Noise Analysis and Characterization of a Sigma-Delta Capacitive Microaccelerometer

Haluk Külah, Member, IEEE, Junseok Chae, Member, IEEE, Navid Yazdi, and Khalil Najafi, Fellow, IEEE

Abstract—This paper reports a high-sensitivity low-noise capacitive accelerometer system with one micro-g/√Hz resolution. The accelerometer and interface electronics together operate as a second-order electromechanical sigma-delta modulator. A detailed noise analysis of electromechanical sigma-delta capacitive accelerometers with a final goal of achieving sub-μg resolution is also presented. The analysis and test results have shown that amplifier thermal and sensor charging reference voltage noises are dominant in open-loop mode of operation. For closed-loop mode of operation, mass-residual motion is the dominant noise source at low sampling frequencies. By increasing the sampling frequency, both open-loop and closed-loop overall noise can be reduced significantly. The interface circuit has more than 120 dB dynamic range and can resolve better than 10 aF. The complete module operates from a single 5-V supply and has a measured sensitivity of 960 mV/g with a noise floor of 1.08 μg/√Hz in open-loop. This system can resolve better than 10 μg/√Hz in closed-loop.

Index Terms—Capacitive readout, inertial sensors, microaccelerometers, micro-g, sigma-delta, switched capacitor.

I. INTRODUCTION

High-precision accelerometers with micro-g (μg, g = 9.8 m/s²) resolution have many applications, including inertial navigation and guidance, microgravity measurements in space, tilt control and platform stabilization, seismometry, and GPS-aided navigators for the consumer market. To achieve μg resolution, a few transduction techniques, device structures, and system approaches have been reported [1]–[5]. Recently, capacitive accelerometers have become very attractive for high-precision μg applications due to their high sensitivity, low temperature sensitivity, low power consumption, wide dynamic range of operation, and simple structure. However, no micromachined capacitive accelerometer system has yet been reported in the literature with sub-μg/√Hz noise floor at atmospheric pressure.

The microaccelerometer system consists of two main parts: the sensing structure and the interface electronics. As well as the sensor structure itself, the interface electronics also plays a critical role in the overall system performance. In fact, noise analysis of the accelerometer, electronic circuit, and the overall system shows that as the device performance improves, the interface electronics limit the overall system resolution.

Sigma-delta (ΣΔ) modulators are very popular for low-frequency analog-to-digital conversion in applications such as speech processing where the oversampling ratio can be considerably high and the noise rejection is very efficient [6]. In micromechanical accelerometers, since the mechanical bandwidth is usually quite small (<2 kHz), sigma-delta conversion can effectively reduce noise and improve overall performance [7]–[14]. In most of the reported systems, the sensor’s mechanical noise is the dominant factor limiting the overall performance. Therefore, the general trend is toward improving the accelerometer itself rather than analyzing the electrical interface electronics and improving the overall system noise performance.

We have previously reported a high-performance silicon microaccelerometer [15] and its open- and closed-loop operation using a switched-capacitor readout circuit [16], [17]. The performance parameters of the system have shown that although the sensor’s mechanical noise floor is less than 1 μg/√Hz, the overall system noise is larger, indicating that the interface electronics is the dominant noise source. In this paper, a detailed noise analysis of the ΣΔ microaccelerometer system is presented and a 1 μg/√Hz accelerometer system is demonstrated. In Section II, a brief overview of the micro-g accelerometer is presented. Then, the front-end circuit operation is described in Section III. The noise analysis of the overall system is presented in Section IV. Finally, measurement results are discussed in Section V.

II. MICRO-G ACCELEROMETER

The accelerometer, shown in Fig. 1, is all-silicon and fabricated on a single silicon wafer using a combined surface and bulk micromachining fabrication process [15]. Fig. 2 shows the cross section of the accelerometer fabricated in this technology. The device consists of a wafer-thick proof mass suspended symmetrically between two stiffened polysilicon electrodes on top and bottom. In the presence of an external acceleration in the z-direction, the silicon frame moves with respect to the proof mass, and the air gaps separating the proof mass from top and bottom electrodes change in opposite directions. Hence, the difference between $C_{\text{top}}$ and $C_{\text{bottom}}$ provides a capacitance change that is a measure of the applied acceleration. The device has a large proof mass (milligrams), controllable/small damping, and narrow air gap that result in large capacitance variation and low mechanical noise floor. It also offers a low offset and long term gain stability as it is all-silicon and no wafer bonding is used in its fabrication process. The
measured differential sensitivity of the sensor with a double clamped-clamped bridge suspension is about 4.9 pF/g on top of a 38 pF rest capacitance for a device with 2 mm $\times$ 1 mm proof mass (2.2 mgr) in a full-bridge configuration and the resonance frequency is around 1 kHz. The sensitivity can be increased by more than an order of magnitude by using a cantilever suspension instead. In order to increase the sensitivity of the microaccelerometer and improve the overall signal-to-noise ratio, a narrow air gap of 1.5 $\mu$m is used. This narrow gap and small resonance frequency result in limited linearity and range in an open-loop mode of operation. However, in this mode the required interface IC is simpler and no stability concerns exist. In order to extend the linearity, range, and bandwidth of the accelerometer, it can be operated in closed-loop.

The interface circuit needs to resolve $<10$ aF capacitance in spite of the large rest capacitance and parasitics (tens of pFs) associated with hybrid packaging of the sensor-interface IC module to attain sub-$\mu$g overall resolution. Also in order to provide closed-loop operation and null the large proof mass motion, the interface chip needs to provide tens of $\mu$N electrostatic force, which is relatively large for microsensors with limited ($<5$ V) power supply. Furthermore, the IC is required to have very low offset, and good gain and offset stability (0.01% full-scale) to qualify the micro-g accelerometer for inertial navigation applications.

III. INTERFACE CIRCUIT

The microaccelerometer is interfaced with a capacitive readout circuitry to form a second-order electromechanical sigma-delta modulator. Interface electronics detect the capacitance change and operate the sensor in open-loop or force-rebalance the proof mass in closed-loop. Fig. 3 shows the block diagram of the interface circuit [18]–[20]. The circuit consists of a switched-capacitor charge integrator, digital feedback (latching comparator and digital compensator), a clock generator, and a start-up circuit. Two fixed reference capacitors are used to form a balanced full-bridge with the sensor capacitive half-bridge, and the sensor top and bottom electrodes are used as the input nodes to the chip front-end.

The readout front-end is a fully differential charge integrator with correlated double sampling (CDS) to cancel 1/f noise, amplifier offset and compensate finite amplifier gain as shown in Fig. 4. Fig. 5 shows the clock diagram for operating this circuit. The operation principle of this circuit has been presented in detail in [18]. The next section discusses the noise sources of this system.

IV. NOISE ANALYSIS

There are several noise sources affecting the overall system resolution of an accelerometer system. These noise sources can be classified in two main groups: mechanical and electrical [19], [20], [22], [23]. Mechanical noise is due to the Brownian motion of the proof mass and is directly related to the sensing structure design and environment. It has been shown that this noise can be decreased down to 0.1 $\mu$g/$\sqrt{\text{Hz}}$ [15], [16], [21]. These accelerometers achieve high device sensitivity, low mechanical noise floor, and controllable damping by combining surface and bulk micromachining. The central idea behind the process is to use the whole wafer thickness to attain a large proofmass, to utilize a sacrificial thin film to form a uniform and conformal gap over a large area, and to create electrodes by depositing polysilicon on the wafer [15], [16], [21]. The electronic noise has different components including the front-end amplifier noise, $kT/C$ noise, noise due to mass residual motion,
sensor charge referencing voltage noise and clock jitter noise. Some of these noise sources are dominant in open-loop operation, whereas the others are critical in closed-loop mode of operation. The following subsections analyze these noise sources individually.

A. Mechanical (Brownian) Noise

Mechanical noise is generated by the proof mass itself. This Brownian noise corresponds to an equivalent acceleration noise of (23), (24):

\[ a_n^2 = \frac{4k_BTb}{m^2} \]  

where \( k_B \) is the Boltzman’s constant, \( T \) is the temperature in Kelvin, and \( b \) is the damping coefficient in (N-m/s), and \( m \) is the proof mass.

As the equation shows, this noise is totally dependent on sensing structure mass and damping coefficient. The z axis accelerometers in the hybrid system tested in this paper have a 0.7 \( \mu \)g/\( \sqrt{\text{Hz}} \) noise floor at atmospheric pressure. This value can be improved further by operating the accelerometer in a vacuum environment, or by increasing the size of the proof mass.

B. Front-End Amplifier Noise

The front-end amplifier noise consists of two parts: thermal and flicker noise. Since CDS is employed in the switched-capacitor circuit, the amplifier flicker noise is reduced considerably, and hence the thermal noise is the dominant source. Fig. 6 shows the schematic of the amplifier used in the front-end of the switched capacitor circuit. This is a fully differential folded-cascode amplifier, and in this structure none of the transistors in the common-mode part contributes to the noise of the amplifier, since the output is taken differentially [6]. Similarly, the transistors in the biasing path do not contribute any noise. Cascode devices \( M_6, M_7, M_{10}, \) and \( M_{11} \) do not affect the total noise either, due to the large impedance in the source leg of these devices. The input-referred noise contribution of the remaining transistors can be derived by multiplying the noise power by the square of the ratio of that device’s transconductance to the input device’s transconductance. Therefore, the input-referred noise can be expressed as [6]

\[ e_n^2 = 2 \left[ e_{M_1}^2 + e_{M_5}^2 \left( \frac{g_{mM_5}}{g_{mM_1}} \right)^2 + e_{M_8}^2 \left( \frac{g_{mM_8}}{g_{mM_1}} \right)^2 \right] \]

where \( g_{mM_1}, g_{mM_5}, \) and \( g_{mM_8} \) are the transistor transconductances and \( e_{M_1}, e_{M_5}, \) and \( e_{M_8} \) are the thermal noise voltages generated by the transistors.

The factor of 2 in this equation results from the fact that the fully differential circuit consists of two matched halves and the noise of those two halves is uncorrelated. Therefore, the total
and is the sensing capacitance, noise compared to (7) can be set such that is the ampli
Noise noise generated by thermal noise sampling of the is the parasitic capacit-
is the output capacitance. By replacing (3) and (5)

\[ e_{\text{out-thermal}} = \sqrt{\frac{16kT}{3C_{\text{int}}f_S}} \left( \frac{V}{\sqrt{Hz}} \right). \]  

It should be noted here that the equivalent noise due to ampli-
thermal noise is independent of transistor parameters. It is
mainly dependent on the capacitance values and the sampling
frequency. By increasing the sampling frequency and the inte-
gration capacitance, it is possible to reduce this noise.

C. $kT/C$ Noise

Another major noise source for the interface electronics is the $kT/C$ noise generated by thermal noise sampling of the
switches. Integration capacitance plays a dominant role in this
noise and the output equivalent noise can be expressed as

\[ e_{\text{out-kt/c}} = \sqrt{\frac{4kT}{f_S C_{\text{int}}}} \left( \frac{V}{\sqrt{Hz}} \right). \]  

As indicated in the equation, this noise component is also
inversely proportional to sampling frequency and integration ca-
pacitance, which means that it can be decreased by increasing
these two factors.

Sensors used in our accelerometer systems have large base
capacitances (tens of pFs) as explained in the previous section.
Therefore, the capacitances employed in the switched-capacitor
circuit are also large resulting in low $kT/C$ noise compared to
other accelerometer systems.

D. Sensor Charging Reference Voltage (SCRV) Noise

Sensor readout is performed by charging the sense capaci-
tance with a fixed reference voltage in each cycle and detecting
this charge by the interface electronics. Therefore, any noise on
this reference voltage directly contributes to the overall noise
performance, which is known as sensor charging reference
have a low bandwidth compared to the sampling frequency, and this is why ΣΔ modulators are so popular for low-frequency bandwidth applications. Since the resonant frequency of the accelerometer is less than 1 kHz, the 1-MHz sampling clock provides a high oversampling ratio, which results in negligible quantization noise. Quantization noise is less than 0.02 μg in 1-Hz bandwidth for 1-MHz sampling clock and Δ = 2.6 g.

F. Mass Residual Motion

This noise source is only effective in closed-loop mode of operation like the quantization noise. It is the result of digital feedback in force-rebalancing [22]. Electrostatic feedback is applied by means of a pulsewidth modulated (PWM) digital pulse train. This pulse train results in a periodic motion of the proof mass around the equilibrium condition, even under zero external acceleration. This movement of the proof mass cannot be separated from an external acceleration and appears as noise in the input. This movement can be represented by the equation [22]

\[ Δx = \frac{a_{\text{max}}}{(2π × (f_s/4))^2} \]

(10)

where \( a_{\text{max}} \) is the maximum acceleration and \( f_s \) is the sampling frequency. For \( a_{\text{max}} = 1.35 \text{ g} \) and \( f_s = 1 \text{ MHz} \), \( Δx \) is equal to \( 5.4 × 10^{-12} \text{ m} \). For a z axis accelerometer with 2 mm × 1 mm area and 1.5 μm gap, this movement creates an equivalent acceleration of 0.05 μg/√Hz. Notice that this noise source is inversely proportional to \( f_s^2 \), whereas the other sources are inversely proportional to \( f_s \). Therefore, for low sampling frequencies, this noise source can rise considerably and become dominant, resulting in tens of μg overall resolution.

Table I presents the individual noise components, their expressions and values for different parameters. As the table shows, most of the electrical noise sources mainly depend on sampling frequency and the value of integration capacitance. Fig. 9 shows the dependence of total electronics noise on integration capacitance and sampling frequency. As seen from the figure, it is possible to minimize the total noise considerably by increasing the sampling frequency and the integration capacitance. However, the sampling frequency cannot be increased arbitrarily due to circuit limitations, such as amplifier

### Table I

<table>
<thead>
<tr>
<th>Noise Source</th>
<th>Expression</th>
<th>( f_s = 100 \text{kHz} )</th>
<th>( f_s = 1 \text{MHz} )</th>
<th>( f_s = 1 \text{MHz} )</th>
</tr>
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<tbody>
<tr>
<td></td>
<td></td>
<td>( C_{\text{int}} = 5 \text{pF} )</td>
<td>( C_{\text{int}} = 5 \text{pF} )</td>
<td>( C_{\text{int}} = 15 \text{pF} )</td>
</tr>
<tr>
<td>Front-end amplifier</td>
<td>( \frac{16 C_s + C_p}{3} \frac{kT}{C_{\text{int}}} )</td>
<td>0.66μV/√Hz</td>
<td>0.21μV/√Hz</td>
<td>0.12μV/√Hz</td>
</tr>
<tr>
<td>kT/C ( \sqrt{f_{C_{\text{int}}}} )</td>
<td>( \frac{4kT}{f_{C_{\text{int}}}} )</td>
<td>0.18μV/√Hz</td>
<td>0.06μV/√Hz</td>
<td>0.03μV/√Hz</td>
</tr>
<tr>
<td>Sensor Charging Reference Voltage</td>
<td>( \frac{2V_s^2}{C_{\text{int}}} )</td>
<td>0.89μV/√Hz</td>
<td>0.28μV/√Hz</td>
<td>0.16μV/√Hz</td>
</tr>
<tr>
<td>Quantization ( n_{\text{rms}} = \frac{e_{\text{rms}}^2}{\sqrt{5M^{2.5}}} )</td>
<td>0.08μV/√Hz</td>
<td>0.0025μV/√Hz</td>
<td>0.0025μV/√Hz</td>
<td></td>
</tr>
</tbody>
</table>

\*\( C_s + C_p = \frac{100}{pF} \), \( C_{\text{int}} = 10 \text{ pF} \), \( V_s^2 \) is the charging reference voltage noise assumed to be white with a spectral density of 10 nV/√Hz, \( C_s = 10 \text{ pF} \).
slew rate and unity gain bandwidth. Increasing the integration capacitance decreases the sensitivity of the front-end charge integrator, and hence decreases the signal-to-noise ratio even though it improves the absolute voltage noise. Therefore, the integration capacitance and the sampling frequency should be optimized to achieve desired resolution and open-loop dynamic range.

According to simulations, it is possible to improve the overall system resolution down to hundreds of nano-g level while achieving a high dynamic range by operating the circuit at 1-MHz sampling clock with a 15-pF integration capacitance. However, operating the system under this condition requires a high-performance front-end circuit capable of driving high capacitive loads with a high slew rate and low noise. In this design, a high-slew-rate front-end amplifier with 85-dB DC gain and 12.3-MHz unity gain bandwidth was implemented. Moreover, the input-referred noise of each individual circuit block has been minimized to achieve a low overall system noise performance. The next section summarizes the implementation of this new circuit and presents the test results.

V. IMPLEMENTATION AND TEST RESULTS

According to the noise analysis summarized in the previous section, the interface electronics was designed for high-frequency operation. The noise analysis shows that increasing the sampling frequency from 200 kHz to 1 MHz improves the noise performance significantly, but a further increase does not provide such a drastic improvement. Therefore, the chip is designed to operate at sampling frequencies higher than 1 MHz. The individual blocks of the circuit, such as the operational amplifier and bias generator, were improved to achieve lower noise floor.

The interface chip was designed in 0.5-μm three-metal two-poly n-well CMOS process. Fig. 10 shows the fabricated circuit. All critical individual blocks of the interface chip were tested extensively and the functionality was verified. It was observed through the noise measurements that the CDS technique eliminates the 1/f noise significantly, as expected theoretically. The circuit dissipates less than 7.2 mW from a single 5-V supply and operates from a 1-MHz clock. It has an adjustable sensitivity between 0.2 and 1.2 V/pF using a laser trimmable capacitance array. Table II summarizes the performance parameters of the interface chip.

The CMOS interface chip is combined with a z-axis accelerometer to verify the performance improvement in the system. Fig. 11 shows the z-axis hybrid system with the sensor and the circuit assembled onto a PC board and mounted inside a standard DIP package. Since the sensor’s mechanical noise is very low, there is no need to use vacuum packaging.
A. Open-Loop Tests

Open-loop tests were performed on a dividing head, in a 1-g gravitational field, by changing the acceleration on the sensor from $-1 \, \text{g}$ to $+1 \, \text{g}$. While changing the applied acceleration, the differential analog output voltage of the interface electronics was measured. Fig. 12 shows a measured open-loop sensitivity of 960 mV/g.

The output noise of the hybrid module is measured at a 1-MHz sampling frequency by using an HP 3561 dynamic signal analyzer with a 50-kΩ reference resistor as shown in Fig. 13. This figure indicates that the resistor has 32 nV$/\sqrt{\text{Hz}}$ noise density which matches well with the estimated thermal noise of the resistor (note that the measurement bandwidth is 11.72 Hz), thus verifying the calibration of the measurement setup. From the measured output, the hybrid module can resolve $1.08 \, \mu\text{g}/\sqrt{\text{Hz}}$. It is believed that the periodic peaks in this measurement are due to environmental factors and are not due to the accelerometer system.

Fig. 14 shows the dependence of the open-loop noise floor on sampling frequency. As shown in the figure, although there is a little difference between the two curves for all frequencies, the theoretical and measured curves have the same trend and the noise floor decreases with increasing sampling frequency as expected.

B. Closed-Loop Tests

The closed-loop test setup uses a shaker table, a data acquisition board, and LABVIEW and MATLAB programs for signal processing. Since the interface electronics uses a high over-sampling sigma-delta modulation technique, the PWM output bit stream has to be processed to obtain a useful signal. This is realized by transferring the digital output to a computer by means of a data acquisition board, and processing the signal (decimating and digital filtering). A sinc$^3$ filter, FIR filter, decimator, and digital-to-analog converter have been implemented in MATLAB for this purpose.

The entire system has been operated in closed-loop and the functionality of the system has been verified through extensive tests. Fig. 15 shows the decimated PWM digital outputs for (a) a pure 1-g DC input, and (b) a 0.25-g sinusoidal input acceleration on top of a 1-g DC signal. As the figure shows, the applied input acceleration is recovered successfully. Note that in Fig. 15(a), the only applied acceleration is the 1-g gravitational field. The output voltage is constant, except for variations due to noise generated in the system and/or picked up from the environment.

Fig. 16 shows the Fourier transform of the processed PWM output for 1-g DC bias for sampling frequencies of 100 kHz and 400 kHz. As the figure shows, by increasing the sampling frequency four times, the noise floor decreases by approximately 16 times. This means that the noise is inversely proportional to...
Fig. 15. Closed-loop measurement results for the hybrid sensor system: (a) for 1-g DC input acceleration, and (b) for 0.25-g sinusoidal input acceleration on top of 1-g DC input.

\( f_0^2 \), and hence the mass residual motion is dominant. It has been observed that this noise source is not dominant for higher sampling frequencies. Moreover, as the span of the measurement increased beyond 15 Hz, the undesired peaks become insignificant and the noise level stays constant at higher frequencies.

These results indicate that at sampling frequencies lower than 400 kHz, the mass residual motion is the dominant noise source in closed-loop mode of operation. As the sampling frequency is increased more than 400 kHz, this noise source becomes insignificant compared to others and the overall noise is improved by the square root of the sampling frequency. The system can resolve better than 10 \( \mu g \) in closed-loop mode for a sampling frequency of 400 kHz. Table III summarizes the measured system parameters.

VI. CONCLUSION

A second-order electromechanical sigma-delta microaccelerometer system has been analyzed in terms of noise to identify the limiting factors and an improved system has been implemented. Brownian noise, front-end amplifier thermal noise, \( kT/C \) noise, mass residual motion, sensor charge referencing voltage (SCRV) noise, and quantization noise are the main noise components affecting the sigma-delta modulator performance. The noise analysis and the test results have shown that in open-loop operation, the front-end amplifier thermal noise and SCRV noise are dominant. In closed-loop mode of operation, the mass residual motion becomes critical especially at low sampling frequencies, whereas the amplifier and SCRV noises become dominant at sampling frequencies higher than 400 kHz. Sensors have 0.7 \( \mu g/\sqrt{Hz} \) sensitivity with 1.08 \( \mu g/\sqrt{Hz} \) noise floor in open-loop. The closed loop operation of the system provides a resolution better than 10 \( \mu g/\sqrt{Hz} \). Since the open-loop noise at the 1-MHz sampling frequency is 1.08 \( \mu g/\sqrt{Hz} \), the expected noise floor

<table>
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<tr>
<th>Sensor Application</th>
<th>Noise Floor [( \mu g/\sqrt{Hz} )]</th>
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<tr>
<td>Open-Loop Noise Floor</td>
<td>1.08</td>
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<tr>
<td>Closed-Loop Noise Floor</td>
<td>&lt;10</td>
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Table III PERFORMANCE PARAMETERS OF THE HYBRID SYSTEM

<table>
<thead>
<tr>
<th>CMOS readout electronics</th>
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<tr>
<td>Sensitivity [0.2-1.2V/pF]</td>
</tr>
<tr>
<td>Dynamic Range [&gt;120dB]</td>
</tr>
<tr>
<td>Resolution [&lt;10aF]</td>
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<tr>
<th>MEMS accelerometers</th>
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<tr>
<td>Sensitivity [pF/g]</td>
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<tr>
<td>Mech. Noise [( \mu g/\sqrt{Hz} )]</td>
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<th>MEMS device and interface circuit module</th>
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<tbody>
<tr>
<td>Sensitivity [V/g]</td>
</tr>
<tr>
<td>Open-Loop Noise Floor [( \mu g/\sqrt{Hz} )]</td>
</tr>
<tr>
<td>Closed-Loop Noise Floor [( \mu g/\sqrt{Hz} )]</td>
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</table>
in the closed-loop mode of operation is around 1.5 μg/Hz. The discrepancy in the measured and theoretical values has not been explicitly determined, but could be due to test setup and environmental factors.

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REFERENCES


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