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Current Regulators for I²L Circuits to be Operated from Low-Voltage Power Supplies

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Abstract—A new bandgap current reference is described which can be used to control the injector current of I²L circuits for supply voltages down to about 1 V. For small currents the total injector current is obtained as a mirror of the reference current. For large injector currents the current control is performed by a series regulator which compares the injector current of one I²L gate to the reference current.

The described reference current can be adjusted to give a variation with temperature of about 60 ppm/°C over the temperature range -10 to +70°C. However, in some applications a nonzero, but well controlled temperature coefficient is desired. It is shown how a temperature stable ring oscillator with I²L gates can be constructed by tailoring the temperature dependence of the supply current appropriately.

INTRODUCTION

ONE OF the potential advantages of I²L circuits is that they can, in principle, be operated from supply voltages less than 1 V. This permits, e.g., the use of a single NiCad cell as the power supply—an attractive feature in such applications as camera controls, watches, and calculators. However, some means for controlling the injector current are required. Normally a series drop resistor R_c is connected from the supply voltage V_{cc} to the injector in which case the total injector current is

$$I_{inj} = \frac{V_{cc} - V_{BE}}{R_c} \quad (1)$$

When V_{cc} approaches V_{BE} this solution becomes useless due to the temperature dependence of the base-emitter voltage V_{BE} . Also variations in V_{cc} give rise to large relative variations in I_{inj} when V_{cc} is only slightly larger than V_{BE} .

These problems have led to the investigation of principles for implementation of a current control circuit which can be used with input voltages down to about V_{BE} .

In the present paper we first discuss the circuit design of a constant reference current source which can operate down to less than 1 V. The current source is implemented using a bandgap reference principle, aiming at a temperature coefficient of zero. As the current is sourced to the load through a lateral p-n-p transistor it is only possible to use it directly as the injector current supply for supply currents below approxi-

mately 1 mA. However, for larger values of injector current the described current source can be used as a reference current in a feedback arrangement which employs an external p-n-p transistor as the series pass element.

While the primary aim in the design of the reference current source has been the establishment of a temperature stable current source operable down to about 1 V the demand in some I²L circuits may be a current source that exhibits a nonzero, but well controlled variation with temperature in order to get a constant gate delay. We, therefore, show how the temperature coefficient of the reference current can be adjusted such that a very small variation with temperature of the oscillator frequency of a three-stage ring oscillator can be achieved.

CONSTANT CURRENT SOURCE

A conventional bandgap reference voltage is established by adding a base-emitter voltage of a transistor to a base-emitter voltage difference of two transistors with different emitter current densities. The base-emitter voltage has a negative temperature coefficient while the base-emitter voltage difference has a positive temperature coefficient. It can be shown that if the base-emitter voltage difference is adjusted such that the sum of the two voltages equals the silicon bandgap voltage a zero temperature coefficient of the sum results [1].

When implementing a constant current source the same principle can be applied by using the sum of a V_{BE} dependent current and a ΔV_{BE} dependent current as the reference current [2].

Fig. 1 shows the diagram of the current source. The reference current is obtained by adding the currents I_2 and I_R . The current I_R is equal to V_{BE1}/R_1 and, thus, has a negative temperature coefficient.

The current I_2 is equal to $(V_{BE1} - V_{BE2})/R_2$ and has a positive temperature coefficient. The transistors Q_3 and Q_4 provide the feedback amplifier which regulates the currents in Q_1 , Q_2 , and R_1 such that $I_{ref} = I_1 = I_2 + I_R$. The resistor R_s is included to ensure a correct start-up of the circuit when the supply voltage is applied, and the capacitor C is added for stability reasons. The reference current is obtained from an extra collector of the lateral p-n-p transistor Q_4 . The reference current is, thus, a source current which can be used directly to supply I²L gates provided that the total current consumption does not exceed the maximum current capability of the transistor Q_4 .

Alternatively a reference sink current can be obtained by adding the transistor Q_5 .

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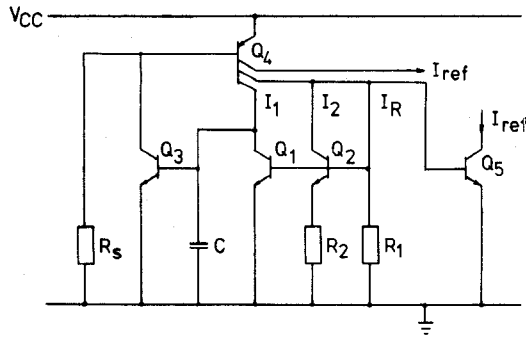


Fig. 1. Schematic of reference current generator.

The temperature dependence of the reference current can be analyzed by examining the currents I_2 and I_R . The base-emitter voltage dependence on temperature is given by [1]

$$V_{BE} = V_{g0} \left(1 - \frac{T}{T_0}\right) + \frac{T}{T_0} V_{BE0} + V_{t0} \frac{T}{T_0} \left((m-4) \ln \frac{T}{T_0} + \ln \frac{I_c}{I_0} \right). \quad (2)$$

In (2) V_{g0} is the silicon bandgap voltage at absolute zero ($V_{g0} = 1185$ mV), T_0 is the room temperature ($T_0 = 300$ K), V_{BE0} is the base-emitter voltage at room temperature and current I_0 , $V_{t0} = kT_0/q = 25.8$ mV, m is a constant related to the base minority carrier mobility ($m \approx 1$ for double diffused n-p-n transistors), and I_0 is the collector current at room temperature T_0 . Hence, we achieve

$$I_R = \frac{V_{BE1}}{R_1} = \frac{1}{R_1} \left(V_{g0} \left(1 - \frac{T}{T_0}\right) + \frac{T}{T_0} V_{BE0} + V_{t0} \frac{T}{T_0} \left((m-4) \ln \frac{T}{T_0} + \ln \frac{I_1}{I_0} \right) \right) \quad (3)$$

and

$$I_2 = \frac{V_{BE1} - V_{BE2}}{R_2} = \frac{V_{t0}}{R_2} \frac{T}{T_0} \left(\ln N + \ln \frac{I_1}{I_2} \right). \quad (4)$$

In (4) N equals the emitter area ratio of Q_2 and Q_1 :

$$N = \frac{A_2}{A_1}.$$

By differentiation we achieve

$$\frac{dI_2}{dT} \Big|_{T=T_0} = \frac{I_{20}}{T_0} \frac{\ln N + \ln z}{1 + \ln N + \ln z} \quad (6)$$

and

$$\frac{dI_R}{dT} \Big|_{T=T_0} = - \frac{V_{g0} - V_{BE0} - V_{t0}(m-4)}{R_1 T_0}. \quad (7)$$

In (6) and (7) we have introduced

$$z = \frac{I_1}{I_2} \Big|_{T=T_0} = \frac{I_{10}}{I_{20}}. \quad (8)$$

Consequently, we obtain a temperature coefficient of zero if

$$\frac{1}{z-1} \frac{\ln N + \ln z}{1 + \ln N + \ln z} = \frac{V_{g0} - V_{BE0} - V_{t0}(m-4)}{V_{BE0}}. \quad (9)$$

This equation determines the relation between N and z . The values of R_1 and R_2 are determined by

$$R_1 (I_{10} - I_{20}) = V_{BE0} \Rightarrow R_1 = \frac{V_{BE0}}{I_{ref}} \frac{z}{z-1} \quad (10)$$

and

$$R_2 I_2 + V_{BE2} = V_{BE1} \Rightarrow R_2 = \frac{z V_{t0}}{I_{ref}} (\ln z + \ln N). \quad (11)$$

Furthermore, it can be shown that the second derivative of the reference current with respect to temperature can be approximated by

$$\frac{T_0^2}{I_{ref}} \frac{d^2 I_{ref}}{dT^2} \Big|_{T=T_0} \approx \frac{V_{t0}(m-4)}{V_{g0} - V_{t0}(m-4)} - \frac{2 + \ln N + \ln z}{(1 + \ln N + \ln z)^2 (z(1 + \ln z + \ln N) - 1)}. \quad (12)$$

When the area ratio N approaches infinity this expression reduces to the results found for a conventional bandgap voltage reference [3].

$$\frac{T_0^2}{V_{ref}} \frac{d^2 V_{ref}}{dT^2} \Big|_{T=T_0} = \frac{V_{t0}(m-4)}{V_{g0} - V_{t0}(m-4)}. \quad (13)$$

For a realistic value of N the bandgap reference described above gives a second-order temperature dependence which is slightly larger than the conventional bandgap reference. As an example we find the second derivative for an area ratio of 18 to be $-9 \cdot 10^{-2}$ compared to the value $-6 \cdot 10^{-2}$ found for the conventional bandgap reference.

In the above calculations we have assumed that the resistors R_1 and R_2 are temperature independent. If these resistors have a nonzero temperature coefficient

$$\alpha = \frac{1}{R} \frac{dR}{dT} \Big|_{T=T_0}$$

(5) it can be shown that the design equation (9) should be modified to

$$\frac{1}{z-1} \frac{\ln N + \ln z}{1 + \ln N + \ln z} = \frac{1}{1 - \alpha T_0} \left(\frac{V_{g0} - V_{BE0} - V_{t0}(m-4)}{V_{BE0}} + \alpha T_0 \right). \quad (14)$$

The circuit in Fig. 1 has been implemented using Exar I²L design kit parts XR-C407 and XR-C408.

Aiming at a reference current of 100 μ A we find $R_1 = 13.130$ k Ω and $R_2 = 1.620$ k Ω .

Fig. 2 shows measured results of I_{ref} versus the supply voltage V_{cc} at different values of the ambient temperature T_a . The measurements were taken with the resistors R_1 and R_2

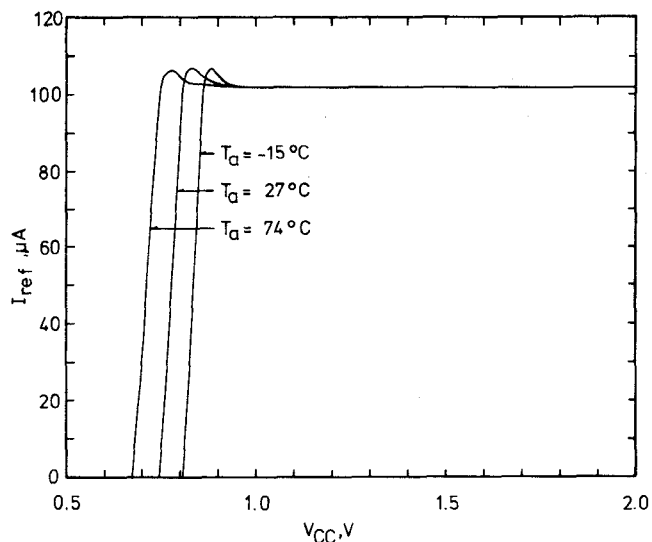


Fig. 2. Reference current I_{ref} versus supply voltage V_{cc} at different ambient temperatures T_a .

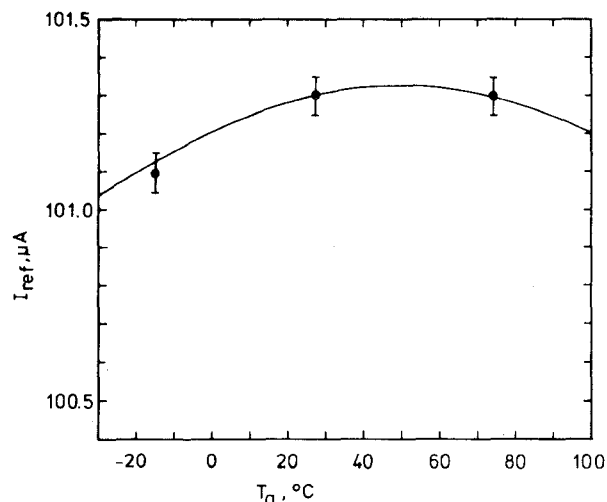


Fig. 3. Calculated and measured temperature dependence of the reference current. For the measured results the measuring uncertainty is indicated in the figure. The measurements were taken with a supply voltage $V_{cc} = 1$ V.

held at a constant temperature in order to avoid variations in I_{ref} due to a nonzero temperature coefficient of the resistors. It is evident from Fig. 2 that the circuit is operable below 1 V over the full temperature range from -10 to $+70^\circ\text{C}$. It is also obvious that the reference current exhibits only a very small variation with the supply voltage. The temperature dependence of the reference current is shown in more detail in Fig. 3 where we have shown the calculated performance (a second derivative of $-9 \cdot 10^{-2}$) together with the measured results. The measured results are found to be in excellent agreement with the theory.

FEEDBACK CIRCUIT FOR HIGH INJECTOR CURRENTS

The current source in Fig. 1 requires the use of a lateral p-n-p transistor as the series pass element. Therefore, the circuit can only be used directly for injector currents below approximately 1 mA. For high current levels an external p-n-p

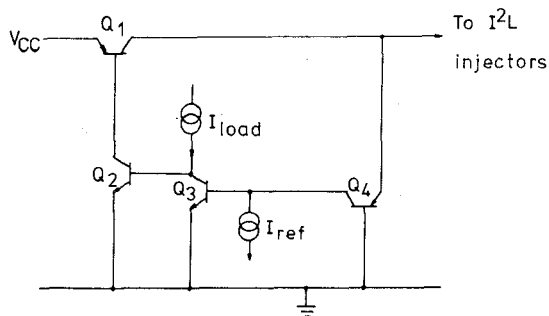


Fig. 4. Series current regulator for high levels of injector current.

series transistor must be used, and the current mirror technique utilized in the circuit in Fig. 1 becomes useless. Instead a feedback arrangement as shown in Fig. 4 may be applied. In this circuit the current reference generated by the circuit in Fig. 1 is compared to the injector current of just one I^2L gate. Transistor Q_4 is, in fact, an I^2L gate used as a p-n-p transistor. The collector of Q_4 is the normal base input of an I^2L gate. Hence, the collector current of Q_4 equals $\alpha_{inj} \cdot I_{inj}$ where α_{inj} is the common base current gain of Q_4 . Assuming a negligible base current of Q_3 we achieve

$$\alpha_{inj} \cdot I_{inj} = I_{ref}. \quad (15)$$

In (15) I_{inj} is the injector current per gate. The base current of Q_3 is negligible compared to I_{ref} if the load current of Q_3 is of the same order of magnitude as I_{ref} . This can be obtained by taking I_{load} from an extra collector of the p-n-p transistor in the current reference circuit in Fig. 1.

TEMPERATURE STABLE RING OSCILLATOR

A convenient method for implementation of an on-chip oscillator in I^2L technique is the use of a chain of inverters as a ring oscillator. As the gate delay of the inverters depends on the injector current the oscillation frequency can be controlled by controlling the injector current. The oscillation frequency is also temperature dependent with the dependency being mainly due to the temperature dependency of the voltage swing.

To a first-order approximation the temperature dependence can be assumed to be linear, and the relative shift in frequency versus temperature can be assumed to be $2 \text{ mV}/V_{BE}$ or about 0.3 percent/ $^\circ\text{C}$.

The oscillation frequency versus temperature is shown in Fig. 5 for a three-stage ring oscillator supplied by a constant current. The oscillator has been implemented using the XR-C401 test circuit.

The temperature dependence can be compensated for by using a V_{BE} -dependent injector current or rather $\alpha_{inj} \cdot I_{inj}$ should vary in proportion to V_{BE} . Such a variation of $\alpha_{inj} \cdot I_{inj}$ can be obtained by controlling the injector current by the circuit shown in Fig. 4 and using a V_{BE} -dependent reference current source.

A V_{BE} -dependent reference current source operable from a supply voltage of less than 1 V can easily be obtained by omitting the transistor Q_2 and the resistor R_2 from the reference current generator shown in Fig. 1. The oscillator performance achievable using this approach is also demonstrated in Fig. 5

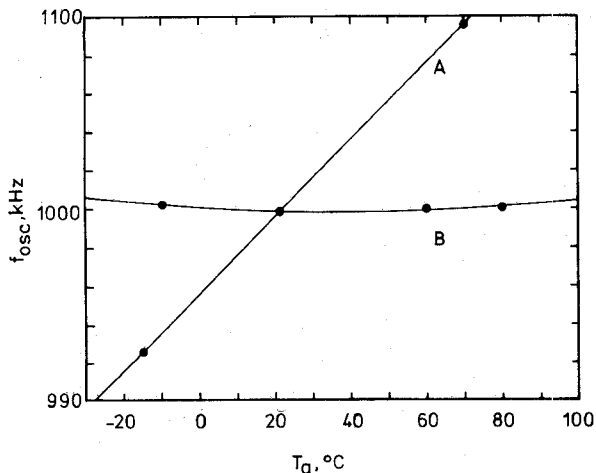


Fig. 5. Oscillation frequency versus temperature for a three-stage ring oscillator supplied from: (a) a constant current source, (b) a series regulator as shown in Fig. 4 and with a V_{BE} -dependent reference current.

which shows the oscillation frequency versus temperature for a ring oscillator the injector current of which is controlled by the circuit in Fig. 4 using a V_{BE} -dependent reference current.

CONCLUSION

We have demonstrated a circuit configuration which can be used as a constant current reference. The circuit can be operated from supply voltages below 1 V and exhibits a temperature stability comparable to the stability achievable from a conventional bandgap voltage reference. The current reference requires two precision resistors for setting of the current level and an external capacitor for stability reasons.

The described current source can directly supply I²L circuits with an injector current up to current levels of about 1 mA. For larger current levels a feedback circuit employing an external p-n-p transistor is proposed. This circuit regulates the current $\alpha_{inj} \cdot I_{inj}$ where I_{inj} is the injector current per gate and α_{inj} is the common base gain of the injector p-n-p-transistor.

We have finally found that this feedback control of $\alpha_{inj} \cdot I_{inj}$ can be used to obtain a temperature independent oscillation frequency from a ring oscillator if $\alpha_{inj} \cdot I_{inj}$ is made to vary in proportion to V_{BE} . This variation of $\alpha_{inj} \cdot I_{inj}$ is obtained from a simplified version of the temperature stable current ref-

erence circuit. The oscillator thus implemented shows a variation of frequency of less than 70 ppm/°C over the temperature range from -10 to +70°C.

REFERENCES

- [1] R. J. Widlar, "New developments in IC voltage regulators," *IEEE J. Solid-State Circuits*, vol. SC-6, pp. 2-7, Feb. 1971.
- [2] R. J. van de Plassche, "Dynamic element matching for high-accuracy monolithic D/A converters," *IEEE J. Solid-State Circuits*, vol. SC-11, pp. 795-800, Dec. 1976.
- [3] A. P. Brokaw, "A simple three-terminal IC bandgap reference," *IEEE J. Solid-State Circuits*, vol. SC-9, pp. 388-393, Dec. 1974.



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