Analog Integrated Circuits and Signal Processing 4, 161–166 (1993)

A Current-Mode Biquadratic Amplitude Equalizer

ADAM WYSZYŃSKI AND ROLF SCHAUMANN
Department of Electrical Engineering, Portland State University, Portland, OR 97207-0751

Received July 28, 1992; Revised

Abstract. A boost biquad transfer function with two symmetrical zeros on the real axis is derived; it provides a technique for amplitude equalization with a second-order continuous-time g_m-C section using current injection. The synthesis procedure yields a differential boost biquad consisting of three double-input OTAs and a single-input high-g_m boost OTA. SPICE simulations show that for boost gain up to 20 dB the phase of the biquad is essentially unaffected. The presented technique is found suitable for amplitude equalization of disk-drive read channels operating in the range of 100 MHz.

1. Introduction

In the design of linear or equiripple phase lowpass filters for disk-drive read channels, an amplitude equalization technique is required for pulse slimming, which improves pulse detection and increases disk write-density [1]. The technique commonly used for monolithic realization of these circuits employs continuous-time filters built with operational transconductance amplifiers (OTAs) and capacitors, g_m-C filters.

The method for equalizing the filter transfer function (a gain boost) proposed in [1] is to lift two capacitors off ground and to feed the signal from the input of the biquad forward to the lifted nodes through an inverting variable-gain voltage amplifier. These designs proved to be reliable for the active filter synthesis in the range up to a few tens of megahertz, but special care must be given to the design of the voltage amplifier. It requires very low output impedance, minimal delay, and low noise, which makes its design difficult. It would be much easier to use an OTA for the boost generation as this building block is already available and it has generally better high-frequency performance than a voltage amplifier. The design of such a boost biquad will be presented in the following.

2. Boost Biquad Transfer Function

The derivation of the boost biquad transfer function starts from a general g_m-C biquad structure given in [2]. By omitting one of the OTAs the simplified structure in figure 1 is obtained which can realize a pair of complex poles. As is well known [2], transmission zeros can be created without disturbing the poles by applying the input voltage V_i to any node lifted off ground (such as nodes a, b, c, d, or e in figure 1), or by injecting a current proportional to V_i into any internal circuit node (such as the nodes labeled V_o, V_1, or V_2). Using the latter method as shown in figure 1 to inject $I = -g_{m5}V_i$ into node V_o and setting $V_a = V_i$, $V_b = V_c = V_d = V_e = 0$ yields

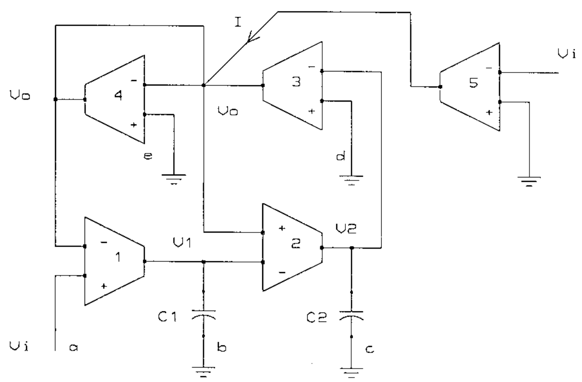

Fig. 1. The circuit diagram of the single-ended g_m-C biquad.

$$sC_1V_1 = g_{m1}(V_i - V_o) \tag{1}$$

$$sC_2V_2 = g_{m2}(V_o - V_1) \tag{2}$$

$$g_{m4}V_o + g_{m5}V_i = -g_{m3}V_2 \tag{3}$$

which solved for V_o results in

$$\frac{V_o}{V_i} = \frac{g_{m5}}{g_{m4}} \frac{-s^2 + g_{m1}g_{m2}g_{m3}/C_1C_2g_{m5}}{s^2 + sg_{m2}g_{m3}/g_{m4}C_2 + g_{m1}g_{m2}g_{m3}/g_{m4}C_1C_2} \tag{4}$$

The corresponding circuit is shown in the top part of figure 2. Setting finally $g_m = g_{m1} = g_{m2} = g_{m3} = g_{m4}$ and $g_{m5} = Kg_m$ yields

$$\frac{V_o}{V_i} = \frac{-Ks^2 + g_m^2/C_1C_2}{s^2 + sg_m/C_2 + g_m^2/C_1C_2} \tag{5}$$

The two real zeros are programmable by varying Kg_m of OTA5, which is equivalent to changing the coefficient K in the numerator of (5).

3. Boost Biquad Synthesis

Since integrated g_m-C filters should have fully-balanced design for the superior noise, distortion, power supply rejection ratio, and common-mode rejection ratio characteristics, it is mandatory to derive a fully balanced structure for the boost biquad. Its synthesis procedure starts with mirroring the single-ended biquad at ground and inverting all inputs and outputs of the OTAs (figure 2). Then the original and mirrored versions of the

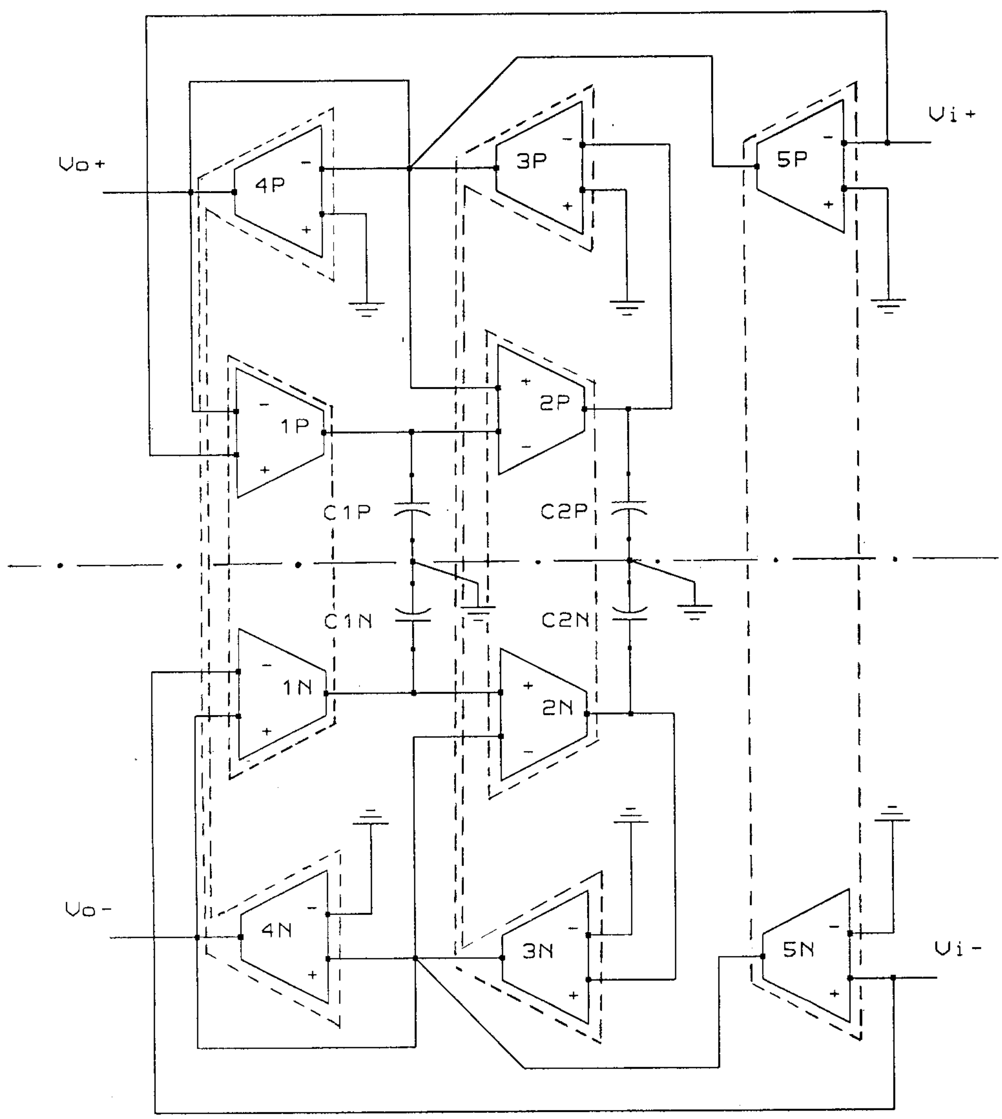

Fig. 2. The synthesis of the fully-balanced version of the boost biquad in figure 1. The dashed lines encircle merged elements. The horizontal symmetry line denotes the ground plane.

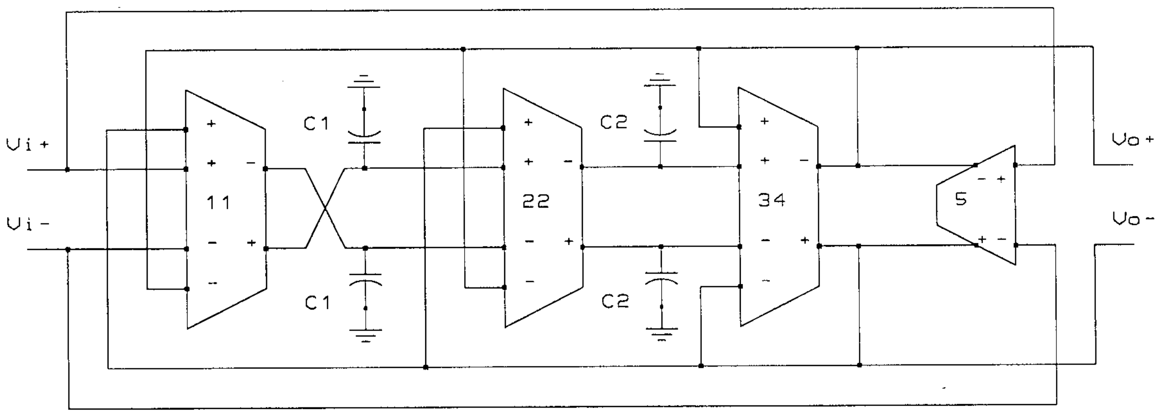

Fig. 3. The circuit diagram of the fully-balanced boost g_m-C biquad, with double-input OTAs and grounded capacitors.

biquad are combined along the ground plane which forms the symmetry axis. Note that all OTAs and capacitors in the upper part of figure 2 are denoted by the symbol P (positive), whereas OTAs in the bottom part are denoted by N (negative). OTA1P and OTA1N merge forming the differential double-input OTA11 (figure 3). Similarly, OTA2P combines with OTA2N yielding the differential double-input OTA22. Because OTA4P and OTA4N have one of their terminals grounded, OTA4P can be merged with OTA4N giving a differential single-input OTA4. A similar procedure applied to OTA3P, OTA3N and OTA5P, OTA5N results in the differential single-input OTA3 and OTA5 respectively. The merged elements in figure 2 are encircled with dashed lines. Since OTA3 and OTA4 have a common output node, these two OTAs can finally be replaced by one differential double-input OTA34 as was postulated in [3]. Also, it would be possible to merge OTA5 with OTA34 forming a differential triple-input OTA, but since the g_m-value of OTA5 should be different from that of the others if the boost is to be adjustable, this solution is not practical. Finally, the two pairs of grounded capacitors C_{1P}, C_{1N} and C_{2P}, C_{2N} can also be merged to form floating capacitors. However, for a monolithic realization grounded capacitors are preferred to floating ones as being free from parasitic bottom-plate capacitance. Thus, the final result of the above procedure is the biquad presented in figure 3 where the capacitors C_{1P}, C_{1N} and C_{2P}, C_{2N} are simply denoted by C_1 and C_2. It consists of three double-input OTAs (OTA11, OTA22, and OTA34) and single-input boost OTA (OTA5). From the structure of the boost biquad in figure 3 it can be seen that its operation involves injecting a current $I \propto -V_{in}$ to the output node (the resistor $1/g_{m4}$), thereby realizing the feedforward path. Note that the output of the biquad is not loaded with a capacitor. However, if g_m of the load resistor is sufficiently high and the parasitic capacitances at the output node are kept low the resulting phase degradation is small.

To illustrate the performance, the circuit in figure 3 was simulated on SPICE with the bipolar OTA in figure 4. Since g_m of the OTA in figure 4 is approximated by

$$g_m = \frac{1}{2R_1}\frac{I_{E2}}{I_{E1}} \tag{6}$$

different values of K are obtained by setting different ratios of bias currents I_{E1}, I_{E2}. I_{E1} is chosen to be 120 μA and I_{E2} varies from 200 to 2000 μA. Figure 5 shows the obtained result for $g_m = 0.985$ mS, $C_1 =$ 0.620 pF, $C_2 = 2.188$ pF, and values 0.84, 2.36, 5.35 for K. Evidently, the boost-gain varies as 4.87 dB, 12.34 dB to 19.60 dB, but the phase is essentially unaffected by the choice of K as required for read channel filters.

4. Conclusions

A synthesis of a useful block in g_m-C filter synthesis, the boost biquad, using only OTAs and capacitors is presented. The new biquad has improved frequency behavior since it does not use a difficult-to-design ideal voltage amplifier to obtain the boost. This in turn results in less distorted group delay characteristics of a filter with boost. The latter can be explained by the fact that phase errors of the amplifier due to its limited band-

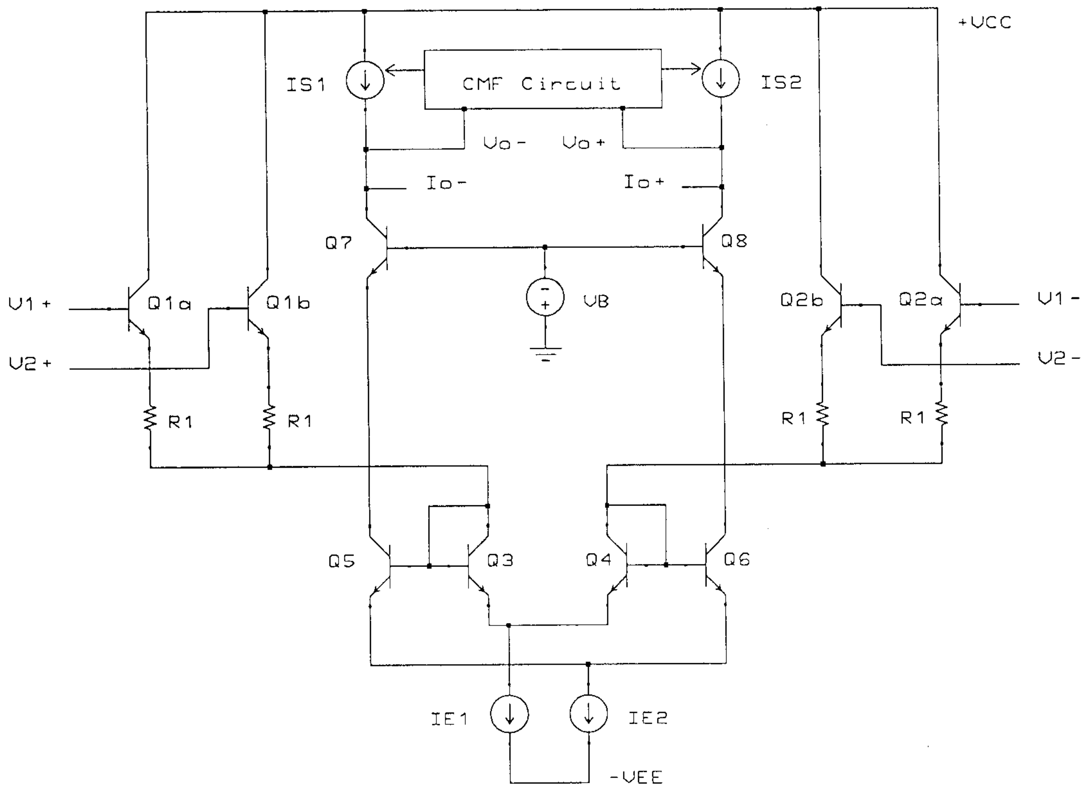

Fig. 4. The bipolar double-input fully-balanced OTA. The boost OTA is obtained simply by removing a pair of inputs V_2^+, V_2^- together with transistors $Q1b$, $Q2b$, and their emitter resistors.

width and nonzero output impedance are responsible for serious degradation of the group delay characteristic. The boost biquad is synthesized using multiple-input OTAs, which may result in substantial component and die area savings [3] as well as in some savings of the total power dissipated. As illustrated in figure 5 and as was demonstrated in [4] for a bipolar Bessel low-pass filter design, this approach is suitable for disk-drive read channel filters operating in the frequency range of 100 MHz.

Acknowledgement

This work was supported by a 1991/92 Tektronix Fellowship and by the NSF under grant HMP-9121360.

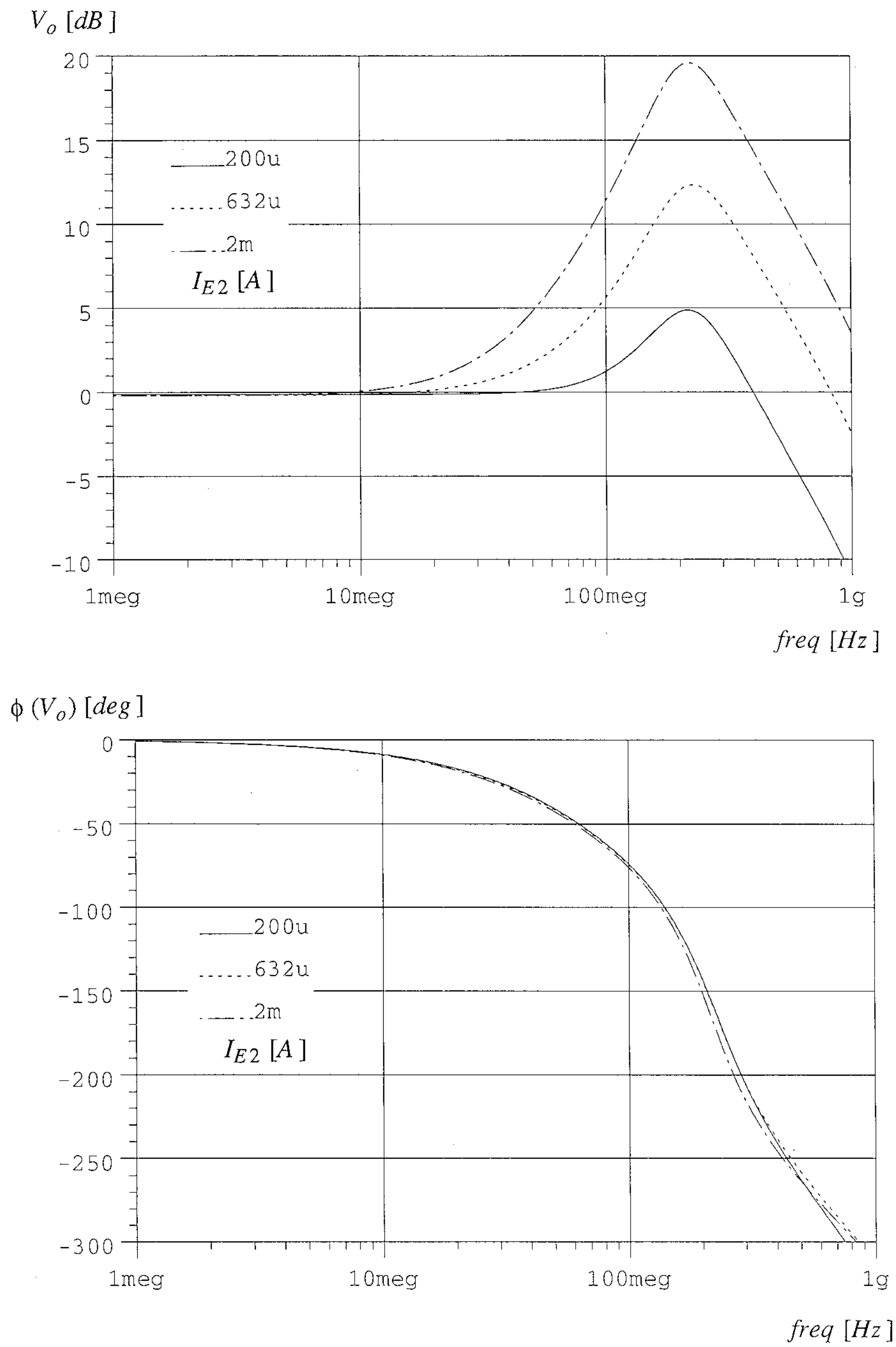

Fig. 5. The simulated frequency performance of the boost g_m-C biquad in figure 3: (a) the simulated gain of the output voltage V_o; (b) the simulated phase ϕ (V_o) of the output voltage. The parameter is the I_{E2} of the boost OTA.

References

1. G.A. De Veirman and R.G. Yamasaki, "Design of a bipolar 10-MHz programmable continuous-time 0.05° equiripple linear phase filter," *IEEE J. Solid-State Circuits*, Vol. 27, pp. 324–331, 1992.
2. R. Schaumann, M.S. Ghausi, and K.R. Laker, *Design of Analog Filters: Passive, Active, and Switched Capacitor*, Prentice Hall: Englewood Cliffs, NJ, 1990.
3. A. Wyszyński and R. Schaumann, "Using multiple-input transconductors to reduce the number of components in OTA-C filter design," *Electron. Lett.*, Vol. 28, pp. 217–220, 1992.
4. A. Wyszyński, R. Schaumann, S. Szczepański, and P. Van Halen, "Design of a 2.7 GHz linear OTA in bipolar transistor-array technology with lateral PNPs," *Proc. IEEE Int. Symp. Circ. Syst.*, pp. 2844–2847, 1992.

Adam Wyszyński received his M.S. degree in electrical engineering from Technical University of Warsaw, Poland, 1977. After graduation he was a research assistant at Technical University of Warsaw working on CAD systems for IC design. In 1980 he joined the Semiconductor Center CEMI in Warsaw as an R&D engineer responsible for development of automatic layout systems for LSI. He was also involved in the design of MOS digital circuits. From 1988 until 1990 he was with Cableship Ltd. in London, UK, as a software engineer for phone network management systems. He is now a research assistant at Portland State University preparing a Ph.D. thesis on monolithic implementation of analog filters. In 1991 and 1992 he served as a consultant for Zeelan Technology, Inc., Portland, OR, and for VTC, Inc., Bloomington, MN, developing analog circuit macromodels, BiCMOS transconductance amplifiers and analog filters. His areas of interest are analog circuit modeling and design, design of bipolar, CMOS, and BiCMOS transconductance amplifiers, design of analog integrated filters, automatic frequency and Q-tuning systems, and CAD for analog circuits. Mr. Wyszyński has published 13 papers and holds one patent, is a member of the IEEE Circuit and Systems Society, an Associate of the British IEE, and a member of Sigma Xi and Eta Kappa Nu.

Rolf Schaumann received the Diplom-Ingenieur degree from the University of Stuttgart, Germany, in 1967 and the Ph.D. degree from the University of Minnesota in 1970, both in electrical engineering. He was on the faculty of the Department of Electrical Engineering of the University of Minnesota, Minneapolis, from 1970 until 1988. In 1989, he accepted the position of professor and chairman of the Department of Electrical Engineering at Portland State University in Portland, Oregon. His teaching and research interests are in circuits and systems, filters, analog integrated circuits, modeling, statistical circuit design, and the realization of fully integrated analog filters. He has some 100 publications, coauthored the chapter "Active Filters" in *Reference Data for Radio Engineers*, 7th and 8th editions, Howard & Sams, 1985 and 1992, is coeditor of *Modern Active Filter Design*, IEEE Press, 1981, and is coauthor of *Analog Filters: Passive LC, Active RC and Switched Capacitor*, Prentice-Hall, 1990. Dr. Schaumann is a Fellow of the Institute of Electrical and Electronics Engineers (IEEE). He was the Editor-in-Chief of the *IEEE Transactions on Circuits and Systems*. He served on the Board of Governors of the Circuits and Systems (CAS) Society, was CAS Vice President for Publications in 1988 and 1989, and in 1990 was President of the CAS Society.