

VARIATIONS ON THE COMPLEMENTARY FOLDED CASCODE TRANSIMPEDANCE STAGE IN DISCRETE AUDIO FREQUENCY POWER AMPLIFIERS

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Introduction

Sometime ago Samuel Groner published an intriguing paper [1, 2] in which he suggested that the shortcomings (in respect of slew asymmetry and power supply rejection) of the classical two-stage voltage gain topology (attributed to J. E. Thompson by Messrs Russell and Solomon [3]) may be ameliorated by adopting a complementary push-pull folded cascode arrangement for the second stage.

Groner's arrangement and novel variations thereof are examined here and their performance evaluated against that of the conventional Thompson topology to establish whether the extra complexity is worthwhile.

The Thompson Topology

This circuit (**fig. 1**) contains the fundamental elements of the majority of modern high-performance voltage amplifiers, and thus constitutes an invaluable reference against which the merits of alternative approaches may be judged [4]. A thorough appreciation of its virtues and limitations is therefore essential.

It consists of a transadmittance input stage (TAS for concision) in the form of differential pair **Q13**, **Q14**, with emitter-degeneration resistors, **R2**, **R4**, Widlar's current mirror [5, 6] **Q15**, **Q16** and a so-called tail in the form of *amplified negative feedback* (ANF) current source **Q29/Q30**. The TAS is effectively a voltage-controlled current source (VCCS) whose output current is proportional to the differential input voltage V_d . At frequencies preceding the amplifier's first non-dominant pole, input stage transadmittance gain may, with negligible error, be assumed to be equal to its DC transconductance.

The impedance at **Q14**'s collector is negligible, virtually eliminating Miller feedback through its collector-base intrinsic capacitance. This transistor is effectively an emitter follower and, consequently, its comparatively low net input capacitance permits the connection of large feedback network impedances before the pole at its base becomes significant [7].

Emitter degeneration in the TAS (**R2** and **R4**) constitutes series-applied local negative feedback, which trades a measure of transconductance for enhanced linearity, and is conducive for trouble-free stabilisation without mandating the use of an inordinately large slew-rate sapping Miller feedback capacitor C_C across the second stage. The gain block K_o represents the output stage, which is usually a unity voltage gain complementary symmetry buffer of substantial current gain.

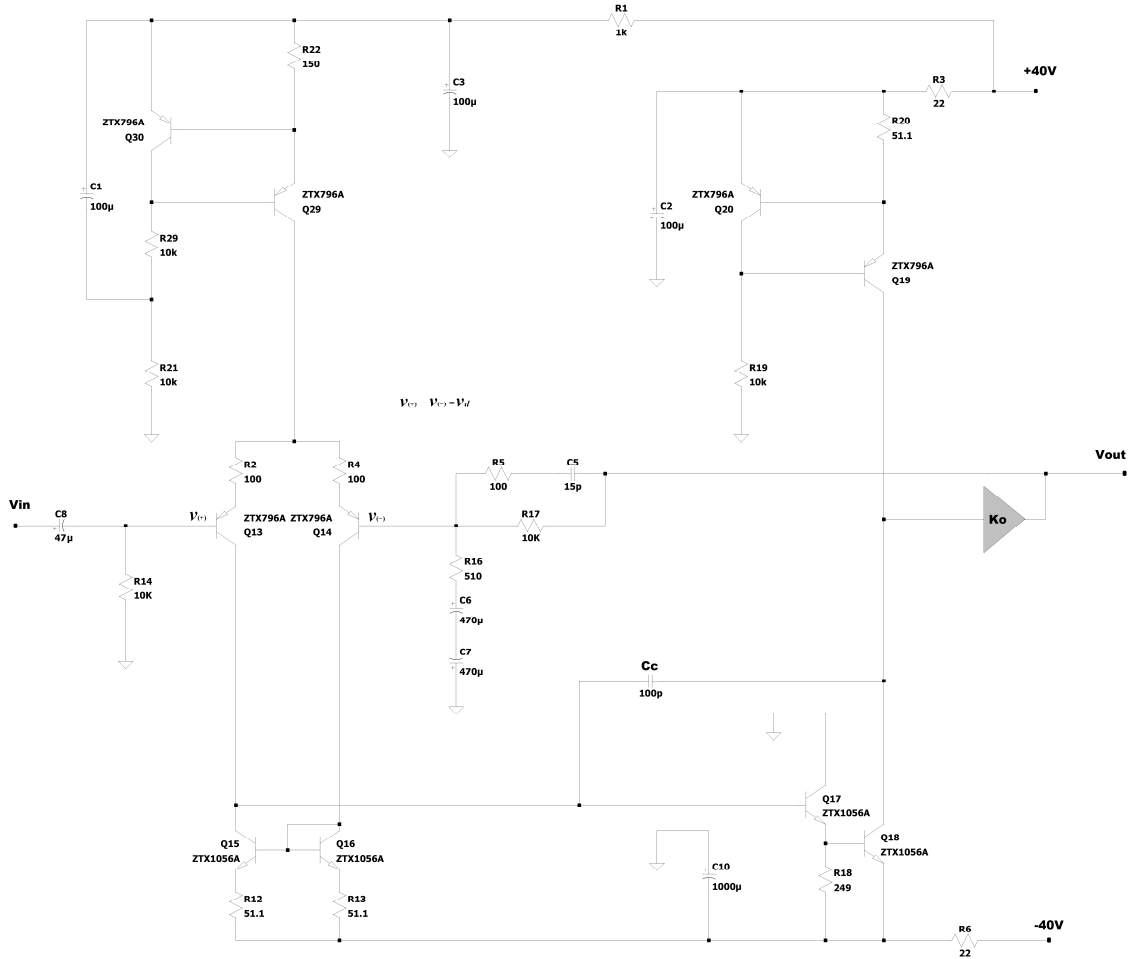


Figure 1. Thompson's two stage topology.

The current mirror facilitates differential-to-single ended conversion and forces equality of collector currents in the differential pair [8]. This minimises DC offset at the output of the amplifier and is a necessary requirement for the elimination of second order distortion generated by the input stage [9].

The mirror also doubles the symmetrical current sourcing and sinking ability of the stage over that obtainable with a resistive load. Degeneration resistors **R12** and **R13** promote equality of currents in the mirror by swamping variations in the base-emitter voltages of its transistors **Q15** and **Q16**.

The ANF active current source **Q29/Q30** vastly improves the common-mode and power supply rail rejection ratios of the stage over those attainable with a simple resistive source. However, a resistive load does not amplify its internal noise, and therefore possess the advantage of producing somewhat less noise than would be generated by the current mirror or active current source.

Capacitor **C1** filters out residual power supply ripple in the bias current established by resistors **R21** and **R29** [10]. The temptation to connect **C7** directly across **Q30** should be resisted as this would simply couple supply ripple directly into the current source.

The second stage, comprising **Q17**, **Q18** and ANF current source **Q19/Q20**, is effectively linearised and converted into a near-ideal transimpedance amplifier stage (TIS) by local shunt (voltage) derived-shunt (current) applied (*viz.* admittance) frequency dependant negative feedback courtesy of the Miller compensation (or stabilising) capacitor C_C .

The TIS is effectively a current-controlled voltage source (CCVS) at the frequencies of interest, and ideally requires infinitely large source and load impedances for maximal transimpedance gain. These conditions are best realised in practice by employing a first-stage current mirror and a high current gain output buffer.

Minor loop negative feedback due to C_C reduces the TIS's input impedance pro rata with increasing frequency, making it negligible (virtually zero) compared to the TAS's output impedance. The local feedback loop also reduces the TIS's output impedance, reducing distortion generated by the non-linear loading of a class-B (or AB) output stage on the second stage [11].

Although the second stage is often [12] referred to as the 'voltage amplifier stage' (VAS), this is technically incorrect, as it implies that the stage is a voltage controlled voltage source (VCVS). In fact, a closed-loop VCVS is synthesised by the application of shunt (voltage) derived-series (voltage) applied negative feedback, which clearly does not obtain with the TIS.

Forward path transimpedance gain local to the second stage is, to a good first approximation, merely the product of its current gain and the effective impedance at its output (see **appendix**). Thus emitter follower **Q17** increases forward-path gain local to the TIS (and therefore minor loop feedback through C_C) by increasing the second stage's effective current gain.

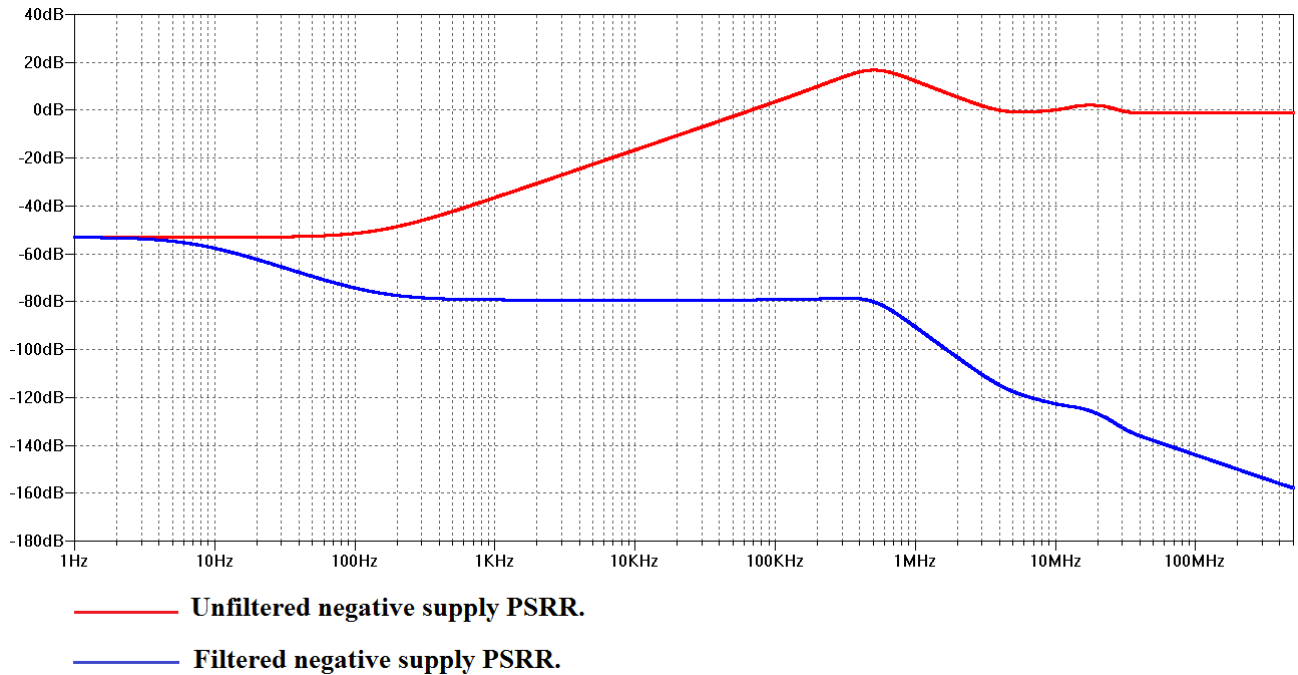


Figure 2. Simulated negative rail PSRR of Thompson's two stage topology of fig. 1.

With this topology and with the polarity of the TIS in **figure 1**, the limiting power supply rejection ratio (PSRR) is defined by the negative supply rail. This is because the TIS, whose low impedance emitter sits on the negative supply rail, couples supply rail ripple to the output of the TIS at the same single pole rate as major loop transmission falls with increasing frequency. This gives a PSRR of only 53dB at ripple frequency (100Hz) which deteriorates dramatically thereafter at a single pole rate to 0dB at ultrasonic frequencies (**fig. 2**).

Power supply rejection was established in LTspice by inserting an AC voltage source in series with the DC power supply and merely plotting the AC frequency response at the output of the amplifier. An ideal unity gain voltage controlled voltage source was used to model the output stage.

The TIS is not as vulnerable to the introduction of ripple from the positive supply because its output is shielded by ANF current source **Q19/Q2**. The ANF current source possesses a high output impedance due in part to the fact that its loop transmission exceeds 40dB to at least 10KHz; the beneficial effect of this is evident in simulation by supplying the TAS with an ideal current source with the TIS's ANF current source in situ: positive supply rejection is found to exceed 80dB across the audio band.

The RC filters **R1/C3** and **R3/C2** in the positive supply rail and **R6/C10** in the negative supply rail not only guarantee a minimum power supply rejection of about 75dB and 95dB for the negative and positive supplies respectively at ripple frequency (100Hz), but they also vastly increase the PSRR at ultrasonic frequencies. This desirable because it significantly attenuates any ultrasonic voltage, such as radio-frequency interference (RFI), on the supply rails which may otherwise drive the amplifier into slew limiting. In the context of this paper, however, it is the unfiltered negative supply PSRR of Thompson's topology (**fig. 2**) that is of interest as it is this that Groner's arrangement seeks to improve.

Another mostly aesthetic inadequacy of the Thompson arrangement that Groner's circuit seeks to eliminate is its slew asymmetry. With the polarity of TIS in **figure 1** positive slew rate is appreciably smaller than the negative slew rate. This deficiency in slew rate for positive voltage swings occurs because a significant amount of current that would otherwise service the compensation capacitor C_C is instead siphoned off by parasitic shunt capacitance to ground at the TIS's output [13, 14]. No such impediment exists for negative slew as the second stage transistor **Q18** is capable of sinking as much current as the TAS can supply through the compensation capacitor.

Groner's Complementary Folded Cascode TIS

Groner sought to improve the mediocre negative PSRR of the Thompson topology by anchoring the TIS to ground by means of a pair of complementary folded cascodes **Q1/Q5** and **Q3/Q6** (**fig. 3**). Emitter followers **Q2/Q4** bias the common emitter transistors **Q1/Q3** and ensure that the input of the TIS is roughly at ground potential. The TAS is, of necessity, an arrangement of differential-input folded cascodes whose current output is level shifted to the potential at the TIS's input.

The inclusion of emitter followers **Q2/Q4** means that there are three transistors in series in the TIS's forward path; this makes the minor (compensation) loop unstable. Consequently, feed-forward capacitors **C9/C10** are required to improve minor loop stability margins by shunting the emitter followers **Q2/Q4** out of the forward path at ultrasonic frequencies.

In general including more than two transistors in the minor loop makes it unstable; compensation of the minor loop is then necessary, typically by means of a small (47pF~1nF) shunt capacitance to ground at the TIS's output. This is not a good idea with the Thompson topology as it merely exacerbates slew asymmetry, but shouldn't be a problem with Groner's arrangement as the push-pull TIS should, in principal, provide all the current required to drive the shunt capacitance.

Alas, the positive and negative rail power supply rejection of this topology (**fig. 4**) is not a significant improvement on the mediocre unfiltered negative rail performance of the Thompson arrangement. For both negative and positive power supply rails, PSRR is roughly 50dB sustained to about 2KHz whereupon it rapidly deteriorates.

Groner maintains that using a regulated power supply for the TAS and emitter followers **Q2/Q4** should improve power supply rejection dramatically. This, taken at face value, seems like an elegant solution not least because, as Groner observes, the TAS and emitter followers **Q2/Q4** need only be energised from lower regulated voltage rails directly derived from the supplies connected to the complementary folded cascodes and the output stage. The amplifier's power supply rejection would then effectively be defined only by the complementary folded cascodes.

This notion was tested in LTspice by merely connecting a pair of ideal DC voltage supplies, independent of those powering the complementary folded cascodes, to the TAS and emitter followers **Q2/Q4**. A copy of the circuit was made and an AC voltage source connected in series with the positive DC voltage source energising the complementary folded cascodes in the first circuit; a second AC voltage source was then connected in series with the negative DC voltage source powering the complementary folded cascodes in the second circuit. The frequency response (effectively the power supply rejection) with respect to the output of each circuit was then obtained (**fig. 5**).

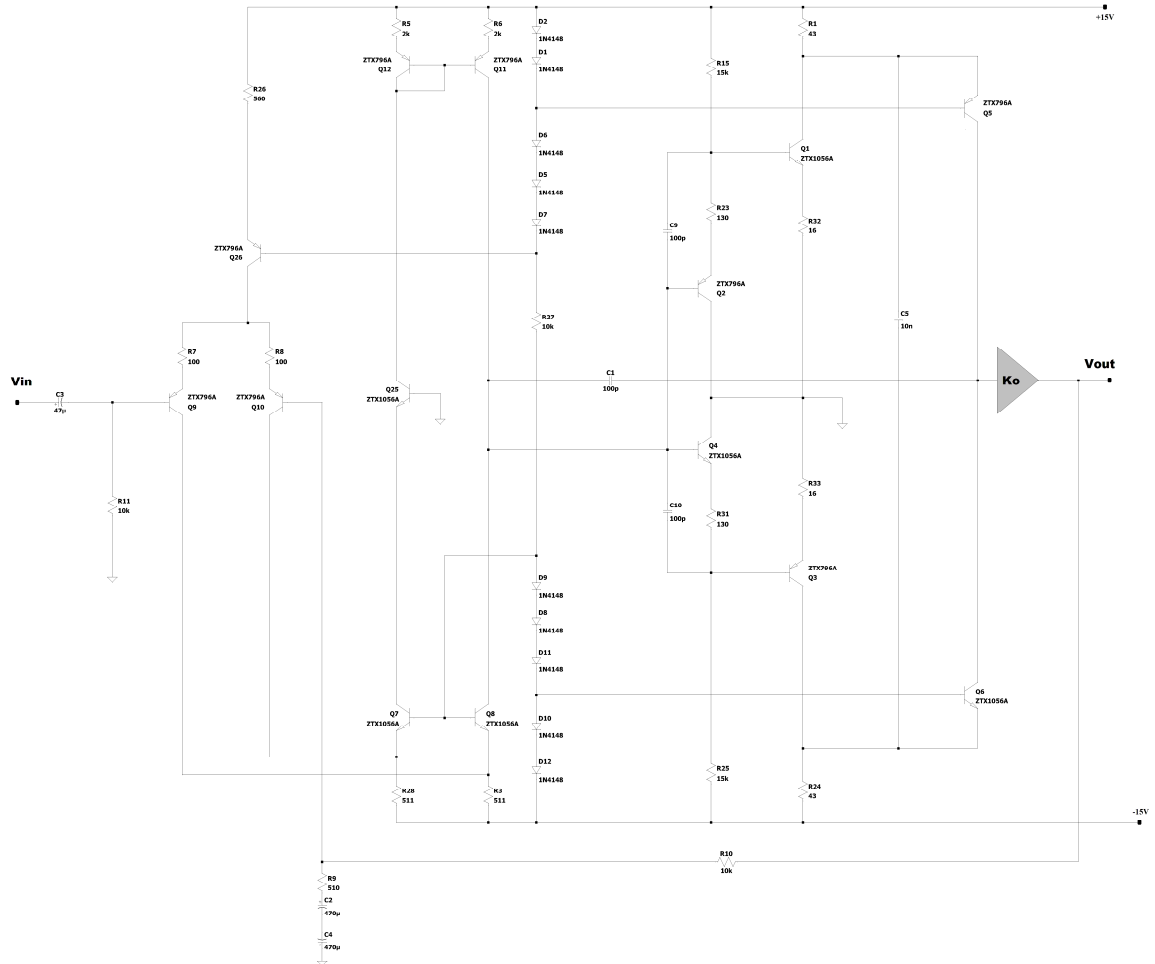


Figure 3. Groner's amplifier with complementary folded cascode TIS.

Regrettably, the improvement in power supply rejection is a barely perceptible 4dB with respect to the positive supply at ripple frequency. There was virtually no change detected in negative supply PSRR.

These results demonstrate that the complementary folded cascodes are the weakest link in the design as far as power supply rejection is concerned. This is because the low impedance emitters of common base stages **Q5/Q6** of the folded cascodes are all but connected directly to the supply rails, since the values of their emitter resistors **R1** and **R24** are negligible. Supply rail ripple is, therefore, coupled virtually unchecked through the emitters of **Q5/Q6** to the output of the TIS, and explains why the positive and negative power supply rejection of Groner's arrangement is just as poor as that due to the unfiltered negative supply of the Thompson topology of **figure 1**.

That the emitters of common base transistors **Q5/Q6** are the means by which supply rail ripple passes to the output of the TIS was demonstrated by replacing emitter resistor **R24** with an ANF current source while retaining the ideal power supplies for the input stage and the emitter followers **Q2/Q4**. Power supply rejection with respect to the negative supply rail powering the negative polarity folded cascode was then found to exceed 100dB to approximately 1KHz whereupon it deteriorates as the current source's parasitic capacitances take effect and its loop transmission, as well as that of the amplifier's major loop, declines (**fig. 6**).

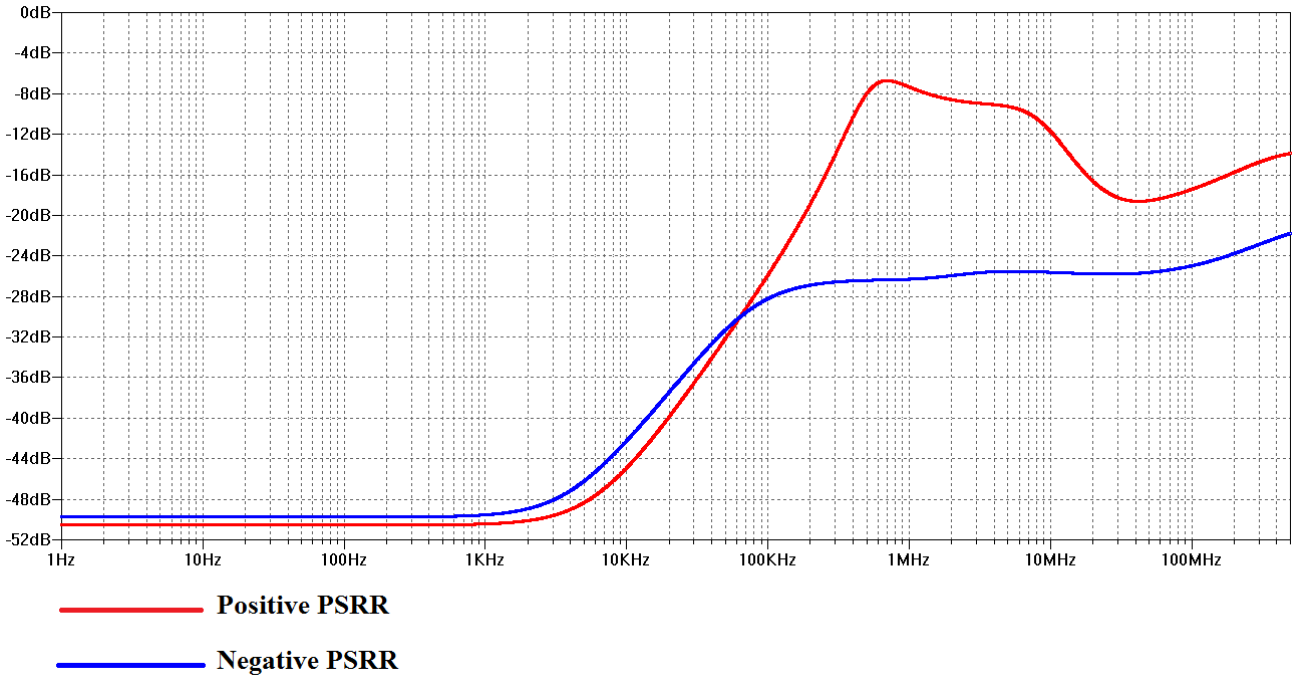


Figure 4. Simulated PSRR of Groner's amplifier.

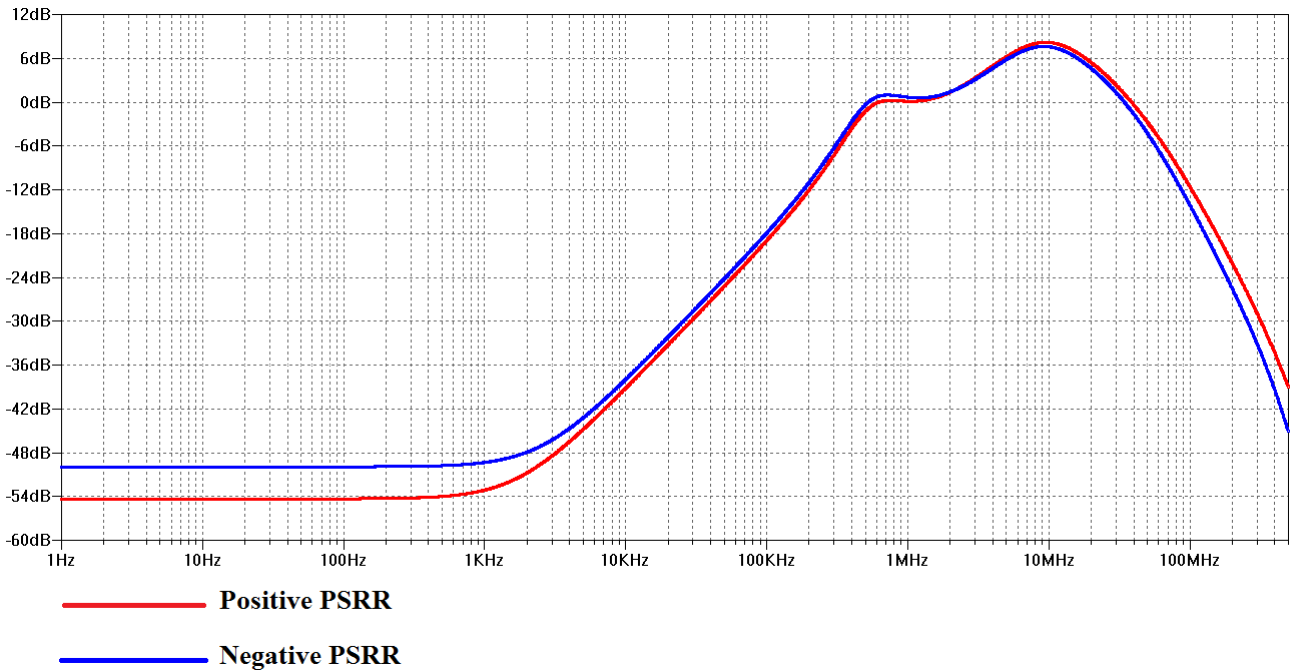


Figure 5. Simulated PSRR of Groner's amplifier with ideal DC voltage supplies powering the input stage and emitter followers isolated from those energising the complementary folded cascodes.

Clearly, as is the case with the Thompson arrangement, Groner's design should benefit from simple RC filtering of the negative and positive power supplies to the TIS in addition to ANF current source biasing of the folded cascodes. Note that although there is another potentially major entry point for negative supply ripple through the emitters of the input stage's common base transistors **Q7/Q8**, there is virtually nothing to be gained in PSRR terms from using regulated supplies for the input stage and emitter followers **Q2/Q4** unless the ingress of ripple through the TIS folded cascodes is attended to by replacing resistors **R1/R24** with ANF current sources.

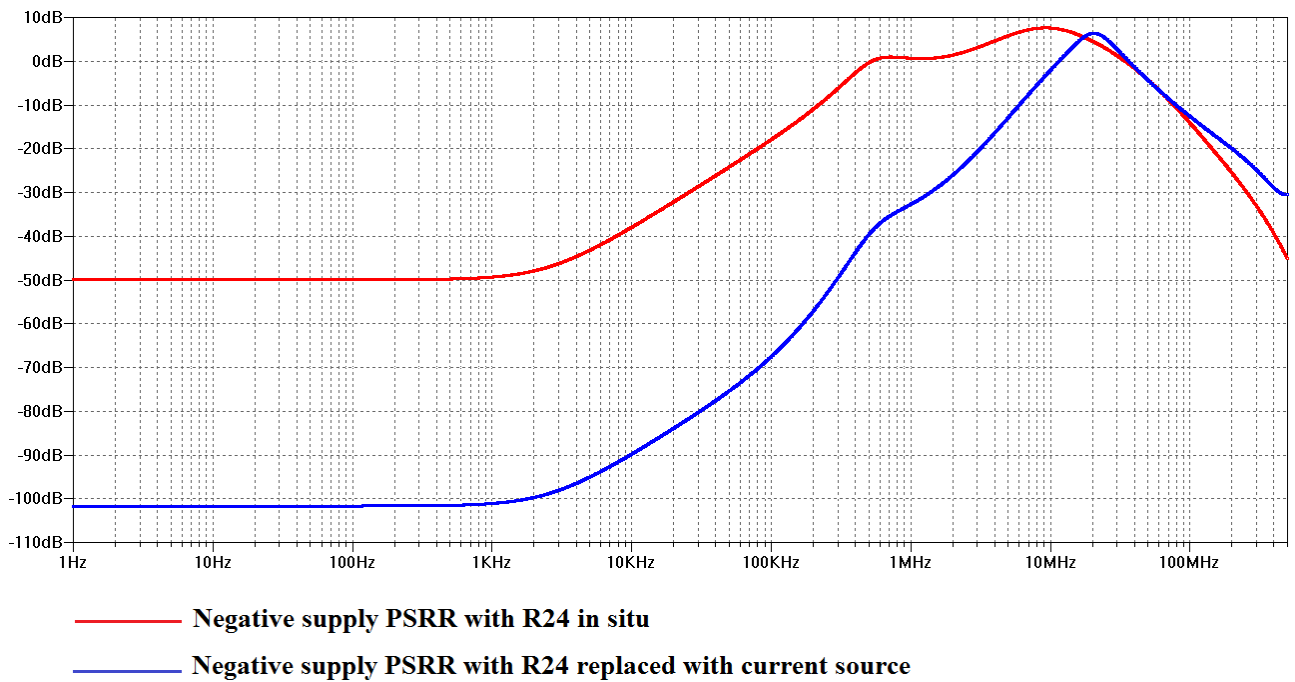


Figure 6. Simulated negative supply PSRR of Groner's amplifier with R4 in situ and with it replaced with an ANF current source. Ideal DC voltage supplies were used for the input stage and emitter followers Q2/Q4 independent of those energising the complementary folded cascodes.

Since resistor **R1** provides a fixed current, push-pull action occurs with Groner's TIS because an increase in the current demanded of common base transistor **Q5** can only be met by an equivalent decrease in the collector current of common emitter transistor **Q1**. Similarly, the required reduction in the collector current of **Q6** is necessarily accompanied by an equivalent increase in the collector current of **Q3**. Therefore, the peak AC current swing available at the output of the TIS is equal to the quiescent current in each of the common emitter transistors **Q1/Q3**.

Consequently, the standing collector current in common base transistors **Q5/Q6** has to be greater than that of the common emitter transistors **Q1/Q3** if the common base transistors are not to be alternately cut off at the limit of their current excursions. It is therefore essential to set the quiescent currents in the common emitter and common base sections of the complementary folded cascodes to the desired relative amounts accurately and precisely.

Resistors **R23**, **R32**, **R31** and **R33** are essential for establishing the quiescent collector currents of common emitter transistors **Q1/Q3**. Without these resistors the standing currents in these transistors would be undefined, and, accordingly, setting the quiescent current in common base transistors **Q5/Q6** would be impossible.

The problem is inserting **R23/R31** in series with the emitters of emitter followers **Q2/Q4** reduces their quiescent current; resistors **R15/R25** need to be reduced by an amount equal to the values of **R23/R31** to maintain a constant current in each of the emitter followers.

This is necessary to accurately establish the voltages across **R23/R31** as it is these voltages which appear across **R32/R33** and, therefore set the quiescent currents in common emitter transistors **Q1/Q3**. However, because resistors are available only in discrete preferred values, maintaining the desired invariant standing currents in emitter followers **Q2/Q4** when **R23/R31** are introduced is a

forlorn hope. Consequently, setting the quiescent collector currents of common emitter transistors **Q1/Q3** with accuracy and precision would be tedious and difficult.

Indeed, Groner acknowledges that the values of the resistors biasing common emitter transistors **Q1/Q3** would have to be determined experimentally. This is inelegant and unwelcome, especially if large scale industrial production is envisaged.

Current source biasing of the emitter followers **Q2/Q4** is, therefore, essential to establish with accuracy the voltages across resistors **R23/R31**. Thus, with the standing currents in the common emitter and common base elements of the complementary folded cascodes accurately set to their optimal relative levels, push-pull action does, in fact, occur in the TIS, and slew asymmetry which is the only putative disadvantage of the Thompson arrangement of **figure 1** is abolished.

Incidentally, Groner contends that capacitor **C5** dramatically enhances the push-pull action of the TIS by causing it to operate in Class AB. That this is not true is established by noting that **C5** is connected to the emitters of the common base transistors in the complementary folded cascodes. These emitters are firmly anchored by their respective voltage references, and, as a result, capacitor **C5** cannot in any way vary the collector currents of the common base transistors.

The Differential Complementary Folded Cascode TIS

When it became clear that the emitter followers **Q2/Q4** in Groner's TIS required current source biasing, it was then apparent that the increased component count could be put to use more elegantly and efficiently by converting the emitter followers to the grounded base configuration instead. The grounded base transistors are then emitter-coupled to the common emitter transistors of the complementary folded cascodes, and the current sources used to bias each emitter-coupled pair (**fig. 7**).

The feed-forward capacitors required to stabilise the minor (compensation) loop in Groner's arrangement are now redundant as there are effectively only two transistors in series in the forward path of the TIS; the grounded base transistors **Q2/Q4** (**fig. 7**) only serve the purpose of anchoring the TIS to ground and do not appear in the signal path. Consequently the stability margins of the minor loop with this arrangement are more than adequate and are of the same order as those of the buffered TIS in the Thompson topology of **figure 1**.

The power supply rejection of this arrangement is a little better than that of Groner's circuit principally because the emitters of common base transistors **Q5/Q6** in the TIS are connected to their respective supply rails through low value resistors **R1/R24**. Replacing these resistors with ANF current sources is mandatory if good power supply rejection (sustained to at least 1KHz) is to be obtained. When this is done, the secondary negative supply ripple entry route through the emitters of common base transistors **Q10/Q15** in the input stage may be attended to by either replacing resistors **R6/R7** with ANF current sources, or using smaller regulated supplies for the input stage derived from those of the TIS and the output stage.

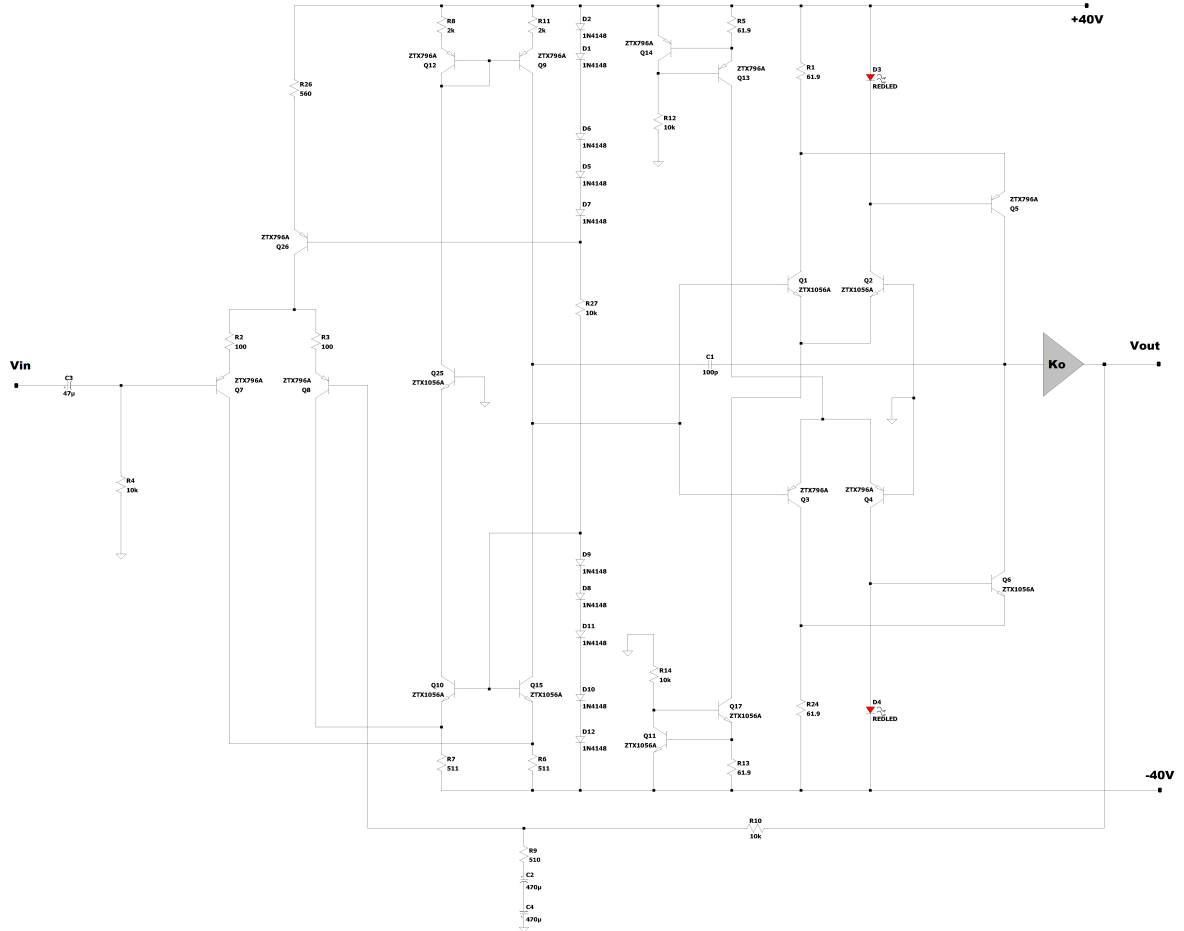


Figure 7. Amplifier with differential complementary folded cascode TIS.

The Differential Current Boosted Complementary Folded Cascode TIS

Current source biasing of the TIS's complementary folded cascodes can be more elegantly accomplished by using current mirrors (**fig. 8**). Quiescent current in the complementary common base transistors **Q5/Q6** is established by cross-coupling resistor **R4**. Contrary to intuition replacing this resistor with an active current source does not significantly improve power supply rejection.

The buffered current mirrors double the peak push-pull output current for a given input compared to Groner's circuit (**fig. 3**) and the circuit of **figure 7**. The current mirrors also double the forward path gain of the TIS compared to that of the differential complementary folded cascode arrangement of **figure 7**. However, the increased number of transistors in the minor loop means that minor loop stability margins (in the absence of compensation elements **R12**, **R13**, **R14**, **R15**, **C5** and **C6**) are significantly reduced compared to those of the buffered TIS in the Thompson arrangement of **figure 1**, but are still more than adequate, with a phase margin of the order of forty-five degrees.

If even greater minor loop stability margins are required, then the TIS can be straightforwardly degenerated with resistors **R12**, **R13**, **R14** and **R15** to reduce minor loop transmission (**fig. 8**). The small emitter-coupling capacitors, **C5** and **C6**, then introduce a zero in the vicinity of the unity minor loop gain frequency, further enhancing minor loop stability margins; minor loop phase margin increases to roughly seventy-five degrees, which is of the same order as that of the buffered TIS in the Thompson circuit.

The differential current boosted complementary folded cascode topology was apparently invented by Shinichi Kamiyo [15] who used it as a single voltage gain stage in his amplifiers. However, it would appear this is the first time its use as a transimpedance stage (TIS) has been published.

Indeed, the differential current boosted complementary folded cascode circuit can also be used as the input stage (TAS) where it gives twice the transconductance of the ordinary differential pair with a current mirror used in the Thompson topology (**fig. 9**). Thus, if the compensation capacitor remains the same and the unity loop gain frequency is to remain unchanged, the degeneration resistors in the input stage of **figure 9** have to be doubled with respect to those used in the input stage of the Thompson circuit and Groner's arrangement.

In contrast to the simple differential pair with a current mirror used in the Thompson circuit, matching of differential pair collector currents in the input stage of **figure 9** is not guaranteed. However, even order distortion generated as a result is cancelled at the stage's output where push-pull action occurs, and the quiescent collector currents of the complementary common base transistors **Q11/Q12** are, for practical purposes, perfectly matched. Note that due to the slight imbalance in differential pair standing currents the DC offset at the output of the amplifier is likely to be somewhat worse than that of the Thompson circuit.

The power supply rejection of the novel circuits of **figure 8** and **figure 9** was found to surpass that of the Thompson arrangement with unfiltered supplies at ripple frequency only if regulated supplies for the input stage are used. The regulated supplies need only consist of a zener diode and a resistor, as Groner pointed out.

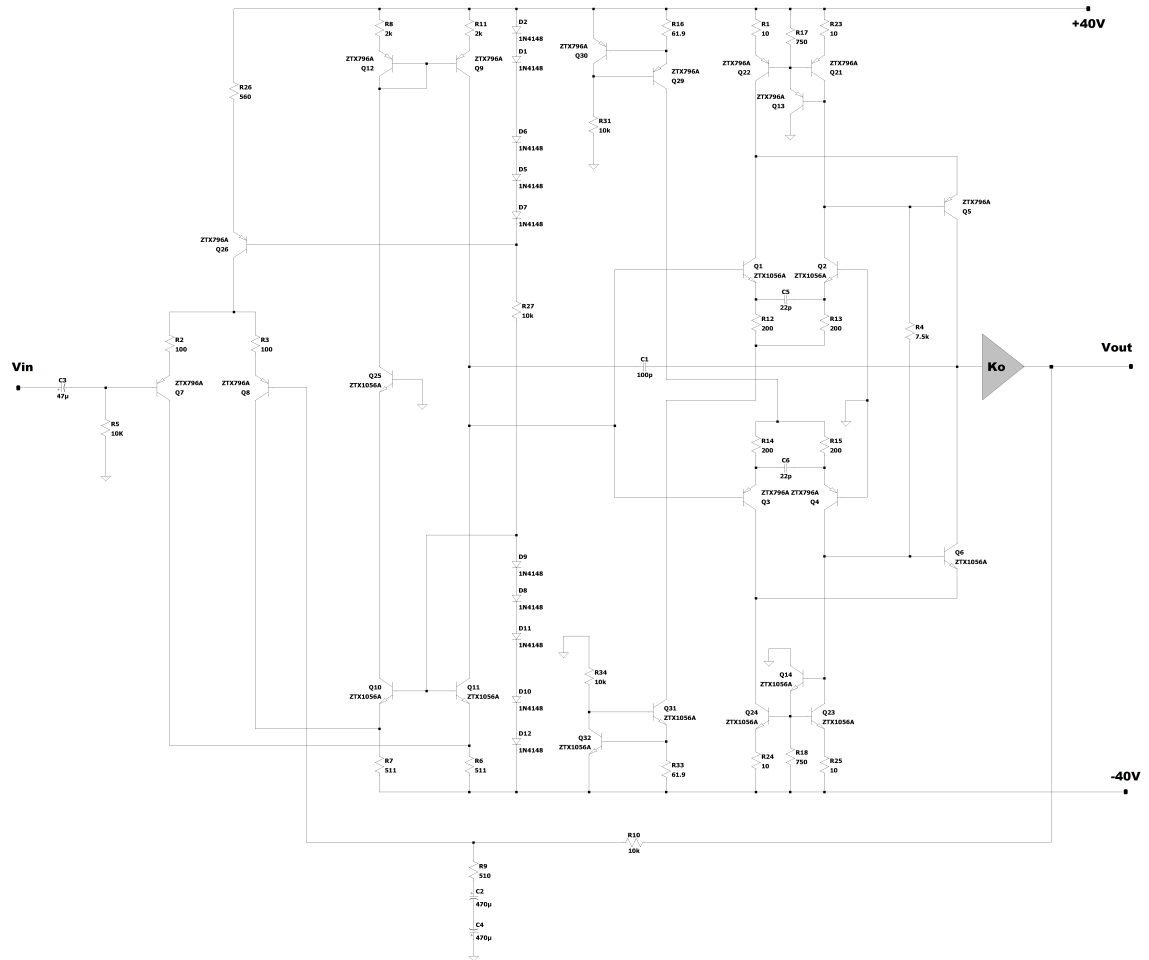


Figure 8. Amplifier with differential current boosted complementary folded cascode TIS.



Conclusion

Groner's circuit and the novel variations thereof achieve somewhat better power supply rejection at ripple frequency than the Thompson arrangement with unfiltered power supplies provided all circuit elements in the TIS are biased with active current sources and regulated power supplies are used for the input stage. However, this is achieved at more than twice the cost of Thompson's arrangement, and the increase in circuit complexity is daunting. Moreover, to extend good power supply rejection to ultrasonic frequencies, simple RC filtering would still have to be used with Groner's circuit and the novel variations discussed here.

Nevertheless, push-pull action in the complementary folded cascode TIS of Groner's circuit and its variants does virtually guarantee symmetrical slew; slew rate can simply be increased by increasing the size of the degeneration resistors in the input stage so that the compensation capacitor can be reduced for the same unity loop gain frequency. This is achieved without running into the Thompson circuit's slew rate brick wall defined by its second stage's inability to supply the requisite current to the compensation capacitor. However, this is not a significant advantage over the Thompson arrangement whose limiting slew rate is more than adequate for the majority of medium power domestic applications.

Appendix

Miller Compensation: A First-order Analysis

The generic two-stage voltage gain block with minor-loop compensation (**fig. 1**) is modelled in **figure A1** by a differential voltage controlled current source (VCCS) driving a TIS consisting of a current controlled current source (CCCS) and load resistor R_{eq} , which is the means by which the TIS's output current is expressed as a voltage.

Resistor R_{eq} represents the modulus of the effective impedance at the output of the TIS, and comprises the parallel combination of the TIS's output impedance and the output buffer's input impedance. The TIS's current gain β_{eq} is merely the product of the current gains of transistors **Q17** and **Q18**.

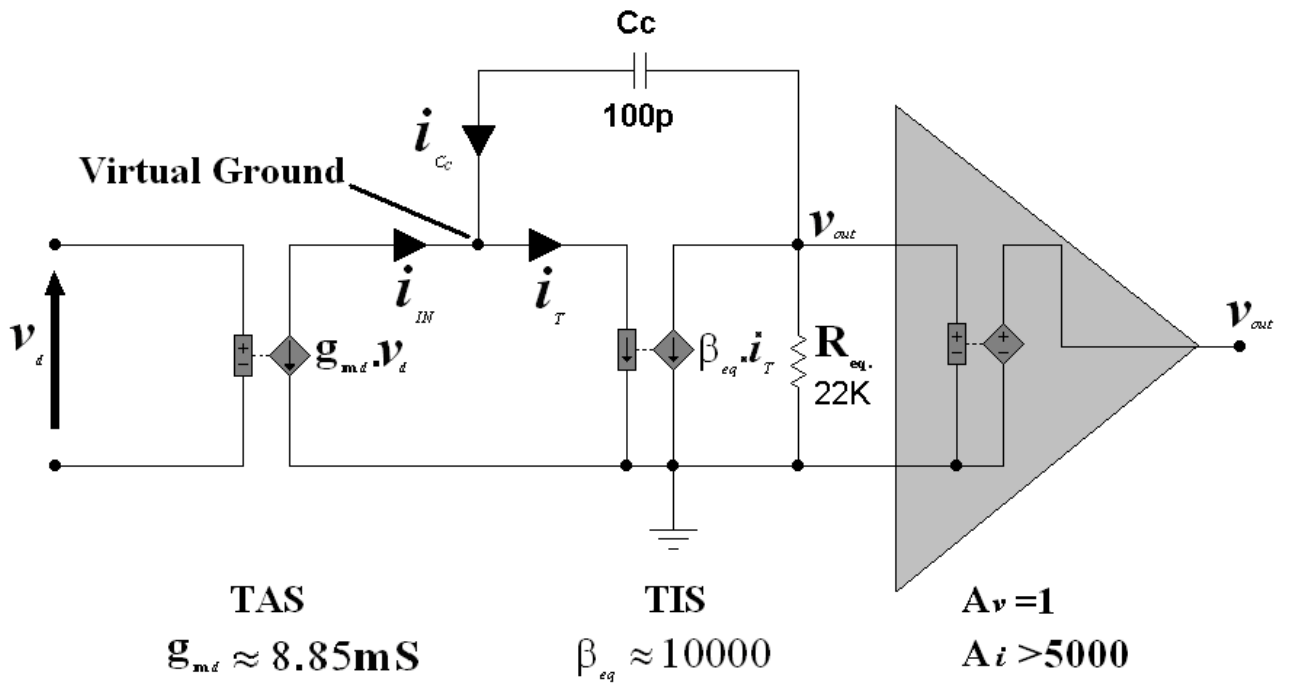


Figure A1. First-order model of the single-pole compensated voltage gain block.

It is assumed here that the minor feedback loop defined by C_c is stable, and that the amplifier's unity-gain frequency f_U is sufficiently low so that non-dominant poles have negligible effect on its open-loop transfer function.

At DC the TIS, comprising **Q17** and **Q18** in **figure 1**, possesses a low input resistance compared to the TAS's output resistance. As local feedback through C_c increases beyond the dominant pole frequency, the TIS's input impedance rapidly tends to zero; the TIS's input is then virtually at ground potential, and the entire output voltage may be deemed to appear across C_c .

Invoking Kirchoff's current Law with respect to the output node (**fig. A1**)

$$-i_{C_c} - \beta_{eq} i_T + \frac{(0 - v_{out})}{R_{eq}} = 0$$

$$\Rightarrow$$

$$i_{C_c} + \beta_{eq} i_T + \frac{v_{out}}{R_{eq}} = 0 \quad (1a)$$

Similarly at the input node

$$i_{in} + i_{C_c} - i_T = 0 \quad (2a)$$

Since shunt-applied negative feedback makes the TIS's input node a virtual ground at the frequencies of interest, then

$$i_{C_c} = sC_c v_{out} \quad (3a)$$

Substituting (3a) into (1a)

$$sC_c v_{out} + \beta_{eq} i_T + \frac{v_{out}}{R_{eq}} = 0 \quad (4a)$$

Substituting (3a) into (2a)

$$i_{in} + sC_c v_{out} - i_T = 0 \quad (5a)$$

Equation (5a) is multiplied by β_{eq} as a prelude to eliminating i_T :

$$i_{in} \beta_{eq} + sC_c v_{out} \beta_{eq} - \beta_{eq} i_T = 0 \quad (6a)$$

Thus, adding equation (4a) to (6a) eliminates i_T :

$$i_{in} \beta_{eq} + sC_c v_{out} \beta_{eq} + sC_c v_{out} + \frac{v_{out}}{R_{eq}} = 0 \quad (7a)$$

\Rightarrow

$$\frac{v_{out}}{i_{in}}(s) = -\frac{\beta_{eq} R_{eq}}{1 + sC_c R_{eq} (\beta_{eq} + 1)}$$

Since $(\beta_{eq} \gg 1)$, then it may be assumed with negligible error that $(\beta_{eq} + 1) \approx \beta_{eq}$, and

$$\frac{v_{out}}{i_{in}}(s) \approx \frac{-\beta_{eq} R_{eq}}{1 + sC_c R_{eq} \beta_{eq}} \quad (8a)$$

First stage transconductance g_{md} in **figure 1** (with degeneration resistors R_2 and R_4) is given by $g_{md} \approx (r_e + R_2)^{-1}$.

The intrinsic emitter resistance r_e in each TAS transistor is merely the reciprocal of the stage's undegenerated transconductance g_{mo} , viz. $r_e \approx 1/g_{mo}$, and $g_{mo} \approx qI_C/KT = I_C/V_T$.

Where K is Boltzmann's constant ($\sim 1.38 \times 10^{-23}$ joules/Kelvin), T the absolute temperature, (Kelvin), q the electronic charge ($\sim 1.6 \times 10^{-19}$ coulomb) and V_T the thermal voltage (~ 26 mV at room temperature).

Thus, at room temperature and with the component values in **figure 1**, $g_{md} \approx [(38.5 \times 2\text{mA})^{-1} + 100\Omega]^{-1} \approx 8.85\text{mS}$.

But

$$i_{IN} = -g_{md}v_d$$

Thus, the amplifier's forward path gain is given by

$$\frac{v_{out}}{v_d}(s) \approx \frac{g_{md}\beta_{eq}R_{eq}}{1 + sC_C\beta_{eq}R_{eq}} \quad (9a)$$

or

$$\frac{v_{out}}{v_d}(s) \approx K \cdot \frac{1}{1 + sC_C\beta_{eq}R_{eq}} \quad (10a)$$

Where K is the forward path gain at DC:

$$K = g_{md}\beta_{eq}R_{eq} \quad (11a)$$

From (10a) the dominant pole frequency f_D is given by

$$f_D \approx \frac{1}{2\pi C_C\beta_{eq}R_{eq}} \quad (12a)$$

Unity-gain frequency f_U is obtained by merely equating (9a) to unity:

$$1 = \frac{g_{md}\beta_{eq}R_{eq}}{1 + \omega_U C_C\beta_{eq}R_{eq}} \quad (13a)$$

\Rightarrow

$$f_U \approx \frac{(g_{md}\beta_{eq}R_{eq} - 1)}{2\pi C_C \beta_{eq} R_{eq}} \quad (14a)$$

In practice, only forward path gain well beyond the dominant pole frequency is of interest and, with respect to **equation (9a)**, the condition $(\beta_{eq}R_{eq} \rightarrow \infty)$ is invoked, so that

$$\left. \frac{v_{out}}{v_d}(s) \right|_{(10f_D \leq f \leq f_U)} \approx \frac{g_{md}}{sC_C} \quad (15a)$$

Equation (14a) becomes

$$f_U \approx \frac{g_{md}}{2\pi C_C} \quad (16a)$$

Equation (15a) is valid only at frequencies well beyond the dominant pole. This is demonstrated by the plot of **figure A2** using typical values (**fig. A1**); the finite gain of the TIS introduces significant error at DC and infrasonic frequencies.

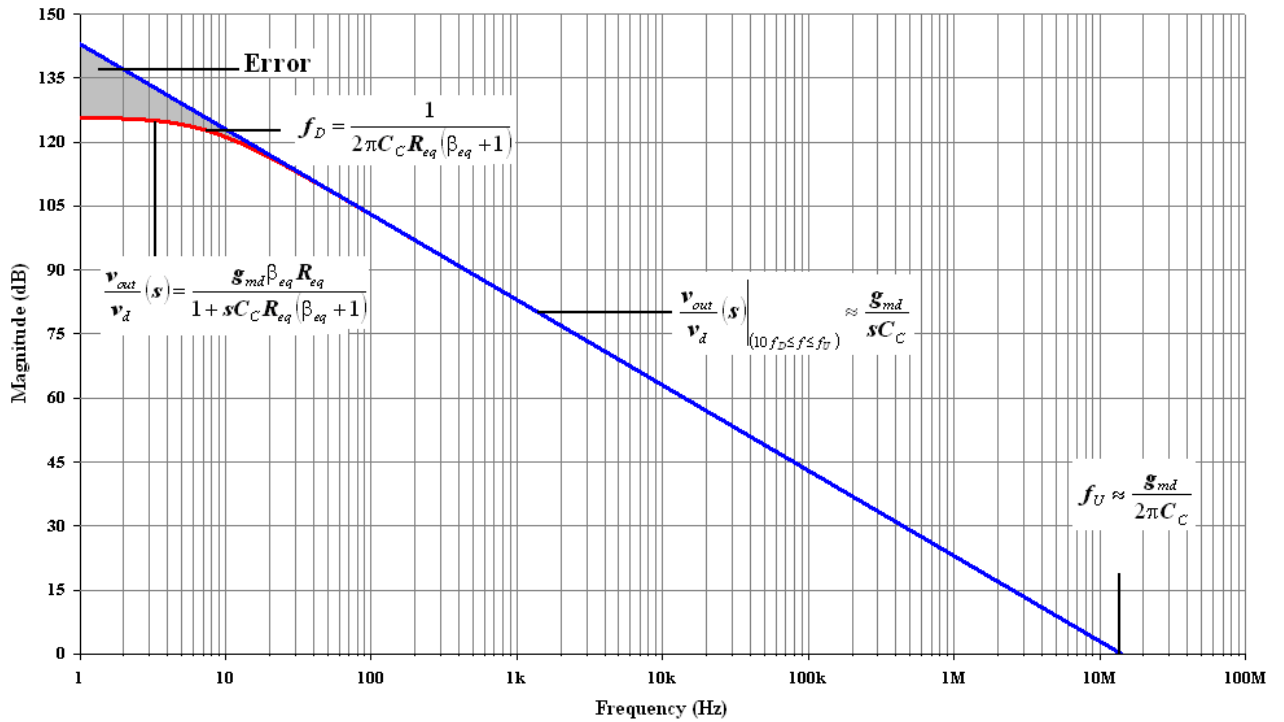


Figure A2. The simplification $(\beta_{eq}R_{eq} \rightarrow \infty)$ gives negligible error in the forward-path transfer function at the frequencies of interest.

Pole Splitting

The presence of the first non-dominant pole may be accommodated by performing a second order analysis in which the TIS is more accurately modelled by a voltage controlled current source (VCCS) with finite input and output shunt impedances [4, 16], which give rise to two dominant poles (**fig. A3**). Capacitors C_{eq1} and C_{eq2} represent the equivalent shunt capacitance at the input and output nodes of the TIS, while the effective shunt resistance is represented by R_{eq1} and R_{eq2} respectively.

Tedious but rudimentary nodal analysis at the input and output of the TIS demonstrates that single-pole feedback compensation causes the first two dominant system poles to move apart, while the finite input voltage v_i generates a so-called feedforward current i_f through C_C . Ultimately, the forward current gives rise to a non-minimum phase (RHP) zero when C_C short-circuits the TIS's load, $R_{eq2} // 1/sC_{eq2}$, so that $i_f = g_{m1} \cdot v_i$ and $v_{out} = 0$.

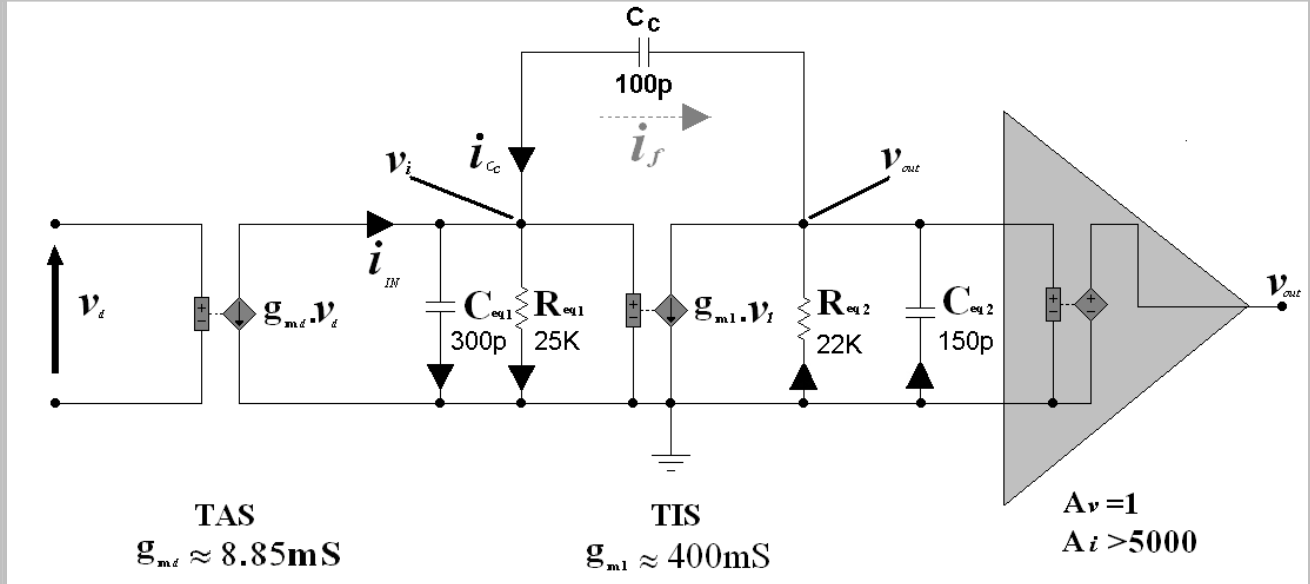


Figure A3. Second-order model of the single-pole compensated generic voltage gain block.

At the TIS's input node

$$-g_{md} \cdot v_d - v_i \cdot sC_C + v_{out} \cdot sC_C - v_i \cdot sC_{eq1} - v_i / R_{eq1} = 0 \quad (17a)$$

And at the TIS's output node

$$v_i \cdot sC_C - v_{out} \cdot sC_C - g_{m1} \cdot v_i - v_{out} \cdot sC_{eq2} - v_{out} / R_{eq2} = 0 \quad (18a)$$

Solving (18a) for v_i and substituting the result into (17a) eliminates v_i :

$$\frac{v_{out}}{v_d}(s) = \frac{g_{md} g_{m1} R_{eq1} R_{eq2} (1 - sC_C / g_{m1})}{s^2 R_{eq1} R_{eq2} (C_C C_{eq2} + C_C C_{eq1} + C_{eq1} C_{eq2}) + s \{ R_{eq1} (C_C + C_{eq1}) + R_{eq2} (C_C + C_{eq2}) + g_{m1} C_C R_{eq1} R_{eq2} \} + 1} \quad (19a)$$

or

$$\frac{v_{out}(s)}{v_d} = \frac{K(1 - sC_C/g_{m1})}{s^2 R_{eq1} R_{eq2} (C_C C_{eq2} + C_C C_{eq1} + C_{eq1} C_{eq2}) + s \{ R_{eq1} (C_C + C_{eq1}) + R_{eq2} (C_C + C_{eq2}) + g_{m1} C_C R_{eq1} R_{eq2} \} + 1} \quad (20a)$$

Where K is the forward path gain at DC:

$$K = g_{md} g_{m1} R_{eq1} R_{eq2} \quad (21a)$$

From **equation (20a)**

$$P_1|_{C_C=0} = -1/(R_{eq1} C_{eq1}) \text{ and } P_2|_{C_C=0} = -1/(R_{eq2} C_{eq2}).$$

The denominator in **(19a)** is a second order polynomial and provided $C_C \gg C_{eq1} \vee C_{eq2}$, the local loop is stable and assuming the poles are real it may be expressed as the product of two first-order factors:

$$\text{Denominator}(s) = (1 + s/P_1)(1 + s/P_2) = 1 + s(1/P_1 + 1/P_2) + s^2/(P_1 P_2) \quad (22a)$$

By merely equating coefficients it is apparent that

$$P_1 \approx -1/(g_{m1} R_{eq1} R_{eq2} C_C), \quad P_2 \approx -g_{m1} C_C / (C_{eq1} C_{eq2} + C_{eq1} C_C + C_{eq2} C_C) \text{ and } Z = g_{m1} / C_C.$$

Where P_1 is the dominant pole, P_2 the first non-dominant pole and Z the right half plane zero.

Therefore, $|P_1| \propto 1/(g_{m1} C_C)$ and decreases as the product $(g_{m1} C_C)$ increases, while $|P_2| \propto (g_{m1} C_C)$ and increases with increasing $(g_{m1} C_C)$. Rough estimates of the variables C_{eq1} , C_{eq2} , R_{eq1} and R_{eq2} in **(fig. A3)** were obtained from a simplified hybrid- π (VCCS) BJT model, derived from **2N5551** datasheet characteristics, with g_{m1} modified to accommodate the current gain provided by emitter-follower **Q17 (fig. 1)**. SPICE simulation of **figure A3** shows that P_1 moves down from 21KHz, in the absence of C_C , to 7Hz with C_C in-situ, while P_2 moves from just 48KHz to over 70MHz **(fig. A4)**.

The system therefore maintains a much wider bandwidth with dominant pole feedback compensation than would accrue if such a characteristic were realised by merely increasing shunt-capacitance at the input or output nodes of the TIS. This is unacceptable [17] as it adversely loads the TIS's collector, severely compromising second-stage linearity.

Moreover, dominant pole shunt compensation at either the TIS's input or output node would leave P_2 virtually unchanged, and, consequently, the system's unity loop-gain bandwidth would necessarily have to be much less than 21KHz to guarantee stability when the major feedback loop is closed.

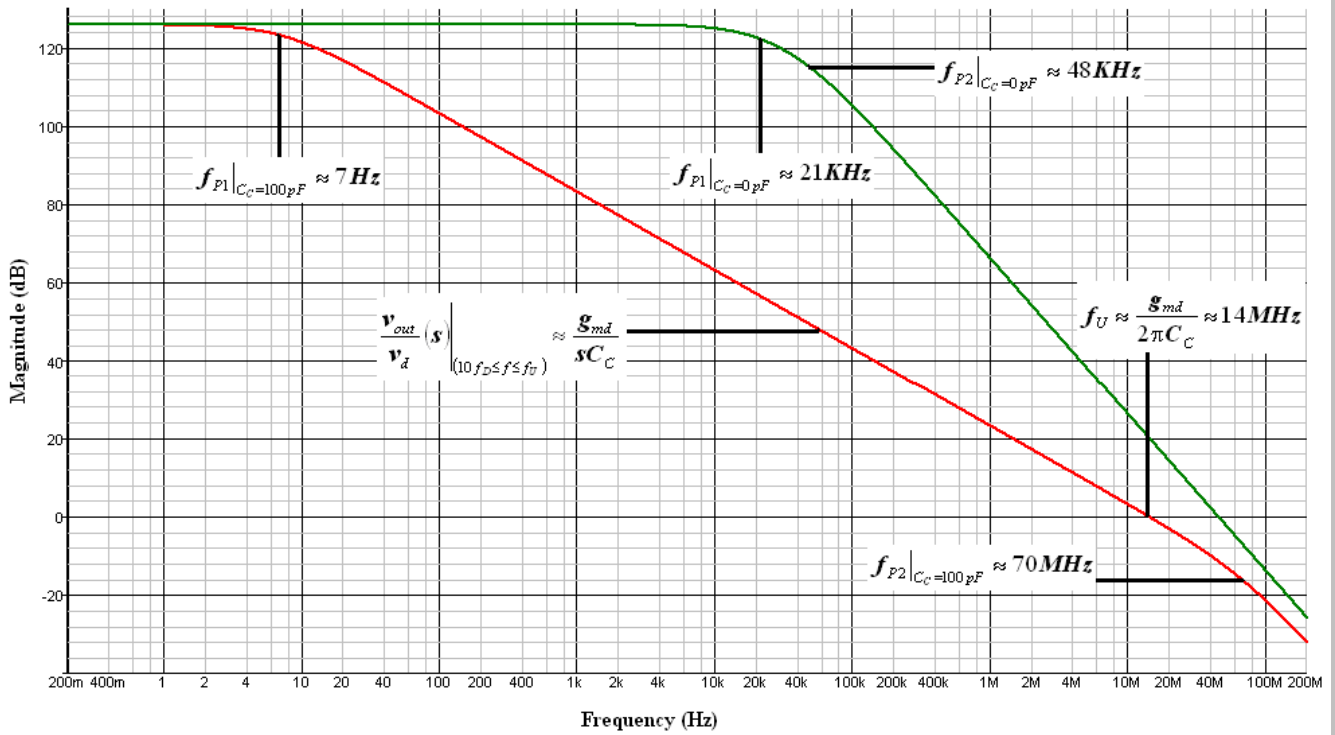


Figure A4. The apparent migration of dominant poles due to Miller-effect compensation on the forward path's frequency response (red trace). The green trace is the uncompensated forward path's frequency response.

The RHP zero (not shown in **fig. A4**) has the magnitude response of an LHP zero (i.e. magnitude response 'breaks up') but the phase response of an LHP pole (i.e. phase response 'breaks down' or tends to -90 degrees in the limit) and might therefore be expected to compromise stability margins. This, however, is of no concern in discrete power amplifiers with an all-BJT second stage as g_{m1} is invariably much larger than C_c which ensures that the RHP zero resides well beyond the unity-gain bandwidth of the amplifier.

Determining system singularities in this fashion helps develop a vivid appreciation of circuit behaviour, but is only of academic interest to the designer of discrete amplifiers, as the solutions depend on imprecisely known variables such as C_{eq1} and C_{eq2} which, moreover, also vary dynamically.

In practice, the first-order approximation of **equation (16a)** is all that is required to determine the value of first stage transconductance and the corresponding size of compensation capacitor needed to ensure that non-dominant system singularities, including output stage poles, are relegated to well beyond the unity loop-gain frequency. For the practicing engineer, the design-stage analysis of second-order circuit behaviour is the province of SPICE simulators.

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THE DIFFERENTIAL FOLDED CASCODE TRANSIMPEDANCE STAGE IN DISCRETE AUDIO FREQUENCY POWER AMPLIFIERS†

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Instead of using a pair of complementary differential stages, driving a pair of complementary common-base stages, as a basis for a push-pull transimpedance stage (TIS), both collectors of a *single* differential stage may be used to drive a pair of common-base stages in a folded cascode arrangement in which the collector outputs of the common-base stages are connected to a current mirror to give a single-ended voltage output (**fig. 1**). This arrangement is identical to that of the input (transadmittance) stage (TAS), and represents a significant reduction in complexity compared with the TIS topology with complementary differential stages.

The push-pull TIS of **figure 1** is just as effective at delivering virtually symmetrical slew as the arrangement with two complementary differential stages, but, regrettably, its performance, in respect of power supply rejection and linearity, is no better than that delivered by the latter, or, indeed, the conventional Thompson topology with a buffered TIS. To obtain a significant improvement in power supply rejection for the circuit of **figure 1**, regulated supplies are required for its input stage, making it even more uneconomic than the much simpler Thompson arrangement which only requires simple RC filters for very good power supply rejection.

