Cascomp BJT Amplifier vs. Traditional Configurations

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Abstract: All transistor circuits introduce distortion. In Radio Frequency (RF) circuits, the third-order distortion components are the most important. The quest for more linear circuits has become more important with complex-modulation as used in modern cellular phone systems. Quinn's Cascomp Amplifier, first reported in the 1970s, promises ideal linearity and can deliver close to that promise. We review the theory and address the question of why the Cascomp has not replaced other configurations in amplifiers where low distortion is important. Calculations are supported by measurements. A new, alternative variant of the Cascomp topology is introduced and compared with the existing configuration. We assert that the improved linearity comes at such a price in gain that it does not make sense to use the configuration in broadband RF circuits.

Keywords: Cascomp, Compensation, Error-correction, Feed-forward, Intermodulation distortion, Linearity, Transconductance amplifiers.

1 INTRODUCTION

The collector current of a Bipolar Junction Transistor (BJT) obeys the well-known nonlinear equation

$$I_C = I_S \, e^{\frac{V_{BE}}{V_T}}. \qquad (1)$$

When a signal is imposed on the base-emitter junction, a nonlinear function of that signal appears in the collector. This results in distortion of the signal. For example, when the input signal is a sinewave, the collector current contains both the original sinewave and its harmonics, and we say that the amplified signal contains "harmonic distortion" [1]. Of more interest to communications engineers is the distortion that arises when the input signal consists of multiple sinewaves. In the case of two sinewave inputs, some algebra shows that the output signal will contain signals at various frequency components as depicted in Figure 1 [1]. The tones whose frequencies lie nearby to the two input frequencies, f_1 and f_2, are the ones considered most objectionable because they cannot easily be filtered out. It is common to measure the extent of the distortion introduced by an amplifier with a measure such as Third-order Intermodulation Distortion ("IM3") or Third-order Intercept (TOI) [1].

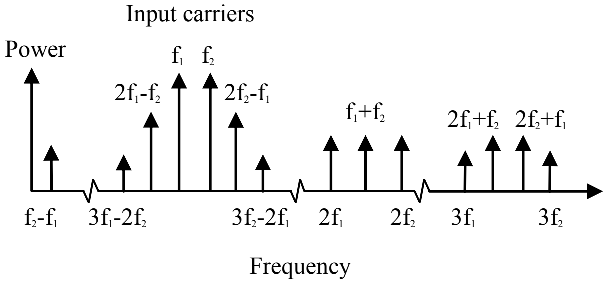

Figure 1. Components involved in harmonic and intermodulation distortion (IMD) from two input tones.

There has been considerable effort spent in recent decades to find a configuration of transistors that produces less distortion. The Differential Pair (DP), shown in Figure 2(b), has a tanh transfer function not an exponential one. This is claimed to be more linear than the traditional Common-Emitter (CE) amplifier, shown in Figure 2(a). Tanh is symmetrical, so it yields no second-order components, and thus produces much lower harmonic distortion figures. However, it offers less advantage if one is interested only in the IM3 figure.

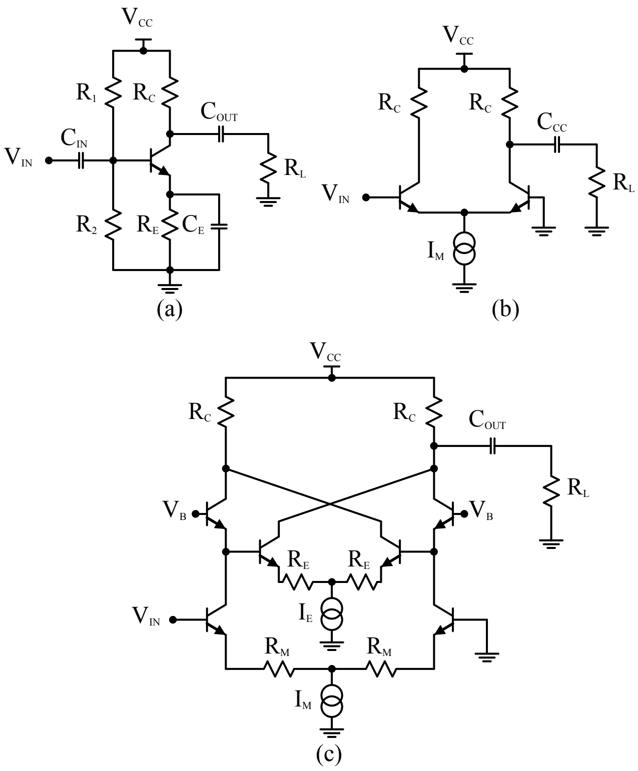

Figure 2. (a) Common-emitter, (b) Differential pair and (c) Cascomp configurations discussed in this paper.

2 COMPARISON OF DP AND CE CONFIGURATIONS

To compare the DP to the CE configuration, we consider a two-tone test applied to each amplifier. In each case, we will consider the input peak voltage at the base of the transistors and the output peak current taken at the collector. The measurements were taken this way because, at small signal, the resistance seen by the collector and the voltage supply have no effect on the output current, and thus our comparison is independent of these.

Collector bias currents of the CE and DP amplifiers are set at 5 mA. The Values of R_1 and R_2 are set low enough to mitigate the effects of β variations on the quiescent collector current. The value of R_C was set low enough that compression did not occur until higher voltages. This was done because the TOI is determined when the output is a linear function of the input. Increasing the value of R_C means that compression occurs earlier and the TOI may need to be acquired at lower input voltages.

Figure 7 and the upper half of Table 2 present a comparison between the CE and DP circuits with respect to gain (g_m) and TOI. Simple theory predicts that the gain of the CE will be 192 mS and the (single-ended) gain of the DP half that compared to the CE or 96 mS. The theory does not allow for parasitic resistance in the BJT nor Early effect or β variations, and so yields slightly optimistic numbers compared with the simulations and measurements. Intersil's CA3083 transistor-array chip was used after carefully determining the SPICE parameters using an Agilent E5270B. Simulation and measurement agree well. The SPICE parameters are shown in Table 1.

Table 1. Extracted SPICE parameters for the transistors in intersil's CA3083.

IS (Saturation current)	9.6E-15 A
NF (Forward ideality factor)	1.004
NR (Reverse ideality factor)	1.012
VAF (Forward Early voltage)	54 V
VAR (Reverse Early voltage)	6 V
BF (Forward β)	224
BR (Reverse β)	25
RE (Terminal emitter resistance)	1.2 Ω

With equivalent collector quiescent currents, the DP has ~6 dB less gain than the CE, and 1.4-1.7 dB worse IIP3. The DP also has lower fifth-order and seventh-order intermodulation products than the CE. If the user is interested in THD there is an improvement, but if TOI is the measure, the DP performs worse in both ways.

To summarise, the DP reduces even order nonlinearities and THD in the output, but has slightly worse TOI with half the g_m. For this reason more elegant circuits were designed using feed-forward and feed-back techniques [3] to cancel nonlinearities. One circuit of particular interest is the Cascomp feed-forward amplifier [4]-[6].

3 QUINN'S CASCOMP AMPLIFIER

The Cascomp feed-forward amplifier is shown in theoretical form in Figure 3. The Cascomp configuration used in this paper is shown in Figure 2(c). The source impedance (R_S in Figure 3) of the input voltage source is not shown in the circuits in Figure 2, but is implicit to the voltage inputs applied to these circuits. R_B (Figure 3) is used to compensate for the beta dependant gain caused by R_S. It is not included in Figure 2(c) because it adds complexity to the circuit and is not essential to our analysis. The π configuration of the circuitry below the emitters of the main differential pair in Figure 3 performs exactly as the T configuration at the same nodes in Figure 2(c).

The Cascomp amplifier uses the cascoding stage of the main amplifier to replicate the nonlinearities produced by $\Delta V_{BE} = V_{BE1} - V_{BE2}$ not being equal to zero. The error amplifier, ideally represented in Figure 3 but appearing as the inner DP in Figure 2(c), then senses the replicated voltage $\Delta V_{BE}^{'} = V_{BE3} - V_{BE4}$ and adds a copy of this to the output of the cascaded DP. This ideally perfectly cancels the distortion in signal at the output. For a more thorough explanation of this profound process the reader is referred to the references cited!

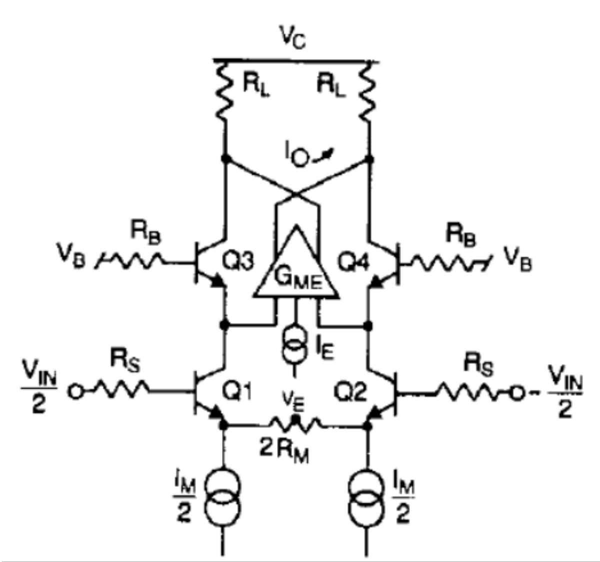

Figure 3. Patrick Quinn's Cascomp feed-forward error correction amplifier.

Nonlinear cancellation occurs in the Cascomp when the transconductance of the error amplifier, $g_{ME} = 1/R_M$, where R_M is the emitter degeneration resistance of the main DP. In Quinn's implementation, the value of I_E is suggested to be half that of I_M. This is because if I_E is too low the DP error amplifier causes spurious nonlinearity through its own tanh characteristic. When I_E is too large, the base currents drawn by the inputs of the error correction amplifier become so large that the current in the transistors of the main DP are no longer the same as the currents in the cascading devices [5]. In Figure 3, the collector currents of Q1 and Q3 would no longer be the same.

3.1 Cascomp Considerations

When designing a Cascomp amplifier many values need to be determined to ensure perfect cancellation. Taking into consideration the conditions for compensation, and after some algebra, the transconductance of a single-ended output of the Cascomp becomes

$$g_M = \frac{1}{2R_M}. \tag{2}$$

It is most common to use a degenerated differential pair as the error amplifier. Assuming $\beta \gg 1$ and satisfying the condition $g_{ME} = 1/R_M$,

$$R_E = R_M - \frac{2V_T}{I_E} \tag{3}$$

where R_E is the degeneration resistor in the error amplifier. It is impossible to obtain more than a certain gain for a given value of tail current I_M, because the required value of R_E would be less than zero. In practice, Quinn's theory requires R_M to have a value of at least $1/g_{ME}$, severely limiting the achievable overall g_M. Refer to Figure 4 for a visual representation of this limit.

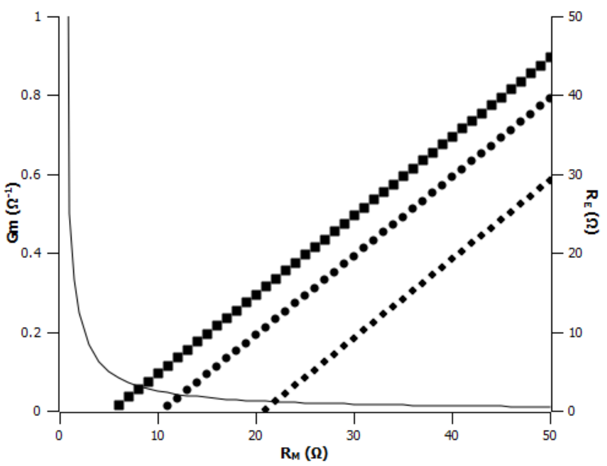

Figure 4. This plot shows the value of R_E required to achieve a given transconductance in the overall amplifier for various ratios of the two tail currents when I_M=20 mA (■), I_M=10 mA (●), I_M=5 mA (♦). Cascomp transconductance (—).

Using this relationship, suitable resistor values were chosen. Simulation and measurement showed that the actual value needed for the degeneration resistors in the error amplifier was slightly lower than the theoretical value. This is attributed to the transistor's terminal resistance in the emitter leads, as well as added resistances from connections in both tail circuits. Finite beta and early effects also contribute to the deviation from theory.

Also note that the Cascomp requires sufficient degeneration to produce significant compensation (cancellation of distortion products). This null occurs when $R_M = 30\ \Omega$ for the currents used in this paper. A value of 33 Ω was used to produce a g_M close to the maximum g_M attainable, while ensuring a high level of compensation. This also meant that the null would be present with shifts due to component tolerances.

4 COMPARISON OF THE CASCOMP AGAINST CE AND DP

To compare the Cascomp with the CE and DP we consider the same two-tone test in section 2. Table 2 and Figure 7 show the comparison between the Cascomp (single-ended), CE and DP circuits with respect to gain (g_m) and TOI. Theory (Equation 2) predicts that the gain of the Cascomp will be $g_m = 15.2\ mS$. Simulation and measurement agree well with theory.

With equivalent collector quiescent currents, the Cascomp has 14 dB less gain than the DP, and 20 dB less than the CE. There is vast improvement in TOI over both the CE and DP.

The TOI of the Cascomp was taken at an output current of -131 dBA. This was the lowest point measurable owing to the limits of dynamic range in the spectrum analyser. Whilst this is not a perfect indication of the true TOI, the values show dramatic improvement over the CE and DP. There is an improvement in IIP3 of over 38 dB compared to the CE and DP. Figure 7 shows the slope of the third-order product as 110 dB/dec instead of the typical 60 dB/dec. This is due to the fifth-order intermodulation product's contribution to the third's coefficient.

To summarise, the Cascomp reduces Intermodulation Distortion (IMD) and increases the TOI. The downfall of the Cascomp is that the degeneration needed for compensation reduces the gain (g_m) to a tenth of the gain possessed by the CE and a fifth the gain of the DP.

5 ALTERBATIVE CASCOMP CONFIGURATIONS

According to Equation 3 there are multiple points where compensation (distortion nulling) occurs for different tail currents in the error amplifier. This theory holds true only when there is sufficient degeneration in the error amplifier that it approximates the ideal of Quinn's theory. When there is low degeneration, the null formed by compensation does not exist, because the error amplifier contributes its own distortion. This circumstance is not taken into account by the original theory.

Simulation and optimisation have shown that there are a range of currents that have compensation at two resistance values. One of these values is the value predicted by Quinn's theory (Equation 3). The second value is the point where the distortion contribution of the error amplifier itself is taken into account, and it corresponds to a higher tail current value. This increased current also means an increase in the Cascomp's gain. It is possible to think of this second cancellation point as the place where the distortion of the error amplifier is additionally pitted against the nonlinearity of the main amplifier.

To compare these two points simulation was done to find the point at which compensation occurs when there is no degeneration in the error amplifier. This point would bring added gain to the Cascomp which it severely needs. The main amplifier used in the previous sections was kept constant with only the error amplifier being altered. The input voltage was kept at a constant 0.02 V (-34 dBV). Figure 5 and Figure 6 display the third-order intermodulation distortion and fundamental gain when the error amplifier is so modified. The inverted spikes in Figure 5 are due to large simulation steps.

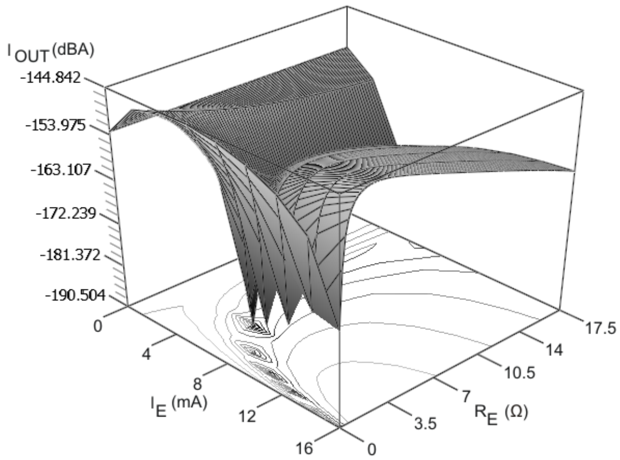

Figure 5. Output level of the Cascomp's third-order intermodulation product as a function of the error amplifiers tail current and degeneration resistance.

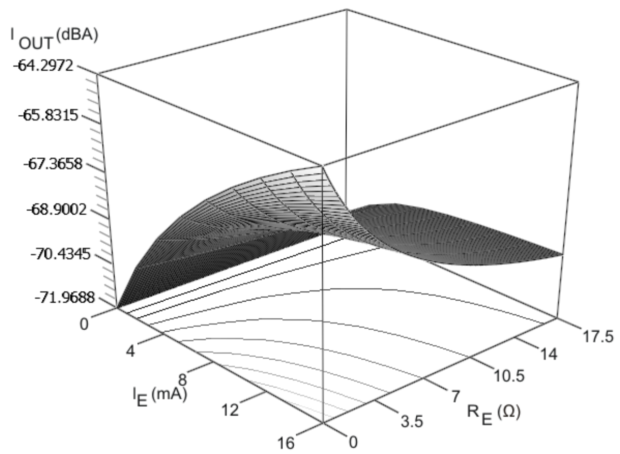

Figure 6. Output level of the Cascomp's carrier as a function of the error amplifiers tail current and degeneration resistance.

The theoretical point from the previous sections occurred when $I_E = 5\,mA$ with a $17.6\,\Omega$ degeneration resistor. The point where no degeneration was needed occurred when $I_E = 16.4\,mA$. This produced a ~6 dB increase in gain and a ~3 dB decrease in TOI compared with the traditional Cascomp. Figure 7 and Table 2 compare the new design with the previously mentioned amplifiers.

In section 3 it was explained why the current should be half that of the main amplifier. Because of the increase in current the output current will be influenced by the base currents from the error amplifier. This effect can be cancelled by a combination of various techniques including of scaling transistors [7].

Table 2. Third-order intercept and transconductance. Values given are rounded to three significant figures.

Common-emitter	**IIP3 (dBV)**	**Gm (mS)**
Simulated	-15.4	147
Measured	-14.8	146
Differential pair		
Simulated	-16.8	76.5
Measured	-16.1	74.3
Cascomp		
Simulated	>22.7	15.3
Measured	>23.3	15.2
Alternative Cascomp		
Simulated	>19.3	30.7

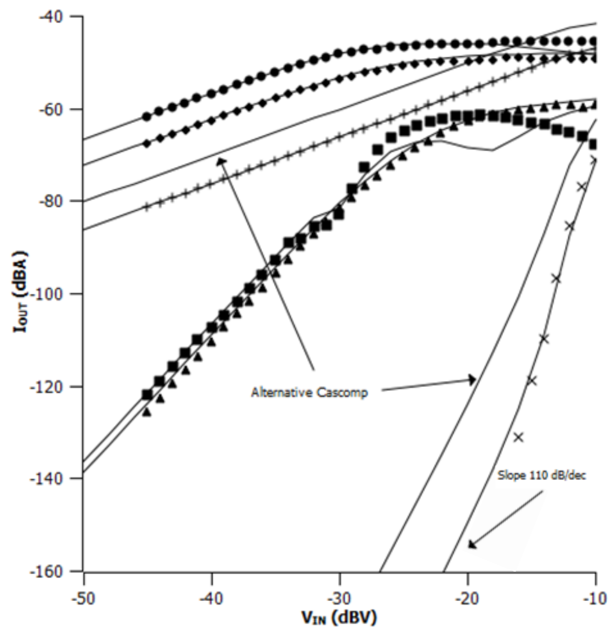

Figure 7. Output level as a function of input level for fundamental and third-order products for CE, DP and Cascomp amplifiers with similar device operating points. Simulation (—), symbols are measured data. CE fundamental (●), CE third-order product (■), DP fundamental (♦), DP third-order product (▲), Cascomp fundamental (+) and Cascomp third-order product (x).

6 REALISATION IN HBT

Although the Cascomp amplifier allows for significant improvement in TOI compared to other topologies, the gain-bandwidth product of the amplifier is much smaller. The compensation trades off gain for linearity at such a rate that the gain is smaller than the CE and DP. In terms of bandwidth, the first pole of the Cascomp is inherently diminished, compared with the other topologies. If all circuits were moved to a RF Heterojunction Bipolar

Transistor (HBT) technology, we would expect a higher overall gain-bandwidth product, but comparably the Cascomp would still be less effective compared to other topologies. If the Cascomp was to be realised to compete in the RF amplifier market, where linearity and gain-bandwidth are important, we cannot say that we expect it to perform better.

7 CONCLUSIONS

The CE outperforms the DP in gain and TOI but not in THD, Signal to Noise and Distortion (SINAD) ratio and Spurious-Free Dynamic Range (SFDR). If the output signal is filtered, leaving only the carriers with the intermodulation products closest to them, then the CE is superior to the DP.

The Cascomp dominated the CE and DP in TOI, THD, SINAD and SFDR, even when the calculated TOI was extremely underestimated.

With the revelation of a secondary compensation point with no degeneration in the error amplifier, an alternative topology was created for the Cascomp. The new topology offers an increase in gain for a small decrease in TOI. This is a new result that has not appeared before in the open literature.

The gain-bandwidth product of the Cascomp is smaller than the CE and DP. The bandwidth of the Cascomp is inherently diminished due to a larger time constant through its signal path. If all circuits were moved to an RF HBT technology, we would expect a higher overall gain-bandwidth product, but the Cascomp will still be less effective than the other topologies, and is likely to have insufficient gain to allow broadbanding.

8 REFERENCES

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