

New UltraFast Comparators: Rail-to-Rail Inputs and 2.4V Operation Allow Use on Low Supplies

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Introduction

The new LT1711 family of UltraFast comparators has fully differential rail-to-rail inputs and outputs and operates on supplies as low as 2.4V, allowing unfettered application on low voltages. The LT1711 (single) and LT1712 (dual) are specified at 4.5ns of propagation delay and 100MHz toggle frequency. The low power LT1713 (single) and LT1714 (dual) are specified at 7ns of propagation delay and 65MHz toggle frequency. All of these comparators are fully equipped to support multiple-supply applications, and have latch-enable pins and complementary outputs like the popular LT1016, LT1671 and LT1394. They are available in MSOP and SSOP packages, fully specified over commercial and industrial temperature ranges on 2.7V, 5V and $\pm 5V$ supplies.

Circuit Description

Figure 1 shows a simplified schematic of the LT1711 through LT1714. The front end consists of eight cur-

rent sources and sinks feeding the NPN and PNP differential pairs formed by Q3-Q4 (protected by fast diodes D11-D12) and Q1-Q2 (protected by D1-D2). This approach makes the inputs truly fully differential and noninteracting, unlike approaches that resort to resistors and diode clamps. Even with the inputs at opposite rails, the input bias currents are still a simple function of the input transistor base currents and remain in the μA region. Both input stages feed the level shifting transistors Q5-Q6, and the remainder of the differential voltage gain circuit flows with a delightful symmetry towards the output. Note that the channels are identical, with polarity yet unassigned, and are therefore interchangeable in layout. The symmetry, broken only by the latch-enable circuit, is enhanced by the fact that all of the transistors are well matched, complementary 6GHz f_T BJTs. Each output stage ends in two Baker-clamped common emitter transistors,

allowing full rail-to-rail output swing. All the comparators guarantee full 5V TTL output capability over temperature, even when supplied with only 3V. Output rise and fall times are fast, at 2ns for the LT1711 and LT1712 and 4ns for the LT1713 and LT1714. Jitter is among the lowest for any monolithic comparator, at 11ps_{RMS} for the LT1711 and LT1712 and 15ps_{RMS} for the LT1713 and LT1714.

Some Applications

Simultaneous Full-Duplex 75MBaud Interface with Only Two Wires

The circuit of Figure 2 shows a simple, fully bidirectional, differential 2-wire interface that gives good results to 75MBaud, using the low power LT1714. Eye diagrams under conditions of unidirectional and bidirectional communication are shown in Figures 3 and 4. Although not as pristine as the unidirectional

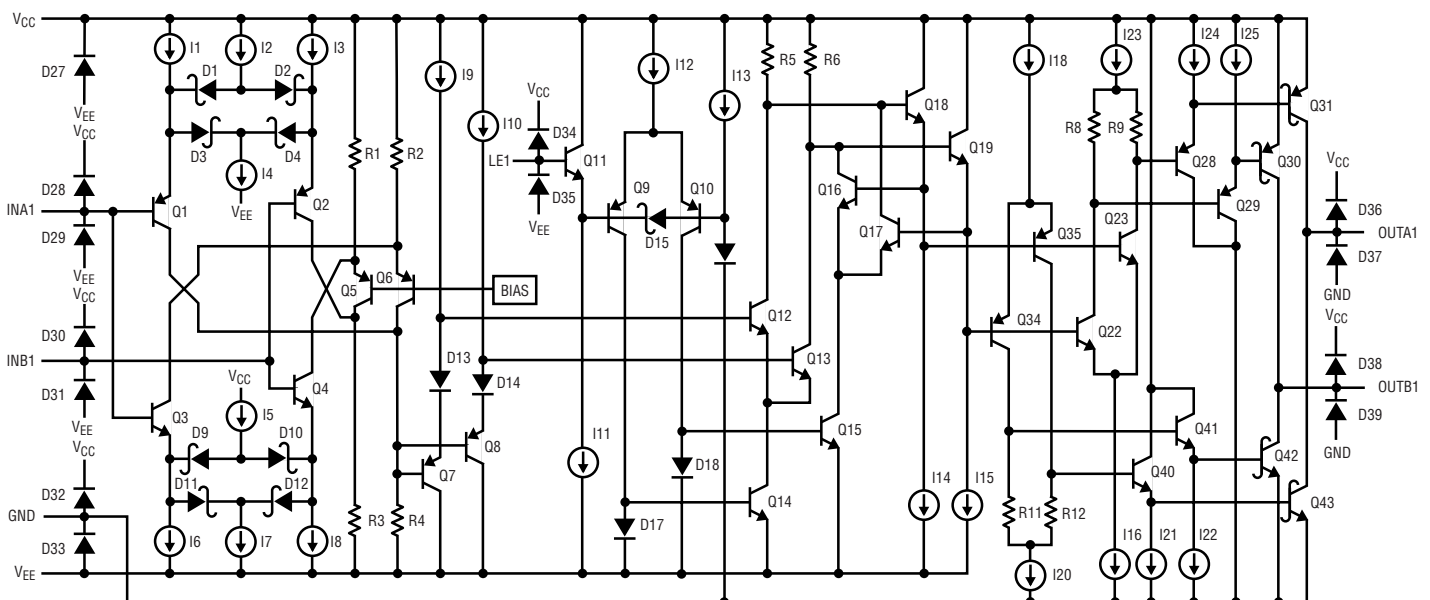


Figure 1. LT1711-LT1714 simplified schematic

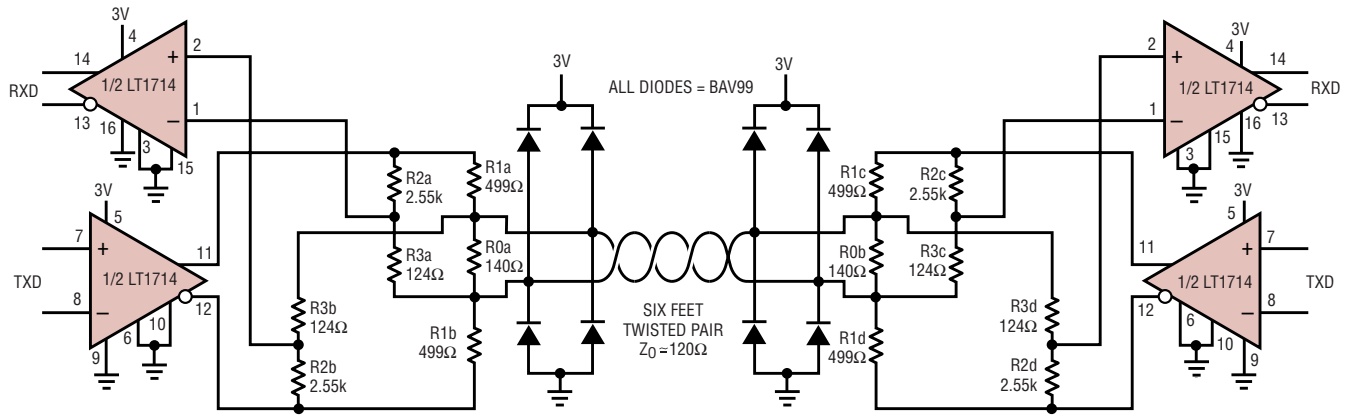


Figure 2. 75MBaud full-duplex interface on two wires

performance of Figure 3, the performance under simultaneous bidirectional operation is still excellent. Because the LT1714 input voltage range extends 100mV beyond both supply rails, the circuit works with a full ±3V of ground potential difference.

The circuit works well with the resistor values shown, but other sets of values can be used. The starting point is the characteristic impedance, Z_0 , of the twisted-pair cable. The input impedance of the resistive network should match the characteristic impedance and is given by:

$$R_{IN} = \frac{2 \cdot R_0 \cdot (R_1 || (R_2 + R_3))}{(R_0 + 2 \cdot (R_1 || (R_2 + R_3)))}$$

This comes out to 120Ω for the values shown. The Thevenin equivalent source voltage is given by:

$$V_{Th} = \frac{V_S \cdot (R_2 + R_3 - R_1)}{(R_2 + R_3 + R_1)} \cdot \frac{R_0}{(R_0 + 2 \cdot (R_1 || (R_2 + R_3)))}$$

This amounts to an attenuation factor of 0.0978 with the values shown. (The actual voltage on the lines will be cut in half again due to the 120Ω Z_0 .) This attenuation factor is important because it is the key to deciding the ratio of the R2, R3 resistor divider in the receiver path. This divider allows the receiver to reject the large signal of the local transmitter and instead sense the attenuated signal of the remote transmitter. Note

that in the above equations, R2 and R3 are not yet fully determined because they only appear as a sum. This allows the designer to now place an additional constraint on their values. The R2, R3 divider ratio should be set to one-half of the attenuation factor mentioned above or $R_3/R_2 = 1/2 \cdot 0.0976$.¹

The author having already designed R2 + R3 to be 2.653kΩ (by allocating input impedance across R_0 , R1, and R2 + R3 to get the requisite 120Ω), R2 and R3 then become 2529Ω and 123.5Ω, respectively. The nearest 1% value for R2 is 2.55k and that for R3 is 124Ω.

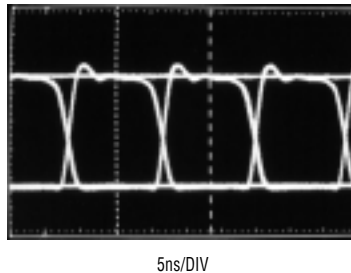


Figure 3. Performance of Figure 1's circuit operated unidirectionally; eye is wide open (cursors show bit interval of 13.3ns or 75MBaud).

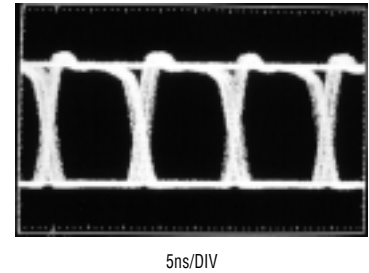


Figure 4. Performance of Figure 1's circuit operated simultaneous-bidirectionally; crosstalk appears as noise. Eye is slightly shut but performance is still excellent.

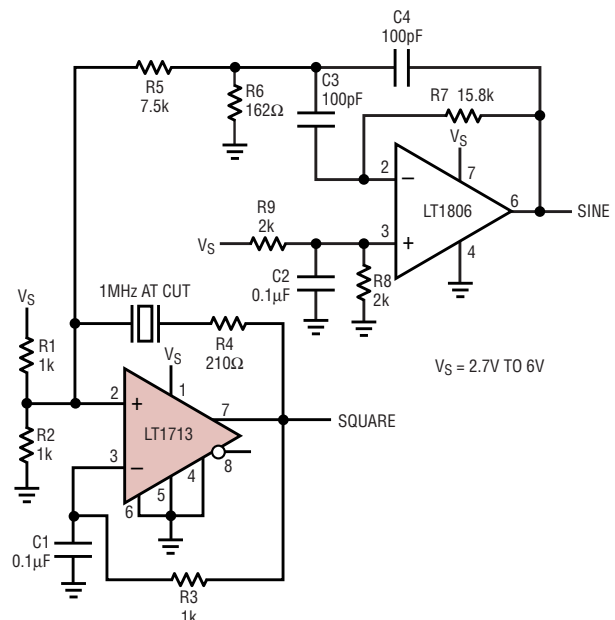


Figure 5. LT1713 comparator configured as a series-resonant crystal oscillator; the LT1806 op amp is configured as a bandpass filter with a Q of 5 and f_c of 1MHz.

1MHz Series-Resonant Crystal Oscillator with Square and Sinusoid Outputs

Figure 5 shows a classic 1MHz series-resonant crystal oscillator. At series resonance, the crystal is a low impedance and the positive feedback connection brings about oscillation at the series resonance frequency. The RC feedback to the $-$ input ensures that the circuit does not find a stable DC operating point and refuse to oscillate. The comparator output is a 1MHz square wave (top trace of Figure 6), with jitter measured at 28ps_{RMS} on a 5V supply and 40ps_{RMS} on a 3V supply. At pin 2 of the comparator, on the other side of the crystal, is a clean sine wave except for the presence of the small high frequency glitch (middle trace of Figure 6). This glitch is caused by the fast edge of the comparator output feeding back through crystal capacitance. Amplitude stability of the sine wave is maintained by the

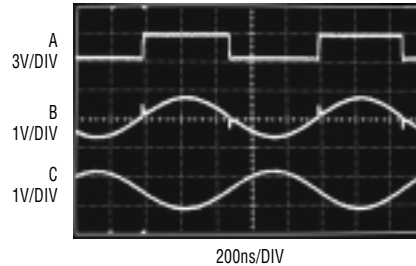



Figure 6. Oscillator waveforms with $V_S = 3\text{V}$: Trace A = comparator output; Trace B = crystal feedback to pin 2 of the LT1713; Trace C = buffered, inverted and bandpass filtered output of LT1806

fact that the sine wave is basically a filtered version of the square wave. Hence, the usual amplitude-control loops associated with sinusoidal oscillators are not necessary.² The sine wave is filtered and buffered by the fast, low noise LT1806 op amp. To remove the glitch, the LT1806 is configured as a bandpass filter with a Q of 5 and unity gain center frequency of 1MHz. The final sinusoidal output

is the bottom trace of Figure 6. Distortion was measured at -70dBc and -55dBc on the second and third harmonics, respectively.

Conclusion

The fully differential rail-to-rail inputs of the new LT1711 family of fast comparators make them useful across a wide variety of applications. The high speed, low jitter performance of this family, coupled with their small package sizes and 2.4V operation, makes them attractive where PCB real estate is at a premium and bandwidth-to-power ratios must be optimized. 

¹ Using the design value of $R2 + R3 = 2.653\text{k}$ rather than the implementation value of $2.55\text{k} + 124\Omega = 2.674\text{k}$.

² Amplitude will be a linear function of comparator output swing, which is supply dependent and therefore adjustable. The important difference here is that any added amplitude stabilization or control loop will not be faced with the classical task of avoiding regions of nonoscillation vs clipping.