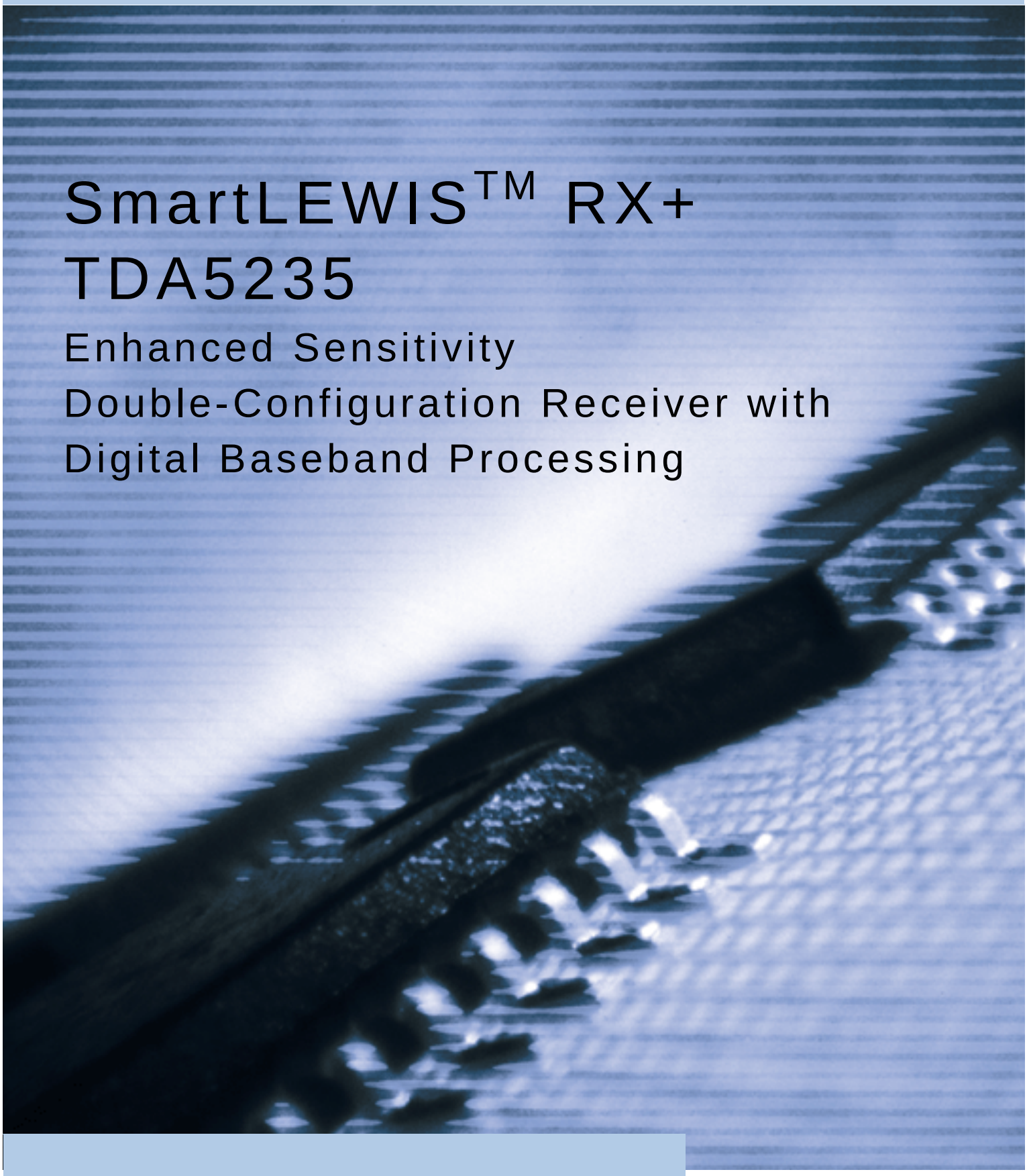

SmartLEWIS™ RX+ TDA5235

Enhanced Sensitivity Double-Configuration Receiver with Digital Baseband Processing

Wireless Control

Never stop thinking.

Edition February 19, 2010

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Enhanced Sensitivity Double-Configuration Receiver with Digital Baseband Processing

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TDA5235

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Page	Subjects (major changes since last revision)
Page 25	Update of Figure 9
Page 27	Update of Figure 10
Page 29	AFC limitation added
Page 31	AGC setting proposal added
Page 32	New Section 2.4.6.5 ADC added
Page 34	Additional information on RSSIPRX register inserted
Page 39	Signal and Noise Detector Procedure adapted
Page 43	x_CDRRI register recommendation changed
Page 47, 50, 54	Data Slicer Modes adapted; limitation added
Page 67	Update of Figure 41
Page 68	Update of Figure 42
Page 76	Additional hint on clock and data recovery algorithm of the user software inserted
Page 82	PLDLEN limitation added
Page 84	Limitation for ISx readout and Burst-read function added
Page 86	Limitation for Burst-read function added
Page 105	Description of “Parallel Wake-up Search” adapted
Page 123	Additional hints added
Page 125	Adaption of Section 4.1
Page 128	New item C7 added
Page 136 f	Comments added for items I6, I7, I8, I9, J11, J12
Page 136	Item J1 updated
Page 139 ff	General test conditions noted for parameters K, L and M
Page 145	BOM components C7, C8, L1, R2 and R3 updated

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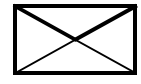

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1 Product Description

1.1 Overview

The IC is a low power ASK/FSK Receiver for the frequency bands 300-320, 425-450, 863-870 and 902-928 MHz. Bi-phase modulation schemes, like Manchester, bi-phase mark, bi-phase space and differential Manchester are supported.

The chip offers best-in-class sensitivity performance at a very high level of integration and needs only a few external components.

The device is qualified to automotive quality standards and operates between -40 and +105°C at supply voltage ranges of 3.0-3.6 Volts or 4.5-5.5 Volts.

The receiver is realized as a double down conversion super-heterodyne/low-IF architecture each with image rejection supplemented by digital signal processing in the baseband. A fully integrated Sigma-Delta Fractional-N PLL Synthesizer allows for high-resolution frequency generation and uses a crystal oscillator as the reference. The on-chip temperature sensor may be utilized for temperature drift compensation via the crystal oscillator.

The digital baseband processing unit together with the high performance down converter is the key element for the exceptional sensitivity performance of the device which take it close to the theoretical top-performance limits. It comprises signal and noise detectors, matched data filter, clock and data recovery, data slicer and a format decoder. It demodulates the received ASK or FSK data stream independently and recovers the data clock out of the received data stream with very fast synchronization times which can then be either accessed via separate pins or used for further processing like frame synchronization and intermediate storage in the on-chip FIFO. The RSSI output signal is converted to the digital domain with an ADC. All these signals are accessible via the 4-wire SPI interface bus. Up to 2 pre-configured telegram formats can be stored into the device offering independent pre-processing of the received data to an extent not available till now. The down converter can be also configured in single-conversion mode at moderately reduced selectivity performance but at the advantage of omitting the IF ceramic filter.

1.2 Features

- Enhanced sensitivity receiver
- Multi-band (300-320, 425-450, 863-870 and 902-928 MHz)
- One crystal frequency for all supported frequency bands
- 21-bit Sigma-Delta Fractional-N PLL synthesizer with high resolution of 10.5 Hz
- Up to 2 parallel parameter sets for autonomous scanning and receiving from different sources reduces significantly host processor power consumption and system standby power consumption
- One frequency channels per parameter set is supported with 10.5 Hz resolution
- Autonomous receive mode leads to reduced noise of host processor and improved system performance
- Ultrafast Wake-up on RSSI
- Fast synchronization on incoming data stream typically within first 4 bits of a telegram
- Selectable IF filter bandwidth and optional external filters possible
- Double down conversion image reject mixer
- ASK and FSK capability
- Automatic Frequency Control (AFC) for carrier frequency offset compensation
- Supports bi-phase line codes like Manchester, bi-phase mark/space and differential Manchester
- NRZ data pre-processing capability
- Digital base band receiver with clock synch, frame synch, format decoding and FIFO
- Separate outputs for recovered data and clock
- RSSI peak detectors
- Wake-up generator and polling timer unit
- Message ID scanning
- Unique 32-bit serial number
- On-chip temperature sensor
- Integrated timer usable for external watch unit
- Integrated 4-wire SPI interface bus
- Supply voltage range 3.0 Volts to 3.6 Volts or 4.5 Volts to 5.5 Volts
- Operating temperature range -40 to +105°C
- ESD protection +/- 2 kV on all pins
- Package PG-TSSOP-28

1.3 Applications

- Remote keyless entry systems
- Remote start applications
- Tire pressure monitoring
- Short range radio data transmission
- Remote control units
- Cordless alarm systems
- Remote metering

2 Functional Description

2.1 Pin Configuration

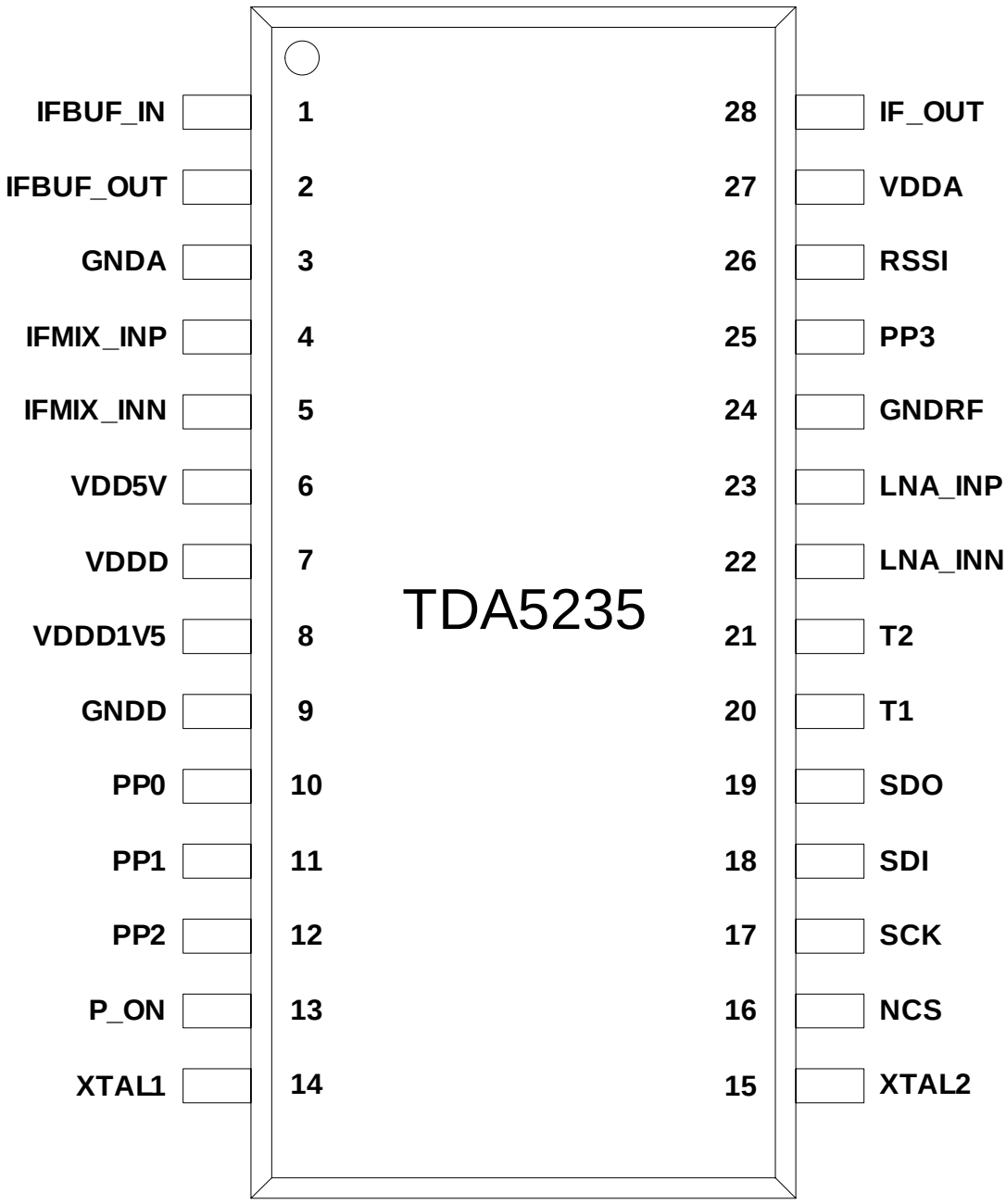
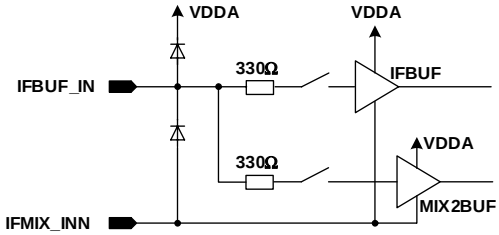
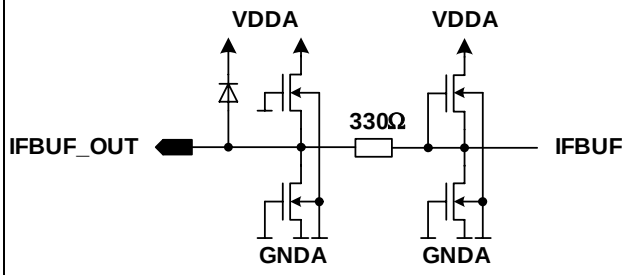
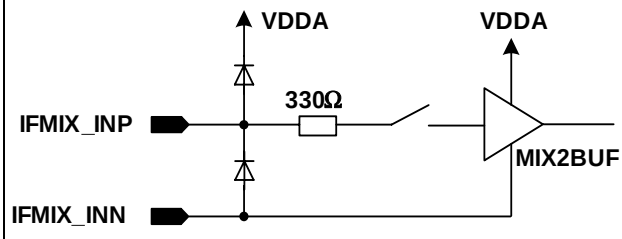
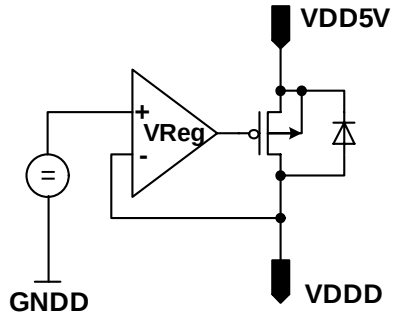
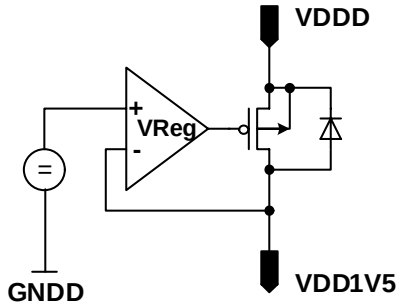
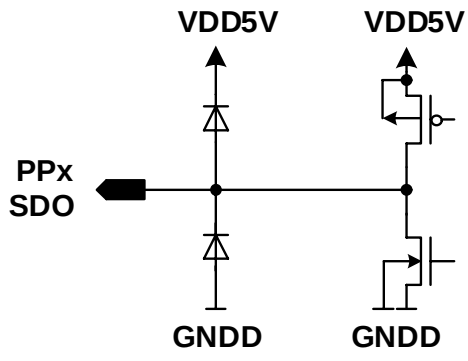


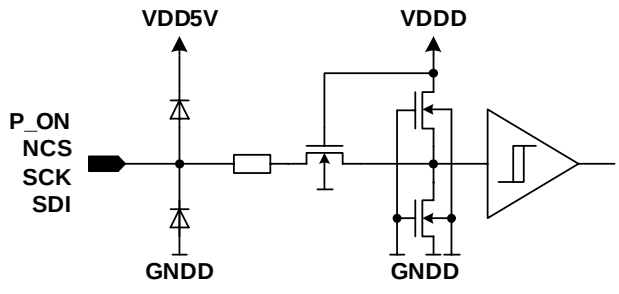
Figure 1 Pin-out

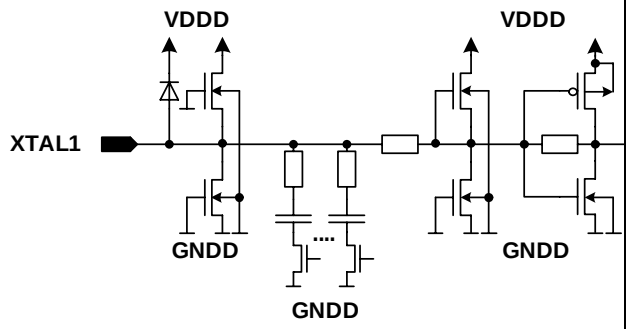
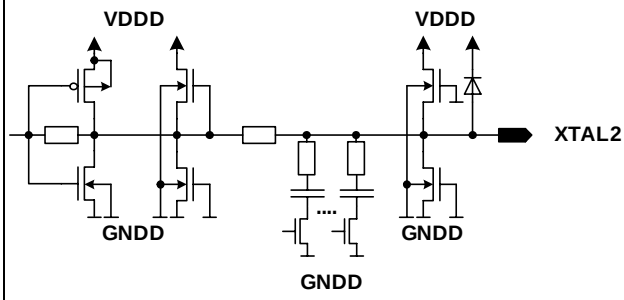
2.2 Pin Definition and Pin Functionality

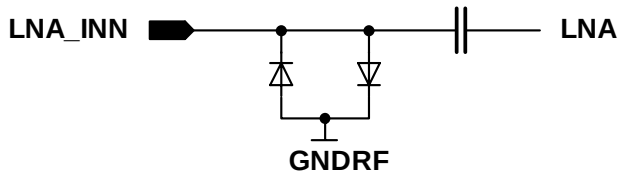
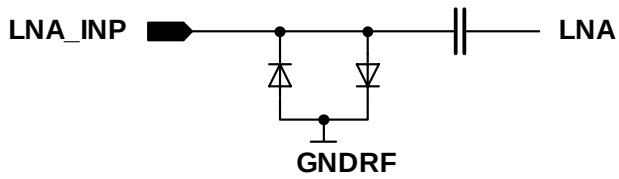
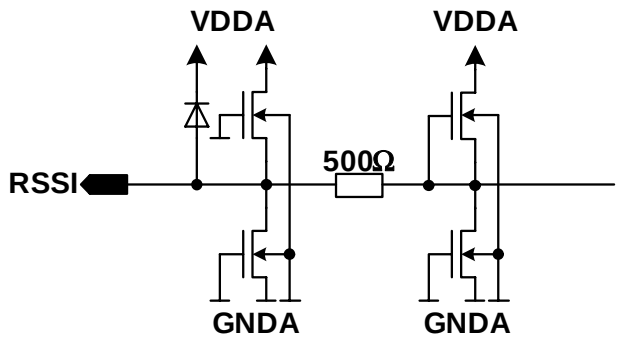
Table 1 Pin Definition and Function

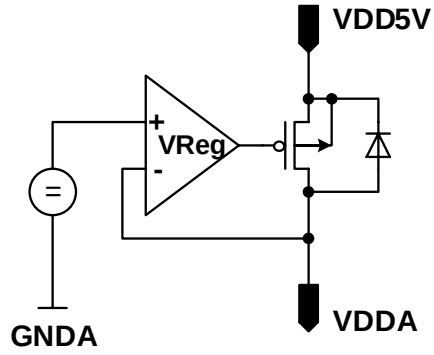
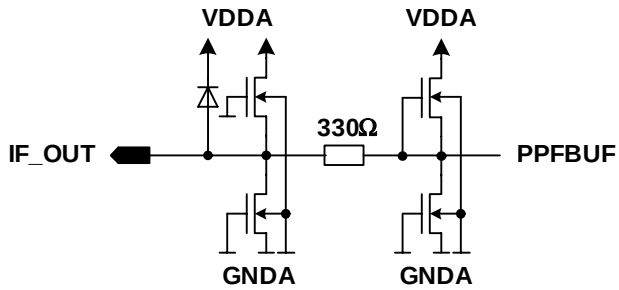
Pin No.	Pad name	Equivalent I/O Schematic	Function
1	IFBUF_IN		Analog input IF Buffer input Note: Input is biased at VDDA/2
2	IFBUF_OUT		Analog output IF Buffer output
3	GNDA		Analog ground
4	IFMIX_INP		Analog input + IF mixer input Note: Input is biased at VDDA/2
5	IFMIX_INN	see schematic of Pin 1 and 4	Analog input. - IF mixer input
6	VDD5V		Analog input 5 Volt supply input

Pin No.	Pad name	Equivalent I/O Schematic	Function
7	VDDD		Analog input digital supply input
8	VDDD1V5		Analog output 1.5 Volt voltage regulator
9	GNDD		Digital ground
10	PP0		Digital output CLK_OUT, RX_RUN, NINT, LOW, HIGH, DATA, DATA_MATCHFIL, CH_DATA, CH_STR, RXD and RXSTR are programmable via a SFR (Special Function Register), default = CLK_OUT

Pin No.	Pad name	Equivalent I/O Schematic	Function
11	PP1	see schematic of Pin 10	Digital output CLK_OUT, RX_RUN, NINT, LOW, HIGH, DATA, DATA_MATCHFIL, CH_DATA, CH_STR, RXD and RXSTR are programmable via a SFR, default = DATA
12	PP2	see schematic of Pin 10	Digital output CLK_OUT, RX_RUN, NINT, LOW, HIGH, DATA, DATA_MATCHFIL, CH_DATA, CH_STR, RXD and RXSTR are programmable via a SFR, default = NINT
13	P_ON		Digital input power-on reset

Pin No.	Pad name	Equivalent I/O Schematic	Function
14	XTAL1		Analog input crystal oscillator input
15	XTAL2		Analog output crystal oscillator output
16	NCS	see schematic of Pin 13	Digital input SPI enable
17	SCK	see schematic of Pin 13	Digital input SPI clock
18	SDI	see schematic of Pin 13	Digital input SPI data in
19	SDO	see schematic of Pin 10	Digital output SPI data out
20	T1		Digital input, connect to Digital Ground
21	T2		Digital input, connect to Digital Ground

Pin No.	Pad name	Equivalent I/O Schematic	Function
22	LNA_INN		Analog input - RF input
23	LNA_INP		Analog input + RF input
24	GNDRF		RF analog ground
25	PP3	see schematic of Pin 10	Digital output RX_RUN, NINT, LOW, HIGH, DATA, DATA_MATCHFIL, CH_DATA, CH_STR, RXD and RXSTR are programmable via a SFR, default = RX_RUN
26	RSSI		Analog output analog RSSI output/ analog test pin ANA_TST

Pin No.	Pad name	Equivalent I/O Schematic	Function
27	VDDA		Analog input Analog supply
28	IF_OUT		Analog output IF output

2.3 Functional Block Diagram

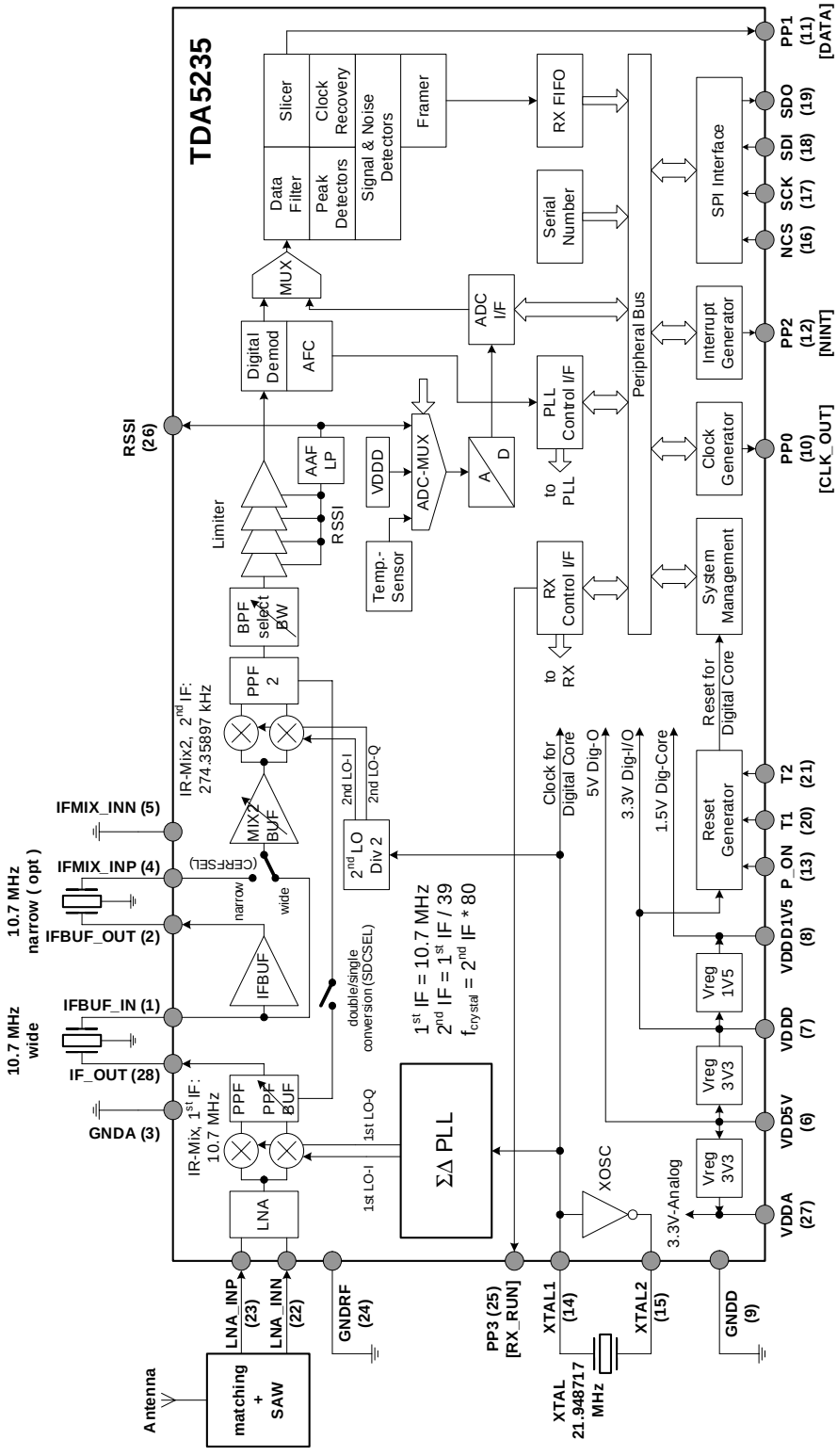


Figure 2 TDA5235 Block Diagram[1)]

1) The function on each PPx port pin can be programmed via SFR (see also **Table 1**). Default values are given in squared brackets in **Figure 2**.

2.4 Functional Block Description

2.4.1 Architecture Overview

A fully integrated Sigma-Delta Fractional-N PLL Synthesizer covers the frequency bands 300-320 MHz, 425-450 MHz, 863-870 MHz, 902-928 MHz with a high frequency resolution, using only one VCO running at around 3.6 GHz.
For Multi-Configuration applications requiring different RF channels a very good channel separation is essential. To achieve the necessary high sensitivity and selectivity a double down conversion super-heterodyne architecture is used. The first IF frequency is located around 10.7 MHz and the second IF frequency around 274 kHz. For both IF frequencies an adjustment-free image frequency rejection feature is realized. In the second IF domain the filtering is done with an on-chip third order bandpass polyphase filter. A multi-stage bandpass limiter completes the RF/IF path of the receiver. For Single-Channel applications with relaxed requirements to selectivity, a single down conversion low-IF scheme can be selected.

For Multi-Configuration systems requiring different RF channels where even higher channel separation is required, up to two (switchable) external ceramic (CER) filters can be used to improve the selectivity.
An RSSI generator delivers a DC signal proportional to the applied input power and is also used as an ASK demodulator. Via an anti-aliasing filter this signal feeds an ADC with 10 bits resolution.
The harmonic suppressed limiter output signal feeds a digital FSK demodulator. This block demodulates the FSK data and delivers an AFC signal which controls the divider factor of the PLL synthesizer.
A digital receiver, which comprises RSSI peak detectors, a matched data filter, a clock and data recovery, a data slicer, a frame synchronization and a data FIFO, decodes the received ASK or FSK data stream. The recovered data and clock signals are accessible via 2 separate pins. The FIFO data buffer is accessible via the SPI bus interface.
The crystal oscillator serves as the reference frequency for the PLL phase detector, the clock signal of the Sigma-Delta modulator and divided by two as the 2^{nd} local oscillator signal. To accelerate the start up time of the crystal oscillator two modes are selectable: a Low Power Mode (with lower precision) and a High Precision Mode.

2.4.2 Block Overview

The TDA5235 is separated into the following main blocks:

- **RF / IF Receiver**
- **Crystal Oscillator and Clock Divider**
- **Sigma-Delta Fractional-N PLL Synthesizer**
- **ASK / FSK Demodulator incl. AFC, AGC and ADC**
- **RSSI Peak Detector**
- **Digital Baseband Receiver**
- **Power Supply Circuitry**
- **System Interface**
- **System Management Unit**

2.4.3 RF/IF Receiver

The receiver path uses a double down conversion super-heterodyne/low-IF architecture, where the first IF frequency is located around 10.7 MHz and the second IF frequency around 274 kHz. For the first IF frequency an adjustment-free image frequency rejection is realized by means of two low-side injected I/Q-mixers followed by a second order passive polyphase filter centered at 10.7 MHz (PPF). The I/Q-oscillator signals for the first down conversion are delivered from the PLL synthesizer. The frequency selection in the first IF domain is done by an external CER filter (optionally by two, decoupled by a buffer amplifier). For moderate or low cost applications, this ceramic filter can be substituted by a simple LC Pi-filter or completely by-passed using the receiver as a single down conversion low-IF scheme with 274 kHz IF frequency. The down conversion to the second IF frequency is done by means of two high-side injected I/Q-mixers together with an on-chip third order bandpass polyphase filter (PPF2 + BPF). The I/Q-oscillator signals for the second down conversion are directly derived by division of two from the crystal oscillator frequency. The bandwidth of the bandpass filter (BPF) can be selected from 50 kHz to 300 kHz in 5 steps. For a frequency offset of -150 kHz to -120 kHz, the AFC (Automatic Frequency Control) function is mandatory. Activated AFC option might require a longer preamble sequence in the receive data stream.

The receiver enable signal (RX_RUN) can be offered at each of the port pins to control external components. Whenever the receiver is active, the RX_RUN output signal is active. Active high or active low is configurable via PPCFG2 register.

The frequency relations are calculated with the following formulas:

$$f_{IF1} = 10.7MHz$$

$$f_{IF2} = \frac{f_{IF1}}{39}$$

$$f_{crystal} = f_{IF2} \times 80$$

$$f_{LO2} = \frac{f_{crystal}}{2}$$

$$f_{LO1} = f_{crystal} \times NF_{divider}$$

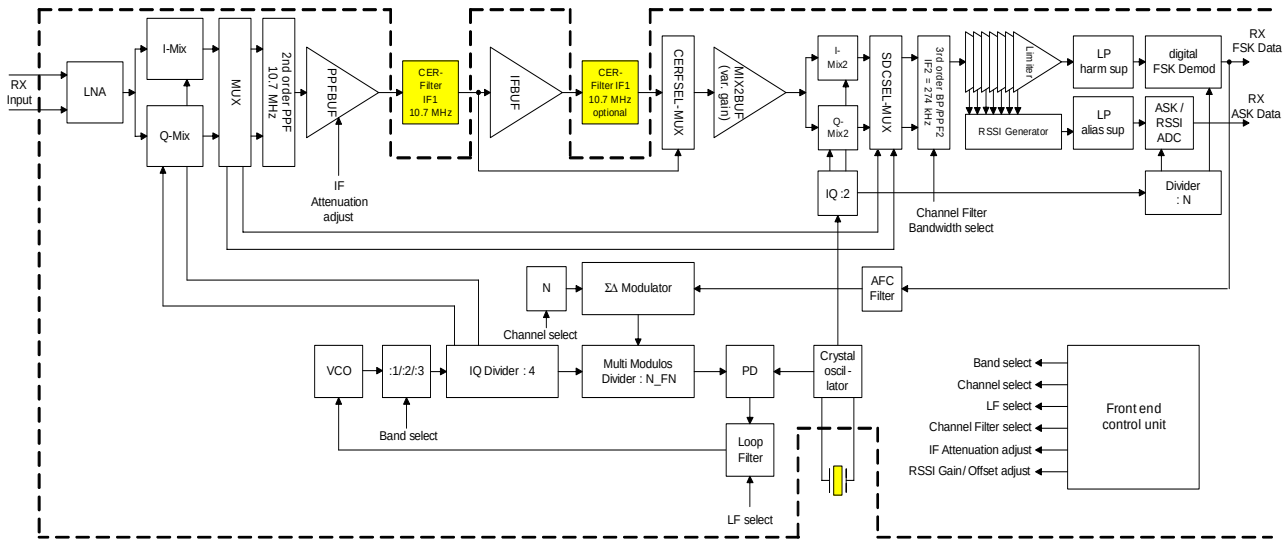


Figure 3 Block Diagram RF Section

The front end of the receiver comprises an LNA, an image reject mixer and a digitally gain controlled buffer amplifier. This buffer amplifier allows the production spread of the on-chip signal strip, of external matching circuitry and RF SAW and ceramic IF filters to be trimmed. The second image reject mixer down converts the first IF to the second IF.

The bandpass filter follows the subsequent formula:

$$f_{center} = \sqrt{f_{corner,\,low} \times f_{corner,\,high}}$$

Therefore asymmetric corner frequencies can be observed. The use of AFC results in more symmetry.

A multi-stage bandpass limiter at a center frequency of 274 kHz completes the receiver chain. The -3dB corner frequencies of the bandpass limiter are typically at 75 kHz and at 520 kHz.

An RSSI generator delivers a DC signal proportional to the applied input power and is also used as an ASK demodulator. Via a programmable anti-aliasing filter this signal is converted to the digital domain by means of a 10-bit ADC.

The limiter output signal is connected to a digital FSK demodulator.

The immunity against strong interference frequencies (so called blockers) is determined by the available filter bandwidth, the filter order and the 3rd order intercept point of the front end stages. For Single-Channel applications with moderate requirements to the selectivity the performance of the on-chip 3rd order bandpass polyphase filter might be sufficient. In this case no external filters are necessary and a single down conversion architecture can be used, which converts the input signal frequency directly to the 2nd IF frequency of 274 kHz.

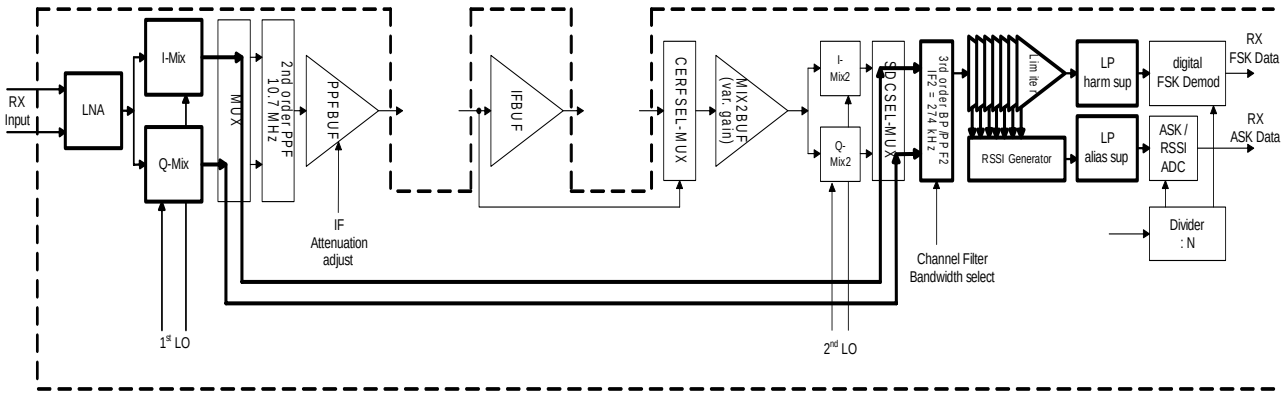


Figure 4 Single Down Conversion (SDC, no external filters required)

For Multi-Configuration applications requiring different RF channels or systems which demand higher selectivity the double down conversion scheme together with one or two external CER filters can be selected. The order of such ceramic filters is in a range of 3, so the selectivity is further improved and a better channel separation is guaranteed.

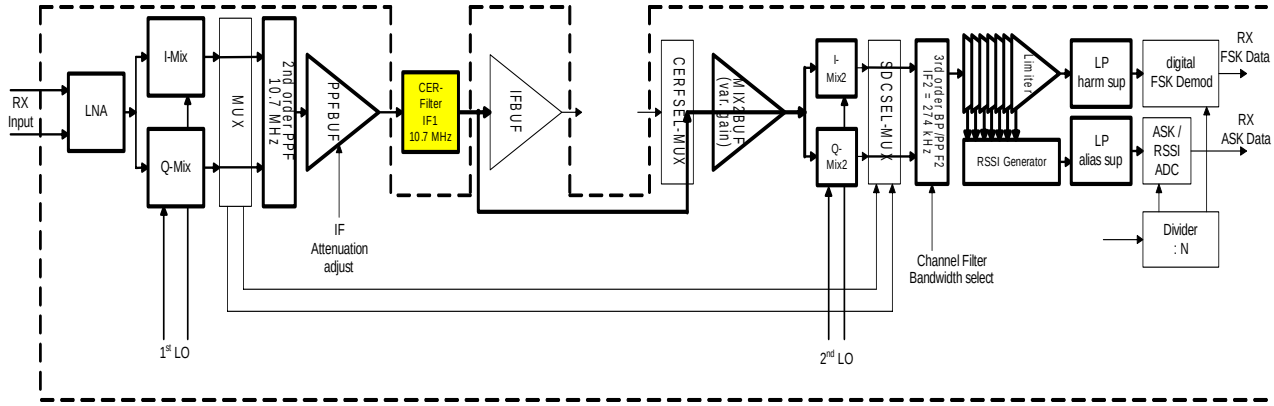


Figure 5 Double Down Conversion (DDC) with one external filter

For applications which demand very high selectivity and/or channel separation even two CER filters may be used. Also in applications where one configuration/channel requires a wider bandwidth than the other (e.g. TPMS and RKE) the second filter can be by-passed.

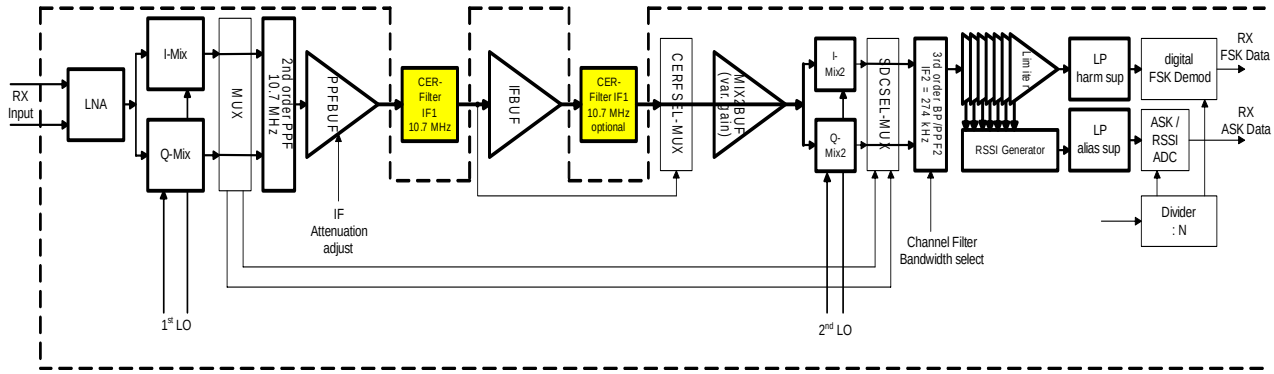


Figure 6 Double Down Conversion (DDC) with two external filters

2.4.4 Crystal Oscillator and Clock Divider

The crystal oscillator is a Pierce type oscillator which operates together with the crystal in parallel resonance mode. An automatic amplitude regulation circuitry allows the oscillator to operate with minimum current consumption. In SLEEP Mode, where the current consumption should be as low as possible, the load capacitor must be small and the frequency is slightly detuned, therefore all internal trim capacitors are disconnected. The internal capacitors are controlled by the crystal oscillator calibration registers XTALCALx. With a binary weighted capacitor array the necessary load capacitor can be selected.

Whenever a XTALCALx register value is updated, the selected trim capacitors are automatically connected to the crystal so that the frequency is precise at the desired value. The SFR control bit XTALHPMS can be used to activate the High Precision Mode also during SLEEP Mode.

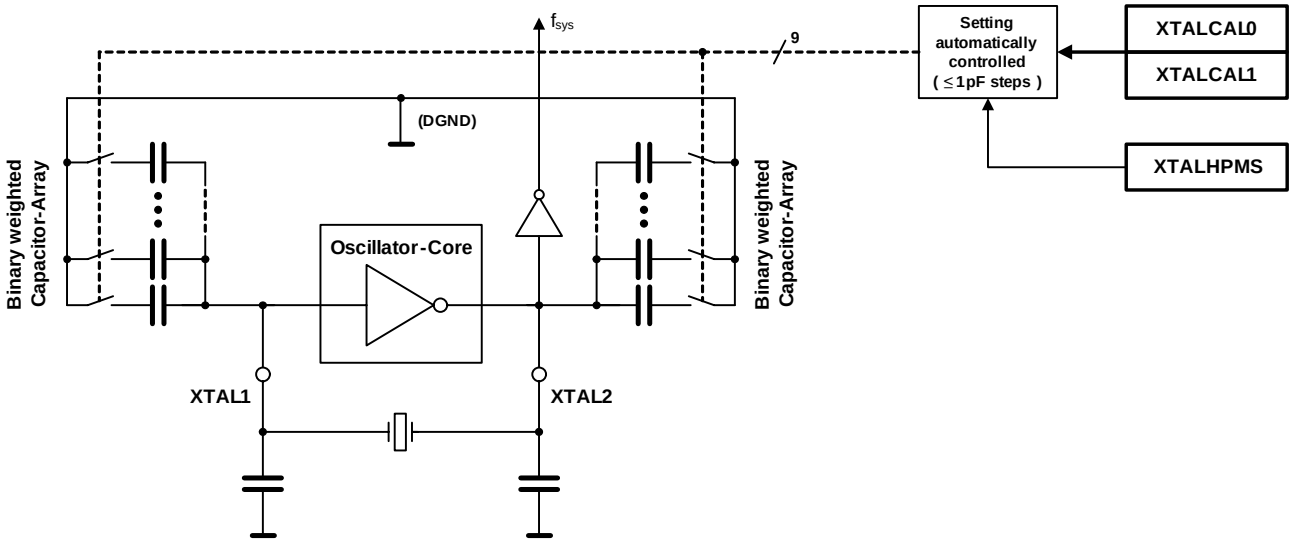


Figure 7 Crystal Oscillator

Recommended Trimming Procedure

- Set the registers XTALCAL0 and XTALCAL1 to the expected nominal values
- Set the TDA5235 to Run Mode Slave
- Wait for 0.5ms minimum
- Trim the oscillator by increasing and decreasing the values of XTALCAL0/1
- Register changes larger than 1 pF are automatically handled by the TDA5235 in 1pF steps
- After the Oscillator is trimmed, the TDA5235 can be set to SLEEP mode and keeps these values during SLEEP mode
- Add the settings of XTALCAL0/1 to the configuration. It must be set after every power up or brownout!

Using the High Precision Mode

As discussed earlier, the TDA5235 allows the crystal oscillator to be trimmed by the use of internal trim capacitors. It is also possible to use the trim functionality to compensate temperature drift of crystals.

During Run Mode (always when the receiver is active) the capacitors are automatically connected and the oscillator is working in the High Precision Mode.

On entering SLEEP Mode, the capacitors are automatically disconnected to save power.

If the High Precision Mode is also required for SLEEP Mode, the automatic disconnection of trim capacitors can be avoided by setting XTALHPMS to 1 (enable XTAL High Precision Mode during SLEEP Mode).

External Clock Generation Unit

A built in programmable frequency divider can be used to generate an external clock source out of the crystal reference. The 20 bit wide division factor is stored in the registers CLKOUT0, CLKOUT1 and CLKOUT2. The minimum value of the programmable frequency divider is 2. This programmable divider is followed by an additional divider by 2, which generates a 50% duty cycle of the CLK_OUT signal. So the maximum frequency at the CLK_OUT signal is the crystal frequency divided by 4. The minimum CLK_OUT frequency is the crystal frequency divided by 2^{21}.

To save power, this programmable clock signal can be disabled by the SFR control bit CLKOUTEN. In this case the external clock signal is set to low.

The resulting CLK_OUT frequency can be calculated by:

$$f_{CLKOUT} = \frac{f_{sys}}{2 \cdot divisionfactor}$$

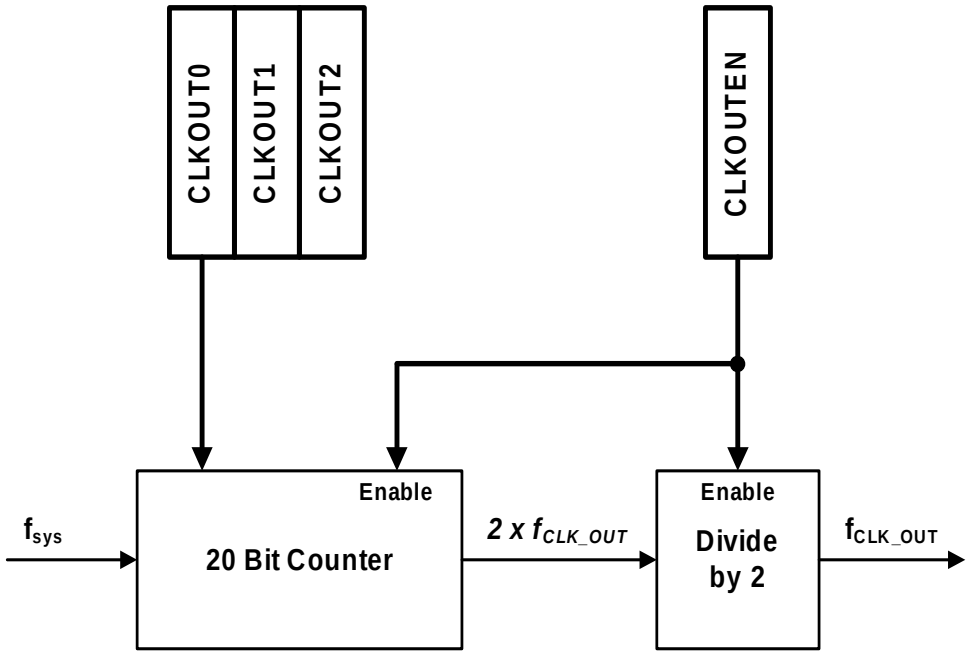


Figure 8 External Clock Generation Unit

The maximum CLK_OUT frequency is limited by the driver capability of the PPx pin and depends on the external load connected to this pin. Please be aware that large loads and/or high clock frequencies at this pin may interfere with the receiver and reduce performance.

After Reset the PPx pin is activated and the division factor is initialized to 11 (equals f_{CLK_OUT} = 998 kHz).

A clock output frequency higher than 1 MHz is not supported.

For high sensitivity applications, the use of the external clock generation unit is not recommended.

2.4.5 Sigma-Delta Fractional-N PLL Block

The Sigma-Delta Fractional-N PLL is fully integrated on chip. The ***V****oltage* ***C****ontrolled* ***O****scillator* (VCO) with on-chip LC-tank runs at approximately 3.6 GHz and is first divided with a band select divider by 1, 2 or 3 and then with an I/Q-divider by 4 which provides an orthogonal local oscillator signal for the first image reject mixer with the necessary high accuracy.
The multi-modulus divider determines the channel selection and is controlled by a 3rd order Sigma-Delta Modulator (*SDM*). A type IV phase detector, a charge pump with programmable current and an on-chip loop filter closes the phase locked loop.

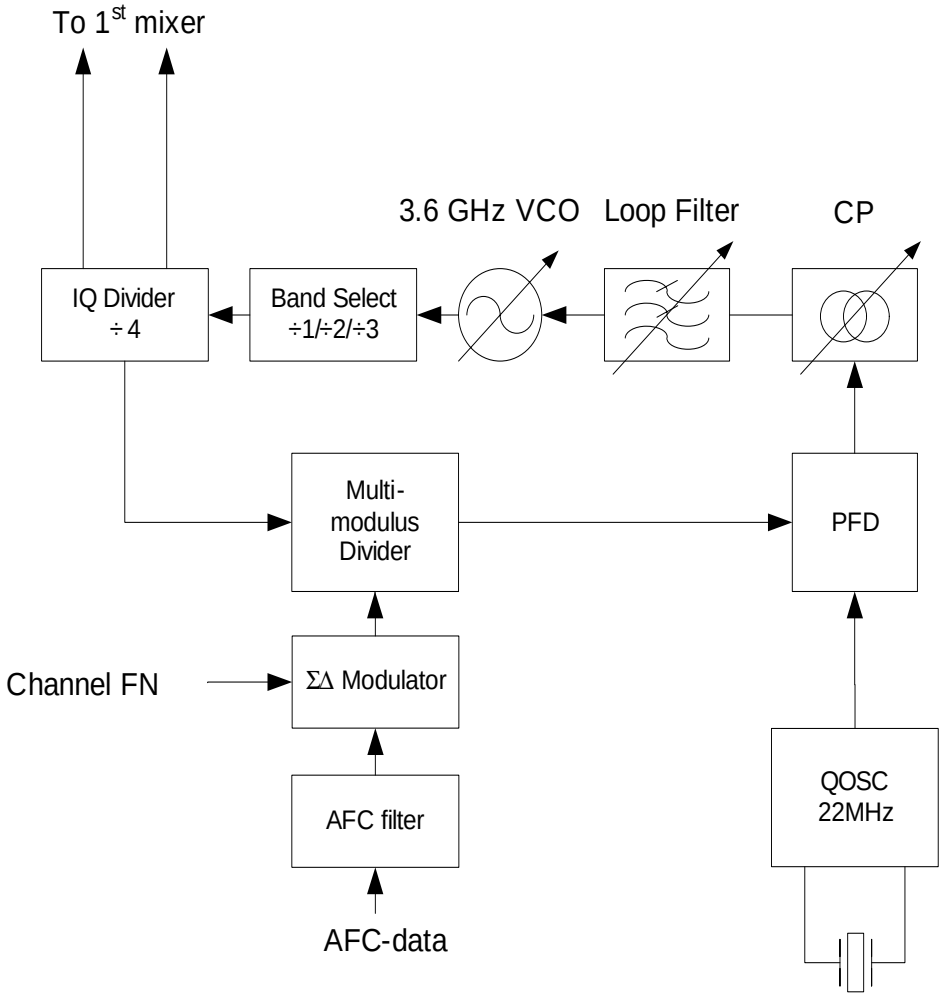


Figure 9 Synthesizer Block Diagram

When defining a Multi-Configuration system requiring different RF channels, the correct selection of channel spacing is extremely important. A general rule is not possible, but following must be considered:

• If an additional SAW filter is used, all channels including their tolerances have to be inside the SAW filter bandwidth.

• The distance between channels must be high enough, that no overlapping can occur. Strong input signals may still appear as recognizable input signal in the neighboring channel because of the limited suppression of IF Filters. Example: a typical 330kHz IF filter has at 10.3 MHz (10.7 MHz - 0.4 MHz) only 30 dB suppression. A -70 dBm input signal appears like a -100 dBm signal, which is inside the receiver sensitivity. In critical cases the use of two IF filters must be considered. See also **Chapter 2.4.3 RF/IF Receiver**.

2.4.5.1 PLL Dividers

The divider chain consists of a band select divider 1/2/3, an I/Q-divider by 4 which provides an orthogonal 1st local oscillator signal for the first image reject mixer with the necessary high accuracy and a multi-modulus divider controlled by the Sigma-Delta Modulator. With the band select divider, the wanted frequency band is selected. Divide by 1 selects the 915 MHz and 868 MHz band, divide by 2 selects the 434 MHz band and divide by 3 selects the 315 MHz band. The ISM band selection is done via bit group BANDSEL in x_PLLINTC1 register.

2.4.5.2 Digital Modulator

The 3rd order ***Sigma-Delta Modulator (SDM)*** has a 22 bit wide input word, however the LSB is always high, and is clocked by the XTAL oscillator. This determines the achievable frequency resolution.

The ***Automatic Frequency Control Unit*** filters the actual frequency offset from the FSK demodulator data and calculates the necessary correction of the divider factor to achieve the nominal IF center frequency.

2.4.6 ASK and FSK Demodulator

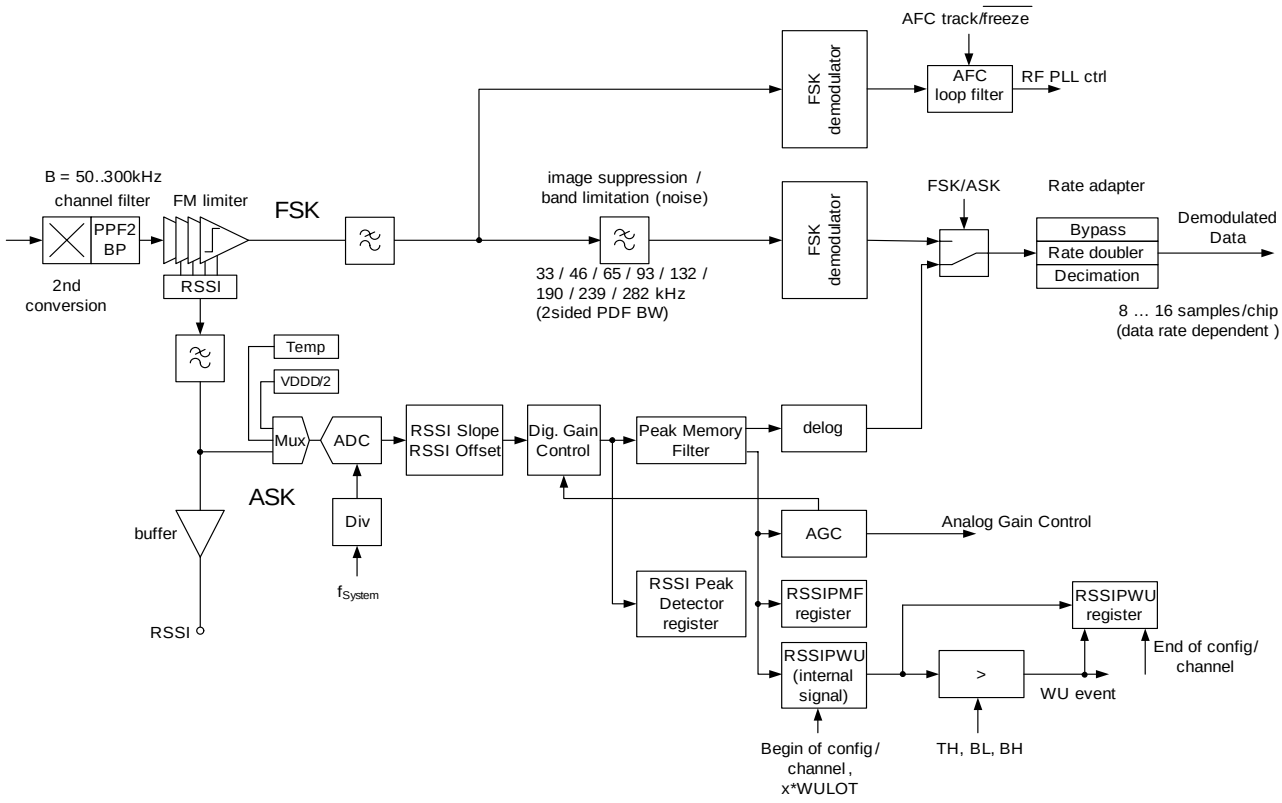


Figure 10 Functional Block Diagram ASK/FSK Demodulator

The IC comprises two separate demodulators for ASK and FSK.
After combining FSK and ASK data path, a sampling rate adaptation follows to meet an output oversampling between 8 and 16 samples per chip. Finally, an oversampling of 8 samples per chip can be achieved using a fractional sample rate converter (SRC) with linear interpolation (for further details see **Figure 15**).

2.4.6.1 ASK Demodulator

The RSSI generator delivers a DC signal proportional to the applied input power at a logarithmic scale (dBm) and is also used as an ASK demodulator. Via a programmable anti-aliasing filter this signal is converted to the digital domain by means of a 10-bit ADC. For the AM demodulation a signal proportional to the linear power is required. Therefore a conversion from logarithmic scale to linear scale is necessary. This is done in the digital domain by a nonlinear filter together with an exponential function. The analog RSSI signal after the anti-aliasing filter is available at the RSSI pin via a buffer amplifier. To enable this buffer the SFR control bit RSSIMONEN must be set. The anti-aliasing filter can be by-passed for visualization on the RSSI pin (see AAFBYP control bit).

2.4.6.2 FSK Demodulator

The limiter output signal, which has a constant amplitude over a wide range of the input signal, feeds the FSK demodulator. There is a configurable lowpass filter in front of the FSK demodulation to suppress the down conversion image and noise/limiter harmonics (FSK Pre-Demodulation Filter, PDF). This is realized as a 3rd order digital filter. The sampling rate after FSK demodulation is fixed and independent from the target data rate.

2.4.6.3 Automatic Frequency Control Unit (AFC)

In front of the image suppression filter a second FSK demodulator is used to derive the control signal for the **Automatic Frequency Control Unit**, which is actually the DC value of the FSK demodulated signal. This makes the AFC loop independent from signal path filtering and allow so a wider frequency capture range of the AFC. The derivation of the AFC control signal is preferably done during the DC free preamble and is then frozen for the rest of the datagram.

Since the digital FSK demodulator determines the exact frequency offset between the received input frequency and the programmed input center frequency of the receiver, this offset can be corrected through the sigma delta control of the PLL. As shown in **Figure 10**, for AFC purposes a parallel demodulation path is implemented. This path does not contain the digital low pass filter (PDF, Pre-Demodulation Filter). The entire IF bandwidth, filtered by the analog bandpass filter only, is processed by the AFC demodulator.

There are two options for the active time of the AFC loop:

- 1. always on
- 2. active for a programmable time relative to a signal identification event (several options can be programmed in SFR).

In the latter case the AFC can either be started or frozen relative to the signal identification. After the active time the offset for the sigma-delta PLL (SD PLL) is frozen.

The programming of the active time is especially necessary in case the expected frame structure contains a gap (noise) between wake-up and payload in order to avoid the AFC from drifting.

AFC works both for FSK and ASK. In the latter case the AFC loop only regulates during ASK data = high.

The maximum frequency offset generated by the AFC can be limited by means of the x_AFCLIMIT register. This limit can be used to avoid the AFC from drifting in the presence of interferers or when no RF input signal is available (AFC wander). A maximum AFC limit of 42 kHz is recommended. AFC wandering needs to be kept in mind especially when using Run Mode Slave.

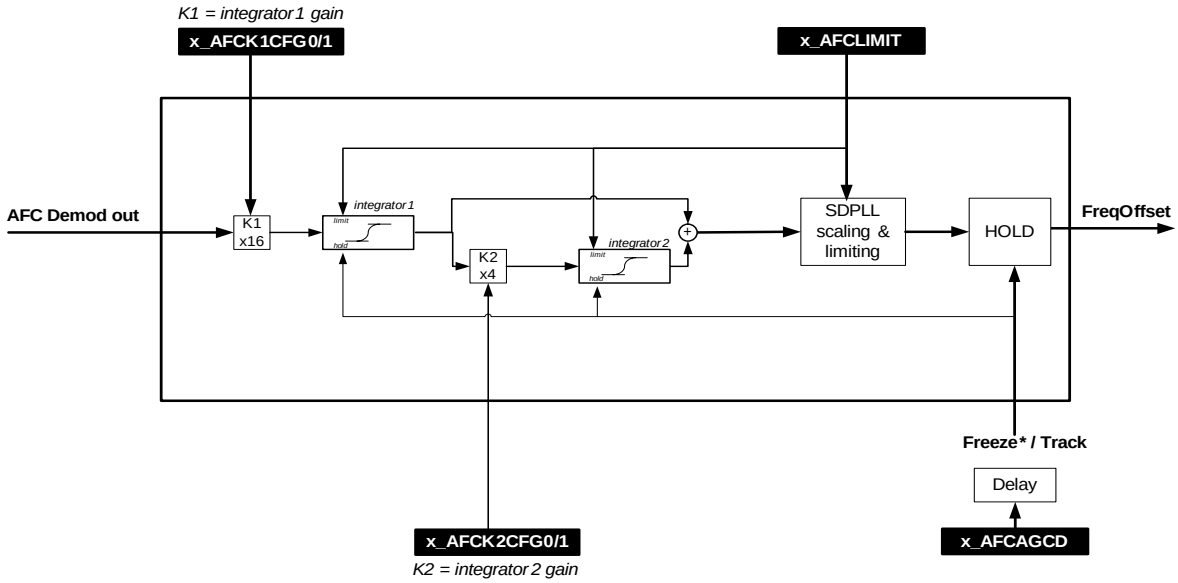


Figure 11 AFC Loop Filter (I-PI Filtering and Mapping)

The bandwidth (and thus settling time) of the loop is programmed by means of the integrator gain coefficients K1 and K2 (x_AFCK1CFG and x_AFCK2CFG register).

K1 mainly determines the bandwidth. K2 influences the dynamics/damping (overshoot) - smaller K2 means smaller overshoot, but slower dynamics. The bandwidth of the AFC loop is approximately 1.3*K1.

To avoid residual FM, limiting the AFC BW to 1/20 ~ 1/40 of the bit rate is suggested, therefore K1 must be set to approximately 1/50 ~ 1/100 of the bit rate. For most applications K2 can be set equal to K1 (overshoot is then <25%).

When very fast settling is necessary K1 and K2 can be increased up to bit rate/10, however, in this case approximately 1dB sensitivity loss is to be expected due to the AFC counteracting the input FSK signal.

AFC limitation at Local Oscillator (LO) frequencies at multiples of reference frequency (f_xtal). When AFC is activated and AFC drives the wanted LO frequency over the integer limit of Sigma Delta (SD) modulator, the SD modulator stucks at frac=1.0 or frac=0.0 due to saturation. So when AFC can change the integer value for the LO (register x_PLLINTC1) within the frequency range LO-frequency +/- AFC-limit, a change of the LO injection side or a smaller AFC-limit is recommended.

The frequency offset found by AFC (AFC loop filter output) can be readout via register AFCOFFSET, when AFC is activated. The value is in signed representation and has a frequency resolution of 2.68 kHz/digit. The output can be limited by the x_AFCLIMIT register.

2.4.6.4 Digital Automatic Gain Control Unit (AGC)

Automatic Gain Control (AGC) is necessary mainly because of the limited dynamic range of the on-chip bandpass filter (BPF). The dynamic range reduces to less than 60dB in case of minimum BPF bandwidth.

AGC is used to cover the following cases:

1. ASK demodulation at large input signals
2. RSSI reading at large input signals
3. Improve IIP3 performance in either FSK or ASK mode

The 1st IF buffer (PPFBUF, see **Figure 3**) can be fine tuned "manually" by means of 4 bits thus optimizing the overall gain to the application (attenuation of 0dB to -12dB by means of IFATT0 to IFATT15 in DDC mode; SDC mode has lower IFATT range). This buffer allows the production spread of external components to be trimmed.

The gain of the 2nd IF path is set to three different values by means of an AGC algorithm. Depending on whether the receiver is used in single down conversion or in double down conversion mode the gain control in the 2nd IF path is either after the 2nd poly-phase network or in front of the 2nd mixer.

The AGC action is illustrated in the RSSI curve below:

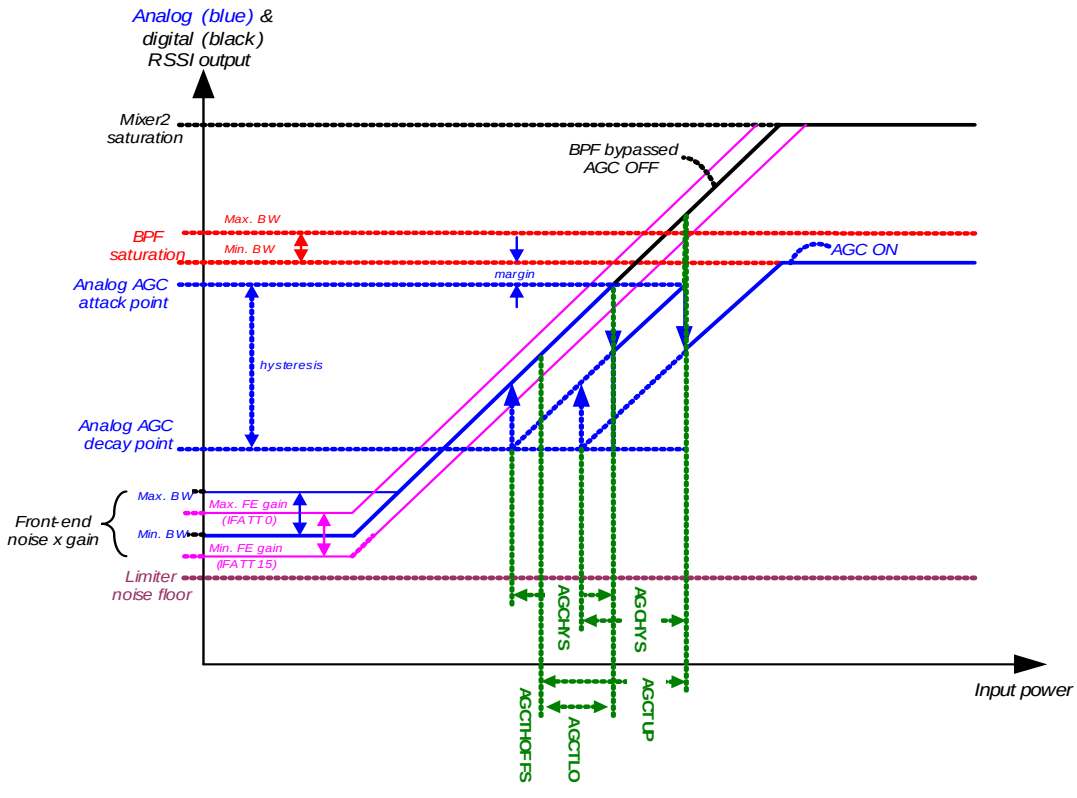


Figure 12 Analog RSSI output curve with AGC action ON (blue) vs. OFF (black)

Digital RSSI, AGC and Delog:

In order to match the analog RSSI signal to the **digital RSSI** output a correction is necessary. It adds an offset (RSSIOFFS) and modifies the slope (RSSISLOPE) such that standardized AGC levels and an appropriate DELOG table can be applied.

Upon entering the **AGC unit** the digital RSSI signal is passed through a Peak Memory Filter (PMF). This filter has programmable up and down integration time constants (PMFUP, PMFDN) to set attack respectively decay time. The integration time for decay time must be significantly longer than the attack time in order to avoid the AGC interfering with the ASK modulation.

The integrator is followed by two digital Schmitt triggers with programmable thresholds (AGCTLO; AGCTUP) - one Schmitt trigger for each of the two attack thresholds (two digital AGC switching points). The hysteresis of the Schmitt triggers is programmable (AGCHYS) and sets the decay threshold. The Schmitt triggers control both the analog gain as well as the corresponding (programmable) digital gain correction (DGC).

The difference ("error") signal in the PMF is actually a normalized version of the modulation. This signal is then used as input for the **DELOG** table.

AGC threshold programming

The SFR description for the AGC thresholds are in dBs. The first value to set is the AGC threshold offset in AGCTHOFFS.

This value is the offset relative to 0 input (no noise, no signal), which for the default setting of gain, and assuming typical insertion loss of matching network and ceramic filter, can be extrapolated to be approximately -143dBm.

In this case the default setting of the AGCTHOFFS of 63.9dB corresponds to an input power of approximately -79dBm (= -143dBm + 63.9dB).

The low (digital) AGC threshold is then -79 + 12.8dB (default AGCTLO) = -66dBm and the upper (digital) AGC threshold is -79 + 25.6 (default AGCTUP) = -53dBm.

Therefore a margin of about 6dB is indicated before a degradation of the linearity of the 2^{nd} IF can be observed when using the 50kHz BPF or even about 16dB when using the 300kHz BPF.

The input power level at which the AGC switches back to maximum gain is -66dBm - 21.3dB (default AGCHYS) = -87dBm. This provides enough margin against the minimum sensitivity.

When AGC is activated, RSSI is untrimmed, IFATT <= 5.6dB and the same RSSI offset should be applied for all bandpass filter settings, then the settings in **Table 2** can be applied, where a small reduction of the RSSI input range can be observed.

Table 2 AGC Settings 1

AGC Threshold Hysteresis = 21.3 dB					
AGC Digital RSSI Gain Correction = 15.5 dB					
BPF	**RSSI Offset Compensation (untrimmed) 1)**	**AGC Threshold Offset**	**AGC Threshold Low**	**AGC Threshold Up**	**RSSI Input Range Reduction**
300 kHz	32	63.9 dB	8	4	5 dB
200 kHz	32	63.9 dB	6	2	5 dB
125 kHz	32	63.9 dB	5	0	5 dB
80 kHz	32	51.1 dB	11	6	2.8 dB
50 kHz	32	51.1 dB	9	5	0 dB

1) Note: This value needs to be used for calculating the register value

For the full RSSI input range, the values in **Table 3** can be applied.

Table 3 AGC Settings 2

AGC Threshold Hysteresis = 21.3 dB				
AGC Digital RSSI Gain Correction = 15.5 dB				
BPF	**RSSI Offset Compensation (untrimmed) 1)**	**AGC Threshold Offset**	**AGC Threshold Low**	**AGC Threshold Up**
300 kHz	-18	63.9 dB	5	1
200 kHz	-18	51.1 dB	11	7
125 kHz	-18	51.1 dB	10	5
80 kHz	4	51.1 dB	9	5
50 kHz	32	51.1 dB	9	5

1) Note: This value needs to be used for calculating the register value

Attack and Decay coefficients PMF-UP & PMF-DOWN:

The settling time of the loop is determined by means of the integrator gain coefficients PMFUP and PMFDN, which need to be calculated from the wanted attack and decay times.

The ADC is running at a fixed sampling frequency of 274kHz. Therefore the integrator is integrating with PMFUP*274k per second, i.e. time constant is 1/(PMFUP*274k). The attack times are typically 16 times faster than the decay times.

Typical calculation of the coefficients by means of an example:

- PMFUP = 2^-round(ln(AttTime / BitRate * 274kHz) / ln(2))
- PMFDN = 2^-round(ln(DecTime / BitRate * 274kHz) / ln(2)) / PMFUP

where AttTime, DecTime = attack, decay time in number of bits

Note: PMFDN = overall_PMFDN / PMFUP

Example:

BitRate = 2kbps

AttTime = 0.1 bits

=> PMFUP = 2^-round(ln(0.1bit/2kbps*274kHz)/ln(2)) = 2^-round(3.8) = 2^-4

DecTime = 2 bits

=> PMFDN = 2^-round(ln(2bit/2kbps*274kHz)/ln(2))/PMFUP = 2^-round(8.1)/2^-4 = 2^-4

Note: In case of ASK with large modulation index the attack time (PMFUP) can be up to a factor 2 slower due to the fact that the ASK signal has a duty cycle of 50% - during the ASK low duration the integrator is actually slightly discharged due to the decay set by PMFDN.

The AGC start and freeze times are programmable. The same conditions can be used as in the corresponding AFC section above. They will however, be programmed in separate SFR registers.

2.4.6.5 Analog to Digital Converter (ADC)

In front of the AD converter there is a multiplexer so that also temperature and VDDD can be measured (see **Figure 10**).

The default value of the ADC-MUX is RSSI (register ADCINSEL: 000 for RSSI; 001 for Temperature; 010 for VDDD/2).

After switching ADC-MUX to a value other than RSSI in SLEEP Mode, the internal references are activated and this ADC start-up lasts 100µs. So after this ADC start-up time the readout measurements may begin. The chip stays in this mode until reconfiguration of register ADCINSEL to setting RSSI. However, it is recommended to measure temperature during SLEEP mode (This is also valid for VDDD).

Readout of the 10-bit ADC has to be done via ADCRESH register (the lower 2 bits in ADCRESL register can be inconsistent and should not be used).

Typical the ADC refresh rate is 3.7 µs. Time duration between two ADC readouts has to be at least 3.7 µs, so this is already achieved due to the maximum SPI rate (16 bit for SPI command and address last 8µs at an SPI rate of 2MBit/s). The EOC bit (end of conversion) indicates a successful conversion additionally. Repetition of the readout measurement for several times is for averaging purpose.

The input voltage of the ADC is in the range of 1 .. 2 V. Therefore VDDD/2 (= 1.65 V typical) is used to monitor VDDD.

Further details on the measurement and calibration procedure for temperature and VDDD can be taken from the corresponding application note.

2.4.7 RSSI Peak Detector

The IC possesses several digital RSSI peak level detectors. The RSSI level is averaged over 4 samples before it is fed to any of the peak detectors. This prevents the evaluated peak values to be dominated by single noise peaks.

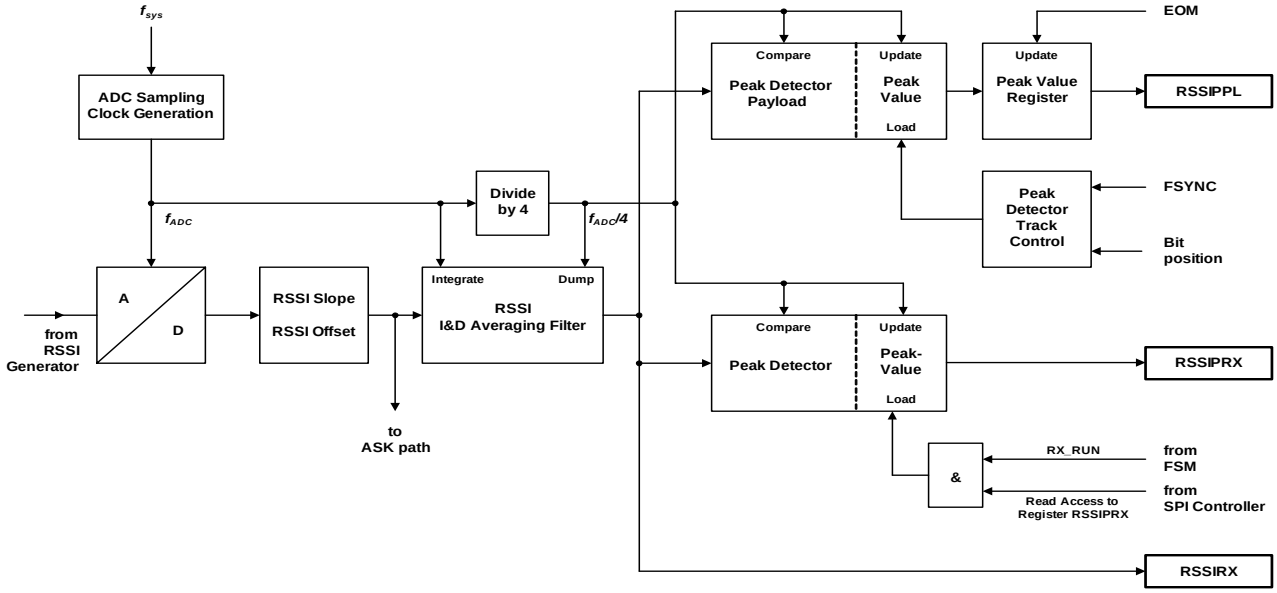


Figure 13 Peak Detector Unit

Peak Detector Payload is used to measure the input signal power of a received and accepted data telegram. It is read via SFR RSSIPPL.

Observation of the RSSI signal starts at the detection of a TSI (FSYNC) and ends with the detection of EOM. The internal RSSIPPL value is cleared after FSYNC. The evaluated RSSI peak level RSSIPPL is transferred to the RSSIPPL register at EOM. Starting the observation of the RSSI level can be delayed by a selectable number of data bits and is controlled by the register x_PKBITPOS. A latency in the generation of FSYNC and EOM of approx. 2..3 bits in relation to the contents of the Peak Detector must be considered. Within the boundaries described, the register RSSIPPL always contains the peak value of the last completely received data telegram. The register RSSIPPL is reset to 0 at power up reset only.

Peak Detector is used to measure RSSI independent of a data transfer and to digitally trim RSSI. It is read via SFR RSSIPRX.

Observation of the RSSI signal is active whenever the RX_RUN signal is high. The RSSIPRX register is refreshed and the Peak Detector is reset after every read access to RSSIPRX.

It may be required to read RSSIPRX twice to obtain the required result. This is because, for example, during a trim procedure in which the input signal power is reduced, after

reading RSSIPRX, the peak detector will still hold the higher RSSI level. After reading RSSIPRX the lower RSSI level is loaded into the Peak Detector and can be read by reading RSSIPRX again.

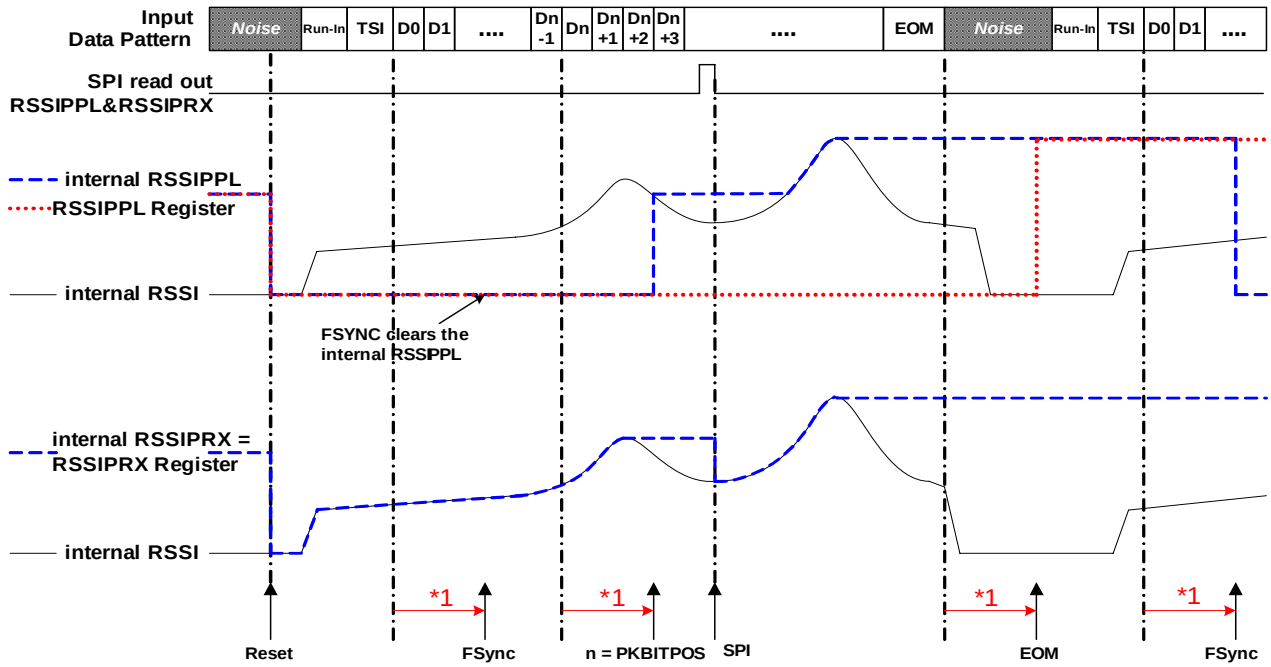
Register RSSIPRX should not be read-out faster than 41µs in case AGC is ON (as register value would not represent the actual, but a lower value).

When the RX_RUN signal is inactive, a read access has no influence to the peak detector value. The register RSSIPRX is reset to 0 at power up reset.

Peak Detector Wake-Up RSSIPWU (see **Figure 10**) is used to measure the input signal power during Wake-Up search. The internal signal RSSIPWU gets initialized to 0 at start of the first observation time window at the beginning of each configuration. The peak value of this signal is tracked during Wake-Up search.

In case of a Wake-Up, the actual peak value is written in the RSSIPWU register. Even in case no Wake-Up occurred, actual peak value is written in the RSSIPWU register at the end of the actual configuration of the Self Polling period. So if no Wake-Up occurred, then the RSSIPWU register contains the peak value of the last configuration of the Self Polling period, even in a Multi-Configuration setup. This functionality can be used to track RSSI during unsuccessful Wake-Up search due to no input signal or due to blocking RSSI detection.

For further details please refer to **Chapter 2.4.8.5 Wake-Up Generator** and **Chapter 2.6.2 Polling Timer Unit**.



*1 Computation Delay due to filtering and signal calculation.

Figure 14 Peak Detector Behavior

Recommended Digital Trimming Procedure

- Download configuration file (Run Mode Slave; RSSISLOPE, RSSIOFFS set to default, i.e. RSSISLOPE=1, RSSIOFFS=0)
- Turn off AGC (AGCSTART=0) and set gain to AGCGAIN=0
- Apply P_{IN1} = -85 dBm RF input signal
- Read RSSIRX eleven times (minimum 10 ms in-between readings), use average of last ten readings (always), store as RSSIM1
- Apply P_{IN2} = -65 dBm RF input signal
- Read RSSIRX eleven times (minimum 10 ms in-between readings), use average of last ten readings (always), store as RSSIM2
- Calculate measured RSSI slope SLOPEM=(RSSIM2-RSSIM1)/(P_{IN2}-P_{IN1})
- Adjust RSSISLOPE for required RSSI slope SLOPER as follows: RSSISLOPE=SLOPER/SLOPEM
- Adjust RSSIOFFS for required value RSSIR2 at P_{IN2} as follows: RSSIOFFS=(RSSIR2-RSSIM2)+(SLOPEM-SLOPER)*P_{IN2}
- The new values for RSSISLOPE and RSSIOFFS have to be added to the configuration!

Notes:

1. The upper RF input level must stay well below the saturation level of the receiver (see **Chapter 2.4.6.4 Digital Automatic Gain Control Unit (AGC)**)

2. The lower RF input level must stay well above the noise level of the receiver

3. If IF Attenuation is trimmed, this has to be done before trimming of RSSI

4. If RSSI needs to be trimmed in a higher input power range the AGCGAIN must be set accordingly

2.4.8 Digital Baseband (DBB) Receiver

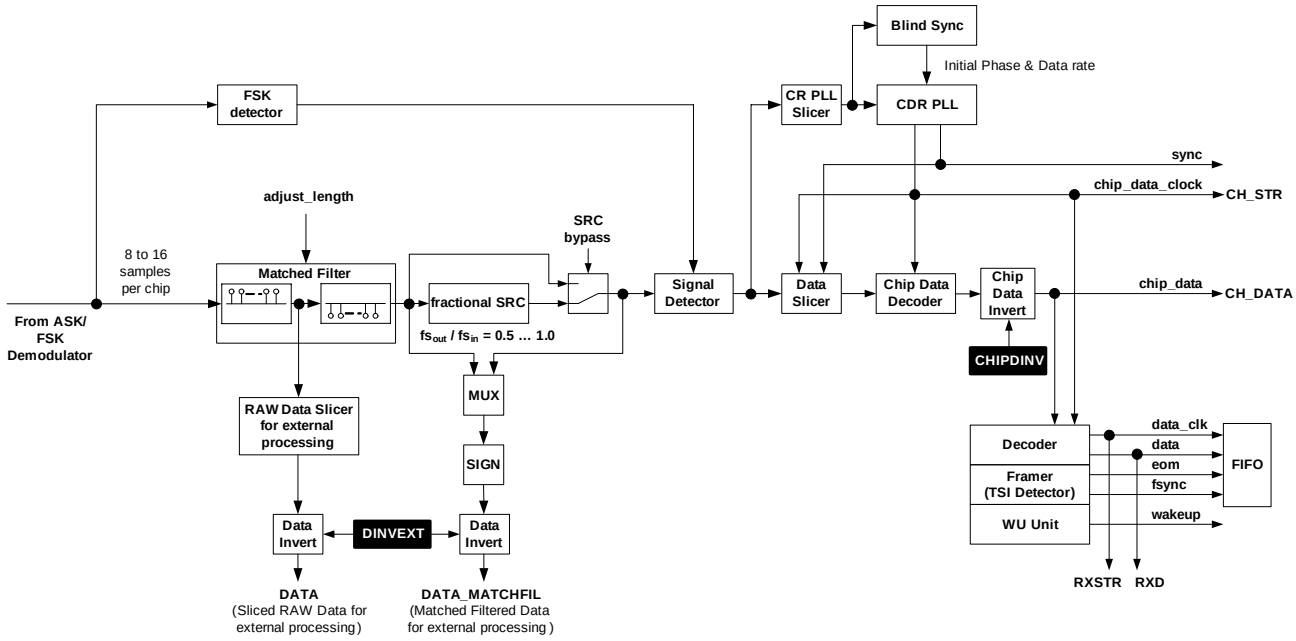


Figure 15 Functional Block Diagram Digital Baseband Receiver

The digital baseband receiver comprises a matched data filter, a clock and data recovery, a data slicer, a line decoder, a wake-up generator, a frame synchronization and a data FIFO. The recovered data and clock signals are accessible via 2 separate pins. The FIFO data buffer is accessible via the SPI bus interface.

2.4.8.1 Data Filter and Signal Detection

The data filter is a matched filter (*MF*). The frequency response of a matched filter has ideally the same shape as the power spectral density (*PSD*) of the originally transmitted signal, therefore the signal-to-noise ratio (*SNR*) at the output of the matched filter becomes maximum. The input sampling rate of the baseband receiver has to be between 8 and 16 samples per chip. The oversampling factor within this range is depending on the data rate (see **Figure 10**). The MF has to be adjusted accordingly to this oversampling. After the MF a fractional sample rate converter (SRC) is applied using linear interpolation. Depending on the data rate decimation is adjusted within the range 1...2. Finally, at the output of the fractional SRC the sampling rate is adjusted to 8 samples per chip for further processing.

To distinguish whether the incoming signal is really a signal or only noise adequate detectors for ASK and FSK are built in.

Signal and Noise Detector

The Signal Detector decides between acceptable and unacceptable data (e.g. noise). This decision is taken by comparing the signal power of the actually received data (register SPWR) with a configurable threshold level (registers x_SIGDET0/1), which must be evaluated. In case the actual signal power is above the threshold, acceptable data has been detected.

To decide in case of FSK whether there is a data signal or simply noise at the output of the rate adapter, there is a Noise Detector implemented. The principle is based on a power measurement of the demodulated signal. The current noise power is stored in the NPWR register and is updated at every SPI controller access. The Noise Detector is useful if data signal is transmitted with small FSK deviations. In case the current noise power (register NPWR) is below the configurable threshold (register x_NDTHRES), a data signal has been detected.

The Signal Recognition mode must be configured based on whether ASK or FSK modulation is used. Signal Recognition can be a combination of Signal Detector and Noise Detector:

- Signal Detector (=Squelch) only (related registers: x_SIGDET0, x_SIGDET1 and SPWR). This mode is generally used for ASK and recommended for FSK.
- Noise Detector only (related registers: x_NDTHRES and NPWR).
- Signal and Noise Detector simultaneously.
- Signal and Noise Detector simultaneously, but the FSK noise detect signal is valid only if the x_SIGDETLO threshold is exceeded. This is the recommended FSK mode, if minimum FSK deviation is not sufficient to use Signal Detector only.

Signal Recognition can also be used as Wake-up on Level criterion (see **Chapter 2.4.8.5**).

Figure 16 shows the system characteristics to consider in choosing the best Signal Detector level. On the one hand, a higher SIGDET threshold level must be set for achieving good FAR (False Alarm Rate) performance, but then the MER/BER (Message Error Rate/Bit Error Rate) performance will decrease. On the other hand, the MER/BER performance can be increased by setting smaller SIGDET threshold levels but then the FAR performance will worsen.

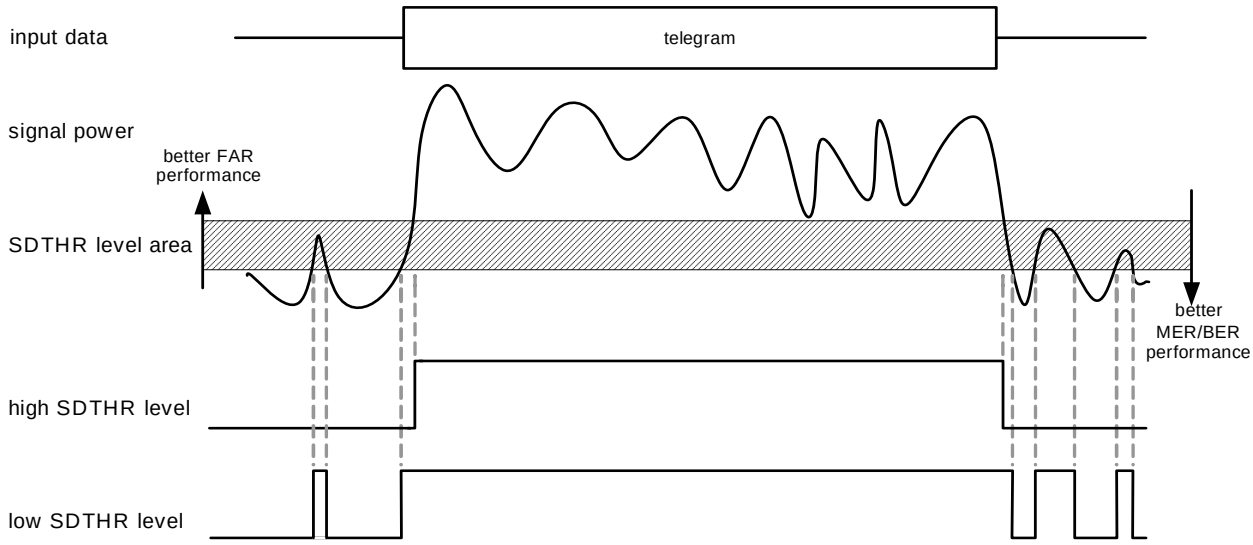


Figure 16 Signal Detector Threshold Level

Quick Procedure to Determine Signal and Noise Detector Thresholds

Preparation

A setup is required with original RF hardware as in the final application. The values of SPWR and NPWR can be read via the final application.

A complete configuration file using right modulation, data rate and Run Mode Slave, must be prepared and downloaded to the TDA5235.

Signal Detector Threshold for ASK

Take 500 readings of SPWR (50 are also possible, but this leads to less accurate results) with no RF input signal applied (=noise only). Calculate average and Standard Deviation. Signal Detector Threshold is average plus 2 times the Standard Deviation. To load the x_SIGDET0/1 register the calculated value must be rounded and converted to hexadecimals. For a final application, the Signal Detector Threshold should be varied to optimize the false alarm rate and the sensitivity.

Signal and Noise Detector Thresholds for FSK

Signal Detector Threshold

Do 500 (50) readings of SPWR with no RF input signal applied (=noise only). Calculate average and Standard Deviation. Signal Detector Threshold is average plus 2 times the Standard Deviation. Of course this value has to be rounded and converted to

hexadecimals. For a final application the Signal Detector Threshold should be varied to optimize the false alarm rate and the sensitivity.

Verification if Squelch only is possible

Apply a bit pattern (e.g. PRBS9) with correct data rate at about -80 dBm input signal power and minimum FSK deviation to the RF input. Do 500 (50) readings of SPWR, calculate average minus three times the Standard Deviation. This value should be higher than the calculated Signal Detector Threshold calculated above. If this is not the case, Signal Detector AND Noise Detector must be used.

Noise Detector Threshold

Do 500 (50) readings of NPWR with no RF input signal applied (=noise only). Calculate average and Standard Deviation. Noise Detector Threshold is average minus the Standard Deviation. Round this value and convert it to hexadecimals. For a final application, the Noise Detector Threshold should be varied to optimize false alarm rate and sensitivity.

Signal Detector Low Threshold

The Signal Detector Low Threshold is always required in combination with the Noise Detector.

Set register bit SDLORE to 1 and set bit group SDLORSEL to 00. Apply a bit pattern (e.g. PRBS9) at correct data rate at about -80 dBm input signal power and minimum FSK deviation to the RF input. Do 500 (50) readings of SPWR, calculate average. If average is larger than 200 dec (=0xC8), SDLORSEL has to be increased to the next larger value until average is smaller than 200 dec. x_SIGDETLO = 0.8 * (average - 3 * Standard Deviation). Set register SDLORE back to 0. The last setting of bit group SDLORSEL must also be used for configuration!

Verification

Threshold settings should be verified by testing receiver sensitivity over the input frequency range, with a step size of 100Hz, at minimum FSK deviation with all combinations of minimum and maximum data rate and duty cycle.

Further detailed information can be taken from the corresponding Application Note.

2.4.8.2 Encoding Modes

The IC supports the following Bi-phase encodings:

- Manchester code
- Differential Manchester code
- Bi-phase space code
- Bi-phase mark code

The encoding mode is set and enabled by bit group CODE in x_DIGRXC configuration register.

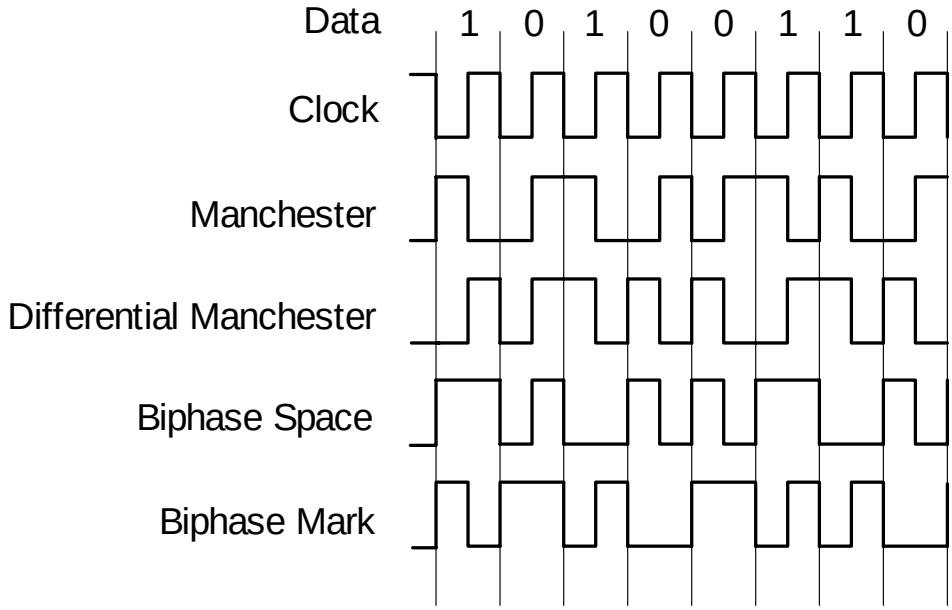


Figure 17 Coding Schemes

The encoding modes Inverted Manchester and Inverted Differential Manchester can also be decoded internally by usage of CHIPDINV bit in x_DIGRXC register (see **Figure 15**).

All the Manchester symbol combinations including Code Violations are shown in **Figure 18**. Digital 0 and 1 are coded with the change of the amplitude in the middle of the symbol period. The Code Violations (CV) M (mark) and S (space), are coded as low/high signal levels.

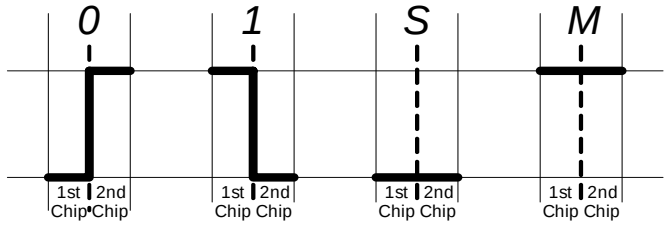


Figure 18 Manchester Symbols including Code Violations

2.4.8.3 Clock and Data Recovery

An all-digital PLL *(*ADPLL*)* recovers the data clock from the incoming data stream. The second main function is the generation of a signal indicating symbol synchronization. Synchronization on the incoming data stream generally occurs within the first 4 bits of a telegram.

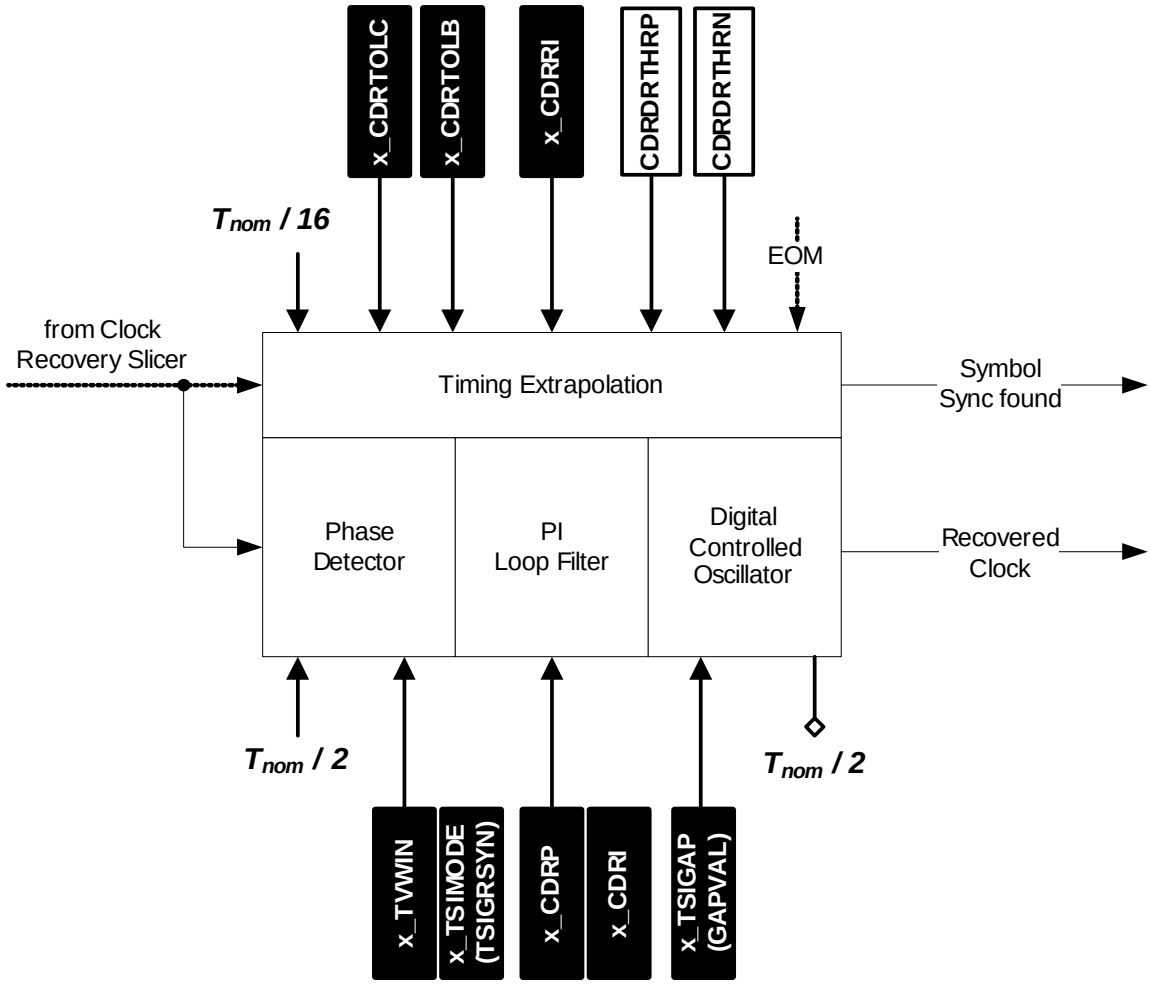


Figure 19 Clock Recovery (ADPLL)

Clock Recovery is implemented as standard ADPLL PI regulator with Timing Extrapolation Unit for fast settling.

In the unlocked state, the Timing Extrapolation Unit calculates the frequency offset for the incoming data stream. If the defined number of Bi-phase encoded bits are detected (the RUNIN length can be set in the x_CDRRI register), the I-part and the PLL oscillator will be set and the PLL will be locked.

When x_CDRRI.RUNLEN is set to small values, then the I-part is less accurate (residual error) and can lead to a longer needed PLL settling time and worse performance in the

first following bits. Therefore the selected default value is a good compromise between fast symbol synchronization and accuracy/performance.

Duty cycle and data rate acceptance limits are adjustable via registers. After locking, the clock must be stable and must follow the reference input. Therefore, a rapid settling procedure (Timing Extrapolation Unit) and a slow PLL are implemented.

If the PLL is locked, the reference signal from the Clock Recovery Slicer is used in the phase detector block to compute the actual error. The error is used in the PI loop filter to set the digital controlled oscillator running frequency. For the P, I and Timing Extrapolation Unit settings, the default values for the x_CDRP and x_CDRI control registers are recommended.

The PLL will be unlocked, if a code violation of more than the defined length is detected, which is set in the x_TVWIN control register. Another criterion for PLL resynchronization is an End Of Message (EOM) signalled by the Framer block.

The PLL oscillator generates the chip clock (2 * f_{data}).

The internal PLL lock signal used by the Framer is generated up to 1 bit before RUNIN ends. The Timing Extrapolation Unit counts the incoming edges and interprets the delay between two edges as a bit or a chip. Due to the fact that the first edge of a “Low” bit, coded as ’0’ and ’1’, rises one chip later than a “High” bit, the PLL locks later in this case (see **Figure 20**). The real needed RUNIN time can be shorter than the configured RUNIN length in the x_CDRRI register by up to two chips. This should be considered when setting the TSI pattern and/or TSI length. See also **Chapter 2.4.8.6 Frame Synchronization**.

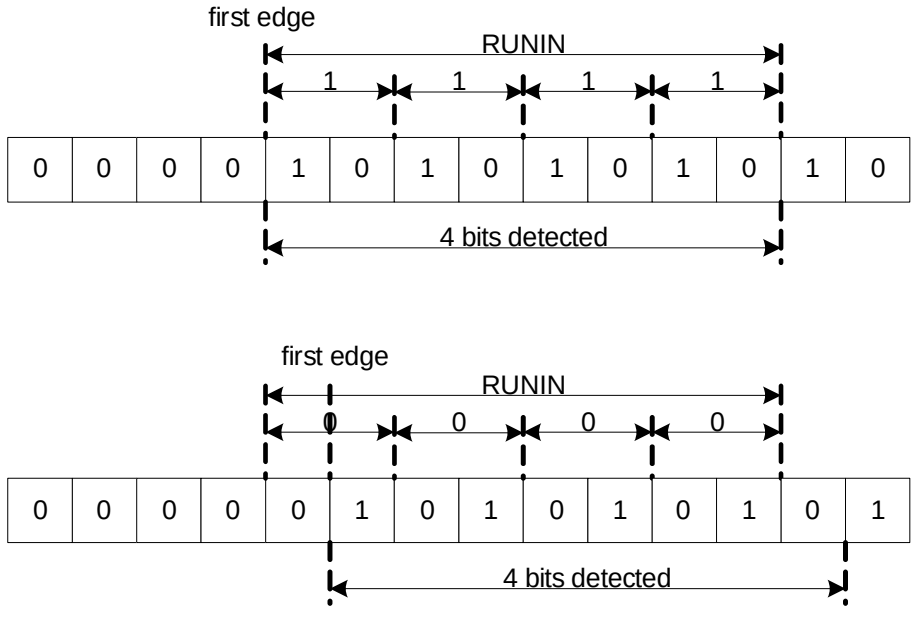


Figure 20 RUNIN Generation Principle

Number of Required RUNIN Bits

The number of RUNIN bits specified in x_CDRRI register should always be 3.0. This setting defines the duration of the internal synchronization. Because of internal processing delays, the pattern length that must be reserved for RUNIN is longer.

The ideal RUNIN pattern is a series of either Manchester 1's or Manchester 0's. This pattern includes the highest number of edges that can be used for synchronization. In this case, the number of physically sent RUNIN bits is 4.

For any other RUNIN pattern, 5.5 bits should be reserved for RUNIN.

TVWIN (Timing Violation WINdow length)

The PLL unlocks if the reference signal is lost for more than the time defined in the x_TVWIN register. During the TSI Gap (see TSI Gap Mode in **Chapter 2.4.8.6 Frame Synchronization**), the PLL and the TVWIN are frozen.

TVWIN time is the time during which the Digital Baseband Receiver should stay locked without any incoming signal edges detected. The time resolution is T/16.

Calculation of TVWIN can be seen at the end of subsection TSI Gap Mode in **Chapter 2.4.8.6 Frame Synchronization**.

Duty Cycle Variation

Ideally, the input signal to the Clock and Data Recovery (CDR) would have a chip width of 8 samples and a bit width of 16 samples and the CDR would not lock onto any input that violates this. However, due to variations in the duty cycle this stringent assumption for the pulse widths will in general not be true. Therefore it is necessary to loosen this requirement by using tolerance windows.

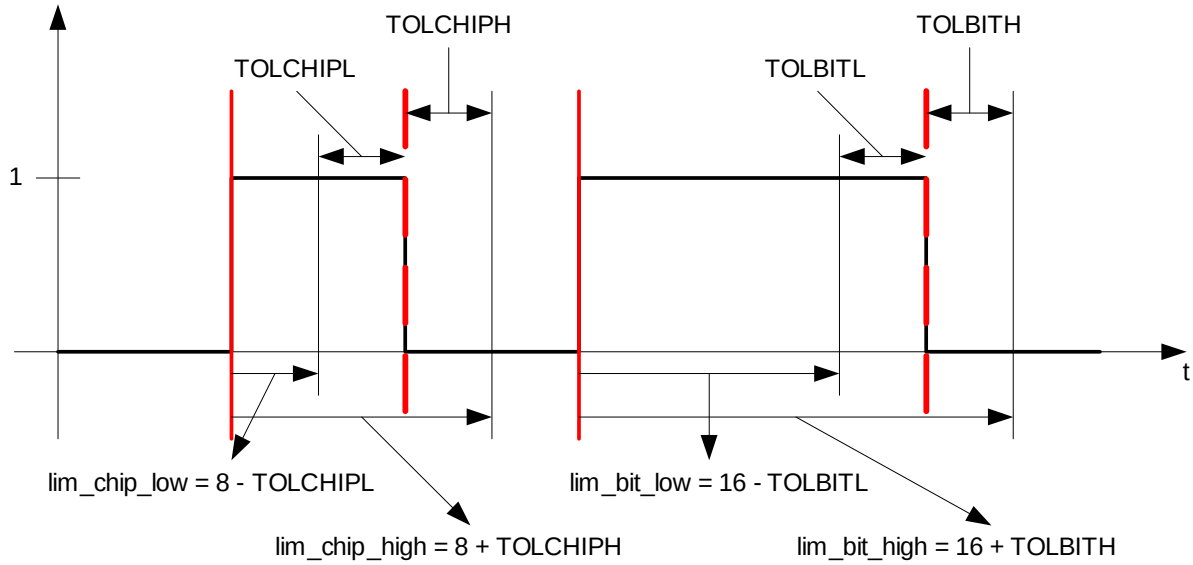


Figure 21 Definition of Tolerance Windows for the CDR

There exist now two registers - x_CDRTOLC for the chip width tolerance and x_CDRTOLB for the bit width tolerance - that can be used to tighten or loosen the windows around the ideal pulse widths. As it can easily be seen from **Figure 21**, tighter windows will result in more stringent requirements for the input data to have a 50% duty cycle and bigger windows will allow the duty cycle to vary more. **Figure 21** also depicts the meaning of the bits in the registers x_CDRTOLC and x_CDRTOLB.

Data Rate Acceptance Limitation

The Clock and Data Recovery is able to accept data rate errors of more than +/-15% with a certain loss of performance. There exist Multi-Configuration applications where the data rate of both configurations are within this range. So the adjacent data rates of these configurations are disturbing each other. The limitation of the data rate acceptance can be activated in this case.

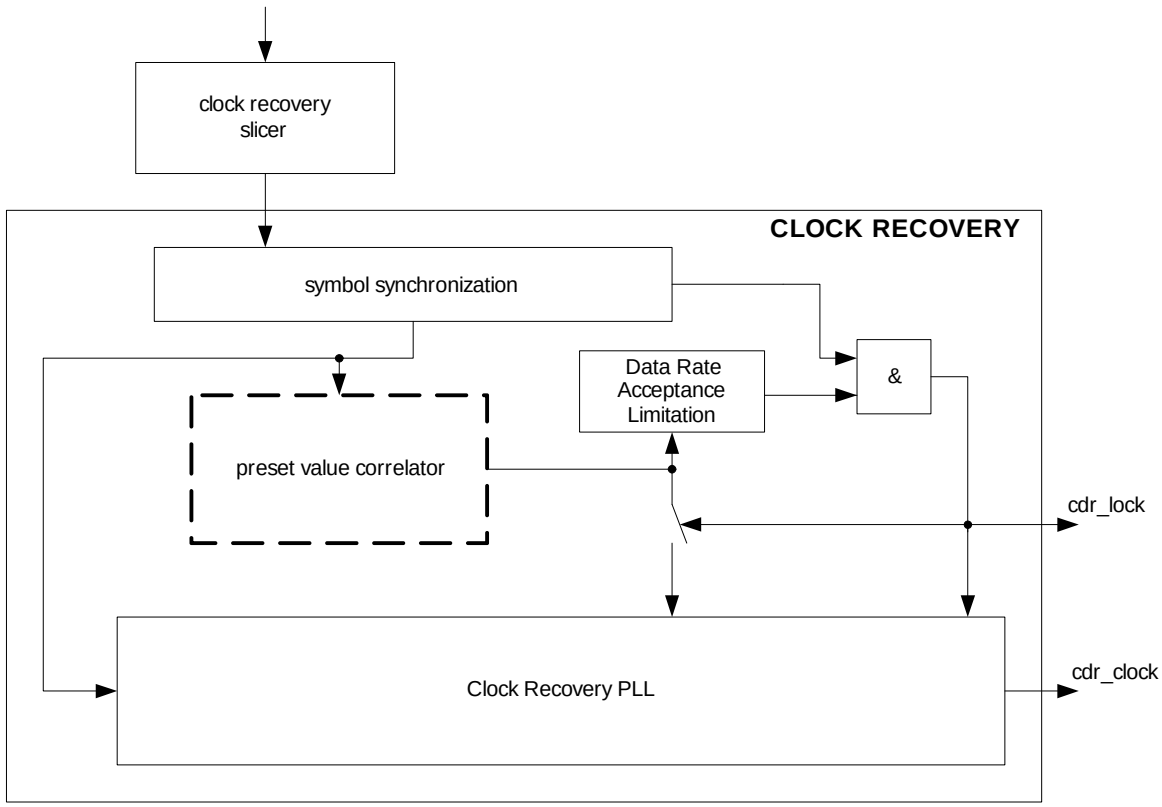


Figure 22 Data Rate Acceptance Limitation

The clock and data recovery (CDR) regenerates the clock based on the input data delivered from the clock recovery (CR) slicer. Symbol synchronization (cdr_lock) is achieved when a specified number of chips (can be set via register x_CDRRI.RUNLEN) has a valid pulse width. In parallel the preset value correlator estimates a preset value for the clock recovery PLL so that a shorter settling time is achieved. This preset value is also proportional to the data rate and is therefore used in the data rate acceptance limitation block. If the preset value is outside a certain range (positive and negative threshold configurable via registers CDRDRTHRP and CDRDRTHRN), the CDR does not go into lock and no symbol synchronization is generated.

For each configuration there exists one bit (register x_CDRRI.DRLIMEN) to switch the data rate acceptance limitation functionality on or off. Data rate acceptance limitation is disabled by default. All configurations share the same threshold registers, the default

thresholds are set so that almost all packets with a data rate error of +/-10% and larger are rejected.

The following statements summarize some important aspects that need to be kept in mind when using the described functionality:

- The output of the estimator must be described on statistical terms - this means that it can not be guaranteed that all packets with a certain data rate outside the allowed range will be rejected
- The quality of the estimated data rate value is mainly influenced by the setting of the signal and noise detectors
- Reducing the RUNIN length in register x_CDRRI reduces the quality of the data rate estimation, resulting in a degradation of the performance of the data rate acceptance limitation block
- The same threshold can be used for FSK and ASK
- If the thresholds are too small it may happen that also packets with a valid data rate are rejected

2.4.8.4 Data Slicer and Line Decoding

The output signal of the matched filter within the internal data processing path is in the range of +x to -x (x is the maximum value of the internal bit width). If Code Violations within a Manchester encoded bitstream have to be detected, the data slicer has to recover the underlying chipstream instead of the bitstream. In this case zero values at the matched filter output lead to an additional slicing threshold and an implicit sensitivity loss. To provide the full reachable sensitivity for applications which do not need the symbols S (space) and M (mark), the data slicer has two different operating modes:

- Chip mode (Code Violations are allowed)
- Bit mode (without Code Violations)

The chip mode introduces an implicit sensitivity loss compared to the bit mode, because a zero-crossing in the 2-chip matched filter signal must be detectable. This is only possible when an additional slicing level is introduced in the data slicer.

The data slicer internally maps a positive value to a 1 and a negative value to a -1. Everything inside the zero thresholds (zero-tube) becomes a 0. After that, the decoding to the chip-level representation is done by mapping the -1 to a "0" chip and the 1 to a "1" chip. A zero out of the data slicer is decoded to chip-level by referencing to the previous chip value.

In bit mode the data slicer has only one threshold (zero) to distinguish between the two levels of the matched filter output. The data slicer internally maps a positive value to a 1 and a negative value to a -1. After that, the selected line decoding is applied.

Summary of data slicer modes in the TDA5235:

Data Slicer Chip mode:

- Code violations detectable (TSI, or EOM)
- Performance loss compared to bit mode
- Activation via setting register x_SLCCFG to a value of
  + 0x90 (Chip Mode EOM-CV: For patterns with code violations in data packet and optimized for activated EOM code violation criterion (and optional EOM data length criterion))
  + 0x94 (Chip Mode EOM-Data length: For patterns with code violations in data packet and optimized for activated EOM data length criterion only)
  + 0x95 (Chip Mode Transparent: When Framer is not used, but CH_DATA / CH_STR are used for data processing)

Data Slicer Bit mode:

- No code violations detectable
- Full performance
- In case of Bi-phase mark and Bi-phase space an additional bit must be sent to ensure correct decoding of the last bit
- Activation via setting register x_SLCCFG to a value of 0x75

In Data Slicer Bit mode an even number of TSI chips needs to be used.

When Data Slicer Bit mode is selected, then the the last chip of RUNIN must be different from first chip of TSI (e.g. Runin-bit sequence 000000 and TSI bit sequence 0xx...xxx is OK). Otherwise the TSI will not be detected correctly.

On using Data Slicer Bit Mode, the Wake-up criteria Equal Bits Detection and Pattern Detection cannot be applied.

A line decoder decodes the incoming data chips according to the encoding scheme (see **Chapter 2.4.8.2**).

2.4.8.5 Wake-Up Generator

A wake-up generation unit is used only in the Self Polling Mode for the detection of a predefined wake-up criterion in the received pattern.

There are two groups of configurable wake-up criteria:

- Wake-up on Level criteria
- Wake-up on Data criteria

The search for the wake-up data criterion is started if data chip synchronization has occurred within the predefined number of symbols, otherwise the wake-up search is aborted. Several different wake-up patterns, like random bit, equal bit, bit pattern or bit synchronization, are programmable.

Additional level criterion fulfilment for RSSI or Signal Recognition can lead to a fast wake-up and to a change to Run Mode Self Polling. Whenever one of these Wake-up Level criteria is enabled and exceeds a programmable threshold, a wake-up has been detected.

The Wake-up Level criterion can be used very effectively in combination with the Ultrafast Fall Back to SLEEP Mode (see **Chapter 2.6.2.3**) for further decreasing the needed active time of the autonomous receive mode. A configurable observation time for Wake-up on Level can be set in the x_WULOT register. The Wake-up on Level criterion can be handled very quickly for FSK modulation, while in case of ASK the nature of this modulation type has to be kept in mind.

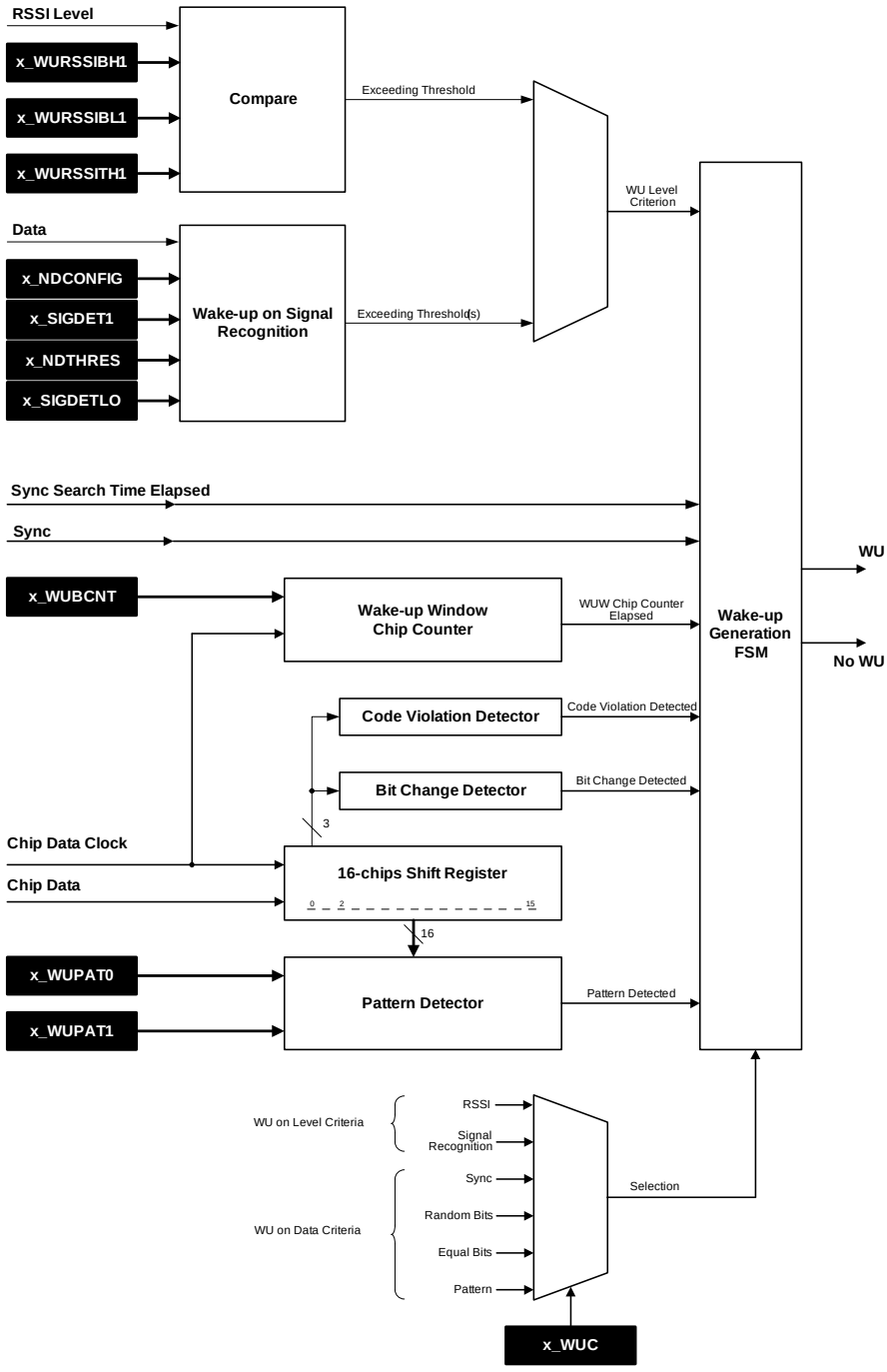


Figure 23 Wake-Up Generation Unit

Wake-Up on RSSI

The threshold x_WURSSITH1 is used to decide whether the actual signal is a wanted signal or just noise. Any kind of interfering RSSI level can be blocked by using an RSSI blocking window. This window is determined by the thresholds x_WURSSIBL1 and

x_WURSSIBH1. These two thresholds can be evaluated during normal operation of the application to handle the actual interferer environment.

The blocking window can be disabled by setting x_WURSSIBH1 to the minimum value and x_WURSSIBL1 to the maximum value.

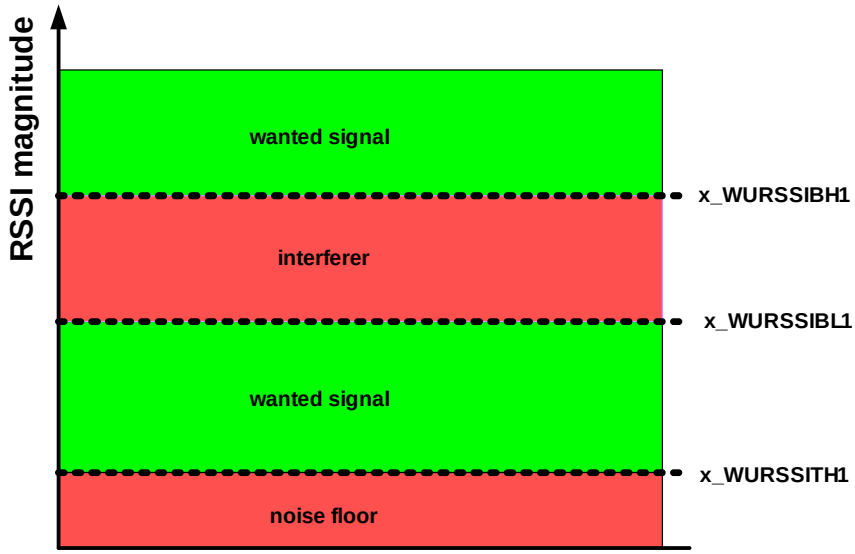


Figure 24 RSSI Blocking Thresholds

Threshold evaluation procedure

A statistical noise floor evaluation using read register RSSIPMF (RMS operation) leads to the threshold x_WURSSITH1. The interferer thresholds x_WURSSIBL1 and x_WURSSIBH1 are disabled when they are set to their default values.

For evaluation of the interferer thresholds, either use register RSSIPMF for RMS operation or during SPM and WU (Wake-Up) on RSSI use register RSSIPWU to statistically evaluate the interferer band. Finally the thresholds x_WURSSIBL1 and x_WURSSIBH1 can be set.

Wake-Up on RSSI can also be applied as additional criterion when already using a Wake-Up on Data criterion in Constant On-Off (COO) Mode.

Further details can be seen in **Figure 10**, **Chapter 2.4.7 RSSI Peak Detector**, **Chapter 2.6.2.2 Constant On-Off Time (COO)** and **Chapter 2.6.2.3 Fast Fall Back to SLEEP (FFB)**.

NOTE: If e.g. an interferer ends/starts too close after/to the beginning/end of the observation time, then a decision level error can arise. This is due to the filter dynamics (settling time). Further, for interferer thresholds evaluation in SPM this changes interferer statistics. Several interferer measurements are recommended to suppress this, what makes sense anyway for a better distribution.

Wake-Up on Signal Recognition

Instead of the previously mentioned RSSI criterion, the Signal Recognition criterion (see **Chapter 2.4.8.1**) can be applied for Wake-Up search. So the x_SIGDET1, x_SIGDETLO and x_NDTHRES threshold registers can be used.

The observation time has to be specified in the register x_WULOT. This observation time has to contain the delay in the signal path (12.5 µs + 2.25*T_{bit}) and the duration for the comparison of the Signal Recognition criterion.

The number of consecutive valid Signal Recognition samples/levels is compared vs. a threshold defined in x_WURSSIBH1 register. Please note that x_WURSSIBH1 register is used for both Wake-Up on RSSI and Wake-Up on Signal Recognition function. This threshold has an influence on the false alarm rate. So x_WURSSIBH1 defines the minimum needed consecutive T/16 samples of the Signal Recognition output to be at high level for a positive Wake-Up event generation.

Wake-Up on Data Criterion

All SFRs configuring the Wake-up Generation Unit support the Multi-Configuration capability. The search for a wake-up data criterion is started if symbol synchronization is given within a certain duration (see **Chapter 2.4.8.8 RUNIN, Synchronization Search Time and Inter-Frame Time**); otherwise the wake-up search is aborted. During the observation period, the wake-up data search is aborted immediately if symbol synchronization is lost. If this is not the case, the wake-up search will last for the number of chips/bits defined in the register x_WUBCNT.

The Wake-up Window (WUW) Chip/Bit Counter counts the number of received chips/bits and compares this number vs. the number of chips/bits defined in the register x_WUBCNT.

The Code Violation Detector checks the incoming chip data stream for being Bi-Phase coded. A Code Violation is given if three consecutive chips are ’One’ or ’Zero’.

The Bit Change Detector checks the incoming Bi-phase coded bit data stream for changes from 'Zero' to 'One' or 'One' to 'Zero'.

The Pattern Detector searches for a pattern with 16 chips/bits length within the Wake-up Window. The pattern is configurable via the registers x_WUPAT0 and x_WUPAT1.

On using Data Slicer Bit Mode, the Wake-up criteria Equal Bits Detection and Pattern Detection cannot be applied. Further details can be seen at the end of **Chapter 2.4.8.4**.

The selection of 1 out of 4 wake-up data criteria is done via the x_WUC register.

Details on the four wake-up data criteria

Pattern Detection

The incoming signal must match a dedicated pattern of up to 8 bits or 16 chips in Wake-Up Pattern Chip Mode. When the WUW chip counter elapses, the search is stopped. The higher the setting of WUBCNT the longer it is possible to search for the wake-up pattern. The minimum for the WUBCNT is 0x11!

The pattern detection is stopped either when WUW elapses, or when symbol synchronization is lost.

The Wake-Up pattern can be extended from 16 chips to 16 bits on activation of WUPMSEL bit (Wake-Up Pattern Bit Mode). In this Bit Mode no Code Violations (CV) are allowed and thus Pattern Detection is aborted, when a CV is detected.

Equal Bits Detection

Wake-up condition is fulfilled if all received bits inside of WUW are either 0 or 1. WUBCNT holds the number of required equal bits. The higher the setting of WUBCNT the lower the number of wrong wake-ups.

Equal bits detection is stopped if a bit change or a CV has been detected, or symbol synchronization is lost.

Random Bits Detection

Wake-up condition is fulfilled if there is no code violation inside of WUW. WUBCNT holds the number of required Bi-phase coded bits. The higher the setting of WUBCNT, the lower the number of wrong wake-ups.

Random bits detection is stopped if a code violation has been detected, or symbol synchronization is lost.

Valid Data Rate Detection

Wake-up condition is fulfilled if symbol synchronization is possible inside of Sync Search Time out (see **Chapter 2.4.8.8 RUNIN, Synchronization Search Time and Inter-Frame Time**). WUBCNT is not used.

This is the weakest wake-up data criterion, and should be avoided.

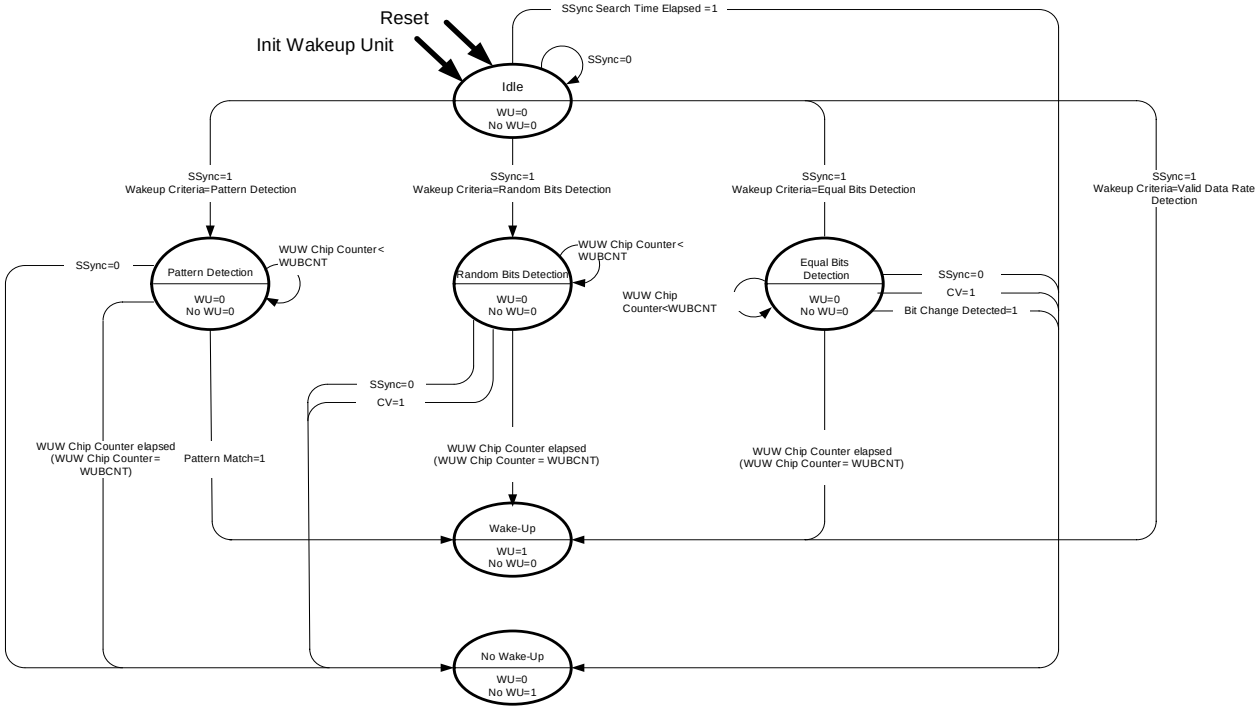


Figure 25 Wake-Up Data Criteria Search

2.4.8.6 Frame Synchronization

The Frame Synchronization Unit (Framer) synchronizes to a specific pattern to identify the exact start of a payload data frame within the data stream. This pattern is called Telegram Start Identifier (TSI).

There are different TSI modes selectable via the configuration:

- 16-Bit TSI Mode, supporting a TSI length of up to 16 bits or 32 chips
- 8-Bit Parallel TSI Mode, supporting two independent TSI pattern of up to 8 bits length each. Different payload length is possible for these two TSI pattern.
- 8-Bit Extended TSI Mode, identical to 8-Bit Parallel TSI Mode, but identifies which pattern matches by adding a single bit at the beginning of the data frame
- 8-Bit TSI Gap Mode, supporting two independent TSI pattern separated by a discontinuity

All SFRs configuring the Frame Synchronization Unit support the Multi-Configuration capability (Config A and B). The Framer starts working in Run Mode Slave after Symbol Sync found and in Self Polling Mode after wake-up found and searches for a frame until TSI is found or synchronization is lost. The input of the Framer is a sequence of Bi-phase

encoded data (chips). Basically the Framer consists of two identical correlators of 16 chips in length. It allows a Telegram Start Identifier (TSI) to be composed of Bi-phase encoded “Zeros” and “Ones”. The active length of each of the 16 chips correlators is defined independently in the x_TSILENA and x_TSILENB registers. The pattern to match is defined as a sequence of chips in the x_TSIPTA0, x_TSIPTA1, x_TSIPTB0 and x_TSIPTB1 registers.

Note that the RUNIN length shown in the figures below is the maximum needed RUNIN with the length of 8 chips. Further details on the needed RUNIN time of the receiver can be seen in **Chapter 2.4.8.3 Clock and Data Recovery**.

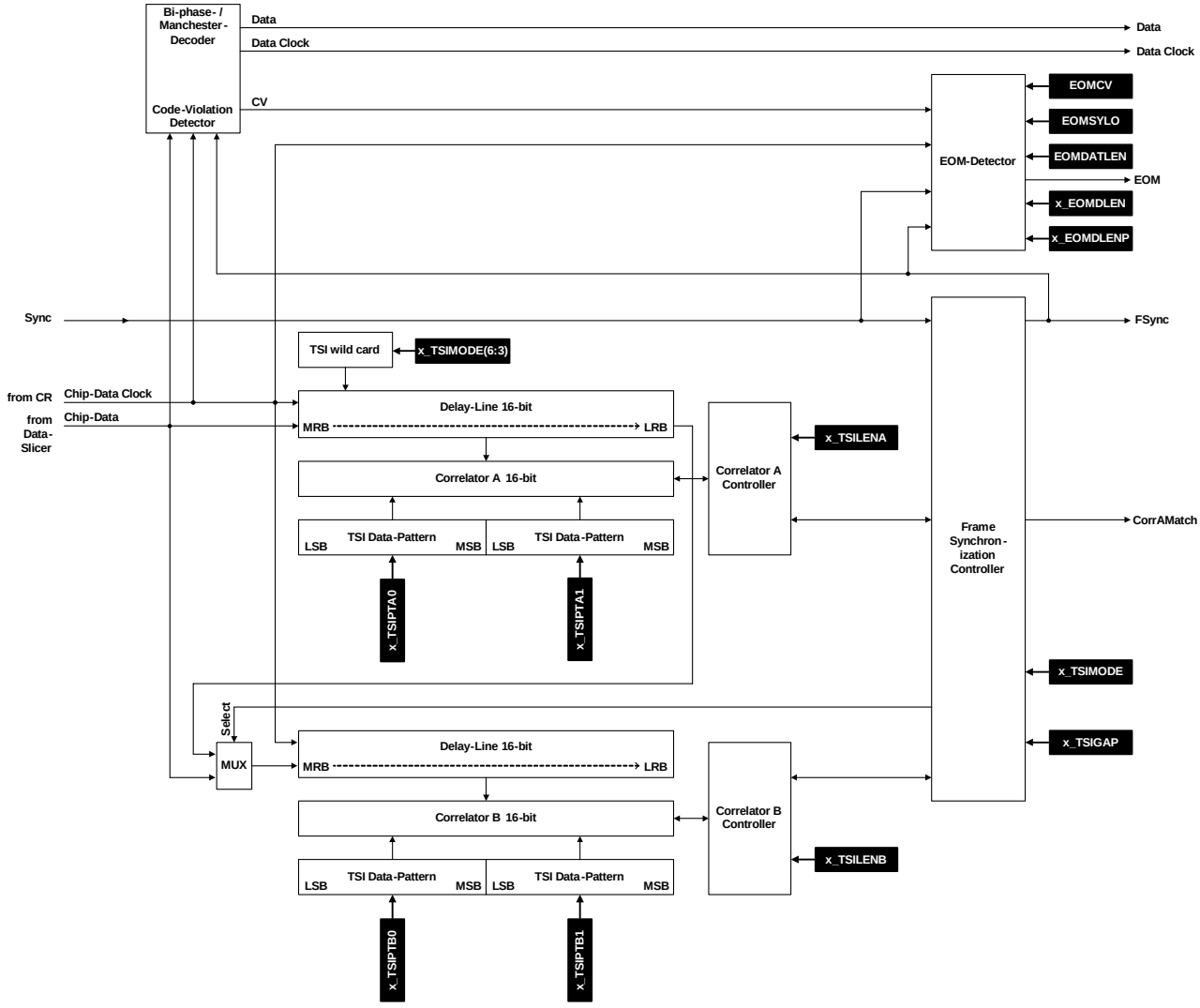


Figure 26 Frame Synchronization Unit

Please note that for Data Slicer Bit Mode a special constellation of RUNIN bits and TSI bits has to be ensured. Further details can be seen at the end of **Chapter 2.4.8.4**.

The two independent correlators can be configured in the x_TSIMODE register to work in one of the following four TSI modes:

16-Bit Mode: As a single correlator of up to 32 chips

The length of the x_TSILENA register must be set to 16d whenever x_TSILENB is higher than 0.

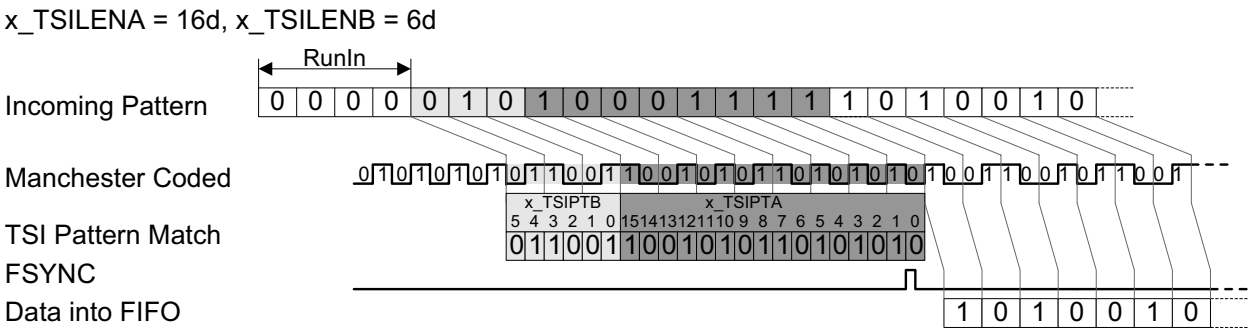


Figure 27 16-Bit TSI Mode

8-Bit Parallel Mode: As two correlators of up to 16 chips length each working simultaneously in parallel

In the following example, TSI Pattern B matches first and generates an FSYNC. The lengths of both TSI Patterns are now independent from each other. The payload length for these two TSI Pattern may be different.

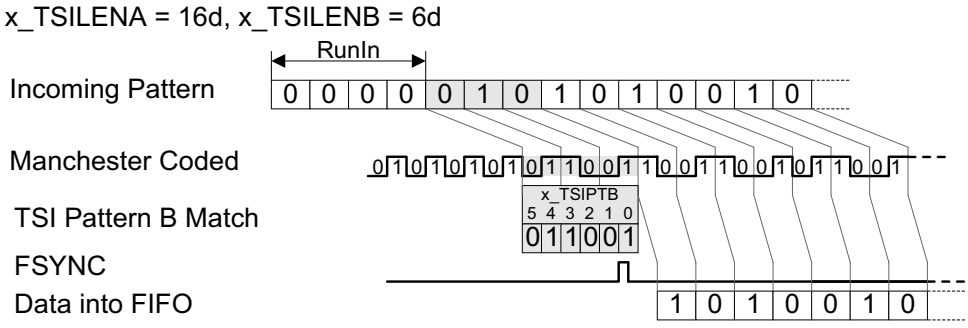


Figure 28 8-Bit Parallel TSI Mode

8-Bit Extended Mode: As two correlators of up to 16 chips length each working simultaneously in parallel, with matching information insertion

This bit is inserted at the beginning of the payload. “0” is inserted, when correlator A has matched and “1” when correlator B has matched. The payload length for these two TSI Pattern may be different.

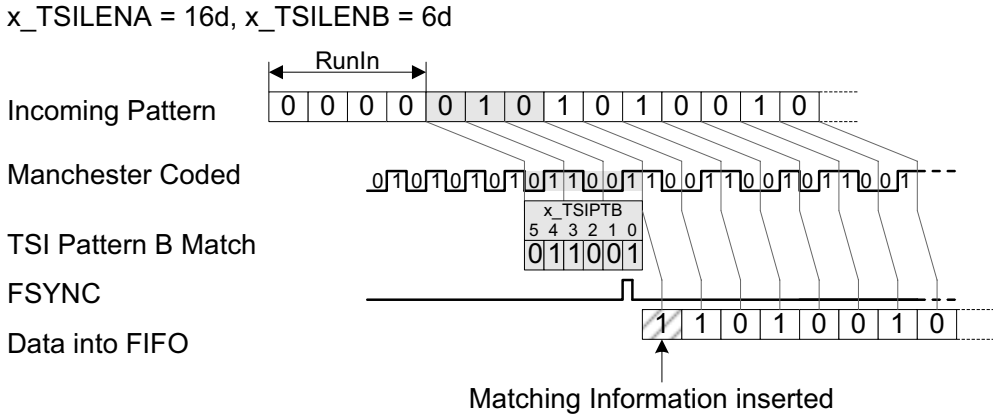


Figure 29 8-Bit Extended TSI Mode

8-Bit Gap Mode: As two sequentially working correlators of up to 16 chips length each

This mode is only used in combination with the TSI Gap Mode shown below!

This mode is used to define a gap between the two patterns which is preset in the x_TSIGAP register. To identify exactly the beginning of the gap it would be helpful on occasion to place the first CV of the gap into the TSI Pattern A. In this case, the gap length needed for the x_TSIGAP register must be shortened and the x_TVWIN length must be extended.

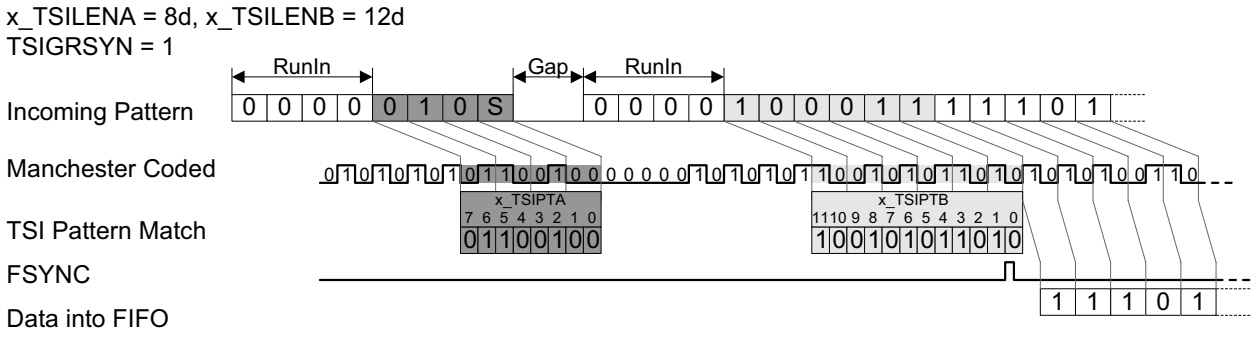


Figure 30 8-Bit Gap TSI Mode

Selection of a TSI Pattern

TSI patterns must be different to the wake-up bit stream and the RUNIN to clearly mark the start of the following payload data frame. It should be considered that the synchronization has a tolerance of about one bit. In addition, synchronization is related to data chips, and may occur in the middle of a data bit. This all must be tolerated by the data framer. Further details can be seen in **Chapter 2.4.8.3 Clock and Data Recovery**.

Ideal TSI patterns have a unique bit combination at their end, which may also contain a number of code violations (CVs), when possible (see **Chapter 2.4.8.4 Data Slicer and Line Decoding**).

Some examples of TSI patterns:

0 0 0 0 0 0 0 0 0 0 0 0 0 0 0 1

0 0 0 0 0 0 0 0 0 0 0 0 0 0 1 1

0 0 0 0 0 0 0 0 0 0 0 0 0 0 1 0

1 1 1 1 1 1 1 1 1 1 1 1 1 1 1 0

When CVs are used:

0 0 0 0 0 0 0 0 0 0 0 0 0 M 1

1 1 1 1 1 1 1 1 1 1 1 1 1 M 0

Note: CVs in a TSI are practical for better differentiation to the real data, especially if repetition of data frames is used for wake-up.

End of Message (EOM) Detection

An End Of Message (EOM) detection feature is provided by the EOM detector. Three criteria can be selected to indicate EOM.

The first is based on the number of received bits since frame synchronization. The number of expected bits is preset in the x_EOMDLEN register. Sending fewer bits as defined in the register will result in no EOM. The EOM counter will be reset after new frame synchronization.

In 8-Bit Parallel TSI Mode and 8-Bit Extended TSI Mode, the payload length for the two independent TSI pattern may be different. Therefore the payload length for TSI B pattern can be preset in the x_EOMDLENP register, while payload length for TSI A pattern can be preset in the x_EOMDLEN register.

The second criterion is the detection of a Code Violation. This EOM criterion is not applicable for Data Slicer Bit mode.

The third criterion is the loss of symbol synchronization. Depending on the x_TVWIN register, the Sync signal persists for a certain amount of time after the end of the pattern has been reached. Therefore, more bits could be written into the FIFO than sent. The three EOM criteria can be combined with each other. If one of the selected EOM criteria is fulfilled, an EOM signal will be generated.

TSI Gap Mode

The TSI Gap Mode is only used if TSI patterns contain a gap that is not synchronous to the data rate, e.g. if a gap is 7.7 data bits, or if a gap is longer than 10 data bits. In all other cases, gaps should be included in the TSI pattern as code violations.

Because of its complexity in configuration, TSI Gap Mode should be only used in applications as noted above!

For these special protocols, it is possible to lock the actual data frequency during a long Code Violation period inside a TSI (x_TSIGAP must have a minimum of 8 chips). TSIGAP is used to lock the PLL after TSI A was found. After the lock period, two different resynchronization modes are available (TSI Gap ReSYNchronization, TSIGRSYN):

- Frequency readjustment (PLL starts from the beginning), TSIGRSYN = 1. In this mode the T/2 gap resolution can be set in the 5 MSB x_TSIGAP register bits. The value in GAPVAL (3 LSB in x_TSIGAP register) is not used. This is the preferred mode in TSI Gap Mode.

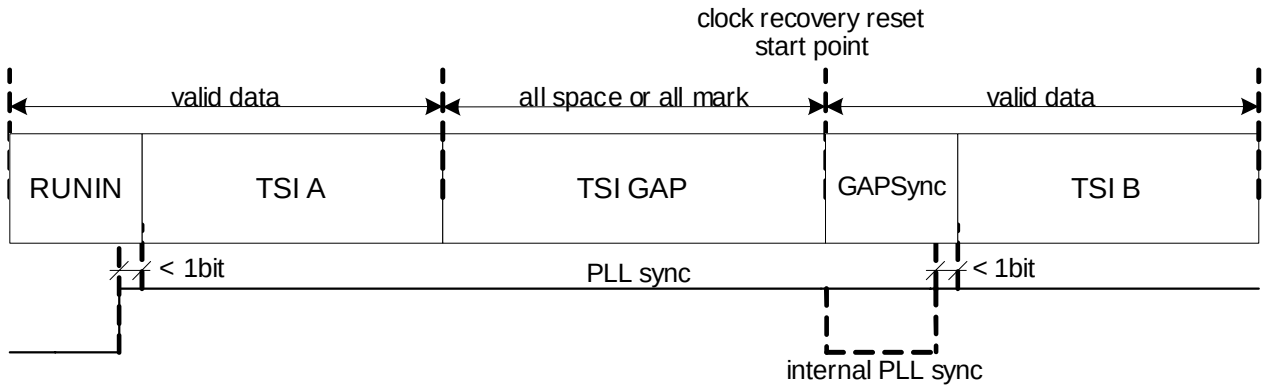


Figure 31 Clock Recovery Gap Resynchronization Mode TSIGRSYN = 1

- Phase readjustment only, TSIGRSYN = 0. In this mode, the value in GAPVAL is used to correct the phase after the gap phase. Overall gap time can be defined in T/16

steps. The 5 MSB bits (TSIGAP) define the real gap time and the 3 LSB bits (GAPVAL) the DCO (digital controlled oscillator) phase correction value.

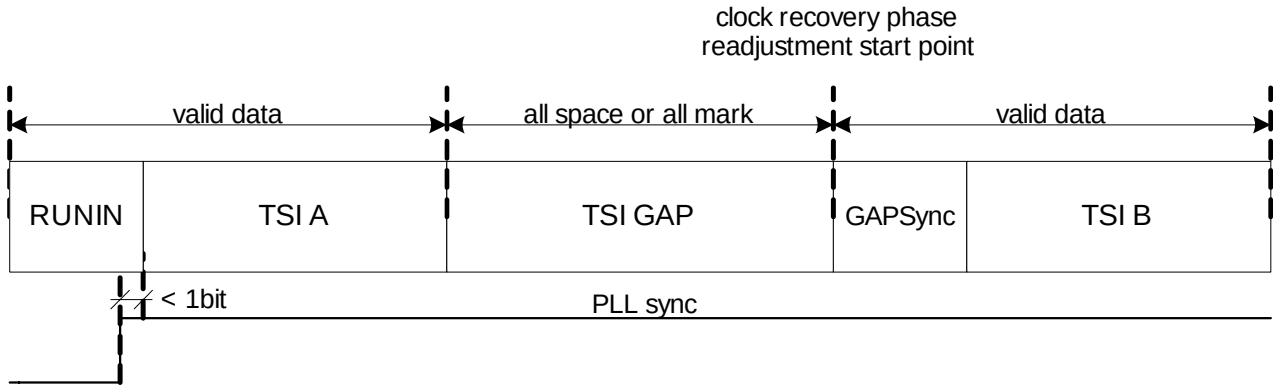


Figure 32 Clock Recovery Gap Resynchronization Mode TSIGRSYN = 0

When the time TSI GAP in the start sequence of the transmitted telegram has elapsed, the receiver needs a certain time (GAPSync = 5...6 chips) to readjust the PLL settings.

Behavior of the system at the starting position of the TSI B:

The starting position (TSI B start) for the TSI B comparison is independent from the RUNIN settings (x_CDRRI register) and the resynchronization mode (x_TSIMODE register):

$$TSIBstart[chips] = TSIGAP[chips] + 6...8$$

The incoming chips at TSI B start and the following incoming chips are compared with the contents of the register TSI B. Please notice that the receiver's PLL runs at the data rate determined before the gap. Therefore, the receiver calculates the gap based on this data rate.

Behavior of the system at the ending position of TSI B:

The system checks for the TSI B to match within a limited time. If there is no match within this time, then the receiver starts again to search for the TSI A pattern at the following incoming chips:

$$TSIBstop[chips] = TSIGAP[chips] + TSILENB[chips] + 11$$

For a successful TSI B pattern match, the defined TSI B pattern must be between "Start of TSI B" and "Stop of TSI B". In the example below, the earliest possible start position would be the 18^{th} chip and the latest possible start position would be the 22^{nd} chip.

Please note that after a gap, the internal TSI comparison register is cleared (all chips set to ’0’). In this case, a TSI B criteria of “0000” would always match at the beginning. To avoid such an unwanted matching, set the highest TSI B match chip to ’1’.

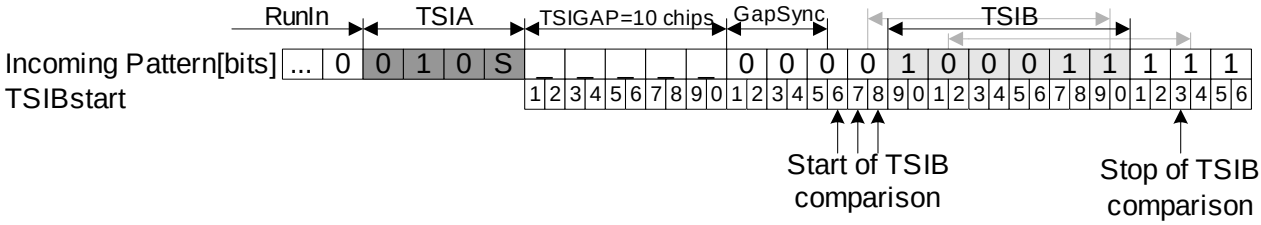


Figure 33 TSIGap TSIB Timing

The TVWIN (Timing Violation WINdow) and TSIGAP dependency is shown in **Figure 34**.

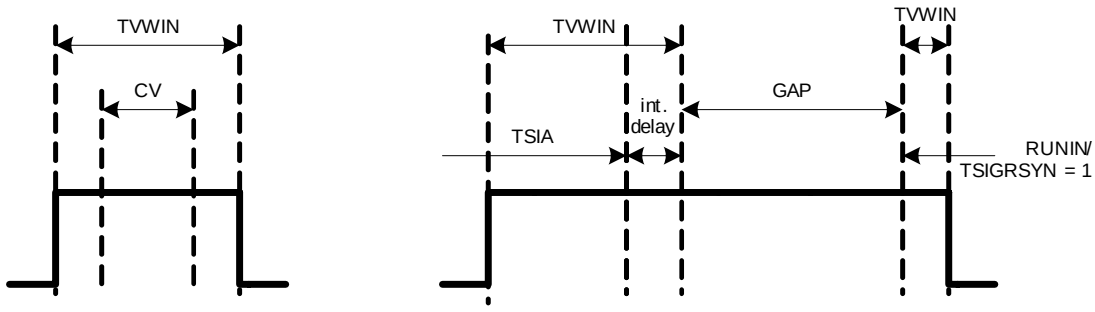


Figure 34 TVWIN and TSIGAP dependency example

TVWIN calculation for pattern without Gap time:

$$TVWIN = round((8 + 16 \cdot CV + 8) \cdot 1.25)$$

The entire TVWIN time is made up of the CV[1)] number itself, the half bit before CV and the half bit after the CV. To reach all frequency and duty cycle errors, 25% of the overall sum must be added.

TVWIN calculation with Gap time:

$$TVWIN = round(max\{((8 + 16 \cdot CV + 8) \cdot 1.25), (8 + 16 \cdot TSIA_{CV} + 16 \cdot 1 + 8) \cdot 1.25\})$$

1) CV...number of bits containing manchester code violations

2.4.8.7 Message ID Scanning

This unit is used to define an ID or special combination of bits in the payload data stream, which identifies the pattern. All SFRs configuring the Message ID Scanning Unit feature the Multi-Configuration capability. Furthermore, it is available in the Slave and Self Polling Mode. The MID Unit can be mainly configured in two modes: 4-Byte and 2-Byte organized Message ID. For each configuration there are 20 8-bit registers designed for ID storage. SFRs are used to configure the MID Unit: Enabling of the MID scanning, setting of the ID storage organization, the starting position of the comparison and number of bytes to scan.

When the Message ID Scanning Unit is activated, the incoming data stream is compared bit-wise serially with all stored IDs. If the Scan End Position is reached and all received data have matched the observed part of at least one MID the Message ID Scanning Unit indicates a successful MID scanning to the Master FSM, which generates an MID interrupt.

Please note that the default register value of the MID registers is set to 0x00. All MID registers must be set to a pattern value to avoid matching to default value 0x00.

If the MID Unit finishes ID matching without success, the data receiving is stopped and the FSM waits again for a Frame Start criterion. The received bits are still stored in the FIFO.

4-Byte Organized Message ID:

In this mode four bytes are merged to define an ID-Pattern. This does not mean that the ID must be exact four bytes long. The number of bytes used is defined in register x_MIDC1. Up to 5 ID Patterns are available.

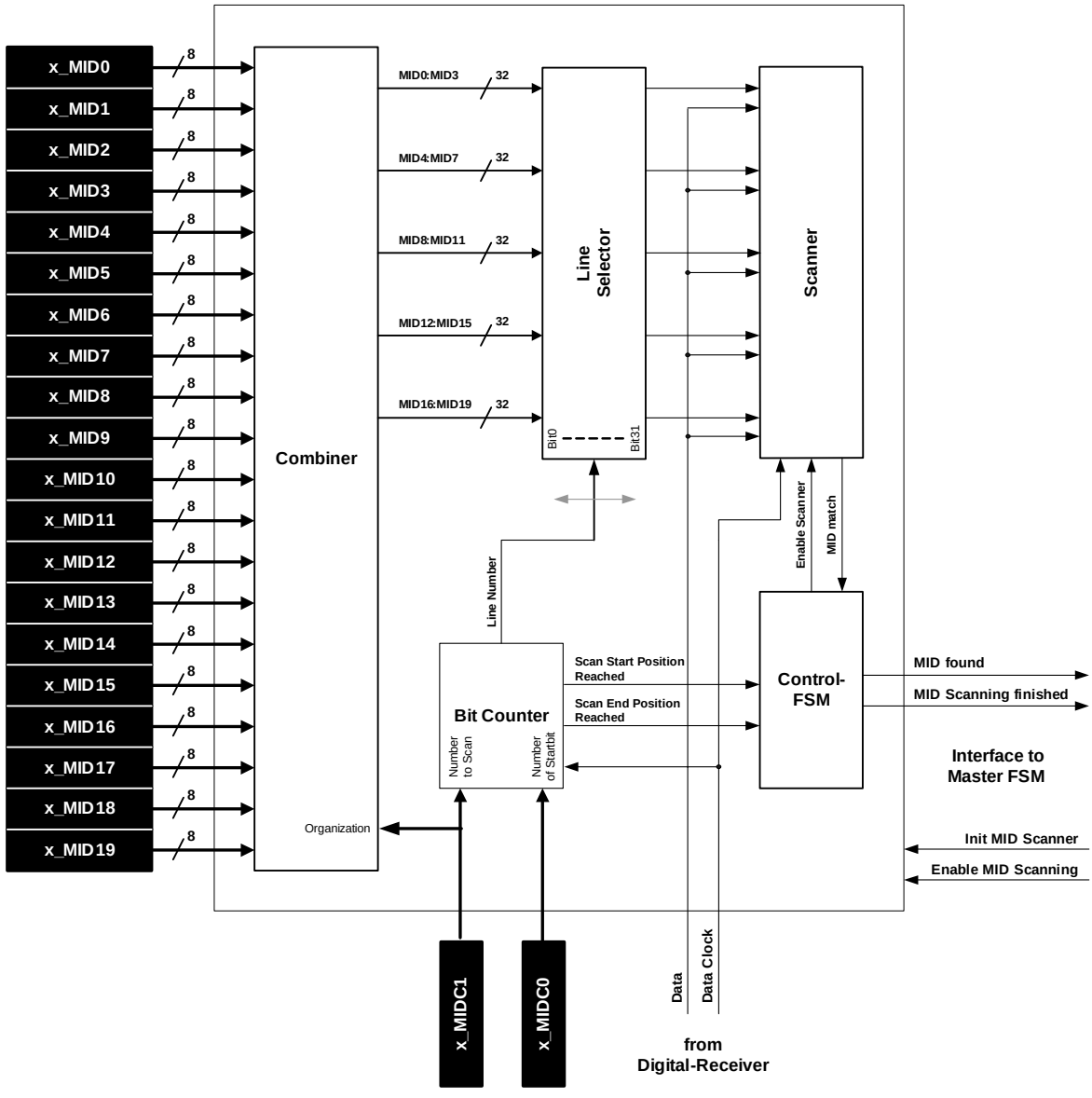


Figure 35 4-Byte Message ID Scanning

2-Byte Organized Message ID:

In this mode two bytes are merged to define an ID Pattern. Up to 10 patterns are possible.

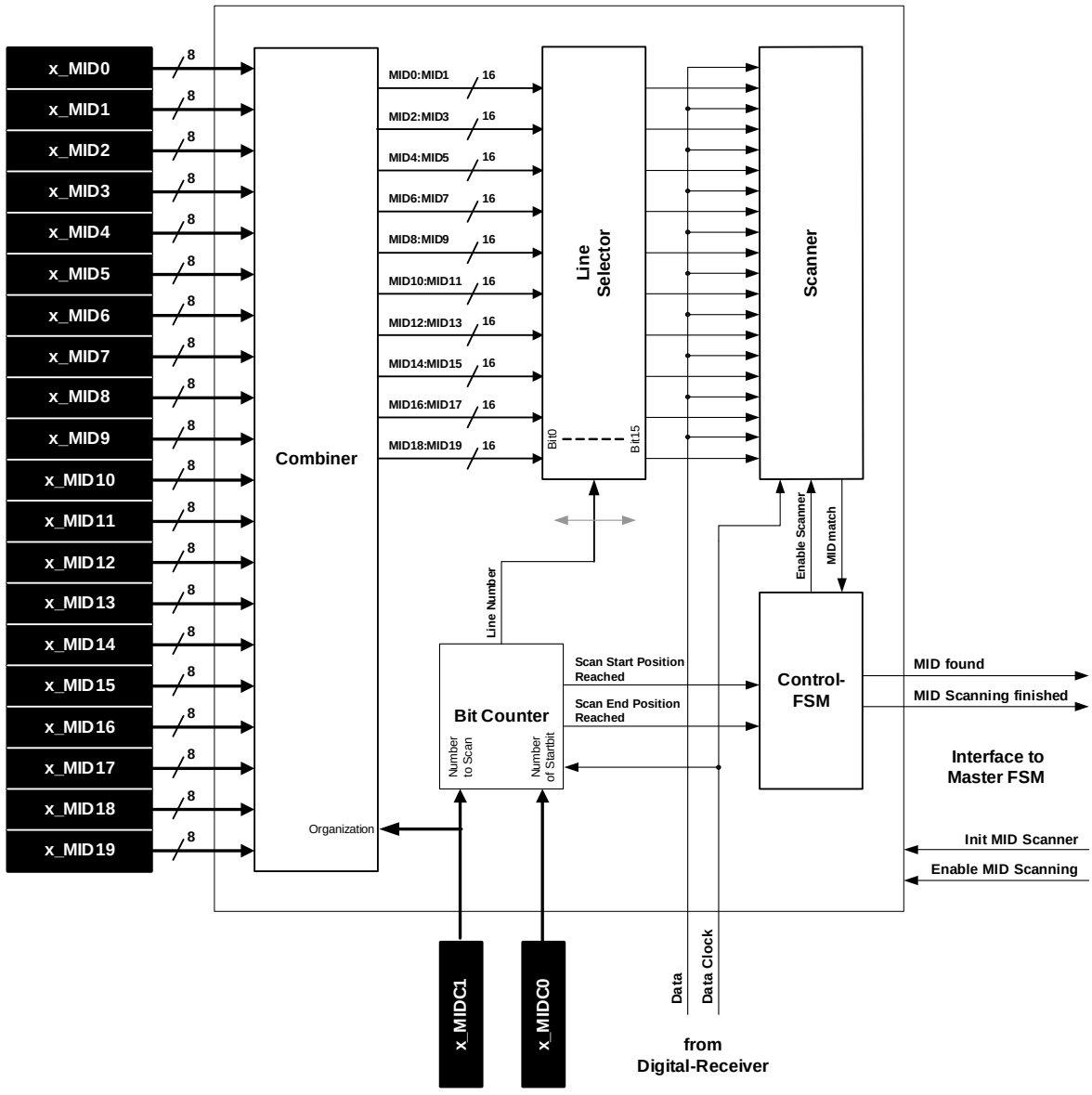


Figure 36 2-Byte Message ID Scanning

ID Position Configuration:

It is possible to choose which part of the incoming data stream is compared against the stored MIDs. The register x_MIDC0 contains the Scan Start Position. If the Bit Counter detects the Scan Start Position, the Control FSM enables the Scanner. The register x_MIDC1 contains the number of bytes to scan. During the observation period, the

Message ID Scanning is aborted immediately by the Master FSM, if symbol synchronization is lost or an EOM (End Of Message) is detected.

Example:

Start Selection: 00010001b

Number to scan: 00b, 01b, 10b, 11b

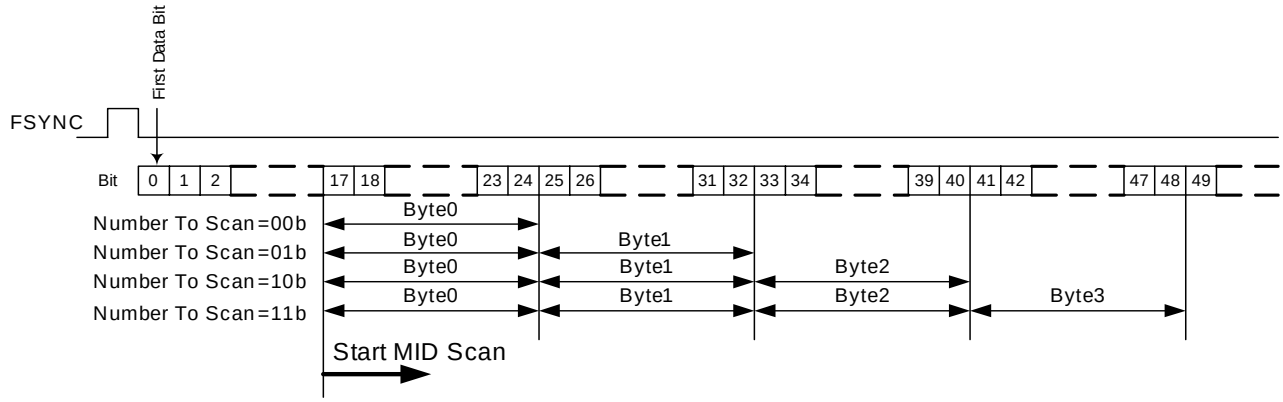


Figure 37 MID Scanning

The starting position in this case is Bit 17. Depending on the number to scan, the corresponding number of bytes is compared with the stored MIDs.

2.4.8.8 RUNIN, Synchronization Search Time and Inter-Frame Time

The functionality of the Digital Baseband Receiver is divided into four consecutive data processing stages; the data filter, clock and data recovery, data slicer and frame synchronization unit. The architecture of the Digital Baseband Receiver is optimized for processing bi-phase coded data streams.

The basic structure of a payload frame is shown in **Figure 38**. The protocol starts with a so called RUNIN. The RUNIN with the minimum length of four bi-phase coded symbols is used for internal filter settling and frequency adjustment. The TSI (Telegram Start Identifier), which is used as framing word, follows the RUNIN sequence. The payload contains the effective data. The length of the valid payload data is defined as the length itself or additional criteria (e.g. loss of Sync).

Please note that almost all transmitted protocols send a wake-up sequence before the payload frame (see also **Figure 72**). This wake-up sequence allows a very fast decision, whether there is a suitable message available or not. Further details on this topic can be gained from **Chapter 2.6.1.5** and **Chapter 2.4.8.5**.

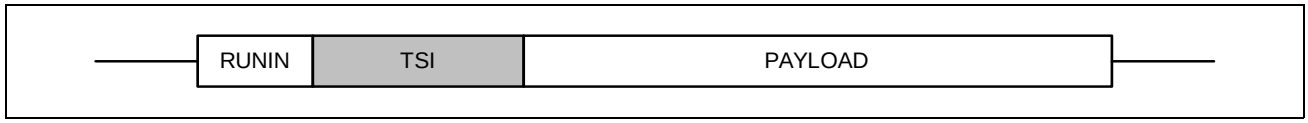


Figure 38 Structure of Payload Frame

Two important system parameters are described in this section: the **Sy**nchronization **Sea**r**ch** **T**ime **O**ut (SYSRCTO) and the Inter-Frame Time. The processing sequence of a payload frame is shown in **Figure 39**.

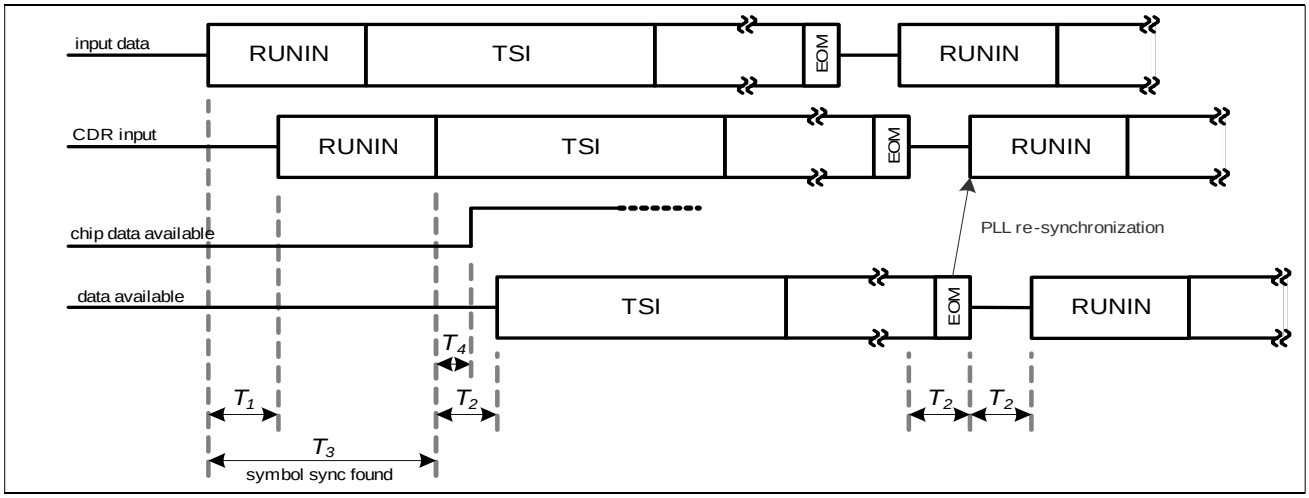


Figure 39 Data Latency

The overall system latency time is calculated in two steps: T_1 is the delay between ADC input (ASK) / limiter output (FSK) and the CDR input, and T_2 is the time between Symbol Sync Found and the Framer output (decoded data available).

T_4 is the time between Symbol Sync Found and Chip Data output (RX mode TMCDS). T_4 = 1 T. T is the nominal duration of one data bit.

T_1 latency time include: (T_1 = 12.5µs + 2 T)

- digital frontend processing delay
- matched filter computation time
- signal detector delay

T_2 latency time include: (T_2 = 1.5 T + 0.5 T[1)])

- Data Slicer computation time
- Framer computation time.

1) The 0.5 T have to be added in case of activation of Bi-phase mark / space decoding mode and Data Slicer Bit mode without Code Violation (see register x_SLCCFG)

The synchronization search time T_3 is the time the receiver requires to search for a pattern in an incoming data stream and needs to be considered in the receivers start-up phase. The minimum value of the search time out length is the consequence of the system latency time T_1, the RUNIN length and the time of asynchronism between transmitter and receiver.

This means, that for the minimum length of register value for SYSRCTO, the value 2 bits plus 12.5 µs plus the RUNIN length, which is set in the x_CDRRI register, plus 2 bits (to consider worst case RUNIN patterns and TX-RX asynchronism) have to be used. To reach data rate and duty cycle errors, 10% of the overall sum must be added.

$$SYSRCT0 = \text{roundup}\left(\frac{12.5\mu s}{T_{bit}} + 2 + RUNLEN + 2\right) \cdot 16 \cdot 1.1$$

A second important system parameter that must be considered, is the minimal Inter-Frame Time (time between two data frames). This time is equal to the time T_2 and has a length of 1.5 or 2 bits[1)]. The EOM to PLL resynchronization time is negligible in case INITDRXES is disabled. Otherwise T_1 has to be added.[2)]

Note that the described Inter-Frame Time is based on the input pattern with equal signal power in the following data frame; in other cases, the Inter-Frame Time can vary from the calculated value.

$$T_{Inter-Frame} = 1.5T_{bit} + \begin{matrix} 0.5T_{bit}^{1)} \\ 0 \end{matrix} + \begin{matrix} T_1^{2)} \\ 0 \end{matrix}$$

1) see previous footnote

2) in case INITDRXES is enabled

2.4.9 Power Supply Circuitry

The chip may be operated within a 5 Volts or a 3.3 Volts environment.

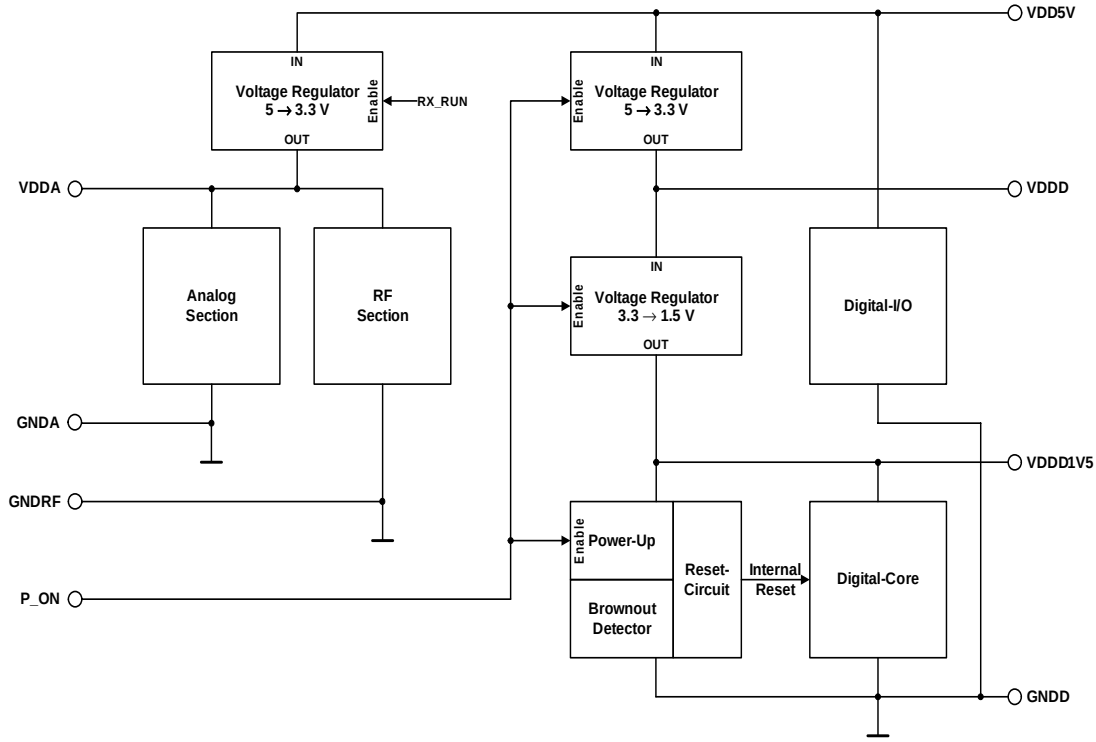


Figure 40 Power Supply

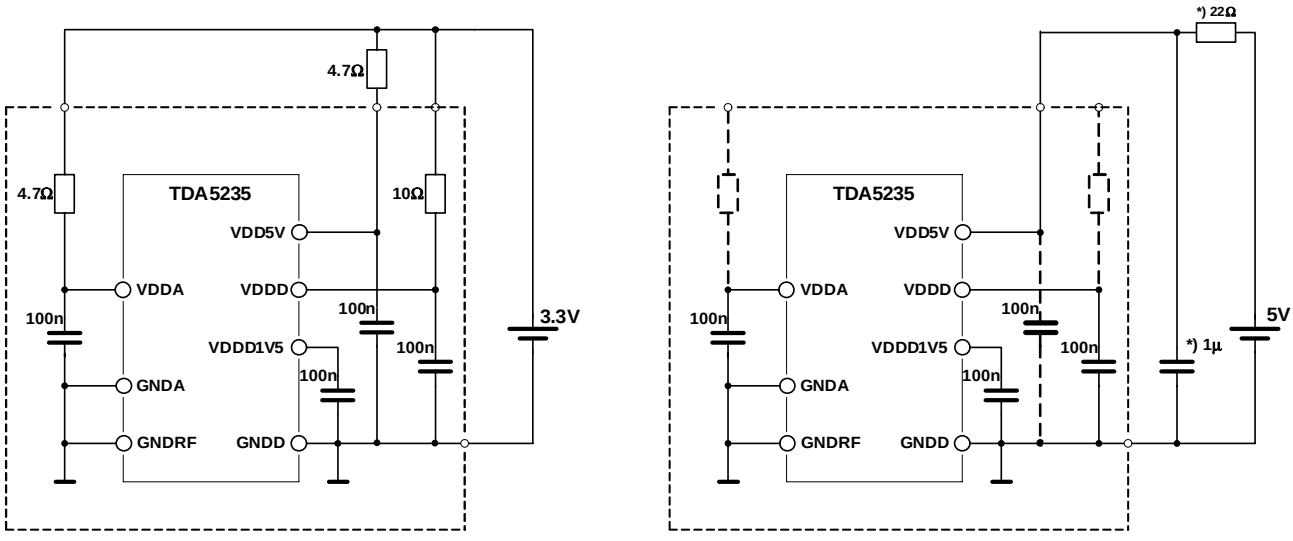
For operation within a 5 Volts environment (supply voltage range 1), the chip is supplied via the VDD5V pin. In this configuration the digital I/O pads are supplied via VDD5V and a 5 V to 3.3 V voltage regulator supplies the analog/RF section (only active in Run Modes).
When operating within a 3.3 Volts environment (supply voltage range 2), the VDD5V, VDDA and VDDD pins must be supplied. The 5 V to 3.3 V voltage regulators are inactive in this configuration.

The internal digital core is supplied by an additional 3.3 V to 1.5 V regulator.

The regulators for the digital section are controlled by the signal at P_ON (Power On) pin. A low signal at P_ON disables all regulators and set the IC in Power Down Mode. A low to high transition at P_ON enables the regulators for the digital section and initiates a power on reset. The regulator for the analog section is controlled by the Master Control Unit and is active only when the RF section is active.

To provide data integrity within the digital units, a brownout detector monitors the digital supply. In case a voltage drop of VDDD below approximately 2.45 V is detected a RESET will be initiated.

A typical power supply application for a 3.3 Volts and a 5 Volts environment is shown in the figure below.



Supply-Application in 3.3V environment

Supply-Application in 5V environment

*) When operating in a 5V environment, the voltage-drop across the voltage regulators 5 → 3.3V has to be limited, to keep the regulators in a safe operating range. Resistive or capacitive loads (in excess to the scheme shown above) on pins VDDA and VDDD are not recommended.

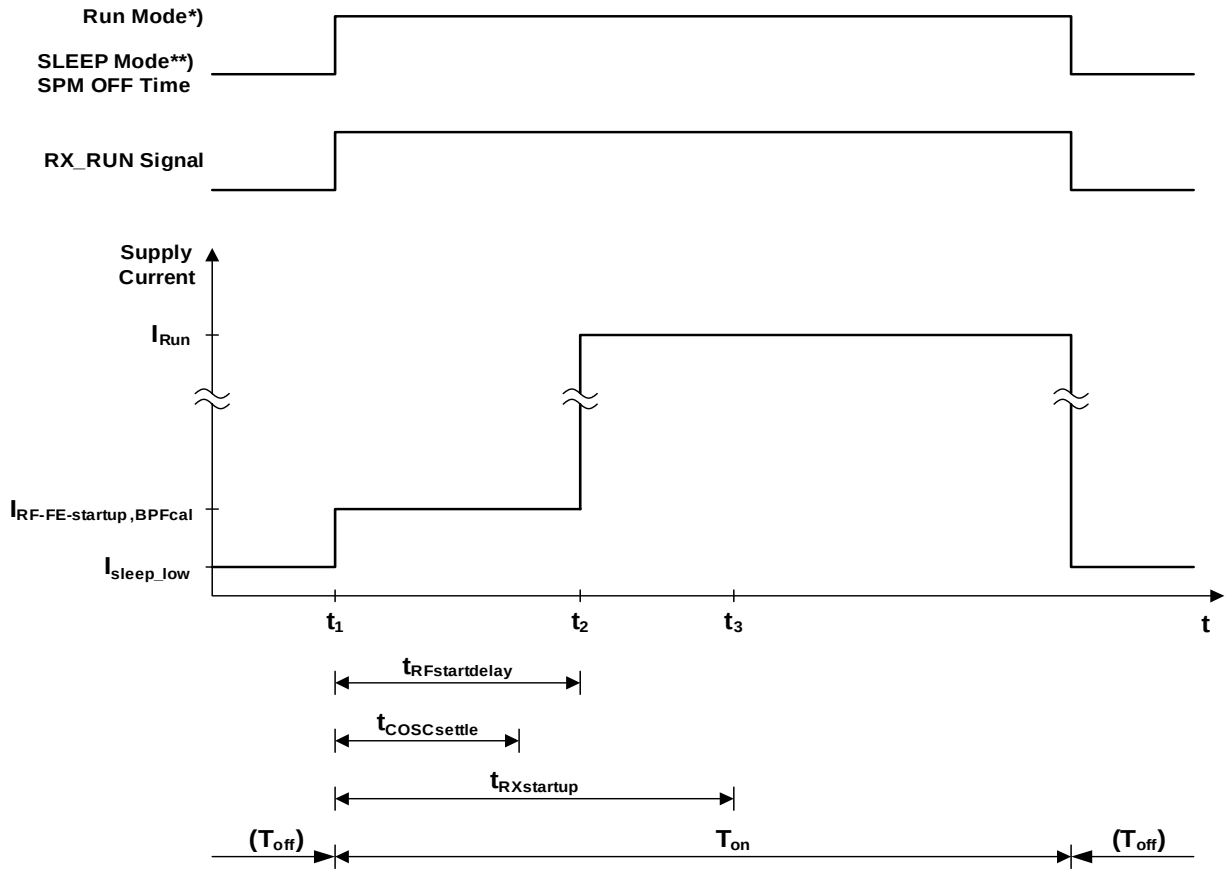
Figure 41 3.3 Volts and 5 Volts Applications

2.4.9.1 Supply Current

In SLEEP Mode, the Master Control Unit switches the crystal oscillator into Low Power Mode (all internal load capacitors are disconnected) to minimize power consumption. This is also valid for Self Polling Mode during Off time (SPM_OFF).

Whenever the chip leaves the SLEEP Mode/SPM_OFF (t_1), the crystal oscillator resumes operation in High Precision Mode and requires $t_{COSCsettle}$ to settle at the trimmed frequency. At t_2 the analog signal path (RF and IF section) and the RF PLL are activated. At t_3 the chip is ready to receive data. The chip requires $t_{RXstartup}$ when leaving SLEEP Mode/SPM_OFF until the receiver is ready to receive data.

A transient supply current peak may occur at t_1, depending on the selected trimming capacitance. The average supply current drawn during $t_{RFstartdelay}$ is $I_{RF\text{-}FE\text{-}startup,BPFcal}$.



*) Run Mode covers the global chip states Run Mode Slave/ Receiver active in Self Polling Mode/ Run Mode Self Polling
**) I_{sleep_low} is valid in the chip states SLEEP / Off time during Self Polling Mode

Figure 42 Supply Current Ramp Up/Down

If the IF buffer amplifier or the clock generation feature (PPx pin active) are enabled, the respective currents must be added.

2.4.9.2 Chip Reset

Power down and power on are controlled by the P_ON pin. A LOW at this pin keeps the IC in Power Down Mode. All voltage regulators and the internal biasing are switched off. A high transition at P_ON pin activates the appropriate voltage regulators and the internal biasing of the chip. A power up reset is generated at the same time.

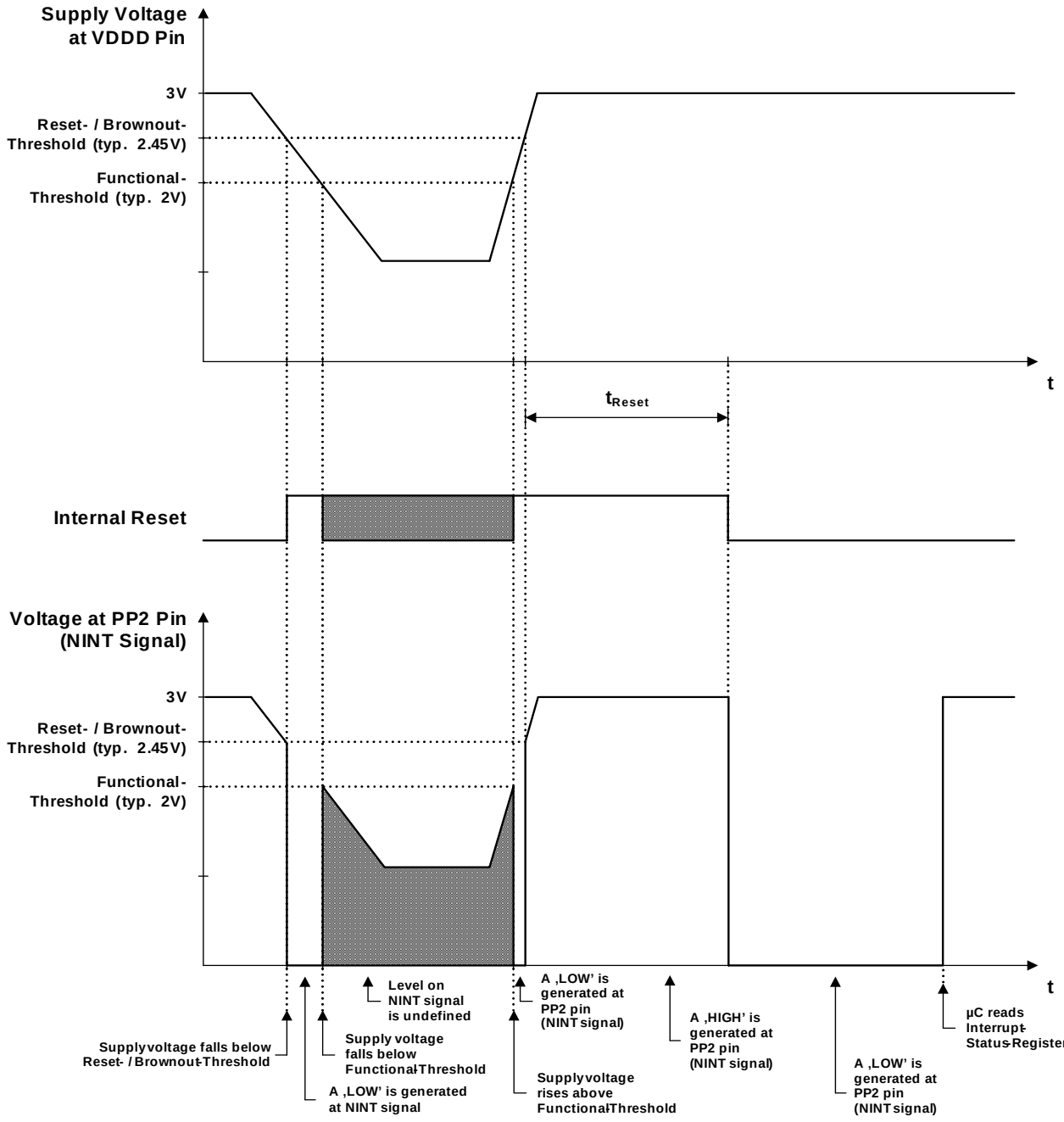


Figure 43 Reset Behavior

A second source that can trigger a reset is a brownout event. Whenever the integrated brownout detector measures a voltage drop below the brownout threshold on the digital

supply, the integrity of the stored data and configuration can no longer be guaranteed; thus a reset is generated. While the supply voltage stays between the brownout and the functional threshold of the chip, the NINT signal is forced to low. When the supply voltage drops below the functional threshold, the levels of all digital output pins are undefined.

When the supply voltage raises above the brownout threshold, the IC generates a high pulse at NINT and remains in the reset state for the duration of the reset time. When the IC leaves the reset state, the Interrupt Status register IS0 is set to 0xFF and the NINT signal is forced to low. Now, the IC starts operation in the SLEEP Mode, ready to receive commands via the SPI interface. The NINT signal will go high, when the Interrupt Status register is read for the first time.

2.5 System Interface

In most applications, the TDA5235 receiver IC is attached to an external microcontroller. This so-called Application Controller executes a firmware which governs the TDA5235 by reading data from the receiver when data has been received on the RF channel and by configuring the receiver device. The TDA5235 features an easy to use System Interface, which is described in this chapter.

Transparent Mode

The TDA5235 supports two levels of integration. In the most elementary fashion, it provides a rather rudimentary interface by which the incoming RF signal is demodulated and the corresponding data is made available to the Application Controller. Optionally, a chip clock is generated by the TDA5235. Since the data signal is always directly the baseband representation of the RF signal, we call this mode the Transparent Mode. The usage of the Transparent Mode will be described in **Chapter 2.5.1.2**.

Packet Oriented Mode

Alternatively, the TDA5235 features the so-called Packet Oriented Mode which supports the autonomous reception of data telegrams. The Packet Oriented Mode provides a high-level System Interface which greatly simplifies the integration of the receiver in data-centric applications. In Packet Oriented Mode, the data interface is based on chunks of synchronous data which are received in packets. In the easiest way, the Application Controller only reacts on the synchronous data it receives. The receiver autonomously handles the line decoding and the deframing of these data, and supports the timed reception of packets. Data is buffered in a receive FIFO and can be read out via the data interface. Further, the receiver provides support for the identification of wake-up signals. Details on the usage of the Packet Oriented Mode of the receiver are given in **Chapter 2.5.1.2**.

2.5.1 Interfacing to the TDA5235

The TDA5235 is interfacing with an application by three logical interfaces, see **Figure 44**. The RF/IF interface handles the reception of RF signals and is responsible for the demodulation. Its physical implementation has been described in **Chapter 2.4.3** and **Chapter 2.4.8**, respectively. The other two logical interfaces establish the connection to the Application Controller. Note that due to the high level of integration of the receiver, these interfaces impose minor requirements on the Application Controller, which can be as simple as an 8-bit microcontroller operated at low clock rate. As will be shown later, the physical implementation of the data interface depends on whether the receiver is operated in Packet Oriented or in Transparent Mode.

For the sake of clarity, the communication between the TDA5235 and the Application Controller is split into **control flow** and **data flow**. This separation leads to an independent definition of the data interface and the control interface, respectively.

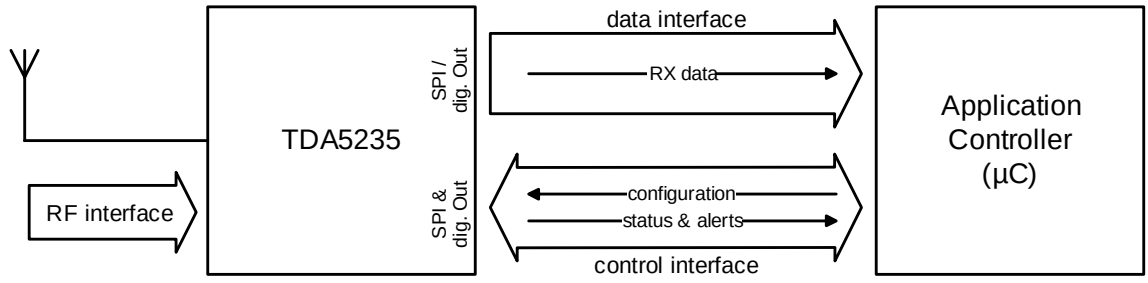


Figure 44 Logical and electrical System Interfaces of the TDA5235

2.5.1.1 Control Interface

The control interface is used in order to configure the TDA5235 after start-up or to re-configure it during run-time, as well as to properly react on changes in the status of the receiver in the Application Controller's firmware. The control interface offers a bi-directional communication link by which

- **configuration data** is sent from the Application Controller to the TDA5235,
- the receiver provides **status information** (e.g. the status of a data reception) as response to a request it has received from the Application Controller, and
- the TDA5235 autonomously **alerts** the Application Controller that a certain, configurable event has occurred (e.g. that a packet has been received successfully).

Configuration and status information are sent via the 4-wire SPI interface as described in **Chapter 2.5.5**. The configuration data determines the behavior of the receiver, which comprises

- scheduling the inactive power-saving phases as well as the active receive phases,
- selecting the properties of the RF/IF interface configuration (e.g. carrier frequency selection, filter settings),
- configuring the properties of the frames (e.g. wake-up patterns, Telegram Start Identifier (TSI), and optionally specifying the position, format and content of patterns within packets that stimulate a certain, configurable alerting behavior (Message ID)).

Note that the TDA5235 receiver IC supports reception of up to two configuration sets on a single channel each in a time-based manner without reconfiguration. Thus, the RF/IF interface as well as the frame format properties support alternative settings, which can be activated autonomously by the receiver as part of the scheduling process.

In contrast to the high-level interface used for communicating configuration instructions and status information, alerts are emitted by the receiver on a digital output pin that may trigger external interrupts in the Application Controller. Note that the alerting conditions as well as the polarity of the output pin are configurable, see **Chapter 2.5.4**.

2.5.1.2 Data Interface

The data interface between the Application Controller and the TDA5235 receiver IC is used for the transport of the received data, see **Figure 44**. The physical implementation as well as the features of the data interface depend on the selected mode of operation.

There are 5 possible receive modes:

- Packet Oriented FIFO Mode (POF)
- Packet Oriented Transparent Payload Mode (POTP)
- Transparent Mode - Chip Data and Strobe (TMCDS)
- Transparent Mode - Matched Filter (TMMF)
- Transparent Mode - Raw Data Slicer (TMRDS)

Access points for these receive modes can be seen in **Figure 15**.

The possible combinations of receive modes and polling mode setup is noted in **Figure 45**.

Self Polling Mode / RX Mode - available signal	Const ON-OFF					Fast Fall Back (UFFB), Mixed, PWUS			
	WU on Level criterion		WU on data criterion			WU on Level criterion		WU on data criterion	
	RSSI	Signal Recognition	Sync	Random, Equal, Pattern		RSSI	Signal Recognition	Sync	Random, Equal, Pattern
POF - FIFO	√	√	√	√		√	√	√	√
POTP - RXD - RXSTR	√	√	√	√		√	√	√	√
TMCDS - CH_DATA - CH_STR	√	√	√	-		√	√	√	-
TMMF - DATA_MATCHFIL	√	-	-	-		-	-	-	-
TMRDS - DATA	√	-	-	-		-	-	-	-

Legend:

√ … available

- … not available

Figure 45 Receive Modes

Packet Oriented FIFO Mode (POF)

In Packet Oriented FIFO Mode, data is transferred via the 4-wire SPI bus. During receive operation, the incoming RF signal is demodulated in the RF/IF interface, the line decoding is performed and the data, of which wake-up frames, data frame headers and optional footers have been stripped off, is stored in the RX FIFO. Then, the received data can be read from the RX FIFO using the “read FIFO” command described in

Chapter 2.5.2 and Chapter 2.5.5. The data which is read from the RX FIFO is accompanied by information which contains the status of the respective receive operation. Note that the availability of received data packets is communicated via alerts in the control interface.

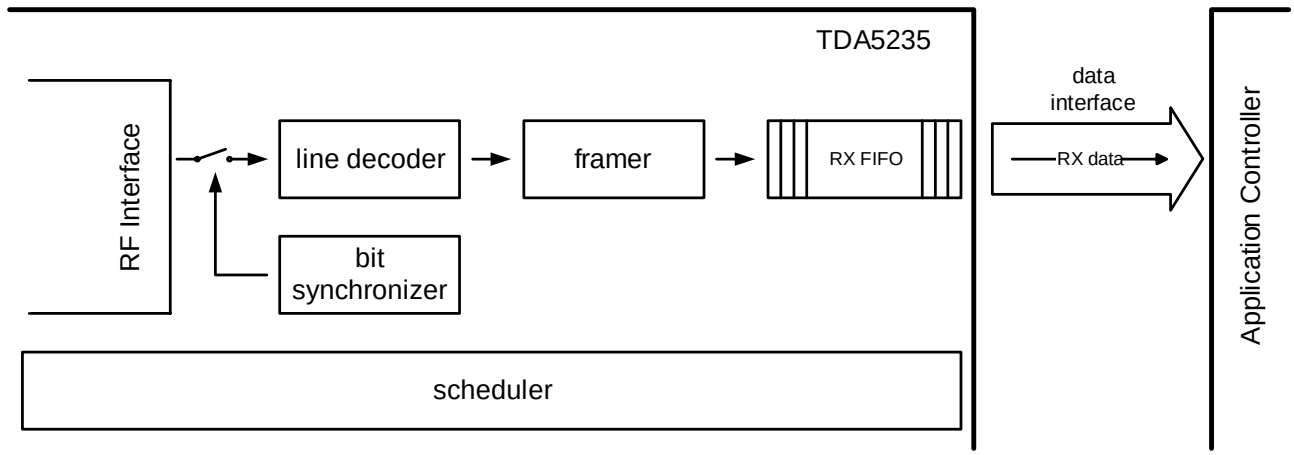


Figure 46 Data interface for the Packet Oriented FIFO Mode

Packet Oriented Transparent Payload Mode (POTP)

This mode is very similar to POF Mode as data which is going into FIFO is also available via RXD and RXSTR signals (see Chapter 2.5.3 Digital Output Pins).

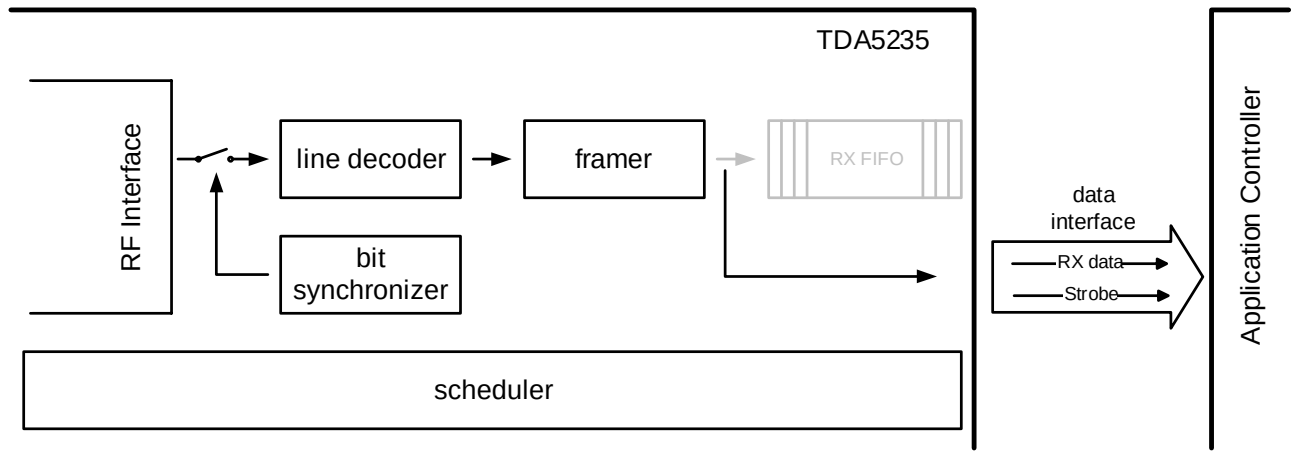


Figure 47 Data interface for the Packet Oriented Transparent Payload Mode

In the TDA5235, there are specific digital output lines (PPx pin) for the Bi-phase decoded data and an appropriate Strobe signal. During inactivity of the receiver, the line is in default mode switched to low.

In default mode the Strobe signal is active high and has a delay of $T_{BIT}/16$ relative to the data bit and a duration of $T_{BIT}/2$. The polarity of the Strobe signal is programmable, this can be done via PPCFG2 register.

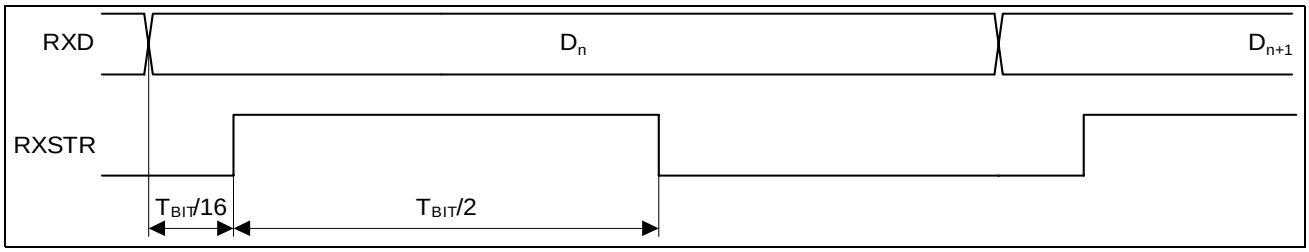


Figure 48 Timing of the Packet Oriented Transparent Payload Mode

Transparent Mode - Chip Data and Strobe (TMCDS)

The receiver's simple plain data interface in this Transparent Mode is shown in **Figure 49**. In this mode, the demodulated data signal is made directly available on the data output pin of the data interface. Concurrently, an estimate of the chip clock is optionally provided on the respective clock output line. Note that a sensible chip clock can only be generated if the selected line encoding exhibits a constant chip rate. The chip clock generation can be significantly improved by using a run-in signal of alternating one-zero chips (maximum number of transitions within a data stream).

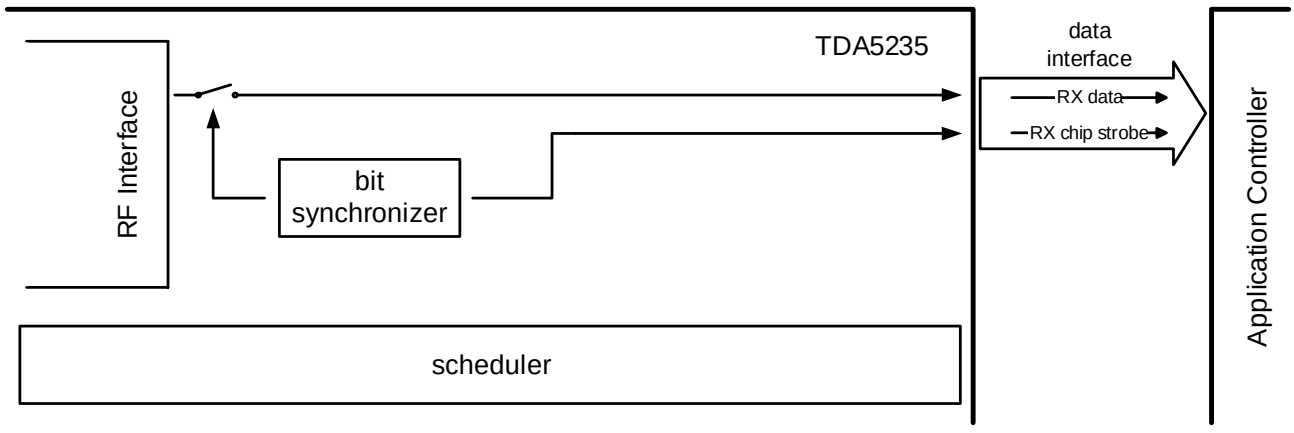


Figure 49 Data interface for the Transparent Mode - Chip Data and Strobe

In the TDA5235, there is a specific digital output line for the chip clock estimate as well as for the data output line, which delivers the encoded chip data. During inactivity of the receiver, the line is in default mode switched to low.

The PPx pin provides the estimated chip clock, if CH_STR is selected. Further details are given in **Chapter 2.5.3**.

In default mode the CH_STR signal is active high and has a delay of $T_{CHIP}/8$ relative to the data chip and a duration of $T_{CHIP}/2$. The polarity of the CH_STR signal is programmable, this can be done via PPCFG2 register.

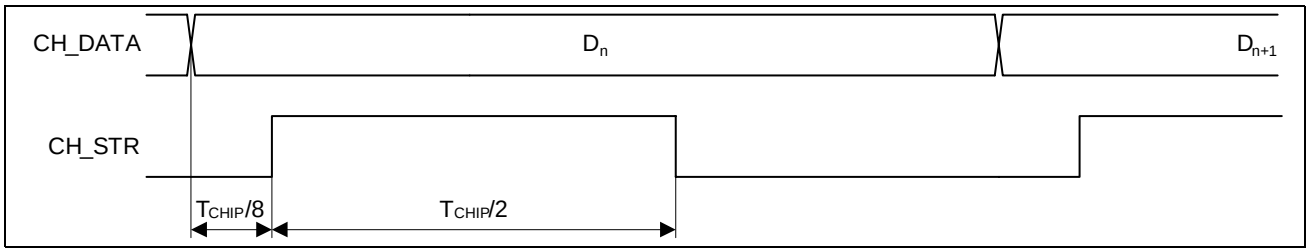


Figure 50 Timing of the Transparent Mode - Chip Data and Strobe

Transparent Mode - Matched Filter (TMMF)

The received data after the Matched Filter (Two-Chip Matched Filter) with an additional SIGN function is provided via the DATA_MATCHFIL signal (PPx pin). In this mode sensitivity measurements with ideal data clock can be performed very simple. For further details see the block diagram in **Figure 15**.

Sensitivity in this transparent mode is significantly depending on the implemented clock and data recovery algorithm of the user software in the application controller.

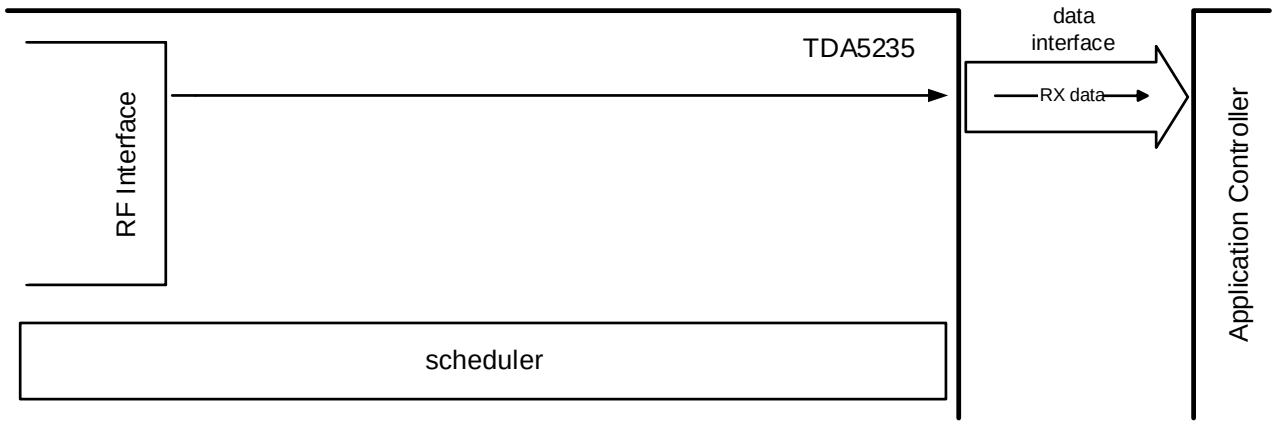


Figure 51 Data interface for the Transparent Modes TMMF / TMRDS

Transparent Mode - Raw Data Slicer (TMRDS)

This mode supports processing of data even without bi-phase encoding (e.g. NRZ coding) by providing the received data via the One-Chip Matched Filter on the DATA signal (PPx pin). See more details in the block diagram in **Figure 15**.

Sensitivity in this transparent mode is significantly depending on the implemented clock and data recovery algorithm of the user software in the application controller.

The data interface can be seen from **Figure 51**.

Self Polling capabilities are possible as well, but only Constant On-Off Mode and Wake-up on RSSI makes sense. Assume one of the TDA5235 configurations (e.g. Configuration B) is set for external data processing mode. See also example in **Figure 52**. The needed On time (latency through TDA5235) is configured in the corresponding On time registers of the chip. The interrupt for Wake-Up Config B (WUB) is enabled and suitable RSSI thresholds are set.

If the RSSI signal is in a valid threshold area, the TDA5235 changes to Run Mode Self Polling and an interrupt can be signaled to the Application Controller.

In case the RSSI signal is outside the valid threshold area, the chip stays in Self Polling Mode and the external controller gets no interrupt (as the desired RSSI level is not reached).

It should be mentioned that all Timeout Timers (TOTIMs) should be disabled in the configuration set of the external processing mode as the microcontroller takes over the control (see SFR bit group EXTPROC in the x_CHCFG register).

It is recommended to put this external configuration at the end of the On time within the polling cycle (so right before the Off time). This is helpful when using the "EXTTOTIM" command (goto Self Polling Mode, next configuration or Configuration A; see **Figure 77**). When the external configuration is the last configuration before the Off time, then the next configuration within the polling cycle would be the sequence of the Off time.

When data is available and the RSSI is within a valid threshold area, an interrupt is generated (NINT). So the Application Controller can process the data and decide about valid data.

In case the controller decides that wrong data was sent, the microcontroller can send the register command "EXTTOTIM" (see **Figure 77** and EXTPCMD register).

When the microcontroller detects valid data, then the controller can send the register command "EXTEOM found" (see **Figure 77** and EXTPCMD register) after completing the data reception.

The functionality described above can also be used for other receive modes (mainly TMMF, TMCDS), where the external microcontroller takes on responsibility for further data processing.

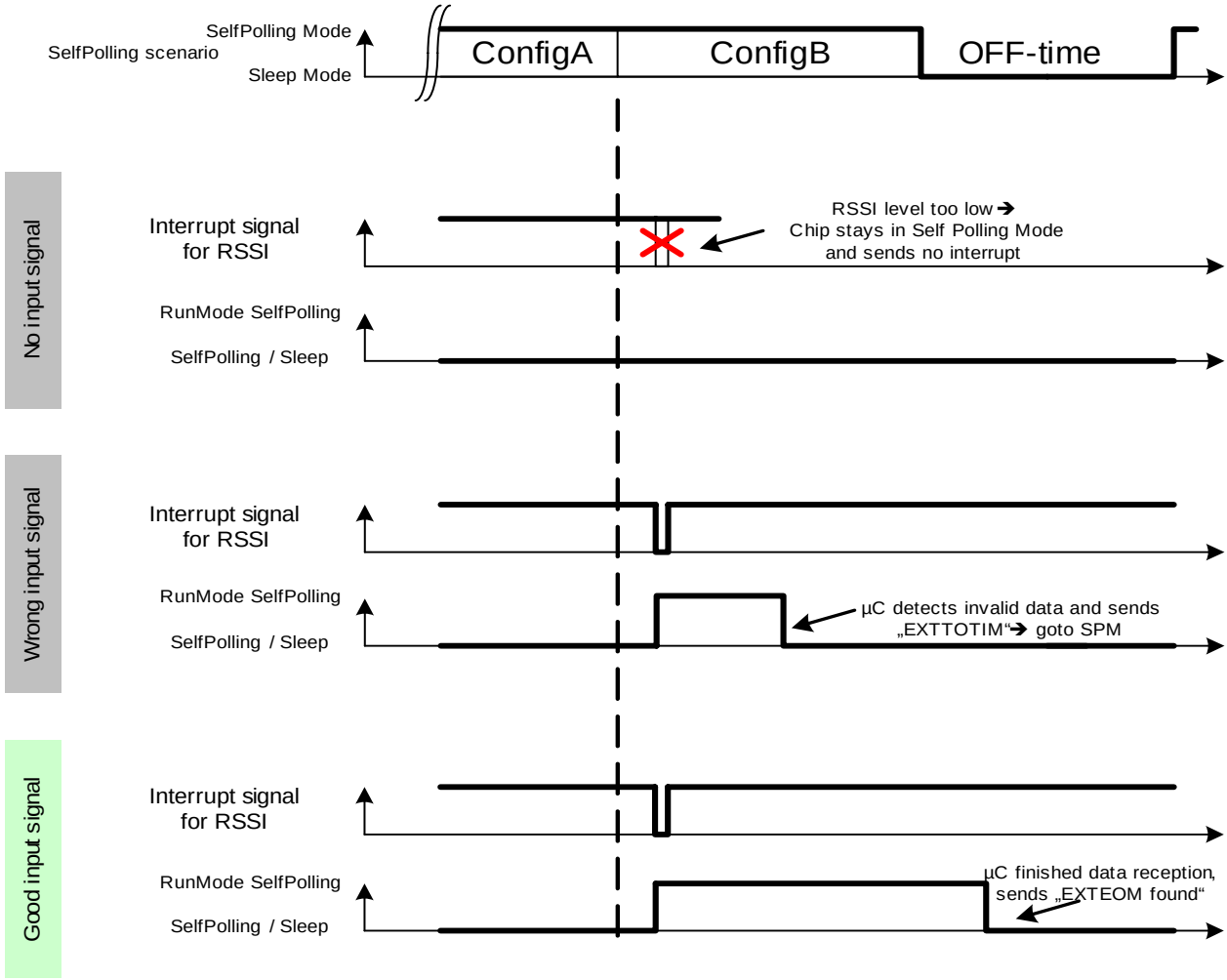


Figure 52 External Data Processing

The SFR bit group EXTPROC in the x_CHCFG register can be activated for each configuration set for an easier handling of external data processing by the Application Controller. Depending on the intended transparent receive mode an activation of this function means:

- Data path in front of Framer Unit is no longer closed (so that no data is going into Framer Unit accidentally)
- Interrupts for FSync, MID and EOM are deactivated internally
- Some/all TOTIM counters are deactivated
- Some/all Wake-up on Data Criteria are disabled
- Wake-up on Signal Recognition is/is not disabled

2.5.2 Receive FIFO

The Receive FIFO is the storage of the received data frames and is only used in the POF Mode. It is written during data reception. The host microcontroller is able to start reading via SPI right after frame sync (interrupt) or in the most common case right after detection of EOM (interrupt). The FIFO can store up to 256 received data bits. If the expected data transmission contains more bits (note that in TSI 8-bit Extended Mode one bit is added in front of the real payload to indicate which of the two TSI pattern has matched), reading from FIFO must start a certain time after frame sync to prevent an overrun.

Architecture

The 256-bit receive FIFO is based on a bit-addressable 2-port memory architecture.

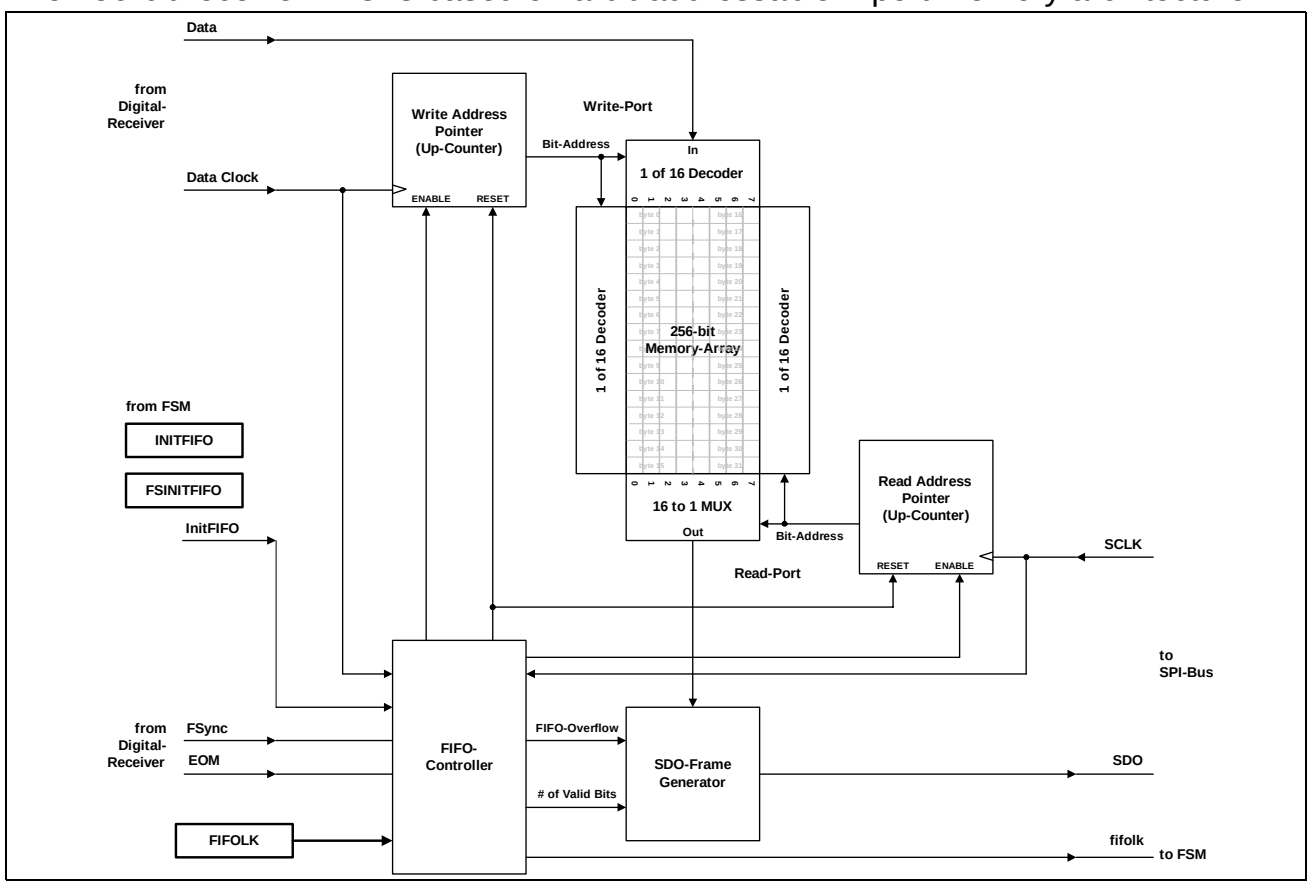


Figure 53 Receive FIFO

The write port is controlled by the Digital Receiver using the Write Address Pointer. Writing data into the FIFO starts with the detection of a TSI. The Write Address Pointer is incremented with each data clock signal generated by the Digital Receiver. The read port is controlled by the SPI controller using the Read Address Pointer. Each bit read from the SPI controller increments the Read Address Pointer. The Read and Write Address Pointers jump from their maximum value (255_d) to address zero. Writing to the FIFO stops at EOM or after Sync loss.

FIFO Lock Behavior

The FIFO possesses a lock mechanism that is enabled via the SFR control bit FIFOLK in the CMC1 register. If this mechanism is enabled, the FIFO will enter a FIFO Lock state at the detection of the EOM criterion. During the time that the FIFO is locked, it is not possible to receive additional data in Run Mode Self Polling. This means that it is only possible to detect another wake-up in the Self Polling Mode, but no more data in the Run Mode Self Polling. This will guarantee that only the first complete data packet is stored in the FIFO. Enabling FIFOLK also locks the digital receive chain at EOM until release from FIFO lock state.

The FIFO will remain locked unless one of three conditions occurs:

1.) The remaining contents of the FIFO are completely read out via the SPI

2.) The SFR control bit FIFOLK is cleared

3.) INITFIFO at Cycle Start is set in the CMC1 register and

a) FSM is switched to Run Mode Slave or

b) FSM switches from Self Polling Mode to Run Mode Self Polling

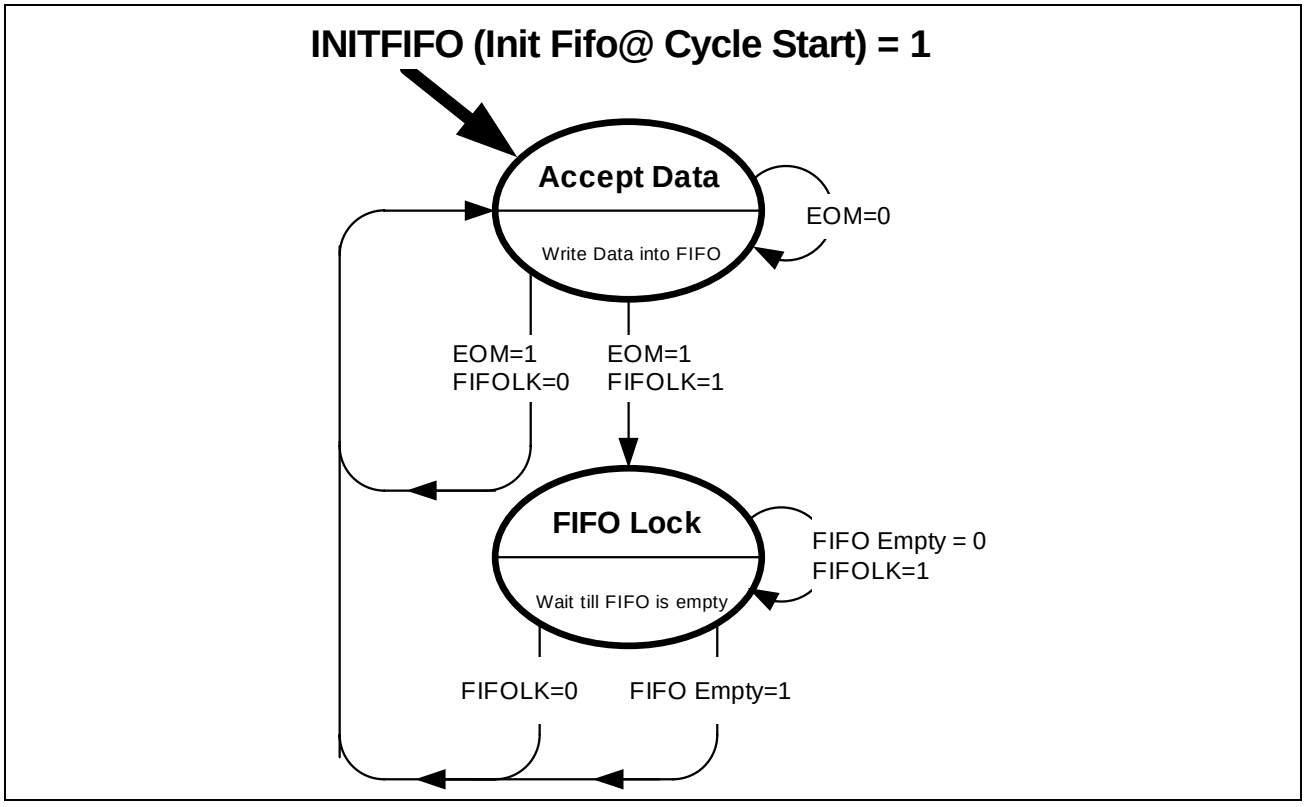


Figure 54 FIFO Lock Behavior

FIFO Status Word

The FIFO Status Word is attached at the end of a FIFO SPI transmission, and shows if there was an overflow, and how many valid data bits were transmitted. The number of valid FIFO bits is indicated at bit positions S0 to S5. S6 of the Status Word is always undefined.

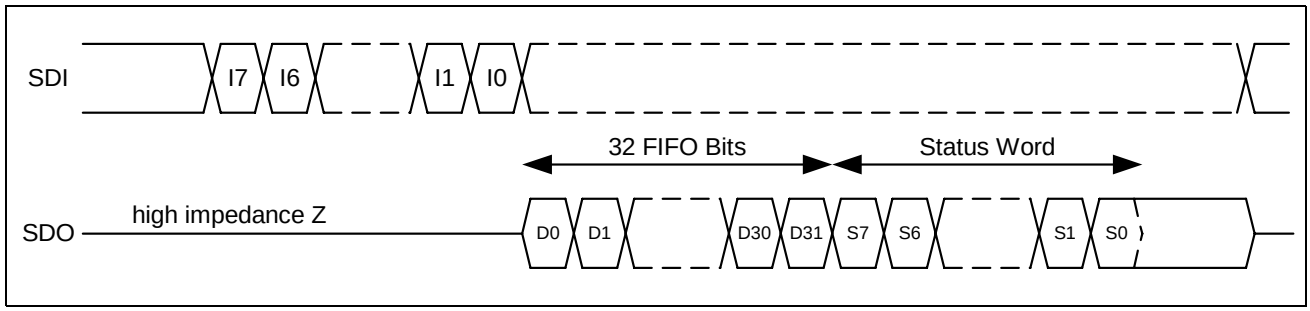


Figure 55 SPI Data FIFO Read

If the Write Address Pointer outruns the Read Address Pointer, an overflow is indicated in the FIFO Overflow Status bit in the FIFO Read Status Word at position S7. All 32 FIFO bits and the bits S5 to S0 of the Status Word are undefined while the Overflow Status bit is set.

If a TSI is detected after an overflow, the FIFO Overflow Status bit is cleared and the entire receive FIFO is initialized.

Initialization

Additionally, there are two possibilities to initialize the receive FIFO.

- If the INITFIFO bit is set in the CMC1 register (“Init FIFO at Cycle Start”) the entire receive FIFO is always initialized

  a.) after switching to Run Mode Slave or

  b.) switching from Self Polling Mode to Run Mode Self Polling.
- If the FSINITFIFO bit in CMC1 register is set, the entire receive FIFO is initialized when a TSI is detected and the receive FIFO is not locked (“Init FIFO at Frame Start”).

Last received message length

For application protocols with several payload frames and only a short pause in-between, the microcontroller would have to read out the FIFO very fast after detection of an EOM. Thus even slow or overloaded Application Controllers have the possibility now to determine the end of the last message, when reading out the FIFO, while the next payload frame gets already received and payload data is further stored in the FIFO.

Therefore the last received message length (e.g. after an EOM event) is stored in register PLDLEN and the upper two bits of register RFPLLACC at TSI detection of the next message. The upper two bits of register RFPLLACC hold the MSBs, thus a message length of 256 up to 1023 payload bits can be depicted. A saturation of the message length at the maximum value of 1023 is realized. Storage at TSI of the next message ensures that even wrong payload data (e.g. if MID is not matching, no EOM will be generated, but payload is kept in FIFO. Or EOM data length criterion is selected only and a sync loss prevents from generating an EOM event) can be identified.

On initialization of the FIFO, the register PLDLEN and the upper two bits of register RFPLLACC are cleared. The corresponding internal counter is cleared with every TSI detection and initialization of the FIFO.

PLDLEN will work correctly in case:

(INITDRXES = 0) AND ((Data rate > 22kBit/s) OR (EOM2SPM = 0))

If the condition above is not fulfilled, then the chip internal state machine can set PLDLEN to 0 and a correct function of PLDLEN cannot be guaranteed.

2.5.3 Digital Output Pins

As long as the P_ON pin is high, all digital output pins operate as described. If the P_ON pin is low, all digital output pins are switched to high impedance mode.

The digital outputs PP0, PP1, PP2 and PP3 are configurable, where each of the signals CLK_OUT, RX_RUN, NINT, a LOW level (GND) and a HIGH level, DATA, DATA_MATCHFIL, CH_DATA, CH_STR, RXD and RXSTR can be routed to any of the four output pins. There is only one exception, CLK_OUT is not available on PP3. The default configuration for these four output pins can be seen in **Table 1**.

Each port pin can be inverted by usage of PPCFG2 register.

The RX_RUN signal is active high for all Configurations by default. It can be deactivated for every Configuration separately. Every PPx can be configured with an individual RX_RUN setup. This can be set in RXRUNCFG0 and RXRUNCFG1 registers.

Interfacing to 3.3V Logic:

The TDA5235 is able to interface directly to a 3.3V logic, when chip is operated in 3.3V environment.

Interfacing to 5V Logic:

The TDA5235 is able to interface directly to a 5V logic, when chip is operated in 5V environment.

EMC Reduction of Digital I/Os:

Because electromagnetic distortion generated by digital I/Os may interfere with the high sensitivity radio receiver, it is recommended that all inputs are filtered by adding an RC low pass circuit.

2.5.4 Interrupt Generation Unit

The TDA5235 is able to signal interrupts (NINT signal) to the external Application Controller on one of the PPx port pins (for further details see **Chapter 2.5.3 Digital Output Pins**). The Interrupt Generation Unit receives all possible interrupts and sets the NINT signal based on the configuration of the Interrupt Mask register IM0. The Interrupt Status register IS0 is set from the Interrupt Generation Unit, depending on which interrupt occurred. The polarity of the interrupt can be changed in the PPCFG2 register. Please note that during power up and brownout reset, the polarity of NINT signal is always as described in **Chapter 2.4.9.2 Chip Reset**.

A Reset event has the highest priority. It sets all bits in the Status register to “1” and sets the interrupt signal to “0”. The first interrupt after the Reset event will clear the Status register and will set the interrupt signal to “1”, even if this interrupt is masked.

A Wake-up interrupt clears the FsyncA, FsyncB and the complementary Wake-up flag. An Fsync interrupt clears the EOMA, EOMB, MIDA, MIDB and the complementary Fsync flag.

The Interrupt Status register is always cleared after read out via SPI.

It is not possible to disable the Power On Reset Indicator Interrupt using the Interrupt Mask register.

Some interrupts are not usable depending on the selected receive mode, which is described in **Chapter 2.5.1.2 Data Interface**.

Interrupts for WU can be used in all receive modes.

Interrupts for FSync, MID and EOM can only be used in the receive modes POTP and POF.

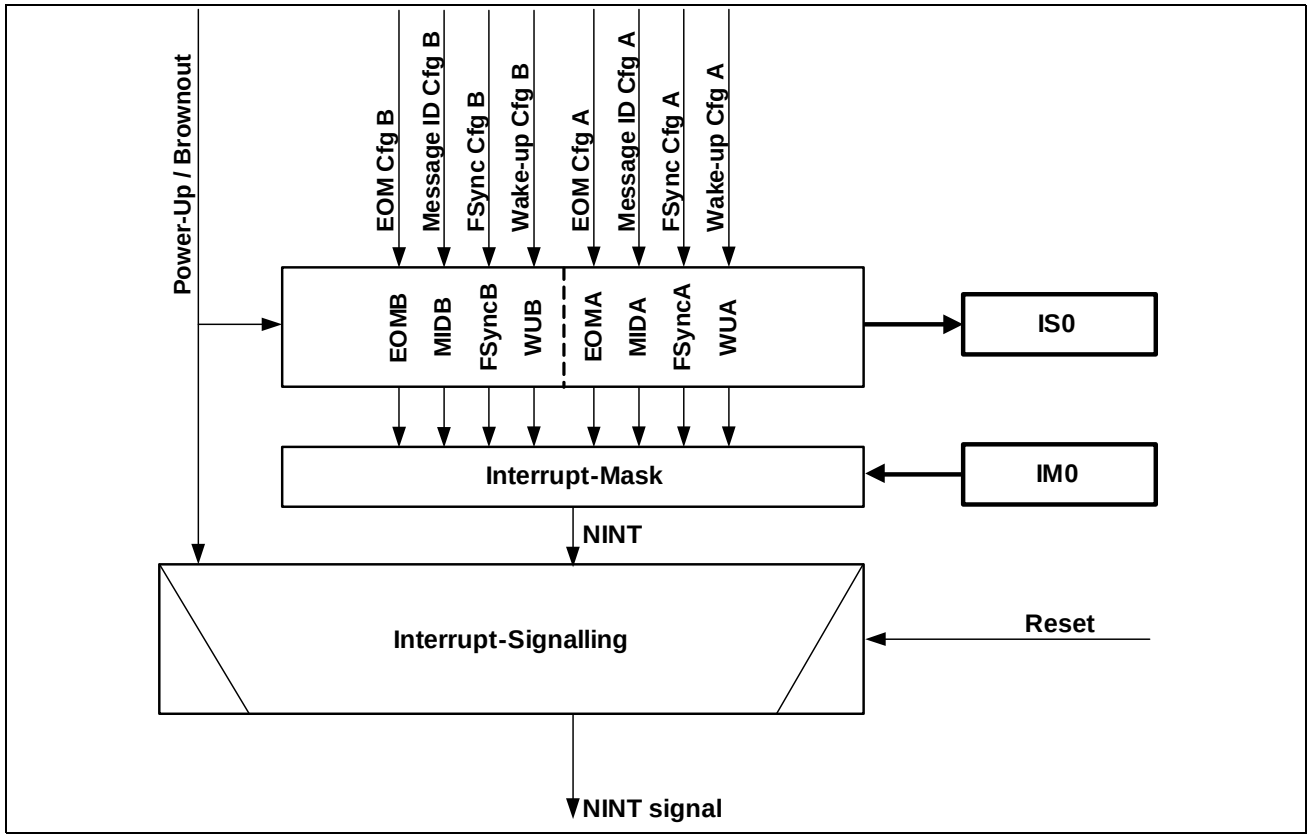


Figure 56 Interrupt Generation Unit

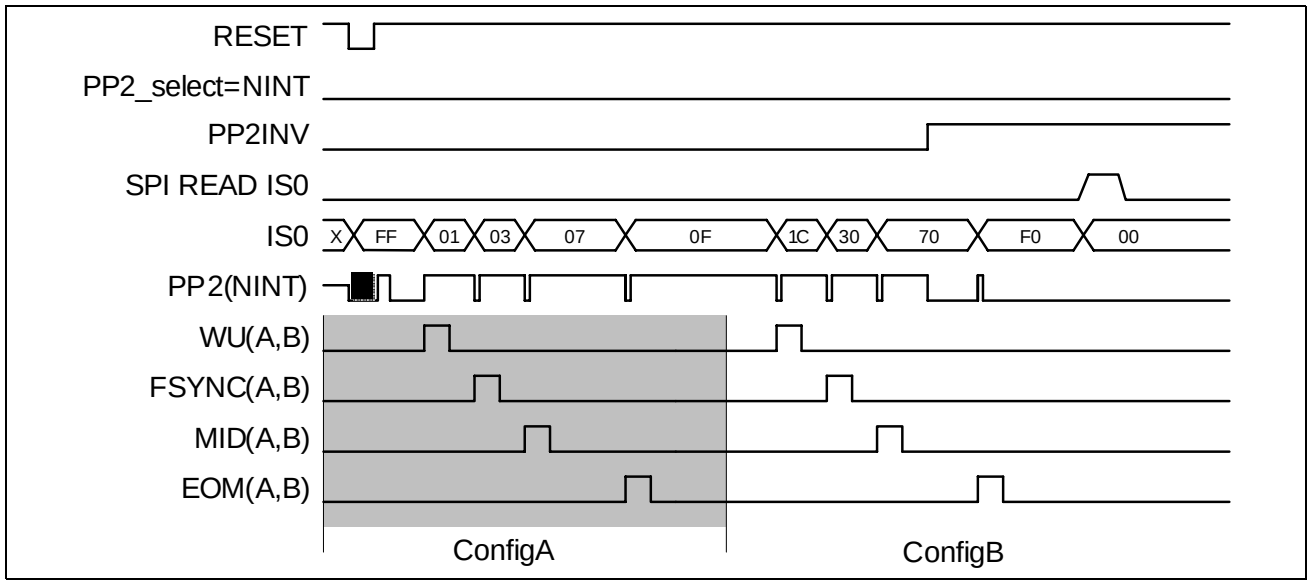


Figure 57 Interrupt Generation Waveform (Example for Configuration A+B)

The following handling mechanism for read-clear registers was chosen due to implementation of the Burst Read command:

- the current Interrupt Status (ISx) register 8-bit content is **latched** into the SPI shift register after the last address bit is clocked-in (**point A** in **Figure 58**)
- the IS register is then **cleared** after last IS register bit is clocked out of the SPI interface (**point B** in **Figure 58**)

Consequence: any interrupt event occurring in the window-time between points A and B is cleared at point B and not stored/shown in an later readout of ISx.

(However: NINT signal is toggling in any case, if occurring interrupt is not masked in IMx register)

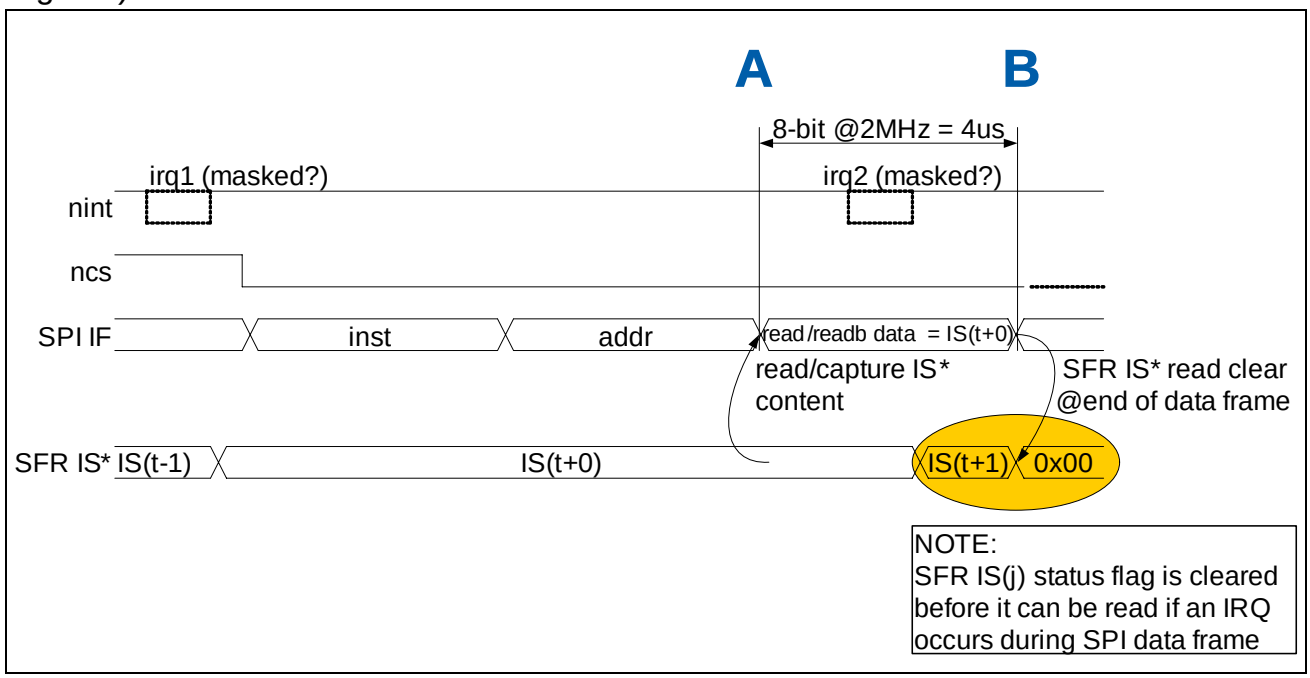


Figure 58 ISx Readout Set Clear Collision

Please see also the **IMPORTANT NOTE** in the Burst Read section !

2.5.5 Digital Control (4-wire SPI Bus)

The control interface used for device control and data transmission is a 4-wire SPI interface.

- *NCS* - select input, active low
- *SDI* - data input
- *SDO* - data output
- SCK - clock input: Data bits on *SDI* are read in at rising *SCK* edges and written out on *SDO* at falling *SCK* edges.

Level definition:
logic 0 = low voltage level
logic 1 = high voltage level

Note for non-Burst modes: It is possible to send multiple frames while the device is selected. It is also possible to change the access mode while the device is selected by sending a different instruction.

Note: In all bus transfers MSB is sent first, except for the received data read from the FIFO. There the bit order is given as first bit received is first bit transferred via the bus.

To **read from the device**, the SPI master has to select the SPI slave unit first. Therefore, the master must set the *NCS* line to low. After this, the instruction byte and the address byte are shifted in on *SDI* and stored in the internal instruction and address register. The data byte at this address is then shifted out on *SDO*. After completing the read operation, the master sets the *NCS* line to high.

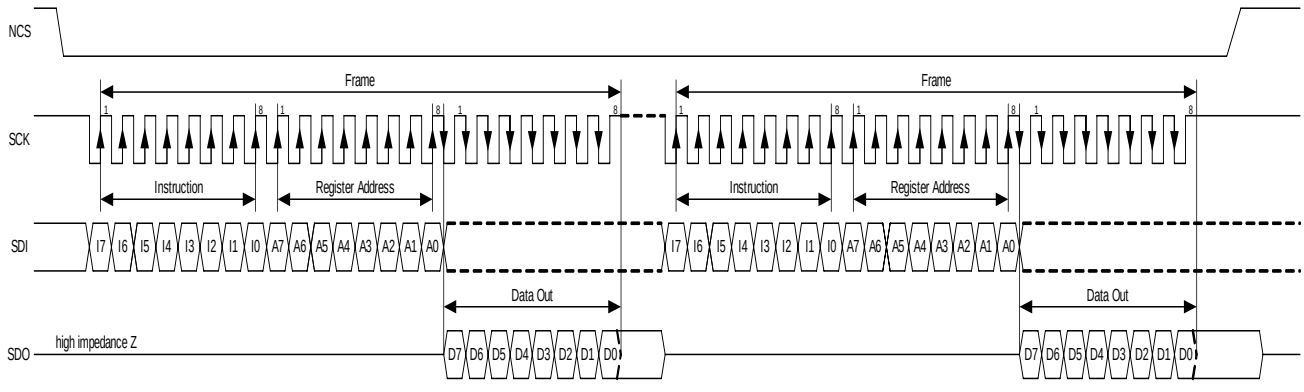


Figure 59 Read Register

To **read from the device in Burst mode**, the SPI master has to select the SPI slave unit first. Therefore the master has to drive the NCS line to low. After the instruction byte and the start address byte have been transferred to the SPI slave (MSB first), the slave unit will respond by transferring the register contents beginning from the given start address (MSB first). Driving the NCS line to high will end the Burst frame.

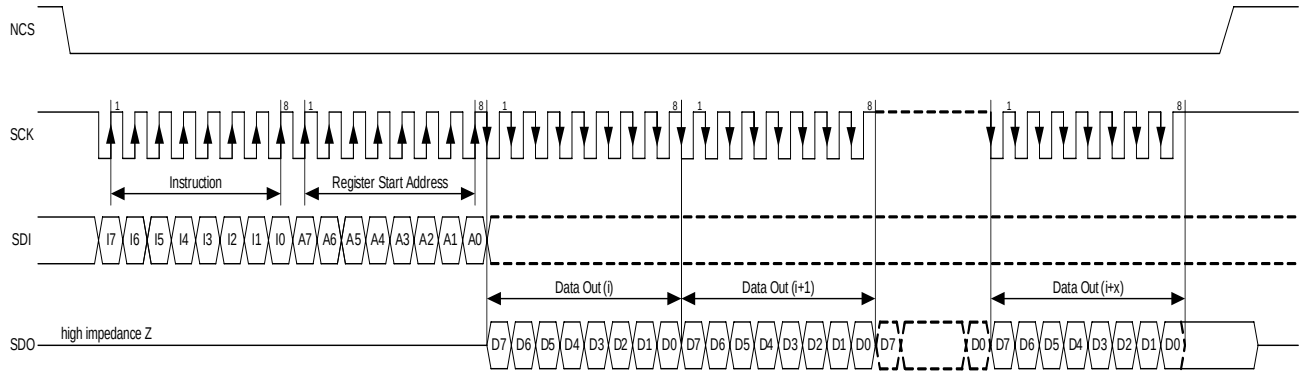


Figure 60 Burst Read Registers

IMPORTANT NOTE - for being upwards compatible with further versions of the product, we give following strong recommendation:

For read-clear registers at address (N), no read-burst access stopping at address (N-1) is allowed, because read-clear register will be cleared without being read out. Use single read command to read out the register at address (N-1) or extend the burst read to include the read-clear register at address (N).

To **write to the device,** the SPI master has to select the SPI slave unit first. Therefore, the master must set the *NCS* line to low. After this, the instruction byte and the address byte are shifted in on *SDI* and stored in the internal instruction and address register. The following data byte is then stored at this address.
After completing the writing operation, the master sets the *NCS* line to high.

Additionally the received address byte is stored into the register *SPIAT* and the received data byte is stored into the register *SPIDT*. These two **trace registers** are readable.

Therefore, an external controller is able to check the correct address and data transmission by reading out these two registers after each write instruction. The trace registers are updated at every write instruction, so only the last transmission can be checked by a read out of these two registers.

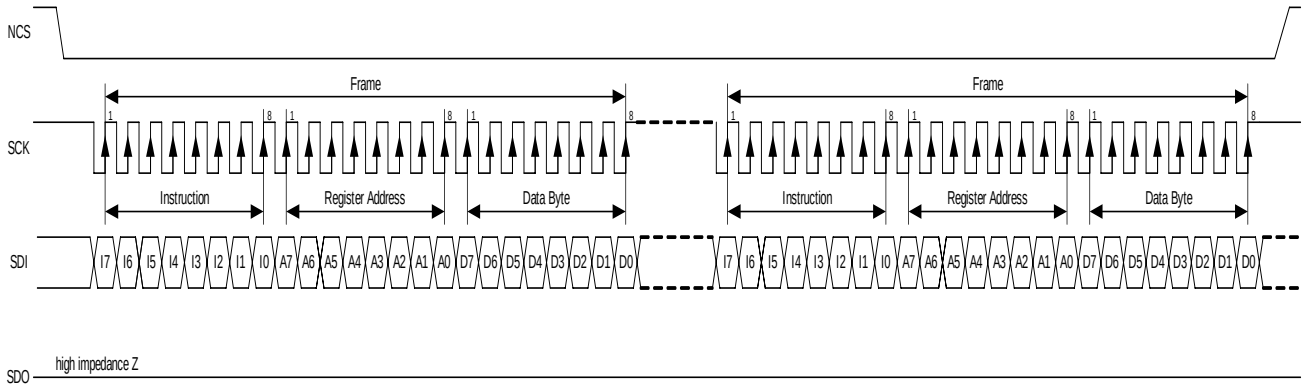


Figure 61 Write Register

To **write to the device in Burst mode**, the SPI master has to select the SPI slave unit first. Therefore the master has to drive the NCS line to low. After the instruction byte and the start address byte have been transferred to the SPI slave (MSB first) the successive data bytes will be stored into the automatically addressed registers.

To verify the SPI Burst Write transfer, the current address (start address, start address + 1, etc.) is stored in register SPIAT and the current data field of the frame is stored in register SPIDT. At the end of the Burst Write frame the latest address as well as the latest data field can be read out to verify the transfer. Note that some error in one of the intermediate data bytes can not be detected by reading SPIDT.

Driving the NCS line to high will end the Burst frame.

A single SPI Burst Write command can be applied very efficiently for data transfer either within a register block of configuration dependent registers or within the block of configuration independent registers.

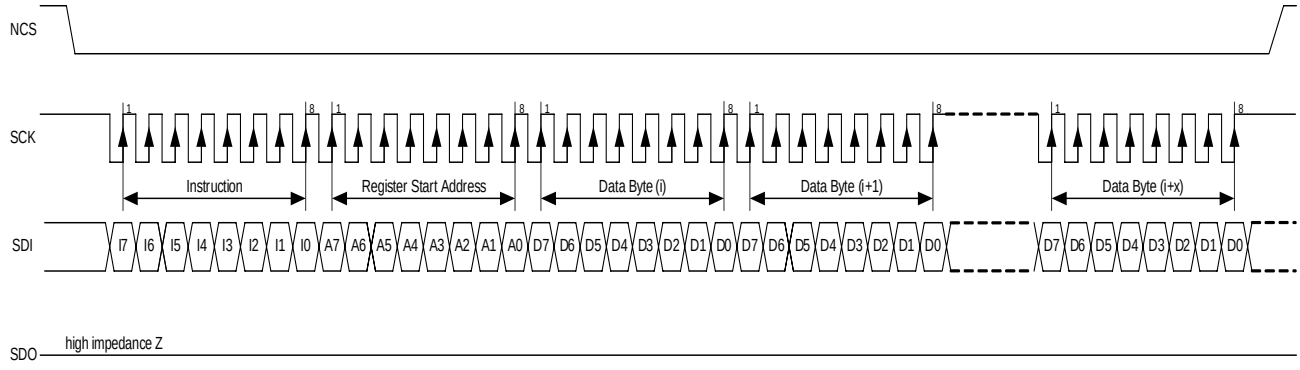


Figure 62 Burst Write Registers

The SPI also includes a safety feature by which the **checksum is calculated** with an XOR operation from the address and the data when writing SFR registers. The checksum is in fact an XOR of the data 8-bitwise after every 8 bits of the SPI write command. The calculated checksum value is automatically written in the SPICHKSUM register and can be compared with the expected value. After the SPICHKSUM register is read, its value is cleared.

In case of an SPI Burst Write frame, a checksum is calculated from the SPI start address and consecutive data fields.

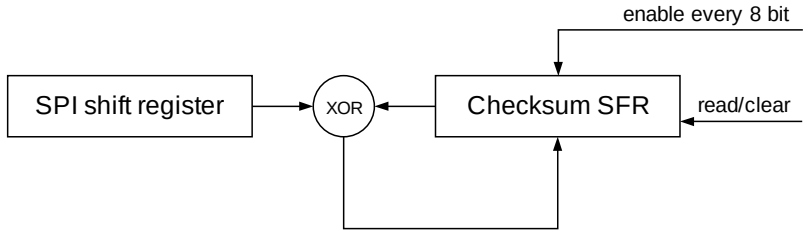


Figure 63 SPI Checksum Generation

To **read the FIFO,** the SPI master has to select the SPI slave unit first. Therefore, the master must set the *NCS* line to low. After this, the instruction byte is shifted in on *SDI* and stored in the internal instruction register. The data bits of the FIFO are then shifted out on *SDO*. The following byte is a status word that contains the number of valid bits in the data packet. After completing the read operation, the master sets the *NCS* line to high.

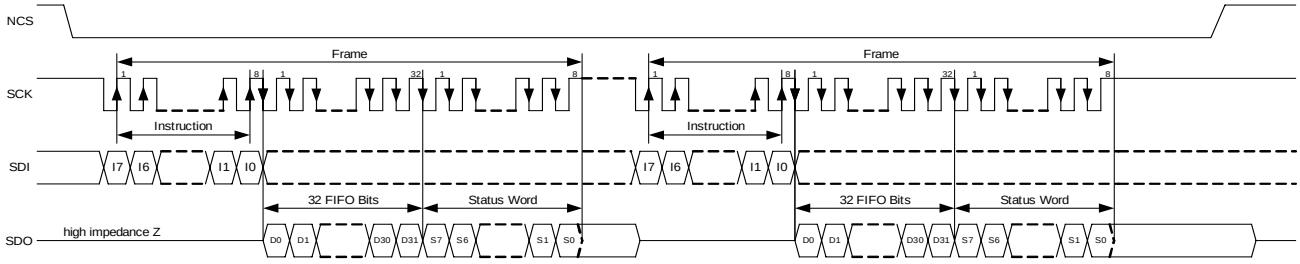


Figure 64 Read FIFO

Table 4 Instruction Set

Instruction	Description	Instruction Format
WR	Write to chip	0000 0010
RD	Read from chip	0000 0011
RDF	Read FIFO from chip	0000 0100
WRB	Write to chip in Burst mode	0000 0001
RDB	Read from chip in Burst mode	0000 0101

2.5.5.1 Timing Diagrams

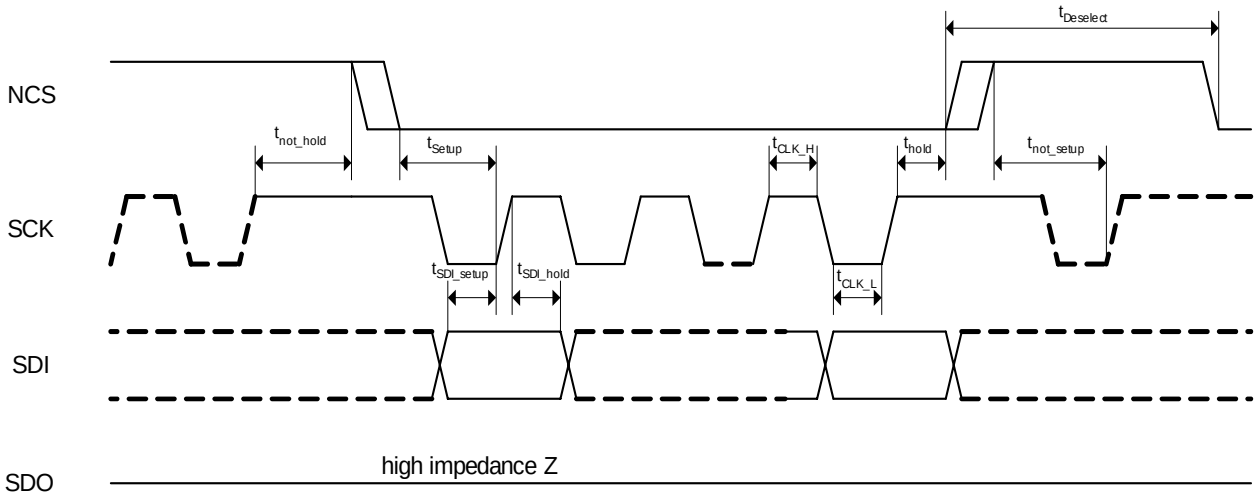


Figure 65 Serial Input Timing

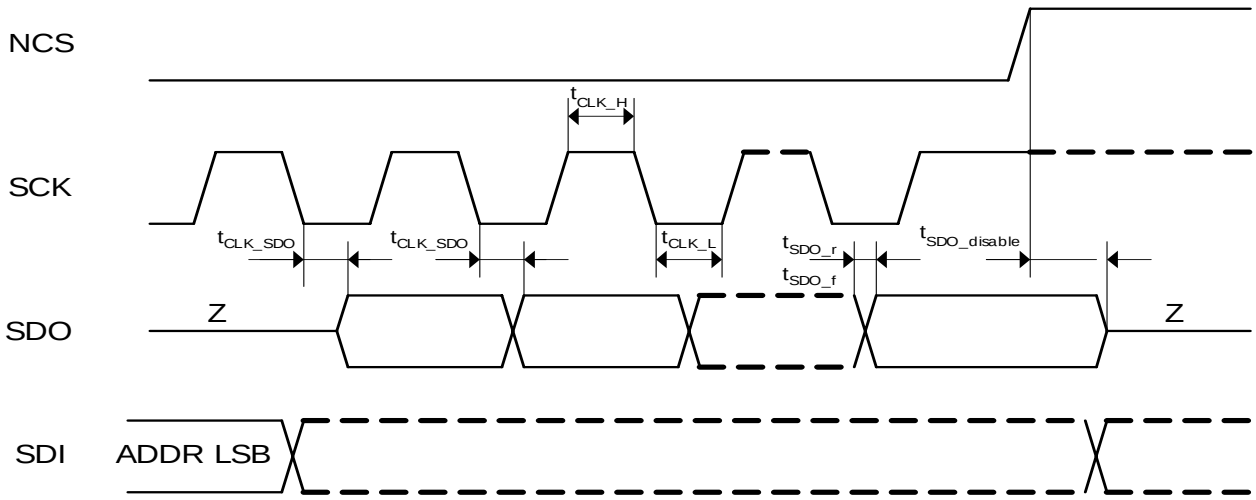


Figure 66 Serial Output Timing

Table 5 SPI Bus Timing Parameter

Symbol	Parameter
f_{clock}	Clock frequency
t_{CLK_H}	Clock High time
t_{CLK_L}	Clock Low time
t_{setup}	Active setup time
t_{not_setup}	Not active setup time
t_{hold}	Active hold time
t_{not_hold}	Not active hold time
$t_{Deselect}$	Deselect time
t_{SDI_setup}	SDI setup time
t_{SDI_hold}	SDI hold time
t_{CLK_SDO}	Clock low to SDO valid
t_{SDO_r}	SDO rise time
t_{SDO_f}	SDO fall time
$t_{SDO_disable}$	SDO disable time

2.5.6 Chip Serial Number

Every device contains a unique, preprogrammed 32-bit wide serial number. This number can be read out from SN3, SN2, SN1 and SN0 registers via the SPI interface. The TDA5235 always has SN0.6 set to 1 and SN0.5 set to 0.

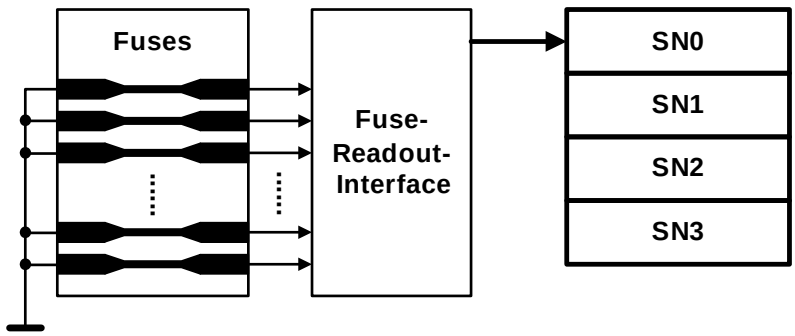


Figure 67 Chip Serial Number

2.6 System Management Unit (SMU)

The System Management Unit consists of two main units:

- **Master Control Unit**, where the various operating modes can be configured.
- **Polling Timer Unit**, where the receiver's On and Off times and modes are defined. The Polling Timer Unit is only working in the Self Polling Mode.

2.6.1 Master Control Unit (MCU)

2.6.1.1 Overview

The Master Control Unit controls the operation modes, the global states, and is generally responsible for automating data reception, verification, identification, extraction, and storage into the FIFO. The payload data without RUNIN, TSI and optional EOM can be read from the FIFO via SPI by the external microcontroller.

Alternatively, a transparent data stream can also be processed externally by the Application Controller (see **Chapter 2.5.1.2 Data Interface**).

The following operation modes and the behavior of the Master Control Unit are fully automatic and only influenced by SFR settings and by incoming RF data streams.

The TDA5235 has two major operation modes, which are switched by SFR bit MSEL.

In **Slave Mode** the device is controlled via SPI by the external microcontroller. This mode supports:

- **Run Mode Slave (RMS)**, where the receiver is continuously active
- **SLEEP Mode**, where the receiver is switched off for power saving. This mode can also be used to change register settings
- **HOLD Mode**, allows register settings to be changed. The change to HOLD Mode and back to RMS is faster than changing to SLEEP Mode and back to RMS.

In Slave Mode, switching between configurations, as well as between Run and SLEEP Mode must be initiated by the microcontroller.

In **Self Polling Mode,** TDA5235 autonomously polls for incoming RF signals. The receiver switches automatically between up to two configurations (Configuration A and B). Further information can be found in **Chapter 2.6.2**.

Between the RF signal scans, the receiver is automatically switched to Low Power Mode for reducing the average power consumption. If an incoming signal fulfills the selected wake-up criterion an interrupt can be generated and Run Mode Self Polling will be entered. If the following received data matches to the TSI pattern, and passes the optional message ID screening, the payload is loaded into the FIFO, and, if not masked, an interrupt is generated. Then the payload data can be read via SPI.

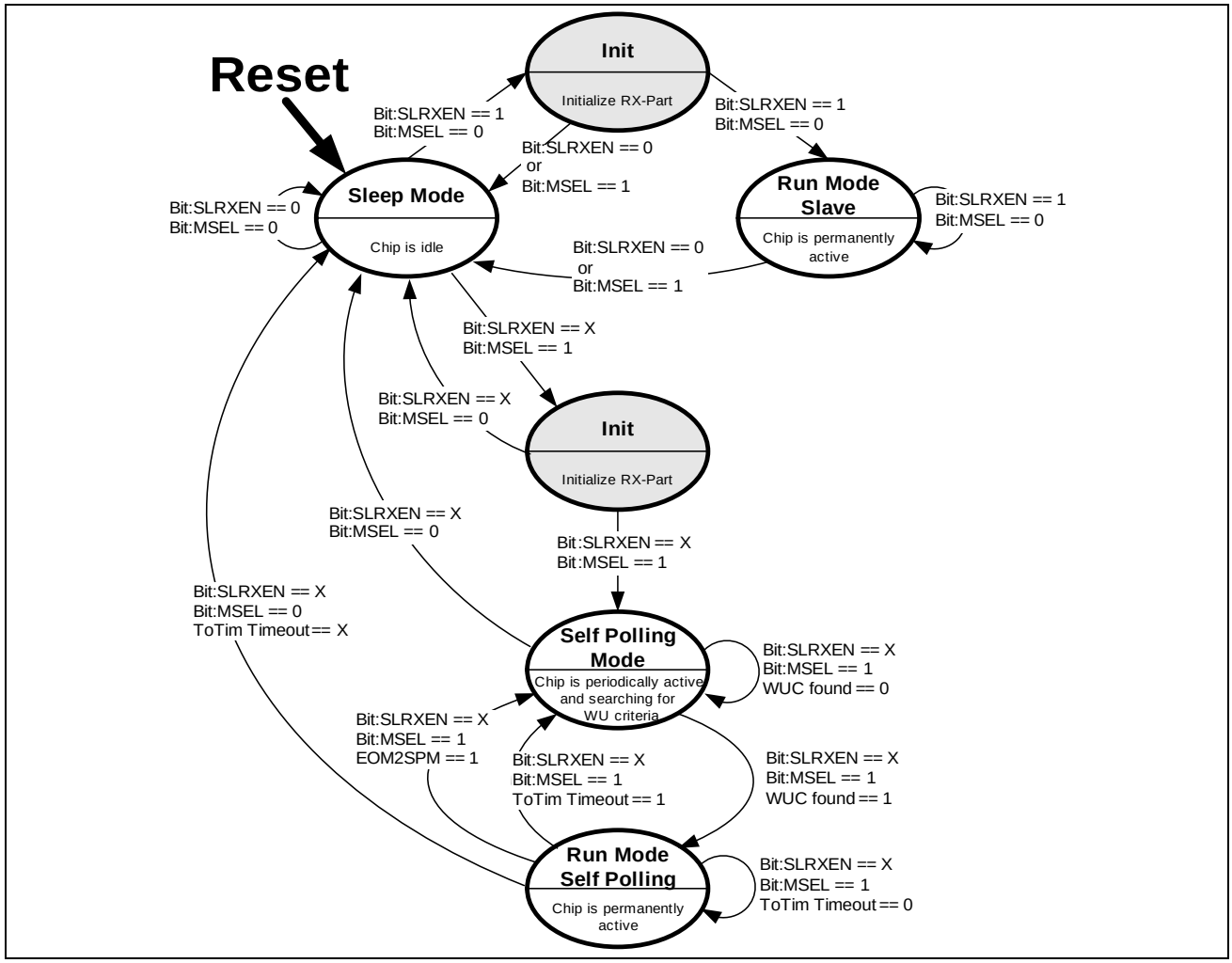


Figure 68 Global State Diagram

2.6.1.2 Run Mode Slave (RMS)

In Run Mode Slave, the receiver is able to continuously scan for incoming data streams. Detection and validation of a wake-up criterion are not performed, but RUNIN and TSI are required.

Recognition of TSI and validation of the optional MID (Message IDentification) are done automatically. The data payload is extracted from the data stream, and moved to the FIFO.

The various recognition steps are communicated by interrupts. Interrupts can be generated at frame-start (when a valid TSI has been detected), when a valid MID has been found and at EOM (End of Message).

Alternatively, a transparent data stream can also be processed externally by the Application Controller (see **Chapter 2.5.1.2 Data Interface**).

Run Mode Slave is entered by setting SFR CMC0 bits MSEL to 0 and SLRXEN to 1.

Configurations are switched via SFR bit MCS in the CMC0 register. The RF channel can be selected by SFRs x_PLLINTC1, x_PLLFRAC0C1, x_PLLFRAC1C1, x_PLLFRAC2C1, where x = A or B.

The configuration may be changed only in SLEEP or in HOLD Mode before returning to the previously selected operation mode. This is necessary to restart the state machine with defined settings at a defined state. Otherwise the state machine may hang up. Reconfigurations in HOLD Mode are faster, because there is no Start-Up sequence.

The following flowchart and explanation show and help to understand the internal behavior of the Finite State Machine (FSM) in Run Mode Slave.

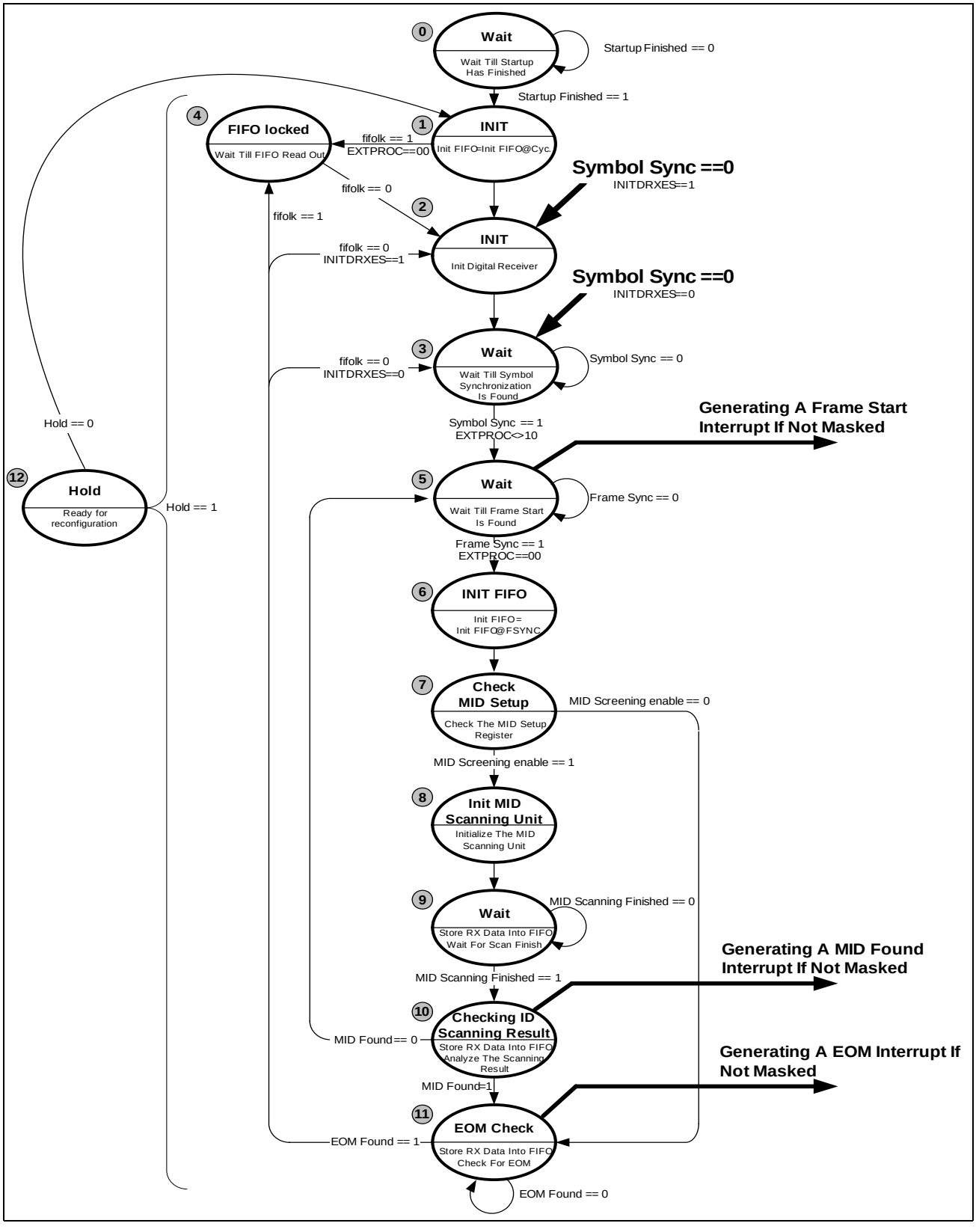


Figure 69 Run Mode Slave

2.6.1.3 HOLD Mode

This state (item 12 in **Figure 69**) is used for fast reconfiguration of the chip in Run Mode Slave. This state can be reached after the Start-Up Sequencer and Initialization of the chip have been completed from any state from 3 to 11. To reconfigure the chip the SFR control bit HOLD must be set. After reconfiguration in this state the SFR control bit HOLD must be cleared again. After leaving the HOLD state, the INIT state is entered and the receiver can work with the new settings. Be aware that the time between changing the configuration and reinitialization of the chip has to be at least 40us. Take note that one SPI command for clearing the SFR control bit HOLD needs 24 bits or 12µs at an SPI data rate of 2.0Mbit/s. The remaining 28µs must be guaranteed by the application.

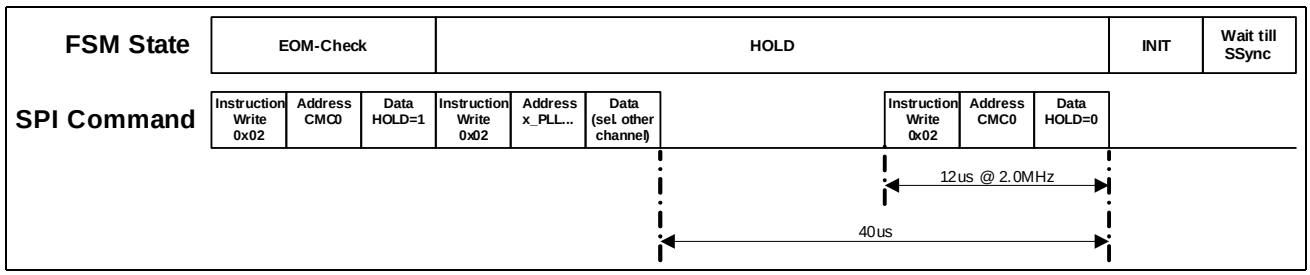


Figure 70 HOLD State Behavior (INITPLLHOLD disabled)

In case of large frequency steps, an additional VAC routine (VCO Automatic Calibration) has to be activated when recovering from HOLD Mode (INITPLLHOLD bit). The maximum allowed frequency step in HOLD Mode without activation of VAC routine is depending on the selected frequency band. The limits are +/- 1 MHz for the 315 MHz band, +/- 1.5 MHz for the 434 MHz band and +/- 3 MHz for the 868/915 MHz band.

When this additional VAC routine is enabled, the TDA5235 starts initialization of the Digital Receiver block after release from HOLD and an additional Channel Hop time.

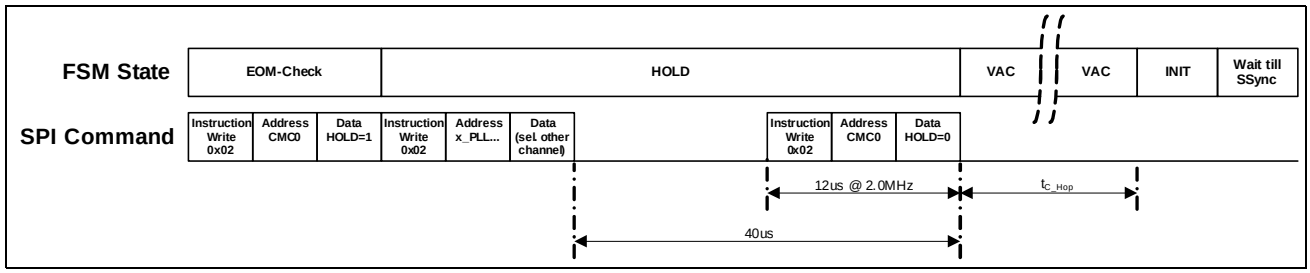


Figure 71 HOLD State Behavior (INITPLLHOLD enabled)

HOLD Mode is only available in Run Mode Slave. Configuration changes in Self Polling Mode have to be done by switching to SLEEP Mode and returning to Self Polling Mode after reconfiguration.

2.6.1.4 SLEEP Mode

The SLEEP Mode is a power save mode. The complete RF part is switched off and the oscillator is in Low Power Mode. As in HOLD Mode, the chip can be reconfigured. When switching from SLEEP to Run Mode Slave, the state machine starts with the internal Start-Up Sequence.

2.6.1.5 Self Polling Mode (SPM)

In Self Polling Mode TDA5235 autonomously polls for incoming RF wake-up data streams. There is no processing load on the host microcontroller. When a wake-up criterion has been found, an interrupt can be generated and the TDA5235 mode is changed to Run Mode Self Polling for automatic verification of TSI, optional MIDs and for transfer of payload data into the FIFO.

A general overview on a typically transmitted protocol and the behaviour of the TDA5235 is given in **Figure 72**.

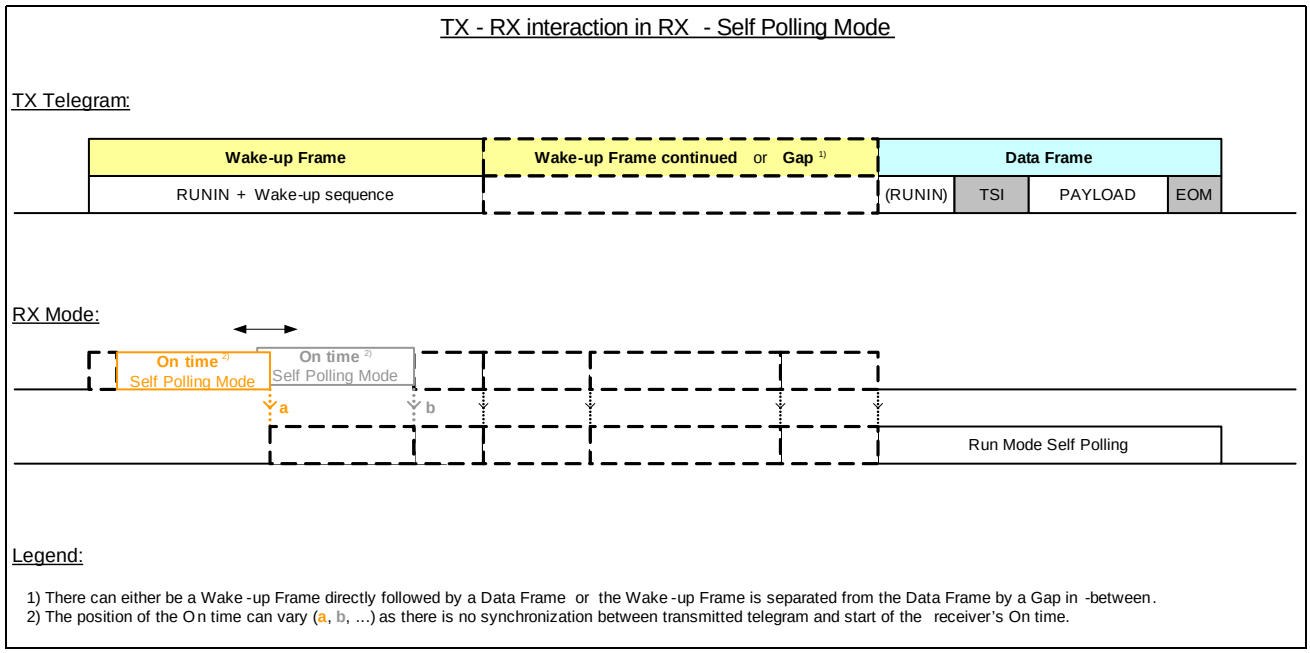


Figure 72 SPM - TX-RX Interaction

Alternatively, a transparent data stream can also be processed externally by the Application Controller (see **Chapter 2.5.1.2 Data Interface**).

Self Polling Mode is entered by setting the MSEL register bit to 1.

Configuration changes are allowed only by switching to SLEEP Mode, and returning to Self Polling Mode after reconfiguration.

The **Polling Timer Unit** controls the timing for scanning (On time) and sleeping (Off time, SPM_OFF). Up to two independent configuration sets (A and B) can automatically be processed, thus enabling scanning from different transmit sources. See also **Chapter 2.6.2 Polling Timer Unit**.

The **Wake-Up Generation Unit** identifies, whether an incoming data stream matches the configurable wake-up criterion.

After fulfillment of the wake-up criterion, modulation can be switched automatically.

See also **Chapter 2.6.1.6 Automatic Modulation Switching**, **Chapter 2.4.8.5 Wake-Up Generator** and **Chapter 2.5.1.2 Data Interface** (in Subsection TMRDS).

The following state diagrams and explanations help to illustrate the behavior during Self Polling Mode. First there is a search for a wake-up criterion according to Configuration A. Then, there is an optional search for a wake-up criterion according to Configuration B.

In applications using only Single-Configuration, settings are always taken from Configuration A.

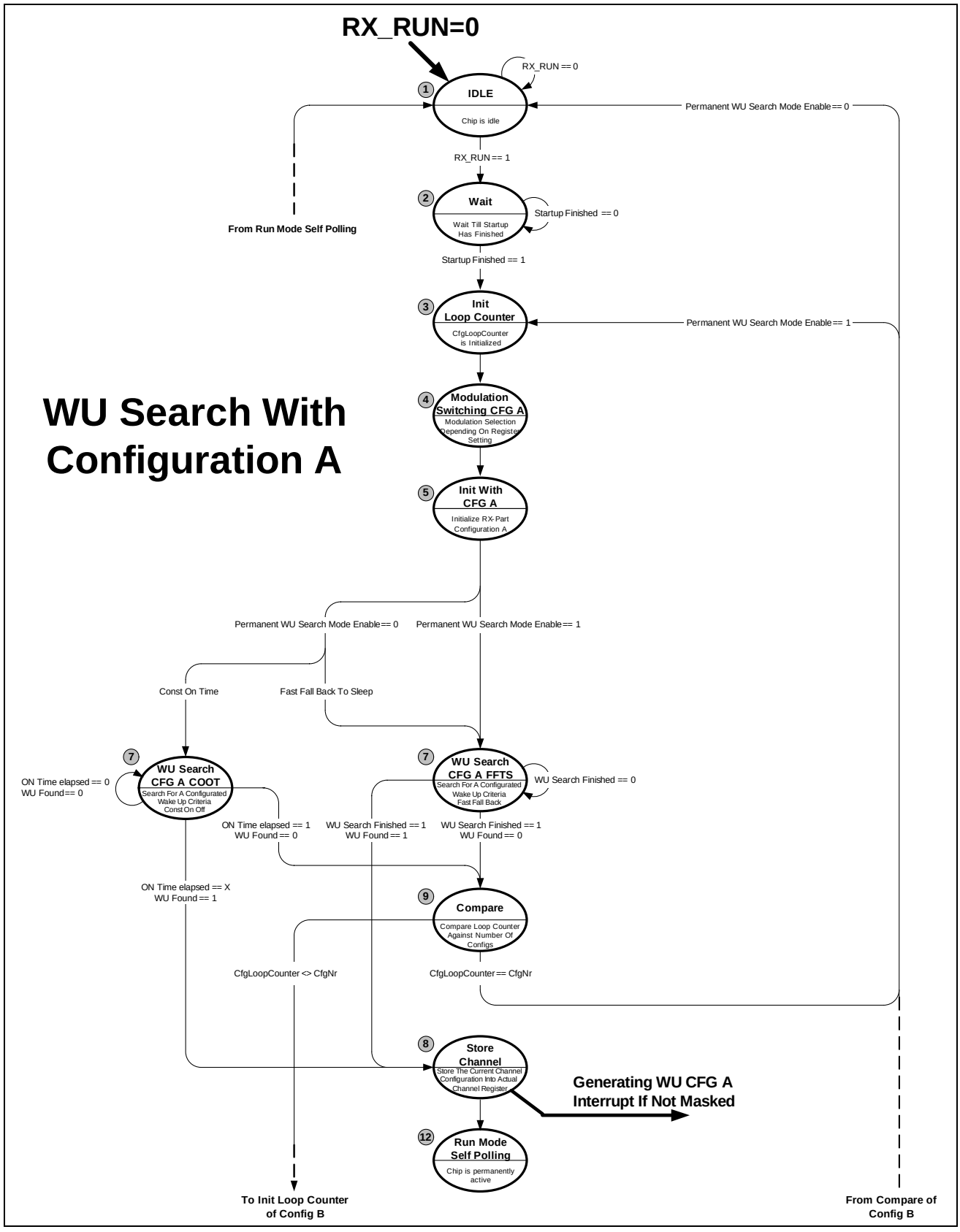


Figure 73 Wake-up Search with Configuration A

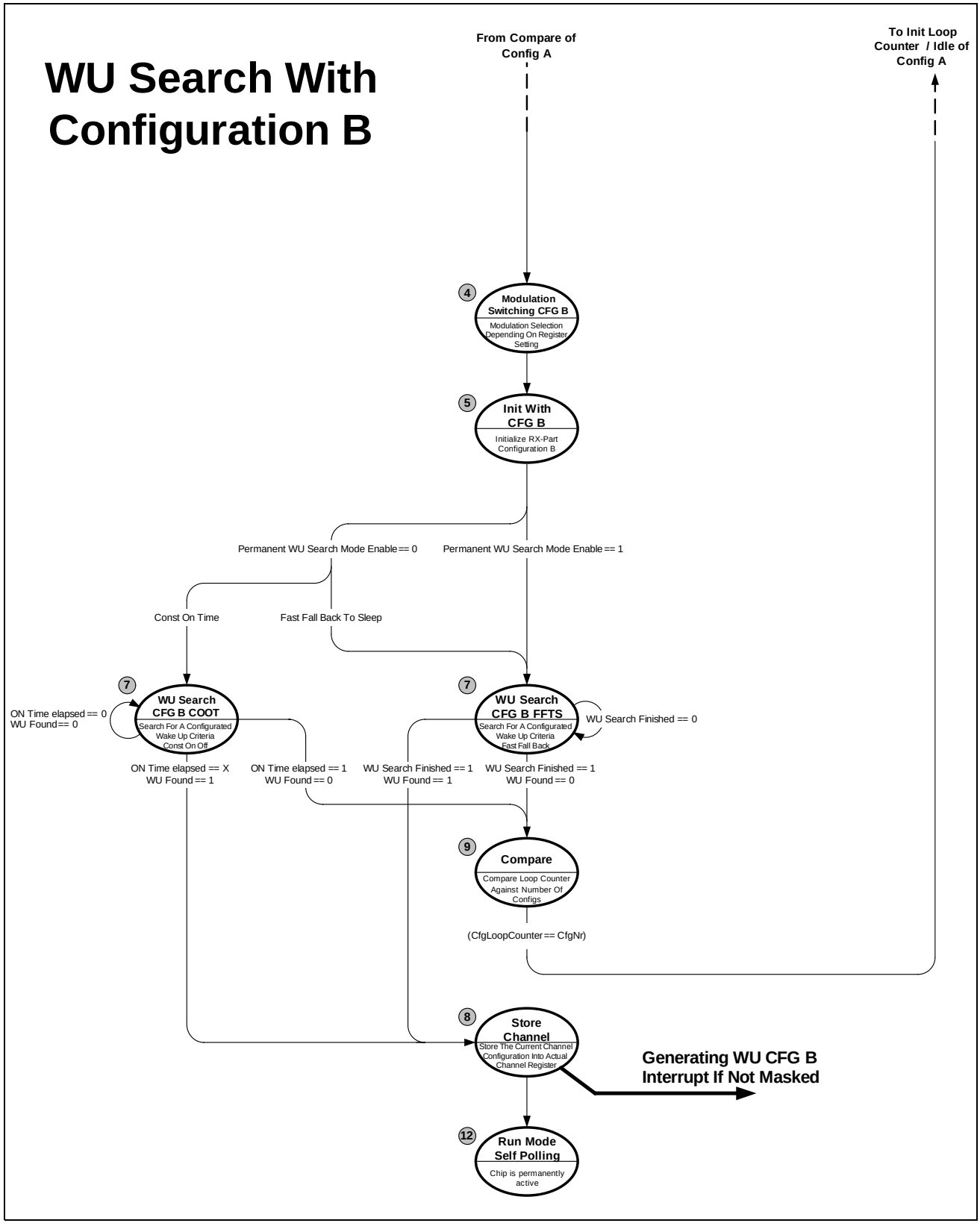


Figure 74 Wake-up Search with Configuration B

2.6.1.6 Automatic Modulation Switching

In **Self Polling Mode**, the chip is able to automatically change the type of modulation after a **wake-up** criterion was fulfilled in a received data stream. The type of modulation used in the different operational modes is selected by the SFR control bit MT.

2.6.1.7 Multi-Channel in Self Polling Mode

A simple Multi-Channel behavior can be realized when the same application parameters are programmed in both Configuration sets and only the RF PLL frequency differs. Channel frequencies are defined in registers x_PLLINTC1, x_PLLFRAC0C1, x_PLLFRAC1C1, x_PLLFRAC2C1, where x = A or B.

See also **Chapter 2.4.5 Sigma-Delta Fractional-N PLL Block**.

2.6.1.8 Run Mode Self Polling (RMSP)

The chip enters **Run Mode Self Polling** after a successful fulfillment of a **wake-up** criterion in Self Polling Mode.

When Wake-Up criterion for RSSI or Signal Recognition (see **Chapter 2.4.8.1**) is selected and fulfilled, this leads to a change to **Run Mode Self Polling**. This will be interesting especially in case of a transparent data stream being processed externally by the Application Controller (see **Chapter 2.5.1.2 Data Interface**).

The following steps are performed automatically, depending on register settings:

- Modulation switching (see **Chapter 2.6.1.6 Automatic Modulation Switching**)
- Wait for valid TSI (see **Chapter 2.4.8.6 Frame Synchronization**)
- Initialize FIFO (see **Chapter 2.5.2 Receive FIFO**) and write data to FIFO
- Scan for MIDs (see **Chapter 2.4.8.7 Message ID Scanning**)

Depending on interrupt masking, the host microcontroller is alerted when

- a data frame has started,
- an MID has been found, (if enabled) or
- EOM (End of Message) has been detected.

See also **Chapter 2.5.4 Interrupt Generation Unit**

Run Mode Self Polling is left, when synchronization is lost and the timeout timer for loss of synchronization (TOTIM_SYNC) has elapsed, or when one of the other timeout timers (TOTIM_TSI, TOTIM_EOM) for each configuration (A, B) has elapsed, or when an EOM occurred and the SFR bit EOM2SPM is activated, or when the operating mode is switched to SLEEP or Run Mode Slave by the host microcontroller.

Timeout timers for getting no TSI or getting no EOM within a certain time period can be used to avoid a deadlock situation, e.g. TOTIM_TSI can be used in case an interfering transmit signal fulfilled the wake-up criterion and keeps on transmitting, but no TSI can be found in this data stream within a certain programmable time period. TOTIM_EOM might be used in case EOM criterion “EOM by payload data length” cannot be applied.

The timeout timer functionality in the absence/presence of an interfering signal is shown in **Figure 75** and **Figure 76**.

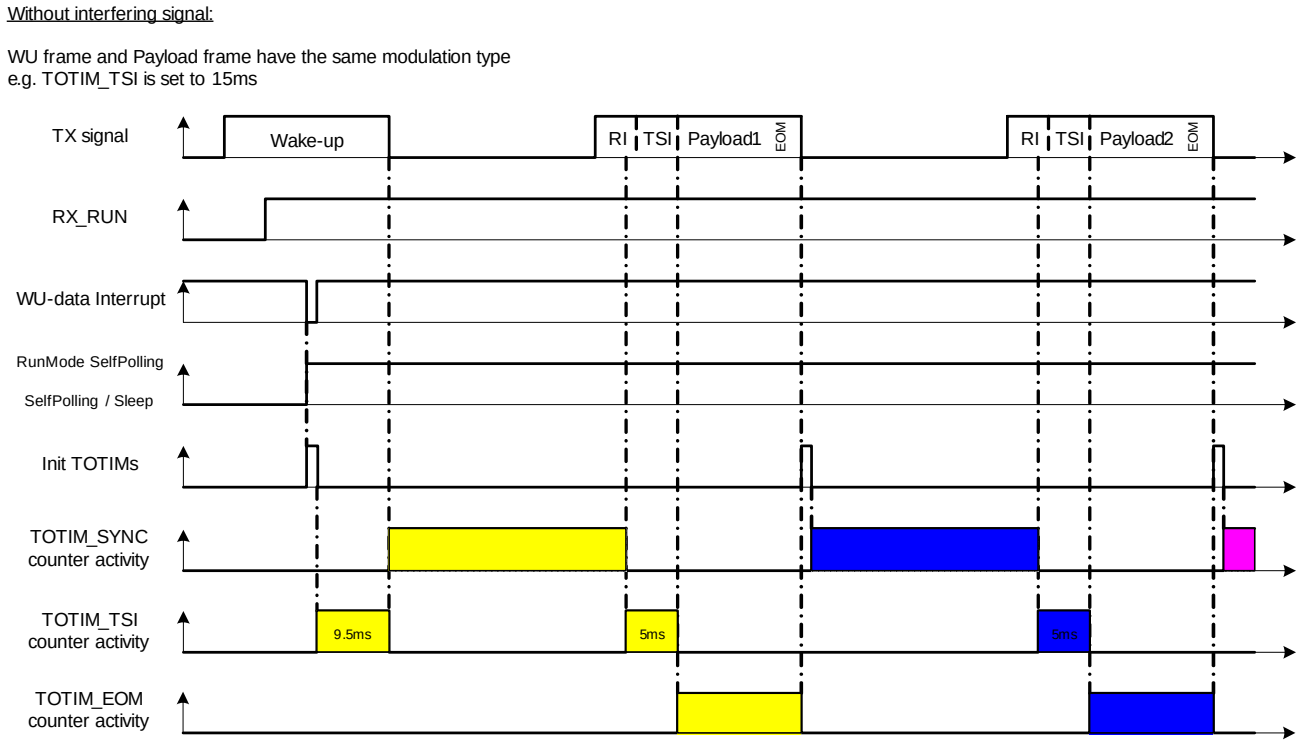


Figure 75 TOTIM Behavior without Presence of Interferer

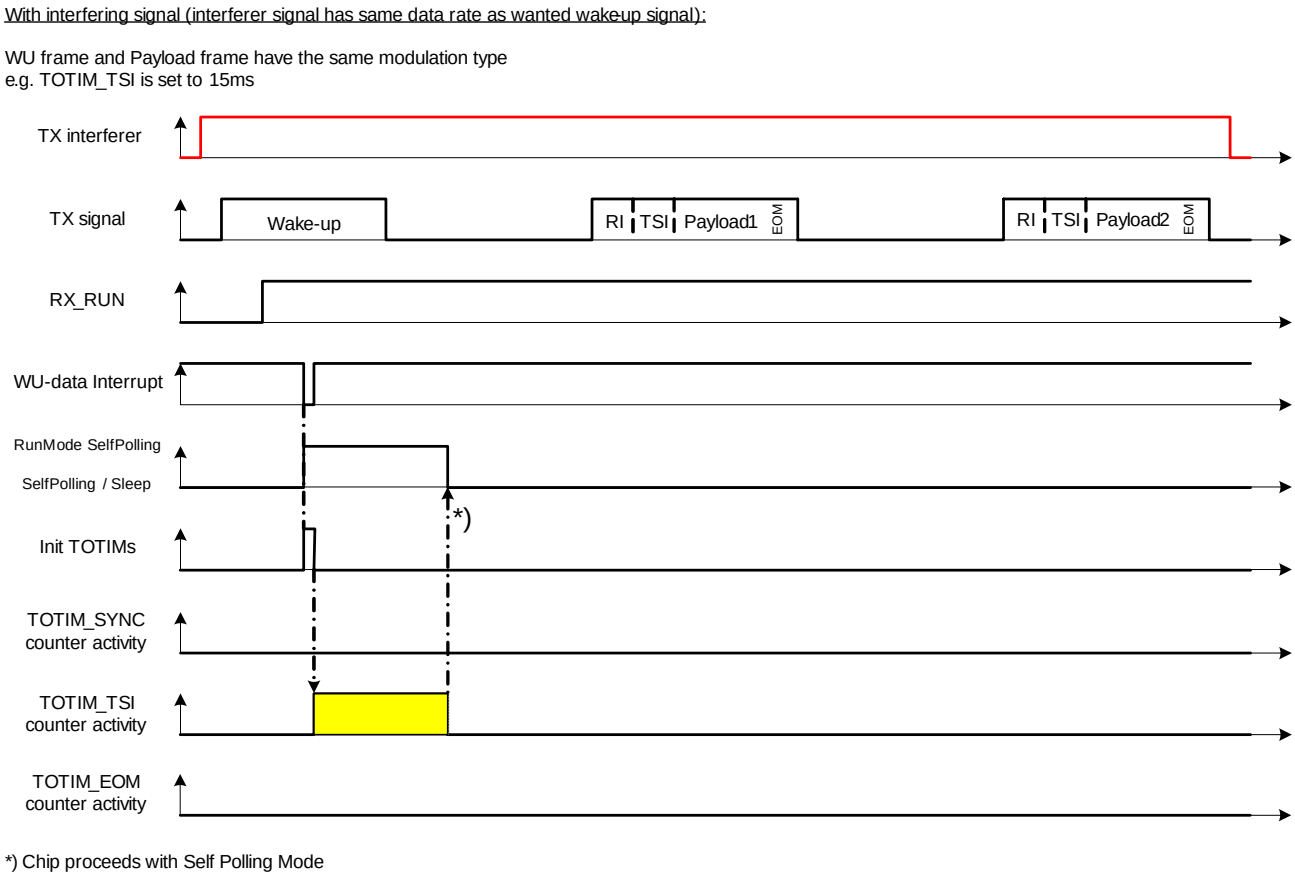


Figure 76 TOTIM Behavior in Presence of Interferer

On expiring of one of the timeout timers, the receiver proceeds with Self Polling Mode and with searching for a suitable wake-up criterion on the next configuration or a search for a wake-up criterion in Configuration A is initiated.

As long as the chip is in Run Mode Self Polling, incoming data frames (including a RUNIN sequence and TSI, but without necessity of additional wake-up patterns) can be received and stored.

The data FIFO can be initialized and cleared either at

- Cycle Start, that means whenever Run Mode Self Polling is entered or
- Frame Start, when a TSI has been successfully identified (and Receive FIFO is not locked).

Further information about the Receive FIFO can be found in the **Chapter 2.5.2 Receive FIFO**.

After an EOM was found, the information about the RF channel and the configuration of the actual payload data is saved in the RFPLLACC register.

After detection of EOM the TDA5235 can either proceed with a search for a wake-up criterion in the next configuration or a search for wake-up in Configuration A can follow or the TOTIMs of the current configuration are reloaded for being prepared to receive another (redundant) payload data frame within the same configuration.

Alternatively, a transparent data stream can also be processed externally by the Application Controller. Therefore the external controller needs the possibility to send following commands, which would normally be generated by the TDA5235 itself (see **Figure 77** and EXTPCMD register as well):

- EXTTOTIM: So the TDA5235 can proceed with Self Polling Mode (either with the next configuration or with Configuration A).
- EXTEOM found: In this case the TDA5235 can either proceed with Self Polling Mode (either with the next configuration or with Configuration A) or stay in Run Mode Self Polling.

EXTTOTIM and EXTEOM are only available, when the external processing mode is deactivating functional blocks (see bit group x_CHCFG.EXTPROC).

When the actual processed configuration is right before the Off time and the Application Controller sends one of the above mentioned commands, then the TDA5235 can proceed with the Off time (in case next configuration is selected).

If the autonomous Wake-up Search with Configuration A follows a TOTIM or EOM event, then also the Polling Timer is initialized, this means a new On period is started. In case the Wake-up Search gets started with Next Configuration (after a TOTIM or EOM event), then the Polling Timer is not initialized. This means that the On time counter proceeds with the old value from leaving the previous Wake-up search period successfully. This is the case for Fast Fall Back to SLEEP Mode.

In Constant On-Off Time Mode the Polling Timer is always initialized after a TOTIM or EOM event.

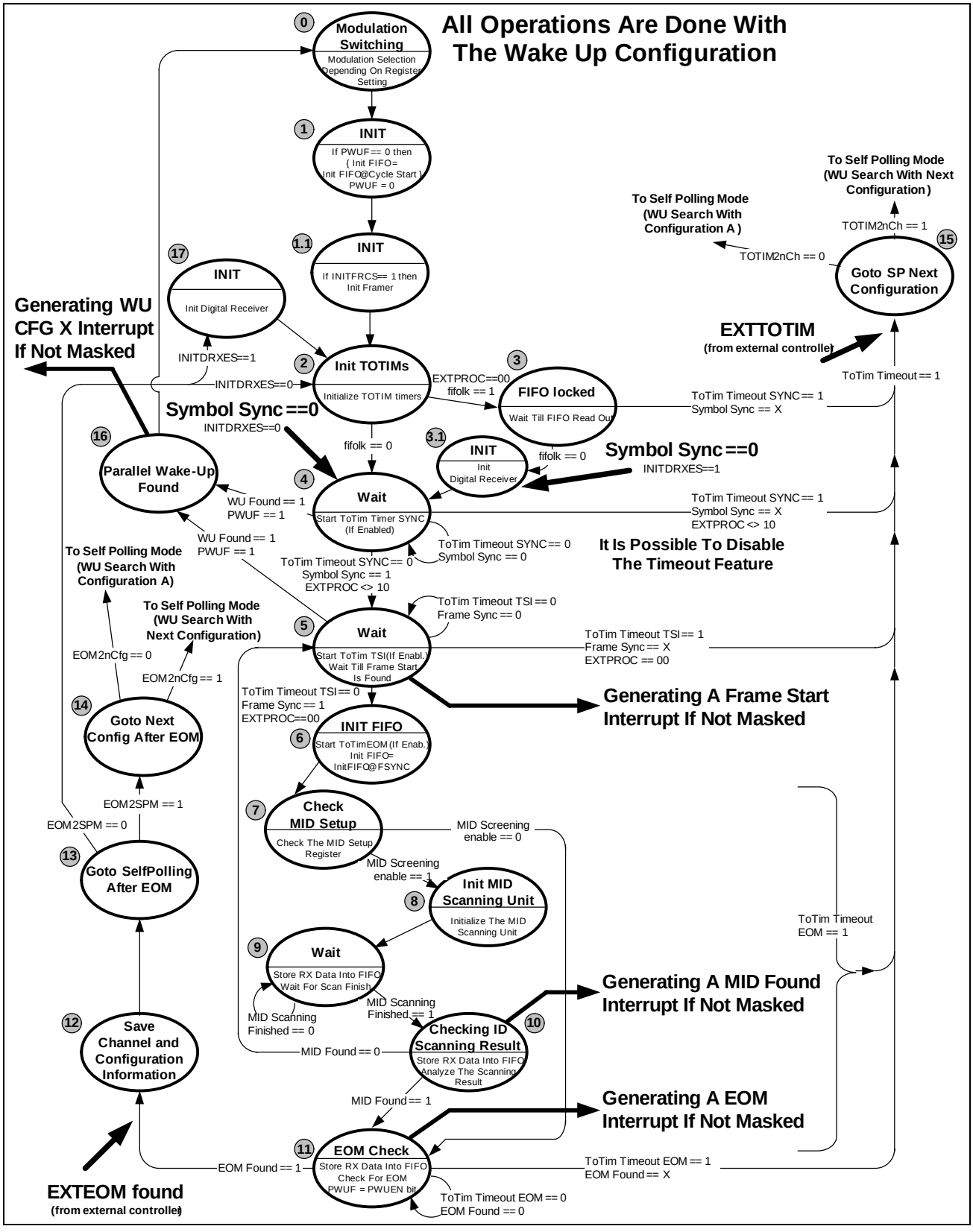


Figure 77 Run Mode Self Polling

While the TDA5235 is in Run Mode Self Polling, further Wake-ups would normally not be detected by the receiver. If the functionality of a parallel Wake-up search during the search for a TSI is desired, this can be activated by the PWUEN bit. In this case the Wake-up search is not active during a recognized payload and is only active after the first received payload frame, as can be seen from **Figure 77**. This feature can only be used, when modulation type is the same for SPM and RMSP.

So after a reception of the EOM from the current payload, the parallel WU search can take place in this mode. The WU search will be active after Symbol Sync has been detected. The WU search will be active until the Synchronization gets lost or wake-up is generated. After the Synchronization gets lost the WU search will be finished and wake-up can not be detected any more (the TSI search continues as usual).

Following procedure can be applied with help of 3 SPI Write command sequences.

The idea is to generate external EOM every time the Symbol Sync goes to inactive state and no interrupt (TSI or WU) has been detected. This will bring the MCU to the cycle start and reinitialize the WU search.

Configuration:

Write x_WUC.PWUEN = 1

// Enable Parallel Wake-up search

Write x_WURSSITH1 0xFF

// Set RSSI threshold to max value (avoid WU during the reinitialization procedure)

Write x_WULOT 0xFF

// Set WULOT to max value (avoid WU during the reinitialization procedure)

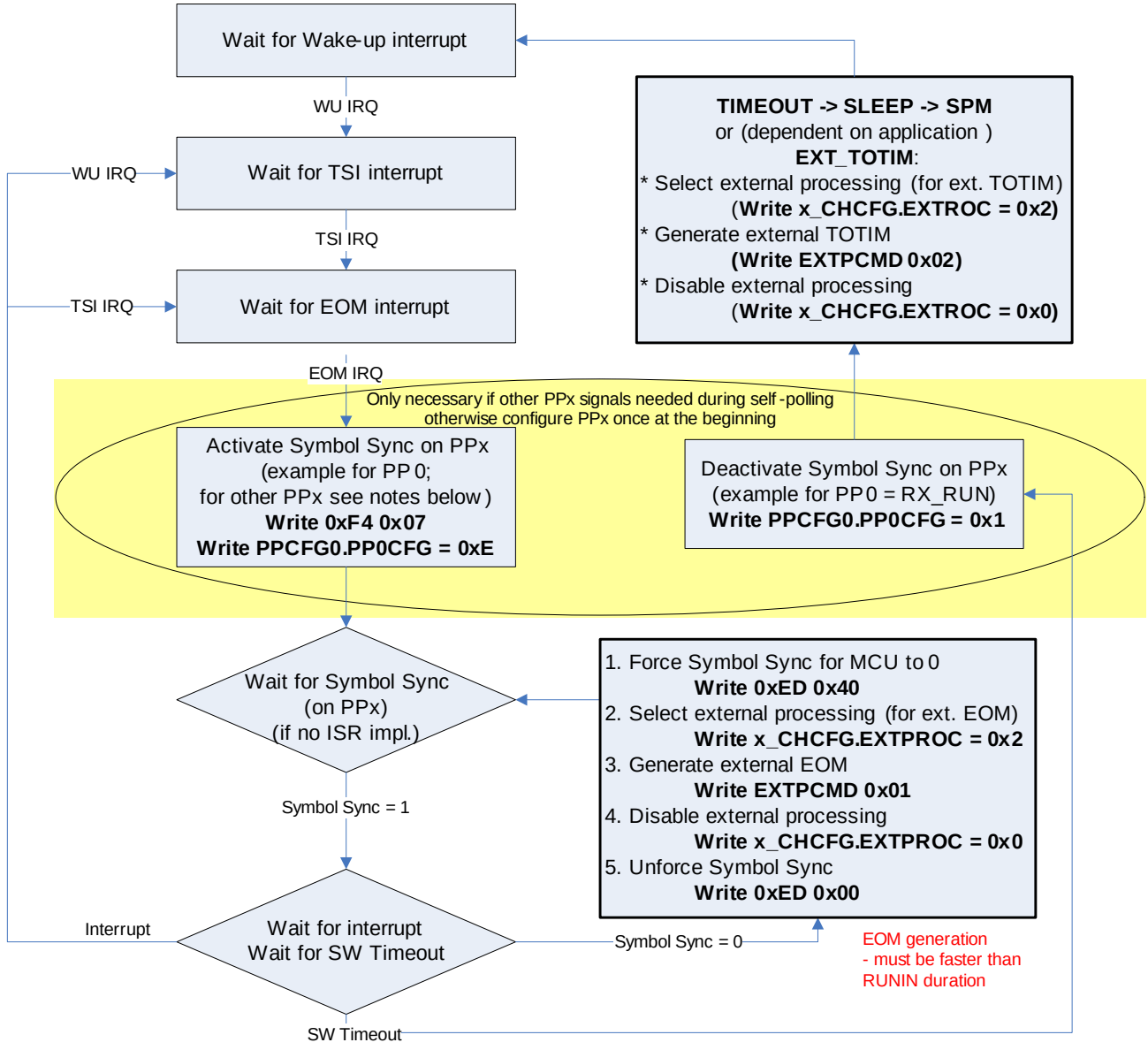


Figure 78 Parallel Wake-up Search

Notes:

- Symbol Sync can be activated on any PPx port

  PP0: Write 0xF4 0x07 & Write PPCFG0 0x0E

  PP1: Write 0xF4 0x70 & Write PPCFG0 0xE0

  PP2: Write 0xF5 0x01 & Write PPCFG1 0x0E

  PP3: Write 0xF5 0x10 & Write PPCFG1 0xE0

- Symbol Sync monitoring necessary only in run mode between frames and WU pattern or till software timeout generated

- generation of external EOM will reinitialize also the TOTIM timers
- external EOM generation period should be smaller than the RUNIN length
  (7 chips RUNIN = ~62 us @ 112 kchip/s , 5 SPI write commands = ~ 60 us @ 2 Mbit)
- minimal Symbol Sync active period = TVWIN, minimal Symbol Sync inactive period = RUNIN

For protocols where no ASK/FSK switching is required between the Wake-up and payload frame, the Wake-up and TSI pattern can share the same bits (e.g. Wake-up pattern = ..00000, TSI = 000001, all bits Manchester encoded). This function can be activated by the INITFRCS bit, so then there is no reset of the framer compare shift register after a Wake-up event, which can shorten the required processing time.

2.6.2 Polling Timer Unit

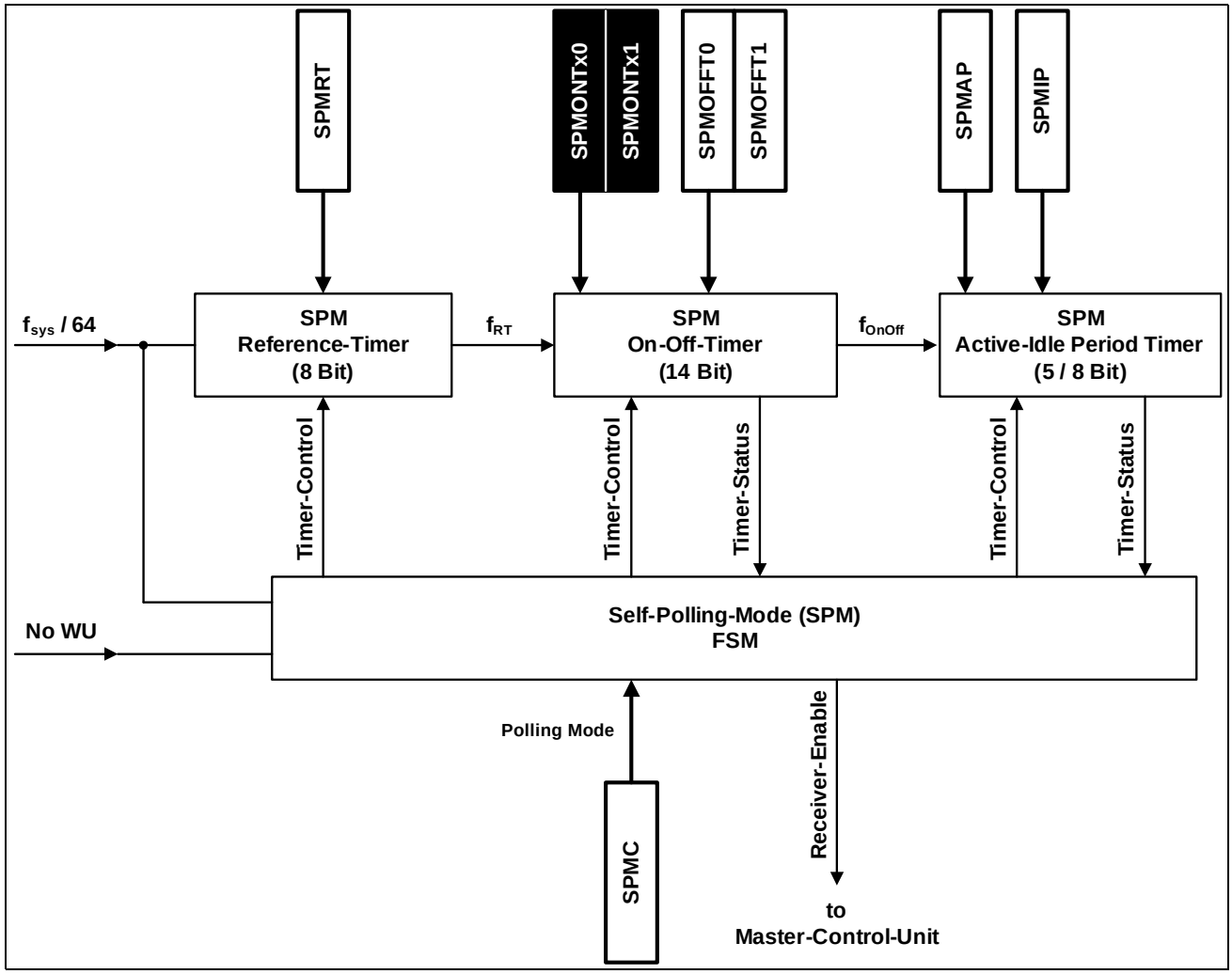


Figure 79 Polling Timer Unit

The Polling Timer Unit consists of a Counter Stage and a Control FSM (Finite State Machine).

The Counter Stage is divided into three sub-modules.

The **Reference Timer** is used to divide the state machine clock (f_{sys}/64) into the slower clock required for the SPM timers.

The **On-Off Timer** and the **Active Idle Period Timer** are used to generate the polling signal. The entire unit is controlled by the SPM FSM.

The TDA5235 is able to handle up to two different sets of configurations automatically.

2.6.2.1 Self Polling Modes

Four polling modes are available to fit the polling behavior to the expected wake-up patterns and to optimize power consumption in Self Polling Mode.

The following 4 Polling Modes are available and can be configured via 2 bits in the configuration register SPMC:

- Constant On-Off (COO)
- Fast Fall Back to SLEEP (FFB)
- Mixed Mode (MM)
- Permanent Wake-Up Search (PWUS)

A detected wake-up data sequence or an actual value for RSSI or Signal Recognition (a combination of Signal Detector and Noise Detector, see **Chapter 2.4.8.1**) exceeding a certain adjustable threshold forces the TDA5235 into Run Mode Self Polling.

In all modes the timing resolution is defined by the Reference Timer, which scales the incoming frequency (f_{sys}/64) corresponding to the value, which is defined in the Self Polling Mode Reference Timer (SPMRT) register. Changing values of SPMRT helps to fit the final On-Off timing to the calculated ideal timing.

2.6.2.2 Constant On-Off Time (COO)

In this mode there is a constant On and a constant Off time. Therefore also the resulting master period time is constant. The On and Off time are set in the SPMONTA0, SPMONTA1, SPMONTB0, SPMONTB1, SPMOFFT0 and SPMOFFT1 registers. The On time configuration is done separately for Configuration A and B.

When **Single-Configuration** is selected then only Configuration A is used. **Multi-Configuration** Mode allows reception of up to 2 different transmit sources. The diagram below shows possible scenarios.

All receive modes described in **Chapter 2.5.1.2 Data Interface** can be used.

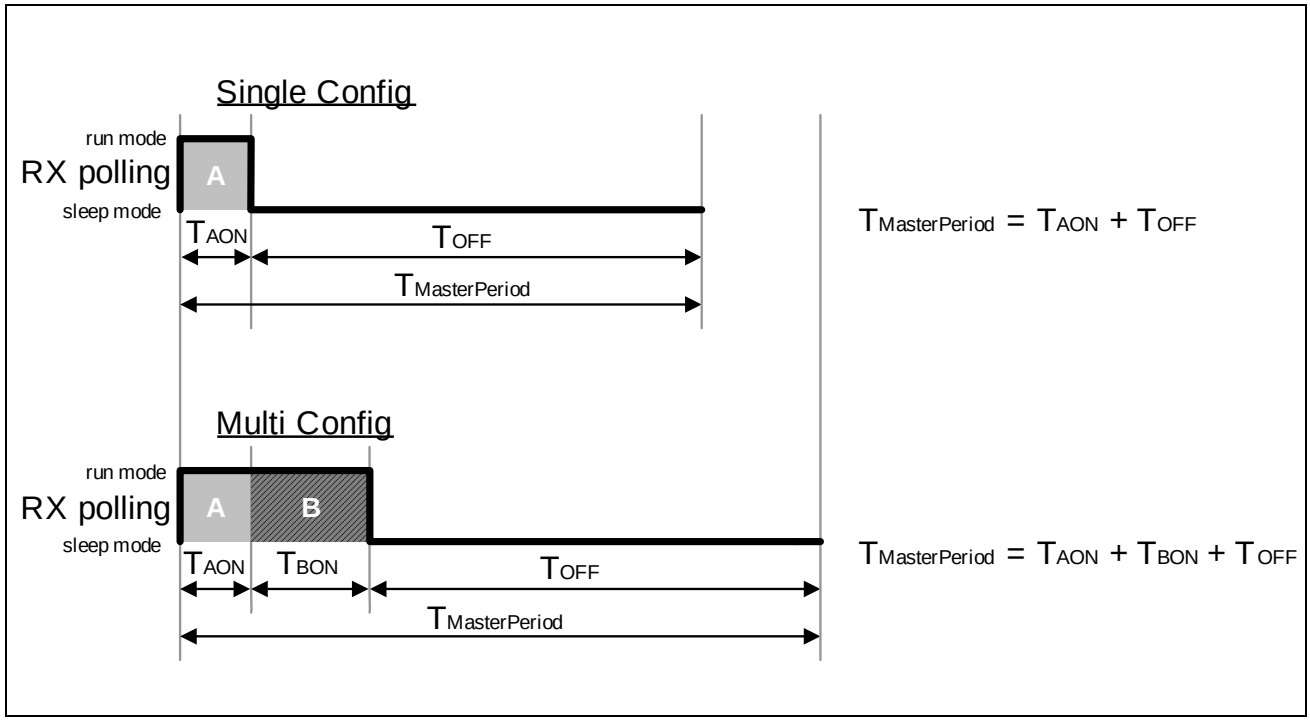


Figure 80 Constant On-Off Time

Calculation of the On time:

The On time must be long enough to ensure proper detection of a specified wake-up criterion. Therefore the On time depends on the wake-up pattern, and the wake-up criterion. It has to include transmitter data rate tolerances.

A widely used wake-up pattern is a sequence of equal Bi-phase encoded bits or a certain Bi-phase encoded bit pattern.

T_{ON} also must include the relevant start-up times. In case of the first configuration after T_{OFF}, this is the Receiver Start-Up Time. In case of the following Configuration B (RF Receiver is already on, there is only a change of the configuration), e.g. if Configuration B is used, this is the Configuration Change Latency Time. In addition, it has to be considered that some data bits are required for synchronization and internal latency, see **Chapter 2.4.8.8 RUNIN, Synchronization Search Time and Inter-Frame Time**.

There are other wake-up patterns in use as well, which have several (up to 10 and more) short wake-up sequences (a few byte each) that are separated by a certain pause (again a few byte each). In this case the On time has to be set, so that a possible wake-up can be found within two wake-up sequences including the pause in-between.

Calculation of the Off time:

The longer the Off time, the lower the average power consumption in Self Polling Mode. On the other hand, the Off time has to be short enough that no transmitted wake-up

pattern is missed. Therefore the Off time depends mainly on the duration of the expected wake-up pattern.

For basic timing of WU on RSSI in COO mode, please see **Figure 81**.

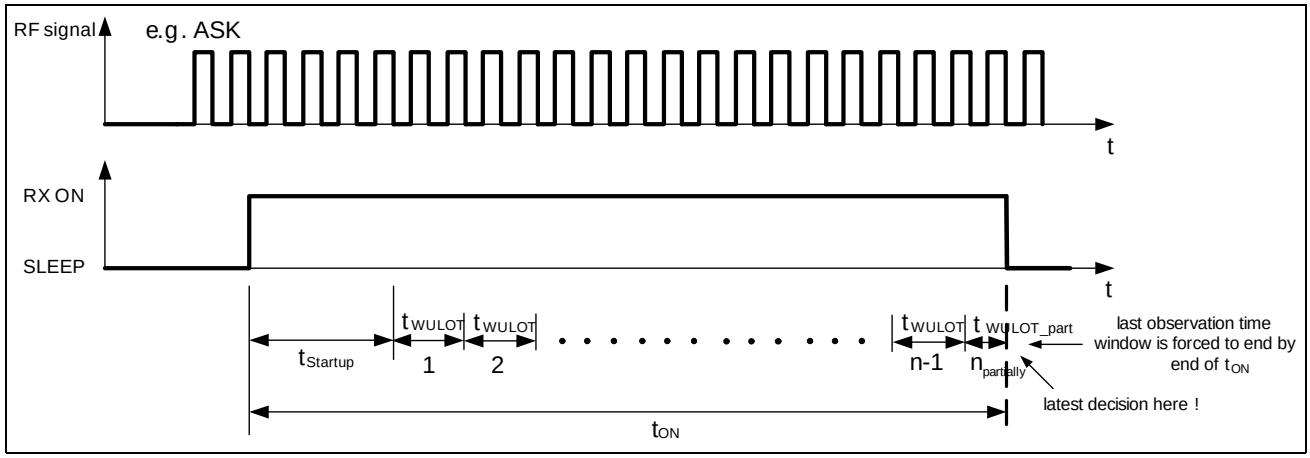


Figure 81 COO Polling in WU on RSSI Mode

Always check at the end of the current observation time window, if there is a WU (Wake-Up) event or NOT. This means, in algorithmic description (see also **Figure 10**, **Chapter 2.4.7 RSSI Peak Detector** and **Chapter 2.4.8.5 Wake-Up Generator**):

if (RSSIPWU_value > x_WURSSITH1) and (RSSIPWU_value > x_WURSSIBH1)

then WU

else NOT

Here, ‘NOT‘ means to keep on evaluating and move on to the next observation time window, also keep on peak value tracking of RSSIPWU signal. Keep on walking through the observation time windows until there is a WU event from the algorithm above or finally decide at the end of the On time with the following algorithm:

if (RSSIPWU_value > x_WURSSITH1) and (RSSIPWU_value < x_WURSSIBL1 or RSSIPWU_value > x_WURSSIBH1)

then WU

else NOT

If there is a WU event at the end of an observation time window while walking through the observation time windows, freeze/hold this decision/peak value in register RSSIPWU for optional read out and switch to run mode self polling.

Instead of the single RSSI criterion also the Signal Recognition criterion can be activated.

Combined Level and Data criterion in COO mode

On using the Wake-Up on Data criterion in COO mode, the RSSI criterion (including the RSSI blocking window) can be applied additionally by setting the bit x_WUC.UFFBLCOO. This means that a Wake-Up interrupt will not be generated, when a blocking RSSI level (e.g. an interfering signal) is detected even when the Data criterion is fulfilled.

The behavior of the additional RSSI criterion is similar to the behavior in Ultrafast Fall Back Mode.

After the level observation time the receiver checks, if the RSSI level is within a valid range. If RSSI is within a valid range, the state machine will go on to check the Data criterion. If the RSSI is within a forbidden range, a new level observation time is started (Note that no parts of the Wake-Up pattern are lost in this case, when the RSSI criterion succeeds within the following observation time).

This will be done as long as the RSSI value is within a forbidden range and the On time is not elapsed.

If the receiver loses synchronization within the search for the Data criterion (e.g. pattern detection), the WU unit will be initialized and checks again for the RSSI criterion.

Instead of the additional RSSI criterion also Signal Recognition criterion can be applied.

When the Signal Recognition threshold (x_WURSSIBH1) is not exceeded at Observation Time, the Wake-Up on Level FSM (finite state machine) and Wake-Up on Data FSM are initialized.

If the threshold is exceeded, then the Wake-Up on Level FSM enters the READY state and has no further impact on Wake-up search until the Wake-up unit is initialized again.

When afterwards a Data Criterion is found to be OK (e.g. pattern matches, number of equal bits or random bits is reached), the Wake-up search is completed positively.

When a Data Criterion is found to be not OK, the Wake-up search is terminated independent of the state of the Wake-Up on Level FSM. Therefore both FSMs are initialized.

2.6.2.3 Fast Fall Back to SLEEP (FFB)

This mode is used to switch off the receiver, if there is no RF signal, as quickly as possible to reduce power consumption.

During the search for wake-up data, there is a check for a bit stream, to which the system can be synchronized. If there is no synchronization to a bit stream within the so-called Sync Search Time Out (SYSRCTO), the wake-up search for this channel is stopped. If synchronization to a bit stream is possible (and not lost again), the TDA5235 waits if the wake-up criterion is fulfilled. If the wake-up criterion is not fulfilled (in worst case, if the last bit of an expected wake-up data pattern is wrong), the wake-up procedure for this configuration is stopped, and the TDA5235 tries to synchronize on the next configuration, or falls back to sleep. That means that the effective search time and, consequently, the receiver active time is significantly shorter, and power consumption is reduced, when no input signal is present. Calculation of Sync Search Time Out can be found in **Chapter 2.4.8.8 RUNIN, Synchronization Search Time and Inter-Frame Time**.

The needed time for detecting that no relevant transmission took place can be further reduced by using Ultrafast Fall Back to SLEEP (UFFB). When there was no Wake-up on Level criterion fulfilled in UFFB Mode during the Observation Time (T_{WULOT}, see **Chapter 2.4.8.5**), then the system goes back to SLEEP (or to next config). This can further reduce the receiver active time, when no data is available. When Wake-up on Level criterion was fulfilled, then the system proceeds with normal FFB functionality (SYSRCTO, optional Wake-up data criterion).

Note: UFFB and FFB start working at the same time!

Ultrafast Fall Back to SLEEP is working, when a Wake-up on Data criterion is selected, the UFFBLCOO bit is enabled and FFB or PWUS mode is selected. The UFFB level criterion can be selected in the x_WUC register.

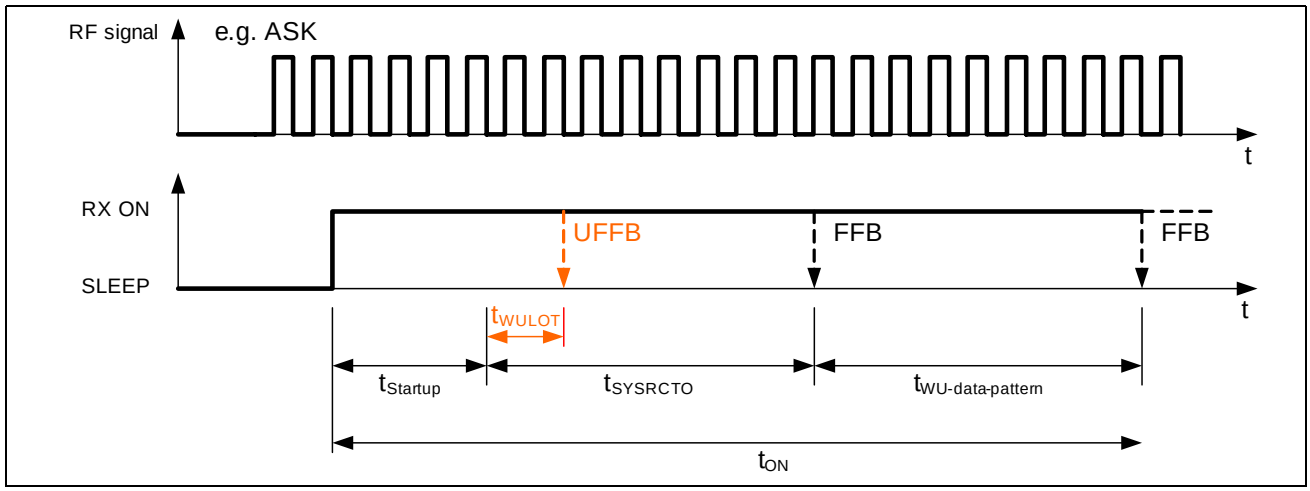


Figure 82 Ultrafast Fall Back to SLEEP

At the end of the observation time the RSSI peak tracking value of RSSIPWU signal is compared to the 3 thresholds. Then the decision is made. The algorithmic description is as follows (see also **Figure 10**, **Chapter 2.4.7 RSSI Peak Detector** and **Chapter 2.4.8.5 Wake-Up Generator**):

if (RSSIPWU_value > x_WURSSITH1) and (RSSIPWU_value < x_WURSSIBL1 or RSSIPWU_value > x_WURSSIBH1)

then WU

else NOT

Instead of the RSSI criterion also Signal Recognition criterion can be applied. When the Signal Recognition threshold (x_WURSSIBH1) is not exceeded at Observation Time, then the system goes back to SLEEP or the Wake-Up on Level FSM (finite state machine) is initialized and a Wake-up search is performed on the next specified configuration.

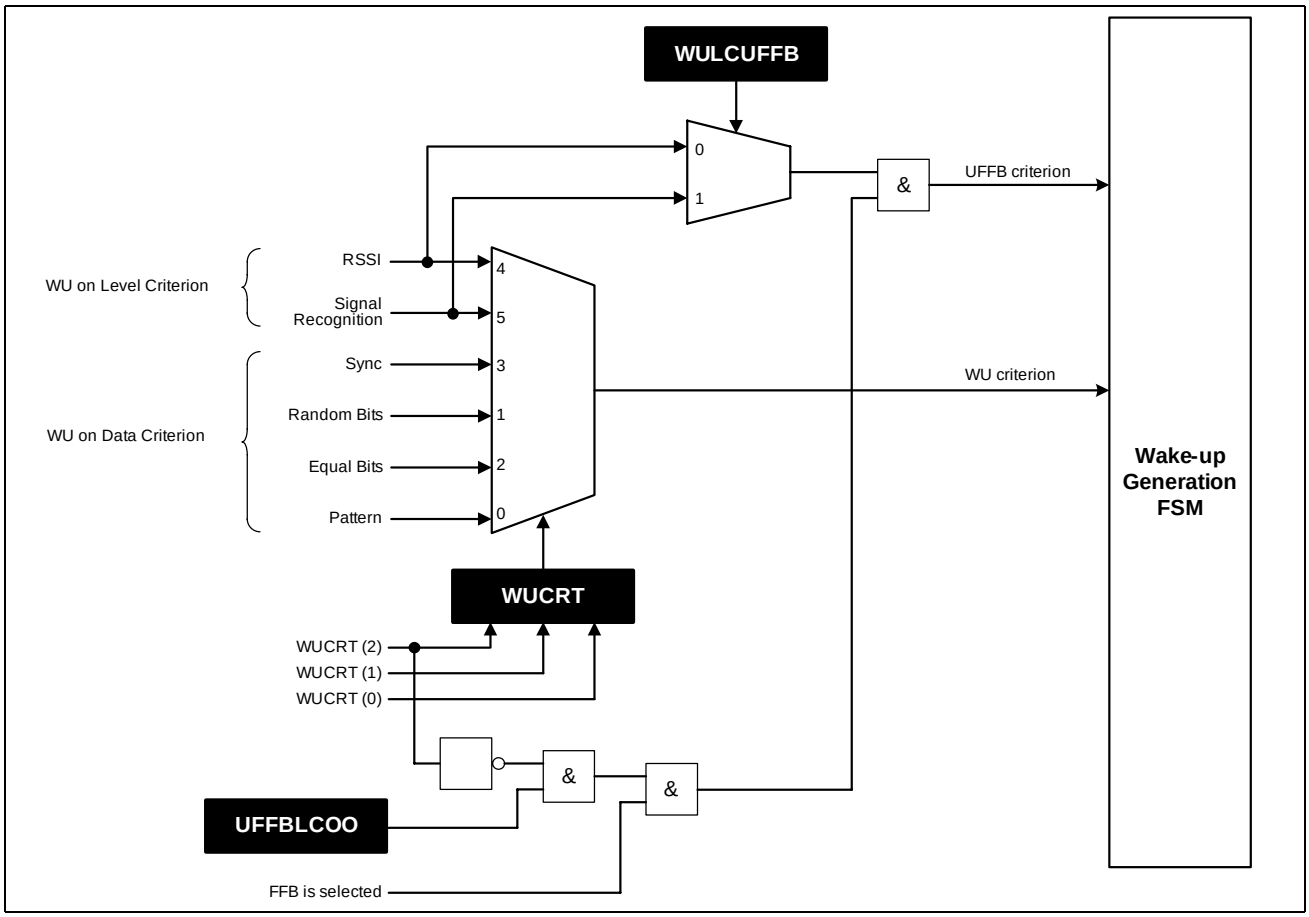


Figure 83 UFFB activation

The On and Off time setting is different from the Constant On-Off Time Mode. The entire On time is defined in the SPMONTA0 and SPMONTA1 registers. Regardless of the numbers of Configurations, the On time is defined with the Configuration A On-Timer. The deactivation of the receiver can happen at different times, but this event does not influence the timer stage, because the On time is still the same. So the master period is constant. The following scenarios are the same as before, but with Fast Fall Back to SLEEP.

Only the following receive modes (see **Chapter 2.5.1.2 Data Interface**) can be used:

- Packet Oriented FIFO Mode (POF)
- Packet Oriented Transparent Payload Mode (POTP)
- Transparent Mode - Chip Data and Strobe (TMCDS)

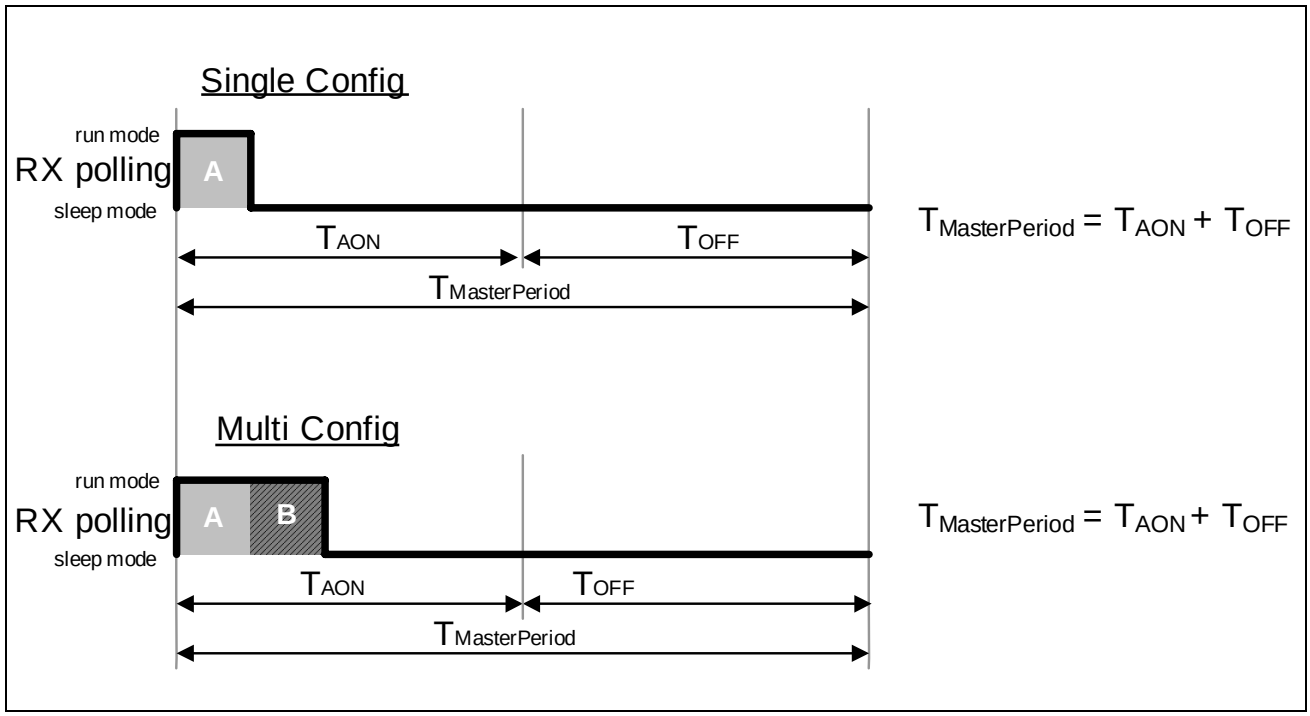


Figure 84 Fast Fall Back to SLEEP

Calculation of the On time:

The On time, which is now a sum for all of the configurations used, must include enough time to ensure proper detection of the specified wake-up pattern on all configurations. To cover the worst case scenario, the maximum time is required on all configurations as in Constant On-Off.

T_{ON} must also include the relevant start-up times. In case of the first configuration after T_{OFF}, this is the Receiver Start-Up Time. In case of the following Configuration B (RF Receiver is already on, there is only a change of the configuration), e.g. if Configuration B is used, this is the Configuration Change Latency Time.

In addition, it has to be considered that some data bits are required for synchronization and internal latency (see **Chapter 2.4.8.8 RUNIN, Synchronization Search Time and Inter-Frame Time**).

Calculation of the Off time:

The same general rules apply as for Constant On-Off Time. The Off time has to be short enough that no wake-up pattern reception is missed.

2.6.2.4 Mixed Mode (MM, Const On-Off & Fast Fall Back to SLEEP)

This mode combines Constant-On Time and Fast Fall Back to SLEEP within different configuration sets: Cfg.A: COO; Cfg.B: FFB

T_{ON} for Configuration A is always calculated according to Const On-Off rules.

T_{ON} for Configuration B is always calculated according to Fast Fall Back to SLEEP rules.

In Mixed Mode the On time of Configuration B is used for the FFB group. Below there are shown the same scenarios as before, but now for Mixed Mode. Note that Single-Configuration can be set, but is not recommended in Mixed Mode.

Only the following receive modes (see **Chapter 2.5.1.2 Data Interface**) can be used:

- Packet Oriented FIFO Mode (POF)
- Packet Oriented Transparent Payload Mode (POTP)
- Transparent Mode - Chip Data and Strobe (TMCDS)

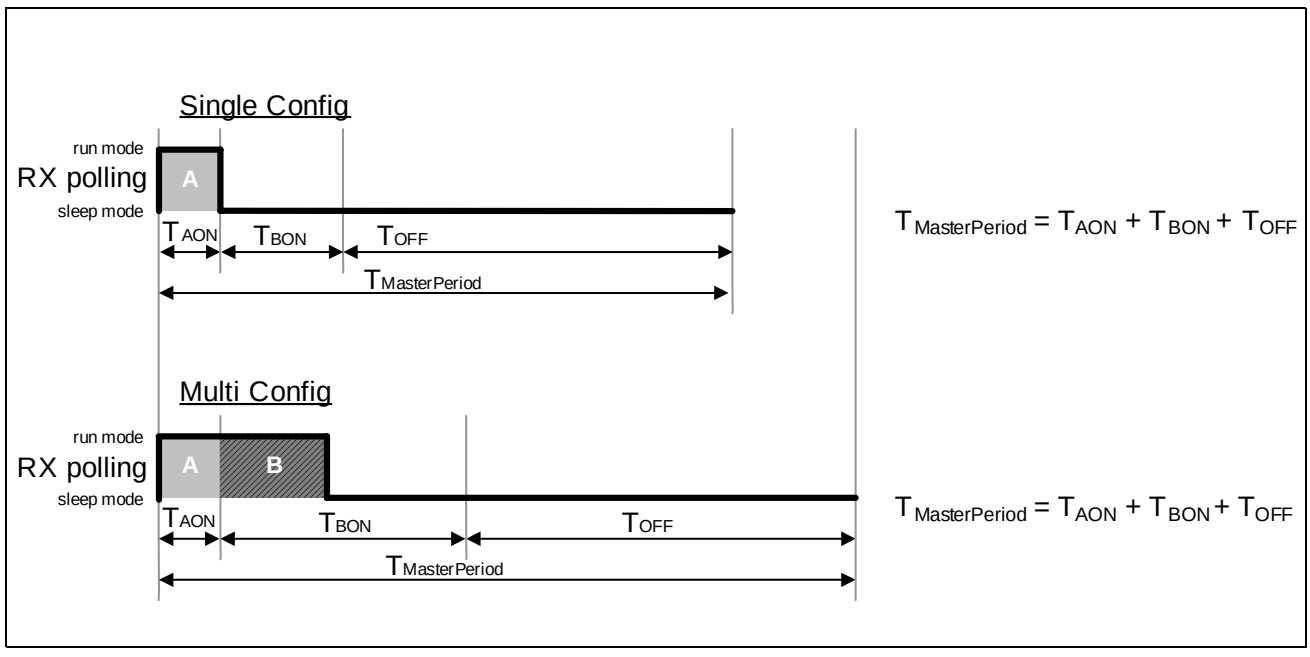


Figure 85 Mixed Mode

2.6.2.5 Permanent Wake-Up Search (PWUS)

In this mode the receiver will work in Fast Fall Back Mode, but it will not go back to the SLEEP state after the last configuration has been searched. Instead, it will start again from the beginning (Configuration A) until the On time has elapsed. The timing calculation can be seen in **Figure 86**. Ultrafast Fall Back to SLEEP functionality can be used as well.

Only the following receive modes (see **Chapter 2.5.1.2 Data Interface**) can be used:

- Packet Oriented FIFO Mode (POF)
- Packet Oriented Transparent Payload Mode (POTP)
- Transparent Mode - Chip Data and Strobe (TMCDS)

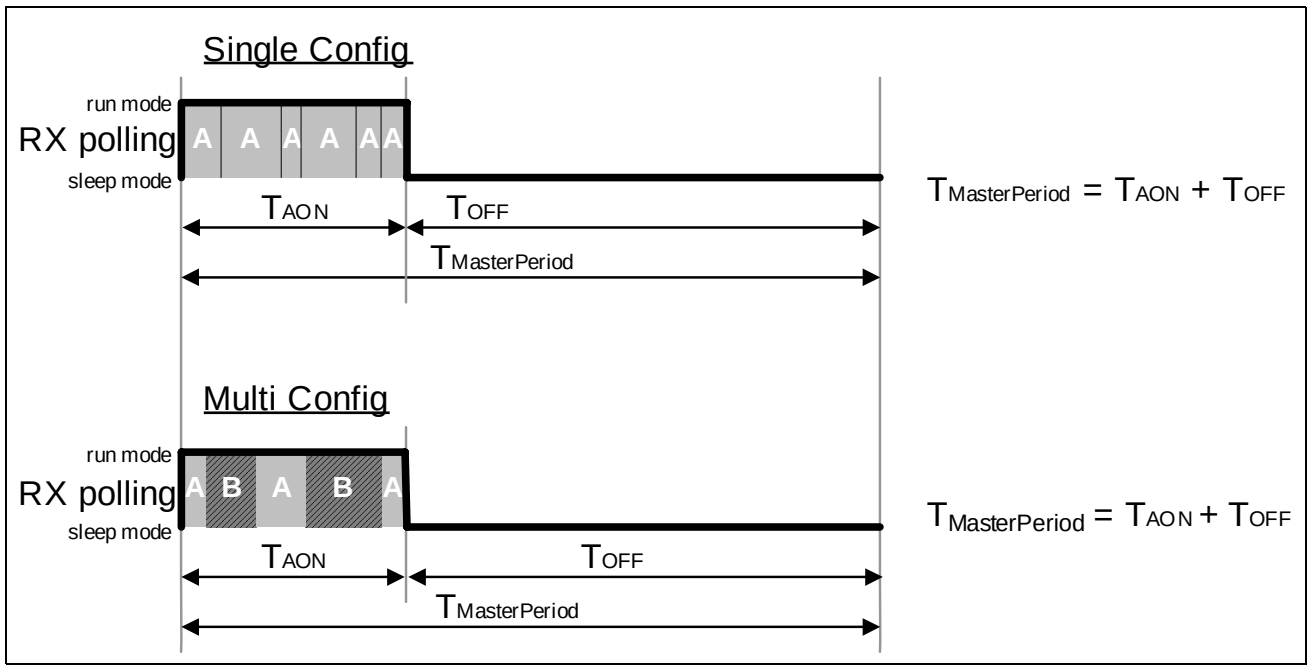


Figure 86 Permanent Wake-Up Search

2.6.2.6 Active Idle Period Selection

This mode is used to deactivate some polling periods and can additionally be applied to each of the above mentioned Polling Modes.

Normally, polling starts again after the $T_{MasterPeriod}$. With this Active Idle Period selection some of the polling periods can be deactivated, independent from the Polling Mode. The active and the idle sequence is set with the SPMAP and the SPMIP registers. The values of these registers determine the factor M and N.

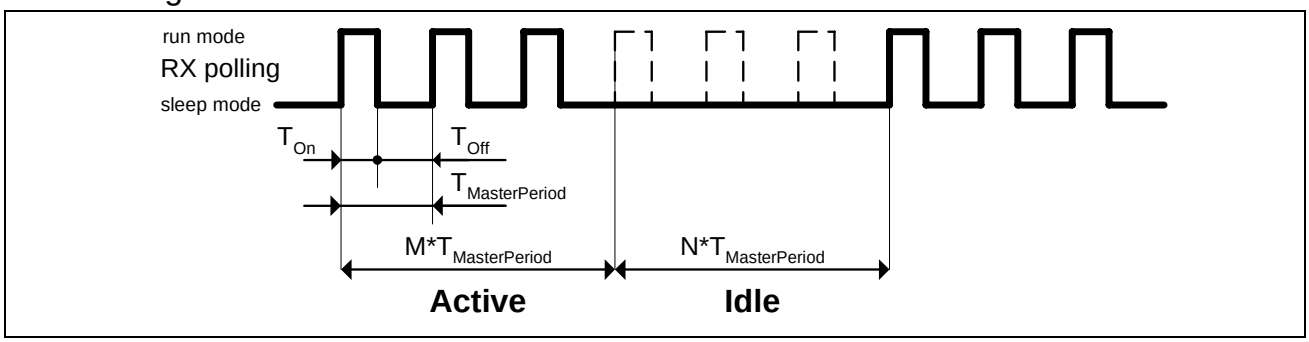


Figure 87 Active Idle Period

2.7 Definitions

2.7.1 Definition of Bit Rate

The definition for the bit rate in the following description is:

$$\text{bitrate} = \frac{\text{symbols}}{\text{s}}$$

If a symbol contains n chips (for Manchester n=2; for NRZ n=1) the chip rate is n times the bit rate:

$$\text{chiprate} = \text{n} \times \text{bitrate}$$

2.7.2 Definition of Manchester Duty Cycle

Several different definitions for the Manchester duty cycle *(MDC)* are in place. To avoid wrong interpretation some of the definitions are given below.

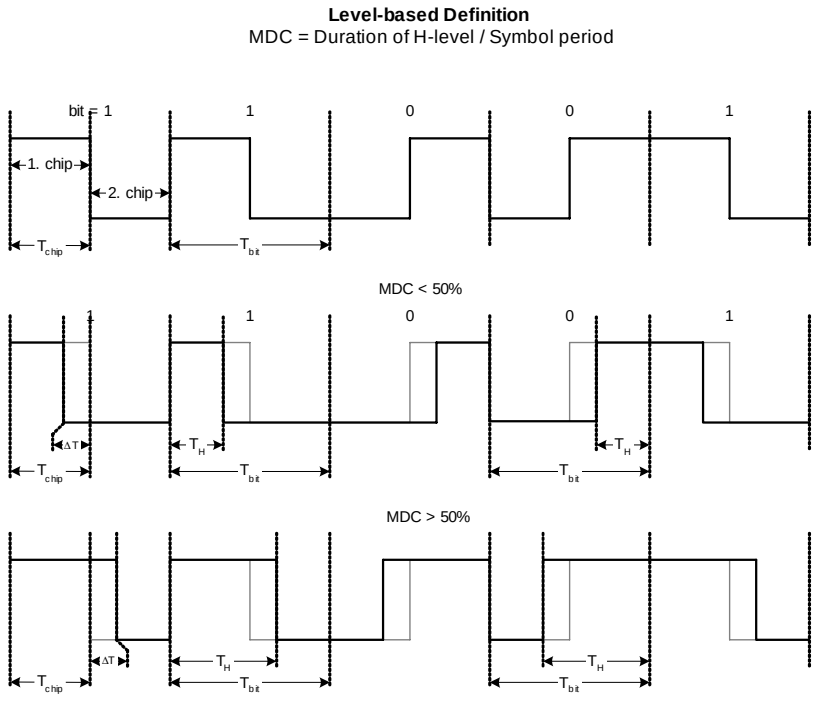


Figure 88 Definition A: Level-based definition

This definition determinates the duty cycle to be the ratio of the high pulse width and the ideal symbol period. The DC content is constant and directly proportional to the specified duty cycle.
For $\Delta T > 0$ the high period is longer than the chip-period and for $\Delta T < 0$ the high period is shorter than the chip-period.

Depending on the bit content, the same type of edge (e.g. rising edge) is sometimes shifted and sometimes not.

With this definition the Manchester duty cycle is calculated to

$$MDC_A = \frac{T_H}{T_{bit}} = \frac{T_{chip} + \Delta T}{T_{bit}}$$

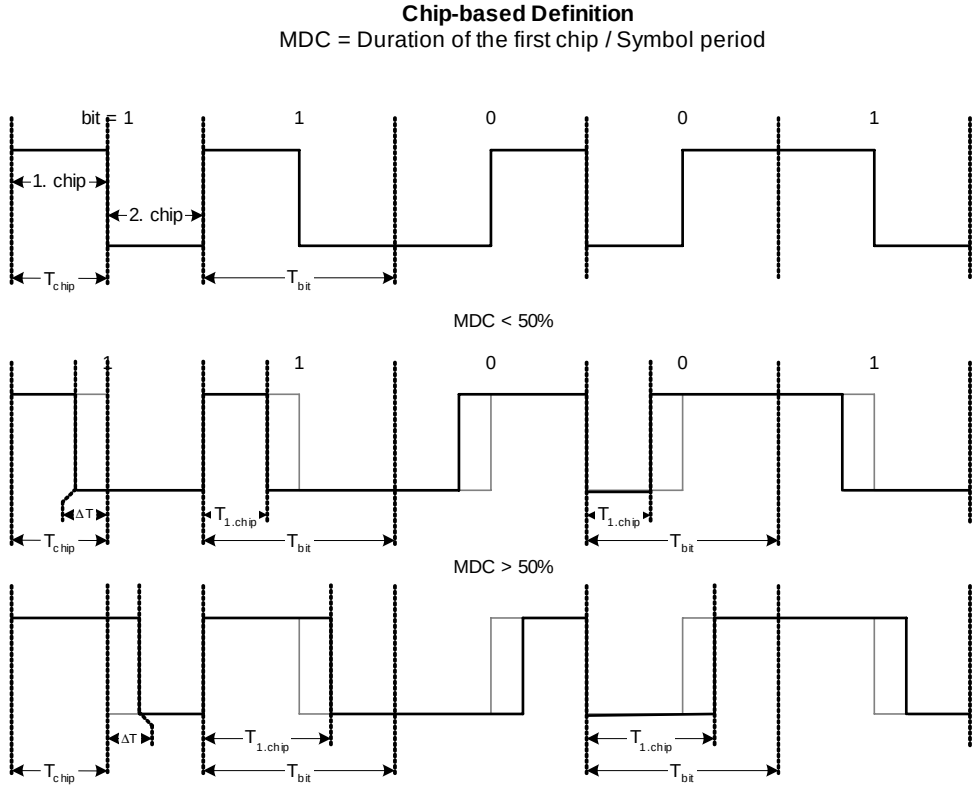


Figure 89 Definition B: Chip-based definition

This definition determinates the duty cycle to be the ratio of the first symbol chip and the ideal symbol period independently of the information bit content. The DC content depends on the information bit and it is balanced only if the message itself is balanced. For $\Delta T > 0$ the first chip-period is longer than the ideal chip-period and for $\Delta T < 0$ the first chip-period is shorter than the ideal chip-period.

Depending on the bit content, the same type of edge (e.g. rising edge) is sometimes shifted and sometimes not.

With this definition the Manchester duty cycle is calculated to

$$MDC_B = \frac{T_{1.chip}}{T_{bit}} = \frac{T_{chip} + \Delta T}{T_{bit}}$$

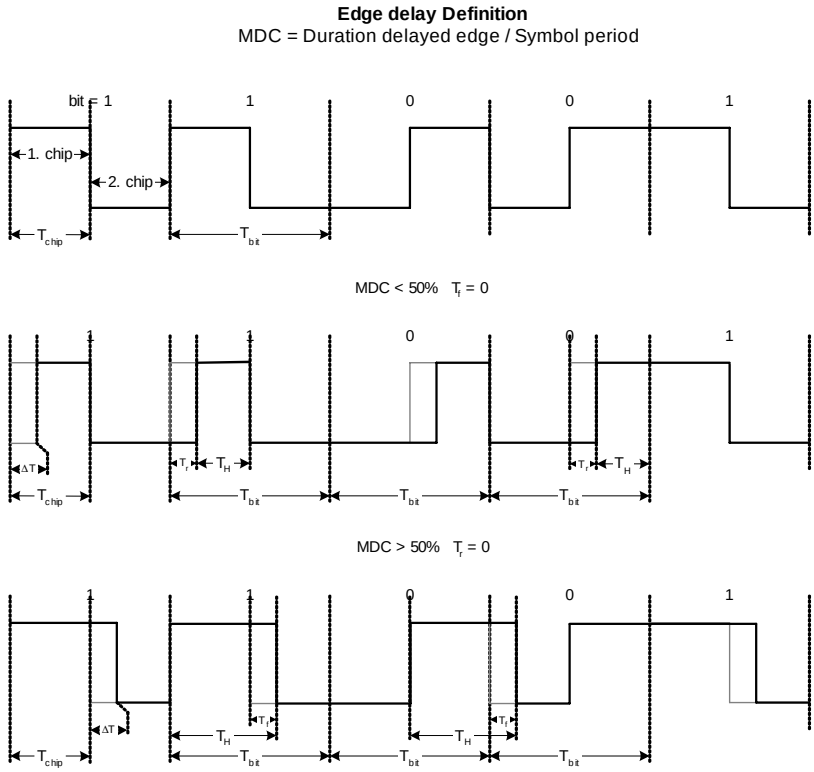


Figure 90 Definition C: Edge delay definition

This definition determinates the duty cycle to be the ratio of the duration of the delayed high-chip and the ideal symbol period independently of the information bit content. The position of the high-chip is determined by the delayed rising edge and/or the delayed falling edge. For $\Delta T = T_{fall} - T_{rise}$ the Manchester duty cycle is calculated to

$$MDC_C = \frac{T_{delayedHighchip}}{T_{bit}} = \frac{T_{chip} + \Delta T}{T_{bit}} = \frac{T_{chip} + T_{fall} - T_{rise}}{T_{bit}}$$

Independent on the bit content, the same type of edge (rising edge and/or falling edge) is shifted.

2.7.3 Definition of Power Level

The reference plane for the power level is the input of the receiver board. This means, the power level at this point (P_r) is corrected for all offsets in the signal path (e.g. attenuation of cables, power combiners etc.).

The specification value of power levels in terms of sensitivity is related to the peak power of P_r in case of On-Off Keying (OOK). This is noted by the unit dBm peak.

Specification value of power levels is related to a Manchester encoded signal with a Manchester duty cycle of 50% in case of ASK modulation.

An RF signal generator usually displays the level of the unmodulated carrier ($P_{carrier}$). This has following consequences for the different modulation types:

Table 6 Power Level

Modulation scheme	Realization with RF signal generator	Power level specification value
ASK	AM 100%	$P_r = P_{carrier} + 6dB$
ASK	Pulse modulation (=OOK)	$P_r = P_{carrier}$
FSK	FM with deviation Δf: $f_1 = f_{carrier} - \Delta f$ $f_2 = f_{carrier} + \Delta f$	$P_r = P_{carrier}$

For power levels in sensitivity parameters given as average power, this is noted by the unit dBm. Peak power can be calculated by adding 3 dB to the average power level in case of ASK modulation and a Manchester duty cycle of 50%.

2.7.4 Symbols of SFR Registers and Control Bits

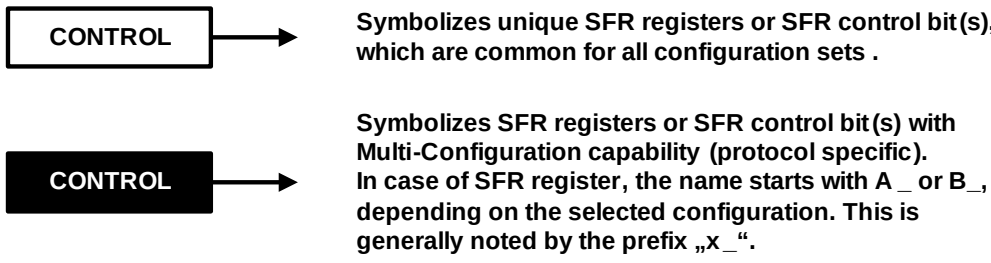
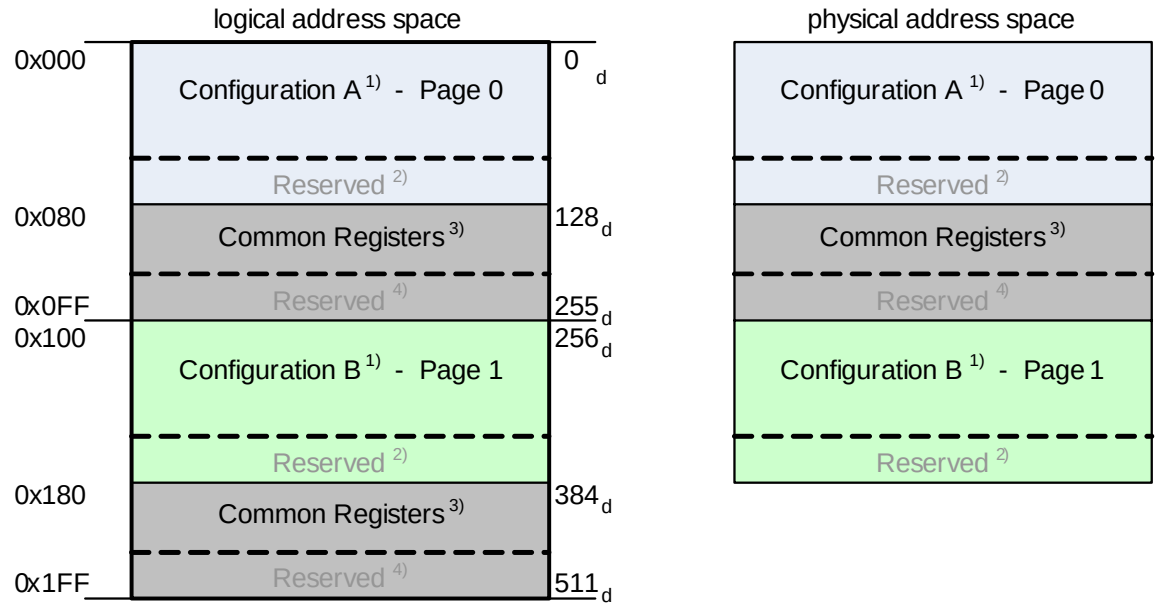


Figure 91 SFR Symbols

2.8 Digital Control (SFR Registers)

2.8.1 SFR Address Paging

An SPI instruction allows a maximum address space of 8 bit. The address space for supporting more than one configuration set is exceeding this 8 bit address room. Therefore a page switch is introduced, which can be applied via register SFRPAGE (see **Figure 92**).



1) Configuration dependent register block (2 protocol specific sets)
page switch via SFRPAGE register

2), 4) Reserved – Forbidden area

3) Configuration independent registers (common for all configurations)
map (“mirror“) to the same physical address space

Figure 92 SFR Address Paging

2.8.2 SFR Register List and Detailed SFR Description

The register list is attached in the Appendix at the end of the document.

Registers for Configuration B are equivalent and not shown in detail.

All registers with prefix “A_” are related to Configuration A. All these registers are also available for Configuration B having the prefix “B_”.

3 Applications

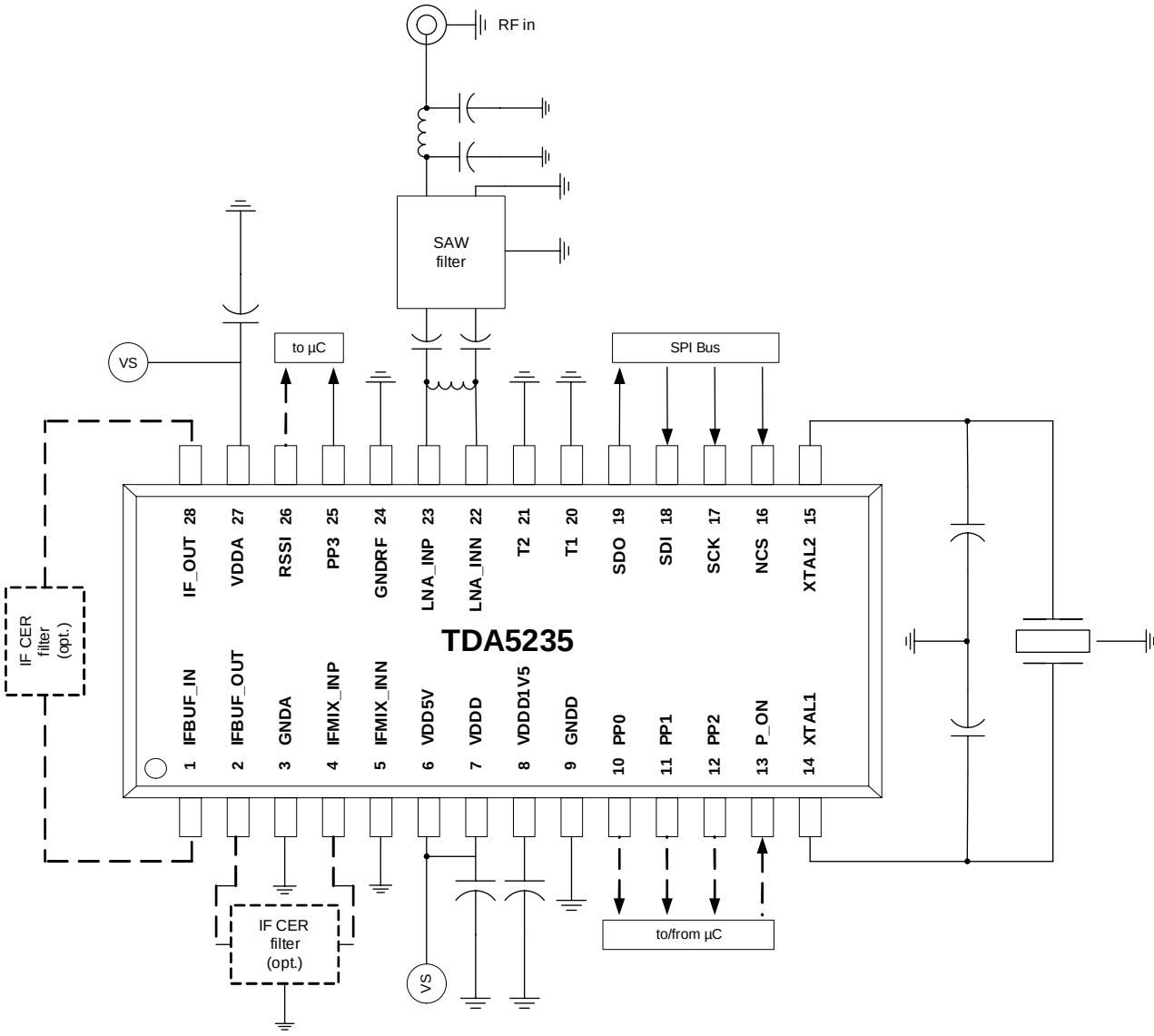


Figure 93 Typical Application Schematic

Note: As a good practice in any RF design, shielding around sensitive nodes can improve the EMC performance of the application.

For achieving the best sensitivity results the following has to be kept in mind. Every digital system generates certain frequencies (f_{SRC}, e.g. the crystal frequency or a microcontroller clock) and harmonics (N * f_{SRC}) of it, which can act as interferer (EMI source) and therefore sensitivity can be reduced.

There are two different cases, which need to be checked for the desired receive channel(s):

Elimination of in-band EMI mixing with (2*M + 1) * f_{LO}, where M > 0:

A square wave is used as LO (Local Oscillator) for the switching-type mixer, which also has odd harmonics. When the harmonics of the EMI source are exactly the IF frequency away from the harmonics of the LO, these spurs will be down-converted to the IF frequency and act as a co-channel interferer within the receiver's channel bandwidth mainly in the 315 MHz band.

In this case a change of the LO injection side (high side or low side injection) can be applied.

Example (Low Side LO-injection):

Wanted channel f_{RF} = 314.233MHz ==> f_{LO} = 303.533MHz ==> 3*f_{LO} = 910.599MHz

f_{XOSC} = 21.948717 MHz ==> 41 * f_{XOSC} = 899.8974 MHz

Resulting IF = 910.599 - 899.8974 MHz = 10.702 MHz ==> co-channel interferer within the receiver's channel bandwidth ==> change LO injection side

Example (High Side LO-injection):

Wanted channel f_{RF} = 314.233 MHz ==> f_{LO} = 324.933 MHz ==> 3*f_{LO} = 974.799 MHz

f_{XOSC} = 21.948717 MHz ==> 44 * f_{XOSC} = 965.744 MHz; 45 * f_{XOSC} = 987.692 MHz

==> both XOSC harmonics are not generating a co-channel interferer at 10.7 MHz

A final sensitivity measurement on the application hardware is recommended.

Elimination of in-band EMI mixing with 1 * f_{LO}:

Assuming a harmonic (N * f_{SRC}) is falling within the BW of the wanted channel and has an impact on the sensitivity there. In this case another XTAL frequency shall be selected, e.g. 10 kHz away

$$| N * f_{SRC} - f_{LocalOscillator} | < BW_{Channel}$$

Example (e.g. EMI source TDA5235 XOSC):

f_{XOSC} = 21.948717 MHz ==> 42 * f_{XOSC} = 921.846114 MHz

For further details please refer to the corresponding application note or to the latest configuration software.

3.1 Configuration Example

Please see configuration files supplied with the Explorer tool.

4 Reference

4.1 Electrical Data

4.1.1 Absolute Maximum Ratings

Attention: The maximum ratings must not be exceeded under any circumstances, not even momentarily and individually, as permanent damage to the IC may result.

Table 7 Absolute Maximum Ratings

#	Parameter	Symbol	Limit Values		Unit	Remarks
			min.	max.		
A1	Supply Voltage at VDD5V pin	V_{smax}	-0.3	+6	V	
A2	Supply Voltage at VDDD, VDDA pin	V_{smax}	-0.3	+4	V	
A3	Voltage between VDD5V vs VDDD and VDD5V vs VDDA	V_{smax}	-0.3	+4	V	
A4	Junction Temperature	T_j	-40	+125	°C	
A5	Storage Temperature	T_s	-40	+150	°C	
A6	Thermal resistance junction to air	$R_{th(ja)}$		140	K/W	
A7	Total power dissipation at T_{amb} = 105°C	P_{tot}		100	mW	
A8	ESD HBM integrity	V_{HBMRF}	-2	2	KV	According to ESD Standard JEDEC EIA / JESD22-A114-B
A9	ESD SDM integrity (All pins except corner pins)	V_{SDM}	-500	500	V	
A10	ESD SDM integrity (All corner pins)	V_{SDM}	-750	750	V	
A11	Latch up	I_{LU}	100		mA	AEC-Q100 (transient current)
A12	Maximum input voltage at digital input pins	V_{inmax}	-0.3	V_{DD5V}+0.5 or 6.0	V	whichever is lower
A13	Maximum current into digital input and output pins	I_{IOmax}		4	mA	

4.1.2 Operating Range

Table 8 Supply Operating Range and Ambient Temperature

#	Parameter	Symbol	Limit Values		Unit	Remarks
			min.	max.		
B1	Supply voltage at pin VDD5V	V_{DD5V}	4.5	5.5	V	Supply voltage range 1
B2	Supply voltage at pin VDD5V=VDDD=VDDA	V_{DD3V3}	3.0	3.6	V	Supply voltage range 2
B3	Ambient temperature	T_{amb}	-40	105	°C	

4.1.3 AC/DC Characteristics

Supply voltage VDD5V = 4.5 to 5.5 Volt or VDD5V = VDDA = VDDD = 3.0 to 3.6 Volt
Ambient temperature T_{amb} = -40...105°C; T_{amb} = +25°C and VDD5V = 5.0V or VDD5V = VDDA = VDDD = 3.3V for typical parameters, unless otherwise specified.

■ not subject to production test - verified by characterization/design

Table 9 AC/DC Characteristics

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
General DC Characteristics								
C1.1	Supply Current in Run Mode and Double Down Conversion Mode	$I_{Run, Double}$		12	15	mA	ASK or FSK mode P_{in} < -50dBm	
C1.2	Supply Current in Run Mode and Single Down Conversion Mode	$I_{Run, Single}$		10.5	14	mA	ASK or FSK mode P_{in} < -50dBm	
C2	Supply current in Sleep Mode	I_{sleep_low}					crystal oscillator in Low Power Mode; clock generator off; valid for SLEEP Mode and during SPM Off time	
	T_{amb} = 25 °C			40	50	µA		
	T_{amb} = 85 °C			60	110	µA		■
	T_{amb} = 105 °C			90	160	µA		■
C3	Supply current in Sleep Mode	I_{sleep_high}		115	350	µA	crystal oscillator in High Precision Mode C_{load} = 25 pF; clock generator off; valid for SLEEP Mode and during SPM Off time	
C4	Supply current in Power Down Mode	I_{PDN}						
	T_{amb} = 25 °C			0.8	1.5	µA		
	T_{amb} = 85 °C			3.7	13	µA		■
	T_{amb} = 105 °C			9.0	27	µA		■
C5	Supply current clock generator	I_{clock}		23	27	µA	$f_{clockout}$ = 1 kHz C_{load} = 10 pF	■
C6	Supply current IF-Buffer	I_{Buffer}		0.5	0.7	mA	f_{IF_1} = 10.7 MHz R_{load} = 330 Ω no AC signal	■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
C7	Supply current during RF-FE startup / BPF calibration	$I_{RF\text{-}FE\text{-}startup,BPFcal}$		2.2	2.9	mA		■
C8	Brownout detector threshold	V_{BOR}	2.3	2.45	2.6	V		
C9	Receiver reset time	t_{Reset}	1.0		3.0	ms	Note: No SPI communication is allowed before XOSC start-up is finished and chip reset is already finished	
C10	Receiver startup time	$t_{RXstartup}$	455	455	455	µs	Time to startup RF frontend (comprises time required to switch crystal oscillator from Low Power Mode to High Precision Mode	■
C11	RF Channel Hop Latency Time and Configuration (Hop) Change Latency Time (e.g. Cfg A to Cfg B)	t_{C_Hop}	111	111	111	µs	Time to switch RF PLL between different RF Channels (does not include settling of Data Clock Recovery) and time to change Configuration	■
C12	RF Frontend startup delay	$t_{RFstartdelay}$	350	350	350	µs	Delay of startup of RF frontend	■
C13	P_ON pulse width	t_{P_ON}	15			µs	Minimal pulse width to reset the chip	■
C14	NINT pulse length	t_{NINT_Pulse}		12		µs	Pulse width of interrupt	■
C15	Accuracy of Temperature Sensor						Valid for temperature range -40°C .. +105°C; using upper 8 ADC bits (ADCRESH)	
C15.1	uncalibrated	$T_{Error, uncal}$			+/- 23	°C	uncalibrated (3 sigma) value	■
C15.2	calibrated	$T_{Error, cal}$			+/- 4.5	°C	after 1-point calibration at room temperature (3 sigma)	■
C16	Accuracy of VDDD readout						Valid for temperature range -40°C .. +105°C; using upper 8 ADC bits (ADCRESH)	
C16.1	uncalibrated	$V_{DDD, Error, uncal}$			+/- 200	mV	uncalibrated (3 sigma) value	■
C16.2	calibrated	$V_{DDD, Error, cal}$			+/- 25	mV	after 1-point calibration at room temperature (3 sigma)	■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
General RF Characteristics (overall)								
D1	Frequency							
	Range 1	f_{band_1}	300		320	MHz	1st Local Oscillator Low Side LO-injection and High Side LO-injection allowed; See also **Chapter 3**	
	Range 2	f_{band_2}	425		450	MHz		
	Range 3	f_{band_3}	863		870	MHz		
	Range 4	f_{band_4}	902		928	MHz		
D2	Frequency step of Sigma-Delta PLL	f_{step}	10.5			Hz	$f_{step} = f_{XTAL} / 2^{21}$	■
D3	ASK Demodulation							
	Data Rate	R_{data}	0.5		40	kchip/s		■
	Data rate tol.	R_{data_tol}	-10		+10	%		■
	Modulation index	m_{ASK}	50		100	%	ASK	■
		m_{OOK}	99		100	%	ON-OFF keying	■
D4	FSK Demodulation							
	Data Rate	R_{data}	0.5		112	kchip/s	including tolerance	■
	Data rate tol.	R_{data_tol}	-10		+10	%		■
	Frequency deviation	Δf	1		64	kHz	frequency deviation zero-peak	■
	Modulation index	m_{FSK}	1.0				m = frequency_deviation$_{zero\text{-}peak}$ / maximum_occuring_data_frequency; m >= 1.25 is recommended at small frequency deviation	■
D5	Decoding schemes							
			Manchester, differential Manchester, Bi-phase Mark / Bi-phase Space					
	Duty cycle ASK	T_{chip}/T_{data}	35		55	%	see **Chapter 2.7.2** Definition C	■
	Duty cycle FSK	T_{chip}/T_{data}	45		55	%	see **Chapter 2.7.2** Definition B	■
D6	Overall noise figure						RF input matched to 50 Ω @ T_{amb} = 25 °C	
	Noise figure	NF		6	8	dB		■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
D7	BER Sensitivity (FSK)						BER = $2*10^{-3}$ RF input matched to 50 Ω @ T_{amb} = 25 °C; Single-Ended Matching without SAW; Insertion loss of input matching network = 1dB; Receive Mode = TMMF (sampled with ideal data clock); Double Down Conversion	
		Manchester coding; for additional test conditions see right after this table						
D7.1	Data Rate 2 kBit/s; Δf = 10 kHz	$SFSK1_{BER}$		-119	-116	dBm	2^{nd} IF BW = 50 kHz PDF = 33 kHz, AFC off, IFATT=0	■
D7.2	Data Rate 10 kBit/s; Δf = 14 kHz	$SFSK2_{BER}$		-114	-111	dBm	2^{nd} IF BW = 50 kHz PDF = 65 kHz, AFC off, IFATT=0	■
D7.3	Data Rate 10 kBit/s; Δf = 50 kHz	$SFSK3_{BER}$		-112	-109	dBm	2^{nd} IF BW = 125 kHz PDF = 132 kHz, AFC off, IFATT=0	■
D7.4	Data Rate 50 kBit/s; Δf = 50 kHz	$SFSK4_{BER}$		-105	-102	dBm	2^{nd} IF BW = 300 kHz PDF = 239 kHz, AFC off, IFATT=0	■
D7.5	Data Rate 2 kBit/s; Δf = 10 kHz	$SFSK5_{BER}$		-110	-107	dBm	2^{nd} IF BW = 300 kHz PDF = 282 kHz, IFATT=7 Note: 3dB sensitivity loss @ f_{offset}=+/-90kHz @ AFC on	■
D7.6	Data Rate 10 kBit/s; Δf = 14 kHz	$SFSK6_{BER}$		-106	-103	dBm	2^{nd} IF BW = 300 kHz PDF = 282 kHz, IFATT=7 Note: 3dB sensitivity loss @ f_{offset}=+/-90kHz @ AFC on	■
D7.7	Data Rate 10 kBit/s; Δf = 50 kHz	$SFSK7_{BER}$		-110	-107	dBm	2^{nd} IF BW = 300 kHz PDF = 282 kHz, IFATT=7 Note: 3dB sensitivity loss @ f_{offset}=+/-90kHz @ AFC on	■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
D8	BER Sensitivity (OOK)						BER = $2*10^{-3}$ RF input matched to 50 Ω @ T_{amb} = 25 °C, peak power level (see **Chapter 2.7.3**); Single-Ended Matching without SAW; Insertion loss of input matching network = 1dB; Receive Mode = TMMF (sampled with ideal data clock); Double Down Conversion	
		Manchester coding; for additional test conditions see right after this table						
D8.1	Data Rate 0.5 kBit/s	$SASK1_{BER}$		-120	-117	dBm peak	m = 100%, IFATT=0 2nd IF BW = 50 kHz	■
D8.2	Data Rate 2 kBit/s	$SASK2_{BER}$		-116	-113	dBm peak	m = 100%, IFATT=0 2nd IF BW = 50 kHz	■
D8.3	Data Rate 10 kBit/s	$SASK3_{BER}$		-111	-108	dBm peak	m = 100%, IFATT=0 2nd IF BW = 50 kHz	■
D8.4	Data Rate 16 kBit/s	$SASK4_{BER}$		-109	-106	dBm peak	m = 100%, IFATT=0 2nd IF BW = 80 kHz	■
D8.5	Data Rate 0.5 kBit/s	$SASK5_{BER}$		-115	-112	dBm peak	m = 100%, IFATT=7 2nd IF BW = 300 kHz; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■
D8.6	Data Rate 2 kBit/s	$SASK6_{BER}$		-112	-109	dBm peak	m = 100%, IFATT=7 2nd IF BW = 300 kHz; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■
D8.7	Data Rate 10 kBit/s	$SASK7_{BER}$		-106	-103	dBm peak	m = 100%, IFATT=7 2nd IF BW = 300 kHz; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■
D8.8	Data Rate 16 kBit/s	$SASK8_{BER}$		-104	-101	dBm peak	m = 100%, IFATT=7 2nd IF BW = 300 kHz; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■
D9.1	Sensitivity increase for Single Down Conversion mode	ΔS_{SDC}	0	0.5	1	dB		■
D9.2	Double Down Conversion sensitivity decrease for higher blocking performance (IFATT=0 => IFATT=7)	$\Delta S_{DDC, IFATT7}$		1	2	dB		■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
D9.3	Single Down Conversion sensitivity decrease for higher blocking performance (IFATT=4 => IFATT=7)	$\Delta S_{SDC, IFATT7}$		0.5	1	dB		■
D10.1	Sensitivity variation due to temperature (-40...+105°C)	ΔP_{in}			2	dB	relative to T_{amb} = 25 °C; temperature drift of crystal not considered	■
D10.2	Sensitivity variation due to frequency offset [1)]	ΔP_{in}			3	dB	AFC inactive; For Sensitivity Bandwidth see **Table 11**	■
D10.3	Sensitivity variation due to frequency offset	ΔP_{in}			3	dB	AFC active, slow AFC; For Sensitivity Bandwidth see **Table 11** and applied AFCLIMIT	■
D10.4	Sensitivity loss when AFC active at center frequency	ΔP_{in}			1	dB	AFC active; center frequency - no AFC wander (see **Chapter 2.4.6.3**)	■
D11	3rd order intercept IIP3	P_{IIP3}	-16	-14		dBm	input matched to 50 Ω; Insertion loss of input matching network = 1dB; IFATT = 7; valid for Single and Double Down Conversion Mode	■
D12	1 dB compression point CP1dB	P_{CP1dB}	-27	-25		dBm	input matched to 50 Ω; Insertion loss of input matching network = 1dB; IFATT = 7; valid for Single and Double Down Conversion Mode	■
D13	1st IF image rejection	d_{image1}	30	40		dB	1st IF = 10.7 MHz without front end SAW filter; valid for Double Down Conversion Mode	
D14	2nd IF image rejection	d_{image2}	30	34		dB	2nd IF = 274 kHz without 1st IF CER filter; valid for Single and Double Down Conversion Mode	

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
RF Front End Characteristics (Unless otherwise noted, all values apply for the specified frequency ranges)								
E1	LNA input impedance							
E1.1	f_{RF} = 315 MHz	$R_{in_p,diff}$		680		Ω	differential parallel equivalent input between LNA_INP and LNA_INN	■
		$C_{in_p,diff}$		1.05		pF		■
E1.2	f_{RF} = 434MHz	$R_{in_p,diff}$		570		Ω		■
		$C_{in_p,diff}$		0.87		pF		■
E1.3	f_{RF} = 868MHz	$R_{in_p,diff}$		550		Ω		■
		$C_{in_p,diff}$		0.63		pF		■
E1.4	f_{RF} = 915MHz	$R_{in_p,diff}$		540		Ω		■
		$C_{in_p,diff}$		0.63		pF		■
E1.5	f_{RF} = 315 MHz	$R_{in_p,\ SE}$		500		Ω	single-ended parallel equivalent input between LNA_INP and GNDRF / LNA_INN and GNDRF	■
		$C_{in_p,\ SE}$		1.87		pF		■
E1.6	f_{RF} = 434MHz	$R_{in_p,\ SE}$		400		Ω		■
		$C_{in_p,\ SE}$		1.63		pF		■
E1.7	f_{RF} = 868MHz	$R_{in_p,\ SE}$		322		Ω		■
		$C_{in_p,\ SE}$		1.59		pF		■
E1.8	f_{RF} = 915MHz	$R_{in_p,\ SE}$		312		Ω		■
		$C_{in_p,\ SE}$		1.56		pF		■
E2	FE output impedance	R_{out_IF}	290	330	380	Ω	f_{IF} = 10.7 MHz	■
E3	FE voltage conversion gain	$AV_{FE,\ max}$	34	36	38	dB	min. IF attenuation (IFATT = 0); input matched to 50 Ω; Insertion loss of input matching network = 1dB R_{load_IF} = 330 Ω; tested at 434 MHz	
E4	FE voltage conversion gain	AV_{FE_7}	29	31	33	dB	IF attenuation (IFATT = 7); input matched to 50 Ω; Insertion loss of input matching network = 1dB R_{load_IF} = 330 Ω; tested at 434 MHz	

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
E5	FE voltage conversion gain	$AV_{FE, min}$	22	24	26	dB	max. IF attenuation (IFATT = 15); input matched to 50 Ω; Insertion loss of input matching network = 1dB R_{load_IF} = 330 Ω; tested at 434 MHz	
E6	FE voltage conversion gain step			0.8		dB	12dB / 15 = 0.8dB/step Double Down Conversion: 16 gain settings (4 bit) Single Down Conversion: 7 gain settings	■
E7	1^{st} Local Oscillator SSB Noise						closed loop	
E7.1	PLL loop Bandwidth	BW	100	150	200	kHz	BW and its tolerances	■
E7.2	f_{in_R1} = 315MHz	d_{SSB_LO}		-81	-76	dBc/Hz	@ f_{offset} = 1 kHz	■
				-85	-80		@ f_{offset} = 10 kHz	■
				-82	-77		@ f_{offset} = 100 kHz	■
				-120	-115		@ f_{offset} = 1 MHz	■
				-130	-125		@ f_{offset} => 10 MHz	■
E7.3	f_{in_R2} = 434MHz	d_{SSB_LO}		-78	-73	dBc/Hz	@ f_{offset} = 1 kHz	■
				-83	-78		@ f_{offset} = 10 kHz	■
				-82	-77		@ f_{offset} = 100 kHz	■
				-117	-112		@ f_{offset} = 1 MHz	■
				-130	-125		@ f_{offset} => 10 MHz	■
E7.4	f_{in_R3} = 868MHz	d_{SSB_LO}		-75	-70	dBc/Hz	@ f_{offset} = 1 kHz	■
				-79	-74		@ f_{offset} = 10 kHz	■
				-77	-72		@ f_{offset} = 100 kHz	■
				-114	-109		@ f_{offset} = 1 MHz	■
				-130	-125		@ f_{offset} => 10 MHz	■
E7.5	f_{in_R4} = 915MHz	d_{SSB_LO}		-71	-66	dBc/Hz	@ f_{offset} = 1 kHz	■
				-79	-74		@ f_{offset} = 10 kHz	■
				-77	-72		@ f_{offset} = 100 kHz	■
				-116	-111		@ f_{offset} = 1 MHz	■
				-130	-125		@ f_{offset} => 10 MHz	■
E8.1	Spurious emission < 1 GHz				-57	dBm		■
E8.2	Spurious emission > 1 GHz				-47	dBm		■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
E9	Inband fractional spur		-40			dBc		■
E10	3dB Overall Analog Frontend Bandwidth	BW_{ANA}		230		kHz	LNA input to Limiter output, excluding external CER filter	■
1st IF Buffer Characteristics								
F1	Input impedance	R_{in_IF}	290	330	370	Ω	f_{IF} = 10...12 MHz	■
F2	Output impedance	R_{out_IF}	290	330	370	Ω	f_{IF} = 10...12 MHz	■
F3	Voltage gain	AV_{Buffer}	3	4	5	dB	f_{IF} = 10...12 MHz Z_{source} = 330 Ω Z_{load} = 330 Ω	
F4	Buffer switch isolation (CERFSEL)	$d_{isolation}$	60			dB	f_{IF} = 10...12 MHz see **Figure 6**	■
2nd IF Mixer, RSSI and Filter Characteristics								
G1	Mixer input impedance	R_{in_IF}	290	330	390	Ω	f_{IF} = 10...12 MHz	■
G2	RSSI						Related to RF input matched to 50 Ω	
G2.1	Dynamic range (Linearity +/- 2 dB)	DR_{RSSI}	-110		-30	dBm	applies for digital RSSI; AGC on	■
			-115		-60	dBm	applies for analog RSSI @ 50kHz BPF, AGC off	■
			-110		-50	dBm	applies for analog RSSI @ 300kHz BPF, AGC off	■
G2.2	Linearity	DR_{LIN}	-1		+1	dB	-95 dBm...-35 dBm; applies for digital RSSI	■
G2.3	Temperature drift within linear dynamic range	DR_{TEMP}	-2.5		+1.5	dB	-95 dBm...-35 dBm; applies for digital RSSI	■
G2.4	Output dynamic range	V_{RSSI+}	0.8		2.0	V		
G2.5	analog RSSI error, untrimmed	$DRSSI_{ana}$	-4		+2	dB	at RSSI pin	
G2.6	analog RSSI slope, untrimmed	dV_{RSSI}/ dV_{mix_in}	8	10	12	mV/dB	at RSSI pin; typical 600 mV/60 dB = 10 mV/dB	
G2.7	digital RSSI error, untrimmed	$DRSSI_{dig_u}$	-4		+2	dB	RSSI register readout	

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
G2.8	digital RSSI error, user trimmed via SFRs RSSISLOPE and RSSIOFFS	$DRSSI_{dig_t}$	-1		+1	dB	RSSI register readout	■
G2.9	digital RSSI slope, untrimmed	dV_{RSSI}/dV_{mix_in}	2	2.5	3	LSB /dB	RSSI register readout; typical 600 mV/60 dB = 10 mV/dB, 1mV = 1 LSB (10-bit ADC) 8-bit readout: 4mV=1LSB	
G2.10	digital RSSI slope, user trimmed via SFRs RSSISLOPE and RSSIOFFS	dV_{RSSI}/dV_{mix_in}	2.35	2.5	2.65	LSB /dB	RSSI register readout; typical 600 mV/60 dB = 10 mV/dB, 1mV = 1 LSB (10-bit ADC) 8-bit readout: 4mV=1LSB	■
G2.11	Resistive load at RSSI pin	$R_{L,RSSImax}$	100			kΩ		■
G2.12	Capacitive load at RSSI pin	$C_{L,RSSI}$			20	pF		■
G3	2nd IF Filter (3rd order Bandpass Filter)							
G3.1	Center frequency	f_{center}	262	274	288	kHz	Asymmetric BPF corners: f_center=sqrt(f_{low} * f_{high}); Use AFC for more symmetry	
G3.2	-3 dB BW	BW_{-3dB}		50 80 125 200 300		kHz		■
G3.3	-3 dB BW tolerance	tol_BW_{-3dB}	-5		+5	%	For BW = 125, 200, 300 kHz	■
G3.4	-3 dB BW tolerance	tol_BW_{-3dB}	-6		+6	%	For BW = 50, 80 kHz	■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
Crystal Oscillator Characteristics								
H1	Frequency range	f_{XTAL}		21.948 717		MHz		
H2	Crystal parameters							
H2.1	Motional capacitance	C_1	3	6	10	fF		■
H2.2	Motional resistance	R_1		18	80	Ω		■
H2.3	Shunt capacitance	C_0		2	4	pF		■
H2.4	Load capacitance	C_{Load}		12		pF	nominal value	■
H2.5	Initial frequency tolerance	f_{XTAL_Tol}	-30		+30	ppm	oscillator untrimmed (trim capacitor default settings, usage of recommended crystal); not including crystal tolerances	■
H2.6	Frequency trimming range	Δf_{XTAL}	-50		+50	ppm	larger trimming range possible via SD PLL	
H2.7	Trimming step	Δf_{X_step}		1	4	ppm	see also step size of SD PLL	■
H3	Clock output frequency at PPx pin	f_{clock_out}	11		5.5M	Hz	10pF load	
H4	Crystal oscillator settling time (switching from Low Power to High Precision Mode)	$t_{COSCsettle}$	292	292	292	µs		■
H5	Start up time	t_{start_up}		0.45	1	ms	crystal type: NDK NX5032SD; See also BOM for ext. load caps; Note: No SPI communication is allowed before XOSC start-up is finished and chip reset is already finished	

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
Digital Inputs/Outputs Characteristics								
I1	High level input voltage	V_{In_High}	0.7* VDDD		VDD5V +0.1	V		
I2	High level input leakage current	I_{In_High}			5	µA		
I3	Low level input voltage (except P_ON pin)	V_{In_Low}	0		0.8	V		
I4	Low level input voltage (at P_ON pin)	$V_{In_Low_PON}$	0		0.5	V		
I5	Low level input leakage current	I_{In_Low}	-5			µA		
I6	High level output voltage 1	V_{Out_High1}	VDD5V -0.4		VDD5V	V	IOH=-500 µA, static driver capability; Normal Pad Mode (see register PPCFG2 and CMC0)	
I7	Low level output voltage 1	V_{Out_Low1}	0		0.4	V	IOL=500 µA, static driver capability; Normal Pad Mode (see register PPCFG2 and CMC0)	
I8	High level output voltage 2	V_{Out_High2}	VDD5V -0.8		VDD5V	V	IOH=-4 mA, static driver capability; High Power Pad Mode (see register PPCFG2 and CMC0)	
I9	Low level output voltage 2	V_{Out_Low2}	0		0.8	V	IOL=4 mA, static driver capability; High Power Pad Mode (see register PPCFG2 and CMC0)	

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
Timing SPI-Bus Characteristics								
J1	Clock frequency	f_{clock}			2.2	MHz	Note: A high SPI clock rate during data reception can reduce sensitivity	
J2	Clock High time	t_{CLK_H}	200			ns		■
J3	Clock Low time	t_{CLK_L}	200			ns		■
J4	Active setup time	t_{setup}	200			ns		■
J5	Not active setup time	t_{not_setup}	200			ns		■
J6	Active hold time	t_{hold}	200			ns		■
J7	Not active hold time	t_{not_hold}	200			ns		■
J8	Deselect time	$t_{Deselect}$	200			ns		■
J9	SDI setup time	t_{SDI_setup}	100			ns		■
J10	SDI hold time	t_{SDI_hold}	100			ns		■
J11	Clock low to SDO valid	t_{CLK_SDO}			145	ns	@ C_{load} = 80 pF High Power Pad not enabled (Normal Mode) (see register PPCFG2 and CMC0)	■
J12	Clock low to SDO valid	t_{CLK_SDO}			40	ns	@ C_{load} = 10 pF High Power Pad not enabled (Normal Mode) (see register PPCFG2 and CMC0)	
J13	SDO rise time	t_{SDO_r}			90	ns	@ C_{load} = 80 pF	■
J14	SDO fall time	t_{SDO_f}			90	ns	@ C_{load} = 80 pF	■
J15	SDO rise time	t_{SDO_r}			15	ns	@ C_{load} = 10 pF	■
J16	SDO fall time	t_{SDO_f}			15	ns	@ C_{load} = 10 pF	■
J17	SDO disable time	$t_{SDO_disable}$			25	ns		■

1) Please note that the system bandwidth is smaller than the smallest bandwidth in the signal path.

Unless explicitly otherwise noted, the following test conditions apply to the given specification values in Table 10 and items D7 and D8:

* Hardware: TDA5240 Platform Testboard V1.0
* Single-Ended Matching for 315.0 MHz / 433.92 MHz / 868.3 MHz / 915.0 MHz
* RF input matched to 50 Ω; Insertion loss of input matching network = 1dB
* Receive Frequency 315.0 MHz / 433.92 MHz / 868.3 MHz / 915.0 MHz; Lo-Side LO-Injection
* Reference Clock: XTAL=21.948717 MHz

* IF-Gain: Attenuation set to default value (IFATT = 7)
* Double Down Conversion
* 1 IF-Filter: Center=10.7MHz; BW=330kHz; Connected between IF_OUT and IFBUF_IN
* 2nd IF Filter BW: Depending on Data Rate and FSK Deviation
* Received Signal at zero Offset to IF Center Frequency
* RSSI trimmed
* FSK Pre-Demodulation Filter (PDF) BW: Depending on Data Rate and FSK Deviation
* No SPI-traffic during telegram reception, CLK_OUT disabled
* AFC and AGC are OFF, unless otherwise noted
* Specification values are in respect to Manchester-coded Infineon-Reference Pattern 1
  (7 Bits '0', 1 Bit '1', 1 Bits '0', 1 Bit '1', 1 Bits '0', 1 Bit '1', PRBS5 (31 Bit), 1 Bit 'M') according to **Figure 18**
  However a Code Violation is not used as EOM criterion

BER sensitivity measurements use Receive Mode TMMF (sampled with ideal data clock)
MER sensitivity measurements use Receive Mode POF

* DRE ... Data Date Error of received telegram vs. adjusted Data Rate
* DC ... Duty Cycle
* MER ... Message Error Rate
  [MER = 1 - (number_of_correctly_received_messages / number_of_transmitted messages)]
* FAR ... False Alarm Rate
  [FAR = number_of_mistakenly_wake_ups / number_of_periods_searching_for_data_on_channel]
* MMR ... Missed Message Rate
  [MMR = number_of_mistakenly_missed_wake_up_patterns /
  number_of_periods_with_wake_up_pattern_transmitted_and_searching_for_wake_up_pattern]
* BER ... Bit Error Rate (using a PRBS9 Pseudo-Random Binary Sequence)
  [BER = 1 - (number_of_correctly_received_bits / number_of_transmitted bits)]

Table 10 MER Characteristics (Receive Mode = POF)

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
Characteristics of Digital Data Filter and Data Clock Recovery Acceptance Criterion is: MER <= 10%. For additional test conditions see right before this table. Double Down Conversion Mode								
K1	Data Rate Error of received Telegram Sensitivity loss < 1dB	Db	-10		10	%	at DC = 50%	■
K2	Duty Cycle Error of Manchester coding of received Telegram							
K2.1	Sensitivity loss < 1dB	tolManch_DefB	45		55	%	According to Definition B in **Chapter 2.7.2**; including DRE of -10% to +10%; Data Rate < 50 kBit/s	■
K2.2	Sensitivity loss < 3.5dB	tolManch_DefC	35		55	%	According to Definition C in **Chapter 2.7.2** including DRE of -10% to +10% Data Rate < 10 kBit/s	■
K2.3	Sensitivity loss < 1.5dB	tolManch_DefB	45		55	%	According to Definition B in **Chapter 2.7.2**; including DRE of -10% to +10%; Data Rate >= 50 kBit/s	■
K2.4	Sensitivity loss < 4dB	tolManch_DefC	35		55	%	According to Definition C in **Chapter 2.7.2** including DRE of -10% to +10% Data Rate >= 10 kBit/s Note: If BPF_BW / Bitrate < 12, the selected data rate in the configuration tool needs to be set 5% higher.	■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
Sensitivity of Receiver Acceptance Criterion is: MER <= 10%. For additional test conditions see right before this table. Double Down Conversion Mode								
L1	Sensitivity Limit in ASK (OOK) Mode; Manchester coding						At DC = 50% and DRE = 0%. T_{amb} = 25 °C, peak power level (see **Chapter 2.7.3**)	
L1.1	Data Rate 0.5 kBit/s	SASK1		-120	-117	dBm peak	m = 100% 2nd IF BW = 50 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode; 868/915MHz: <=1dB loss	■
L1.2	Data Rate 2 kBit/s	SASK2		-116	-113	dBm peak	m = 100% 2nd IF BW = 50 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
L1.3	Data Rate 10 kBit/s	SASK3		-111	-108	dBm peak	m = 100% 2nd IF BW = 50 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
L1.4	Data Rate 16 kBit/s	SASK4		-109	-106	dBm peak	m = 100% 2nd IF BW = 80 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode; 868/915MHz: <=1dB loss	■
L1.5	Data Rate 0.5 kBit/s	SASK5		-115	-112	dBm peak	m = 100% 2nd IF BW = 300 kHz; IFATT = 7, CDR = fast; Data Slicer Bit Mode; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■
L1.6	Data Rate 2 kBit/s	SASK6		-112	-109	dBm peak	m = 100% 2nd IF BW = 300 kHz; IFATT = 7, CDR = fast; Data Slicer Bit Mode; 868/915MHz: <=1dB loss; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■
L1.7	Data Rate 10 kBit/s	SASK7		-106	-103	dBm peak	m = 100% 2nd IF BW = 300 kHz; IFATT = 7, CDR = fast; Data Slicer Bit Mode; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
L1.8	Data Rate 16 kBit/s	SASK8		-104	-101	dBm peak	m = 100% 2nd IF BW = 300 kHz; IFATT = 7, CDR = fast; Data Slicer Bit Mode; Note: 3dB sensitivity loss @ f_{offset} = +/-100 kHz	■
L2	Sensitivity Limit in FSK Mode; Manchester coding						At DC = 50% and DRE = 0%. T_{amb} = 25 °C	
L2.1	Data Rate 2 kBit/s; Δf = 10 kHz	SFSK1		-118	-115	dBm	2nd IF BW = 50 kHz; PDF = 33 kHz, AFC off; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
L2.2	Data Rate 10 kBit/s; Δf = 14 kHz	SFSK2		-113	-110	dBm	2nd IF BW = 50 kHz; PDF = 65 kHz, AFC off; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
L2.3	Data Rate 10 kBit/s; Δf = 50 kHz	SFSK3		-112	-109	dBm	2nd IF BW = 125 kHz; PDF = 132 kHz, AFC off; IFATT = 0, CDR = normal; Data Slicer Bit Mode; 868/915MHz: <=1dB loss	■
L2.4	Data Rate 50 kBit/s; Δf = 50 kHz	SFSK4		-106	-103	dBm	2nd IF BW = 300 kHz; PDF = 239 kHz, AFC off; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
L2.5	Data Rate 2 kBit/s; Δf = 10 kHz	SFSK5		-108	-105	dBm	2nd IF BW = 300 kHz; PDF = 282 kHz; IFATT = 7, CDR = fast; Data Slicer Bit Mode; 868/915MHz: <=1dB loss; Note: 3dB sensitivity loss @ f_{offset}=+/-90kHz @ AFC on	■
L2.6	Data Rate 10 kBit/s; Δf = 14 kHz	SFSK6		-107	-104	dBm	2nd IF BW = 300 kHz; PDF = 282 kHz; IFATT = 7, CDR = fast; Data Slicer Bit Mode; Note: 3dB sensitivity loss @ f_{offset}=+/-90kHz @ AFC on	■
L2.7	Data Rate 10 kBit/s; Δf = 50 kHz	SFSK7		-109	-106	dBm	2nd IF BW = 300 kHz; PDF = 282 kHz; IFATT = 7, CDR = fast; Data Slicer Bit Mode; Note: 3dB sensitivity loss @ f_{offset}=+/-90kHz @ AFC on	■

#	Parameter	Symbol	Limit Values			Unit	Test Conditions Remarks	
			min.	typ.	max.			
Dynamic Range of Receiver Acceptance Criteria are: MER <= 1E-3, FAR < 1E-5, MMR < 1E-4 (Criterion: 8 Equal Bits), Manchester coding. For additional test conditions see right before this table. Double Down Conversion Mode								
M1	Dynamic Range in ASK (OOK) Mode, AGC on						At DC = 50% and DRE = 0%. T_{amb} = 25 °C, peak power level (see Chapter 2.7.3)	
M1.1	Data Rate 2 kBit/s	DR2,OOK	-10		-109	dBm peak	m = 100% 2nd IF BW = 50 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
M1.2	Data Rate 10 kBit/s	DR10,OOK	-10		-105	dBm peak	m = 100% 2nd IF BW = 50 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
M1.3	Data Rate 2 kBit/s	DR2,ASK50	-45		-103	dBm peak	m = 50% 2nd IF BW = 50 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
M1.4	Data Rate 10 kBit/s	DR10,ASK50	-60		-99	dBm peak	m = 50% 2nd IF BW = 50 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
M2	Dynamic Range in FSK Mode, AGC on Data Rate 10 kBit/s & Δf = 50 kHz						At DC = 50% and DRE = 0%. T_{amb} = 25 °C	
M2.1	0% AM Modulation	DR10,AM0	-10		-106	dBm	2nd IF BW = 125 kHz PDF = 132 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■
M2.2	90% AM Modulation, 100 Hz	DR10,AM90	-10		-90	dBm	2nd IF BW = 125 kHz PDF = 132 kHz; IFATT = 0, CDR = normal; Data Slicer Bit Mode	■