

### FEATURES

- Serial data input: 6.5 Mbps to 8.5 Gbps
- No reference clock required
- Exceeds SONET/SDH requirements for jitter transfer/generation/tolerance
- Quantizer sensitivity: 6.3 mV typical (limiting amplifier mode)
- Optional limiting amplifier, equalizer (EQ), and 0 dB EQ inputs
- Programmable jitter transfer bandwidth to support G.8251 OTN
- Programmable slice level
- Sample phase adjust (5.65 Gbps or greater)
- Output polarity invert
- Programmable LOS threshold via I<sup>2</sup>C
- I<sup>2</sup>C interface to access optional features
- Loss of signal (LOS) alarm (limiting amplifier mode only)
- Loss of lock (LOL) indicator
- PRBS generator/detector
- Application-aware power
  - 352 mW at 8.5 Gbps, equalizer mode, no clock output
  - 380 mW at 6.144 Gbps, limiting amplifier mode, no clock output
  - 340 mW at 622 Mbps, 0 dB EQ mode, no clock output
- Power supplies: 1.2 V, flexible 1.8 V to 3.3 V, and 3.3 V
- 4 mm × 4 mm, 24-lead LFCSP

### APPLICATIONS

- SONET/SDH OC-1/OC-3/OC-12/OC-48 and all associated FEC rates
- 1GE, 1GFC, 2GFC, 4GFC, 8GFC, CPRI OS/L.6 up to OS/L.60
- Any rate regenerators/repeaters

### GENERAL DESCRIPTION

The ADN2913 provides the receiver functions of quantization, signal level detection, and clock and data recovery for continuous data rates from 6.5 Mbps to 8.5 Gbps. The ADN2913 automatically locks to all data rates without the need for an external reference clock or programming. ADN2913 jitter performance exceeds all jitter specifications required by SONET/SDH, including jitter transfer, jitter generation, and jitter tolerance.

The ADN2913 provides manual or automatic slice adjust and manual sample phase adjusts. Additionally, the user can select a limiting amplifier, equalizer, or 0 dB EQ at the input. The equalizer is adaptive or it can be manually set.

The receiver front-end loss of signal (LOS) detector circuit indicates when the input signal level falls below a user-programmable threshold. The LOS detection circuit has hysteresis to prevent chatter at the LOS output. In addition, the input signal strength can be read through the I<sup>2</sup>C registers.

The ADN2913 also supports pseudorandom binary sequence (PRBS) generation, bit error detection, and input data rate readback features.

The ADN2913 is available in a compact 4 mm × 4 mm, 24-lead lead frame chip scale package (LFCSP). All ADN2913 specifications are defined over the ambient temperature range of -40°C to +85°C, unless otherwise noted.

### FUNCTIONAL BLOCK DIAGRAM

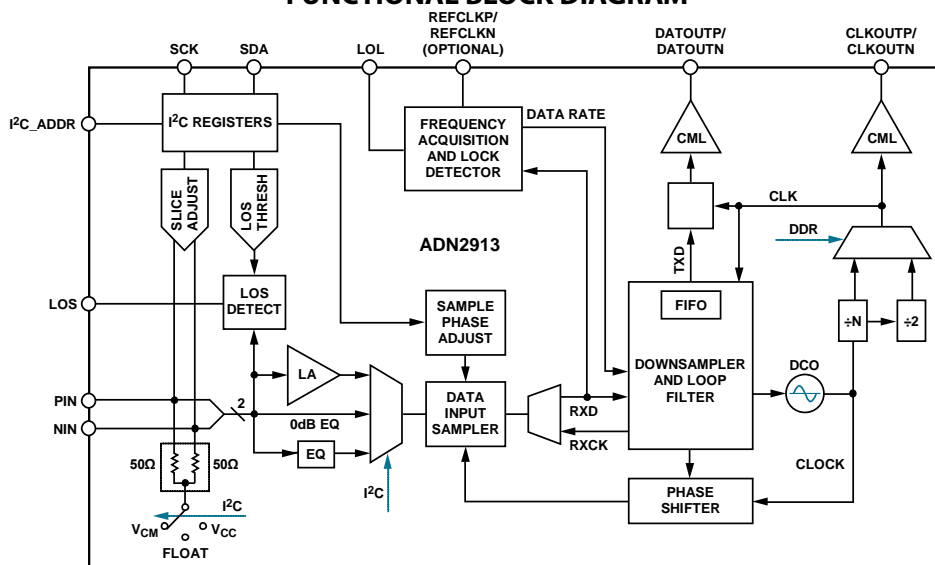


Figure 1.

Rev. 0

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## REVISION HISTORY

12/13—Revision 0: Initial Version

## SPECIFICATIONS

$T_A = T_{MIN}$  to  $T_{MAX}$ ,  $V_{CC} = V_{CC_{MIN}}$  to  $V_{CC_{MAX}}$ ,  $V_{CC1} = V_{CC1_{MIN}}$  to  $V_{CC1_{MAX}}$ ,  $V_{DD} = V_{DD_{MIN}}$  to  $V_{DD_{MAX}}$ ,  $V_{EE} = 0$  V, input data pattern: PRBS  $2^{23} - 1$ , ac-coupled, I<sup>2</sup>C register default settings, unless otherwise noted.

Table 1.

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
DATA RATE SUPPORT RANGE		0.0065		8.5	Gbps
INPUT—DC CHARACTERISTICS					
Peak-to-Peak Differential Input <sup>1</sup>	PIN – NIN			1.0	V
Input Resistance	Differential	95	100	105	$\Omega$
0 dB EQ PATH—CML INPUT					
Input Voltage Range	At PIN or NIN, dc-coupled, RX_TERM_FLOAT = 1 (float)	0.5		VCC	V
Input Common-Mode Level	DC-coupled (see Figure 33), 600 mV p-p differential, RX_TERM_FLOAT = 1 (float)	0.65		VCC – 0.15	V
Differential Input Sensitivity					
OC-48			22		mV p-p
8GFC <sup>2</sup>	Jitter tolerance scrambled pattern (JTSPAT), ac-coupled, RX_TERM_FLOAT = 0 ( $V_{CM} = 1.2$ V), BER = $1 \times 10^{-12}$		200		mV p-p
LIMITING AMPLIFIER INPUT PATH					
Differential Input Sensitivity					
OC-48	BER = $1 \times 10^{-10}$		6.3		mV p-p
8GFC <sup>2</sup>	JTSPAT, BER = $1 \times 10^{-12}$		8.3		mV p-p
EQUALIZER INPUT PATH					
Differential Input Sensitivity	15 inch FR-4, 100 $\Omega$ differential transmission line, adaptive equalizer (EQ) on				
8GFC <sup>2</sup>	JTSPAT, BER = $1 \times 10^{-12}$		115		mV p-p
INPUT—AC CHARACTERISTICS					
S11	At 7.5 GHz, differential return loss, see Figure 14		–12		dB
LOSS OF SIGNAL (LOS) DETECT					
Loss of Signal Detect			10		mV p-p
	Loss of signal minimum program value		5		mV p-p
	Loss of signal maximum program value		128		mV p-p
Hysteresis (Electrical)			5.7		dB
LOS Assert Time	AC-coupled <sup>3</sup>		135		$\mu$ s
LOS Deassert Time	AC-coupled <sup>3</sup>		110		$\mu$ s
LOSS OF LOCK (LOL) DETECT					
DCO Frequency Error for LOL Assert	With respect to nominal, data collected in lock to reference (LTR) mode		1000		ppm
DCO Frequency Error for LOL Deassert	With respect to nominal, data collected in LTR mode		250		ppm
LOL Assert Response Time	10.0 Mbps		10		ms
	2.5 Gbps		51		$\mu$ s
	8.5 Gbps, JTSPAT		25		$\mu$ s
ACQUISITION TIME					
Lock to Data (LTD) Mode					
	10.0 Mbps		24		ms
	2.5 Gbps		0.5		ms
	8.5 Gbps, JTSPAT		0.5		ms
Optional LTR Mode <sup>4</sup>			6.0		ms
DATA RATE READBACK ACCURACY					
Coarse Readback			$\pm 5$		%
Fine Readback	In addition to reference clock accuracy		$\pm 100$		ppm

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
POWER SUPPLY VOLTAGE					
VCC		1.14	1.2	1.26	V
VDD		2.97	3.3	3.63	V
VCC1		1.62	1.8	3.63	V
POWER SUPPLY CURRENT	Limiting amplifier mode, clock output enabled				
VCC	1.25 Gbps		277.1	311.0	mA
	3.125 Gbps		256.2	288.3	mA
	4.25 Gbps		270.1	304.0	mA
	6.144 Gbps		303.1	340.4	mA
	8GFC, <sup>2</sup> JTSPAT		319.1	359.5	mA
VDD	1.25 Gbps		7.24	8.28	mA
	3.125 Gbps		7.21	8.21	mA
	4.25 Gbps		7.23	8.33	mA
	6.144 Gbps		7.26	8.17	mA
	8GFC, <sup>2</sup> JTSPAT		7.20	8.1	mA
VCC1	1.25 Gbps		35.6	46.8	mA
	3.125 Gbps		19.0	24.1	mA
	4.25 Gbps		22.2	28.2	mA
	6.144 Gbps		19.4	24.6	mA
	8GFC, <sup>2</sup> JTSPAT		22.2	28.4	mA
TOTAL POWER DISSIPATION					
Clock Output Enabled	Limiting amplifier mode, 1.25 Gbps		420.4		mW
	Limiting amplifier mode, 3.125 Gbps		365.5		mW
	Limiting amplifier mode, 4.25 Gbps		388		mW
	Limiting amplifier mode, 6.144 Gbps		422.5		mW
	Limiting amplifier mode, 8GFC, <sup>2</sup> JTSPAT		446.6		mW
Clock Output Disabled	Equalizer mode, 8.5 Gbps		352		mW
	Limiting amplifier mode, 6.144 Gbps		380		mW
	0 dB EQ mode, 622 Mbps		340		mW
OPERATING TEMPERATURE RANGE		-40		+85	°C

<sup>1</sup> See Figure 34.

<sup>2</sup> Fibre Channel Physical Interface-4 standard, FC-PI-4, Rev 8.00, May 21, 2008.

<sup>3</sup> When ac-coupled, the LOS assert and deassert times are dominated by the RC time constant of the ac coupling capacitor and the 100 Ω differential input termination of the ADN2913 input stage.

<sup>4</sup> This typical acquisition specification applies to all selectable reference clock frequencies in the range of 11.05 MHz to 176.8 MHz.

## JITTER SPECIFICATIONS

T<sub>A</sub> = T<sub>MIN</sub> to T<sub>MAX</sub>, VCC = VCC<sub>MIN</sub> to VCC<sub>MAX</sub>, VCC1 = VCC1<sub>MIN</sub> to VCC1<sub>MAX</sub>, VDD = VDD<sub>MIN</sub> to VDD<sub>MAX</sub>, VEE = 0 V, input data pattern: PRBS 2<sup>23</sup> - 1, ac-coupled to 100 Ω differential termination load, I<sup>2</sup>C register default settings, unless otherwise noted.

Table 2.

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
PHASE-LOCKED LOOP CHARACTERISTICS					
Jitter Transfer Bandwidth (BW) <sup>1</sup>					
8GFC <sup>2</sup>			1242	1676	kHz
OC-48	TRANBW[2:0] = 4 (default)		663	896	kHz
	OTN mode, TRANBW[2:0] = 1		157	181	kHz
OC-12			175		kHz
OC-3			44		kHz
Jitter Peaking					
8GFC <sup>2</sup>	20 kHz to 80 MHz		0.004	0.021	dB
OC-48	20 kHz to 10 MHz		0.004	0.023	dB
OC-12			0.01		dB
OC-3			0.01		dB

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
Jitter Generation					
8GFC <sup>2</sup>	Unfiltered		0.005		UI rms
	Unfiltered		0.044		UI p-p
OC-48	12 kHz to 20 MHz		0.0025		UI rms
	Unfiltered			0.0046	UI rms
	12 kHz to 20 MHz		0.0156		UI p-p
	Unfiltered			0.0276	UI p-p
OC-12	12 kHz to 5 MHz		0.0007		UI rms
	Unfiltered			0.0011	UI rms
	12 kHz to 5 MHz		0.0038		UI p-p
	Unfiltered			0.0076	UI p-p
OC-3	12 kHz to 1.3 MHz		0.0002		UI rms
	Unfiltered			0.0003	UI rms
	12 kHz to 1.3 MHz		0.0008		UI p-p
	Unfiltered			0.0018	UI p-p
Jitter Tolerance	TRANBW[2:0] = 4 (default)				
8GFC, <sup>2</sup> JTSPAT					
Sinusoidal Jitter at 340 kHz			6.7		UI p-p
Sinusoidal Jitter at 5.098 MHz			0.53		UI p-p
Sinusoidal Jitter at 80 MHz			0.59		UI p-p
Rx Jitter Tracking Test <sup>3</sup>	Voltage modulation amplitude (VMA) = 170 mV p-p at 100 MHz, 425 mV p-p at 100 MHz, 170 mV p-p at 2.5 GHz, and 425 mV p-p at 2.5 GHz excitation frequency <sup>4</sup>				
510 kHz, 1 UI		10 <sup>-12</sup>	<10 <sup>-12</sup>		BER
100 kHz, 5 UI		10 <sup>-12</sup>	<10 <sup>-12</sup>		BER
OC-48	600 Hz		1528		UI p-p
	6 kHz		378		UI p-p
	100 kHz		16.6		UI p-p
	1 MHz		0.70		UI p-p
	20 MHz		0.63		UI p-p
OC-12	30 Hz		193		UI p-p
	300 Hz		44		UI p-p
	25 kHz		19.2		UI p-p
	250 kHz		0.82		UI p-p
	5 MHz		0.60		UI p-p
OC-3	30 Hz		50.0		UI p-p
	300 Hz		24.0		UI p-p
	6500 Hz		14.4		UI p-p
	65 kHz		0.80		UI p-p
	1.3 MHz		0.61		UI p-p

<sup>1</sup> Jitter transfer bandwidth is programmable by adjusting TRANBW[2:0] in the DPLLA register (Address 0x10).

<sup>2</sup> Fibre Channel Physical Interface-4 standard, FC-PI-4, Rev 8.00, May 21, 2008.

<sup>3</sup> Conditions of FC-PI-4, Rev 8.00, Table 27, 800-DF-EL-5 apply.

<sup>4</sup> Must have zero errors during the tests for an interval of time that is  $\leq 10^{-12}$  BER to pass the tests.

## OUTPUT AND TIMING SPECIFICATIONS

$T_A = T_{MIN}$  to  $T_{MAX}$ ,  $V_{CC} = V_{CC_{MIN}}$  to  $V_{CC_{MAX}}$ ,  $V_{CC1} = V_{CC1_{MIN}}$  to  $V_{CC1_{MAX}}$ ,  $V_{DD} = V_{DD_{MIN}}$  to  $V_{DD_{MAX}}$ ,  $V_{EE} = 0$  V, input data pattern: PRBS 2<sup>23</sup> - 1, ac-coupled to 100  $\Omega$  differential termination load, I<sup>2</sup>C register default settings, unless otherwise noted.

Table 3.

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
<b>CML OUTPUT CHARACTERISTICS</b>					
Data Differential Output Swing	8GFC, <sup>1</sup> DATA_SWING[3:0] setting = 0xC (default)	540	600	666	mV p-p
	8GFC, <sup>1</sup> DATA_SWING[3:0] setting = 0xF (maximum)	662	725	778	mV p-p
	8GFC, <sup>1</sup> DATA_SWING[3:0] setting = 0x4 (minimum)	190	214	245	mV p-p
Clock Differential Output Swing	8GFC, <sup>1</sup> CLOCK_SWING[3:0] setting = 0xC (default)	426	518	588	mV p-p
	8GFC, <sup>1</sup> CLOCK_SWING[3:0] setting = 0xF (maximum)	489	603	680	mV p-p
	8GFC, CLOCK_SWING[3:0] setting = 0x4 (minimum)	166	213	245	mV p-p
Output High Voltage	$V_{OH}$ , dc-coupled	$V_{CC} - 0.05$	$V_{CC} - 0.025$	$V_{CC}$	V
Output Low Voltage	$V_{OL}$ , dc-coupled	$V_{CC} - 0.36$	$V_{CC} - 0.325$	$V_{CC} - 0.29$	V
<b>CML OUTPUT TIMING CHARACTERISTICS</b>					
Rise Time	20% to 80%, at 8GFC, <sup>1</sup> DATOUTN/DATOUTP	20.4	33.1	44	ps
	20% to 80%, at 8GFC, <sup>1</sup> CLKOUTN/CLKOUTP	23.1	29.7	35.8	ps
Fall Time	80% to 20%, at 8GFC, <sup>1</sup> DATOUTN/DATOUTP	23	34.2	46.8	ps
	80% to 20%, at 8GFC, <sup>1</sup> CLKOUTN/CLKOUTP	25	31.3	37.1	ps
Setup Time, Full Rate Clock	$t_s$ (see Figure 2)		0.5		UI
Hold Time, Full Rate Clock	$t_H$ (see Figure 2)		0.5		UI
Setup Time, Half Rate/DDR Clock	$t_s$ (see Figure 3)		0.5		UI
Hold Time, Half Rate/DDR Clock	$t_H$ (see Figure 3)		0.5		UI
<b>I<sup>2</sup>C INTERFACE DC CHARACTERISTICS</b>					
Input High Voltage	$V_{IH}$	2.0			V
Input Low Voltage	$V_{IL}$			0.8	V
Input Current	$V_{IN} = 0.1 \times V_{DD}$ or $V_{IN} = 0.9 \times V_{DD}$	-10.0		+10.0	$\mu$ A
Output Low Voltage	$V_{OL}$ , $I_{OL} = 3.0$ mA			0.4	V
<b>I<sup>2</sup>C INTERFACE TIMING</b>					
SCK Clock Frequency	See Figure 22			400	kHz
SCK Pulse Width High	$t_{HIGH}$	600			ns
SCK Pulse Width Low	$t_{LOW}$	1300			ns
Start Condition Hold Time	$t_{HD,STA}$	600			ns
Start Condition Setup Time	$t_{SU,STA}$	600			ns
Data Setup Time	$t_{SU,DAT}$	100			ns
Data Hold Time	$t_{HD,DAT}$	300			ns
SCK/SDA Rise/Fall Time	$t_R/t_F$	$20 + 0.1 C_b^2$		300	ns
Stop Condition Setup Time	$t_{SU,STO}$	600			ns
Bus Free Time Between Stop and Start Conditions	$t_{BUF}$	1300			ns
<b>LVTTL DC INPUT CHARACTERISTICS (I<sup>2</sup>C_ADDR Pin)</b>					
Input Voltage					
High	$V_{IH}$	2.0			V
Low	$V_{IL}$			0.8	V
Input Current					
High	$I_{IH}$ , $V_{IN} = 2.4$ V			5	$\mu$ A
Low	$I_{IL}$ , $V_{IN} = 0.4$ V	-5			$\mu$ A
<b>LVTTL DC OUTPUT CHARACTERISTICS (LOS/LOL Pins)</b>					
Output Voltage					
High	$V_{OH}$ , $I_{OH} = +2.0$ mA	2.4			V
Low	$V_{OL}$ , $I_{OL} = -2.0$ mA			0.4	V

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
REFERENCE CLOCK CHARACTERISTICS	Optional LTR mode				
Input Compliance Voltage (Common-Mode Voltage Referred to Ground)	$V_{CM}$ (no input offset, no input current), see Figure 30, ac-coupled input	0.55		1.0	V
Minimum Input Drive	See Figure 30, ac-coupled, differential input		100		mV p-p diff
Reference Frequency		11.05		176.8	MHz
Required Accuracy <sup>3</sup>	AC-coupled, differential input		100		ppm

<sup>1</sup> Fibre Channel Physical Interface-4 standard, FC-PI-4, Rev 8.00, May 21, 2008.

<sup>2</sup>  $C_b$  is the total capacitance of one bus line in picofarads (pF). If mixed with high speed (HS) mode devices, faster rise/fall times are allowed (refer to the Philips *I<sup>2</sup>C Bus Specification, Version 2.1*).

<sup>3</sup> Required accuracy in dc-coupled mode is guaranteed by design as long as the clock common-mode voltage output matches the reference clock common-mode voltage range.

TIMING DIAGRAMS

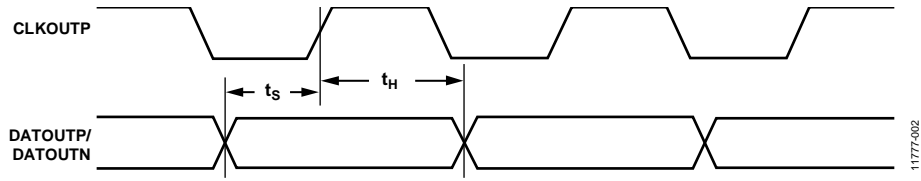


Figure 2. Data to Clock Timing (Full Rate Clock Mode)

11777-002

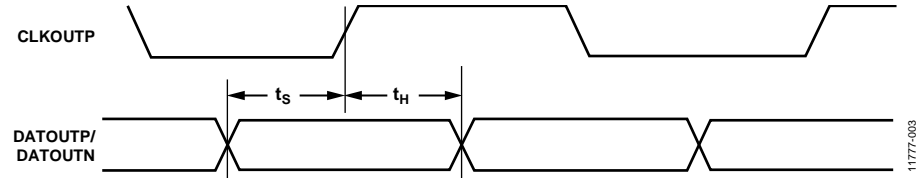
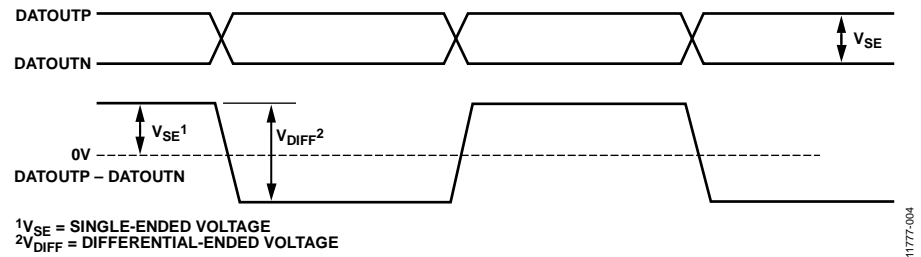


Figure 3. Data to Clock Timing (Half Rate Clock/DDR Mode)

11777-003



<sup>1</sup> $V_{SE}$  = SINGLE-ENDED VOLTAGE  
<sup>2</sup> $V_{DIFF}$  = DIFFERENTIAL-ENDED VOLTAGE

11777-004

Figure 4. Single-Ended vs. Differential Output Amplitude Relationship



## ABSOLUTE MAXIMUM RATINGS

Table 4.

Parameter	Rating
Supply Voltage (VCC = 1.2 V)	1.26 V
Supply Voltage (VDD and VCC1 = 3.3 V)	3.63 V
Maximum Input Voltage (REFCLKP/REFCLKN, NIN/PIN)	1.26 V
Minimum Input Voltage (REFCLKP/REFCLKN, NIN/PIN)	$V_{EE} - 0.4 V$
Maximum Input Voltage (SDA, SCK, I <sup>2</sup> C_ADDR)	3.63 V
Minimum Input Voltage (SDA, SCK, I <sup>2</sup> C_ADDR)	$V_{EE} - 0.4 V$
Maximum Junction Temperature	125°C
Storage Temperature Range	-65°C to +150°C
Lead Temperature (Soldering, 10 sec)	300°C

Stresses above those listed under Absolute Maximum Ratings may cause permanent damage to the device. This is a stress rating only; functional operation of the device at these or any other conditions above those indicated in the operational section of this specification is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

## THERMAL CHARACTERISTICS

### Thermal Resistance

Thermal resistance is specified for the worst-case conditions, that is, a device soldered in a circuit board for surface-mount packages, for a 4-layer board with the exposed pad soldered to VEE.

Table 5. Thermal Resistance

Package Type	$\theta_{JA}$ <sup>1</sup>	$\theta_{JB}$ <sup>2</sup>	$\theta_{JC}$ <sup>3</sup>	Unit
24-Lead LFCSP	45	5	11	°C/W

<sup>1</sup> Junction to ambient.

<sup>2</sup> Junction to board.

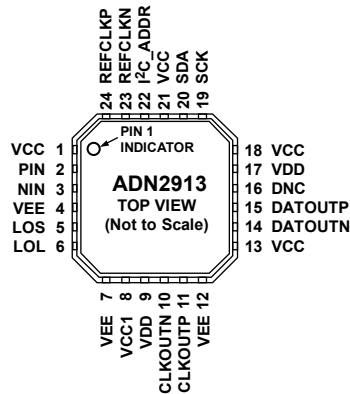
<sup>3</sup> Junction to case.

### ESD CAUTION



**ESD (electrostatic discharge) sensitive device.** Charged devices and circuit boards can discharge without detection. Although this product features patented or proprietary protection circuitry, damage may occur on devices subjected to high energy ESD. Therefore, proper ESD precautions should be taken to avoid performance degradation or loss of functionality.

## PIN CONFIGURATION AND FUNCTION DESCRIPTIONS



### NOTES

1. DNC = DO NOT CONNECT. TIE OFF TO GROUND.
2. THE EXPOSED PAD ON THE BOTTOM OF THE DEVICE PACKAGE MUST BE CONNECTED TO VEE ELECTRICALLY. THE EXPOSED PAD WORKS AS A HEAT SINK.

11777-005

Figure 5. Pin Configuration

Table 6. Pin Function Descriptions

Pin No.	Mnemonic	Type <sup>1</sup>	Description
1	VCC	P	1.2 V Supply for Limiting Amplifier.
2	PIN	AI	Positive Differential Data Input (CML).
3	NIN	AI	Negative Differential Data Input (CML).
4	VEE	P	Ground for Limiting Amplifier.
5	LOS	DO	Loss of Signal Output (Active High).
6	LOL	DO	Loss of Lock Output (Active High).
7	VEE	P	Digital Control Oscillator (DCO) Ground.
8	VCC1	P	1.8 V to 3.3 V DCO Supply.
9	VDD	P	3.3 V High Supply.
10	CLKOUTN	DO	Negative Differential Recovered Clock Output (CML).
11	CLKOUTP	DO	Positive Differential Recovered Clock Output (CML).
12	VEE	P	Ground for CML Output Drivers.
13	VCC	P	1.2 V Supply for CML Output Drivers.
14	DATOUTN	DO	Negative Differential Retimed Data Output (CML).
15	DATOUTP	DO	Positive Differential Retimed Data Output (CML).
16	DNC	DI	Do Not Connect. Leave this pin unconnected or tie it to VEE (ground).
17	VDD	P	3.3 V High Supply.
18	VCC	P	1.2 V Core Digital Supply.
19	SCK	DI	Clock for I <sup>2</sup> C Interface.
20	SDA	DIO	Bidirectional Data for I <sup>2</sup> C Interface.
21	VCC	P	1.2 V Core Digital Supply.
22	I <sup>2</sup> C_ADDR	DI	I <sup>2</sup> C Address Setting. Sets the device I <sup>2</sup> C address = 0x80 when I <sup>2</sup> C_ADDR = 0. Sets the device I <sup>2</sup> C address = 0x82 when I <sup>2</sup> C_ADDR = 1.
23	REFCLKN	DI	Negative Reference Clock Input (Optional).
24	REFCLKP	DI	Positive Reference Clock Input (Optional).
	EPAD	P	Exposed Pad (VEE). The exposed pad on the bottom of the device package must be connected to VEE electrically. The exposed pad works as a heat sink.

<sup>1</sup> P = power, AI = analog input, DI = digital input, DO = digital output, DIO = digital input/output.

### TYPICAL PERFORMANCE CHARACTERISTICS

T<sub>A</sub> = 25°C, VCC = 1.2 V, VCC1 = 1.8 V, VDD = 3.3 V, VEE = 0 V, input data pattern: PRBS 2<sup>15</sup> - 1, ac-coupled inputs and outputs, unless otherwise noted.

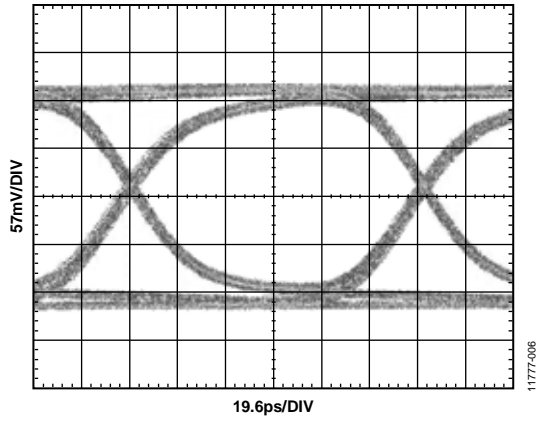


Figure 6. Output Eye Diagram at 8GFC

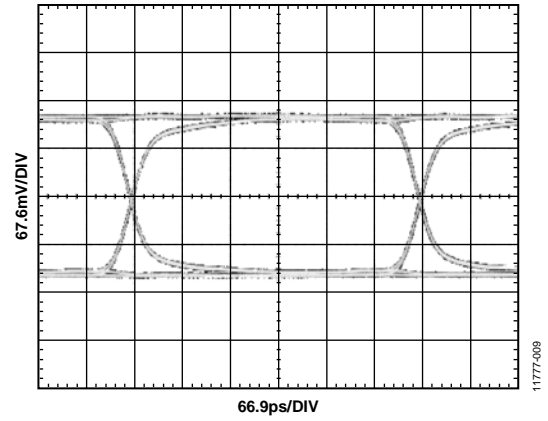


Figure 9. Output Eye Diagram at OC-48

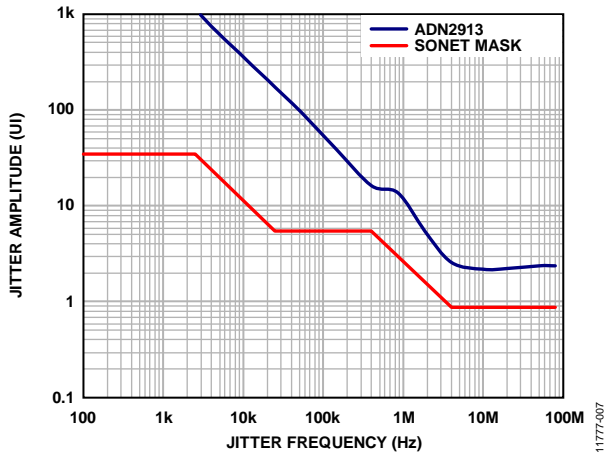


Figure 7. Jitter Tolerance: 8GFC

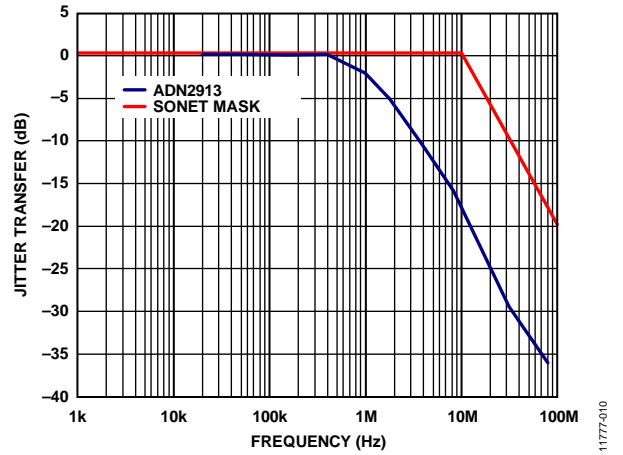


Figure 10. Jitter Transfer: 8GFC (TRANBW[2:0] = 4)

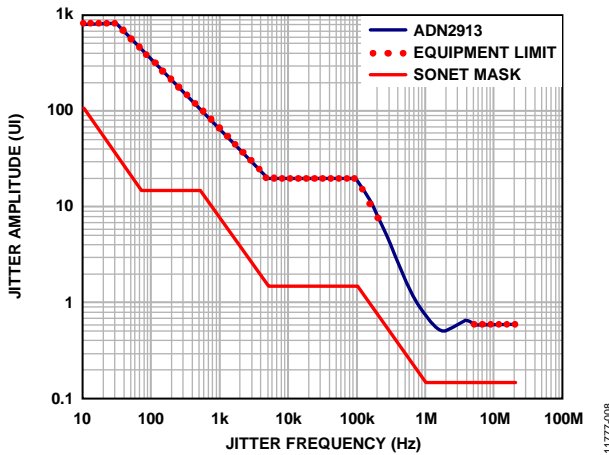


Figure 8. Jitter Tolerance: OC-48

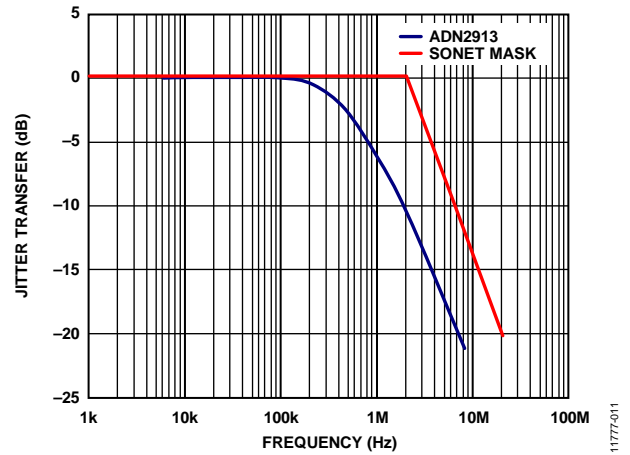


Figure 11. Jitter Transfer: OC-48

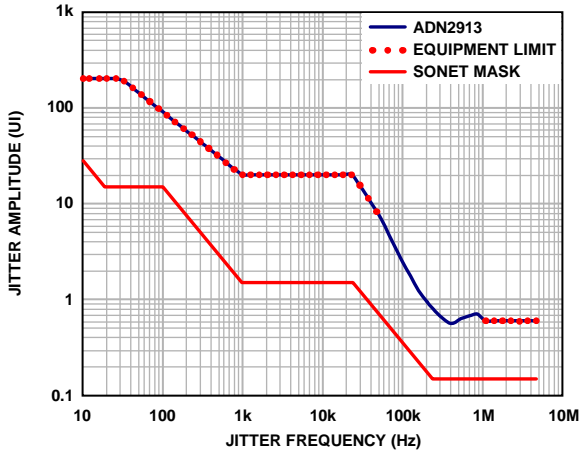


Figure 12. Jitter Tolerance: OC-12

11777-012

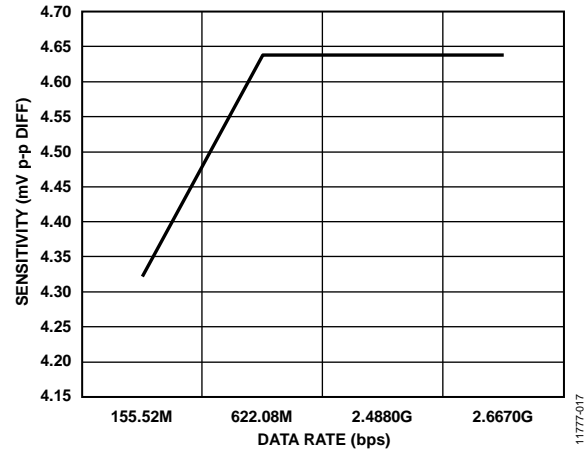


Figure 15. Sensitivities of SONET/SDH Data Rates ( $BER = 10^{-10}$ )

11777-017

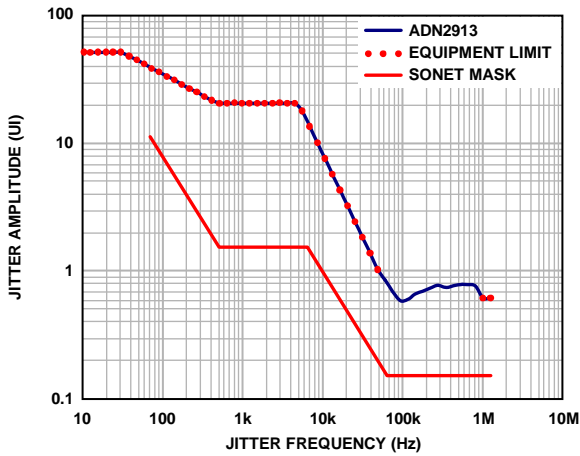


Figure 13. Jitter Tolerance: OC-3

11777-013

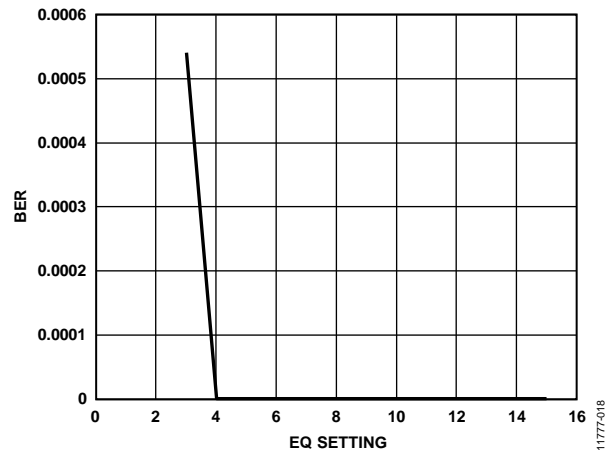


Figure 16. BER in Equalizer Mode vs. EQ Compensation at 8GFC (Measured with an 8GFC Signal of 400 mV p-p diff, on 15-Inch FR4 Traces, with Variant EQ Compensation, Including Adaptive EQ)

11777-018

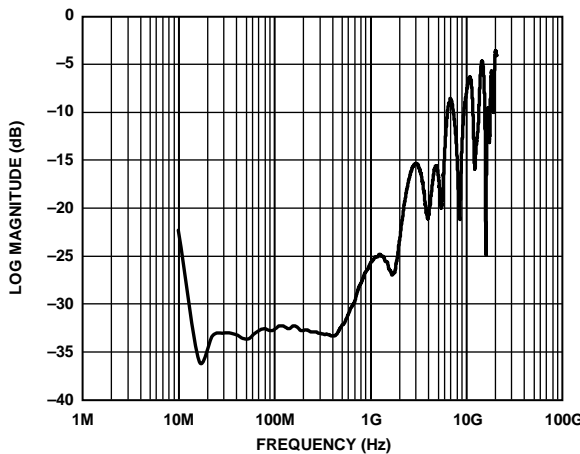


Figure 14. Typical S11 Spectrum Performance

11777-014

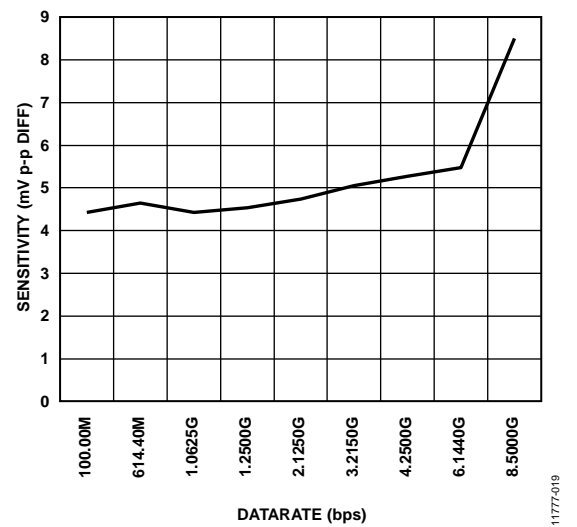


Figure 17. Sensitivities of non-SONET/SDH Data Rates ( $BER = 10^{-12}$ )

11777-019

# I<sup>2</sup>C INTERFACE TIMING AND INTERNAL REGISTER DESCRIPTIONS

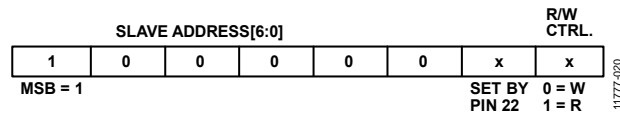


Figure 18. Slave Address Configuration

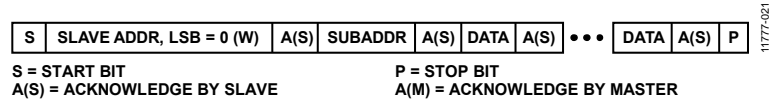


Figure 19. I<sup>2</sup>C Write Data Transfer

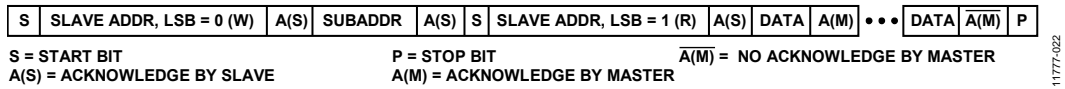


Figure 20. I<sup>2</sup>C Read Data Transfer

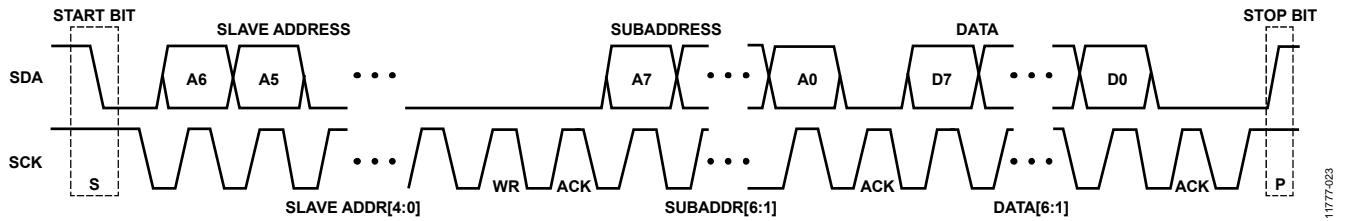


Figure 21. I<sup>2</sup>C Data Transfer Timing

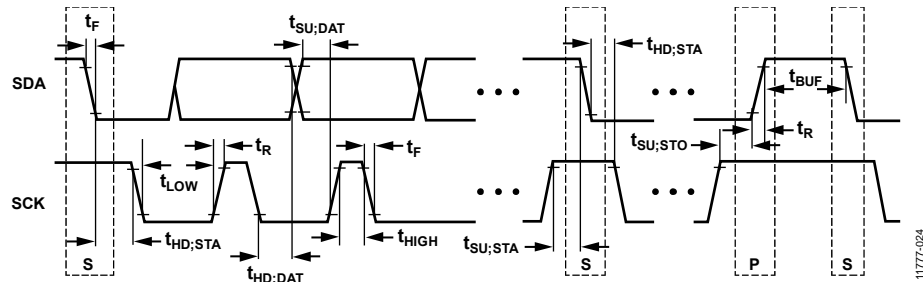


Figure 22. I<sup>2</sup>C Interface Timing Diagram

## REGISTER MAP

Writing to register bits other than those labeled in Table 7 is not recommended and may cause unintended results.

Table 7. Internal Register Map

Reg Name	R/W	Addr (Hex)	Default (Hex)	D7	D6	D5	D4	D3	D2	D1	D0	
<b>Readback/Status</b>												
FREQMEAS0	R	0x0	N/A	FREQ0[7:0] (RATE_FREQ[7:0])								
FREQMEAS1	R	0x1	N/A	FREQ1[7:0] (RATE_FREQ[15:8])								
FREQMEAS2	R	0x2	N/A	FREQ2[7:0] (RATE_FREQ[23:16])								
FREQ_RB1	R	0x4	N/A	VCOSEL[7:0]								
FREQ_RB2	R	0x5	N/A		FULLRATE	DIVRATE[3:0]				VCOSEL[9:8]		
STATUSA	R	0x6	N/A			LOS status	LOL status	LOS done	Static LOL		RATE_MEAS_COMP	
<b>General Control</b>												
CTRLA	R/W	0x8	0x10	0	CDR_MODE[2:0]			0	Reset static LOL	RATE_MEAS_EN	RATE_MEAS_RESET	
CTRLB	R/W	0x9	0x00	SOFTWARE_RESET	INIT_FREQ_ACQ	CDR bypass	LOL_CONFIG	LOS_PDN	LOS polarity	0	0	
CTRLC	R/W	0xA	0x05	0	0	0	0	0	REFCLK_PDN	0	1	
<b>FLL Control</b>												
LTR_MODE	R/W	0xF	0x00	0	LOL data	FREQ_RANGE[1:0]		DATA_TO_REF_RATIO[3:0]				
<b>D/PLL Control</b>												
DPLLA	R/W	0x10	0x1C	0	0	0	EDGE_SEL[1:0]		TRANBW[2:0]			
DPLLD	R/W	0x13	0x06	0	0	0	0	0	ADAPTIVE_SLICE_EN	DLL_SLEW[1:0]		
Phase	R/W	0x14	0x00	0	0	0	0	SAMPLE_PHASE[3:0]				
Slice	W	0x15	N/A	Extended slice	Slice[6:0]							
LA_EQ	R/W	0x16	0x08	RX_TERM_FLOAT	INPUT_SEL[1:0]		ADAPTIVE_EQ_EN	EQ_BOOST[3:0]				
Slice Readback	R	0x73	N/A	SLICE_RB[7:0]								
<b>Output Control</b>												
OUTPUTA	R/W	0x1E	0x00	0	0	Data squelch	DATOUT_DISABLE	CLKOUT_DISABLE	DDR_DISABLE	DATA_POLARITY	CLOCK_POLARITY	
OUTPUTB	R/W	0x1F	0xCC	DATA_SWING[3:0]				CLOCK_SWING[3:0]				
<b>LOS Control</b>												
LOS_DATA	R/W	0x36	0x00	LOS_DATA[7:0]								
LOS_CTRL	R/W	0x74	0x00	0	0	LOS_WRITE	LOS_ENABLE	LOS_RESET	LOS_ADDRESS[2:0]			
LOS_THRESH	R/W	0x38	0x0A	LOS_THRESHOLD[7:0]								
<b>PRBS Control</b>												
PRBS Gen 1	R/W	0x39	0x00	0	0	DATA_CID_BIT	DATA_CID_EN	0	DATA_GEN_EN	DATA_GEN_MODE[1:0]		
PRBS Gen 2	R/W	0x3A	0x00	DATA_CID_LENGTH[7:0]								
PRBS Gen 3	R/W	0x3B	0x00	PROG_DATA[7:0]								
PRBS Gen 4	R/W	0x3C	0x00	PROG_DATA[15:8]								
PRBS Gen 5	R/W	0x3D	0x00	PROG_DATA[23:16]								
PRBS Gen 6	R/W	0x3E	0x00	PROG_DATA[31:24]								
PRBS Rec 1	R/W	0x3F	0x00	0	0	0	0	DATA_RECEIVER_CLEAR	DATA_RECEIVER_ENABLE	DATA_RECEIVER_MODE[1:0]		
PRBS Rec 2	R	0x40	0x00	PRBS_ERROR_COUNT[7:0]								
PRBS Rec 3	R	0x41	0x00								PRBS_ERROR	
PRBS Rec 4	R	0x42	N/A	DATA_LOADED[7:0]								

Reg Name	R/W	Addr (Hex)	Default (Hex)	D7	D6	D5	D4	D3	D2	D1	D0
PRBS Rec 5	R	0x43	N/A	DATA_LOADED[15:8]							
PRBS Rec 6	R	0x44	N/A	DATA_LOADED[23:16]							
PRBS Rec 7	R	0x45	N/A	DATA_LOADED[31:24]							
<b>ID/Revision</b>											
REV	R	0x48	0x54	REV[7:0]							
ID	R	0x49	0x15	ID[7:0]							
HI_CODE	R	0x20	0xA8	Reserved							
LO_CODE	R	0x21	0x00	Reserved							

Table 8. Status Register, STATUSA (Address 0x6)

Bits	Bit Name	Bit Description
D5	LOS status	0 = no loss of signal 1 = loss of signal
D4	LOL status	0 = locked 1 = frequency acquisition mode
D3	LOS done	0 = LOS action not completed 1 = LOS action completed
D2	Static LOL	0 = no LOL event since last reset 1 = LOL event since last reset; clear using reset static LOL bit, Bit D2 in Register 0x8
D0	RATE_MEAS_COMP	Rate measurement complete 0 = frequency measurement incomplete 1 = frequency measurement complete; clear using RATE_MEAS_RESET bit, Bit D0 in Register 0x8

Table 9. Control Register, CTRLA (Address 0x8)

Bits	Bit Name	Bit Description
D7		Reserved to 0.
D6:D4	CDR_MODE[2:0]	CDR modes. 001 = lock to data (LTD). 011 = lock to reference (LTR). 000, 010, 1xx = reserved.
D3		Reserved to 0.
D2	Reset static LOL	Set to 1 to clear static LOL (Bit D2 in Register 0x6).
D1	RATE_MEAS_EN	Fine data rate measurement enable. Set to 1 to initiate a rate measurement.
D0	RATE_MEAS_RESET	Rate measurement reset. Set to 1 to clear a rate measurement.

Table 10. Control Register, CTRLB (Address 0x9)

Bits	Bit Name	Bit Description
D7	SOFTWARE_RESET	Software reset. Write a 1 followed by a 0 to reset the device.
D6	INIT_FREQ_ACQ	Initiate frequency acquisition. Write a 1 followed by a 0 to initiate a frequency acquisition (optional).
D5	CDR bypass	CDR bypass. 0 = CDR enabled. 1 = CDR bypassed.
D4	LOL_CONFIG	LOL configuration. 0 = normal LOL. 1 = static LOL.
D3	LOS_PDN	LOS power-down. 0 = normal LOS. 1 = LOS powered down.
D2	LOS polarity	LOS polarity. 0 = active high LOS pin. 1 = active low LOS pin.
D1:D0		Reserved to 0.

Table 11. Control Register, CTRLC (Address 0xA)

Bits	Bit Name	Bit Description
D7:D3		Reserved to 0.
D2	REFCLK_PDN	Reference clock power-down. Write a 0 to enable the reference clock.
D1		Reserved to 0.
D0		Reserved to 1.

Table 12. Lock to Reference Clock Mode Programming Register, LTR\_MODE (Address 0xF)

Bits	Bit Name	Bit Description
D7		Reserved to 0
D6	LOL data	LOL data 0 = valid recovered clock vs. reference clock during tracking 1 = valid recovered clock vs. data during tracking
D5:D4	FREF_RANGE[1:0]	$f_{REF}$ range 00 = 11.05 MHz to 22.1 MHz 01 = 22.1 MHz to 44.2 MHz 10 = 44.2 MHz to 88.4 MHz 11 = 88.4 MHz to 176.8 MHz
D3:D0	DATA_TO_REF_RATIO[3:0]	Data to reference ratio <sup>1</sup> ( $N \geq 2^{(N-1)}$ , where N is the decimal equivalent of the binary code) 0000 = 1/2 0001 = 1 0010 = 2 0011 = 4 0100 = 8 ... 1010 = 512

<sup>1</sup>  $Data \div DIV_{f_{REF}}$ , where  $DIV_{f_{REF}}$  is the divided down reference referred to the 11.05 MHz to 22.1 MHz band (see the Reference Clock (Optional) section).  
 $Data\ Rate / 2^{(LTR\_MODE[3:0] - 1)} = REFCLK / 2^{LTR\_MODE[5:4]}$

Table 13. D/PLL Control Register, DPLLA (Address 0x10)

Bits	Bit Name	Bit Description
D7:D5		Reserved to 0.
D4:D3	EDGE_SEL[1:0]	Edge for phase detection. See the Edge Select section for more information. 00 = rising and falling edge data. 01 = rising edge data. 10 = falling edge data. 11 = rising and falling edge data.
D2:D0	TRANBW[2:0]	Transfer bandwidth. Scales the transfer bandwidth. Default value is 4, resulting in the 8GFC default BW shown in Table 2. See the Transfer Bandwidth section for more information. $Transfer\ BW = Default\ BW \times (TRANBW[2:0]/4)$

Table 14. D/PLL Control Register, DPLLD (Address 0x13)

Bits	Bit Name	Bit Description
D7:D3		Reserved to 0.
D2	ADAPTIVE_SLICE_EN	Adaptive slice enable. 1 = enables automatic slice adjust.
D1:D0	DLL_SLEW[1:0]	DLL slew. Sets the BW of the DLL. See the DLL Slew section for more information.

Table 15. Phase Control Register, Phase (Address 0x14)

Bits	Bit Name	Bit Description
D7:D4		Reserved to 0.
D3:D0	SAMPLE_PHASE[3:0]	Adjust the phase of the sampling instant for data rates above 5.65 Gbps in steps of 1/32 UI. This register is in twos complement format. See the Sample Phase Adjust section for more information.



Table 16. Slice Level Control Register, Slice (Address 0x15)

Bits	Bit Name	Bit Description
D7	Extended slice	Extended slice enable. 0 = normal slice mode. 1 = extended slice mode.
D6:D0	Slice[6:0]	Slice is a digital word that sets the input threshold. See the Slice Adjust section for more information. When slice[6:0] = 0000000, the slice function is disabled.

Table 17. Input Stage Programming Register, LA\_EQ (Address 0x16)

Bits	Bit Name	Bit Description
D7	RX_TERM_FLOAT	Rx termination float. 0 = termination common-mode driven. 1 = termination common-mode floated ( $V_{CC} = 1.2 V$ ).
D6:D5	INPUT_SEL[1:0]	Input stage select. 00: limiting amplifier. 01: equalizer. 10: 0 dB EQ. 11: undefined.
D4	ADAPTIVE_EQ_EN	Enable adaptive EQ. 0 = manual EQ control. 1 = adaptive EQ enabled.
D3:D0	EQ_BOOST[3:0]	Equalizer gain. These bits set the EQ gain. See the Passive Equalizer section for more information.

Table 18. Output Control Register, OUTPUTA (Address 0x1E)

Bits	Bit Name	Bit Description
D7:D6		Reserved to 0
D5	Data squelch	Squelch 0 = normal data 1 = squelch data
D4	DATOUT_DISABLE	Data output disable 0 = data output enabled 1 = data output disabled
D3	CLKOUT_DISABLE	Clock output disable 0 = clock output enabled 1 = clock output disabled
D2	DDR_DISABLE	Double data rate 0 = DDR clock enabled 1 = DDR clock disabled
D1	DATA_POLARITY	Data polarity 0 = normal data polarity 1 = flip data polarity
D0	CLOCK_POLARITY	Clock polarity 0 = normal clock polarity 1 = flip clock polarity

Table 19. Output Swing Register, OUTPUTB (Address 0x1F)

Bits	Bit Name	Bit Description
D7:D4	DATA_SWING[3:0]	Adjust data output amplitude. Step size is approximately 50 mV differential. Default register value is 0xC. Typical differential data output amplitudes are 0x1 = invalid. 0x2 = invalid. 0x3 = invalid. 0x4 = 200 mV. 0x5 = 250 mV. 0x6 = 300 mV. 0x7 = 345 mV. 0x8 = 390 mV. 0x9 = 440 mV. 0xA = 485 mV. 0xB = 530 mV. 0xC = 575 mV. 0xD = 610 mV. 0xE = 640 mV. 0xF = 655 mV.
D3:D0	CLOCK_SWING[3:0]	Adjust clock output amplitude. Step size is approximately 50 mV differential. Default register value is 0xC. Typical differential clock output amplitudes are 0x1 = invalid. 0x2 = invalid. 0x3 = invalid. 0x4 = 200 mV. 0x5 = 250 mV. 0x6 = 300 mV. 0x7 = 345 mV. 0x8 = 390 mV. 0x9 = 440 mV. 0xA = 485 mV. 0xB = 530 mV. 0xC = 575 mV. 0xD = 610 mV. 0xE = 640 mV. 0xF = 655 mV.

## THEORY OF OPERATION

The ADN2913 implements clock and data recovery for data rates between 6.5 Mbps and 8.5 Gbps. A front end is configurable to either amplify or equalize the nonreturn-to-zero (NRZ) input waveform to full-scale digital logic levels, or to bypass a full digital logic signal.

The user can choose one of three input stages to process the data: a high gain limiting amplifier with better than 10 mV sensitivity, a high-pass passive equalizer with up to 10 dB of boost at 5 GHz, or a 0 dB EQ buffer with 600 mV sensitivity.

An on-chip loss of signal (LOS) detector works with the high sensitivity limiting amplifier. The default threshold for the LOS detector is the sensitivity of the device, with a maximum threshold level of 128 mV p-p. The limiting amplifier slice threshold can use a factory trim setting, a user defined threshold set by the I<sup>2</sup>C interface, or an adjusted level for the best eye opening at the phase detector.

When the input signal is corrupted due to FR-4 or other impairments in the PCB traces, a passive equalizer can be one of the signal integrity options. The equalizer high frequency boost is configurable through the I<sup>2</sup>C registers. A user enabled adaptation is included that automatically adjusts the equalizer to achieve the widest eye opening. The equalizer can be manually set for any data rate, but adaptation is available only at data rates greater than 5.5 Gbps.

When a signal is presented to the clock and data recovery (CDR) system, the ADN2913 acts as a delay-locked and phase-locked loop circuit for clock recovery and data retiming from an NRZ encoded data stream. Input data is sampled by a high speed clock. A digital downsampler accommodates data rates spanning three orders of magnitude. Downsampled data is applied to a binary phase detector (see Figure 23).

The phase of the input data signal is tracked by two separate feedback loops. A high speed delay-locked loop path combines a digital integrator with a digitally controlled phase shifter (PSH) on the digital control oscillator (DCO) clock to track the high frequency components of jitter. A separate phase-locked loop composed of a digital integrator and DCO tracks the low

frequency components of jitter. The initial frequency of the DCO is set by a third loop that compares the DCO frequency with the input data frequency. This third loop also sets the decimation ratio of the digital downsampler.

The delay-locked and phase-locked loops together track the phase of the input data. For example, when the clock lags the input data, the phase detector drives the DCO to a higher frequency and decreases the delay of the clock through the phase shifter; both of these actions serve to reduce the phase error between the clock and data. Because the loop filter is an integrator, the static phase error is driven to zero.

Another view of the circuit is that the phase shifter implements the zero required for frequency compensation of a second-order phase-locked loop. This zero is placed in the feedback path and, therefore, does not appear in the closed-loop transfer function. Because this circuit has no zero in the closed-loop transfer, jitter peaking is eliminated.

The combination of the delay-locked and phase-locked loops simultaneously provide wideband jitter tolerance and narrow-band jitter filtering. The simplified block diagram in Figure 23 shows that  $Z(s)/X(s)$  is a second-order low-pass jitter transfer function that provides excellent filtering. The low frequency pole is formed by dividing the gain of the PLL by the gain of the DLL, where the upsampling and zero-order hold in the DLL has a gain approaching N at the transfer bandwidth of the loop. Note that the jitter transfer has no zero, unlike an ordinary second-order phase-locked loop. This means that the main PLL loop has no jitter peaking, making the circuit ideal for signal regenerator applications, where jitter peaking in a cascade of regenerators can contribute to hazardous jitter accumulation.

The error transfer,  $e(s)/X(s)$ , has the same high-pass form as an ordinary phase-locked loop up to the slew rate limit of the DLL with a binary phase detector. This transfer function can be optimized to give excellent wideband jitter tolerance because the jitter transfer function,  $Z(s)/X(s)$ , provides the narrow-band jitter filtering.

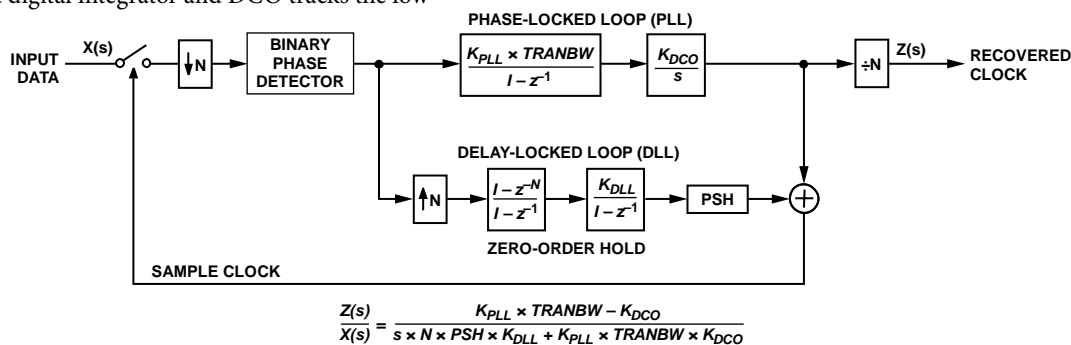


Figure 23. CDR Jitter Block Diagram

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The delay-locked and phase-locked loops contribute to overall jitter tolerance. At low frequencies of input jitter on the data signal, the integrator in the loop filter provides high gain to track large jitter amplitudes with small phase error. In this case, the oscillator is frequency modulated and jitter is tracked as in an ordinary phase-locked loop. The amount of low frequency jitter that can be tracked is a function of the DCO tuning range. A wider tuning range gives larger tolerance of low frequency jitter. The internal loop control word remains small for small jitter frequency so that the phase shifter remains close to the center of its range and, thus, contributes little to the low frequency jitter tolerance.

At medium jitter frequencies, the gain and tuning range of the DCO are not large enough to track input jitter. In this case, the DCO control word becomes large and saturates. As a result, the DCO frequency remains at an extreme of its tuning range. The

size of the DCO tuning range, therefore, has only a small effect on the jitter tolerance. The delay-locked loop control range is now larger; therefore, the phase shifter tracks the input jitter. An infinite range phase shifter is used on the clock. Consequently, the minimum range of timing mismatch between the clock at the data sampler and the retiming clock at the output is limited by the depth of the FIFO to 32 UI.

There are two ways to acquire the data rate. The default mode is for the frequency to lock to the input data, where a finite state machine extracts frequency measurements from the data to program the DCO and loop division ratio so that the sampling frequency matches the data rate to within 250 ppm. The PLL is enabled, driving this frequency difference to 0 ppm. The second mode is to lock to the reference, in which case the user provides a reference clock between 11.05 MHz and 176.8 MHz. Division ratios must be written to a serial port register.

## FUNCTIONAL DESCRIPTION

### FREQUENCY ACQUISITION

The ADN2913 acquires its frequency from the input data over a range of data frequencies from 6.5 Mbps to 8.5 Gbps. The lock detector circuit compares the frequency of the DCO and the frequency of the incoming data. When these frequencies differ by more than 1000 ppm, the LOL pin is asserted and a new frequency acquisition cycle is initiated. The DCO frequency is reset to the lowest point of its range, and the internal division rate is set to its lowest value of  $N = 1$ , which is the highest octave of data rates. The frequency detector then compares this sampling rate frequency to the data rate frequency and either increases  $N$  by a factor of 2 if the sampling rate frequency is greater than the data rate frequency, or increases the DCO frequency if the data rate frequency is greater than the sampling rate frequency. Initially, the DCO frequency is incremented in large steps to aid fast acquisition. As the DCO frequency approaches the data frequency, the step size is reduced until the DCO frequency is within 250 ppm of the data frequency, at which point LOL is deasserted.

When LOL is deasserted, the frequency-locked loop is turned off. The PLL or DLL pulls in the DCO frequency until the DCO frequency equals the data frequency.

### LIMITING AMPLIFIER

The limiting amplifier has differential inputs (PIN and NIN) that are each internally terminated with  $50\ \Omega$  to an on-chip voltage reference ( $V_{CM} = 0.95\ \text{V}$  typically). The inputs must be ac-coupled. Input offset is factory trimmed to achieve better than 10 mV p-p typical sensitivity with minimal drift. The limiting amplifier can be driven differentially or single-ended. DC coupling of the limiting amplifier is not possible because the user must supply a common-mode voltage to exactly match the internal common-mode voltage; otherwise, the internal  $50\ \Omega$  termination resistors absorb the difference in common-mode voltages.

Another reason that the limiting amplifier cannot be dc-coupled is that the factory trimmed input offset becomes invalid. The offset is adjusted to zero by differential currents from the slice adjust DAC (see Figure 1). With ac coupling, all of the current goes to the  $50\ \Omega$  termination resistors on the ADN2913. However, with dc coupling, this current is shared with the external drive circuit, and calibration of the offset is lost. In addition, the slice adjust must have all the current from the slice adjust DAC go to the resistors; otherwise, the calibration is lost (see the Slice Adjust section).

### SLICE ADJUST

The quantizer slicing level can be offset by  $\pm 100\ \text{mV}$  in 1.6 mV steps or by  $\pm 15\ \text{mV}$  in 0.24 mV steps to mitigate the effect of amplified spontaneous emission (ASE) noise or duty cycle distortion. The quantizer slice adjust level is set by the slice[6:0] bits in Register 0x15.

Accurate control of the slice threshold requires the user to read back the factory trimmed offset, which is stored as a 7-bit number in the slice readback register (Register 0x73). Use Table 20 to decode the measured offset of the device, where an LSB corresponds to 0.24 mV.

**Table 20. Program Slice Level, Normal Slice Mode (Extended Slice = 0)**

Slice[6:0]	Decimal Value	Offset
0000000	0	Slice function disabled
0000001	1	-15 mV
...	...	...
1000000	64	0 mV
...	...	...
1111111	127	+14.75 mV

The amount of offset required for manual slice adjustment is determined by subtracting the offset of the device from the desired slice adjust level. Use Table 20 or Table 21 to determine the code word to be written to the slice register.

An extended slice with coarser granularity for each LSB step is found in Table 21. Setting the extended slice bit (Bit 7) = 1 in Register 0x15 scales the full-scale range of the slice adjust by a factor of 6.

**Table 21. Program Slice Level, Extended Slice Mode (Extended Slice = 1)**

Slice[6:0]	Decimal Value	Offset
0000000	128	Slice function disabled
0000001	129	-100 mV
...	...	...
1000000	192	0 mV
...	...	...
1111111	255	+100 mV

When manual slice is desired, disable the dc offset loop, which drives duty cycle distortion on the data to 0. Adaptive slice is disabled by setting ADAPTIVE\_SLICE\_EN = 0 in Register 0x13.

### EDGE SELECT

A binary, or Alexander phase, detector drives both the DLL and PLL at all division rates. Duty cycle distortion on the received data leads to a dead band in the phase detector transfer function if phase errors are measured on both rising and falling data transitions. This dead band leads to jitter generation of unknown spectral composition whose peak-to-peak amplitude is potentially large.

The recommended usage of the device when the dc offset loop is disabled is to compute phase errors exclusively on either the rising data edges with EDGE\_SEL[1:0] (Bits[D4:D3] in Register 0x10) = 1 (decimal) or falling data edges with EDGE\_SEL[1:0] = 2. The alignment of the clock to the rising data edges with EDGE\_SEL[1:0] = 1 is represented by the top two curves in Figure 24. Duty cycle distortion with narrow 1s moves the

significant sampling instance where data is sampled to the right of center. The alignment of the clock to the falling data edges with EDGE\_SEL[1:0] = 2 is represented by the first and third curves in Figure 24. The significant sampling instance moves to the left of center. Sample phase adjustment for rates above 5.65 Gbps can be used to move the significant sampling instance to the center of the narrow 1 (or narrow 0) for best jitter tolerance.

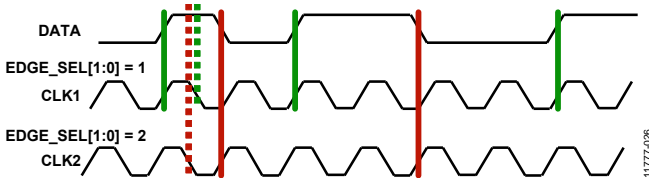


Figure 24. Phase Detector Timing

**DLL Slew**

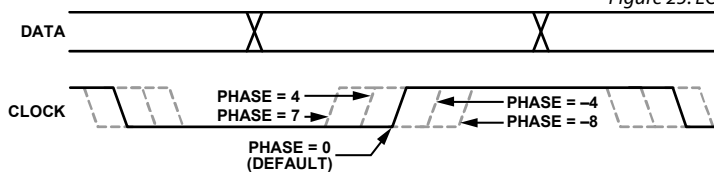
Jitter tolerance beyond the transfer bandwidth of the CDR is determined by the slew rate of the delay-locked loop implementing a delta modulator on phase. Setting DLL\_SLEW[1:0] = 2 (the default value) in Register 0x13, configures the DLL to track 0.75 UI p-p jitter at the highest frequency breakpoint in the SONET/SDH jitter tolerance mask. This frequency scales with the rate as  $f_{p4} = \text{Rate (Hz)}/2500$  (for example, 1.0 MHz for OC-48). Peak-to-peak tracking in UI at  $f_{p4}$  obeys the expression  $(1 + \text{DLL\_SLEW})/4$  UI p-p.

In some applications, full SONET/SDH jitter tolerance is not needed. In this case, DLL\_SLEW[1:0] can be set to 0, giving lower jitter generation on the recovered clock and better high frequency jitter tolerance.

**Sample Phase Adjustment**

The phase of the sampling instant can be adjusted using the I<sup>2</sup>C interface when the devices operate at data rates of 5.65 Gbps or higher by writing to SAMPLE\_PHASE[3:0] (Bits[D3:D0] in Register 0x14). This feature allows the user to adjust the sampling instant to improve the BER and jitter tolerance. Although the default sampling instant chosen by the CDR is sufficient in most applications, when dealing with degraded input signals, the BER and jitter tolerance performance can be improved by manually adjusting the phase.

A total adjustment range of 0.5 UI is available, with 0.25 UI in each direction, in increments of 1/32 UI. SAMPLE\_PHASE[3:0] is a two's complement number. The relationship between data and the sampling clock is shown in Figure 26.



NOTES  
1. PHASE REFERS TO SAMPLE\_PHASE[3:0]

Figure 26. Data vs. Sampling Clock

**Transfer Bandwidth**

The transfer bandwidth can be adjusted using the I<sup>2</sup>C interface by writing to TRANBW[2:0] in Register 0x10. The default value is 4. When set to values below 4, the transfer bandwidth is reduced. When set to values above 4, the transfer bandwidth is increased. The resulting transfer bandwidth is based on the following formula:

$$\text{Transfer BW} = (\text{Default Transfer BW}) \times \left( \frac{\text{TRANBW}[2:0]}{4} \right)$$

For example, at OC-48, the default transfer bandwidth is 650 kHz. The resulting transfer bandwidth when TRANBW[2:0] is changed is

- TRANBW[2:0] = 1: transfer BW = 162.5 kHz
- TRANBW[2:0] = 2: transfer BW = 325 kHz
- TRANBW[2:0] = 3: transfer BW = 487.5 kHz
- TRANBW[2:0] = 4: transfer BW = 650 kHz (default)
- TRANBW[2:0] = 5: transfer BW = 812.5 kHz
- TRANBW[2:0] = 6: transfer BW = 975 kHz
- TRANBW[2:0] = 7: transfer BW = 1137.5 kHz

Reducing the transfer bandwidth is commonly used in OTN applications. Never set TRANBW[2:0] = 0 because this value makes the CDR open loop. Also, note that setting TRANBW[2:0] to a value greater than 4 may cause a slight increase in jitter generation and potential jitter peaking.

**LOSS OF SIGNAL (LOS) DETECTOR**

The receiver front-end LOS detector circuit detects when the input signal level falls below a user adjustable threshold.

There is typically 6 dB of electrical hysteresis on the LOS detector to prevent chatter on the LOS pin. Therefore, if the input level falls below the programmed LOS threshold, causing the LOS pin to assert, the LOS pin is not deasserted until the input level increases to 6 dB (2x) above the LOS threshold (see Figure 25).

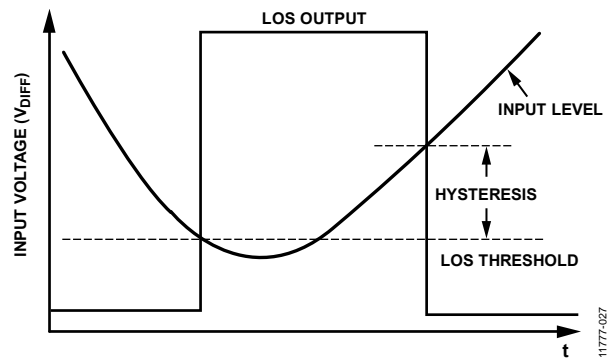


Figure 25. LOS Detector Hysteresis

The LOS detector and the slice level adjust can be used simultaneously on the ADN2913. Therefore, any offset added to the input signal by the slice[6:0] bits does not affect the LOS detector measurement of the absolute input level.

### LOS Power-Down

By default, the LOS detector is enabled and consumes power. The LOS detector is placed in a low power mode by setting LOS\_PDN = 1 (Bit D3 in Register 0x9).

### LOS Threshold

The LOS threshold has a range between 0 mV and 128 mV and is set by writing the number of millivolts (mV) to Register 0x36 followed by toggling the LOS\_ENABLE bit in Register 0x74 while LOS\_ADDRESS is set to 1. The following is a procedure for writing the LOS threshold:

1. Write 0x21 to LOS\_CTRL (Register 0x74).
2. Write the desired threshold in millivolts to LOS\_DATA (Register 0x36).
3. Write 0x31 to LOS\_CTRL (Register 0x74).
4. Write 0x21 to LOS\_CTRL (Register 0x74).

The LOS threshold can be set to a value between 0 mV and 63 mV in 1 mV steps and from 64 mV to 128 mV in 2 mV steps. In the lower range, all of the bits are active, giving 1 mV/LSB resolution, where Bit D0 is the LSB. In the upper range, Bit D0 is disabled (that is, D0 = 0), making Bit D1 the LSB and resulting in 2 mV/LSB resolution.

The LOS\_CTRL register contains the necessary address and write enable bits to program this LOS threshold.

### Signal Strength Measurement

The LOS detector measures and digitizes the peak-to-peak amplitude of the received signal. A single shot measurement is taken by writing the following sequence of bytes to LOS\_CTRL at Address 0x74: 0x7, 0x17, 0x7. When LOS\_ENABLE goes low, the peak-to-peak amplitude in millivolts is loaded into LOS\_DATA[7:0] (Register 0x36). The contents of LOS\_DATA change only when LOS\_ENABLE (Bit D4 in Register 0x74) is toggled low to high to low while LOS\_ADDRESS[2:0] (Bits[D2:D0] in Register 0x74) is set to 7.

### PASSIVE EQUALIZER

A passive equalizer is available at the input to equalize large signals that have undergone distortion due to PCB traces, vias, or connectors. The adaptive EQ functions only at data rates greater than 5.5 Gbps. Therefore, at rates less than 5.5 Gbps, the EQ must be manually set.

The equalizer can be manually set using the LA\_EQ register (Register 0x16). An adaptive loop is also available to optimize the EQ setting based on characteristics of the received eye at the phase detector. If the channel is known in advance, set the EQ manually to obtain the best performance; however, the adaptive EQ finds the best setting in most cases.

Table 22 lists typical EQ settings for several trace lengths. The values in Table 22 are based on measurements taken on a test board with simple FR-4 traces. Table 23 lists the typical maximum reach in inches of FR-4 of the EQ at several data rates. If a real channel includes lossy connectors or vias, the FR-4 reach length is lower. For any real-world system, it is highly recommended to test several EQ settings with the real channel to ensure the best signal integrity.

**Table 22. EQ Settings vs. Trace Length on FR-4**

Trace Length (Inches)	Typical EQ Setting
6	10
10	12
15	14
20 to 30	15

**Table 23. Typical EQ Reach on FR-4 vs. Maximum Data Rates Supported**

Maximum Data Rate (Gbps)	Typical EQ Reach on FR-4 (Inches)
4	30
8	20
10	15
11	10

### 0 dB EQ

The 0 dB EQ path connects the input signal directly to the digital logic inside the ADN2913. The 0 dB EQ is useful at lower data rates where the signal is large (therefore, the limiting amplifier is not needed and power can be saved by deselecting the limiting amplifier) and unimpaired (therefore, the equalizer is not needed). The signal swing of the internal digital circuit is 600 mV p-p differential, the minimum signal amplitude that must be provided as the input in 0 dB EQ mode.

In 0 dB EQ mode, the internal 50  $\Omega$  termination resistors can be configured in one of two ways, either floated or tied to  $V_{CC} = 1.2$  V (see Figure 31 and Table 27). By setting the RX\_TERM\_FLOAT bit (Bit D7 in Register 0x16) to 1, these 50  $\Omega$  termination resistors are floated internal to the ADN2913 (see Figure 35). By setting the RX\_TERM\_FLOAT bit to 0, these 50  $\Omega$  termination resistors are connected to  $V_{CC} = 1.2$  V (see Figure 36). In both termination cases, the user must ensure a valid common-mode voltage on the input.

When the termination is floated, the two 50  $\Omega$  resistors are a purely differential termination. The input must conform to the range of signals shown in Figure 33.

When the termination is connected to a 1.2 V VCC power supply (see Figure 36 and Figure 37), the common-mode voltage is created by the driver circuit and the 50  $\Omega$  resistors on the ADN2913. For example, the driver can be an open-drain switched current (see Figure 36), and the 50  $\Omega$  resistors return this current to  $V_{CC}$ . In Figure 36, the common-mode voltage is created by both the current and the resistors. In this case, ensure that the current is a minimum of 6 mA, which gives a single-ended swing of 300 mV or a differential swing of 600 mV p-p

differential, with  $V_{CM} = 1.05\text{ V}$  (see Figure 33). The maximum current is 10 mA, which gives a single-ended 500 mV swing and differential 1.0 V p-p, with  $V_{CM} = 0.95\text{ V}$  (see Figure 34).

Another possibility is to back terminate the switched current driver, as shown in Figure 37, with the two VCC supplies having the same potential. In this example, the current is returned to VCC by the two 50  $\Omega$  resistors in parallel, or 25  $\Omega$ , so that the minimum current is 12 mA and the maximum current is 20 mA.

## LOCK DETECTOR OPERATION

The lock detector on the ADN2913 has three modes of operation: normal mode, LOL detector operation using a reference clock (LTR mode), and static LOL mode.

### Normal Mode

In normal mode, the ADN2913 is a continuous rate CDR that locks onto any data rate from 6.5 Mbps to 8.5 Gbps without the use of a reference clock as an acquisition aid. In this mode, the lock detector monitors the frequency difference between the DCO and the input data frequency, and deasserts the loss of lock signal, which appears on LOL, Pin 6, when the DCO is within 250 ppm of the data frequency. This enables the digital PLL (D/PLL), which pulls the DCO frequency in the remaining amount and acquires phase lock. If the input frequency error exceeds 1000 ppm (0.1%), the loss of lock signal is reasserted and control returns to the frequency loop, which begins a new frequency acquisition. The LOL pin remains asserted until the DCO locks onto a valid input data stream to within 250 ppm frequency error. This hysteresis is shown in Figure 27.

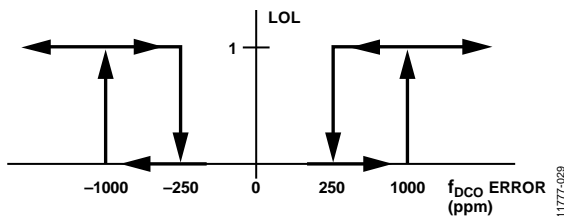


Figure 27. Transfer Function of LOL

### LOL Detector Operation Using a Reference Clock (LTR Mode)

In lock to reference (LTR) mode, a reference clock is used as an acquisition aid to lock the ADN2913 DCO. LTR mode is enabled by setting  $CDR\_MODE[2:0]$  to 3 (Bits[D6:D4] in Register 0x8). The user must also write to  $FREF\_RANGE[1:0]$  and  $DATA\_TO\_REF\_RATIO[3:0]$  (Bits[D5:D0] in Register 0xF) to set the reference frequency range and the divide ratio of the data rate with respect to the reference frequency. Finally, the reference clock power-down to the reference clock buffer must be deasserted by writing a 0 to  $REFCLK\_PDN$  (Bit D2 in Register 0xA). To maintain fastest acquisition, keep Bit D0 in Register 0xA set to 1.

For more information, see the Reference Clock (Optional) section. In LTR mode, the lock detector monitors the difference in frequency between the divided down DCO and the divided down reference clock. The loss of lock signal, which appears on LOL (Pin 6), is deasserted when the DCO is within 250 ppm of the desired frequency. This enables the D/PLL, which pulls in the DCO frequency by the remaining amount with respect to the input data and acquires phase lock. If the frequency error exceeds 1000 ppm (0.1%), the loss of lock signal is reasserted and control returns to the frequency loop, which reacquires lock with respect to the reference clock. The LOL pin remains asserted until the DCO frequency is within 250 ppm of the desired frequency. This hysteresis is shown in Figure 27.

### Static LOL Mode

The ADN2913 implements a static LOL feature that indicates whether a loss of lock condition has occurred and remains asserted, even if the ADN2913 regains lock, until the static LOL bit (Bit D2 in Register 0x6) is manually reset. If a loss of lock condition occurs, this bit is internally asserted to logic high. The static LOL bit remains high even after the ADN2913 reacquires lock to a new data rate. This bit can be reset by writing 1, followed by 0, to the reset static LOL bit (Bit D2 in Register 0x8). When reset, the static LOL bit remains deasserted until another loss of lock condition occurs.

Writing a 1 to  $LOL\_CONFIG$  (Bit D4 in Register 0x9) causes the LOL pin, Pin 6, to become a static LOL indicator. In this mode, the LOL pin mirrors the contents of the static LOL bit (Bit D2 in Register 0x6) and has the functionality described previously. The  $LOL\_CONFIG$  bit (Bit D4 in Register 0x9) defaults to 0. In this mode, the LOL pin operates in the normal operating mode; that is, it is asserted only when the ADN2913 is in acquisition mode and is deasserted when the ADN2913 has reacquired lock.

## HARMONIC DETECTOR

The ADN2913 provides a harmonic detector that detects whether the input data has changed to a lower harmonic of the data rate than the one that the sampling clock is currently locked onto. For example, if the input data instantaneously changes from OC-12, 622.08 Mbps, to an OC-3, 155.52 Mbps bit stream, this change can be perceived as a valid OC-12 bit stream because the OC-3 data pattern is exactly 4 $\times$  slower than the OC-12 pattern. Therefore, if the change in data rate is instantaneous, a 101 pattern at OC-3 is perceived by the ADN2913 as a 111100001111 pattern at OC-12. If the change to a lower harmonic is instantaneous, a typical inferior CDR may remain locked at the higher data rate.

The ADN2913 implements a harmonic detector that automatically identifies whether the input data has switched to a lower harmonic of the data rate than the DCO is currently locked onto. When a new harmonic is identified, the LOL pin is asserted, and a new frequency acquisition is initiated. The ADN2913 automatically locks onto the new data rate, and the LOL pin is deasserted.



The time to detect a lock to harmonic is

$$2^{16} \times (T_d/\rho)$$

where:

$1/T_d$  is the new data rate. For example, if the data rate is switched from OC-12 to OC-3, then  $T_d = 1/155.52$  MHz.

$\rho$  is the data transition density. Most coding schemes seek to ensure that  $\rho = 0.5$ , for example, PRBS and 8B/10B.

When the ADN2913 is placed in lock to reference mode, the harmonic detector is disabled.

## OUTPUT DISABLE AND SQUELCH

The ADN2913 has two types of output disable/squelch. The DATOUTP/DATOUTN and CLKOUTP/CLKOUTN outputs can be disabled by setting DATOUT\_DISABLE and CLKOUT\_DISABLE (Bits[D4:D3] in Register 0x1E) high, respectively. When an output is disabled, it is fully powered down, saving approximately 30 mW per output. Disabling DATOUTP/DATOUTN also disables the CLKOUTP/CLKOUTN outputs, saving a total of about 60 mW of power.

If it is desired to set the data output while leaving the clock on, the output data can be squelched by setting the data squelch bit (Bit D5 in Register 0x1E) high. In this mode, the data driver remains powered, but the data itself is forced to a value of 0 (or 1, depending on the setting of DATA\_POLARITY (Bit D1 in Register 0x1E)).

## I<sup>2</sup>C INTERFACE

The ADN2913 supports a 2-wire, I<sup>2</sup>C-compatible serial bus driving multiple peripherals. Two inputs, serial data (SDA) and serial clock (SCK), carry information between any devices connected to the bus. Each slave device is recognized by a unique address. The slave address consists of the seven MSBs of an 8-bit word. The upper six bits (Bits[6:1]) of the 7-bit slave address are factory programmed to 100000. The LSB of the slave address (Bit 0) is set by Pin 22, I<sup>2</sup>C\_ADDR. The LSB of the word specifies either a read or write operation (see Figure 18). Logic 1 corresponds to a read operation, whereas Logic 0 corresponds to a write operation.

To control the device on the bus, the following protocol must be used:

1. The master initiates a data transfer by establishing a start condition, defined as a high to low transition on SDA while SCK remains high. This indicates that an address/data stream follows.
2. All peripherals respond to the start condition and shift the next eight bits (the 7-bit address and the R/W bit). The bits are transferred from MSB to LSB.
3. The peripheral that recognizes the transmitted address responds by pulling the data line low during the ninth clock pulse. This is known as an acknowledge bit.
4. All other devices withdraw from the bus at this point and maintain an idle condition. In the idle condition, the

device monitors the SDA and SCK lines waiting for the start condition and the correct transmitted address.

The R/W bit determines the direction of the data. Logic 0 on the LSB of the first byte means that the master writes information to the peripheral. Logic 1 on the LSB of the first byte means that the master reads information from the peripheral.

The ADN2913 acts as a standard slave device on the bus. The data on the SDA pin is eight bits long, supporting the 7-bit addresses plus the R/W bit. The ADN2913 has subaddresses to enable the user accessible internal registers (see Table 7).

The ADN2913, therefore, interprets the first byte as the device address and the second byte as the starting subaddress. Auto-increment mode is supported, allowing data to be read from or written to the starting subaddress and each subsequent address without manually addressing the subsequent subaddress. A data transfer is always terminated by a stop condition. The user can also access any unique subaddress register on a one-by-one basis without updating all registers.

Stop and start conditions can be detected at any stage of the data transfer. If these conditions are asserted out of sequence with normal read and write operations, they cause an immediate jump to the idle condition. During a given SCK high period, issue one start condition, one stop condition, or a single stop condition followed by a single start condition. If an invalid subaddress is issued by the user, the ADN2913 does not issue an acknowledge and returns to the idle condition. If the user exceeds the highest subaddress while reading back in auto-increment mode, the highest subaddress register contents continue to be output until the master device issues a no acknowledge. This indicates the end of a read. In a no acknowledge condition, the SDA line is not pulled low on the ninth pulse. See Figure 20 and Figure 19 for sample read and write data transfers, respectively, and Figure 21 for a more detailed timing diagram.

## REFERENCE CLOCK (OPTIONAL)

A reference clock is not required to perform clock and data recovery with the ADN2913. However, support for an optional reference clock is provided. The reference clock can be driven differentially or single-ended. If the reference clock is not used, float both the REFCLKP and REFCLKN pins.

Two 50  $\Omega$  series resistors present a differential load between REFCLKP and REFCLKN. Common mode is internally set to  $0.56 \times V_{CC}$  by a resistor divider between VCC and VEE. See Figure 28, Figure 29, and Figure 30 for sample configurations.

The reference clock input buffer accepts any differential signal with a peak-to-peak differential amplitude of greater than 100 mV. The phase noise and duty cycle of the reference clock are not critical, and 100 ppm accuracy is sufficient.

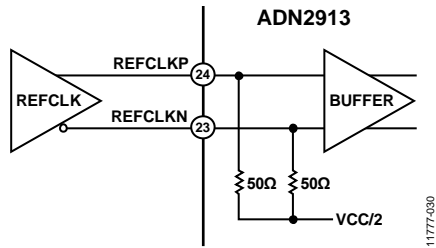


Figure 28. DC-Coupled, Differential REFCLKx Configuration

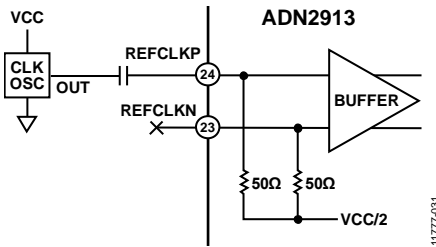


Figure 29. AC-Coupled, Single-Ended REFCLKx Configuration

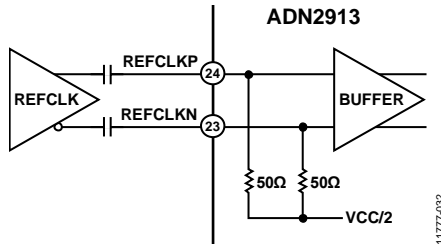


Figure 30. AC-Coupled, Differential REFCLKx Configuration

The reference clock can be used either as an acquisition aid for the ADN2913 to lock onto data, or to measure the frequency of the incoming data to within 0.01%. The modes are mutually exclusive because, in the first use, the user can force the device to lock onto only a known data rate; in the second use, the user can measure an unknown data rate.

Lock to reference mode is enabled by writing a 3 to CDR\_MODE[2:0] (Bits[D6:D4] in Register 0x8). An on-chip clock buffer must be powered on by writing a 0 to REFCLK\_PDWN (Bit D2 in Register 0xA). Fine data rate readback mode is enabled by writing a 1 to RATE\_MEAS\_EN (Bit D1 in Register 0x8). Enabling lock to reference and data rate readback at the same time causes an indeterminate state and is not supported.

### Using the Reference Clock to Lock Onto Data

In LTR mode, the ADN2913 locks onto a frequency derived from the reference clock according to the following equation:

$$\text{Data Rate}/2^{L\text{TR\_MODE}[3:0]-1} = \text{REFCLK}/2^{L\text{TR\_MODE}[5:4]}$$

The user must know exactly what the data rate is and provide a reference clock that is a function of this rate. The ADN2913 can still be used as a continuous rate device in this configuration if the user can provide a reference clock that has a variable frequency (see the AN-632 Application Note).

The reference clock can have a frequency from 11.05 MHz to 176.8 MHz. By default, the ADN2913 expects a reference clock of between 11.05 MHz and 22.1 MHz. If the reference clock is between 22.1 MHz and 44.2 MHz, 44.2 MHz and 88.4 MHz, or

between 88.4 MHz and 176.8 MHz, the user must configure the ADN2913 to use the correct reference frequency range by setting the two bits of FREQ\_RANGE[1:0] (Bits[5:4] in Register 0xF).

Table 24. LTR\_MODE Register Settings

FREQ_RANGE[1:0] Bits	Range (MHz)	DATA_TO_REF_RATIO[3:0] Bits	Ratio
00	11.05 to 22.1	0000	$2^{-1}$
01	22.1 to 44.2	0001	$2^0$
10	44.2 to 88.4	N	$2^{n-1}$
11	88.4 to 176.8	1010	$2^9$

The user can specify a fixed integer multiple of the reference clock to lock onto using DATA\_TO\_REF\_RATIO[3:0] (Bits[D3:D0] in Register 0xF), as follows:

$$\text{DATA\_TO\_REF\_RATIO}[3:0] = \text{data rate} \div \text{DIV}_{f_{\text{REF}}}$$

where  $\text{DIV}_{f_{\text{REF}}}$  represents the divided down reference referred to the 11.05 MHz to 22.1 MHz band.

For example, if the reference clock frequency is 38.88 MHz and the input data rate is 622.08 Mbps, then FREQ\_RANGE[1:0] is set to 01 to give a divided down reference clock of 19.44 MHz. DATA\_TO\_REF\_RATIO[3:0] is set to 0110, that is, 6, because

$$622.08 \text{ Mbps}/19.44 \text{ MHz} = 2^{(6-1)}$$

If the ADN2913 is operating in lock to reference mode and the user changes the reference frequency, that is, the  $f_{\text{REF}}$  range or the  $f_{\text{REF}}$  ratio (Bits[D5:D4] or Bits[D3:D0], respectively, in Register 0xF), this change must be followed by writing a low to high to low transition to the INIT\_FREQ\_ACQ bit (Bit D6 in Register 0x9) to initiate a new lock to reference command.

By default in lock to reference clock mode, when lock has been achieved and the ADN2913 is in tracking mode, the frequency of the DCO is compared to the frequency of the reference clock. If this frequency error exceeds 1000 ppm, lock is lost, LOL is asserted, and the device relocks to the reference clock while continuing to output a stable clock.

An alternative configuration is enabled by setting LOL data (Bit D6 in Register 0xF) = 1. In this configuration, when the device is in tracking mode, the frequency of the DCO is compared to the frequency of the input data rather than the frequency of the reference clock. If the frequency error exceeds 1000 ppm, lock is lost, LOL is asserted, and the device relocks to the reference clock while continuing to output a stable clock.

### Using the Reference Clock to Measure Data Frequency

The user can also provide a reference clock to measure the recovered data frequency. In this case, the ADN2913 compares the frequency of the incoming data to the incoming reference clock and returns a ratio of the two frequencies to 0.01% (100 ppm). The accuracy error of the reference clock is added to the accuracy error of the ADN2913 data rate measurement. For example, if a 100 ppm accuracy reference clock is used, the total accuracy of the measurement is 200 ppm.

The reference clock can range from 11.05 MHz to 176.8 MHz. Prior to reading back the data rate using the reference clock, the FREF\_RANGE[1:0] bits (Bits[D5:D4] in Register 0xF) must be set to the appropriate frequency range with respect to the reference clock being used, according to Table 24. A fine data rate readback is then executed as follows:

1. Apply the reference clock.
2. Write a 0 to REFCLK\_PDN (Bit D2 in Register 0xA) to enable the reference clock circuit.
3. Write to FREF\_RANGE[1:0] (Bits[D5:D4] in Register 0xF) to select the appropriate reference clock frequency circuit.
4. Write a 1 to RATE\_MEAS\_EN (Bit D1 in Register 0x8). This enables the fine data rate measurement capability of the ADN2913. This bit is level sensitive and does not need to be reset to perform subsequent frequency measurements.
5. Write a low to high to low transition to RATE\_MEAS\_RESET (Bit D0 in Register 0x8). This initiates a new data rate measurement.
6. Read back RATE\_MEAS\_COMP (Bit D0 in Register 0x6). If the bit is 0, the measurement is not complete. If it is 1, the measurement is complete and the data rate can be read back on RATE\_FREQ[23:0], FULLRATE, and DIVRATE[3:0] (see Table 7). The approximate time for a data rate measurement is given in Equation 2.

Use the following equation to determine the data rate:

$$f_{\text{Datarate}} = \frac{(\text{RATE\_FREQ}[23:0]) \times f_{\text{REFCLK}}}{2^{\text{LTR}[5:4]} \times 2^7 \times 2^{\text{FULLRATE}} \times 2^{\text{DIVRATE}[3:0]}} \quad (1)$$

where:

$f_{\text{Datarate}}$  is the data rate (Mbps).  
 $\text{RATE\_FREQ}[23:0]$  is from FREQ2[7:0] (most significant byte), FREQ1[7:0], and FREQ0[7:0] (least significant byte). See Table 7.  
 $f_{\text{REFCLK}}$  is the reference clock frequency (MHz).  
 $\text{FULLRATE} = \text{FREQ\_RB2}[6]$ .  
 $\text{DIVRATE}[3:0] = \text{FREQ\_RB2}[5:2]$ .

MSB		LSB
D23 to D16	D15 to D8	D7 to D0
FREQ2[7:0]	FREQ1[7:0]	FREQ0[7:0]

Consider the example of a 1.25 Gbps (GbE) input signal and a reference clock source of 32 MHz at the PIN/NIN and REFCLKP/REFCLKN ports, respectively. In this case, FREF\_RANGE[1:0] (Bits[D5:D4] in Register 0xF) = 01 and the reference frequency falls into the range of 22.1 MHz to 44.2 MHz. After following Step 1 through Step 6, the readback value of RATE\_FREQ[23:0] is 0x13880, which is equal to  $8 \times 10^4$ . The readback value of FULLRATE (Bit D6 in Register 0x5) is 1, and the readback value of DIVRATE[3:0] (Bits[D5:D2] in Register 0x5) is 2. Inserting these values into Equation 1 yields

$$((8 \times 10^4) \times (32 \times 10^6)) / (2^1 \times 2^7 \times 2^1 \times 2^2) = 1.25 \text{ Gbps}$$

If subsequent frequency measurements are required, keep RATE\_MEAS\_EN (Bit D1 in Register 0x8) set to 1. It does not need to be reset. The measurement process is reset by writing a 1 followed by a 0 to RATE\_MEAS\_RESET (Bit D0 in Register 0x8).

This initiates a new data rate measurement. Follow Step 2 through Step 6 to read back the new data rate. Note that a data rate readback is valid only if the LOL pin is low. If LOL is high, the data rate readback is invalid.

Initiating a frequency measurement by writing a low to high to low transition into RATE\_MEAS\_RESET (Bit D0 in Register 0x8) also resets the RATE\_MEAS\_COMP bit (Bit D0 in Register 0x6). The approximate time to complete a frequency measurement from RATE\_MEAS\_RESET being written with a low to high to low transition to when the RATE\_MEAS\_COMP bit returns high is given by

$$\text{Measurement Time} = \frac{2^{11} \times 2^{\text{LTR}[5:4]}}{f_{\text{REFCLK}}} \quad (2)$$

### LOS Configuration

The LOS detector output, LOS (Pin 5), can be configured to be either active high or active low. If LOS polarity (Bit 2 in Register 0x9) is set to Logic 0 (default), the LOS pin is active high when a loss of signal condition is detected.

### ADDITIONAL FEATURES AVAILABLE VIA THE I<sup>2</sup>C INTERFACE

#### Coarse Data Rate Readback

The data rate can be read back over the I<sup>2</sup>C interface to approximately  $\pm 5\%$  without using an external reference clock according to the following formula:

$$\text{Data} = \frac{f_{\text{DCO}}}{2^{\text{FULLRATE}} \times 2^{\text{DIVRATE}[3:0]}} \quad (3)$$

where

$\text{FULLRATE} = \text{FREQ\_RB2}[6]$ .  
 $\text{DIVRATE}[3:0] = \text{FREQ\_RB2}[5:2]$ .  
 $f_{\text{DCO}}$  is the frequency of the DCO, derived as shown in Table 25.

Four oscillator cores defined by VCOSSEL[9:8] (Bits[D1:D0] in Register 0x5) span the highest octave of data rates according to Table 25.

Table 25. DCO Center Frequency vs. VCOSSEL[9:8]

Core = VCOSSEL[9:8]	Min Frequency (MHz) = Min_f(core)	Max Frequency (MHz) = Max_f(core)
0	5570	7105
1	7000	8685
2	8610	10,330
3	10,265	11,625

$f_{\text{DCO}}$  is determined from VCOSSEL[9:0] (Bits[D7:D0] in Register 0x4, and Bits[D1:D0] in Register 0x5), according to the following formula:

$$f_{\text{DCO}} = \text{Min}_f(\text{core}) + \frac{\text{Max}_f(\text{core}) - \text{Min}_f(\text{core})}{256} \times \text{VCOSSEL}[7:0]$$

**Worked Example**

Read back the contents of the `FREQ_RB1` and `FREQ_RB2` registers. For example, with an OC-48 signal presented to the PIN/NIN ports,

```
FREQ_RB1 = 0xCD
FREQ_RB2 = 0x46
FULLRATE (FREQ_RB2[6]) = 1
DIVRATE[3:0] (FREQ_RB2[5:2]) = 1
VCOSEL[9:8] core (FREQ_RB2[1:0]) = 2
```

Then

$$f_{DCO} = 8610 \text{ Mbps} + \frac{10330 \text{ Mbps} - 8610 \text{ Mbps}}{256} \times 205 = 9987.34 \text{ Mbps}$$

and

$$f_{data} = \frac{9987.34 \text{ Mbps}}{2^1 \times 2^1} = 2496.84 \text{ Mbps}$$

**Initiate Frequency Acquisition**

A frequency acquisition can be initiated by writing a 1 followed by a 0 to `INIT_FREQ_ACQ` (Bit D6 in Register 0x9). This initiates a new frequency acquisition while keeping the [ADN2913](#) in the operating mode that was previously programmed in Register 0x8, Register 0x9, and Register 0xA (`CTRLA`, `CTRLB`, and `CTRLC` registers, respectively).

**PRBS Generator/Receiver**

The [ADN2913](#) has an integrated PRBS generator and detector for system testing purposes. The device is configurable as either a PRBS detector or a PRBS generator.

The following steps configure the PRBS detector:

1. Set `DATA_RECEIVER_ENABLE` (Bit D2 in Register 0x3F) to 1 and set `DATA_RECEIVER_MODE[1:0]` (Bits[D1:D0] in Register 0x3F) according to the desired PRBS pattern (0 = PRBS7; 1 = PRBS15; 2 = PRBS31). Setting `DATA_RECEIVER_MODE[1:0]` to 3 leads to a one-shot sampling of recovered data into `DATA_LOADED[15:0]`.
2. Set `DATA_RECEIVER_CLEAR` (Bit D3 in Register 0x3F) to 1 followed by 0 to clear `PRBS_ERROR` and `PRBS_ERROR_COUNT[7:0]`.
3. States of `PRBS_ERROR` (Bit D1 in Register 0x41) and `PRBS_ERROR_COUNT[7:0]` (Bits[D7:D0] in Register 0x40) can be frozen by setting `DATA_RECEIVER_ENABLE` (Bit D2 in Register 0x3F) to 0.

The following steps configure the PRBS generator:

1. Set `DATA_GEN_EN` (Bit D2 in Register 0x39) = 1 to enable the PRBS generator and set `DATA_GEN_MODE[1:0]` (Bits[D1:D0] in Register 0x39) for the desired PRBS output pattern (0 = PRBS7; 1 = PRBS15; 2 = PRBS31). An arbitrary 32-bit pattern stored as `PROG_DATA[31:0]` is activated by setting `DATA_GEN_MODE[1:0]` to 3.
2. Strings of consecutive identical digits (CIDs) of sensed `DATA_CID_BIT` (Bit D5 in Register 0x39) can be introduced in the generator by setting `DATA_CID_EN` (Bit D4 in Register 0x39) set to 1. The length of CIDs is  $8 \times \text{DATA\_CID\_LENGTH}$ , which is set via Bits[D7:D0] in Register 0x3A.

**Table 26. PRBS Settings**

PRBS Pattern	DATA_GEN_MODE[1:0]	PRBS Polynomial
PRBS7	00	$1 + X^6 + X^7$
PRBS15	01	$1 + X^{14} + X^{15}$
PRBS31	10	$1 + X^{28} + X^{31}$
PROG_DATA[31:0]	11	N/A

**Double Data Rate Mode**

The default output clock mode is a double data rate (DDR) clock, where the output clock frequency is one-half the data rate. DDR mode allows direct interfacing to FPGAs that support clocking on both rising and falling edges. Setting `DDR_DISABLE` (Bit D2 in Register 0x1E) = 1 enables full data rate mode. Full data rate mode is not supported for data rates in the highest octave between 5.6 Gbps and 8.5 Gbps.

**CDR Bypass Mode**

The CDR in the [ADN2913](#) can be bypassed by setting the CDR bypass bit (Bit D5 in Register 0x9) = 1. In this mode, the [ADN2913](#) feeds the input directly through the input amplifiers to the output buffer, bypassing the CDR. The CDR bypass path is intended for use in testing or debugging a system. Use the CDR bypass path at data rates at or below 3.0 Gbps only.

**Disable Output Buffers**

The [ADN2913](#) provides the option of disabling the output buffers for power savings. The clock output buffer can be disabled by setting Bit `CLKOUT_DISABLE` (Bit D3 in Register 0x1E) = 1. This setting reduces the total output power by 30 mW. For a total of 60 mW of power savings, such as in a low power standby mode, both the `CLKOUT` and `DATOUT` buffers can be disabled together by setting `DATOUT_DISABLE` (Bit D4 in Register 0x1E) = 1.

**INPUT CONFIGURATIONS**

The ADN2913 input stage can work with the signal source in an ac-coupled or dc-coupled configuration. The ADN2913 supports one of the following input modes: limiting amplifier, equalizer, or 0 dB EQ. The ADN2913 can be configured to use any required input configuration through the I<sup>2</sup>C bus. Figure 31 shows a block diagram of the input stage circuit.

The input signal path is configurable with the INPUT\_SEL[1:0] bits (Bits[D6:D5] in Register 0x16). Table 27 shows the INPUT\_SEL[1:0] bits and the input signal configuration.

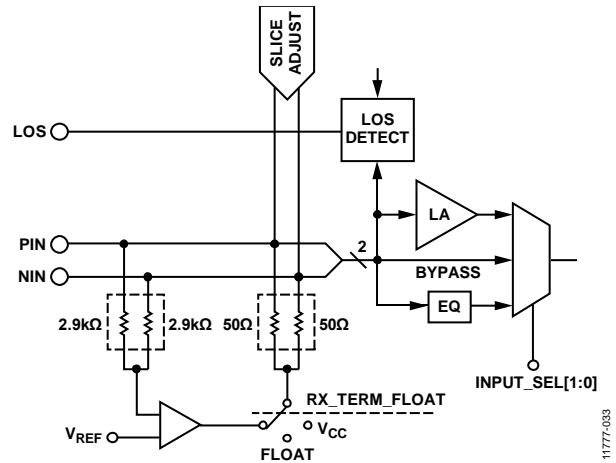


Figure 31. Input Stage Block Diagram

Table 27. Input Signal Configuration

Selected Input	INPUT_SEL[1:0]	RX_TERM_FLOAT = 0	RX_TERM_FLOAT = 1
Limiting Amplifier	00	V <sub>REF</sub>	Not defined
Equalizer	01	V <sub>REF</sub>	Not defined
0 dB EQ (0 dB Buffer)	10	V <sub>CC</sub>	Float
Not Defined	11	Not defined	Not defined

**Choosing AC Coupling Capacitors**

AC coupling capacitors at the inputs (PIN, NIN) and outputs (DATOUTP, DATOUTN) of the ADN2913 must be chosen such that the device works properly over the full range of data rates used in the application. When choosing the capacitors, the time constant formed with the two 50 Ω resistors in the signal path must be considered. When a large number of consecutive identical digits (CIDs) are applied, the capacitor voltage can droop due to baseline wander (see Figure 32), causing pattern dependent jitter (PDJ).

The user must determine how much droop is acceptable and choose an ac coupling capacitor based on that amount of droop. The amount of PDJ can then be approximated based on the capacitor selection. The actual capacitor value selection may require some trade-offs between droop and PDJ.

For example, assuming that 2% droop is acceptable, the maximum differential droop is 4%.

Normalizing to V p-p,

$$Droop = \Delta V = 0.04 V = 0.5 V \text{ p-p} (1 - e^{-t/\tau})$$

Therefore,

$$\tau = 12t$$

where:

$\tau$  is the RC time constant (C is the ac coupling capacitor, R = 100 Ω seen by C).

t is the total discharge time.

$$t = nT$$

where:

n is the number of CIDs.

T is the bit period.

Calculate the capacitor value by combining the equations for  $\tau$  and t.

$$C = 12nT/R$$

When the capacitor value is selected, the PDJ can be approximated as

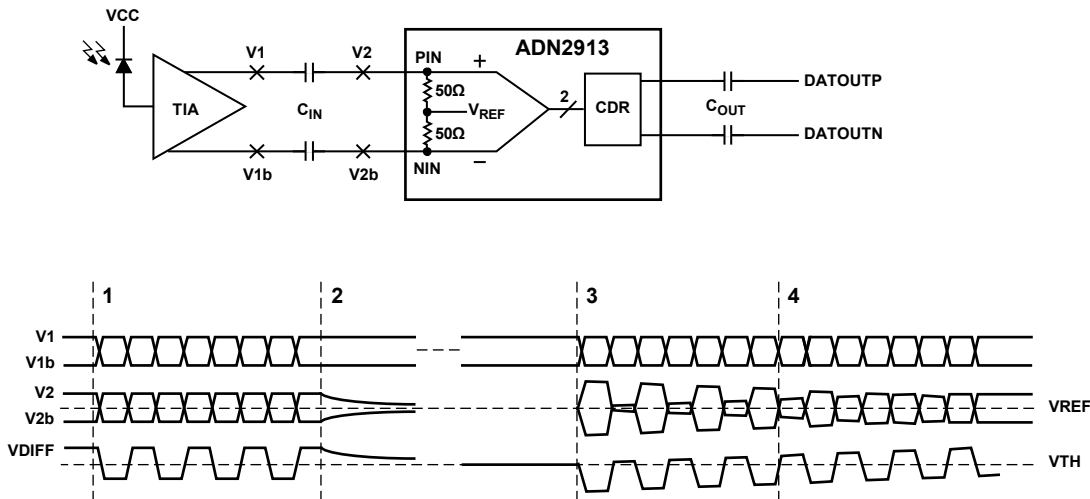
$$PDJ_{ps \text{ p-p}} = 0.5t_R(1 - e^{(-nT/RC)})/0.6$$

where:

$PDJ_{ps \text{ p-p}}$  is the amount of pattern dependent jitter allowed, <0.01 UI p-p typical.

$t_R$  is the rise time, which is equal to  $0.22/BW$ ;  $BW \approx 0.7$  (bit rate).

Note that this expression for  $t_R$  is accurate only for the inputs. The output rise time for the ADN2913 is ~30 ps regardless of data rate.



$V_{DIFF} = V2 - V2b$   
 $V_{TH} = \text{ADN2913 QUANTIZER THRESHOLD}$

**NOTES**

1. DURING THE DATA PATTERNS WITH HIGH TRANSITION DENSITY, THE DIFFERENTIAL DC VOLTAGE AT V1 AND V2 IS ZERO.
2. WHEN THE TIA OUTPUTS CONSECUTIVE IDENTICAL DIGITS, V1 AND V1b ARE DRIVEN TO DIFFERENT DC LEVELS. V2 AND V2b DISCHARGE TO THE  $V_{REF}$  LEVEL, WHICH EFFECTIVELY INTRODUCES A DIFFERENTIAL DC OFFSET ACROSS THE AC COUPLING CAPACITORS.
3. WHEN THE BURST OF DATA STARTS AGAIN, THE DIFFERENTIAL DC OFFSET ACROSS THE AC COUPLING CAPACITORS IS APPLIED TO THE INPUT LEVELS, CAUSING A DC SHIFT IN THE DIFFERENTIAL INPUT. THIS SHIFT IS LARGE ENOUGH SUCH THAT ONE OF THE STATES, EITHER HIGH OR LOW, DEPENDING ON THE LEVELS OF V1 AND V1b WHEN THE TIA BEGAN DETECTING AND OUTPUTTING A CID DATA SYSTEM, IS CANCELLED OUT. THE QUANTIZER DOES NOT RECOGNIZE THIS AS A VALID STATE.
4. THE DC OFFSET SLOWLY DISCHARGES UNTIL THE DIFFERENTIAL INPUT VOLTAGE EXCEEDS THE SENSITIVITY OF THE ADN2913. THE QUANTIZER RECOGNIZES BOTH HIGH AND LOW STATES AT THIS POINT.

Figure 32. Example of Baseline Wander

**DC-Coupled Application**

The inputs to the ADN2913 can also be dc-coupled. This can be necessary in burst mode applications with long periods of CIDs and where baseline wander cannot be tolerated. If the inputs to the ADN2913 are dc-coupled, care must be taken not to violate the input range and common-mode level requirements of the ADN2913 (see Figure 33 and Figure 34). If dc coupling is required, and the output levels of the transimpedance amplifier (TIA) do not adhere to the levels shown in Figure 33 and Figure 34, level shifting and/or attenuation must occur between the TIA outputs and the ADN2913 inputs.

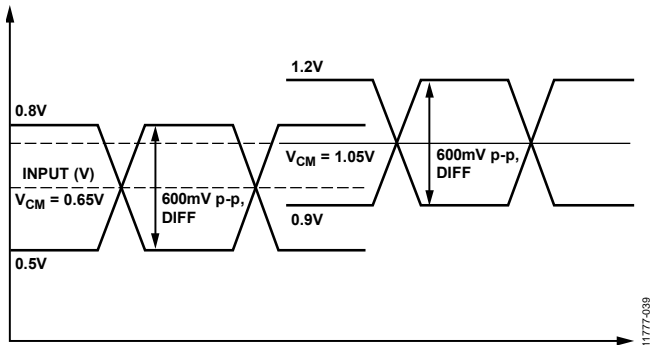


Figure 33. Minimum Allowed DC-Coupled Input Levels

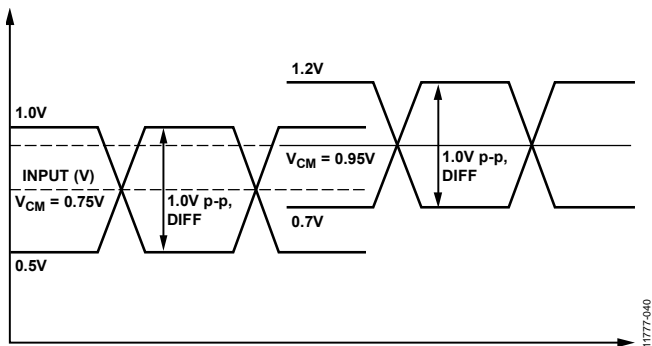


Figure 34. Maximum Allowed DC-Coupled Input Levels

Figure 35 shows 0 dB EQ mode with 50 Ω termination resistors floated internal to the ADN2913.

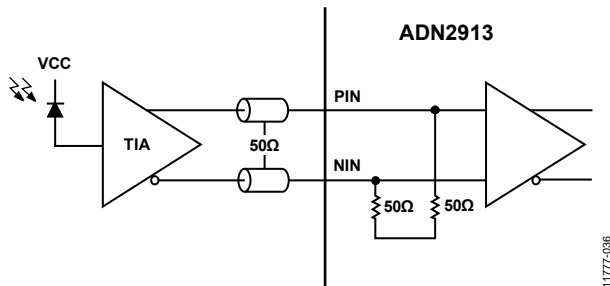


Figure 35. DC-Coupled Application, 0 dB EQ Input (Rx Termination Float Mode)

Figure 36 shows the default dc-coupled configuration when using the 0 dB EQ input. The two terminations are connected directly to VCC in a normal CML fashion, giving a common mode that is set by the dc signal strength from the driving chip. The 0 dB EQ input has a high common-mode range and can tolerate  $V_{CM}$  up to and including VCC.

Figure 37 shows back terminated mode. The switched current driver is back terminated, with the two VCC supplies having the same potential.

See the 0 dB EQ section for more information.

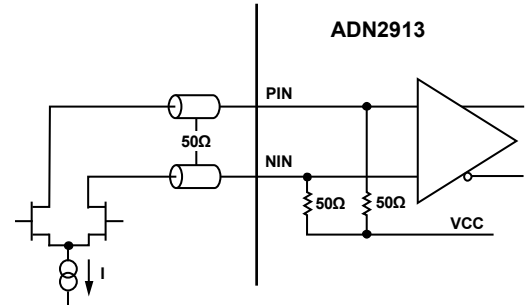


Figure 36. DC-Coupled Application, 0 dB EQ Input (Normal Mode)

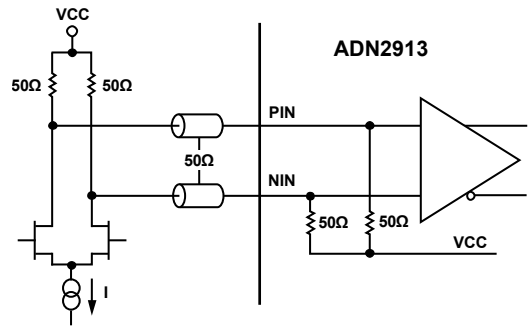


Figure 37. DC-Coupled Application, 0 dB EQ Input (Back Terminated Mode)

## APPLICATIONS INFORMATION

### TRANSMISSION LINES

Use of 50  $\Omega$  transmission lines is required for all high frequency input and output signals to minimize reflections: PIN, NIN, CLKOUTP, CLKOUTN, DATOUTP, and DATOUTN (also REFCLKP and REFCLKN, if using a high frequency reference clock, such as 155 MHz). The PIN and NIN input traces must be matched in length, and the CLKOUTP, CLKOUTN, DATOUTP, and DATOUTN output traces must match in length to avoid skew between the differential traces.

The high speed inputs (PIN and NIN) are each internally terminated with 50  $\Omega$  to an internal reference voltage (see Figure 31). As with any high speed, mixed-signal circuit, take care to keep all high speed digital traces away from sensitive analog nodes.

The high speed outputs (DATOUTP, DATOUTN, CLKOUTP, and CLKOUTN) are internally terminated with 50  $\Omega$  to VCC.

### SOLDERING GUIDELINES FOR LEAD FRAME CHIP SCALE PACKAGE

The lands on the 24-lead LFCSP are rectangular. The printed circuit board pad for the lands is 0.1 mm longer than the package land length, and 0.05 mm wider than the package land width. Center the land on the pad to ensure that the solder joint size is maximized. The bottom of the lead frame chip scale package has a central exposed pad. The pad on the printed circuit board must be at least as large as this exposed pad. The user must connect the exposed pad to VEE using plugged vias to prevent solder from leaking through the vias during reflow. This ensures a solid connection from the exposed pad to VEE.

It is highly recommended to include as many vias as possible when connecting the exposed pad to VEE. This minimizes the thermal resistance between the die and VEE, and minimizes the die temperature. It is recommended that the vias be connected to a VEE plane, or planes, rather than a signal trace, to improve heat dissipation, as shown in Figure 38.

Placing an external VEE plane on the backside of the board opposite the ADN2913 provides an additional benefit because this allows easier heat dissipation into the ambient environment.

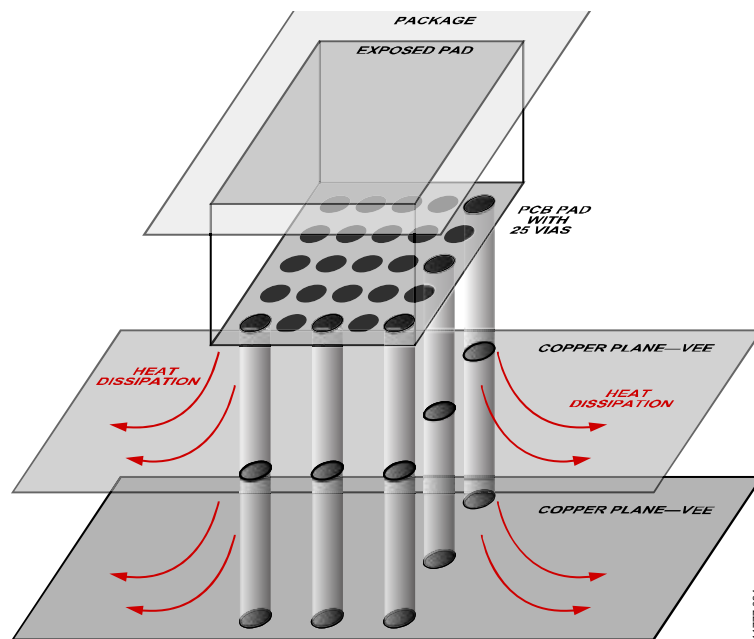
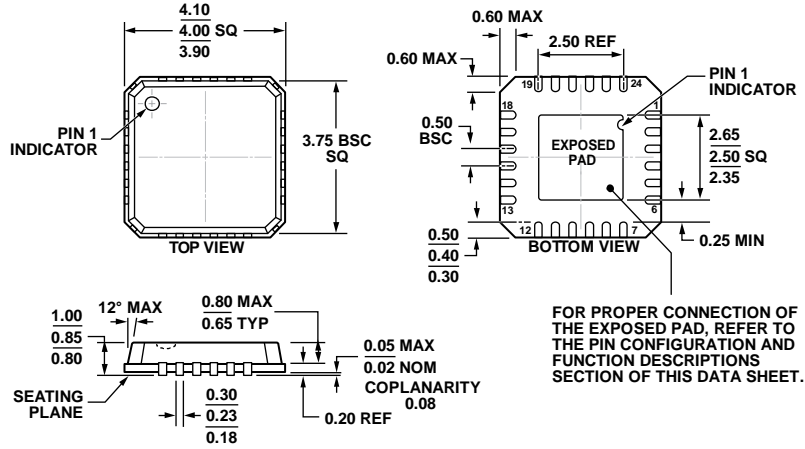


Figure 38. Connecting Vias to VEE



### OUTLINE DIMENSIONS



COMPLIANT TO JEDEC STANDARDS MO-220-VGGD-8

Figure 39. 24-Lead Lead Frame Chip Scale Package [LFCSP\_VQ]  
 4 mm × 4 mm Body, Very Thin Quad  
 (CP-24-3)  
 Dimensions shown in millimeters

04-11-2012-A

### ORDERING GUIDE

Model <sup>1</sup>	Temperature Range	Package Description	Package Option	Ordering Qty
ADN2913ACPZ	-40°C to +85°C	24-Lead LFCSP_VQ	CP-24-3	490
ADN2913ACPZ-RL7	-40°C to +85°C	24-Lead LFCSP_VQ, 7" Tape and Reel	CP-24-3	1500
EVALZ-ADN2913				

<sup>1</sup> Z = RoHS Compliant Part.

**NOTES**

**NOTES**

**NOTES**

I<sup>2</sup>C refers to a communications protocol originally developed by Philips Semiconductors (now NXP Semiconductors).