

THE PROJECT “MY_REFERENCE”

Technical description and design criteria for the amplifier “My_reference” based on low cost integrated chip solutions

Project, analysis and description by Mauro Penasa

Introduction

At about the middle eighties I had the possibility to listen to a solid state amplifier [Mosfet] of English origin, the Musical Fidelity A370. It's sound had impressed me a lot, because it was so different from the typical SS sound of this period, and was generally considered to be “tube like” in that it had a good level of “musicality”, most probably because of it's huge and realistic sound stage. Upon a closer scrutiny I've discovered that this amp was something really original, and, to top it up, was using an OPAMP as a front-end! Another strange feature was the battery of Mosfets in inverting configuration in the output stage, which had effectively transformed it's output stage into a “current pump”. I got curious about this technology, or rather, about the highly musical sound that resulted from it. So I decided to conduct a series of tests so as to understand it better, involving other techniques as well. Some characteristic features of that original circuit, like for example the use of the LM318 in inverting config, I had left unchanged, and for this reason one might discover similarities between my circuit and that of the M.F. A370.

On the other hand, I'm in debt with Tim de Parravicini [the designer of the MF A370], for shedding light on this rather rarely used technique. Practically every development in electronics is based on “previous art”, and this is the reason why I 'm rather sceptic about the concept of “authorship” of circuits, and I believe it's always honest and intellectually correct to give due acknowledge to the source of the ideas on which a particular new design is based. In this circuit here I am just taking a step further the concepts pioneered by others [and in particular, by M.F.]. Though obviously my activity here was quite different from the usual cut and paste so typical to certain “creative” ambients..

Theory of operation

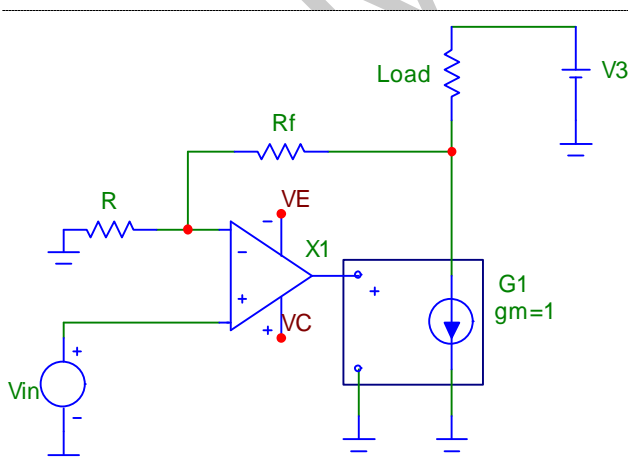


Fig1

Fig 1 shows the principal circuit. The block G1 is a transconductance stage, [input in tension / output in current] and can be realized in either discrete or integrated format, as needed. This stage puts out all the power that will drive the load, that is, this is the power stage, and here for convenience let's assume it has a $g_m=1$ [that is, unity transconductance; 1Volt input = 1Amper output] With these conditions, the stage G1, while generates the output current in the load, automatically determines also the tension dropping on it, so we get also an output stage voltage amplification factor, which is proportional to the load impedance:

$$A_v(G1) \sim Z_{load}/g_m.$$

For this reason, the differential input stage X1 is liberated from the task of generating a full scale output signal, so a standard op-amp can be used here. The open loop voltage gain in this circuit can reach quite high levels, because it's equivalent to:

$$A_{ol} = A_v(X1) * A_v(G1) \text{ [where } A_v(X1) = \text{open loop gain of X1]}$$

Note that A_{ol} depends on the load characteristics [on count of the $A_v(G1)$ term], and because of this the final tuning, with special respect to the frequency compensation, becomes very important. It is of particular importance to maintain a good phase margin, so as to avoid that the variations in the load could lead to an unstable condition. ($A_{ol} > 1 @ 180$ phase)

Naturally, the closed loop voltage gain will be calculated as usual:

$$A_v = 1 + (R_f/R)$$

This circuit has two interesting features:

1. The output stage is a current generator, and the structure used to obtain this property is quite different from the usual. Especially the applied bias technique, the phase characteristics and the dynamics of the G1 stage are different from the usual. This element is really important for what regards some forms of THD & IMD, typical of the "standard" configurations. Particularly, it's possible to exploit some peculiarity of some circuits, as we will see it in the next chapters.
2. Probably the most important feature, from an "audio" point of view, is that:

As a whole, the complete circuit works like if it were a common, normal Voltage Opamp, because the high open loop output impedance becomes compensated by the applied negative feedback, according to the following:

$$R_{out} = R_{int} / (1 + A_{ol} * \beta) \text{ where } \beta = R / (R + R_f) \text{ (}\beta = \text{feedback factor)}$$

But, looking closer at the above equations and the circuit, we can observe that the Load, because of the unique functionality of G1, **becomes an integral part of the transfer function**, because the signal that feeds the NFB network is not simply $V_{out}(G1)$, but $I_{out}(G1) Z_{load}$!

All the nonlinearities of the Z_{load} become highly apparent at the input terminals of the differential pair, which now can proceed to compensate for them. [by driving the output current of G1 accordingly.] In a conventional setup, the open loop output impedance is generally much lower with respect to the load, and **so the NFB network works by sampling the amps output voltage**. ($V_{out} = A_v * V_{in}$) In this way the circuit, though it will be highly immune to the load variations, it will become too much self-referent and as such will not be able to compensate for the dynamic interfacing problems between the loudspeaker and amplifier. This might be the real reason why a lot of modern SS power-Opamp performs less satisfying than some older amps with other technologies.

My_ref, output stage

I am experimenting with solid state audio circuits since a lot of years, and start to conclude that one of the basic problems with both of BJT & Mosfet power stages is the biasing, both from static and dynamic point of view. We can attribute to it either directly or indirectly the generation of harmonic distortions and also different instabilities. The most antique [and simplistic] mode of obtaining a linear power stage is the Class A biasing. [a classification which I think evokes direct associations in audiophiles with a certain “class” of the sound..] This method, apart from its inefficiency, forces the usage of a large amount of output devices mounted on enormous metallic structures with fins.. Getting tired of being a metalworker, I have decided to survey the world of negative feedback, even if it is usually considered off-limits for the High End territory.

As an interesting alternative, the use of power chips came into sight. These are normal integrated operational amplifiers, only with a higher voltage / power output. Their differential structure gives quite a high level of freedom to the constructor. One of my design goals was to try to “hide” their inherent THD & IMD in the resulting audio circuit.

One of my “discoveries” was the current bridge, that is, “Bilateral Current Source”, as it is called by National Semiconductor [AN-29 Linear Applications Databook], or “Howland Current Pump”, as it is referred to by some.

The main (nice) feature of this bridge, as I discovered, (apart from being an optimal transconductance building block) is the capacity of reducing that negative impact on the sound, which accompanies so much of the usual high feedback circuits, especially when the unit is included in a global negative feedback. The reason for it can be found in the “bridge” structure. In such a configuration, all of the Aol (open loop gain) of the unit will be used for stabilizing (balancing) the bridge, so it will – in a certain sense – remain “outside”, will be left out from the dynamic processes trying to linearize the output signal. This feature is quite typical to the transconductance type circuits, which, when well implemented, have a minor impact on the dynamics of the distortion processes. (at least in audio range) We could say that all the transconductance output stages, when included in a global voltage NFB loop, are potentially “neutral” with respect to the final result, because of the different relationship between current and tension, generated on reactive loads.

The “current pump”:

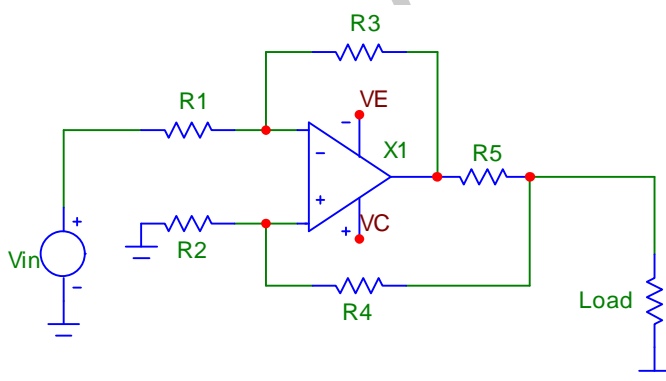


Fig 2

The design formulas for the “current pump”:

$$I_{out} = - ((R3 \text{ Vin}) / (R1 \text{ R5}))$$

If: $R3 = R4 + R5$ (or $R5 \ll R4$) & $R1 = R2$

$R_{out} = R5(R/\lambda R)$ $R = R4$ or $R3$ and $\lambda R =$ bridge res. Error ($R3$ compare $R4$ & $R1$ compare $R2$)

Another nice feature of this circuit is the possibility to use it in a differential configuration and, in case of $R1 = R2 = R3 = R4$ [that is, the bridge balanced], it is possible to connect to the output any kind of tension [floating], without afflicting in any way the functionality of the current mode operation. Further, it is possible to use as an input both $R1$ or $R2$ [inverting or noninverting input].

So as to keep the bridge stable, it is necessary that $X1$ be unity gain stable. If it can not be fulfilled, then one could apply a pole / zero compensation network between the two differential inputs, which will work by degenerating the bridge characteristics above a certain frequency limit. One possible technique, if we use chips which are stable with gains between 2 and 5, is to “design in” $R1$ and $R2$, as elements of the compensation network, together with the chip’s input capacity and differential input impedance.

One should not forget that such a bridge configuration [and the current mode operation] frequently creates a “group delay” which can become a problem in a “closed loop” situation. For this reason the compensation networks should be carefully thought over..

I have done my first experiments with this type of bridge using the TDA2030, then I passed to the more powerful types, like the LM3875 /3886. Each time the compensation should be carefully re-stabilized, in accordance with the actual chip type applied.

The application of this bridge in My_ref:

Initially, I’ve used a bridge with LM3886, configured with an R_{out} quite small, at about 4-10 ohms (see My_ampli), because my main target was to exploit the good “NFB neutrality” characteristics of this configuration. The result was quite OK right away, because the final sound reflected almost exclusively the tonal characteristics of the input stage (lm318). One basic problem remained, namely that in this configuration the resulting g_m of the bridge was quite high, which in consequence raised both the level of the output damping factor and also its frequency dependency.

After having conducted a series of studies on the problems connected to the damping Factor (DF; the relation between Z_{int} of the amp and Z_{load}) and after having re-considerated the way how global NFB works (shown in the theoretical circuit description),

I have found it beneficial to raise the Z_{out} of the bridge to about 500 ohm, which, in consequence, also reduced its g_m to $g_m = 1$.

My choice of the LM3886 is based on its low cost, high reliability, and good power output level (68Wrms) even at 4 ohms load level. A general efficiency problem remained, though: it’s caused by the value of the resistor in the output, $R5 = 47\text{ohm}$.

This value was needed to get the high Z_{out} and low g_m levels described above. Under heavy load conditions the dissipated power in $R5$ gets quite high, but I consider it acceptable in the light of the obtained acoustical results, especially if we consider that the losses provoked in this way are much smaller than in case of class A circuits – which would be the direct competitors here..

My_ref, the input stage

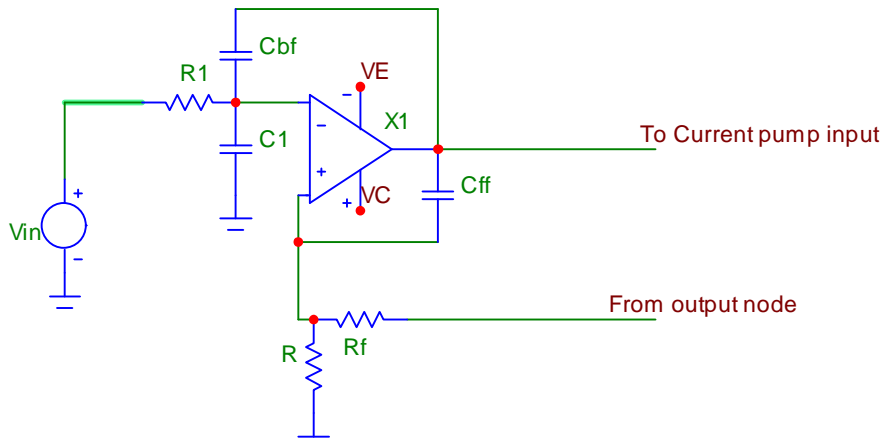


Fig.3

As it had been shown in the theoretical circuit description, it is highly possible to use a standard, low voltage operational amplifier in the input circuit, given the fact that even in case of a small value g_m , ($g_m=1$) the output buffer will still have a small but sufficient output voltage gain margin. The input circuit can be considered, in all the topologies, the most important part of an audio amplifier, because it is here where the “linearization” of the input differential signal happens, and an eventual malfunction here can lead to raised THD & IMD levels which can not be corrected any more. In case of the actual “My_ref” circuit, the structure of the input diff. amplifier is a little bit different from that “typical” configuration shown in the theoretical circuit (Fig.1). In particular, I’ve chosen the same structure that I discovered by scrutinizing the M.F. A370, and what I use since then with success. The input signal, after the input filter / decoupler network $R1\ C1$, is connected to the inverting input of the LM318N. It’s for many years that this chip accompanies me in my designs, because I believe that still today it has a price/quality ratio almost unequalled, moreover it’s innards are accessible in the key points, so it’s easy to compensate it “to full tilt”, both in frequency & in phase. The non inverting input, differently from the standard solutions, is connected to the NFB network. Applying after this the current bridge also in inverting config, will sum up to a globally non-inverting configuration. This type of connection, which we can define as “floating”, because of the missing “virtual ground” on the inverting input, has some advantages over a “classic” non-inverting config:

- 1.) The sequel of two active inverting stages in series permits to apply diverse techniques of compensation, which permit to optimize for global stability as well as “align” relative phase (for example, the inverse driven phase test)
- 2.) The fact that the input signal is applied to the inverting (-) input, allows to profit the most from the opamps inner characteristics, which are more linear in this configuration. (though this is not an absolute rule, and varies from case to case)
- 3.) This “floating” condition helps the front end in the hard job to follow, “hunt down” the non-linearities in tension presented on the output load.

Compensation techniques:

Apart from other, local compensation elements, which I added to or eliminated from the different circuit versions, I have basically used two different techniques, which are actually complementary in their function. In “MY_REF RevA (the “official” version) I’ve applied Cbf (Cap. Back-feed or feedback). This component creates a high frequency dominant pole (integration), together with R1, but inversely proportional (in frequency) to the output signal of the LM318. In this way, it’s possible to maintain unconditional stability of the circuit.

In the Rev3 version I've used the inverse approach, based on Cff (Cap. Feed-forward).

This component executes a positive compensation of the diminishing gain of the input differential with higher frequencies (it “differentiates” the output of the LM318) This method results in a notable linearization of the DF... (with frequency).

Further features:

The Zout open loop output impedance of X1 participates in the balancing act of the current bridge – so it's value should be taken into account. Not doing it can upset the bridge balance and can lead to a decrease of the bridge's final output impedance. In case of the LM318 this value is at about 20 –30 ohm, so with the given values, the bridge will be balanced close to 0.1%, which corresponds to about 400 – 500 ohm Zout.

The complete circuit

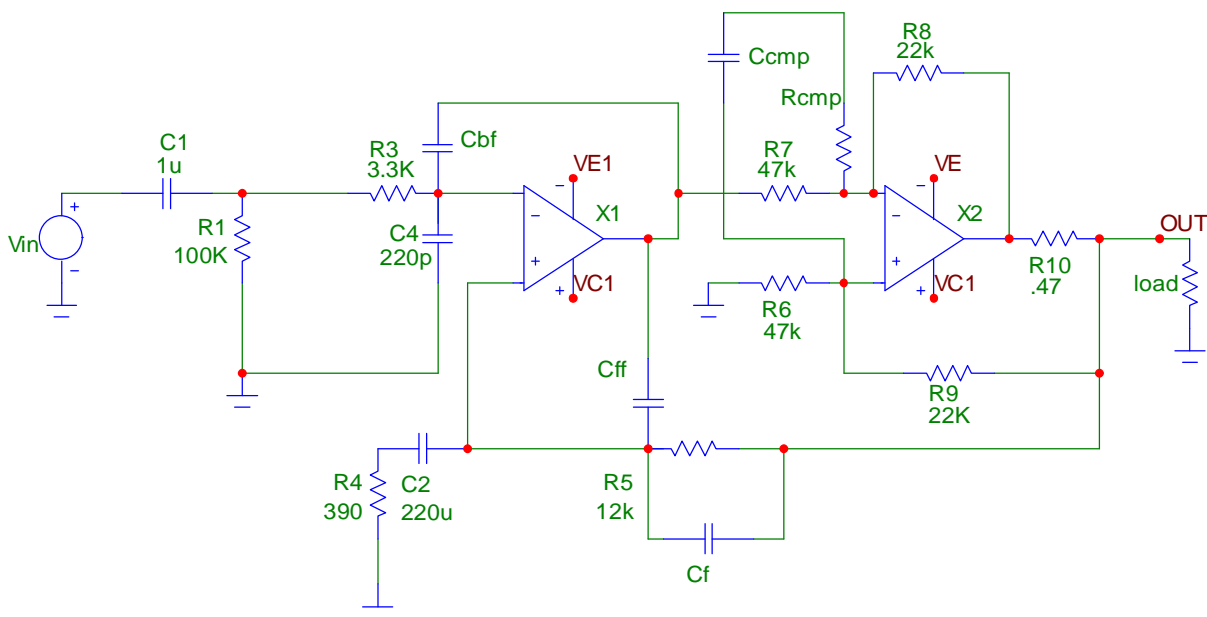


Fig4

Now, the resulting complete circuit is quite simple, does not need too much further explanations. The input impedance is determined by R1, which is fixed in 100 kohm.

R3 /C4 function as input LP filter against high frequency spikes, it's pass band is at about 220kHz – though the value is influenced by the internal poles of the actual compensation network, too. R5 & R4 forms the negative feedback network, and determines the closed loop gain, which is about ~30 dB ((Av

= $1+(R5/R4) \Rightarrow 31$). C2 provides unity gain at DC, so as to reduce the output offset. The low corner frequency of the network C2 /R4 is at around 2Hz. This corner frequency should always be chosen very low, because of the special characteristics of the electrolytics applied here, which give a strongly rising THD in the vicinity of the corner frequency. Another point is the phase linearity. To keep the phase shift low in the audio range (20Hz – 20Khz), it's necessary to apply an intervention point at about 10 times lower, 1-2 Hz (6 degree @20 Hz with a 2Hz pole).

To this phase shift it will be added also that one caused by the subsonic input filter, formed by C1 / R1, which with the given values result to be -3dB at about 1,5Hz.

Altogether we will get ~10degree phase shift at 20 Hz, which is still acceptable.

The combination of these two HP filters results in a 12dB/Oct., 2nd order attenuation below 1,5Hz.

The different networks Cf, Ccmp-Rcmp, Cff and Cbf are there to realize the different type of compensations in frequency & phase, with which I was experimenting during these past months, trying to pinpoint the possible acoustic effects of the different compensations.

General observations:

The open loop gain of this circuit is (or can be) very high, and mostly depends on X1.

This aspect is an essential element concerning the tonal character which discerns this type of circuit, that is, the characteristics of the element with the highest open loop gain (Aol) are predominating, given that the NFB loop works with tensions. The other face of the medal is that when this gain becomes too high, instability problems raise their ugly heads, frequently as a direct consequence of the reactive nature of the load – either capacitive or inductive. Different techniques exist which help to limit these problems, and which I have applied accordingly with the particular compensation type applied. One, quite critical condition is the clipping. In this case in fact, oscillations can be generated, caused by the energy bounced back and forth, coming from the spikes generated either by the LM318 or the LM3886. If & when the compensation network is well tuned, this phenomenon becomes well damped, and so does not create stability problems. So I felt not so convenient to insert further, complex elements which could have “absorbed” the clipping, but would have increased the complexity of the circuit, and could compromise the final quality. On the other hand, I consider **the clipping to be not an operational condition but only an emergency situation**, in which occasion the most important thing is to have a quick and non destructive recovery, neither for the loudspeakers nor for the amplifier itself. For similar reasons I did not put any Zobel circuits on the output. In particular, the current mode output of the power stage, and the associated NFB in tension “sensible” to the load convinced me to reduce all the compensations on the load, so as to help the “feedback” from the cable / loudspeaker system. During start-up and power down conditions there can be instabilities, and they depend on the actual compensation applied. In my circuits I am always inserting an output relay, with the appropriate delay circuit, so as to avoid these “abnormal” conditions. This way we gain a good protection level also against black-out thumps, which can be destructive for the woofers, independently of amp quality..

General Characteristics:

Some of them might change in accordance with the type of compensation, but some other are just common in all the cases:

Bandwith (typical): 2Hz-70Khz

Output power(±37V): 40W/8ohm – 56W/4ohm

Damping Factor (8ohm): >200 (frequency and compensation depended)

S/N 600ohm: >90 dB (compensation depended)

THD e IMD (typical), 20Hz-20Khz, 1-40W/8ohm: <0.05%