

### FEATURES

Low power, low IF transceiver

Frequency bands:

433 MHz to 464 MHz

862 MHz to 928 MHz

Data rates supported:

0.3 kbps to 200 kbps, FSK

0.3 kbps to 64 kbps, ASK

2.3 V to 3.6 V power supply

Programmable output power:

-16 dBm to +13 dBm in 0.3 dBm steps

Receiver sensitivity:

-117.5 dBm at 1 kbps, FSK

-110.5 dBm at 9.6 kbps, FSK

-106.5 dBm at 9.6 kbps, ASK

Low power consumption:

19 mA in receive mode

22 mA in transmit mode (10 dBm output)

On-chip VCO and fractional-N PLL

On-chip 7-bit ADC and temperature sensor

±1 ppm RF output frequency accuracy possible from  
low cost 100 ppm crystal

Digital RSSI

Leakage current <1 µA in power-down mode

48-lead ultrasmall MLF package (chip scale)

### APPLICATIONS

Low cost wireless data transfer

Remote control/security systems

Wireless metering

Keyless entry

Home automation

Process and building control

Wireless voice

### FUNCTIONAL BLOCK DIAGRAM

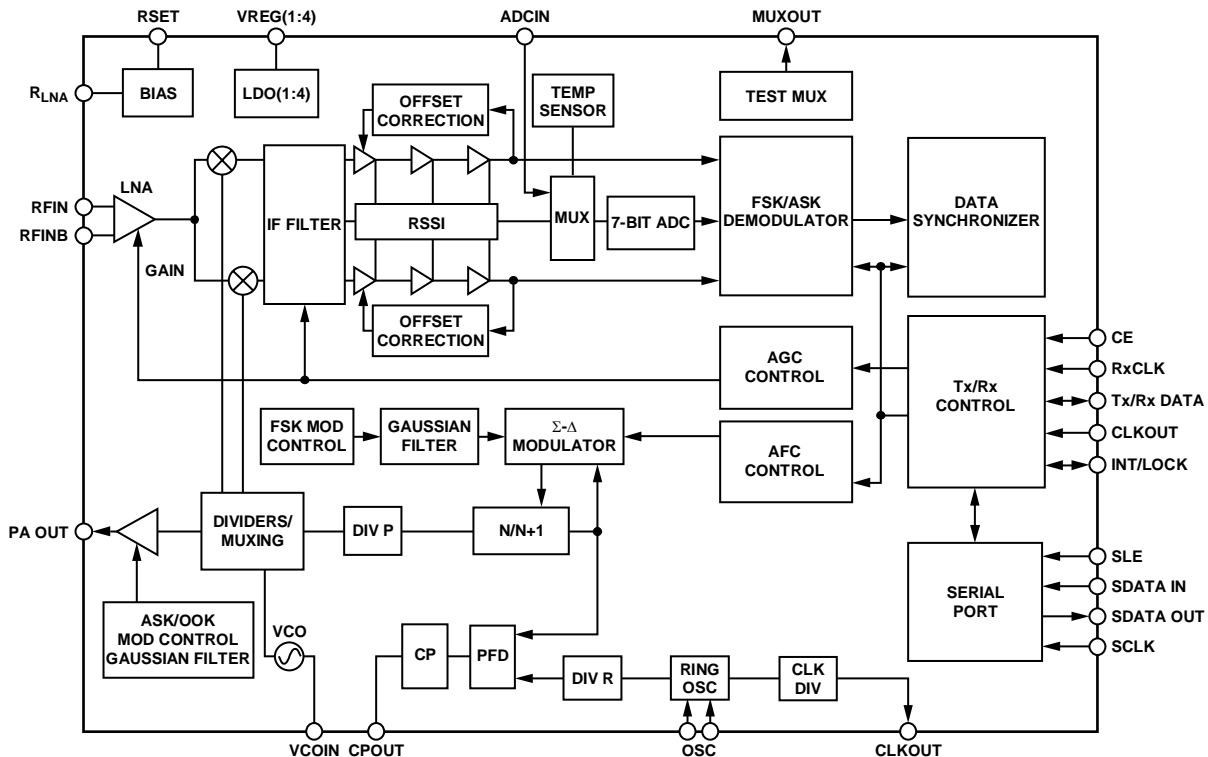


Figure 1.

01975-PG-001

### Rev. PrH

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## REVISION HISTORY

Revision PrH: Preliminary Version

## GENERAL DESCRIPTION

The ADF7020 is a low power, highly integrated FSK/GFSK/ASK/OOK/GASK transceiver designed for operation in the license-free ISM bands at 433 MHz, 868 MHz, and 915 MHz. It is suitable for circuit applications that meet either the European ETSI-300-220 or the North American FCC (Part 15) regulatory standards. A complete transceiver can be built using a small number of external discrete components, making the ADF7020 very suitable for price-sensitive and area-sensitive applications.

The transmit section contains a VCO and low noise fractional-N PLL with output resolution of <1 ppm. The VCO operates at twice the fundamental frequency to reduce spurious emissions and frequency pulling problems.

The transmitter output power is programmable in 0.3 dB steps from -16 dBm to +13 dBm. The transceiver RF frequency, channel spacing, and modulation are programmable using a simple 3-wire interface. The device operates with a power supply range of 2.3 V to 3.6 V and can be powered down when not in use.

A low IF architecture is used in the receiver (200 kHz), minimizing power consumption and the external component count and avoiding interference problems at low frequencies. The ADF7020 supports a wide variety of programmable features including Rx linearity, sensitivity, and IF bandwidth, allowing the user to trade off receiver sensitivity and selectivity against current consumption, depending on the application. The receiver also features a patent-pending automatic frequency control (AFC) loop, allowing the PLL to track out the frequency error in the incoming signal.

An on-chip ADC provides readback of an integrated temperature sensor, an external analog input, the battery voltage, or the RSSI signal, which provides savings on an ADC in some applications. The temperature sensor is accurate to  $\pm 5^{\circ}\text{C}$  over the full operating temperature range of  $-40^{\circ}\text{C}$  to  $+85^{\circ}\text{C}$ .

## SPECIFICATIONS

$V_{DD} = 2.3 \text{ V}$  to  $3.6 \text{ V}$ ,  $GND = 0 \text{ V}$ ,  $T_A = T_{MIN}$  to  $T_{MAX}$ , unless otherwise noted.

Typical specifications are at  $V_{DD} = 3 \text{ V}$ ,  $T_A = 25^\circ\text{C}$ .

All measurements are performed using the test circuit in Figure TBD using PN9 data sequence, unless otherwise noted.

**Table 1.**

Parameter	Min	Typ	Max	Unit	Test Conditions
<b>RF CHARACTERISTICS</b>					
Frequency Ranges	862		928	MHz	
Frequency Ranges (Divide-by-2 Mode)	433		464	MHz	
Phase Frequency Detector Frequency	RF/256		20	MHz	
<b>TRANSMISSION PARAMETERS</b>					
Data Rate FSK/GFSK	0.3		200	kbps	
Data Rate OOK/ASK	0.3		64 <sup>1</sup>	kbps	
Frequency Shift Keying					
GFSK/FSK Frequency Deviation <sup>2,3</sup>	1		110	kHz	PFD = 3.625 MHz
	4.88		620	kHz	PFD = 20 MHz
Deviation Frequency Resolution	100			Hz	PFD = 3.625 MHz
Gaussian Filter BT		0.5			
Adjacent Channel Power, GFSK		TBD		dBc	Channel spacing = 25 kHz, measured in adjacent channel $\pm 8.5$ kHz from center; DR = 4.8 kbps, $F_{DEV} = 2.4$ kHz, FRF = 868 MHz
		-50		dBc	868.95 MHz $\pm 250$ kHz, DR = 38.4 kbps, $F_{DEV} = 19.2$ kHz
Amplitude Shift Keying					
ASK Modulation Depth			30	dB	
OOK -PA Off Feedthrough		-50		dBm	
Transmit Power <sup>4</sup>	-20		+13	dBm	FRF = 915 MHz, $V_{DD} = 3.0 \text{ V}$ , $T_A = 25^\circ\text{C}$
			TBD	dBm	FRF = 868 MHz, $V_{DD} = 3.0 \text{ V}$ , $T_A = 25^\circ\text{C}$
			TBD	dBm	FRF = 433 MHz, $V_{DD} = 3.0 \text{ V}$ , $T_A = 25^\circ\text{C}$
Transmit Power Variation					
Highest Power Setting		TBD		dBm	FRF = 915 MHz, $V_{DD} = 3.6 \text{ V}$
		13		dBm	FRF = 915 MHz, $V_{DD} = 3.0 \text{ V}$
		TBD		dBm	FRF = 915 MHz, $V_{DD} = 2.3 \text{ V}$
Transmit Power Flatness		TBD		dB	From 902 MHz to 928 MHz
Programmable Step Size					
-20 dBm to +13 dBm		0.3125		dB	
Spurious Emissions during PLL Settling			-57	dBm	Mute PA until lock enabled (R2_DB5 = 1)
Integer Boundary		-55		dBc	50 kHz loop BW
Reference		-65		dBc	
Harmonics					
Second Harmonic		-27	-18	dBc	
Third Harmonic		-21	-18	dBc	
All Other Harmonics			-35	dBc	
VCO Frequency Pulling, OOK mod		TBD		kHz rms	DR = 9.6 kbps
Optimum PA Load Impedance <sup>5</sup>		TBD		$\Omega$	FRF = 915 MHz
		TBD		$\Omega$	FRF = 868 MHz
		TBD		$\Omega$	FRF = 433 MHz
<b>RECEIVER PARAMETERS</b>					
FSK Input Sensitivity <sup>6</sup>					At BER = 1E - 3, FRF = 915 MHz
High Sensitivity Mode		-117.5		dBm	DR = 1 kbps, $F_{DEV} = 5$ kHz
Low Current Mode		-TBD		dBm	DR = 1 kbps, $F_{DEV} = 5$ kHz

See notes at end of table.

Parameter	Min	Typ	Max	Unit	Test Conditions
High Sensitivity Mode		-110.5		dBm	DR = 9.6 kbps, FDEV = 10 kHz
Low Current Mode		-104		dBm	DR = 9.6 kbps, FDEV = 10 kHz
High Sensitivity Mode		-99		dBm	DR = 200 kbps, FDEV = 50 kHz
Low Current Mode		-TBD		dBm	DR = 200 kbps, FDEV = 50 kHz
OOK Input Sensitivity					At BER = 1E - 3, FRF = 915 MHz
High Sensitivity Mode		-TBD		dBm	DR = 1 kbps
Low Current Mode		-TBD		dBm	DR = 1 kbps
High Sensitivity Mode		-106.5		dBm	DR = 9.6 kbps
Low Current Mode		-TBD		dBm	DR = 9.6 kbps
LNA and Mixer					
Input IP3 <sup>7</sup>					
Enhanced Linearity Mode		6.8		dBm	Pin = -20 dBm, 2 CW interferers
Low Current Mode		-3.2		dBm	FRF = 915 MHz, f1 = FRF + 3 MHz
High Sensitivity Mode		-35		dBm	F2 = FRF + 6 MHz, maximum gain
Rx Spurious Emissions <sup>8</sup>			-57	dBm	<1 GHz at antenna input
			-47	dBm	>1 GHz at antenna input
AFC					
Pull-In Range		±50		kHz	IF_BW = 200 kHz
Response time			TBD	Bits	
Accuracy		1	TBD	kHz	
Channel Filtering					
Adjacent Channel Rejection (Offset = ±1 × IF Filter BW Setting)		27		dB	IF filter BW settings = 100 kHz, 150 kHz, 200 kHz
Second Adjacent Channel Rejection (Offset = ±2 × IF Filter BW Setting)		50		dB	Desired signal 3 dB above the input sensitivity level, CW interferer power level increased until BER = 10 <sup>-3</sup> , image channel excluded
Third Adjacent Channel Rejection (Offset = ±3 × IF Filter BW Setting)		TBD		dB	
Image Channel Rejection (Image Channel = FRF - 400 kHz)		30		dB	Uncalibrated
		TBD		dB	Calibrated <sup>9</sup>
Co-channel Rejection		-3		dB	
Wide-Band Interference Rejection		TBD		dB	Swept from 100 MHz to 2 GHz, measured as channel rejection
Saturation (Maximum Input Level)		12		dBm	FSK mode, BER = 10 <sup>-3</sup>
Input Impedance		TBD		Ω	FRF = 915 MHz, RFIN, RFIN to GND
		TBD		Ω	FRF = 868 MHz
		TBD		Ω	FRF = 433 MHz
RSSI					
Range at Input		-100 to -36		dBm	
Linearity		±3		dB	
Absolute Accuracy		TBD		dB	
Response Time		350		μs	Maximum input step change, AGC included, RSSI ready for readback
PHASE LOCKED LOOP					
VCO Gain		65		MHz/V	902 MHz to 928 MHz band, VCO adjust = 0
		130		MHz/V	860MHz to 870 MHz band, VCO Adjust = 0
		TBD		MHz/V	At 433MHz, VCO Adjust = 0
Phase Noise (In-Band)		-92		dBc/Hz	PA = 0 dBm, V <sub>DD</sub> = 3.0 V, PFD = 10 MHz, FRF = 915 MHz, VCO BIAS = 4
Phase Noise (Out-of-Band)		-110		dBc/Hz	At 1 MHz offset
Residual FM		TBD		Hz	From 300 Hz to 5 kHz

See notes at end of table.

Parameter	Min	Typ	Max	Unit	Test Conditions
PLL Settling Time		40		μs	Measured for a 10 MHz frequency step to within 5 ppm accuracy, PFD = 20 MHz, LBW = TBD
REFERENCE INPUT					
Crystal Reference	3.625		24	MHz	See the Reference Input Section
External Oscillator	3.625		24	MHz	
Load capacitance		TBD		pF	
Input Level				CMOS levels	
TIMING INFORMATION					
Chip Enabled to Regulator Ready			TBD	μs	C <sub>REG</sub> = 100 nF With 19.2 MHz XTAL Time to synchronized data, includes AGC settling
Crystal Oscillator Startup time		1		ms	
Tx to Rx Turnaround Time		350 μs + (5 × T <sub>BIT</sub> )			
LOGIC INPUTS					
V <sub>INH</sub> , Input High Voltage	0.7 × V <sub>DD</sub>			V	
V <sub>INL</sub> , Input Low Voltage			0.2 × V <sub>DD</sub>	V	
I <sub>INH</sub> /I <sub>INL</sub> , Input Current			±1	μA	
C <sub>IN</sub> , Input Capacitance			10	pF	
Control Clock Input			50	MHz	
LOGIC OUTPUTS					
V <sub>OH</sub> , Output High Voltage	DV <sub>DD</sub> – 0.4			V	I <sub>OH</sub> = 500 μA
V <sub>OL</sub> , Output Low Voltage			0.4	V	I <sub>OL</sub> = 500 μA
CLK <sub>OUT</sub> Rise/Fall			5	ns	
CLK <sub>OUT</sub> Load			10	pF	
TEMPERATURE RANGE—T <sub>A</sub>	–40		+85	°C	
POWER SUPPLIES					
Voltage Supply					FRF = 915 MHz, V <sub>DD</sub> = 3.0 V, PA is matched in to 50 Ω  VCO_BIAS_SETTING = 3
AV <sub>DD</sub>	2.3		3.6	V	
DV <sub>DD</sub>	AV <sub>DD</sub>		AV <sub>DD</sub>		
Transmit Current Consumption					
–20 dBm		TBD		mA	
–10 dBm		12		mA	
0 dBm		15		mA	
10 dBm		22		mA	
Receive Current Consumption					
Low Current Mode		19	TBD	mA	
High Sensitivity Mode		21	TBD	mA	
Power-Down Mode					
Low Power Sleep Mode		0.1	1	μA	

<sup>1</sup> Higher data rates are achievable depending on local regulations.

<sup>2</sup> For definition of frequency deviation, see the Register 2—Transmit Modulation Register (FSK Mode) section.

<sup>3</sup> For definition of GFSK frequency deviation, see the Register 2—Transmit Modulation Register (GFSK/GOOK Mode) section.

<sup>4</sup> Measured as maximum unmodulated power. Output power varies with both supply and temperature.

<sup>5</sup> For matching details, see the LNA/PA Matching section.

<sup>6</sup> See Table 5 for description of different receiver modes.

<sup>7</sup> See Table 5 for description of different receiver modes.

<sup>8</sup> Follow the matching and layout guidelines to achieve the relevant FCC/ETSI specifications.

<sup>9</sup> See the Image Rejection Calibration section.

# TIMING CHARACTERISTICS

$V_{DD} = 3V \pm 10\%$ ;  $V_{GND} = 0V$ ,  $T_A = 25^\circ C$ , unless otherwise noted.  
 Guaranteed by design, but not production tested.

Table 2.

Parameter	Limit at $T_{MIN}$ to $T_{MAX}$	Unit	Test Conditions/Comments
$t_1$	<10	ns	SDATA to SCLK Setup Time
$t_2$	<10	ns	SDATA to SCLK Hold Time
$t_3$	<25	ns	SCLK High Duration
$t_4$	<25	ns	SCLK Low Duration
$t_5$	<10	ns	SCLK to SLE Setup Time
$t_6$	<20	ns	SLE Pulse Width
$t_7$	<TBD	ns	SLE to SCLK Setup Time, Readback
$t_8$	<TBD	ns	SCLK to SREAD Data Valid, Readback
$t_9$	<TBD	ns	SREAD Hold Time after SCLK, Readback
$t_{10}$	<TBD	ns	SCLK to SLE Disable Time, Readback

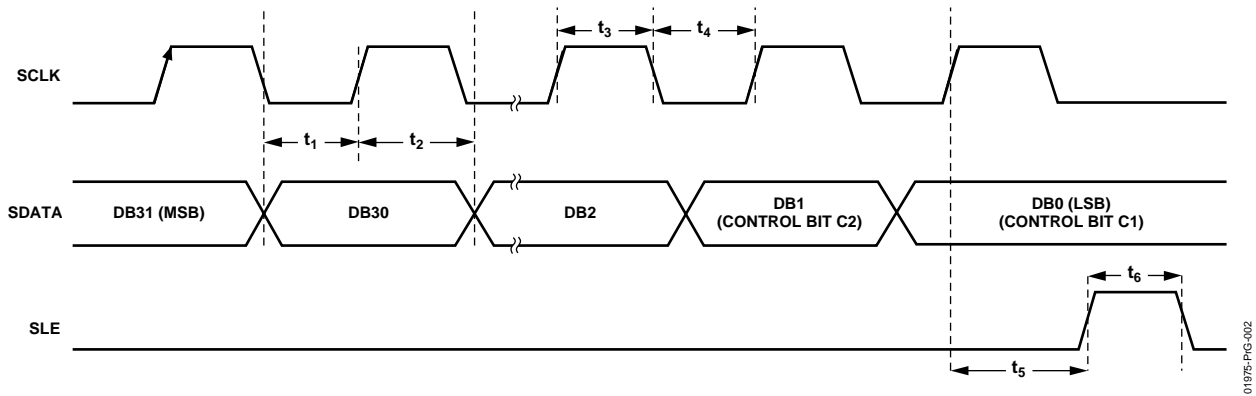


Figure 2. Serial Interface Timing Diagram

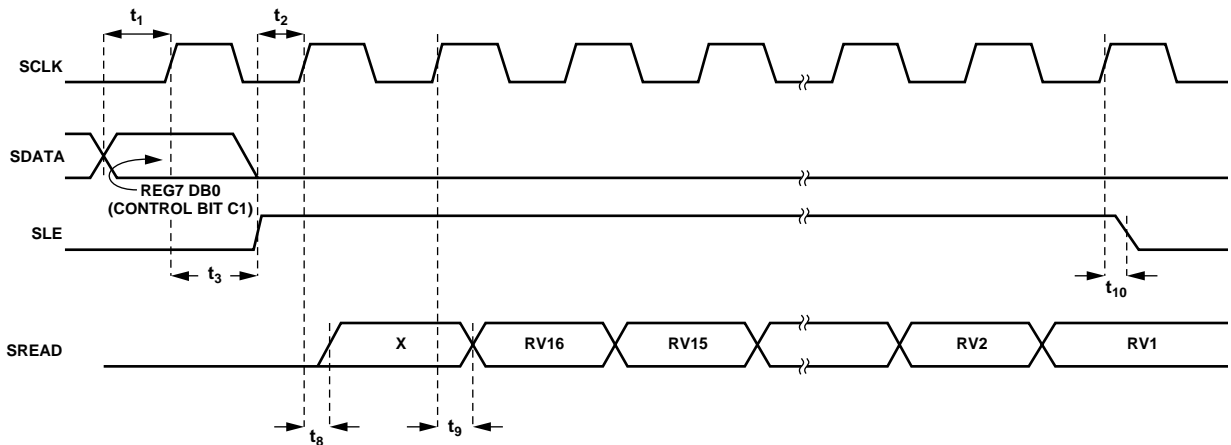


Figure 3. Readback Timing Diagram

## ABSOLUTE MAXIMUM RATINGS

$T_A = 25^\circ\text{C}$ , unless otherwise noted.

**Table 3.**

Parameter	Rating
$V_{DD}$ to GND <sup>1</sup>	-0.3 V to +5 V
Analog I/O Voltage to GND	-0.3 V to $AV_{DD} + 0.3$ V
Digital I/O Voltage to GND	-0.3 V to $DV_{DD} + 0.3$ V
Operating Temperature Range	
Industrial (B Version)	-40°C to +85°C
Storage Temperature Range	-65°C to +125°C
Maximum Junction Temperature	125°C
MLF $\theta_{JA}$ Thermal Impedance	TBD°C/W
Lead Temperature Soldering	
Vapor Phase (60 s)	235°C
Infrared (15 s)	240°C

<sup>1</sup> GND = CPGND = RFGND = DGND = AGND = 0 V.

## ESD 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 this product 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.



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

This device is a high-performance RF integrated circuit with an ESD rating of <2 kV and it is ESD sensitive. Proper precautions should be taken for handling and assembly.



# PIN CONFIGURATION AND FUNCTION DESCRIPTIONS

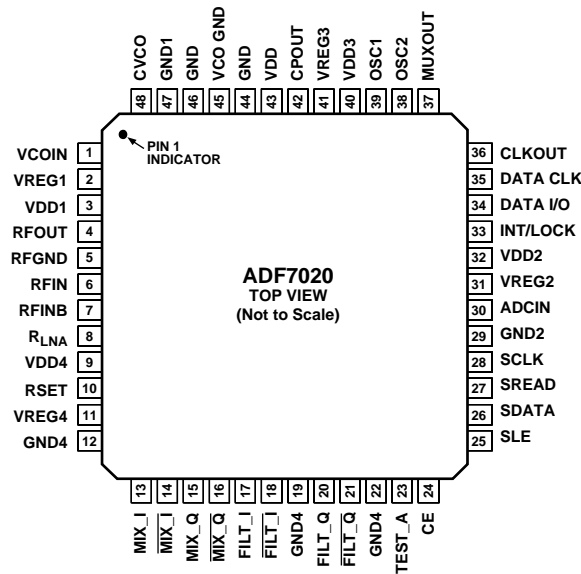


Figure 4. Pin Configuration

Table 4. Pin Function Descriptions

Pin No.	Mnemonic	Function
1	VCOIN	The tuning voltage on this pin determines the output frequency of the voltage controlled oscillator (VCO). The higher the tuning voltage, the higher the output frequency.
2	VREG1	Regulator Voltage for PA Block. A 100 nF capacitor should be placed between this pin and ground for regulator stability and noise rejection.
3	VDD1	Voltage Supply for PA Block. Decoupling capacitors (X7R or Tantalum) of 0.1 μF and 0.01 μF should be placed as close as possible to this pin.
4	RFOUT	The modulated signal is available at this pin. Output power levels are from –20 dBm to +13 dBm. The output should be impedance matched to the desired load using suitable components. See the Transmitter section.
5	RFGND	Ground for Output Stage of Transmitter.
6	RFIN	LNA Input for Receiver Section. Input matching is required between the antenna and the differential LNA input to ensure maximum power transfer. See the LNA/PA Matching section.
7	RFINB	Complementary LNA Input. See the LNA/PA Matching section.
8	RLNA	External bias resistor for LNA. Optimum resistor is 1.1 kΩ with 5% tolerance.
9	VDD4	Voltage supply for LNA/MIXER block. This pin should be decoupled to ground with a 0.01 μF capacitor.
10	RSET	External Resistor to Set Charge Pump Current and Some Internal Bias Currents. Use 3.6 kΩ with 5% tolerance.
11	VREG4	Regulator Voltage for LNA/MIXER block. A 100 nF capacitor should be placed between this pin and GND for regulator stability and noise rejection.
12	GND4	Ground for LNA/MIXER block.
13–18	MIX/FILT	Signal Chain Test Pins. These pins are high impedance under normal conditions and should be left unconnected.
19, 22	GND4	Ground for LNA/MIXER block.
20, 21, 23	FILT/TEST_A	Signal Chain Test Pins. These pins are high impedance under normal conditions and should be left unconnected.
24	CE	Chip Enable. Bringing CE low puts the ADF7020 into complete power-down. Register values are lost when CE is low, and the part must be reprogrammed once CE is brought high.
25	SLE	Load Enable, CMOS Input. When LE goes high, the data stored in the shift registers is loaded into one of the four latches. A latch is selected using the control bits.
26	SDATA	Serial Data Input. The serial data is loaded MSB first with the two LSBs as the control bits. This pin is a high impedance CMOS input.
27	SREAD	Serial Data Output. This pin is used to feed readback data from the ADF7020 to the microcontroller. The SCLK input is used to clock each readback bit (AFC, ADC readback) from the SREAD pin.
28	SCLK	Serial Clock Input. This serial clock is used to clock in the serial data to the registers. The data is latched into the 24-bit shift register on the CLK rising edge. This pin is a digital CMOS input.

Pin No.	Mnemonic	Function
29	GND2	Ground for Digital Section.
30	ADCIN	Analog-to-Digital Converter Input. The internal 7-bit ADC can be accessed through this pin. Full scale is 0 to 1.9 V. Readback is made using the SREAD pin.
31	VREG2	Regulator Voltage for Digital Block. A 100 nF capacitor should be placed between this pin and ground for regulator stability and noise rejection.
32	VDD2	Voltage Supply for Digital Block. A decoupling capacitor (X7R or Tantalum) of 0.01 $\mu$ F should be placed as close as possible to this pin.
33	INT/LOCK	Bidirectional Pin. In output mode (INTerrupt mode), the ADF7020 asserts the INT/ LOCK pin when it has found a match for the preamble sequence. In input mode (lock mode), the microcontroller can be used to lock the demodulator threshold when a valid preamble has been detected. Once the threshold is locked, NRZ data can be reliably received. In this mode, a demod lock can be asserted with minimum delay.
34	DATA I/O	Transmit Data Input/Received Data Output. This is a digital pin and normal CMOS levels apply.
35	DATA CLK	In receive mode, the pin outputs the synchronized data clock. The positive clock edge is matched to the center of the received data. In GFSK transmit mode, the pin outputs an accurate clock to latch the data from the microcontroller into the transmit section at the exact required data rate. See the Gaussian Frequency Shift Keying (GFSK) section.
36	CLKOUT	A Divided-Down Version of the Crystal Reference with Output Driver. The digital clock output can be used to drive several other CMOS inputs such as a microcontroller clock. The output has a 50:50 mark-space ratio.
37	MUXOUT	This pin provides the Lock_Detect signal, which is used to determine if the PLL is locked to the correct frequency. Other signals include Regulator_Ready, which is an indicator of the status of the serial interface regulator.
38	OSC2	The reference crystal should be connected between this pin and OSC1. A TCXO reference can be used by driving this pin with CMOS levels and disabling the crystal oscillator.
39	OSC1	The reference crystal should be connected between this pin and OSC2.
40	VDD3	Voltage Supply for the Charge Pump and PLL Dividers. This pin should be decoupled to ground with a 0.01 $\mu$ F capacitor.
41	VREG3	Regulator Voltage for Charge Pump and PLL Dividers. A 100 nF capacitor should be placed between this pin and ground for regulator stability and noise rejection.
42	CPOUT	Charge Pump Output. This output generates current pulses that are integrated in the loop filter. The integrated current changes the control voltage on the input to the VCO.
43	VDD	Voltage Supply for VCO Tank Circuit. This pin should be decoupled to ground with a 0.01 $\mu$ F capacitor.
44–47	GND	Grounds for VCO Block.
48	CVCO	A 22 nF capacitor should be placed between this pin and VREG1 to reduce VCO noise.

# FREQUENCY SYNTHESIZER

## REFERENCE INPUT SECTION

The on-board crystal oscillator circuitry (Figure 5) can use an inexpensive quartz crystal as the PLL reference. The oscillator circuit is enabled by setting R1\_DB12 high. It is enabled by default on power-up and is disabled by bringing CE low. Errors in the crystal can be corrected using the automatic frequency control (see the AFC Section) feature or by adjusting the fractional-N value (see the N Counter section). A single-ended reference (TCXO, CXO) can also be used. The CMOS levels should be applied to OSC2 with R1\_DB12 set low.

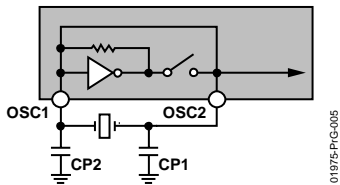


Figure 5. Oscillator Circuit on the ADF7020

Two parallel resonant capacitors are required for oscillation at the correct frequency; their values are dependent on the crystal specification. They should be chosen so that the series value of capacitance added to the PCB track capacitance adds up to the load capacitance of the crystal, usually 20 pF. Track capacitance values vary from 2 pF to 5 pF, depending on board layout. Where possible, choose capacitors that have a very low temperature coefficient to ensure stable frequency operation over all conditions.

## CLKOUT Divider and Buffer

The CLKOUT circuit takes the reference clock signal from the oscillator section, shown in Figure 5, and supplies a divided-down 50:50 mark-space signal to the CLKOUT pin. An even divide from 2 to 30 is available. This divide number is set in R1\_DB(8:11). On power-up, the CLKOUT defaults to divide-by-8.

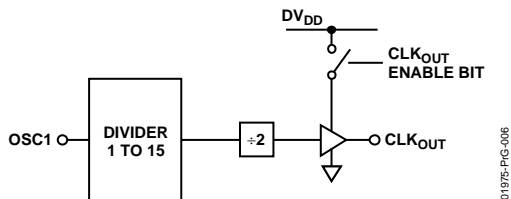


Figure 6. CLKOUT Stage

To disable CLKOUT, set the divide number to 0. The output buffer can drive up to a 20 pF load with a 10% rise time at 4.8 MHz. Faster edges can result in some spurious feedthrough to the output. A small series resistor (50 Ω) can be used to slow the clock edges to reduce these spurs at F<sub>CLK</sub>.

## R Counter

The 3-bit R counter divides the reference input frequency by an integer from 1 to 7. The divided-down signal is presented as the reference clock to the phase frequency detector (PFD). The divide ratio is set in Register 1. Maximizing the PFD frequency reduces the N value. This reduces the noise multiplied at a rate of 20 log(N) to the output, as well as reducing occurrences of spurious components. The R Register defaults to R = 1 on power-up:

$$PFD [Hz] = XTAL/R$$

## MUXOUT and Lock Detect

The MUXOUT pin allows the user to access various digital points in the ADF7020. The state of MUXOUT is controlled by Bits R0\_DB(29:31).

## Regulator Ready

REGULATOR READY is the default setting on MUXOUT after the transceiver has been powered up. The power-up time of the regulator is typically 50 μs. Because the serial interface is powered from the regulator, the regulator must be at its nominal voltage before the ADF7020 can be programmed. The status of the regulator can be monitored at MUXOUT. When the REGULATOR READY signal on MUXOUT is high, programming of the ADF7020 can begin.

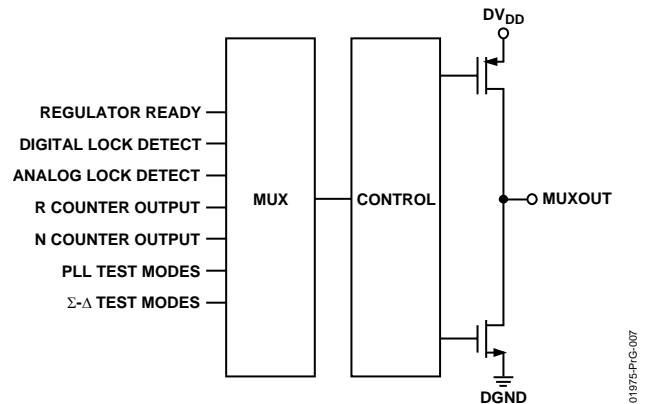


Figure 7. MUXOUT Circuit

## Digital Lock Detect

Digital lock detect is active high. The lock detect circuit is located at the PFD. When the phase error on five consecutive cycles is less than 15 ns, lock detect is set high. Lock detect remains high until 25 ns phase error is detected at the PFD. Because no external components are needed for digital lock detect, it is more widely used than analog lock detect.

### Analog Lock Detect

This N-channel open-drain lock detect should be operated with an external pull-up resistor of 10 k $\Omega$  nominal. When a lock has been detected, this output is high with narrow low-going pulses.

### Voltage Regulators

The ADF7020 contains four regulators to supply stable voltages to the part. The nominal regulator voltage is 2.3 V. Each regulator should have a 100 nF capacitor connected between VREG and GND. When CE is high, the regulators and other associated circuitry are powered on, drawing a total supply current of 2 mA. Bringing the chip-enable pin low disables the regulators, reduces the supply current to less than 1  $\mu$ A, and erases all values held in the registers. The serial interface operates off a regulator supply; therefore, to write to the part, the user must have CE high and the regulator voltage must be stabilized. Regulator status (VREG4) can be monitored using the regulator ready signal from MUXOUT.

### Loop Filter

The loop filter integrates the current pulses from the charge pump to form a voltage that tunes the output of the VCO to the desired frequency. It also attenuates spurious levels generated by the PLL. A typical loop filter design is shown in Figure 8.

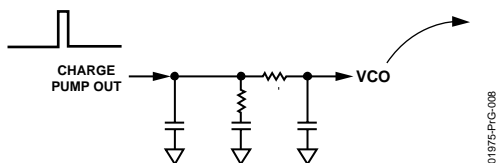


Figure 8. Typical Loop Filter Configuration

In FSK, the loop should be designed so that the loop bandwidth (LBW) is approximately five times the data rate. Widening the LBW excessively reduces the time spent jumping between frequencies, but can cause insufficient spurious attenuation.

For ASK systems, a wider LBW is recommended. The sudden large transition between two power levels might result in VCO pulling and can cause a wider output spectrum than is desired. By widening the LBW to more than 10 times the data rate, the amount of VCO pulling is reduced, because the loop settles quickly back to the correct frequency. The wider LBW might restrict the output power and data rate of ASK-based systems compared with FSK-based systems.

Narrow-loop bandwidths can result in the loop taking long periods of time to attain lock. Careful design of the loop filter is critical to obtaining accurate FSK/GFSK modulation.

For GFSK, it is recommended that an LBW of 2.0 to 2.5 times the data rate be used to ensure that sufficient samples are taken of the input data while filtering system noise. The free design tool ADIsimPLL can be used to design loop filters for the ADF7020.

### N Counter

The feedback divider in the ADF7020 PLL consists of an 8-bit integer counter and a 14-bit  $\Sigma$ - $\Delta$  fractional-N divider. The integer counter is the standard pulse-swallow type common in PLLs. This sets the minimum integer divide value to 31. The fractional divide value gives very fine resolution at the output, where the output frequency of the PLL is calculated as

$$F_{OUT} = \frac{XTAL}{R} \times \left( Integer-N + \frac{Fractional-N}{2^{14}} \right)$$

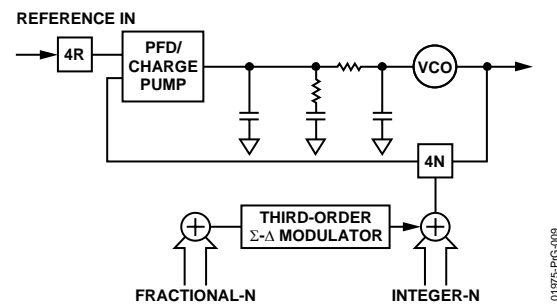


Figure 9. Fractional-N PLL

The combination of the integer-N (maximum = 255) and the fractional-N (maximum = 16383/16384) give a maximum N divider of 255 + 1. Therefore, the minimum usable PFD is

$$PDF_{MIN} [Hz] = \text{Maximum Required Output Frequency} / (255 + 1)$$

For example, when operating in the European 868 MHz to 870 MHz band,  $PDF_{MIN}$  equals 3.4 MHz.

### Voltage Controlled Oscillator (VCO)

To minimize spurious emissions, the on-chip VCO operates from 1732 MHz to 1856 MHz. The VCO signal is then divided by 2 to give the required frequency for the transmitter and the required LO frequency for the receiver.

The VCO should be recentered, depending on the required frequency of operation, by programming the VCO adjust bits R1\_DB(20:21).

The VCO is enabled as part of the PLL by the PLL-enable bit, R0\_DB28.

A further frequency divide-by-2 is included to allow operation in the lower 433 MHz and 460 MHz bands. To enable operation in the these bands, R1\_DB13 should be set to 1. The VCO needs an external 22 nF between the VCO and the regulator to reduce internal noise.

**VCO Bias Current**

VCO bias current can be adjusted using Bits R1\_DB19 to R1\_DB16. To ensure VCO oscillation, the minimum bias current setting under typical conditions is 2.5 mA.

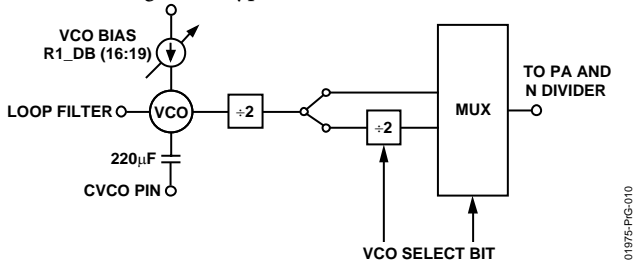


Figure 10. Voltage Controlled Oscillator (VCO)

**CHOOSING CHANNELS FOR BEST SYSTEM PERFORMANCE**

The fractional-N PLL allows the selection of any channel within 868 MHz to 928 MHz (and 433MHz using divide-by-2) to a resolution of <100 Hz. This also facilitates frequency hopping systems.

Careful selection of the RF transmit channels must be made to achieve best spurious performance. The architecture of fractional-N results in some level of the nearest integer channel moving through the loop to the RF output. These “beat-note” spurs are not attenuated by the loop, if the desired RF channel and the nearest integer channel are separated by a frequency of less than the LBW.

The occurrence of beat-note spurs is rare, because the integer frequencies are at multiples of the reference, which is typically >10 MHz.

Beat-note spurs can be significantly reduced in amplitude by avoiding very small or very large values in the fractional register, using the frequency doubler. By having a channel 1 MHz away from an integer frequency, a 100 kHz loop filter can reduce the level to <-45 dBc. When using an external VCO, the fast lock (bleed) function reduces the spurs to <-60 dBc for the same conditions.

# TRANSMITTER

## RF OUTPUT STAGE

The PA of the ADF7020 is based on a single-ended, controlled current, open-drain amplifier that has been designed to deliver up to 13 dBm into a 50 Ω load at a maximum frequency of 928 MHz.

The PA output current and, consequently, the output power are programmable over a wide range. The PA configurations in FSK/GFSK and ASK/OOK modulation modes are shown in Figure 11 and Figure 12, respectively. In FSK/GFSK modulation mode, the output power is independent of the state of the DATA\_IO pin. In ASK/OOK modulation mode, it is dependent on the state of the DATA\_IO pin and Bit R2\_DB29, which selects the polarity of the TxData input. For each transmission mode, the output power can be adjusted as follows:

- FSK/GFSK: The output power is set using bits R2\_DB(9:14).
- ASK: The output power for the inactive state of the TxData input is set by Bits R2\_DB(15:20). The output power for the active state of the TxData input is set by Bits R2\_DB(9:14).
- OOK: The output power for the active state of the TxData input is set by Bits R2\_DB(9:14). The PA is muted when the TxData input is inactive.

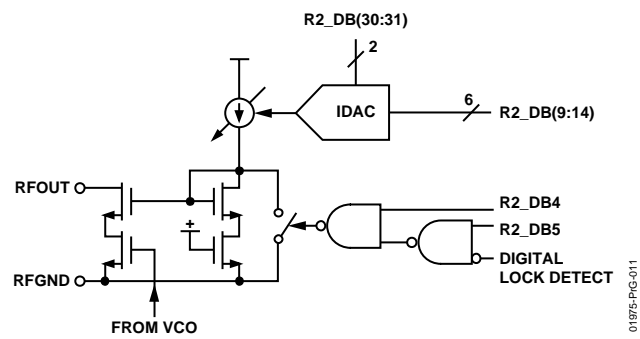


Figure 11. PA Configuration in FSK/GFSK Mode

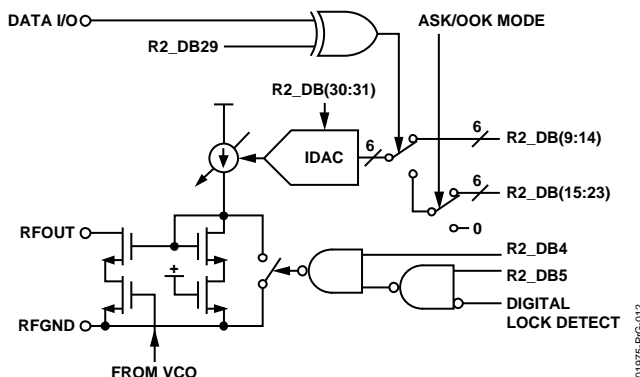


Figure 12. PA Configuration in ASK/OOK Mode

The PA is equipped with overvoltage protection, which makes it robust in severe mismatch conditions. Depending on the application, one can design a matching network for the PA to exhibit optimum efficiency at the desired radiated output power level for a wide range of different antennas, such as loop or monopole antennas. See the LNA/PA Matching section for details.

### PA Bias Currents and Mute PA until Lock Bit

Control Bits R2\_DB(30:31) facilitate an adjustment of the PA bias current to further extend the output power control range, if necessary. If this feature is not required, the default value of 7 μA is recommended. The output stage is powered down by resetting Bit R2\_DB4. To reduce the level of undesired spurious emissions, the PA can be muted during the PLL lock phase by setting Bit R2\_DB5 (mute PA until lock bit).

## MODULATION SCHEMES

### Frequency Shift Keying (FSK)

Frequency shift keying is implemented by setting the N value for the center frequency and then toggling this with the TxData line. The deviation from the center frequency is set using Bits R2\_DB(15:23). The deviation from the center frequency in Hz is

$$FSK_{DEVIATION} [Hz] = \frac{PFD \times Modulation\ Number}{2^{14}}$$

where *Modulation Number* is a number from 1 to 511 (R2\_DB(15:23)).

Select FSK using Bits R2\_DB(6:8).

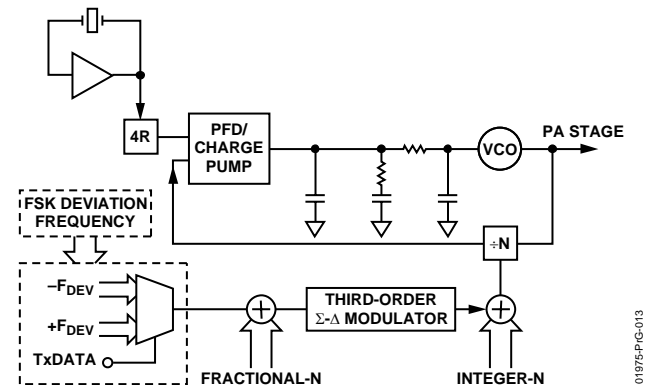


Figure 13. FSK Implementation

**Gaussian Frequency Shift Keying (GFSK)**

Gaussian frequency shift keying reduces the bandwidth occupied by the transmitted spectrum by digitally prefiltering the TxData. A TxCLK output line is provided from the ADF7020 for synchronization of TxData from the microcontroller. The TxCLK line can be connected to the clock input of a shift register that clocks data to the transmitter at the exact data rate.

**Setting Up the ADF7020 for GFSK**

To set up the frequency deviation, set the PFD and the mod control bits:

$$GFSK_{DEVIATION} [\text{Hz}] = \frac{PFD \times 2^m}{2^{12}}$$

where  $m$  is GFSK\_MOD\_CONTROL set using R2\_DB(24:26).

To set up the GFSK data rate:

$$DR [\text{bps}] = \frac{PFD}{DIVIDER\_FACTOR \times INDEX\_COUNTER}$$

For further information, see the application note, *Using GFSK on the ADF7010*, in the EVAL-ADF7010EB1 Technical Note.

**Amplitude Shift Keying (ASK)**

Amplitude shift keying is implemented by switching the output stage between two discrete power levels. This is accomplished by toggling the DAC, which controls the output level between two 6-bit values set up in Register 2. A zero TxData bit sends Bits R2\_DB(15:20) to the DAC. A high TxData bit sends Bits R2\_DB(9:14) to the DAC. A maximum modulation depth of 30 dB is possible.

**On-Off Keying (OOK)**

On-off keying is implemented by switching the output stage to a certain power level for a high TxData bit and switching the output stage off for a zero. For OOK, the transmitted power for a high input is programmed using Bits R2\_DB(9:14).

**Gaussian On-Off Keying (G-OOK)**

Gaussian on-off keying represents a prefiltered form of OOK modulation. The usually sharp symbol transitions are replaced with smooth Gaussian filtered transitions, the result being a reduction in frequency pulling of the VCO. Frequency pulling of the VCO in OOK mode can lead to a wider than desired BW, especially if it is not possible to increase the loop filter BW > 300 kHz. The G-OOK sampling clock samples data at the data rate. (See the Setting Up the ADF7020 for GFSK section.)

## RECEIVER SECTION

### RF FRONT END

The ADF7020 is based on a fully integrated, low IF receiver architecture. The low IF architecture facilitates a very low external component count and does not suffer from power-line-induced interference problems.

Figure 14 shows the structure of the receiver front end. The many programming options allow users to trade off sensitivity, linearity, and current consumption against each other in the way best suitable for their applications. To achieve a high level of resilience against spurious reception, the LNA features a differential input. Switch SW2 shorts the LNA input when transmit mode is selected (R0\_DB27 = 0). This feature facilitates the design of a combined LNA/PA matching network, avoiding the need for an external Rx/Tx switch. See the LNA/PA Matching section for details on the design of the matching network.

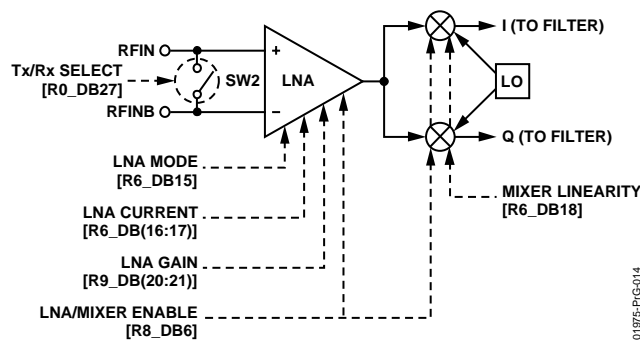


Figure 14. ADF7020 RF Front End

The LNA is followed by a quadrature downconversion mixer, which converts the RF signal to the IF frequency of 200 kHz. It is important to consider that the output frequency of the synthesizer must be programmed to a value 200 kHz below the center frequency of the received channel.

The LNA has two basic operating modes: high gain/low noise mode and low gain/low power mode. To switch between these two modes, use the LNA\_mode bit, R6\_DB15. The mixer is also configurable between a low current and an enhanced linearity mode using the mixer\_linearity bit, R6\_DB18.

Based on the specific sensitivity and linearity requirements of the application, it is recommended to adjust control bits LNA\_mode (R6\_DB15) and mixer\_linearity (R6\_DB18) as outlined in Table 5.

The gain of the LNA is configured by the LNA\_gain field, R9\_DB(20:21), and can be set by either the user or the AGC logic.

### IF Filter Settings/Calibration

Out-of-band interference is rejected by means of a fourth-order Butterworth polyphase IF filter centered around a frequency of 200 kHz. The bandwidth of the IF filter can be programmed between 100 kHz and 200 kHz by means of Control Bits R1\_DB(22:23), and should be chosen as a compromise between interference rejection, attenuation of the desired signal, and the AFC pull-in range.

To compensate for manufacturing tolerances, the IF filter should be calibrated once after power-up. The IF filter calibration logic requires that the IF filter divider in Bits R6\_DB(20:28) be set dependent on the crystal frequency. Once initiated by setting Bit R6\_DB19, the calibration is performed automatically without any user intervention. The calibration time is 200 μs, during which the ADF7020 should not be accessed. It is important not to initiate the calibration cycle before the crystal oscillator has fully settled. If the AGC loop is disabled, the gain of IF filter can be set to three levels using the filter\_gain field, R9\_DB(20:21). The filter gain is adjusted automatically, if the AGC loop is enabled.

The signal in the image channel of the low IF mixer, located at a frequency of 400 kHz below the desired channel, is rejected due to the image rejection of the polyphase filter. The image rejection performance of the IF filter is subject to manufacturing tolerances, and, to some extent, temperature drift. To improve the image rejection, a calibration procedure can be performed as outlined in the Image Rejection Calibration section.

Table 5. LNA/Mixer Modes

Receiver Mode	LNA Mode (R6_DB15)	LNA Gain Value R9_DB(21:20)	Mixer Linearity (R6_DB18)	Sensitivity (DR = 9.6 kbps, f <sub>DEV</sub> = 10 kHz)	Rx Current Consumption (mA)	Input IP3 (dBm)
High Sensitivity Mode (default)	0	30	0	-110.5	22	-35
RxMode2	1	10	0	-104	20	-15.9
Low Current Mode	1	3	0	-91	19	-3.2
Enhanced Linearity Mode	1	3	1	-101	19	6.8
RxMode5	1	10	1	TBD	TBD	-8.25
RxMode6	0	30	1	TBD	TBD	-28.8



**RSSI/AGC SECTION**

The RSSI is implemented as a successive compression log amp following the base-band channel filtering. The log amp achieves ±3 dB log linearity. It also doubles as a limiter to convert the signal-to-digital levels for the FSK demodulator. The RSSI itself is used for amplitude shift keying (ASK) demodulation. In ASK mode, extra digital filtering is performed on the RSSI value. Offset correction is achieved using a switched capacitor integrator in feedback around the log amp. This uses the BB offset clock divide. The RSSI level is converted for user readback and digitally controlled AGC by an 80-level (7-bit) flash ADC. This level can be converted to input power in dBm.

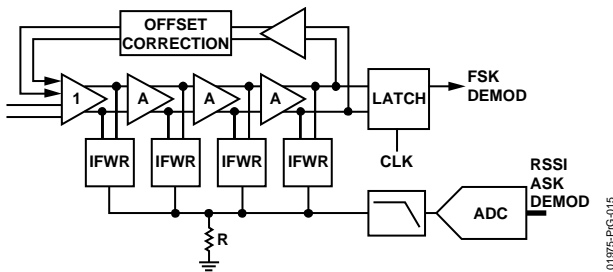


Figure 15. RSSI Block Diagram

**RSSI Thresholds**

When the RSSI is above AGC\_HIGH\_THRESHOLD, the gain is reduced. When the RSSI is below AGC\_LOW\_THRESHOLD, the gain is increased. A delay (AGC\_DELAY) is programmed to allow for settling of the loop. The user programs the two threshold values (recommended defaults, 27 and 76) and the delay (default, 10). The default AGC setup values should be adequate for most applications. The threshold values must be chosen to be more than 30 apart for the AGC to operate correctly.

**Offset Correction Clock**

In Register 3, the user should set the BB offset clock divide bits R3\_DB(4:5) to give an offset clock between 1 MHz and 2 MHz, where:

$$BBOS\_CLK [Hz] = XTAL / (BBOS\_CLK\_DIVIDE)$$

BBOS\_CLK\_DIVIDE can be set to 4, 8, or 16.

**AGC Information**

In Register 9, the user should select automatic gain control by selecting auto in R9\_DB18 and auto in R9\_DB19. The user should then program AGC low threshold R9\_DB(4:10) and AGC high threshold R9\_DB(11:17). The recommended/default values for the low and high thresholds are 30 and 70, respectively. In the AGC2 register the user should program the AGC delay to be long enough to allow the loop to settle. The recommended value is 10.

**RSSI Formula (Converting to dBm)**

$$Input\_Power [dBm] = -110 dBm + (Readback\_Code + Gain\_Mode\_Correction) \times 0.5$$

where:

Readback\_Code is given by Bits RV7 to RV1 in the readback register (see Readback Format section).

Gain\_Mode\_Correction is given by the values in Table 6.

LNA gain and filter gain (LG2/LG1, FG2/FG1) are also obtained from the readback register.

Table 6. Gain Mode Correction Table

LNA Gain (LG2, LG1)	Filter Gain (FG2, FG1)	Gain Mode Correction
H (10)	H (10)	0
M (01)	H (10)	11
M (01)	M (01)	19 + 11 = 30
M (01)	L (00)	19 + 19 + 11 = 49
L (00)	L (00)	19 + 19 + 19 + 11 = 68

An additional factor should be introduced to account for losses in the front-end matching network/antenna.

**FSK DEMODULATORS ON THE ADF7020**

The two FSK demodulators on the ADF7020 are

- FSK correlator/demodulator
- Linear demodulator

Select these using the demod select bits, R4\_DB(4:5).

**FSK CORRELATOR/DEMODULATOR**

The quadrature outputs of the IF filter are first limited and then fed to a pair of digital frequency correlators that perform band-pass filtering of the binary FSK frequencies at (IF + F<sub>DEV</sub>) and (IF - F<sub>DEV</sub>). Data is recovered by comparing the output levels from each of the two correlators. The performance of this frequency discriminator approximates that of a matched filter detector, which is known to provide optimum detection in the presence of AWGN.

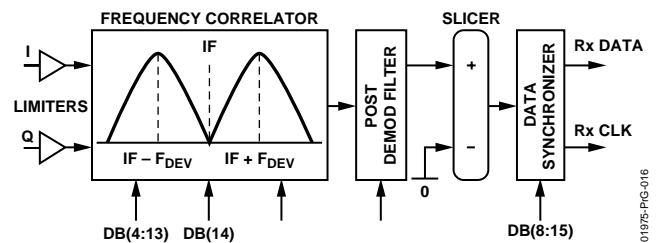


Figure 16. FSK Correlator/Demodulator Block Diagram

**Postdemodulator Filter**

A second-order, digital low-pass filter removes excess noise from the demodulated bit stream at the output of the discriminator. The bandwidth of this postdemodulator filter is programmable and must be optimized for the user’s data rate. If the bandwidth is set too narrow, performance is degraded due to intersymbol interference (ISI). If the bandwidth is set too wide, excess noise degrades the receiver’s performance. Typically, the 3 dB bandwidth of this filter is set at approximately 0.75 times the user’s data rate, using Bits R4\_DB(6:15).

**Bit Slicer**

The received data is recovered by threshold detecting the output of the postdemodulator low-pass filter. In the correlator/demodulator, the binary output signal levels of the frequency discriminator are always centered on zero. Therefore, the slicer threshold level can be fixed at zero and the demodulator performance is independent of the run-length constraints of the transmit data bit stream. This results in robust data recovery, which does not suffer from the classic baseline wander problems that exist in the more traditional FSK demodulators.

Frequency errors are removed by an internal AFC loop that measures the average IF frequency at the limiter output and applies a frequency correction value to the fractional-N synthesizer. This loop should be activated when the frequency errors are greater than approximately 40% of the transmit frequency deviation (see the AFC Section).

**Data Synchronizer**

An oversampled digital PLL is used to resynchronize the received bit stream to a local clock. The oversampled clock rate of the PLL (CDR\_CLK) must be set at 32 times the data rate. See the notes for the Register 3—Receiver Clock Register section for a definition of how to program. The clock recovery PLL can accommodate frequency errors of up to ±2%.

**FSK Correlator Register Settings**

To enable the FSK correlator/demodulator, Bits R4\_DB(5:4) should be set to [01]. To achieve best performance, the bandwidth of the FSK correlator must be optimized for the specific deviation frequency that is used by the FSK transmitter.

The discriminator BW is controlled in Register 6 by R6\_DB(4:13) and is defined as

$$Discriminator\_BW = (DEMODO\_CLK \times K) / (800 \times 10^3)$$

where:

DEMODO\_CLK is as defined in the Register 3—Receiver Clock Register section, Note 2.

$$K = Round(200e3 / FSK\ Deviation)$$

To optimize the coefficients of the FSK correlator, two additional bits, R6\_DB14 and R6\_DB29, must be assigned. The value of these bits depends on whether K (as defined above) is odd or even. These bits are assigned according to Table 7 and Table 8.

**Table 7. When K Is Even**

K	K/2	R6_DB14	R6_DB29
Even	Even	0	0
Even	Odd	0	1

**Table 8. When K Is Odd**

K	(K + 1)/2	R6_DB14	R6_DB29
Odd	Even	1	0
Odd	Odd	1	1

**Postdemodulator Bandwidth Register Settings**

The 3 dB bandwidth of the postdemodulator filter is controlled by Bits R4\_DB(6:15) and is given by

$$Post\_Demod\_BWSetting = \frac{2^{10} \times 2\pi \times F_{CUTOFF}}{DEMODO\_CLK}$$

where F\_CUTOFF is the target 3 dB bandwidth in Hz of the postdemodulator filter. This should typically be set to 0.75 times the data rate (DR).

Some sample settings for the FSK correlator/demodulator are

- DEMODO\_CLK = 5 MHz
- DR = 9.6 kbps
- F\_DEV = 20 kHz

Therefore,

$$F_{CUTOFF} = 0.75 \times 9.6 \times 10^3 \text{ Hz}$$

$$Post\_Demod\_BW = 2^{11} \pi \times 7.2 \times 10^3 \text{ Hz} / (5 \text{ MHz})$$

$$Post\_Demod\_BW = Round(9.26) = 9$$

and

$$K = Round(200 \text{ kHz} / 20 \text{ kHz}) = 10$$

$$Discriminator\_BW = (5 \text{ MHz} \times 10) / (800 \times 10^3) = 62.5 = 63 \text{ (rounded to nearest integer)}$$

**Table 9.**

Setting Name	Register Address	Value
Post_Demod_BW	R4_DB(6:15)	0x09
Discriminator BW	R6_DB(4:13)	0x3F
Dot Product	R6_DB14	0
Rx Data Invert	R6_DB29	0

### LINEAR FSK DEMODULATOR

A block diagram of the linear FSK demodulator is shown in Figure 17.

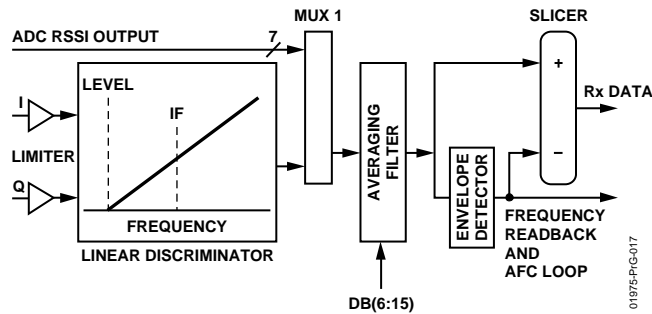


Figure 17. Block Diagram of Frequency Measurement System and ASK.OOK/Linear FSK Demodulator

This method of frequency demodulation is useful when very short preamble length is required and the system protocol cannot support the overhead of the settling time of the internal feedback AFC loop settling.

A digital frequency discriminator provides an output signal that is linearly proportional to the frequency of the limiter outputs. The discriminator output is then filtered and averaged using a combined averaging filter and envelope detector. The demodulated FSK data is recovered by threshold-detecting the output of the averaging filter, as shown in Figure 17. In this mode, the slicer output shown in Figure 17 is routed to the data synchronizer PLL for clock synchronization. To enable the linear FSK demodulator, set Bits R4\_DB(4:5) to [00].

The 3 dB bandwidth of the postdemodulation filter is set in the same way as the FSK correlator/demodulator, which is set in R4\_DB(6:15) and is defined as

$$Post\_Demod\_BW\_Setting = \frac{2^{10} \times 2\pi \times F_{CUTOFF}}{DEMOM\_CLK}$$

where:

$F_{CUTOFF}$  is the target 3 dB bandwidth in Hz of the postdemodulator filter.

$DEMOM\_CLK$  is as defined in the Register 3—Receiver Clock Register section, Note 2.

#### ASK/OOK Operation

ASK/OOK demodulation is activated by setting Bits R4\_DB(4:5) to [10].

Digital filtering and envelope detecting the digitized RSSI input via MUX 1, as shown in Figure 17, perform ASK/OOK demodulation. The bandwidth of the digital filter must be optimized to remove any excess noise without causing ISI in the received ASK/OOK signal.

The 3 dB bandwidth of this filter is typically set at approximately 0.75 times the user data rate and is assigned by R4\_DB(6:15) as

$$Post\_Demod\_BW\_Setting = \frac{2^{10} \times 2\pi \times F_{CUTOFF}}{DEMOM\_CLK}$$

where  $F_{CUTOFF}$  is the target 3 dB bandwidth in Hz of the postdemodulator filter.

### AFC SECTION

The ADF7020 supports a real-time AFC loop, which is used to remove frequency errors that can arise due to mismatches between the transmit and receive crystals. This uses the frequency discriminator block, as described in the Linear FSK Demodulator section (see Figure 17). The discriminator output is filtered and averaged using a combined averaging filter and envelope detector. In FSK mode, the output of the envelope detector provides an estimate of the average IF frequency.

Two methods of AFC, external and internal, are supported on the ADF7020 (in FSK mode only).

#### External AFC

The user reads back the frequency information through the ADF7020 serial port and applies a frequency correction value to the fractional-N synthesizer's N divider.

The frequency information is obtained by reading the 16-bit signed AFC\_readback, as described in the Readback Format section, and applying the following formula:

$$FREQ\_RB [Hz] = (AFC\_READBACK \times DEMOM\_CLK) / 2^{15}$$

Note that while the  $AFC\_READBACK$  value is a signed number, under normal operating conditions it is positive. In the absence of frequency errors, the  $FREQ\_RB$  value is equal to the IF frequency of 200 kHz.

#### Internal AFC

The ADF7020 supports a real-time internal automatic frequency control loop. In this mode, an internal control loop automatically monitors the frequency error and adjusts the synthesizer N divider using an internal PI control loop.

The internal AFC control loop parameters are controlled in Register 11. The internal AFC loop is activated by setting R11\_DB20 to 1. A scaling coefficient must also be entered, based on the crystal frequency in use. This is set up in R11\_DB(4:19) and should be calculated using

$$AFC\_Scaling\_Coefficient = (500 \times 2^{24}) / XTAL$$

Therefore, using a 10 MHz XTAL yields an AFC scaling coefficient of 839.

**Maximum AFC Range**

The maximum AFC frequency range is  $\pm 100$  kHz. This is set by the maximum IF filter bandwidth of 200 kHz. Using the minimum IF filter bandwidth of 100 kHz, the AFC range is  $\pm 50$  kHz.

When AFC errors have been removed using either the internal or external AFC, further improvement in the receiver's sensitivity can be obtained by reducing the IF filter bandwidth using Bits R1\_DB(22:23).

**AUTOMATIC SYNC WORD RECOGNITION**

The ADF7020 also supports automatic detection of the sync or ID fields. To activate this mode, the sync (or ID) word must be preprogrammed into the ADF7020. In receive mode, this preprogrammed word is compared to the received bit stream and, when a valid match is identified, the external pin INT/LOCK is asserted by the ADF7020.

This feature can be used to alert the microprocessor that a valid channel has been detected. It relaxes the computational requirements of the microprocessor and reduces the overall power consumption. The INT/LOCK is automatically de-asserted again after nine data clock cycles.

The automatic sync/ID word detection feature is enabled by selecting demod mode 2 or 3 in the demodulator setup register. Do this by setting R4\_DB(25:23) = [010] or [011]. Bits R5\_DB(4:5) are used to set the length of the sync/ID word, which can be either 12, 16, 20, or 24 bits long. The transmitter must transmit the MSB of the sync byte first and the LSB last to ensure proper alignment in the receiver sync byte detection hardware.

For systems using FEC, an error tolerance parameter can also be programmed that accepts a valid match when up to three bits of the word are incorrect. The error tolerance value is assigned in R5\_DB(6:7).

## APPLICATIONS SECTION

### LNA/PA MATCHING

The ADF7020 exhibits optimum performance in terms of sensitivity, transmit power, and current consumption only if its RF input and output ports are properly matched to the antenna impedance. For cost-sensitive applications, the ADF7020 is equipped with an internal Rx/Tx switch, which facilitates the use of a simple combined passive PA/LNA matching network. Alternatively, an external Rx/Tx switch such as the Analog Devices ADG919 can be used, which yields a slightly improved receiver sensitivity and lower transmitter power consumption.

#### External Rx/Tx Switch

Figure 18 shows a configuration using an external Rx/Tx switch. This configuration allows an independent optimization of the matching and filter network in the transmit and receive path, and is, therefore, more flexible and less difficult to design than the configuration using the internal Rx/Tx switch. The PA is biased through inductor L1, while C1 blocks dc current. Both elements, L1 and C1, also form the matching network, which transforms the source impedance into the optimum PA load impedance,  $Z_{OPT\_PA}$ .

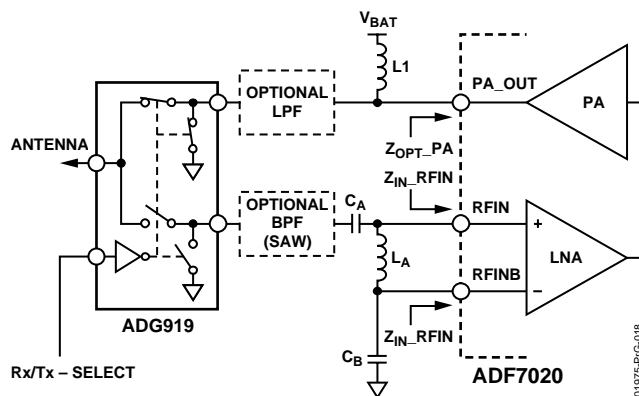


Figure 18. ADF7020 with External Rx/Tx Switch

$Z_{OPT\_PA}$  depends on various factors such as the required output power, the frequency range, the supply voltage range, and the temperature range. Selecting an appropriate  $Z_{OPT\_PA}$  helps to minimize the Tx current consumption in the application. This datasheet contains a number of  $Z_{OPT\_PA}$  values for representative conditions. Under certain conditions, however, it is recommended to obtain a suitable  $Z_{OPT\_PA}$  value by means of a load-pull measurement.

Due to the differential LNA input, the LNA matching network must be designed to provide both a single-ended to differential conversion and a complex conjugate impedance match. The network with the lowest component count that can satisfy these requirements is the configuration shown in Figure 18, which consists of two capacitors and one inductor. A first-order implementation of the matching network can be obtained by understanding the arrangement as two L-type matching

networks in a back-to-back configuration. Due to the asymmetry of the network with respect to ground, a compromise between the input reflection coefficient and the maximum differential signal swing at the LNA input must be established. The use of appropriate CAD software is strongly recommended for this optimization.

Depending on the antenna configuration, the user might need a harmonic filter at the PA output to satisfy the spurious emission requirement of the applicable government regulations. The harmonic filter can be implemented in various ways, such as a discrete LC-filter. Dielectric low-pass filter components such as the LFL18924MTC1A052 (for operation in the 915 MHz band), or LFL18869MTC2A160 (for operation in the 868 MHz band), both by Murata Mfg. Co., Ltd., represent an attractive alternative to discrete designs. The immunity of the ADF7020 to strong out-of-band interference can be improved by adding a band-pass filter in the Rx path. Apart from discrete designs, SAW or dielectric filter components such as the SAFCH869MAM0T00B0S, SAFCH915MAL0N00B0S, DCFB2869MLEJAA-TT1, or DCFB3915MLDJAA-TT1, all by Murata Mfg. Co., Ltd., are well suited for this purpose.

#### Internal Rx/Tx Switch

Figure 19 shows the ADF7020 in a configuration where the internal Rx/Tx switch is used with a combined LNA/PA matching network. Depending on the application, the slight performance degradation caused by the internal Rx/Tx switch might be acceptable, allowing the user to take advantage of the cost-saving potential of this solution. The design of the combined matching network must compensate for the reactance presented by the networks in the Tx and the Rx paths, taking the state of the Rx/Tx switch into consideration.

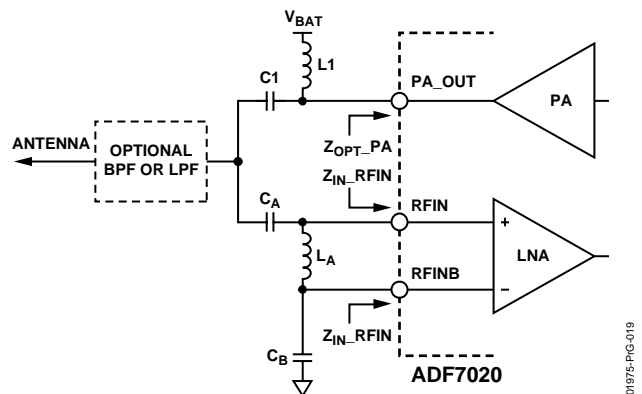


Figure 19. ADF7020 with Internal Rx/Tx Switch

The procedure typically requires several iterations until an acceptable compromise has been reached. The successful implementation of a combined LNA/PA matching network for the ADF7020 is critically dependent on the availability of an

accurate electrical model for the PC board. In this context, the use of a suitable CAD package is strongly recommended. To avoid this effort, the reference design provided for the ADF7020 RF module can be used.

As with the external Rx/Tx switch, an additional LPF or BPF might be required to suppress harmonics in the transmit spectrum or to improve the resilience of the receiver against out-of-band interferers.

**TRANSMIT PROTOCOL AND CODING CONSIDERATIONS**

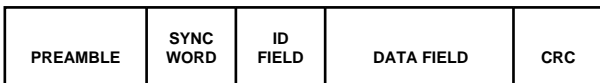


Figure 20. Typical Format of a Transmit Protocol

A dc-free preamble pattern is recommended for FSK/ASK/OOK demodulation. The recommended preamble pattern is a dc-free pattern such as a 10101010... pattern. Preamble patterns with longer run-length constraints such as 11001100.... can also be used. However, this results in a longer synchronization time of the received bit stream in the receiver.

Manchester coding can be used for the entire transmit protocol. However, the remaining fields that follow the preamble header do not have to use dc-free coding. For these fields, the ADF7020 can accommodate coding schemes with a run-length of up to 6 bits without any performance degradation.

If longer run-length coding must be supported, the ADF7020 has several other features that can be activated. These involve a range of programmable options that allow the envelope detector output to be frozen after preamble acquisition.

**IMAGE REJECTION CALIBRATION**

The image channel in this receiver, with an IF at 200 kHz, is at -200 kHz or +400 kHz below the desired signal. The polyphase filter rejects this image with an asymmetric frequency response. The image rejection performance of the receiver is dependent on how well matched in amplitude the I and Q signals are, and how perfect the quadrature is between them, that is, how close to 90° apart they are. The uncalibrated image rejection performance is approximately 30 dB. However, it is possible to improve on this performance by adjusting the I/Q phase/gain adjust bits in Register 10, resulting in an image rejection of approximately 45 dB.

Bits R10\_DB(24:27) adjust the relative phase of the signal and Bits R10\_DB(16:20) adjust the relative amplitude (see Figure 20).

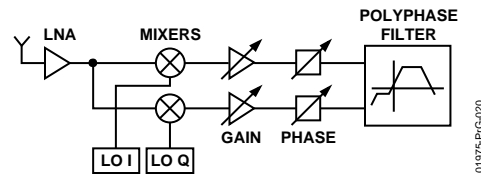


Figure 20. Phase/Gain Adjustment on ADF7020

**DEVICE PROGRAMMING AFTER INITIAL POWER-UP**

Basic mode is the minimum number of write sequences to power up the device. Enhanced mode uses the additional features of the ADF7020 to tailor the part to a particular application such as setting up a sync byte sequence or doing automatic frequency control.

The sample setting is for the following setup:

$$FRF = 915 \text{ MHz, FSK, DR} = 9.868 \text{ kbps, } I_{CP} = 1.44 \text{ mA}$$

$$F_{DEV} = 50 \text{ kHz, XTAL} = 10 \text{ MHz, Correlator/Demodulator}$$

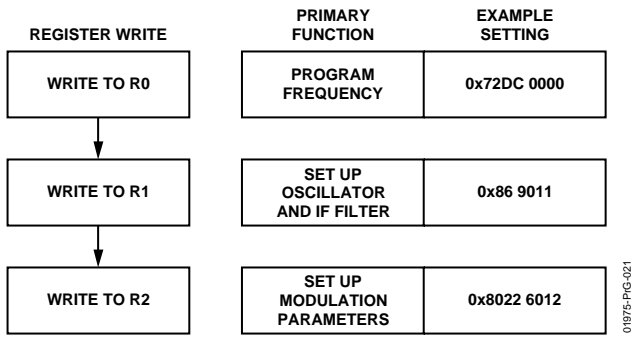


Figure 21. Basic Mode—Tx

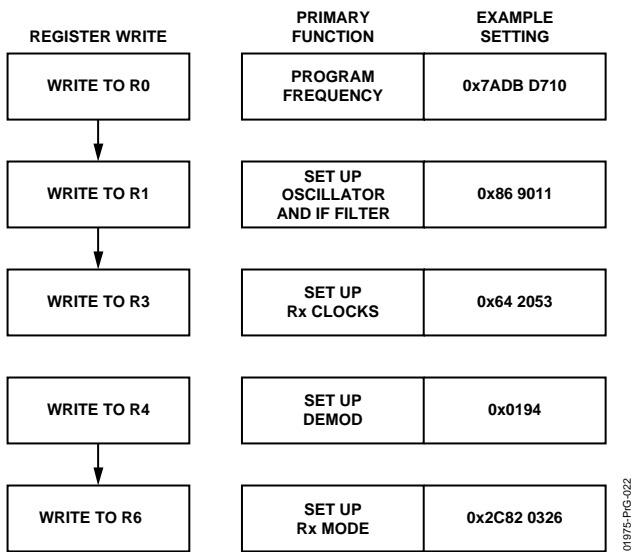


Figure 22. Basic Mode—Rx

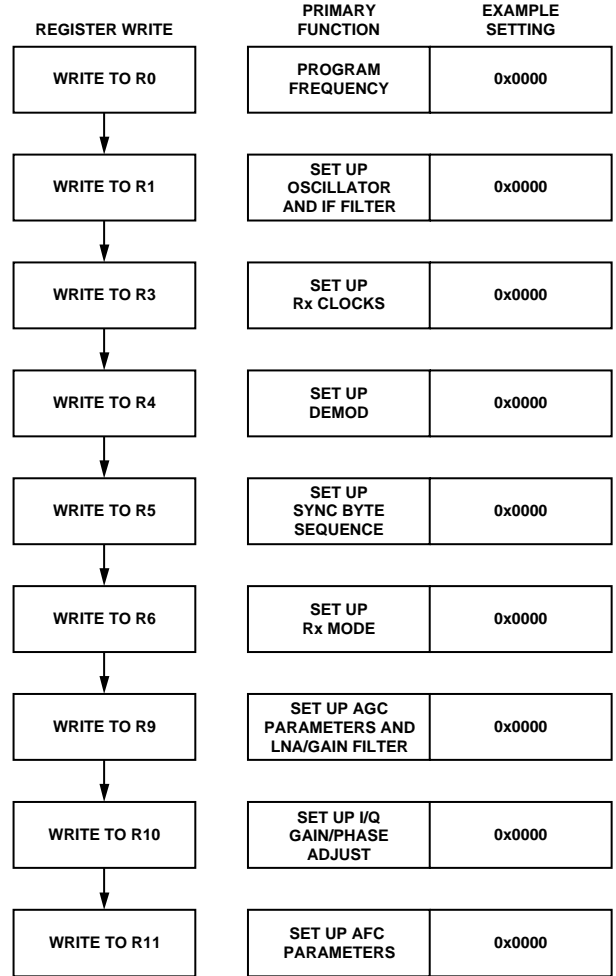


Figure 23. Enhanced Mode—Rx

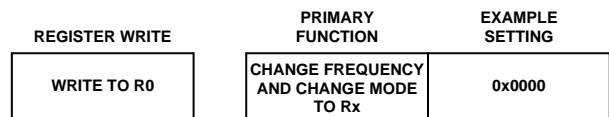


Figure 24. Change Mode from Tx to Rx

## SERIAL INTERFACE

The serial interface allows the user to program the eleven 32-bit registers using a 3-wire interface (SCLK, SDATA, and SLE). It consists of a level shifter, 32-bit shift register and eleven latches. Signals should be CMOS compatible. The serial interface is powered by the regulator, and, therefore, is inactive when CE is low.

Data is clocked into the register, MSB first, on the rising edge of each clock (SCLK). Data is transferred to one of eleven latches on the rising edge of SLE. The destination latch is determined by the value of the four control bits (C4 to C1). These are the bottom four LSBs, DB3 to DB0, as shown in the timing diagram in Figure 2. Data can also be read back on the SREAD pin.

### READBACK FORMAT

The readback operation is initiated by writing a valid control word to the readback register and setting the readback-enable bit ( $R7\_DB8 = 1$ ). The readback can begin after the control word has been latched with the SLE signal. SLE must be kept high while the data is being read out. Each active edge at the SCLK pin clocks the readback word out successively at the SREAD pin, as shown in Figure 25, starting with the MSB first. The data appearing at the first clock cycle following the latch operation must be ignored.

#### AFC Readback

The AFC readback is valid only during the reception of FSK signals with either the linear or correlator demodulator active. The AFC readback value is formatted as a signed 16-bit integer comprised of Bits RV1 to RV16, and is scaled according to the following formula:

$$FREQ\_RB [Hz] = (AFC\_READBACK \times DEMOD\_CLK)/2^{15}$$

In the absence of frequency errors, the FREQ\_RB value is equal to the IF frequency of 200 kHz. Note that, for the AFC readback to yield a valid result, the down-converted input signal must not fall outside the bandwidth of the analogue IF filter. At low-input signal levels, the variation in the readback value can be improved by averaging.

#### RSSI Readback

The RSSI readback operation yields valid results in Rx mode with ASK or FSK signals. The format of the readback word is shown in Figure 25. It is comprised of the RSSI level information (Bits RV1 to RV7), the current filter gain (FG1, FG2), and the current LNA gain (LG1, LG2) setting. The filter and LNA gain are coded in accordance with the definitions in Register 9. With the reception of ASK modulated signals, averaging of the measured RSSI values improves accuracy. The input power can be calculated from the RSSI readback value as outlined in the RSSI/AGC Section.

#### Battery Voltage ADCIN/Temperature Sensor Readback

The battery voltage is measured at Pin VDD4. The readback information is contained in Bits RV1 to RV7. This also applies for the readback of the voltage at the ADCIN pin and the temperature sensor. From the readback information, the battery or ADCIN voltage can be determined using

$$V_{BATTERY} = (Battery\_Voltage\_Readback)/21.1$$

$$V_{ADCIN} = (ADCIN\_Voltage\_Readback)/42.1$$

#### Silicon Revision Readback

The silicon revision readback word is valid without setting any other registers, especially directly after power-up. The silicon revision word is coded with four quartets in BCD format. The product code (PC) is coded with two quartets extending from Bits RV9 to RV16. The revision code (RV) is coded with two quartets extending from Bits RV1 to RV8. The product code should read back as PC = #20h. The current revision code should read as RC = #30h.

#### Filter Calibration Readback

The filter calibration readback word is contained in Bits RV1 to RV8, and is for diagnostic purposes only. Using the automatic filter calibration function, accessible through Register 6, is recommended.

READBACK MODE	READBACK VALUE															
	DB15	DB14	DB13	DB12	DB11	DB10	DB9	DB8	DB7	DB6	DB5	DB4	DB3	DB2	DB1	DB0
AFC READBACK	RV16	RV15	RV14	RV13	RV12	RV11	RV10	RV9	RV8	RV7	RV6	RV5	RV4	RV3	RV2	RV1
RSSI READBACK	X	X	X	X	X	LG2	LG1	FG2	FG1	RV7	RV6	RV5	RV4	RV3	RV2	RV1
BATTERY VOLTAGE/ADCIN/ TEMP. SENSOR READBACK	X	X	X	X	X	X	X	X	X	RV7	RV6	RV5	RV4	RV3	RV2	RV1
SILICON REVISION	RV16	RV15	RV14	RV13	RV12	RV11	RV10	RV9	RV8	RV7	RV6	RV5	RV4	RV3	RV2	RV1
FILTER CAL READBACK	0	0	0	0	0	0	0	0	RV8	RV7	RV6	RV5	RV4	RV3	RV2	RV1

01975-PG-025

Figure 25. Readback Value Table



REGISTER 0—N REGISTER

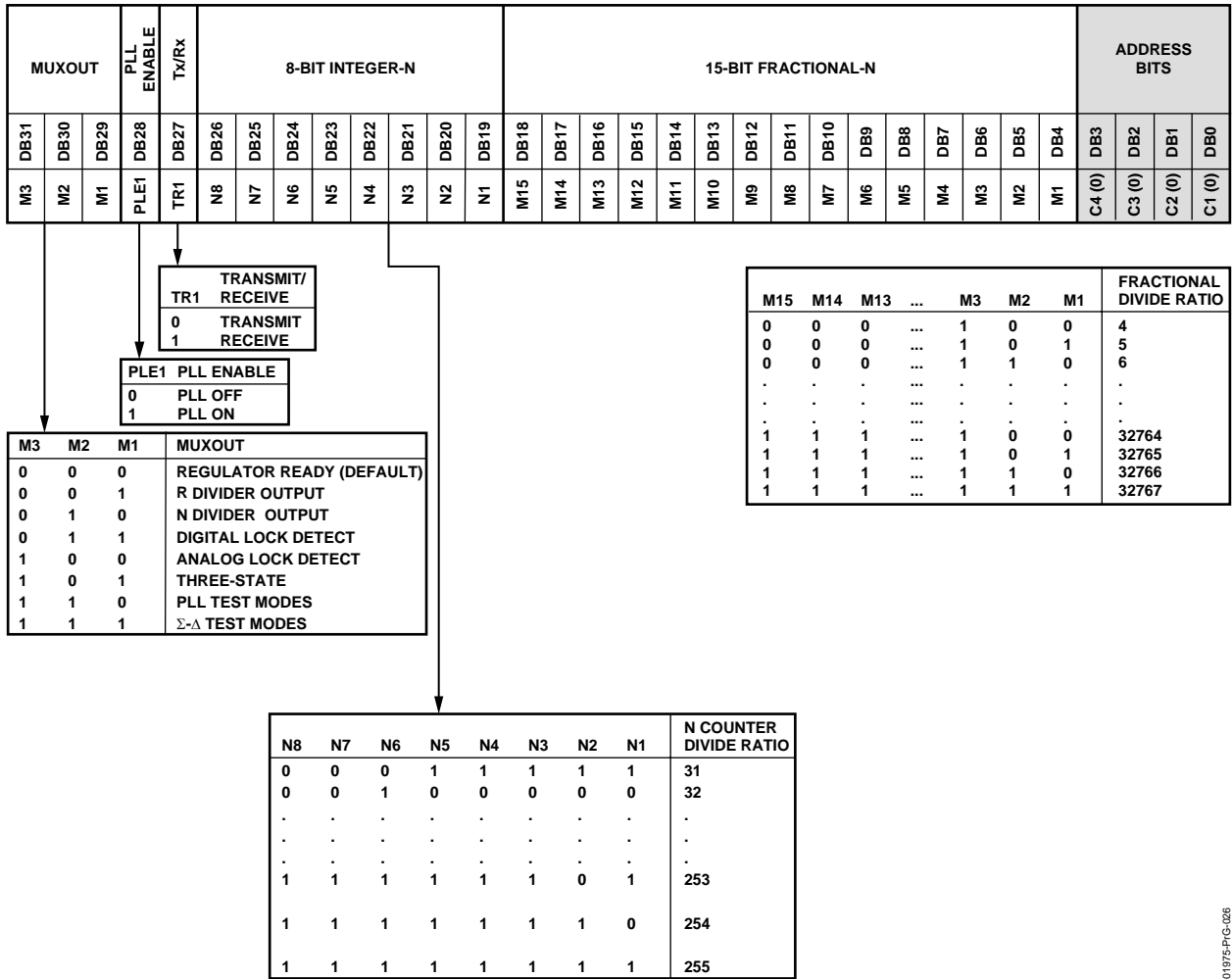


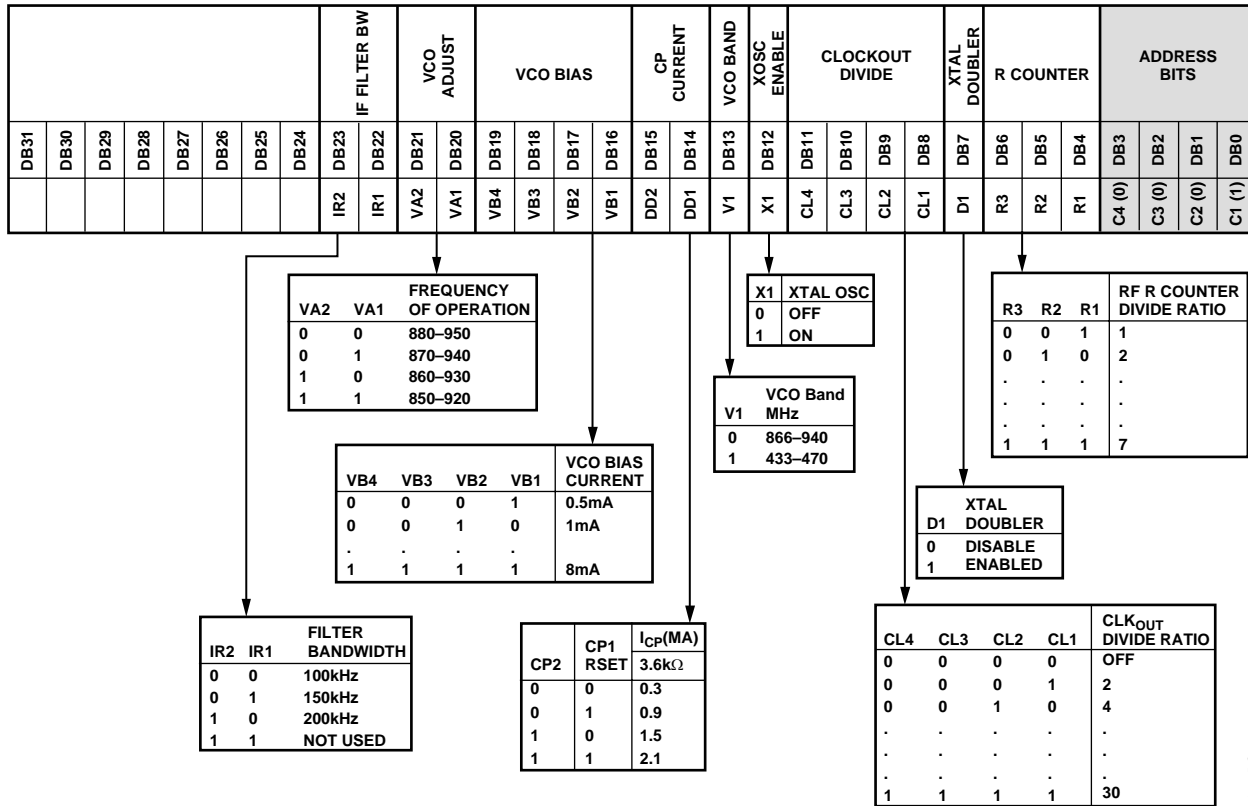
Figure 26.

Notes:

- The Tx/Rx bit (R0\_DB27) configures the part in Tx or Rx mode and also controls the state of the internal Tx/Rx switch.
- $$F_{OUT} = \frac{XTAL}{R} \times (Integer-N + \frac{Fractional-N}{2^{15}})$$

01975-PC-026

REGISTER 1—OSCILLATOR/FILTER REGISTER



01975-PKG-027

Figure 27.

Notes:

1. Set the VCO adjust bits (R1\_DB(20:21)) to 0 for normal operation.

2. 
$$F_{OUT} = \frac{XTAL}{R} \times \left( Integer-N + \frac{Fractional-N}{2^{15}} \right).$$

**REGISTER 2—TRANSMIT MODULATION REGISTER (ASK/OOK MODE)**

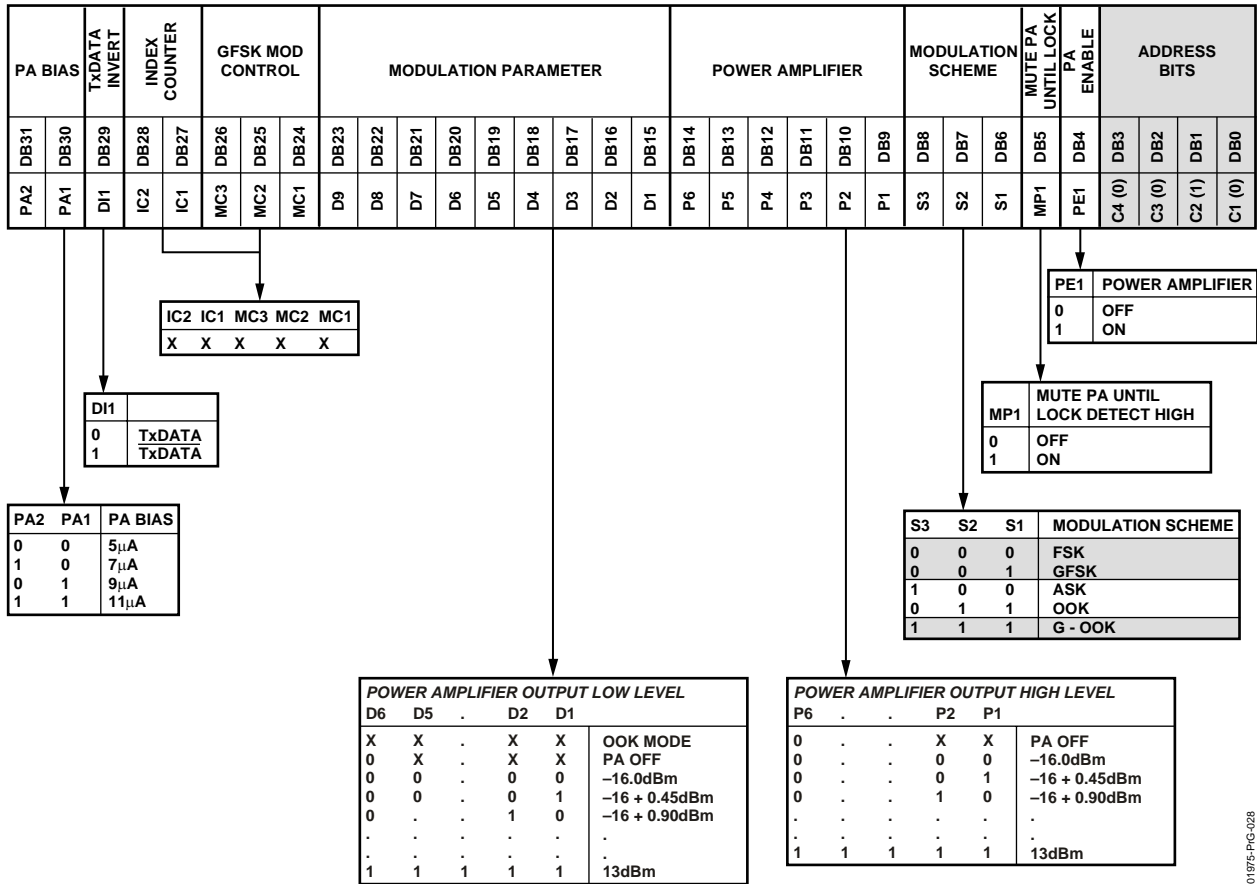


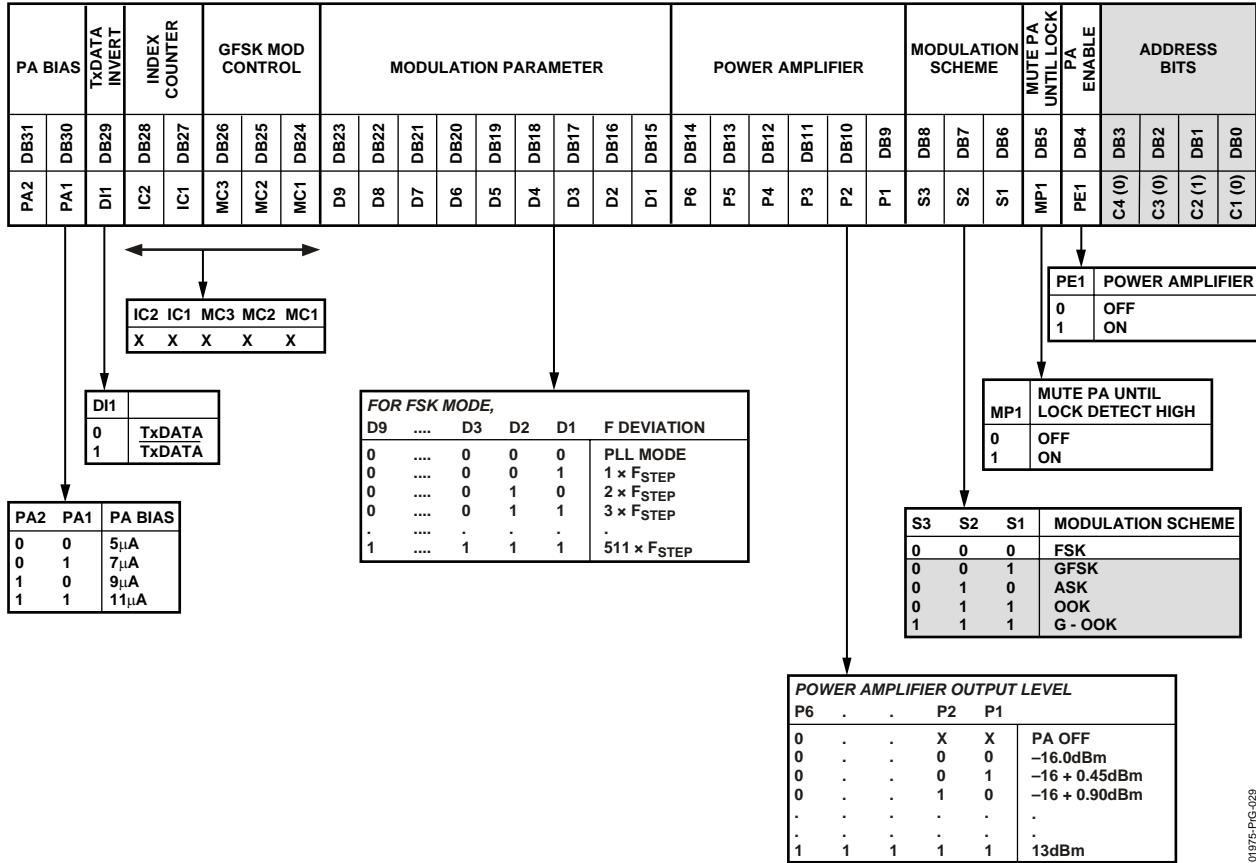
Figure 28.

01975-PKG-028

Note:

1. See the Transmitter section for a description of how the PA bias affects power amplifier level. Default level is 9 µA.

REGISTER 2—TRANSMIT MODULATION REGISTER (FSK MODE)



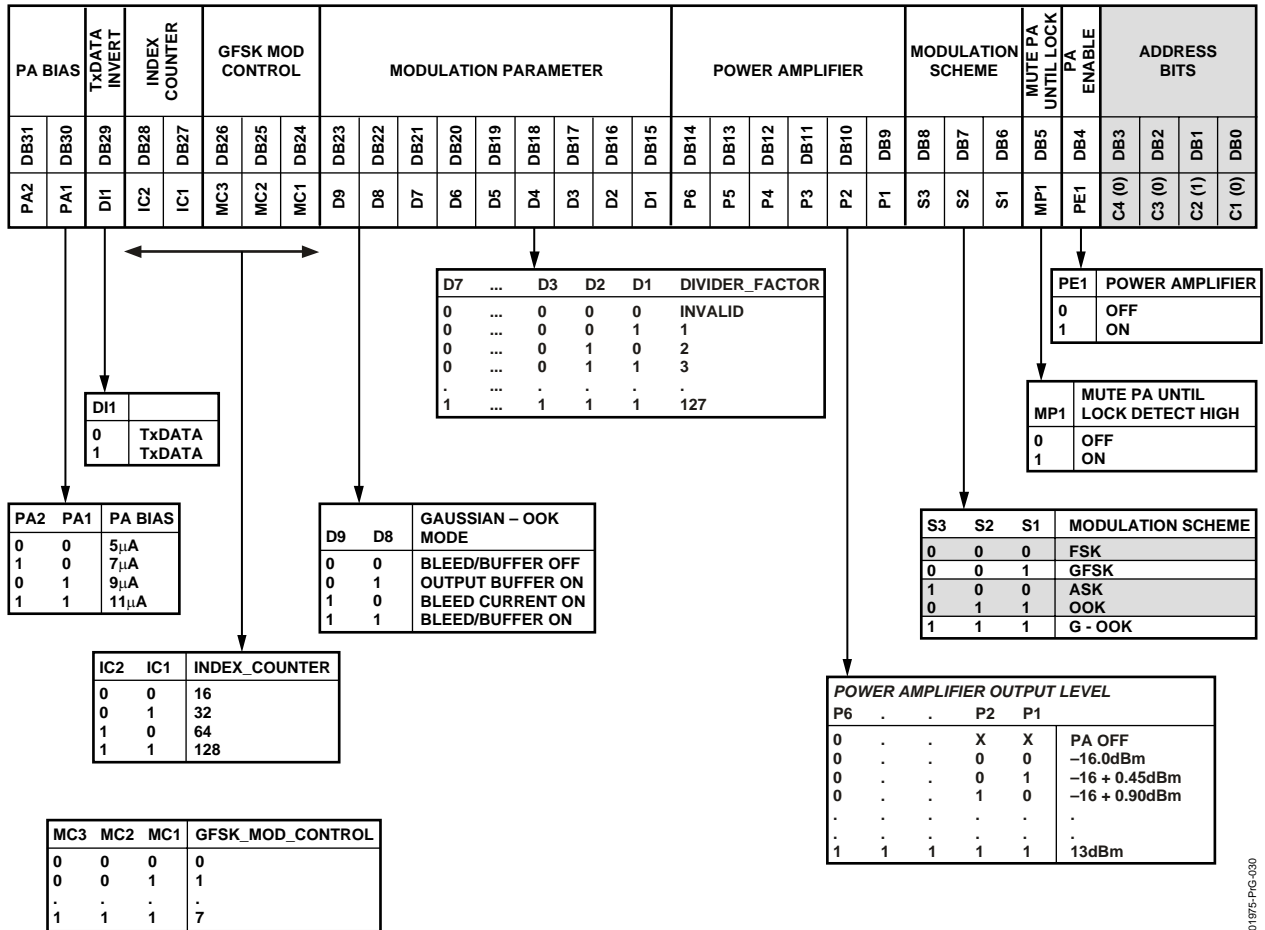
01875-PC-029

Figure 29.

Notes:

1.  $F_{STEP} = PFD/2^{14}$
2. PA Bias default = 9 µA.

REGISTER 2—TRANSMIT MODULATION REGISTER (GFSK/GOOK MODE)



01975-PG-030

Figure 30.

Notes:

1.  $GFSK\_DEVIATION = (2^{GFSK\_MOD\_CONTROL} \times PFD) / 2^{12}$ .
2.  $DR = PFD / (INDEX\_COUNTER \times DIVIDER\_FACTOR)$ .
3. PA Bias default = 9  $\mu$ A.

REGISTER 3—RECEIVER CLOCK REGISTER

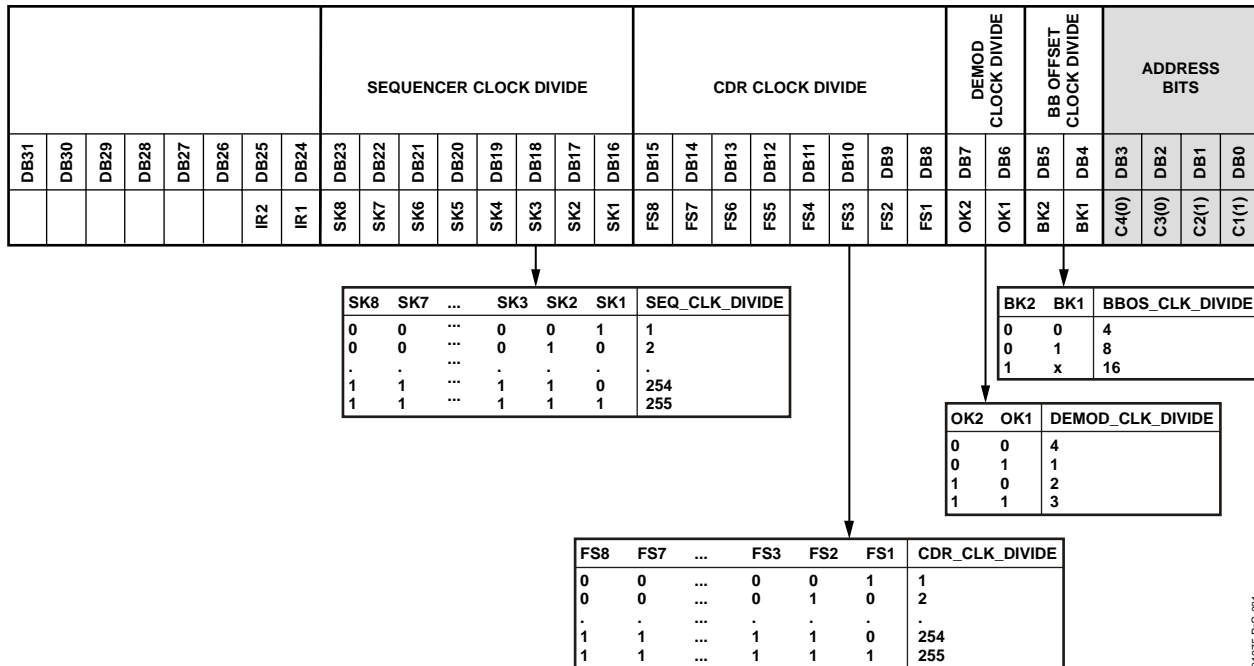


Figure 31.

01975-PrH-001

Notes:

1. Baseband offset clock frequency (BBOS\_CLK) must be greater than 1 MHz and less than 2 MHz, where:

$$BBOS\_CLK = \frac{XTAL}{BBOS\_CLK\_DIVIDE}$$

2. The demodulator clock (DEMOM\_CLK) must be < 12 MHz for FSK and < 6 MHz for ASK, where:

$$DEMOM\_CLK = \frac{XTAL}{DEMOM\_CLK\_DIVIDE}$$

3. Data/clock recovery frequency (CDR\_CLK) should be within 2% of (32 × data rate), where:

$$CDR\_CLK = \frac{DEMOM\_CLK}{CDR\_CLK\_DIVIDE}$$

Note that this might affect your choice of XTAL, depending on the desired data rate.

4. The sequencer clock (SEQ\_CLK) supplies the clock to the digital receive block. It should be close to 100 kHz for FSK and close to 40 kHz for ASK:

$$SEQ\_CLK = \frac{XTAL}{SEQ\_CLK\_DIVIDE}$$

**REGISTER 4—DEMODULATOR SETUP REGISTER**

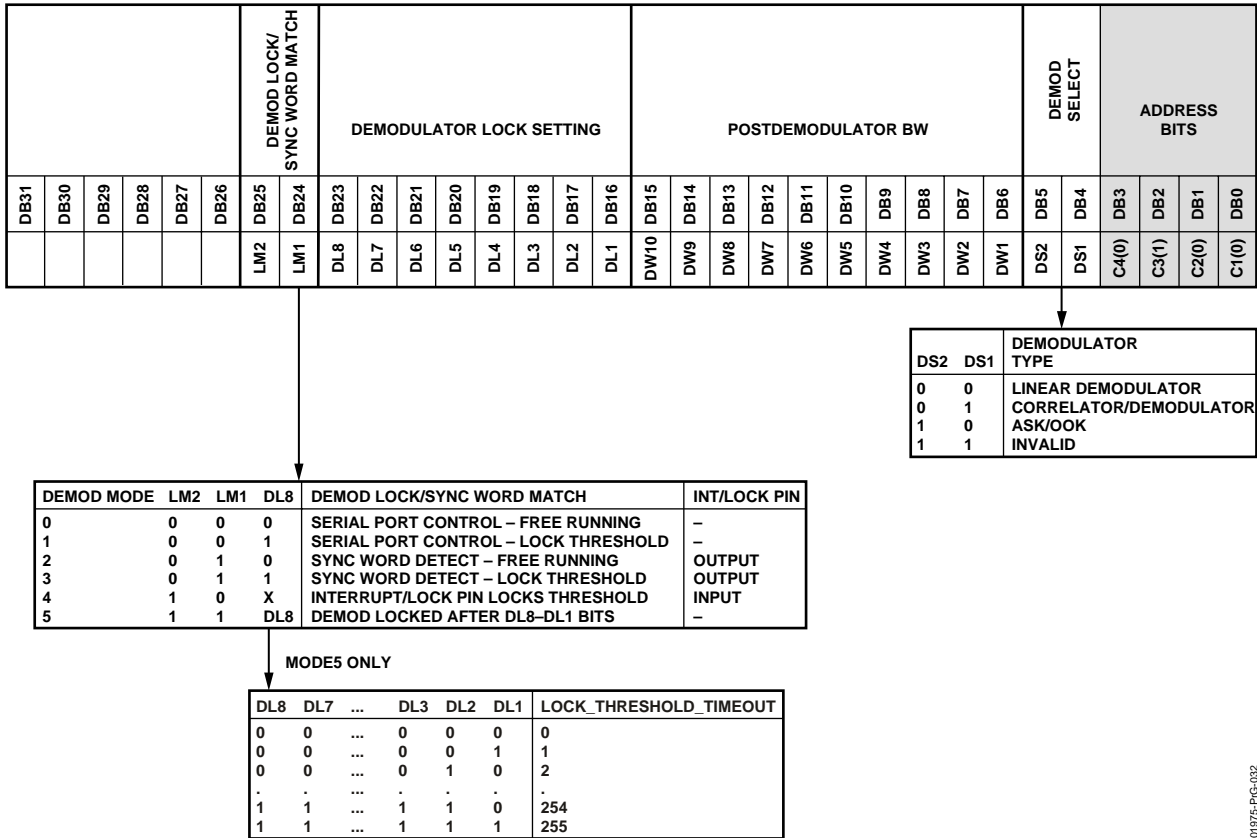


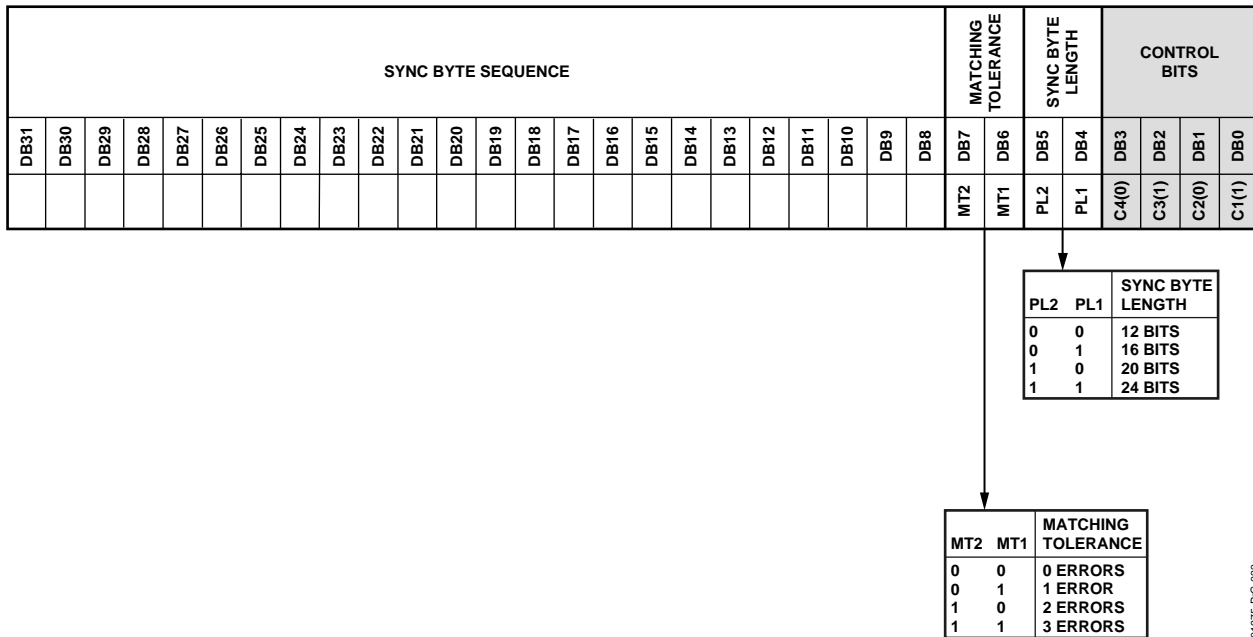
Figure 32.

Notes :

1. The cutoff frequency of the postdemodulator filter should typically be 0.75 times the data rate.
2. Demodulator modes 1, 3, 4, and 5 are modes that can be activated to allow the ADF7020 to demodulate data-encoding schemes that have run-length constraints greater than 7.
3.  $Post\_Demod\_BW = 2^{11} \pi F_{Cutoff}/DEMOD\_CLK$ .
4. For Mode 5, the timeout delay to lock threshold = (LOCK\_THRESHOLD\_SETTING)/SEQ\_CLK, where SEQ\_CLK is defined in the Register 3—Receiver Clock Register section.

01975-PC-002

REGISTER 5—SYNC BYTE REGISTER



01975-PG-033

Figure 33.

- Notes:
1. Sync byte detect is enabled by programming Bits R4\_DB(25:23) to [010] or [011].
  2. This register allows a 28-bit sync byte sequence to be stored internally. If the sync byte detect mode is selected, then the INT/LOCK pin goes high when the sync byte has been detected in Rx mode. Once the sync word detect signal has gone high, it goes low again after nine data bits.
  3. The transmitter must Tx the MSB of the sync byte first and the LSB last to ensure proper alignment in the receiver sync byte detection hardware.



**REGISTER 6—CORRELATOR/DEMODULATOR REGISTER**

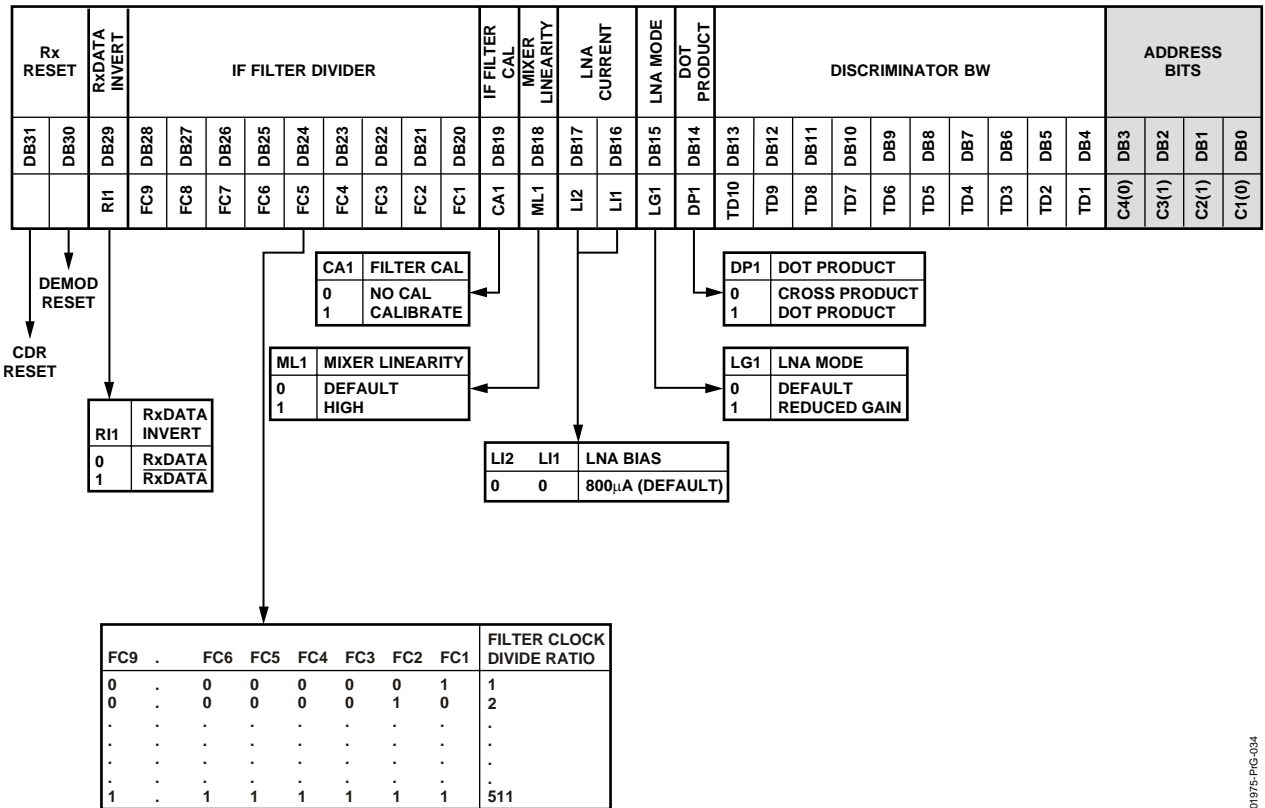


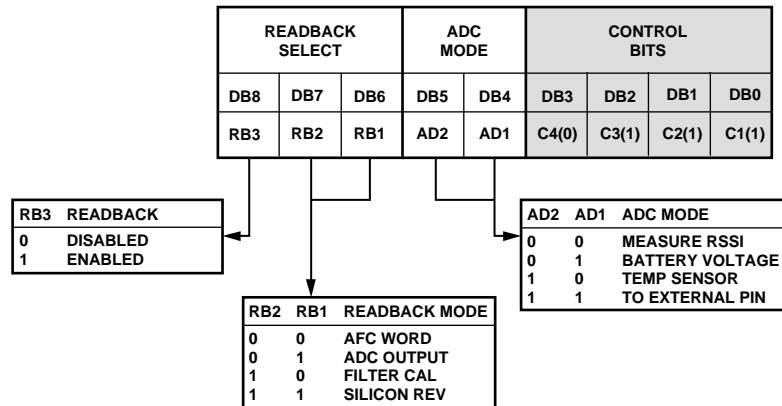
Figure 34.

Notes:

1. See the FSK Correlator/Demodulator section for an example of how to determine register settings.
2. Nonadherence to correlator programming guidelines results in poorer sensitivity.
3. The filter clock is used to calibrate the IF filter. The filter clock divide ratio should be adjusted so that the frequency is 50 kHz. The formula is XTAL/FILTER\_CLOCK\_DIVIDE.
4. The filter should be calibrated only when the crystal oscillator is settled. The filter calibration is initiated every time Bit R6\_DB19 is set high.
5.  $Discriminator\_BW = (DEMODO\_CLK \times K) / (800 \times 10^3)$ . See the FSK Correlator/Demodulator section.

01975-PG-034

REGISTER 7—READBACK SETUP REGISTER



01975-FRG-035

Figure 35.

Notes:

1. Readback of the measured RSSI value is valid only in Rx mode. Readback of the battery voltage, the temperature sensor, and the voltage at the external pin is not available in Rx mode, if the ASK demodulator is active or if AGC is enabled.
2. Readback of the ADC value is valid in Tx mode only if the log amp/RSSI has not been disabled through the power-down bits R8\_DB10. The log amp/RSSI section is active per default upon enabling Tx mode.
3. Readback of the AFC word is valid in Rx mode only if either the linear demodulator or the correlator/demodulator is active.
4. See the Readback Format section for more information.

**REGISTER 8—POWER-DOWN TEST REGISTER**

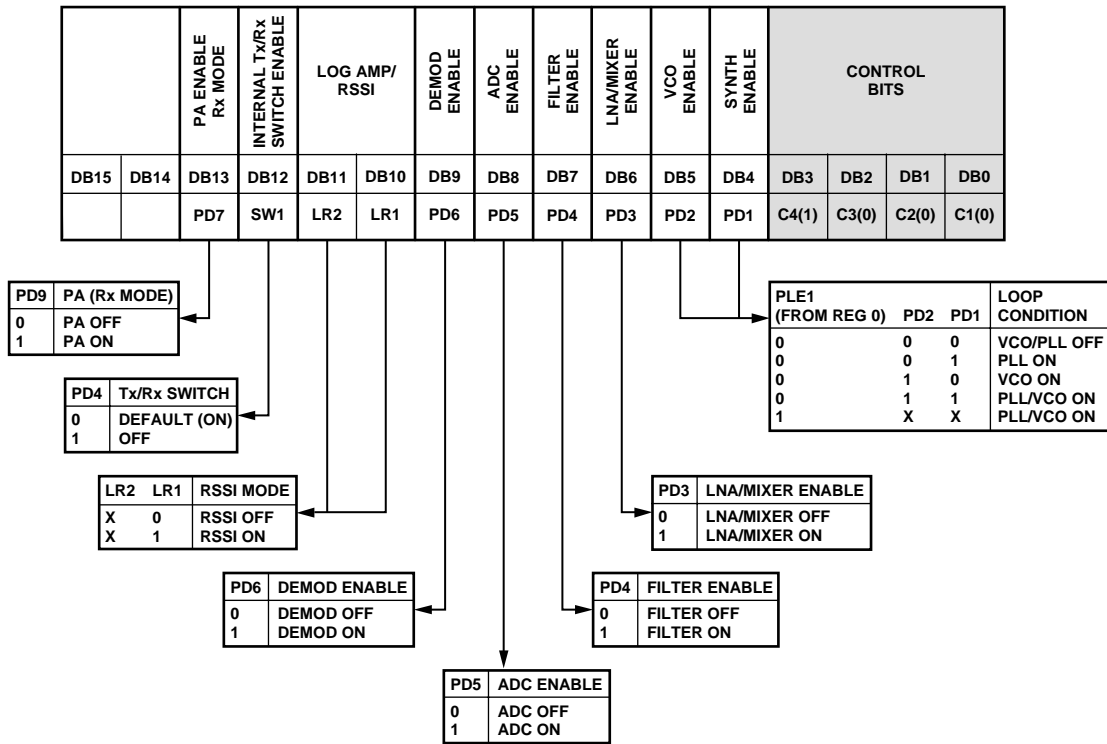


Figure 36.

Notes:

1. For a combined LNA/PA matching network, Bit R8\_DB12 should always be set to 0. This is the power-up default condition.
2. It is not necessary to write to this register under normal operating conditions.

01975-FPG-036

REGISTER 9—AGC REGISTER

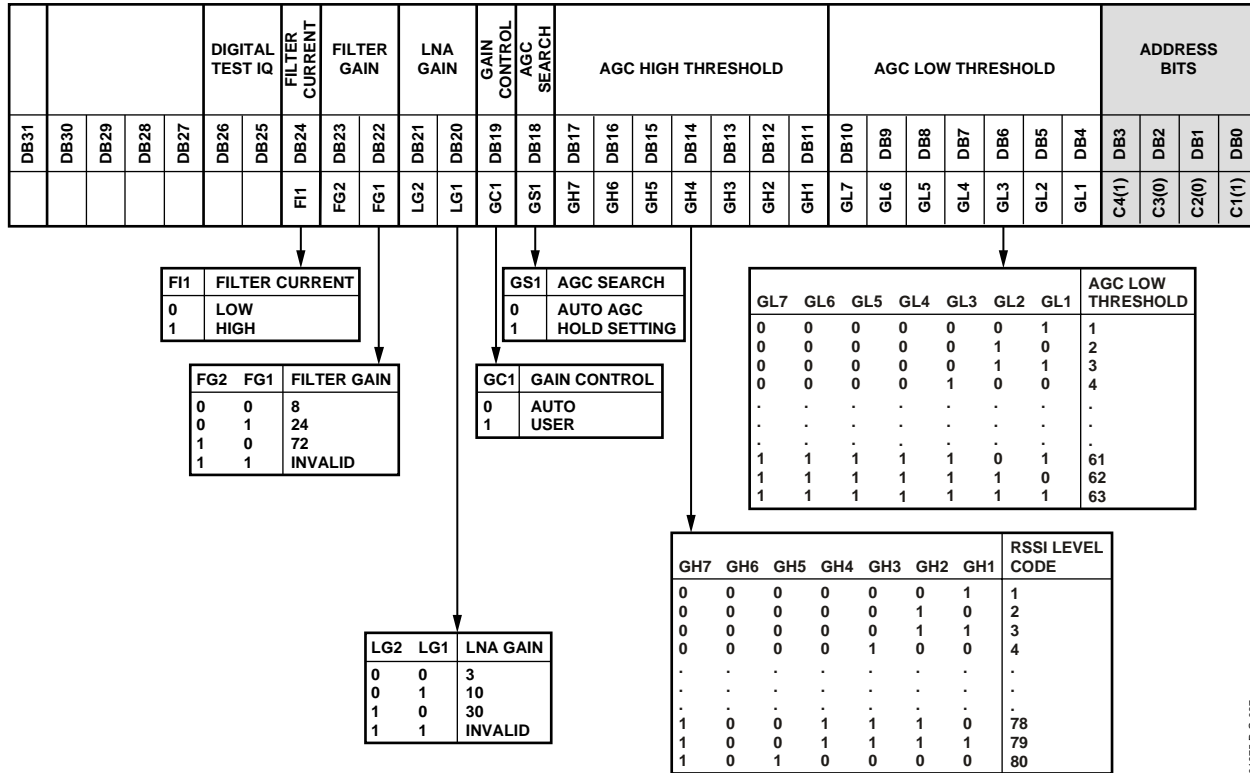


Figure 37.

Notes :

1. Default AGC\_LOW\_THRESHOLD = 27, default AGC\_HIGH\_THRESHOLD = 76. See the RSSI/AGC Section for more details.
2. AGC high and low settings must be more than 30 apart to ensure correct operation.
3. LNA gain of 30 is available only if LNA mode, R6\_DB15, is set to zero.

01975-PFG-037

**REGISTER 10—AGC 2 REGISTER**

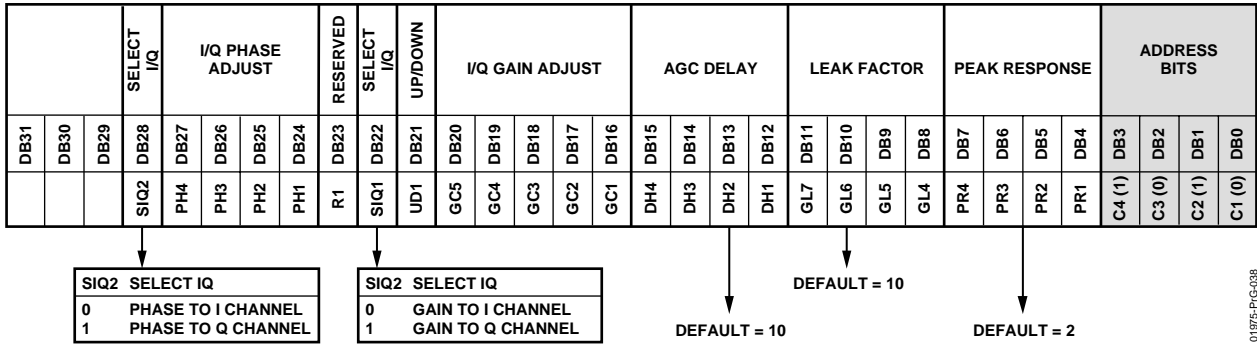


Figure 38.

Note:

- This register is not used under normal operating conditions.

**REGISTER 11—AFC REGISTER**

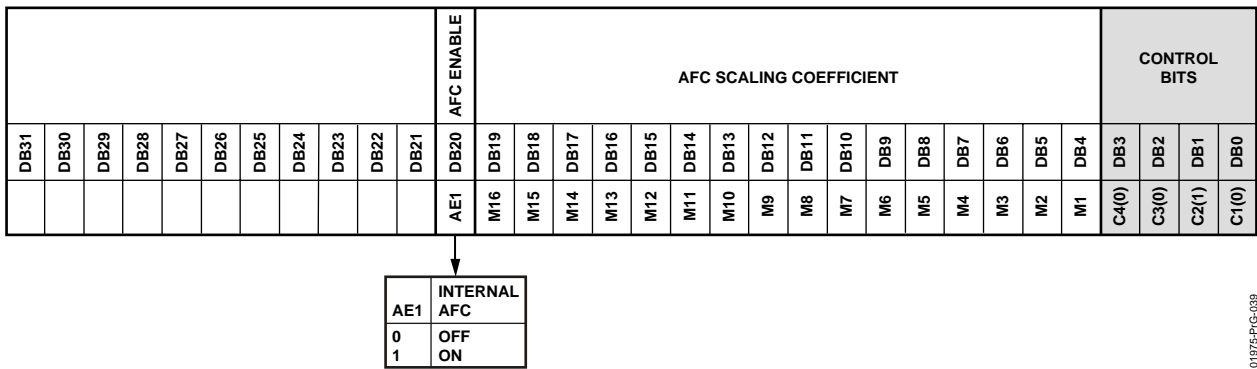


Figure 39.

Notes:

- See the Internal AFC section to program AFC scaling coefficient bits.
- The AFC scaling coefficient bits can be programmed using the following formula:

$$AFC\_Scaling\_Coefficient = Round((500 \times 2^{24})/XTAL)$$

REGISTER 12—TEST REGISTER

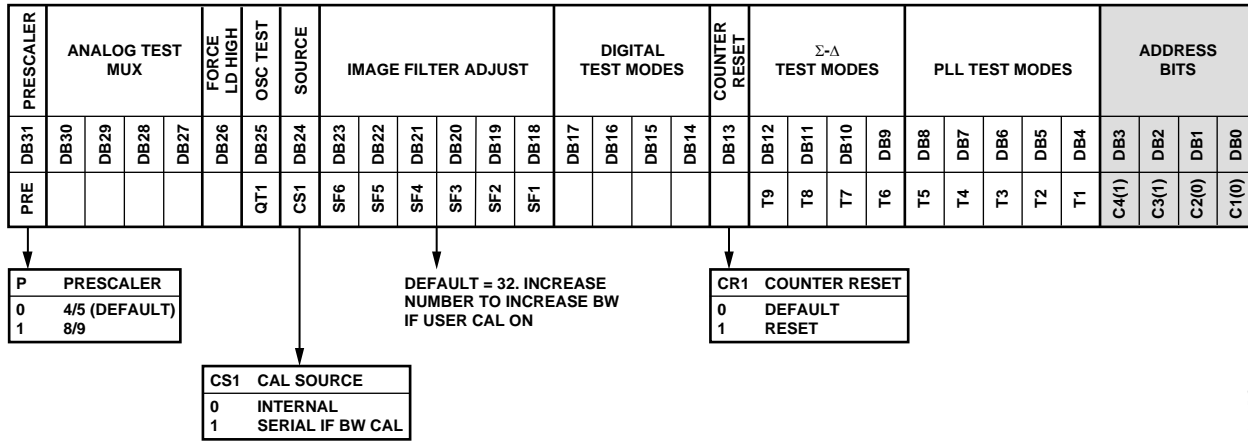


Figure 40.

**Using the Test DAC on the ADF7020 to Implement Analog FM DEMOD and Measuring SNR**

The test DAC allows the output of the postdemodulator filter for both the linear and correlator/demodulators (Figure 16 and Figure 17) to be viewed externally. It takes the 16-bit filter output and converts it to a high frequency, single-bit output using a second-order error feedback Σ-Δ converter. The output can be viewed on the XCLK<sub>OUT</sub> pin. This signal, when IF filtered appropriately, can then be used to

- Monitor the signals at the FSK/ASK postdemodulator filter output. This allows the demodulator output SNR to be measured. Eye diagrams can also be constructed of the received bit stream to measure the received signal quality.
- Provide analog FM demodulation.

While the correlators and filters are clocked by DEMOD\_CLK, CDR\_CLK clocks the test DAC. Note that, although the test DAC functions in a regular user mode, the best performance is achieved when the CDR\_CLK is increased up to or above the frequency of DEMOD\_CLK. The CDR block does not function when this condition exists.

Programming the test register, Register 12, enables the test DAC. Both the linear and correlator/demodulator outputs can be multiplexed into the DAC.

Register 13 allows a fixed offset term to be removed from the signal (to remove the IF component in the ddt case). It also has a signal gain term to allow the usage of the maximum dynamic range of the DAC.

**Setting Up the Test DAC**

- Digital test modes = 7: enables the test DAC, with no offset removal.
- Digital test modes = 10: enables the test DAC, with offset removal.

The output of the active demodulator drives the DAC, that is, if the FSK correlator/demodulator is selected, the correlator filter output drives the DAC.

**REGISTER 13—OFFSET REMOVAL AND SIGNAL GAIN REGISTER**

TEST DAC GAIN						TEST DAC OFFSET REMOVAL						PULSE EXTENSION				KI				KP				CONTROL BITS											
DB31	DB30	DB29	DB28	DB27	DB26	DB25	DB24	DB23	DB22	DB21	DB20	DB19	DB18	DB17	DB16	DB15	DB14	DB13	DB12	DB11	DB10	DB9	DB8	DB7	DB6	DB5	DB4	DB3	DB2	DB1	DB0				
																PE4	PE3	PE2	PE1													C4(1)	C3(1)	C2(0)	C1(1)

PE4	PE3	PE2	PE1	PULSE EXTENSION
0	0	0	0	NORMAL PULSE WIDTH
0	0	0	1	2 x PULSE WIDTH
0	0	1	0	3 x PULSE WIDTH
.	.	.	.	.
.	.	.	.	.
1	1	1	1	16 x PULSE WIDTH

01975-PRG-044

Figure 41.

Note:

1. Because the linear demodulator's output is proportional to frequency, it usually consists of an offset combined with a relatively low signal. The offset can be removed, up to a maximum of 1.0 and gained to use the full dynamic range of the DAC:

$$DAC\_input = (2^{\wedge} Test\_DAC\_Gain) \times (Signal - Test\_DAC\_Offset\_Removal/4096)$$

