A Fully Independently Adjustable, Integrable Simple Current Controlled Oscillator and Derivative PWM Signal Generator

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SUMMARY A simple circuit scheme, able to generate square/triangular wave, is proposed. Its advantages are that oscillation frequency and amplitudes of the proposed circuit do not have a small range of temperature drift. Electronic adjustments of them can be obtained with a wide sweep range and DC offset adjustment available. In addition, the proposed scheme can produce frequency-constant derivative of PWM signal without an additional device requirement. It is very appropriate for, with simple circuit details, not only circuit implementation but also monolithic fabrication. The PSPICE simulation results through bipolar technology are given here, they show good performance of the proposed circuit.

key words: OTA, PWM, CCO, bipolar, temperature-insensitive

1. Introduction

Current/Voltage Controlled Oscillators (CCOs/VCOs) and Pulse Width Modulators (PWMs) play important rules in many fields such as instrumentation, electronic and communication systems. In traditional, the circuit performances rely on environment condition: i.e., temperature variation and power supply voltage. A simple circuit of square/triangular wave VCO with a wide sweep capability can be easily realized based on using Operational Transconductance Amplifiers (OTAs) as a switching current source to charge and discharge a grounded timing capacitor followed by a Schmitt trigger [1]. This scheme has poor frequency stability because the oscillation frequency is dependent not only on power supply voltage but also on temperature due to temperature-sensitive of transconductance gain of OTA. The novel VCO based on OTA has been subsequently proposed [2]. Its advantage is that the oscillation frequency is independent of the temperature. However, it has much complicated scheme and a small range of the oscillation frequency due to comprising of operational amplifiers, which have narrow bandwidth relative to an OTA [3]. Especially the amplitude adjustments of square/triangular wave can not be achieved.

The square/triangular wave generator using a Current Feedback Operational Amplifier (CFOA) has been later proposed [4], it provides the signal with a wide range of frequencies. However the oscillation frequency and amplitudes can not be adjusted via this circuit. Recently, current controlled oscillator over a wide range has been introduced [5], this circuit gives square/triangular wave with electronic controllability only frequency, not include its amplitudes. In addition, temperature stability of the circuit was not investigated. Hence these schemes still do not provide sufficient stability to implement them as a precise component in the design of instrumentation and communication systems, particularly under varying environment conditions.

For PWM signal, the output frequency of conventionally simple PWM signal generators depends on modulating signal [6]–[9]. This makes it to complicate in control systems [10]. Furthermore, some of PWM signal generators can not provide the duty cycle of PWM output signal, which is linearly dependent on a modulating signal, this causes a distorted modulating signal after the PWM demodulation.

The purpose of this article is to present a simple square/triangular wave-generating scheme based on OTA Schmitt trigger [11]. The features are that the oscillation frequency and amplitudes are frequency-stable and a relatively wide range. Both of the oscillation frequency and amplitudes can be independently electronically adjusted. In addition, with the proposed scheme, it can originate the derivative PWM signal without an additional device requirement and frequency variation. The circuit description is simple scheme, which is composed of 3 simple OTAs, 2 resistors and 1 capacitor. The circuit performances can be demonstrated through the simulation results with bipolar technology.

2. Circuit Description and Operation

2.1 The Current Controlled Oscillator

The proposed CCO is illustrated in Fig.1 whereas $v_i(t)$ is firstly connected to ground to serve as a simple CCO for easily understanding the circuit. From the circuit, OTA1 and the timing capacitor $C$ function as an inte-
grator whose time constant is proportional to the bias current. The OTA2, OTA3 and the resistors $R_1$ and $R_2$ work as a Schmitt trigger whose threshold voltage is proportional to the bias current $I_{B3}$ and saturation voltage is proportional to the bias current $I_{B2}$.

The maximum output current of the OTAs is respectively $I_{B1}$, $I_{B2}$ or $I_{B3}$. At the first instant when the power supplies are turned ON the capacitor $C$ is discharged and $v_{o1}(t) = 0$. Positive feedback in OTA2 and OTA3 results in the maximum output current in these amplifiers. Assume that these currents are flowing out of OTA2 and OTA3, then the voltage $v_{o3}(t)$ at the threshold resistor jumps to its maximum value of $V_{o3H} = I_{B3}R_1$ (1)

Simultaneously, OTA1 will provide the maximum output current of $I_{B1}$ and the capacitor $C$ is charged by this current. The voltage $v_{o1}(t)$ is linearly increasing and when it is close to its highest value $V_{o1H}$ (the exact value will be calculated in Sect. 2.2) which is close to $V_{o3H}$. The output current of OTA2 starts to turn its direction. As a result, the output current will quickly achieve its maximum value flowing into OTA2. Now the voltage $v_{o3}(t)$ has jumped to its minimum value of $V_{o3L} = -I_{B3}R_1$ (2)

And the output current of OTA1 changes its direction. The current $I_{B1}$ flows into OTA1 and the capacitor $C$ is discharged by this current. The voltage $v_{o1}(t)$ is linearly decreasing now. When it becomes approximately equal to $V_{o1L}$, the output current of OTA2 starts to turn its direction again. Therefore $v_{o3}(t)$ jumps up to $V_{o3H}$. The proposed circuit is now in its periodic operation, consequently the waveforms generated from the proposed circuit; $v_{o1}(t)$, $v_{o2}(t)$ and $v_{o3}(t)$ are shown in Fig. 2. The circuit description of proposed scheme is shown in Fig. 3. It uses simple bipolar OTAs to obtain more simple circuit.

2.2 The Oscillation Frequency

The operating currents $I_{B1}$, $I_{B2}$ and $I_{B3}$ and the threshold resistors $R_1$ and $R_2$ should be chosen so that OTA1, OTA2 and OTA3 are not saturated when their output voltages are maximal. Then the maximum oscillation can be achieved [12].

The oscillation frequency in such non-saturated CCO can be calculated with sufficient precision if the exact values of capacitor voltage (i.e. $V_{o1H}$ and $V_{o1L}$) at the instants of jumps are known. Owing to the rapid change of states the variation of the charging or dis-
charging current during the short period of time just before a jump can be neglected [13], [14]. In theoretical derivation, with ideal components and all of matching active elements, it is easy to display that period of oscillation consists of time interval of saturation operation ($T_{\text{sat}}$) plus that of non-saturation operation ($T_{\text{non-sat}}$) as Fig. 2, which can be shown as

$$T = 2T_{\text{sat}} + 4T_{\text{non-sat}}$$

(3)

Where $T_{\text{sat}}$ can be calculated as

$$T_{\text{sat}} = \frac{2C(I_{B3}R_1 - \Delta V)}{I_{B1}}$$

(4)

and $T_{\text{non-sat}}$ is shown as

$$T_{\text{non-sat}} = \frac{C\Delta V}{I_{o1}}$$

(5)

Where $\Delta V$ is a slightly voltage difference between the inputs of OTA2 ($v_{o2} - v_{o1}$) due to the moment of switching. Because $v_{o2}$ in this interval is much small than $2V_T$, we can yield output currents of OTAs as follows

$$I_{o1} = I_{B1} \tanh \left( \frac{v_{o2}}{2V_T} \right)$$

(6)

$$I_{o2} = I_{B2} \tanh \left( \frac{v_{o3} - v_{o1}}{2V_T} \right)$$

(7)

$$I_{o3} = I_{B3} \tanh \left( \frac{v_{o2}}{2V_T} \right)$$

(8)

Where $I_{o1}$, $I_{o2}$ and $I_{o3}$ are respectively the output currents of OTA1, OTA2 and OTA3. We can thus use Taylor’s series as following form

$$\tanh x = x \left( 1 - x^2/3 + (2/15) x^4 + \ldots \right)$$

(9)

Usually, a first-order approximation is proposed to describe the evolution of the output currents of OTAs. It concisely gives

$$T_{\text{non-sat}} = \frac{C\Delta V}{I_{B1} \tanh \frac{\Delta V}{2V_T}} \approx \frac{C\Delta V}{I_{B1} \frac{\Delta V}{2V_T}}$$

$$= \frac{I_{B1}I_{B2}R_2 \tanh \frac{\Delta V}{2V_T}}{I_{B1}I_{B2}R_2}$$

(10)

So,

$$T_{\text{non-sat}} \approx \frac{C\Delta V}{2V_T I_{B1}I_{B2}R_2} = \frac{4CV_T^2}{4R_1CI_{B3} R_1}$$

(11)

Substituting Eqs. (4) and (10) into Eq. (3) and providing $\Delta V = 2V_T$ [15], it yields

$$T = \frac{4R_1CI_{B3}}{I_{B1}} \left[ 1 - \frac{2V_T}{I_{B3}R_1} + \frac{4V_T^2}{I_{B2}I_{B3}R_1R_2} \right]$$

(12)

The oscillation frequency is then given by

$$f_o = \frac{4R_1CI_{B3}}{I_{B1}} \left[ 1 - \frac{2V_T}{I_{B3}R_1} + \frac{4V_T^2}{I_{B2}I_{B3}R_1R_2} \right]^{-1}$$

(13)

For a usual application, from Eq. (12), if $4V_T^2 \ll I_{B2}I_{B3}R_1R_2$ and $2V_T \ll I_{B3}R_1$, it can be approximately reduced to

$$f_o = \frac{4R_1CI_{B3}}{I_{B1}}$$

(14)

Thus its period is approximately depicted as

$$T = 4R_1CI_{B3}$$

(15)

Subsequently, amplitudes of output voltages are exactly given by

$$v_{o1}(p-p) = 2(I_{B3}R_1)$$

(16)

$$V_{o2}(p-p) = 2(I_{B2}R_2)$$

(17)

The relationship among Eqs. (13)–(17) shows that its frequency and amplitudes can be independently tunable. It means that the carrier oscillation frequency can be adjusted by $I_{B1}$, the triangular wave ($v_{o1}$) amplitude adjustment can be done by $I_{B3}$ and the square wave ($v_{o2}$) amplitude can be tuned by $I_{B2}$. In addition, if temperature coefficients of $R_1$ and $C$ are assumed to be zero, it should be observed that the oscillation frequency is independent of temperature because of compensation of current ratio $I_{B1}/I_{B3}$. However, the precise analysis will be given in Sect. 3.

2.3 DC Offset Adjustment

The DC offset adjustment of output signal of presented square/triangular wave generating scheme can be achieved by tuning a DC voltage source as $V_{\text{offset}}$ connected at $v_t(t)$ node in Fig. 1, then the voltage of triangular and square wave with adding DC offset: $v_{o1}(t)$ and $v_{o3}(t)$, can be respectively shown as

$$v_{o1}(t) = v_{\text{tri}}(t) + V_{\text{offset}}$$

(18)

$$v_{o3}(t) = v_{\text{square}}(t) + V_{\text{offset}}$$

(19)

Where $v_{\text{tri}}(t)$ and $v_{\text{square}}(t)$ are triangular and square wave signals of $v_{o1}(t)$ and $v_{o3}(t)$, as shown in Fig. 2, respectively.

2.4 Derivative PWM Signal Generator

2.4.1 The Conventional PWM Signal Generator

The conventional PWM signal generation used principally in communication systems is schematically depicted in Fig. 4 [16]–[19]. If we assume that a sinusoidal signal is a modulating (information) signal, the amplitude summation of the modulating signal and triangular carrier wave is illustrated in the middle trace of...
Fig. 4 Conventional PWM signal generation in a communication system.

Fig. 5 Different signals of the PWM signal generation in Fig. 4.

Fig. 6 Principle of proposed circuit.

Fig. 5. Then, PWM output signal is generated by comparing the summation result to reference level demonstrated in lower trace of Fig. 5. We can found that although carrier frequency is not much higher than information frequency, the PWM output frequency is rather not constant. So in typical applications, the carrier frequency is provided to be much higher than ten times of the information frequency [20], [21].

2.4.2 Principle of Proposed PWM Signal Generation

The PWM signal generation, which comes from differentiated result of modulating signal, has scheme shown in Fig. 6. Its principle is to establish a differential result of modulating signal during \( t_1 \) and \( t_2 \) which are switched times of carrier voltage to change its polarity. The duty cycle \( (D) \) is then depend on differential result of modulating signal at \( t_1 \) and \( t_2 \) as signal shown in Fig. 2, which can be controlled by carrier frequency as Eq. (13).

\[
D = \frac{1}{2} \left[ 1 + a \frac{\Delta v_i (t)}{T/2} \right] \times 100\%
\]  

(20)

Where \( a \) is a circuit constant value and \( \Delta v_i (t) = v_i (t_2) - v_i (t_1) \). We can see that, from Eq. (20), if \( \Delta v_i (t) = 0 \), this means that in case of no variation of modulating signal, the duty cycle will be 50%. In the case of the carrier frequency is much greater than the modulating frequency, we can assume that

\[
\frac{\Delta v_i (t)}{T/2} = \frac{d}{dt} v_i (t)
\]

(21)

It is clearly seen that, from Eqs. (20) and (21), the duty cycle of PWM signal is dependent on a differentiated result or a slope of the information signal. We can anticipate that if common mode noise involves in the modulating signal, the noise can be eliminated because only differential result of modulating signal is transformed to be a pulse width of PWM output signal. From the proposed circuit in Fig. 1, if \( v_i (t) \) is assumed as a modulating (information) signal, time intervals of positive and negative saturation voltage of \( v_o (t) \) depend on the modulating signal respectively shown as

\[
T_1 = 2 \frac{R_1 C I_{B3}}{I_{B1}} \left[ 1 + \frac{v_i (t_2) - v_i (t_1)}{2 I_{B3} R_1} \right]
\]

(22a)

\[
T_2 = 2 \frac{R_1 C I_{B3}}{I_{B1}} \left[ 1 + \frac{v_i (t_1) - v_i (t_2)}{2 I_{B3} R_1} \right]
\]

(22b)

We found that, in theoretical, from Eqs. (22a) and (22b), the period time of oscillation (carrier) which comes from summation of \( T_1 \) and \( T_2 \) is equal to Eq. (14), hence the oscillation frequency still be equal to Eq. (13) while its duty cycle depends on the differential result of modulating signal as follows

\[
D = \frac{1}{2} \left[ 1 + \frac{\Delta v_i (t)}{2 I_{B3} R_1} \right] \times 100\%
\]

(23)

Where \( \Delta v_i (t) \) is derivative of the modulating voltage \( (v_i (t_2) - v_i (t_1)) \). If the circuit has condition as (21), it yields

\[
D = \frac{1}{2} \left( 1 + \frac{C}{I_{B1} \frac{d}{dt} v_i (t)} \right) \times 100\%
\]

(24)

It clearly demonstrates that the proposed scheme, without an additional element requirement, can function as the differentiated-input or slope-input PWM signal generator. The features are, independency of frequency against the modulating signal variation, and the relatively simple scheme due to integration of various stages of PWM signal generator such as triangular wave generator, adder and comparator into the same circuit. The accurate modulating signal can be recovered by using integrator circuit after demodulation of the PWM signal.
3. Performance Analysis

3.1 Third-Order Effects

The first-order approximation of \( \tanh x \) as Eq. (9) is only correct for very low value of \( x \). So, an OTA must be considered as a nonlinear device for transition-state to obtain more precise derivation. In that case, the higher \( n \) order approximation defined by Eq. (9) must be used to describe the output currents of OTAs. However, as the analysis difficulty increases with \( n \), we will suppose that \( x \) will be small enough to use a third-order development. In this case, as similar deviation to the first-order approximation, the following equations of period and oscillation frequency will be respectively obtained

\[
T = \frac{4R_1 C I_{B3}}{I_{B1}} \left[ 1 - \frac{2V_T}{I_{B3} R_1} + \frac{24V_T^2}{I_{B3} I_{B1} R_1 R_2 (6V_T - 1)} \right] \tag{25}
\]

\[
fo = \frac{I_{B1}}{4R_1 C I_{B3}} \left[ 1 - \frac{2V_T}{I_{B3} R_1} + \frac{24V_T^2}{I_{B3} I_{B1} R_1 R_2 (6V_T - 1)} \right]^{-1} \tag{26}
\]

However the last two terms in bracket of Eq. (26) can be approximated as Eq. (13).

3.2 Temperature Stability of Proposed CCO

The temperature stability of the proposed oscillator depends strongly on current levels and voltage thresholds of the Schmitt trigger as well as on the external components. In the following discussion, however, we will examine only those temperature drift sources that come from transition time delays as shown in Fig. 4, and not those contributed by the external components and switching time delay due to parasitic capacitances of active elements in the proposed circuit. Independent-temperature external components, which are capacitor, resistors including external current sources, are supplied by the user [15]. Temperature coefficient of oscillation frequency with the first-order approximation can be obtained as

\[
\frac{1}{f_o} \frac{\partial f_o}{\partial T} = \left( \frac{2}{I_{B3} R_1} - \frac{8V_T}{I_{B3} I_{B1} R_1 R_2} \right) \frac{\partial V_T}{\partial T} \tag{27}
\]

In similar to the first-order approximation, the temperature coefficient of oscillation frequency with the third-order approximation is

\[
\frac{1}{f_o} \frac{\partial f_o}{\partial T} = \left( \frac{2}{I_{B3} R_1} - \frac{72V_T^2}{I_{B3} I_{B1} I_{B2} R_1 R_2 (6V_T - 1)} + \frac{144V_T^3}{I_{B3} I_{B1} I_{B2} R_1 R_2 (6V_T - 1)^2} \right) \frac{\partial V_T}{\partial T} \tag{28}
\]

If \( I_{B1} = I_{B2} = I_{B3} = 100 \mu A \) and \( R_1 = R_2 = 10 \text{ k\ohm} \), for example, we can find that the temperature coefficients of oscillation frequency at room temperature range with the first-order and third-order approximation are \(+16.336 \text{ ppm/}^\circ C \) and \(+17.732 \text{ ppm/}^\circ C \), respectively. These values can be eligible for the most applications [15].

3.3 Switching Time Delay

Parasitic capacitances of active elements affect the performance of the CCO at high frequencies in two ways: 1) in the form of switching time delay; and 2) modifying the square and triangular waveforms. Although the switching time delay is always present, its effect at low frequency are negligible since the period duration is much longer than the magnitude of the delay [22]. The switching time delay affects not only linearity, but also the high-frequency temperature coefficient. In the presence of the delay, the oscillation frequency is

\[
f_{\text{actual}} = \frac{1}{T_{\text{desired}} + t_{\text{delay}}} \tag{29}
\]

Computer simulations through PSPICE using specific parameters, as depicted in Sect. 4, of the CCO were made to determine more accurately the switching delay and to help important contributing sources. To estimate the time delay, the following method was used. The magnitude of the voltage across the timing capacitance \( V_{LF} \) during low-frequency (\( \approx 1 \text{ kHz} = f_{LF} \)) oscillations was measured. Changing just the timing capacitor, the high-frequency (\( \approx 10 \text{ MHz} = f_{HF} \)) voltage across the capacitor was again measured \( V_{HF} \). Due to the same time delay and different capacitance values where \( V_{\text{delay}} = (I_{o1}/C) t_{\text{delay}} \), \( V_{HF} \) is larger than \( V_{LF} \). Hence, the time delay is given by

\[
\frac{V_{HF} - V_{LF}}{2V_{HF} f_{HF}} \tag{30}
\]

in which \( 2V_{HF} f_{HF} \) is simply the slew rate of the triangular wave at a frequency \( f_{HF} \). The results show a switching time delay of 5.2ns. The switching time delay is strongly temperature-dependent due to the parasitic resistances, especially \( r_b \), the resistor loads, particularly of the parasitic capacitances whose temperature-variation is a direct consequence of the junction built-in voltage temperature dependence.

4. Simulation Results

4.1 The Current Controlled Oscillators

The completely proposed circuit shown in Fig.3 was simulation tested using PSPICE through PNP and NPN transistor models of the PR200N and NR200N bipolar transistors, respectively [23]. The circuit was operated with \( \pm 5 \text{ V} \) supply with all of Resistors \( 10 \text{ k\ohm} \).
Where $C = 100 \, \text{nF}$, $I_{B1} = I_{B3} = 100 \, \mu\text{A}$ and $I_{B2} = 200 \, \mu\text{A}$.

(b) Where $C = 10 \, \text{pF}$, $I_{B1} = I_{B3} = 100 \, \mu\text{A}$ and $I_{B2} = 200 \, \mu\text{A}$.

Fig. 7 Different signals obtained from simulation.

First results of proposed CCO shown in Fig. 7 were obtained with different specific values. There are the different signals in different component values in the circuit, as expected in Eq. (13). Where the upper trace is $v_{o1}(t)$, middle trace shown is $v_{o2}(t)$ and lower trace is $v_{o3}(t)$.

In addition, the behaviors of the proposed circuit working as electronically adjustable square/triangular wave generator are also investigated in Figs. 8(a)–(c). They demonstrate simulation results of input bias current variations of OTAs, they show good agreement with the theoretical anticipation. It means that the proposed scheme can work as electronically adjustable square/triangular wave generator. A wide range adjustment of oscillation frequency by $I_{B1}$, as shown in Fig. 8(a). The simulation result shows that the proposed circuit can operate over a wide range of frequency due to $I_{B1}$ adjustment without changing any elements. At high frequency, it has an error due to the switching time delay, as explained in Sect. 3.

Figure 9 shows the temperature stability of the output frequency for low frequency, 10 MHz and 20 MHz. The temperature coefficient is maintained to about $\pm 130 \, \text{ppm/}^\circ\text{C}$ throughout the whole frequency range. The low frequency curve shows a positive TC while the high frequency ones have a negative TC. This was done due to the extra TC added by the switching time delay at high frequencies.

4.2 The Derivative PWM Signal Generator

First simulation result of proposed circuit working as PWM shown in Fig. 10 was obtained with capacitor 150 pF, $I_{B1} = 200 \, \mu\text{A}$, $I_{B2} = 300 \, \mu\text{A}$ and $I_{B3} = 100 \, \mu\text{A}$. From Fig. 10, during A period the input signal has a positive slope, then the PWM output signal has duty cycle more than 50%. During B period, the input signal has a zero slope, then the PWM output signal has duty cycle equal to 50%. For C interval, the input signal has a negative slope, then the PWM output...
Fig. 9  Simulated oscillation frequency deviations against temperature variations.

Fig. 10  The simulation result with upper trace is modulating signal and lower trace is PWM output signal.

Fig. 11  The simulation result of modulated PWM output signal relative to sinusoidal modulating signal in (a) Time domain (b) Frequency domain.

Fig. 12  The result of simulated duty cycle relative to $dV/dt$ of input signal at about 50 kHz PWM frequency: $I_{B1} = 100 \mu A$, $I_{B2} = I_{B3} = 50 \mu A$ and capacitor 1 nF.

signal has duty cycle less than 50%. It is clearly noted that the PWM output signal is modulated by a differentiated (or a slope) input signal followed by Eq. (24) while the frequency of PWM output signal is constant and independent of the magnitude of modulating signal.

Figures 11(a), (b) show results of the modulated PWM output signal against modulating signal of 2Vp sinusoidal signal with 500 kHz frequency in time and frequency domain respectively, which were achieved with capacitor 10 pF, $I_{B1} = 200 \mu A$ and $I_{B2} = I_{B3} = 100 \mu A$ for receiving carrier frequency of approximately 5 MHz. It is clearly found that it consists of modulating signal component, so it can be recovered by using an appropriate low pass filter [24]. In addition, Fig. 12 shows simulation result of duty cycle of PWM output signal against $dV/dt$ of input signal at 50 kHz PWM frequency, which is relative to theoretical duty cycle as Eq. (24). It shows that the proposed circuit provides a duty cycle linearly dependent on a slope of input signal, which is consistent to Eq. (24).

5. Conclusions

A simple current controlled oscillator, which is able to generate square/triangular waves has been proposed. Its features are that the oscillation frequency and amplitudes are independent of temperature. However, from Eqs. (13)–(17), we can also see that it is insensitive to power supply voltage, which is a desirable condition [15]. The oscillation frequency and amplitudes can be independently adjusted by the bias currents, which can be easily modified to a voltage form [1], [2] to function as VCO. The proposed circuit can provide electronic frequency, amplitude and DC offset control over a relatively wide range of frequencies. In addition, without any additional components, the same circuit can function as simple PWM signal generator of differentiated result of information signal. The features
of proposed PWM signal generation scheme are that, the frequency of PWM output signal is independent of magnitude of information signal which differs from conventional PWM signal generation, it can generate a precise PWM signal even adjacent of carrier frequency and modulating frequency is involved. Furthermore, with the proposed circuit, its carrier frequency and amplitude can be electronically adjusted with insensitivity of temperature. Therefore, the proposed circuit is especially suitable for further fabricating to use in communication systems over a wide range of operating frequencies. Our future work is to develop the circuit to obtain a lower temperature coefficient.

References

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