



2 MHz PWM Boost Switching Regulator

Preliminary Technical Data

ADP3031

FEATURES

- Up to 2 MHz PWM Frequency
- Fully Integrated 1.5 A Power Switch
- 3% Output Regulation Accuracy
- Adjustable Output Voltage from 3 V to 12 V
- Simple Compensation
- Small Inductor and MLC Capacitors
- 90% Efficiency
- Under-voltage Lockout
- Shutdown

APPLICATIONS

- TFT LCD Bias Supplies
- Portable Electronic Instruments
- Industrial/Instrumentation Equipment

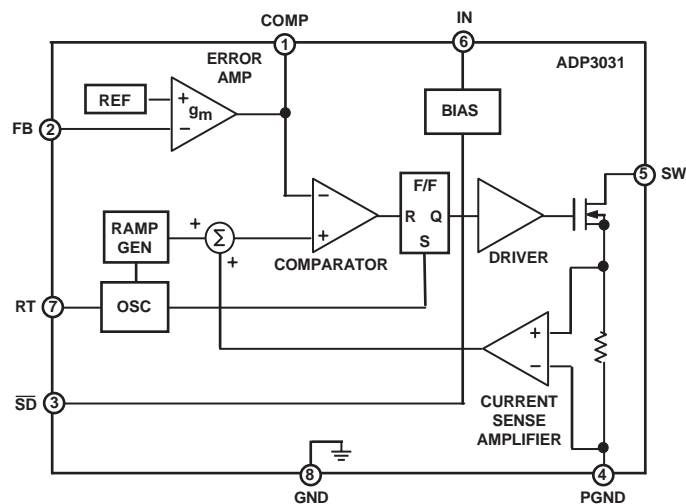
GENERAL DESCRIPTION

The ADP3031 is a high frequency, step-up DC-DC switching regulator capable of 12 V boosted output voltage in a space saving MSOP-8 package. It provides high efficiency, low-noise operation, and is easy to use. Capable of operating up to 2 MHz, the high switching frequency and PWM current mode architecture allow for excellent transient response, ease of noise filtering, and also small, cost-saving external inductive and capacitive components. The current limit and the power switch are integrated completely on chip.

Capable of operating from 2.5 V to 5.5 V input, the ADP3031 is ideal for Thin-Film Transistor (TFT) Liquid Crystal Display (LCD) module applications, where local point-of-use power regulation is required. Featuring an adjustable output that supports voltages down to 3 V, the ADP3031 is ideal to generate today's low voltage rails, delivering power efficiently, and simply with minimal printed circuit board area.

The device is specified over the industrial temperature range of -40°C to $+85^{\circ}\text{C}$.

Functional Block Diagram



REV. PrB 8/2/02

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PRELIMINARY TECHNICAL DATA

ADP3031—SPECIFICATIONS¹ ($V_{IN} = +3.3\text{ V}$, $T_A = -40^\circ\text{C}$ to 85°C , unless otherwise noted)

Parameter	Symbol	Conditions	Min	Typ	Max	Units
SUPPLY						
Input Voltage	V_{IN}		2.5		5.5	V
Quiescent Current						
Switching State ³	I_{QSW}	$f = 600\text{ kHz}$, light load		2	5	mA
Non-switching State	I_Q			300	500	μA
Shutdown	I_{SD}				1	μA
ERROR AMPLIFIER						
Feedback Voltage	V_{FB}		TBD	1.233	TBD	V
Line Regulation		$V_{IN} = 2.5\text{ V}$ to 5.5 V	-0.1		+0.1	%/V
Bias Current				100		nA
Overall Regulation		Line, Load, Temp	-3		3	%
Transconductance	g_m			100		$\mu\text{A/V}$
Gain	A_v			1000		V/V
SWITCH						
On Resistance	R_{ON}	at $I_{SW} = 1.5\text{ A}$, $V_{IN} = 3.3\text{ V}$		200	300	m Ω
Output Load Current ²	I_{LOAD}	continuous operation	200	300		mA
Leakage Current		$V_{IN} = 3.3\text{ V}$, $V_{OUT} = 10\text{ V}$			5	μA
Efficiency ²		$V_{SWITCH} = 12\text{ V}$		85		%
Efficiency ²		$I_{LOAD} = 200\text{ mA}$, $V_{OUT} = 10\text{ V}$, $f = 600\text{ kHz}$		90		%
Peak Current Limit	I_{CLSET}	$I_{LOAD} = 100\text{ mA}$, $V_{OUT} = 10\text{ V}$, $f = 600\text{ kHz}$	1.5	1.8		A
Output Voltage Range	V_{OUT}		3	7	12	V
OSCILLATOR						
Oscillator Frequency	FOSC	RT = Open	0.5	0.6	0.7	MHz
		RT = GND	1.7	2	2.3	MHz
Maximum Duty Cycle	D_{MAX}	COMP = open, FB = 1 V		80	85	%
SHUTDOWN						
Shutdown Input Voltage Low					0.8	V
Shutdown Input Voltage High			2.2			V
UNDER-VOLTAGE LOCKOUT						
UVLO Threshold			2.2	2.45	2.5	V
UVLO Hysteresis				100		mV

NOTES

¹ All limits at temperature extremes are guaranteed via correlation and characterization using standard Statistical Quality Control (SQC).

² See Figure xx.

³ This is the average current while switching.

Specifications subject to change without notice.

ABSOLUTE MAXIMUM RATINGS*

Input Voltage	-0.3 V to +6 V
SW Voltage	14 V
COMP Voltage	-0.3 V to +2.5 V
FB Voltage	-0.3 V to +1.3 V
\overline{SD} Voltage	-0.3 V to +6 V
RT Voltage	-0.3 V to +1.2 V
PGND TO GND	± 200 mV
Operating Ambient Temperature Range	-40°C to 85°C
Operating Junction Temperature Range	-40°C to 125°C
Storage Temperature Range	-65°C to +150°C
θ_{JA} Two Layer	206°C/W
θ_{JA} Four Layer	142°C/W
Lead Temperature Range (Soldering, 60 sec.)	300°C

*This is a stress rating only; operation beyond these limits can cause the device to be permanently damaged. Unless otherwise specified, all voltages are referenced to GND

PIN FUNCTION DESCRIPTIONS

Pin	Mnemonic	Function
1	COMP	Compensation Input.
2	FB	Feedback voltage sense input.
3	\overline{SD}	Shutdown Input.
4	PGND	Power Ground. Ground return for power transistor.
5	SW	Switching Output.
6	IN	Main Power Supply Input.
7	RT	Frequency Setting Input. A resistor between this pin and GND sets the switching frequency of the device.
8	GND	Analog Ground. The control circuitry is referenced to this ground.

ORDERING GUIDE

Model	Voltage Output	Package Option	Branding Information
ADP3031ARM	ADJ	MSOP-8	P8A

PIN CONFIGURATION

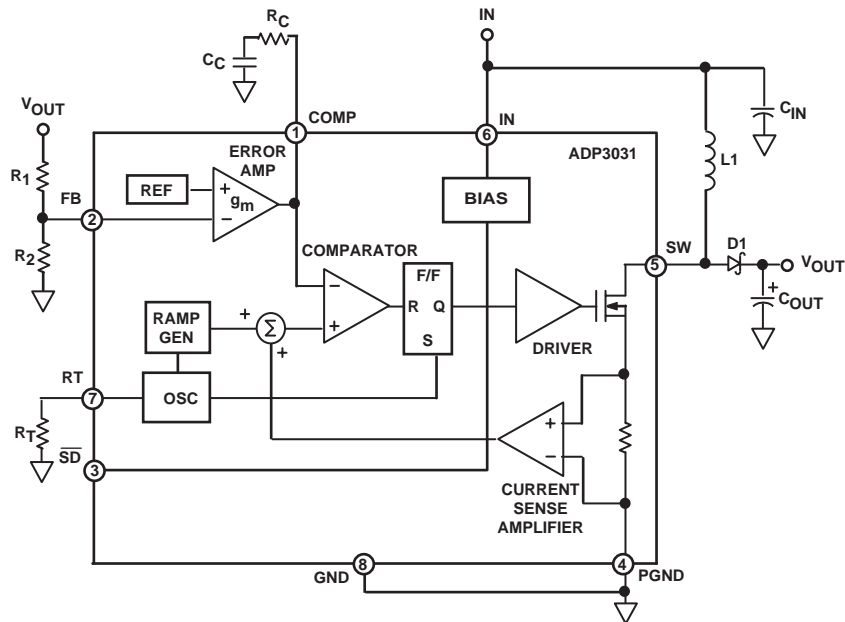
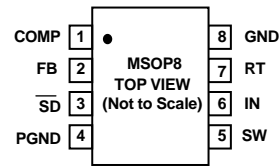


Figure 1. Typical Application

CAUTION

ESD (electrostatic discharge) sensitive device. Electrostatic charges as high as 4000 V readily accumulate on the human body and test equipment and can discharge without detection. Although the device features proprietary ESD protection circuitry, permanent damage may occur on devices subjected to high energy electrostatic discharges. Therefore, proper ESD precautions are recommended to avoid performance degradation or loss of functionality.



ADP3031

THEORY OF OPERATION

The ADP3031 is a boost converter driver which stores energy from an input voltage in an inductor, and delivers that energy, augmented by the input, to a load at a higher output voltage. It includes a voltage reference and error amplifier to compare some fraction of the load voltage to the reference, and amplify any difference between them. The amplified error signal is compared to a dynamic signal produced by an internal ramp generator incorporating switch current feedback. The comparator output timing sets the duty ratio of a switch driving the inductor to maintain the desired output voltage.

Referring to Figure 1, a typical application will power both the IC and the inductor from the same input voltage. The on chip MOSFET will be driven on, pulling pin SW close to PGND. The resulting voltage across the inductor will cause its current to increase approximately linearly, with respect to time.

When the MOSFET switch is turned off the inductor current cannot drop to zero, and so this current drives the SW node capacitance rapidly positive until the diode becomes forward biased. The inductor current will now begin to charge the load capacitor, causing a slight increase in output voltage. Generally, the load capacitor is made large enough that this increase is very small during the time the switch is off. During this time inductor current is also delivered to the load. In steady state operation, the inductor current will exceed the load current, and the excess will be what charges the load capacitor. The inductor current will fall during this time, though not necessarily to zero.

During the next cycle, initiated by the on-chip oscillator, the switch will again be turned on so that the inductor current will be ramped back up. The charge on the load capacitor will provide load current, during that interval. The remainder of the chip is arranged to control the duty ratio of the switch, to maintain a chosen output voltage despite changes in input voltage or load current.

The output voltage is scaled down by a resistor voltage divider and presented to the g_m amplifier. This amplifier operates on the difference between an on-chip reference and the voltage at the FB pin so as to bring them to balance. This will be when the output voltage equals the reference voltage, multiplied by the resistor voltage divider ratio.

The g_m amplifier drives an internal comparator, which has at its other input a positive going ramp produced by the Oscillator and modified by the current sense amplifier. The MOSFET switch is turned on as the modified ramp voltage rises. When this voltage exceeds the output of the g_m amplifier the comparator will turn off the switch, by resetting the flip-flop, previously set by the oscillator. The output of the flip-flop is buffered by a high current driver which turned on the MOSFET switch at the beginning of the Oscillator cycle.

In the steady state with constant load and input voltage, the current in the inductor will cycle around some average current level. The increasing ramp of current will depend on input voltage and t_1 , the switch on-time, while the decreasing ramp will depend on the difference between input and output voltage and t_2 , the remainder of the cycle. In order for the peaks of these two ramps to be equal and opposite to

maintain steady state we can say that $t_1 \cdot V_{IN}$ will equal $t_2 \cdot (V_{OUT} - V_{IN})$, if we neglect the effect of resistance in the inductor and switch, and the forward voltage drop of the diode. From this equality we can derive $t_1/T = 1 - V_{IN}/V_{OUT}$, where T is the period of a cycle, $t_1 + t_2$. This result gives us the switch duty ratio, t_1/T in terms of the input and output voltages.

In practice the duty ratio will need to be slightly higher than this calculation. Because of series resistance in the inductor and the switch, the voltage across the actual inductance is somewhat less than applied V_{IN} , and the actual output voltage is less than our approximation by the amount of the diode forward voltage drop. However, the feedback control within the ADP3031 will adjust the duty ratio to maintain the output voltage. Changes in load current and input voltage are also accommodated by the feedback control.

Changes in load current alone require a change in duty ratio, in order to change the average inductor current. But once the inductor current adapts to the new load current, the duty ratio should return to nearly its original value, as we see from the duty cycle calculation which depends on input and output voltages, but not on current. Increasing the switch duty ratio initially reduces the output voltage, until the average inductor current increases enough to offset the reduction of the t_2 interval. By limiting the duty ratio we prevent this effect from regeneratively increasing the duty ratio to 100%, which would cause the output to fall and the switch current to rise without limit. The duty ratio is limited to about 80% by the design of the Oscillator and an additional flip-flop reset.

A comparator compares the current sense amplifier output to a factory set limit which resets the flip-flop, turning off the switch. This prevents runaway or overload conditions from damaging the switch and reflecting fault overloads back to the input. Of course, the load is directly connected to the input by way of the diode and inductor, so protection against short circuited loads must be done at the power input.

The g_m amplifier has high voltage gain, to insure the output voltage accuracy and invariance with load and input voltage. However, because it is a g_m amplifier with a specified current response to input signal voltages, its high frequency response can be controlled by the compensation impedance. This permits the high frequency gain of the g_m amplifier to be optimized for the best compromise between speed of response and frequency stability.

The stable closed loop bandwidth of the system can be extended by the current feedback shown. A signal representing the magnitude of the switch current is added to the ramp. This dynamically reduces the duty ratio, as the current in the inductor increases, until the g_m amplifier restores it, improving the closed loop frequency stability.

The ADP3031 is intended to operate over a range of frequencies, set by the RT pin. If the pin is open, the oscillator runs at its lowest frequency: if the pin is

“grounded” it runs at its highest frequency. A resistor from RT to ground can be used to set intermediate operating frequencies.

Because of the large currents which flow in the main MOSFET switch, it is provided with a separate PGND return to the negative supply terminal, to avoid corrupting the small signal return, GND, that can be used as a sense line at the output load point.

APPLICATION INFORMATION

Frequency Selection

The ADP3031's frequency can be user selected to operate at either 600 KHz or 2 MHz and programmable by setting the RT pin. Tie RT to GND for 2 MHz operation. For 600 KHz operation, float the RT pin.

The nominal resistance at the RT pin to get a switching frequency, f_{sw} , is given by:

$$RT (\Omega) = 320,000 \times (2,000,000 - f_{sw}) / (3.6667 \times f_{sw} - 2,000,000) \quad (1)$$

Output Voltage

The ADP3031 features an adjustable output voltage range of V_{IN} to 12 V. The output voltage is fed back to the ADP3031 via resistor dividers R1 and R2 (Figure 1.). The feedback voltage is 1.233 V, so the output voltage is set by the formula:

$$V_{OUT} = 1.233 \times (1 + R1/R2) \quad (2)$$

Since the feedback bias current is 100 nA maximum, R2 may have a value up to 100 K Ω with minimum error due to the bias current.

Inductor Selection

For most of the applications, the inductor used with the ADP3031 should be in the range of 2 μ H to 22 μ H. Several inductor manufacturers are listed in Table 1. When selecting an inductor, it is important to make sure that the inductor used with the ADP3031 is able to handle the peak current without saturation and that the peak current is below the current limit of the ADP3031.

As a rule, powdered iron cores saturate softly, whereas Ferrite cores saturate abruptly. Open drum core inductors tend to saturate gradually, are low cost and are small in size, making these types of inductors attractive in many applications. However, care must be exercised in their placement because they have high magnetic fields. In applications that are sensitive to magnetic fields, shielded geometries are recommended.

In addition, inductor losses must be considered. Both core and copper losses contribute to loss in converter efficiency. To minimize core losses, look for inductors rated for operation at high switching frequencies. To minimize copper losses, it is best to use low dc resistance inductors. Typically, it is best to use an inductor with a dc resistance lower than 20 m Ω per μ H.

The inductor value can be estimated using the following:

$$L = (V_{OUT} - V_{IN}) \times M_{SLOPE}$$

Where M_{SLOPE} = scaling factor for proper slope compensation.

$$M_{SLOPE} = \frac{1.456}{f_{SW}}$$

Choose the closest standard inductor value as a starting point.

The corresponding peak inductor current can then be calculated:

$$I_L (PEAK) = \left(I_{OUT} \times \frac{V_{OUT}}{V_{IN}} \right) + \frac{1}{2} \left(\frac{V_{IN} \times (V_{OUT} - V_{IN})}{L \times V_{OUT} \times f_S} \right) \quad (3)$$

It is recommended to try several different inductor values, sizes and types to find the best inductor for the application. In general, large inductor values lead to lower ripple current, less output noise, and either larger size or higher DC resistance. Conversely, low inductor values lead to higher ripple current, more noise, and either smaller size or lower DC resistance. The final inductor selection should be based on the best trade-off of size, cost, and performance.

Table 1. Inductor Manufacturers

Part	L(μ H)	Max DC Current	Max DCR m Ω	Height (mm)	Vendor
CMD4D11-2R2MC	2.2	0.95	116	1.2	Sumida
CMD4D11-4R7MC	4.7	0.75	216	1.2	847-956-0666
CDRH4D28-100	10	1.00	128	3.0	www.sumida.com
CDRH5D18-220	22	0.80	290	2.0	
CR43-4R7	4.7	1.15	109	3.5	
CR43-100	10	1.04	182	3.5	
DS1608-472	4.7	1.40	60	2.9	Coilcraft
DS1608-103	10	1.00	75	2.9	847-639-6400 www.coilcraft.com
D52LC-4R7M	4.7	1.14	87	2.0	Toko
D52LC-100M	10	0.76	150	2.0	847-297-0070 www.tokoam.com

ADP3031

Capacitor Selection

The ADP3031 requires an input capacitor to reduce the switching ripple and noise on the IN pin. The value of the input capacitor will be dependent on the application. For most applications, a minimum of 10 μF is required. For applications that are running close to current limit or that have large transient loads, input capacitors in the range of 22 μF to 47 μF are required. The selection of the output capacitor is also dependent on the application. Given the allowable output ripple voltage, ΔV_{OUT} , the criteria for selecting the output capacitor can be calculated using the following equations:

$$C_{OUT} \geq 8 \times I_{OUT} \left(\frac{(V_{OUT} - V_{IN})}{f_s \times V_{OUT} \times \Delta V_{OUT}} \right) \quad (4)$$

$$ESR_{C_{OUT}} \leq \frac{\Delta V_{OUT}}{I_L (PEAK)} \quad (5)$$

When selecting an output capacitor, make sure that the ripple current rating is sufficient to cover the rms switching current of the ADP3031.

The ripple current in the output capacitor is given by the following:

$$I_{RMS}(C_{OUT}) = I_{OUT} \sqrt{\frac{V_{OUT} - V_{IN}}{V_{IN}}} \quad (6)$$

Multi-layer ceramic capacitors are a good choice, as they have low ESR, high ripple current rating and a very small package size. Tantalum or OS-CON capacitors can be used, however they have a larger package size and have higher ESR. Table 2 shows a list of several capacitor manufacturers. Consult the manufacturer for more information.

Table 2. Capacitor Manufacturers

Vendor	Phone #	Web Address
AVX	408-573-4150	www.avxcorp.com
Murata	714-852-2001	www.murata.com
Sanyo	408-749-9714	www.sanyovideo.com
Taiyo-Yuden	408-573-4150	www.t-yuden.com

Diode Selection

In specifying a diode, consideration must be given to speed, the forward current, the forward voltage drop, reverse leakage current, and the breakdown voltage. The output diode should be rated to handle the maximum output current. If the output can be subjected to accidental short circuits then the diode must be rated to handle currents up to the current limit of the ADP3031. The breakdown rating of the diode must exceed the output voltage. A high-speed diode with low forward drop, and low leakage will help improve the efficiency of the converter by lowering the losses of the diode. Schottky diodes are recommended.

Loop Compensation

Like most current programmed PWM converters, the ADP3031 needs compensation to maintain stability over the operating conditions of the particular application. For operation at duty cycles above 50%, the choice of inductor is critical in maintaining stability. If the slope of the inductor current is too small or too large, the circuit will be unstable. See Inductor Selection for more information on choosing the proper inductor.

The ADP3031 provides a pin (COMP) for compensating the voltage feedback loop. This is done by connecting a series RC network from the COMP pin to GND. See Figure 2. For most applications, the compensation resistor, R_C , should be in the range of $30 \text{ k}\Omega \leq RC \leq 300 \text{ k}\Omega$ and the compensation capacitor, C_C , in the range of $100 \text{ pF} \leq C_C \leq 4 \text{ nF}$.

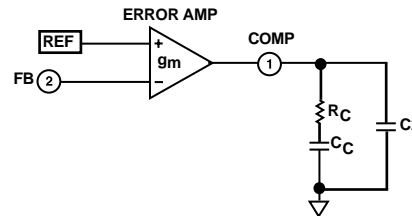


Figure 2. Compensation Components

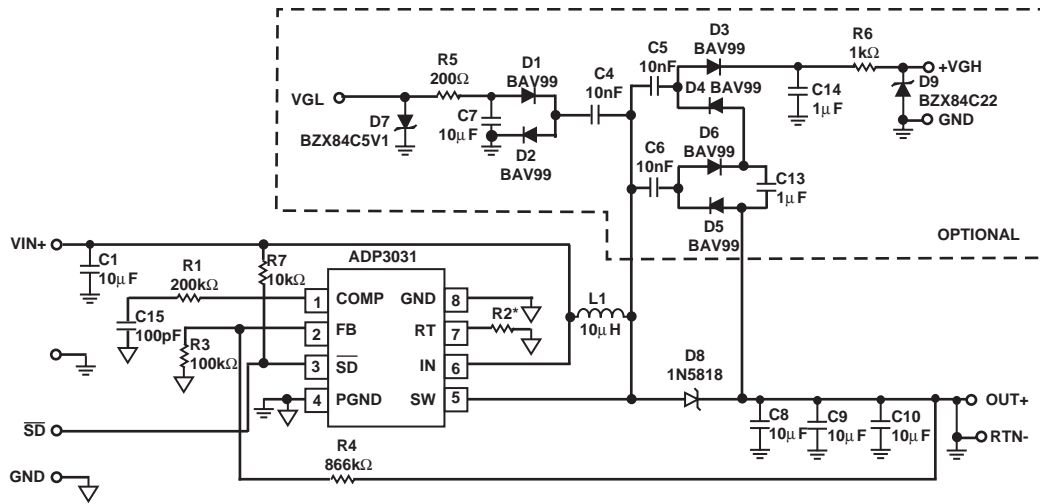
Shutdown

The ADP3031 shuts down to reduce the supply current to 1 μA maximum when the shutdown pin is pulled low. In this mode, the internal reference, error amplifier, comparator, biasing circuitry, and the internal MOSFET switch are turned off. Note that the output is still connected to the input via the inductor and Schottky diode when in shutdown.

Layout Procedure

In order to get high efficiency, good regulation and stability, a good printed circuit board layout is required. It is strongly recommended that the evaluation board layout be followed as closely as possible. Use the following general guidelines when designing printed circuit boards (refer to Figure 1):

1. Keep C_{IN} close to the IN pin of the ADP3031.
2. Keep the high current path from C_{IN} , through L1, to the SW pin and PGND pin as short as possible.
3. Similarly, keep the high current path from C_{IN} , through L1, D1, and C_{OUT} as short as possible.
4. High current traces should be kept as short and as wide as possible.
5. Place the compensation components as close to the COMP pin as possible.
6. Place the feedback resistors as close to the FB pin as possible to prevent noise pickup.
7. Avoid routing noise sensitive traces near the high current traces and components.



* FREQUENCY ADJUST

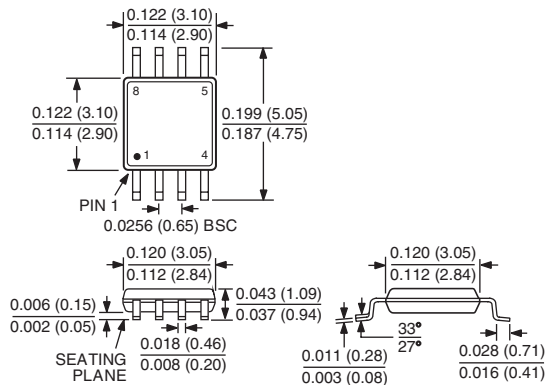
Figure 3. Typical Application for TFT LCD

OUTLINE DIMENSIONS

Dimensions shown in inches and (mm).

RM-8

8-Lead Mini/micro SOIC Package [Mini_SO]



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ADP3031

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