

A Novel Approach for Realization of Higher Order Filter using Bipolar and MOS Current-Mirrors

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Abstract: In this paper, we have presented an approach to realize higher order current-mode filters using cascade technique of lower order filters realized using CMOS current-mirrors. This paper presents the realization of higher-order low-pass (LP) active current-mode filters using cascade technique of second and third order filters. Interest in current-mode (CM) filters has been growing due to the fact that current-mode devices have wider dynamic range, improved linearity, and extended bandwidth as compared to their voltage-mode counterparts. In the sub-micron era of semi-conductor technology, the minute structure of devices causes a high electric field and thus the power supply voltage is limited to a certain low value to protect the device from destruction. Current-mode signal processing is quite attractive for operation at high frequencies and low power supply voltage operation from the viewpoint of sub-micron era. The realized filters do not contain any resistors, therefore simplifying the pattern of integrated circuits and thus reducing the chip area and the cost. The proposed circuits presented in the paper using bipolar technology and CMOS technology were tested in SPICE using BJT model NR100N and PR100N and the MOS structure using 0.5 μ m CMOS process parameters provided by MOSIS (AGILENT) and the results thus obtained were in accordance with the theoretical values.

Keywords: Continuous-time current-mode filters, Current Adders and Integrators, Bipolar and CMOS Current-Mirrors, Higher Order Filters, Cascade realization.

I. Introduction

Recent advancement in the field of integrated circuit technologies have miniaturized the devices used in an IC causing high internal electric field to develop. Therefore, the power supply voltage of the circuits must be reduced extremely to low value. In addition to this, the use of portable equipments has restricted the operation to a single battery [1], [2]. Also operation at high frequencies is a key desirable feature for these devices. In this respect, realizations of low-voltage and high-frequency active filters are an attractive choice in the field of signal processing.

Active filters are the best choice because they have the possibilities of integration and they are suitable for high frequency and low voltage operation. Among the numerous techniques available for the realization of active RC filters, we present two different methods in this paper namely (i) Cascade realization and (ii) Leapfrog realization [3]. These techniques of filter realizations are quite popular and can generate any type of transfer function. Traditionally, such circuits work in the continuous-time domain and have been designed as resistively terminated lossless LC filters where resonance could be employed to achieve complex poles and the desired steep transition region between the pass bands and the stop bands. With the growing pressure towards micro-miniaturization, inductors were found to be too bulky so that designers started to replace passive RLC filters by their active RC counterpart circuits where gain was obtained from operational amplifiers, and with the presence of resistors and capacitors in feedback networks, complex poles can be achieved. Active filters designed using the ubiquitous op-amp have found wide acceptance throughout the literature over the past and their mature development and technology guarantee their continued popularity in many applications.

One disadvantage of these op-amp based active RC filters is the limited frequency range of operation over which these circuits can be used: the finite bandwidth of op-amps usually confined the application to be below 100 KHz with performance becoming deteriorating with increasing frequency [4], [5]. For operation in MHz range the main design challenges for the analog filters are: (i) Reliable high-frequency performance. (ii) Automatic on-chip tuning against fabrication tolerances and (iii) changing operating conditions [6] and [7].

For application as continuous-time filters, the transconductance [8], [9] should satisfy the following properties: (i) Circuits must be simple, linear and should have a wider frequency response. (ii) Must have large input and output impedances to prevent undesirable interactions and simplify circuit design. (iii) They should preferably work with low-voltage power supplies to conserve power and to be compatible with the prevalent digital technologies on the same chip. (iv) Their transconductance parameter must depend on some dc bias voltage or current to facilitate electronic tuning against environmental or processing variations [10].

In many useful active simulation of filters, we may even insist that all the transconductors are identical and all the capacitors are grounded for specially simple IC layout and processing, and implementation of

integrated analog filters based on analog gate arrays appears to be a distinct possibility. In addition, the wider useful bandwidth of transconductances coupled with the reduced effects of circuit and device parasitics on filter performance result in far higher operating frequencies at which the circuits can function.

In the stop-band the gain of the filter changes at the rate of 20 db/decade for first-order filter which implies that for n^{th} -order filter the gain changes at the rate of $20n$ db/decade. The simplest way to build a fifth-order low-pass filter is by cascading a second order filter with a third-order section. Although there is no limit to the order of the filter that can be formed, as the order of the filter increases, so does its size. Also the accuracy declines, in that the difference between the actual stop-band response and the theoretical stop-band response increases with an increase in the order of the filter. Generally, the minimum-order filter required depends on the application specifications. Although a high-order filter necessarily provides a better stop-band response, the higher-order filters are more complex, occupy more space and are more expensive. Higher order filters provide greater roll-off rates between pass-band and stop-band, and can be necessary to achieve the required levels of stop-band attenuation or sharpness of cutoff frequency. They are more complex, requiring more components when implemented as analogue filters, or more processing when implemented digitally. The filter design process can be described as an optimization problem where each requirement contributes with a term to an error function which should be minimized. Typical requirements which are considered in the design process are: the filter should have a specific frequency response, a specific phase shift or group delay, a specific impulse response, causal, stable, localized, the computational complexity of the filter should be low and the filter should be implemented in particular hardware or software. Another important parameter of great concern is the required frequency response of the filter. In particular, the steepness and complexity of the response curve is a deciding factor for the filter order and feasibility. For many purposes, this is not sufficient. To achieve steeper slopes, higher order filters are required.

II. Cascade Design

This approach is based on the factorization of a transfer function $H(s)$ into the product of lower-order terms. If the order n is even, the decomposition consists of $n/2$ second order terms.

$$H(s) = H_1(s) \times H_2(s) \times \dots \times H_{n/2}(s)$$

If n is odd, the factorization includes also a first order term. Sometimes this term is combined with one of the second-order terms to create a third-order filter stage. The first-order term, if any, can be implemented with a simple RC or CR network, so all we need to know is its required cut-off frequency ω_0 . The second-order terms can be implemented with any of the filters of second-order. For each of these stages we need to know ω_0 and Q , and ω_z if the stage is a notch stage. The cascade approach offers a number of advantages. The design of these sections is relatively simple, and the component count is usually low. The low-output impedance of the individual sections eliminates inter-stage loading, so each section can be regarded as isolated from the others and can be tuned independently, if needed. The inherent modularity of this approach is also attractive from the economic stand point, since one can use a few standardized blocks to design a variety of more complex filters.

Mathematically, the order in which the various sections are cascaded in series is irrelevant. In practice, to avoid loss of dynamic range and filter accuracy due to possible signal clipping in the high- Q sections, the sections are cascaded in order of ascending Q 's, with the low- Q stages first in the signal path. This ordering, however, does not take into account internal noise, which may be of concern in the high- Q stages, where any noise component falling under the resonance peak may be amplified significantly. So, to minimize noise, high- Q stages should go first in the cascade. In general, the optimum ordering depends on the input spectrum, the filter type and the noise characteristics of its components.

III. Proposed Circuits

Current-mode signal processing is quite attractive for low power supply voltage operation and high frequency application. In this paper design of current-mode filters using nMOS transistor current mirrors and pMOS transistor current sources as an active load, as reported in [4], which is a counterpart of the MOS realization [6], [7] are presented. In such designs, pMOS transistors were only used for DC current sources which provide bias currents to the current mirrors and also behave as an active load for the current mirror.

The currents of these sources must match with each other and also match with the DC currents of each current mirror to give a proper DC bias to transistors. However the matching is sometimes difficult due to the parameter mismatch between nMOS and pMOS transistors [7], [8]. Here, we present design of CMOS continuous-time filters using nMOS and pMOS transistor current mirrors. Using pMOS transistors as signal processing elements, the circuits can be simplified and problems resulting from difficulty of DC biasing can be avoided.

Using integrators and adders, any kind of transfer function can be realized; therefore this method can be easily applied for realizing wide variety of filters. Generally, monolithic integrated filters have a difficulty on setting the filter frequency due to absolute value error of elements. In these filters, the tuning of the filter frequency can be easily achieved by adjusting the current of a single DC current source.

2.1 Realization of CMOS Current-Mode Fifth-Order Low Pass Active Filter using Cascade Realization Technique:

Fig. 1 shows the practical realization of the proposed fifth-order low pass filter where the current-mode second-order section of Fig. 2 [11] and the current-mode third-order section of Fig. 2 [12] are cascaded together to obtain the proposed circuit of a fifth-order filter.

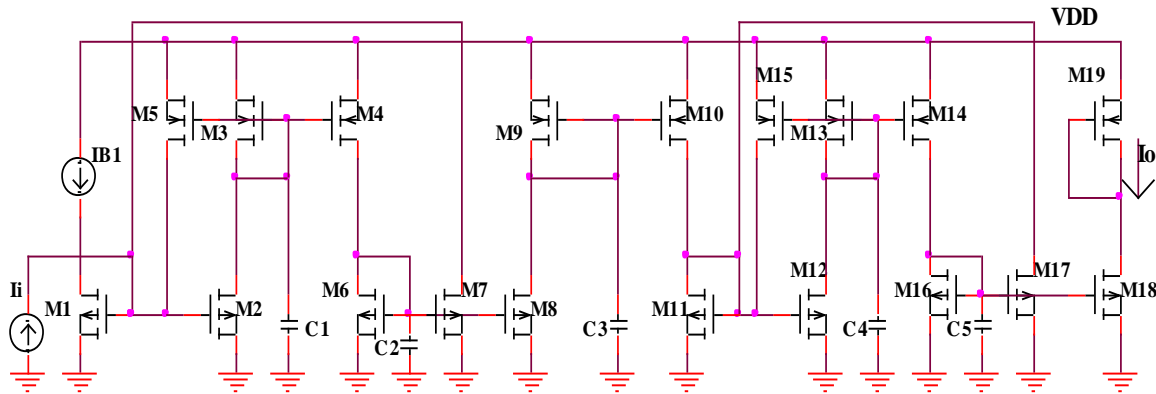


Fig. 1: Cascaded Current Mode Fifth-Order Section using CMOS Realization

Assuming that all transistors of Fig. 2 have an equal aspect ratio's that is, all the current mirrors have same (W/L) ratios, so we obtain the following transfer function:

$$T_{LP}(s) = \left[\frac{1}{s^3 C_3 C_4 C_5 R^3 + s^2 C_3 R^2 (C_4 + C_5) + s R (C_3 + C_5) + 1} \right] \left[\frac{1}{s^2 C_1 C_2 R^2 + s C_1 R + 1} \right] \quad (2.1)$$

Since the DC bias currents of all transistors are equal to the bias current I_B , the resistances are given by,

$$R_1 = R_2 = R_3 = R_4 = R_5 \cong \frac{kT}{q} \frac{1}{I_{B1}} \cong \frac{0.0259}{I_{B1}} \quad (2.2)$$

Equation-2.2 shows that I_B controls R_1, R_2, R_3, R_4 and R_5 simultaneously and the simultaneous control of R_1, R_2, R_3, R_4 and R_5 do not affect the Q factor. Thus we can control the cut-off frequency of the filter ω_c without changing the Q factor.

2.2 Realization of Current-Mode Fifth-Order Low Pass Active Filter using Bipolar Realization:

Fig. 2 shows the practical realization of the proposed fifth-order low pass filter using bipolar technology where we have used the cascaded realization of the current-mode second-order and the current-mode third-order section to obtain the circuit of a fifth-order filter.

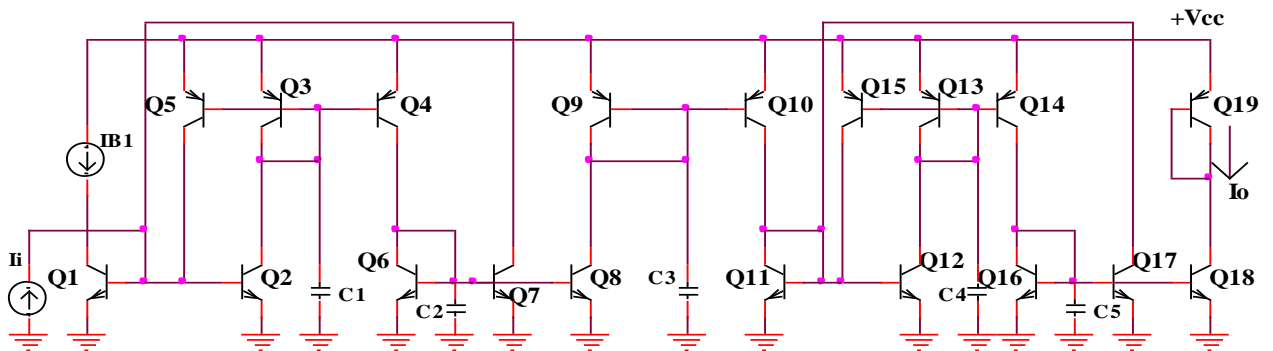


Fig. 2: Cascaded Current Mode Fifth-Order Section using Bipolar Realization

IV. Simulation Results

The workability of the proposed circuits were tested and verified in SPICE using 0.5 μ m CMOS process parameters provided by MOSIS (AGILENT) and the Bipolar parameters for NR100N and PR100N as listed in Table-1.

TABLE-1: CMOS AND BIPOLAR PROCESS PARAMETERS

TRANSISTOR	PROCESS PARAMETERS
nMOS	LEVEL=3 UO=460.5 TOX=1.0E-8 TPG=1 VTO=0.62 JS=1.08E-6 XJ=0.15U RS=417 RSH=2.73 LD=0.04U VMAX=130E3 NSUB=1.17E17 PB=0.761 ETA=0.00 THETA=0.129 PHI=0.905 GAMMA=0.69 KAPPA=0.10 CJ=76.4E-5 MJ=0.357 CJSW=5.68E-10 MJSW=0.302 CGSO=1.38E-10 CGDO=1.38E-10 CGBO=3.45E-10 KF=3.07E-28 AF=1 WD=0.11U DELTA=0.42 NFS=1.2E11
pMOS	LEVEL=3 UO=100 TOX=1.0E-8 TPG=1 VTO=0.58 JS=0.38E-6 XJ=0.10U RS=886 RSH=1.81 LD=0.03U VMAX=113E3 NSUB=2.08E17 PB=0.911 ETA=0.00 THETA=0.120 PHI=0.905 GAMMA=0.76 KAPPA=2 CJ=85E-5 MJ=0.429 CJSW=4.67E-10 MJSW=0.631 CGSO=1.38E-10 CGDO=1.38E-10 CGBO=3.45E-10 KF=1.08E-29 AF=1 WD=0.14U DELTA=0.81 NFS=0.52E11
NR100N	IS = 121E-18 BF = 137.5 VAF = 159.4 IKF = 6.974E-3 ISE = 36E-16 NE = 1.713 BR = 0.7258 VAR = 10.73 IKR = 2.198E-3 RE = 1 RB = 524.6 RBM = 25 RC = 50 CJE = 0.214E-12 VJE = 0.5 MJE = 0.28 CJC = 0.983E-13 VJC = 0.5 MJC = 0.3 XCJC = 0.034 CJS = 0.913E-12 VJS = 0.64 MJS = 0.4 FC = 0.5 TF = 0.425E-8 TR = 0.5E-8 EG = 1.206 XTB = 1.538 XTI = 2.0
PR100N	IS = 73.5E-18 BF = 110 VAF = 51.8 IKF = 2.359E-3 ISE = 25.1E-16 NE = 1.650 BR = 0.4745 VAR = 9.96 IKR = 6.478E-3 RE = 3 RB = 327 RBM = 24.55 RC = 50 CJE = 0.18E-12 VJE = 0.5 MJE = 0.28 CJC = 0.164E-12 VJC = 0.8 MJC = 0.4 XCJC = 0.037 CJS = 1.03E-12 VJS = 0.55 MJS = 0.35 FC = 0.5 TF = 0.610E-9 TR = 0.610E-8 EG = 1.206 XTB = 1.866 XTI = 1.7

4.1 Simulation Results for the CMOS Current-Mode Fifth-Order Low Pass Filter using Cascade Realization Technique:

For the circuit shown in Fig. 1 the ac analysis were carried out with the value of dc bias current $I_B = 2\mu$ A, $C_1 = C_3 = C_5 = 0.5$ pF, $C_2 = C_4 = 1$ pF, $(W/L)_P$ ratio = 1μ m/ 1μ m, $(W/L)_N$ ratio = 1μ m/ 1μ m and supply voltage $V_{DD} = 1.7$ V. The value of cut-off frequency is found to be $f_0 = 4.0536$ MHz which is very well in agreement with the calculated theoretical value of $f_0 = 4.05$ MHz. SPICE simulation results are shown in Fig. 3.

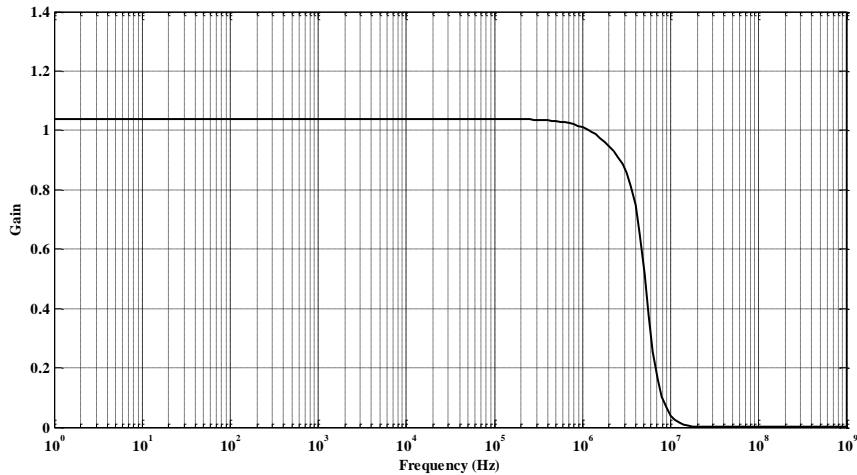


Fig. 3: Low Pass Response of Current-Mode Fifth Order Section using CMOS technology

Fig. 4 shows the change in the value of cut-off frequency with respect to the change in the values of the capacitor C_1 from 0.5 pF to 75 pF keeping the value of capacitors C_2 and C_4 fixed at 1 pF and capacitors C_1 and C_3 are kept fixed at 0.5 pF. The value of cut-off frequency changes from 34.28 KHz to 4.0536 MHz which shows that cut-off frequency varies inversely with capacitance C_1 .

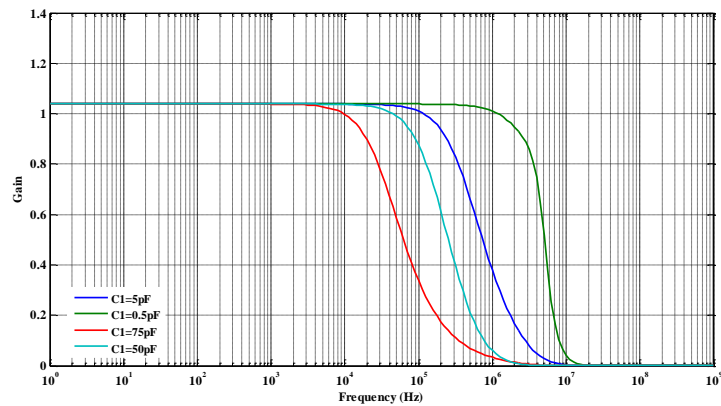


Fig. 4: Variation in the cut-off frequency with change in Capacitor value

Fig. 5 shows the change in the value of cut-off frequency with change in the value of bias current when varied from 10 μ A to 80 μ A. Due to this variation the cut-off frequency varies from 4.0858 MHz to 235.612 KHz, thus cut-off frequency varies inversely with bias current.

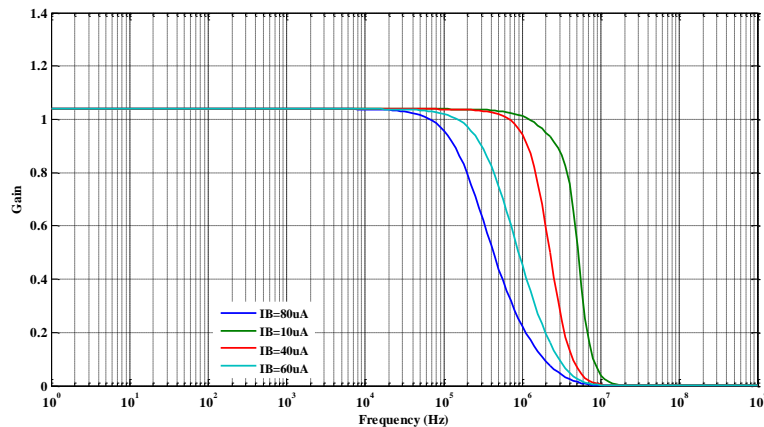


Fig. 5: Variation in the Gain with change in bias current I_B

Fig. 6 shows the change in the value of cut-off frequency with respect to the change in the values of the capacitor C_2 and C_4 from 5 pF to 75 pF keeping the value of capacitors C_1 , C_3 and C_5 fixed at 0.5 pF. The value of cut-off frequency changes from 21.419 KHz to 358.783 KHz which shows that cut-off frequency varies inversely with capacitance C_2 and C_4 .

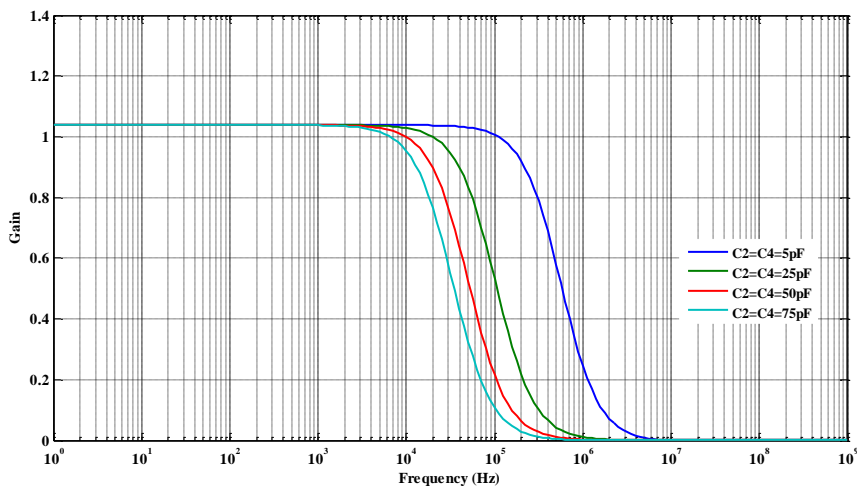


Fig. 6: Variation in the cut-off frequency with change in Capacitor values

4.2 Simulation Results for the Current-Mode Fifth-Order Low Pass Filter using Bipolar Technology:

For the circuit shown in Fig. 2 the ac analysis were carried out with the value of dc bias current $I_B = 2\mu\text{A}$, $C_1 = C_3 = C_5 = 0.5 \text{ pF}$ and $C_2 = C_4 = 5 \text{ pF}$ and supply voltage $V_{DD} = 1.7 \text{ V}$. The value of cut-off frequency is found to be $f_0 = 1.1526 \text{ MHz}$ which is very well in agreement with the calculated theoretical value of $f_0 = 1.16 \text{ MHz}$. SPICE simulation results are shown in Fig. 7.

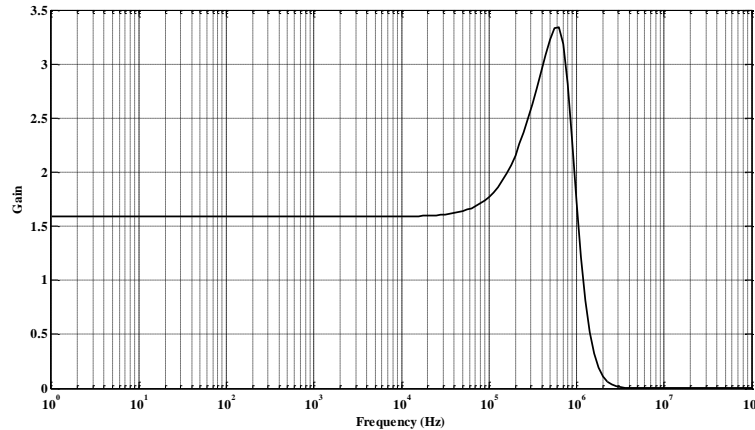


Fig. 7: Low Pass Response of Fifth-Order Current-Mode Filter using bipolar technology

V. Conclusion

In the given paper high frequency fifth order current-mode filter have been presented which are quite suitable for the realization in high frequencies till 10 MHz and these filters can operate at a voltage as low as 2V or less. Also, it has been verified that variations in the value of either the capacitor or the bias current improves the gain as well as the operating frequency of the filters. The frequency of this filter can be easily and widely controlled by a single DC bias current and thus provides good tunability. All the circuits were tested using SPICE and the verified results confirms the theoretical values.

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