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A Low-Power AGC with Level-Independent Phase Margin

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***Abstract*— We describe a 32-μW 77dB-dynamic-range power-law-compression VLSI AGC circuit for use in audio applications particularly for hearing aids and cochlear implants. We demonstrate that in spite of the nonlinearities in the circuit, the feedback-loop phase margin is invariant with input level. This ensures that robust performance at one level guarantees robust performance at any level.**

I. INTRODUCTION

Gain control circuits for audio applications have two main purposes. First, adaptive compression limits the overall instantaneous input dynamic range to prevent distortion of the audio signal. Second, amplification at the lowest amplitude allows the listener to hear faint sounds even with reduced hearing capability. An additional benefit of compression is reduced instantaneous dynamic range making possible low-power signal processing for portable applications.

Unlike communications applications, gain control systems for audio require gentle compression to prevent audible distortion [1]. Previous work has examined stability and transient dynamics of AGC loops for RF applications[2], [3]. As these systems are intended for processing of radio-frequency signals at a fixed level, these AGCs exhibit highly compressive characteristics. Audio applications require a more gentle compressive function to permit patients to perceive loud and soft levels. Gain control systems for speech signals require carefully chosen adaptive control strategies to maintain audible and clear speech signals. The time constants in the feedback control must be controlled carefully to ensure good distortion performance and resilience to loud transients.

The analysis presented in this paper is a simplified linear adaptation of the general linear theory presented in [4]. Section 2 describes the overall system and circuits. Section 3 describes the linear analysis performed. Section 4 presents experimental results from a chip. Section 5 concludes the paper.

II. NONLINEAR MODEL

Figure 1 shows a block diagram for the translinear feedback AGC of interest. The variable-gain amplifier (VGA) acts to multiply the input signal v_{IN} by the gain-control signal a_V. This VGA is shown in Figure 2. Two coupled wide-linear range transconductors [5] are configured to produce voltage gain proportional to a_V/I_{Ref2}. The first transconductor operates as a programmable voltage-to-current converter. This output current drives the second transconductor which is configured in negative feedback to implement a resistance. Thus, an amplifier with variable gain $G_M R$ is obtained. The dynamic range of this structure is not explored in detail here. However, the noise and distortion are limited so as to provide 77dB of dynamic range at the input.

The output voltage v_{OUT} is rectified and converted to a current with transconductance G_{ED}[6], [7]. This rectified output is then filtered with asymmetric-attack-and-release low-pass filters. These filters apply an attack time constant τ_a when the envelope amplitude is increasing, and a release time constant τ_r when the envelope amplitude is decreasing

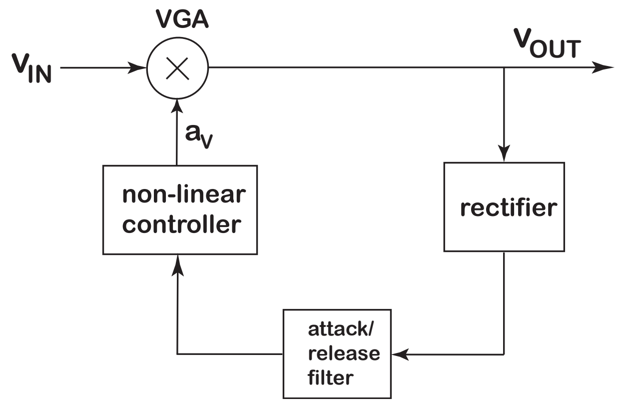

Fig. 1. The feedback AGC consists of a variable-gain amplifier, a rectifier, an asymmetric-attack-and-release filter, and a nonlinear control element.

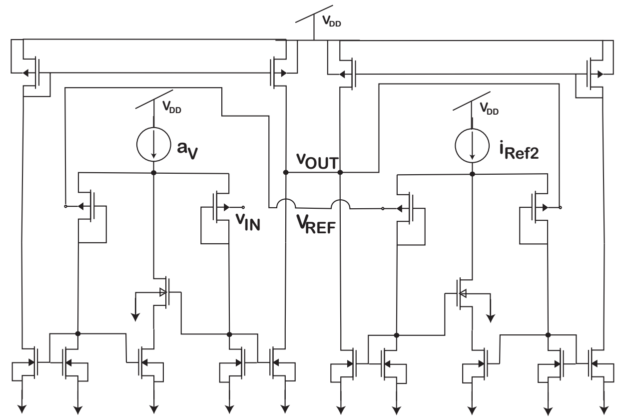

Fig. 2. The variable gain amplifier topology is shown. The transconductors are designed after [5].

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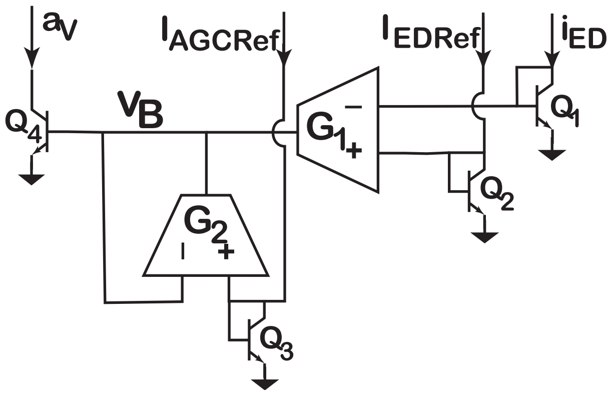

Fig. 3. The translinear control circuit consists of an input from the envelope detector i_{ED} and several reference currents. The output current is employed directly in the variable gain amplifier.

with a current-mode dynamic translinear circuit discussed in [7]. Detecting the output envelope rather than the input envelope reduces dynamic range requirements as the output dynamic range of the AGC is less than the input dynamic range.

The nonlinear controller operates on the resulting peak-detected current output of the attack-and-release filters, i_{ED}, to yield a gain a_V given by,

$$a_V = K\left(\frac{I_{EDRef}}{i_{ED}}\right)^{\alpha}. \tag{1}$$

The circuits of Figure 2 and 3 implement $\alpha = G_1/G_2$ and $K = I_{AGCRef}/I_{Ref2}$. With this choice of translinear control, we can determine the amplitude response of the system, E_{OUT}, to constant-amplitude sinusoidal inputs with amplitude E_{IN} to be,

$$E_{OUT} = (E_{IN}K)^{\frac{1}{1+\alpha}}\left(\frac{I_{EDRef}}{G_{ED}}\right)^{\frac{\alpha}{1+\alpha}}. \tag{2}$$

For a compressive transfer characteristic we set $\alpha > 0$, which is equivalent to choosing transconductances G_1 and G_2 both greater than zero. Note that when $i_{ED} = I_{EDRef}$, the transconductor, G_1, contributes no current to the integrating node v_B in Figure 3. This property ensures that the gain a_V is K which we typically choose to be 1. This decouples our minimum gain at large amplitudes where $i_{ED} = I_{EDRef}$ from the compression ratio.

III. Linear Analysis

To evaluate the response of the AGC to transients in the envelope amplitude, we need a simple formulation for the closed-loop properties of our system. The behavior of similar non-linear feedback systems for gain-control have been studied for applications in audio, sonar [8], [9], and narrowband RF applications [2]. In each of these works, the use of a logarithmic-exponential feedback element allows conclusions about the behavior of closed-loop feedback time-constants to be drawn by working in the log-domain. In our paper we simply linearize the system about an operating point with a certain input and output amplitude.

In this work, we have two low-pass elements in the feedback path, the desired peak-detector time-constant τ_a, and a time-constant associated with our translinear control system, τ_t. For this analysis we assume the input is a sinusoid with amplitude $e_{IN}(t)$ and the output is a sinusoid with amplitude $e_{OUT}(t)$.

A. Variable Gain Section

The variable gain section performs a multiplication of v_{IN} with a_V. Each of the variables involved in the multiplication may be decomposed into their large-signal and small-signal components. That is,

$$v_{OUT} = (V_{IN} + v_{in})(A_V + a_v) \tag{3}$$
$$v_{OUT} = (V_{IN}A_V) + (V_{IN}a_v + v_{in}A_V) + (v_{in}a_v) \tag{4}$$
$$V_{OUT} = V_{IN}A_V \tag{5}$$
$$v_{out} \simeq V_{IN}a_v + v_{in}A_V \tag{6}$$

Note that we have ignored the small-signal cross term $v_{in}a_v$ for our linear model and used IEEE notation for large-signal and small-signal variables. For constant amplitude sinusoids we can also write,

$$e_{out} \simeq E_{IN}a_v + e_{in}A_V. \tag{7}$$

The linearized feedback model of the multiplier described in (7) is shown as the adder block of Fig. 4.

B. Translinear control circuit

A linearized model of the translinear circuit can also be computed, in terms of the input-envelope amplitude by differentiating (1):

$$\left[\frac{da_V}{di_{ED}}\right]_{I_{ED}} = \frac{d}{di_{ED}}K\left(\frac{I_{EDRef}}{i_{ED}}\right)^{\alpha} \tag{8}$$
$$= \frac{-\alpha K I_{EDRef}^{\alpha}}{I_{ED}^{\alpha+1}} \tag{9}$$
$$= \frac{-\alpha}{I_{ED}}A_V. \tag{10}$$

The form in Eq. (10) corresponds to the dependence on the input from the $-\alpha/I_{ED}$ term and the operating point A_V. The operating point term I_{ED} can be written in terms of the circuit input envelope level and the operating gain,

$$I_{ED} = G_{ED}E_{OUT} = G_{ED}E_{IN}A_V. \tag{11}$$

Some algebraic manipulations allow simplification of (10) to

$$\left[\frac{da_V}{di_{ED}}\right]_{I_{ED}} = -\frac{\alpha}{G_{ED}E_{IN}}. \tag{12}$$

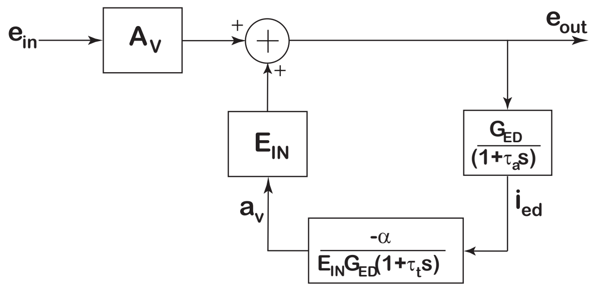

Fig. 4. Linearized loop model.

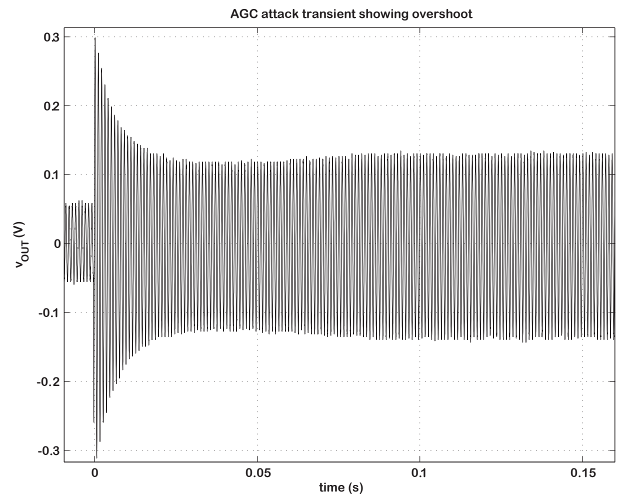

Fig. 5. An attack transient showing some slight overshoot in the envelope control.

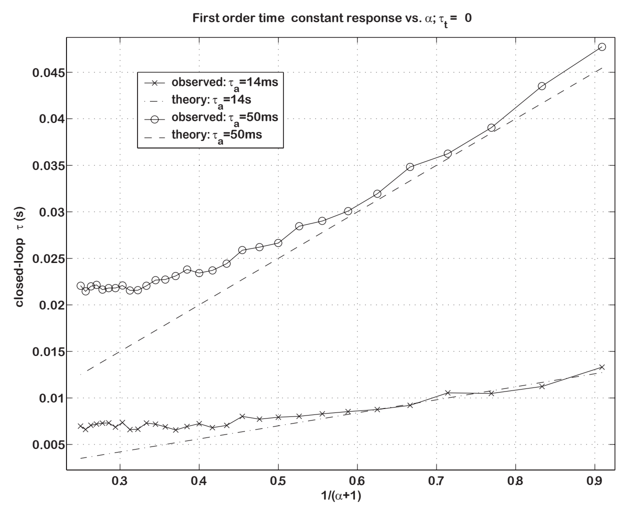

Fig. 6. Transient attack time constant varies with the compression ratio $1/(\alpha+1)$. At high compression ratios, near the left side of the plot, the peak detector time constant varies with amplitude resulting in deviations from the fit.

C. Model of the Loop Transmission

A complete model of the linearized system, excluding the asymmetric-release dynamics, is shown in Figure 4. The system yields a loop transmission $L(s)$ independent of both the input level E_{IN} and the envelope detector sensitivity G_{ED}. The loop transmission depends only on α and the time constants τ_a and τ_t,

$$L(s) = \frac{-\alpha}{(1+\tau_a s)(1+\tau_t s)}. \tag{13}$$

The form in Eq. 13 indicates a loop-gain dependence on the compression factor, α. When the translinear circuit time constant is very fast, i.e. $\tau_t \simeq 0$, the effective closed-loop time constant is $\tau_a/(1+\alpha)$. If τ_a and τ_t are both significant, a second-order response results. Critical damping arises when

$$\alpha = \frac{(\tau_a+\tau_t)^2}{4\tau_a\tau_t} - 1. \tag{14}$$

It is worth noting why the loop transmission $L(s)$ has such a simple form and is independent of G_{ED} and E_{IN}: A high value of either G_{ED} or E_{IN} attempt to increase the loop transmission by increasing multiplier or rectifier gain. However, the $1/i_{ED}^{\alpha}$ power-law gain-control causes the circuit to equilibrate at a point where the derivative of the translinear controller drops in inverse proportion to G_{ED} or E_{IN} thus exactly cancelling the attempted increase in loop transmission.

IV. Results

Experimental observations of the transient behavior of the loop were carried out. We fabricated the VLSI circuitry in a MOSIS AMI 1.5 μm process. A representative attack transient is shown in Figure 5. We have intentionally shown a case with second-order dynamics, i.e. with overshoot in the envelope response.

A. First-Order Dynamics ($\tau_t \simeq 0$)

We extracted the attack transient envelope and fit it to a first order response while varying α and τ_a. The translinear time constant τ_t was set artificially fast, to ensure that it had little effect on the response. Figure 6 shows two representative curves for varying α. The deviation from predicted linear behavior is due to a change in τ_a in the peak detector versus level, assumed by us to be constant.

B. Second-Order Dynamics (τ_t and τ_a significant)

Figure 7 shows the overshoot performance of the loop for a variety of input levels. The upward trend in these data reflect changes in τ_t, which decreases slightly as the input amplitude decreases owing to changes in the gain current.

V. Conclusions

The transient response properties for small changes in the envelope amplitudes of our AGC system were predicted from a linearized feedback analysis to be amplitude invariant. We confirmed these predictions experimentally over all input amplitudes where the open-loop time constants were roughly constant.

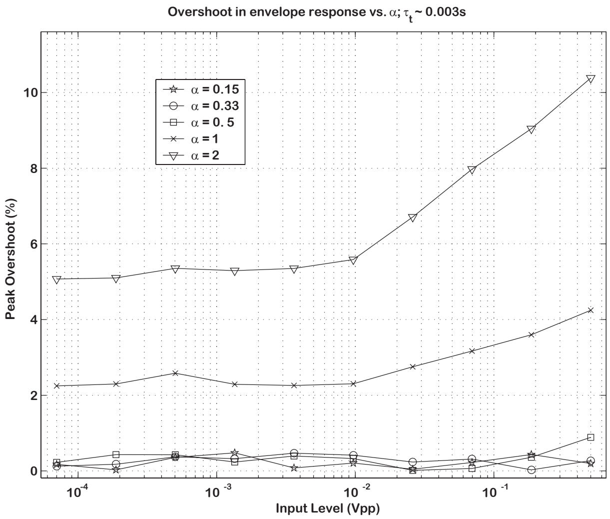

Fig. 7. The envelope overshoot is roughly invariant with input level over 40dB of dynamic range. Above 10mVpp, the translinear control pole $(1/\tau_t)$ becomes slower as the gain is decreased, leading to increased overshoot especially at high compression ratios where output amplitudes are consistently large.

References

[1] B. C. J. Moore "An Introduction to the Psychology of Hearing," *Academic Press, Cambridge, UK,* 3rd Ed. 1989.

[2] E. Bertran, J. M. Palacin, "Control theory applied to the design of AGC circuits," *Proceedings of the 6th Mediterranean Electrotechnical Conference,* IEEE, 1991.

[3] Douglas N. Green, "Global stability analysis of automatic gain control circuits," *IEEE Transactions on Circuits and Systems,* Vol. CAS-30, No. 2, Feb. 1983.

[4] John E. Ohlson, "Exact dynamics of automatic gain control," *IEEE Transactions on Communications,* Jan. 1974.

[5] R. Sarpeshkar, R. F. Lyon, C. A. Mead, "A low-power wide-linear-range transconductance amplifier," *Analog Integrated Circuits and Signal Processing,* Vol. 13, 123–151, 1997.

[6] M. W. Baker, S. Zhak, R. Sarpeshkar "A micropower envelope detector for audio applications," *International Symposium on Circuits and Systems 2003,* Vol. 5, May 25-28, 2003.

[7] S. Zhak, M. W. Baker, R. Sarpeshkar "A low-power wide dynamic range envelope detector," *IEEE Journal of Solid State Circuits,* Vol. 38 N. 10, Oct. 2003.

[8] J. Moskowitz "Linear feedback AGC with input-level-invariant response times," *Journal of the Acoustical Society of America,* Vol.62, No.6, 1977.

[9] D. W. Ricker "Nonlinear feedback system for the normalization of active sonar returns," *Journal of the Acoustical Society of America,* Vol.59, No.2, 1977.