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Wide-range precision RF peak detectors☆

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ABSTRACT

A significant increase in the accuracy of a conventional bipolar monolithic radio-frequency bipolar peak detector is robustly achieved over a wide amplitude range using a simple nonlinear loading circuit. The loading circuit provides two voltage-controlled auxiliary currents which cancel the errors by altering the original DC currents flowing through the core transistors, according to the level of the detected voltage. Expressions are derived for the auxiliary currents for perfect error cancellation; a good approximation is provided by a simple cross-coupled transistor pair, but more complex circuit structures can be employed for enhanced range and/or robustness. Because the loading circuit operates essentially at DC, the high-frequency nature of the original peak detector is preserved. The presented concept and its various circuit implementations are discussed and validated in simulation, and are expected to work in a large variety of modern bipolar or BiCMOS integrated-circuit technologies, where parasitics-free, simple exponential-law models accurately represent the operation of the transistors up to very high frequencies.

1. Introduction

Peak detectors are widely used in communication systems for functions such as amplitude stabilization in oscillators, level measurement, demodulation, automatic tuning, automatic gain control, and power amplifier envelope tracking [1–4]. Typically, signals encountered in high-frequency narrowband systems are either unmodulated sinusoids (in the case of oscillators), or modulated signals using a sinusoidal carrier. Peak detectors operating with such signals must usually meet conflicting requirements of precision, speed, dynamic range, and power consumption, but existing circuit topologies have limitations. A high-performance monolithic low-power radio-frequency (RF) peak detector was presented and mathematically analyzed in Ref. [5], revealing the difficulty of ensuring accuracy at even moderately-low input levels; more complicated circuits attempting mathematical corrections do improve detector slope, but not necessarily accuracy [6]. This paper introduces a robust concept for improving the accuracy of the conventional circuit over a wide range, which in its simplest form requires only a small number of additional components.

2. Conventional architectures and their limitations

The starting point for the present work is the virtually ubiquitous bipolar peak detector shown in Fig. 1: the circuit is simple, low-power, capable of operating up to very high frequencies, and accurate mathematical expressions for its output voltage—although not necessarily easy to use—are available. Moreover, based on the subthreshold and bipolar similarities, the same topology is also widely used in MOS technologies, where operating the transistors in the saturation rather than the subthreshold region can also be exploited [7].

With commonly-encountered input sinusoidal signals, however, it has been shown that the conventional detector of Fig. 1 has limited accuracy due to the presence of an amplitude-dependent term in the expression of the output voltage. Thus, with $Q_1 \equiv Q_2$, $I_1 = I_2$, for a sinusoidal input amplitude V_{IN}, and with C_{IN} a virtual short circuit at the frequencies of interest, the DC output voltage can be accurately calculated as in Ref. [5]:

$$V_{OUT} = V_T \ln\left(I_{Bessel0}\left(\frac{V_{IN}}{V_T}\right)\right), \tag{1}$$

where $I_{Bessel0}$ is the modified Bessel function of order 0, and $V_T = kT/q$ is the thermal voltage. Alternatively, V_{OUT} can be written in the more convenient form:

$$V_{OUT} = V_{IN} - V_{error}, \tag{2}$$

where the amplitude-dependent error term V_{error} is:

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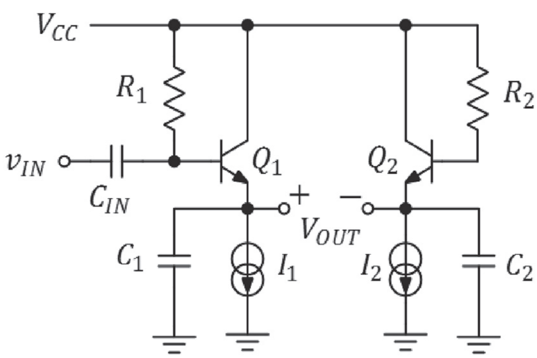

Fig. 1. Conventional low-power RF peak detector with AC-coupled input.

$$V_{error} = V_T \ln \frac{e^{\frac{V_{IN}}{V_T}}}{I_{Bessel0}\left(\frac{V_{IN}}{V_T}\right)}. \tag{3}$$

If the large-argument approximation is made for the Bessel function (not entirely appropriate at V_{IN} approaching $2V_T$), then V_{error} can be approximated by the closed-form expression $V_T \ln \sqrt{2\pi V_{IN}/V_T}$; however, this approximation is avoided in this work because of the rather significant inaccuracies at low input amplitudes. Capacitor C_1 sets the allowable droop for the output voltage, and optional capacitor C_2 acts as a filter for the noise/perturbations that may be present on the supply line. It is also assumed that resistor R_1 is large enough to ensure a large input impedance and proper input AC-coupling with a reasonably small capacitance C_{IN}, but at the same time both R_1 and R_2 must be sufficiently small to ensure that the DC voltage drops across them caused by the transistor base currents are negligible and do not generate large offsets.

The error term given in (3) can be compensated to some degree by mismatching Q_1 and Q_2 and/or I_1 and I_2 [5]. The usefulness of this technique is limited to situations where V_{IN} assumes values in a relatively narrow range. For example, Fig. 2 shows the output voltage and relative error of the detector of Fig. 1, where the transistors have been mismatched in such a way that the error term is perfectly compensated at $V_{IN} = 0.3\,\text{V}$ (with $I_1 = I_2$, and emitter area ratio $A_1/A_2 = 8.5$); while the relative error is still somewhat small in a V_{IN} range around 0.3 V, it increases rapidly at smaller input amplitudes (e.g., for $V_{IN} \le 0.1\,\text{V}$). Techniques for compensating the error term over a wider amplitude range employ significant additional circuitry around the configuration of Fig. 1, including duplicate peak detectors and translinear circuits for implementing mathematical functions such as multiplication and square rooting. Underscoring the difficulty of the task, such techniques do produce uniform detector slope, but – despite circuit complexity – exhibit (uniformly) large relative errors [6].

3. Proposed concept

It is useful at this point to take another look at the mechanics and implications of compensating the error term by mismatching Q_1 and Q_2 and/or I_1 and I_2 as described above. This can be done with the aid of the circuit in Fig. 3, which will also provide some direction for improving the detector accuracy over a wider amplitude range. In Fig. 3, we assume a virtual branch ("vb") consisting of Q_{2vb}, I_{2vb}, R_{2vb}, and C_{2vb}, placed between the original branches comprising transistors Q_1 and Q_2 of Fig. 1 as shown. We also assume that all the devices in the virtual branch are identical to their counterparts in the input branch, i.e., $I_1 = I_{2vb}$, $R_1 = R_{2vb}$, and $Q_1 \equiv Q_{2vb}$.

As a consequence, for a sinusoidal input voltage v_{IN}, the theoretical output voltage V_{OUTth} between the otherwise identical branches in Fig. 3 is given by (2), and contains the previously-noted amplitude-dependent error term V_{error} according to (3); including the compensation voltage V_{comp} shown in Fig. 3, the output voltage is then written as:

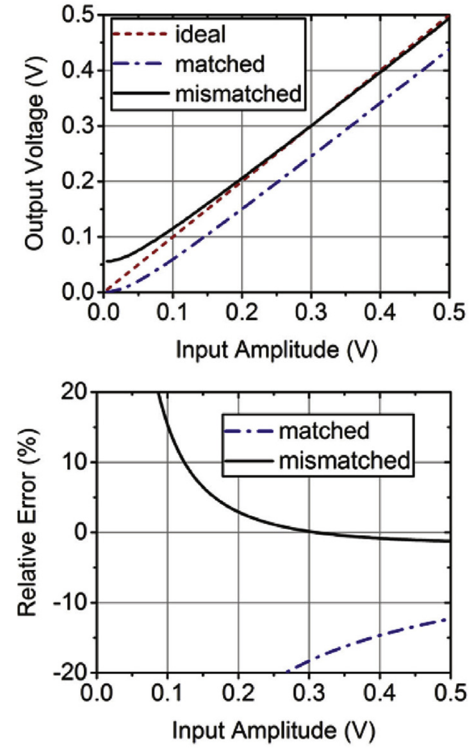

Fig. 2. Detector output voltage and relative error vs. input amplitude for conventional detector of Fig. 1, using simple transistor-area mismatching for error compensation.

$$V_{OUT} = V_{OUTth} + V_{comp} = V_{IN} - V_{error} + V_{comp}, \tag{4}$$

where

$$V_{comp} = V_{BE2} - V_{BE2vb} = V_T \ln\left(\frac{I_2}{I_{2vb}} \cdot \frac{I_{S2vb}}{I_{S2}}\right). \tag{5}$$

In (5), I_{S2} and I_{S2vb} are the base-emitter saturation currents of Q_2 and Q_{2vb}, respectively, being proportional to their emitter areas; because $I_1 = I_{2vb}$ and $Q_1 \equiv Q_{2vb}$, (5) can be re-written as:

$$V_{comp} = V_T \ln\left(\frac{I_2}{I_1} \cdot \frac{I_{S1}}{I_{S2}}\right) = V_T \ln\left(\frac{I_2}{I_1} \cdot \frac{A_1}{A_2}\right), \tag{6}$$

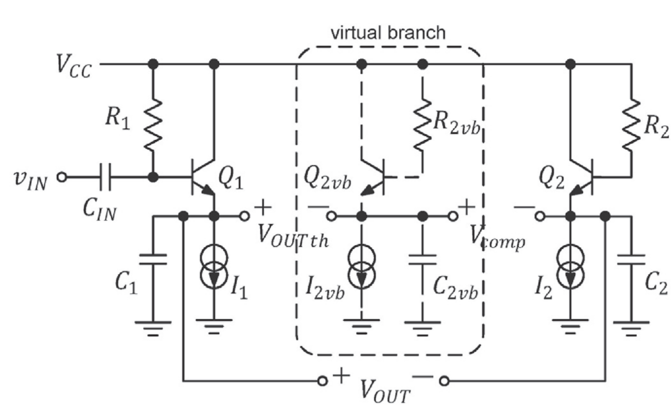

Fig. 3. Visualizing the compensation of the error term in the conventional peak detector using a virtual branch.