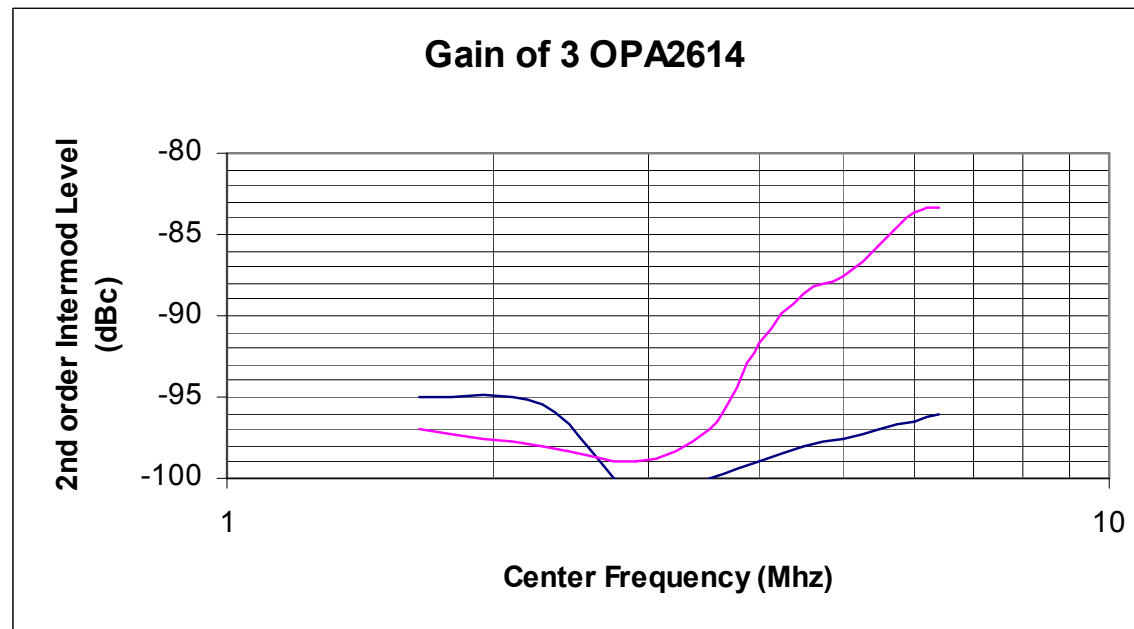
# Applications:

- – Transmit Function

# Example Design #1

Here – the input match was not done in the gain resistors – the circuit was eventually intended to show 1kohm differential load to the two outputs of a mixer. The 500ohm to ground on the outputs increased the supply current by adding a 2.5V/500Ω = 5mA in each output stage or another 10mA total – this class A current improved the 3rd order intermod.
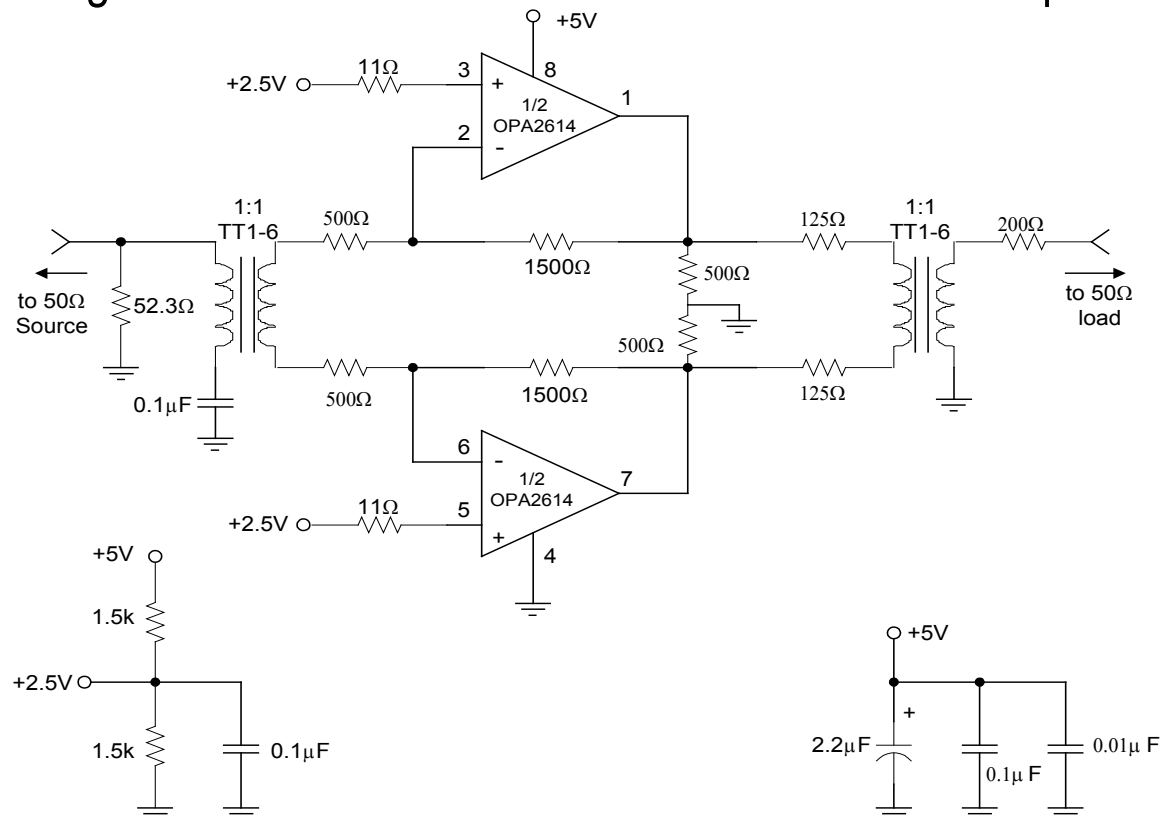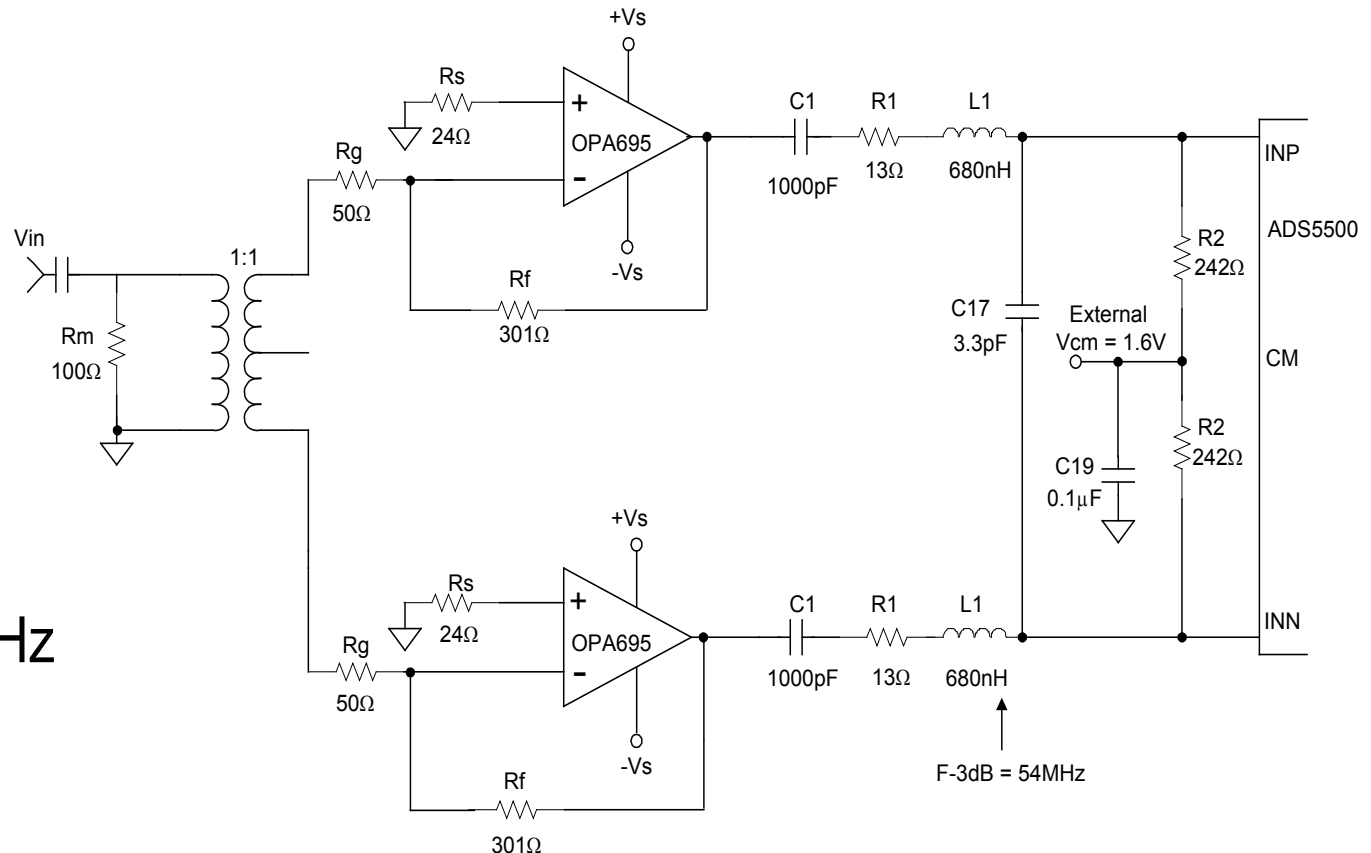
# Example Designs

This first example needed very low noise to low frequencies (low 1/f corner) , single +5V operation, moderate gain and <-85dBc SFDR to 5Mhz for 2 tones at 2Vpp total envelope. The test circuit showed the differential source is emulated with a transformer and the output differential signal is converted to single ended to measure through another transformer. Neither transformer would have been used in the actual application. The lower trace is the 2-tone, even order intermod while the higher one is 3$^{rd}$ order. This dual uses 10.5mA on +5V along with another 10mA of output stage class A current



**Gain of 3 OPA2614**

TEXAS INSTRUMENTS

# Example Design #2

1st Nyquist zone design for low frequency through 40MHz with 122MSPS clock rate



Input referred noise voltage is approx. 2.9nV/√Hz

This 2nd order RLC filter design is described in "RLC Filter Design for ADC Interface Applications" SBAA108A

**TEXAS INSTRUMENTS**

# Example Design #2

This 2nd example was similar to #1 but looking to achieve >80dB SFDR for an amplifier/converter combination through 40MHz at moderate gains (6V/V). Here a current feedback device was used to get good noise to high frequencies and better slew rate margin for the desired 2Vpp output signal. A 2nd order low pass at the output controlled the noise power bandwidth to hold minimal SNR degradation. This is using a 14-bit, 125MSPS converter.

**Performance of ADS5500 - OPA695**
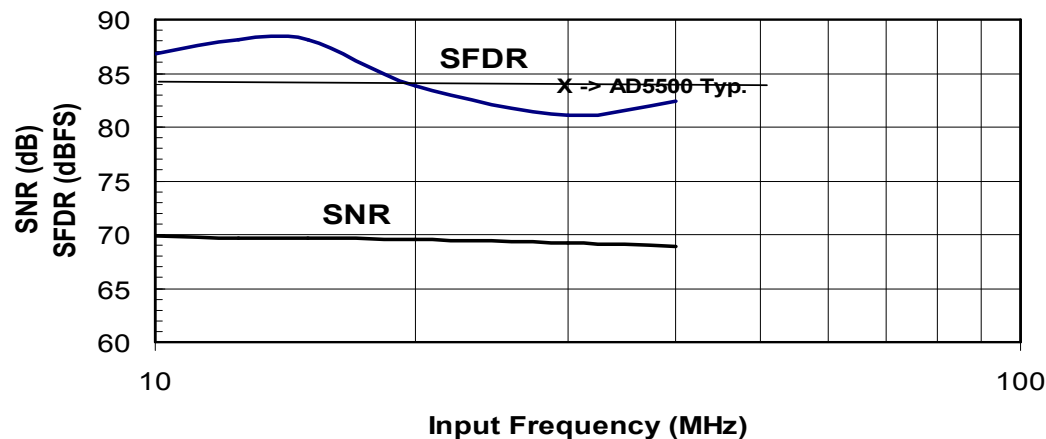


The ADS5500 has
SNR = 70.5dBFS

This circuit, layout, and performance is described in TI User Guide, SBOU028.

# Summary

– Transmit Function

# Conclusions

- Most VFB devices are low gain stable and can give the lowest noise and distortion at low gains and frequencies. Non-inverting differential I/O stages work pretty good here. For moderate performance targets, all CFB devices are low gain stable and do well to very high output powers.

- CFB devices at higher gains, and particularly inverting, are probably lower noise and can deliver a lower distortion to higher gains. Inverting differential I/O are the best for HD2 suppression.

- FDA devices at low gains can push the frequency envelope up for very low distortion. Also provide an easy way to get DC coupled single ended to differential conversion with a common mode level shift. Should carefully consider matching the feedback ratios over frequency for best results.