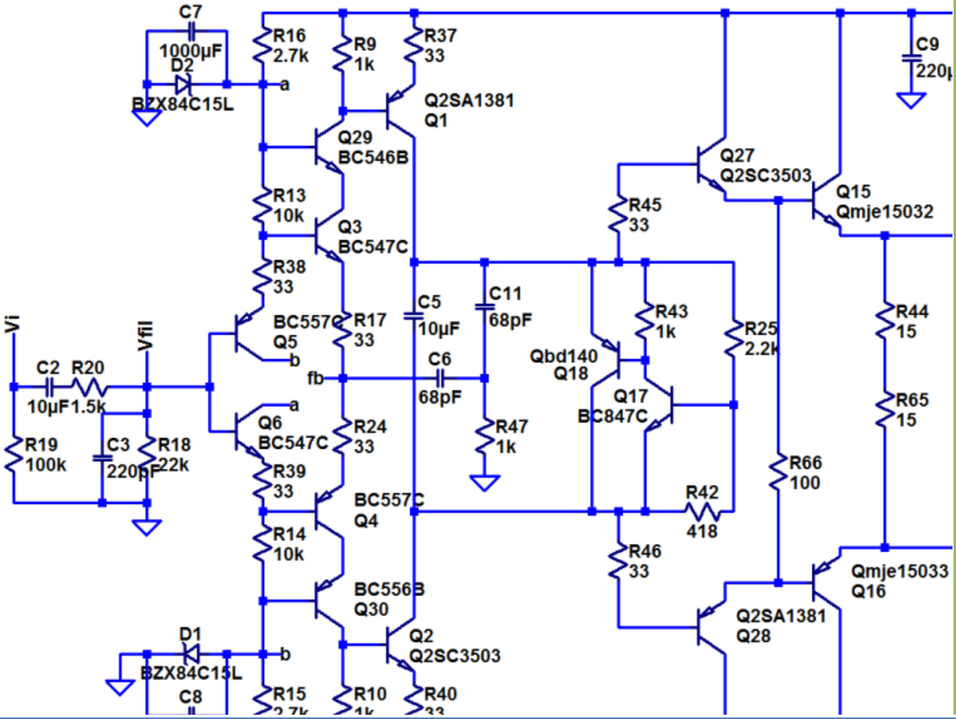

CFA vs. VFA: a short primer for the uninitiated

A concise introduction to VFA and CFA audio power amplifiers, in which the two topologies are compared and their strengths and weaknesses evaluated.

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CFA vs. classic Lin VFA topology:

A short primer for the uninitiated

Introduction. CFA topology amplifiers have been around in the IC industry for 30 years. Following a patent claim by David Nelson, the earliest commercial offering was a module from Comlinear in 1982 and a few years later, IC's from both Comlinear and Elantec. Prior to this, they were also described and analyzed in a number of papers. With regard to discrete based audio amplifiers, the topology has been used by a few esoteric brands in audio, with Accuphase, a Japanese company based in Yokohama, being a notable exponent. Cyrus, a small UK company, has also marketed CFA based power amplifiers. There are examples of Pioneer amplifiers from the early 1970's that used CFA, which apparently even pre-date the IC offerings and Mark Alexander published a design as an ADI application note in the 1980's, while Marantz have also marketed CFA power amplifiers. CFA topology audio amplifiers continue to be somewhat upstaged by their more widely understood and deployed VFA counterparts – a situation not helped by the fact that neither Cordell nor Self touched the subject in their otherwise wide ranging audio design books. A CFA's operation is not as intuitive as a VFA and there are some subtleties in regard to whether a transimpedance (TIS) or transadmitance (TAS) second stage is used and certainly the guidelines used by power amplifier designers to set the ULGF on VFA's do not apply to CFA's. The upshot of these and other factors meant designers preferred to go with something that is generally more widely documented and traditional – i.e. VFA. There is a lot of misinformation out in the audio industry and DIY community about CFA's, with some notable commentators dismissing them altogether, which is a pity since they do bring very specific properties to the table that are of benefit in audio power amplifiers.

There are many explanations about IC CFA topologies like this or this. Some plunge into math, loop gain equations and so forth, leaving the reader none the wiser, while this one (equations 1~4 and associated gain plots) from Hans Palouda is altogether easier to understand, as is ADI's here. For VFA's, Bruno Putzeys' explanation is by far the most succinct, even though the main thrust of his article is to dispel some enduring myths about negative feedback. Which brings me to the reason for this short document: 'CFA vs. classic VFA[1] – a short primer for the uninitiated' in which I will try to explain the differences between the two topologies, dispel the myths and hopefully encourage more audio power amplifier designers to experiment with this technique.

So, how do you tell if an amplifier is VFA or CFA?

[1] I have deliberately used the term 'classic VFA' to mean MC or *dominant pole compensated* VFA which will be used a vehicle to explain the fundamental differences between the two topologies. There are alternative VFA compensation schemes that allow the it=CV SR limit in MC to be broken so that similar SR performance is attainable by VFAs, while TPC allows the loop gain BW to be widened to that of CFA's as well. You can read about some of these techniques in Bob Cordell's book 'Designing Audio Power Amplifiers' Chapter 4.

Test/Pointer	VFA	CFA
Test 1	Peak input current to TIS = LTP tail current	Peak input current to TIS/TAS = $Vo_{peak}/R_{feedback}$
Test 2	Closed loop -3 dB bandwidth constrained by constant gain bandwidth product	Closed loop -3 dB bandwidth independent of closed loop gain* (See footnote 2 below)
Pointer 1	Both + and – inputs are high impedance nodes	+input is high impedance, -input is low impedance
Pointer 2	Two gain stages (LTP+TIS) = higher OLG	One gain stage – 2nd stage TIS/TAS = lower OLG

Table 1 – How to identify CFA from VFA – two tests and two pointers method

The two tests and two pointers method will allow you in *most cases* to accurately identify whether an amplifier circuit is CFA or not. Other than the mathematical derivations of the loop gains (see references) which are very different, the defining behavioral characteristic of classic CFA amplifiers are their gain-bandwidth product independence[2] and the fact that peak TIS input current (a key factor in SR performance) is not limited by the an input stage tail current, as is the case in a VFA. The detailed descriptions of the tests and pointers will be evident in the discussion of the two topologies that follow in this document. What is important here is that the above approach covers most variants of the two topologies – so single ended types, balanced, unbalanced circuits, JFET or bipolar inputs. If a circuit behaves like a CFA (or a VFA) then the assumption here is that it is a CFA (or VFA as the case may be). The pointers act as secondary guides, if identification is still difficult – in most cases however, tests 1 and 2 are sufficient to accurately categorize an amplifier topology.

Some amplifier designs are more difficult to identify – for example VFA's using folded cascode techniques are single gain stage VFA's; similarly, there are CFA's with two gain stages, and Jean Hiraga's famous 20W class A design from the early 1980's had an output stage with gain – so it was a two gain stages CFA . However, in both of these cases, they would pass Test 1 and Test 2 correctly for their specific topologies, allowing accurate identification. There are exceptions to the rule. H bridge input amplifiers appear topologically similar to a classic CFA with the inverting feedback network input buffered by a second diamond stage, mirroring the non-inverting input diamond buffer. A resistor connected between the summing junction of the two buffers sets the front end stage *gm*, allowing very wide bandwidths and high slew rates. The H bridge input stage would test out using the postulates in Table 1 as a CFA – the peak TIS current is set by the buffer coupling resistor, and it is *not* constant gain bandwidth limited like a VFA – a fact I easily confirmed in simulation. However, the IC industry classifies it as a VFA and so we will leave it at that – in certain cases the debate as to whether an amplifier circuit is VFA or CFA will remain a contentious one.

[2] Note this applies at low gains and at reduced PM's – the so called 'gain range sweet spot' often referred to in IC CFA application notes which is up to about 25 dB. At higher gains and or loop PM's, CFA's tend to degenerate into constant gain bandwidth behavior, albeit at higher closed loop bandwidths than VFA's. We will return to discuss these points later in the document.

VFA Overview. On the right hand side diagram of Fig. 1 you see a conceptual drawing of a classic VFA – differential input loaded with a current mirror, driving an integrator (a TIS with C_{dom} wrapped around it) followed by a unity gain buffer. A VFA thus described has two active gain stages - the LTP and the TIS; the input stage LTP is usually designed to provide gains of 20 to 40 dB depending on the design specifics, with most of the open loop gain coming from the TIS, with a gain of 60~85 dB. A VFA has two predominant poles in its transfer function – the TIS and the output stage. In the open loop condition, the TIS pole can lie anywhere between 10 kHz and a few 100 kHz and is caused primarily by the input capacitance load of the output stage on the TIS, and the TIS intrinsic input capacitance. The output stage pole is at about 30 MHz if you are using modern bipolar devices, but older devices like the MJ21193/21194 would show a pole at ~4 MHz. In a discrete design, which is what we are discussing here, the LTP and mirror load pole is considered to be much higher in frequency (perhaps up at 100 MHz). Both the inverting and non-inverting inputs are high impedance nodes in a VFA.

Because of the additional LTP gain stage and mirror load, VFA OLG is greater, but these active stages introduce more phase shift before the OL UGF[3] compared to a CFA. As we shall see later, this has a marked effect on compensation design between the two topologies.

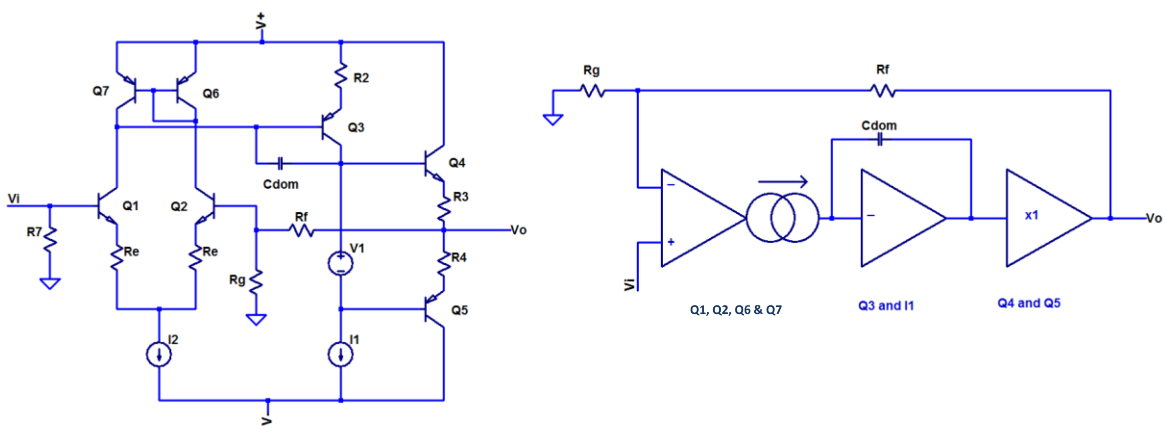


Figure 1 - VFA Generic Circuit and Conceptual Model RHS

[3] Note, this is not saying that phase shift is a property of gain – VFA's and CFA's are minimum phase systems

Hifisonix amplifiers from front to rear: 100 W class AB CFA nx-Amplifier, 15 W CFA Class A sx-Amplifier; Rear LHS 280 W class AB VFA Ovation 250 and rear RHS the 180 W class AB VFA e-Amp

In a Lin VFA topology, the input pair tail current is fixed by a current source I1 with the signals on the LTP input essentially steering a portion of this fixed current into or away from the TIS input node at the base of Q3 - hence the current source output depiction in the conceptual VFA in Fig. 1. The maximum output current of the diff amp stage available to drive the TIS (Q3 loaded by I1 in the circuit on the left in Fig. 1) and any compensation networks (MC, TPC, TMC, shunt, etc. but in this classic Lin VFA example, C_{dom})[4] is equal to this tail current I1, assuming the LTP is loaded with a mirror. If it's resistively loaded it's lower and in a correctly balanced LTP about *half* the tail current.

CFA Overview. In a CFA (Fig. 2) , the input devices are arranged in a diamond buffer configuration (Q1~Q4) with unity gain – the non-inverting input is a high impedance node, and the buffer output is connected to a *low impedance* inverting input node at the junction of Rg and Rf. Note that the front end buffer transistors (Q1 and Q2) are not inside the global feedback loop, as in the case of the VFA. The output current of the diamond stage appears at the collectors of the level shifters Q3 and Q4 and is not limited by a current source as is the case in a VFA, but instead set by the output voltage level and the *value of the feedback resistor* + Ro. Ro is usually small - in IC's a fraction of an Ω, but in practical power amplifiers in order to stabilize the DC operating point, usually up to 10's of Ω's.

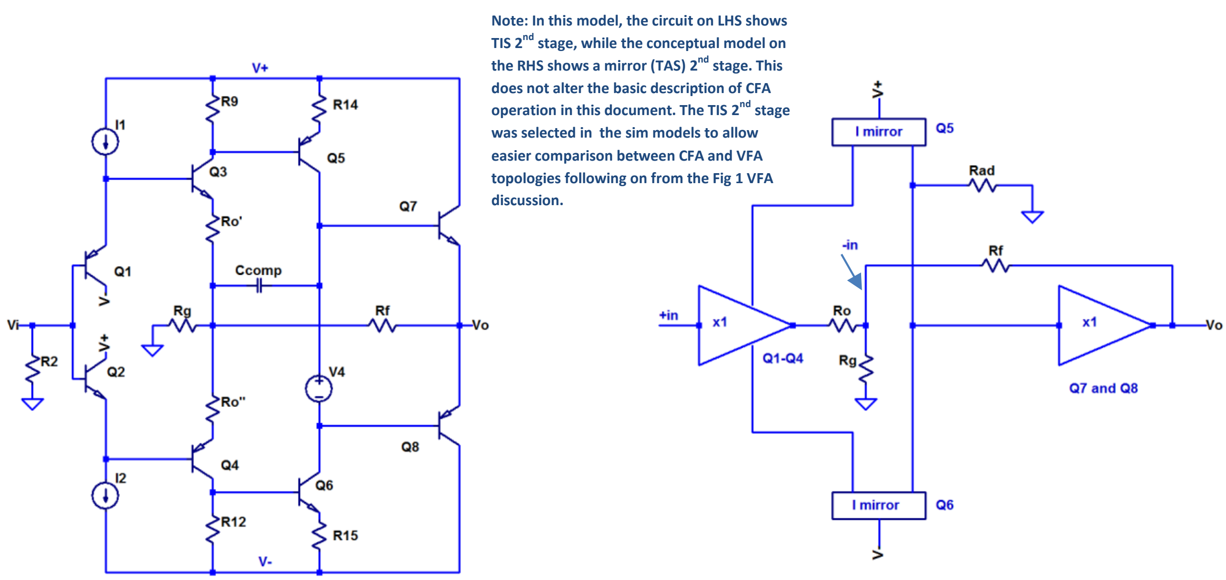


Figure 2 - CFA Generic Circuit and Conceptual Model on RHS

In IC opamps, a current mirror (TAS) is almost always used to convert the front end diamond buffer output current to a voltage that is then buffered by the output stage. This has the advantage, in general, of providing high gains from I_{mirror} x $R_{ad,}$ and isolating the front end stage

[4] MC = Miller compensation; TPC = Two Pole compensation; TMC = Transitional Miller Compensation; OIC = Output Inclusive Compensation

from the output stage much more so than a TIS. Further, since the intrinsic mirror bandwidth is very wide compared to a typical TIS configuration – MHz in the case of an IC opamp because there is little or no Miller effect – the stage pole is therefore also high. In the small signal regime of an IC opamp, this works well because the output load is a few mA and well defined with minimal load reflection back into the 2nd stage. However, the situation in an audio power amplifier is very different: the output stage input impedance varies significantly over the voltage swing and the output load impedance (most often reactive with big swings in load impedance) is reflected back onto its input to a much greater degree, thus, the load on the output of the TAS mirror is highly non-linear and the overall impact in terms of distortion reduction and bandwidth is less than one would expect. Therefore, in current feedback discrete power amplifiers, a conventional TIS makes much more sense, and as a result, instead of the uncompensated 2nd stage pole lying in the 100's of kHz as in the case of a TAS mirror, it typically lies in the 10's of kHz range – i.e. about an order of magnitude lower. Unlike VFA's, the phase shift accumulation in a CFA proceeds more slowly due to fewer active gain stages, affording greater PM and GM at HF.

TIS vs TAS. It is interesting here to contrast the behavior of an MC VFA, where the TIS output impedance (and stage gain) actually *decreases* with frequency due to the increasing local feedback provided by the reduction in C_{dom} X_c - thus the OPS is driven from a low source impedance at HF, mitigating somewhat the issues alluded to earlier. In a TAS CFA, you ideally want the mirror output to be flat in order to preserve bandwidth – difficult in practice on a power amplifier unless you are prepared to carry the burden of extra circuit complexity – and then you are still left with the output stage phase shift to deal with (see later). One solution to this problem is to configure the main gain stage in a CFA as a TIS, but to preserve the SR performance, apply Alexander compensation[5] where the local feedback loop is taken from the TIS output back to the inverting input – somewhat analogous to MIC in a VFA.

Setting Closed Loop Gain Magnitude. For both the VFA and CFA, the non-inverting closed loop gain is defined as $A_{vcl} = 1+(Rf/Rg)$, and for inverting simply as Rf/Rg, where Rf is the resistor connected between the output and the inverting input, and Rg is the resistor between the inverting input and ground as shown in Figs. 1 and 2. Although the closed loop gain for both configurations is expressed the same way, the underlying derivations (see Hans Palouda's article for example) are *not* the same, and this explains the differences in the loop gain behavior.

[5] See the appendix of the application note for the full derivation of this compensation technique

Loop Gain and Compensation Compared. As we have seen, VFA's (see figures 7 & 8) have higher open loop gains, and hence loop gains at LF because of the additional gain provided by the LTP stage, but phase shift accumulation also proceeds more rapidly as a result. To deal with this and ensure closed loop stability, *dominant pole compensation* (MC) is employed. In MC, the ULGF intercept is located by design (see the formula below) somewhere between 1 MHz and 3 MHz where there is adequate PM (60 degrees or more in a practical power amplifier) with an assumed slope of 20 dB/decade which then intercepts the loop LF gain at a frequency from a few hundred Hz down to a few 10s of Hz – the exact figure dependent upon the LF OLG. 'Dominant' pole compensation pushes the first open loop pole *down* in frequency, and the second open loop pole *up* beyond the ULGF resulting in so called 'pole splitting' - which you can see demonstrated graphically in curves 3 and 4 in Fig. 3 on the next page. The PM at HF is also thus improved and in the ideal case is 90 degrees at the ULGF. The result is a *constant gain bandwidth product* closed loop response which is a feature of dominant pole compensated amplifiers such that if we fix the ULGF and the required closed loop gain, the value of C_{dom} in Fig. 1 (assuming 20 dB/decade roll off)

$$Cdom=\left[\frac{1}{4\times\pi\times fulgf\times Acl\times(Rdegen+re')}\right]$$ [6]

Where f_{ulgf} *= the unity loop gain frequency (ULGF)*

A_{cl} *= is the closed loop gain magnitude below the -3db roll off point – i.e. low frequency gain*

R_{degen} *is the LTP emitter degeneration resistor - in Fig. 1 these are not shown as Re*

*re' is the internal emitter resistance of the LTP transistors from [0.026/(2*LTP tail current)]*

You can see the constant gain bandwidth term above from f_{ulgf} x *Acl.*

For Alexander compensated CFA's, assuming a 20 dB/decade response roll off, C_{comp} in Fig 2 can be estimated from

$$Ccomp = 1/[(2\pi \times (Rf + Ro + re') \times fulgf]$$

Note from this formula, there is no gain term Acl as in the VFA example. Shunt compensation from the TIS output to ground can also be used but the same ULGF in a CFA requires about five times the C_{comp} value compared to using Alexander compensation which also preserves the high slew rate performance of this topology. Some practitioners consider this type of compensation very sub-optimal. In my designs, I prefer to use Alexander compensation.

ULGF Intercepts in CFA and VFA compared. In CFA amplifiers, the fewer active gain stages and lower open loop and loop gain, mean that phase accumulation is less than in VFA's: the designer therefore does not have to employ dominant pole compensation to push the HF pole further below the unity gain frequency intercept to improve the PM for stability, instead trading the greater gain and PMs for wider loop gains.

[6] This formula is referenced in many texts e.g. D. self, R. Cordell, M. Leach etc.

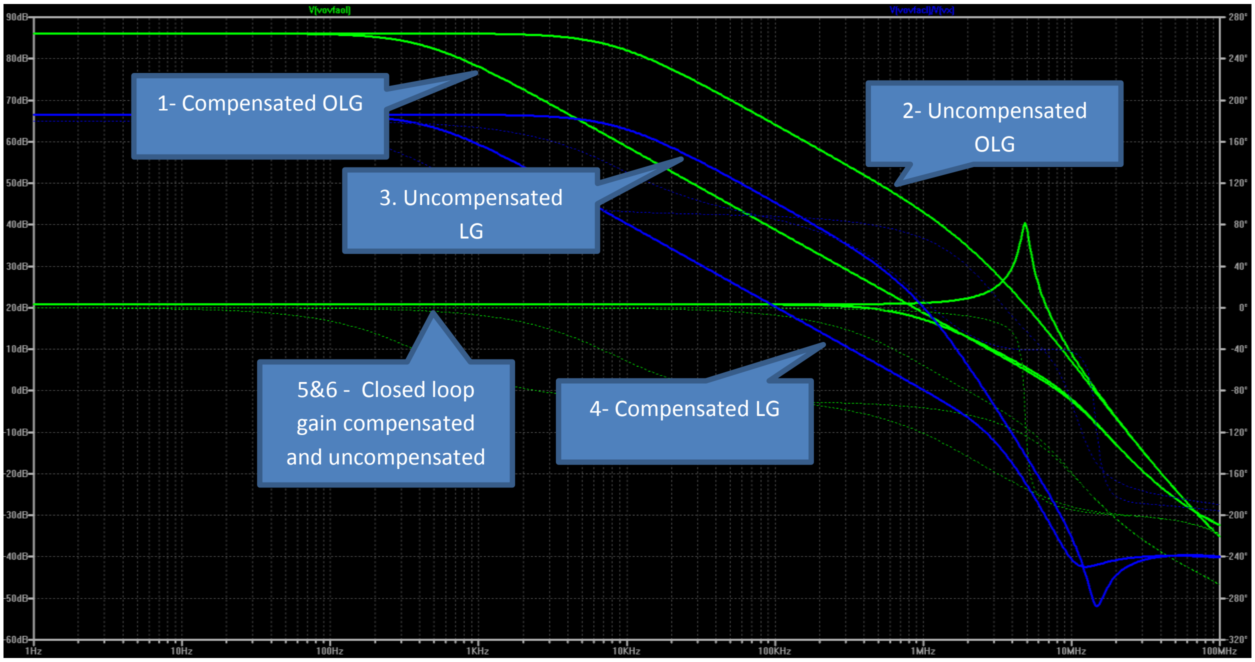


Figure 3 - VFA Pole Splitting

In Fig. 3 you can see the action of pole splitting in a VFA by comparing LG response curves 3 & 4 – the dominant pole at ~8 kHz is pushed down to ~200 Hz, while the HF pole at ~500 kHz is pushed up to ~2 MHz. The ULGF in this example is 1 MHz. Pole splitting reduces the effect of HF phase shift, ensuring stability.

The closed loop response (curves 5 and 6) show little deviation from each other until ~200 KHz, after which the uncompensated CL response diverges, peaking at 40dB at 5 MHz with rapid phase change. The Compensated LG roles off at 20 dB/decade with a CL UG frequency of about 15 MHz.

Note, if a CFA uses MC, the pole splitting behavior also results – so it's not unique to VFA's, but simply a property of dominant pole compensation.

It should be noted at this point that audio power amplifier applications are quite unique in that they demand PM's of at least 60 degrees in order to cater for a wide range of reactive loads. Therefore, no matter what compensation technique is deployed (MIC, TPC, OIC etc), the ULGF PM should always be in the region of 60 degrees or more to ensure the design is capable of dealing with real world loads. I usually incorporate an output coupling inductor of between 0.6uH and 1uH in my designs which is extremely effective in isolating the amplifier output from capacitive loads, ensuring stability.

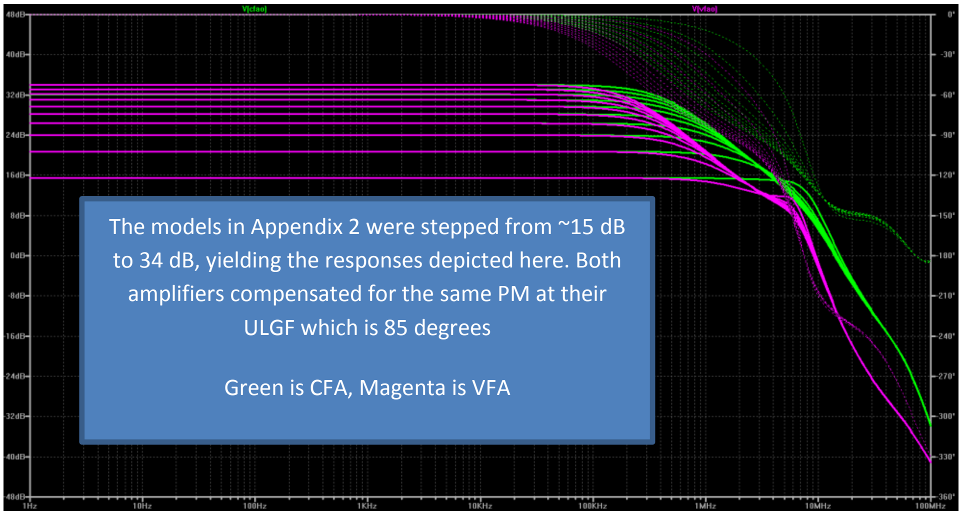


Figure 4 - VFA vs CFA closed loop Responses

The gain 'sweet spot' sometimes mentioned in IC applications notes refers to the gain bandwidth independence noted in CFA's. It's called a sweet spot because this characteristic only holds for loop PM's in the 30-50 degree range and *low CLG's*. Once the application requires high PM's – like the 60-90 degrees required for in an audio power amplifier - this characteristic is less evident as the loop compensation has to be conservative - see Fig. 4. In the IC application realm, it is for reasons of maximizing bandwidth that CFA's are generally compensated for much lower gain and phase margins thus preserving bandwidth – typical applications being video amplifiers and high speed data converters where the focus in terms of compensation design is on overshoot and settling performance, rather than PM.

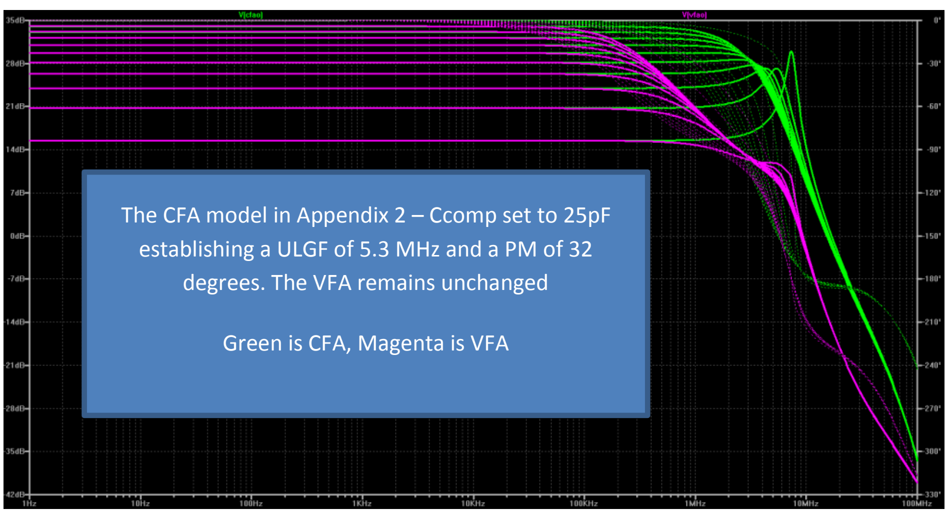


Figure 5 - CFA comp'd for ~30 degree PM showing gain bandwidth independence

In Fig.5 you see this characteristic demonstrated where the PM on the CFA model has been deliberately set to 32 degrees. Here the gain bandwidth independence is clearly visible at gains of up to about 28 dB, thereafter the bandwidth is linked to the gain.

On designs where the CLG is low but ULGF PM still rather high, the gain bandwidth independence is also better maintained, and you can see an example of this in a practical low CLG amplifier, the sx-Amp, on page 8 of that write-up. Lets be clear here: 30 degree PM's in practical audio power amplifiers will lead to problems – at least 60 degrees is required, with many designers targeting even higher figures in the 80~90 degree region.

In low PM and or low gain CFA systems, CLG can be varied over a wide range, provided Rf is kept constant to minimize any disturbance of the compensation, and Rg varied instead with reduced or little impact on the -3 dB bandwidth of the amplifier as shown in Fig. 4. In general, for low closed loop gain designs (typical in audio), the CFA is notable for its wide loop bandwidths – often > 10 kHz, and in the sx-Amp for example, it's ~60 kHz (see Fig. 13 red trace in the sx-Amplifier write up). Some designers claim that setting the loop gain -3 dB point *above* the audio bandwidth reduces Phase Intermodulation Distortion (PIMD), but this has been contested – see R. Cordell's TIM I and TIM II for example.

The conventional way to compensate an IC CFA is to adjust the value of Rf to achieve the optimum gain and PMs, usually by observing the overshoot and settling time to a fast rise/fall time square wave input stimulus. Part of the reason for this approach, rather than using some type of external compensation capacitor or network, and especially so in very high performance IC CFA's, is to avoid parasitic capacitance or inductance creeping into the TAS/TIS node which is what would happen if a compensation connection were brought out to one of the IC pins – remember, we are talking about devices with gain bandwidth products in the GHz region. In practical power amplifiers, capacitive shunt compensation from the TIS output to ground is often employed, although there are more advanced techniques like Alexander compensation as used in the sx-Amplifier. CFA amplifiers - and especially discrete power amplifiers - almost always exhibit gain peaking when the closed loop response is plotted, and this is linked of course to the output stage pole[7]. The cure is to bandwidth limit the input signal with a simple RC filter – you can see how I did this in the sx-Amplifier design (see Fig 14 in that document).

It's also important to note that you can apply dominant pole MC to CFA's as well – but to do so one ends up with an amplifier in which the response morphs into that of a classic MC pole splitting VFA, but with lower HF gain. For this reason, and SR performance, MC on a CFA is not recommended.

Slew Rate (SR). The other major difference between the two topologies is the slew rate (SR), set in a VFA by the compensation capacitor value and the LTP tail current from *SR= i/c*. As already pointed out, in a CFA the peak current available to the input of the TIS is set by the maximum output voltage and the value of the feedback resistor (assuming Ro is very low in value compared to Rg). In a correctly compensated CFA using TAS for the 2nd stage, this can be a factor of 10 higher than a classic VFA, and explains the big differences in SR between the two topologies, with 200 V/us the norm in a CFA audio power amplifier.

[7] Unlike discrete power amplifiers, the output stage device Ft's in IC devices are very high and peaking in IC's (CFA and VFA) is therefore more generally associated with stray capacitance from the –input to ground, and/or stray capacitance across Rf. Layout is critical to minimize these effects and to preserve the bandwidth performance – follow data sheet recommendations carefully to avoid these problems.

Output Stage Pole Impact on Loop Gain Bandwidth in Discrete Power Amplifiers. Importantly, in both topologies, the output stage phase shift ultimately limits the amount of feedback and the loop bandwidth of the amplifier. In a bipolar VFA, one usually sets the ULGF based on the output stage response; a good rule of thumb is to set it at between 5-10% of the Ft of the output devices *up to a maximum of 3 MHz* depending on the type of output stage[8] – EF2 or EF3. For example, If you use a MJL1302/3281 bipolar output stage with Ft's of 30 MHz, set the ULGF for an EF3 at 1~1.5 MHz, while for an EF2, you can go to 3 MHz. Mosfet output devices have Ft's at about 300 MHz, but in practice you cannot set the amplifier ULGF at 30 MHz because of circuit parasitics (layout inductances and capacitances) and the high input capacitance of these devices[9]. In this case, one would set the ULGF at 3 MHz, which is a practical upper limit for VFA audio power amplifiers. As already touched upon earlier in this document, there is greater gain and PM at HF in CFA's, and the loop *can be closed at higher frequencies.* Whereas in VFA's the general approach is usually to select a ULGF (see the formula on page 7 for example) on the assumption that around the 1-3 MHz region, the phase margin will be adequate, in a CFA its better to adjust the ULGF to meet a specific PM – i.e its PM that is prioritized – as already noted, the minimum recommended for audio power amplifiers is 60 degrees. Using this approach, the designer is able to exploit the greater gain and PM and the net result is wider loop bandwidths and lower HF distortion than would be the case if the same ULGF as a VFA were targeted. My investigations show that the improvements in HF loop gain can be as much as 12 dB, with 6~8 dB more the norm.

Practical Compensation Design and Optimization. In a practical VFA or CFA power amplifier, extensive testing of the final system enables the designer to determine the stability envelope. With purely resistive loads and the output inductor shorted, there must be no overshoot or ringing with a small signal (2~3 V pk-pk) 1 us rise time square wave stimulus *and no front end filter fitted*. If ringing is noted, the ULGF has to be lowered until the square wave response is clean. The second phase of testing involves the application of a wide range of capacitive and resistive parallel loads with the output coupling inductor in-situ. Ringing caused by the output inductance and the capacitive load will be observed, but the amplifier must not break into oscillation – if it does, the ULGF must be lowered and/or the output inductor value increased. I cover this subject from a practical perspective in the e-Amp, nx-Amp and already mentioned sx-Amp write ups. As a side note, it's also important that the designer is able to clearly distinguish between loop gain related stability issues and parasitic instability – the two are quite separate, and the cures different. However, one can trigger the other and this should also always be borne in mind. Some designers eschew the output inductor. If this is the design choice made, testing needs to ensure that the amplifier remains stable with the worst case expected capacitive load. It should be noted that the speaker cable inductance can help isolate the capacitive load, provided the cable capacitance is low. However, layout, decoupling and awareness of the impact of parasitic board and device elements is critical if you are to exploit this additional bandwidth.

[8] For both the EF2 and the EF3 ULGF figures quoted in this text, an output coupling inductor of 1uH or higher and Zobel network are mandatory.
[9] The 300 MHz Ft is only available if the devices is driven from a low source impedance such that the input pole thus formed is not the limiting factor in the response – this is difficult to do in a practical amplifier.

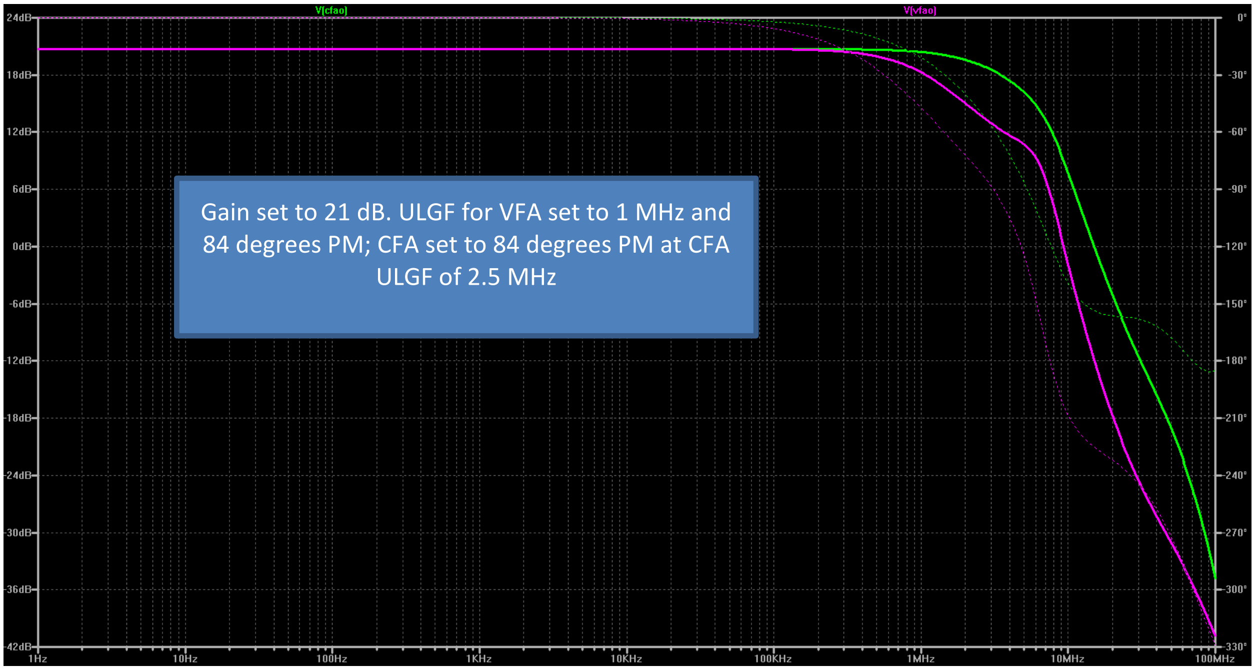


Figure 6 - The closed loop response of the model amplifiers at a gain of 21 dB

Figure 6 shows the closed loop response for the two model amplifiers used in this article. Despite the lower loop gain on the CFA, the UGF is twice that of the VFA and this is directly attributable to the greater PM available, allowing the loop to be closed at a higher frequency than the VFA. In a practical amplifier, a bandwidth limiting filter would be placed in front of both amplifiers to ensure their input stages were not exposed to fast input transients, keeping the input stage in the linear portion of their transfer curve, and providing RFI protection.

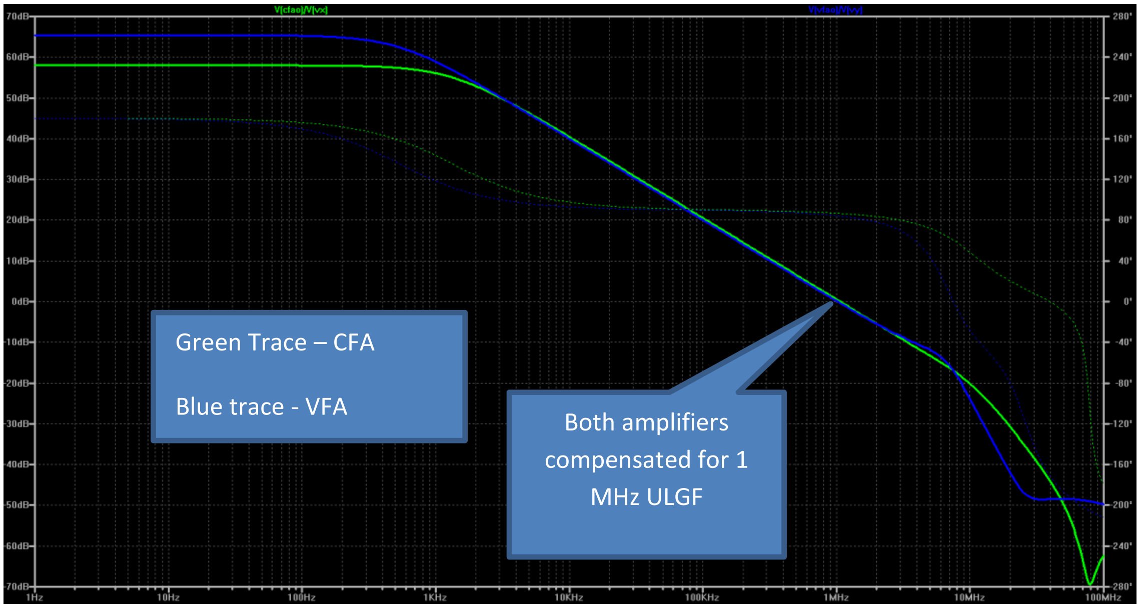


Figure 7 - Setting ULGF

ULGF's Compared. In a CFA, if you close the loop at the same ULGF as you would a VFA, the response after the loop -3 dB breakpoints *are very similar*, and this is reflected in Fig 7. However, in CFA's, simply closing the loop at the 'traditional' 1-3 MHz like you would with a VFA is suboptimal and ignores the additional gain and PM available which should instead be traded for higher ULGF. By closing the CFA loop at higher frequencies, (a) more feedback is made available at HF, and this is often the main reason for the lower HF distortion often observed in practical CFA amplifiers when compared to MC VFA designs and (b) the closed loop bandwidths are wider – see Fig. 8. For example, on the original [sx-amplifier] prototype design, the -3 dB closed loop bandwidth was in excess of 8 MHz.

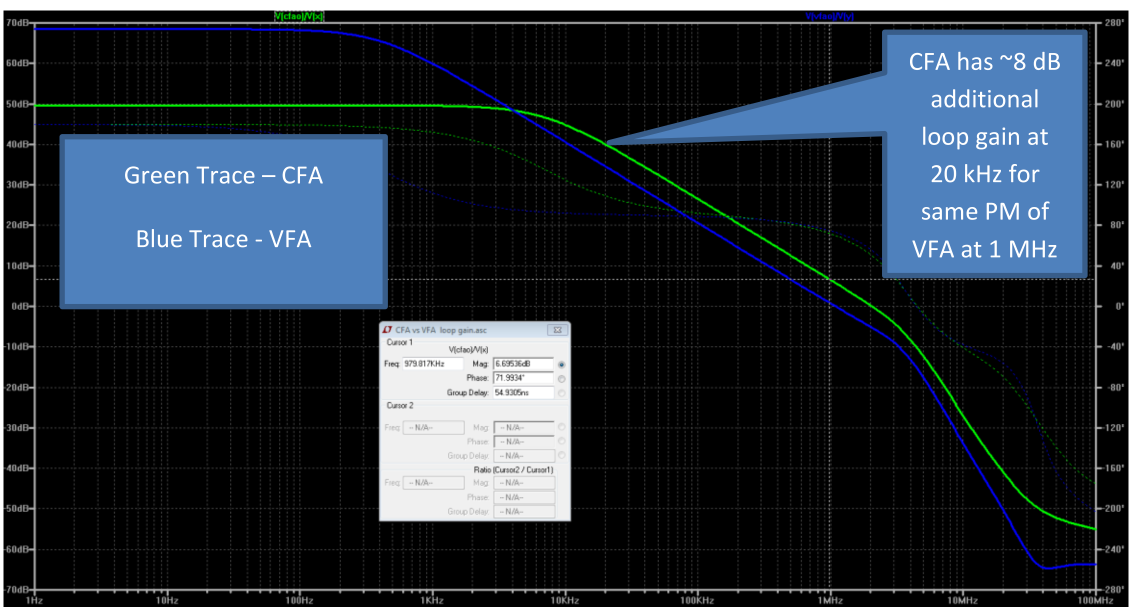
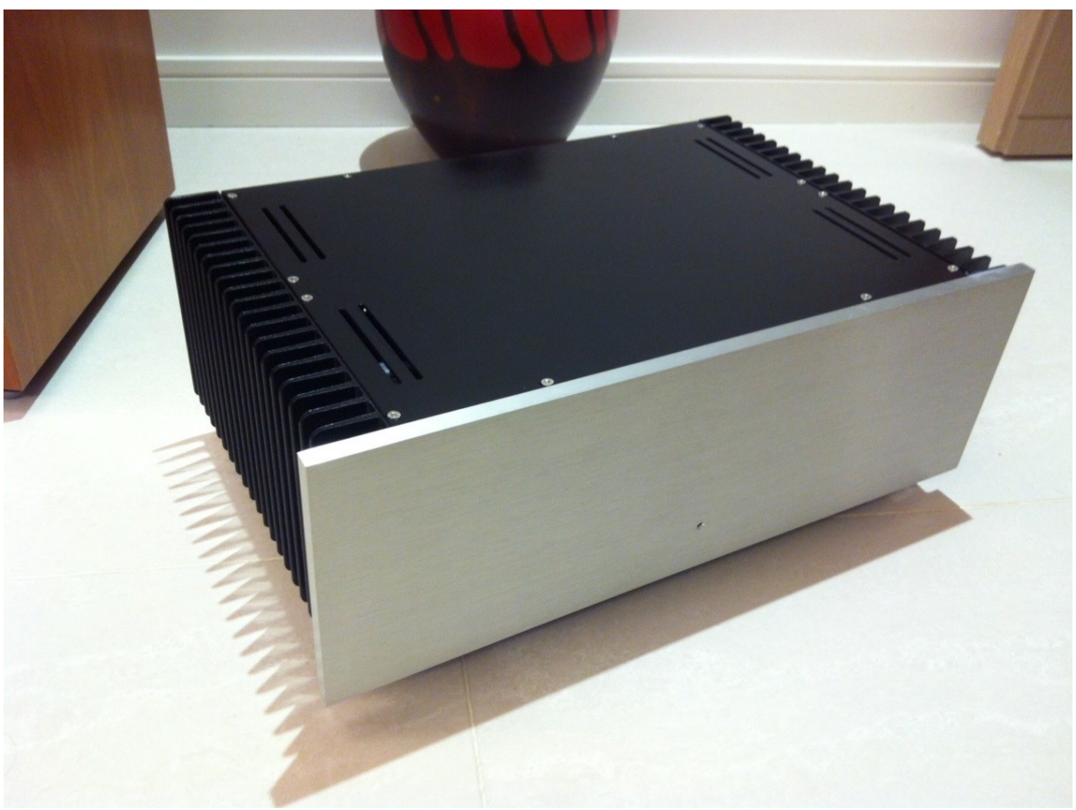


Figure 8 – CFA Compensated for Same PM as VFA ULGF of 1 MHz's (~70 degrees)

Topology Summary: VFA's have two gain stages, higher open loop gain and 2 major poles (TIS and OPS) in their open loop response; CFA's have one gain stage, lower open loop gains and also feature 2 major poles in in their response (TIS/TAS and OPS). VFA's require dominant pole compensation in order deal with greater phase shift in their response due to their additional gain stages (but providing higher OLG), and this compensation links the closed loop -3 dB bandwidth to the closed loop gain i.e. constant gain bandwidth closed loop response. In a correctly compensated classic VFA audio power amplifiers, the loop gain starts dropping off at -20 dB/decade from the dominant pole – usually a few 10's or 100's of Hz, such that the PM is 60 degrees or higher at the ULGF. CFA's on the other hand do not require dominant pole compensation, because the PM is greater and typically yields wider closed and feedback loop bandwidths; furthermore, the SR in CFA's is not limited by the tail current, but by the total feedback resistance (Rf + Ro in Fig. 2) and any capacitive load connected to the TIS output, allowing very high slew rates to be achieved as a matter of course. Because the loop gain bandwidths in CFA's are wider, this can translate into lower distortion at HF compared to VFA's (see Fig. 8). At LF, VFA's exhibit lower distortion because of the higher loop gain.

You can also draw the conclusion at this point that in practical audio amplifiers, there are few circuit differences between CFA and VFA topologies beyond their respective input stages.



Applicability to audio Power Amplifier Design. Both topologies can be exploited successfully to create high performance, practical audio amplifiers, provided their associated shortcomings are suitably mitigated. Table 2 below summarizes *indicative* class AB power amplifier performance parameters to give a feel for the differences in the topologies.

Parameter	VFA (classic MC)	CFA (TIS)	Notes
Open Loop -3 dB Bandwidth	500 Hz~5 KHz	1 kHz~10 KHz	
Open Loop Gain at 100 Hz	90~110 dB	60~75 dB	Additional gain provided by LTP in VFA
Loop Gain -3 dB	500~1kHz	1 kHz~100 kHz	CFA sx-Amp is ~60 kHz ; CFA nx-Amp is 8 kHz while VFA e-Amp[10] is about 1 kHz
SR	Easy to achieve 50~80 V/us	Easy to achieve 100~200 V/us	For audio power amplifiers, guide is minimum 1 V/us per peak output voltage
PSRR @ 1 kHz	70~90 dB	50~60 dB	Filter the supply rails; Use AFEC or Cap multiplier to improve CFA PSRR; Use Cap multiplier to improve VFA PSRR
THD @ 1 kHz	15 ppm	25 ppm	Greater VFA loop gain at 1 kHz results in lower distortion
THD @ 20 kHz	30 ppm	25 ppm	Loop gains at HF are often higher in CFA's
Closed loop – 3 dB response	150~200 kHz. Input filter may be required to ensure LTP remains in linear portion of transfer function on fast input transients	500~700 kHz often with response peaking, requiring input BW limiting filter	Both topologies may require input filters, but for different reasons as noted; RF ingress is also an issue in both cases but not considered here

Table 2 – CFA/VFA Indicative Performance Characteristics

Improving the Classic Topologies. In classic VFA designs, the ULGF limit usually sets the maximum feedback (loop gain) at 20 kHz to around 35 dB, assuming the 20 dB/decade loop gain roll-off required for stability. If higher loop gains and/or lower distortion at HF are desired, the designer has to employ more advanced compensation techniques like TMC, TPC, OIC and so forth. These can allow an additional 25~30 dB more feedback to be applied at 20 kHz without causing stability problems - approaches that are now considered mainstream in VFA topology amplifier design. With practical CFA power amplifiers, the second stage is usually in the form of a TIS rather than a mirror (TAS). For reasons already discussed, as the closed loop gain is increased, practical CFA power amplifiers can degenerate into VFA like CLG response. For this reason (and it applies to IC CFA designs as well), CFA's are *not* suited to very high gain applications - the lower open loop gains make that obvious in any event. However, they are imminently suitable for power amplifiers where the gains are 20~35 dB, and the wider loop gain bandwidths can be an advantage for HF distortion reduction. TPC and TMC can also be applied successfully to TIS CFA's and in simulation, TPC for example allows the feedback to be safely raised by a further ~30 dB, such that the total loop gain at 20 kHz is in excess of 55 dB, yielding low single digit 20 kHz ppm performance at full power.

[10] When configured for standard MC without TIS loading

However, based on the comments passed earlier about the higher possible ULGF in CFA's, if the overall OLG of a CFA is raised (and therefore loop gain as well), the PM will degrade, requiring that the ULGF be lowered if instability is to be avoided. It therefore appears that for CFA power amplifier designs, there is a tradeoff to be made if you are to avoid having a design that morphs into a low loop gain dominant pole amplifier – when it comes to OLG this is a case of less is more. Similarly, low open loop gain CFA's allow the designer to close the loop at very high frequencies if they so choose. In the sx-Amplifier for example, Cdom was deliberately set to 220pF for a ULGF of 3 MHz – however, the amplifier is perfectly stable with Cdom = 100pF, indicating a ULGF of > 4.5 MHz (figures from simulations).

In the dialog below, I capture some of the points raised between protagonists in the VFA vs CFA debate.

VFA's can achieve sub 1ppm and CFA's only 3 or 4ppm. If you are chasing sub 1ppm distortion then a VFA with advanced compensation and high open loop gains may allow a lower absolute number to be achieved. When applied to VFA's, advanced compensation techniques rather than the dominant pole MC we have discussed in this document allow the designer to make better use of the higher available loop gains to achieve very low distortion by managing the PM at the ULGF intercept. If one is prepared to trade the wider loop bandwidths and greater PM's of CFA's for greater OLG, these same techniques can be applied to CFA's, such that they then also achieve sub 1ppm distortion figures at HF. A few very advanced designs using this approach are demonstrated by DIYAudio forum members Damir Verson and Ostripper. AFEC, a simple feedback augmentation technique can meanwhile confer an additional 30~50 dB PSRR improvement on a CFA such that it performs better in this regard than a VFA (I suspect by the way that the true value add of AFEC is not in reducing distortion, but in improving PSRR and removing DC offsets). It is important to note that these are all 'simulation' figures, and say nothing about the real world performance of the final amplifier. You can read about some of the potential pitfalls in Douglas Self's 8 Distortions and a Few More.

I have remarked elsewhere that in audio, the nonsense of subjectivism has been supplanted by the tyranny of objectivism in which designers blindly pursue sub/single digit ppm distortion performance: its important to keep things in perspective since the source material, speakers and room acoustics all produce *orders of magnitude higher distortion* . Differences between a 1ppm and a 100ppm amplifier will still be inaudible, provided there are no glaring problems elsewhere. This same argument of course applies to those that chase excessive bandwidth or slew rates. Better to spend ones effort on the PSU, layout, wiring, housing design, construction and speakers than chase meaningless numbers.

In CFA's, the input stage transitions to class B, therefore they must sound inferior. This is not correct on both the technical and the sonic counts. In discrete CFA power amplifiers, the diamond input stage can be designed to remain in class A for any conceivable audio input signal. In CFA IC opamps, being able to have the input stage transition from (a very narrow) class A region to class B on demand is seen as a benefit, and exploited by CFA IC designers to keep power consumption at a minimum, whilst still being able to offer high slew rates and wide bandwidths.

As I show in the nx-Amplifier write-up (pages 7 & 8), all you have to do to ensure the front end always remains in class A is to design the peak feedback current under full power output slewing conditions to be lower than the level shifter standing current so that the 'off' side of the level shifter stage still remains on – i.e. it never actually goes off. This is best checked in a simulation model by feeding a very fast (10~50 ns) rise time signal into the amplifier input such that the output is operating in the large signal domain (so at least three quarters of the full output swing of the amplifier, but below clipping), and observing one half of the diamond stage's output current. It should not go to zero current. If it does, either increase the diamond stage output standing current (eg by reducing R_{degen}), increase the feedback resistor value, or apply some front end filtering (any combination of the three, or singly). This issue by the way, is analogous to a VFA LTP exiting its linear operating region, or slewing, which in both cases run the risk of one of the LTP devices switching off, whilst the other is driven into saturation. In a VFA, unless properly compensated, this runs the risk of SID – however, in a properly designed CFA (as is the case in a properly designed VFA) there is no risk of SID. In the case of VFA's, a combination of high tail current, degeneration and input filtering ensure that the front end never saturates – I discuss this in some depth in my VFA design e-Amp write up (see pages 23~26).

Secondly, there are already many IC CFA products (One example here from TI) that are being used successfully in high end audio with no reported 'sound' issues, and certainly there are no distortion anomalies like increased higher order distortion or cross over artifacts apparent in the specifications that could be attributed to the input stage class A to class B switching phenomena.

The PSRR on a CFA is up to 30 dB worse than a VFA. Use AFEC or a cap multiplier if you are concerned about it - In practice, it's inaudible. AFEC will take 1 kHz PSRR performance to >120 dB, while cap multipliers can confer PSRR performance of >100 dB. However, as a first step, decent layout and RC filtering of the supply rails go a long way to mitigating these issues.

The 500+ kHz closed loop bandwidth and 300 V/us slew rates of CFA are not needed. The same argument applies to those pursuing single digit or sub 1 ppm distortion – it's not needed either.

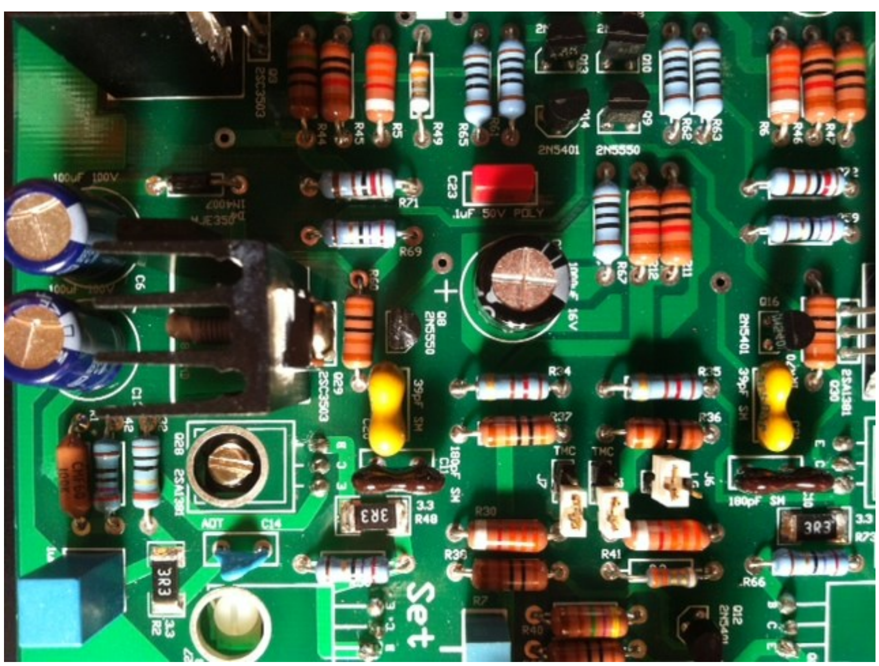
The lower loop gains at LF in CFA are bad (See Fig. 8 for an example). Presumably, this comment is based on the fact that most music signal energy is below 1 kHz, therefore more feedback will mean lower distortion and better sound. Firstly, there is no study that proves any correlation between higher feedback and better sound at frequencies below a few kHz – or any other audio bandwidth you care to choose. There are plenty of *zero global feedback* amplifiers that are highly rated – and of course, many designs with high feedback that also garner accolades. Despite perhaps clear measurement result differences, could a listener reliably identify a zero feedback from high feedback amplifier, assuming both were good designs in their own right? Is it likely that a listener would reject the zero feedback amplifier as inferior in a DBT? I do not think so and have to conclude this claim simply as conjecture. Secondly, the human ear cannot tell the difference between 0.1% and 0.01% THD on single tones, let alone a complex passage of music. What is important for pleasing sound, is that distortion spectra are reasonably low to begin with, and low order 2nds and 3rds. Simple CFA designs like the sx-Amp or nx-Amp, easily achieve sub 100 ppm at HF at full power and almost exclusively 2nd and 3rd harmonics, while alternative high loop gain CFA designs in DIYaudio.com have demonstrated 1~2ppm at high power. The correct conclusion is that below the threshold of human perception, measurements cannot be relied upon as a predictor of sonic performance which is an entirely subjective matter. This does not mean measurements are not important.

There is no such thing as a CFA – they are just sub-optimal VFA's. Firstly, apply the two tests and two pointers – there is a clear difference between the two topologies. Secondly, although yielding the same closed loop gain equations, the derivation of the closed loop gains for both topologies is *very different* – see Hans Palouda's article and Appendix 3 of this document. If a CFA was a suboptimal VFA, I would expect the loop gain derivations to be the same, which clearly they are not. There is nothing sub-optimal about a CFA's performance. They achieve very low distortion, wide bandwidth and high SR's. Their PSRR is not as high as VFA's, but then with VFA's you have to apply advanced forms of compensation (MIC or TPC example) to achieve similar SR performance anyway. Even then, they cannot match the inherently wide bandwidths of CFA's. Pick your tradeoff's - there is no free lunch with either topology.

I suspect the other reason for making this claim is that the proponents have confused canonical feedback forms with amplifier circuit topology – they are two completely separate matters - see appendices 4 and 5 which explain the differences clearly and concisely.

In Closing . . .

I hope this short article has steered a sensible path, highlighting the key differences, potential pitfalls, and benefits in the two topologies. There is no doubt that a bit more care is required to design and build a good power CFA – there is less latitude for error than in a VFA, where the higher loop gains at LF and better PSRR can help to cover a multitude of sins. However, tube amplifiers also require a lot of effort and care to create high performance designs, so in this respect CFA power amplifiers are not unique. CFA's bring to the designer, as a matter of course, high slew rates, wide bandwidths and circuit simplicity (single gain stage). They are easy to compensate, and dare I say it, *easily a match for VFA's in the sound department* . I will let page 4 of this document and the associated amplifier write-ups on www.hifisonix.com stand as testament to that. What is there not to like about them? I view the CFA topology as another instrument in my tool box in the pursuit of audio excellence, to be used alongside VFA topologies as and when circumstances dictate. At the end of the day, its one's ears and musical enjoyment that should decide, and nothing else.



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Prof. Marshal Leach 'Feedback Amplifiers – collection of solved problems'

sx-Amplifier

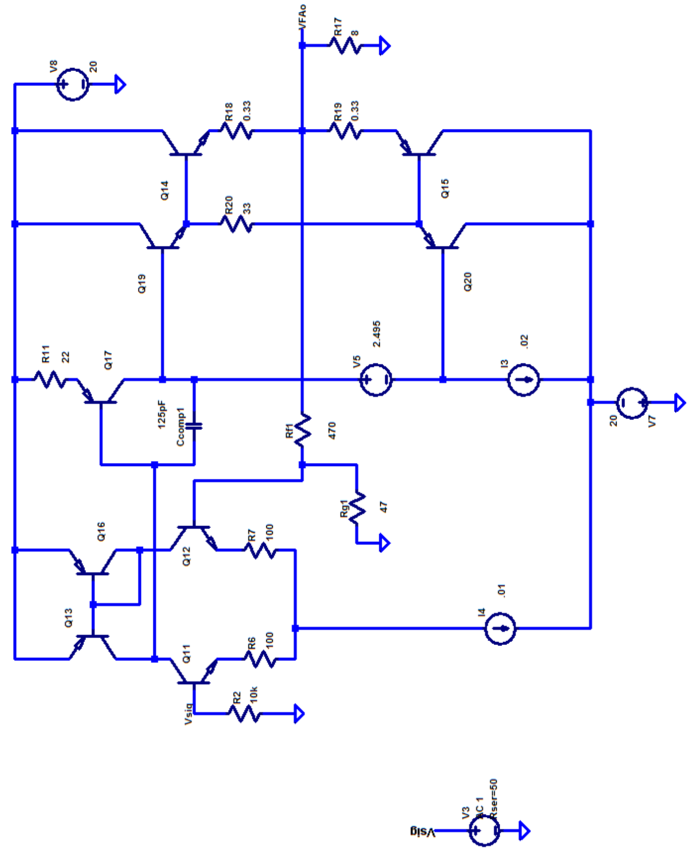
nx-Amplifier

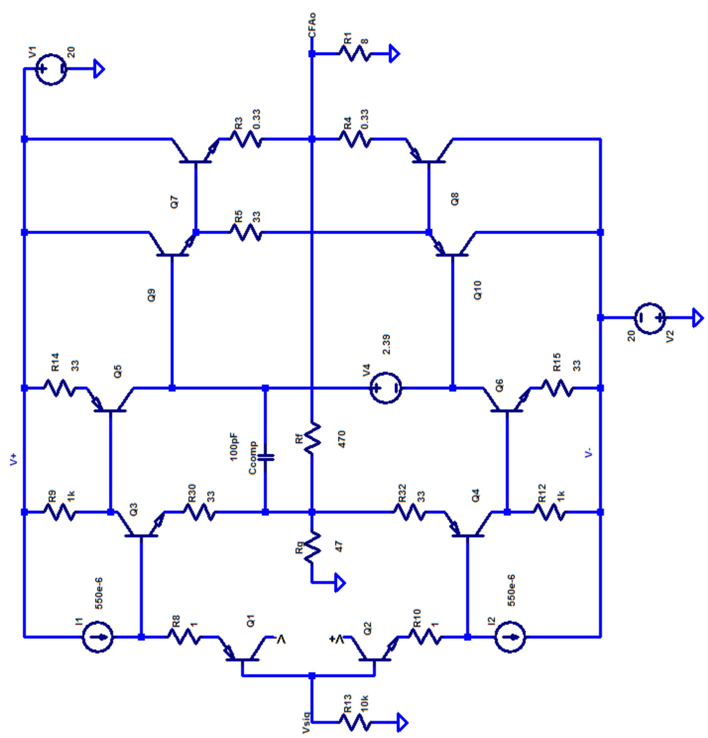
e-Amp

Document History

V1.00	**November 2013**	**Initial Release**
V1.01	**December 2013**	**Minor grammatical corrections, punctuation etc**
V2.00	**January 2014**	**Updated compensation and output stage poles discussion;**
		Changed/added new figures; corrected Fig 8 title; minor corrections; clarification of PM on page 8; based all plots and commentary on same models; corrected errors in models – used different device models for OPS on original draft; clarified filter comment on page 12;

Appendix 1 – Model Amplifiers Used in this Document



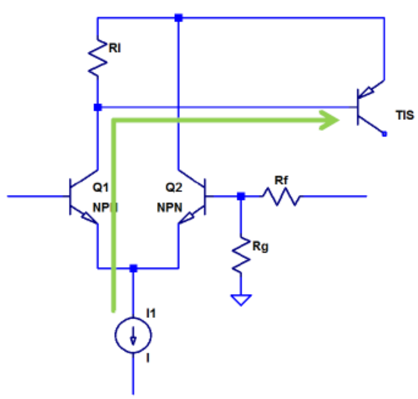


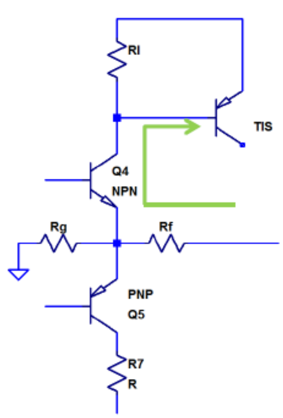
Appendix 2 – How to Tell CFA from VFA

How to tell if an amplifier is VFA or CFA – two tests and two pointers

Test	Detail	VFA	CFA
1	What determines the peak current into the TIS/TAS?	The LTP current	~Vo_{peak}/Rf
2	Is the amplifier constant gain bandwidth constrained	Yes: -3dB CL BW linked to CLG	No: -3 dB CL BW independent of CLG

Pointer	Detail	VFA	CFA
1	Are both inv and non-inv inputs high Z?	Yes – both + IP and – IP are Hi Z	No. + IP is Hi-Z; - IP is Lo Z
2	How many gain stages in the basic (classic) topology	At least 2 – LTP and TIS	1 – TIS or TAS structure





- VFA: Peak current into TIS cannot be higher than I1
- CFA: Peak current into TIS determined by value of Rf

Appendix 3: Table of CFA and VFA Gain Equations

Courtesy Texas Instruments Application note SLOA021 1999

CIRCUIT CONFIGURATION	CURRENT FEEDBACK AMPLIFIER	VOLTAGE FEEDBACK AMPLIFIER
NONINVERTING		
Direct gain	$\frac{Z(1 + Z_F/Z_G)}{Z_F(1 + Z_B/Z_F \parallel Z_G)}$	a
Loop gain	$Z/Z_F(1 + Z_B/Z_F \parallel Z_G)$	$aZ_F/(Z_G + Z_F)$
Closed loop gain	$1 + Z_F/Z_G$	$1 + Z_F/Z_G$
INVERTING		
Direct gain	$\frac{Z}{Z_G(1 + Z_B/Z_F \parallel Z_G)}$	$aZ_F/(Z_F + Z_G)$
Loop gain	$Z/Z_F(1 + Z_B/Z_F \parallel Z_G)$	$aZ_G/(Z_G + Z_F)$
Closed loop gain	Z_F/Z_G	Z_F/Z_G

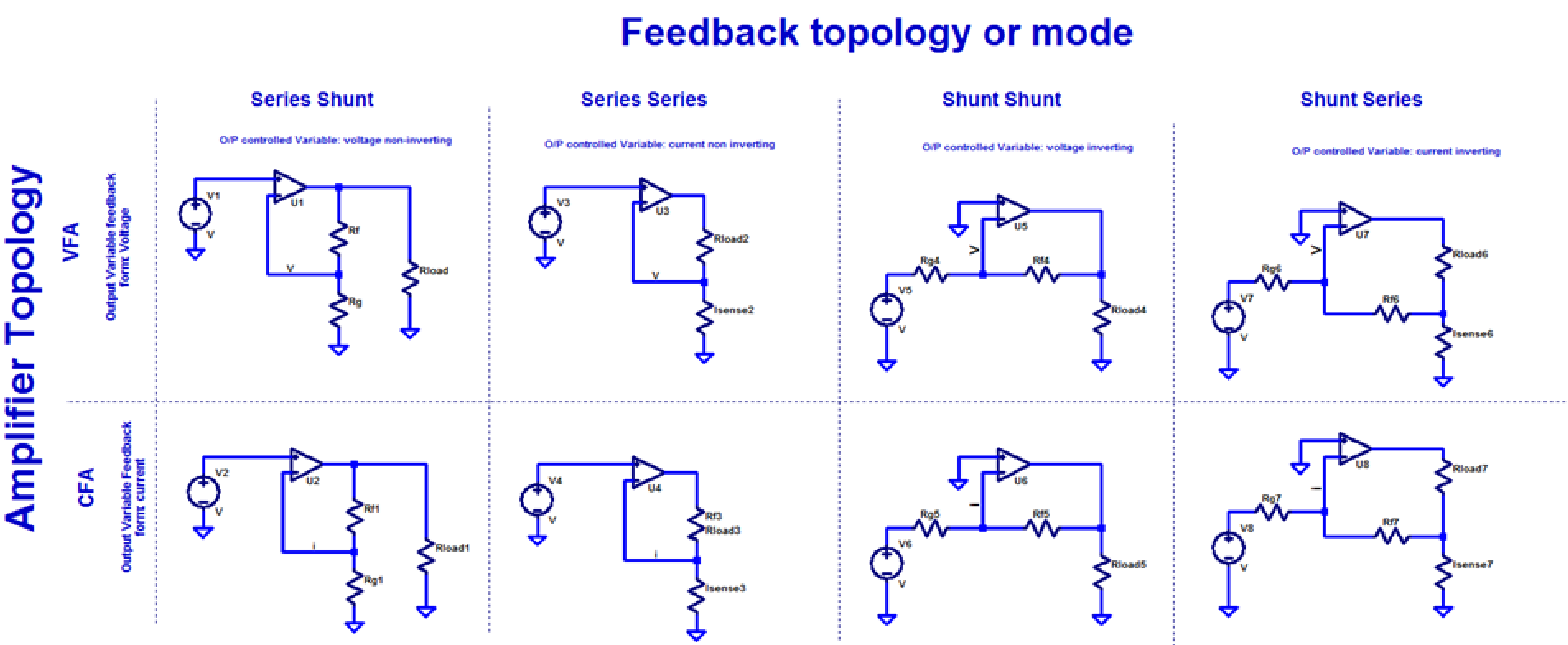
Appendix 4 – Canonical Feedback Topology vs Amplifier Topology Summary

(courtesy Prof. Marshal Leach, Georgia Tech, 2009)

Table 1. The Four Types of Feedback

Name	Input Variable x	Output Variable y	Error Variable z	Forward Gain A	Feedback Factor b
Series-Shunt	Voltage v	Voltage v	Voltage v	Voltage Gain	Dimensionless
Shunt-Shunt	Current i	Voltage v	Current i	Transresistance	siemens ($\mho$)
Series-Series	Voltage v	Current i	Voltage v	Transconductance	ohms (Ω)
Shunt-Series	Current i	Current i	Current i	Current Gain	Dimensionless

Appendix 5 (Overleaf) – Feedback Topology Summary



This table shows how the canonical feedback mode relates to amplifier topology. The feedback topology or mode axis is concerned with two things: what the (1) amplifier output controlled variable is, and (2) how the feedback network is sampling the controlled variable.

Series-Shunt – Input signal in series with feedback; feedback from output is in parallel (i.e shunts) the signal and load

Series-Series – input signal is in series with feedback; feedback from output is in series with load

Shunt-Shunt – feedback input shunts the input source; feedback from the output is in parallel with the load

Shunt-Series – Input signal is shunted by the feedback signal; feedback from output is in series with the load

Note the inverting input electrical quantity for each of the amplifiers, labelled V or I. Note that both CFA and VFA can be configured for *any* of the four feedback modes, as shown above. You therefore cannot use canonical feedback analysis to support the notion that CFA=VFA. Further, see the derivation of the VFA and CFA gain equations which explain the fundamental differences in the operation of CFA and VFA amplifiers