AMPLIFIER DESIGN CONSIDERATIONS FOR HIGH FREQUENCY MONOLITHIC FILTERS

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ABSTRACT

The fundamental limits of the most common CMOS amplifier structures for high frequency monolithic continuous–time filter applications are examined. Applications of these amplifiers in integrator–based filters are investigated. It is shown that transconductance amplifiers (TA's) have inherently better high frequency response than operational voltage amplifiers, and that TA based open–loop integrators are better suited for high frequency applications than are closed loop feedback integrator structures.

I. INTRODUCTION

Implementation of high frequency ($\gg$ 100 kHz) continuous–time filters in a CMOS process has applications in AM and FM IF filters, TV signal processing, as well as other telecommunication and instrumentation applications. In these frequency ranges, the aliasing problem inherent in sampled data filters along with other known limitations of existing sampled–data technologies make a continuous–time filter approach attractive. Recent publications report simulated designs [1],[2] and experimental results [3],[4] of monolithic CMOS continuous–time filters designed to operate at frequencies up to 1 MHz. Devices termed "operational amplifiers" (op amps) and transconductance amplifiers (TA's) have been reported as basic gain blocks. Analysis of the basic gain block topologies provides insight into which are best suited for high frequency filtering as applications enter the MHz range. In this paper, the term *op amp* will be used to denote any amplifier with very high gain which is intended for use in applications where the characteristics of the circuit using the device are ideally independent of the amplifier gain. When distinction of *op amp* variety is necessary, the amplifier type will be included, such as Operational Transconductance Amplifier (OTA) and Operational Voltage Amplifier (OVA).

In the following section, the frequency response of published high frequency gain block topologies is analyzed. Emphasis is placed on small signal performance at high frequencies. Two families of gain blocks provide excellent bandwidth and little excess phase. In Section III, we examine the two families of high frequency gain blocks in integrator topologies. One structure is found useful as a high frequency building block.

II. GAIN BLOCK COMPARISON

Most published high frequency gain blocks can be divided into 3 categories: transconductance amplifiers (including both the TA and OTA), traditional two–stage op amps, and the newer single–stage "op amps". Each category will be identified by a basic topology and transfer function. The transconductance category covers all transconductance circuits from the simple differential amplifier [3] to the more complex TAs [1],[4]. These can be constructed from transconductance subcircuits and current mirrors, excluding any internal high impedance nodes. The traditional two–stage op amp (OVA) incorporates an internal high impedance node used for compensation. The newer single–stage "op amp" category is comprised of operational amplifiers with high impedance outputs which are compensated by load capacitance. Included in this category are the stacked mirror [5] and the folded–cascode op amps [6]. These latter structures also find applications as transconductance amplifiers in circuits where the transfer characteristics of the circuits depend directly on the gain of the amplifiers [7].

TAs and OTAs

A generic TA block diagram is shown in Fig. 1. This topology utilizes an input transconductance stage followed by current mirrors which steer currents to a high impedance output. The transconductance and current mirror subcircuits have wide transfer function bandwidths and no internal high impedance nodes, so the overall TA bandwidth is very large. The transconductance gain can be modelled by

$$g_m(s) = \frac{g_{mo}}{1 + s/\omega_g} \tag{1}$$

where ω_g is an "effective" pole characterizing several very high frequency internal mirror poles. This lumped model is mathematically tractable and reasonably accurate to nearly ω_g. These parasitic poles are sufficiently large to justify using the single pole model well into the MHz range. Unfortunately, a single pole model cannot be justified in this range for the remaining gain blocks discussed in this section.

Traditional Two–Stage Op amps

A conventional two–stage op amp (OVA) is shown in Fig. 2a and its small signal equivalent circuit in Fig. 2b. The two–stage OVA is characterized by an internal high impedance node used for compensation. When no nulling resistor (R_Z) is used it is well known that the op amp has a dominant pole (p_1), a load–dependent second pole (p_2) and a RHP zero :

$$p_1 \approx \frac{-1}{C_C R_1 g_{m2} R_2}\ , \quad p_2 \approx \frac{-g_{m2}}{C_2}\ , \quad z = \frac{g_{m2}}{C_C} \tag{2}$$

assuming that C_1 is much smaller than either C_2 or C_C, and that the poles are widely spaced. The RHP zero is often close to the non–dominant pole (p_2) which seriously degrades the phase margin. Addition of a series nulling resistor (with value $R_Z = 1/g_{m2}$) removes the RHP zero and creates a third pole at $p_3 \approx -1/R_Z C_1 = -g_{m2}/C_1$. Since C_1 is usually much smaller than C_C, the phase margin improves significantly.

The use of two–stage OVAs (even with the improved compensation technique) in high frequency filters is limited by their gain–bandwidth product (GB). OVA based active RC filters show large GB–induced filter characteristic shifts when GBs of the OVAs are close to the desired filter critical frequency. Increasing the OVA GB, given by $GB = g_{m1}/C_C$, requires a very large input stage transconductance (g_{m1}) and a small compensation capacitor (C_C). But both large g_{m1} and small C_C make stabilizing the resultant op amp very difficult. The parasitic pole positions (p_2 and p_3) relative to the OVA GB determine stability. The normalized poles are:

$$\left|\frac{p_2}{GB}\right| = \frac{g_{m2}}{g_{m1}}\frac{C_C}{C_2}\ , \quad \left|\frac{p_3}{GB}\right| = \frac{C_C}{g_{m1} R_Z C_1} = \frac{g_{m2}}{g_{m1}}\frac{C_C}{C_1} \tag{3}$$

The non-dominant OVA poles (p_2 and p_3) must be far above the filter critical frequency to avoid excess phase, and well above the OVA GB for good op amp phase margin. The C_2 capacitor includes the load capacitance and the drain-bulk capacitances of the (typically very wide) output devices. Its substantial size tends to constrain the p_2/GB ratio near unity, limiting the useful frequency range of the two-stage OVA.

The bandwidth achievable from a two-stage OVA is inferior to that of the OTA because of the additional internal high impedance node and compensation capacitor. The single parasitic pole in the OTA depends only on device parasitics, and is located well above the unity gain frequency (assuming a capacitive load) at ω_g. In contrast, the first (non-dominant) parasitic pole in the two-stage OVA depends on the large C_2 capacitance. There is also a pole due to the OVA current mirror near $2\omega_g$ (not included in the model of Fig. 2b), and a third (p_3) pole (usually around ω_g). Consequently, the design of a two-stage OVA with large GB and sufficient phase margin for critical frequency applications above 1 MHz appears to be very difficult.

Single-Stage Op amps

The newer single-stage op amps are characterized by high impedance outputs and compensation by load capacitance. Very significant improvements in bandwidth over what is attainable with the two-stage op amps have been reported in the literature with gain-bandwidth products beyond 50 MHz being common [7]. These circuits are topologically equivalent to the TA block diagram of Fig. 1. No internal compensation capacitor is required because there are no internal high impedance nodes. Differences between the various single-stage op amps are mainly due to choice of current mirrors, and the style of common mode feedback used in fully differential designs. The frequency response of the single-stage op amp is similar to that of the topologically equivalent TA and is given by (1). The bandwidth is limited by the same parasitic (mirror) poles affecting the TAs, with some differences in ω_g due to the type of current mirror selected.

In summary, from a high frequency performance perspective there is no preference between the topologically identical TAs and the single-stage op amps as both exhibit very good high frequency performance and have no internal high impedance nodes. The major difference is due to the names applied to these devices by the authors. The two-stage OVAs were shown to have much smaller usable bandwidths than the single-stage op amps, the TAs and the OTAs because of their internal high impedance node and compensation capacitor.

III. INTEGRATOR COMPARISON

The fundamental building block used in most active filters is the integrator. Most integrators discussed in the literature are either based upon the Miller (feedback) structure of Fig. 3a or the transconductance (open loop) structure of Fig. 3b or some minor variant of these structures. For either type , the unity gain frequency of the integrator is the key parameter which determines filter characteristics. The Miller structure of Fig. 3a requires a very high gain operational amplifier. The structure employs a large amount of feedback to render a gain expression $A_V = V_o/V_i = -1/sR_SC_F$, where the unity gain frequency, $\omega_o = 1/R_SC_F$, is independent of the gain of the op amp. The TA based integrator of Fig. 3b has a gain expression of the form $A_V = V_o/V_i = g_m/sC_L$. This is an open loop rather than a feedback structure, in which the voltage gain is directly dependent upon the transconductance gain g_m.

Closed Loop (Miller) Integrator

A traditional closed loop (Miller) integrator is shown in Fig. 3a, and the corresponding small signal model is in Fig. 4a. Loading by an identical stage is assumed, which introduces the load resistor (R_L) and the load capacitor (C_L). Two options for the op amp will be considered. If the two-stage internally compensated amplifier (OVA) is used, it is well known from basic active filter theory that practical unity gain frequencies are typically limited to around .01GB. Even if modifications of the basic integrator structure are made, it is difficult to obtain satisfactory operation above 0.1GB. Since the GB of the OVA is typically limited to the MHz range, the Miller integrator with an OVA is not useful at high frequencies.

As a second option, the transconductance amplifiers or the OTA type op amp such as the single–stage "op amps" will be considered. In this case, an expression for the voltage gain of the closed loop integrator is

$$A_V = \frac{V_o}{V_i} = \frac{\frac{1}{R_S C_L}[s^2 + s\omega_g - g_{mo}\omega_g/C_F]}{s^3 + s^2\alpha_2 + s\alpha_1 + g'_L\omega_g/R_S C_L C_F} \tag{4}$$

with $\alpha_2 = \omega_g + \frac{g'_L}{C_L} + \frac{1}{R_S C_{eq}}$, $\alpha_1 = \omega_g\left(\frac{g_{mo} + g'_L}{C_L} + \frac{1}{R_S C_{eq}}\right) + \frac{g'_L}{R_S C_F C_L}$

where $1/g'_L = R_L \parallel r_o$ and $1/C_{eq} = (1/C_F) + (1/C_L)$. By assuming a dominant low frequency pole the closed loop integrator poles and zeros can be approximated by :

$$p_1 = \frac{-g'_L}{R_S g_{mo} C_F} \quad , \quad p_{2,3} = \frac{-\omega_g}{2}\left[1 \pm j\sqrt{\frac{4g_{mo}/C_L}{\omega_g} - 1}\right] \tag{5a}$$

$$z_{1,2} = \frac{-\omega_g}{2}\left[1 \pm \sqrt{1 + \frac{4g_{mo}/C_F}{\omega_g}}\right] \tag{5b}$$

Open Loop Integrator

The open loop integrator is constructed from a transconductance amplifier with a grounded load capacitor. It is depicted with a TA gain block in Fig. 3b, and a corresponding small signal model is shown in Fig. 4b. The stacked current mirrors gives rise to a large output resistance (r_o) and the frequency dependence of the transconductance gain is modeled in (1). Loading by an identical stage will not affect the circuit model, only the effective load capacitance. The integrator voltage gain is :

$$A_V = \frac{V_o}{V_i} = \frac{g_{mo} r_o}{(1 + s/\omega_g)(1 + s r_o C_L)} \tag{6}$$

and the unity gain frequency is approximately g_{mo}/C_L.

The closed loop (Miller) integrator and the open loop integrator can now be compared based on their pole/zero distributions. The position and number of high frequency integrator singularities relative to the filter critical frequency determines the excess phase and the derivative of the phase at the critical frequency. It is well known that very small amounts of excess phase ($< 1°$) can cause serious shifts in filter characteristics. The open loop integrator has a single high frequency parasitic pole at ω_g. The closed loop integrator has two high frequency parasitic poles, a LHP zero and a RHP zero given by (5). Note that large values of g_{mo}/C_L and g_{mo}/C_F give high frequency poles ($p_{2,3}$) and zeros ($z_{1,2}$) well above ω_g. Unfortunately, this makes the Miller integrator unstable, as can be seen by the loop gain

$$A_V(o.l.) = \frac{-s g_{mo}/C_L}{(1 + s/\omega_g)\left[s^2 + s\left\{g'_L + \frac{1}{R_S C_L}(1 + \frac{C_L}{C_F})\right\} + \frac{g'_L}{R_S C_L C_F}\right]} \tag{7}$$

It is clear that large values of ω_g produce a 90° phase margin, $\omega_g = g_{mo}/C_L$ produces a 45° phase margin, and ω_g much below g_{mo}/C_L may cause instability if higher–order poles are present. To facilitate comparison between the open and closed loop integrators, values for C_F and C_L in the closed loop equations must be specified. The maximum high frequency performance from the closed loop integrator will be obtained as the op amp phase margin reaches its minimum acceptable value (practically 45°) and for minimum C_F. The lowest realistic value for the feedback capacitor will be defined as C_L. This is probably overly optimistic, but will establish an upper performance bound for the closed loop integrator.

Based on these assumptions, the pole/zero root locus plots for both integrator topologies are shown in Fig. 5 relative to the ω_g of the OTA. Aside from the low frequency dominant pole, the open loop integrator has one high frequency parasitic pole at ω_g. The closed loop integrator has a RHP zero at $.62\omega_g$, a pair of complex poles ($Q = 1$) with magnitude ω_g, and a LHP zero at $1.62\omega_g$. For a given critical frequency it may be possible to design both with zero excess phase, but the slope of the closed loop integrator phase function will be at least double that of the corresponding phase function in the open loop integrator because of the additional singularities near ω_g. Process parameter variations and statistical mismatches between supposed matched devices will cause phase slope to be a significant limiting factor in the realizability and tunability of high frequency monolithic designs. In practice, having fewer significant high frequency singularities (and at higher frequencies) means that open loop integrators will exhibit less excess phase at a given critical frequency and be feasible at higher frequencies than closed loop integrators. For the pole/zero positions of Fig. 5, it is possible to calculate the frequencies at which the excess phase reaches .5° (called f_{ep}). For the open loop integrator, $f_{ep} \approx \omega_g/100$, and for the closed loop integrator, $f_{ep} \approx \omega_g/200$. Since it was observed that the stated conditions for the closed integrator provide only marginally acceptable stability to improve the excess phase of the overall structure, it can be stated that the .5° phase error frequency for the open loop integrator is at least twice the .5° frequency of the closed loop integrator.

V. CONCLUSION

An examination of the reported high frequency gain blocks, including OTAs, the newer single–stage OVAs, and the traditional two–stage OVAs shows results consistant with previous observations. Because of the internal high impedance node and accompanying compensation capacitor, the two–stage OVA is an inferior high frequency gain block. The OTA and single–stage OVA are topologically equivalent, differing only by the definition of the output quantity (current or voltage). Minor differences occur due to the choice of internal current mirrors, but they generally have better bandwidth and less excess phase than the two–stage OVAs.

Using a gain block with the TA topology (Fig. 1) allows construction of both open and closed loop integrators. The open loop integrator was shown superior to the closed loop (Miller) integrator using reported high frequency gain blocks. The open loop (TA style) integrator most naturally utilizes a transconductance based description of the gain block, implying that the "best" suited high frequency gain blocks are TAs (employed in open loop integrators).

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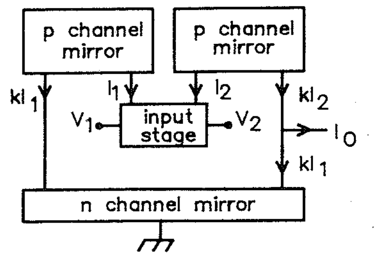

Fig. 1 Generic TA block diagram.

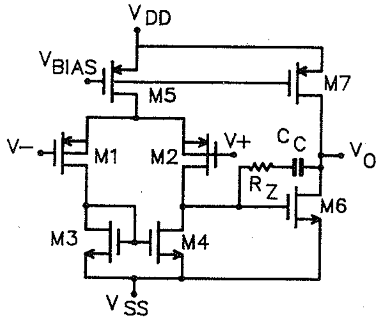

Fig. 2 (a) Conventional two–stage op amp.

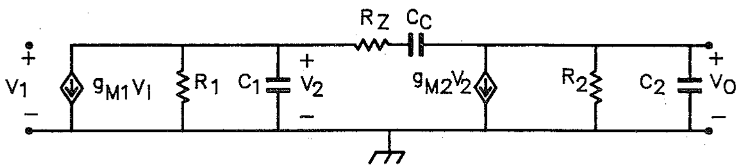

Fig. 2 (b) Small signal equivalent circuit.

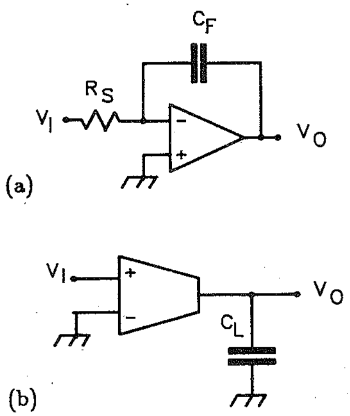

Fig. 3 (a) Closed loop (Miller) integrator.
(b) Open loop integrator.

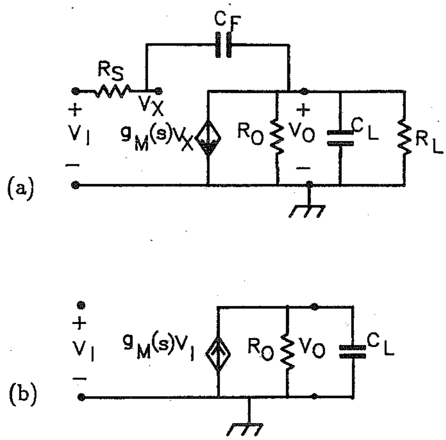

Fig. 4 Small signal equivalent circuits :
(a) Miller integrator.
(b) Open loop integrator.

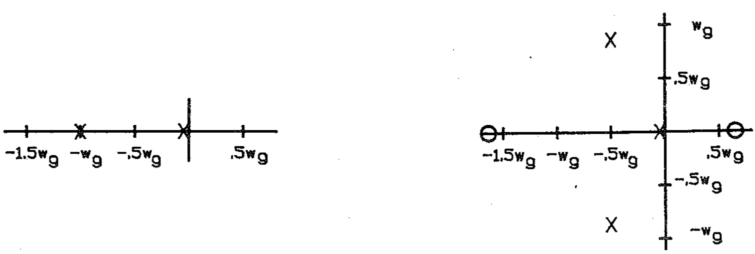

Fig. 5 (a) OTA integrator pole plot. (b) Miller integrator pole/zero plot.