

## Voltage-to-Frequency and Frequency-to-Voltage Converter

### FEATURES

- V/F Conversion to 1 MHz**
- Reliable Monolithic Construction**
- Very Low Nonlinearity**
  - 0.002% typ at 10 kHz
  - 0.005% typ at 100 kHz
  - 0.07% typ at 1 MHz
- Input Offset Trimmable to Zero**
- CMOS or TTL Compatible**
- Unipolar, Bipolar, or Differential V/F**
- V/F or F/V Conversion**
- Available in Surface Mount**
- MIL-STD-883 Compliant Versions Available**

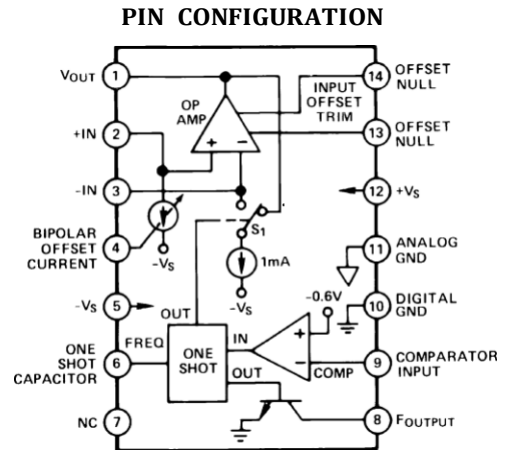
### PRODUCT DESCRIPTION

The HT8650 V/F/V (voltage-to-frequency or frequency-to-voltage converter) provides a combination of high frequency operation and low nonlinearity previously unavailable in monolithic form. The inherent monotonicity of the V/F transfer function makes the HT8650 useful as a high-resolution analog-to-digital converter. A flexible input configuration allows a wide variety of input voltage and current formats to be used, and an open-collector output with separate digital ground allows simple interfacing to either standard logic families or opto-couplers.

The linearity error of the HT8650 is typically 20 ppm (0.002% of full scale) and 50 ppm (0.005%) maximum at 10 kHz full scale. This corresponds to approximately 14-bit linearity in an analog-to-digital converter circuit. Higher full-scale frequencies or longer count intervals can be used for higher resolution conversions. The HT8650 has a useful dynamic range of six decades allowing extremely high resolution measurements. Even at 1 MHz full scale, linearity is guaranteed less than 1000 ppm (0.1%) on the HT8650KN, BD, and SD grades.

In addition to analog-to-digital conversion, the HT8650 can be used in isolated analog signal transmission applications, phased locked-loop circuits, and precision stepper motor speed controllers. In the F/V mode, the HT8650 can be used in precision tachometer and FM demodulator circuits.

The input signal range and full-scale output frequency are user-programmable with two external capacitors and one resistor. Input offset voltage can be trimmed to zero with an external potentiometer.



The HT8650JN and HT8650KN are offered in a plastic 14-lead DIP package. The HT8650JP is available in a 20-lead plastic leaded chip carrier (PLCC). Both plastic packaged versions of the HT8650 are specified for the commercial (0°C to +70°C) temperature range. For industrial temperature range (-25°C to +85°C) applications, the HT8650AD and HT8650BD are offered in a ceramic package. The HT8650SD is specified for the full -55°C to +125°C extended temperature range.

### PRODUCT HIGHLIGHTS

1. In addition to very high linearity, the HT8650 can operate at full-scale output frequency up to 1 MHz. The combination of these two features makes the HT8650 an inexpensive solution for applications requiring high resolution monotonic A/D conversion.
2. The HT8650 has a very versatile architecture that can be configured to accommodate bipolar, unipolar, or differential input voltages, or unipolar input currents.
3. TTL or CMOS compatibility is achieved using an open collector frequency output. The pull-up resistor can be connected to voltages up to +30 V, or +15 V or +5 V for conventional CMOS or TTL logic levels.
4. The same components used for V/F conversion can also be used for F/V conversion by adding a simple logic biasing network and reconfiguring the HT8650.
5. The HT8650 provides separate analog and digital grounds. This feature allows prevention of ground loops in real-world applications.
6. The HT8650 is available in versions compliant with MIL-STD-883. Refer to the Analog Devices Military Products Databook or current HT8650/883B data sheet for detailed specifications.

Model	HT8650J/HT8650A			HT8650K/HT8650B			HT8650S			Units
	Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
<b>DYNAMIC PERFORMANCE</b>										
Full-Scale Frequency Range			1			1			1	MHz
Nonlinearity <sup>1</sup> $f_{MAX} = 10$ kHz		0.002	0.005	0.002	0.005	0.002	0.005	0.005	0.005	%
$f_{MAX} = 100$ kHz		0.005	<b>0.02</b>	0.005	<b>0.02</b>	0.005	0.005	<b>0.02</b>	<b>0.02</b>	%
$f_{MAX} = 500$ kHz		0.02	0.05	0.02	0.05	0.02	0.05	0.02	0.05	%
$f_{MAX} = 1$ MHz		0.1		0.05	<b>0.1</b>		0.05	<b>0.1</b>	<b>0.1</b>	%
Full-Scale Calibration Error <sup>2</sup> , 100 kHz		$\pm 5$		$\pm 5$		$\pm 5$	$\pm 5$			%
1 MHz		$\pm 10$		$\pm 10$		$\pm 10$	$\pm 10$			%
vs. Supply <sup>3</sup>	-0.015		<b>+0.015</b>	-0.015		<b>+0.015</b>	-0.015		<b>+0.015</b>	% of FSR/V
vs. Temperature										
A, B, and S Grades										
at 10 kHz			$\pm 75$			$\pm 75$			$\pm 75$	ppm/°C
at 100 kHz			$\pm 150$			$\pm 150$			$\pm 150$	ppm/°C
J and K Grades										
at 10 kHz		$\pm 75$		$\pm 75$		$\pm 75$			$\pm 75$	ppm/°C
at 100 kHz		$\pm 150$		$\pm 150$		$\pm 150$			$\pm 150$	ppm/°C
<b>BIPOLAR OFFSET CURRENT</b>										
Activated by 1.24 k $\Omega$ Between Pins 4 and 5	0.45	0.5	0.55	0.45	0.5	0.55	0.45	0.5	0.55	mA
<b>DYNAMIC RESPONSE</b>										
Maximum Settling Time for Full Scale Step Input	1 Pulse of New Frequency Plus 1 $\mu$ s			1 Pulse of New Frequency Plus 1 $\mu$ s			1 Pulse of New Frequency Plus 1 $\mu$ s			
Overload Recovery Time Step Input	1 Pulse of New Frequency Plus 1 $\mu$ s			1 Pulse of New Frequency Plus 1 $\mu$ s			1 Pulse of New Frequency Plus 1 $\mu$ s			
<b>ANALOG INPUT AMPLIFIER (V/F Conversion)</b>										
Current Input Range (Figure 1)	0		+0.6	0		+0.6	0		+0.6	mA
Voltage Input Range (Figure 5)	-10		0	-10		0	-10		0	V
Differential Impedance	2 M $\Omega$ 10 pF			2 M $\Omega$ 10 pF			2 M $\Omega$ 10 pF			
Common-Mode Impedance	1000 M $\Omega$ 10 pF			1000 M $\Omega$ 10 pF			1000 M $\Omega$ 10 pF			
Input Bias Current										
Noninverting Input		40	<b>100</b>	40	<b>100</b>	40	<b>100</b>	<b>100</b>	<b>100</b>	nA
Inverting Input		$\pm 8$	<b><math>\pm 20</math></b>	$\pm 8$	<b><math>\pm 20</math></b>	$\pm 8$	<b><math>\pm 20</math></b>	<b><math>\pm 20</math></b>	<b><math>\pm 20</math></b>	nA
Input Offset Voltage (Trimable to Zero)			<b><math>\pm 4</math></b>			<b><math>\pm 4</math></b>			<b><math>\pm 4</math></b>	mV
vs. Temperature ( $T_{MIN}$ to $T_{MAX}$ )		$\pm 30$		$\pm 30$		$\pm 30$	$\pm 30$		$\pm 30$	$\mu$ V/°C
Safe Input Voltage		$\pm V_S$		$\pm V_S$		$\pm V_S$	$\pm V_S$		$\pm V_S$	C
<b>COMPARATOR (F/V Conversion)</b>										
Logic "0" Level	$-V_S$		-1	$-V_S$		-1	$-V_S$		+1	V
Logic "1" Level	0		$+V_S$	0		$+V_S$	0		$+V_S$	V
Pulse Width Range <sup>4</sup>	0.1		( $0.3 \times t_{OS}$ )	0.1		( $0.3 \times t_{OS}$ )	0.1		( $0.3 \times t_{OS}$ )	$\mu$ s
Input Impedance		250		250		250	250		250	k $\Omega$
<b>OPEN COLLECTOR OUTPUT (V/F Conversion)</b>										
Output Voltage in Logic "0"										
$I_{SINK} \leq 8$ mA, $T_{MIN}$ to $T_{MAX}$			0.4			0.4			0.4	V
Output Leakage Current in Logic "1"			<b>100</b>			<b>100</b>			<b>100</b>	nA
Voltage Range <sup>5</sup>	0		+36	0		+36	0		+36	V
<b>AMPLIFIER OUTPUT (F/V Conversion)</b>										
Voltage Range (1500 $\Omega$ min Load Resistance)	0		+10	0		+10	0		+10	V
Source Current (750 $\Omega$ max Load Resistance)	10			10			10			mA
Capacitive Load (Without Oscillation)			100			100			100	pF
<b>POWER SUPPLY</b>										
Voltage, Rated Performance	$\pm 9$		<b><math>\pm 18</math></b>	$\pm 9$		<b><math>\pm 18</math></b>	$\pm 9$		<b><math>\pm 18</math></b>	V
Quiescent Current			<b>8</b>			<b>8</b>			<b>8</b>	mA
<b>TEMPERATURE RANGE</b>										
Rated Performance - N Package	0		+70	0		+70				°C
D Package	-25		+85	-25		+85	-55		+125	°C

**NOTES**
<sup>1</sup>Nonlinearity is defined as deviation from a straight line from zero to full scale, expressed as a fraction of full scale.

<sup>2</sup>Full-scale calibration error adjustable to zero.

<sup>3</sup>Measured at full-scale output frequency of 100 kHz.

<sup>4</sup>Refer to F/V conversion section of the text.

<sup>5</sup>Referred to digital ground.

Specifications subject to change without notice.

 Specifications shown in **boldface** are tested on all production units at final electrical test. Results from those test are used to calculate outgoing quality levels. All min and max specifications are guaranteed, although only those shown in **boldface** are tested on all production units.

**ABSOLUTE MAXIMUM RATINGS**

Total Supply Voltage ..... 36 V  
 Storage Temperature ..... -55°C to +150°C  
 Differential Input Voltage ..... ±10 V  
 Maximum Input Voltage ..... ±V<sub>S</sub>  
 Open Collector Output Voltage Above Digital GND ..... 36 V  
 Current ..... 50 mA  
 Amplifier Short Circuit to Ground ..... Indefinite  
 Comparator Input Voltage ..... ±V<sub>S</sub>

**PIN CONFIGURATION**

PIN NO.	D-14	N-14	P-20A
1	V <sub>OUT</sub>	V <sub>OUT</sub>	NC
2	+IN	+IN	V <sub>OUT</sub>
3	-IN	-IN	+IN
4	BIPOLAR OFFSET CURRENT	BIPOLAR OFFSET CURRENT	-IN
5	-V <sub>S</sub>	-V <sub>S</sub>	NC
6	ONE SHOT CAPACITOR	ONE SHOT CAPACITOR	BIPOLAR OFFSET CURRENT
7	NC	NC	NC
8	F <sub>OUTPUT</sub>	F <sub>OUTPUT</sub>	-V <sub>S</sub>
9	COMPARATOR INPUT	COMPARATOR INPUT	ONE SHOT CAPACITOR
10	DIGITAL GND	DIGITAL GND	NC
11	ANALOG GND	ANALOG GND	NC
12	+V <sub>S</sub>	+V <sub>S</sub>	F <sub>OUTPUT</sub>
13	OFFSET NULL	OFFSET NULL	COMPARATOR INPUT
14	OFFSET NULL	OFFSET NULL	DIGITAL GND
15			NC
16			ANALOG GND
17			NC
18			+V <sub>S</sub>
19			OFFSET NULL
20			OFFSET NULL

Model	Gain Tempco ppm/°C 100 kHz	1 MHz Linearity	Specified Temperature Range °C	Package Description	Package Option
HT8650AD	150 max	0.1% typ	-25 to +85	Ceramic DIP	D-14

## CIRCUIT OPERATION

### UNIPOLAR CONFIGURATION

The HT8650 is a *charge balance* voltage-to-frequency converter. In the connection diagram shown in Figure 1, or the block diagram of Figure 2a, the input signal is converted into an equivalent current by the input resistance  $R_{IN}$ . This current is *exactly* balanced by an internal feedback current delivered in short, timed bursts from the switched 1 mA internal current source. These bursts of current may be thought of as precisely defined packets of charge. The required number of charge packets, each producing one pulse of the output transistor, depends upon the amplitude of the input signal. Since the number of charge packets delivered per unit time is dependent on the input signal amplitude, a linear voltage-to-frequency transformation will be accomplished. The frequency output is furnished via an open collector transistor.

A more rigorous analysis demonstrates how the charge balance voltage-to-frequency conversion takes place.

A block diagram of the device arranged as a V-to-F converter is shown in Figure 2a. The unit is comprised of an input integrator, a current source and steering switch, a comparator and a one-shot. When the output of the one-shot is low, the current steering switch  $S_1$  diverts all the current to the output of the op amp; this is called the Integration Period. When the one-shot has been triggered and its output is high, the switch  $S_1$  diverts all the current to the summing junction of the op amp; this is called the Reset Period. The two different states are shown in Figure 2 along with the various branch currents. It should be noted that the output current from the op amp is the same for either state, thus minimizing transients.

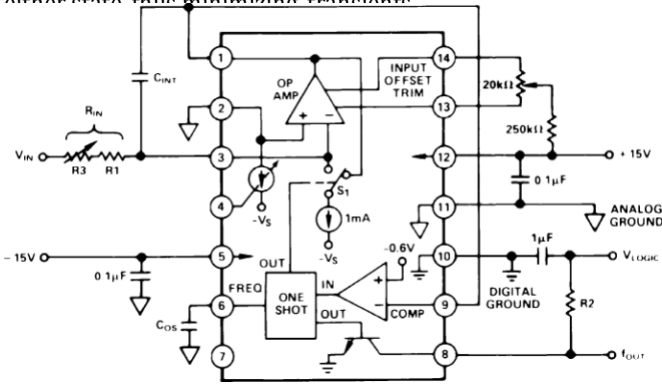


Figure 1. Connection Diagram for V/F Conversion, Positive Input Voltage

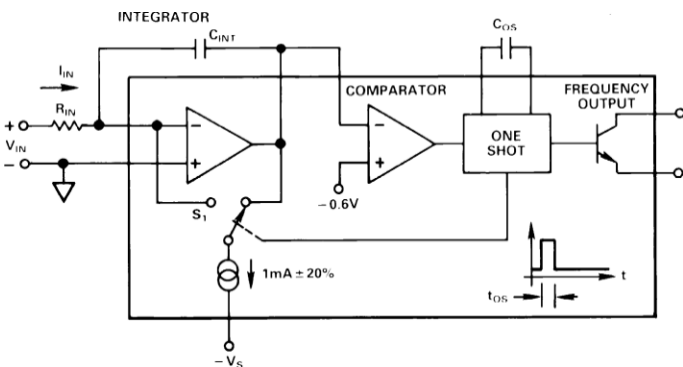


Figure 2a. Block Diagram

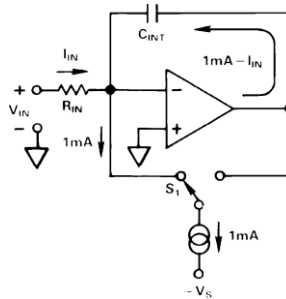


Figure 2b. Reset Mode

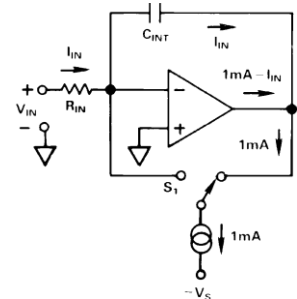


Figure 2c. Integrate Mode

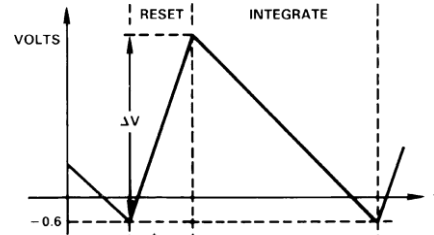


Figure 2d. Voltage Across  $C_{INT}$

The positive input voltage develops a current ( $I_{IN} = V_{IN}/R_{IN}$ ) which charges the integrator capacitor  $C_{INT}$ . As charge builds up on  $C_{INT}$ , the output voltage of the integrator ramps downward towards ground. When the integrator output voltage (Pin 1) crosses the comparator threshold (-0.6 volt) the comparator triggers the one shot, whose time period,  $t_{OS}$  is determined by the one shot capacitor  $C_{OS}$ .

Specifically, the one shot time period is:

$$t_{OS} = C_{OS} \times 6.8 \times 10^3 \text{ sec} / F + 3.0 \times 10^{-7} \text{ sec} \quad (1)$$

The Reset Period is initiated as soon as the integrator output voltage crosses the comparator threshold, and the integrator ramps upward by an amount:

$$\Delta V = t_{OS} \cdot \frac{dV}{dt} = \frac{t_{OS}}{C_{INT}} (1\text{mA} - I_{IN}) \quad (2)$$

After the Reset Period has ended, the device starts another Integration Period, as shown in Figure 2, and starts ramping downward again. The amount of time required to reach the comparator threshold is given as:

$$T_I = \frac{\Delta V}{\frac{dV}{dt}} = \frac{t_{OS}/C_{INT} (1\text{mA} - I_{IN})}{I_{IN}/C_{INT}} = t_{OS} \left( \frac{1\text{mA}}{I_{IN}} - 1 \right) \quad (3)$$

The output frequency is now given as:

$$f_{OUT} = \frac{1}{t_{OS} + T_I} = \frac{I_{IN}}{t_{OS} \times 1\text{mA}} = 0.15 \frac{F \cdot \text{Hz}}{A} \frac{V_{IN}/R_{IN}}{C_{OS} + 4.4 \times 10^{-11} F} \quad (4)$$

Note that  $C_{INT}$ , the integration capacitor has no effect on the transfer relation, but merely determines the amplitude of the sawtooth signal out of the integrator.

### One Shot Timing

A key part of the preceding analysis is the one shot time period that was given in equation (1). This time period can be broken down into approximately 300 ns of propagation delay, and a second time segment dependent linearly on timing capacitor  $C_{OS}$ . When the one shot is triggered, a voltage switch that holds Pin 6

at analog ground is opened allowing that voltage to change. An internal 0.5 mA current source connected to Pin 6 then draws its current out of  $C_{OS}$ , causing the voltage at Pin 6 to decrease linearly. At approximately  $-3.4$  V, the one shot resets itself, thereby ending the timed period and starting the V/F conversion cycle over again. The total one shot time period can be written mathematically as:

$$t_{OS} = \frac{\Delta V C_{OS}}{I_{DISCHARGE}} + T_{GATE DELAY} \quad (5)$$

substituting actual values quoted above,

$$t_{OS} = \frac{-3.4 V \times C_{OS}}{-0.5 \times 10^{-3} A} + 300 \times 10^{-9} \text{ sec} \quad (6)$$

This simplifies into the timed period equation given above.

### COMPONENT SELECTION

Only four component values must be selected by the user. These are input resistance  $R_{IN}$ , timing capacitor  $C_{OS}$ , logic resistor  $R_2$ , and integration capacitor  $C_{INT}$ . The first two determine the input voltage and full-scale frequency, while the last two are determined by other circuit considerations.

Of the four components to be selected,  $R_2$  is the easiest to define. As a pull-up resistor, it should be chosen to limit the current through the output transistor to 8 mA if a TTL maximum  $V_{OL}$  of 0.4 V is desired. For example, if a 5 V logic supply is used,  $R_2$  should be no smaller than 5 V/8 mA or 625  $\Omega$ . A larger value can be used if desired.

$R_{IN}$  and  $C_{OS}$  are the only two parameters available to set the full-scale frequency to accommodate the given signal range. The "swing" variable that is affected by the choice of  $R_{IN}$  and  $C_{OS}$  is nonlinearity. The selection guide of Figure 3 shows this quite graphically. In general, larger values of  $C_{OS}$  and lower full-scale input currents (higher values of  $R_{IN}$ ) provide better linearity. In Figure 3, the implications of four different choices of  $R_{IN}$  are shown. Although the selection guide is set up for a unipolar configuration with a zero to 10 V input signal range, the results can be extended to other configurations and input signal ranges. For a full scale frequency of 100 kHz (corresponding to 10 V input), you can see that among the available choices,  $R_{IN} = 20$  k and  $C_{OS} = 620$  pF gives the lowest nonlinearity, 0.0038%. Also, if you wish to use the highest frequency that will give the 20 ppm minimum nonlinearity, it is approximately 33 kHz (40.2 k $\Omega$  and 1000 pF).

For input signal spans other than 10 V, the input resistance must be scaled proportionately. For example, if 100 k $\Omega$  is called out for a 0 V–10 V span, 10k would be used with a 0 V–1 V span, or 200 k $\Omega$  with a  $\pm 10$  V bipolar connection.

The last component to be selected is the integration capacitor  $C_{INT}$ . In almost all cases, the best value for  $C_{INT}$  can be calculated using the equation:

$$C_{INT} = \frac{10^{-4} F / \text{sec}}{f_{MAX}} \quad (1000 \text{ pF minimum}) \quad (7)$$

When the proper value for  $C_{INT}$  is used, the charge balance architecture of the HT8650 provides continuous integration of the input signal, hence large amounts of noise and interference

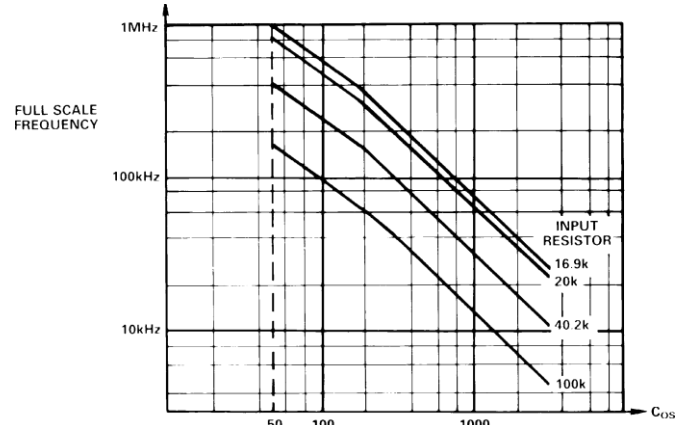


Figure 3a. Full-Scale Frequency vs.  $C_{OS}$

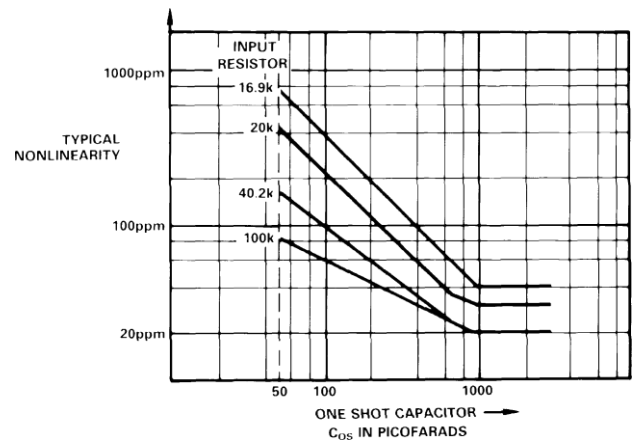


Figure 3b. Typical Nonlinearity vs.  $C_{OS}$

can be rejected. If the output frequency is measured by counting pulses during a constant gate period, the integration provides infinite normal-mode rejection for frequencies corresponding to the gate period and its harmonics. However, if the integrator stage becomes saturated by an excessively large noise pulse, the continuous integration of the signal will be interrupted, allowing the noise to appear at the output. If the approximate amount of noise that will appear on  $C_{INT}$  is known ( $V_{NOISE}$ ), the value of  $C_{INT}$  can be checked using the following inequality:

$$C_{INT} > \frac{t_{OS} \times 1 \times 10^{-3} A}{+V_S - 3V - V_{NOISE}} \quad (8)$$

For example, consider an application calling for a maximum frequency of 75 kHz, a 0 volt–1 volt signal range, and supply voltages of only  $\pm 9$  volts. The component selection guide of Figure 3 is used to select 2.0 k $\Omega$  for  $R_{IN}$  and 1000 pF for  $C_{OS}$ . This results in a one shot time period of approximately 7  $\mu$ s. Substituting 75 kHz into equation 7 yields a value of 1300 pF for  $C_{INT}$ . When the input signal is near zero, 1 mA flows through the integration capacitor to the switched current sink during the reset phase, causing the voltage across  $C_{INT}$  to increase by approximately 5.5 volts. Since the integrator output stage requires approximately 3 volts head room for proper operation, only 0.5 volt margin remains for integrating extraneous noise on the signal line. A negative noise pulse at this time might saturate the integrator, causing an error in signal integration. Increasing  $C_{INT}$  to 1500 pF or 2000 pF will provide much more noise margin, thereby eliminating this potential trouble spot.



### BIPOLAR V/F

Figure 4 shows how the internal bipolar current sink is used to provide a half-scale offset for a  $\pm 5$  V signal range, while providing a 100 kHz maximum output frequency. The nominally 0.5 mA ( $\pm 10\%$ ) offset current sink is enabled when a 1.24 k $\Omega$  resistor is connected between Pins 4 and 5. Thus, with the grounded 10 k $\Omega$  nominal resistance shown, a -5 V offset is developed at Pin 2. Since Pin 3 must also be at -5 V, the current through  $R_{IN}$  is  $10 \text{ V}/40 \text{ k}\Omega = +0.25 \text{ mA}$  at  $V_{IN} = +5 \text{ V}$ , and 0 mA at  $V_{IN} = -5 \text{ V}$ .

Components are selected using the same guidelines outlined for the unipolar configuration with one alteration. The voltage across the total signal range must be equated to the maximum

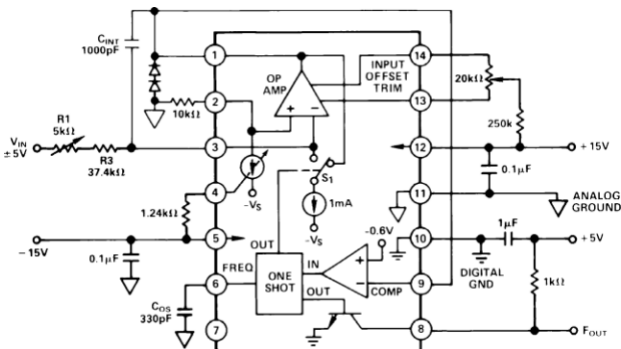


Figure 4. Connections for  $\pm 5$  V Bipolar V/F with 0 to 100 kHz TTL Output

input voltage in the unipolar configuration. In other words, the value of the input resistor  $R_{IN}$  is determined by the input voltage span, not the maximum input voltage. A diode from Pin 1 to ground is also recommended. This is further discussed in the Other Circuit Conditions section.

As in the unipolar circuit,  $R_{IN}$  and  $C_{OS}$  must have low temperature coefficients to minimize the overall gain drift. The 1.24 k $\Omega$  resistor used to activate the 0.5 mA offset current should also have a low temperature coefficient. The bipolar offset current has a temperature coefficient of approximately -200 ppm/ $^{\circ}\text{C}$ .

### UNIPOLAR V/F, NEGATIVE INPUT VOLTAGE

Figure 5 shows the connection diagram for V/F conversion of negative input voltages. In this configuration full-scale output frequency occurs at negative full-scale input, and zero output frequency corresponds with zero input voltage.

A very high impedance signal source may be used since it only drives the noninverting integrator input. Typical input impedance at this terminal is 1 G $\Omega$  or higher. For V/F conversion of positive input signals using the connection diagram of Figure 1, the signal generator must be able to source the integration current to drive the HT8650. For the negative V/F conversion circuit of Figure 5, the integration current is drawn from ground through  $R_1$  and  $R_3$ , and the active input is high impedance.

Circuit operation for negative input voltages is very similar to positive input unipolar conversion described in a previous section. For best operating results use component equations listed in that section.

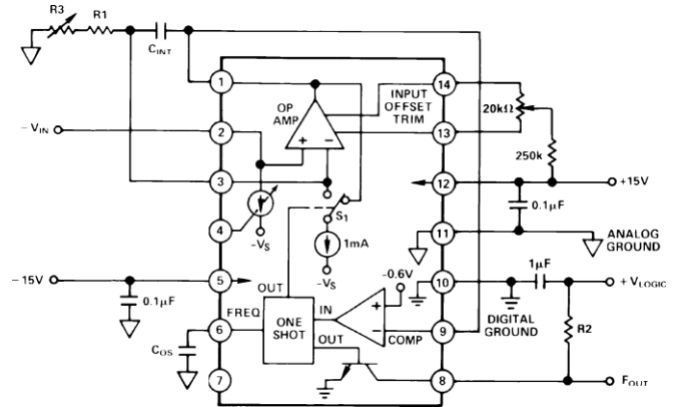


Figure 5. Connection Diagram for V/F Conversion, Negative Input Voltage

### F/V CONVERSION

The HT8650 also makes a very linear frequency-to-voltage converter. Figure 6 shows the connection diagram for F/V conversion with TTL input logic levels. Each time the input signal crosses the comparator threshold going negative, the one shot is activated and switches 1 mA into the integrator input for a measured time period (determined by  $C_{OS}$ ). As the frequency increases, the amount of charge injected into the integration capacitor increase proportionately. The voltage across the integration capacitor is stabilized when the leakage current through  $R_1$  and  $R_3$  equals the average current being switched into the integrator. The net result of these two effects is an average output voltage which is proportional to the input frequency. Optimum performance can be obtained by selecting components using the same guidelines and equations listed in the V/F Conversion section.

The reader is referred to Analog Devices' Application Note AN-279 where a more complete description of this application can be found.

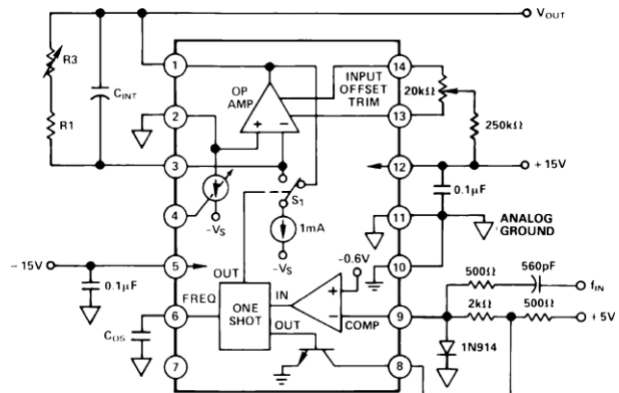


Figure 6. Connection Diagram for F/V Conversion

### HIGH FREQUENCY OPERATION

Proper RF techniques must be observed when operating the HT8650 at or near its maximum frequency of 1 MHz. Lead lengths must be kept as short as possible, especially on the one shot and integration capacitors, and at the integrator summing junction. In addition, at maximum output frequencies above 500 kHz, a 3.6 k $\Omega$  pull-down resistor from Pin 1 to  $-V_S$  is required (see Figure 7). The additional current drawn through the pull-down resistor reduces the op amp's output impedance and improves its transient response.

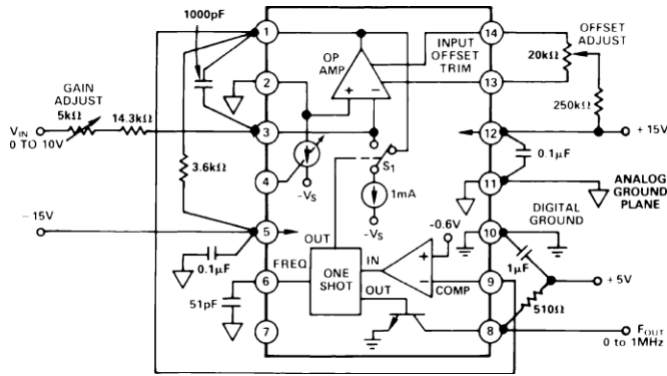


Figure 7. 1 MHz V/F Connection Diagram

### DECOUPLING AND GROUNDING

It is good engineering practice to use bypass capacitors on the supply-voltage pins and to insert small-valued resistors (10 Ω to 100 Ω) in the supply lines to provide a measure of decoupling between the various circuits in a system. Ceramic capacitors of 0.1 μF to 1.0 μF should be applied between the supply-voltage pins and analog signal ground for proper bypassing on the HT8650.

In addition, a larger board level decoupling capacitor of 1 μF to 10 μF should be located relatively close to the HT8650 on each power supply line. Such precautions are imperative in high resolution data acquisition applications where one expects to exploit the full linearity and dynamic range of the HT8650. Although some types of circuits may operate satisfactorily with power supply decoupling at only one location on each circuit board, such practice is strongly discouraged in high accuracy analog design.

Separate digital and analog grounds are provided on the HT8650. The emitter of the open collector frequency output transistor is the only node returned to the digital ground. All other signals are referred to analog ground. The purpose of the two separate grounds is to allow isolation between the high precision analog signals and the digital section of the circuitry. As much as several hundred millivolts of noise can be tolerated on the digital ground without affecting the accuracy of the VFC. Such ground noise is inevitable when switching the large currents associated with the frequency output signal.

At 1 MHz full scale, it is necessary to use a pull-up resistor of about 500 Ω in order to get the rise time fast enough to provide well defined output pulses. This means that from a 5 volt logic supply, for example, the open collector output will draw 10 mA.

This much current being switched will surely cause ringing on long ground runs due to the self inductance of the wires. For instance, #20 gauge wire has an inductance of about 20 nH per inch; a current of 10 mA being switched in 50 ns at the end of 12 inches of 20 gauge wire will produce a voltage spike of 50 mV. The separate digital ground of the HT8650 will easily handle these types of switching transients.

A problem will remain from interference caused by radiation of electro-magnetic energy from these fast transients. Typically, a voltage spike is produced by inductive switching transients; these spikes can capacitively couple into other sections of the circuit. Another problem is ringing of ground lines and power supply lines due to the distributed capacitance and inductance of the wires. Such ringing can also couple interference into sensitive analog circuits. The best solution to these problems is proper bypassing of the logic supply at the HT8650 package. A

1 μF to 10 μF tantalum capacitor should be connected directly to the supply side of the pull-up resistor and to the digital ground—Pin 10. The pull-up resistor should be connected directly to the frequency output—Pin 8. The lead lengths on the bypass capacitor and the pull up resistor should be as short as possible. The capacitor will supply (or absorb) the current transients, and large ac signals will flow in a physically small loop through the capacitor, pull up resistor, and frequency output transistor. It is important that the loop be physically small for two reasons: first, there is less self-inductance if the wires are short, and second, the loop will not radiate RFI efficiently.

The digital ground (Pin 10) should be separately connected to the power supply ground. Note that the leads to the digital power supply are only carrying dc current and cannot radiate RFI. There may also be a dc ground drop due to the difference in currents returned on the analog and digital grounds. This will not cause any problem. In fact, the HT8650 will tolerate as much as 0.25 volt dc potential difference between the analog and digital grounds. These features greatly ease power distribution and ground management in large systems. Proper technique for grounding requires separate digital and analog ground returns to the power supply. Also, the signal ground must be referred directly to analog ground (Pin 11) at the package. All of the signal grounds should be tied directly to Pin 11, especially the one-shot capacitor. More information on proper grounding and reduction of interference can be found in Reference 1.

### TEMPERATURE COEFFICIENTS

The drift specifications of the HT8650 do not include temperature effects of any of the supporting resistors or capacitors. The drift of the input resistors R1 and R3 and the timing capacitor C<sub>OS</sub> directly affect the overall temperature stability. In the application of Figure 2, a 10 ppm/°C input resistor used with a 100 ppm/°C capacitor may result in a maximum overall circuit gain drift of:

$$150 \text{ ppm}/^{\circ}\text{C} (\text{HT8650A}) + 100 \text{ ppm}/^{\circ}\text{C} (\text{C}_{\text{OS}}) + 10 \text{ ppm}/^{\circ}\text{C} (\text{R}_{\text{IN}}) 260 \text{ ppm}/^{\circ}\text{C}$$

In bipolar configuration, the drift of the 1.24 kΩ resistor used to activate the internal bipolar offset current source will directly affect the value of this current. This resistor should be matched to the resistor connected to the op amp noninverting input (Pin 2), see Figure 4. That is, the temperature coefficients of these two resistors should be equal. If this is the case, then the effects of the temperature coefficients of the resistors cancel each other, and the drift of the offset voltage developed at the op amp non-inverting input will be determined solely by the HT8650. Under these conditions the TC of the bipolar offset voltage is typically -200 ppm/°C and is a maximum of -300 ppm/°C. The offset voltage always decreases in magnitude as temperature is increased.

Other circuit components do not directly influence the accuracy of the VFC over temperature changes as long as their actual values are not so different from the nominal value as to preclude operation. This includes the integration capacitor, C<sub>INT</sub>. A change in the capacitance value of C<sub>INT</sub> simply results in a different rate of voltage change across the capacitor. During the Integration Phase (refer to Figure 2), the rate of voltage change across C<sub>INT</sub> has the opposite effect that it does during the Reset Phase. The result is that the conversion accuracy is unchanged

by either drift or tolerance of  $C_{INT}$ . The net effect of a change in the integrator capacitor is simply to change the peak to peak amplitude of the sawtooth waveform at the output of the integrator.

The gain temperature coefficient of the HT8650 is not a constant value. Rather the gain TC is a function of both the full-scale frequency and the ambient temperature. At a low full-scale frequency, the gain TC is determined primarily by the stability of the internal reference—a buried Zener reference. This low speed gain TC can be quite good; at 10 kHz full scale, the gain TC near 25°C is typically  $0 \pm 50$  ppm/°C. Although the gain TC changes with ambient temperature (tending to be more positive at higher temperatures), the drift remains within a  $\pm 75$  ppm/°C window over the entire military temperature range. At full-scale frequencies higher than 10 kHz dynamic errors become much more important than the static drift of the dc reference. At a full-scale frequency of 100 kHz and above, these timing errors dominate the gain TC. For example, at 100 kHz full-scale frequency ( $R_{IN} = 40$  k and  $C_{OS} = 330$  pF) the gain TC near room temperature is typically  $-80 \pm 50$  ppm/°C, but at an ambient temperature near +125°C, the gain TC tends to be more positive and is typically  $+15 \pm 50$  ppm/°C. This information is presented in a graphical form in Figure 8. The gain TC always tends to become more positive at higher temperatures. Therefore, it is possible to adjust the gain TC of the HT8650 by using a one-shot capacitor with an appropriate TC to cancel the drift of the circuit. For example, consider the 100 kHz full-scale frequency. An average drift of  $-100$  ppm/°C means that as temperature is increased, the circuit will produce a lower frequency in response to a given input voltage. This means that the one-shot capacitor must decrease in value as temperature increases in order to compensate the gain TC of the HT8650; that is, the capacitor must have a TC of  $-100$  ppm/°C. Now consider the 1 MHz full-scale frequency.

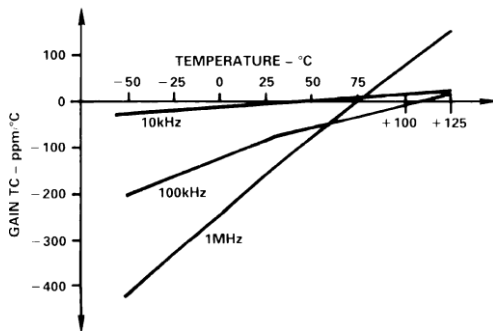


Figure 8. Gain TC vs. Temperature

It is not possible to achieve very much improvement in performance unless the expected ambient temperature range is known. For example, in a constant low temperature application such as gathering data in an Arctic climate (approximately  $-20^\circ\text{C}$ ), a  $C_{OS}$  with a drift of  $-310$  ppm/°C is called for in order to compensate the gain drift of the HT8650. However, if that circuit should see an ambient temperature of  $+75^\circ\text{C}$ , the  $C_{OS}$  cap would change the gain TC from approximately 0 ppm to  $+310$  ppm/°C.

The temperature effects of the components described above are the same when the HT8650 is configured for negative or bipolar input voltages, and for F/V conversion as well.

### NONLINEARITY SPECIFICATION

The linearity error of the HT8650 is specified by the endpoint method. That is, the error is expressed in terms of the deviation from the ideal voltage to frequency transfer relation after calibrating the converter at full scale and “zero”. The nonlinearity will vary with the choice of one-shot capacitor and input resistor (see Figure 3). Verification of the linearity specification requires the availability of a switchable voltage source (or a DAC) having a linearity error below 20 ppm, and the use of very long measurement intervals to minimize count uncertainties. Every HT8650 is automatically tested for linearity, and it will not usually be necessary to perform this verification, which is both tedious and time consuming. If it is required to perform a nonlinearity test either as part of an incoming quality screening or as a final product evaluation, an automated “bench-top” tester would prove useful. Such a system based on the Analog Devices’ LTS-2010 is described in Reference 2.

The voltage-to-frequency transfer relation is shown in Figure 9 with the nonlinearity exaggerated for clarity. The first step in determining nonlinearity is to connect the endpoints of the

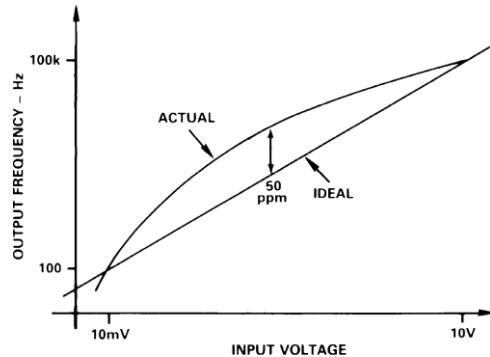


Figure 9a. Exaggerated Nonlinearity at 100 kHz Full Scale

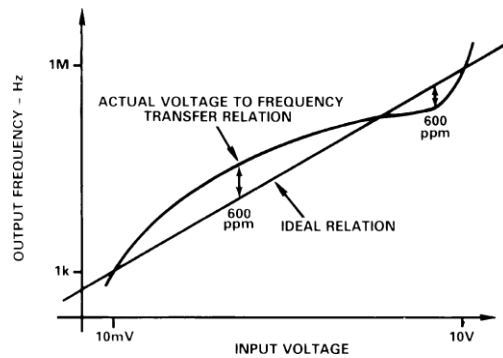


Figure 9b. Exaggerated Nonlinearity at 1 MHz Full Scale

operating range (typically at 10 mV and 10 V) with a straight line. This straight line is then the ideal relationship which is desired from the circuit. The second step is to find the difference between this line and the actual response of the circuit at a few points between the endpoints—typically ten intermediate points will suffice. The difference between the actual and the ideal response is a frequency error measured in hertz. Finally, these frequency errors are normalized to the full-scale frequency and expressed either as parts per million of full scale (ppm) or parts per hundred of full scale (%). For example, on a 100 kHz full



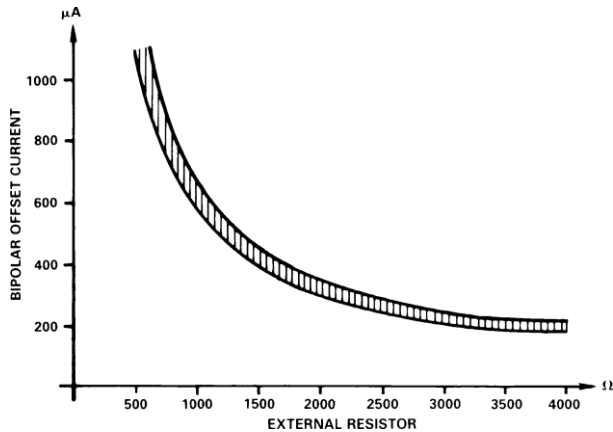


Figure 11. Bipolar Offset Current vs. External Resistor

## APPLICATIONS

### DIFFERENTIAL VOLTAGE-TO-FREQUENCY CONVERSION

The circuit of Figure 12 accepts a true floating differential input signal. The common-mode input,  $V_{CM}$ , may be in the range +15 to -5 volts with respect to analog ground. The signal input,  $V_{IN}$ , may be  $\pm 5$  volts with respect to the common-mode input. Both inputs are low impedance: the source which drives the common-mode input must supply the 0.5 mA drawn by the bipolar offset current source and the source which drives the signal input must supply the integration current.

If less common-mode voltage range is required, a lower voltage Zener may be used. For example, if a 5 volt Zener is used, the  $V_{CM}$  input may be in the range +10 to -5 volt. If the Zener is not used at all, the common-mode range will be  $\pm 5$  volts with respect to analog ground. If no Zener is used, the 10k pull-down resistor is not needed and the integrator output (Pin 1) is connected directly to the comparator input (Pin 9).

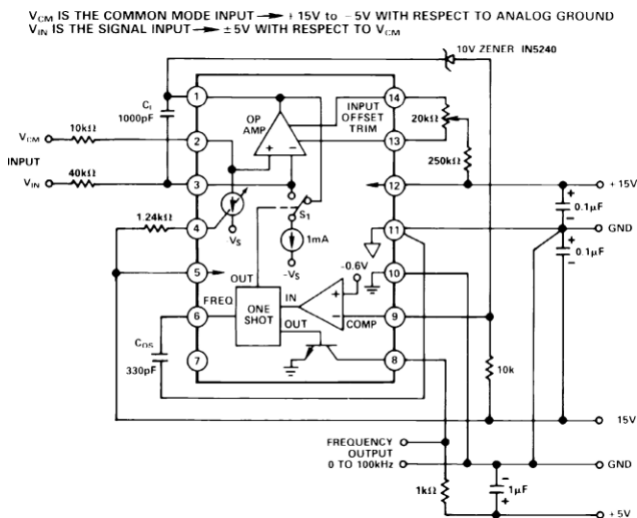


Figure 12. Differential Input

### AUTOZERO CIRCUIT

In order to exploit the full dynamic range of the HT8650 VFC, very small input voltages will need to be converted. For example, a six decade dynamic range based on a full scale of 10 volts will require accurate measurement of signals down to  $10\mu\text{V}$ . In these situations a well-controlled input offset voltage is imperative. A

constant offset voltage will not affect dynamic range but simply shift all of the frequency readings by a few hertz. However, if the offset should change, then it will not be possible to distinguish between a small change in a small input voltage and a drift of the offset voltage. Hence, the usable dynamic range is less. The circuit shown in Figure 13 provides automatic adjustment of the op amp offset voltage. The circuit uses an AD582 sample and hold amplifier to control the offset and the input voltage to the VFC is switched between ground and the signal to be measured via an AD7512DI analog switch. The offset of the HT8650 is adjusted by injecting a current into or drawing a current out of Pin 13. Note that only one of the offset null pins is used. During the "VFC Norm" mode, the SHA is in the hold mode and the hold capacitor is very large, 0.1  $\mu\text{F}$ , to hold the HT8650 offset constant for a long period of time.

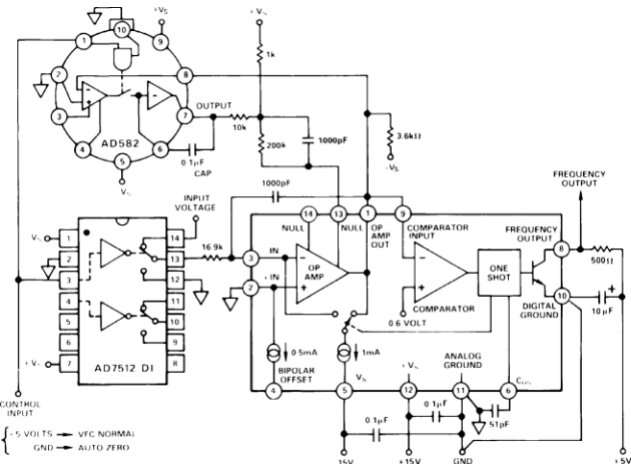


Figure 13. Autozero Circuit

When the circuit is in the "Autozero" mode the SHA is in sample mode and behaves like an op amp. The circuit is a variation of the classical two amplifier servo loop, where the output of the Device Under Test (DUT)—here the DUT is the HT8650 op amp—is forced to ground by the feedback action of the control amplifier—the SHA. Since the input of the VFC circuit is connected to ground during the autozero mode, the input current which can flow is determined by the offset voltage of the HT8650 op amp. Since the output of the integrator stage is forced to ground it is known that the voltage is not changing (it is equal to ground potential). Hence if the output of the integrator is constant, its input current must be zero, so the offset voltage has been forced to be zero. Note that the output of the DUT could have been forced to any convenient voltage other than ground. All that is required is that the output voltage be known to be constant. Note also that the effect of the bias current at the inverting input of the HT8650 op amp is also nulled in this circuit. The 1000 pF capacitor shunting the 200 kΩ resistor is compensation for the two amplifier servo loop. Two integrators in a loop requires a single zero for compensation. Note that the 3.6 kΩ resistor from Pin 1 of the HT8650 to the negative supply is *not* part of the autozero circuit, but rather it is required for VFC operation at 1 MHz.

### PHASE LOCKED LOOP F/V CONVERSION

Although the F/V conversion technique shown in Figure 6 is quite accurate and uses only a few extra components, it is very limited in terms of signal frequency response and carrier feed-through. If the carrier (or input) frequency changes instantaneously, the

output cannot change very rapidly due to the integrator time constant formed by  $C_{INT}$  and  $R_{IN}$ . While it is possible to decrease the integrator time constant to provide faster settling of the F-to-V output voltage, the carrier feedthrough will then be larger. For signal frequency response in excess of 2 kHz, a phase locked F/V conversion technique such as the one shown in Figure 14 is recommended.

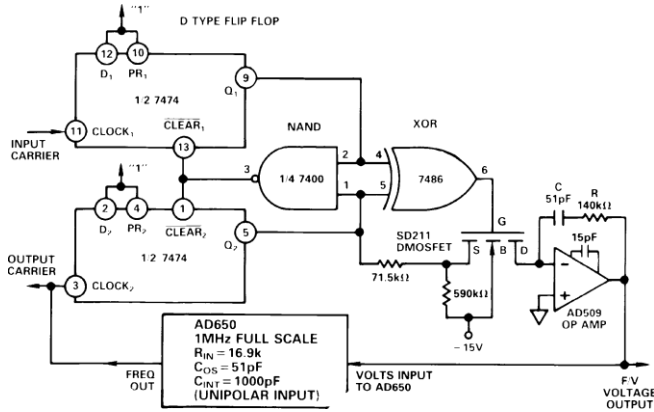


Figure 14. Phase Locked Loop F/V Conversion

In a phase locked loop circuit, the oscillator is driven to a frequency and phase equal to an input reference signal. In applications such as a synthesizer, the oscillator output frequency is first processed through a programmable “divide by N” before being applied to the phase detector as feedback. Here the oscillator frequency is forced to be equal to “N times” the reference frequency and it is this frequency output which is the desired output signal and not a voltage. In this case, the HT8650 offers compact size and wide dynamic range.

In signal recovery applications of a PLL, the desired output signal is the voltage applied to the oscillator. In these situations a linear relationship between the input frequency and the output voltage is desired; the HT8650 makes a superb oscillator for FM demodulation. The wide dynamic range and outstanding linearity of the HT8650 VFC allow simple embodiment of high performance analog signal isolation or telemetry systems. The circuit shown in Figure 14 uses a digital phase detector which also provides proper feedback in the event of unequal frequencies. Such phase-frequency detectors (PFDs) are available in integrated form. For a full discussion of phase lock loop circuits see Reference 3.

An analysis of this circuit must begin at the 7474 dual D flip flop. When the input carrier matches the output carrier in both phase and frequency, the Q outputs of the flip flops will rise at exactly the same time. With two zeros, then two ones on the inputs of the exclusive or (XOR) gate, the output will remain low keeping the DMOS FET switched off. Also, the NAND gate will go low resetting the flip-flops to zero. Throughout the entire cycle just described, the DMOS integrator gate remained off, allowing the voltage at the integrator output to remain unchanged from the previous cycle. However, if the input carrier leads the output carrier by a few degrees, the XOR gate will be turned on for the small time span that the two signals are mismatched. Since  $Q_2$  will be low during the mismatch time, a negative current will be fed into the integrator, causing its output voltage to rise. This in turn will increase the frequency of

the HT8650 slightly, driving the system towards synchronization. In a similar manner, if the input carrier lags the output carrier, the integrator will be forced down slightly to synchronize the two signals.

Using a mathematical approach, the  $\pm 25 \mu A$  pulses from the phase detector are incorporated into the phase detector gain,  $K_d$ . Using a mathematical approach, the  $\pm 25 \mu A$  pulses from the phase detector are incorporated into the phase detector gain,  $K_d$ .

$$K_d = \frac{25 \mu A}{2\pi} = 4 \times 10^{-6} \text{ amperes/radian} \quad (9)$$

Also, the V/F converter is configured to produce 1 MHz in response to a 10 volt input, so its gain  $K_o$ , is:

$$K_o = \frac{2\pi \times 1 \times 10^6 \text{ Hz}}{10 \text{ V}} = 6.3 \times 10^5 \frac{\text{radians}}{\text{volt} \cdot \text{sec}} \quad (10)$$

The dynamics of the phase relationship between the input and output signals can be characterized as a second order system with natural frequency  $\omega_n$ :

$$\omega_n = \sqrt{\frac{K_o K_d}{C}} \quad (11)$$

and damping factor

$$\zeta = \frac{\sqrt{C K_o K_d}}{2} \quad (12)$$

For the values shown in Figure 14, these relations simplify to a natural frequency of 35 kHz with a damping factor of 0.8.

For those desiring a simple approach to determining component values for other PLL frequencies and VFC full-scale voltage, the following cookbook steps can be used:

1. Determine  $K_o$  (in units of radians per volt second) from the maximum input carrier frequency  $F_{MAX}$  (in hertz) and the maximum output voltage  $V_{MAX}$ .

$$K_o = \frac{2\pi \times F_{MAX}}{V_{MAX}} \quad (13)$$

2. Calculate a value for C based upon the desired loop bandwidth,  $f_n$ . Note that this is the desired frequency range of the output signal. The loop bandwidth ( $f_n$ ) is *not* the maximum carrier frequency ( $f_{MAX}$ ); the signal may be very narrow even though it is transmitted over a 1 MHz carrier.

$$C = \frac{K_o \cdot 1 \times 10^{-7} \text{ V} \cdot \text{F}}{f_n^2 \text{ Rad} \cdot \text{sec}} \quad \begin{array}{l} \text{C units FARADS} \\ f_n \text{ units HERTZ} \\ K_o \text{ units RAD/VOLT} \cdot \text{SEC} \end{array} \quad (14)$$

3. Calculate R to yield a damping factor of approximately 0.8 using this equation:

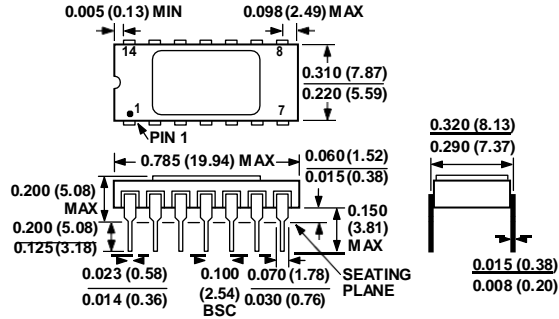
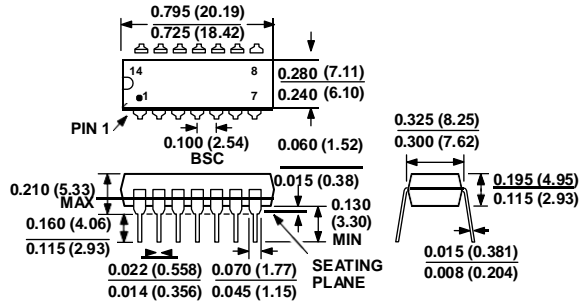
$$R = \frac{f_n \cdot 2.5 \times 10^6 \text{ Rad} \cdot \Omega}{K_o \text{ V}} \quad \begin{array}{l} \text{R units OHMS} \\ f_n \text{ units HERTZ} \\ K_o \text{ units RAD/VOLT} \cdot \text{SEC} \end{array} \quad (15)$$

If in actual operation the PLL overshoots or hunts excessively before reaching a final value, the damping factor may be raised by increasing the value of R. Conversely, if the PLL is overdamped, a smaller value of R should be used.

<sup>3</sup>“Phase lock Techniques,” 2nd Edition, by F.M. Gardner, (John Wiley and Sons, 1979)

**OUTLINE DIMENSIONS**

Dimensions shown in inches and (mm).

**14-Lead Ceramic DIP  
(D-14)**

**14-Lead Plastic DIP  
(N-14)**

**20-Lead PLCC  
(P-20A)**
