Super Cheap dc/dc
Problem: implement the cheapest possible dc/dc controller

Solution: use a 555 timer—the highest volume IC in history—to implement a peak current controlled, constant off-time converter.

The 555 timer is insanely high volume and thus incredibly cheap. It also serves as the perfect core for a hysteretic converter.

The heart of the timer is the set-dominant R-S latch. When the TRIG input falls below REF/2, the output Q goes high and the DI5 pin tri-states.
THRESH exceeds REF, assuming that TRIG is no longer below REF/2 (since the latch is set dominant), the output goes low and the DIS pin is pulled down. At any time the RESET pin can be pulled low to force the output and DIS pins low.

The above circuit turns the transistor M "on" until the voltage across Rsns equals Vref. Thereafter, Ctrig discharges through Rtrig until the voltage on Ctrig is below Vref/2, at which time the output goes high and Ctrig charges through D, starting the cycle again.

Note that if \( \frac{V_{out}}{V_{in}} \) is great enough that the length of the "on" time is more than about \( 2 \times \) the off time, D is unnecessary because Rtrig has time to charge Ctrig during the on time.
The constant off time aspect of this control method means that the switching frequency changes with input-output voltage ratio. In a boost converter in continuous conduction,

\[
\frac{V_{out}}{V_{in}} = \frac{1}{1-D}
\]

where D is the ratio of "on" time to total cycle length.
A final practical consideration in the current loop is that the trigger requirements for the circuit go into the base of a PNP transistor (Darlington pair). 

The trigger, however, is not an TRIAC. We could change an NPN to a PNP, but this is not recommended because the current into the base of the PNP is just 1.3V, and thus probably does not work very well. 

Now, the exact value of the trigger actually changes somewhat as Vref changes, but a good estimate is that the trigger happens in one time constant, i.e.

\[ \frac{f_{sw}}{V_{n}} = \frac{T_{sw}}{V_{n}} \]

Rearranging,

\[ 1 - D = \frac{V_{n}}{V_{not}} \]

This means that the trigger happens at a current specified by the voltage difference.
Instead, we employ a "floating battery" implemented with a $V_{be}$ multiplier. Consider:

Assuming that the current through $R_{bias}$ is large enough, some current leaks through $R_1$ & $R_2$.

As soon as the voltage across $R_1$ is sufficient to turn on the transistor, the voltage across $R_1$ & $R_2$ (assuming $g 	o \infty$) are fixed:

$$V_{R_1} = V_{BE}$$

$$I_{R_2} = \frac{V_{BE}}{R_1}$$

$$V_{R_2} = R_2 I_{R_2} = \frac{R_2}{R_1} V_{be}$$

$$\therefore V_{out} = \left(1 + \frac{R_2}{R_1}\right) V_{BE}$$

Using this circuit, we add an offset to $V_{sense}$ while leaking an inconsequential current through $R_{sense}$. 
Voltage loop

Now that we can set the current based on a reference, we need another slower feedback loop to set that reference current to produce the desired output voltage.

Since the dynamics of the current loop are very fast (each cycle the output current goes precisely to the reference; effectively this is a pole at \( f_{SW} \)), we can just throw an integrator into our voltage loop to get high DC gain and thus low steady-state error.

The loop crossover is related to the integrator gain, \( \frac{1}{R_{i}C_{i}T} \).
Because of non-zero $R_{DS(on)}$ of the switching transistor, parasitic $R$ of the inductor, and the output resistance of the power source, there is a maximum achievable inductor current, $I_{\text{max}}$. During large transients, if $I_{\text{ref}}$ exceeds $I_{\text{max}}$, the current loop latches up. Such transients occur, for example, at startup.

To prevent this, we clamp the maximum $I_{\text{ref}}$ to a value less than $I_{\text{max}}$. This guarantees that the circuit always oscillates.

In its simplest form, the clamp can just be another $V_{BE}$ multiplier:
However, this circuit is somewhat impractical because the input impedance of the clamp is rather high, requiring a large \( R_{\text{clamp}} \). If the \( \text{REF} \) input on the 555 were high impedance, this would be acceptable; however, the \( \text{REF} \) pin looks into the stack of 5kΩ resistors, and thus has an impedance of only 3.3kΩ. If \( R_{\text{clamp}} \) becomes too large compared to this value, the achievable \( T_{\text{ref}} \) range becomes limited. In systems with a relatively constant load, this is acceptable, but this circuit can be improved rather easily:

Now we can choose \( R_{\text{bias}} \) to be rather large and get better control over clamping while retaining \( T_{\text{ref}} \) range.
However, we can get an even sharper clamping response slightly differently:

Now, we're still using two op-amps, but our clamp uses the op-amp directly, meaning it has much lower input impedance than the BJT-based one.

**LED current control**

In this application, we are driving LEDs. Thus, it's really the current we want to control. To do this, we control the voltage across the tail resistor in the LED string, effectively giving us a constant current output.
Since $V_{\text{tail}} = V_{\text{Diode}}$, $I_{\text{LED}} = \frac{V_{\text{Diode}}}{R_{\text{tail}}}$.

Practically, this requires an op-amp whose input common-mode range includes the negative supply (unless we want to supply the op-amp with a negative voltage), but fortunately one of the cheapest op-amps obtainable, the LM358, not only has PNP inputs, but is also a dual, giving us the ability to implement an improved current reference clamp for free!  

One final consideration: since the clamp current is dependent on the $V_{\text{BE}}$ of the THRESH offset generator and on $V_{\text{ref}}$, both are derived from 2N3904s in the shape that this will track somewhat better than, say, a 2N3904 & a 1N4148.