LINEAR AUDIO FREQUENCY POWER AMPLIFIERS WITH COMMON-EMITTER OR COMMON-SOURCE OUTPUT STAGES†

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Introduction

Most extant high quality linear audio frequency power amplifiers consist of a complementary compound emitter-follower output stage driven by a two-stage voltage gain block. The latter consists of a transadmittance stage (TAS) driving a transimpedance stage (TIS). The TAS is a voltage-controlled current source (VCCS) whose single-ended output current is proportional to its differential input voltage, while the TIS is a current-controlled voltage source (CCVS) whose single-ended output voltage is proportional to its single-ended input current from the TAS. **Figure 1** is a novel example of such an amplifier designed by the author.

However, some designers [1], notably of professional amplifiers [2], use the complementary common-emitter (or common-source) output stage of **figure 2**. This arrangement was apparently first published by Cherry and Hooper [3].

The complementary common-emitter output stage of **figure 2** possesses the advantage that it makes more efficient use of its power supply than an emitter-follower output stage in respect of the ratio of maximum peak output voltage swing to the magnitude of the power supply. This ratio is always larger with the complementary common-emitter output stage of **figure 2** than with a complementary emitter-follower output stage, unless the latter has an error correction feedback loop around it which improves its voltage swing efficiency.

Another advantage of a common-emitter output stage is the preceding voltage gain block need only operate from modest power supply rails compared with those that power the output stage, and this is because the output stage has a voltage gain much greater than unity [4]. In contrast, the supply rails powering the voltage gain block preceding an emitter-follower output stage must be at least equal in magnitude to those powering the output stage. Consequently, higher quality small-signal bipolar junction transistors of superior gain-bandwidth product, which are typically of modest voltage ratings, may be used in the voltage gain block driving a complementary common-emitter output stage than are available for use in the voltage gain block preceding an emitter-follower output stage.

One disadvantage of a common-emitter output stage of the form of **figure 2** is its power supply floats, and, consequently, each channel of amplification must have its own split-rail power supply [4]. Additionally, the voltage gain block preceding a common-emitter output stage of the form of **figure 2** must have its own grounded split-rail power supply independent of that powering the output stage. Because it is grounded, this power supply can power the voltage gain blocks of more than one channel of the amplifier.

† This work was expedited by the accuracy, precision, versatility and robustness afforded by Mike Engelhardt's QSPICE, qualities that are notably absent from other SPICE implementations, including LTspice.

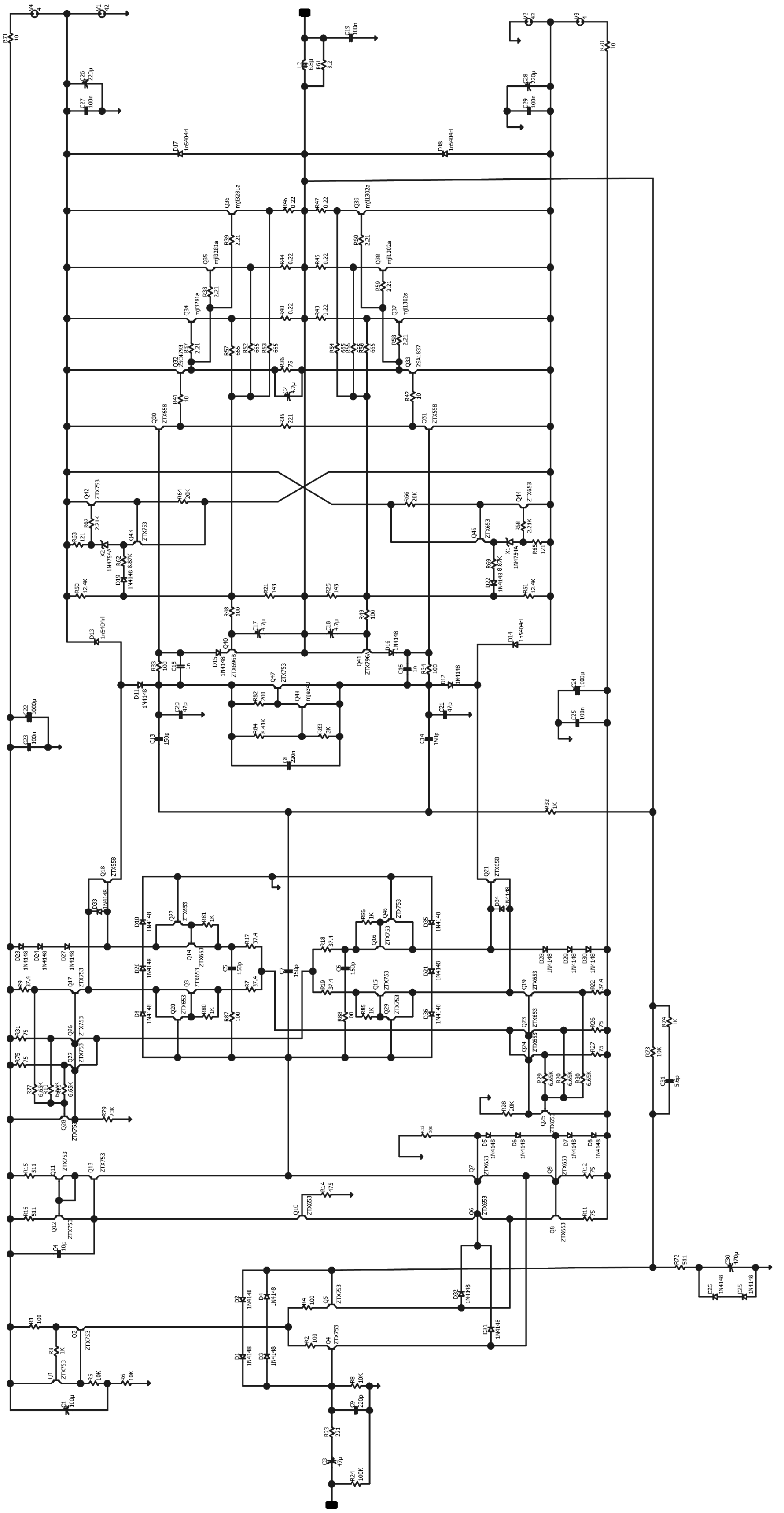

Figure 1. A novel audio frequency power amplifier with a complementary compound emitter-follower output stage. This amplifier has a positive major negative feedback loop phase and gain margin of about 95 degrees and 20 dB respectively, while its primary Miller minor negative feedback frequency compensation loop has a positive phase and gain margin of about 100 degrees and 20 dB respectively (QSPICE).

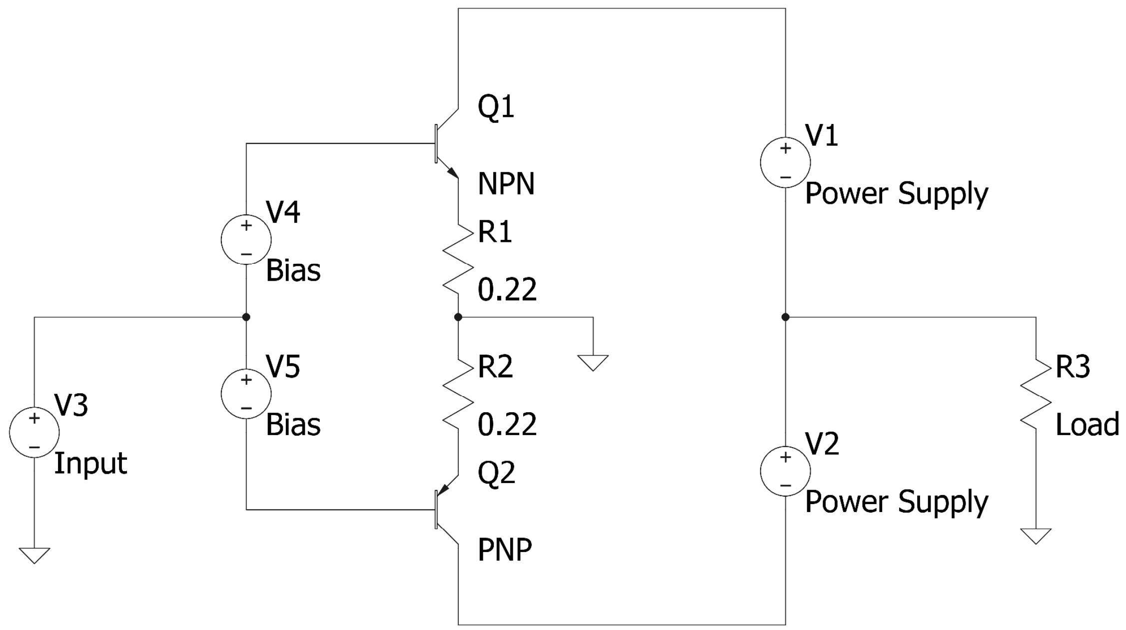

Figure 2. A complementary common-emitter output stage with a floating split-rail power supply.

A three-stage amplifier with a common-emitter output stage

The voltage-gain block preceding the complementary compound common-emitter output stage of the amplifier of **figure 3** is virtually identical to that of the amplifier of **figure 1** with one important caveat: the transadmittance stage (TAS) of the amplifier of **figure 1** is inverting of necessity while that of the amplifier of **figure 3** is necessarily non-inverting.

The voltage gain of the complementary compound common-emitter output stage of the amplifier of **figure 3** increases the amplifier's forward-path gain, and, therefore, increases its major negative feedback loop gain, without the common-emitter output stage broadbanding its own dominant pole to anywhere near the same extent as an emitter-follower output stage. Consequently, draconian minor negative feedback loop frequency compensation must be applied to both the transadmittance stage (TAS), by means of series-derived, series-applied minor negative feedback, and the transimpedance stage (TIS), by means of shunt-derived, shunt-applied minor negative feedback, to prevent the extra voltage gain generated by the common-emitter output stage from destabilising the amplifier's major negative feedback loop. This is consistent with Gift's findings [5] and contradicts Cherry and Cambrell's contention [6] that subject to the output stage having a high current gain the stability margins of an amplifier with a common-emitter output stage are indistinguishable from those of an amplifier with an emitter-follower output stage.

The amplifier of **figure 3** has a mediocre positive major negative feedback loop phase margin of about 20 degrees and an equally unsatisfactory positive major negative feedback loop gain margin of about 6 dB (QSPICE) despite the application of very aggressive minor negative feedback loop frequency compensation in both the transadmittance stage (TAS) and the transimpedance stage (TIS) to attenuate the amplifier's forward-path gain and, therefore, to attenuate its major negative feedback loop gain. Moreover, the amplifier's major negative feedback loop frequency compensation is complicated by the fact that the complementary compound common-emitter output stage's voltage gain, and, therefore, the amplifier's major negative feedback loop gain, varies significantly with the magnitude of the load connected to its output (**fig. 4**), with instability likely to ensue if the amplifier's output is open-circuited [5]. This problem is most unlikely to arise in amplifiers with a complementary emitter-follower output stage.

The draconian frequency compensation required by the amplifier of **figure 3** degrades its linearity as well as the slew rate of the two-stage voltage gain block preceding its output stage. However, this problem is alleviated by the fact that the slew rate of the amplifier of **figure 3** is the product of the slew rate of its two-stage voltage gain block and the voltage gain of its complementary compound common-emitter output stage, which varies roughly between ten and fifteen times depending on the magnitude of the amplifier's output load. Thus, although the amplifier's slew rate varies somewhat with changes in the amplifier's output load, it is more than adequate for most applications.

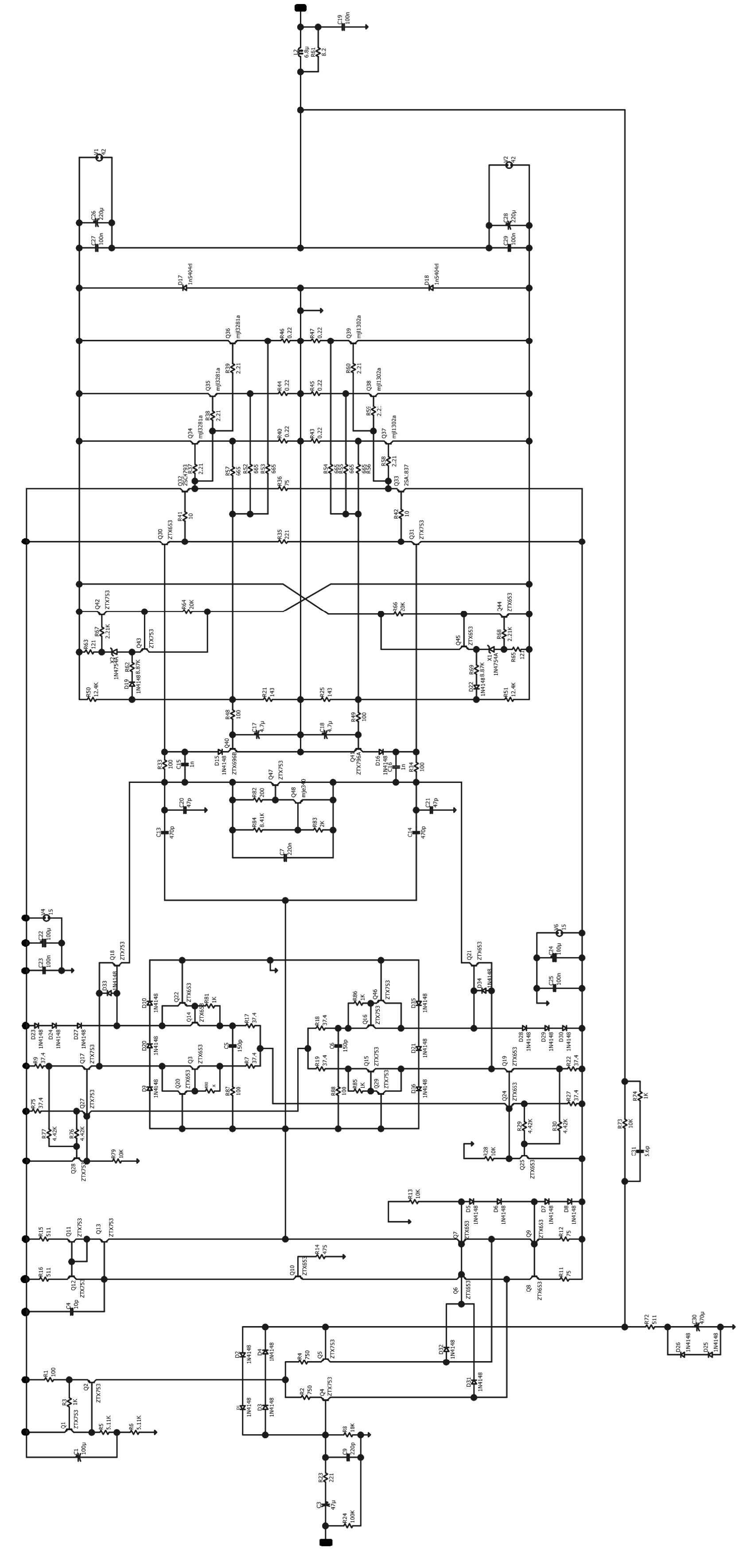

Figure 3. A novel three-stage audio frequency power amplifier with a complementary compound common-emitter output stage. This amplifier has a mediocre positive major negative feedback loop phase margin of about 20 degrees and an equally unsatisfactory positive major negative feedback loop gain margin of about 6 dB (QSPICE).

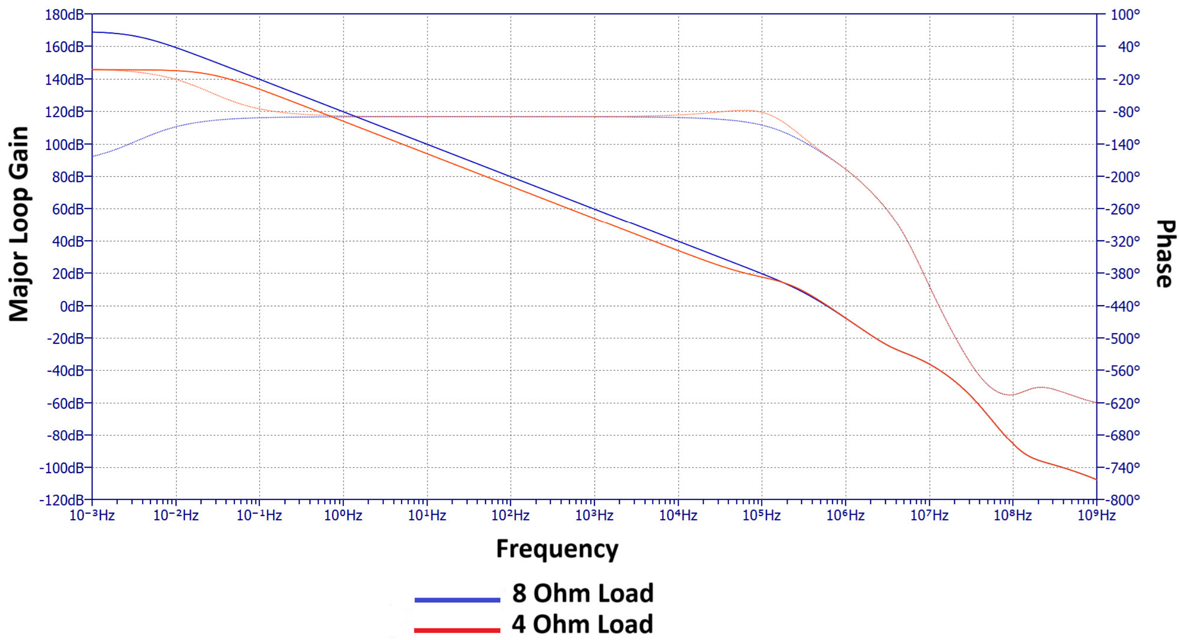

Figure 4. The major negative feedback loop gain of the amplifier of figure 3 varies significantly with changes in the magnitude of the load connected to its output (QSPICE).

A novel two-stage amplifier with a common-emitter output stage

The onerous minor negative feedback loop frequency compensation mandated by the three stages of gain in the amplifier of **figure 3** may be greatly ameliorated by reducing the number of stages generating the amplifier's forward-path gain to two. This is done by deleting the transimpedance stage (TIS) so that the transadmittance stage (TAS), a voltage-controlled current source (VCCS) — which must be inverting of necessity — drives the complementary compound common-emitter output stage directly as shown in **figure 5**. The complementary compound common-emitter output stage then becomes a transimpedance stage (TIS), a current-controlled voltage source (CCVS), with Miller minor negative feedback loop frequency compensation shunt-derived from its output and shunt-applied to its input courtesy of capacitor **C21**.

Wrapping the complementary compound common-emitter output stage in a Miller minor negative feedback frequency compensation loop lowers the frequency of its dominant pole while broadbanding its first non-dominant pole. This stabilises the amplifier's major negative feedback loop by ensuring that the amplifier's major loop gain rolls off at a single pole rate at the unity major loop gain frequency. Additionally, the minor negative feedback frequency compensation loop around the complementary compound common-emitter output stage improves its linearity and makes the amplifier's major negative feedback loop gain less susceptible to changes in the magnitude of the load connected to the amplifier's output at ultrasonic frequencies and across most of the audio band.

The amplifier of **figure 5** has a perfectly respectable positive major negative feedback loop phase and gain margin of about 85 degrees and 14 dB respectively. The minor negative feedback frequency compensation loop around the output stage has a positive phase margin of about 60 degrees, which is satisfactory (QSPICE).

The compensating Miller minor negative feedback loop around the output stage of the amplifier of **figure 5** would likely be unstable or have inadequate stability margins in the absence of feedforward capacitors **C5** and **C6** around the output stage's complementary emitter-follower drivers. Regrettably, these feedforward capacitors rather degrade the linearity of the amplifier at the top end of the audio band. A potentially unstable minor negative feedback loop, with a negative or inadequate phase and/or gain margin, typically introduces a resonant peak in the major negative feedback loop's response beyond its unity loop gain frequency, and this would likely degrade the major negative feedback loop's gain margin (**fig. 6**). This is contrary to Cherry's assertion [7] that "a substantial phase margin for the inner loop is...[not] a requirement for system stability."

Whereas the slew rate of the amplifier of **figure 3** is simply the product of the slew rate of its two-stage voltage gain block and the voltage gain of its complementary compound common-emitter output stage (which varies with changes in the magnitude of the amplifier's output load), the slew rate of the amplifier of **figure 5** is roughly equal to the value of its transadmittance stage's tail current divided by the value of the Miller minor negative feedback loop frequency compensation capacitor **C21** wrapped around its complementary compound common-emitter output stage. The slew rate of both amplifiers of **figure 3** and **figure 5** is of the order of 60V/uS with an 8 Ohm output load.

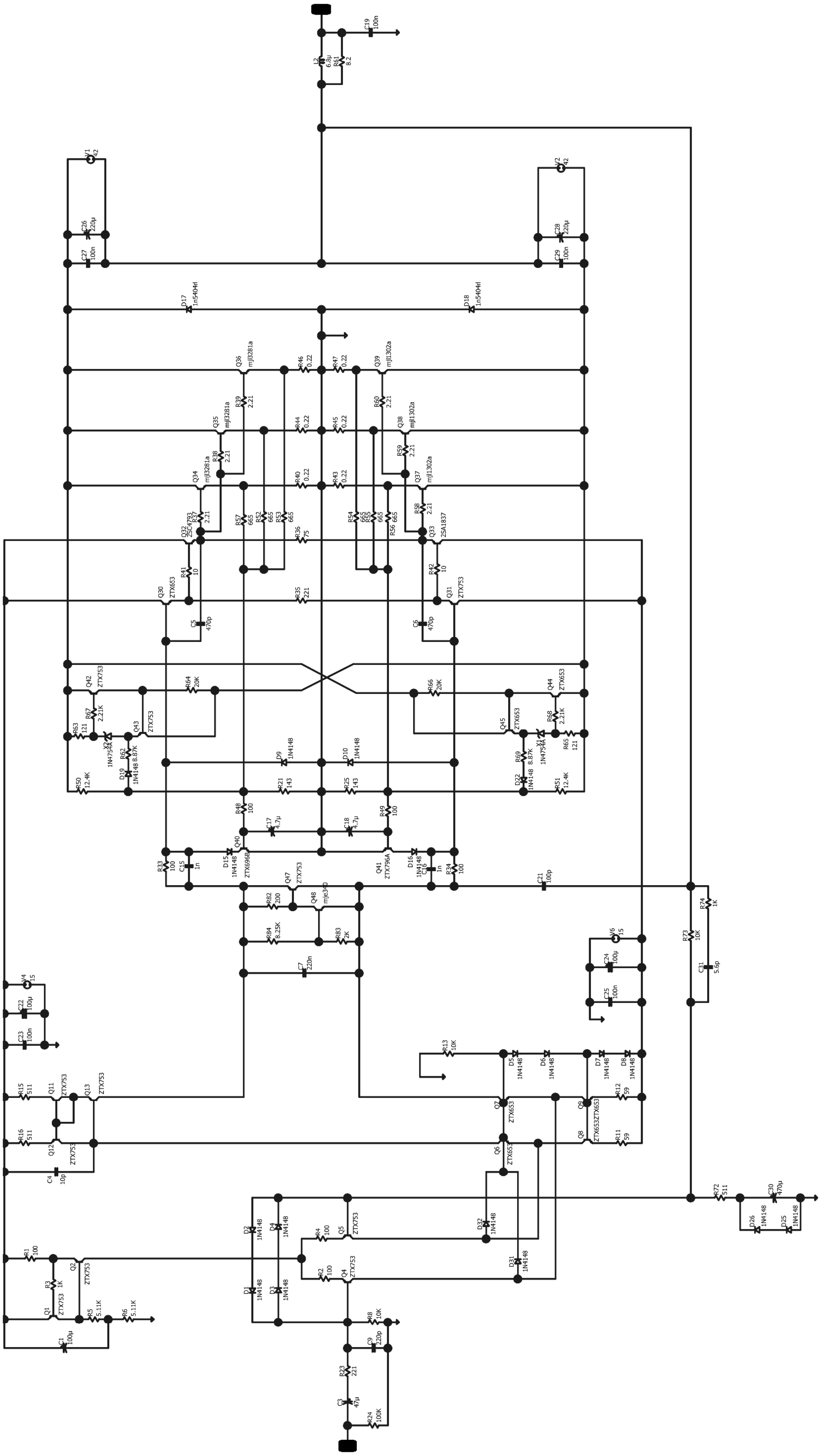

Figure 5. A novel two-stage audio frequency power amplifier with a complementary compound common-emitter output stage.

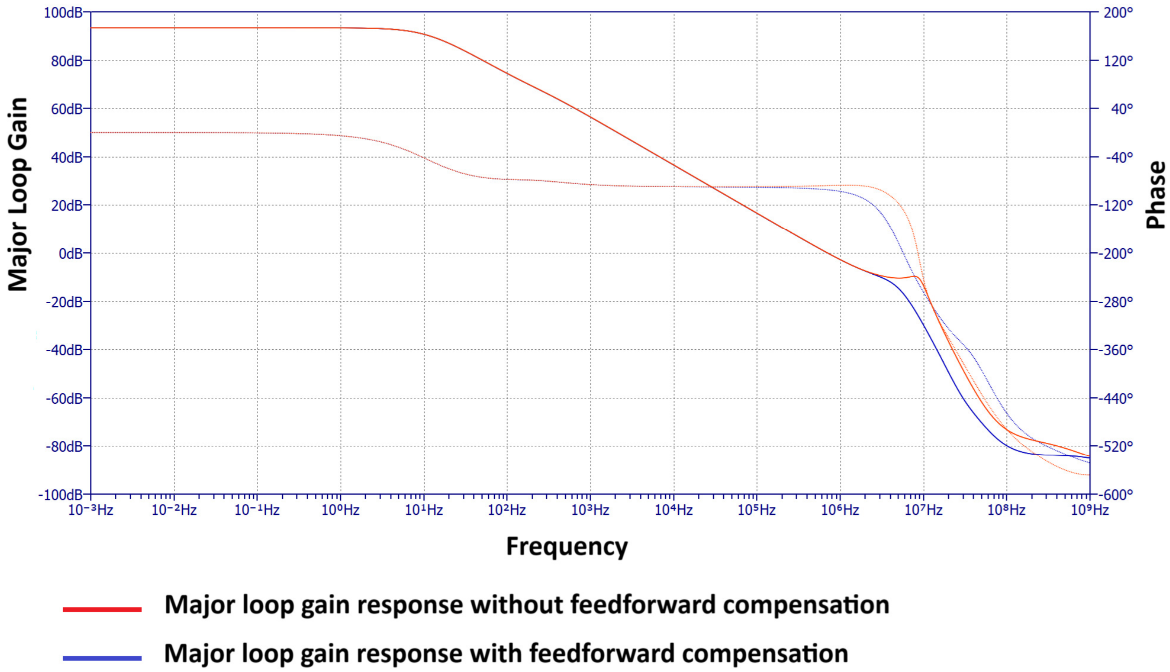

Figure 6. The feedforward capacitors around the complementary emitter-follower output stage drivers in the amplifier of figure 5 increase the stability margins of the amplifier's Miller minor negative feedback frequency compensation loop, and this eliminates the resonant peak in the amplifier's major negative feedback loop gain response (QSPICE).

A novel composite amplifier with a common-emitter output stage

The linearity, the DC offset and the output impedance of the simple two-stage power amplifier of **figure 5** may be improved by making it part of a composite power amplifier in which it is driven by a low-distortion operational amplifier as shown in **figure 7**. In this arrangement the operational amplifier is a voltage error amplifier, while the amplifier of **figure 5** is used as a voltage booster with a closed-loop gain of fifteen. This reduced closed-loop gain compared with that of the amplifier of **figure 5** is necessary to reduce the major loop gain and, therefore, to reduce the unity major loop gain frequency of the composite amplifier and to maximise the closed-loop bandwidth of its booster. Ideally, the closed-loop bandwidth of the booster should exceed the unity major loop gain frequency of the composite amplifier to guarantee its stability. In general, the booster amplifier's closed-loop voltage gain should be made equal to or less than the composite amplifier's required closed-loop voltage gain so that the operational amplifier's closed-loop voltage gain is forced to be equal to or slightly greater than unity to maximise its linearity. Note that all the negative feedback loops in a composite amplifier must themselves be independently stable.

The minor negative feedback loop defined by capacitor **C7** around the operational amplifier stabilises the composite amplifier by increasing its major negative feedback loop's stability margins from negative values to positive values (**fig. 8**). Capacitor **C7** shunts the booster amplifier out of the operational amplifier's feedback loop at high frequencies, and this has the effect of lowering the frequency of the major negative feedback loop's dominant pole and, therefore, lowering the unity major loop gain frequency. This reduces the phase shift the booster introduces at the unity major loop gain frequency. Note that the operational amplifier must be designed for stability at unity closed-loop gain since the minor negative feedback loop defined by capacitor **C7** imposes 100% negative feedback around it at high frequencies.

Regrettably, the unity major loop gain frequency of the composite amplifier with capacitor **C7** *in situ* is close to the frequency of the booster's dominant closed-loop pole which appears in the major negative feedback loop's response, and, consequently, this pole significantly degrades the major negative feedback loop's phase margin. Thus, resistor **R22** is inserted in series with capacitor **C7** to introduce a left half-plane zero in the vicinity of the major negative feedback loop's unity loop gain frequency to cancel the booster's dominant closed-loop pole, and this increases the major negative feedback loop's phase margin (**fig. 8**). The frequency f_0 of the left half-plane zero in the composite amplifier's major negative feedback loop gain response is given by $f_0 = 1/(2 \times \pi \times C7 \times R22)$. Note that ideally the operational amplifier's gain-bandwidth product should be at least 10MHz to maximise the composite amplifier's major negative feedback loop gain and stability margins without requiring draconian minor negative feedback loop frequency compensation around the operational amplifier. Additionally, note that the frequency of the major negative feedback loop's dominant pole is not only a function of the value of capacitor **C7** but also of the values of the major negative feedback network's resistors.

The small capacitor **C2** in parallel with the major loop's feedback resistor **R8** introduces a left half-plane zero at a slightly higher frequency than the composite amplifier's unity major loop gain frequency in order to obtain the extra phase advance provided by the zero without incurring the penalty of a significant attendant increase in the unity major loop gain frequency, while resistor **R7** in series with capacitor **C2** controls the location of the first non-dominant pole after the left half-plane zero introduced by capacitor **C2**: the larger the value of resistor **R7**, the closer the frequency of this pole to that of the left half-plane zero introduced by capacitor **C2**. Resistor **R7** also helps attenuate any ultrasonic interference that might otherwise be introduced by **C2** directly into the inverting input of the operational amplifier from the composite amplifier's output.

Although it is significantly more linear than the amplifier of **figure 5**, the composite design of **figure 7** generates roughly three times the distortion at 20KHz of the amplifier of **figure 1**. Additionally, as shown in **figure 9**, although much lower than that of the amplifier of **figure 5**, courtesy of the extra loop gain to which its output stage is subject, the closed-loop output impedance of the amplifier of **figure 7** is rather higher across the audio band than that of the amplifier of **figure 1**. Indeed, all other parameters being equal, the closed-loop output impedance of any linear audio frequency power amplifier with a complementary common-emitter output stage of the form of **figure 2** is invariably higher than that of an amplifier with an emitter-follower output stage, contrary to Cherry [4].

The relatively high closed-loop output impedance of an amplifier with a common-emitter output stage of the form of **figure 2** suggests that it is more susceptible to instability provoked by a reactive load than an amplifier with an emitter-follower output stage. This is consistent with Gift's analysis [5] and is contrary to Cherry and Cambrell's assertion [6] that "there is no truth whatsoever in the claim that [an amplifier with an emitter-follower output stage], because of its lower open-loop output [impedance], is more tolerant toward changes in load capacitance [than an amplifier with a common-emitter output stage]." To improve the linearity and to lower the output impedance of the amplifier of **figure 5** so that they are even better than those of the amplifier of **figure 1**, the amplifier of **figure 5** is made part of the novel triple composite nested feedback arrangement shown in **figure 10**.

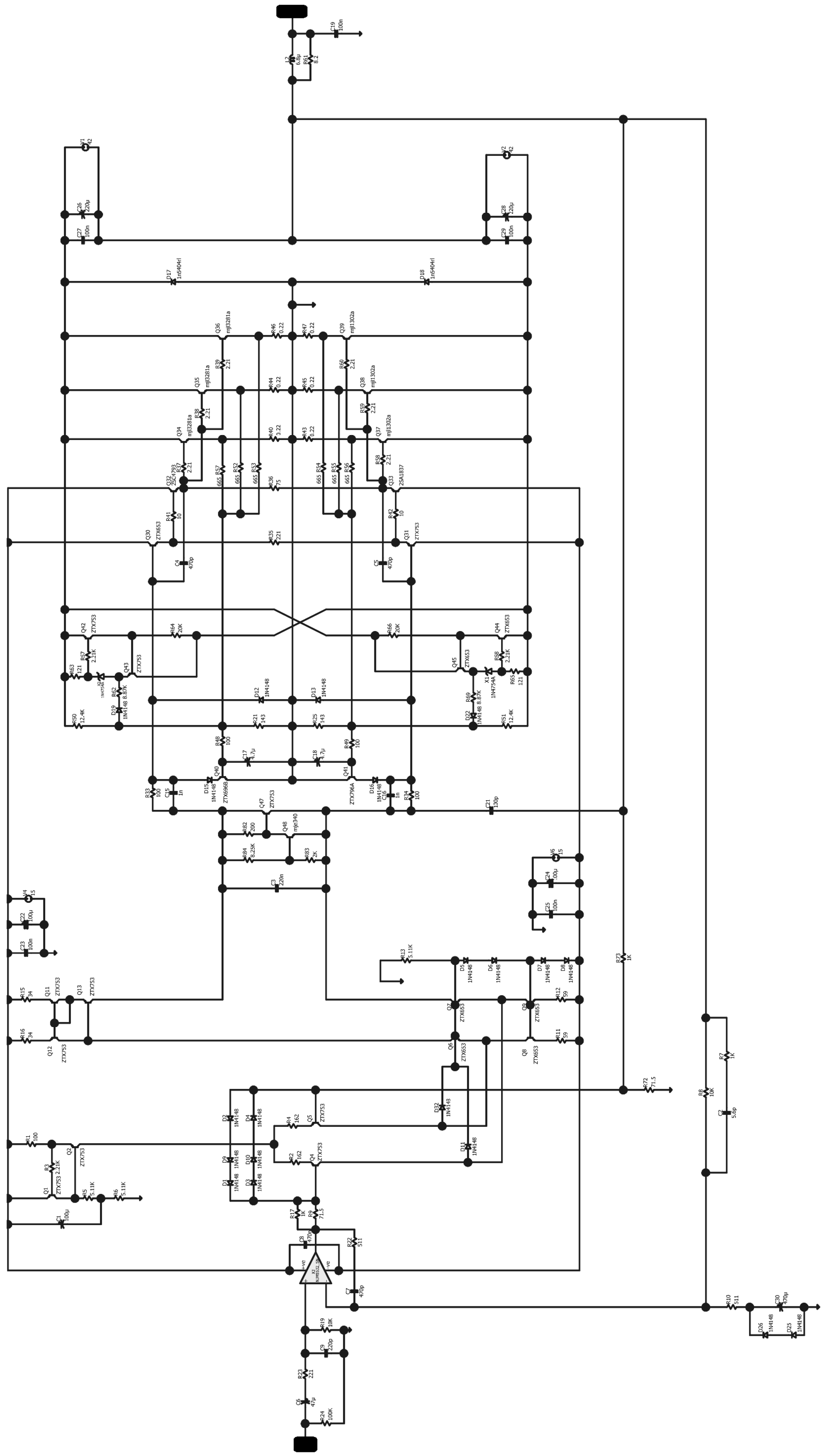

Figure 7. A novel composite linear audio frequency power amplifier with a complementary compound common-emitter output stage.

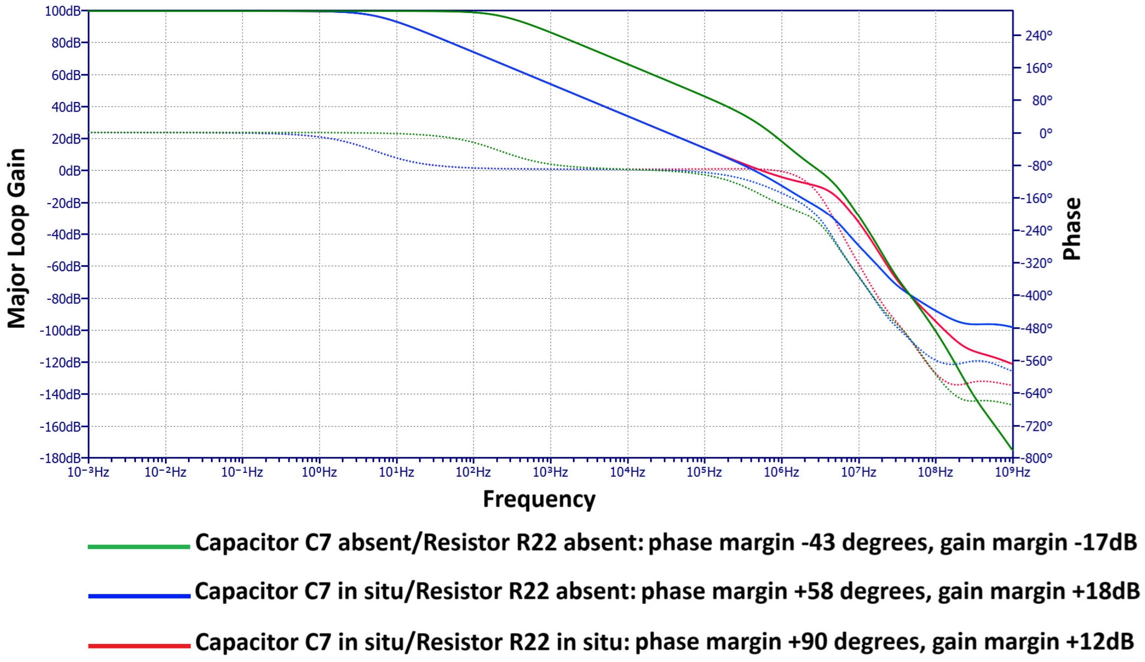

Figure 8. The effect that the two compensator elements C7 and R22 in the amplifier of figure 7 have on its major negative feedback loop gain response (QSPICE).

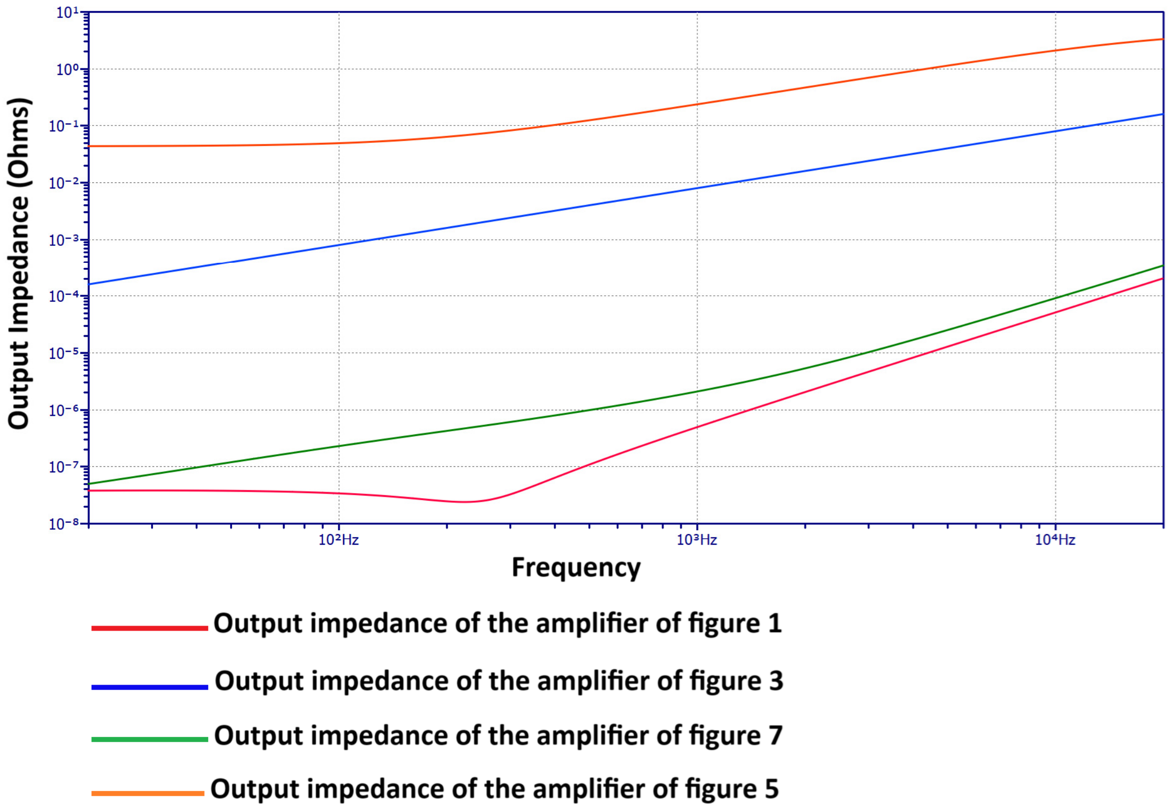

Figure 9. The closed-loop output impedance of an amplifier with a complementary common-emitter output stage of the form of figure 2 is invariably higher across the audio band than that of an amplifier with a complementary emitter-follower output stage (QSPICE).

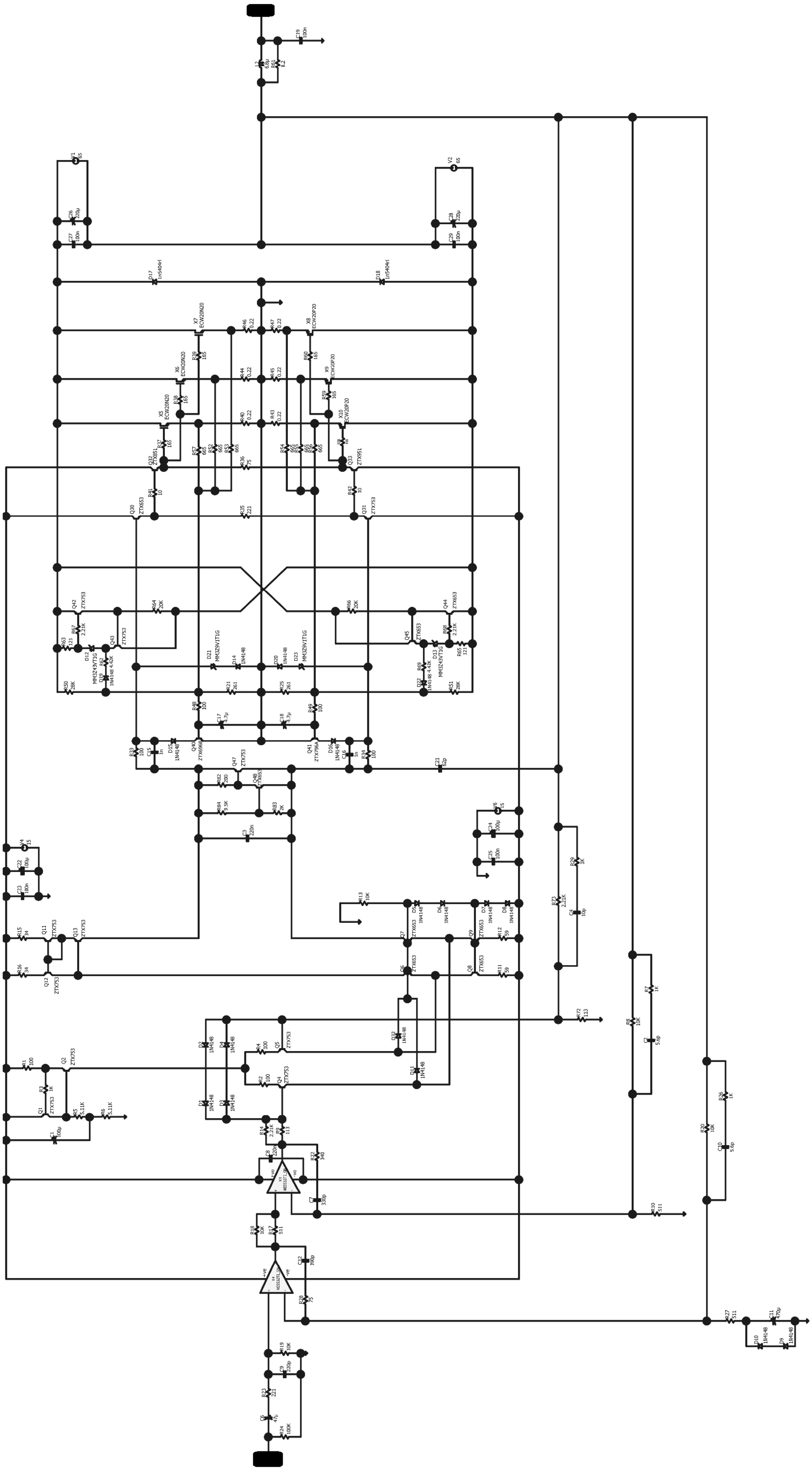

Figure 10. A novel triple composite linear audio frequency power amplifier with a complementary compound lateral MOSFET common-source output stage.

Conclusion

A linear audio frequency power amplifier with the complementary common-emitter output stage of **figure 2** has a number of disadvantages. Apart from the fact that a single split-rail floating power supply cannot power more than one channel of such an amplifier, supply rail power semiconductor switches would be difficult to use as part of the amplifier's DC fault protection system because the power supply is not grounded. With the exception of the amplifiers of **figure 7** and **figure 10**, an amplifier with a complementary common-emitter or common-source output stage of the form of **figure 2** typically suffers from poorer linearity and generates significantly greater DC offset than an amplifier with a complementary emitter-follower output stage. Moreover, unlike an emitter-follower output stage, an error feedback correction loop cannot be applied around the complementary common-emitter output stage of **figure 2** to reduce its distortion because its voltage gain is ill-defined and much greater than unity. Additionally, all other parameters being equal, the closed-loop output impedance of an amplifier with a complementary common-emitter output stage of the form of **figure 2** is invariably higher across the audio band than that of an amplifier with an emitter-follower output stage.

The dominant forward-path pole of a common-emitter output stage of the form of **figure 2** is always at a significantly lower frequency than that of an emitter-follower output stage given the same output devices, and the voltage gain generated by the common-emitter output stage of the three-stage amplifier of **figure 3**, for example, severely degrades its major negative feedback loop's stability margins. These two problems together call for far more draconian minor negative feedback loop frequency compensation if the common-emitter output stage is driven by a two-stage voltage gain block than is required for an amplifier with an emitter follower output stage. This means a linear audio frequency power amplifier with an emitter-follower output stage reaps the benefit in respect of linearity of being driven by a two-stage voltage gain block with modest minor negative feedback loop frequency compensation that an amplifier with a common-emitter output stage driven by a vastly more aggressively-compensated, but otherwise identical, two-stage voltage gain block does not.

Clearly, the disadvantages of a three-stage amplifier with a complementary common-emitter output stage, such as that of **figure 3**, outweigh one of its two advantages: it is possible to use low-voltage small-signal transistors of relatively high current-gain-bandwidth product in the two-stage voltage gain block preceding the complementary common-emitter output stage. This should improve the amplifier's major and minor negative feedback loop stability margins in principle, but in practice this is not the case because the aforementioned shortcomings of a common-emitter output stage driven by a two-stage voltage gain block have an overwhelmingly deleterious effect on the amplifier's major negative feedback loop stability margins if not its minor negative feedback loops' stability margins.

The use of a common-emitter output stage of the form of **figure 2** is only tenable if the number of gain stages preceding the output stage is limited to one, a transadmittance stage (TAS), with the output stage converted into a transimpedance stage (TIS) in order to guarantee the stability of the amplifier's major negative feedback loop with modest frequency compensation (**fig. 5**). The amplifiers of **figure 7** and **figure 10** are an elaboration of that of **figure 5** that improve its performance.

Ultimately, all other parameters being equal, a linear audio frequency power amplifier with a complementary emitter-follower output stage is superior in every meaningful way, except output voltage swing efficiency, to that with a complementary common-emitter or common-source output stage.

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