# ANALOG DEVICES

# 4-Channel, 104 MSPS Digital Transmit Signal Processor (TSP)

# AD6623

## FEATURES
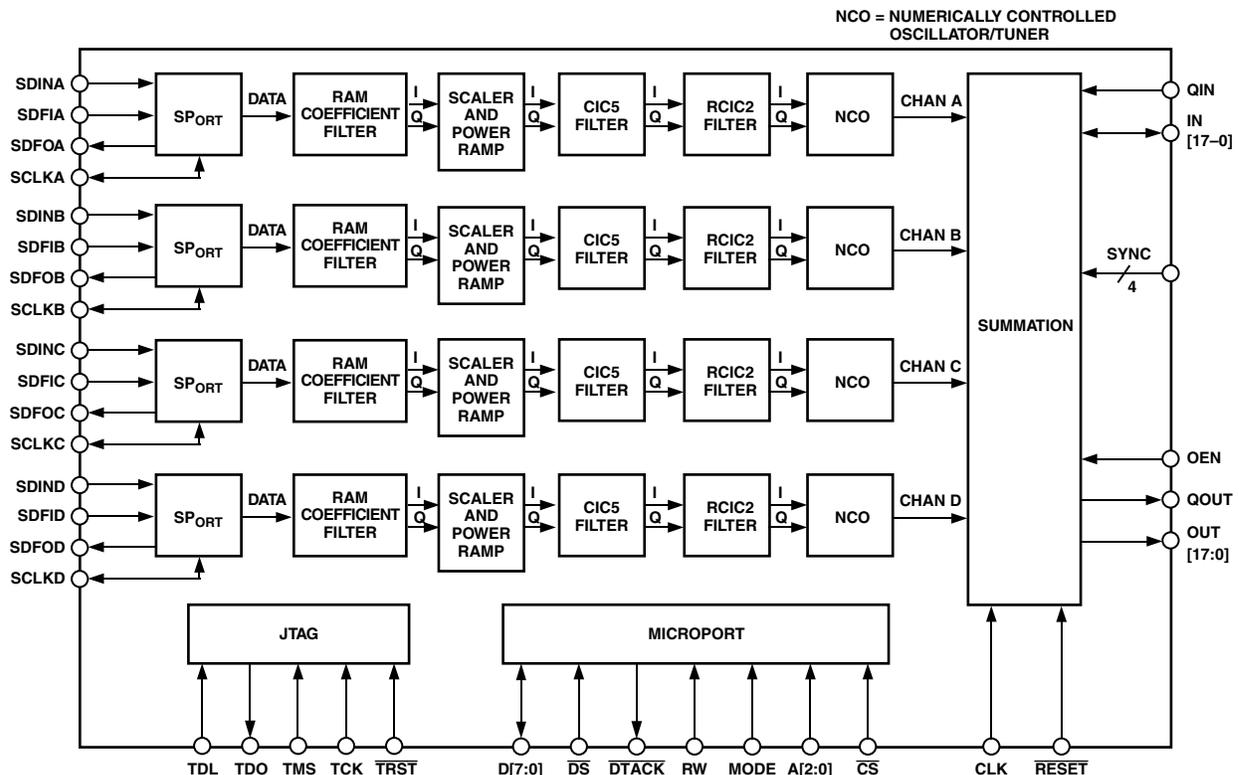**Pin Compatible to the AD6622**
**18-Bit Parallel Digital IF Output**
  Real or Interleaved Complex
**18-Bit Bidirectional Parallel Digital IF Input/Output**
  Allows Cascade of Chips for Additional Channels
  Clipped or Wrapped Over Range
  Two's Complement or Offset Binary Output
**Four Independent Digital Transmitters in Single Package**
**RAM Coefficient Filter (RCF)**
  Programmable IF and Modulation for Each Channel
  Programmable Interpolating RAM Coefficient Filter
  p/4-DQPSK Differential Phase Encoder
  3p/8-PSK Linear Encoder
  8-PSK Linear Encoder
  Programmable GMSK Look-Up Table
  Programmable QPSK Look-Up Table
  All-Pass Phase Equalizer
  Programmable Fine Scaler
  Programmable Power Ramp Unit
**High Speed CIC Interpolating Filter**

**Digital Resampling for Noninteger Interpolation Rates**
**NCO Frequency Translation**
  Spurious Performance Better than –100 dBc
**Separate 3-Wire Serial Data Input for Each Channel**
  Bidirectional Serial Clocks and Frames
**Microprocessor Control**
**2.5 V CMOS Core, 3.3 V Outputs, 5 V Inputs**
**JTAG Boundary Scan**

## APPLICATIONS
**Cellular/PCS Base Stations**
**Micro/Pico Cell Base Stations**
**Wireless Local Loop Base Stations**
**Multicarrier, Multimode Digital Transmit**
  GSM, EDGE, IS136, PHS, IS95, TDS CDMA, UMTS, CDMA2000
**Phased Array Beam Forming Antennas**
**Software Defined Radio**
  Tuning Resolution Better than 0.025 Hz
  Real or Complex Outputs

## FUNCTIONAL BLOCK DIAGRAM



NCO = NUMERICALLY CONTROLLED OSCILLATOR/TUNER

REV. 0

# AD6623

## TABLE OF CONTENTS

**PRODUCT DESCRIPTION**

The AD6623 is a 4-channel Transmit Signal Processor (TSP) that creates high bandwidth data for Transmit Digital-to-Analog Converters (TxDACs) from baseband data provided by a Digital Signal Processor (DSP). Modern TxDACs have achieved sufficiently high sampling rates, analog bandwidth, and dynamic range to create the first Intermediate Frequency (IF) directly. The AD6623 synthesizes multicarrier and multistandard digital signals to drive these TxDACs. The RAM-based architecture allows easy reconfiguration for multimode applications. Modulation, pulse-shaping and anti-imaging filters, static equalization, and tuning functions are combined in a single, cost-effective device. Digital IF signal processing provides repeatable manufacturing, higher accuracy, and more flexibility than comparable high dynamic range analog designs.

The AD6623 has four identical digital TSPs complete with synchronization circuitry and cascadable wideband channel summation. AD6623 is pin compatible to AD6622 and can operate in AD6622-compatible control register mode.

The AD6623 utilizes a 3.3 V I/O power supply and a 2.5 V core power supply. All I/O pins are 5 V tolerant. All control registers and coefficient values are programmed through a generic microprocessor interface. Intel and Motorola microprocessor bus modes are supported. All inputs and outputs are LVCMOS compatible.

**FUNCTIONAL OVERVIEW**

Each TSP has five cascaded signal processing elements: a programmable interpolating RAM Coefficient Filter (RCF), a programmable Scale and Power Ramp, a programmable fifth order Cascaded Integrator Comb (CIC5) interpolating filter, a flexible second order Resampling Cascaded Integrator Comb filter (rCIC2), and a Numerically Controlled Oscillator/Tuner (NCO).

The outputs of the four TSPs are summed and scaled on-chip. In multicarrier wideband transmitters, a bidirectional bus allows the Parallel (wideband) IF Input/Output to drive a second DAC. In this operational mode two AD6623 channels drive one DAC and the other two AD6623 channels drive a second DAC. Multiple AD6623s may be combined by driving the INOUT[17:0] of the succeeding with the OUT[17:0] of the preceding chip. The

INOUT[17:0] can alternatively be masked off by software to allow preceding AD6623's outputs to be ignored.

Each channel accepts input data from independent serial ports that may be connected directly to the serial port of Digital Signal Processor (DSP) chips.

The RCF implements any one of the following functions: Interpolating Finite Impulse Response (FIR) filter, $\pi/4$-DQPSK modulator, 8-PSK modulator, or 3 $\pi/8$-8-PSK modulator, GMSK modulator, and QPSK modulator. Each AD6623 channel can be dynamically switched between the GMSK modulation mode and the 3 $\pi/8$-8-PSK modulation mode in order to support the GSM/EDGE standard. The RCF also implements an Allpass Phase Equalizer (APE) which meets the requirements of IS-95-A/B standard (CDMA transmission).

The programmable Scale and Power Ramp block allows power ramping on a time-slot basis as specified for some air-interface standards (e.g., GSM, EDGE). A fine scaling unit at the programmable FIR filter output allows an easy signal amplitude level adjustment on time slot basis.

The CIC5 provides integer rate interpolation from 1 to 32 and coarse anti-image filtering. The rCIC2 provides fractional rate interpolation from 1 to 4096 in steps of 1/512. The wide range of interpolation factors in each CIC filter stage and a highly flexible resampler incorporated into rCIC2 makes the AD6623 useful for creating both narrowband and wideband carriers in a high-speed sample stream.

The high resolution 32-bit NCO allows flexibility in frequency planning and supports both digital and analog air interface standards. The high speed NCO tunes the interpolated complex signal from the rCIC2 to an IF channel. The result may be real or complex. Multicarrier phase synchronization pins and phase offset registers allow intelligent management of the relative phase of independent RF channels. This capability supports the requirements for phased array antenna architectures and management of the wideband peak/power ratio to minimize clipping at the DAC.

The wideband Output Ports can deliver real or complex data. Complex words are interleaved into real (I) and imaginary (Q) parts at half the master clock rate.

# AD6623

## RECOMMENDED OPERATING CONDITIONS

| Parameter | Test Level | AD6623 Min | Typ | Max | Unit |
|---|---|---|---|---|---|
| VDD | IV | 2.25 | 2.5 | 2.75 | V |
| VDDIO | IV | 3.0 | 3.3 | 3.6 | V |
| T$_{AMBIENT}$ | IV | −40 | +25 | +70 | °C |

## ELECTRICAL CHARACTERISTICS

| Parameter (Conditions) | Temp | Test Level | Min | Typ | Max | Unit |
|---|---|---|---|---|---|---|
| LOGIC INPUTS (5 V TOLERANT) | | | | | | |
|   Logic Compatibility | Full | | | 3.3 V CMOS | | |
|   Logic "1" Voltage | Full | IV | 2.0 | | 5.0 | V |
|   Logic "0" Voltage | Full | IV | −0.3 | | +0.8 | V |
|   Logic "1" Current | Full | IV | | 1 | 10 | µA |
|   Logic "0" Current | Full | IV | | 0 | 10 | µA |
|   Input Capacitance | 25°C | V | | 4 | | pF |
| LOGIC OUTPUTS | | | | | | |
|   Logic Compatibility | Full | | | 3.3 V CMOS/TTL | | |
|   Logic "1" Voltage (I$_{OH}$ = 0.25 mA) | Full | IV | 2.0 | VDD − 0.2 | | V |
|   Logic "0" Voltage (I$_{OL}$ = 0.25 mA) | Full | IV | | 0.2 | 0.4 | V |
| IDD SUPPLY CURRENT | | | | | | |
|   CLK = 104 MHz, VDD = 2.75 V* | Full | IV | | 422 | TBD* | mA |
|   CLK = 104 MHz, VDDIO = 3.6 V* | Full | IV | | 193 | | mA |
|   GSM Example: CORE | | V | | 232 | | mA |
|     I/O | | | | 56 | | mA |
|   IS-136 Example: CORE | | V | | 207 | | mA |
|     I/O | | | | 55 | | mA |
|   WBCDMA Example | | V | | Tbd | | mA |
|   Sleep Mode | Full | IV | | Tbd | TBD | mA |
| POWER DISSIPATION | | | | | | |
|   GSM Example | | V | | 740 | | mW |
|   IS-136 Example | | V | | 700 | | mW |
|   WBCDMA Example | | V | | Tbd | | mW |
|   Sleep Mode | Full | IV | | Tbd | TBD | mW |

*This specification denotes an absolute maximum supply current for the device. The conditions include all channels active, minimum interpolation in both CIC stages, and maximum switching of input data. In an actual application the power will be less.

See the Thermal Management section of the data sheet for further details.

# GENERAL TIMING CHARACTERISTICS[1, 2]

| Parameter (Conditions) | | Temp | Test Level | AD6623AS | | | Unit |
|---|---|---|---|---|---|---|---|
| | | | | Min | Typ | Max | |
| *CLK Timing Requirements:* | | | | | | | |
| $t_{CLK}$ | CLK Period | Full | I | 9.6 | | | ns |
| $t_{CLKL}$ | CLK Width Low | Full | IV | 3 | | | ns |
| $t_{CLKH}$ | CLK Width High | Full | IV | 3 | $0.5 \times t_{CLK}$ | | ns |
| $\overline{RESET}$ *Timing Requirement:* | | | | | | | |
| $t_{RESL}$ | $\overline{RESET}$ Width Low | Full | I | 30.0 | | | ns |
| *Input Data Timing Requirements:* | | | | | | | |
| $t_{SI}$ | INOUT[17:0], QIN to ↑CLK Setup Time | Full | IV | 1 | | | ns |
| $t_{HI}$ | INOUT[17:0], QIN to ↑CLK Hold Time | Full | IV | 2 | | | ns |
| *Output Data Timing Characteristics:* | | | | | | | |
| $t_{DO}$ | ↑CLK to OUT[17:0], INOUT[17:0], QOUT Output Delay Time | Full | IV | 2 | | 6 | ns |
| $t_{DZO}$ | OEN HIGH to OUT[17:0] Active | Full | IV | 3 | | 7.5 | ns |
| *SYNC Timing Requirements:* | | | | | | | |
| $t_{SS}$ | SYNC(0, 1, 2, 3) to ↑CLK Setup Time | Full | IV | 1 | | | ns |
| $t_{HS}$ | SYNC(0, 1, 2, 3) to ↑CLK Hold Time | Full | IV | 2 | | | ns |
| *Master Mode Serial Port Timing Requirements (SCS = 0):* *Switching Characteristics[3]* | | | | | | | |
| $t_{DSCLK1}$ | ↑CLK to ↑SCLK Delay (divide by 1) | Full | IV | 4 | | 10.5 | ns |
| $t_{DSCLKH}$ | ↑CLK to ↑SCLK Delay (for any other divisor) | Full | IV | 5 | | 13 | ns |
| $t_{DSCLKL}$ | ↑CLK to ↓SCLK Delay (divide by 2 or even number) | Full | IV | 3.5 | | 9 | ns |
| $t_{DSCLKLL}$ | ↓CLK to ↓SCLK Delay (divide by 3 or odd number) | Full | IV | 4 | | 10 | ns |
| | Channel is Self-Framing | | | | | | |
| $t_{SSDI0}$ | SDIN to ↑SCLK Setup Time | Full | IV | 1.7 | | | ns |
| $t_{HSDI0}$ | SDIN to ↑SCLK Hold Time | Full | IV | 0 | | | ns |
| $t_{DSFO0A}$ | ↑SCLK to SDFO Delay | Full | IV | 0.5 | | 3.5 | ns |
| | Channel is External-Framing | | | | | | |
| $t_{SSFI0}$ | SDFI to ↑SCLK Setup Time | Full | IV | 2 | | | ns |
| $t_{HSFI0}$ | SDFI to ↑SCLK Hold Time | Full | IV | 0 | | | ns |
| $t_{SSDI0}$ | SDIN to ↑SCLK Setup Time | Full | IV | 2 | | | ns |
| $t_{HSDI0}$ | SDIN to ↑SCLK Hold Time | Full | IV | 0 | | | ns |
| $t_{DSFO0B}$ | ↑SCLK to SDFO Delay | Full | IV | 0.5 | | 3 | ns |
| *Slave Mode Serial Port Timing Requirements (SCS = 1):* *Switching Characteristics[3]* | | | | | | | |
| $t_{SCLK}$ | SCLK Period | Full | IV | | $2 \times t_{CLK}$ | | ns |
| $t_{SCLKL}$ | SCLK Low Time | Full | IV | 3.5 | | | ns |
| $t_{SCLKH}$ | SCLK High Time | Full | IV | 3.5 | | | ns |
| | Channel is Self-Framing | | | | | | |
| $t_{SSDH}$ | SDIN to ↑SCLK Setup Time | Full | IV | 1 | | | ns |
| $t_{HSDH}$ | SDIN to ↑SCLK Hold Time | Full | IV | 2.5 | | | ns |
| $t_{DSFO1}$ | ↑SCLK to SDFO Delay | Full | IV | 4 | | 10 | ns |
| | Channel is External-Framing | | | | | | |
| $t_{SSFI1}$ | SDFI to ↑ SCLK Setup Time | Full | IV | 2 | | | ns |
| $t_{HSFI1}$ | SDFI to ↑SCLK Hold Time | Full | IV | 1 | | | ns |
| $t_{SSDI1}$ | SDIN to ↑SCLK Setup Time | Full | IV | 1 | | | ns |
| $t_{HSDI1}$ | SDIN to ↑SCLK Hold Time | Full | IV | 2.5 | | | ns |
| $t_{DSFO1}$ | ↓SCLK to SDFO Delay | Full | IV | 10 | | | ns |

NOTES
[1]All Timing Specifications valid over VDD range of 2.375 V to 2.675 V and VDDIO range of 3.0 V to 3.6 V.
[2]$C_{LOAD}$ = 40 pF on all outputs (unless otherwise specified).
[3]The timing parameters for SCLK, SDIN, SDFI, SDFO, and SYNC apply to all four channels (A, B, C, and D).

Specifications subject to change without notice.

# AD6623

## MICROPROCESSOR PORT TIMING CHARACTERISTICS[1, 2]

| Parameter (Conditions) | | Temp | Test Level | AD6623AS | | | Unit |
|---|---|---|---|---|---|---|---|
| | | | | Min | Typ | Max | |
| **MICROPROCESSOR PORT, MODE INM (MODE = 0)** | | | | | | | |
| *MODE INM Write Timing*: | | | | | | | |
| $t_{SC}$ | Control[3] to ↑CLK Setup Time | Full | IV | 4.5 | | | ns |
| $t_{HC}$ | Control[3] to ↑CLK Hold Time | Full | IV | 2.0 | | | ns |
| $t_{HWR}$ | $\overline{WR}$(RW) to RDY($\overline{DTACK}$) Hold Time | Full | IV | 8.0 | | | ns |
| $t_{SAM}$ | Address/Data to $\overline{WR}$(RW) Setup Time | Full | IV | 3.0 | | | ns |
| $t_{HAM}$ | Address/Data to RDY($\overline{DTACK}$) Hold Time | Full | IV | 2.0 | | | ns |
| $t_{DRDY}$ | $\overline{WR}$(RW) to RDY($\overline{DTACK}$) Delay | Full | IV | 4.0 | | | ns |
| $t_{ACC}$ | $\overline{WR}$(RW) to RDY($\overline{DTACK}$) High Delay | Full | IV | $4 \times t_{CLK}$ | $5 \times t_{CLK}$ | $9 \times t_{CLK}$ | ns |
| *MODE INM Read Timing*: | | | | | | | |
| $t_{SC}$ | Control[3] to ↑CLK Setup Time | Full | IV | 4.5 | | | ns |
| $t_{HC}$ | Control[3] to ↑CLK Hold Time | Full | IV | 2.0 | | | ns |
| $t_{SAM}$ | Address to $\overline{RD}$($\overline{DS}$) Setup Time | Full | IV | 3.0 | | | ns |
| $t_{HAM}$ | Address to Data Hold Time | Full | IV | 2.0 | | | ns |
| $t_{ZOZ}$ | Data Three-State Delay | Full | IV | | | | ns |
| $t_{DD}$ | RDY($\overline{DTACK}$) to Data Delay | Full | IV | | | | ns |
| $t_{DRDY}$ | $\overline{RD}$($\overline{DS}$) to RDY($\overline{DTACK}$) Delay | Full | IV | 4.0 | | | ns |
| $t_{ACC}$ | $\overline{RD}$($\overline{DS}$) to RDY($\overline{DTACK}$) High Delay | Full | IV | $8 \times t_{CLK}$ | $10 \times t_{CLK}$ | $13 \times t_{CLK}$ | ns |
| **MICROPROCESSOR PORT, MOTOROLA (MODE = 1)** | | | | | | | |
| *MODE MNM Write Timing*: | | | | | | | |
| $t_{SC}$ | Control[3] to ↑CLK Setup Time | Full | IV | 4.5 | | | ns |
| $t_{HC}$ | Control[3] to ↑CLK Hold Time | Full | IV | 2.0 | | | ns |
| $t_{HDS}$ | $\overline{DS}$($\overline{RD}$) to $\overline{DTACK}$(RDY) Hold Time | Full | IV | 8.0 | | | ns |
| $t_{HRW}$ | RW($\overline{WR}$) to $\overline{DTACK}$(RDY) Hold Time | Full | IV | 8.0 | | | ns |
| $t_{SAM}$ | Address/Data to RW($\overline{WR}$) Setup Time | Full | IV | 3.0 | | | ns |
| $t_{HAM}$ | Address/Data to RW($\overline{WR}$) Hold Time | Full | IV | 2.0 | | | ns |
| $t_{DDTACK}$ | $\overline{DS}$($\overline{RD}$) to $\overline{DTACK}$(RDY) Delay | | | | | | ns |
| $t_{ACC}$ | RW($\overline{WR}$) to $\overline{DTACK}$(RDY) Low Delay | Full | IV | $4 \times t_{CLK}$ | $5 \times t_{CLK}$ | $9 \times t_{CLK}$ | ns |
| MODE MNM Read Timing: | | | | | | | |
| $t_{SC}$ | Control[3] to ↑CLK Setup Time | Full | IV | 4.0 | | | ns |
| $t_{HC}$ | Control[3] to ↑CLK Hold Time | Full | IV | 2.0 | | | ns |
| $t_{HDS}$ | $\overline{DS}$($\overline{RD}$) to $\overline{DTACK}$(RDY) Hold Time | Full | IV | 8.0 | | | ns |
| $t_{SAM}$ | Address to $\overline{DS}$($\overline{RD}$) Setup Time | Full | IV | 3.0 | | | ns |
| $t_{HAM}$ | Address to Data Hold Time | Full | IV | 2.0 | | | ns |
| $t_{ZD}$ | Data Three-State Delay | Full | IV | | | | ns |
| $t_{DD}$ | $\overline{DTACK}$(RDY) to Data Delay | Full | IV | | | | ns |
| $t_{DDTACK}$ | $\overline{DS}$($\overline{RD}$) to $\overline{DTACK}$(RDY) Delay | Full | IV | | | | ns |
| $t_{ACC}$ | $\overline{DS}$($\overline{RD}$) to $\overline{DTACK}$(RDY) Low Delay | Full | IV | $8 \times t_{CLK}$ | $10 \times t_{CLK}$ | $13 \times t_{CLK}$ | ns |

NOTES
[1]All Timing Specifications valid over VDD range of 2.375 V to 2.675 V and VDDIO range of 3.0 V to 3.6 V.
[2]$C_{LOAD}$ = 40 pF on all outputs (unless otherwise specified).
[3]Specification pertains to control signals: RW, ($\overline{WR}$), $\overline{DS}$, ($\overline{RD}$), $\overline{CS}$.

Specifications subject to change without notice.

**TIMING DIAGRAMS**
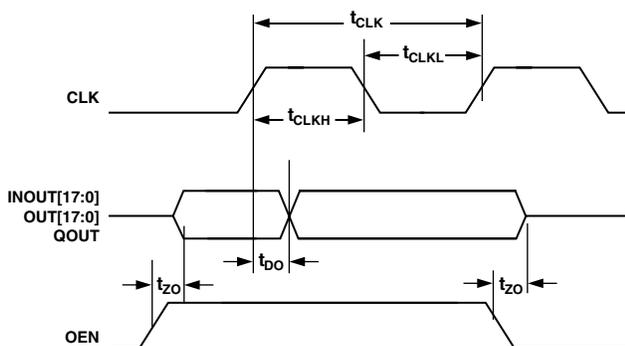


Figure 1.  Parallel Output Switching Characteristics



Figure 2.  Wideband Input Timing



Figure 3.  SYNC Timing Inputs



Figure 4.  $\overline{RESET}$ Timing Requirements



Figure 5.  SCLK Switching Characteristics (Divide by 1)
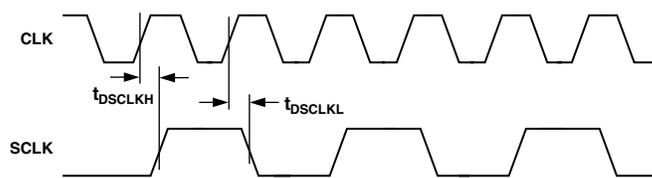


Figure 6.  SCLK Switching Characteristic (Divide by 2 or EVEN Integer)
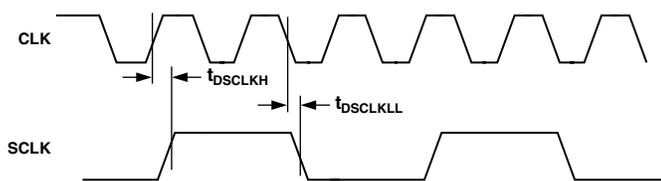


Figure 7.  SCLK Switching Characteristic (Divide by 3 or ODD Integer)

**AD6623**
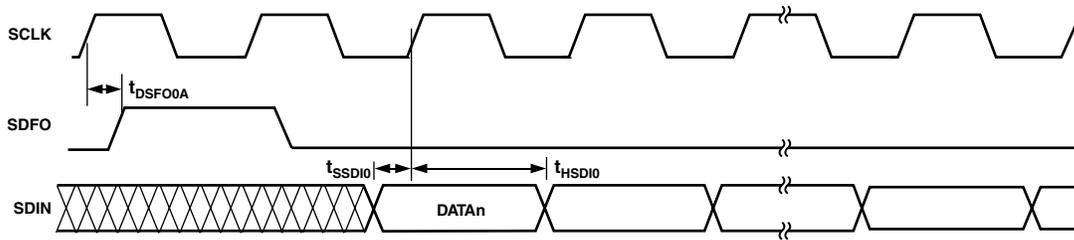


Figure 8. Serial Port Timing, Master Mode (SCS = 0), Channel is Self-Framing
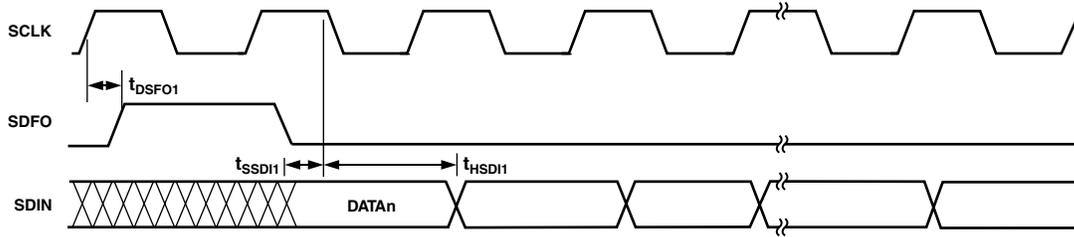


Figure 9. Serial Port Timing, Slave Mode (SCS = 1), Channel is Self-Framing
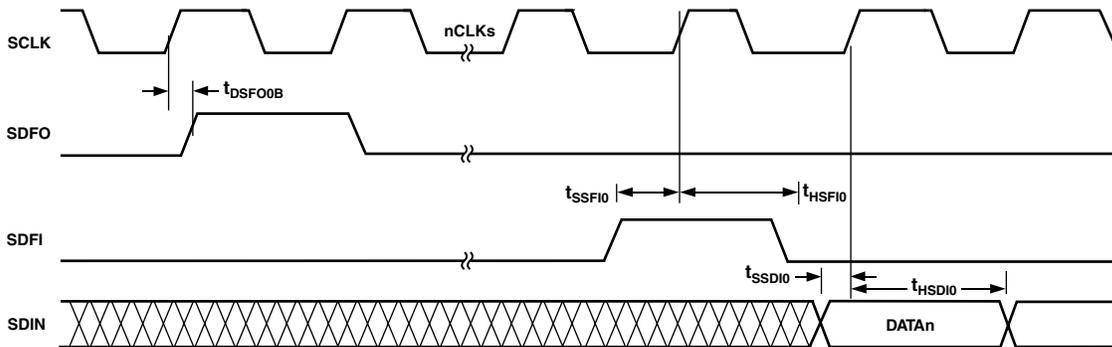


Figure 10. Serial Port Timing, Master Mode (SCS = 0), Channel is External-Framing
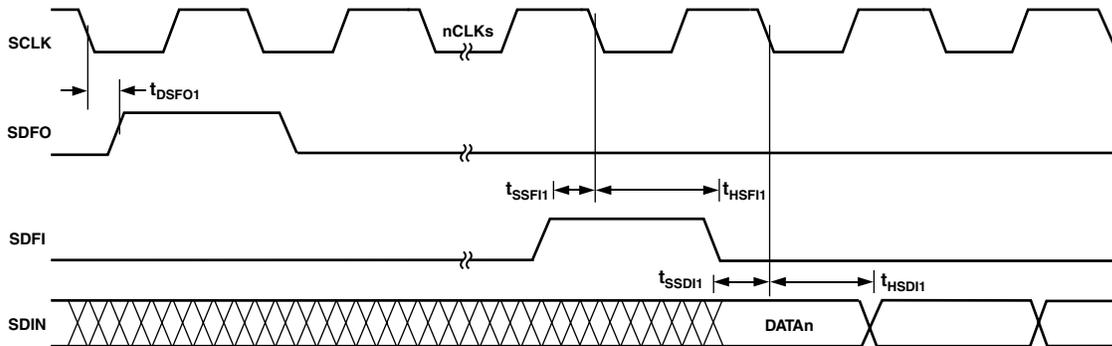


Figure 11. Serial Port Timing, Slave Mode (SCS = 1), Channel is External-Framing

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## TIMING DIAGRAMS—INM MICROPORT MODE
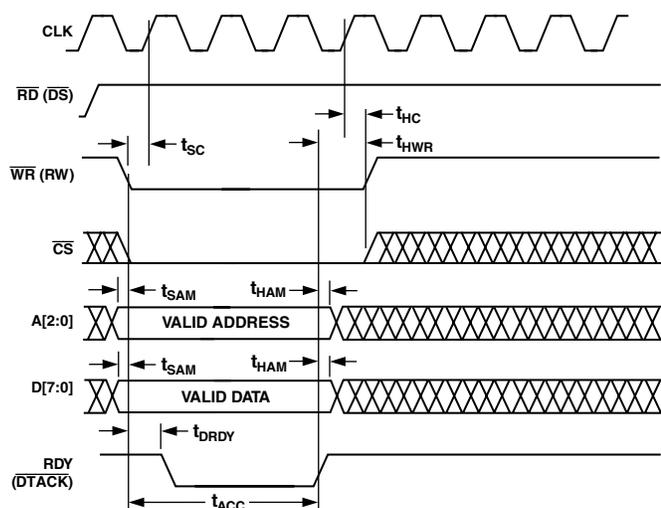


NOTES
1. $t_{ACC}$ ACCESS TIME DEPENDS ON THE ADDRESS ACCESSED. ACCESS TIME IS MEASURED FROM FE OF $\overline{WR}$ TO THE RE OF RDY.
2. $t_{ACC}$ REQUIRES A MAXIMUM 9 CLK PERIODS.

Figure 12. INM Microport Write Timing Requirements



NOTES
1. $t_{ACC}$ ACCESS TIME DEPENDS ON THE ADDRESS ACCESSED. ACCESS TIME IS MEASURED FROM FE OF $\overline{WR}$ TO THE RE OF RDY.
2. $t_{ACC}$ REQUIRES A MAXIMUM OF 13 CLK PERIODS AND APPLIES TO A[2:0] = 7, 6, 5, 3, 2, 1

Figure 13. INM Microport Read Timing Requirements
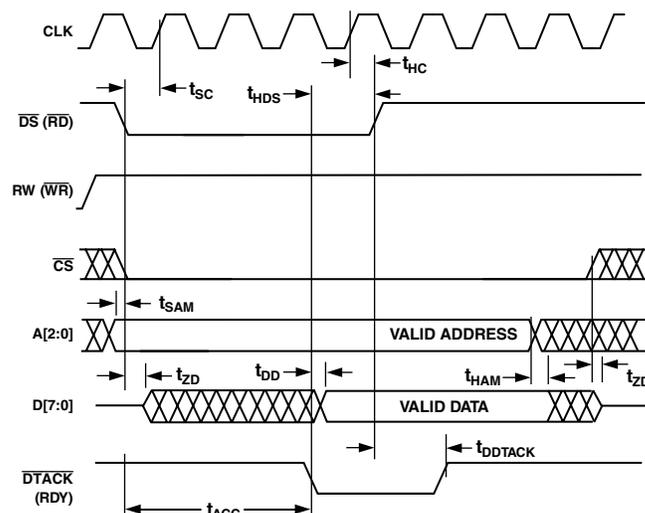
## TIMING DIAGRAMS—MNM MICROPORT MODE



NOTES
1. $t_{ACC}$ ACCESS TIME DEPENDS ON THE ADDRESS ACCESSED. ACCESS TIME IS MEASURED FROM FE OF $\overline{DS}$ TO THE FE OF $\overline{DTACK}$.
2. $t_{ACC}$ REQUIRES A MAXIMUM 9 CLK PERIODS.

Figure 14. MNM Microport Write Timing Requirements



NOTES
1. $t_{ACC}$ ACCESS TIME DEPENDS ON THE ADDRESS ACCESSED. ACCESS TIME IS MEASURED FROM FE OF $\overline{DS}$ TO THE FE OF $\overline{DTACK}$.
2. $t_{ACC}$ REQUIRES A MAXIMUM 13 CLK PERIODS.

Figure 15. Motorola Microport Read Timing Requirements

# AD6623

**ABSOLUTE MAXIMUM RATINGS\***

Supply Voltage . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . . 3.6 V
Input Voltage . . . . . . . . . . . . . . −0.3 V to +5 V (5 V Tolerant)
Output Voltage Swing . . . . . . . . . −0.3 V to VDDIO + 0.3 V
Load Capacitance . . . . . . . . . . . . . . . . . . . . . . . . . . . 200 pF
Junction Temperature Under Bias . . . . . . . . . . . . . . . . 125°C
Storage Temperature Range . . . . . . . . . . . −65°C to +150°C
Lead Temperature (5 sec) . . . . . . . . . . . . . . . . . . . . . 280°C

\*Stresses greater than those listed above may cause permanent damage to the device. These are stress ratings only; functional operation of the devices at these or any other conditions greater than those indicated in the operational sections of this specification is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

**THERMAL CHARACTERISTICS**

128-Lead MQFP:
$\theta_{JA}$ = 33°C/W, no airflow
$\theta_{JA}$ = 27°C/W, 200 lfpm airflow
$\theta_{JA}$ = 24°C/W, 400 lfpm airflow

196-Lead BGA:
$\theta_{JA}$ = 26.3°C/W, no airflow
$\theta_{JA}$ = 22°C/W, 200 lfpm airflow

Thermal measurements made in the horizontal position on a 2-layer board.

**EXPLANATION OF TEST LEVELS**
I.   100% Production Tested
II.  100% Production Tested at 25°C, and Sample Tested at Specified Temperatures
III. Sample Tested Only
IV.  Parameter Guaranteed by Design and Analysis
V.   Parameter is Typical Value Only

## ORDERING GUIDE

| Model | Temperature Range | Package Description | Package Option |
|---|---|---|---|
| AD6623AS | −40°C to +70°C (Ambient) | 128-Lead MQFP (Metric Quad Flatpack) | S-128A |
| AD6623ABC | −40°C to +85°C (Ambient) | 196-Lead BGA (Ball Grid Array) | BC-196 |
| AD6623S/PCB | | MQFP Evaluation Board with AD6623 and Software | |
| AD6623BC/PCB | | BGA Evaluation Board with AD6623 and Software | |

**CAUTION**
ESD (electrostatic discharge) sensitive device. Electrostatic charges as high as 4000 V readily accumulate on the human body and test equipment and can discharge without detection. Although the AD6623 features proprietary ESD protection circuitry, permanent damage may occur on devices subjected to high-energy electrostatic discharges. Therefore, proper ESD precautions are recommended to avoid performance degradation or loss of functionality.

**WARNING!**
ESD SENSITIVE DEVICE

# AD6623

## PIN CONFIGURATION
### 128-Lead MQFP



AD6623
TOP VIEW
(Not to Scale)

**Top pins (left to right):**
102 GND, 101 TCK, 100 TRST, 99 GND, 98 GND, 97 INOUT0, 96 GND, 95 GND, 94 INOUT1, 93 INOUT2, 92 INOUT3, 91 INOUT4, 90 VDDIO, 89 INOUT5, 88 INOUT6, 87 INOUT7, 86 INOUT8, 85 GND, 84 GND, 83 GND, 82 INOUT9, 81 INOUT10, 80 INOUT11, 79 INOUT12, 78 VDDIO, 77 INOUT13, 76 INOUT14, 75 INOUT15, 74 INOUT16, 73 SYNC3, 72 GND, 71 INOUT17, 70 QIN, 69 SYNC2, 68 GND, 67 CLK, 66 VDD, 65 GND

**Left pins (top to bottom):**
103 GND, 104 VDD, 105 SDFIA, 106 TMS, 107 TDO, 108 TDI, 109 SCLKA, 110 VDDIO, 111 SDFOA, 112 SDINA, 113 SCLKB, 114 SDFOB, 115 SDFIB, 116 GND, 117 SDFIC, 118 SDINB, 119 SCLKC, 120 SDFOC, 121 SDINC, 122 VDDIO, 123 SCLKD, 124 SDFOD, 125 SDIND, 126 SDFID, 127 VDD, 128 GND

**Right pins (top to bottom):**
64 GND, 63 SYNC1, 62 SYNC0, 61 RESET, 60 CS, 59 VDD, 58 A0, 57 A1, 56 A2, 55 MODE, 54 GND, 53 GND, 52 GND, 51 RW(WR), 50 DTACK(RDY), 49 DS(RD), 48 D0, 47 VDD, 46 D1, 45 D2, 44 D3, 43 D4, 42 GND, 41 VDDIO, 40 D5, 39 GND

**Bottom pins (left to right):**
1 GND, 2 OEN, 3 GND, 4 GND, 5 GND, 6 OUT0, 7 OUT1, 8 OUT2, 9 GND, 10 OUT3, 11 OUT4, 12 OUT5, 13 OUT6, 14 VDDIO, 15 OUT7, 16 OUT8, 17 OUT9, 18 OUT10, 19 GND, 20 GND, 21 GND, 22 OUT11, 23 OUT12, 24 OUT13, 25 OUT14, 26 VDDIO, 27 OUT15, 28 OUT16, 29 OUT17, 30 QOUT, 31 GND, 32 GND, 33 D7, 34 GND, 35 GND, 36 GND, 37 D6, 38 GND

# AD6623

| Pin Number | Mnemonic | Type | Description |
|---|---|---|---|
| 1, 3–5, 9, 19–21, 31, 32, 34–36, 38, 39, 42, 52–54, 64–65, 68, 72, 83–85, 95, 96, 98, 99, 102, 103, 116, 128 | GND | P | Ground Connection |
| 2 | OEN[1] | I | Active High Output Enable Pin |
| 29, 28, 27, 25, 24, 23, 22, 18, 17, 16, 15, 13, 12, 11, 10, 8, 7, 6 | OUT[17:0] | O/T | Parallel Output Data |
| 47, 59, 66, 104, 127 | VDD | P | 2.5 V Supply |
| 14, 26, 41, 78, 90, 110, 122 | VDDIO | P | 3.3 V Supply |
| 30 | QOUT | O/T | When HIGH indicates Q Output Data (Complex Output Mode) |
| 33, 37, 40, 43, 44, 45, 46, 48 | D[7:0] | I/O/T | Bidirectional Microport Data |
| 49 | $\overline{DS}$ ($\overline{RD}$) | I | INM Mode: Read Signal, MNM Mode: Data Strobe Signal |
| 50 | $\overline{DTACK}$ (RDY) | O | Acknowledgment of a Completed Transaction (Signals when μP Port Is Ready for an Access) Open Drain, Must Be Pulled Up Externally |
| 51 | RW ($\overline{WR}$) | I | Active HIGH Read, Active Low Write |
| 55 | MODE | I | Sets Microport Mode: MODE = 1, MNM Mode; MODE = 0, INM Mode |
| 56, 57, 58 | A[2:0] | I | Microport Address Bus |
| 60 | $\overline{CS}$ | I | Chip Select, Active low enable for μP Access |
| 61 | $\overline{RESET}$[2] | I | Active Low Reset Pin |
| 62 | SYNC0[1] | I | SYNC Signal for Synchronizing Multiple AD6623s |
| 63 | SYNC1[1] | I | SYNC Signal for Synchronizing Multiple AD6623s |
| 67 | CLK[1] | I | Input Clock |
| 69 | SYNC2[1] | I | SYNC Signal for Synchronizing Multiple AD6623s |
| 70 | QIN[1] | I | When HIGH indicates Q input data (Complex Input Mode) |
| 71, 74–77, 79–82, 86–89, 91–94, 97 | INOUT[17:0][1] | I/O | Wideband Input/Output Data (Allows Cascade of Multiple AD6623 Chips In a System) |
| 73 | SYNC3[1] | I | SYNC Signal for Synchronizing Multiple AD6623s |
| 100 | $\overline{TRST}$[2] | I | Test Reset Pin |
| 101 | TCK[1] | I | Test Clock Input |
| 105 | SDFIA | I | Serial Data Frame Input—Channel A |
| 106 | TMS[2] | I | Test Mode Select |
| 107 | TDO | O | Test Data Output |
| 108 | TDI[1] | I | Test Data Input |
| 109 | SCLKA | I/O | Bidirectional Serial Clock—Channel A |
| 111 | SDFOA | O | Serial Data Frame Sync Output—Channel A |
| 112 | SDINA[1] | I | Serial Data Input—Channel A |
| 113 | SCLKB | I/O | Bidirectional Serial Clock—Channel B |
| 114 | SDFOB | O | Serial Data Frame Sync Output—Channel B |
| 115 | SDFIB | I | Serial Data Frame Input —Channel B |
| 117 | SDFIC | I | Serial Data Frame Input—Channel C |
| 118 | SDINB[1] | I | Serial Data Input—Channel B |
| 119 | SCLKC | I/O | Bidirectional Serial Clock—Channel C |
| 120 | SDFOC | O | Serial Data Frame Sync Output—Channel C |
| 121 | SDINC[1] | I | Serial Data Input—Channel C |
| 123 | SCLKD | I/O | Bidirectional Serial Clock—Channel D |
| 124 | SDFOD | O | Serial Data Frame Sync Outpu—Channel D |
| 125 | SDIND[1] | I | Serial Data Input—Channel D |
| 126 | SDFID | I | Serial Data Frame Input—Channel D |

NOTES
[1]Pins with a Pull-Down resistor of nominal 70 kΩ.
[2]Pins with a Pull-Up resistor of nominal 70 kΩ.

REV. 0

## PIN CONFIGURATION
### 196-Lead BGA

**TOP VIEW**



15mm sq.

**BALL LEGEND**
○ I/O
● GROUND
◍ CORE POWER
◉ RING POWER

←1.0mm
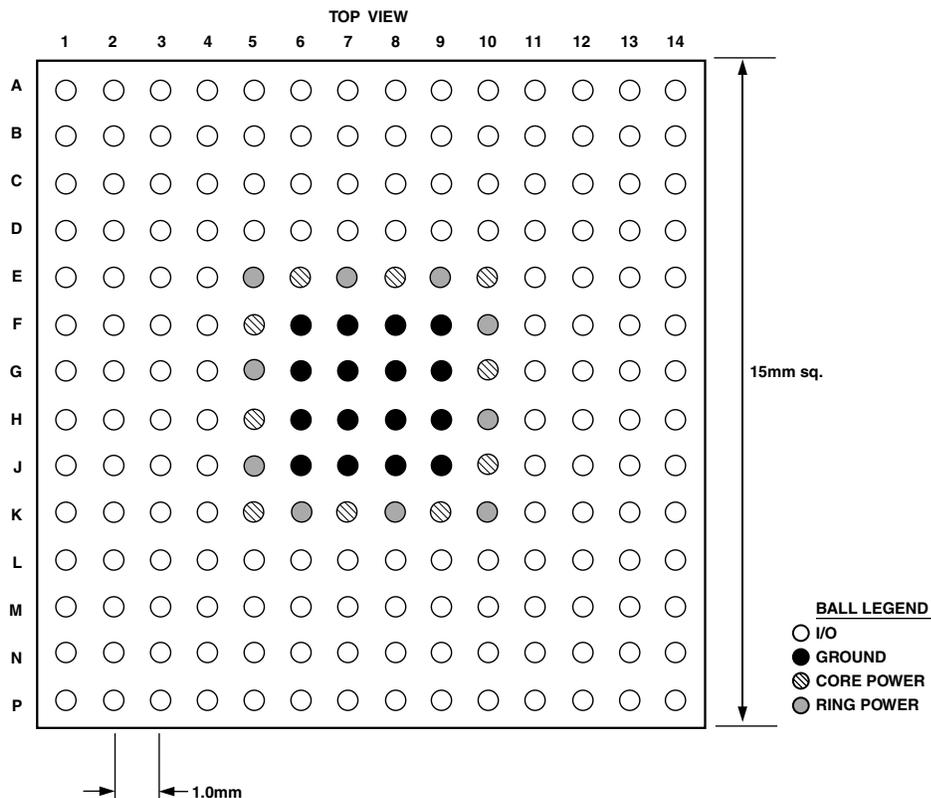
| | 1 | 2 | 3 | 4 | 5 | 6 | 7 | 8 | 9 | 10 | 11 | 12 | 13 | 14 |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| A | NC | | | SDFID | SDINC | SDINB | SDFOB | SCLKB | SCLKA | TDO | SDFIA | TCK | | NC |
| B | | | | OEN | SDIND | SDFOC | SDFIC | SDINA | TDI | | $\overline{\text{TRST}}$ | | | |
| C | | | | SDFOD | | SKLKD | SKLKC | SDFIB | SDFOA | TMS | | | | IN0 |
| D | OUT2 | | OUT0 | | | | | | | | | IN2 | | IN1 |
| E | | OUT1 | OUT3 | | VDDIO | VDD | VDDIO | VDD | VDDIO | VDD | | | | IN4 |
| F | OUT5 | OUT4 | OUT6 | | VDD | GND | GND | GND | GND | VDDIO | | IN3 | IN5 | IN7 |
| G | OUT8 | OUT4 | | | VDDIO | GND | GND | GND | GND | VDD | | IN6 | IN8 | IN9 |
| H | OUT9 | OUT10 | OUT12 | | VDD | GND | GND | GND | GND | VDDIO | | | IN11 | IN10 |
| J | OUT11 | OUT13 | | | VDDIO | GND | GND | GND | GND | VDD | | IN12 | IN14 | IN13 |
| K | OUT14 | | OUT17 | | VDD | VDDIO | VDD | VDDIO | VDD | VDDIO | | IN16 | IN17 | IN15 |
| L | OUT16 | | OUT15 | | | | | | | | | QIN | | SYNC3 |
| M | QOUT | | D7 | D4 | D1 | $\overline{\text{DTACK}}$ (RDY) | MODE (ALE) | A1 | $\overline{\text{RESET}}$ | | | | | SYNC2 |
| N | | | | D5 | D3 | D0 | RW($\overline{\text{WR}}$) | | A0 | SYNC0 | | | | |
| P | NC | D6 | | D2 | | $\overline{\text{DS}}(\overline{\text{RD}})$ | | | A2 | $\overline{\text{CS}}$ | SYNC1 | | CLK | NC |

NC = NO CONNECT

# AD6623

| Mnemonic | Type | Function |
|---|---|---|
| **POWER SUPPLY** | | |
| VDD | P | 2.5 V Supply |
| VDDIO | P | 3.3 V IO Supply |
| GND | G | Ground |
| **INPUTS** | | |
| INOUT[17:0][1] | I/O | A Input Data (Mantissa) |
| QIN[1] | I | When HIGH Indicates Q Input Data (Complex Input Mode) |
| $\overline{\text{RESET}}$[2] | I | Active LOW Reset Pin |
| CLK[1] | I | Input Clock |
| SYNC0[1] | I | All Sync Pins Go to All Four Output Channels |
| SYNC1[1] | I | All Sync Pins Go to All Four Output Channels |
| SYNC2[1] | I | All Sync Pins Go to All Four Output Channels |
| SYNC3[1] | I | All Sync Pins Go to All Four Output Channels |
| SDINA[1] | I | Serial Data Input—Channel A |
| SDINB[1] | I | Serial Data Input—Channel B |
| SDINC[1] | I | Serial Data Input—Channel C |
| SDIND[1] | I | Serial Data Input—Channel D |
| $\overline{\text{CS}}$ | I | Active LOW Chip Select |
| **CONTROL** | | |
| SCLKA | I/O | Bidirectional Serial Clock—Channel A |
| SCLKB | I/O | Bidirectional Serial Clock—Channel B |
| SCLKC | I/O | Bidirectional Serial Clock—Channel C |
| SCLKD | I/O | Bidirectional Serial Clock—Channel D |
| SDFOA | O | Serial Data Frame Sync Output—Channel A |
| SDFOB | O | Serial Data Frame Sync Output—Channel B |
| SDFOC | O | Serial Data Frame Sync Output—Channel C |
| SDFOD | O | Serial Data Frame Sync Output—Channel D |
| SDFIA | I | Serial Data Frame Input—Channel A |
| SDFIB | I | Serial Data Frame Input—Channel B |
| SDFIC | I | Serial Data Frame Input—Channel C |
| SDFID | I | Serial Data Frame Input—Channel D |
| OEN[1] | I | Active High Output Enable Pin |
| **MICROPORT CONTROL** | | |
| D[7:0] | I/O/T | Bidirectional Microport Data |
| A[2:0] | I | Microport Address Bus |
| $\overline{\text{DS}}$ ($\overline{\text{RD}}$) | I | Active Low Data Strobe (Active Low Read) |
| $\overline{\text{DTACK}}$ (RDY)[2] | O/T | Active Low Data Acknowledge (Microport Status Bit) |
| RW ($\overline{\text{WR}}$) | I | Read Write (Active Low Write) |
| MODE | I | Intel or Motorola Mode Select |
| **OUTPUTS** | | |
| OUT[17:0] | O | Wideband Output Data |
| QOUT | O | When HIGH Indicates Q Output Data (Complex Output Mode) |
| **JTAG AND BIST** | | |
| $\overline{\text{TRST}}$[2] | I | Test Reset Pin (Active Low) |
| TCK[1] | I | Test Clock Input |
| TMS[2] | I | Test Mode Select Input |
| TDO | O/T | Test Data Output |
| TDI[1] | I | Test Data Input |

NOTES
[1]Pins with a Pull-Down resistor of nominal 70 kΩ.
[2]Pins with a Pull-Up resistors of nominal 70 kΩ.

**SERIAL DATA PORT**
The AD6623 has four independent Serial Ports (A, B, C, and D), and each accepts data to its own channel (A, B, C, or D) of the device. Each Serial Port has four pins: SCLK (Serial CLocK), SDFO (Serial Data Frame Out), SDFI (Serial Data Frame In), and SDIN (Serial Data INput). SDFI and SDIN are inputs, SDFO is an output, and SCLK is either input or output depending on the state of SCS (Serial Clock Slave: 0xn16, Bit 4). Each channel can be operated either as a Master or Slave channel depending upon SCS. The Serial Port can be self-framing or accept external framing from the SFDI pin or from the previous adjacent channel (0xn16, Bits 7 and 6).

**Serial Master Mode (SCS = 0)**
In master mode, SCLK is created by a programmable internal counter that divides CLK. When the channel is "sleeping," SCLK is held low. SCLK becomes active on the first rising edge of CLK after Channel sleep is removed (D0 through D3 of external address 4). Once active, the SCLK frequency is determined by the CLK frequency and the SCLK divider, according to the equations below.

AD6623 mode:

$$f_{SCLK} = \frac{f_{CLK}}{SCLKdivider + 1} \qquad (1)$$

AD6622 mode:

$$f_{SCLK} = \frac{f_{CLK}}{2 \times (SCLKdivider + 1)} \qquad (2)$$

The SCLK divider is a 5-bit unsigned value located at Internal Channel Address 0xn0D (Bits 4–0), where "n" is 1, 2, 3, or 4 for the chosen channel A, B, C, or D, respectively. The user must select the SCLK divider to insure that SCLK is fast enough to accept full input sample words at the input sample rate. See the design example at the end of this section. The maximum SCLK frequency is equal to the CLK when operating in AD6623 mode serial clock master. When operating in AD6622 compatible mode, the maximum SCLK frequency is one-half the CLK. The minimum SCLK frequency is 1/32 of the CLK frequency in AD6623 mode or 1/64 of the CLK frequency when in AD6622 mode. SDFO changes on the positive edge of SCLK when in master mode. SDIN is captured on positive edge when SCLK is in master mode.

**Serial Slave Mode (SCS = 1)**
Any of the AD6623 serial ports may be operated in the serial slave mode. In this mode, the selected AD6623 channel requires that an external device such as a DSP to supply the SCLK. This is done to synchronize the serial port to meet an external timing requirement. SDIN is captured on negative edge of SCLK when in slave mode.

**Self-Framing Mode**
In this mode Bit 7 of register 0xn16 is set low. The serial data frame output, SDFO, generates a self-framing data request and is pulsed high for one SCLK cycle at the input sample rate. In this mode, the SDFI pin is not used, and the SDFO signal would be programmed to be a serial data frame request (0xn16, Bit 5 = 0).

SDFO is used to provide a sync signal to the host. The input sample rate is determined by the CLK divided by channel interpolation factor. If the SCLK rate is not an integer multiple of the input sample rate, then the SDFO will continually adjust the period by one SCLK cycle to keep the average SDFO rate equal to the input sample rate. When the channel is in sleep mode, SDFO is held low. The first SDFO is delayed by the channel reset latency after the Channel Reset is removed. The channel reset latency varies dependent on channel configuration.

**External Framing Mode**
In this mode Bit 7 of register 0xn16 is set high. The external framing can come from either the SDFI pin (0xn16, Bit 6 = 0) or the previous adjacent channel (0xn16, Bit 6 = 1). In the case of external framing from a previous channel, it uses the internal frame end signal for serial data frame syncing. When in master mode, SDFO and SDFI transition on the positive edge of SCLK, and SDIN is captured on the positive edge of SCLK. When in slave mode, SDFO and SDFI transition on the negative edge of SCLK, and SDIN is captured on the negative edge of SCLK.

**Serial Port Cascade Configuration**
In this case the SDFO signal from the last channel of the first chip would be programmed to be a serial data frame end (SFE: 0xn16, Bit 5 = 1). This SDFO signal would then be fed as an input for the second cascaded chip's SDFI pin input. The second chip would be programmed to accept external framing from the SDFI pin (0xn16, Bit 7 = 1, Bit 6 = 0).

**Serial Data Format**
The format of data applied to the serial port is determined by the RCF mode selected in Control Register 0xn0C. Below is a table showing the RCF modes and input data format that it sets.

**Table I. Serial Data Format**

| 0xn0C Bit 6 | 0xn0C Bit 5 | 0xn0C Bit 4 | Serial Data Word Length | RCF Mode |
|---|---|---|---|---|
| 0 | 0 | 0 | 32 | FIR |
| 0 | 0 | 1 | | π/4-DQPSK |
| 0 | 1 | 0 | | GMSK |
| 0 | 1 | 1 | | MSK |
| 1 | 0 | 0 | 24 (Bit 9 is high) 16 (Bit 9 is low) | FIR, compact |
| 1 | 0 | 1 | | 8-PSK |
| 1 | 1 | 0 | | 3π/8-8-PSK |
| 1 | 1 | 1 | | QPSK |

The serial data input, SDIN, accepts 32-bit words as channel input data. The 32-bit word is interpreted as two 16-bit two's complement quadrature words, I followed by Q, MSB first. This results in linear I and Q data being provided to the RCF. The first bit is shifted into the serial port starting on the next rising edge of SCLK after the SDFO pulse. Figure 16 shows a timing diagram for SCLK master (SCS = 0) and SDFO set for frame request (SFE = 0).

# AD6623
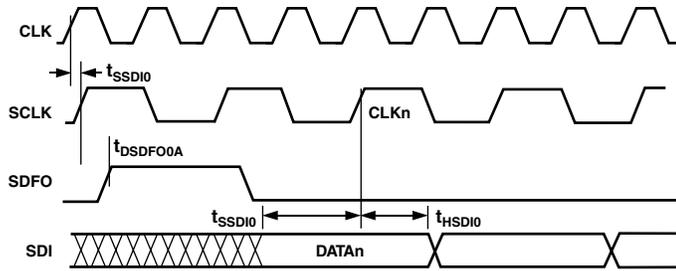


*Figure 16. Serial Port Switching Characteristics*

As an example of the Serial Port operation, consider a CLK frequency of 62.208 MHz and a channel interpolation of 2560. In that case, the input sample rate is 24.3 kSPS (62.208 MHz/2560), which is also the SDFO rate. Substituting, $f_{SCLK} \geq 32 \times f_{SDFO}$ into the equation and solving for SCLKdivider, we find the minimum value for SCLKdivider according to the equation below.

$$SCLKdivider \leq \frac{f_{CLK}}{32 \times f_{SFDO}} \qquad (3)$$

Evaluating this equation for our example, SCLKdivider must be less than or equal to 79. Since the SCLKdivider channel register is a 5-bit unsigned number it can only range from 0 to 31. Any value in that range will be valid for this example, but if it is important that the SDFO period is constant, then there is another restriction. For regular frames, the ratio $f_{SCLK}/f_{SDFO}$ must be equal to an integer of 32 or larger. For this example, constant SDFO periods can only be achieved with an SCLK divider of 31 or less. See Table II for usable SCLK divider values and the corresponding SCLK and $f_{SCLK}/f_{SDFO}$ ratio for the example of L = 2560.

In conclusion, SDFO rate is determined by the AD6623 CLK rate and the interpolation rate of the channel. The SDFO rate is equal to the channel input rate. The channel interpolation is equal to RCF interpolation times CIC5 interpolation, times CIC2 interpolation:

$$\left( L = L_{RCF} \times L_{CIC5} \times \frac{L_{CRIC2}}{M_{CRIC2}} \right) \qquad (4)$$

The SCLK divide ratio is determined by SCLKdivider as shown in the previous equation. The SCLK must be fast enough to input 32 bits of data prior to the next SDFO. Extra SCLKs are ignored by the serial port.

### Table II. Example of Usable SCLK Divider Values and $f_{SCLK}/f_{SDPO}$ Ratios for L = 2560

| SCLKdivisor | $f_{SCLK}/f_{SDFO}$ |
|---|---|
| 0 | 2560 |
| 1 | 1280 |
| 3 | 640 |
| 4 | 512 |
| 7 | 320 |
| 9 | 256 |
| 15 | 160 |
| 19 | 128 |
| 31 | 80 |

## PROGRAMMABLE RAM COEFFICIENT FILTER (RCF)

Each channel has a fully independent RAM Coefficient Filter (RCF). The RCF accepts data from the Serial Port, processes it, and passes the resultant I and Q data to the CIC filter. A variety of processing options may be selected individually or in combination, including PSK and MSK modulation, FIR filtering, all-pass phase equalization, and scaling with arbitrary ramping. See Table III.

### Table III. Data Format Processing Options

| Processing Block | Input Data | Output Data |
|---|---|---|
| Interpolating FIR Filter<br>PSK Modulator | I and Q<br>2 or 3 bits<br>per symbol | I and Q<br><br>Unfiltered I<br>and Q:<br>$\pi/4$-QPSK,<br>8-PSK, or<br>$3\pi/8$-8-PSK |
| MSK Modulator | 1 bit per symbol | Filtered MSK<br>or GSM I and Q |
| QPSK | 2 bits per symbol | Filtered QPSK<br>I and Q |
| All-pass Phase Equalizer | I and Q | I and Q |
| Scale and Ramp | I and Q | I and Q |

## OVERVIEW OF THE RCF BLOCKS

The Serial Port passes data to the RCF with the appropriate format and bit precision for each RCF configuration, see Figure 17. The data may be modulated vectors or unmodulated bits. I and Q vectors are sent directly to the Interpolating Fir Filter. Unmodulated bits may be sent to the PSK Modulator, the Interpolating MSK Modulator, or the Interpolating QPSK Modulator. The PSK Modulator produces unfiltered I and Q vectors at the symbol rate which are then passed through the Interpolating FIR Filter. The Interpolating MSK Modulator and the Interpolating QPSK Modulator produce oversampled, pulse-shaped vectors directly without employing the Interpolating FIR Filter. When possible, the MSK and QPSK modulators are recommended for increased throughput and decreased power consumption compared to Interpolating FIR Filter. In addition, the Interpolating MSK Modulator can realize filters with nonlinear inter-symbol interference, achieving excellent accuracy for GMSK applications.

After interpolation, an optional Allpass Phase Equalizer (APE) can be inserted into the signal path. The APE can realize any real, stable, two-pole, two-zero all-pass filter at the RCF's interpolated rate. This is especially useful to precompensate for nonlinear phase responses of receive filters in terminals, as specified by IS-95. When active, the APE utilizes shared hardware with the interpolating modulators and filter, which may reduce the allowed RCF throughput, inter-symbol interference, or both. See Figure 18.

| 31 | 30 | 29 | 28 | 27 | 26 | 25 | 24 | 23 | 22 | 21 | 20 | 19 | 18 | 17 | 16 | 15 | 14 | 13 | 12 | 11 | 10 | 9 | 8 | 7 | 6 | 5 | 4 | 3 | 2 | 1 | 0 | BIT |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | | | | | | < msb, I, lsb > | | | | | | | | | | | | | | | < msb, Q, lsb > | | | | | | | | | | FIR |

| 23 | 22 | 21 | 20 | 19 | 18 | 17 | 16 | 15 | 14 | 13 | 12 | 11 | 10 | 9 | 8 | 7 | 6 | 5 | 4 | 3 | 2 | 1 | 0 | BIT |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | | | < msb, I, lsb > | | | | | | | | | | | < msb, Q, lsb > | | | | | | | | | COMPACT FIR |

| 15 | 14 | 13 | 12 | 11 | 10 | 9 | 8 | 7 | 6 | 5 | 4 | 3 | 2 | 1 | 0 | BIT |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | < msb, I, lsb > | | | | | | | | < msb, Q, lsb > | | | | | | COMPACT FIR |

| 4 | 3 | 2 | 1 | 0 | BIT |
|---|---|---|---|---|---|
| m | s | D1 | D2 | D0 | 8PSK |

| 4 | 3 | 2 | 1 | 0 | BIT |
|---|---|---|---|---|---|
| m | s | X | D1 | D0 | QPSK |

SERIAL SYNC RAMP

| 4 | 3 | 2 | 1 | 0 | BIT |
|---|---|---|---|---|---|
| M | S | X | X | D0 | MSK/GSM |

| 2 | 1 | 0 | BIT |
|---|---|---|---|
| 0 | D1 | D0 | 8PSK |

| 1 | 0 | BIT |
|---|---|---|
| D1 | D0 | QPSK |

| 0 | BIT |
|---|---|
| D0 | MSK/GSM |

*Figure 17. Data Formats Supported by the AD6623 when SCLK Master (SCS = 0), and SFDO Set for Frame Request (SFE = 0)*
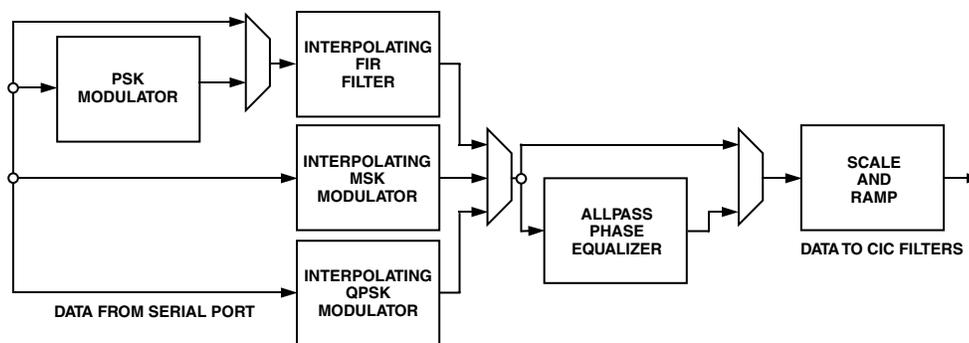


*Figure 18. RCF Block Diagram*

**Table IV. FIR Filter Internal Precision**

| Signal | x × y Notation | Minimum | | Maximum | |
|---|---|---|---|---|---|
| | | Decimal | Hexadecimal (h) | Decimal | Hexadecimal (h) |
| I and Q Inputs | 1.15 | –1.00000 | +1.00000 | 0.999969 | 0.FFFE |
| Coefficients | 1.15 | –1.00000 | +1.00000 | 0.999969 | 0.FFFE |
| Product | 2.18 | –0.99969 | +3.00020 | 1.000000 | 1.00000 |
| Sum | 4.18 | –7.00000 | +8.0000 | 7.999996 | 7.FFFFC |
| FIR Output | 1.17 | –1.00000 | +1.00000 | 0.999992 | 0.FFFF8 |

The Scale and Ramp block adjusts the final magnitude of the modulated RCF output. A synchronization pulse from the SYNC0–3 pins or serial words can be used to command this block to ramp down, pause, and ramp up to a new scale factor. The shape of the ramp is stored in RAM, allowing complete sample by sample control at the RCF interpolated rate. This is particularly useful for time division multiplexed standards such as GSM/EDGE. Modulator configurations can be updated while the ramp is quiet, allowing for GSM and EDGE timeslots to be multiplexed together without resetting or reconfiguring the channel. Each of the RCF processing blocks is discussed in greater detail in the following sections.

**INTERPOLATING FIR FILTER**
The Interpolating FIR Filter realizes a real, sum-of-products filter on I and Q inputs using a single interleaved Multiply-Accumulator (MAC) running at the CLK rate. The input signal is interpolated by integer factors to produce arbitrary impulse responses up to 256 output samples long.

Each bus in the data path carries bipolar two's complement values. For the purpose of discussion, we will arbitrarily consider the radix point positioned so that the input data ranges from –1 to just below 1. In Figure 19, the data buses are marked x × y to denote finite precision limitations. A bus marked x × y has x bits above the radix and y bits below the radix, which implies a range from

# AD6623

$-2^{x-1}$ to $2^{x-1} - 2^{-y}$ in $2^{-y}$ steps. The range limits are tabulated in Table IV for each bus. The hexadecimal values are bit-exact and each MSB has negative weight. Note that the Product bus range is limited by result of the multiplication and the two most significant bits are the same except in one case.
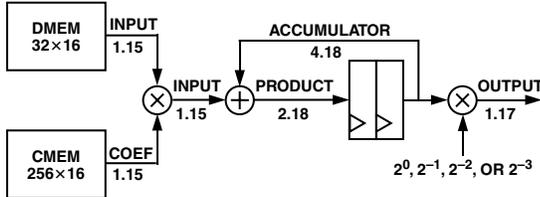


*Figure 19.  Interpolating FIR Filter Block Diagram*

The RCF realizes a FIR filter with optional interpolation. The FIR filter can produce impulse responses up to 256 output samples long. The FIR response may be interpolated up to a factor of 256, although the best filter performance is usually achieved when the RCF interpolation factor ($L_{RCF}$) is confined to eight or below. The $256 \times 16$ coefficient memory (CMEM) can be divided among an arbitrary number of filters, one of which is selected by the Coefficient Offset Pointer (channel address 0x0B). The polyphase implementation is an efficient equivalent to an integer up-sampler followed FIR filter running at the interpolated rate.

The AD6623 RCF realizes a sum-of-products filter using a polyphase implementation. This mode is equivalent to an interpolator followed by a FIR filter running at the interpolated rate. In the functional diagram below, the interpolating block increases the rate by the RCF interpolation factor ($L_{RCF}$) by inserting $L_{RCF}-1$ zero valued samples between every input sample. The next block is a filter with a finite impulse response length ($N_{RCF}$) and an impulse response of h[n], where n is an integer from 0 to $N_{RCF}-1$.

The difference equation for Figure 20 is written below, where h[n] is the RCF impulse response, b[n] is the interpolated input sample sequence at point 'b' in the diagram above, and c[n] is the output sample sequence at point 'c' in Figure 20.
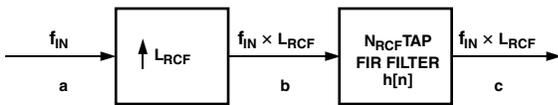


*Figure 20.  RCF Interpolation*

$$c[n] = \sum_{k=0}^{N_{RCF}-1} h[n] \times b[n-k] \qquad (5)$$

This difference equation can be described by the transfer function from point 'b' to 'c' as:

$$H_{bc}(z) = \sum_{k=0}^{N_{RCF}-1} h[n] \times z^{-1} \qquad (6)$$

The actual implementation of this filter uses a polyphase decomposition to skip the multiply-accumulates when b[n–k] is zero. Compared to the diagram above, this implementation has the benefits

of reducing by a factor of $L_{RCF}$ both the time needed to calculate an output and the required data memory (DMEM). The price of these benefits is that the user must place the coefficients into the coefficient memory (CMEM) indexed by the interpolation phase. The process of selecting the coefficients and placing them into the CMEM is broken into three steps shown below.

The FIR accepts two's complement I and Q samples from the serial port with a fixed-point resolution of 16 bits each. When the serial port provides data with less precision, the LSBs are padded with zeroes.

The Data-Mem stores the most recent 16 I and Q pairs for a total of 32 words. The size of the Data-Mem limits the RCF impulse response to $16 \times L_{RCF}$ output samples. When the data words from the Serial Port have fewer than 16 bits, the LSBs are padded with zeroes. The Data-Mem can be accessed through the Microport from 0x20 to 0x5F above the processing channel's base internal address, while the channel's Prog bit is set (external address 4). In order to avoid start-up transients, the Data-Mem should be cleared before operation. The Prog bit must then be reset to enable normal operation.

The Coef-Mem stores up to 256 16-bit filter coefficients. The Coef-Mem can be accessed through the Microport from 0x800 to 0x8FF above the processing channel's base internal address, while the channel's Prog bit is set (external address 4). For AD6622 compatibility, the lower 128 words are also mirrored from 0x080 to 0x0FF above the processing channel's base internal address, while the Prog bit is set. To avoid start-up transients, the Data-Mem should be cleared before operation. The Prog bit must then be reset to enable channel operation.

There is a single Multiply-Accumulator (MAC) on which both the I and Q operations must be interleaved. Two CLK cycles are required for the MAC to multiply each coefficient by an I and Q pair. The MAC is also used for four additional CLK cycles if the All-pass Phase Equalizer is active.

The size of the Data-Mem and Coef-Mem combined with the speed of the MAC determine the total number of the taps per phase ($T_{RCF}$) that may be calculated. $T_{RCF}$ is the number of RCF input samples that influence each RCF output sample. The maximum available $T_{RCF}$ is calculated by the equation below.

$$T_{RCF} \leq least\ of\left(16, floor\left(\frac{256}{L_{RCF}}\right), floor\left(\frac{f_{CLK}}{2 \times f_{SDO}} - 2 \times APE\right)\right) \quad (7)$$

The impulse response length at the output of the RCF is determined by the product of the number of interfering input samples ($T_{RCF}$) and the RCF interpolation factor ($L_{RCF}$), as shown by equation (8) below. The values of $N_{RCF}$ and $T_{RCF}$ are programmed into control registers. $L_{RCF}$ is not a control register, but $N_{RCF}$ and $T_{RCF}$ must be set so that $L_{RCF}$ is an integer. If the integer interpolation by the RCF results in an inconvenient sample rate at the output of the RCF, the desired output rate can usually be achieved by selecting non-integer interpolation in the resampling CIC[2] filter.

$$N_{RCF} = T_{RCF} \times L_{RCF} \qquad (8)$$

**Table V.  RCF Control Registers**

| Channel Address | Bit Width | Description |
|---|---|---|
| 0x0A | 16 | 15–8: $N_{RCF}-1$ B; 7–0: $N_{RCF}-1$ A |
| 0x0B | 8 | 7–0: $O_{RCF}$ |
| 0x0C | 10 | 9: Ch.  A Compact FIR Input Word Length |
|  |  |     0: 16 bits–8 I followed by 8 Q |
|  |  |     1: 24 bits–12 I followed by 12 Q |
|  |  | 8: Ch. A RCF PRBS Enable |
|  |  | 7: Ch A RCF PRBS Length |
|  |  |     0: 15 |
|  |  |     1: 8,388,607 |
|  |  | 6–4: Ch. A RCF Mode Select |
|  |  | 000 = FIR |
|  |  | 001 = p/4-DQPSK Modulator |
|  |  | 010 = GMSK Look-Up Table |
|  |  | 011 = MSK Look-Up Table |
|  |  | 100 = FIR compact mode |
|  |  | 101 = 8-PSK |
|  |  | 110 = 3p/8-8PSK Modulator |
|  |  | 111 = QPSK Look-Up Table |
|  |  | 3–0: Ch. A RCF Taps per Phase |
| 0x0D | 8 | 7–6: RCF Coarse Scale (g): |
|  |  |     00 = 0 dB |
|  |  |     01 = –6 dB |
|  |  |     10 = –12 dB |
|  |  |     11 = –18 dB |
|  |  | 5: Ch.  A Allpass Ph. Eq. Enable |
|  |  | 4–0: Serial Clock Divider (1, ..., 32) |
| 0x0E | 16 | 15–2: Ch. A Unsigned Scale Factor |
|  |  | 1–0: Reserved |
| 0x0F | 18 | 17–16: Ch. A Time Slot Sync Select |
|  |  |     00: Sync0 (See 0x001 Time Slot) |
|  |  |     01: Sync1 |
|  |  |     10: Sync2 |
|  |  |     11: Sync3 |
|  |  | 15–0: Ch. A RCF Scale Hold-Off Counter |
|  |  | 1) Ramp Down (if Ramp is enabled) |
|  |  | 2) Update Scale and Mode |
|  |  | 3) Ramp Up (if Ramp is enabled) |
| 0x110 | 16 | 15–0: Ch. A RCF Phase EQ Coef1 |
| 0x111 | 16 | 15–0: Ch. A RCF Phase EQ Coef2 |
| 0x112 | 16 | 15–0: Ch. A RCF MPSK Magnitude 0 |
| 0x113 | 16 | 15–0: Ch. A RCF MPSK Magnitude 1 |
| 0x114 | 16 | 15–0: Ch. A RCF MPSK Magnitude 2 |
| 0x115 | 16 | 15–0: Ch. A RCF MPSK Magnitude 3 |
| 0x116 | 8 | 7: Reserved |
|  |  | 6: Ch. A Serial Data Frame Select |
|  |  |     0: Serial Data Frame Request |
|  |  |     1: Serial Data Frame End |

| Channel Address | Bit Width | Description |
|---|---|---|
|  |  | 5: Ch. A External SDFI Select |
|  |  | 0: Internal SDFI |
|  |  | 1: External SDFI |
|  |  | 4: Ch. A SCLK Slave Select |
|  |  | 0: Master |
|  |  | 1: Slave |
|  |  | 3: Ch.  A Serial Fine Scale Enable |
|  |  | 2: Ch. A Serial Time Slot Sync Enable (ignored in FIR mode) |
|  |  | 1: Ch. A Ramp Interpolation Enable |
|  |  | 0: Ch. A Ramp Enable |
| 0x117 | 6 | 5–0: Ch. A Mode 0 Ramp Length, R0–1 |
| 0x118 | 6 | 5–0: Ch. A Mode 1 Ramp Length, R1–1 |
| 0x119 | 5 | 4–0: Ch. A Ramp Rest Time, Q |
| 0x11A–0x11F |  | Reserved |
| 0x120–0x13F | 16 | 15–0: Ch. A Data Memory |
| 0x140–0x17F | 16 | 15–14: Reserved |
|  |  | 13–0: Ch. A Power Ramp Memory |
| 0x180–0x1FF | 16 | 15–0: Ch. A Coefficient Memory |
|  |  | This address is mirrored at 0x900–0x97F and contiguously extended at 0x980–0x9FF |

**PSK MODULATOR**

The PSK Modulator is an AD6623 extension feature that is only available when the control register bit 0x000:7 is high. The PSK Modulator creates 32-bit complex inputs to the Interpolating FIR Filter from two or three data bits captured by the serial port. The FIR Filter operates exactly as if the 32-bit word came directly from the serial port. There are three PSK modulation options to choose from: π/4-DQPSK, 8-PSK, and 3π/8-8-PSK. Every symbol of any of these modulations can be represented by one of the 16 phases shown in Figure 21.
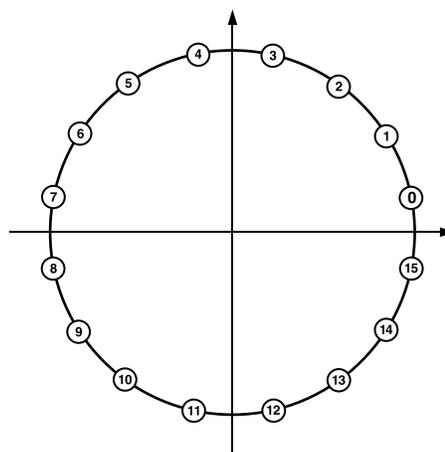


*Figure 21.  16-Phase Modulations*

# AD6623

All of these phase locations are represented in rectangular coordinates by only four unique magnitudes in the positive and negative directions. These four values are read from four channel registers that are programmed according to the following table, which gives the generic formulas and a specific example. The example is notable because it is only 0.046 dB below full-scale and the 16-bit quantization is so benign at that magnitude, that the rms error is better than –122 dBc. It is also worth noting that because none of the phases are aligned with the axes, magnitudes slightly beyond 0.16 dB above full-scale are achievable.

**Table VI. Program Registers**

| Channel Register | Magnitude M | Magnitude E 0x7F53 |
|---|---|---|
| 0x12 | M 3 cos(p/16) | 0x7CE1 |
| 0x13 | M 3 cos(3p/16) | 0x69DE |
| 0x14 | M 3 cos(5p/16) | 0x46BD |
| 0x15 | M 3 cos(7p/16) | 0x18D7 |

Using the four channel registers from the preceding table, the PSK Modulator assembles the 16 phases according to Table VII.

**Table VII. PSK Modulator Phase**

| Phase | I Value | Q Value |
|---|---|---|
| 0 | 0x12 | 0x15 |
| 1 | 0x13 | 0x14 |
| 2 | 0x14 | 0x13 |
| 3 | 0x15 | 0x12 |
| 4 | –0x15 | 0x12 |
| 5 | –0x14 | 0x13 |
| 6 | –0x13 | 0x14 |
| 7 | –0x12 | 0x15 |
| 8 | –0x12 | –0x15 |
| 9 | –0x13 | –0x14 |
| 10 | –0x14 | –0x13 |
| 11 | –0x15 | –0x12 |
| 12 | 0x15 | –0x12 |
| 13 | 0x14 | –0x13 |
| 14 | 0x13 | –0x14 |
| 15 | 0x12 | –0x15 |

The following three sections show how the phase values are created for each PSK modulation mode.

## π/4-DQPSK Modulation

IS-136 compliant π/4-DQPSK modulation is selected by setting the channel register 0x0C: 6–4 to 001b. The phase word is calculated according to the following diagram. The two LSBs of the serial input word update the payload bits once per symbol. The QPSK Mapper creates a data dependent static phase word (Sph) which is added to a time dependent rotating phase word (Rph). The Rph starts at zero when the RCF is reset or switches modes via a sync pulse. Otherwise, the Rph increments by two on every symbol.
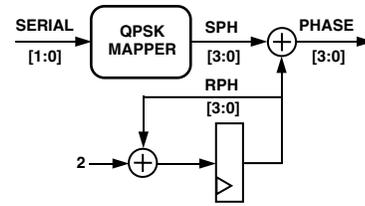


*Figure 22. QPSK Mapper*

The Sph word is calculated by the QPSK Mapper according to the following truth table.

**Table VIII. QPSK Mapper Truth Table**

| Serial [1:0] | Sph [3:0] |
|---|---|
| 00b | 0 |
| 01b | 4 |
| 11b | 8 |
| 10b | 12 |

## 8-PSK Modulation

IS-136+ compliant 8-PSK modulation is selected by setting the channel register 0x0C: 6–4 to 101b. The Phase word is calculated according to the following diagram. The three LSBs of the serial input word update the payload bits once per symbol.



*Figure 23. 8-PSK Mapper*

The Phase word is calculated by the 8-PSK Mapper according to the following truth table:

**Table IX. 8-PSK Mapper Truth Table**

| Serial [2:0] | Sph [3:0] |
|---|---|
| 111b | 0 |
| 011b | 2 |
| 010b | 4 |
| 000b | 6 |
| 001b | 8 |
| 101b | 10 |
| 100b | 12 |
| 110b | 14 |

## 3 π/8-8-PSK Modulation

EDGE compliant 3 π/8-8-PSK modulation is selected by setting the channel register 0x0C: 6–4 to 110b. The phase word is calculated according to the following diagram. The three LSBs of the serial input word update the payload bits once per symbol. The 8-PSK Mapper creates a data-dependent static phase word (Sph) which is added to a time-dependent rotating phase word (Rph). The 8-PSK Mapper operates exactly as described in the preceding 8-PSK Modulation section. The Rph starts at zero when the RCF is reset or switches modes via a sync pulse. Otherwise, the Rph increments by three on every symbol.
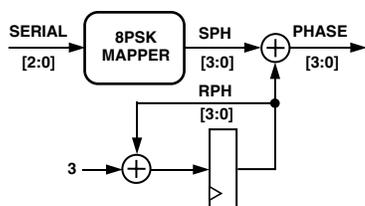
Figure 24. 3 π/8-8-PSK Mapper

**MSK Look-Up Table**
The MSK Look-Up Table mode for the RCF is selected in Control Register 0x10C. In the MSK Mode, the RCF performs arbitrary pulse-shaping based on four symbols of impulse response. For the MSK Mode (3, [16] bits), the serial input format is 11 bits of scaling (MSB first) followed by 1 bit of data. The 11 bits can be used to scale the input data.

**GMSK Look-Up Table**
The GMSK Look-Up Table mode for the RCF is selected in Control Register 0x10C. In the GMSK Mode, the RCF performs arbitrary pulse-shaping based on four symbols of impulse response. For the GMSK Mode (3, [16] bits), the serial input format is 11 bits of scaling (MSB first) followed by 1 bit of data. The 11 bits can be used to scale the input data.

**QPSK Look-Up Table**
The QPSK Filter mode for the RCF is selected in Control Register 0xX0C. In the QPSK Mode, the RCF performs baseband linear pulse-shaping based on filter impulse response up to 12 symbols. For the QPSK Mode (3, [16]bits), the serial input format is 13 bits of scaling (MSB first) followed by one bit I and then one bit Q. The 13 bits can be used to scale the input data.

**PHASE EQUALIZER**
The IS-95 Standard includes a phase equalizer after matched filtering at the baseband transmit side of a base station. This filter pre-distorts the transmitted signal at the base station in order to compensate for the distortion introduced to the received signal by the analog baseband filtering in a handset. The AD6623 includes this functionality in the form of an Infinite Impulse Response (IIR) all-pass filter in the RCF. This Phase Equalizer pre-distort filter has the following transfer function:

$$H(z) = \frac{Y(z)}{X(z)} = \frac{1 + b1z + b2z^2}{z^2 + b1z + b2} \qquad (9)$$
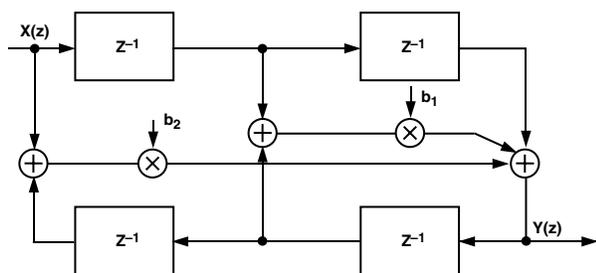


Figure 25. Second Order All-Pass IIR Filter

The Phase Equalizer is enabled/disabled in Control Register 0xn0D Bit 5. The coefficients $b_1$ and $b_2$ are located in Control Registers 0xn10 and 0xn11 respectively

The format for $b_1$ and $b_2$ is two's complement fractional binary with a range of (–2, 2). With one bit for sign at most significant bit position there are 15 bits for magnitude. The value of one bit is

$(2^{-15}) \times 2$, or 0.00006103515625. The register values, in hexadecimal, and the corresponding coefficient weight from positive full-scale through zero to negative full-scale is illustrated in Table X.

**Table X. Coefficient Weights**

| Register Value | Coefficient Weight |
|---|---|
| 0x7FFF | 1.999938964844 |
| .. | |
| 0x0001 | 0.00006103515625 |
| 0x0000 | 0 |
| 0xFFFF | –0.00006103515625 |
| .. | |
| 0x8001 | –1.999938964844 |
| 0x8000 | –2 |

Table XI shows the recommended $b_1$ and $b_2$ coefficients for the respective oversampling rate.

**Table XI. $b_1$ and $b_2$ Coefficients**

| Oversampling | $b_0$ | $b_1$ | $b_2$ |
|---|---|---|---|
| 4 | 1 | –1.45514 | 0.57832 |
| 5 | 1 | –1.56195 | 0.64526 |
| 6 | 1 | –1.63412 | 0.69414 |
| 8 | 1 | –1.72513 | 0.76047 |

**SCALE AND RAMP**
Scale factors can be range from 0 to [CHF]–1/[CHF] with a resolution 1/[CHF]

**FINE SCALING**
AD6623 allows fine scaling of the RCF output signal. A scale factor of 12 to 14 bits is available through the Microport. The Microport fine scale factor is located in Channel Register 0xn0E.

**RCF POWER RAMPING**
The output of the RCF will be multiplied by a 14-bit ramping profile before entering the CIC filters. It is a RAM programmable engine that starts indexing through ramping coefficients when the RAMP bit works its way through the chain. It will then count a programmable number of samples and then RAMP down in reverse order. This will allow the ramping values to update at a modest rate relative to the DAC and still contain the spectral leakage associated with the ramping. A user should provide through the MicroPort the ramping coefficient values, the number of samples to ramp up, and one bit to define the air-interface standard. The programmable power ramp up/down unit allows power ramping on time-slot basis as specified from some wireless transmission technologies (e.g, TDMA).

**CASCADED INTEGRATOR COMB (CIC) INTERPOLATING FILTERS**
The I and Q outputs of the RCF stage are interpolated by two cascaded integrator comb (CIC) filters. The CIC section is separated into three discrete blocks: a fifth order filter (CIC5), a second order resampling filter (rCIC2), and a scaling block (CIC Scaling). The CIC5 and rCIC2 blocks each exhibit a gain that changes with respect to their rate change factors, $L_{rCIC2}$, $M_{rCIC2}$ and $L_{CIC5}$. The product of these gains must be compensated for in a shared CIC Scaling block and can be done to within 6 dB The remaining compensation can come from the RCF (in the form of coefficient scaling) or the fine scaling unit.

# AD6623

## CIC Scaling

The scale factor $S_{CIC}$ is a programmable unsigned integer between 4 and 32. This is a combined scaler for the CIC5 and rCIC2 stages. The overall gain of the CIC section is given by the equation below

$$CIC\_Gain = L_{CIC5}{}^4 \times L_{rCIC2} \times 2^{-S_{CIC}} \tag{10}$$

## CIC5

The first CIC filter stage, the CIC5, is a fifth order interpolating cascaded integrator comb whose impulse response is completely defined by its interpolation factor, $L_{CIC5}$. The value $L_{CIC5}-1$ can be independently programmed for each channel at location 0xn09.
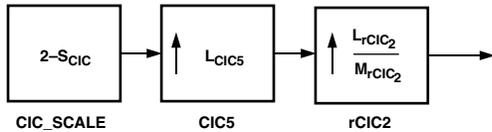
*Figure 26. CIC5*

While this control register is 8 bits wide, $L_{CIC5}$ should be confined to the range from 1 to 32 to avoid the possibility of internal overflow for full scale inputs. The output rate of this stage is given by the equation below.

$$f_{CIC2} = f_{CIC5} \times L_{CIC5} \tag{11}$$

The transfer function of the CIC5 is given by the following equations with respect to the CIC5 output sample rate, $f_{samp5}$.

$$CIC5(z) = \left( \frac{1 - z^{-L_{CIC5}}}{1 - z^{-1}} \right)^5 \tag{12}$$

The SCIC value can be independently programmed for each channel at Control Register 0xn06. $S_{CIC}$ may be safely calculated according to equation (13) below to ensure the net gain through the CIC stages.

SCIC serves to frame which bits of the CIC output are transferred to the NCO stage. This results in controlling the data out of the CIC stages in 6 dB Increments. For the best dynamic range, $S_{CIC}$ should be set to the smallest value possible (lowest attenuation) without creating an overflow condition. This can be safely accomplished using the equation below. To ensure the CIC output data is in range, equation (13) must always be met. The maximum total interpolation rate may be limited by the amount of scaling available.

$$S_{CIC} \geq ceil\left( 4 \times \log_2\left( L_{CIC5} \right) + \log_2\left( L_{CIC2} \right) \right) \tag{13}$$

$$0 \leq S_{CIC} \leq 58 \tag{14}$$

This polynomial fraction can be completely reduced as follows demonstrating a finite impulse response with perfect phase linearity for all values of $L_{CIC5}$.

$$CIC5(z) = \left( \sum_{k=0}^{L_{CIC5}-1} z^{-k} \right)^5 = \sum_{k=1}^{L_{CIC5}-1}\left( z^{-1} - e^{j2\pi \frac{k}{L_{CIC5}}} \right)^5 \tag{15}$$

The frequency response of the CIC5 can be expressed as follows. The initial $1/L_{CIC5}$ factor normalizes for the increased rate, which is appropriate when the samples are destined for a DAC with a zero order hold output. The maximum gain is $L_{CIC5}{}^4$ at

baseband, but internal registers peak in response to various dynamic inputs. As long as $L_{CIC5}$ is confined to 32 or less, there is no possibility of overflow at any register.

$$CIC5(f) = \frac{1}{L_{CIC5}} \left( \frac{sin\left( \pi \frac{L_{CIC5} \times f}{f_{CIC5}} \right)}{sin\left( \pi \frac{f}{f_{CIC5}} \right)} \right)^5 \tag{16}$$

The pass band droop of CIC5 should be calculated using this equation and can be compensated for in the RCF stage. The gain should be calculated from the CIC scaling section above.

As an example, consider an input from the RCF whose bandwidth is 0.141 of the RCF output rate, centered at baseband. Interpolation by a factor of five reveals five images, as shown below.
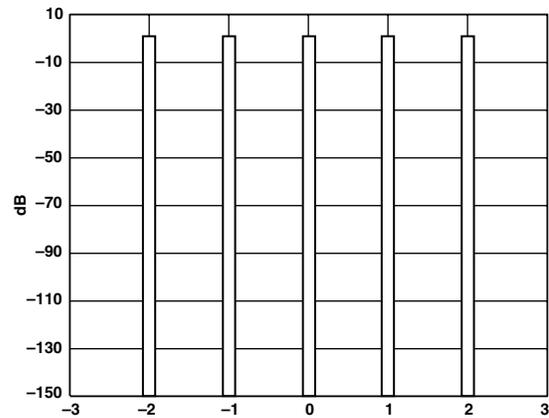
*Figure 27. Interpolation Images*

The CIC5 rejects each of the undesired images while passing the image at baseband. The images of a pure tone at channel center (DC) are nulled perfectly, but as the bandwidth increases the rejection is diminished. The lower band edge of the first image always has the least rejection. In this example, the CIC5 is interpolating by a factor of five and the input signal has a bandwidth of 0.141 of the RCF output sample rate. The plot below shows −110 dBc rejection of the lower band edge of the first image. All other image frequencies have better rejection.

*Figure 28. −110 dBc Rejection*

Table XII lists maximum bandwidth that will be rejected to various levels for CIC5 interpolation factors from 1 to 32. The example above corresponds to the listing in the –110 dB column and the $L_{CIC5} = 5$ row. It is worth noting here that the rejection of the CIC5 improves as the interpolation factor increases.

**Table XII. Max Bandwidth of Rejection for $L_{CIS}$ Values**

| –110 dB | –100 dB | –90 dB | –80 dB | –70 dB |
|---------|---------|--------|--------|--------|
| Full | Full | Full | Full | Full |
| 0.101 | 0.127 | 0.160 | 0.203 | 0.256 |
| 0.126 | 0.159 | 0.198 | 0.246 | 0.307 |
| 0.136 | 0.170 | 0.211 | 0.262 | 0.325 |
| 0.136 | 0.175 | 0.217 | 0.269 | 0.333 |
| 0.143 | 0.178 | 0.220 | 0.272 | 0.337 |
| 0.144 | 0.179 | 0.222 | 0.275 | 0.340 |
| 0.145 | 0.180 | 0.224 | 0.276 | 0.341 |
| 0.146 | 0.181 | 0.224 | 0.277 | 0.342 |
| 0.146 | 0.182 | 0.225 | 0.278 | 0.343 |
| 0.147 | 0.182 | 0.226 | 0.278 | 0.344 |
| 0.147 | 0.182 | 0.226 | 0.279 | 0.344 |
| 0.147 | 0.183 | 0.226 | 0.279 | 0.345 |
| 0.147 | 0.183 | 0.226 | 0.279 | 0.345 |
| 0.148 | 0.183 | 0.227 | 0.280 | 0.345 |
| 0.148 | 0.183 | 0.227 | 0.280 | 0.345 |
| 0.148 | 0.183 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.183 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.183 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.280 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.281 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.281 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.281 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.281 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.281 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.281 | 0.346 |
| 0.148 | 0.184 | 0.227 | 0.281 | 0.346 |
| 0.148 | 0.184 | 0.228 | 0.281 | 0.346 |

### rCIC2

The rCIC2 filter is a second order Cascaded Resampling Integrator Comb filter whose impulse response is completely defined by its rate change factors, $L_{rCIC2}$ and $M_{rCIC2}$. The resampler is implemented using a unique technique that does not require a high-speed clock (but it's still completely jitter-free), thus simplifying the design and saving power. The resampler allows for noninteger relationships between the input data rate and the master clock. This allows easier implementation of systems that are either multimode or require a master clock that is not a multiple of the input data rate to be used.

The value of $L_{rCIC2}$ is limited from 1 to 4096 subject to some restrictions based upon the chosen LCIC5 value as shown in Table XIII.

**Table XIII. Maximum Permissible $L_{RC1C2}$ Values**

| Chosen $L_{CIC5}$ Value | Maximum Allowed $L_{rCIC2}$ |
|---|---|
| 32 | 4096 |
| 31 | 1162 |
| 24–30 | $L_{rCIC2} = 2^{[30 - 4\log_2(L_{CIC3})]}$ |
| 23 | 3836 |
| 1–22 | 4096 |

Two parameters determine the rate change in this block. They are the interpolation factor ($L_{rCIC2}$, 12 bits) and the decimation factor (M, 9 bits). When combined, and subject to the maximum values of $L_{rCIC2}$, the total rate change can be any fraction in the form of:

$$R_{rCIC2} = \frac{L}{M} \geq 1, 1 \leq L \leq 4096, 1 \leq M \leq 512 \qquad (17)$$

The only constraint is that the ratio L/M must be greater than or equal to one. This implies that the rCIC2 has a net interpolation of 1 or more.

Resampling is implemented by apparently increasing the input sample rate by the factor L, using zero stuffing for the new data samples. Following the resampler is a second order cascaded integrator comb filter. Filter characteristics are determined only by the fractional rate change (L/M).

The filter can produce output signals at the full CLK rate of the AD6623. The output rate of this stage is given by the equation below.

$$f_{out} = \frac{L_{rCIC2}}{M_{rCIC2}} f_{rCIC2} \qquad (18)$$

Both $L_{rCIC2}$ and $M_{rCIC2}$ are unsigned integers. The interpolation rate ($L_{rCIC2}$) may be from 1 to 4096 and the decimation ($M_{rCIC2}$) may be between 1 and 512. The stage can be bypassed by setting the L and M to 1.

The transfer function of the rCIC2 is given by the following equations with respect to the rCIC2 output sample rate, $f_{out}$.

$$rCIC2(z) = \left( \frac{1 - z^{-L_{rCIC2}}}{1 - z^{-1}} \right)^2 \qquad (19)$$

The frequency response of the rCIC2 can be expressed as follows. The maximum gain is $L_{rCIC2}$ at baseband. The initial $M_{rCIC2}/L_{rCIC2}$ factor normalizes for the increased rate, which is appropriate when the samples are destined for a DAC with a zero order hold output.

$$rCIC2(f) = \frac{M_{rCIC2}}{L_{rCIC2}} \left( \frac{sin\left( \pi \frac{L_{rCIC2} \times f}{f_{out}} \right)}{sin\left( \pi \frac{f}{f_{out}} \right)} \right)^2 \qquad (20)$$

The pass-band droop of CIC5 should be calculated using this equation and can be compensated for in the RCF stage. The gain should be calculated from the CIC scaling section above.

The values $M_{rCIC2}-1$, $L_{rCIC2}-1$ can be independently programmed for each channel at locations 0xn07, 0xn08. While these control registers are nine bits and 12 bits wide respectively, $M_{rCIC2}-1$ and

# AD6623

$L_{rCIC2}-1$ should be confined to the ranges shown by Table XIII according to the interpolation factor of the CIC5. Exceeding the recommended guidelines may result in overflow for input sequences at or near full scale. While relatively large ratios of $L_{rCIC2}/M_{rCIC2}$ allow for the larger overall interpolations with minimal power consumption, $L_{rCIC2}/M_{rCIC2}$ should be minimized to achieve the best overall image rejection.

As an example, consider an input from the CIC5 whose bandwidth is 0.0033 of the CIC5 rate, centered at baseband. Interpolation by a factor of five reveals five images, as shown below.
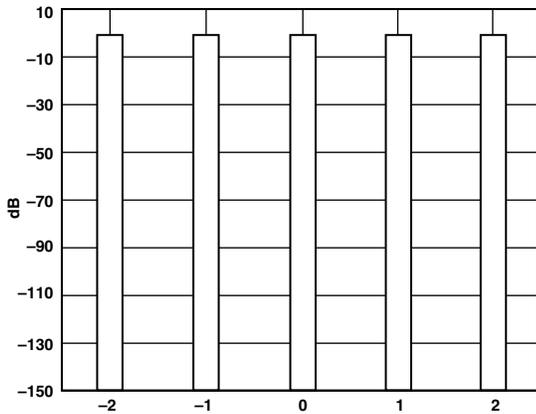
Figure 29. CIC5 Interpolation Images

The rCIC2 rejects each of the undesired images while passing the image at baseband. The images of a pure tone at channel center (DC) are nulled perfectly, but as the bandwidth increases the rejection is diminished. The lower band edge of the first image always has the least rejection. In this example, the rCIC2 is interpolating by a factor of five and the input signal has a bandwidth of 0.0033 of the CIC5 output sample rate. Figure 30 below shows –110 dBc rejection of the lower band edge of the first image. All other image frequencies have better rejection.
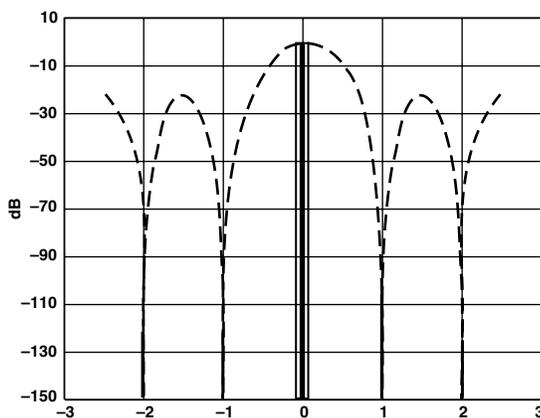
Figure 30. rCIC2 –110 dBc

Table XIV lists maximum bandwidth that will be rejected to various levels for CIC2 interpolation factors from 1 to 32. The example above corresponds to the listing in the –110 dB column and the $L_{CIC2} = 5$ row. The rejection of the CIC2 improves as the interpolation factor increases.

**Table XIV. Maximum Bandwidth of Rejection for $L_{CIC2}$ Values**

| –110 dB | –100 dB | –90 dB | –80 dB | –70 dB |
|---------|---------|--------|--------|--------|
| Full | Full | Full | Full | Full |
| 0.0023 | 0.0040 | 0.0072 | 0.0127 | 0.0226 |
| 0.0029 | 0.0052 | 0.0093 | 0.0165 | 0.0292 |
| 0.0032 | 0.0057 | 0.0101 | 0.0179 | 0.0316 |
| 0.0033 | 0.0059 | 0.0105 | 0.0186 | 0.0328 |
| 0.0034 | 0.0060 | 0.0107 | 0.0189 | 0.0334 |
| 0.0034 | 0.0061 | 0.0108 | 0.0192 | 0.0338 |
| 0.0035 | 0.0062 | 0.0109 | 0.0193 | 0.0341 |
| 0.0035 | 0.0062 | 0.0110 | 0.0194 | 0.0343 |
| 0.0035 | 0.0062 | 0.0110 | 0.0195 | 0.0344 |
| 0.0035 | 0.0062 | 0.0110 | 0.0195 | 0.0345 |
| 0.0035 | 0.0062 | 0.0111 | 0.0196 | 0.0346 |
| 0.0035 | 0.0062 | 0.0111 | 0.0196 | 0.0346 |
| 0.0035 | 0.0063 | 0.0111 | 0.0196 | 0.0347 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0347 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0347 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0348 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0348 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0348 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0348 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0348 |
| 0.0035 | 0.0063 | 0.0111 | 0.0197 | 0.0348 |
| 0.0035 | 0.0063 | 0.0112 | 0.0197 | 0.0348 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |
| 0.0035 | 0.0063 | 0.0112 | 0.0198 | 0.0349 |

## NUMERICALLY CONTROLLED OSCILLATOR/TUNER (NCO)

Each channel has a fully independent tuner. The tuner accepts data from the CIC filter, tunes it to a digital Intermediate Frequency (IF), and passes the result to a shared summation block. The tuner consists of a 32-bit quadrature NCO and a Quadrature Amplitude Mixer (QAM). The NCO serves as a local oscillator and the QAM translates the interpolated channel data from baseband to the NCO frequency. The worst case spurious signal from the NCO is better than –100 dBc for all output frequencies. The tuner can produce real or complex outputs as requested by the shared summation block.

In the complex mode, the NCO serves as a quadrature local oscillator running at $f_{CLK}/2$ capable of producing any frequency step between $-f_{CLK}/4$ and $+f_{CLK}/4$ with a resolution of $f_{CLK}/2^{33}$ (0.0121 Hz for $f_{CLK} = 104$ MHz).

In the real mode, the NCO serves as a quadrature local oscillator running at $f_{CLK}$ capable of producing any frequency step between $-f_{CLK}/2$ and $+f_{CLK}/2$ with a resolution of $f_{CLK}/2^{32}$ (0.0242 Hz for $f_{CLK} = 104$ MHz). The quadrature portion of the output is discarded. Negative frequencies are distinguished from positive
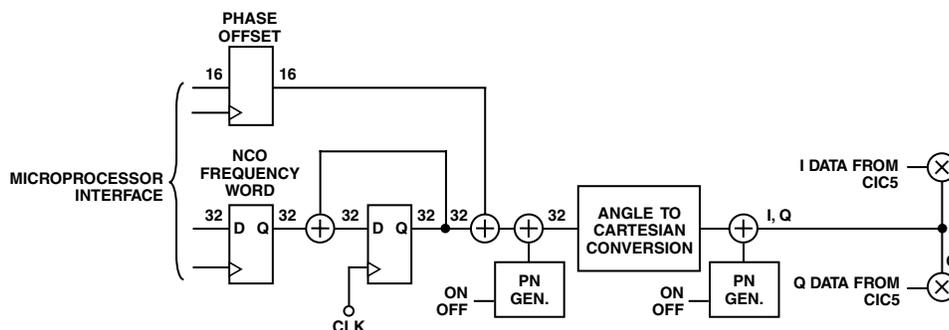
*Figure 31. Numerically Controlled Oscillator and QAM Mixer*

frequencies solely by spectral inversion. The digital IF is calculated using the equation:

$$f_{IF} = f_{NCO} \times \frac{NCO\_frequency}{2^{32}} \qquad (21)$$

where:

$NCO\_frequency$ is the value written to 0xn02,

$f_{IF}$ is the desired intermediate frequency, and

$f_{NCO}$ is $f_{CLK}/2$ for complex outputs and $f_{CLK}$ for real outputs.

**Phase Dither**
The AD6623 provides a phase dither option for improving the spurious performance of the NCO. Phase dither is enabled by writing a "1" to Bit 3 of Channel Register 0xn01. When phase dither is enabled, spurs due to phase truncation in the NCO are randomized. The choice of whether phase dither is used in a system will ultimately be decided by the system goals and the choice of IF frequency. The 18 most significant bits of the phase accumulator are used by the angle to Cartesian conversion. If the NCO frequency has all zeroes below the 18th bit, then phase dither has no effect. If the fraction below the 18th bit is near a 1/2 or 1/3 of the 18th bit, then spurs will accumulate separated from the IF by 1/2 or 1/3 of the CLK frequency. The smaller the denominator of this residual fraction, the larger the spurs due to phase truncation will be. If the phase truncation spurs are unacceptably high for a given frequency, then the phase dither can reduce these at the penalty of a slight elevation in total error energy. If the phase truncation spurs are small, then phase dither will not be effective in reducing them further, but a slight elevation in total error energy will occur.

**Amplitude Dither**
Amplitude dither can also be used to improve spurious performance of the NCO. Amplitude dither is enabled by writing a "1" to Bit 4 of Channel Register at 0xn01. When enabled, amplitude dither can reduce spurs due to truncation at the input to the QAM. If the entire frequency word is close to a fraction that has a small denominator, the spurs due to amplitude truncation will be large and amplitude dither will spread these spurs effectively. Amplitude dither also will increase the total error energy by approximately 3 dB. For this reason amplitude dither should be used judiciously.

**Phase Offset**
The phase offset (Channel Register 0xn04) adds an offset to the phase accumulator of the NCO. This is a 16-bit register that is interpreted as a 16-bit unsigned integer. Phase offset ranges from 0 to nearly $2\pi$ radians with a resolution of $\pi/32768$ radians.

This register allows multiple NCOs to be synchronized to produce sine waves with a known phase relationship.

**NCO Frequency Update and Phase Offset Update Hold-Off Counters**
The update of both the NCO frequency and phase offset can be synchronized with internal Hold-Off counters. Both of these counters are 16-bit unsigned integers and are clocked at the master CLK rate. These Hold-Off counters used in conjunction with the frequency or phase offset registers, allow beam forming and frequency hopping. See the Synchronization section of the data sheet for additional details. The NCO phase can also be cleared on Sync (set to 0x0000) by setting Bit 2 of Channel Register 0xn01 high.

**NCO Control Scale**
The output of the NCO can be scaled in four steps of 6 dB each via Channel Register 0xn01, Bits 1–0. Table XV show a breakdown of the NCO Control Scale. The NCO always has loss to accommodate the possibility that both the I and Q inputs may reach full-scale simultaneously, resulting in a 3 dB input magnitude.

**Table XV. NCO Control Scale**

| 0xn01 Bit 1 | 0xn01 Bit 0 | NCO Output Level |
|---|---|---|
| 0 | 0 | –6 dB (no attenuation) |
| 0 | 1 | –12 dB attenuation |
| 1 | 0 | –18 dB attenuation |
| 1 | 1 | –24 dB attenuation |

**SUMMATION BLOCK**
The Summation Block of the AD6623 serves to combine the outputs of each channel to create a composite multicarrier signal. The four channels are summed together and the result is then added with the 18-bit Wideband Input Bus (IN[17:0]). The final summation is then driven on the 18-bit Wideband Output Bus (OUT[17:0]) on the rising edge of the high speed clock. If the OEN input is high then this output bus is three-stated. If the OEN input is low then this bus will be driven by the summed data. The OEN is active high to allow the Wideband Output Bus to be connected to other busses without using extra logic. Most other busses (like 374 type registers) require a low output enable, which is opposite of the AD6623 OEN, thus eliminating extra circuitry.

The wideband parallel input IN[17:0] is defined as bidirectional, to support dual parallel outputs. Each parallel output produces the sum of two of the four internal TSPs and AD6623 that can

# AD6623

drive two DACs. Channels are added in pairs (A + B), (C + D) as shown in Figure 32.
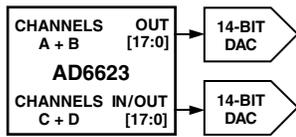


*Figure 32. AD6623 Driving Two DACs*

The Wideband Output Bus may be interpreted as a two's complement number or as an offset binary number as defined by bit 1 of the Summation Mode Control Register at address 0x000. When this bit is high, then the Wideband Output is in two's complement mode and when it is low it is configured for offset binary output data.

The MSB (bit 17) of the Wideband Output Bus is typically used as a guard bit for the purpose of clipping the wideband output bus when bit 0 of the Summation Mode Control Register at address 0x000 is high. If clip detection is enabled then bit 17 of the output bus is not used as a data bit. Instead, bit 16 will become the MSB and is connected to the MSB of the DAC. Configuring the DAC in this manner gives the summation block a gain of 0 dB. When clip detection is not enabled and bit 17 is used as a data bit then the summation block will have a gain of –6.02 dB.

There are two data output modes. The first is offset binary. This mode is used only when driving offset binary DACs. Two's complement mode may be used in one of two circumstances. The first is when driving a DAC that accepts two's complement data. The second is when driving another AD6623 in cascade mode.

When clipping is enabled, the two's complement mode output bus will clip to 0x2FFFF for output signals more positive than the output can express and it will clip to 0x3000 for signals more negative than the output can express. In offset binary mode the output bus will clip to 0x3FFFF for output signals more positive than the output can express and it will clip to 0x2000 for signals more negative than the output can express.

The Wideband Input is always interpreted as an 18-bit two's complement number and is typically connected to the Wideband Output Bus of another AD6623 in order to send more than four carriers to a single DAC. The Output Bus of the preceeding AD6623 should be configured in two's complement mode and clip detection disabled. The 18-bit resolution insures that the noise and spur performance of the wideband data stream does not become the limiting factor as large numbers of carriers are summed.

There is a two-clock cycle latency from the Wideband Input Bus to the Wideband Output Bus. This latency may be calibrated out of the system by use of the Start Hold-Off counter. The preceding AD6623 in a cascaded chain can be started two CLK cycles before the following AD6623 is started and the data from each AD6623 will arrive at the DAC on the same clock cycle. In systems where the individual signals are not correlated, this is usually not necessary.

The AD6623 is capable of outputting both real and complex data. When in Real mode the QIN input is tied low signaling that all inputs on the Wideband Input Bus are real and that all outputs on the Wideband Output Bus are real. The Wideband Input Bus will be pulled low and no data will be added to the composite signal if this port is unused (not connected).

If complex data is desired there are two ways this can be obtained. The first method is to simply set the QIN input of the AD6623 high and to set the Wideband Input Bus low. This allows the AD6623

to output complex data on the Wideband Output Bus. The I data samples would be identified when QOUT is low and the Q data samples would be identified when QOUT is high. The second method of obtaining complex data is to provide a QIN signal that toggles on every rising edge of the CLK. This could be obtained by connecting the QOUT of another AD6623 to QIN as shown in Figure 33. In a cascaded system the QIN of the first AD6623 in the chain would typically be tied high and the QOUT of the first AD6623 would be connected to the QIN of the following part. All AD6623s will synchronize themselves to the QIN input so that the proper samples are always paired and the Wideband Output bus represents valid complex data samples. Table XV shows different parallel input and output data bus formats as a function of QIN and QOUT.

**Table XV.  Valid Output Bus Data Modes**

| QIN | Wideband Input IN[17:0] | Output Data Type OUT[17:0], QOUT |
|---|---|---|
| Low | Real | Real |
| High | Zero | Complex |
| Pulsed | Complex | Complex |
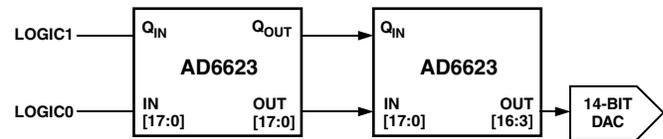


*Figure 33.  Cascade Operation of Two AD6623s*

## SYNCHRONIZATION

Three types of synchronization can be achieved with the AD6623. These are Start, Hop, and Beam. Each is described in detail below. The synchronization is accomplished with the use of a shadow register and a Hold-Off counter. See Figure 34 for a simplistic schematic of the NCO shadow register and NCO Frequency Hold-Off counter to understand basic operation. Enabling the clock (AD6623 CLK) for the Hold-Off counter can occur with either a Soft_Sync (via the micro port), or a Pin Sync (via the AD6623 Sync pin, Pin 62). The functions that include shadow registers to allow synchronization include:

1. Start

2. Hop (NCO Frequency)

3. Beam (NCO Phase Offset)

**Start**

Refers to the start-up of an individual channel, chip, or multiple chips. If a channel is not used, it should be put in the Sleep Mode to reduce power dissipation. Following a hard reset (low pulse on the AD6623 RESET pin), all channels are placed in the Sleep Mode.

**Start With No Sync**

If no synchronization is needed to start multiple channels or multiple AD6623s, the following method should be used to initialize the device.

1. To program a channel, it must first be set to the Program Mode (bit high) and Sleep Mode (bit high) (Ext Address 4). The Program Mode allows programming of data memory and coefficient memory (all other registers are programmable whether in Program Mode or not). Since no synchronization is used, all

Sync bits are set low (External Address 5). All appropriate control and memory registers (filter) are then loaded. The Start Update Hold-Off Counter (0xn00) should be set to 0.

2. Set the appropriate program and sleep bits low (Ext Address 4). This enables the channel. The channel must have Program and Sleep Mode low to activate a channel.
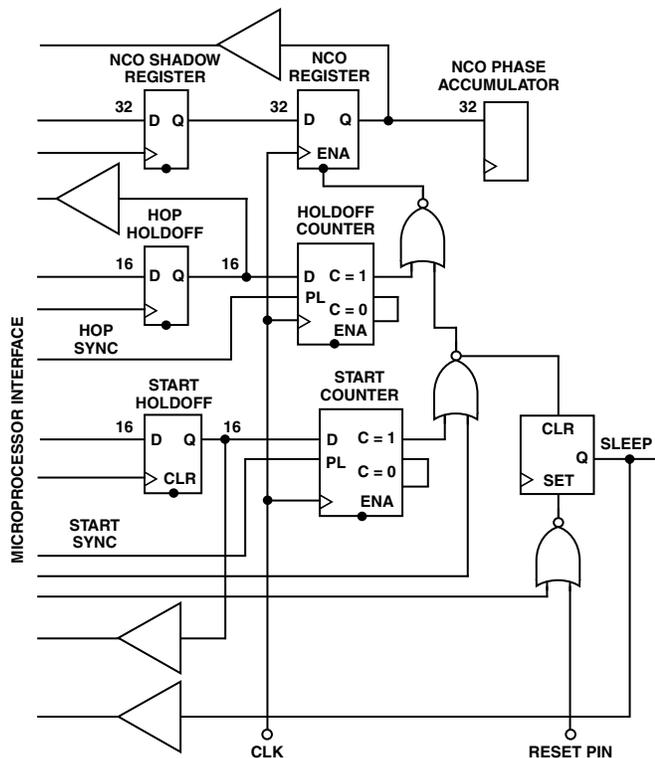


Figure 34. NCO Shadow Register and Hold-Off Counter

**Start with Soft Sync**
The AD6623 includes the ability to synchronize channels or chips under microprocessor control. One action to synchronize is the start of channels or chips. The Start Update Hold-Off Counter (0xn00) in conjunction with the Start bit and Sync bit (Ext Address 5) allow this synchronization. Basically the Start Update Hold-Off Counter delays the Start of a channel(s) by its value (number of AD6623 CLKs). The following method is used to synchronize the start of multiple channels via microprocessor control.

1. Set the appropriate channels to sleep mode (a hard reset to the AD6623 Reset pin brings all four channels up in sleep mode).

2. Write the Start Update Hold-Off Counter(s) (0xn00) to the appropriate value (greater than 1 and less than $2^{16}-1$). If the chip(s) is not initialized, all other registers should be loaded at this step.

3. Write the Start bit and the SyncX(s) bit high (Ext Address 5).

4. This starts the Start Update Hold-Off Counter counting down. The counter is clocked with the AD6623 CLK signal. When it reaches a count of one the sleep bit of the appropriate channel(s) is set low to activate the channel(s).

**Start with Pin Sync**
A Sync pin is provided on the AD6623 to provide the most accurate synchronization, especially between multiple AD6623s. Synchronization of start with an external signal is accomplished with the following method.

1. Set the appropriate channels to sleep mode (a hard reset to the AD6623 Reset pin brings all four channels up in sleep mode).

2. Write the Start Update Hold-Off Counter(s) (0xn00) to the appropriate value (greater than 1 and less than $2^{16}-1$). If the chip(s) is not initialized, all other registers should be loaded at this step.

3. Set the Start on Pin Sync bit and the appropriate Sync Pin Enable high (0xn01).

4. When the Sync pin is sampled high by the AD6623 CLK this enables the count down of the Start Update Hold-Off Counter. The counter is clocked with the AD6623 CLK signal. When it reaches a count of one the sleep bit of the appropriate channel(s) is set low to activate the channel(s).

**Hop**
A jump from one NCO frequency to a new NCO frequency. This change in frequency can be synchronized via microprocessor control or an external Sync signal as described below.

To set the NCO frequency without synchronization the following method should be used.

**Set Frequency No Hop**
1. Set the NCO Frequency Hold-Off counter to 0.

2. Load the appropriate NCO frequency. The new frequency will be immediately loaded to the NCO.

**Hop with Soft Sync**
The AD6623 includes the ability to synchronize a change in NCO frequency of multiple channels or chips under microprocessor control. The NCO Frequency Hold-Off counter (0xn03) in conjunction with the Hop bit and the Sync bit (Ext Address 5) allow this synchronization. Basically the NCO Frequency Hold-Off counter delays the new frequency from being loaded into the NCO by its value (number of AD6623 CLKs). The following method is used to synchronize a hop in frequency of multiple channels via microprocessor control.

1. Write the NCO Frequency Hold-Off (0xn03) counter to the appropriate value (greater than 1 and less then $2^{16}-1$).

2. Write the NCO Frequency register(s) to the new desired frequency.

3. Write the hop bit and the Sync(s) bit high (Ext Address 5).

4. This starts the NCO Frequency Hold-Off counter counting down. The counter is clocked with the AD6623 CLK signal. When it reaches a count of one the new frequency is loaded into the NCO.

**Hop with Pin Sync**
A Sync pin is provided on the AD6623 to provide the most accurate synchronization, especially between multiple AD6623s. Synchronization of hopping to a new NCO frequency with an external signal is accomplished with the following method.

1. Write the NCO Frequency Hold-Off counter(s) (0xn03) to the appropriate value (greater than 1 and less than $2^{16}-1$).

2. Write the NCO Frequency register(s) to the new desired frequency.

3. Set the Hop on Pin Sync bit and the appropriate Sync Pin Enable high (0xn01).

4. When the Sync pin is sampled high by the AD6623 CLK this enables the count down of the NCO Frequency Hold-Off counter. The counter is clocked with the AD6623 CLK signal. When it reaches a count of one the new frequency is loaded into the NCO.

# AD6623

**Beam**
A change in phase for a particular channel and can be synchronized with respect to other channels or AD6623s. This change in phase can be synchronized via microprocessor control or an external Sync signal.

To set the amplitude without synchronization the following method should be used.

**Set Phase No Beam**
1. Set the NCO Phase Offset Update Hold-Off Counter (0xn05) to 0.
2. Load the appropriate NCO Phase Offset (0xn04). The NCO Phase Offset will be immediately loaded.

**Beam with Soft Sync**
The AD6623 includes the ability to synchronize a change in NCO phase of multiple channels or chips under microprocessor control. The NCO Phase Offset Update Hold-Off Counter in conjunction with the Beam bit and the Sync bit (Ext Address 5) allow this synchronization. Basically the NCO Phase Offset Update Hold-Off Counter delays the new phase from being loaded into the NCO/RCF by its value (number of AD6623 CLKs). The following method is used to synchronize a beam in phase of multiple channels via microprocessor control.

1. Write the NCO Phase Offset Update Hold-Off Counter (0xn05) to the appropriate value (greater than 1 and less then $2^{16}-1$).
2. Write the NCO Phase Offset register(s) to the new desired phase and amplitude.
3. Write the beam bit and the Sync(s) bit high (Ext Address 5).
4. This starts the NCO Phase Offset Update Hold-Off Counter counting down. The counter is clocked with the AD6623 CLK signal. When it reaches a count of one the new phase is loaded into the NCO.

**Beam with Pin Sync**
A Sync pin is provided on the AD6623 to provide the most accurate synchronization, especially between multiple AD6623s. Synchronization of beaming to a new NCO Phase Offset with an external signal is accomplished using the following method.

1. Write the NCO Phase Offset Hold-Off (0xn05) counter(s) to the appropriate value (greater than 1 and less than $2^{16}-1$).
2. Write the NCO Phase Offset register(s) to the new desired phase and amplitude.
3. Set the Beam on Pin Sync bit and the appropriate Sync Pin Enable high (0xn01).
4. When the Sync pin is sampled high by the AD6623 CLK this enables the count down of the NCO Phase Offset Hold-Off counter. The counter is clocked with the AD6623 CLK signal. When it reaches a count of one the new phase is loaded into the NCO registers.

**JTAG INTERFACE**
The AD6623 supports a subset of IEEE Standard 1149.1 specification. For additional details of the standard, please see *IEEE Standard Test Access Port and Boundary-Scan Architecture*, IEEE-1149 publication from IEEE.

The AD6623 has five pins associated with the JTAG interface. These pins are used to access the on-chip Test Access Port and are listed in Table XVII.

**Table XVII.  Test Access Port Pins**

| Name | Pin Number | Description |
|------|-----------|-------------|
| $\overline{TRST}$ | 100 | Test Access Port Reset |
| TCK | 101 | Test Clock |
| TMS | 106 | Test Access Port Mode Select |
| TDI | 108 | Test Data Input |
| TDO | 107 | Test Data Output |

Note that TCK and TDI are internally pulled down which is opposite of IEEE Standard 1149.1. These pins may be connected to external pull-up resistors, with the associated additional current draw through the pull-ups, or left unconnected.

The AD6623 supports four op codes are shown in Table XVIII. These instructions set the mode of the JTAG interface.

**Table XVIII.  Op Codes**

| Instruction | Op Code |
|-------------|---------|
| IDCODE | 10 |
| BYPASS | 11 |
| SAMPLE/PRELOAD | 01 |
| EXTEST | 00 |

The Vendor Identification Code (Table XIX) can be accessed through the IDCODE instruction and has the following format.

**Table XIX.   Vendor Identification Code**

| MSB<br>Version | Part<br>Number | Manufacturer<br>ID Number | LSB<br>Mandatory |
|------|------|------|------|
| 0000 | 0010 0111 1000 0000 | 000 1110 0101 | 1 |

A BSDL file for this device is available from Analog Devices, Inc. Contact Analog Devices for more information.

**SCALING**
Proper scaling of the wideband output is critical to maximize the spurious and noise performance of the AD6623. A relatively small overflow anywhere in the data path can cause the spurious free dynamic range to drop precipitously. Scaling down the output levels also reduces dynamic range relative to an approximately constant noise floor. A well-balanced scaling plan at each point in the signal path will be rewarded with optimum performance. The scaling plan can be separated into two parts: multicarrier scaling and single-carrier scaling.

**Multicarrier Scaling**
An arbitrary number of AD6623s can be cascaded to create a composite digital IF with many carriers. As the number of carriers increases, the peak to RMS ratio of the composite digital IF will increase as well. It is possible and beneficial to limit the peak to RMS ratio through careful frequency planning and controlled phase offsets. Nevertheless, in most cases with a large number of carriers, the worst-case peak is an unlikely event.

The AD6623 immediately preceding the DAC can be programmed to clip rather than wrap around (see the Summation Block description). For a large number of carriers, a rare but finite chance of clipping at the AD6623 wideband output will result in superior dynamic range compared to lowering each carrier level until clipping is impossible. This will also be the case for most DACs. Through analysis or experimentation, an optimal output level of individual carriers can be determined for any particular DAC.

**Single-Carrier Scaling**

Once the optimal power level is determined for each carrier, one must determine the best way to achieve that level. The maximum SNR can be achieved by maximizing the intermediate power level at each processing stage. This can be done by assuming the proper level at the output and working along the following path: Summation, NCO, CIC, Ramp, RCF, and finally, Fine Scaler Unit.

The Summation Block is intended to combine multiple carriers with each carrier at least 6 dB below full scale. For this configuration, the AD6623 driving the DAC should have clip detection enabled. OUT17 becomes a clip indicator that reports clipping in both polarities. If the DAC requires offset binary outputs, then the internal offset binary conversion should be enabled as well. Any preceding cascaded AD6623s should disable clip detection and offset binary conversion. The IN17–IN0 of the first AD6623 in the cascade should be grounded. See the Summation Block section for details. In this configuration, intermediate OUT17s will serve as guard bits that allow intermediate sums to exceed full scale. As long as the final output does not exceed 6 dB over full scale, the clip detector will perform correctly.

If a single carrier needs to exceed –6 dB full scale, hardwired scaling can be accomplished according to Table XX. This is most useful when the AD6623 is processing a Single Wideband Carrier such as UMTS or CDMA 2000.

**Table XX. Hardwired Scaling**

| Max. Single Carrier Level | Connect to DAC MSB | Clip Detect | Offset Binary Compensation |
|---|---|---|---|
| –12.04 dB | OUT17 | N/A | Internal |
| –6.02 dB | OUT16 | ± | Internal |
| 0 dB | OUT15 | + only | 0x08000 |
| +6.02 dB | OUT14 | + only | 0x0C000 |

The NCO/Tuner is equipped with an output scaler that ranges from –6.02 dB to –24.08 dB below full scale, in 6.02 dB steps. See the NCO/Tuner section for details. The best SNR will be achieved by maximizing the input level to the NCO and using the largest possible NCO attenuation. For example, to achieve an output level –20 dB below full scale, one should set the CIC output level to –1.94 dB below full scale and attenuate by –18.06 dB in the NCO.

The CIC is equipped with an output scaler that ranges from 0 dB to –186.64 dB below full scale in 6.02 dB steps. This large attenuation is necessary to compensate for the potentially large gains associated with CIC interpolation. See the CIC section for details. For example to achieve an output level of –1.94 dB below full scale, with a CIC5 interpolation of 27 (114.51 dB gain) and a CIC2 interpolation of 3 (9.54 dB gain), one should set the CIC_Scale to 20 and the Fine Scale Unit output level to –5.59 dB below full scale.

$$-1.94 - 9.54 - 114.51 + 20 \times 6.02 = -5.59 \tag{22}$$

The ramp unit when bypassed will have exactly 0 dB of gain and can be ignored. When in use, the gain is dependant on what value is stored in the last valid RMEM location. RMEM words are 14 bits [0–1], so when the value is positive full scale, the gain is about –0.0005 dB; probably neglectable.

The RCF coefficients should be normalized to positive full scale. This will yield the greatest dynamic range. The RCF is equipped with an output scaler that ranges from 0 dB to –18.06 dB below full scale in 6.02 dB steps. This attenuation can be used to partially compensate for filter gain in the RCF. For example, if the maximum gain of the RCF coefficients is 11.26 dB, the RCF coarse scale should be set to 2 (12.04 dB). This yields an RCF output level and fine scale input level of –0.78 dB

$$11.26 - 12.04 = -0.78 \tag{23}$$

The fine scale unit is left to turn a –0.78 dB level into a –5.59 dB level. This requires a gain of –4.81 dB, which corresponds to a 14-bit [0–2] scale value of 1264h. All subsequent rescalings during chip operation should be relative to this maximum.

$$-5.59 - 0.78 = -4.81 \tag{24}$$

$$floor\left(10^{\frac{-4.81}{20}} \times 2^{13}\right) = 1264h \tag{25}$$

Finally, as described in the RCF section, there may be a worst-case peak of a phase that is larger than the channel center gain. In the preceding example, if the worst case to channel center ratio is larger than 4.59 dB (potentially overflowing the RCF), then the RCF_Coarse_Scale should be reduced by one and the CIC_Scale should be increased by one. In the preceding example, if the worst case to channel center ratio is larger than 5.59 dB (potentially overflowing the RCF and CIC), then the RCF_Coarse_Scale should be reduced by one and the NCO_Output_Scale should be increased by one.

**MICROPORT INTERFACE**

The MicroPort interface is the communications port between the AD6623 and the host controller. There are two modes of bus operation: Intel nonmultiplexed mode (INM), and Motorola nonmultiplexed mode (MNM) that is set by hard-wiring the MODE pin to either ground or supply. The mode is selected based on the use of the MicroPort control lines (DS or RD, DTACK or RDY, RW or WR) and the capabilities of the host processor. See the timing diagrams for details on the operation of both modes.

The External Memory Map provides data and address registers to read and write the extensive control registers in the Internal Memory Map. The control registers access global chip functions and multiple control functions for each independent channel.

**MicroPort Control**

All accesses to the internal registers and memory of the AD6623 are accomplished indirectly through the use of the microprocessor port external registers shown in Table XXI. Accesses to the External Registers are accomplished through the 3 bit address bus (A[2:0]) and the 8-bit data bus (D[7:0]) of the AD6623 (MicroPort). External Address [3:0] provides access to data read from or written to the internal memory (up to 32 bits). External Address [0] is the least significant byte and External Address [3] is the most significant byte. External Address [4] controls the Sleep Mode of each channel. External Address [5] controls the sync status of each channel. External Address [7:6] determines the Internal Address selected and whether this address is incremented after subsequent reads and/or writes to the internal registers.

**EXTERNAL MEMORY MAP**

The External Memory Map is used to gain access to the Internal Memory Map described below. External Address [7:6] sets the Internal Address to which subsequent reads or writes will be performed. The top two bits of External Address [7] allow the user to set the address to auto increment after reads, writes, or both. All internal data words have widths that are less than or equal to 32 bits. Accesses to External Address [0] also triggers access to the AD6623's internal memory map. Thus during writes to the

# AD6623

**Table XXI. External Registers**

| External Address | External Data | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| | **D7** | **D6** | **D5** | **D4** | **D3** | **D2** | **D1** | **D0** |
| 7:UAR | Wrinc | Rdinc | – | – | IAII | IAIO | IA9 | IA8 |
| 6:LAR | IA7 | IA6 | IA5 | IA4 | IA3 | IA2 | IA1 | IA0 |
| 5:Sync | – | Beam | Hop | Start | Sync D | Sync C | Sync B | Sync A |
| 4:Sleep | Prog D | Prog C | Prog B | Prog A | Sleep D | Sleep C | Sleep B | Sleep A |
| 3:Byte3 | ID31 | ID30 | ID29 | ID28 | ID27 | ID26 | ID25 | ID24 |
| 2:Byte2 | ID23 | ID22 | ID21 | ID20 | ID19 | ID18 | ID17 | ID16 |
| 1:Byte1 | ID15 | ID14 | ID13 | ID12 | ID11 | ID10 | ID9 | ID8 |
| 0:Byte0 | ID7 | ID6 | ID5 | ID4 | ID3 | ID2 | ID1 | ID0 |

internal registers, External Address [0] must be written last to insure all data is transferred. Reads are the opposite in that External Address [0] must be the first data register read (after setting the appropriate internal address) to initiate an internal access.

External Address [5:4] reads and writes are transferred immediately to internal control registers. External Address [4] is the sleep register. The sleep bits can be set collectively by the address. The sleep bits can be cleared by operation of start syncs (described below).

External Address [5] is the sync register. These bits are write only. There are three types of syncs: start, hop, and beam. Each of these can be sent to any or all of the four channels. For example, a write of X0010100 would issue a start sync to channel C only. A write of X1101111 would issue a beam sync and a hop sync to all channels.

The internal address bus is 12 bits wide and the internal data bus is 32 bits wide. External address 7 is the UAR (Upper Address Register) and stores the upper four bits of the address space in UAR[3:0]. UAR[7:6] define the auto-increment feature. If Bit 6 is high, the internal address is incremented after an internal read. If Bit 7 is high, the internal address is incremented after an internal write. If both bits are high, the internal address in incremented after either a write or a read. This feature is designed for sequential access to internal locations. External address 6 is the LAR (Lower Address Register) and stores the lower 8 bits of the internal address. External addresses 3 through 0 store the 32 bits of the internal data. All internal accesses are two clock cycles long.

Writing to an internal location with a data width of 16 bits is achieved by first writing the upper four bits of the address to bits 3 through 0 of the UAR (bits 7 and 6 of the UAR are written to determine whether or not the auto increment feature is enabled). The LAR is then written with the lower eight bits of the internal address (it doesn't matter if the LAR is written before the UAR as long as both are written before the internal access). Since the data width of the internal address is 16 bits, only data register 1 and data register 0 are needed. Data register 1 must be written first because the write to data register 0 triggers the internal access. Data register 0 must always be the last register written to initiate the internal write.

Reading from the MicroPort is accomplished in a similar manner. The internal address is first written. A read from data register 0 activates the internal read, thus register 0 must always be read first to initiate an internal read. This provides the 8 LSBs of the internal read through the MicroPort (D[7:0]). Additional bytes are then read by changing the external address (A[2:0]) and performing additional reads. If data register 3 (or any other) is read before data register 0, incorrect data will be read. Data register 0 must be read first in order to transfer data from the core memory to the external memory locations. Once the data register is read, the remaining locations may be examined in any order.

Access to the external registers of Table XX is accomplished in one of two modes using the $\overline{CS}$, $\overline{DS}(\overline{RD})$, RW($\overline{WR}$), and $\overline{DTACK}$(RDY) inputs. The access modes are Intel Nonmultiplexed mode and Motorola Nonmultiplexed mode. These modes are controlled by the MODE input (MODE = 0 for INM, MODE = 1 for MNM).

**Intel Nonmultiplexed Mode (INM)**
MODE must be tied low to operate the AD6623 MicroPort in INM mode. The access type is controlled by the user with the chip select ($\overline{CS}$), read ($\overline{RD}$), and write ($\overline{WR}$) inputs. The ready (RDY) signal is produced by the MicroPort to communicate to the user the MicroPort is ready for an access. RDY goes low at the start of the access and is released when the internal cycle is complete. See the timing diagrams for both the read and write modes in the Specifications.

**Motorola Nonmultiplexed Mode (MNM)**
MODE must be tied high to operate the AD6623 MicroPort in MNM mode. The access type is controlled by the user with the chip select ($\overline{CS}$), data strobe ($\overline{DS}$), and read/write (RW) inputs. The data acknowledge ($\overline{DTACK}$) signal is produced by the MicroPort to acknowledge the completion of an access to the user. $\overline{DTACK}$ goes low when an internal access is complete and then will return high after $\overline{DS}$ is deasserted. See the timing diagrams for both the read and write modes in the Specifications.

The $\overline{DTACK}$(RDY) pin is configured as an open drain so that multiple devices may be tied together at the microprocessor/microcontroller without contention.

The MicroPort of the AD6623 allows for multiple accesses while $\overline{CS}$ is held low ($\overline{CS}$ can be tied permanently low if the MicroPort is not shared with additional devices). The user can access multiple locations by pulsing the RW($\overline{WR}$) or $\overline{DS}(\overline{RD})$ lines and changing the contents of the external three bit address bus (A[2:0]).

**External Address 7 Upper Address Register (UAR)**
Sets the four most significant bits of the internal address, effectively selecting channels 1, 2, 3, or 4 (D2:D0). The autoincrement of read and write are also set (D7:D6).

**External Address 6 Lower Address Register (LAR)**
Sets the internal address 8 LSBs (D7:D0).

**External Address 5 Sync**
This register is write only. Bits in this address control the synchronization of the AD6623 channels. If the user intends to bring up channels with no synchronization requirements then all bits of this register should be written low. Two types of sync signals are available with the AD6623. The first is Soft Sync. Soft Sync is software synchronization enabled through the MicroPort. The second synchronization method is Pin Sync. Pin Sync is enabled by a signal

applied to the Sync pin (Pin 62). See the Synchronization section for detailed explanations of the different modes.

## External Address 4 Sleep

Bits in this register determine how the chip is programmed and enables the channels. The program bits (D7:D4) must be set high to allow programming of CMEM and DMEM for each channel. Sleep bits (D3:D0) are used to activate or sleep channels. These can be used manually by the user to bring up a channel by simply writing the required channel high. These bits can also be used in conjunction with the Start and Sync signals available in External Address 5 to synchronize the channels. See the Synchronization section for a detailed explanation of different modes.

## External Address 3:0 (Data Bytes)

These registers return or accept the data to be accessed for a read or write to internal addresses

## INTERNAL COUNTER REGISTERS AND ON-CHIP RAM
## AD6623 and AD6622 Compatibility

The AD6623 functionality is a superset of the AD6622 functionality. The AD6623 is pin-compatible with AD6622.

AD6622 compatibility is selected when bit 7 of Internal Control Register 0x000 is low. In this state, all AD6623 extended control registers are cleared. While in the AD6622 mode the unused AD6623 pins are three-stated.

Listed below is the mapping of internal AD6623 registers. AD6622 compatibility is selected by setting 0x000:7 low. In this state, all AD6623 extended control registers are cleared. Registers marked as "Reserved" must be written low.

**Common Function Registers (not associated with a particular channel)**

| Internal Address | Bit | AD6622 Compatible Description | AD6623 Extensions Description |
|---|---|---|---|
| 0x000 | 7 | AD6623 Extension = 0[1] | AD6623 Extension = 1[1] |
| | 6–5 | Reserved | No Change |
| | 4 | Reserved | Wideband Input Disable[3] |
| | 3 | Reserved | Dual Output Enable[3] |
| | 2 | Reserved | No Change |
| | 1 | Offset Binary Outputs[1] | No Change |
| | 0 | Clip Wideband I/O[3] | No Change |
| 0x001 | 7 | First Sync Only[3] | No Change |
| | 6 | Beam on Pin Sync[3] | No Change |
| | 5 | Hop on Pin Sync[3] | No Change |
| | 4 | Start on Pin Sync[3] | No Change |
| | 3 | Ch. D Sync0 Pin Enable[3] | No Change |
| | 2 | Ch. C Sync0 Pin Enable[3] | No Change |
| | 1 | Ch. B Sync0 Pin Enable[3] | No Change |
| | 0 | Ch. A Sync0 Pin Enable[3] | No Change |
| 0x002 | 23–0 | Unused | BIST Counter[1, 3] |
| 0x003 | 15–0 | Unused | BIST Value (read only) |

**Channel Function Registers (0x1XX = Ch. A, 0x2XX = Ch. B, 0x3XX = Ch. C, 0x4XX = Ch. D)**

| Internal Address | Bit | AD6622 Compatible Description | AD6623 Extensions Description |
|---|---|---|---|
| 0x100 | 17–16 | Unused | Ch. A Start Sync Select[3] |
| | | | 00: Sync0 (See 0x001) |
| | | | 01: Sync1 |
| | | | 10: Sync2 |
| | | | 11: Sync3 |
| | 15–0 | Ch. A Start Hold–Off Counter[3] | No Change |
| 0x101 | 7–5 | Reserved | No Change |
| | 4 | Ch. A NCO Amplitude Dither Enable | No Change |
| | 3 | Ch. A NCO Phase Dither Enable | No Change |
| | 2 | Ch. A NCO Clear Phase Accumulator on Sync | No Change |
| | 1–0 | Ch. A NCO Scale | No Change |
| | | 00: –6 dB | No Change |
| | | 01: –12 dB | No Change |
| | | 10: –18 dB | No Change |
| | | 11: –24 dB | No Change |
| 0x102 | 31–0 | Ch. A NCO Frequency Value[3] | No Change |
| 0x103 | 17–16 | Unused | Ch. A Hop Sync Select[3] |
| | | | 00: Sync0 (See 0x001 Hop) |
| | | | 01: Sync1 |
| | | | 10: Sync2 |
| | | | 11: Sync3 |

# AD6623

**Channel Function Registers (continued)**

| Internal Address | Bit | AD6622 Compatible Description | AD6623 Extensions Description |
|---|---|---|---|
| | 15–0 | Ch. A NCO Frequency Update Hold–Off Counter[3] | No Change |
| 0x104 | 15–0 | Ch. A NCO Phase Offset[2] | No Change[2] |
| 0x105 | 17–16 | Reserved | Ch. A Phase Sync Select[3] |
| | | |    00: Sync0 (See 0x001 Beam) |
| | | |    01: Sync1 |
| | | |    10: Sync2 |
| | | |    11: Sync3 |
| | 15–0 | Ch. A NCO Phase Offset Update Hold–off Counter[3] | No Change |
| 0x106 | 7–5 | Reserved | No Change |
| | 4–0 | Ch. A CIC Scale, $S_{CIC}$ | No Change |
| 0x107 | 8–0 | Reserved | Ch. A CIC2 Decimation, $M_2 - 1$ |
| 0x108 | 11–8 | Reserved | Ch. A CCI2 Interpolation, $L_2 - 1$, extended |
| | 7–0 | Ch. A C1C2 Interpolation, $L_2 - 1$ | No Change |
| 0x109 | 7–0 | Ch. A C1C5 Interpolation, $L_5 - 1$ | No Change |
| 0x10A | 15–8 | Reserved | Ch. A RCF TapsB, $N_{RCF} - 1$ (8 bits)[3] |
| | 7 | Reserved | Ch. A RCF TapsA, $N_{RCF} - 1$ (new MSB)[2] |
| | 6–0 | Ch. A RCF TapsA, $N_{RCF} - 1$ (7 bits)[1] | No Change[2] |
| 0x10B | 7 | Reserved | Ch. A RCF Coef Offset, $O_{RCF}$ (new MSB)[2] |
| | 6–0 | Ch. A RCF Coefficient Offset, $O_{RCF}$ (7 bits)[1] | No Change[2] |
| 0x10C | 15–10 | Unused | Reserved |
| | 9 | Unused | Ch. A Compact FIR Input Word Length |
| | | |    0: 16 bits—8 I followed by 8 Q |
| | | |    1: 24 bits—12 I followed by 12 Q |
| | 8 | Unused | Ch. A RCF PRBS Enable |
| | 7 | Ch. A PRBS Length[3] | Ch. A RCF PRBS Length[3] |
| | |    0: 15 |    0: 15 |
| | |    1: 8,388,607 |    1: 8,388,607 |
| | 6 | Ch. A RCF PRBS Enable | Ch. A RCF Mode Select (1 of 3)[2] |
| | 5 | Ch. A RCF Mode Select (1 of 2)[3] | Ch. A RCF Mode Select (2 of 3)[2] |
| | 4 | Ch. A RCF Mode Select (2 of 2)[3] | Ch. A RCF Mode Select (3 of 3)[2] |
| | |    00: FIR |    000: FIR |
| | |    01: FIR |    001: $\pi/4$–DQPSK |
| | |    10: QPSK |    010: GMSK |
| | |    11: MSK |    011: MSK |
| | | |    100: FIR, Compact Input Resolution |
| | | |    101: 8–PSK |
| | | |    110: $3\pi/8$–8PSK |
| | | |    111: QPSK |
| | 3–0 | Ch. A RCF (Taps per Phase[3]) –1 | No Change[2] |
| 0x10D | 7–6 | Ch. A RCF Coarse Scale (a) | No Change[2] |
| | |    00: 0 dB | |
| | |    01: –6 dB | |
| | |    10: –12 dB | |
| | |    11: –18 dB | |
| | 5 | Ch. A RCF Phase EQ Enable | No Change |
| | 4–0 | Ch. A Serial Clock Divisor (2, 4, …64) | Ch. A Serial Clock Divisor (1, 2,…32) |
| 0x10E | 15 | Ch. A Fine Scale Factor Enable | Ch. A Unsigned Scale Factor[2] |
| | | | This is extended to allow values in the range (0–2). |
| | 14–2 | Ch. A RCF Unsigned Scale Factor [2] | No Change[2] |
| | 1–0 | Reserved | Reserved |
| 0x10F | 17–16 | Unused | Ch. A Time Slot Sync Select |
| | | |    00: Sync0 (See 0x001 Time Slot) |
| | | |    01: Sync1 |
| | | |    10: Sync2 |
| | | |    11: Sync3 |
| | 15–0 | Ch. A RCF Scale Hold–Off Counter[1] | The counter is unchanged, but instead of just scale update, when the counter hits one, the following sequence is initiated: |
| | | |    1. Ramp Down (if Ramp is enabled) |

**Channel Function Registers (continued)**

| Internal Address | Bit | AD6622 Compatible Description | AD6623 Extensions Description |
|---|---|---|---|
| | | | 2. Update RCF Mode Select registers marked with "2". |
| | | | 3. Ramp Up (if Ramp is enabled) |
| 0x110 | 15–0 | Ch. A RCF Phase EQ Coef1 | No Change |
| 0x111 | 15–0 | Ch. A RCF Phase EQ Coef2 | No Change |
| 0x112 | 15–0 | Unused | Ch. A RCF FIR–PSK Magnitude 0 |
| 0x113 | 15–0 | Unused | Ch. A RCF FIR–PSK Magnitude 1 |
| 0x114 | 15–0 | Unused | Ch. A RCF FIR–PSK Magnitude 2 |
| 0x115 | 15–0 | Unused | Ch. A RCF FIR–PSK Magnitude 3 |
| 0x116 | 7–6 | Unused | Ch. A Serial Data Frame Input Select |
| | | | 0x: Internal Frame Request |
| | | | 10: External SDFI Pad |
| | | | 11: Previous Channel's Frame End |
| | 5 | Unused | Ch. A Serial Data Frame Output Select |
| | | | 0: Serial Data Frame Request |
| | | | 1: Serial Data Frame End |
| | 4 | Unused | Ch. A Serial Clock Slave (SCS) |
| | | | SCS = 0: Master Mode |
| | | | (SCLK is an output) |
| | | | SCS = 1: Slave Mode |
| | | | (SCLK is an input) |
| | 3 | Unused | Ch. A Serial Fine Scale Enable[1] |
| | 2 | Unused | Ch. A Serial Time Slot Sync Enable (ignored in FIR mode) |
| | 1 | Unused | Ch. A Ramp Interpolation Enable |
| | 0 | Unused | Ch. A Ramp Enable |
| 0x117 | 5–0 | Unused | Ch. A Mode 0 Ramp Length, R0–1 |
| 0x118 | 4–0 | Unused | Ch. A Mode 1 Ramp Length, R1–1 |
| 0x119 | 4–0 | Unused | Ch. A Ramp Rest Time, Q (No inputs requested during rest time.) |
| 0x11A–11F | | Unused | No Change |
| 0x120–13F | 15–0 | Ch. A Data RAM | No Change |
| 0x140–17F | 15–14 | Unused | No Change |
| | 13–0 | Unused | Ch. A Ramp RAM |
| 0x180–1FF | 15–0 | Ch. A Coefficient RAM | No Change |
| | | | This address is mirrored at 0x900–0x97F and contiguously extended at 0x980–0x9FF |

NOTES
[1]Clear on RESET.
[2]These bits update after a Start or a Beam Sync. See CR 0x10F.
[3]Allows dynamic updates.

**(0x000) Summation Mode Control**
Controls features in the summation block of the AD6623.
Bit 5–6:   Reserved.
Bit 4:   Low: Wideband Input Enabled.
High: Wideband Input Disabled.
Bit 3:   Low: Dual Output Disabled.
High: Dual Output Enabled.
Bit 2:   Reserved.
Bit 1:   Low: Output data will be in two's complement.
High: Output data will be in offset binary.
Bit 0:   Low: Over-range will wrap.
High: Over-range will clip to full scale.

**(0x001) Sync Mode Control**
Bit 7:   Ignores all but the first pin sync.
Bit 6:   Beam on pin Sync.
Bit 5:   Hop on pin Sync.

Bit 4:   High enables the count down of the Start Hold-Off Counter. The counter is clocked with the AD6623 CLK signal. When it reaches a count of one the Sleep bit of the appropriate channel(s) is set low to activate the channel(s).

Bit 3–0:   High enables synchronization of these channels. See the Synchronization section of the data sheet for detailed explanation.

**(0x002) BIST Counter**
Sets the length, in CLK cycles, of the built-in self test.

**(0x003) BIST Result**
A read-only register containing the result after a self test. Must be compared to a known good result for a given setup to determine pass/fail.

# AD6623

## Channel Function Registers

The following registers are channel-specific. '0xn' denotes that these values are represented as hexadecimal numbers; 'n' represents the specified channel. Valid channels are n = 1, 2, 3, and 4.

### (0xn00) Start Update Hold-Off Counter

See the Synchronization section for detailed explanation. If no synchronization is required, this register should be set to 0.

Bit 17–16: The Start Sync Select bits are used to set which sync pin will initiate a start sequence.

Bit 15–0: The Start Update Hold-Off Counter is used to synchronize start–up of AD6623 channels and can be used to synchronize multiple chips. The Start Update Hold-Off Counter is clocked by the AD6623 CLK (master clock).

### (0xn01) NCO Control

Bit 1:0 Set the NCO scaling per Table XXI.

**Table XXI. NCO Control (0xn01)**

| Bit 1 | Bit 0 | NCO Output Level |
|-------|-------|------------------|
| 0 | 0 | –6 dB (no attenuation) |
| 0 | 1 | –12 dB attenuation |
| 1 | 0 | –18 dB attenuation |
| 1 | 1 | –24 dB attenuation |

Bit 2: High clears the NCO phase accumulator to 0 on either a Soft Sync or Pin Sync (see Synchronization for details).

Bit 3: High enables NCO phase dither.

Bit 4: High enables NCO amplitude dither.

Bit 7–5: Reserved and should be written low.

### (0xn02) NCO Frequency

This register is a 32-bit unsigned integer that sets the NCO Frequency. The NCO Frequency contains a shadow register for synchronization purposes. The shadow can be read back directly, the NCO Frequency cannot.

$$NCO_{FREQUENCY} = 2^{32} \times \left( \frac{f_{CHANNEL}}{CLK} \right) \qquad (26)$$

### (0xn03) NCO Frequency Update Hold-Off Counter

See the Synchronization section for detailed explanation. If no synchronization is required, this register should be set to 0.

Bit 17–16: The Hop Sync Select bits are used to set which sync pin will initiate a hop sequence.

Bit 15–0: The Hold-Off Counter is used to synchronize the change of NCO frequencies.

### (0xn04) NCO Phase Offset

This register is a 16-bit unsigned integer that is added to the phase accumulator of the NCO. This allows phase synchronization of multiple channels of the AD6623(s). The NCO Phase Offset contains a shadow register for synchronization purposes. The shadow can be read back directly, the NCO Phase Offset cannot. See the Synchronization section for details.

### (0xn05) NCO Phase Offset Update Hold-Off Counter

See the Synchronization section for a detailed explanation. If no synchronization is required, this register should be set to 0.

Bit 17–16: The Phase Sync Select bits are used to set which sync pin will initiate a phase sync sequence.

Bit 15–0: The Hold-Off Counter is used to synchronize the change of NCO phases.

### (0xn06) CIC Scale

Bits 4–0: Sets the CIC scaling per the equation below.

$$CIC\_Scale = ceil \times \left( log_2 \left( L_{CIC5}^4 \times L_{CIC2} \right) \right) \qquad (27)$$

See the CIC section for details.

### (0xn07) CIC2 Decimation – 1 ($M_{CIC2} - 1$)

This register is used to set the decimation in the CIC2 filter. The value written to this register is the decimation minus one. The CIC2 decimation can range from 1 to 512 depending upon the interpolation of the CIC2. There is no timing error associated with this decimation. See the CIC2 section for further details.

### (0xn08) CIC2 Interpolation – 1 ($L_{CIC2} - 1$)

This register is used to set the interpolation in the CIC2 filter. The value written to this register is the interpolation minus one. The CIC2 interpolation can range from 1 to 4096. $L_{rCIC2}$ must be chosen equal to or larger than $M_{rCIC2}$ and both must be chosen such that a suitable CIC2 Scalar can be chosen. For more details the CIC2 section should be consulted.

### (0xn09) CIC5 Interpolation – 1

This register sets the interpolation rate for the CIC5 filter stage (unsigned integer). The programmed value is the CIC5 Interpolation – 1. Maximum interpolation is limited by the CIC scaling available (See the CIC section).

### (0xn0A) Number of RCF Coefficients – 1

This register sets the number of RCF Coefficients and is limited to a maximum of 256. The programmed value is the number of RCF Coefficients – 1. There is an A register and a B register at this memory location. Value A is used when the RCF is operating in mode 0 and value B is used when in mode 1. The RCF mode bit of interest here is bit 6 of address 0xn0C.

### (0xn0B) RCF Coefficient Offset

This register sets the offset for RCF Coefficients and is normally set to 0. It can be viewed as a pointer which selects the portion of the CMEM used when computing the RCF filter. This allows multiple filters to be stored in the Coefficient memory space, selecting the appropriate filter by setting the offset.

### (0xn0C) Channel Mode Control 1

Bit 9: High, selecting compact FIR mode results in 24-bit serial word length (12 I followed by 12 Q). When low, selecting compact FIR mode results in 16-bit serial word length (8 I followed by 8 Q).

Bit 8: High enables RCF Pseudo-Random Input Select.

Bit 7: High selects a Pseudo-Random sequence length of 8,388,607. Low selects a Pseudo-Random Sequence length of 15.

Bits 6–4: Sets the channel input format as shown in Table XXII.

**Table XXII. Channel Inputs**

| Bit 6 | Bit 5 | Bit 4 | Input Mode |
|-------|-------|-------|------------|
| 0 | 0 | 0 | FIR |
| 0 | 0 | 1 | π/4-DQPSK |
| 0 | 1 | 0 | GSM |
| 0 | 1 | 1 | MSK |
| 1 | 0 | 0 | Compact FIR |
| 1 | 0 | 1 | 8PSK |
| 1 | 1 | 0 | 3π/8-8PSK |
| 1 | 1 | 1 | QPSK |

Bit 6        Can be set through the serial port (see section on serial word formats).

Bits 3–0:    Sets $(N_{RCF}/L_{RCF}) -1$

**(0xn0D) Channel Mode Control 2**
Bits 7–6:    Sets the RCF Coarse Scale as shown in Table XXIII.

**Table XXIII.  RCF Coarse Scale**

| Bit 7 | Bit 6 | RCF Coarse Scale (dB) |
|-------|-------|------------------------|
| 0 | 0 | 0 |
| 0 | 1 | –6 |
| 1 | 0 | –12 |
| 1 | 1 | –18 |

Bit 5:       High enables the RCF phase equalizer.

Bits 4–0:    Sets the serial clock divider (SDIV) that determines the serial clock frequency based on the following equation.

$$f_{SCLK} = \frac{CLK}{SDIV + 1} \qquad (28)$$

**(0xn0E) Fine Scale Factor**
Bits 15–2:   Sets the RCF Fine Scale Factor as an unsigned number representing the values (0,2). This register is shadowed for synchronization purposes. The shadow can be read back directly, the Fine Scale Factor can not.
Bits 1–0:    Reserved.

**(0xn0F) RCF Time Slot Hold-Off Counter**
Bits 17–16: The Time Slot Sync Select bits are used to set which sync pin will initiate a time slot sync sequence.
Bits 15–0:   The Hold-Off Counter is used to synchronize the change of RCF Fine Scale. See the Synchronization section for a detailed explanation.  If no synchronization is required, this register should be set to 0.

**(0xn10–0xn11) RCF Phase Equalizer Coefficients**
See the RCF section for details.

**(0xn12–0xn15)  FIR-PSK Magnitudes**
See the RCF section for details.

**(0xn16) Serial Port Setup**
Bits 7–6:    Serial Data Frame Start Select

**Title XXIV.  Serial Port Setup**

| Bit 7 | Bit 6 | Serial Data Frame Start |
|-------|-------|--------------------------|
| 0 | X | Internal Frame Request |
| 1 | 0 | External SDFI Pad |
| 1 | 1 | Previous Channel's Frame End |

Bit 5:       High means SDFO is a frame end, low means SDFO is a frame request.

Bit 4:       High selects serial slave mode. SCLK is an input in serial slave mode.

Bit 3:       High enables Fine Scaling through the Serial Port (not available in FIR Mode).

Bit 2:       High enables Serial Time Slot Syncs (not available in FIR Mode).

Bit 1:       High enables Power Ramp coefficient interpolation.

Bit 0:       High enables the Power Ramp.

**(0xn17) Power Ramp Length 0**
This is the length of the ramp for mode 0, minus one.

**(0xn18) Power Ramp Length 1**
This is the length of the ramp for mode 1, minus one. Setting this to zero disables dual ramps.

**(0xn19) Power Ramp Rest Time**
This is the number of RCF output samples to rest for between a ramp down and a ramp up.

**(0xn1A–0xn1F) Unused**

**(0xn20–0xn3F) Data Memory**
This group of registers contain the RCF Filter Data.  See the RCF section for additional details.

**(0xn40–0xn17F) Power Ramp Coefficient Memory**
This group of registers contain the Power Ramp Coefficients. See the Power Ramp section for additional details.

**(0xn80–0xn1FF) Coefficient Memory**
This group of registers contain the RCF Filter Coefficients. See the RCF section for additional details.

**PSEUDOCODE**
**Write Pseudocode**
```
Void Write_Micro(ext_address, int data);

Main()
{

/* This code shows the programming of the
NCO frequency register using the Write_Micro
function defined above. The variable
address is the External Address A[2:0] and
data is the value to be placed in the
external interface register.

Internal Address = 0x102, channel 1
*/

/*Holding registers for NCO byte wide
access data*/
int d3, d2, d1, d0;
/*NCO frequency word (32 bits wide)*/
NCO_FREQ=0x1BEFEFFF;
/*write Chan */
Write_Micro(7, 0x01);
/*write Addr */
Write_Micro(6,0x02);
/*write Byte 3*/
d3=(NCO_FREQ & 0xFF02Y∞00)>>24;
Write_Micro(3,d3);
/*write Byte 2*/
d2=(NCO_FREQ & 0xFF0000)>>16;
Write_Micro(2,d2);
/*write Byte 1*/
d1=(NCO_FREQ & 0xFF00)>>8;
Write_Micro(1,d1);
/*write Byte 0, Byte 0 is written last and
causes an internal write to occur*/
d0=NCO_FREQ & 0xFF;
Write_Micro(0,d0);
}
```

# AD6623

**Read Pseudocode**

```
Void Read_Micro(ext_address);
Main()
{
/* This code shows the reading of the NCO
frequency register using the Read_Micro
function defined above.  The variable
address is the External Address A[2:0]

Internal Address = 0x102, channel 1
*/

/*Holding registers for NCO byte wide
access data*/
int d3, d2, d1, d0;
/*NCO frequency word (32 bits wide)*/
/*write Chan */
Write_Micro(7, 0x01);
/*write Addr*/
Write_Micro(6,0x02);
/*read Byte 0,  all data is moved from the
Internal Registers to the interface
registers on this access, thus Byte 0 must
be accessed first for the other Bytes to be
valid*/
d0=Read_Micro(0) & 0xFF;
/*read Byte 1*/
d1=Read_Micro(1) & 0xFF;
/*read Byte 2*/
d2=Read_Micro(2) & 0xFF;
/*read Byte 0 */
d3=Read_Micro(3) & 0xFF;
}
```

## APPLICATIONS

The AD6623 provides considerable flexibility for the control of the synchronization, relative phasing, and scaling of the individual channel inputs. Implementation of a multichannel transmitter invariably begins with an analysis of the output spectrum that must be generated.

## USING THE AD6623 TO PROCESS UMTS CARRIERS

The AD6623 may be used to process two UMTS carriers, each with an output oversampling rate of 24× (i.e., 92.16 MSPS). The AD6623 configuration used to accomplish this consists of using two processing channels in parallel to process each UMTS carrier. Please refer to the Technical Note: Processing Two UMTS Carriers with 24× Oversampling Using the AD6623.

## DIGITAL TO ANALOG CONVERTER (DAC) SELECTION

The selection of a high-performance DAC depends on a number of factors. The dynamic range of the DAC must be considered from a noise and spectral purity perspective. The 14-bit AD9772A is the best choice for overall bandwidth, noise, and spectral purity.

In order to minimize the complexity of the analog interpolation filter which must follow the DAC, the sample rate of the master clock is generally set to at least three times the maximum analog frequency of interest.

In the case where a 15 MHz band of interest is to be up-converted to RF, the lowest frequency might be 5 MHz and the upper band edge at 20 MHz (offset from dc to afford the best image reject filter after the first digital IF). The minimum sample rate would be set to 65 MSPS.

Consideration must also be given to data rate of the incoming data stream, interpolation factors, and the clock rate of the DSP.

## MULTIPLE TSP OPERATION

Each of the four Transmit Signal Processors (TSPs) of the AD6623 can adequately reject the interpolation images of narrow bandwidth carriers such as AMPS, IS-136, GSM, EDGE, and PHS. Wider bandwidth carriers such as IS-95 and IMT2000 require a coordinated effort of multiple processing channels.

This section demonstrates how to coordinate multiple TSPs to create wider bandwidth channels without sacrificing image rejection. As an example, a UMTS carrier is modulated using four TSP channels (an entire AD6623). The same principles can be applied to different designs using more or fewer TSPs. This section does not explore techniques for using multiple TSPs to solve problems other than Serial Port or RCF throughput.

Designing filter coefficients and control settings for de-interleaved TSPs is no harder than designing a filter for a single TSP. For example, if four TSPs are to be used, simply divide the input data rate by four and generate the filter as normal. For any design, a better filter can always be realized by incrementing the number of TSPs to be used. When it is time to program the TSPs, only two small differences must be programmed. First, each channel is configured with exactly the same filter, scalers, modes and NCO frequency. Since each channel receives data at one-quarter the data rate and in a staggered fashion, the Start Hold-Off Counters must also be staggered (see "Programming Multiple TSPs" section). Second, the phase offset of each NCO must be set to match the demultiplexed ratio (in this example). Thus the phase offset should be set to 90 degrees (16384 which is one-quarter of a 16-bit register).

**Determining the Number of TSPs to Use**

There are three limitations of a single TSP that can be overcome by deinterleaving an input stream into multiple TSPs: Serial Port bandwidth, the time restriction to the RCF impulse response length (NRCF), and the DMEM restriction to NRCF.

If the input sample rate is faster than the Serial Port can accept data, the data can be de-interleaved into multiple Serial Ports. Recalling from the Serial Port description, the SCLK frequency ($f_{SCLK}$) is determined by the equation below. To minimize the number of processing channels, SCLKdivider should be set as low as possible to get the highest $f_{SCLK}$ that the serial data source can accept.

$$f_{SCLK} = \frac{f_{CLK}}{SCLKdivider + 1} \qquad (29)$$

A minimum of 32 SCLK cycles are required to accept an input sample, so the minimum number of TSPs ($N_{TSP}$) due to limited Serial Port bandwidth is a function of the input sample rate ($f_{IN}$), as shown in the equation below.

$$N_{TSP} \geq ceil\left(\frac{32 \times f_{IN}}{f_{SCLK}}\right) \qquad (30)$$

For example for a UMTS system, we will assume $f_{CLK}$ = 76.8 MHz, and the serial data source can drive data at 38.4 Mbps (SCLKdivider = 0). To achieve $f_{IN}$ = 3.84 MHz, the minimum $N_{TSP}$ is 3 with a Serial Clock $f_{SCLK}$ = 52 MHz which is a limitation of the Serial Port (This is TSP channels, not TSP ICs).

Multiple TSPs are also required if the RCF does not have enough time or DMEM space to calculate the required RCF filter. Recalling the maximum $N_{TAPS}$ equation from the RCF description, are three restrictions to the RCF impulse response length, $N_{RCF}$.
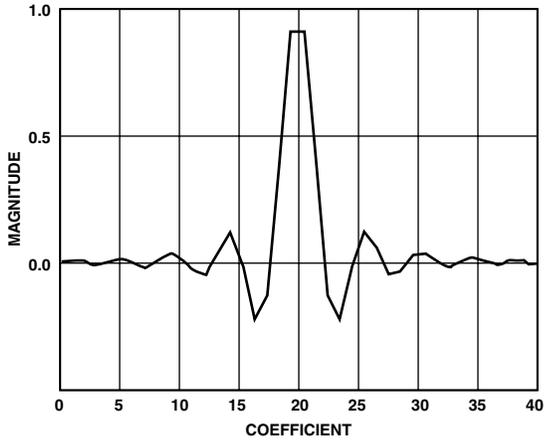
Time Restriction   CMEM Restriction

$$N_{RCF} \leq min\left(\frac{1}{2}, 16 \times L_{RCF}\ 256\right) \qquad (31)$$

DMEM Restriction

where:

$$L = L_{RCF} \times L_{CIC5} \times \frac{L_{CIC2}}{M_{CIC2}} = \frac{N_{TSP} \times f_{CLK}}{f_{IN}} \qquad (32)$$

De-interleaving the input data into multiple TSPs extends the time restriction and may possibly extend the DMEM restriction, but will not extend the CMEM restriction. Deinterleaving the input stream to multiple TSPs divides the input sample rate to each TSP by the number of TSPs used ($N_{TSP}$). To keep the output rate fixed, L must be increased by a factor of $N_{CH}$, which extends the time restriction. This increase in L may be achieved by increasing any one or more of $L_{RCF}$, $L_{CIC5}$, or $L_{CIC2}$ within their normal limits. Achieving a larger L by increasing $L_{RCF}$ instead of $L_{CIC5}$ or $L_{CIC2}$ will relieves the DMEM restriction as well.

In a UMTS example, $N_{TSP} = 4$, $f_{CLK} = 76.8$ MHz, and $f_{IN} = 3.84$ MHz, resulting in L = 80. Factoring L into $L_{RCF} = 10$, $L_{CIC} = 8$, and $L_{CIC2} = 1$ results in a maximum $N_{RCF} = 40$ due to the time restriction. Figure 37 shows an example RCF impulse response which has a frequency response as shown in Figure 38 from 0 Hz to 7.68 MHz ($f_{IN} \times L_{RCF}/N_{TSP}$). The composite RCF and CIC frequency response is shown in Figure 38, on the same frequency scale. This figure demonstrates a good approximation to a root-raised-cosine with a roll-off factor of 0.22, a passband ripple of 0.1 dB, and a stopband ripple better than –70 dB until the lobe of the first image which peaks at –60 dB about 7.68 MHz from the carrier center. This lobe could be reduced by shifting more of the interpolation towards the RCF, but that would sacrifice near in performance. As shown, the first image can be easily rejected by an analog filter further up the signal path.

Scaling must be considered as normal with an interpolation factor of L, to guarantee no overflow in the RCF, CIC, or NCOs. The output level at the summation port should be calculated using an interpolation factor of L/$N_{TSP}$.

**Programming Multiple TSPs**
Configuring the TSPs for de-interleaved operation is straight forward. All the Channel Registers and the CMEM of each TSP are programmed identically, except the Start Hold-Off Counters and NCO Phase Offset.

In order to separate the input timing to each TSP, the Hold-Off Counters must be used to start each TSP successively in response to a common Start SYNC. The Start SYNC may originate from the SYNC pin or the MicroPort. Each subsequent TSP must have a Hold-Off Counter value L/$N_{TSP}$ larger than its predecessor's. If the TSPs are located on cascaded AD6623s, the Hold-Off Counters of the upstream device should be incremented by an additional one.

In the UMTS example, L = 80 and $N_{TSP} = 4$, so to respond as quickly as possible to a Start SYNC, the Hold-Off Counter values should be 1, 21, 41, and 61.

**Driving Multiple TSP Serial Ports**
When configured properly, the AD6623 will drive each SDFO out of phase. Each new piece of data should be driven only into the TSP that pulses its SDFO pin at that time.

In the UMTS example in Figure 35, L = 80 and $N_{TSP} = 4$, so each serial port need only accept every fourth input sample. Each serial port is shifting at peak capacity, so sample 1, 2, and 3 begin shifting into Serial Ports B, C, and D before sample 0 is completed into Serial Port A.



Figure 35.  UMTS Example



Figure 36.  Summation Block

# AD6623



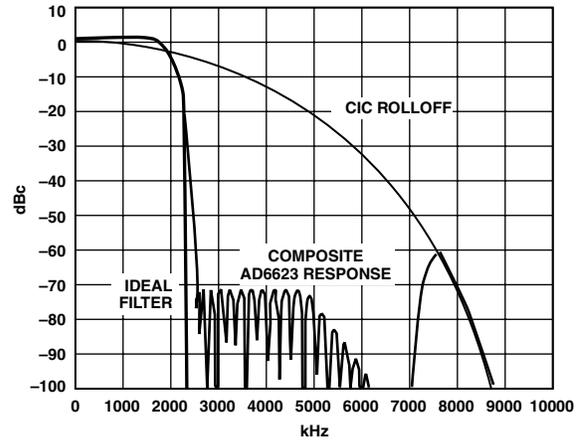Figure 37. Typical Impulse Response for WBCDMA



Figure 38. RAM Coefficient Filter, Frequency Response for WBCDMA
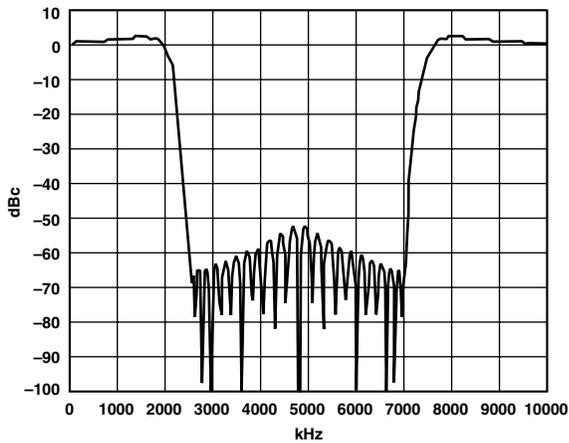


Figure 39. RCF and CIL, Frequency Response for WBCDMA

## THERMAL MANAGEMENT

The power dissipation of the AD6623 is primarily determined by three factors: the clock rate, the number of channels active, and the distribution of interpolation rates. The faster the clock rate the more power dissipated by the CMOS structures of the AD6623 and the more channels active the higher the overall power of the chip. Low interpolation rates in the CIC stages (CIC5, CIC2) results in higher power dissipation. All these factors should be analyzed as each application has different thermal requirements.

The AD6623 128-Lead MQFP is specially designed to provide excellent thermal performance. To achieve the best performance the power and ground leads should be connected directly to planes on the PC board. This provides the best thermal transfer from the AD6623 to the PC board.
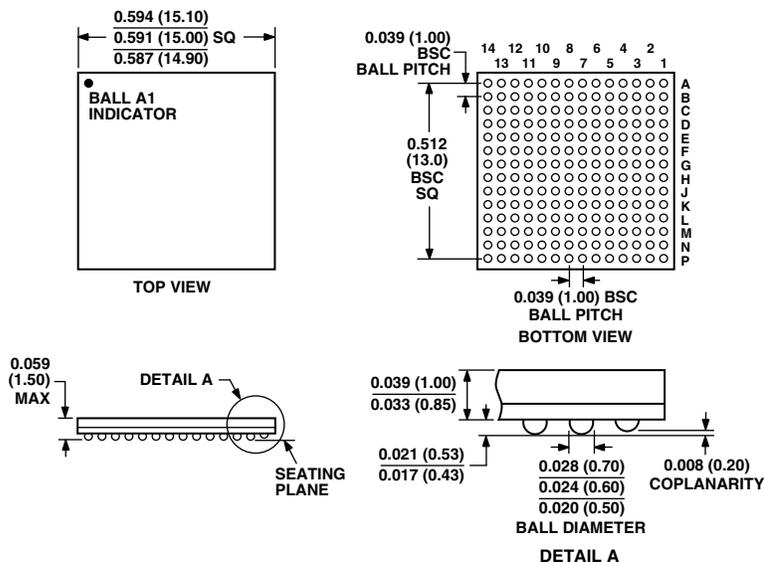
## OUTLINE DIMENSIONS

Dimensions shown in inches and (mm).

### 128-Lead Terminal Metric Quad Flatpack (MQFP)
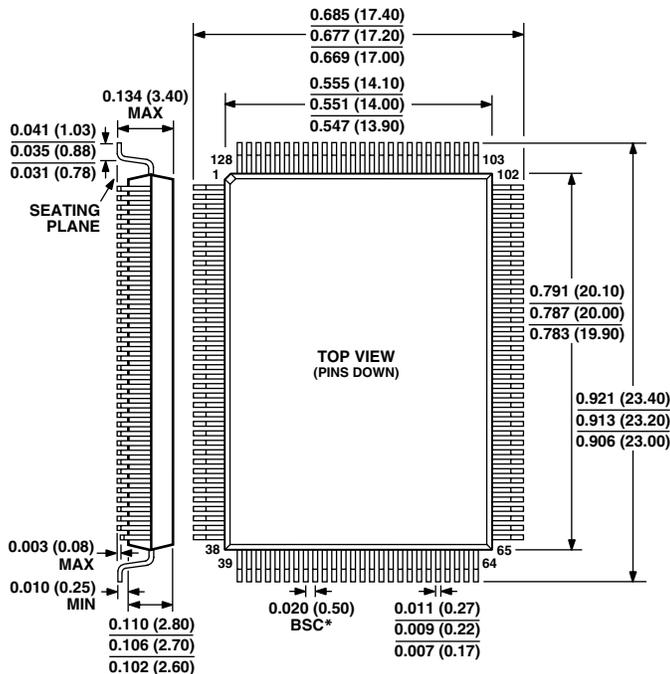


*THE ACTUAL POSITION OF EACH LEAD IS WITHIN 0.00315 (0.08) FROM ITS IDEAL
POSITION WHEN MEASURED IN THE LATERAL DIRECTION. CENTER FIGURES ARE
TYPICAL UNLESS OTHERWISE NOTED

### 196-Lead Ball Grid Array (BGA)



NOTES
1. CONTROLLING DIMENSIONS ARE IN MILLIMETERS
2. THE ACTUAL POSITION OF THE BALL GRID IS WITHIN
   0.008 (0.20) OF ITS IDEAL POSITION RELATIVE TO THE
   PACKAGE EDGES
3. THE ACTUAL POSITION OF EACH BALL IS WITHIN
   0.004 (0.10) OF ITS IDEAL POSITION TO THE BALL GRID
4. CENTER FIGURES ARE NOMINAL DIMENSIONS