nRF Performance Test Instructions

nRF24L01+

Application Note

November 2008

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1 Introduction

RF front end performance plays a vital role in the performance of any RF product, it is also the main area of testing when qualifying your nRF application for regulatory standards for RF emissions (like FCC, ETSI and TELEC). The test instructions presented in this document are the device setup, recommended instrument setup and necessary MCU routines required to perform the corresponding tests described in the nRF Performance Test Guidelines white paper available on our website.

The following nRF devices are covered by the instructions in this document:

- nRF24L01+: standalone 2.4GHz ISM band RF transceiver.
- nRF24LU1+: USB flash microcontroller with nRF24L01+ embedded.
- nRF24LE1

Note: The tests described in this document are intended for performance testing in the development stages of a product. Some of these tests can also be used during production testing while others are too complex and take too much time to be feasible for use in a production line.

2 Output power

The following setup enables you to run the output power tests described in chapter two of the RF Prototype Test Guidelines white paper.

2.1 TX carrier wave output

2.1.1 Device Configuration and data input

The configuration registers listed in Table 1. must be updated from their reset values to the test values to perform the output power test on the Device Under Test (DUT).

Register address	Mnemonic	Reset value (Hex)	Test value (Hex)	Comments
00	CONFIG	0x08	0x02	Bit 1: POWER_UP = 1
05	RF_CH	0x02	XX	RF_CH must be set depending on which frequency (F0) you want the carrier on. Formula: F0 = 2400 + RF_CH (MHz)
06	RF_SETUP	0x0F	0x9F	Bit 7: Cont wave = 1 Bit 4: PLL_lock = 1 If other than max. output power is wanted, bit2:1 must be set according to wanted output power, please refer to device product specification.

Table 1. Configuration registers required to be updated for TX carrier wave output

Note: Please refer to the Product Specifications for details on accessing the configuration registers.

2.1.2 Test routine

To run the test, the system MCU (external or embedded) must implement the following routine:

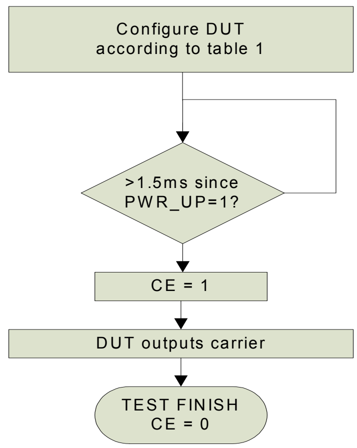

Figure 1. TX carrier wave output test routine

2.1.3 Test instrument setup

Optimal setup of the spectrum analyzer may vary from instrument to instrument, but the following recommendations are a good indication of a typical setup:

Parameter	Value	Comment
Centre frequency	F_0	Must be set to the same value as configured in the device RF_CH register following formula: F0 = 2400 + RF_CH (MHz).
Span	10 MHz	
Amplitude	5 dBm	
Resolution Band Width (RBW)	Auto	Can usually be set to auto to follow the span used. With a span of 10MHz this should give an RBW of 30-100 kHz.

Table 2. TX carrier wave test spectrum analyzer setup

2.2 TX carrier wave sweep

This section describes how you extend the TX carrier wave test described in section 2.1 on page 5 to a TX carrier wave sweep covering the 2.4GHz band.

2.2.1 Device configuration

Device configuration in this test is identical to the TX carrier wave (section 2.1 on page 5) except for register 05, which should be set to its minimum as stated in Table 3.

Register address	Mnemonic	Reset value (Hex)	Test value (Hex)	Comments
05	RF_CH	0x02	0x00	First channel is the lowest possible F0 (2400 MHz)

Table 3. Configuration registers required to be updated for TX carrier wave sweep

2.2.2 Test routine

This test builds on the TX carrier wave test routine:

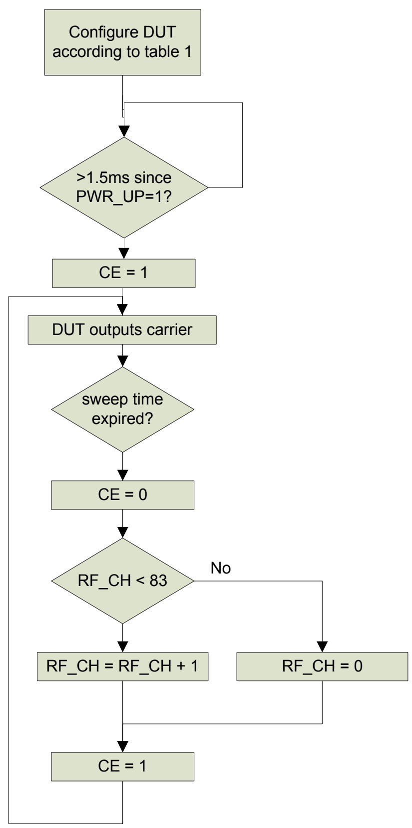

Figure 2. TX carrier wave sweep test routine

2.2.3 Test instrument setup

Optimal setup of the spectrum analyzer may vary from instrument to instrument, but the following recommendations are a good indication of a typical setup:

Parameter	Value	Comment
Centre frequency	2.42GHz	
Span	100 MHz	
Amplitude	5 dBm	
Resolution Band Width (RBW)	Auto	Can usually be set to auto to follow the span used. With a Span of 100MHz this should give a RBW of ~300kHz -1 MHz
Capture mode	Peak hold	Setting usually found in spectrum analyzer TRACE or AVERAGE menus

Table 4. TX carrier wave sweep test spectrum analyzer setup

3 Frequency accuracy

The following setup enables you to run the frequency accuracy test described in chapter four of the RF Prototype Test Guidelines white paper. Use the configurations and the MCU routine described in section 2.1 on page 5 of this document.

3.1 Test instrument setup

Optimal setup of the spectrum analyzer may vary from instrument to instrument, but the following recommendations are a good indication of a typical setup:

Parameter	Value	Comment
Centre frequency	F_0	Must be set to the same value as configured in the device RF_CH register following formula: F0 = 2400 + RF_CH (MHz)
Span	5 MHz	
Amplitude	5 dBm	
Resolution Band Width (RBW)	10kHz	Must be set according to the accuracy you want to measure

Table 5. Frequency accuracy test spectrum analyzer setup

4 Spurious emissions

The following setup enables you to run the spurious emission tests described in chapter three of the nRF Performance Test Guidelines white paper.

4.1 Harmonic output power

This test can use the device configuration and MCU routine from section 2.1 or section 2.2. By using the setup from section 2.2 you get a test showing the power on all possible harmonic output frequencies.

4.1.1 Test instrument setup

Optimal setup of the spectrum analyzer may vary from instrument to instrument, but the following recommendations are a good indication of a typical setup:

Parameter	Value	Comment
Centre frequency	N*2.42GHz	N=2, 3, 4 etc. depending on which harmonic you want to measure (2^{nd} = 4.84GHz, 3^{rd} = 7.26GHz etc.)
Span	2^{N-1} * 100 MHz	N=2, 3, 4 depending on which harmonic to be measured (2. = 200MHz, 3^{rd} = 400 MHz)
Amplitude	5 dBm	
Resolution Band Width (RBW)	Auto	Can usually be set to auto to follow the span used. This should give an RBW of ~1-3MHz
Capture mode	Peak hold	Setting usually found in spectrum analyzer TRACE or AVERAGE menus

Table 6. Harmonic output power test spectrum analyzer setup

4.2 RX local oscillator leakage

Keep the MCU routine from section 2.2 but update the following configuration registers compared to what is listed in (Table 1.):

Register address	Mnemonic	Reset value (Hex)	Test value (Hex)	Comments
00	CONFIG	0x08	0x03	Bit 0: PRIM_RX = 1, enable receive Bit 1: POWER_UP = 1
05	RF_CH	0x02	0x00	First channel is the lowest possible F0 (2400 MHz)

Table 7. Updated configuration registers required for RX local oscillator leakage test

4.2.1 Test instrument setup

Optimal setup of the spectrum analyzer may vary from instrument to instrument, but the following recommendations are a good indication of a typical setup:

Parameter	Value	Comment
Centre frequency	2.77GHz	nRF24L01+ LO frequency for each RX channel is given by: F_{LO} = 8/7 (F_{RXCH} + 2 MHz) where F_{RXCH} is the frequency of your receive channel
Span	100 MHz	
Amplitude	-40 dBm	
Resolution Band Width (RBW)	Auto	This might need reduced in order to distinguish the LO leakage from the noise floor of the spectrum analyzer
Capture mode	Peak hold	Setting usually found in spectrum analyzer TRACE or AVERAGE menus

Table 8. RX local oscillator leakage test spectrum analyzer setup

5 Modulation Bandwidth

The following setup enables you to run the modulation bandwidth tests described in chapter five of the RF Performance Test Guidelines white paper.

5.1 Frequency deviation

To measure frequency deviation on the device you must send a sequence of identical packets creating a low frequency shift between the low and high FSK frequency.

5.1.1 Device Configuration and data input

The configuration registers listed in Table 9. must be updated from their reset values to the test values to perform the output power test.

Register address (Hex)	Mnemonic	Reset value (Hex)	Test value (Hex)	Comments
00	CONFIG	0x08	0x02	Bit 1: POWER_UP = 1 Bit 3: CRC_EN = 0
01	EN_AA	0x3F	0x00	Disable all auto acknowledge
04	SETUP_RETR	0x03	0x00	Disable all auto retransmit
05	RF_CH	0x02	xx	RF_CH must be set depending on which frequency (F0) the output is wanted on. Formula: F0 = 2400 + RF_CH (MHz)
06	RF_SETUP	0x0F	0x0F	Bit 5: Low air data rate 1 = 250kbps, 0 = bit 3 controls Bit 3: High air data rate 0 = 1Mbps, 1 = 2Mbps Given test value will set nRF24L01+ to 0dBm output @ 2 Mbps. If other than max. output power is wanted, bit2:1 must be set according to wanted output power, please refer to device product specification
10	TX_ADDR	0xE7E7E7E7E7	0xFF FF FF FF FF	

Table 9. Configuration registers required to be updated for frequency deviation

Data input (TX_FIFO content) in this test must be set to:

Number of bytes	Content
13 (0x00-0x0C)	0xFF
18 (0x0D-0x1F)	0x00

Table 10. Settings for data input in the frequency deviation test

5.1.2 Test routine

To run the test, the system MCU (external or embedded) must implement the following routine:

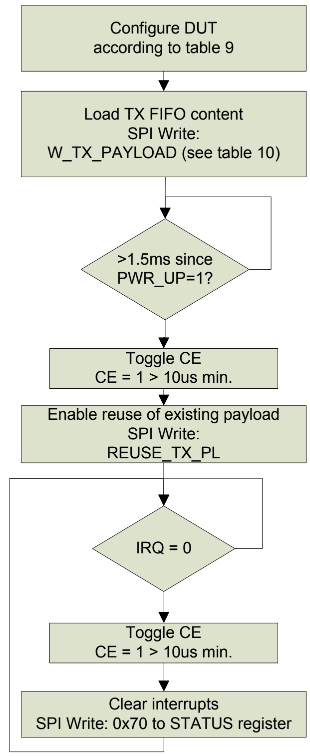

Figure 3. Frequency deviation test routine

By using the nRF interrupt features, the RF front end tells the MCU when it has finished sending the packet (default enabled after power on). The routine makes the radio send the same packet repeatedly as fast as possible. The benefit is that there is no need to change the MCU routine if the data rate is changed. If the RF front end interrupt features are not used the CE can toggle at a maximum rate of ~0.2 ms @ 2 Mbps, ~0.5 ms @ 1 Mbps and 1.5 ms @ 250 kbps.

5.1.3 Test instrument setup

Optimal setup of the spectrum analyzer may vary from instrument to instrument, but the following recommendations are a good indication of a typical setup:

Parameter	Value	Comment
Centre frequency	F_0	Must be set to same value as configured in the device RF_CH register following formula: F0 = 2400 + RF_CH (MHz)
Span	10 MHz	
Amplitude	5 dBm	
Resolution Band Width (RBW)	Auto	Can usually be set to auto to follow the span used. With a Span of 10MHz this should give an RBW of 30-100 kHz
Capture mode	Peak hold	Setting usually found in spectrum analyzer TRACE or AVERAGE menus

Table 11. TX deviation test spectrum analyzer setup

5.2 TX modulation bandwidth

Use the MCU test routine and instrument setup described in section 5.1. The device's TX address and payload content must be changed according to the following tables.

Register address (Hex)	Mnemonic	Reset value (Hex)	Test value (Hex)	Comments
10	TX_ADDR	0xE7E7E7E7E7	0x80 42 30 9C AB	5 byte PRBS data

Table 12. Configuration register required to be updated for TX modulation bandwidth

Data input (TX_FIFO content) in this test must be set to:

Number of bytes	Content (Hex)	Comment
32	(hex) 0xA6 E4 50 AD 3F 64 96 FC 9A 99 80 C6 51 A5 FD 16 3A CB 3C 7D D0 6B 6E C1 6B EA A0 52 BC BB 81 CE 93 D7 51 21 9C 2F 6C D0 EF 0F F8 3D F1 73 20 94 ED 1E 7C D8 A 9 1C 6D 5C 4C 44 02	32 bytes PRBS data

Table 13. Settings for data input in the TX modulation bandwidth test

6 Receiver sensitivity

The following set up enables you to run the receiver sensitivity test described in chapter six of the RF Performance Test Guidelines white paper. To run this test, the system MCU (external or embedded) must implement the general receiver sensitivity routine found in chapter six of the RF Performance Test Guidelines white paper.

6.1 Device Configuration

The configuration registers listed in Table 14. must be updated from their reset values to the test values to perform the output power test.

Register address (Hex)	Mnemonic	Reset value (Hex)	Test value (Hex)	Comments
00	CONFIG	0x08	0x03	Bit 3: CRC_EN = 0 Bit 1: POWER_UP = 1 Bit 0: PRIM_RX = 1
01	EN_AA	0x3F	0x00	Disable all auto acknowledge
04	SETUP_RETR	0x03	0x00	Disable all auto retransmit
05	RF_CH	0x02	XX	RF_CH must be set depending on which frequency (F0) the device is to receive on. Formula: F0 = 2400 + RF_CH (MHz)
06	RF_SETUP	0x0F	0x07	Bit 5: Low air data rate 1 = 250kbps, 0 = bit 3 controls Bit 3: High air data rate 0 = 1Mbps, 1 = 2Mbps Given test value will set nRF24L01+ to 2 Mbps
0A	RX_ADDR_P0	0xE7E7E7E7E7	0x80 42 30 9C AB	5 bytes random address pattern
11	RX_PW_P0	0	0x20	32 bytes payload length

Table 14. Configuration registers required to be updated for Receiver sensitivity

6.2 Test Instrument setup

This is the RF generator setup you need to generate a valid input signal:

Parameter	Value	Comment
RF Frequency	F_0	Must be set to the same value as configu-red in the device RF_CH register following formula: F0 = 2400 + RF_CH (MHz)
Modulation	2 GFSK	
Data rate	1 or 2 Mbps	Must match DUT RF_SETUP register
Deviation	±160 or ±320 kHz	Must match DUT: ±160 kHz @ 250kbps and 1 Mbps, ±320 kHz @ 2 Mbps
BT	0.5 Gaussian	
Recommended packet build	**Preamble:** (hex)AA **Address:** (hex) 0x08 C2 72 AC 37 **Payload:** (hex) 0xA6 E4 50 AD 3F 64 96 FC 9A 99 80 C6 51 A5 FD 16 3A CB 3C 7D D0 6B 6E C1 6B EA A0 52 BC BB 81 CE 93 D7 51 21 9C 2F 6C D0 EF 0F F8 3D F1 73 20 94 ED 1E 7C D8 A 9 1C 6D 5C 4C 44 02	38 byte sequence sent over and over again from the RF generator. MCU routine in DUT must compare received data to Payload defined here and count number of errors in each received payload

Table 15. Receiver sensitivity test instrument setup

7 Receiver selectivity

The following setup enables you to run the receiver selectivity test described in chapter seven of the RF Performance Test Guidelines white paper. To run this test, the system MCU (external or embedded) must implement the general receiver selectivity routine found in chapter seven of the RF Performance Test Guidelines white paper. Use all RF generator 1 device configuration and instrument setup from the receiver sensitivity test in chapter six of the RF Performance Test Guidelines white paper.

> **Note:** The receiver selectivity in nRF devices is decided by the nRF device design. Therefore, this test only verifies the numbers already listed in nRF product specifications.

7.1 Test Instrument setup

The second RF generator in this test can send a carrier, using the following configuration:

Parameter	Value	Comment
RF Frequency	xx	Varied in steps to find co-, adjacent channel and wide band blocking, please refer to the RF Performance Test Guidelines white paper
Output power	xx	Varied in steps to find co-, adjacent channel and wide band blocking, please refer to the RF Performance Test Guidelines white paper
Modulation	OFF	

Table 16. Receiver selectivity test setup

Alternatively, the second RF generator can use the same configuration as the first generator to mimic an interfering second nRF24L01+ device or it can be set up to mimic other common radio systems like Bluetooth.

> **Note:** In the case of mimicking a second nRF24L01+ device, the ‘address’ pattern sent from generator 1 and 2 MUST be different.

Software Examples Using ShockBurst™ Modes in nRF24L01 and nRF24LU1

nAN24-12

Application Note v1.0

February 2008

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1 Introduction

This document uses software examples to describe how to use ShockBurst™ (SB), Enhanced ShockBurst™ (ESB) and Enhanced ShockBurst™ with Bidirectional data (PL), to transfer button status from one device to another. Also described in this document is how to port code to other hardware platforms.

The nAN24-12SW package contains source code that helps you set up the nRF24L01 and nRF24LU1 in the ShockBurst™ modes. The source code is written in C and contains all the files necessary for setting up the devices using only the Hardware Abstraction Layer (HAL). The source code can also be used as a reference when you are developing your own application.

You will need two Basic Feature Boards (BFB), which are included in the nRF24L01 and nRF24LU1 development kits, programmed with the nAN24-12 firmware to run the application.

nRF24L01 and nRF24LU1 use the following hex files for demonstration when plugged into the BFB:

- Basic Feature Board with C8051F320 + nRF24L01: nRF24L01.hex
- Basic Feature Board with nRF24LU1: nRF24LU1.hex

The BFB must be powered on, but does not need to be connected to any peripherals to run the application example. The buttons and LEDs on the BFB are used for user input and output.

1.1 Prerequisites

In order to fully understand the application note, a good knowledge of electronic and software engineering is required.

1.2 Writing conventions

This application note follows a set of typographic rules that makes the document consistent and easy to read. The following writing conventions are used:

- Commands and code are written in `Courier`.
- File names are written in **bold**.
- Cross references are underlined and highlighted in blue.

2 nAN24-12 contents and requirements

This chapter describes the contents of the nAN24-12SW package and the hardware that is required for demonstrating the different operational modes. Refer to chapter 6 for information on porting the source code to other hardware architectures.

2.1 Contents

The software package contains the following files:

- Documentation (/doc)
  - nAN24-12 Application Note
  - Windows help file (**help.chm**)
- Source code (/src)
  - The Keil project files for both nRF24L01 and nRF24LU1
  - Complete source code for the application
  - Complete source code for the necessary HAL libraries
- Firmware files (/hex)
  - The pre-compiled hex files for nRF24L01 and nRF24LU1

2.2 Required software

The application has been tested and compiled on KEIL µVision3 3.53.

2.2.1 nRF24L01 specific software

To program the SiLabs C8051F320 MCU, which is part of the nRF24L01 development kit, you can use either the built in functionality in the Keil software package or you can download the Flash Programming Utilities from SiLabs: http://www.silabs.com

2.2.2 nRF24LU1 specific software

Refer to the documentation for the nRF24LU1 Development Kit on how to program the nRF24LU1.

2.3 Required hardware

To work out of the box, this application needs either the development kit for nRF24L01 or nRF24LU1.

2.3.1 nRF24L01 specific hardware

To program the nRF24L01 development kit you need a programmer that supports the SiLabs C8051F320 MCU. The low cost USB debug adapter from SiLabs has been used during the development: http://www.silabs.com

You need to solder the four pin programming/debug interface on the dongle to a simple interface to support the 10 pin connector on the USB debug adapter. See Figure 1. for the schematic.

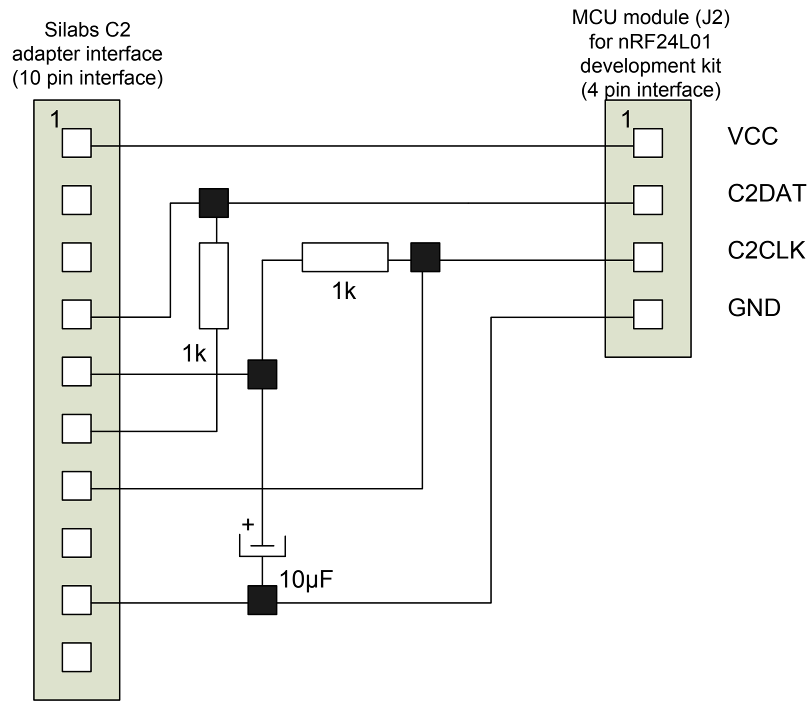


Figure 1. SiLabs modification

2.3.2 nRF24LU1 specific hardware

All the necessary hardware for the nRF24LU1 is included in the nRF24LU1 Development Kit.

3 SW architecture

The software (SW) architecture is shown in Figure 2. A source and header file represent each block. See the help documentation for more information on the prototypes and macros.

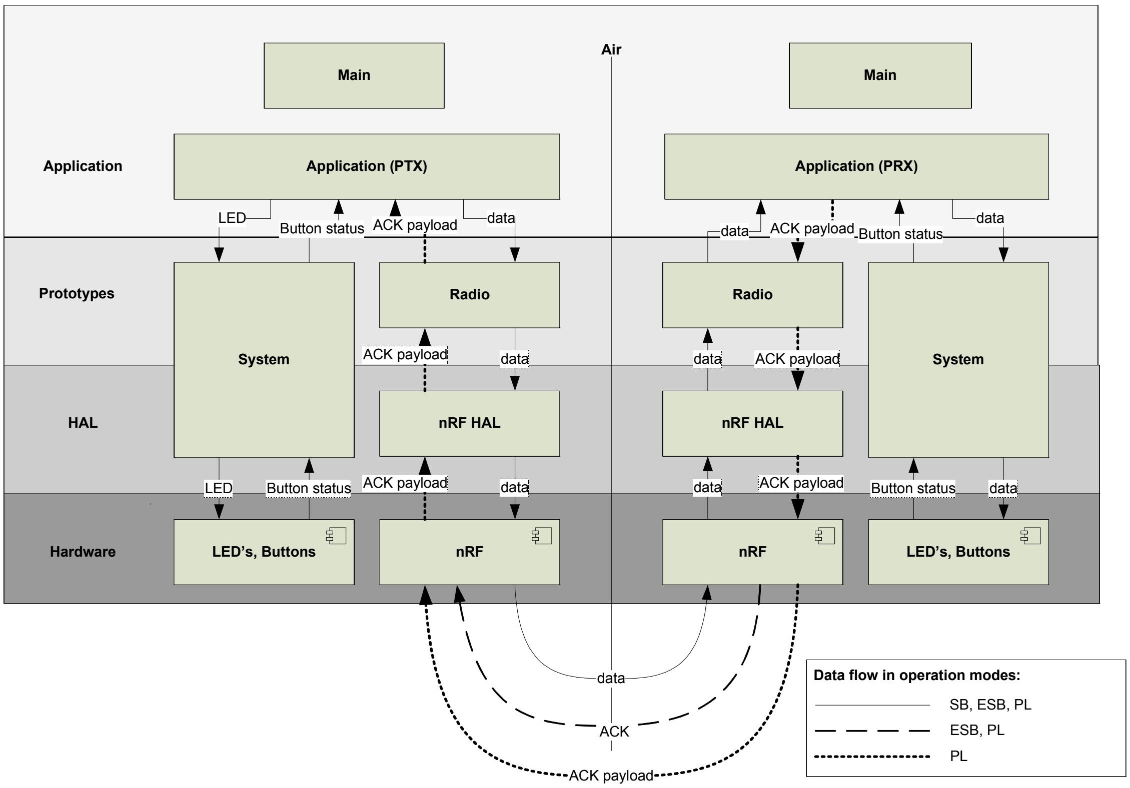


Figure 2. SW Architecture

3.1 Main

The main application is a state machine controlling your choices of operational modes and initialization. It starts up, initializes the microcontroller, and waits for you to select the application mode before initializing the radio and starting the application. After starting the application, the main application releases all control.

The main application defines the following prototypes and typedefs:

Prototypes:

```
state_t get_next_state (state_t current_state);
void wait_for_button_release (void);
void show_status (state_t operation);
```

Typedef:

```
enum {
  DEVICE_IDLE = 0,
  DEVICE_PRX_IDLE,
  DEVICE_PTX_IDLE,
  DEVICE_PRX_SB,
  DEVICE_PRX_ESB,
  DEVICE_PRX_PL,
  DEVICE_PTX_SB,
  DEVICE_PTX_ESB,
  DEVICE_PTX_PL,
  NO_CHANGE
} state_t;
```

3.2 Application

The application handles the logic of the program. It contains the logic for both the Primary Transceiver (PTX) and the Primary Receiver (PRX) in separate functions. There is one implementation for each of the operational modes.

The application defines the following external prototypes:

Prototypes (ShockBurst™):

```
void device_ptx_mode_sb (void);
void device_prx_mode_sb (void);
```

Prototypes (Enhanced ShockBurst™):

```
void device_ptx_mode_esb (void);
void device_prx_mode_esb (void);
```

Prototypes (Enhanced ShockBurst™ with bidirectional data):

```
void device_ptx_mode_pl (void);
void device_prx_mode_pl (void);
```

3.3 System

The system controls the interface to the MCU and the development kit. All hardware interaction is performed through these functions.

The following external prototypes and macros are defined in the system:

Prototypes:

```
void device_boot_msg (void);
void delay_10ms (void);
void delay_100ms (void);
void start_timer (uint16_t time);
void wait_for_timer (void);
bool timer_done (void);
```

Macros:

```
LEDx_ON (); (One each for LED1, LED2 and LED3)
LEDx_OFF (); (One each for LED1, LED2 and LED3)
LEDx_BLINK (); (One each for LED1, LED2 and LED3)
LED_ALL_OFF ();
T0_START ();
T0_STOP ();
T1_START ();
T1_STOP ();
```

3.4 Radio

The radio contains all the necessary functions to send and receive packages. Each operation mode (ShockBurst™, Enhanced ShockBurst™, and Enhanced ShockBurst™ with bidirectional data) has a separate initiation function.

The radio defines the following external prototypes, constants and typedefs:

Prototypes:

```
void radio_send_packet (uint8_t *packet, uint8_t length);
radio_status_t radio_get_status (void);
void radio_set_status (void);
uint8_t radio_get_pload_byte (uint8_t byte_index);
void radio_irq (void);
```

ShockBurst™:

```
void radio_init_sb (uint8_t *address, hal_nrf_operation_mode_t
operational_mode);
```

Enhanced ShockBurst™:

```
void radio_init_esb (uint8_t *address, hal_nrf_operation_mode_t
operational_mode);
```

Enhanced ShockBurst™ with bidirectional data:

```
void radio_init_pl (uint8_t *address, hal_nrf_operation_mode_t
operational_mode);
```

Constants:

```
RF_CHANNEL 40;
RF_POWER_UP_DELAY 2;
PAYLOAD_LENGTH 2;
RF_RETRANSMITS 15;
RF_TRANS_DELAY 250;
```

Typedef:

```
enum {
  RF_IDLE,
  RF_MAX_RT,
  RF_TX_DS,
  RF_RX_DR,
  RF_TX_AP,
  RF_BUSY
} radio_status_t;
```

3.5 nRF HAL

The nRF HAL (Hardware Abstraction Layer) contains all the functions that abstracts communication with the selected radio. These are specific for the chosen radio, but implement the same interface, **hal_nrf.h**.

4 Structure

The source code is divided into logical parts for easy reading. To enable easy porting to different hardware platforms, the hardware specific files are placed in separate subfolders for each architecture. The default supported platforms are:

- SiLabs C8051F320+nRF24L01
- nRF24LU1

There are also separate folders for each operational mode (ShockBurst™, Enhanced ShockBurst™, and Enhanced ShockBurst™ with bidirectional data)

For information on porting, see chapter 6.

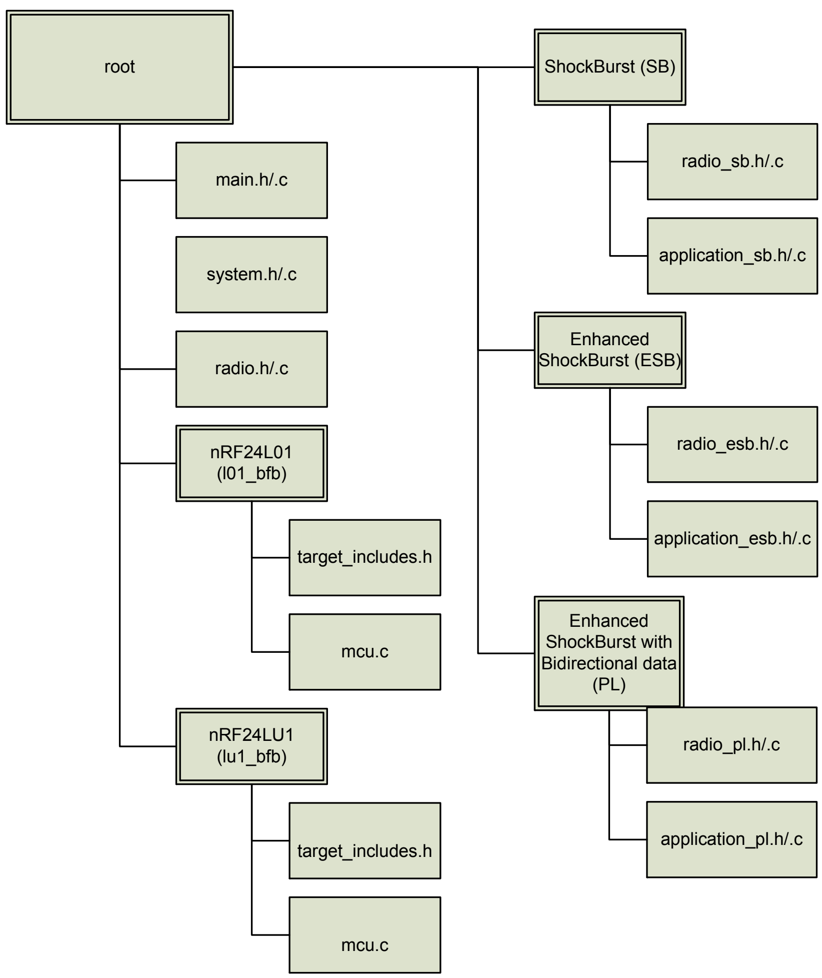


Figure 3. File and folder structure

5 Application

This chapter describes how to operate the application. How to choose an operational mode is described in section 5.1.The different operational modes are described in sections 5.2 to section 5.4.

5.1 Easy setup for demonstration

Figure 4. shows how to set up a board in each mode. The first step is to choose between PTX and PRX mode and the second step is to set up the desired operation mode. For example, the board can be set up to PTX ShockBurst™ mode by pressing Button 1 twice. After you set up the operation mode, the LED stays lit for three seconds to indicate that the set up is successful.

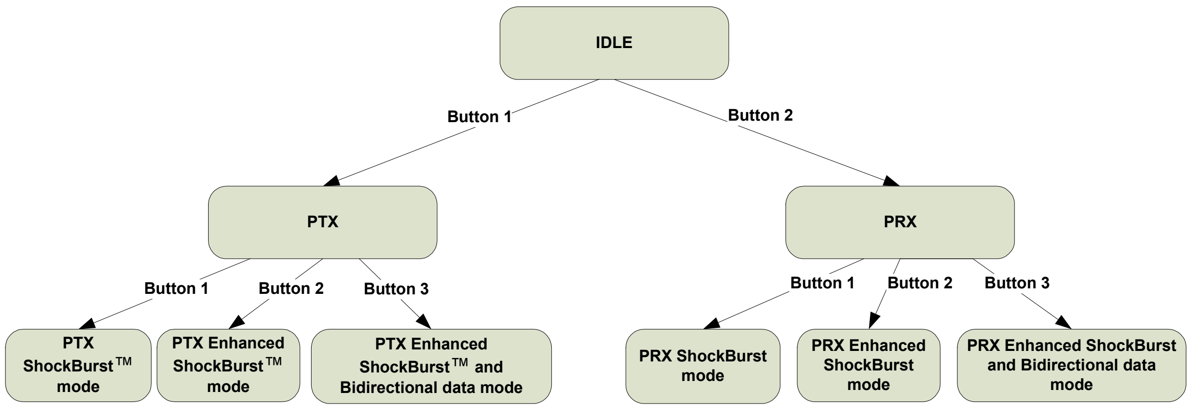
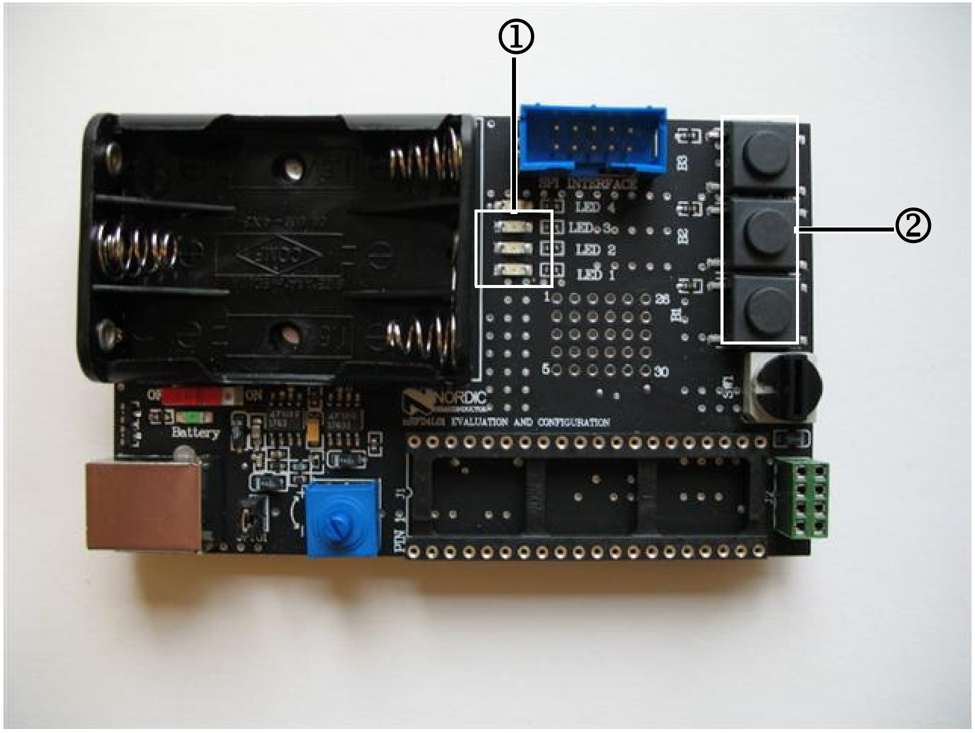


Figure 4. Setting up the board for demonstration

Mode	LED status		
	LED 1	**LED 2**	**LED 3**
Idle mode	ON	ON	ON
PTX	OFF	OFF	OFF
PRX	OFF	OFF	ON
PTX ShockBurst™	ON	OFF	OFF
PRX ShockBurst™	ON	OFF	ON
PTX Enhanced ShockBurst™	OFF	ON	OFF
PRX Enhanced ShockBurst™	OFF	ON	ON
PTX Enhanced ShockBurst™ and Bidirectional data	ON	ON	OFF
PRX Enhanced ShockBurst™ and Bidirectional data	ON	ON	ON

Table 1. LED status of each mode



① LEDs (1,2,3) ② Buttons (1,2,3)

Figure 5. nRF24LU1 Basic Feature Board (BFB)

5.2 ShockBurst™

You can set up the PTX and PRX to ShockBurst™ with no advanced features enabled. The PTX is configured without auto retransmit, and the PRX is configured without auto ack. ShockBurst™ is used to broadcast data.

ShockBurst™ is demonstrated by pressing Button 1 on the PTX which then turns on the LED 1 on the PRX.

See Figure 6. (a) for the flow of PTX in ShockBurst™. Press Button 1 twice to get to this mode.

See Figure 6. (b) for the flow of PRX in ShockBurst™. Press Button 2 followed by Button 1 to get to this mode.

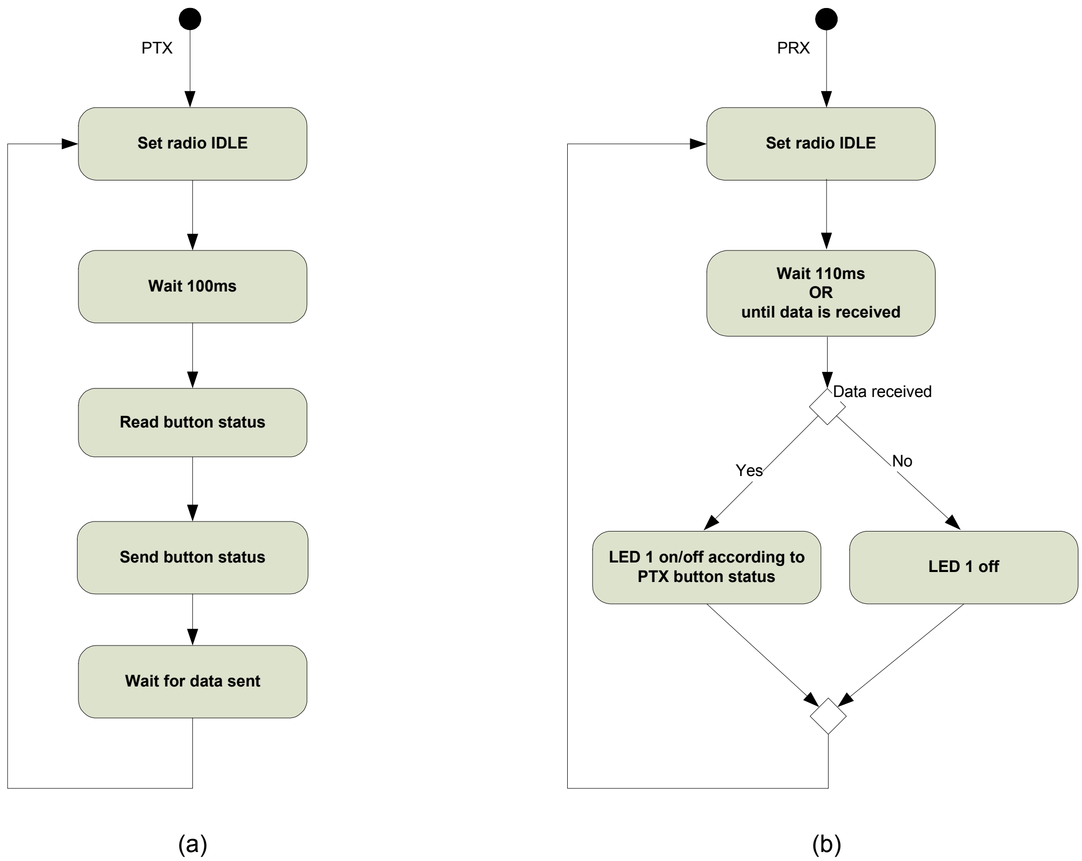


Figure 6. Application flow in ShockBurst™

5.3 Enhanced ShockBurst™

You can set up the PTX and PRX to Enhanced ShockBurst™ with advanced features enabled. The PTX is configured with auto retransmit, and the PRX is configured with auto acknowledgement.

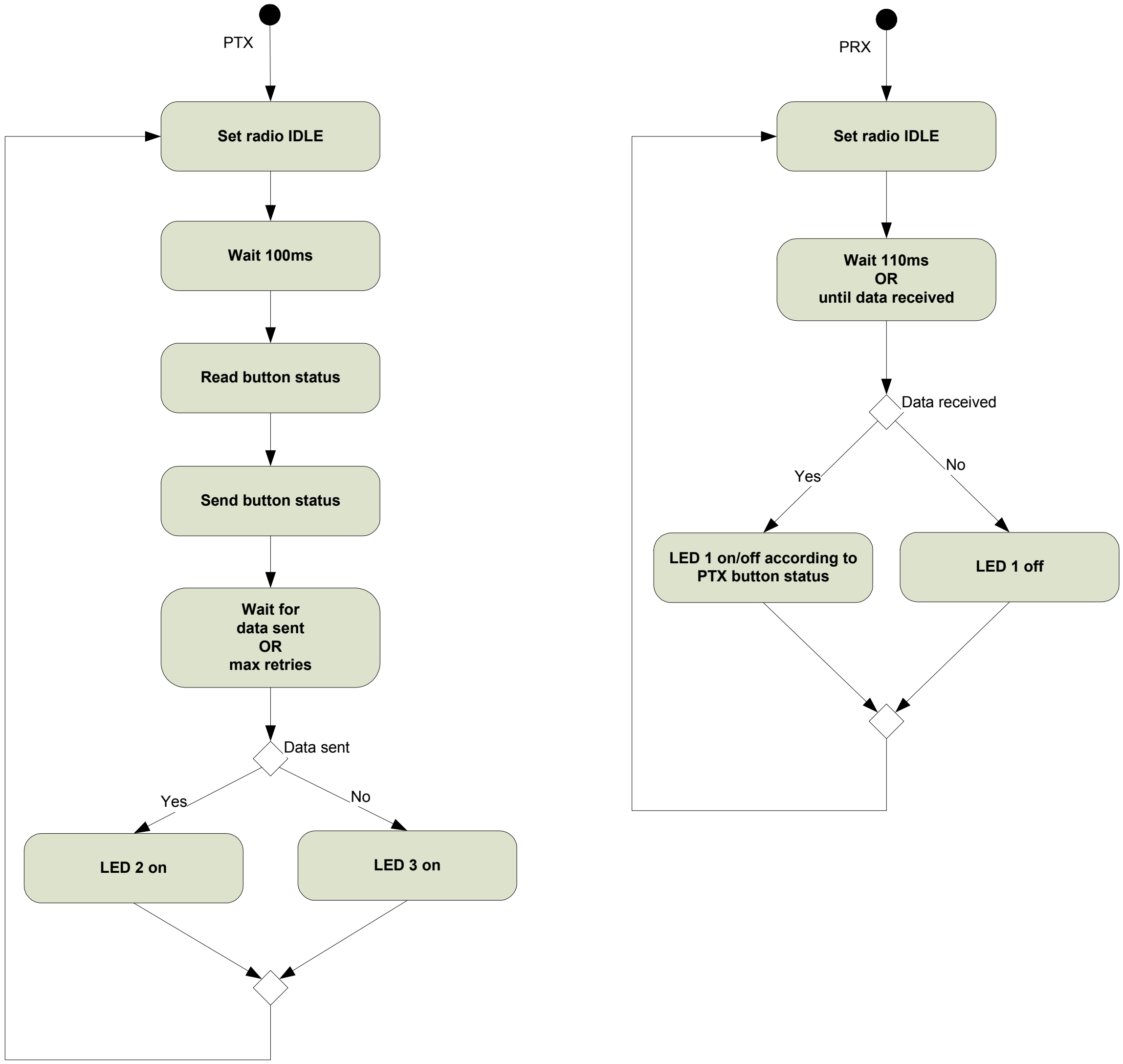
The advantages of using Enhanced ShockBurst™ are:

- One way communication.
- Auto acknowledge and auto retransmit data.

Enhanced ShockBurst™ is demonstrated by pressing Button 1 on the PTX, which then turns on LED 1 on the PRX. The PTX indicates successfully sent data by turning LED 2 on, or if max retransmission occurs this is indicated by the PTX turning LED 3 on.

See Figure 7. (a) for the flow of PTX in Enhanced ShockBurst™. Press Button 1 followed by Button 2 to get to this mode.

See Figure 7. (b) for the flow of PRX in Enhanced ShockBurst™. Press Button 2 twice to get to this mode.



(a) (b)

Figure 7. Application flow in Enhanced ShockBurst™

5.4 Enhanced ShockBurst™ with bidirectional data

You can set up the PTX and PRX to Enhanced ShockBurst™ with advanced features enabled. The PTX is configured with auto retransmit and ack payload, and the PRX is configured with auto ack and ack payload.

The advantages of using the Enhanced ShockBurst™ with bidirectional data are:

- Bidirectional communication
- Auto acknowledge and auto retransmit data
- Dynamic payload length
- Payload with acknowledgement

Enhanced ShockBurst™ with Bidirectional data is demonstrated by pressing either Button 1 on the PTX, which then turns on LED 1 on the PRX or, by pressing Button 1 on the PRX, which then turns on LED1 on the PTX. Both the PTX and the PRX indicate successfully sent data by turning LED 2 on, or if max retransmission occurs this is indicated by LED 3 on.

See Figure 8. (a) for the flow of PTX in Enhanced ShockBurst™ with Bidirectional data. Press Button 1 followed by Button 3 to get to this mode.

See Figure 8. (b) for the flow of PRX in Enhanced ShockBurst™ with Bidirectional data. Press Button 2 followed by Button 3 to get to this mode.

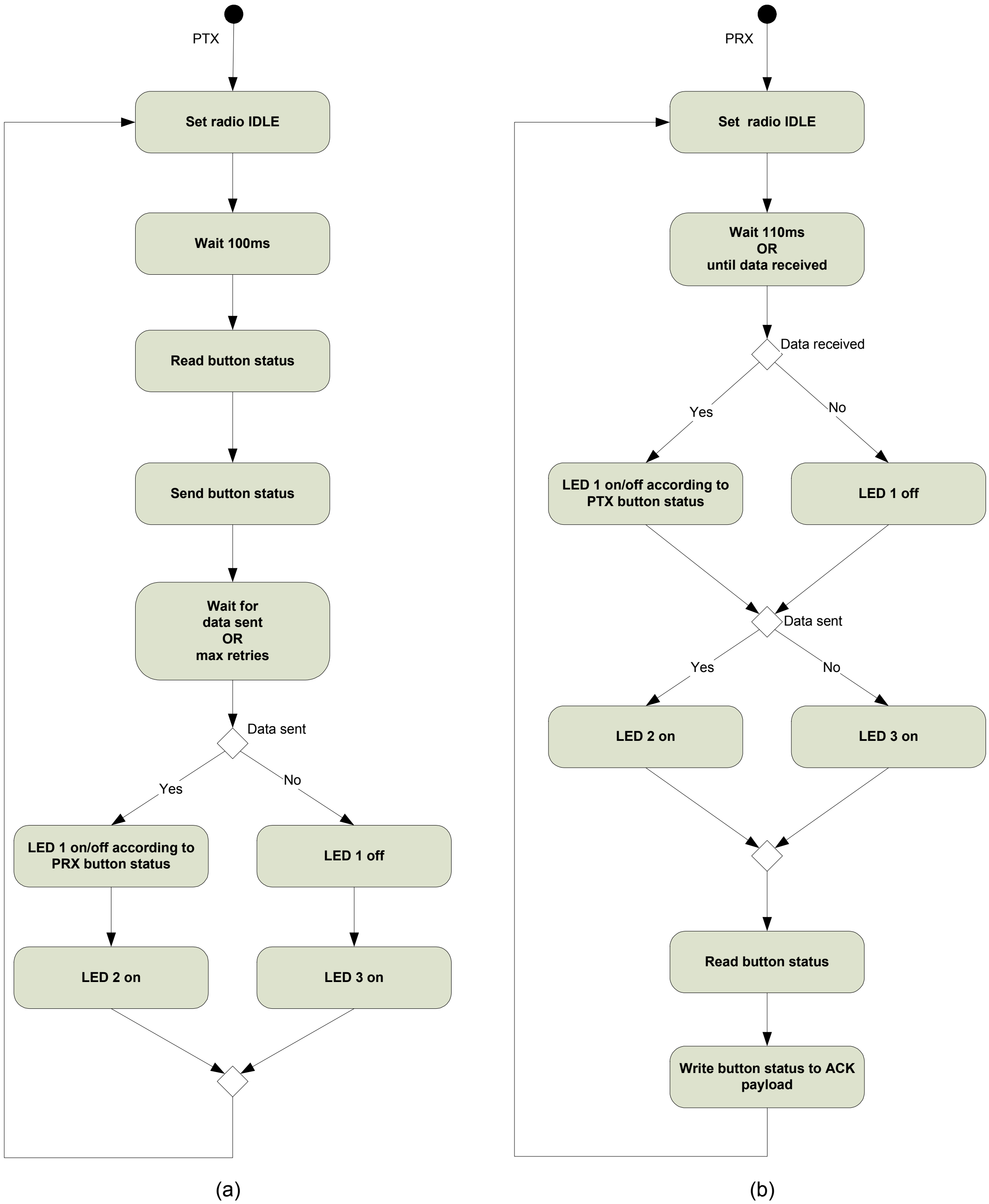


Figure 8. Application flow in Enhanced ShockBurst™ with bidirectional data

6 Porting the source code to another hardware architecture

You must re-map the hardware to port this application to another hardware architecture. Most of the hardware dependent parts of the code are in the MCU specific subfolders. The hardware dependent files for the nRF24LU1 are located in the **lu1_bfb** folder. The hardware dependent files for the nRF24L01 are located in the **l01_bfb** folder. When porting to another MCU, a new folder must be made containing the following files:

- mcu.c
- target_includes.h

If you want a new Keil project, you should base it on the **nRF24LU1.uV2** with the following changes:

1. Update the device architecture to your target device.
2. Name of the compiled hex file should reflect the architecture.
3. Paths should include:
   a. The new folder for the MCU.
   b. The correct HAL library for the radio architecture.
4. Update the HAL group to contain the HAL files for the architecture.
   a. `hal_nrf_hw.c` (implementation of `hal_nrf_rw` from **hal_nrf.h**).
   b. `hal_nrf_x.c` (*x* indicating MCU. Implementation of **hal_nrf.h**).

6.1 mcu.c

This file needs one function: `void system_init (void);` This function should set up three inputs and three outputs (B1-3 and LED1-3 respectively).

6.2 target_includes.h

This file contains most of the hardware dependent settings. The hardware abstractions that must be ported are in Table 2.

Name	Description
LED_ON	The value that should be sent on the output to turn on a LED
LED_OFF	The value that should be sent on the output to turn off a LED
LED1	Address of LED1 (bit address)
LED2	Address of LED2 (bit address)
LED3	Address of LED3 (bit address)
B1	Address of B1 (bit address)
B2	Address of B2 (bit address)
B3	Address of B3 (bit address)
B_PRESSED	Value on input that indicate button pressed
RADIO_ACTIVITY	The way to check if the radio is active without using interrupts
RESET_RADIO_ACTIVITY	How to reset the radio detection so that a new packet can be detected
TIMER1_OVERFLOW	Check for Timer 1 overflow
CYCLES_PR_MS	The number clock cycles a timer counts that make up a millisecond
MAX_RUNTIME	How many milliseconds the timer can run on a 16-bit counter (0xFF / CYCLES_PR_MS)
GLOBAL_INT_ENABLE	Enables all interrupts
GLOBAL_INT_DISABLE	Disables all interrupts

Name	Description
T0_START	Starts Timer 0
T0_STOP	Stops Timer 0
T1_START	Starts Timer 1
T1_STOP	Stops Timer 1
T1_MODE1	Sets timer 1 up in mode 1 (16-bit timer)
T1_SET_LB	Sets the low byte on timer 1
T1_SET_HB	Sets the high byte on timer 1
INTERRUPT_T0	The interrupt vector for timer 0

Table 2. Hardware dependencies

6.3 Interrupt

If the MCU does not have an interrupt vector for timer 0, some of the LED functionality will not work. The workaround for this problem is to change the `LEDx_BLINK` functions in **system.h** to call `LEDx_TOGGLE` instead and remove the `t0_ov_interrupt()` function from **system.c**.

No other interrupts are used in this application. The interrupt status for timer 1 and radio are polled, but not handled as interrupts.

7 Troubleshoot

Q: The board does not flash after I reset. What could be the problem?

A: There could be two reasons why this is happening:

> **i.** The board may not be powered up. In this case, ensure that the board is powered up through either the USB or the battery.
>
> **ii.** The program may not be loaded. In this case, flash the development kit with the correct HEX file for your MCU and radio.

Q: LED 1 on the receiver does not light up when I press Button 1 on the transmitter. What could be the problem?

A: There could be three reasons why this is happening:

> **i.** The button may not be fully pressed. In this case, press the button harder.
>
> **ii.** The receiver or transmitter may not be powered up. In this case, power up and configure.
>
> **iii.** The receiver and transmitter are in non-compatible operational modes. Restart and configure to the correct operational mode.

Q: ACK is not receiving on the transmitter (LED 3 flashing).

A: There could be two reasons why this is happening:

> **i.** The receiver may be out of range. In this case, bring the receiver into range.
>
> **ii.** The receiver is in a different operational mode. In this case, restart and configure to correct operational mode.

8 Glossary of terms

Term	Definition
ACK	Acknowledgment
BFB	Basic Feature Board
Enhanced ShockBurst™ (ESB)	A packet based data link layer that features automatic packet assembly and timing, automatic acknowledgement and retransmissions of packets.
Enhanced ShockBurst™ with Bidirectional data (PL)	A packet based data link layer that features automatic packet assembly and timing, automatic acknowledgement, retransmissions of packets and transmission of data from PRX to PTX in ACK so that no mode change is required.
HAL	Hardware Abstraction Layer
HW	Hardware
MCU	Microcontroller
PRX	Primary receiver
PTX	Primary transceiver
SDK	Software Development Kit
ShockBurst™ (SB)	This mode gives you automatic packet transmission at a high data rate.

Benefits of Total Integration of Large RF Circuits. Market Requirements in Terms of Cost and Current. Commercial Standards vs. Optimum Solutions

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Abstract. - This paper details the implementation of a complete radio system for both 2.4GHz and 868/915/433 MHz in a 0.18 micron standard CMOS process. Topics explored include a brief market requirement analysis of standards such as Bluetooth and Zigbeee vs. optimally designed devices. Such devices are typically employed in ultra-low cost (< $1), low-power designs and the governing factors and influences of this shall be discussed. The basic architecture of the Nordic Semiconductor nRF2401 and nRF905 and the ability to implement additional digital functionality of a radio system in the CMOS process is presented. Key system functionality and performance is covered.

INTRODUCTION

There are several world wide open standards for use in the ISM bands - 433MHz, 868MHz, 915MHz and 2.4GHz. The 2 main open standards currently used for medium data rate transfer are Bluetooth and Zigbee. There are several advantages in adopting an open standard including good connectivity between equipment from different vendors, multiple vendors ensure customers have more choice and a multi-purpose device can be used for many different applications. There are however also several disadvantages; open standards contain a number of negative tradeoffs after years of compromises to set a final standard acceptable for all. A significant increase in competition between manufacturers, where all they can compete on is price, can lead to very low profit margins. Finally, optimum performance is not obtainable for a specific application with an open standard as too many demands are placed on protocol that is often irrelevant for that particular application.
Most products consist of a direct link between a company's own products and under these circumstances it is often better to use an application specific circuit which provides optimum performance, but at the cost of interoperability. This paper shows that the nRF2401 and nRF905 designed by Nordic Semiconductor offers two way communications at optimum price and current consumption, while maintaining excellent RF performance. In addition the devices are designed to be very simple to use with minimum BOM (Bill of Materials) cost.

MARKET REQUIREMENTS

There is an increasing market requirement focus on cost and current consumption while maintaining RF performance. Bluetooth and Zigbee do not meet these requirements as they are designed for interoperability between manufacturers, this leads to an increase in complexity that results in increased cost and higher current consumption.

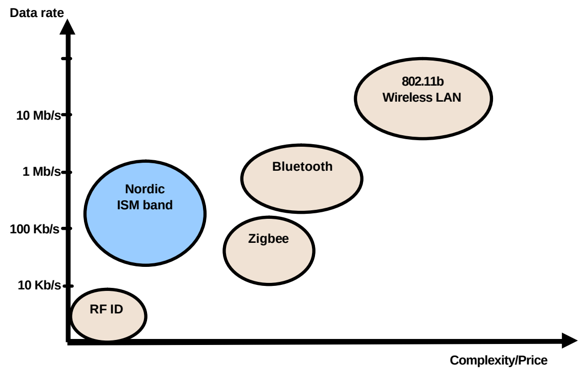

Figure 1: Nordic Semiconductor Price Location

Figure 1 shows a comparison of the data rate and complexity (price) of the nRF2401 and nRF905 with other standards. Due to the fact that these devices are optimised for a specific radio link and do not conform to a standard, the operating complexity and hence price are much lower than compared to a fully integrated Bluetooth or Zigbee circuit. Figure 2 shows how the Nordic Semiconductor circuits perform in relation to power consumption.

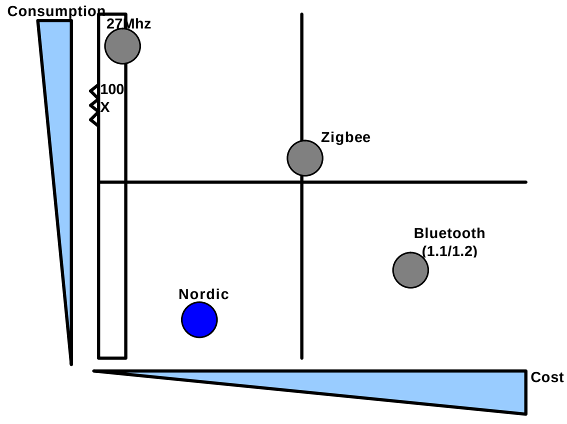

Figure 2: Nordic Semiconductor Power Location

NRF905 MULTI BAND TRANSCEIVER

The nRF905 [1] is a complete radio system incorporated onto a single piece of silicon. It uses 0.18µm CMOS technology and is designed for the 433MHz, 868MHz and 915MHz radio bands [5]. The three main aims for this product were cost, current consumption and simplicity of use while maintaining good RF performance. The main performance figures are given in TABLE 1.

Parameter	Typical
GENERAL	
Voltage Supply	1.9 – 3.6V
Data Rate (Manchester Encoded, GFSK)	50kbps
Operating temperature	-40 to 85C
Standby Current	20µA
Power Down Current	2.5µA
RECEIVER	
Sensitivity	-100dBm
Current Consumption RX Mode	13mA
TRANSMITTER	
Maximum Output Power	10dBm
Power Control	20dB
Current Consumption @ Max Power	28mA
Current Consumption @ Min Power	9mA
COST	
nRF905	$2.10
Crystal	$0.15
Passive Components	$0.05
TOTAL COST	**$2.30**

TABLE 1: General Performance for nRF905

Digital Functionality

The digital functionality in the nRF905 provides the protocol for a complete radio system. The packet that is transmitted is constructed in four parts which are; pre-amble (10 bits), address (1 byte to 4 bytes), payload (1 byte to 32 bytes) and CRC (1 or 2 bytes) see Figure 3. The pre-amble is added automatically by the transmitter and stripped by the receiver; the same is true for the address and CRC.

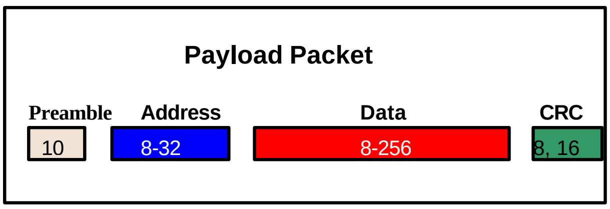

Figure 3: Internal Packet Construction

The address and the payload is set up by the user via a standard SPI (Serial to Parallel Interface) and kept in internal registers. When the packet is to be transmitted, a single external pin must be switched to the positive supply rail (Vdd) to enable the radio. The radio is automatically powered up; the packet is constructed internally and then transmitted. As soon as the packet has been sent the device enters a low current consumption mode. When a device receives a packet, it checks that the address and CRC is correct. It then strips the packet of pre-amble, address, CRC and places the data into a register ready to be clocked out of the device via the SPI interface. An external pin is set high to indicate that a valid packet has been received. The packet sits in the data register ready to be clocked out by a micro-controller. Since so much of the basic packet construction is handled internally on the nRF905, a very cheap micro-controller can be used to drive the device. The total result is an ultra low cost radio solution that provides excellent RF performance for a fraction of the current consumption used for devices that follow standards such as Zigbee and Bluetooth.

Physical Characteristics of the nRF905

The nRF905 is produced in a 32 pin QFN green package. The external component count is minimal as seen in Figure 4. The critical components are a crystal with two tuning capacitors and a single reference resistor to set the internal current. If a 50 ohm match is required, then a low cost passive balun can be implemented. Decoupling capacitors on the supply rails are recommended for optimal RF performance but if cost is the most important factor, then these too can be removed to save cost. All internal power supplies are from internal regulators which provide excellent immunity to noise. An optional external clock can be activated to drive the clock of a micro-processor; the aim again is to save cost by reducing the number of crystals needed for a complete application to just one.

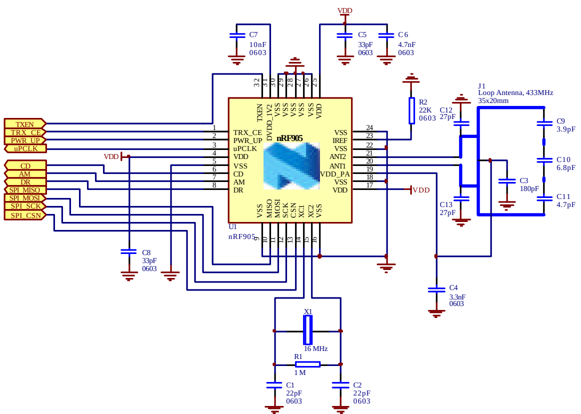

Figure 4: External Components with Loop Antenna

RF Implementation of nRF905

The nRF905 consists of a fully integrated PLL, receiver chain and transmitter chain. All RF components such as VCO, LNA, PA and mixers are included on the silicon. No external components are needed for tuning of the VCO or loop filter. By full integration large savings are achievable by customers both in design time but also in external component count.

The receiver architecture is a single conversion superheterodyne with image rejection via an internal poly phase filter. The VCO lies 1MHz above the wanted frequency and the image 2MHz above. An internal IF filter provides excellent blocking which meets the ETSI EN 300 220 blocking standard for Class 2 receivers [5].
The transmitter functions by direct modulation of the VCO. An internal digital circuit Manchester encodes the data packet to maintain a zero average DC point on the VCO. Modulation of the VCO may not be the most advanced method of modulation implementation, but it is by far the optimal method for lowest current consumption. The power amplifier is a dual stage Class A amplifier with internal current regulation. Four available output powers are available ranging from +10dBm to -10dBm.

The PLL is run at 1.8GHz and is divided to obtain the correct frequencies for the receiver and transmitter. Start up times and PLL switching times are in the order of 650µS from standby mode. The VCO incorporates an automatic calibration routine that selects the right frequency of operation. Internal inductors are used for the VCO which allows excellent control of frequency operation and phase noise. Channel resolution is 200kHz at 868MHz and 915MHz and 100kHz at 433MHz.

The crystal oscillator which is used to provide the reference for the entire circuit may be run using a 4MHz, 8MHz, 12MHz, 16MHz or 20MHz crystal. No extreme tolerance high cost crystals are needed, a simple 30ppm crystal is more than adequate to provide a stable reference for the device. An optional output clock is also available to drive a micro-controller, this may be set to 0.5MHz, 1MHz, 2MHz or 4MHz depending on the micro-controller requirements.

RF parameters such as frequency and output power are programmable and are contained in a stand alone RF configuration register. The devices own address and the address that is to be transmitted is also contained in this register. Programming is byte based via the SPI interface.

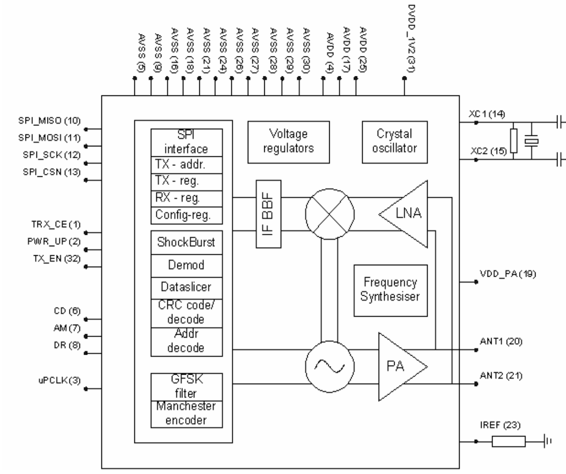

Figure 5: Internal Architecture

Typical Transmitter Operation. When the application MCU has data for a remote node, the address of the receiving node (TX-address) and payload data (TX-payload) are clocked into nRF905 via the SPI interface. The address only needs to be clocked into the device once; it is held in the RF configuration register and is used every time a payload is sent. A separate payload register is used for the payload. The application protocol or MCU sets the speed of the interface. The MCU sets TRX_CE and TX_EN high (two external pins on the device see Figure 4), this activates a nRF905 ShockBurst™ transmission. The radio is automatically powered up. The data packet is completed with preamble and address is added to the payload, CRC is calculated and also added. The data packet is transmitted using a real data rate of 50kbps, this is then Manchester encoded. After the data has been sent the device automatically enters standby mode.

Typical Receiver Operation. ShockBurst™ RX is selected by setting TRX_CE high and TX_EN low. After 650us nRF905 is monitoring the air for incoming communication. When the nRF905 senses a carrier at the receiving frequency, Carrier Detect (CD) pin (external pin on the device see Figure 4) is set high. When a valid address is received, the Address Match (AM) pin (external pin on the device see Figure 4) is set high. When a valid packet has been received (correct CRC found), nRF905 removes the preamble, address and CRC bits, and the Data Ready (DR) pin (external pin on the device see Figure 4) is set high. The MCU sets the TRX_CE low to enter standby mode (low current mode). The MCU can clock out the payload data at a suitable rate via the SPI interface. When all payload data is retrieved, nRF905 sets Data Ready (DR) and Address Match (AM) low again. The chip is now ready for entering ShockBurst™ RX, ShockBurst™ TX or power down mode.

NRF2401 2.4GHz TRANSCEIVER

The nRF2401 [2] is a complete radio system incorporated onto a single piece of silicon. It uses 0.18µm CMOS technology and is designed for the 2.4GHz to 2.5GHz radio band. The three main aims for this product were the same as for the nRF905 i.e. cost, current consumption and simplicity of use, while maintaining good RF performance. The main performance figures are given in TABLE 2.

Parameter	Typical
GENERAL	
Voltage Supply	1.9V – 3.6V
Data Rate (GFSK)	0.25/1Mbps
Operating temperature	-40 to 85C
Standby Current	20µA
Power Down Current	2.5µA
RECEIVER	
Sensitivity	-85dBm
Current Consumption RX Mode	19mA
TRANSMITTER	
Maximum Output Power	0dBm
Power Control	20dB
Current Consumption @ Max Power	13mA
Current Consumption @ Min Power	8.8mA
COST	
nRF2401	$1.40
Crystal	$0.15
Passive Components	$0.05
TOTAL COST	**$1.60**

TABLE 2: General Performance for nRF2401

General Functionality

The digital functionality in the nRF2401 is very similar to nRF905, the same shock burst protocol is used and a similar packet is constructed i.e. pre-amble, address, payload and CRC. The main difference apart from the frequency of operation is the data rate. The nRF2401 transmits and receives at data rates of up to 1Mbps.

The nRF2401 is produced in a 24 pin QFN green package. The external component count is very similar to nRF905, see Figure 6. Like the nRF905 all power supplies are via internal regulators which produce very high resistance to external noise.

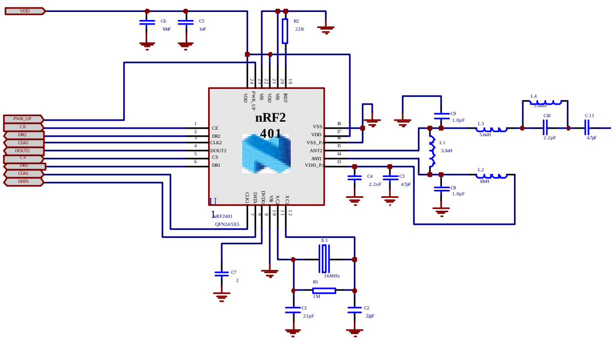

Figure 6: External Components with 50 ohm Balun

RF Implementation of nRF2401

The nRF2401 is similar in architecture to the nRF905. It consists of two fully integrated PLLs, receiver chain and transmitter chain. All RF components such as VCOs, LNA, PA and mixers are included on the silicon. No external components are needed for tuning of the VCO or loop filter.

The receiver architecture is a double conversion superheterodyne with image rejection via an internal poly phase filter at the second IF. The first IF frequency is at 350MHz and the second is at 3MHz. The device incorporates two receive channels so both the wanted 3MHz channel and an image with a frequency offset resulting in a 5MHz channel can be used independently. Two independent receiver chains allow multiple channel operation. Internal IF filters provide excellent blocking rejection. With the addition of an easily realisable frequency agility algorithm, excellent resistance to multiple possible blockers such as Bluetooth, Zigbee and WLAN is achievable. The device has been operated within a mouse while in the presence of two Bluetooth devices, one 2.4GHz frequency hopping cordless telephone and one WLAN operating on the same computer without noticeable delay in the operation of the mouse.

The transmitter functions by direct modulation of the VCO. When the packet is ready to be sent, the PLL is opened and the VCO is directly modulated at 1Mbps. The power amplifier is a dual stage class AB amplifier with internal voltage regulation. Four available output powers are available ranging from 0dBm to -20dBm. With the transmitters low peak current consumption, (see TABLE 2) the resulting average current consumption for the transmitter is approximately 30nJ/bit. This includes start up time and all energy needed to transmit one packet of data with pre-amble/address/payload and CRC.

The PLL is run at 2.45GHz with a 1MHz channel resolution. Start up times is in the order of 200µS from standby mode. The VCO incorporates an automatic calibration routine that selects the right frequency of operation. Internal inductors are used for the VCOs which allows excellent control of frequency operation and phase noise. All RF parameters are programmable and are contained in a stand alone RF configuration register. Programming is via a 3 wire SPI.

Transmitter Typical Operation. The external MCU interfaces with the following pins: CE (Chip Enable), CLK1 and DATA. When the application MCU has data to send the CE pin is set high. This activates nRF2401 on-board data processing. The address of the receiving node (RX address) and payload data is clocked into the nRF2401. The application protocol or MCU sets the speed. The MCU sets the CE pin low; this activates a nRF2401 ShockBurst™ transmission. The RF front end is powered up, the RF packet is completed (preamble added, CRC calculated) and the data is transmitted at high speed (250 kbps or 1 Mbps configured by user), finally the device returns to stand-by mode when finished.

Receiver Typical Operation. The MCU interface pins: CE, DR1 (Data Ready 1), CLK1 and DATA (one RX channel receive mode). To activate RX, the CE pin is set high, after 200µs settling the device is monitoring the air for incoming communication. When a valid packet has been received (correct address and CRC found), nRF2401 removes the preamble, address and CRC bits. It then notifies (interrupts) the MCU by setting the DR1 pin high. The data is available for the MCU to clock out. When all of the payload data is retrieved the device sets DR1 low again and is ready for a new incoming data packet.

CONCLUSIONS

There exists a very large market for optimized circuits that do not conform to an open standard such as Bluetooth or Zigbee. Cost, complexity and current consumption can all be significantly decreased when not complying with an open standard. For companies whose products communicate solely with its own products an open standard can be overly expensive and complicated, it is therefore often better to use a specialized circuit.

Integration of the entire RF circuitry not only minimizes external cost but also greatly increases ease of use for the final customer; this greatly reduces the time to manufacture. With a fully integrated RF circuit it is relatively easy to add digital functionality such as internal protocols which include pre-amble, address and CRC. By incorporating these protocol functions as part of the device a very low cost MCU can be used hence further reducing the total solution cost.

Further digital functionality can be incorporated by including an onboard micro-controller and ADC. Nordic Semiconductor has already implemented this with products nRF24E1 [4] and nRF9E5 [3]. These products include the RF cores discussed in this paper, ie nRF905 and nRF2401 but also include an 8051 MCU and 10 bit ADC in a single chip solution. The additional cost to the customer is approximately 25 cents.

REFERENCES

1. Nordic Semiconductor. nRF905 Datasheet

2. Nordic Semiconductor. nRF2401 Datasheet

3. Nordic Semiconductor. nRF9E5 Datasheet

4. Nordic Semiconductor. nRF24E1 Datasheet

5. ETSI EN 300 220-1. V1.3.1. 2001. Electromagnetic compatibility and radio matters. Short range device. Radio equipment to be used in the 25MHz to 1000MHz frequency range with power levels ranging up to 500mW. Part 1.

Introduction to wireless networks

1. Preface

This white paper raises some fundamental issues the design engineer should address before deciding upon a communication approach for a wireless network. As no universal wireless network solution exists, it should be custom tailored to suit the application demands.

Defining your application communication characteristics is the key to ensure optimal communication reliability and resistance to interfering noise sources.

2. The network concept

In multiple node applications where one or several nodes need to be able to communicate with all, or a subset of the other nodes in the system, some form of network approach is needed.

A general definition of a network is:

"Any arrangement of elements that are interconnected"

The fundamental purpose of the network realization is simply providing a defined procedure of interconnection and information flow between nodes. Secondary issues are ensuring the necessary communication reliability in order to obtain satisfactory application functionality.

Typical network scenarios are:

- The nodes in a system are scattered to such an extent that the nodes are not within radio range of all the other nodes in the system.

  Example: Consider an industrial sensor application where the nodes are positioned at various locations in a large building. All sensor nodes report status to the master control node at predestined time intervals, or if the measured value exceeds given alarm thresholds.
  If a sensor node is outside radio range of the master control node, the information must be relayed to the master control node by means of networking. This is achieved by other sensor nodes located between the two nodes retransmitting packages not addressed to themselves.

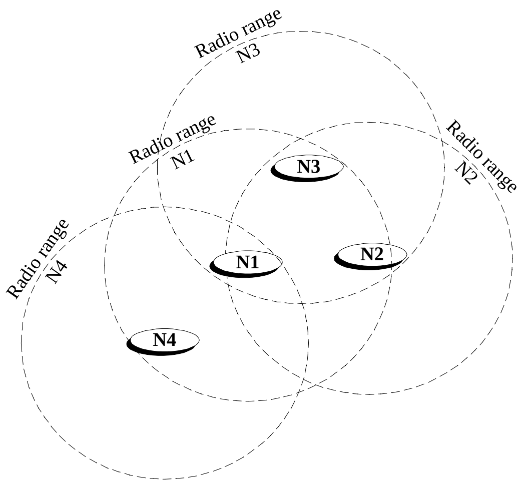

Figure 1 - Non overlapping radio ranges

- Systems where the inherent mobility of nodes, transfer nodes in and out of radio range of other nodes in the system.

  Example: Consider a system consisting of nodes attached to mobile personnel working at a location larger than the radio range. In order to exchange information between any two nodes, retransmission of some packages is likely. As nodes are considered mobile, the available nodes capable of relaying the package to the correct recipient are changing over time.

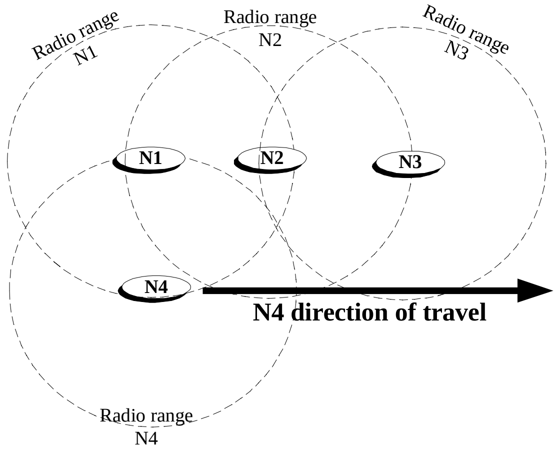

Figure 2 - Time varying radio range overlap

- Systems where it is natural to divide the system nodes into groups. A center node in a group may communicate and exchange status information with center nodes in other groups.

  Example: Consider a system with multiple nodes grouped at separate locations. This might be sensors in an alarm system with multiple sectors. In this case it might be natural that the nodes report their status to a sector master node. In case of sector alarm, the sector master node transmits an alarm status message to the alarm master node.

3. Network approaches

3.1. Star network

Perhaps the easiest network approach is the star topology illustrated in Figure 3.
All communication is directed via the central node, which retransmits the information to the destination node. The central node acts as a relay station and must therefore be positioned within radio range of all nodes in the network. Theoretically, radio range of the network nodes may be as much as doubled.

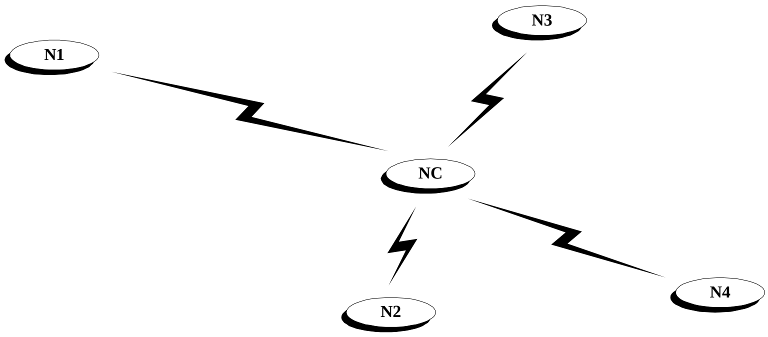

Figure 3 - Star network principle

Below is described a basic network scenario where node N4 needs to transmit information to node N1.

The course of action is as follows;

- N4 needs to alert N1
- N4 generates and transmits a package to node NC, requesting acknowledgement
- Nodes N2 and N3 ignore the package as they are not the designated recipients
- NC recognizes itself as recipient and transmits an acknowledgement package addressed to node N4
- Nodes N1, N2 and N3 ignore the package as they are not the designated recipients
- N4 recognizes itself as recipient of the acknowledge package from NC, ending the communication with N1
- NC retransmits the package to N1 without delay (provided there is no other channel traffic)
- N1 recognizes itself as recipient of the retransmitted package and transmits an acknowledgement package addressed to node NC
- Nodes N2 and N3 ignore the package as they are not the designated recipients
- NC recognizes itself as recipient of the acknowledge package from N1, ending communication

In total, 4 packages are sent in order to achieve a successful acknowledgement of a transmitted package. In a point-to-point system (assuming the nodes are within radio range of each other), only two packages are generated.

The obvious bottleneck of the system is the communication capacity of the center node. The communication intensity of the network must therefore not exceed the maximum throughput of the center node.

An advantage of using a single node to control all traffic, is that the system communication delay is kept at a minimum. The center node may retransmit any package without delay, as long as no other traffic occupies the operating frequency. The system delay issue described in Chapter 4.7 is therefore not applicable for this network approach.

The use of this topology is limited to applications where the node positions are fixed or where node mobility is limited. The placement of the center node is dictated by the application environment and node distribution. As the center node is performing the 'lionsshare' of the work, it is often desirable that it is a stationary mains-fed unit so that the current consumption is no longer an issue.

3.2. Single retransmission of received packages

A simple network approach is that all nodes are to retransmit received packages not addressed to themselves once. The link layer makes sure that previously received packages are identified, avoiding infinite retransmissions. This necessitates a memory function where recently received package identity information is stored.

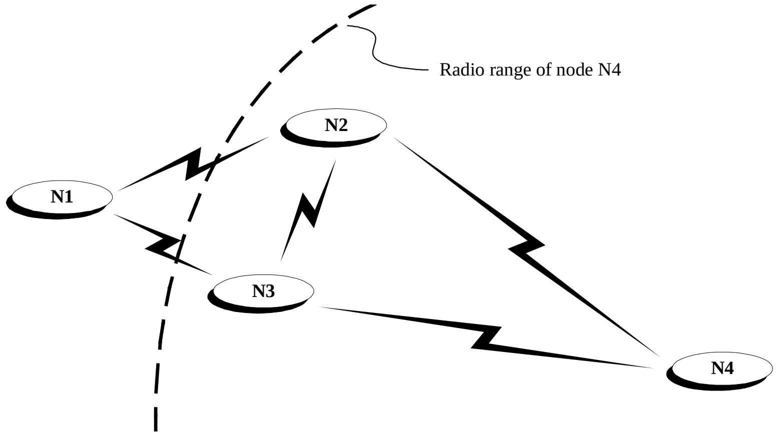

Figure 4 - Basic network communication scenario

Figure 4 illustrates a basic network scenario where node N4 needs to transmit information to node N1 outside radio range.

The course of action is as follows;

- N4 needs to alert N1
- N4 generates and transmits a package to node N1, requesting acknowledgement
- The nodes N2 and N3 are the only recipients, as N1 is outside radio range of N4
- N2 and N3 identify the recipient address to be another node
- N2 and N3 retransmits the package at a random time instant, remembering the package ID to prevent multiple retransmissions
- N4 receives the retransmitted package and discards it as a retransmitted version of it's original transmitted package
- N1 receives a retransmitted package from N2 or N3 (depending on which of the two nodes that retransmitted the package first)
- N1 recognizes itself as recipient and transmits an acknowledgement package addressed to node N4
- N1 receives the second retransmitted package from N2 or N3, and ignores it as already being processed
- The nodes N2 and N3 are the only recipients of the acknowledge package from N4, as N1 is outside radio range of N4
- N2 and N3 identify recipient address to be another node
- N2 and N3 retransmits the acknowledge package to node N4 at a random time instant
- N1 receives the retransmitted acknowledge package and discards it as a copy of the recently transmitted package
- N4 receives a retransmitted acknowledge package from N2 or N3 (depending on which of the two nodes that retransmitted the package first)
- N4 recognizes itself as recipient of the acknowledge package, ending the communication with N1
- N4 receives the second retransmitted acknowledgement package and discards it as a copy of the previously received package

In total, 6 packages are sent in order to achieve a successful acknowledgement of a transmitted package. Using the single dedicated relay node as described in the star network chapter, would in turn generate 4 packages.
A network inevitability introduces more traffic than a point-to-point system. The network strategy described is based on the principle that all received packages are retransmitted once unless the receiving node is the recipient.

Assuming all the nodes in the network are within radio range of each other, the total number of packages generated for a simple package transmission with acknowledge is N+N-2. N is the number of nodes in the system.

The timeslot in which retransmission is to take place, must be long enough to allow all nodes to retransmit any received package in order to avoid collision. *System response time is hence proportional to the number of nodes in the system* (see Chapter 4.7).
As can be seen from Figure 7; in order to avoid blocking or traffic jamming, the average package rate (initiated by a node and not by retransmission), must be less than 1/[retransmission timeslot].

This simple networking approach is suitable for systems where a limited number of nodes coexist within the radio-range of any given node, resulting in a limited system delay/response time. The approach is also robust against individual movement of network nodes.

3.3. Mapping of gateways through neighboring nodes

A more bandwidth efficient, but also more complex approach is illustrated in Figure 5. This solution is based on the assumption that the nodes have the ability to 'learn' the existence of the other nodes in the system, and not only the nodes being within its own radio range. This information is stored in a table, which is updated immediately as the network is established, and when communication is lost between two nodes during normal operation.
The table has the following principal organization and information content:

Active nodes in the application network	Which node within radio range is the gateway to the network node in the left column
1	X
2	Y
...	...
N	Z

Two different package types are used in the network;

- Network mapping packages
- Application communication packages

When a network mapping package is sent, all nodes intercepting the package answers with its address and a list of nodes of which it is able to communicate with. This enables the node to build a table where gateways to nodes outside radio range can be identified. If two nodes both provide a gateway to the same distant node, the first network mapping package received decides the gateway node.
The mapping sequence involves sharing information of the network topology by asking: "Who can hear me, and who can you communicate with?"

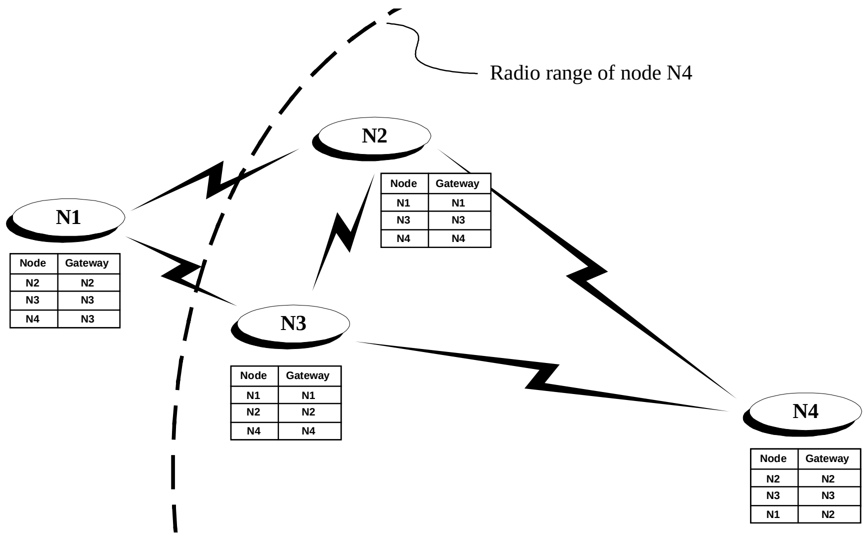

Figure 5 - Network mapping of neighboring nodes reducing retransmission traffic

When a node transmits a package to a dedicated node, the node has two options; either sending the package directly to the recipient (if it is within radio range) or via a gateway node. If the recipient node is within range, network traffic is limited to only two packages; the information package, and the resulting acknowledge package from the recipient.

If the package has to pass through a gateway node, the number of packages is doubled. Normal procedure is that the gateway node sends an acknowledgement package to the package originator, taking over the responsibility of the transmission. The gateway node then transmits the package to the recipient (or via another gateway node) closing the transmission upon reception of an acknowledgement package.

If a recipient node has moved out of range, or is obscured by a noise source, the package transmission chain is broken. If a node fails to receive an acknowledge package within a predefined time period, it initiates the mapping procedure in order to rebuild a valid communication table. This enables the establishment of an alternative package transmission path through the network. For a relatively stable network, mapping traffic is significantly lower than the normal communication traffic.

Note that as the communication path is known at transmission, system response time is kept at a minimum, as opposed to the network approach described in Chapter 3.2.

The course of action is as follows;

Network topology mapping-sequence (performed by all nodes);

- Node N transmits a network mapping package asking: "Who can hear me, and who can you communicate with?"
- All nodes receiving the package responds with a 'I can hear you, and I can communicate with nodes; X, Y, Z ...'-package
- Node N updates the network gateway table
- Sequence is repeated until the content of all tables is stable

Communication sequence;

- N4 needs to alert N1
- N4 checks its network communication table and finds that N2 is the gateway to N1
- N4 generates and transmits a package to node N2 (with final designator address N1), requesting acknowledgement
- N3 ignores the package as it is not a recipient or a gateway node
- N2 receives the package and finds that the designated recipient is N1
- N2 sends an acknowledgement package to N4
- N2 checks its network communication table and finds that N1 is within its radio range
- N4 receives the acknowledge package from N2, closing the transmission as N2 has taken over responsibility of the transmission
- N2 relays the package to the final recipient, N1
- N1 receives the package and transmits an acknowledgement package addressed to node N2
- N2 receives the acknowledge package, ending communication with N1

In total 4 packages are sent in addition to the packages generated in the network mapping sequence, where 8 is the absolute minimum number of network mapping packages. The total number is depending on the node communication order and may vary.

4. Parameters to consider in a networking application

Numerous networking approaches exist and most are custom tailored to a given application or system. It is useful to define some fundamental characteristics of the application before deciding upon a network solution. *Keeping the complexity as low as possible without compromising the application functionality* is the primary design challenge.

The following aspects of the application should be considered and defined:

4.1. Total number of nodes in the system

The number of system has to be seen in conjunction with the application communication activity. A high number of nodes within radio range of each other will cause significant traffic if a simple retransmission strategy is to be used. Lets assume a package size of 100 bits and a datarate of 100Kbit/sec. The duration of a package is then 1ms. Assuming 100 nodes are within radio range, retransmission will cause 99 packages to be transmitted, resulting in 100ms of transmission time for a single package. As the system is not synchronous, guard-time needed to avoid collisions must be added (See Figure 8).
For systems with low communication activity, this is generally not a problem, but as the information flow increase, the network may soon be jammed due to retransmitted packages.

4.2. Network traffic intensity

The retransmission approach must be used with care in systems with high communication intensity. Unnecessary traffic generated by the network may jam the application as a result. Key information is how often nodes within range of each other is active (TX) and the duration of each subsequent package. This is decided by the application and most often involves statistical considerations.

In cases where a large amount of data is to be transferred, the network may be designed to process data transfer on different frequencies. By organizing communication this way, the network is still operational even during lengthy data transmissions. The nodes first establish contact on the signaling channel, then shifting the communication to an available traffic channel.

4.3. Geographical distribution of nodes

In systems where the distance between nodes is small compared to the radio-range of the radio transceiver, a more elaborate network approach is needed in order to minimize traffic resulting from retransmissions. If the number of nodes is large, both described methods generate unnecessary traffic, as a high number of retransmissions within a limited area are superfluous.

4.4. Inherent mobility of nodes

Where nodes are to be mobile after initial installation of the system, they might move out of radio range of some nodes, whilst moving into radio range of others. If a minimum traffic network approach (as described in Chapter 3) is to be used, the system nodes must update the *nodes-within-range*-table at constant intervals. This will in turn introduce extra traffic load and should be taken into consideration.

4.5. Application node hierarchy

Where it is natural to divide the system nodes into subgroups, the network solution may be designed to inhibit inter-group communication except at key node level as shown in Figure 6 below.

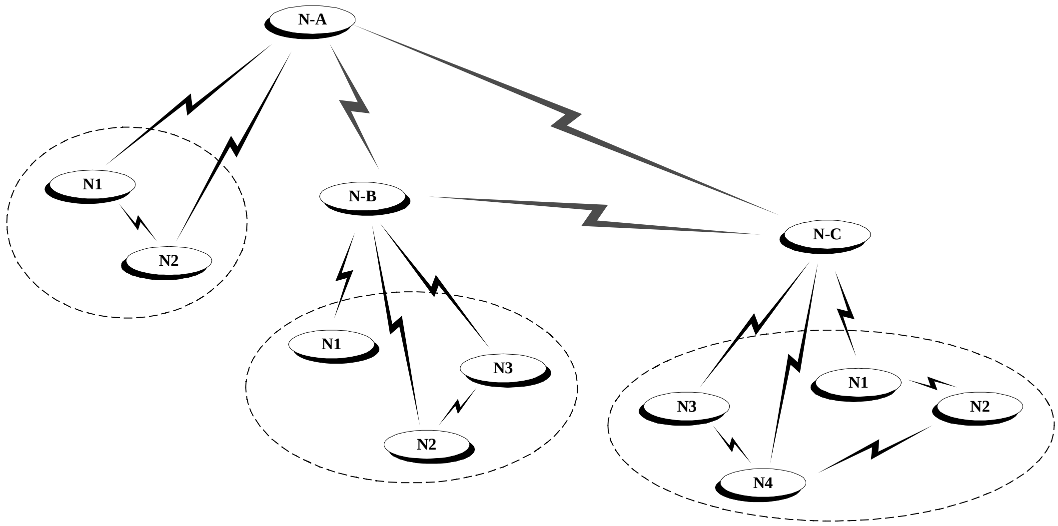

Figure 6 - Hierarchy organized network

4.6. Communication reliability requirement

The communication reliability, often referred to as *Quality of service* (QoS), must be defined for the application. In this context, the question is; what information loss can be tolerated whilst still achieving satisfactory operation?

For systems depending on file transfer, the answer is *none*; every package sent must be received in order for the application to work. The network approach must therefore be based on repetitive package transmissions until an acknowledge-package from the recipient is received.

For a sensor system, one may normally accept the loss of a larger amount of packages when the application is repeating sensor information at regular intervals.

4.7. System response time

An important parameter is the system response time required. In this context: "What is the maximum time between a package is received before it may be retransmitted?"

Consider the imaginary network shown in the figure below.

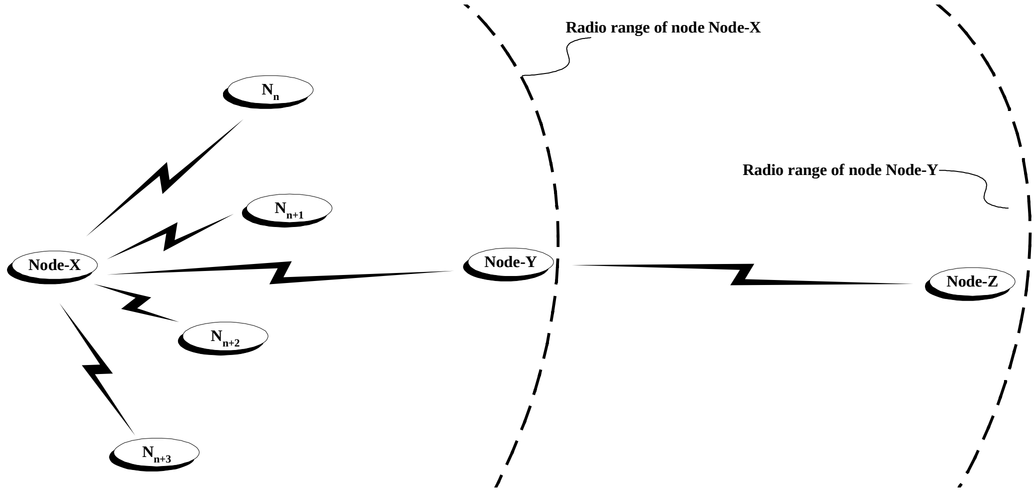

Figure 7 - Network example; System response delay due to retransmission

Node X on the extreme left transmits a package to the node on the extreme right, Node Z. These two nodes are assumed to be out of radio range of each other. Also; the only node within range of Node Z, is Node Y. The message must therefore be relayed through Node Y. For the retransmission approach described in Chapter 3.1, the transmission procedure is sketched in Figure 8. Note that that retransmission is performed at a random time instant to avoid collisions. Worst case package delay is governed by the defined retransmission period, t_{reTX}. This is a system parameter proportional to the number of nodes expected to be within range of each other.

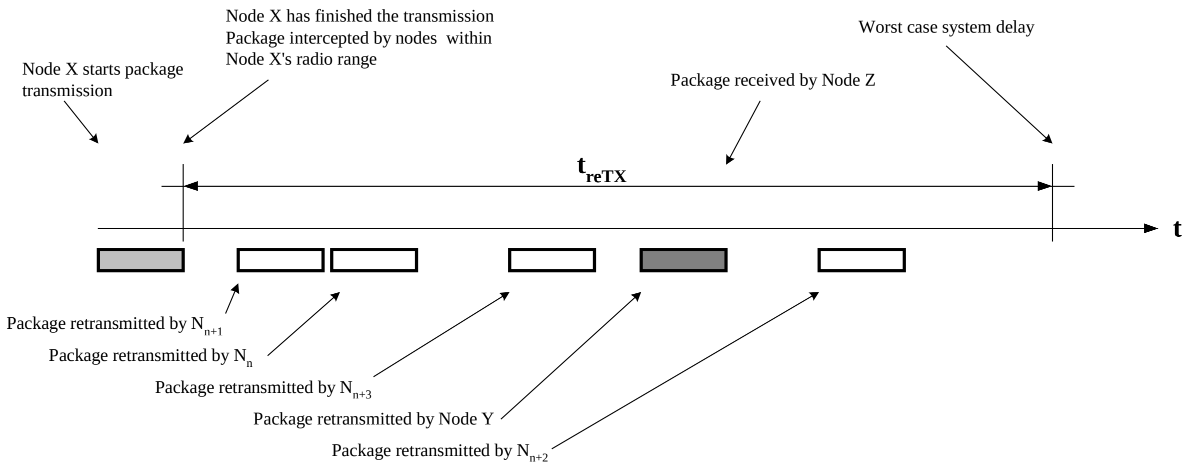

Figure 8 - Network example; System response delay due to retransmission

For the network approach described in Chapter 3.3, this is not an issue as the relay node retransmits the message immediately after reception.

4.8. Presence of noise or potential jammers in the field of operation

It is important to comprehend that the power density of the signal transmitted from any system node, as well as noise sources, decrease with 1/range^2. Presence of noise is thus *generally not a problem if the noise power density at the receiving node is less than the power density of the incoming package at the same location*.

Typical noise sources are other systems operating in the same frequency band. A networking approach generally enhances overall system noise rejection as other nodes provide alternative communication paths between nodes, both geographically and in the time domain. Figure 9 illustrates the general principle.

Assume a noise source is positioned at a certain location:
If Node 1 sends a package to Node 2 during a period when the noise source is active (t=1), two scenarios are likely when the direct transmission path is lost due to interference:

A: Node 2, being positioned outside the noise source interference range, intercepts the package and retransmits it via Node 3, reaching Node 4.

B: The package is intercepted by Node 2 at t=1, and then retransmitted at time instant t=2 when the noise source is silent.

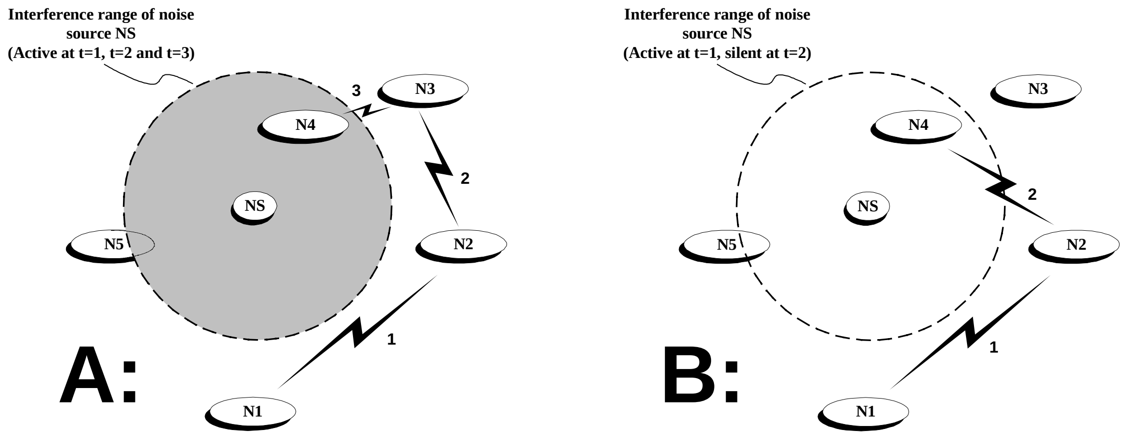

Figure 9 - Alternative package propagation paths in the presence of noise

It is important to be aware of the fact that interference is likely in certain environments, but mostly only for short periods of time. The key questions are:

- What are the likely interference sources?
  (other ISM applications, parallel networks etc.)
- What is the probability of interference over time?
  (ISM applications are almost always active for only a short time)
- How will temporary jamming affect the application?
- How may the system communication protocol be designed so that temporary jamming does not affect application functionality?

5. Summary

"There is more than one way to skin a cat...". The same applies to network implementation. As communication networks are almost always custom tailored to a specific application, there is no "all purpose" network solution. The key to efficient communication and Quality of service is to fully understand the dynamics of your application communication requirements. Define your application requirements and characteristics in terms of;

- maximum response time
- maximum number of system nodes
- package loss tolerance
- network peak traffic (packages/sec or bits/sec)
- node mobility behaviour
- application operation area infrastructure (attenuation, mains availability etc.)
- definition of probable interference scenarios

Once defined, the network solution capable of meeting your specifications may be chosen.

The characteristics of the examples described in Chapter 3 are listed in Table 1.

	Star network	**Retransmission**	**Gateway mapping**
Maximum radio range increase	2 · node radio range	(N*-1) · node radio range	(N*-1) · node radio range
System communication delay	Low	High	Low
Complexity	Low	Medium	High
Node mobility	Low	High	High
Communicaton efficiency	Medium	Low/Medium**	High/Medium**

*: N is the number of nodes in the system
**: Depending on the number of nodes in the system

Table 1 - Network examples, summary of characteristics

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RF Performance Test Guidelines

White Paper

November 2008

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1 Introduction

RF front end performance plays a vital role in any RF product and it is also the main area of testing when qualifying your RF application for regulatory standards for RF emissions (like FCC, ETSI and TELEC). This document aims to give you the necessary background for understanding the terminology, how different RF parameters are tested and why it is important to test these parameters. The following are the main tests when verifying the frontend performance of a nRF device:

- Output power
- Frequency accuracy
- Spurious emissions:
  - Harmonic output power
  - RX local oscillator leakage
- Deviation and modulation bandwidth
- Receiver sensitivity
- Receiver selectivity
- Current consumption

The general guidelines presented in this document are complemented by device specific application notes that give a brief guide on how to run the various tests on our different radio devices (nRF family). Performance tests must be conducted on more than one unit to avoid misleading conclusions. Running the same test with the same setup on multiple units should give you similar results, if not there may be something wrong with the unita or with your test set-up.

> **Note:** The tests described in this document are intended for prototype stage testing. Some of these tests can also be used during production testing while others are too complex and take too much time to be feasible for use in a production line.

2 Output power

The output power of a radio is a critical factor for achieving your desired communication range. Output power is also the first test criteria to qualify for regulatory standards for RF emissions (like FCC, ETSI and TELEC).

2.1 Carrier wave output power

Although all nRF devices send short packages of data in normal operation, output power is best measured while sending continuously on one frequency. To get this frequency output, modulation (see chapter 5 on page 18) is turned off, this concentrates all the output power from the nRF device into one frequency called a transmit carrier wave or short TX carrier. With the power concentrated at one frequency instead of spread across the modulation bandwidth you get the absolute maximum power readings (see chapter 5 on page 18).

Measuring TX carrier output power is the most straightforward of all RF measurements.

2.1.1 Test method and setup

The output power from a radio device in transmit mode is easily measured with a spectrum analyzer in the simple hardware (HW) test setup shown in Figure 1. You should use cables for all output power measurements on a radio device (this type of setup is called a conducted measurement). Using cables is the only way to have a controlled transfer of power from the radio to the measurement equipment. Measuring power accurately over the air requires a fully controlled environment and calibrated antennas, this means you need a calibrated reflection and an echo free (anechoic) chamber.

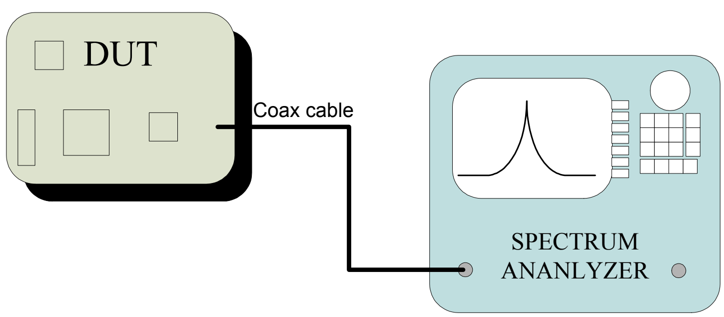


Figure 1. TX carrier hardware test setup

The radio circuitry, generally referred to as the Device Under Test (DUT) and the spectrum analyzer are connected with a coaxial or COAX cable. The spectrum analyzer must cover the frequency range you want to measure. The frequency range depends on the operation frequency of the radio you want to measure.

Spectrum analyzers have the following four main configurations that must be set up before testing:

- **Center frequency:** Equal to the RF frequency configured in the DUT.
- **Frequency span:** 1-10 MHz (depending on the DUT).
- **Amplitude:** Must be set higher than the maximum output power of the DUT.
- **Resolution Bandwidth (RBW):** 30kHz- 300kHz depending on the frequency span used. Generally, it is best to set the spectrum analyzer RBW to auto so that it follows the span.

Recommendations on the above parameters are found in the nRF performance test instructions for each nRF device.

The amplitude setting of a spectrum analyzer controls how much gain is set in the input stages of the instrument. Setting the amplitude setting wrong may saturate the input stages and distort the measurements. For example, if the DUT transmits 0 dBm (1 mW) the spectrum analyzer amplitude should be set to +5 dBm (that is, the input signal can be up to +5dBm strong). If set to 0 dBm or lower, the input amplifiers of the spectrum analyzer may saturate. This causes unwanted frequency components showing on the spectrum analyzer screen, making the DUT seem to have a lot of unwanted output frequencies.

The RBW has an impact on how easy it is to read the spectrum and on the accuracy of the output power reading. Using the auto setting (found on most spectrum analyzers) on RBW usually resolves this. Figure 2. is an example of a carrier output spectrum.

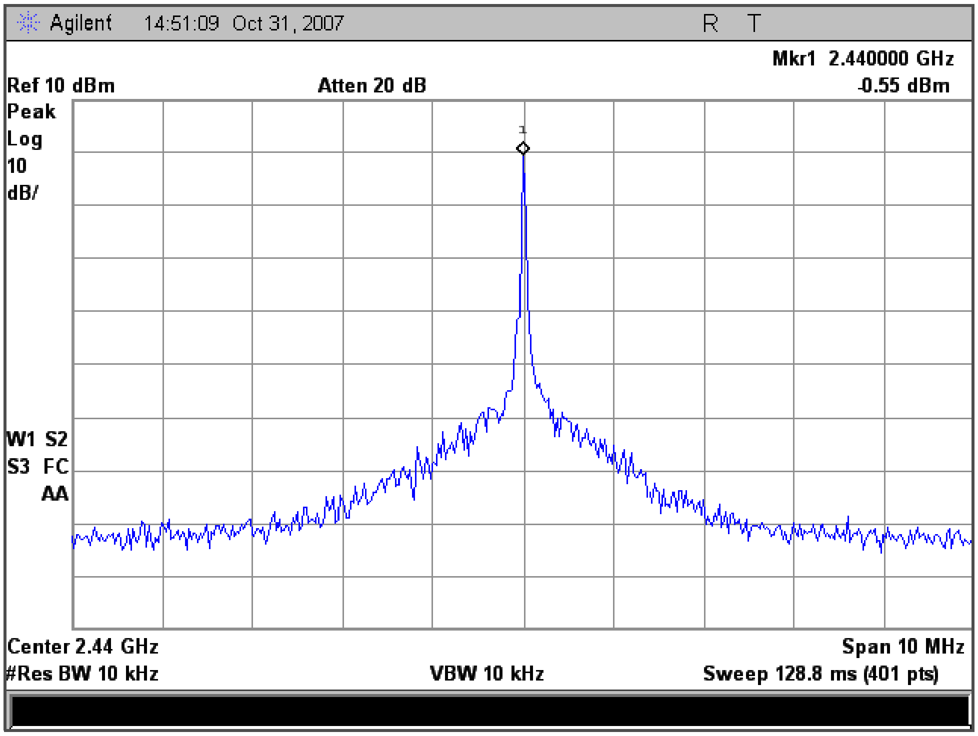


Figure 2. nRF2401 single carrier output spectrum

How to set different nRF devices in a constant carrier transmit mode is described in the nRF performance test instructions for each nRF device.

2.1.2 Calibration of measurements

You always lose some of the power the DUT provides in the test cables and connectors when working on high frequency systems like RF. This is called insertion loss and how much loss you have depends on frequency and the type of cable and connectors you are using.

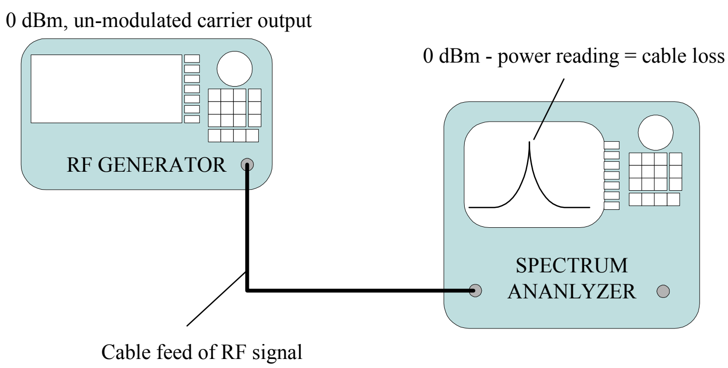


Figure 3. Calibrating DUT output power measurement

To compensate for this loss, measure the insertion loss from the DUT to the spectrum analyzer. You can do this by replacing the DUT with a calibrated RF generator sending a carrier wave with a known output power of, for instance, 0 dBm (Figure 3.). The power level you now measure on the spectrum analyzer should be slightly lower than the output set on the generator, this gives you the insertion loss of the cable and connectors. The insertion loss value should then be added to all your later measurement results. Adding the insertion loss value is called calibrating the test setup because this gives the true output from the DUT.

Note: If you do not have a RF generator you can use an estimated loss in RF cables of ~0.5 dB per meter. However, this is only a rough estimate.

The type of connectors you are using must be designed for the frequency you are measuring on. For example, BNC COAX connectors are usually specified for use only up to 1.8 GHz. On a 2.4GHz system these connectors may give a significant loss while SMA COAX connectors, which are still within specification at 2.4GHz (and above), give you very little insertion loss.

2.2 Carrier wave sweep

If the DUT can operate over a range of frequencies, called a frequency band (for example, ~80 MHz in the case of a 2.4GHz ISM band radio), it is important that you measure the output power at different frequencies across the DUT's frequency band. This will show you the overall output power performance. If the radio output power is the same across all frequencies, you have verified that your communication range stays the same no matter which frequency is used.

Measuring output power across a frequency band can be done by running the TX carrier test on a few frequencies manually. But, the best way to measure the output power is to program the DUT to make a frequency sweep covering all the frequencies of the band and set up the spectrum analyzer slightly differently from the TX carrier test in section 2.1. The necessary routine to implement a TX carrier wave sweep is shown in Figure 4. This routine runs consecutive TX carrier wave tests on all available RF channels in the specified band.

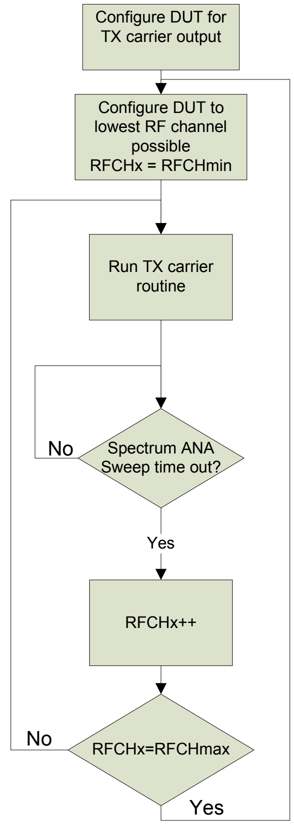


Figure 4. DUT TX carrier sweep flowchart

On some nRF devices the carrier wave sweep must run in an external MCU while other nRF devices have it built in. Please refer to the 'nRF performance test instructions' document for each device.

The spectrum analyzer setup should be as follows:

1. **Center Frequency:** Middle of frequency band in use. For example, 2.4GHz: 2.42GHz.
2. **Span:** Wide enough to cover the entire band in use. For example, 2.4GHz: > 83 MHz.
3. **RBW:** Auto like in the TX carrier wave test.
4. **Amplitude:** Higher than the maximum DUT output power, same as in the TX carrier wave test.
5. Peak or max hold measurement enabled.

Since the DUT only transmits on one frequency at a time, the spectrum analyzer must utilize a feature called peak hold measurement to gradually capture all the carrier frequencies across the band. This feature enables the spectrum analyzer to store all the maximum power readings it has captured. The peak hold feature is found under the AVERAGE or TRACE menu on most spectrum analyzers.

As shown in Figure 4. the carrier wave sweep depends on a spectrum analyzer sweep time out. This means that to capture all TX carriers fast, the DUT should keep the carrier on one channel for as long as it takes the spectrum analyzer to sweep across the whole measurement span. The reason it takes the spectrum analyzer a little time to measure the whole frequency span is because it does many consecutive power measurements on small segments of the span set by the RBW.

This sweep time varies a lot from analyzer to analyzer, you can usually find it listed on the spectrum analyzer's main screen directly or in the SWEEP menu of each instrument. In Figure 5. you can find the sweep time (4 ms) listed in the lower right corner of the screen. This sweep time may be up to several hundred milliseconds depending on the processing speed of the spectrum analyzer.

What you want to achieve is shown in Figure 5. This figure shows the TX carrier output on all the possible frequencies this radio device can use. From the envelope formed by the TX carrier wave sweep you can see if the output power is level across the band or is stronger in particular areas.

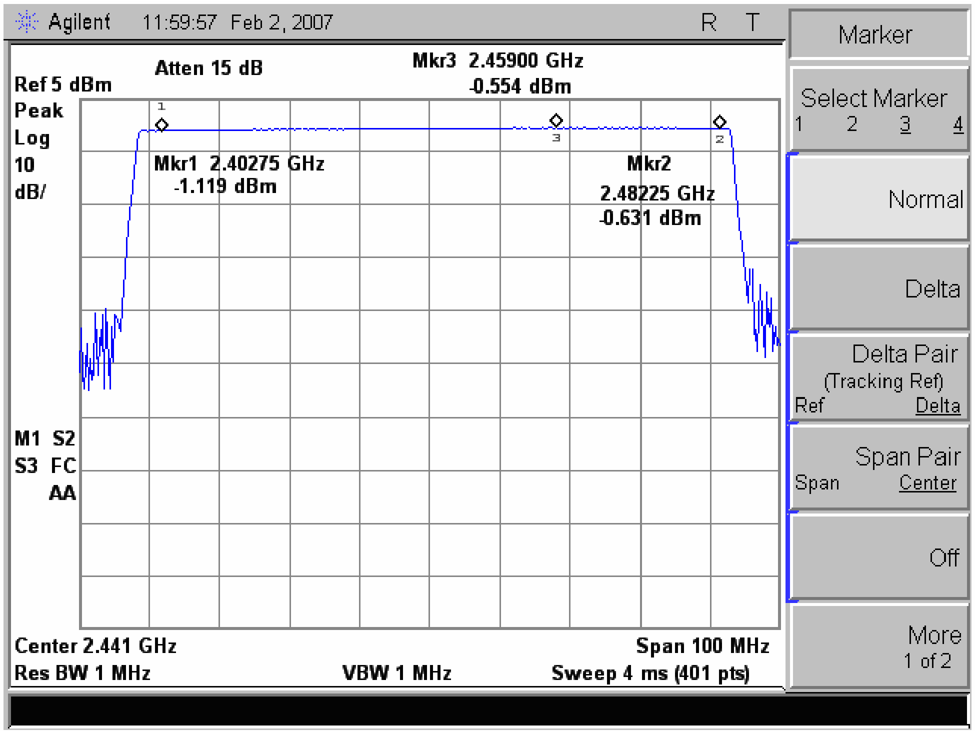


Figure 5. TX carrier frequency sweep across the 2.4GHz band

The TX output power of all nRF devices is tested during our production to guarantee that the devices provide the same power across the band. Therefore, the carrier wave sweep test primarily shows the frequency response of the antenna matching circuitry found between the nRF device and the antenna on your PCB. You can use this information to optimize the antenna matching network, a vital task in any RF design.

3 Spurious emissions

Spurious emissions is radiation from a RF device that is outside the appropriate frequency band. These emissions act as noise for RF systems operating in other frequency bands. The most important spurious emissions are harmonics of the RF output in TX mode and Local Oscillator (LO) leakage in RX mode. Keeping these unwanted RF emissions under control is one of the main criteria for qualifying for governmental regulations like FCC, ETSI or TELEC.

3.1 Harmonic output power

Harmonic output is spurious (unwanted) TX power that appears on harmonics or multiples of the RF frequency used. In the case of a 2.4GHz system these appear at 2 x 2.4GHz = 4.8GHz, 3 x 2.4GHz = 7.2GHz, 4 * 2.4GHz, and so on. In Figure 6. a wide band spectrum analyzer measurement shows the desired output power on 2.4GHz and the unwanted harmonic outputs from a nRF device.

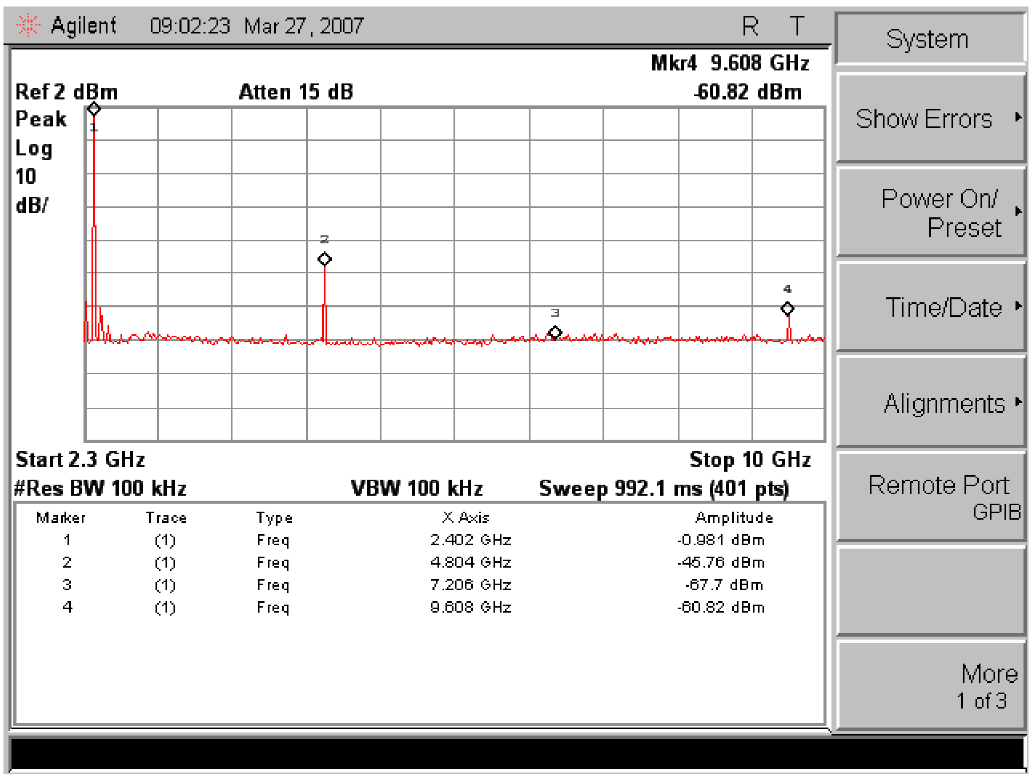


Figure 6. TX harmonic output power

The maximum power allowed on the harmonics depends on regulatory standards and you must identify which regulations you need to pass. Each nRF device is made to comply with certain regulatory standards.

nRF devices suppress harmonics to a certain degree while the remaining filtering is done in the antenna matching network found on the PCB outside the nRF device. When you are testing and optimizing an application for maximum RF output power you must always check the harmonic output. An application with good output power is useless unless you also meet regulatory requirements on maximum harmonic output power level.

3.1.1 Test method and setup

The spurious emissions tests use the same HW setup as the TX carrier wave output power tests. To get accurate measurements you must calibrate for cable insertion loss on each of the frequencies you conduct tests on.

The DUT configuration and test routine (single channel or carrier sweep) from the output power measurements is used in this test. You need to configure the spectrum analyzer to the correct frequency range.

The spectrum analyzer setup should be as follows:

- **Center frequency:**
  - TX carrier wave: N times the RF frequency configured in the DUT where N = 2, 3, and so on.
  - TX carrier wave sweep: N times center of RF band in use. N = 2, 3, and so on.
- **Frequency span:**
  - TX carrier wave setup: the same as TX carrier wave output power.
  - TX carrier wave sweep setup: N times the span used at the TX carrier sweep test.
- **Amplitude:** the same as TX output power test.

Keeping the amplitude at the same level that is used for the TX output power test is very important. Although the main carrier is now outside the view on the spectrum analyzer screen and the output power on the harmonic you measure on should be at least 30 to 40 dB lower than the main carrier output, the spectrum analyzer input is still subjected to the power from the main carrier. Reducing the amplitude setting may lead to saturation of the spectrum analyzer input stages.

Figure 7. shows a close up of the 2. harmonic (2x TX frequency) of a 2.4GHz nRF device. The device is running a TX carrier sweep and Figure 7. shows all the 2. harmonic frequencies the device can emit.

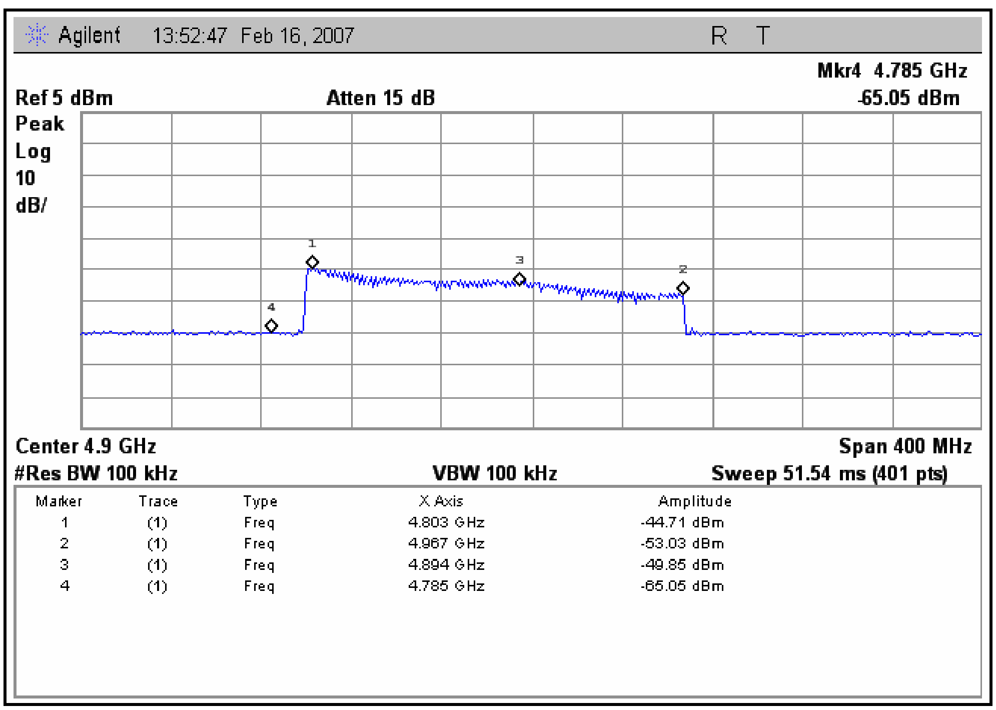


Figure 7. Harmonic output power

3.1.2 Measurement accuracy

When conducting spectrum analyzer measurements on weak spurious components it is important to remember that power levels measured on frequency components that are less than 6 dB above the noise floor of the spectrum analyzer (the line you see on the spectrum analyzer screen without any input, see marker 4 in Figure 7.), will be very inaccurate.

If any spurious components are less than 6 dB above the noise floor in your measurements you must reduce the frequency span and/or RBW to reduce the spectrum analyzer noise floor. You may need to measure segments of the harmonic spectrum individually.

In Figure 8. the RBW of the spectrum analyzer is set deliberately high so the noise floor of the spectrum analyzer is too high. The level of some of the harmonic frequency components as shown in Figure 7. are now, in Figure 8.) less than 6 dB higher than the noise floor and consequently the power level measured is different from the correct setup used in Figure 7. Setting the RBW high raises all the measured power levels slightly higher. This is because the wider RBW measures power over a wider bandwidth for each point drawn on the spectrum analyzer screen. What is important to note is that in Figure 8. the error made by the spectrum analyzer becomes bigger the closer it is to the noise floor.

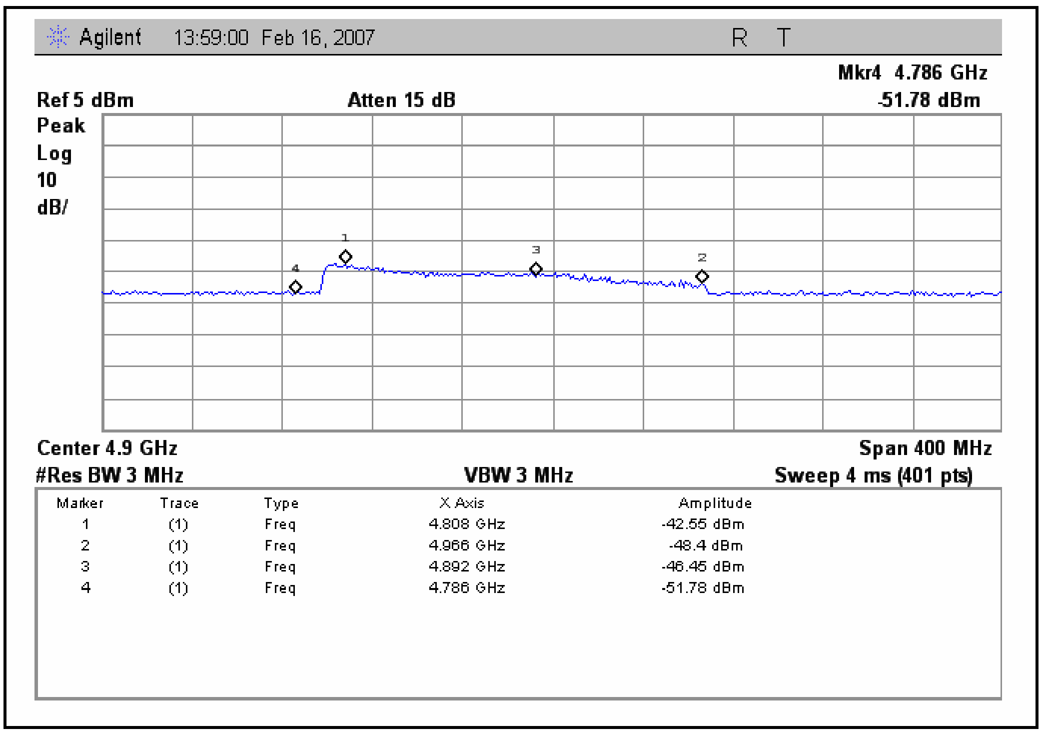


Figure 8. Harmonic output power measured with incorrect spectrum analyzer configuration

3.2 RX local oscillator leakage

RX local oscillator leakage is an unwanted RF power emission from a RF device when it is in receive mode. It is one of the critical parameters you must pass in order to meet the regulatory standards (FCC, ETSI/CE) for your end product.

All nRF receivers first multiply (mix) the incoming RF signal with a RF signal generated locally in the receiver to convert the received RF signal to a base band signal. This local RF signal is made in the Local

Oscillator (LO) and some of the output power from this oscillator leaks out to the antenna interface through the receiver front end low noise amplifiers. So, even in receive mode a RF device emits some very weak spurious frequency components.

This is a common effect in radios and the maximum level on these emitted frequencies is set in regulatory standards. The level of these spurious emissions depends mainly on the actual output from the device, but a poorly tuned antenna matching network also contributes. This means that you need to conduct the RX local oscillator test as well to verify the full performance of the antenna matching network.

The frequencies where this spurious output is found depends on the design of each nRF device family and can be found in the nRF performance test instructions document for each nRF device.

3.2.1 Test methods and set up

Use the same HW setup and routines used for the TX output power measurements, but configure the DUT to receive mode. To measure LO leakage on all possible RF channels run the TX carrier sweep test routine.

The spectrum analyzer setup should be as follows:

- **Center frequency and span:** According to the LO frequency range of each radio device.
- **Amplitude:** Since there is no powerful carrier output in this test, the amplitude only has to be higher than the max LO leakage output. Start with -30 to -40 dBm and reduce this until you have a good view of the LO spurious components on the analyzer screen. To get accurate measurements you need to have a 6 dB margin from the spectrum analyzer noise floor to the output power level of the LO leakage.
- **RBW:** You may need to set a RBW lower than the one you get on the auto setting to get the necessary 6 dB margin to the spectrum analyzer noise floor.

Figure 9. shows an example of the LO leakage measured on the nRF24L01 device. The device scans through all RX channels and the spectrum analyzer captures the LO leakage in each RF channel with a peak hold measurement. For this device the spectrum analyzer is set up with center frequency: 2.8 GHz, Span: 200MHz, Amplitude -40 dBm and RBW: 300kHz to give a good picture of the LO leakage.

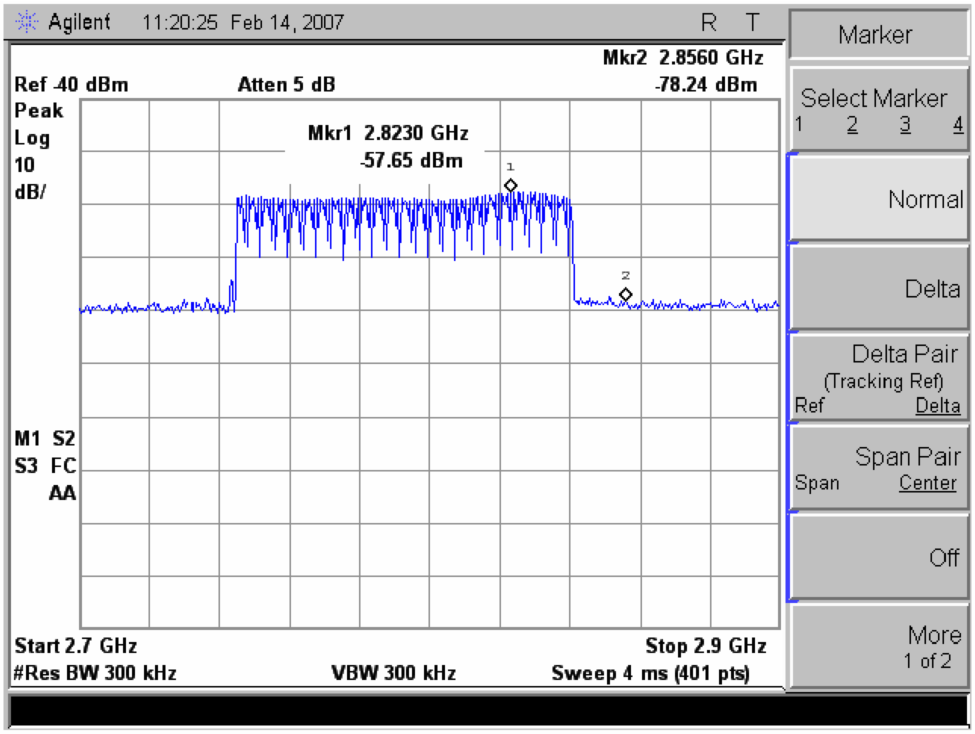


Figure 9. nRF24L01 Receiver LO leakage

4 Frequency accuracy

Through the TX carrier wave measurement you can also measure the frequency accuracy of the radio. Frequency accuracy is part of the telecommunication standards in some areas of the world (Japan and South Korea). But, the main reason for measuring frequency accuracy is to ensure a stable range performance in your application.

All radio devices have a specification on the output frequency error they can tolerate. In nRF devices the external crystal oscillator accuracy is the only contributing factor on frequency accuracy. This means frequency accuracy in nRF designs is directly linked to the crystal specification. Therefore, the frequency accuracy specification of a nRF device is given through a maximum part per million (ppm) offset requirement on the reference crystal.

> **Note:** The crystal accuracy specification in nRF product specifications is the sum of absolute tolerance at 25 degrees celsius, temperature drift and aging in the crystal. These three points are usually separated in crystal data sheets.

Since the frequency accuracy of a nRF device is decided only by crystal reference frequency accuracy, measuring the frequency accuracy actually verifies that the external crystal circuitry design is correct.

4.1 Test method and setup

This test uses the single TX carrier test setup as described in section 2.1 on page 5. Because this test is only concerned with frequency it can also be done over the air using antennas.

Once you have set up the HW and DUT, the most critical thing is to set the spectrum analyzer resolution bandwidth low enough to measure the frequency with sufficient accuracy. Sufficiently low Resolution Bandwidth (RBW) is essential to get a frequency measurement with good resolution.

Setting a RBW of 100 KHz on a spectrum analyzer means that the peak power can be found with a resolution of 100 kHz (+/-50 kHz). If we take as an example a 2.4GHz nRF device with a crystal accuracy specification of maximum +/- 30 ppm, the carrier on this device should not vary more than 2.4GHz x +/-30e-6 ppm = +/- 72 kHz. This means that if you try to measure frequency accuracy on this +/-30 ppm radio with a RBW of 100kHz the accuracy in the measurement is heavily influenced by the accuracy of the spectrum analyzer reading, not the actual accuracy of the DUT.

Generally, a measurement accuracy 10 times the actual variation expected is a good guideline, so a RBW of maximum 10 kHz (2x72kHz/10 = 14.4 kHz) should be used to measure this parameter.

When you have set the measurement accuracy correctly it is easy to verify if the TX carrier is outside the accepted range.

Frequency accuracy is OK if:

$$\frac{|F_0 - F_{DUT}|}{F_{DUT}} \leq M \bullet 10^{-6}$$

Variable	Description
F_0	Measured TX carrier frequency
F_{DUT}	The frequency the DUT is configured to send (For example: 2.403 GHz)
M[a]	Absolute frequency tolerance of crystal at room temperature in ppm (For example: 20)

a. M must be the absolute tolerance of the crystal at room temperature and not the total crystal accuracy requirement of the nRF device. This is because this measurement is conducted at room temperature and crystal frequency offset due to temperature drift should not be present.

Table 1. Frequency accuracy parameters

Figure 10. shows a crystal accuracy measurement on a nRF device with a +/-20ppm tolerance crystal programmed to 2.44GHz. The frequency offset is the difference between the center frequency (2.44GHz) and the marker frequency. The frequency offset in Figure 10. is 2.44Ghz-2.43999425 GHz = 5.75 kHz which is comfortably within the crystal accuracy (2.44GHz* +/-20e-6 = +/-48.8kHz).

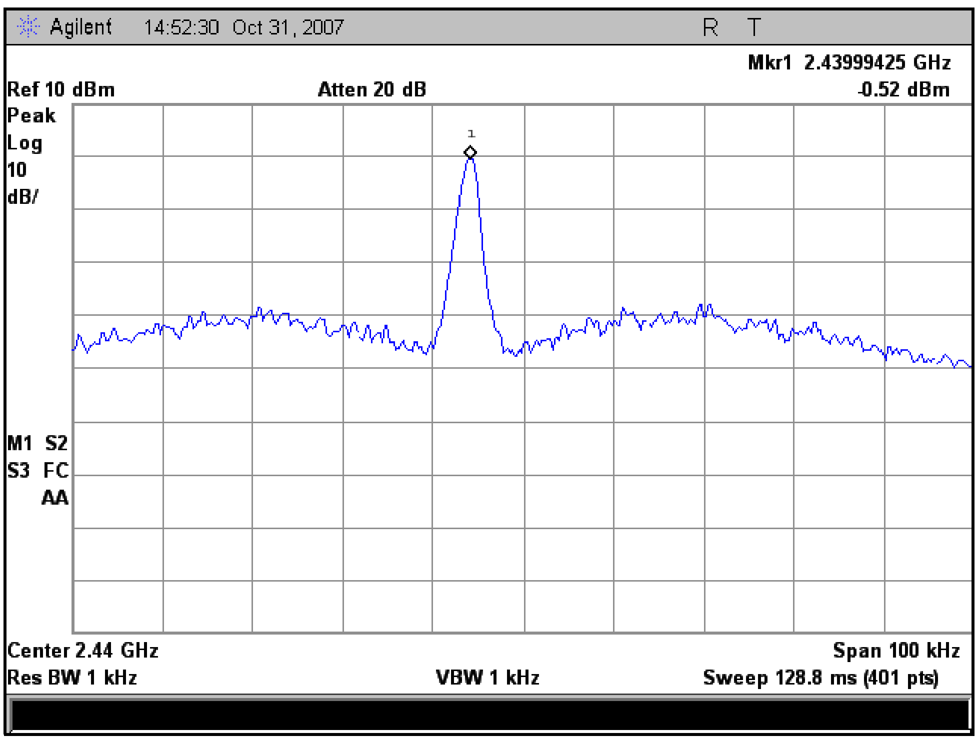


Figure 10. TX frequency accuracy measurement

If the measurement fulfills the frequency accuracy equation then your crystal oscillator circuitry is functioning as intended.

But, what is wrong if you measure a TX carrier frequency outside the window set by the equation?

- **Crystal accuracy:** The first issue to check is the accuracy specification of the crystal fitted. If this is too poor; replace it with a crystal that complies with our crystal specifications (available in the nRF product specifications).
- **Crystal load:** Wrong capacitive loading of the crystal also causes offsets in the TX frequency.

5 Modulation bandwidth

In the frequency bands where the nRF devices operate the modulation bandwidth is generally not regulated if all transmission power stays within the allocated band (2.400 – 2.4835 GHz in case of the 2.4GHz band).

There are two situations where it is important to verify the modulation bandwidth:

- Operation close to the edge of an allocated band (for example, close to 2.400 or 2.4835 GHz).
- If you add an external PA. In this situation the margin to the band edge must be increased and as the output power is increased all regulatory standards require that frequency spreading schemes like frequency hopping (FHSS) are used. The exact power level where these requirements take effect varies (FCC@ -1 dBm average ERP, ETSI/CE@10dBm ERP). The number of hopping positions that must be used in the FHSS scheme depends on the modulation bandwidth of the system. This means that the standards to follow are set by the modulation bandwidth.

5.1 Modulation bandwidth theory

To transfer data, a RF transmitter has to code the data it sends onto the pure carrier wave (explained in the previous chapters). All nRF devices use Gaussian Frequency Shift Keying (GFSK), a filtered version of basic Frequency Shift Keying (FSK). In FSK the logic level of the data sent is translated to a low and a high frequency as shown in Figure 11. Varying the output frequency is called modulation.

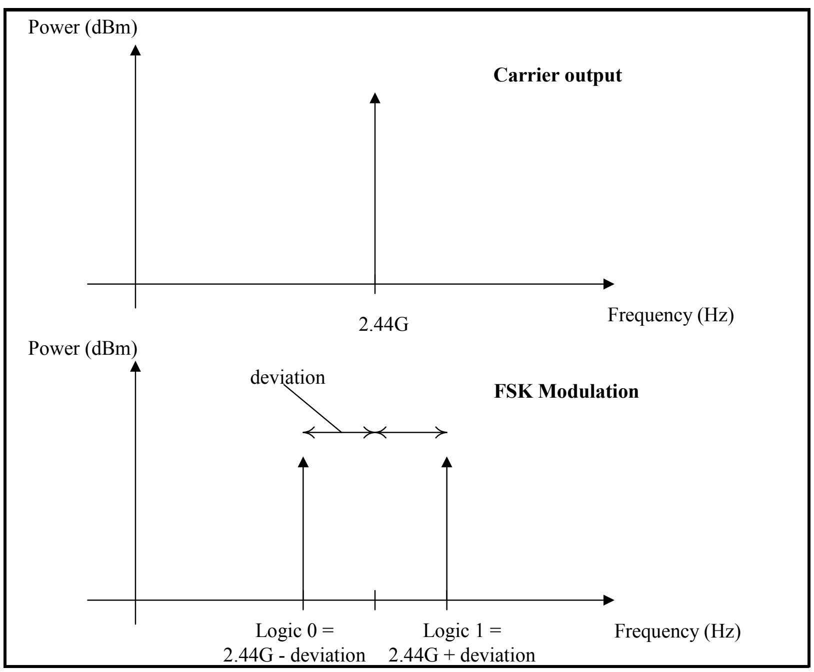


Figure 11. FSK modulation

A FSK transmitter never sends anything on the center frequency (given by the carrier frequency), only on the high and low frequency and which of these is used depends on the logic level of the incoming data. The offset from the center frequency to the frequency the FSK transmitter uses is called deviation.

The succession of changes between the high and low frequency creates a frequency spectrum called the transmit (TX) spectrum of the radio. If the modulation data changes with a periodical low rate the two FSK frequencies dominate the TX spectrum, see Figure 12. This setup makes it easy to measure the deviation of a nRF device (indicated by the markers in Figure 12.) and as shown, the bandwidth increases but the peak power level decreases compared to the pure carrier wave spectrum. This is because the carrier power is now spread over a wider frequency area called the modulation bandwidth.

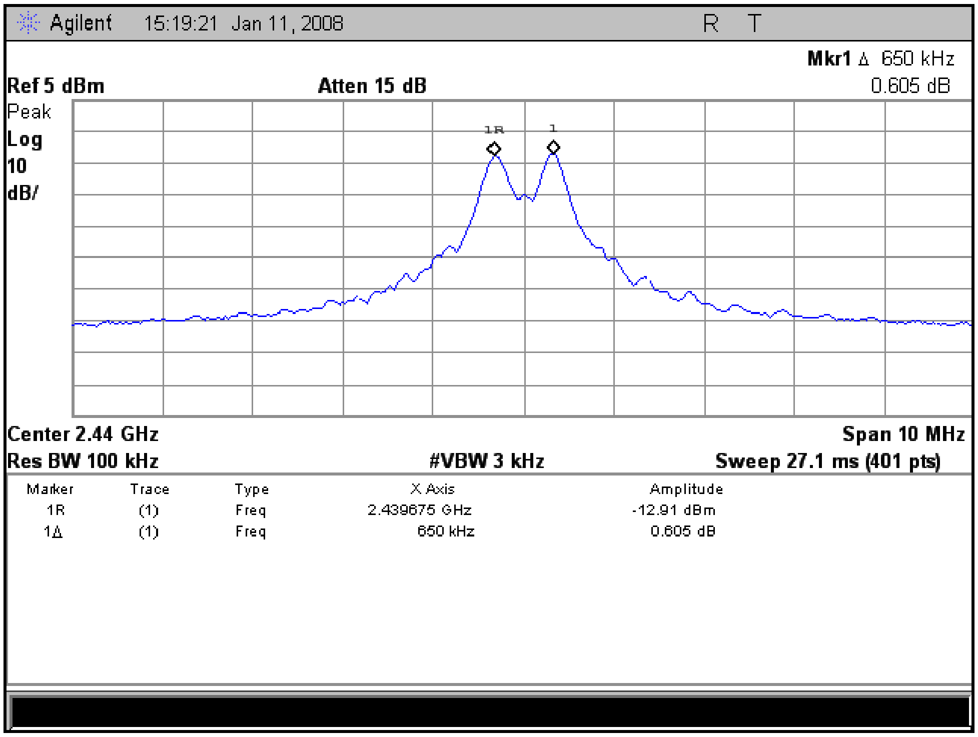


Figure 12. Spectrum showing nRF device deviation

The modulation bandwidth depends on the modulation type (for example, FSK or GFSK), deviation, data rate and the actual data sent. With a nRF device, modulation type and deviation are set by the chip design, so the spectrum in Figure 12. is of limited interest. What is interesting is the maximum modulation bandwidth of the radio in normal operation. To get a representative picture of the maximum modulation bandwidth on a specific data rate a Pseudo Random Bit Sequence (PRBS) should be used as actual payload data and the device should be set to send data packets as in a normal application. The PRBS sequence ensures an equal number of logic 0 and 1 in a random sequence is sent, mimicking the normal behavior of a transmitter.

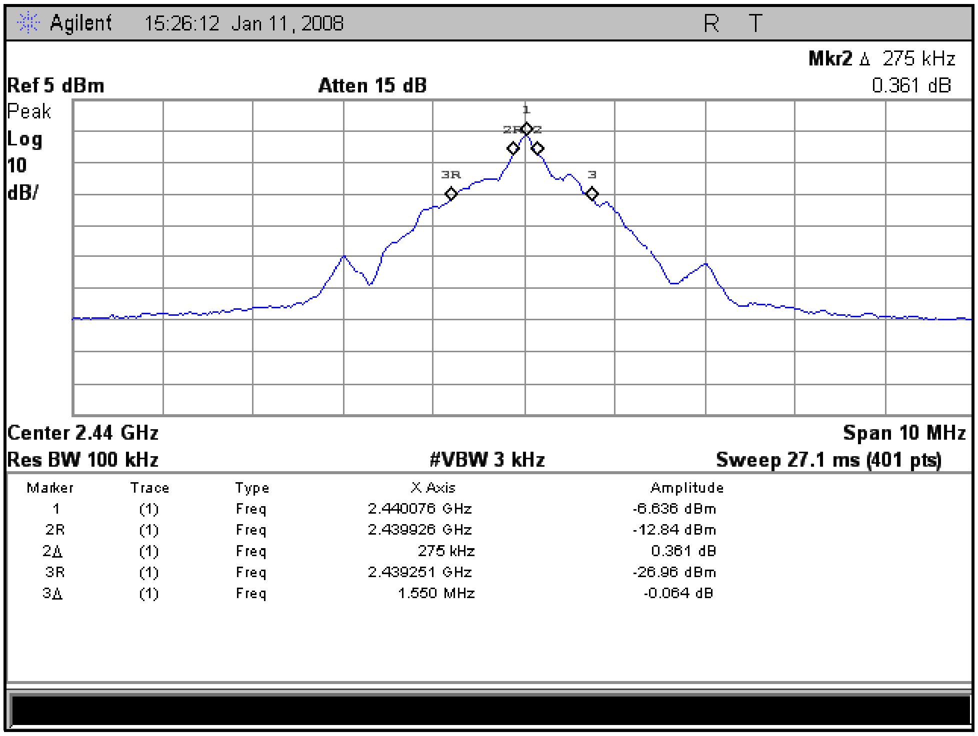


Figure 13. Modulated carrier bandwidth measurements

An example of the modulation bandwidth of a nRF device used in 2 Mbit/s mode is shown in Figure 13. Both 6 dB and 20 dB bandwidths (BW) are commonly used to specify the width of this TX spectrum. The 6 dB bandwidth is, as indicated in Figure 13., the frequency offset between a point (marker 2 and 2R) on each side of the spectrum that is 6 dB lower than the center of the spectrum indicated by the peak output power (marker 1). The 20 dB bandwidth is similarly the offset (marker 3 and 3R) between two points on each side of the spectrum, 20 dB below the peak output power.

A TX bandwidth specification does not mean that there is no power emitted outside this bandwidth. As can be seen in Figure 13. the power in the TX spectrum rolls off outside the 20 dB bandwidth. This is a fact in all radio devices and the rate of the drop is mainly decided by the modulation type used in the RF device.

The 20 dB bandwidth of a transmitter is often denoted as the channel bandwidth of the system because this is the bandwidth the system needs for operating a channel of communication. The power emitted outside this bandwidth is emitted in channels next to the one in use and is called a neighbor channel emission, one of the parameters often specified in a device product specification. The neighbor channel emission shows you how much energy leaks out and can affect a second system operating on one of the neighboring channels.

The modulation bandwidth also impacts how close to the allowed band edges you can operate. All emission outside the frequency band you operate in (for example, 2.4GHz) is seen as spurious (unwanted) emission and must conform to the regulatory standards. Figure 14. shows an example of a radio device set for operation at the lower band edge of the 2.4GHz band (2.400 GHz). This leads to a large amount of too powerful emissions below 2.400 GHz indicated by the shaded area in Figure 14. At 2.400 GHz the out of band spurious limit of the regulatory standards comes into effect. For instance, the maximum spurious emission limit in the regulatory standard for Europe (ETSI) is - 30 dBm. This means that all emissions in the shaded area (that is, everything between the markers) in Figure 14. are illegal.

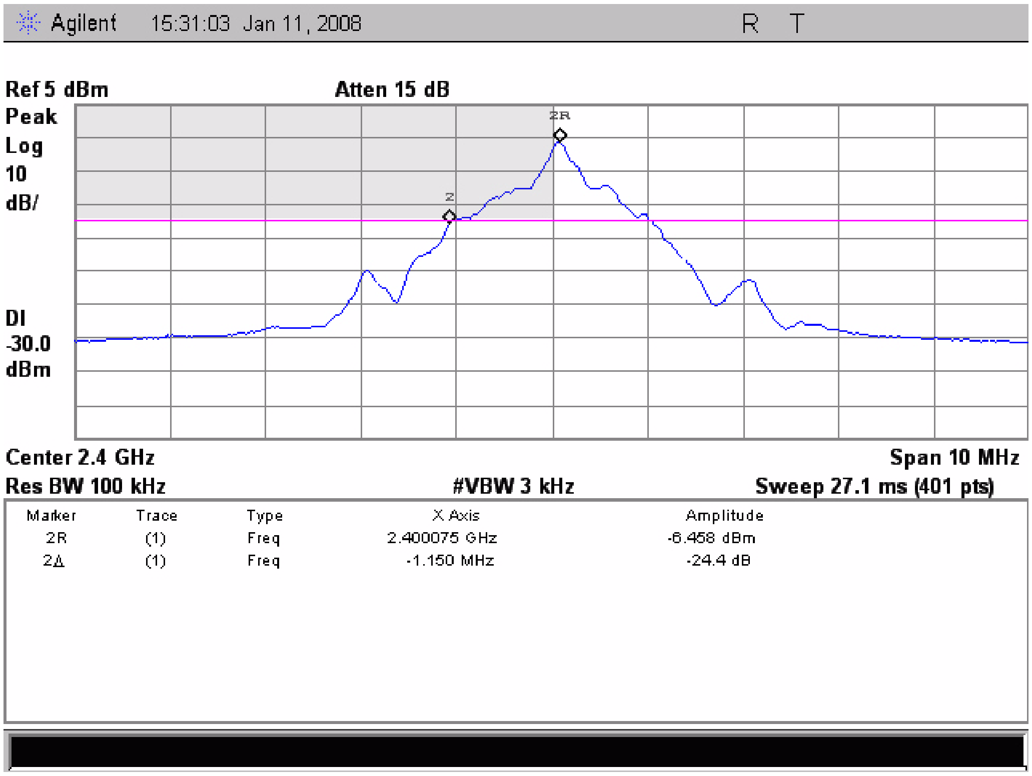


Figure 14. Emission limits at the band edge

This effect is called band edge spurious emission and is one of the main test criteria for regulatory RF emission standards approval. To comply with regulatory standards the lowest frequency this radio device can be set to is 2.402 GHz as shown in Figure 15. It is visible from this figure that the output power at 2.400 GHz (marker 2) and below now has an ample margin to the limit at -30 dBm (marked by the red horizontal line).

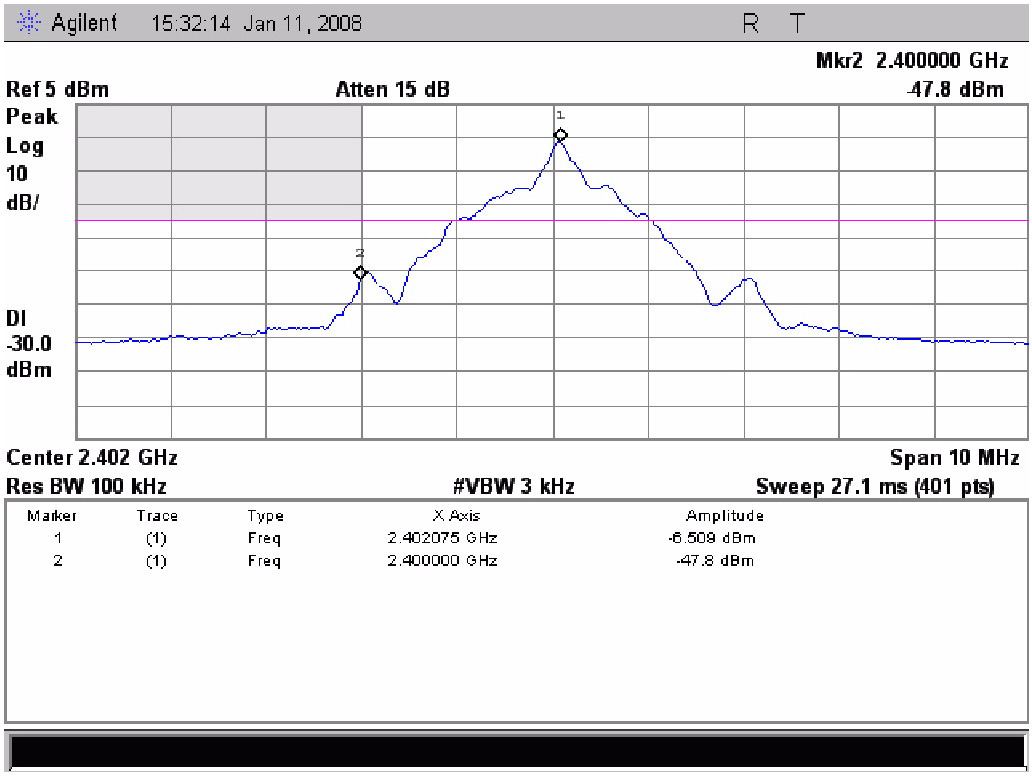


Figure 15. Emission complying with band edge spurious requirement

5.2 Test method and setup

The HW setup for this test is identical to the TX carrier wave setup.

You must set the Device Under Test (DUT) to send data, most nRF devices are optimized to send packets of data in short bursts. The payload of each packet should contain a PRBS sequence. How to set up a specific nRF device for this test can be found in the nRF performance test instructions for each nRF device.

The Spectrum analyzer setup is the same as for the TX carrier test:

- **Center frequency:** Equal to the RF frequency configured in the DUT.
- **Frequency span:** 1 - 10 MHz depending on the DUT.
- **Amplitude:** Must be set higher than the maximum output power capable of the DUT.
- **Resolution Bandwidth (RBW):** 30kHz- 300kHz depending on the frequency span used. In most cases simply setting the spectrum analyzer RBW to auto so that it follows the span is the best option.

Since all nRF devices are sending packets of information, the spectrum analyzer must be set to peak hold measurement and measure over several packets to capture the full TX spectrum. How long this takes depends on the processing speed of your spectrum analyzer.

The modulation bandwidth is found by adding the frequency markers as described in Figure 13. Some spectrum analyzers can do this measurement automatically, to check if your spectrum analyzer can do this refer to its manual.

All nRF devices send packets in this test, this means the spectrum may not be as symmetrical as you expect. Since the measurement is done using peak hold, the measured spectrum also holds some frequency components originating from the frequency settling of the transmitter prior to the actual data modulation. This measurement gives the true picture of the TX spectrum of a nRF device in normal operation.

6 Receiver Sensitivity

The receiver sensitivity level is one of the vital receiver parameters that decides the maximum range you can achieve on a RF link. Measuring sensitivity in a receiver unit is very important since it not only relies on the nRF device, but also on the quality of the PCB design supporting the nRF device.

This section presents the terms related to receiver sensitivity and describes how radio devices are tested to find the receiver sensitivity numbers presented in the electrical specification tables in each device product specification.

6.1 Receiver sensitivity theory

Receiver sensitivity is the minimum input power level from the antenna that a radio receiver can handle without making a critical number of errors when decoding (demodulating) incoming data.

As the power received from your transmitter decreases (Figure 16.) with increasing physical distance between TX and RX, so does the margin between the incoming signal and the receiver's noise floor. The receiver noise floor is decided by the inherent noise found in all electrical circuitry. At a certain level this margin becomes too small for the receiver's demodulator to interpret the received data accurately leading to a sharp increase in the number of errors the receiver makes. The minimum margin between the receiver noise floor and the input power level where it can process an incoming signal properly is called the demodulator threshold. The noise floor level and demodulator threshold equals the sensitivity level of the receiver as shown in Figure 16..

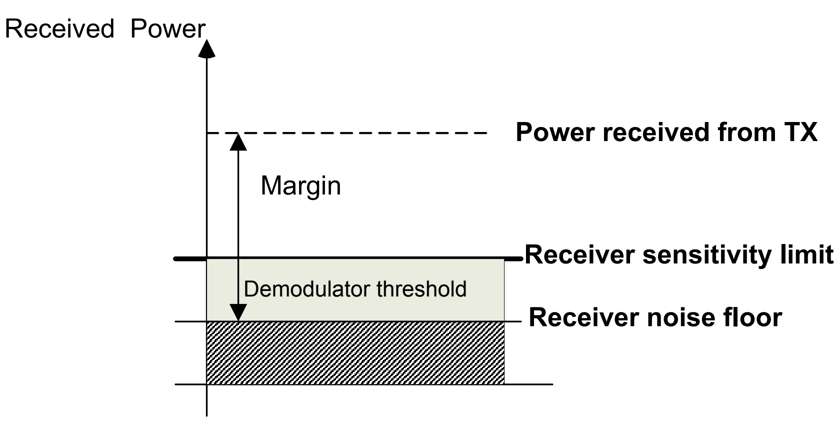


Figure 16. Receiver input power budget

The demodulator threshold depends on what modulation type is used, the noise floor is set by the bandwidth of the receiver and design trade offs (for example, current consumption in the receiver front end).

To design a radio with the best possible sensitivity you must minimize the bandwidth of the radio (less background noise received) and increase the current consumption in the receiver front end (less self generated noise). But, narrow receiver bandwidth limits the data rate you can achieve over the air and, for example, a too high current consumption in the receiver may render the radio unsuitable for battery applications.

Different radios have different sensitivity levels based on the trade offs made to make the radio the best overall solution for a set of end applications.

6.1.1 Bit Error Rate

Before measuring the RX sensitivity level, you must define poor receiver performance. As the received power drops towards and below the sensitivity limit, the average number of errors made by the receiver increases sharply. You must decide how many errors on average are acceptable to deem the receiver performance good. How many errors a receiver, on average, makes is given by the Bit Error Rate calculated by the following equation:

$$BER = \frac{Number\ of\ errors\ made}{total\ number\ of\ bits\ received}$$

A BER limit often used on low cost radios to define the sensitivity limit is 10^{-3} or 0.1%. This means that if the input power is set at the sensitivity limit (for example -82 dBm) the receiver makes an average of 1 error for every 1000 bit it receives.

> **Note:** This means that the radio does make a steady stream of errors when operating on the sensitivity limit.

Specifying sensitivity is always inter-connected with what bit error rate you are willing to accept and the data rate you operate on. Specifying a sensitivity limit without specifying the BER does not make sense. So, comparing the sensitivity limit of two receivers without making sure the BER level and data rate are set to the same, will lead to the wrong conclusions.

6.1.2 Sensitivity accuracy

To measure RX sensitivity with a good amount of accuracy you must first calculate the BER with a good enough accuracy. The noise that causes the receiver to fail at the sensitivity limit is stochastic (random), because of this the number of errors done by the receiver will also have a stochastic variation over time. You can see this effect in real life by the fact that when you measure the number of errors the receiver does over a given number of received bits, the number of errors will be slighlty different for each test run. This of course also results in a slightly different BER for each test run. The only way to calculate a 100% accurate BER is to do an infinite number of test runs. This is, of course, not feasible, so when doing sensitivity measurements you must use statistical methods to set up a test that will give you an accurate BER.

Statistical measurements are aimed at measuring in a way that gives you a good confidence level. A confidence level of 99% means that you can be 99% sure that the number of errors found in one sample volume (one test run) are differing from the ‘true’ average number of errors (the mean found from an infinite numer of test runs) with less than 3 times the variance (3*sigma). The +/-3*sigma interval is called the confidence interval.

The variance is a variable telling you how much the variable (number of errors) can differ from test run to test run. The variance takes on different characteristics depending on what kind of statistical distribution the variable you measure has. Since each error made by the receiver has the same probability, the statistical distribution of the number of errors in one sample volume will follow a Gaussian distribution. This is due to the central limit theorem. Due to these facts the variance in these tests will simply be the square root of the mean value of errors in a given sample volume.

In Table 2 on page 26 you can see some examples of these relations for different sample volumes:

Total # of received bits (Sample volume)	Expected # of errors in sample volume (Mean)	99% Confidence interval (3*sigma)	Possible variation in # of erros on consecutive tests
10,000	10	+/- 9.5	+/- 95%
100,000	100	+/- 30	+/- 30%
1,000,000	1000	+/- 95	+/- 9.5%

Table 2. Accuracy of BER measurement on different sample volumes

Here you can see that if you plan to run a BER test with a total number of received bits of 10.000, the expected number of errros should be:

$$10000 \times 1e-3 = 10$$

But due to the variance of the normal distribution:

$$SQRT(10) = 3.16$$

The only thing you can say with 99% confidence is that actual measured number of errors in each test run will be somewhere between:

$$\pm 3 * sigma = \pm 3 * 3.16 = \pm 9.5 \ldots 0.5 - -19.5 errors$$

This means that you can get a variation in the number of errors on consecutive test runs of +/- 95%. Clearly establishing the BER from a sample volume of 10,000 received bits does not give you a very accurate result. But as you can see increasing the total number of received bits in each test run will give you ever better accuracy in the numbers you base the BER calculation on.

The next step is to decide how accurate the BER actually needs to be to give you a good mesurement of the RX sensitivity. In Figure 16. you can see a typical BER versus received RF power relation for a nRF device measured over a very large number of received bits. This is a very time consuming test but has good BER accuracy as a result.

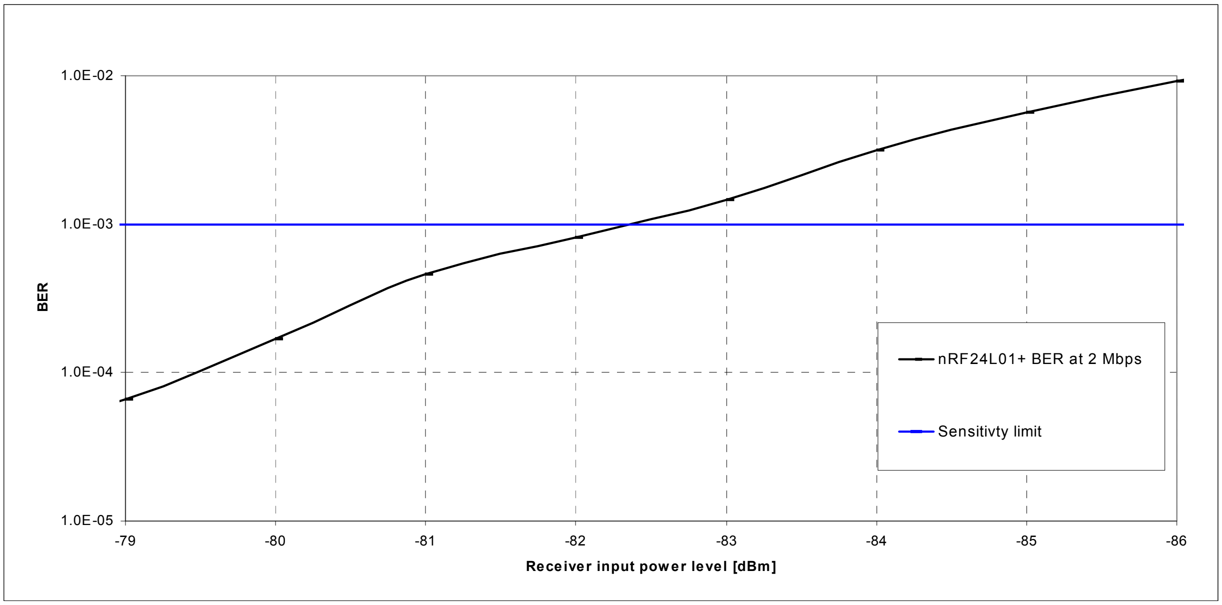


Figure 17. nRF24L01+ BER vs. RF input power

From this figure you can see that measuring the sensitivty limit of a nRF device (Example: -82 dBm for nRF24L01+@ 2 Mbps) with an accuracy of +/- 0.5 dB you must be able to accurately measure and calculate a BER between 8e-4 and 1.4e-3 or 1e-3 -20%/+40%, which we can approximate to 1e-3 +/-30%.

Referring back to Table 2 on page 26, this gives us the conclusion that to measure the sensitivity limit of a nRF device with an accuracy of +/-0.5 dB, you must calculate the BER from the number of errors the receiver does in no less than 100.000 received bits.

6.1.3 What do sensitivity measurements tell you?

Sensitivity on all nRF devices is tested in our production. All our shipped devices comply with specified sensitivity levels in the nRF product specifications. But, in the final application, sensitivity is not decided only by the nRF device.

The deciding factor for actual sensitivity in your application is the incoming signal strength from the transmitter compared to the total noise level in the receiver. Any losses in the receiver signal path on the PCB (from antenna to the radio device antenna pins) or noise introduced from other PCB circuitry into the radio device degrades the sensitivity.

This means that measuring sensitivity on a nRF evaluation or development board enables you to verify our sensitivity limit specification, while measuring it on your own prototypes reveals the quality of your own PCB design. If your design or layout increases the losses in the RF path or level of noise in the RF circuitry, it will be seen as a degradation of the sensitivity limit.

Note: Optimal receiver performance can only be achieved if the PCB design is also optimal.

The reference layouts we provide address both these problems in the RF circuitry. The recommended antenna matches ensure minimal input loss when properly tuned. For details on antenna match tuning, please refer to Tuning the nRF24xx matching network white paper found on our web site. However, the total noise level on any given PCB is something that must be managed in each design through good circuit design and layout (especially power supply and GND). Measuring sensitivity level on your prototype shows you if the noise level in your design is low enough. Use of GND planes and star routing of the power supply is essential for making the radio and other parts of the design perform at their best.

6.2 Test Method and Configuration

To measure sensitivity you must:

1. Apply a modulated RF signal that the receiver can demodulate.
2. Vary the power of this source in a controlled way.
3. Count the number of errors the receiver makes.

When conducting sensitivity measurements the receiver is tested on its performance limit. Low signal levels make this test the most demanding of all RF prototype tests. It is very important to control the test environment closely as any additional noise directly affects the measurement. Following the guidelines in section 6.2.1 is essential. Also, putting the DUT into a shield box prevents neighboring RF systems affecting the stability of your measurements.

6.2.1 Test Setup

Always use cables when measuring sensitivity, using cables to measure is called conducted measurements. Any attempt to measure over the air, using antennas, is impossible because you no longer have exact knowledge of the power received by the DUT.

The test setup for a sensitivity measurement is shown in Figure 18.

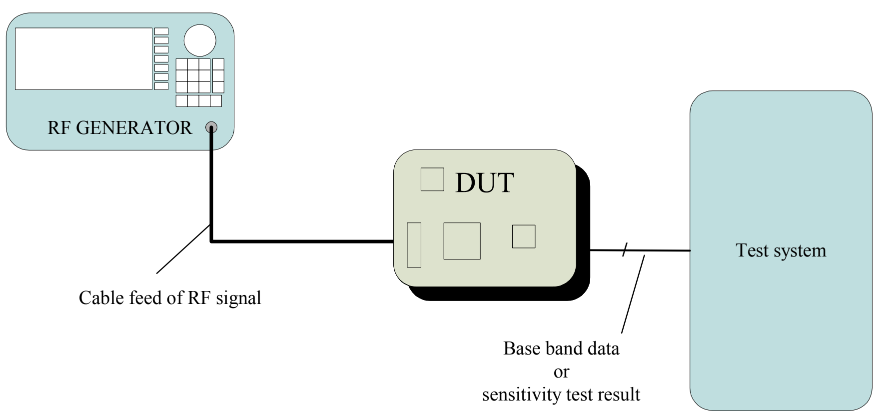


Figure 18. Receive sensitivity level test setup

To get a controlled RF input, use a RF signal generator to generate the input for the DUT. The required test routine is shown in Figure 19.

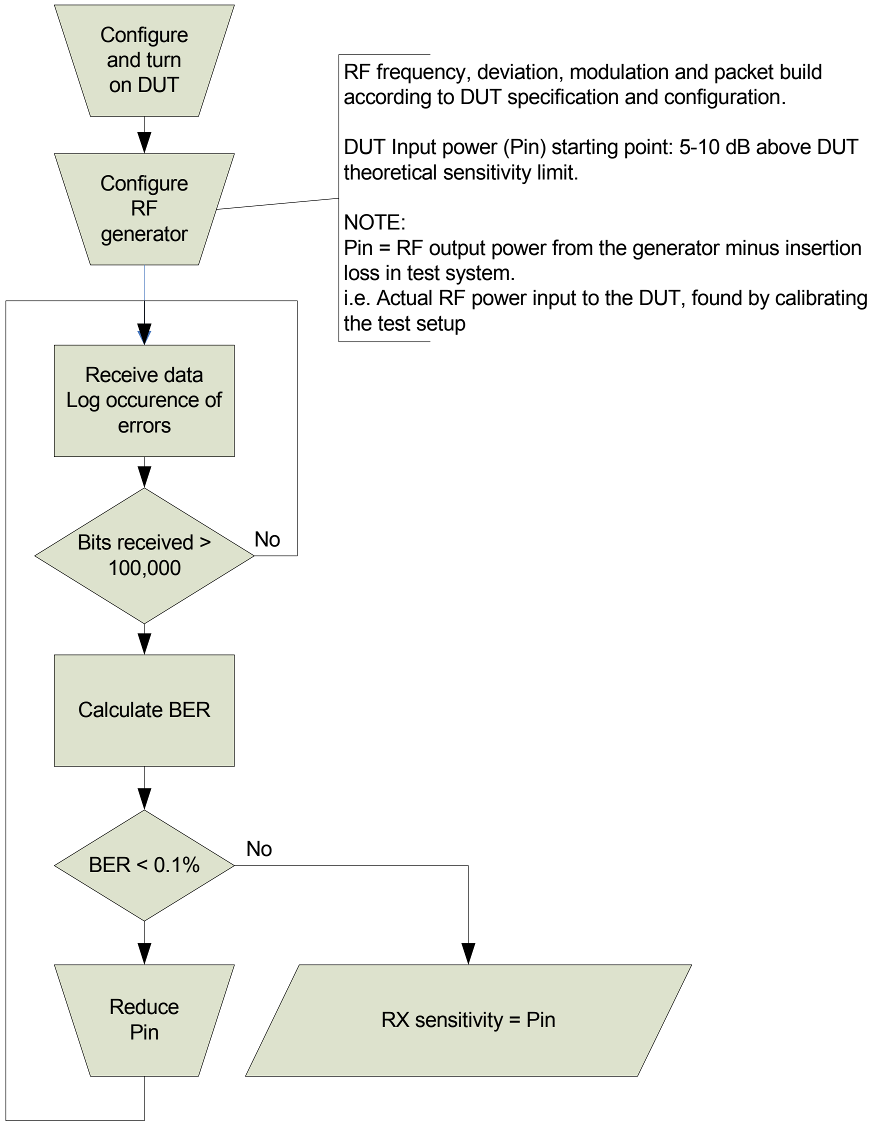


Figure 19. Receiver sensitivity test flowchart

RF generator frequency, modulation, deviation, data rate and possible data format (packet build with preamble, address and so on) are all decided by the DUT product specification. See details of the RF generator setup for specific nRF devices in the nRF performance test instructions for each nRF device.

On the base band side of the receiver you must have a system that can decide if the DUT has made an error. To do this the test system must compare the data sent from the RF source with the data output from the DUT. This can be done by feeding the modulation data from the RF generator to the test system which compares it to the DUT output while receiving. Alternatively, you can set a predefined data pattern (RF packets) in the RF generator to which the test system can compare the received data.

If you test by feeding the modulation data from the RF generator to the test system you need a RF generator that sends out the baseband data to the test system at the same time it sends data to the DUT. The test system must also compensate for the different delays in the path directly from the RF generator and through the DUT. This is often done through auto correlation of two long sequences of data after the test is run. This requires a rather big and complex test system.

Instead, if you test using a predefined data pattern (data package), you run the same packet over and over again. The base band test system can be a simple microcontroller, preferably the one fitted to control the radio in your end application. This significantly reduces complexity. The payload you use should be a PRBS sequence. The results using this method are adequate for identifying noise problems in your design. How to make this sensitivity measurement test firmware on a simple MCU is described for each nRF device in the nRF performance test instructions.

> **Note:** Using off the shelf BER testers is not always straightforward because they generally are made to test devices running on open standard protocol, such as Bluetooth, and key RF parameters usually differ from nRF proprietary ones.

6.2.2 Calibration of measurements

RF generator output power is the varying input parameter in the test. Remember that losses in the coax cable and connectors must be subtracted from the power level indicated on the RF generator to get the true power level applied to the DUT.

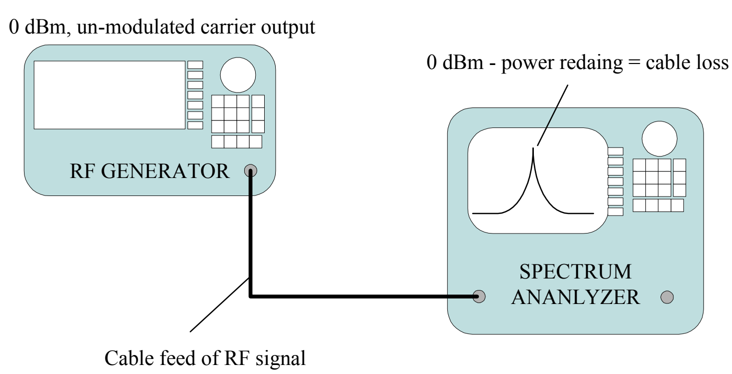


Figure 20. Calibrating DUT power input

The insertion loss from the RF generator to the input of the DUT can be measured by sending a carrier of, for instance, 0 dBm from the RF generator and measuring the actual power level at the DUT input with a spectrum analyzer. 0 dBm is far from the level you are going to use in the sensitivity test, but these levels (-80 to -100 dBm) are very difficult to measure on the spectrum analyzer, so a higher level is usually needed. To add accuracy to the calibration you can also measure the RF generator level on lower output power levels to verify that the linearity of the RF generator output is good, that is, that setting the RF generator to 0 dBm equals 0 dBm output, and setting it to -50 dBm equals -50 dBm output, and so on. For details on spectrum analyzer setup refer to Section 2.1 on page 5.

6.2.3 Low cost set up

If you do not have a RF generator you can get an approximate level of the sensitivity limit through the set up shown in Figure 21.

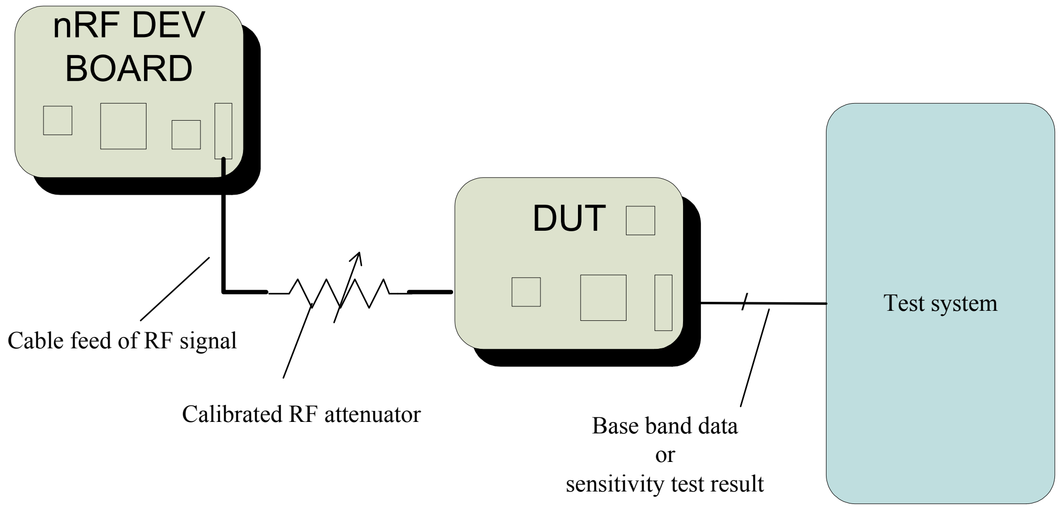


Figure 21. Low cost set up with nRF development board

In this set up you use a Nordic Semiconductor development board as a signal source and a RF attenuator to control your input power. The input power of the DUT must be verified using a spectrum analyzer. The rest of the test routine is as described in Figure 19.

> **Note:** The measurement of the input power at the low levels will not be as accurate as the test setup using a calibrated RF generator.

7 Receiver selectivity

Receiver selectivity quantifies how a RF receiver performs when subjected to unwanted radiation (interference) from other RF systems operating either on the same frequency (co-channel) or frequencies close (adjacent channel) to your operating channel.

The selectivity in nRF devices is fully controlled by the chip design which means this test can only be used to verify the selectivity numbers given in different nRF device product specifications. Performing the following complex test will not give you any information that can improve your PCB design. Receiver selectivity is included to show how the selectivity parameters found in nRF product specifications are measured.

> **Note:** Receiver selectivity is generally not part of governmental regulations and consequently not critical for getting a nRF design approved for sale.

7.1 Theory

Receiver selectivity specifies how much input power from an interfering/unwanted system the receiver can handle while still receiving a signal from its own wanted transmitter. The receiver's ability to reject input signals from other systems is also commonly called co-channel rejection and adjacent channel rejection. The ability of the receiver to reject power from unwanted transmitters operating on frequencies more than 2-3 channels away is called blocking or wideband blocking.

The RF channel is defined as the bandwidth (BW) that the radio under test requires for communication. The first adjacent (neighboring) channel is one bandwidth offset from the frequency used by the radio under test, the second is 2*BW offset and so on. This is illustrated in Figure 22.

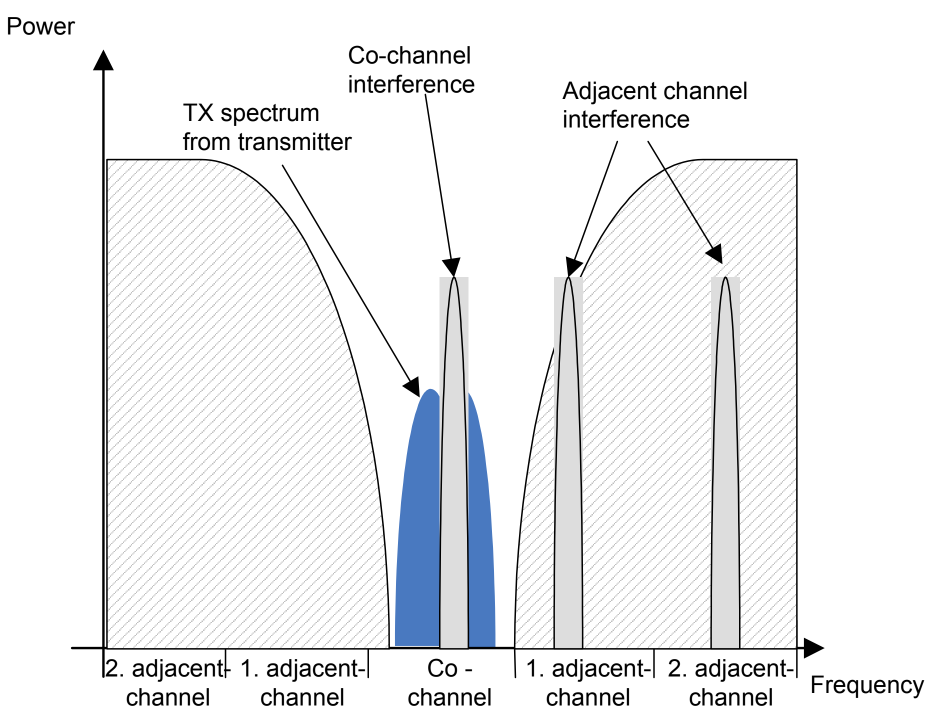


Figure 22. Co and adjacent channel interference

Receiver selectivity is indicated by the shaded area in Figure 22.

If the input power received from an interfering system exceeds the selectivity of the receiver then the interference causes the receiver to make errors and you have an increase in the Bit Error Rate (BER). Figure 22. shows how a received interference signal with a certain power level may affect your receiver if it is the co-channel or 1st adjacent channel interference, while it is no problem if it is found in the 2nd adjacent channel or further away.

7.2 Test Method and Configuration

7.2.1 Test Setup

To measure RX selectivity you must apply the wanted RX input signal and an interfering signal at the same time.

When measuring receiver selectivity, the test environment must be closely controlled. This is only achieved by using cables between the transmitter, noise source and receiver (conducted measurement) instead of antennas. Using antennas is not reliable because of multipath fading effects, constructive and destructive interference and unpredictable signal levels. This is why conducted measurements are the only way to conduct precise receiver parameter tests.

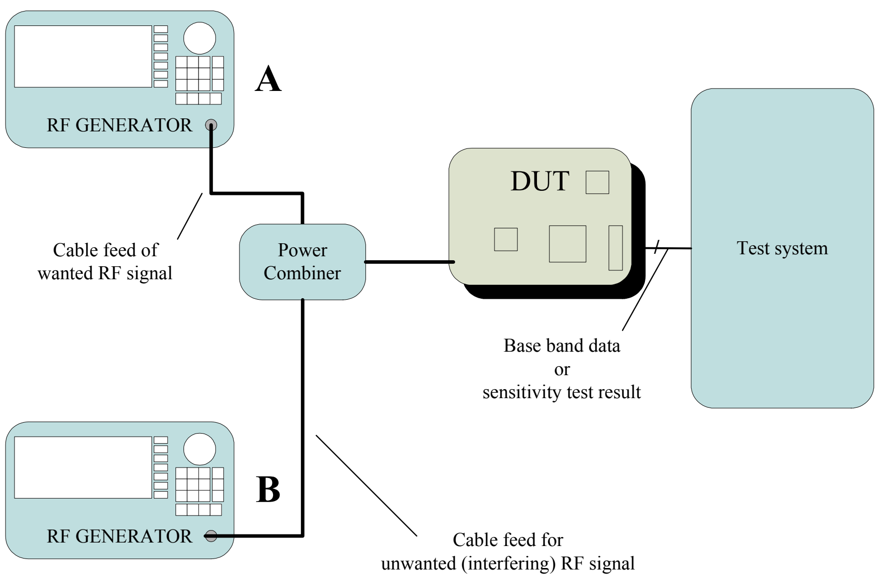


Figure 23. RX selectivity test setup

Figure 23. shows the required hardware test setup. Signal generator A generates the wanted input signal for the Device Under Test (DUT). Signal generator B acts as the interference source. The power from the two generators is then combined and applied to the DUT.

See nRF performance test instructions for each nRF device for details on instrument set up.

RX selectivity test is based on the same core routine as the RX sensitivity test but, with the interference from generator B being the variable input. The test routine for finding RX selectivity is shown in Figure 24..

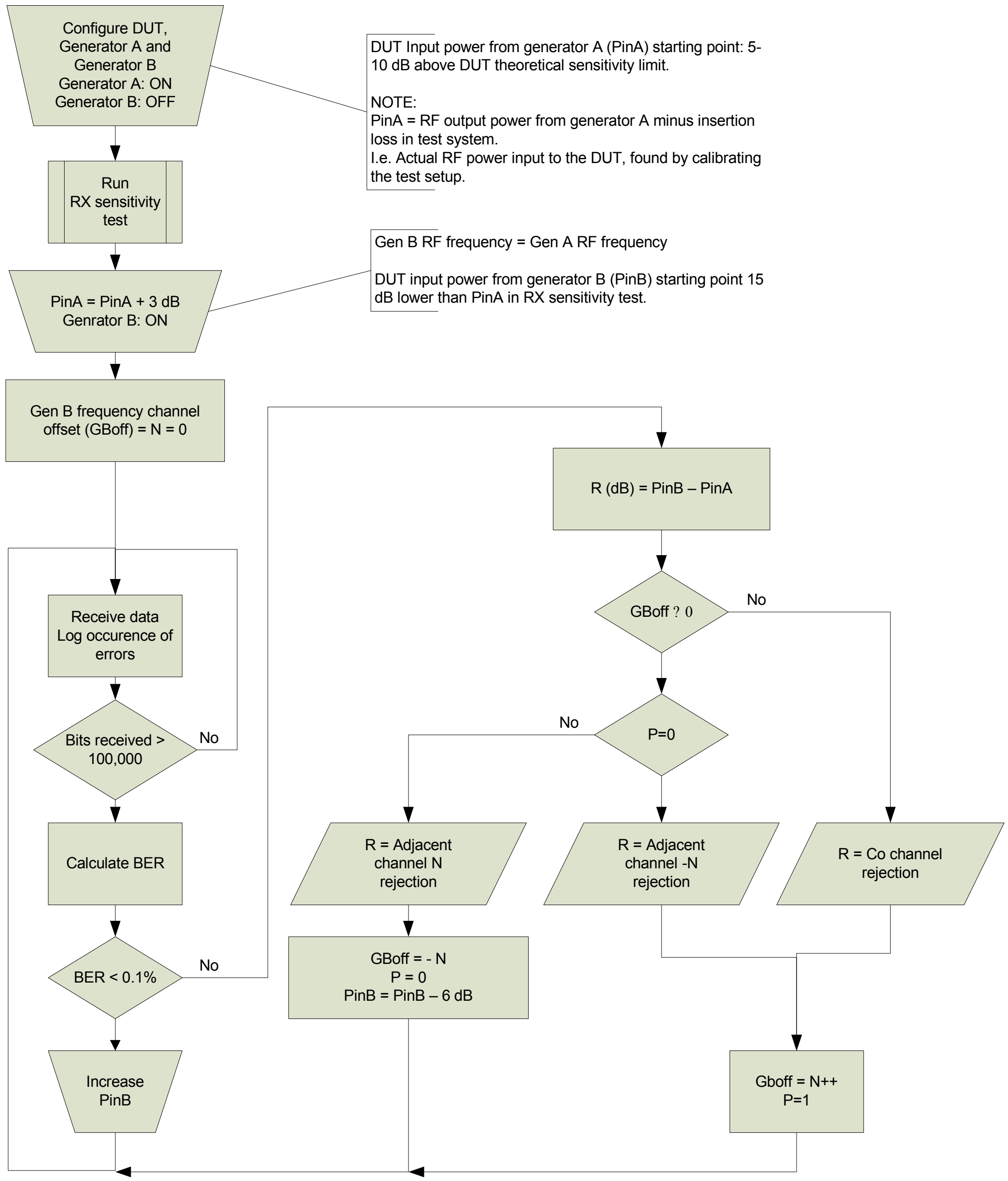


Figure 24. RX selectivity test flowchart

In the RX selectivity test the sensitivity of the DUT is first established and a 'wanted' input signal is applied slightly (3 dB) above this. Then the RX performance of the DUT is measured with generator B applying noise on co- and adjacent channels. How many channels (how big N can be) depends on how much information you want and how much time you have to do the test. Usually in product specifications numbers for up to 3 adjacent channels are used giving N < +/-3. The wanted signal generator is set to send the same information as in the RX sensitivity test, but at a power level 3 dB above the sensitivity limit found in the RX sensitivity test. As long as RF generator B is off, the DUT should make no errors receiving the signal from generator A.

The RX selectivity data can be presented in two ways:

- Receiver selectivity: PinA – PinB.
- Receiver noise rejection: PinB – PinA.

The only different results are the polarity of the numbers presented in device product specifications.

7.2.2 Calibration of measurements

To calibrate this test setup use the same routine that is used for the sensitivity measurements (see chapter 6.). With the power combiner, the insertion loss between the RF generators and the DUT increases. The insertion losses from each RF generator must be measured separately, but with both generators connected. Figure 25. shows the setup used to find the insertion loss from generator A. The difference between the RF output power set on generator A and what you read on the spectrum analyzer informs you of how much the wanted signal level (that is, RF generator A output power) must be increased in order to apply a power of PinA on the DUT. Repeat the measurement with generator A OFF and B on, this reading informs you of how much you must add to the RF power setting of generator B to get a correct level of PinB.

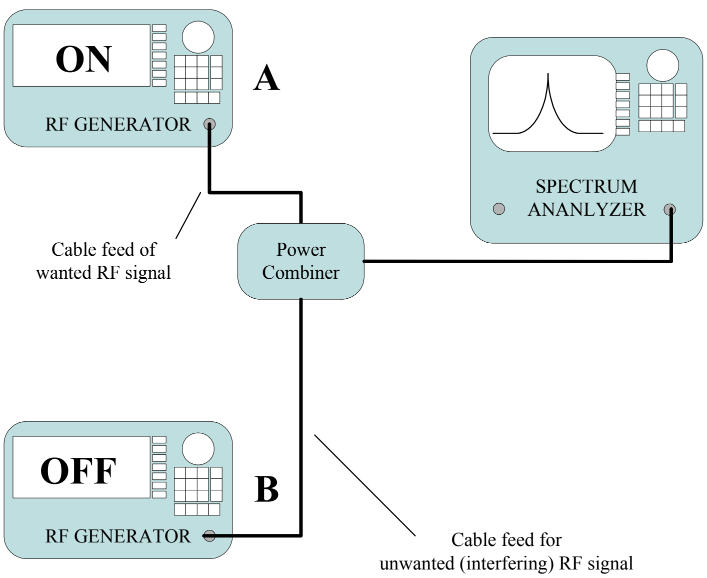


Figure 25. Insertion loss test set up

7.3 Types of interference

Radio devices are exposed to several kinds of interference sources. To compare with other receivers, the nRF devices product specifications have two types of interference listed.

As defined in European regulation ETSI EN 300 440-1 V1.3.1 (2001-09):

- Interfering source is an unmodulated carrier in the center of each channel (Chapter 8.2)
  - This gives a selectivity number purely for comparison with different receivers.

As defined in Bluetooth Core Ver. 2.0:

- Interfering source is a modulated signal similar to the one used by the DUT itself (Ch.4.1.2)
  - This gives a more realistic picture of how several units of the same radio manage.

8 Current consumption

Current consumption is one of the main design parameters in any electronic design. It is important, especially in battery operated applications, to keep current consumption tightly controlled. Current consumption calculations form the basis for your choice of power supply circuitry, battery technology and estimating battery lifetime.

This section presents the method and test setup for accurate measurements of the current consumption (static, average and dynamic) in nRF devices. It also describes how you can use these measurements to find average current consumption in the radio and how you can use it to estimate battery lifetime.

8.1 Static and average current consumption

The simplest approach to measuring current consumption is to use a standard ampere meter. Using an ampere meter or multimeter with the nRF device, as shown in Figure 26., makes measuring the current consumption of a nRF device while it is in a steady state straightforward.

The following points are important for measuring static current consumption accurately:

- Ensure that your measurement setup isolates the current drawn by the nRF device, that is, put it on a VDD branch supplying only the nRF device.
- Ensure that the nRF device is in a steady state. All nRF devices contain HW link layer protocol engines (ShockBurst™ engines), which means that they return to standby as soon as they have finished a task. The most convenient time to measure the current consumption stated in nRF product specifications is during the tests described in this document.
    - TX current consumption: while running any of the TX tests
    - RX current consumption: during the sensitivity test (or normal operation)

In power saving modes the current consumption can be measured at any time but be aware of the accuracy requirements on these tests. High accuracy ampere meters or multi-meters are required to measure in these modes. The current consumption of nRF devices is usually << 100 uA in standby and < 1 uA in power down.

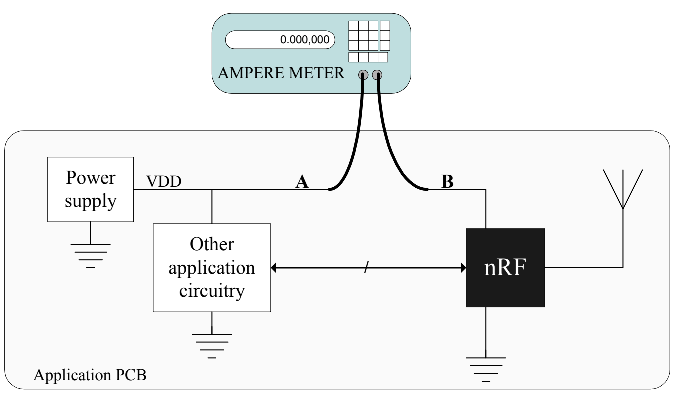


Figure 26. Power down and standby current consumption HW setup

If your ampere meter can also take RMS readings, you can monitor the average current consumption drawn by the nRF device in normal operation, that is, running communication and changing between TX, RX and power saving modes. All ampere meters rely on a shunt resistor and the current is converted to a voltage drop that can be measured. This voltage drop decreases the voltage supplied to the radio device. A typical high accuracy, low current ampere meter has a 5 ohm shunt resistor. You must check the specification for the ampere meter used and make sure this shunt resistor does not cause a too high voltage drop.

If the application has a very low duty cycle (spends little time in active mode), measuring average current consumption with an ampere meter gives you little or no information, especially if you find that your average current consumption is not as expected. There is another way to measure the dynamic current consumption in a nRF device which is described in the following sections.

8.2 Dynamic current consumption

Measuring dynamic current consumption gives you accurate numbers on the current consumption in a nRF device as it switches between different modes. Based on these measurements you can calculate the average current consumption based on measured values. Dynamic current consumption measurements can also be a good tool if you find that unexpected currents are drawn. Even if you have a logical error in your design (it does not do what was expected) a dynamic current consumption test shows clearly which modes the nRF device enters and when it enters them. There are two main challenges when measuring the dynamic current consumption of radio devices:

- Transient current consumption measurements are needed to accurately capture peak current levels when radio devices are in active (TX or RX) mode. All radio devices utilize hardware protocol engines (ShockBurst™) and high data rate to minimize the time used to transfer data. This means short bursts of high and low levels of current consumption. Capturing the peak current levels in different modes is a task that normal ampere meters cannot do.
- High accuracy measurements when radio devices are in standby or power down. High accuracy ampere meters are needed, but how accurate and what should you look out for if the measurements are not as expected?

8.2.1 Test method and HW setup active current consumption

Because of the transient nature of a radio device sending packets, an oscilloscope must be used to measure the voltage drop over a small test resistor put in series with the radio device power supply. On most modern (digital) oscilloscopes you can use mathematics on the measured wave forms and then use this to show the current consumption. See Figure 27. for this type of test setup.

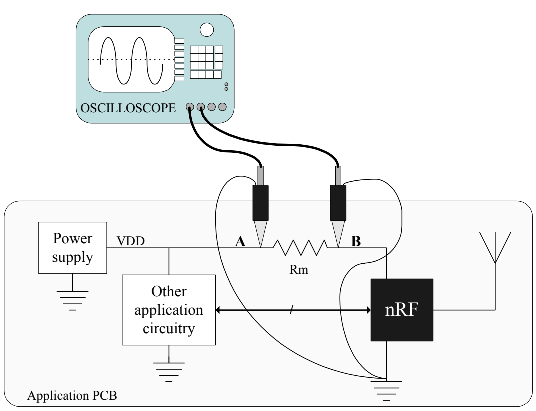


Figure 27. Radio active current consumption measurement

The two probes from the oscilloscope measure the voltage on each side of a test resistor Rm. The current drawn by the radio device gives rise to a voltage drop in the resistor and the difference between the voltage in point A and B gives the following current consumption:

$$I = \frac{V_A - V_B}{Rm}$$

With the mathematics option on the oscilloscope, Channel A minus Channel B ($V_A - V_B$) can be shown on the oscilloscope screen directly. The value read when divided with Rm gives you the instantaneous current consumption of the radio device.

The value of Rm should be in the range of 1 to 10 ohms. A larger resistor gives you a larger voltage drop over the resistor, but a larger resistor also increases the VDD ripple (noise) in the RF device. A large resistor together with the de-coupling capacitors on the RF device will form a low pass filter and smooth the transition to current out, making the measurement difficult to read. To ensure good RF performance the power supply noise should be kept under 50 mVpp. Power supply ripple on the radio device is verified by measuring the voltage in point B.

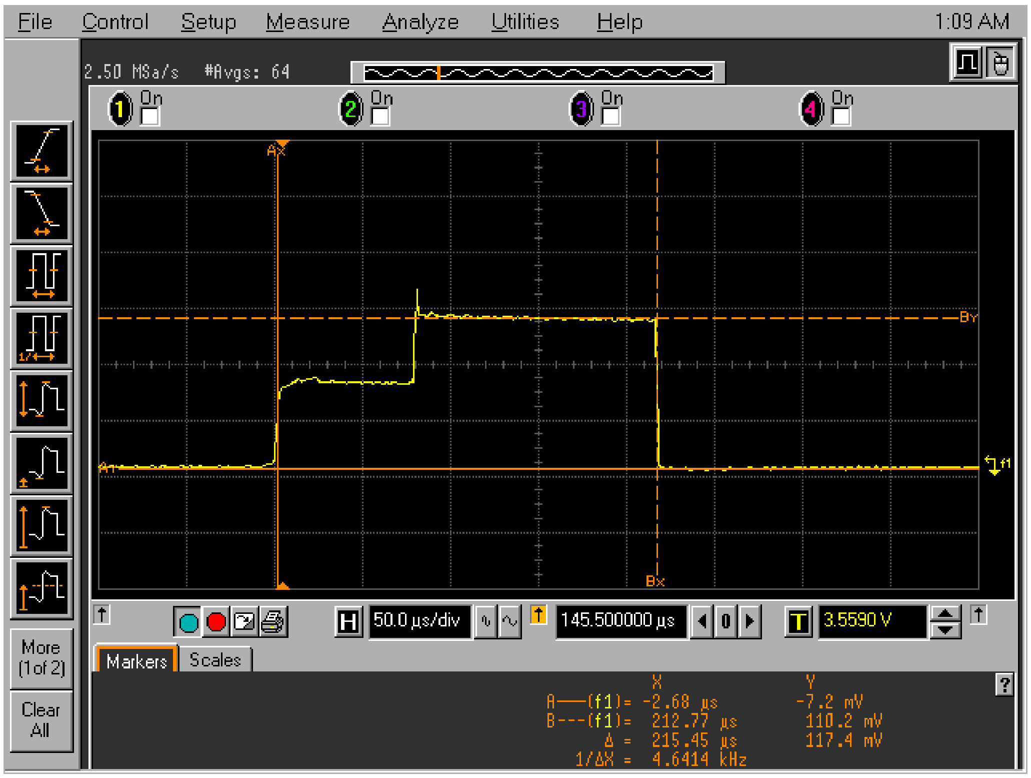


Figure 28. nRF24L01 transient current consumption measurement

Figure 28. shows an example of the current consumption variation during a transmission of a packet using nRF24L01. Initially, the device is in standby and the current consumption is low. From marker AX it increases the current drawn as it powers up internal modules preparing to send data. At t = 80 us the final modules (the PA) are powered up and the device consumes the peak TX current set in the device product specification. After the final modules have stabilized (nRF24L01: t=130us) the actual packet is sent (no change in current consumption). When sending the packet is finished the device automatically returns to standby at t=215 µs.

Note: To capture these transient changes in the DUT current consumption, it is vital that Rm is placed as close to the device as possible.

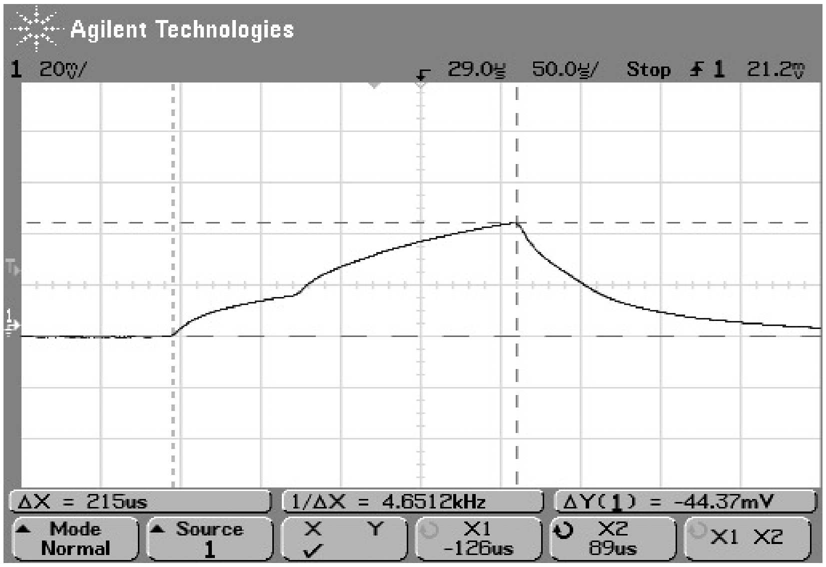


Figure 29. Transient current consumption measurement filtered

Any routing and de-coupling capacitors between Rm and the DUT act as a low pass filter and affect the measurement. Figure 29. shows how the transient current consumption output may look if you have too much filtering between Rm and the DUT. All the transient edges are smoothed out and it is impossible to measure the exact current consumption in the different modes.

If you do not have an oscilloscope with a mathematic option you can do the same measurement with the set up in Figure 30. In this way the $V_A - V_B$ voltage is shown directly on the oscilloscope screen.

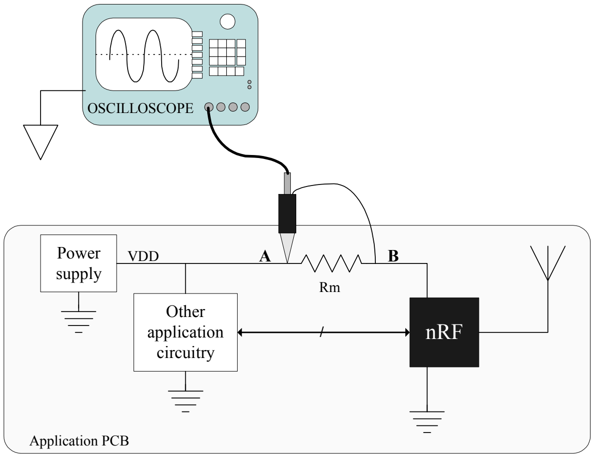


Figure 30. An alternate way to measure current consumption

Now the GND reference of the oscilloscope is no longer the same as the GND in the application. Point B in Figure 30. is an oscilloscope GND reference and is separated from the application VDD by Rm. If your application is mains supplied, ensure that the GND of the oscilloscope is isolated from the GND of your application. If you do not do this massive currents may flow from your power supply through Rm and the oscilloscope to GND.

Warning: This can destroy the power supply in your application and can also be harmful to the oscilloscope.

This isolation is achieved by putting an isolating transformer in series on the oscilloscope mains supply. Isolating transformers can be bought from a number of sources that provide transformers and laboratory power supplies.

8.3 Calculation of average current consumption

If you have an application with a fixed report rate, the average current consumption in a radio circuit can be calculated from this measurement using the following formula:

$$I_{AV} = \frac{\sum_n Ipeak_n \cdot t_n}{\sum_n t_n}$$

As an example let us look at the TX current consumption measured in Figure 28.. Let us assume that the device sends a packet like this every 10 ms and stays in standby using 20 uA between each packet. The overall average current consumption can then be calculated by finding the average current consumption over one of the 10 ms periods. The below equation assumes that the current has peaked from the start of the packet tranmission:

$$I_{AV} = \frac{\sum_n Ipeak_n \cdot t_n}{\sum_n t_n} = \frac{\frac{V_A - V_B}{Rm} \cdot t_{burst} + I_{standby} \cdot t_{standby}}{t_{burst} + t_{standby}}$$

$$= \frac{11,7mA \cdot 0,215ms + 0,02mA \cdot 9,785ms}{10ms} = 0,27mA$$

9 Glossary of Terms

Term	Description
BER	Bit Error Rate: The number of errors a receiver makes.
Conducted measurements	Using cables to conduct all output power measurements on a radio device.
DUT	Device Under Test
FSK	Frequency Shift Keying
Insertion loss	Loss of some power from the DUT in the test cables and connectors. Occurs when working on high frequency systems.

Table 3. Glossary

$\lambda/4$ printed monopole antenna for 2.45GHz

1. Preface

Taking the demand for small size, easy fabrication and low cost into account in the development of low-power radio devices for short-range 2.4GHz applications, a quarter wavelength monopole antenna implemented on the same printed circuit board as the radio module is a good solution. A printed quarter wavelength monopole antenna is very easy to design and can be tuned simply by slight changes in length.

This article presents basic guidelines on how to design such an antenna for use together with the 2.4GHz transceiver and transmitter devices from Nordic Semiconductor.

The described antenna should be fabricated on standard 1.6mm, low cost FR4 printed circuit board (PCB).

2. Basic properties of a quarterwave monopole antenna

A quarterwave monopole is a ground plane dependent antenna that must be fed single-ended. The antenna must have a ground plane to be efficient, and ideally the ground plane should spread out at least a quarter wavelength, or more, around the feed-point of the antenna. The size of the ground plane influences the gain, resonance frequency and impedance of the antenna.

The length of the monopole PCB trace mainly determines the resonant frequency of the antenna, but because of the very wide gain bandwidth of a quarterwave monopole, the antenna length is not too critical. But like any other antenna types, the gain of a quarterwave monopole will vary if parameters in the surroundings, such as case/box materials, distance to the ground plane, size of the ground plane, width and thickness of the PCB trace are varied. If any of these parameters are changed, a retuning of the monopole PCB trace length may be necessary for optimum performance in each application.

3. Determining the length of the printed monopole antenna

The antenna is fabricated on a standard 1.6mm FR4 substrate material with a typical dielectric constant ε_r of 4.4 at 2.45GHz.

The width of the monopole trace is W = 1.5mm. The wavelength in free air is λ_0 = 122mm. It may be approximated that the guided wavelength λ_g on the FR4 substrate is about

$$\lambda_g \sim 0.75 \cdot \lambda_0 = 0.75 \cdot 122\text{mm} \sim 92\text{mm}$$

The approximate, physical length of a printed quarterwave monopole antenna is then

$$L = 92\text{mm} / 4 = 23\text{mm}$$

provided that the size of the available ground plane is close to the ideal as discussed above and that the antenna trace is uniformly surrounded by the FR4 substrate.

When implementing the monopole as a trace on the PCB, the length of the trace should be extended somewhat to allow for some fine-tuning of the antenna to resonance at 2.45GHz. If the size of available ground plane is approaching the ideal size and the antenna trace is uniformly surrounded by the FR4 substrate, then the length of the trace should be extended by about 20%. For an example, see Figure 1a.

If the ground plane size is considerably smaller than the ideal size and/or much of the antenna trace is routed close to the edge of the PCB, then the length of the antenna trace should be extended by about 30%. For an example, see Figure 1b.

4. Combining the printed quarterwave monopole with the nRF24xx device RF-layout

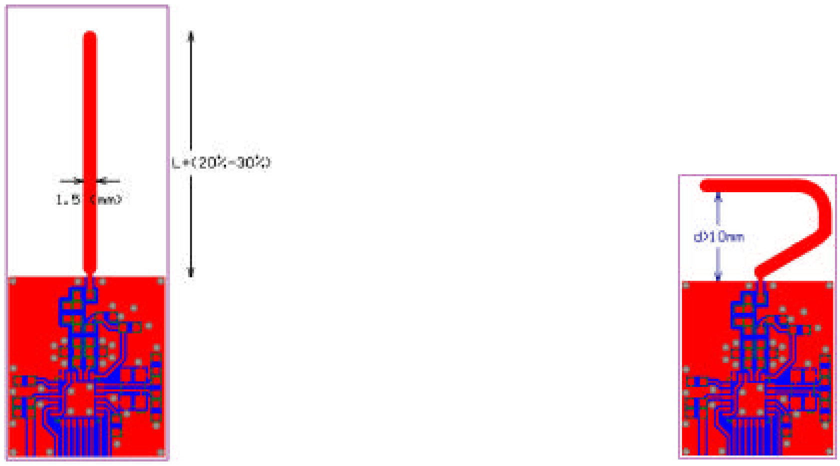
The quarterwave monopole must be fed single-ended, hence a differential to single-ended matching network must be used between the nRF24xx antenna interface ANT1/ANT2 and the monopole feed-point. A suggestion on a differential to single-ended matching network can be found in the nRF24xx datasheet.

Figure 1 shows two examples on how a printed quarterwave monopole can be combined with the nRF2401 RF-layout on the same PCB. Figure 1a shows the optimum placement of the antenna trace. With this placement, the antenna is allowed to radiate freely in all directions. The monopole has maximum radiation in the plane normal to the antenna axis, and minimum radiation along the axis. To be omni-directional, the monopole antenna should be placed vertically.

Figure 1b shows a more compact layout of the antenna trace. This layout may have lower antenna gain in the direction of maximum radiation than for the layout shown in Figure 1a, but it will exhibit a more uniform radiation in the horizontal plane if vertical placement is not possible.

When bending the antenna trace like in Figure 1b, be sure to keep the distance (d) between the open end of the antenna trace and the ground plane as large as possible, preferably 10mm or more. Reducing this distance will reduce the gain of the antenna.

There shall be no ground plane on the PCB layer(s) beneath the antenna trace. No ground plane, PCB traces or components should be placed close to the antenna trace.



a) b)

Figure 1. Examples of nRF2401 RF-layout combined with a printed λ/4 monopole antenna.

Tuning of the antenna is done simply by cutting the length of the PCB antenna trace until resonance at 2.45GHz is obtained. The antenna must be tuned with the PCB placed inside the case/box (if any) and hand-held/body-worn (if this is a hand-held/body-worn application).

For applications where range performance is not critical the antenna can be tuned by measuring radiated power from the antenna with a spectrum analyzer. For more accurate tuning a vector network analyzer must be used for impedance and SWR (Standing Wave Ratio) measurements.

As shown Figure 1 the PCB antenna trace should be made 20%-30% longer than the estimated theoretical length in order to make tuning possible on prototypes. For the production version of the PCB, the optimum antenna length found on the prototype should be used.

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Regulatory and Compliance Standards for RF Devices

White Paper

May 2007

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1 Introduction

This document is an overview of the most important regulations used for RF approval from around the world.

Every device containing a radio sends out RF signals. To limit the signals, regulations are enforced by committees in various countries and regions around the world. The regulations state the maximum permitted radio parameters and regulate the use of frequencies. A product must be certified to ensure it complies with the regulations for the region where it will be sold.

The most important of these regulatory standards are:

- USA: The radio must be certified with the FCC.
- Canada: Industry Canada issues a certification certificate. It uses the same regulations as FCC.
- Europe: Self-certification by the manufacturer based on the R&TTE directive. All products must be CE marked. Harmonized standards from ETSI. Frequency allocation managed by CEPT/ERC rec 70-03.
- Japan: ARIB STD-T66.
- Australia/New Zealand: Self-certification managed by an ACMA auditing process. Uses the AS/NZS 4268:2003. All products must be C-TICK marked.

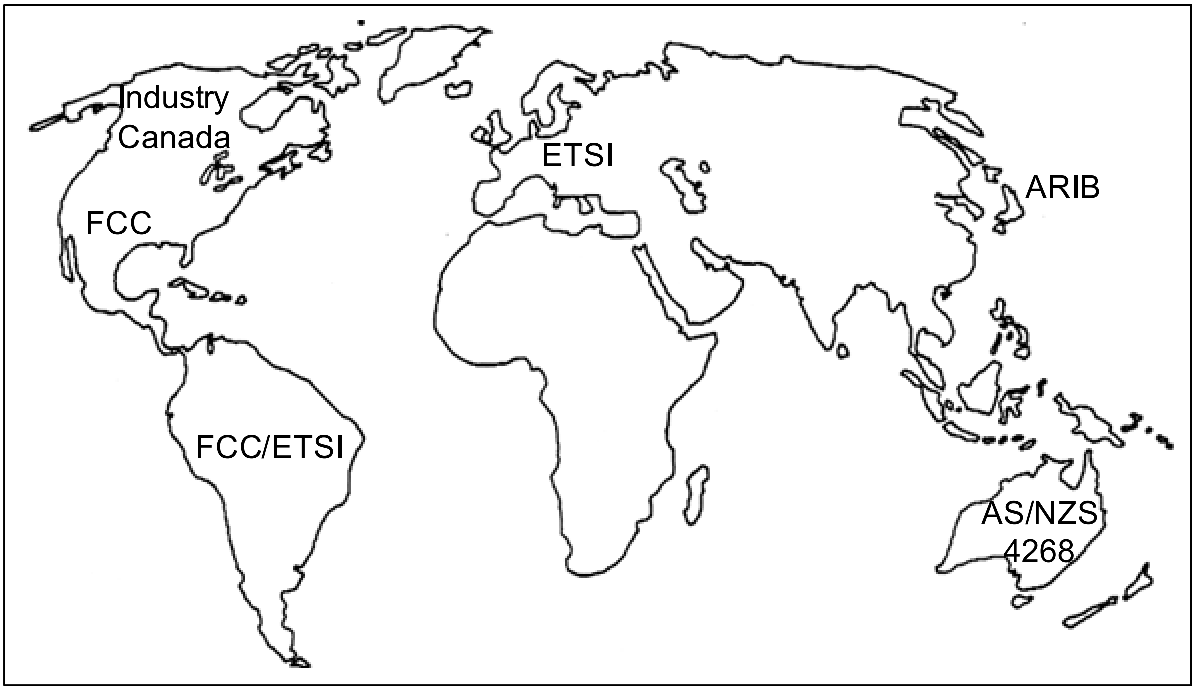

Figure 1. Distribution of world's most important regulations

Depending on the application and product, a number of directives must be followed. This document covers only the RF part of these directives. In addition, there are regulations for electrical safety and so on.

This document discusses the most widespread regulations, with emphasis on the radios produced by Nordic Semiconductor.

The use of the ISB band is license free, but the regulations dictate certain limitations on output power, frequency, duty cycle and, application.

When the product is finished, a EMC (Electomagnetic Compability) test lab must be contacted to conduct the testing. They perform a series of measurements and declare if the product is approved. It is useful to do measurements in the design phase to make the product easier to pass the approval. This document includes some guidelines on how to do such measurements.

It is the finished product that must be approved by the appropriate regulative authorities. This means your product must have the finished housing, PCB and, firmware. This can not be changed after the approval is given. It also means that the radio chip can not be approved individually. There is an exception for modules that have an FCC approval, but some tests are still necessary on the finished product.

Note: The information found in this document can be changed at any time by the different regulative authorities so check the details in the newest official regulations or standards.

Note: Nordic Semiconductor does not take any responsibility for the consequences of usage of this information. All information is believed to be correct at the time of writing.

2 FCC

In the USA, the FCC (Federal Communications Commission) regulates the use of frequencies for RF equipment. CFR 47 part 15 (Code of Federal Regulations) covers the unlicensed ISM bands. This is usually referred to as FCC part 15. All equipment must be certified with the FCC with the issuance of a Grant of Authorization by the FCC. When the product is approved, the FCC issues an identification number which the product must be marked with.

The part 15 sections for ISM band radios are:

- 15.35: General rules for certification measurements
- 15.109: Radiated emission limits for unintentional radiators
- 15.205: Restricted bands of operation
- 15.209: Radiation limits for intentional radiators
- 15.231: Periodic operation in the band 40.66 - 40.70 MHz and above 70 MHz.
- 15.247: Frequency Hopping and Spread Spectrum (FHSS), operation within the bands 902 – 928MHz, 2400 – 2483.5 MHz
- 15.249: Single frequency systems and non FHSS systems, operation within the bands 902 – 928MHz, 2400 – 2483.5 MHz

2.1 Spurious emissions and restricted bands

A large part of the FCC Part 15 is about the limit of spurious emissions (harmonic component) and restricted bands.

The main concept behind part 15 is that a general set of rules must be followed, but with exceptions for different application and frequency bands.

Emissions are divided into two parts:

- Unintended radiators, like receivers and transmitters in standby or other radiators.
- Intended radiators, like active transmitters.

2.2 Receivers

For unintentional radiators the general requirements are given in 15.109. For intentional radiators the general requirements are given in 15.209. The levels are given in EIRP, Electrical Field Strength Power, at a distance of 3 meters.

'Table 1. on page 6' shows the spurious limits at different frequencies.

Frequency	Electrical Field Strength	EIRP
30-88MHz	100µV/m	-55.2dBm
88-216MHz	150µV/m	-51.7dBm
216-960MHz	200µV/m	-49.2dBm
>960MHz	500µV/m	-41.2dBm

Table 1. Spurious emissions for unintentional radiators according to 15.109 and 15.209

There is a number of restricted frequencies according to 15.205 which can not be used for active transmissions, but spurious emissions can be produced. None of the Nordic Semiconductor radios can intentionally transmit in any of these frequencies but, care must be taken if spurious or harmonics falls into these frequency bands.

2.3 Periodic applications

Section 15.231(a) covers Periodic operation in the band 40.66 - 40.70 MHz and above 70 MHz.

The provisions of this section are restricted to periodic operation within the band 40.66 - 40.70 MHz and above 70 MHz. The intentional radiator is restricted to the transmission of a control signal such as those used with alarm systems, door openers, remote switches, and so on. Continuous transmissions, voice, video and the radio control of toys are not permitted. Data is permitted to be sent with a control signal.

The following conditions must be met:

- A manually operated transmitter shall employ a switch that automatically deactivates the transmitter within not more than 5 seconds of being released.
- A transmitter activated automatically shall cease transmission within 5 seconds after activation.
- Periodic transmissions at regular predetermined intervals are not permitted. However, polling or supervision transmissions, including data, to determine system integrity of transmitters used in security or safety applications are allowed if the total duration of transmissions does not exceed more than two seconds per hour for each transmitter. There is no limit on the number of individual transmissions, provided the total transmission time does not exceed two seconds per hour.
- Transmitters which are employed for radio control purposes during emergencies involving fire, security, and safety of life, when activated to signal an alarm, may operate during the pendency of the alarm condition.

The maximum power levels are given in 'Table 2. on page 7'.

Frequency	Electrical Field Strength	EIRP
260-470MHz	3750 to 12500µV/m [a]	-23.7 to -13.2dBm
> 470MHz	12500µV/m	-13.2dBm

a. Linear interpolation

Table 2. Emissions from Intentional Radiators according to 15.231

The spurious levels are 20dB below these levels, unless the spurious falls within the restricted band given in section 15.205.

The maximum bandwidth of the signal is defined in 13.231(c), where the maximum bandwidth below 900MHz is 0.25% of the center frequency and 0.5% above 900MHz. All Nordic Semiconductor radios meet these requirements.

Other periodic applications that do not meet the restrictions in 15.231(a) can still be used, but have a lower transmit power as defined in 15.231(e). In addition, devices operated under the provisions of this paragraph shall be provided with a means for automatically limiting operation so that the duration of each transmission shall not be greater than one second and the silent period between transmissions shall be at least 30 times the duration of the transmission but in no case less than 10 seconds.

The power limit is 5000µV/m or -13.3dBm (EIRP) for all frequencies above 470MHz.

2.4 The 915MHz band

The nRF9x5 radio can operate in the 433, 868 and, 915MHz bands. 'Table 2. on page 7' shows that using these bands allows very low output power. But section 15.249 gives the opportunity to use the 902-928MHz band with a higher output power, 50mV/m at 3 meters or -1.2dBm. The harmonics are limited to 500µV/m or -41.2dBm. This band is commonly referred to as the 915MHz band in the USA. There are no restrictions on duty cycle or the application.

Even higher output power can be used according to section 15.247. This implies the use of frequency hopping. The following requirements must be fulfilled for the use of the 902-928MHz band under section 15.247:

- Hopping channels shall be separated by minimum 25 kHz or the 20dB bandwdith of the hopping channel, whichever is greater.
- The system shall hop to channel frequencies that are selected at the system hopping rate from a pseudorandomly ordered list of hopping frequencies.
- If the 20 dB bandwidth of the hopping channel is less than 250 kHz, the system shall use at least 50 hopping frequencies and the average time of occupancy on any frequency shall not be greater than 0.4 seconds within a 20 second period.
- If the 20 dB bandwidth of the hopping channel is 250 kHz or greater, the system shall use at least 25 hopping frequencies and the average time of occupancy on any frequency shall not be greater than 0.4 seconds within a 10 second period.
- The maximum allowed 20 dB bandwidth of the hopping channel is 500 kHz.

The nRF905 has a 20dB bandwidth of a little more than 200 kHz. This means that the channel separation shall be more than 200 kHz. The closest is 400 kHz, giving 65 hopping channels in the 902-928MHz band.

Maximum peak conducted output power is 1W in the 902-928MHz band if more than 50 hopping channels are used. The limit is 0.125W for 25 to 50 hopping channels.

2.5 The 2.4GHz band

The same two sections, 15.247 and 15.249 apply for the 2.4 GHz band. The permitted frequency is 2400 to 2483.5MHz.

For non-frequency hopping systems (15.249), the power limit is 50mV/m at 3 meters or -1.2dBm and the harmonics are limited to 500µV/m or -41.2dBm. The latter also applies for the radiation outside the frequency band (15.109). Remember that transmissions close to the band edge causes a too high level outside the band. It is therefore necessary to avoid using channels close to the band edge.

15.249(e) states that the field strength limits for frequencies above 1GHz are based on average limits. However, the peak field strength of any emission shall not exceed the maximum permitted average limits specified above by more than 20 dB under any condition of modulation.

This means that the permitted output power and the harmonics levels can be increased with up to 20dB depending on the duty cycle, after the following formula:

$$P_{increase} = 20\log\left(\frac{T_{on}}{T_{on} + T_{off}}\right)$$

where T_{on} is the time of transmission and T_{off} is the pause between transmissions.

Section 15.247 also applies for the 2.4GHz band. The nRF24xx series of radios have a channel spacing of 1MHz. This gives the following requirements for operation under section 15.247:

- Frequency hopping systems operating in the 2400-2483.5 MHz band may have hopping channel carrier frequencies that are separated by 25 kHz or two-thirds of the 20 dB bandwidth of the hopping channel, whichever is greater, provided the systems operate with an output power no greater than 125 mW.
- The system shall hop to channel frequencies that are selected at the system hopping rate from a pseudorandomly ordered list of hopping frequencies.
- Frequency hopping systems in the 2400-2483.5 MHz band shall use at least 15 channels. The average time of occupancy on any channel shall not be greater than 0.4 seconds within a period of 0.4 seconds multiplied by the number of hopping channels employed. Frequency hopping systems may avoid or suppress transmissions on a particular hopping frequency provided that a minimum of 15 channels are used.
- Maximum peak conducted power for systems with more than 75 non overlapping hopping channels is 1W (+30dBm) and 0.125W (+20.9dBm)

Harmonics must be 20dB below the peak in-band emission in any 100 kHz bandwidth for FHSS systems.

Note: Section 15.247 (a)(2) specifies “Systems using digital modulation techniques may operate in the 902 - 928 MHz, 2400 - 2483.5 MHz, and 5725 - 5850 MHz bands. The minimum 6 dB bandwidth shall be at least 500kHz”. Digital modulation is here DSSS (Direct Sequence Spread Spectrum) and does not apply for the Nordic Semiconductor radios.

3 ETSI

In Europe, the procedure of qualification of wireless equipment in the European Economic Area is explained in the R&TTE directive (Directive 199/5/EC of the European Parliament and Council). The standards used to define tests and requirements are made by standardisation bodies, like CEPT (The European Conference of Postal and Telecommunications Administration) and ETSI (European Telecommunications Standards Institute).

Low power devices are referred to as Short Range Devices (SRD).

CEPT has responsibility for frequency allocation and the use of the different bands. This is covered by ERC (European Radiocommunications Committee) recommendation CEPT/ERC 70-03.

ETSI develops and administrates standards for testing and type approval of RF equipment.

General EMC requirements for RF equipment are outlined in EN 300 683 or EN 301 489-3.

The most used ETSI standards for SRD radios are:

- EN 300 220 – 25MHz to 1GHz
- EN 300 440 – 1 to 40GHz
- EN 300 328 – 2.4GHz, Frequency hopping / Spread Spectrum

The manufacturer must self-declare that the product complies with the R&TTE directive in order to mark it with the CE logo.

3.1 Frequency plan

The frequency plan used in the European countries for SRD radios are covered in the CEPT/ERC rec 70-03. The CEPT member countries must ratify the recommendation but there can be different frequency limits between CEPT member countries.

The frequency band allowed for SRD radio use is listed in 'Table 3. on page 10'.

Frequency band	Power	Duty Cycle	Channel Spacing	Notes
433.05-434.79 MHz	+10 dBm	<10%	No limits	
433.05-434.79 MHz	0 dBm	No limits	No limits	
433.05-434.79 MHz	+10 dBm	No limits	< 25 kHz	
865-868 MHz	+20dBm	LBT	200 kHz	RFID[a]
868-868.6 MHz	+14 dBm	<1% or LBT[b]	No limits	
868.7-869.2 MHz	+14 dBm	<0.1% or LBT	No limits	
869.3-869.4 MHz	+10 dBm	No limits	< 25 kHz	
869.4-869.65 MHz	+27 dBm	<10% or LBT	< 25 kHz	
869.7-870 MHz	+7 dBm	No limits	No limits	Channels may be combined
2400-2483.5 MHz	+7.85 dBm	No limits	No limits	

a. Not implemented in all European countries at time of writing, according to ERC/REC 70-03 Annex 11B1
b. LBT = Listen Before talk

Table 3. List of frequency bands in Europe

3.2 Below 1GHz

The ETSI EN 300 220 defines the method of measurements for frequencies below 1GHz.

3.2.1 Receivers

The receivers are divided into three classes, each having its own set of minimum performance criteria. The classification is based upon the impact the equipment has on persons in case of a failure.

Receiver class	Risk assessment of receiver performance
1	Failure may result in a physical risk of a person
2	Failure causes loss of function but not physical risk of a person
3	Failure causes loss of function but can be overcome by parallel means

Table 4. Receiver classes

The Nordic Semiconductor radios fulfill class 2. A class 2 receiver does not have any Adjacent Band Selectivity requirements.

The blocking levels must be better than -69dBm at +/-2MHz from the center frequency and better than -44dBm at +/-10 MHz. The method of measurements is described in chapter 9.4 in the ETSI standards.

The nRF9x5 fulfills all requirements for a class 2 receiver.

Spurious emissions for the receiver must be less than -57dBm for frequencies below 1GHz and less than -47dBm above 1GHz.

3.2.2 Transmitters

The method of measurement for transmitters is found in chapter 8 in the ETSI standards. These are the relevant clauses for the nRF9x5:

8.1: The frequency error must be better than +/-100ppm.

8.7: The nRF9x5 is defined as a wide band transmitter. This means transmitters with a modulation bandwith over 200kHz.

8.8: Spurious emissions must comply with the limits in 'Table 3. on page 10'.

Frequency State	47 to 74MHz 87.5 to 118MHz 174 to 230MHz 470 to 862MHz	Other frequencies below 1GHz	Above 1GHz
Operating	4nW / -54dBm	250nW / -36dBm	1µW / -30dBm
Standby	2nW / -57dBm	2nW / -57dBm	-20nW / -47dBm

Table 5. Spurious limits

8.10: Duty cycle class is defined in CEPT/ERC 70-03

8.11: Listen before talk (LBT)

If the transmitters use LBT, the duty cycle requirement does not apply. The principle is to use the receiver to see if the channel is clear before a transmission is started. The Carrier Detect function of the nRF9x5 can be used for this. The following requirements must be fulfilled:

- The receiver must have a LTB threshold defined in clause 9.2.
- The receiver blocking requirements in clause 9.4.
- Minimum TX off time is 100ms.
- The total listen time, t_L, concists of a fixed part, t_F, and a pseudorandom part, t_{PS}, as the following: $t_L=t_F+t_{PS}$
  - The fixed part is minimum 5ms.
  - The pseudorandom time shall vary between 0ms and 5ms in steps of 0.5ms.
  - If the channel is free at the start of the listening time, and remains free throughout the fixed part of the time, the pseudorandom time can be set to 0.
  - If the channel is occupied at the start of the listening time, or during the period, the pseudo random time must be included to the total listening time.
  - An acknowledge packet can be sent as a receipt for the received message. No listening time is required for this.
  - The limit for a transmission is 1s.
  - The limit for a transmission sequence is 4s.

3.3 Usable frequencies for the nRF9x5

The nRF9X5 is a wide band radio with 100 kHz channel spacing in the 433 MHz band and 200 kHz in the 868MHz band, so the usable frequencies is limited to those in 'Table 6. on page 12'.

Frequency	Power	Duty cycle
433.05-434.79 MHz	+10dBm	< 10%
433.05-434.79 MHz	0dBm	No limits
868.2-868.4 MHz	+14dBm	<1% or LBT
869.0 MHz	+14dBm	0.1% or LBT

Table 6. Usable frequencies for the nRF9x5

The frequency band 865-868MHz can be used according to ERC/REC 70-03, annex 11B1, but this is not implemented in all European countries. Confirm with the ERC/REC 70-03 before you choose this frequency.

3.4 **The 2.4GHz band**

For non-frequency hopping systems, the ETSI EN 300 440 defines the method of measurements for radios operating from 1 to 40GHz.

The Nordic Semiconductor radios in the nRF24xx series are made to operate within the 2.4GHz band and can be used at full output power (1mW) in the 2400 - 2483 MHz band.

'Table 6. on page 12' shows the spurious limits for the radio in standby or RX mode, and active or TX mode.

For FHSS devices, the ETSI EN 300 328 applies.

The main requirements for a FHSS system are:

- Make use of minimum 15 non-overlapping channels
- Minimum channel separation is 1MHz
- Dwell time pr. channel shall not exceed 0.4s
- For adaptive FHSS systems, minimum 90% of the band shall be used and minimum 20 channels.

4 Other regulatory standards

4.1 ARIB

In Japan, the ARIB STD-T66 applies for equipment operating in the 2.4GHz band.

Communication is limited to digital transmission systems, including spread spectrum. The allowed frequency band is 2400 to 2483.5MHz.

Maximum allowed output power is 10mW measured in a 1MHz bandwidth for a non-FHSS system. If FHSS is used, maximum power is 3mW measured in a 1MHz bandwidth if the whole band is used. 10mW can be used for FHSS systems, but not from 2.427 to 2.47075GHz, where the limit is 3mW.

The power to the antenna shall be within +20% and -80% of the rated power.

Maximum frequency tolerance is +/-50ppm. This is directly linked to the tolerance of the crystal used on the nRF devices.

Spurious emissions shall be measured on the antenna connector and must be below 25μW / -16dBm inside the band and 2.5μW / -26dBm outside.

Dwell time per channel shall not exceed 0.4s for FHSS systems.

For receivers, the spurious limit is 4nW / -53dBm below 1GHz and 20nW / -47dBm above 1GHz.

4.2 Australian/New Zealand Standard

The Australian/New Zealand standards - AS/NZS 4268:2003 are mostly based on the measurements defined in the ETSI stardards EN 300 220 (25 to 1000MHz) and EN 300 440 (1 to 40GHz). But the permitted frequency bands are different from Europe. 'Table 7. on page 14' shows an overview for the frequencies that can be used with the Nordic Semiconductor radios.

Frequency	Max. EIRP	Spurious EIRP
433.05 to 434.79MHz	25mW / 24dBm	0.1μW / -40dBm
915 to 928MHz	3mW / 4.7dBm	0.1μW / -40dBm
2400 to 2483.5MHz	4W / 36dBm	Note 1

Table 7. Frequency bands usable with the Nordic Semiconductor radios

Note: In any 100 kHz bandwidth outside the frequency band in which the transmitter is operating, the power shall be at least 20 dB below that in the 100 kHz bandwidth within the band that contains the highest level of the desired power. The minimum 6 dB bandwidth shall be at least 500 kHz.

4.3 Korea

A special precaution must be taken if the device is being used in Korea. The spurious in RX mode must be below -54dBm. This shall be a conducted measurement, that means the spectrum analyser is directly connected to the radio instead of the antenna.

5 Measurements on Nordic radios

5.1 Emission conversion

There is a number of ways to specify the radiated power from an antenna. E (Electrical Field Strength) defines the energy in a distance from the antenna. This is highly dependent on the antenna. If the energy is radiated from an ideal, isotropic antenna, where the radiation is uniform in all directions, the power can be expressed as EIRP (Effective Isotropic Radiated Power). If the antenna is a half wave dipol, the same can be expressed as ERP (Effective Radiated Power). It can be useful to convert between E and EIRP since the latter expresses what output power the radio must have to generate an electrical field of a specific strength with an ideal antenna. The following formula can be used to convert between the two:

$$EIRP = 10 * \log\left(\frac{4 * \pi * E^2 * r^2}{0.377\left[V^2\right]}\right) = 10 * \log\left(\frac{E^2 * r^2}{0.030\left[V^2\right]}\right)$$

Where ***r*** is the distance, ***EIRP*** is in dBm and ***V*** is the unit of measurements.

FCC normally states the distance to 3 meters, so the equation can be simplified to:

$$EIRP = 20\log\left(\frac{E}{\mu V / m}\right) - 95.23$$

Where ***E*** is the Electrical Field Strength at 3 meter and ***EIRP*** is in dBm.

A half wave dipol has a gain that is 2.15dB higher than an isotropic antenna, so:

$$EPR = EIRP - 2.15dB$$

5.2 General measurement procedure

This section describes how to do initial testing before sending the device to the EMC lab.

To properly test the RF parameters, there are two tests that need to be done:

- TX test for the channels in use, or through the band.
- RX test, look for spurious.

The tests should first be conducted, where a coaxial cable is connected to the DUT (Device Under Test) instead of the antenna. If the antenna is connected directly to the matching network, the track needs to be cut. Do not connect the antenna and the coaxial cable in parallel.

If there are components between the matching network and the antenna, they can be removed to disconnect the antenna.

A spectrum analyzer is connected to the cable. It is recommended to have a spectum analyzer with a maximum frequency of more than the 4^{th} harmonic.

The TX is then set to transmit on the chosen channel. If the whole band is to be used, the lowest and highest channel must be checked, in addition to the middle channel. It can be useful to let the TX sweep through the band.

The transmission can either be packets that are sent as fast as possible or you can set the radio to transmit a fixed carrier.

Measure the output power and all the harmonic components. The harmonics can be found at $n*f_c$, where n is an integer and f_c is the carrier frequency.

Note the values and compare with the radio standard or regulations. There should be sufficient margins to safely assume the device passes the final test.

The RX test is performed in a similar manner. Here the whole spectrum must be analyzed to see if there is any spurious components above the limit.

In RX mode, there will be some leakage from the local oscillator. The frequency of this is both device specific and dependent on the frequency the RX is set on.

For the nRF9x5, the LO can be found at f_0+1MHz. This will be very weak and does not represent an issue you need to monitor.

For the nRF24xx series radios, the LO is found using the equation:

$$f_{LO} = \frac{8}{7} * f_0 + IF$$

Where f_{LO} is the local oscillator, f_0 is the RX frequency and IF is:

- nRF24Z1 IF=4
- nRF24L01IF=2
- nRF2401, nRF24E1, nRF24AP1 IF=3

All frequencies are in MHz.

You should also conduct the same measurement in an anechoic chamber. This lets you see the effect of the antenna and also lets you check if the board itself radiates.

A real antenna differs from an isotropic antenna. The radiation pattern will in most cases not be circular, but have some gain in one or more directions. At the approval process, the lab will look for the maximum output power by turning the device in both the horizontal and vertical plane. Unwanted components can be problematic if the antenna has more than 0dBi gain (gain with reference to an isotropic antenna). This means it is important to have some margins on the regulations to account for the antenna gain.

To measure the frequency accuracy, either a fixed carrier is needed or an inductive probe to measure the crystal. If the crystal reference oscillator is probed, the frequency changes due to the capacitance in the probe. It is a better solution to measure the carrier frequency instead or use an inducive probe and measure the leakage from the oscillator. Such a probe can be made of a coiled wire soldered on the end of a coaxial cable. The coil is held close to the crystal can and a spectrum analyzer is used to measure the frequency. For all Nordic Semiconductor radios, the RX and TX frequency is directly connected to the oscillator. It is important that both the TX and RX end of the link uses the exact frequency.

5.3 Common measurements mistakes

5.3.1 Spectrum analyzer resolution bandwidth

When measuring on low level signals, like harmonics and spurious, it is important to adjust the spectrum analyzer so that the signal is well above the noise floor. When the signal is too close to the noise floor, the level will read out higher than it actually is. Decreasing the span and the resolution bandwidth lowers the level of the noise floor and gives more accurate readings. The noise floor should be more than 10dB below the signal. 'Figure 2. on page 17' shows a signal at 2.4GHz with a level of -70dBm. The signal comes from a signal generator. Using a resolution bandwidth of 100kHz brings the noise floor up to around -72dBm, giving only 2dB margin between the signal and the noise floor. The marker reads out -66.6dBm, an error of 3.4dB.

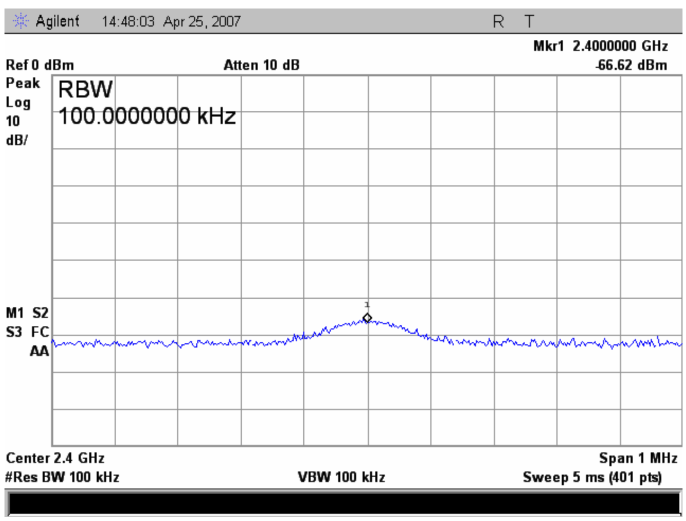

Figure 2. 2.4GHz signal, -70dBm, RBW=100kHz

Decreasing the resolution bandwidth to 3 kHz brings the noise floor down and the marker reads a correct signal level, see 'Figure 3. on page 18'.

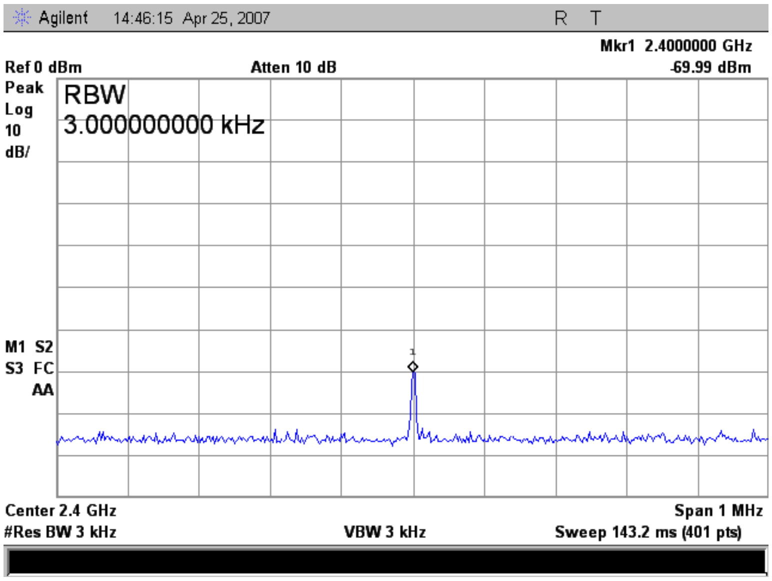

Figure 3. 2.4GHz signal, -70dBm, RBW=3kHz

Using a resolution bandwidth lower than the bandwidth that is measured gives too low readings. If a fixed, unmodulated carrier is output from the DUT, a low resolution bandwidth can safely be used to measure the carrier power and the harmonics. The same applies for the LO in RX mode, this signal is a single tone with low bandwidth, this means a low resolution bandwidth can be used.

In all cases, the resolution bandwidth is too low when the peak power decreases as the resolution bandwidth is decreased.

5.3.2 Intermodulation

Intermodulation in the input stage in the spectrum analyzer occurs when the input signal level is too high. The front end LNA is saturated and false frequency components are generated. This can be a problem if the DUT is set to TX mode and the spectrum analyzer is used to look for spurious. Especially when the frequency span is set to a range below the carrier. A typical property of intermodulation is that the signals vary with a factor k of the input signal, where k > 1. So if the input signal is decreased by 1 dB, the intermodulated signal decreases by more than 1 dB.

Always check this if a signal is suspected to be spurious. Also, intermodulation is less of a problem if the input of the spectrum analyzer is not saturated. 'Figure 4. on page 19' and 'Figure 5. on page 19' show what happens when the input signal is decreased by 3dB. The intermodulated signal drops by 18dB.

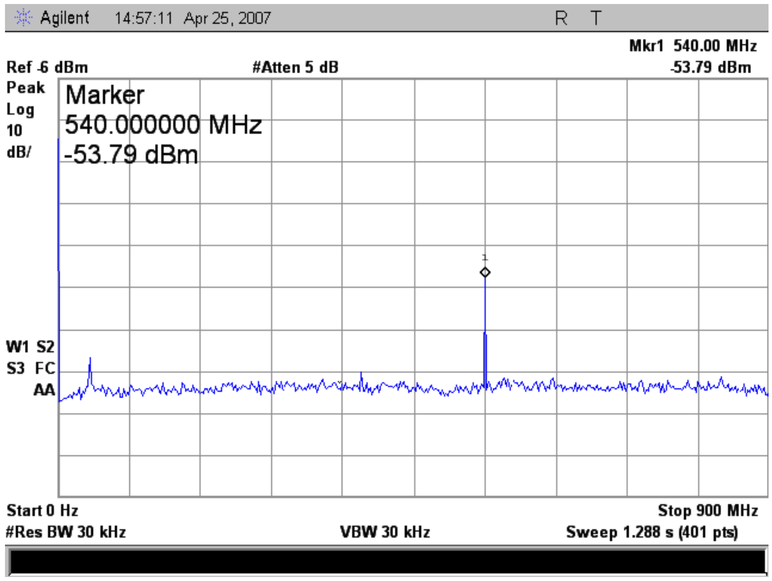

Figure 4. Input signal 1GHz, +13dBm

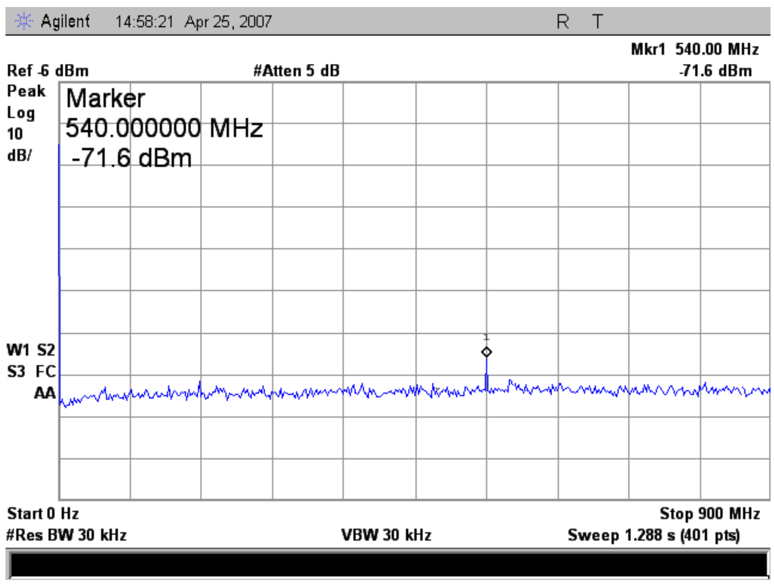

Figure 5. Input signal 1GHz, +10dBm

5.3.3 Application specific standards

Many radio standards and regulations clearly specify the application and how it is meant to be used. The European frequency plan has a number of exceptions and requirements for the use of different frequencies. Always make sure the application fits the requirements, or find the correct standard, regulation and/or frequency plan for the application.

6 Preparation for test

Before the device is sent to the test lab, there are a number of things that must be done:

- All hardware and software must be the final version; no changes can be made after qualification.
- The PCB must be in the final enclosure or housing.
- Some test firmware must be made, it is helpful to have a test mode where the radio is run on different channels, usually low, mid and high channels:
  - The TX is set to a fixed carrier at different channel.
  - Modulated carrier, done by sending Shockburst packets as fast as possible.
  - RX in constant on mode.

The normal operation of the equipment must be stated, in case the duty cycle has an impact of the test limits.

Some regulation standards require conducted measurements, so an extra device must be delivered with a connector instead of the antenna.

Alternatively, the qualification or test lab should be able to tell you about the information that must be submitted.

7 Further information

More information regarding the FCC rules and requirements are found at:

http://www.fcc.gov

ETSI standards are found at:

http://www.etsi.org

The CEPT/ERC rec 70-03 are found at:

http://www.ero.dk

The ARIB STD-T66 regulation is found at:

http://www.arib.or.jp/english/index.html

The Australian/New Zealand Standard AS/NZS 4268:2003 are found at:

http://www.standards.org.au

Sharing crystal with an MCU

nWP-011

White paper v1.1

2011-06-29

1 Introduction

This white paper gives guidelines on how a crystal can be shared between an nRF transceiver from Nordic Semiconductor and an external microcontroller unit (MCU). The nRF24L01+ (and the nRF24L01) has been used as an example in this paper, and all the data is based on the datasheet of that device. However, we refer only to the nRF24L01+ in the rest of this document. In case other nRF devices are to be used, all necessary data can be found in their corresponding datasheets.

2 Crystal from MCU used to feed nRF24L01+

In this setup the MCU controls the startup of crystal, feed the clock to nRF device, and also specify most of the crystal requirements. This is the recommended setup, based on the advantages listed at the end of this chapter.

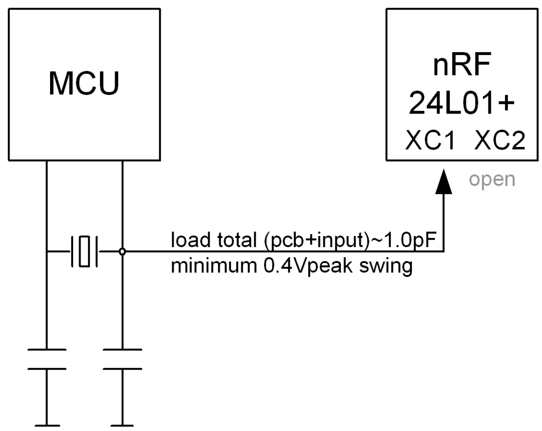

Figure 1. MCU feed nRF24L01+

The nRF24L01+ crystal oscillator is amplitude regulated. To achieve low current consumption and also good signal-to-noise ratio when using an external clock, it is recommended to use an input signal larger than 0.4 V-peak. When clocked externally, XC2 is not used and must not be connected.

The input signal must not have amplitudes exceeding any rail voltage, but any DC voltage within this is OK. Exceeding rail voltage will activate the ESD structure and the radio performance is degraded below specification.

When the MCU drives the nRF24L01+ clock input, the requirement of load capacitance C_L is set by the microcontroller only. The frequency accuracy of ±60 ppm[1] is still required to achieve a functional radio link. The nRF24L01+ will load the crystal by 1.0pF[1] at XC1 in addition to the PCB routing.

Advantages with this kind of setup are summed up in the following bullet points:

- The MCU can control the startup of the crystal oscillator.
- The MCU will ensure a large signal (> 0.4 V-peak) from crystal.
- C_L can be chosen freely depending on MCU requirements (a high value of C_L leads to higher current consumption).

1. This data is nRF24L01+ specific, and must be checked if other nRF devices are to be used.

Requirements	MCU	nRF24L01+
Frequency		16 MHz[a]
Tolerance		+/- 60 ppm[a]
$C_{L\ max}$	X	
$C_{O\ max}$	X	
ESR_{max}	X	

a. This data is nRF24L01+ specific, and must be checked if other nRF devices are to be used.

Table 1. Crystal requirements set by respective device (see Figure 1.)

3 Crystal from nRF24L01+ used to feed MCU

In this setup the nRF24L01+ controls the crystal startup, feeds the MCU clock, and also sets all requirements to crystal specifications.

The nRF device needs to be set in power up mode before the crystal oscillator starts. This means that the MCU must be able to do this before it receives the clock from the nRF device. This is not a recommended setup and there are no advantages with this kind of setup.

4 Conclusion

The advantages of sharing a crystal with an MCU are fewer components and smaller PCB board, which means lower cost to bill of materials. Also it's possible to use a crystal with other specifications, by feeding the clock from an MCU.

An MCU can be used to feed clock to the nRF24L01+, however the opposite is not recommended. This is mainly due to the fact that the nRF24L01+ is by default in power down mode, and needs to be configured with power up bit[1] high, before it will start up its crystal oscillator.

1. This data is nRF24L01+ specific, and other nRF devices might have power up as an input pin.

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Revision History

Date	Version	Description
June 2011	1.1	Updated template, chapter 2 and replaced the product name “nRF24L01” with “nRF24L01+”
May 2006	1.0	

Tuning the nRF24xx matching network

1. Preface

This white paper describes how to tune the matching network of the nRF24xx series radios for optimum performance. The nRF24L01 is used as an example, but the method shown is applicable for all Nordic 2.4GHz radios. A spectrum analyzer is needed to monitor the output signal and measure the performance while the components in the matching network are changed.

This white paper does not cover antenna tuning.

2. How it works

The antenna connection for the radio chip is differential. The impedance is different from 50 ohm. The components between the radio and the antenna are called the matching network and consist of the components shown in Figure 1.

The matching network has 4 main tasks:

1. Transform the balanced output of the radio to an unbalanced connection to the antenna (balun).
2. Transform the output impedance of the radio to a 50 ohm antenna.
3. Suppress harmonics to a level below the regulations level in TX mode.
4. Suppress the LO leakage in RX mode.

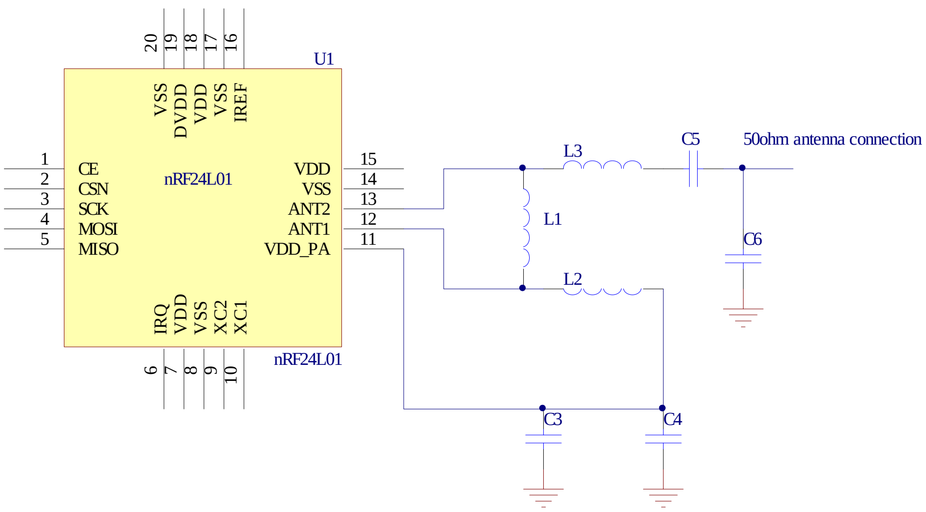

Figure 1- Matching network

The balun-function is done by the 180° degrees phase shift in L1. This causes the signals from the two ANT pins, 180° out of phase, to be added in the point between L1 and L3.

The impedance transformation is done with L3 and C5.

Harmonic suppression is good because the matching network forms a low pass filter (series L and shunt C).

The LO leakage in RX mode is mainly a common mode signal. The LO leakage is shorted by C4 via L2.

Both ANT ports will see the same impedance due to the symmetric matching network. This ensures that the output PA in the TX will operate as designed.

VDD-PA is considered as AC ground. C3 and C4 forms a DC and AC decoupling. VDD-PA is the power supply to the output PA and there must be a DC path between VDD-PA and both ANT pins. The DC will flow trough L2 and L1.

All components have parasitic effects. For instance, an inductor will have series R and shunt C and a capacitor will have a series L and shunt C in addition to the wanted C. The circuit board will also have some parasitic effects. These effects are included in the matching network. It is therefore difficult to calculate the component values. Complex simulation models, where all parasitic effects are included, are often necessary.

Components with different size and from different manufacturers will have different parasitic effects. A layout different from the Nordic recommended layout will also have different parasitic effects.

3. Tuning

To measure the performance of the matching network, two things are needed: Test firmware to run the radio, and a spectrum analyzer. The frequency range of the spectrum analyzer must be higher than the highest harmonic of interest. Normally it is sufficient to test up to the 3rd harmonic, but to it is preferable to test up to the 5th harmonic, e.g. 12,25GHz.

For the nRF24L01, it is easiest to connect the SPI-bus of the radio directly to the evaluation kit and use the test software to run the radio. It simplifies the channel selection and setting of a constant carrier. The MCU connected to the radio could be programmed with a test firmware which sets the channel and the TX mode to a constant carrier. Please refer to the datasheet of the device for details about this. It is also a good idea to have the test firmware sweep trough the band to see if the performance changes over frequency.

The matching network will be optimal in both RX and TX mode if the TX output power is maximized at the same time as the harmonics and the RX LO are minimized. All these parameters must be measured every time a component value is changed. The RX sensitivity is directly proportional with the TX output power.

The spectrum analyzer must be connected to the 50ohm output of the matching network. If a PCB antenna is used, the trace to this must be cut in order to disconnect it. Use a suitable cable, small enough to solder to the board. The screen of the cable must be firmly connected to the ground plane of the board. The center connector is soldered to the 50 ohm point, between C5 and C6. Keep the center conductor short and solder the screen to the ground plane close to the 50 ohm point.

Set the center frequency for the spectrum analyzer, the reference level and the frequency span based on the frequency to be measured. The resolution bandwidth should be 1MHz.

Always start with the component values from the recommended layout.

Follow the flow chart in Figure 2. Adjust the components if the wanted performance is not reached. Adjustments of one component will affect the performance for all parameters. For instance if the output power is too low, and the value of L1 is increased, the harmonic or the LO level might be too high. It is an iterative process. Never change more than one component at a time; else it will be impossible to see the effects of each component. Too low output power and a too high 2^{nd} harmonic are the most common problems. A small increase in L1 normally gives a higher output power and an increase of C6 will decrease the 2^{nd} harmonic. But a too high value of C6 will also affect the output power level.

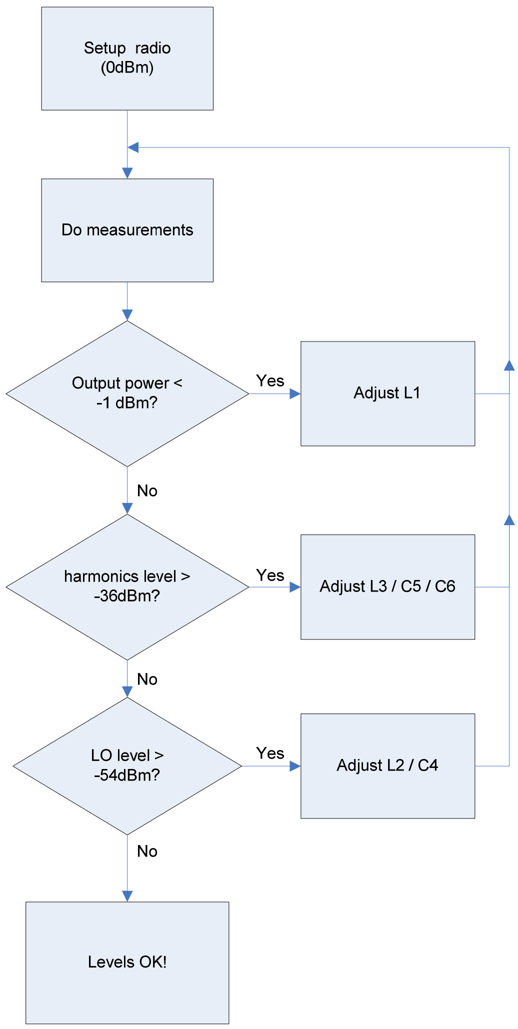

Figure 2- Measurement flow chart

The component values given in the recommended layout on the Nordic Semiconductor web-page can normally be used if the recommended layout is used. But if the layout or the board thickness is changed, some of the components must be changed due to changes in the parasitic effects. The value can be either higher or lower, depending of the stray capacitance and inductance on the board. This also applies if a board with different thickness is used, or if the component size is different from the recommended layout.

The levels for the harmonics and LO given in the flow chart ensures world wide use. The carrier level must be below 0dBm. The harmonics level must be below the outside band spurious level defined by the radio regulations. For world wide use, this is -36dBm. The LO level is defined by the spurious level in RX mode, -54dBm. Please refer to the regulations in the area where the equipment is intended to be used.

The LO frequency in RX mode can be found with this equation:

$$LO = \frac{8}{7}(RF + x) \tag{3.1}$$

Where x is:
nRF2401A 3
nRF24L01 2
nRF24Z1 6

All frequencies are in MHz. RF=Receiving frequency.

Check if the output power level is the same across the band, i.e. measure the power at ch0, ch40 and ch80.

Table 1 shows a suggestion for a measuring series. Write down the changed component value in the same row as the measured values. This makes it easy to track the changes and see the effects of the change. It is necessary to do all the measurements at every component change to see the effects of the change.

When a component is changed, never increase or decrease more than one E12 value at a time.

Some coils are marked in one end, depending on the make. Make sure the direction does not change, as this will affect the performance. The inductor should be tested in both directions and the results noted to find the best orientation. Clearly mark the schematic and the layout with the coil direction for production.

Chip inductors are preferable for mass production due to low cost. Wire wound inductors will have a better performance, but at a higher cost. The reference matching network from Nordic Semiconductor specifies chip inductors. A wire wound inductor will have a higher Q-value and a higher SRF (Self Resonation Frequency). Make sure the inductors have a SRF that is higher than 3GHz. This is especially important if an inductor with a high value is used, as the SRF decreases with the value.

For the capacitors, use NP0 or better quality (+/- 2%) to minimize the variation between samples.

Always use the same component brand/type as in the production. The value of a component is set by the manufacturer, measured at a specific frequency. Different manufacturers are using different methods for this, so even if the value is the same, the effect the component has on the matching network will be different. This is especially important for the inductors.

4. Similarities between devices

Some of the Nordic devices have different matching networks. They all use the same principle, but some have additional tasks.

The nRF2401 matching network is shown in Figure 3. Compared with the nRF24L01 network, C8 and C9 is added and there is a trap formed by C10 and L4. The two caps, C8 and C9, is part of the harmonic filtering. They have been included on the nRF24L01 chip so they have been removed from the matching network for the device. The trap is tuned at LO/2, i.e. half of the frequency from equation 3.1. The LO rejection is improved on the nRF24L01 so the component are not needed.

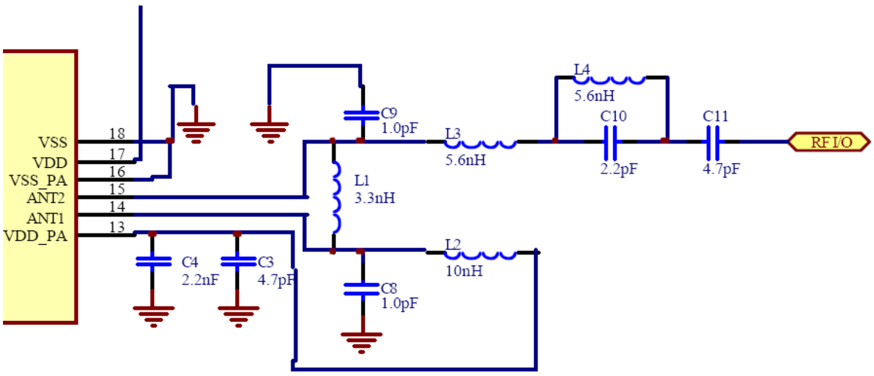

Figure 3- nRF2401 matching network

The same applies for the nRF24Z1 matching network, see Figure 4. The component values are not equal for the different devices. This is because the radio front end is different and different load impedance is needed.

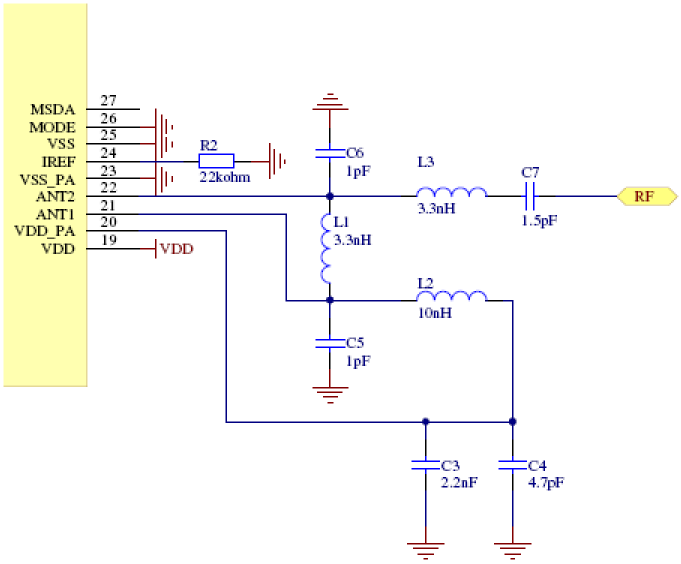

Figure 4- nRF24Z1 matching network

5. Summary

If the performance of the radio module is not as expected, the performance of the matching network must be measured. A decreased performance has several reasons: changed layout of the matching network, different brand/type inductors and so.

The matching network has several roles, like impedance transformation and harmonic rejection, and a number of measurement must be done to be sure the radio complied with the radio regulations.

- Always start with the component values from the recommended layout, found in the datasheet.
- Do the initial measurement to evaluate the performance.
- Follow the flowchart in Figure 2.
- Fill in all the measured values in Table 1.
- Evaluate the changes in performance.
- Change the component values if necessary.
- Always use the same component brand/type as in production.

	Component values						Measured values [dBm]								
Series	**L1**	**L2**	**L3**	**C4**	**C5**	**C6**	**RX LO**	**TX ch0**	**TX ch40**	**TX ch80**	**2.harm**	**3.Harm**	**4. Harm**	**5. Harm**	
1															
2															
3															
4															
5															
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Table 1- Measuring series

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White paper. Revision date: 2006-07-10.

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Crystal Oscillator Design Considerations

White Paper

November 2008

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1 Introduction

An accurate crystal oscillator is needed in all RF designs. This document presents:

- the main RF parameters affected by the crystal oscillator accuracy.
- how to measure crystal oscillator accuracy.
- how to correctly design the crystal circuitry of nRF devices.

2 Background

All RF devices have a maximum offset in RF frequency that they can tolerate while still performing according to the device specification. All RF receivers have a limited frequency window where they can receive incoming RF signals with a minimum power equal to the specified receiver's sensitivity limit. You can find a device's sensitivity limit in its product specification. The limited frequency window is decided by the bandwidth of the filters in the receiver and is called the RX channel bandwidth as shown in Figure 1. on page 4. This bandwidth must be limited to reduce the impact of background noise and to separate the wanted RF signal from unwanted power received from other transmitters operating on frequencies that are in close proximity (called RX selectivity).

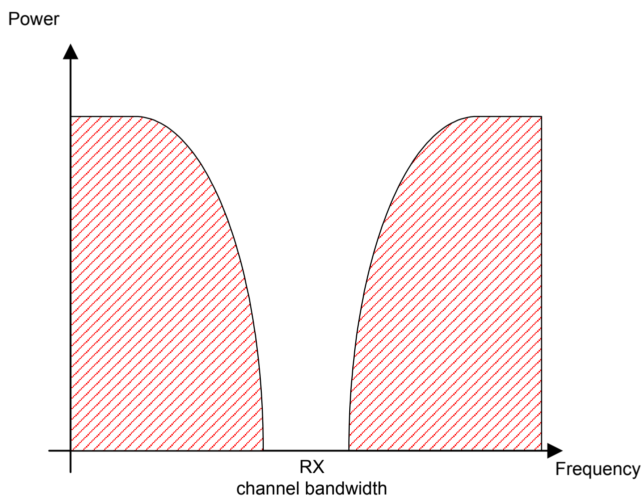

Figure 1. Receiver sensitivity vs. frequency

Incoming RF signals that fall inside the receive bandwidth will pass through the receiver. RF signals outside the RX bandwidth (shaded area in the 'Figure 1. Receiver sensitivity vs. frequency') will be filtered out. It is essential in all RF systems that the incoming RF signal from the transmitter is found within the receive bandwidth of the receiver. This is so that it can utilize the full sensitivity of the receiver. Any offset that brings the TX spectrum outside the RX bandwidth results in the receiver filtering out the incoming signal as noise, see Figure 2. on page 5.

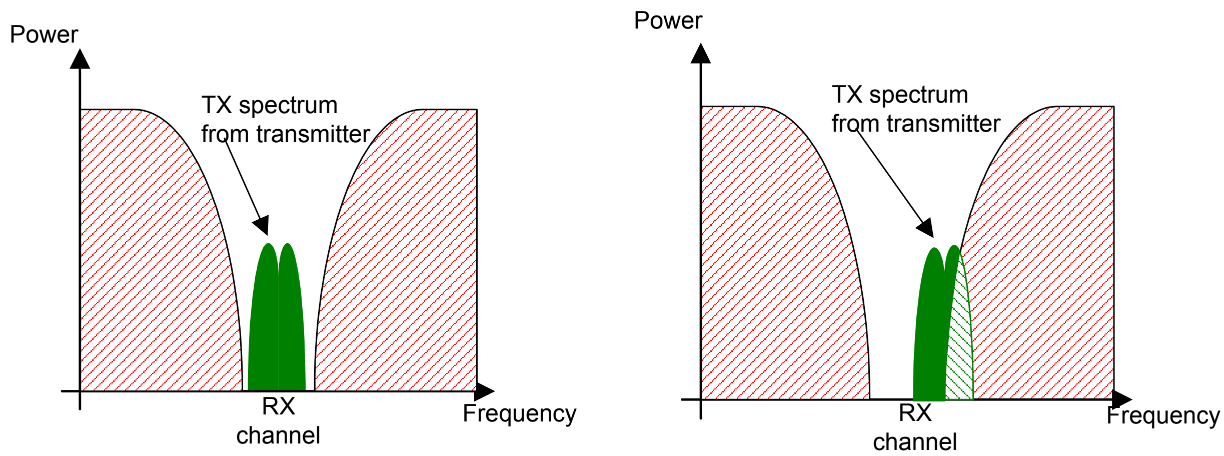

Figure 2. TX frequencies outside the RX BW will be filtered out

The accuracy of the crystal reference is the only deciding factor of this frequency accuracy on all nRF devices. Any offset in crystal frequency is directly mirrored in offset in TX frequency and RX windows. As shown in Figure 2. on page 5, if the offset between the crystal in a TX and RX unit becomes too large, the signal from the transmitter will be filtered out by the receiver. In the system it will look like the receiver has lower sensitivity than specified in the device's product specification and this cuts the range of the application.

The frequency accuracy needed to ensure a good match between transmitted frequency and receiver window position is mirrored directly to a crystal accuracy specification for each nRF device. This accuracy is given as a maximum part per million (ppm) offset requirement and is a standard part of any crystal specification.

Note: The crystal accuracy specification for nRF products is limiting the SUM of:

1. Absolute accuracy at 25 degree Celsius.
2. Temperature drift.
3. Aging.

These three parameters are usually separated in crystal data sheets.

3 Measuring crystal oscillator accuracy

Measuring the accuracy of a fixed TX carrier wave transmitted from a nRF device will verify crystal oscillator accuracy directly since the crystal frequency is the only variable. Measuring the RX window of a nRF device is a time consuming task, but by measuring the TX carrier wave frequency accuracy the RX window follows by design. You should also measure TX carrier accuracy on units that are going to be receivers only.

The crystal oscillator accuracy is OK if:

$$\frac{|F_0 - F_{DUT}|}{F_{DUT}} \leq M \bullet 10^{-6}$$

Variable	Description
F_0	Measured TX carrier frequency
F_{DUT}	The frequency Device Under Test (DUT) is programmed to send (EX: 2.403 GHz)
M	Frequency accuracy of crystal at room temperature in ppm (Ex: 20)

Table 1. Frequency accuracy parameters

Note: M must be the accuracy of the crystal at room temperature and not the total crystal accuracy requirement of the nRF device. This is because this measurement is conducted at room temperature and crystal frequency offset due to temperature drift should not be present.

Please refer to the Nordic Semiconductor white paper ‘nRF TX measurements’ for procedures on how to measure TX frequency accuracy on nRF devices.

If the TX carrier wave frequency is found to be outside the specified window, the first parameter to check is the crystal specification. If the crystal accuracy is not according to the Nordic Semiconductor specifications, it must be replaced by one that is.

4 Crystal oscillator design

Another issue that causes offsets in the crystal oscillator frequency is wrong capacitive loading of the crystal. All nRF devices utilize parallel load compensated crystals and all such crystals are specified for operation with a specific external capacitive load (C_L).

The crystal will NOT oscillate on the frequency specified in the crystal data sheet if the passive crystal load circuitry does not provide the C_L that is specified for the crystal.

Too low capacitive load will result in a crystal oscillator frequency higher than specified, or too high a load results in a lower oscillation frequency. This frequency offset will be translated directly to an offset in the operating frequency of the nRF device.

4.1 Choosing C_L value

When buying crystals you can usually choose from a range of capacitive loads (ex: 8-20pF). When ordering a crystal you must choose a C_L (ex: 12 pF) that is within the range of the C_L specified for the nRF device. The C_L range that nRF devices can operate in is only limited by the maximum C_L (ex: C_L < 16 pF). This maximum C_L depends on the on-chip crystal oscillator design and is specified in the device data sheet.

Choosing a low value C_L gives a lower current consumption in the on-chip crystal oscillator because of the lower load. Crystal oscillator current consumption is the dominant parameter in device stand by modes, where the device operation is suspended but the crystal oscillator is kept active. A low value on C_L should be your preferred choice.

However, the effects of errors in the C_L that you load the crystal with will be more severe on the crystals. To avoid any please read section 4.2 carefully.

4.2 The crystal circuitry

To see what influences the capacitive crystal load has, we must study the crystal circuitry. 'Figure 3. nRF crystal circuit diagram' shows the crystal circuit diagram as you will find it in Nordic Semiconductor documentation. The crystal X1 is loaded with capacitors C1 and C2 and the resistor R1 provides bias for the crystal oscillator found inside the nRF device. R1 is not critical and is usually set to a large value, for example, 1 Mohm.

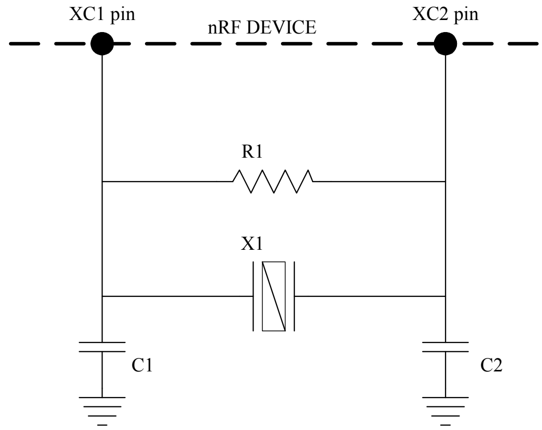

Figure 3. nRF crystal circuit diagram

To see how the capacitive load of the crystal is realized, you must take a look at the small signal equivalent of the circuit in 'Figure 3. nRF crystal circuit diagram'; this is shown in 'Figure 4. Small signal equivalent crystal circuitry'. In this representation you can see that C1 and C2 are found in series and form the capacitive load (C_L) parallel to the crystal. Consequently, we have the term 'parallel load compensated crystal'.

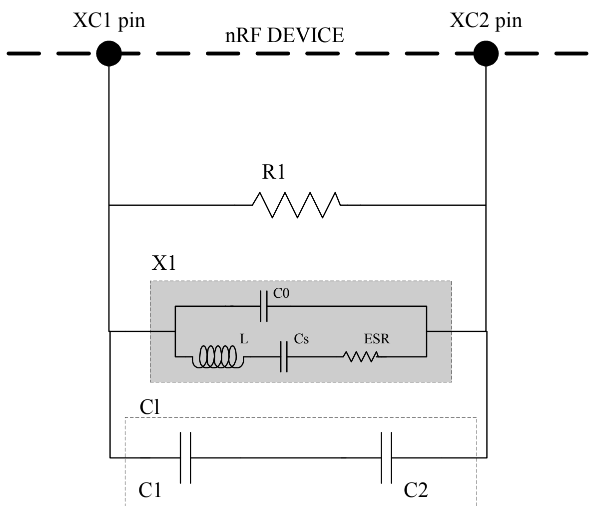

Figure 4. Small signal equivalent crystal circuitry

But this circuit does not present the whole picture. Since the actual C_L seen by the crystal is ALL capacitance outside the crystal pins we must also account for all parasitic capacitive loads found in the crystal circuitry. The crystal circuitry including parasitic capacitances can be seen in 'Figure 5. nRF Crystal circuit diagram including parasitic capacitances'.

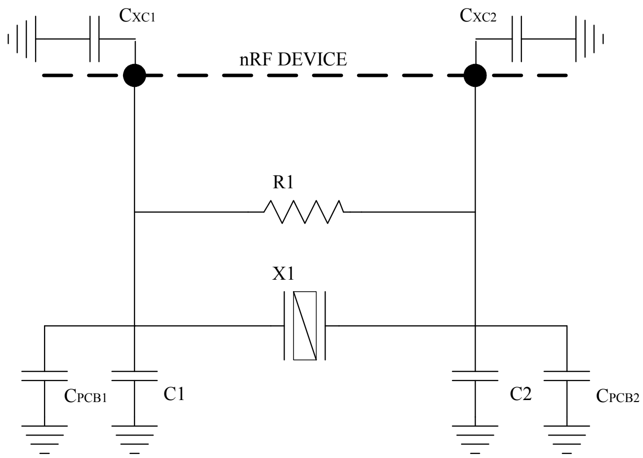

Figure 5. nRF Crystal circuit diagram including parasitic capacitances

In Figure 5. on page 9 the input capacitances of the nRF device crystal oscillator pins are added as C_{XC1} and C_{XC2}. C_{PCB1} and C_{PCB2} represent the parasitic capacitances between the PCB pads of the crystal, the tracks in the crystal circuitry and, the pads of the nRF device to ground.

These parasitic capacitances will be significant in all RF designs because RF layouts need to use ground planes for good performance. Ground will hence surround all tracks and pads on the PCB. When we add these parasitic components to C1 and C2, the total load capacitance on each side of the crystal will increase.

The total load capacitance is now given by:

$$C1' = C1 + C_{XC1} + C_{PCB1}$$
$$C2' = C2 + C_{XC2} + C_{PCB2}$$

Referring to 'Figure 4. Small signal equivalent crystal circuitry', C1' and C2' replace C1 and C2 and form the real C_L:

$$Cl = \frac{C1' \cdot C2'}{C1' + C2'}$$

The layout of any crystal circuitry should be symmetrical (that is, the same capacitive load present on both crystal pins) to ensure stable oscillation. If we set C1'= C2' = C_t the equation simplifies to:

$$Cl = \frac{C_t \cdot C_t}{2 \cdot C_t}$$
$$Cl = \frac{C_t}{2}$$
$$or$$
$$C1' = C2' = 2 \cdot Cl$$

The total capacitive load to ground on each crystal needs to be two times the C_L chosen when selecting the crystal.

The input capacitance of the nRF device crystal pins are equal and ~1 pF. C1 and C2 are off the shelf capacitors. Obtaining a symmetrical crystal load is straightforward as long as you select C1 = C2. The parasitic capacitances found on the PCB must be made as equal as possible through symmetrical PCB routing.

The actual value of the parasitic PCB capacitance may vary from layout to layout and is very difficult to calculate accurately. Fortunately, knowing the exact value of the parasitic PCB capacitance is not needed as you can easily measure the influence of all the parasitic capacitances on the PCB through the TX frequency accuracy measurement.

If you do find an offset outside the frequency accuracy limits, this can be corrected by adjusting the value of C1 and C2. The TX carrier frequency will also be within specification when the total C_L seen by the crystal is according to the crystal specification.

> **Note:** It is very important to do this measurement on several crystals and/or PCB’s (that is, different capacitors) to take variations in crystal and load capacitors into account.

> **Note:** The frequency error that arises from wrong capacitive load on the crystal will be the same on all PCB’s with the same layout, crystal and, load capacitors. A transmitter and a receiver that have identical layout and components fitted will communicate correctly even if the actual RF frequency they operate on differs from the RF frequency they are programmed to operate on.

If you do not pay attention to crystal oscillator accuracy, you may have problems when communicating between a transmitter and a receiver that have differences in crystal, capacitive load or, crystal layout. Now the frequency on the TX and RX may be different. The RF frequency generated by the transmitter will not match the receiver window and you will experience a **loss of range**.

5 Conclusion

As we have seen the frequency accuracy of a nRF design relies solely on the accuracy of the crystal frequency. It is essential to follow the crystal accuracy specification set in each nRF device product specification. The crystal oscillator frequency is not only set by the crystal specification but, the total capacitance the crystal circuitry loads the crystal with must match the specification of the crystal.

Measuring TX frequency accuracy on a prototype will immediately tell you if the crystal oscillator design is correct. If required, the values of the crystal load capacitors (C1 and C2 in this document) must be adjusted to compensate for parasitic capacitive loads in each crystal circuitry layout. Once the frequency accuracy of a prototype is found to be within specified limits, no more tuning is needed in production.

Consequence:

The capacitor values found in the crystal circuitry of Nordic Semiconductor documentation is correct ONLY if:

1. You use a crystal with the same C_L specification as Nordic Semiconductor.
2. The layout, that is, the crystal footprint and PCB routing are equal.

The value of the load capacitors you fit your crystal circuitry MUST match the C_L specification of YOUR selected crystal and complement the parasitic capacitances found in YOUR layout. It must NOT be a blind copy of Nordic Semiconductor schematics.

Antenna tuning

nWP-017

White Paper v1.0

The antenna is an important component in any radio design. This means that it is very important to tune the antenna correctly to achieve the best performance for the radio. If the antenna is not correctly tuned, a lot of transmitter power is lost, the operating range is shortened and, interference problems increase.

This white paper explains how to tune an antenna and subsequently improve radio communication range.

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1 Antenna resonance explained

The capacitance and inductance of an antenna is determined by its physical properties and its environment. The most important parameter for the antenna resonance is the length. A long antenna resonates at a lower frequency than a shorter antenna because the wavelength decreases when the frequency increases. This means the length of the antenna is directly proportional to the frequency and the wavelength.

Since an antenna is designed around a resonant frequency, its bandwidth is limited to that specific resonant frequency. For example, an antenna designed for 900 MHz cannot be used at 2.4 GHz.

The capacitance and inductance can be changed either by changing the length, or by using external components, inductors, or capacitors.

An antenna with a different impedance from the system impedance (an un-tuned antenna) reflects some of the RF power back into the transmitter. The reflected power is lost in the tuning network or the power amplifier. The ratio between the power absorbed by the antenna and the reflected power is called Standing Wave Ratio (SWR). Perfect radiation of the power gives an SWR = 1.0 and increasing numbers means more reflected power. This can be measured with an SWR meter or a network analyzer.

An antenna needs to resonate at the operating frequency to maximize its radiation. The resonance frequency is where the impedance of the inductance X_L equals the impedance of the capacitance X_C. So at the resonance frequency the antenna appears purely resistive. The resistance is a combination of loss resistance and radiation resistance. For an optimized system, the impedance seen into the antenna should match the characteristic impedance (system impedance), which is normally 50 ohm.

2 Why tune the antenna?

The efficiency and radiation pattern of an antenna depends on:

- Size and shape of the antenna element
- The housing
- Proximity to metal
- Shape and size of the ground plane

A single-ended monopole antenna works together with the ground plane to form a dipole antenna. Consequently, the ground plane can be described as the 'second half' of the antenna.

This means that even if the antenna performs well in one application, it will almost never get the same performance in another application. Therefore, a cut-and-paste approach is not recommended. The antenna needs to be adapted to different applications.

An un-tuned antenna has less efficiency than a tuned antenna. Higher efficiency means more radiated power and increased operation range. For optimum performance, the antenna has to be tuned to increase the operation range.

3 Smith chart explained

The Smith chart is a graphical aid that helps calculating transmission lines and antenna tuning circuits. It can be used to represent parameters like impedance, admittance, reflection coefficients, scattering parameters (S_{nn}), noise figure circles, constant gain circles, and regions for unconditional stability. This chapter concentrate on impedance and admittance plots and the reflection coefficient only.

The Smith chart is a plot of complex reflection overlaid with an impedance and/or admittance grid. All impedances (Z) or admittances (Y), real or imaginary, can be plotted in a Smith chart. The impedance is in ohm (Ω) and the admittance in Siemens (S).

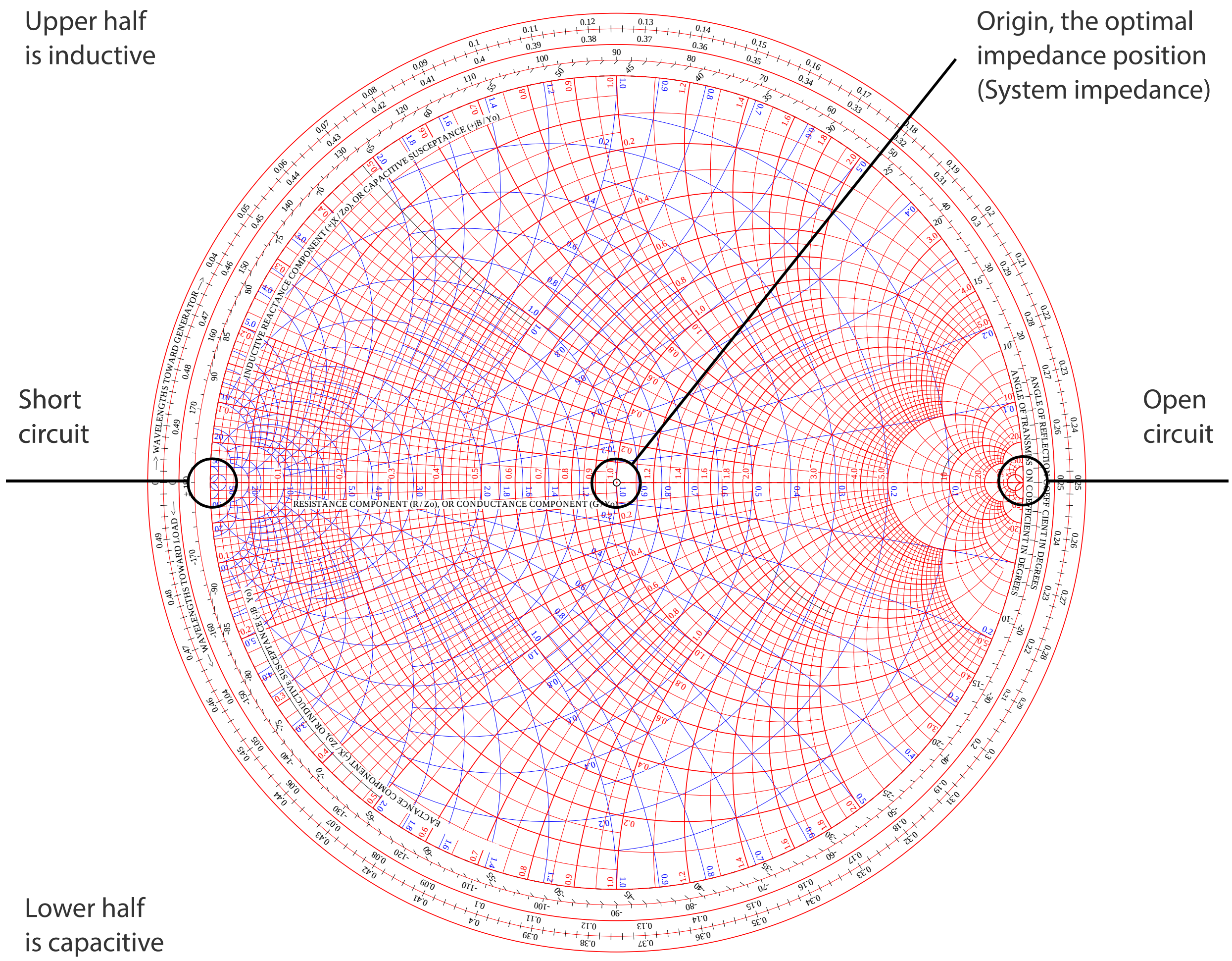


Figure 1 *The Smith chart*

3.1 The Resistance circles and the Reactance arcs

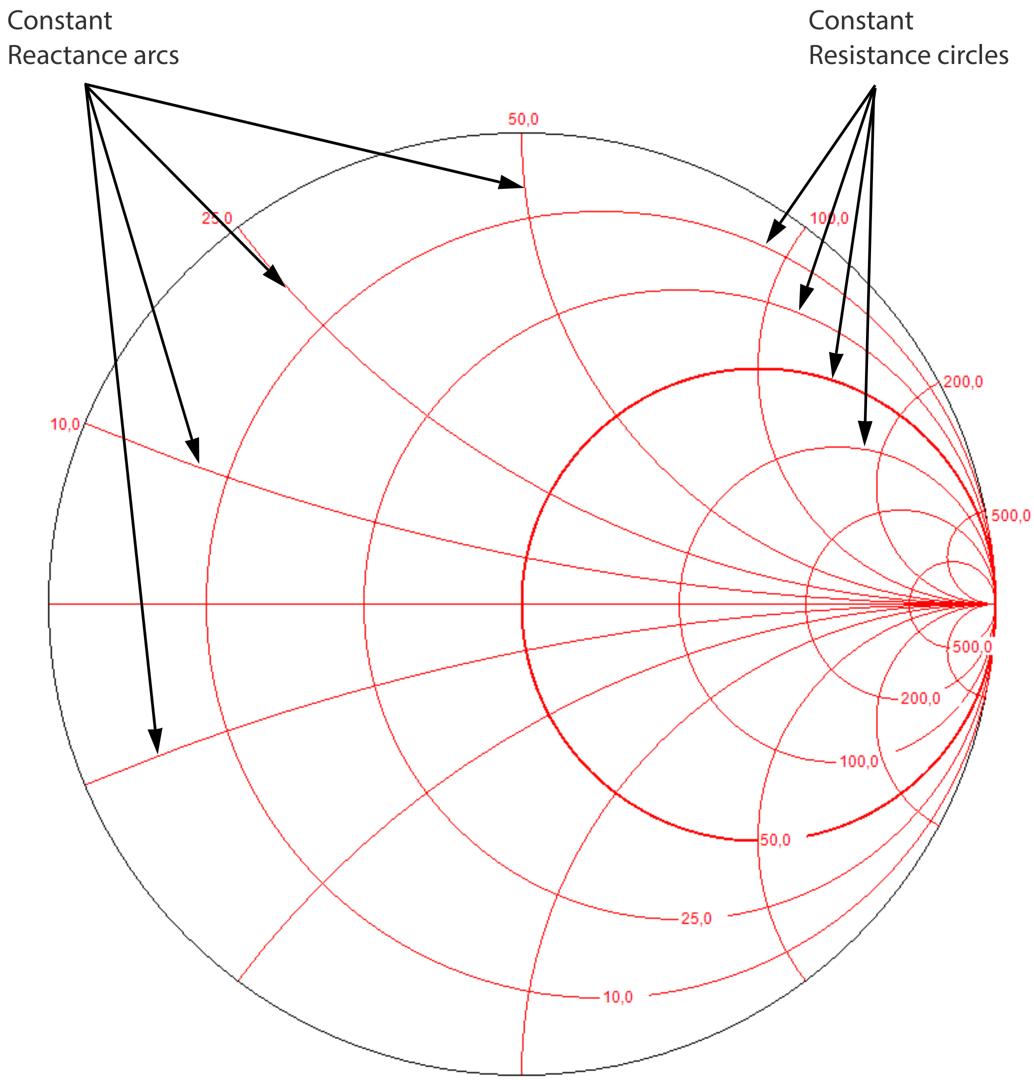


Figure 2 *Impedance plot with Reactance arcs and Resistance circles*

3.2 The Conductance circles and the Susceptance arcs

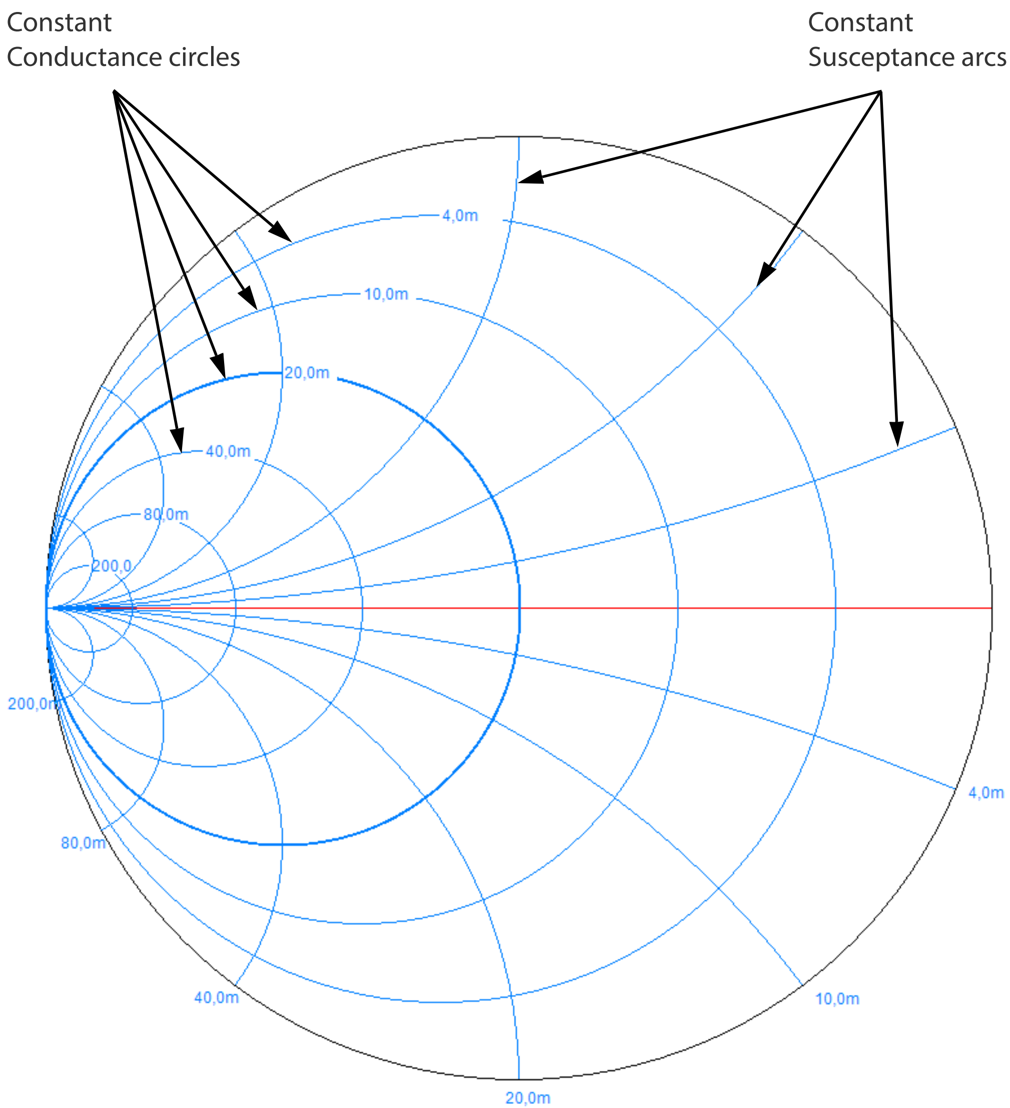


Figure 3 *Admittance plot with Conductance circles and Susceptance arcs*

3.3 Moving the impedance point

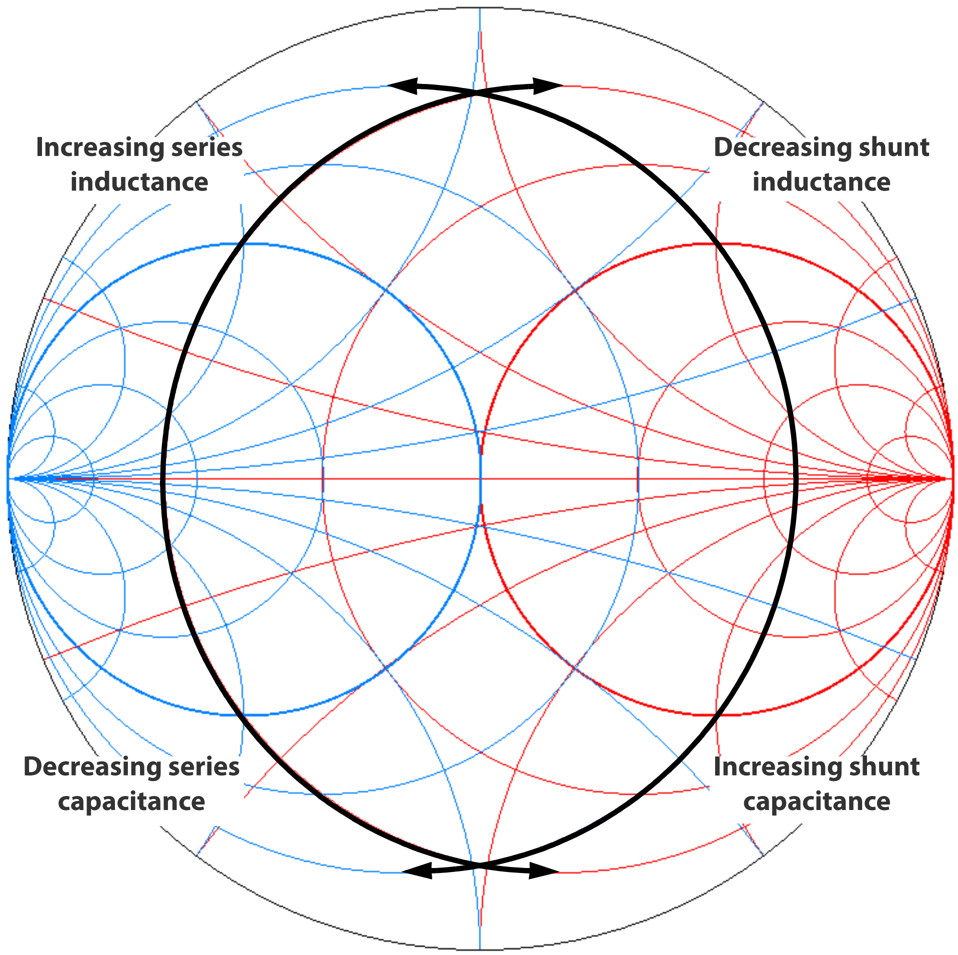


Figure 4 *Moving the impedance point*

You can move the impedance along the constant arcs or circles in the Smith chart in the following way:

- Inserting a reactive series component moves the impedance along the resistance circles
- Inserting a reactive shunt component moves the impedance along the conductance circles
- An inductor moves the impedance upwards
- A capacitor moves the impedance downwards
- A piece of transmission line moves the impedance clockwise around the origin

Impedance equations:

$$\boxed{Z = R + jX}$$

Equation 1 *Impedance represented by resistance (R) and the reactance (X).*

$$\boxed{X_C = \frac{1}{j\omega C}}$$

Equation 2 *Capacitor impedance (where ω is 2πf)*

$$\boxed{X_L = j\omega L}$$

Equation 3 *Inductor impedance*

The impedance is the reciprocal of the admittance:

$$\boxed{Z = \frac{1}{Y} = \frac{1}{G + jB} = \left(\frac{G}{G^2 + B^2}\right) + j\left(\frac{-B}{G^2 + B^2}\right) = R + jX}$$

Equation 4 *Impedance*

Admittance equations:

$$\boxed{Y = G + jB}$$

Equation 5 *Admittance represented by the conductance (G) and susceptance (B):*

The admittance is the reciprocal of the impedance:

$$\boxed{Y = \frac{1}{Z} = \frac{1}{R + jX} = \left(\frac{R}{R^2 + X^2}\right) + j\left(\frac{-X}{R^2 + X^2}\right) = G + jB}$$

Equation 6 *Admittance*

$$\boxed{B_C = j\omega C}$$

Equation 7 *Capacitor admittance*

$$\boxed{B_L = \frac{1}{j\omega L}}$$

Equation 8 *Inductor admittance*

The Smith chart is referenced to a characteristic impedance of 1, called normalization. The Smith chart can be used with any characteristic impedance, but is mostly used in a 50 ohm system. Any impedance, Z, may be normalized by dividing it by the characteristic impedance, so the normalized impedance z is given by:

$$\boxed{z = \frac{Z}{Z_0}}$$

Equation 9 *Normalized impedance*

Similarly, for normalized admittance y:

$$\boxed{y = \frac{Y}{Y_0}}$$

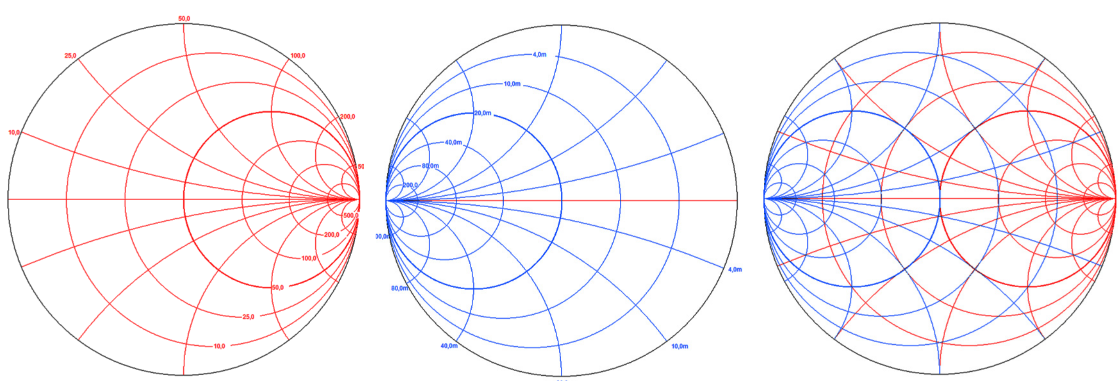
Equation 10 *Normalized admittance*

> **Note:** For a series-connected circuit, the impedances can be added, and for a shunt-connected circuit, the admittances can be added.

When working with a series-connected circuit or inserting elements in series with an existing circuit or transmission line, the resistance and reactance components are easily manipulated on the 'impedance' Smith chart.

Similarly, when working with a parallel-connected circuit or inserting elements in parallel with an existing circuit or transmission line, the conductance and susceptance components are easily manipulated on the 'admittance' Smith chart.

Normally, the Smith chart has both the impedance lines and the admittance lines in the same chart, see ***Figure 5*** below.



Impedance lines　　Admittance lines　　Impedance and admittance lines combined

Figure 5 *Impedance and admittance lines*

4 Methods of tuning an antenna

There are the following two methods to tune an antenna:

- If the physical dimensions of the antenna can be altered, for example, with a PCB antenna, adjusting the length will be one part of the tuning. Another part is to add a component, inductor, or capacitor, to pull the antenna impedance towards the 50 ohm center point.
- If the antenna cannot be altered physically, more external components must be used to tune the antenna. These external components are called the matching network.

4.1 Length adjustment and a single component

The antenna length must be adjusted so the impedance seen into the antenna falls somewhere on the conductance circle that passes through the origin.

Then a component needs to be added to get the impedance moved to the center. If the initial impedance is on the upper half, add a shunt capacitor, see ***Figure 7***. If the initial impedance is on the lower half, add a shunt inductor, see ***Figure 8***.

For convenience, a shunt capacitor is preferred. A shunt can be removed without changing the rest of the circuit and a capacitor is cheaper than an inductor.

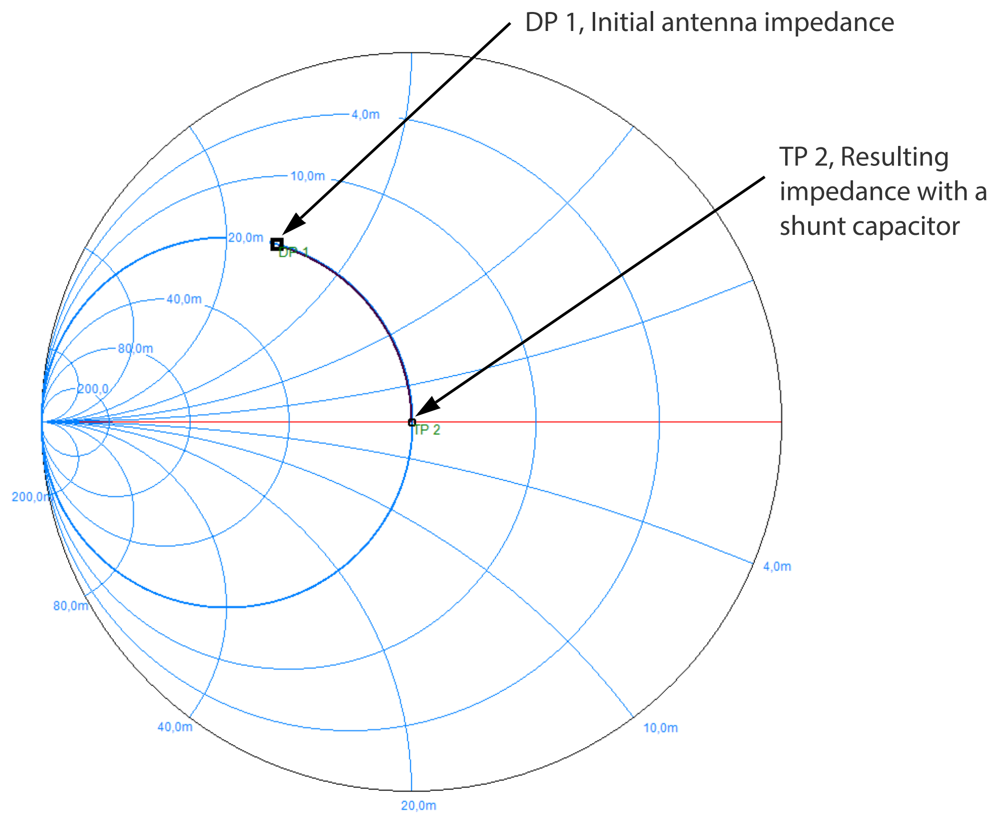


Figure 6 *Shunt capacitor*

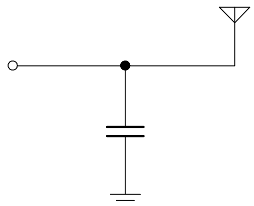


Figure 7 *Shunt capacitor*

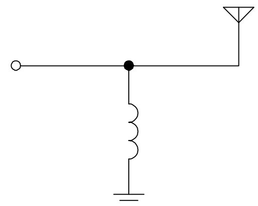


Figure 8 *Shunt inductor*

If the antenna impedance is on the same circle, but on the bottom half of the Smith chart, a shunt inductor pulls the impedance upwards.

If a series component is preferred, the same applies, but here the impedance is moved along the constant resistance circles, as in ***Figure 9***.

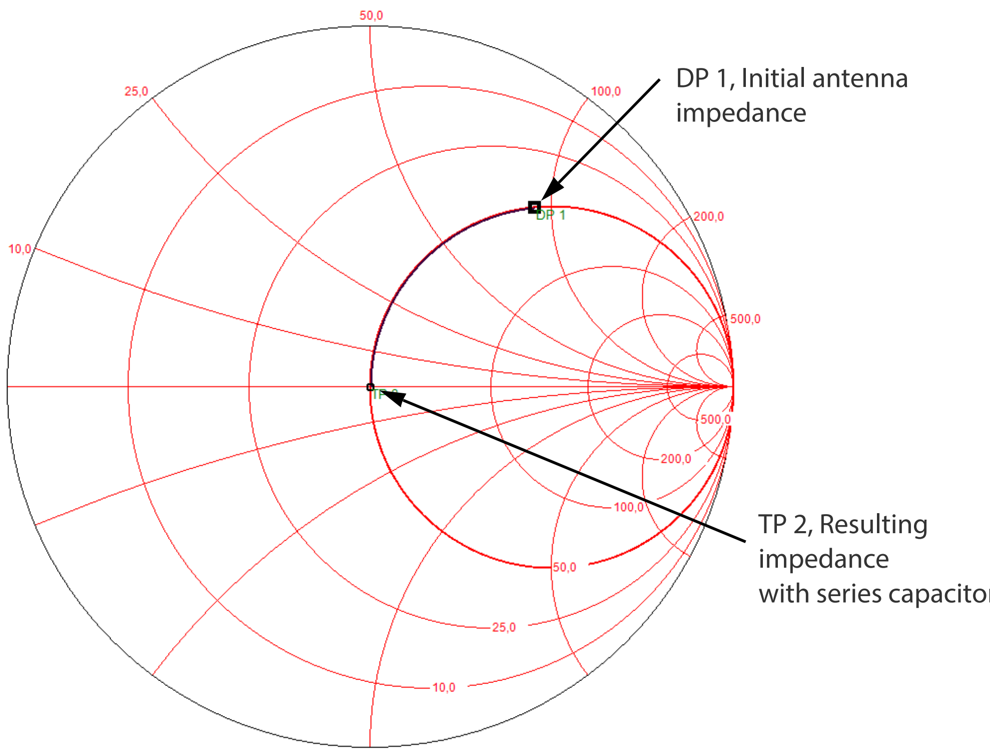


Figure 9 *Series capacitor*

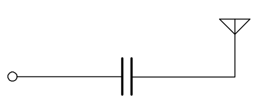


Figure 10 *Series capacitor*

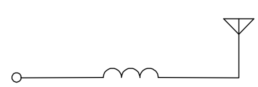


Figure 11 *Series inductor*

4.2 Pi-network

Adding a pi-network is a very flexible way to tune an antenna. It gives the possibility to change the antenna impedance from any value to 50 ohm. The pi-network consists of three components in a pi-configuration. First, one shunt, then one in series, and finally another shunt, see ***Figure 12***.

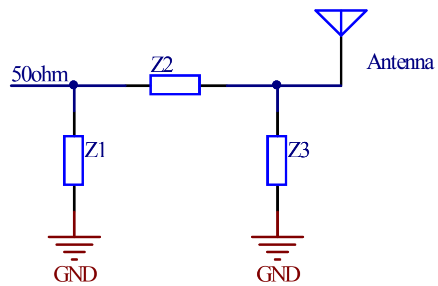


Figure 12 *Pi-network*

At least two components are needed to get the antenna impedance to 50 ohm. This depends on where in the Smith chart the initial antenna impedance is placed. ***Figure 13*** to ***Figure 16*** shows the different configurations with inductors and capacitors. The shaded area is the forbidden region for the shown configuration. In the forbidden region the shown configuration cannot transform the antenna impedance to 50 ohm.
Z_S is seen into the source and Z_L is seen into the antenna.

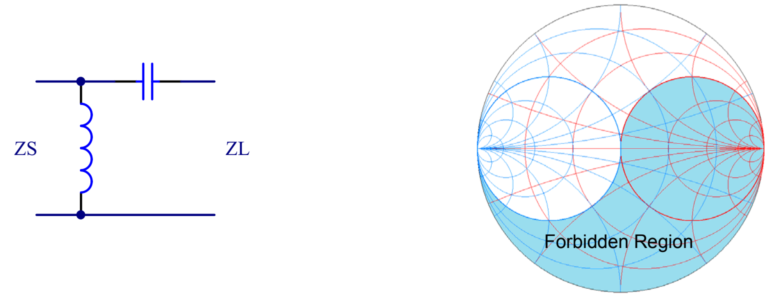


Figure 13 Network configuration with corresponding forbidden region

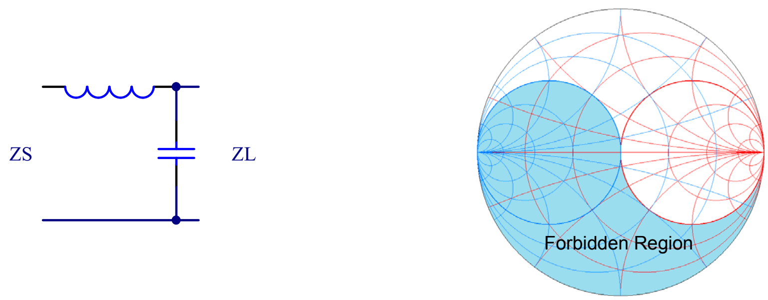


Figure 14 Network configuration with corresponding forbidden region

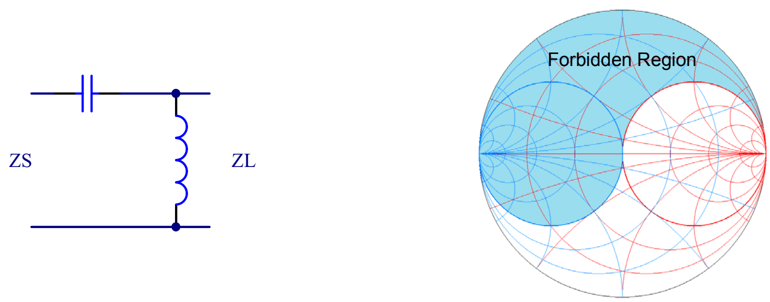


Figure 15 Network configuration with corresponding forbidden region

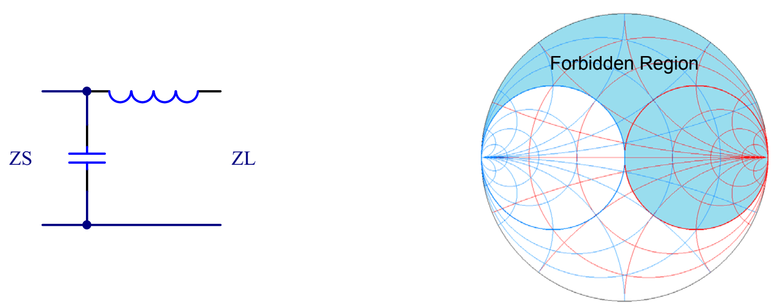


Figure 16 Network configuration with corresponding forbidden region

5 Calculating the component values

This chapter describes how to calculate the values on the capacitors and/or inductors needed for antenna tuning.

The examples in ***Section 5.1***, ***section 5.2 on page 17***, and ***section 5.3 on page 18*** explain how to calculate the values.

5.1 Adding a series capacitor

Adding a series capacitor (or inductor) only changes the imaginary part of the impedance, the reactance.

For example, the initial impedance is measured on a network analyzer to be 50+j100 ohm, see ***Figure 17***. This is the point where the 50 ohm resistance circle meets the 100 ohm reactance arc.

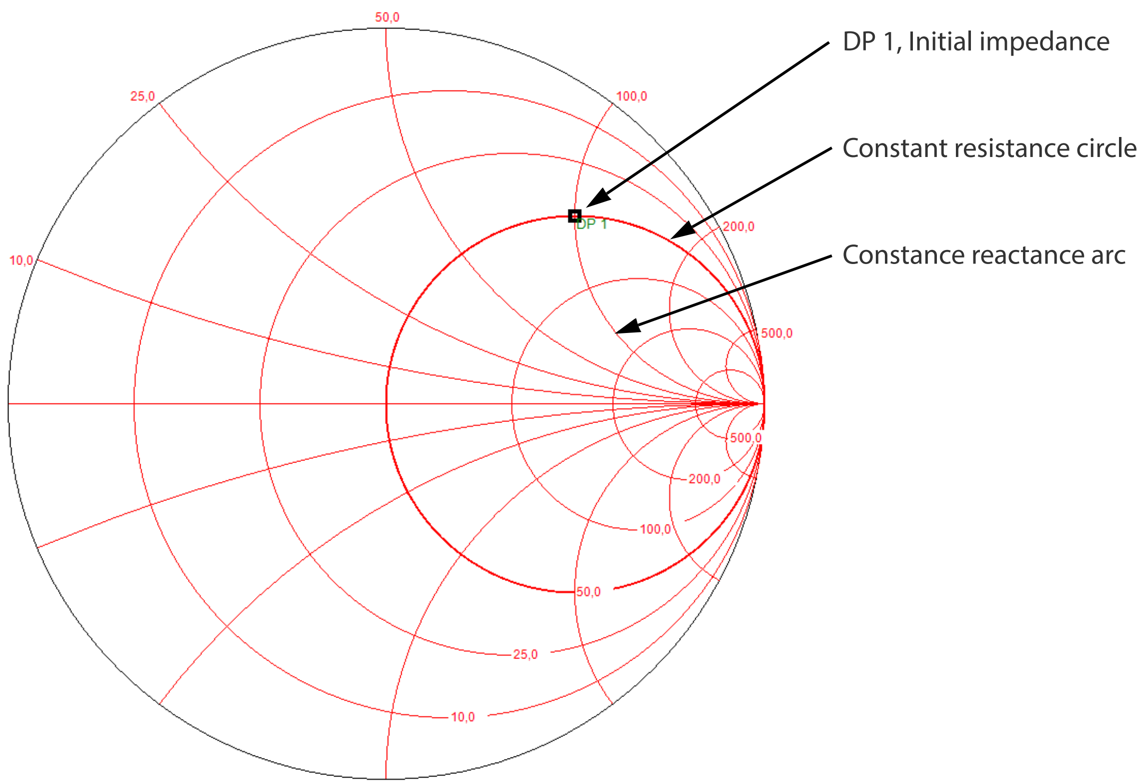


Figure 17 *Initial impedance*

To align the impedance to the 50 ohm point, a series capacitor is needed. The capacitor moves the impedance down the resistance circle. A capacitor that adds -j100 ohm is then needed to cancel the reactance. If the frequency is 2.4 GHz, the value of the capacitor becomes:

$$\boxed{X_C = \frac{1}{j2\pi f C}}$$

Equation 11 *Capacitor impedance*

And by rearranging Equation 11, the capacitor value will be:

$$\boxed{C = \frac{1}{j2\pi f X_C} = \frac{1}{j2 \cdot \pi \cdot 2.4\ \text{GHz} \cdot -j100} = 0.66\ \text{pF}}$$

Equation 12 *Capacitor value*

Note that $\boxed{j \cdot -j = 1}$

Figure 18 shows the impedance after the series capacitor is added.

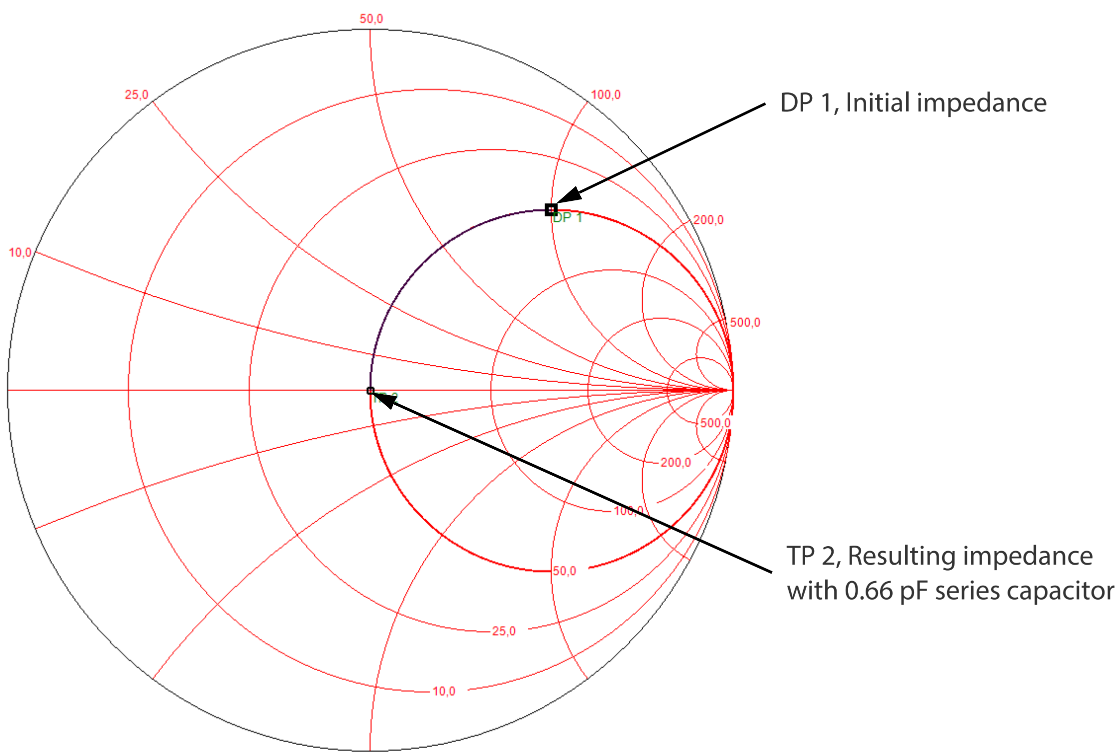


Figure 18 *Adding a series capacitor*

5.2 Adding a shunt capacitor

Adding a shunt capacitor (or inductor) only changes the imaginary part of the admittance, the susceptance.

For example, the initial impedance of the antenna is measured to be 10+j20 ohm. Adding a shunt capacitor moves the impedance along the conductance circles in the admittance part of the Smith chart.

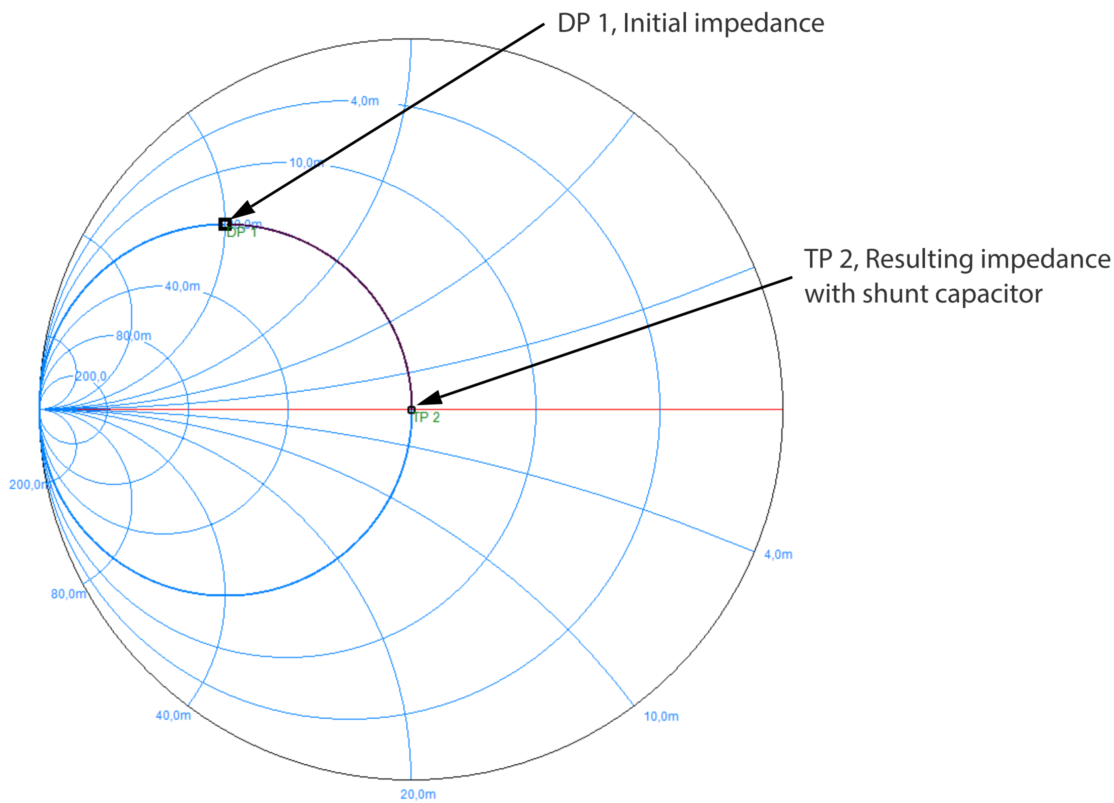


Figure 19 *Adding a shunt capacitor*

From ***Equation 6***, the admittance of the impedance 10+j20 ohm is calculated to be:

$$Y = \frac{1}{Z} = \frac{1}{R+jX} = \left(\frac{R}{R^2+X^2}\right) + j\left(\frac{-X}{R^2+X^2}\right) = \left(\frac{10}{10^2+20^2}\right) + j\left(\frac{-20}{10^2+20^2}\right) = 0.02 - j0.04S$$

Equation 13 *Calculation of the admittance of the impedance*

The admittance point can also be found directly on the network analyzer.

The capacitor moves the admittance along the susceptance line so it needs to add j0.04 S to reach the 50 ohm point.

To calculate the capacitor value, use the equation from ***Equation 7***:

$$B_C = j\omega C \Rightarrow C = \frac{B_C}{j\omega} = \frac{j0.04}{j \cdot 2 \cdot \pi \cdot 2.4\ GHz} = 2.65\ pF$$

Equation 14 *Calculating the capacitor value*

This means the shunt capacitor should be 2.65 pF to get 50 ohm impedance.

5.3 Using more than one component

In this example, the initial antenna impedance is measured to be 13.6-j29.3 ohm. The normalized impedance is 0.272-j0.586 ohm. The Smith chart in ***Figure 20*** shows that the antenna is capacitive.

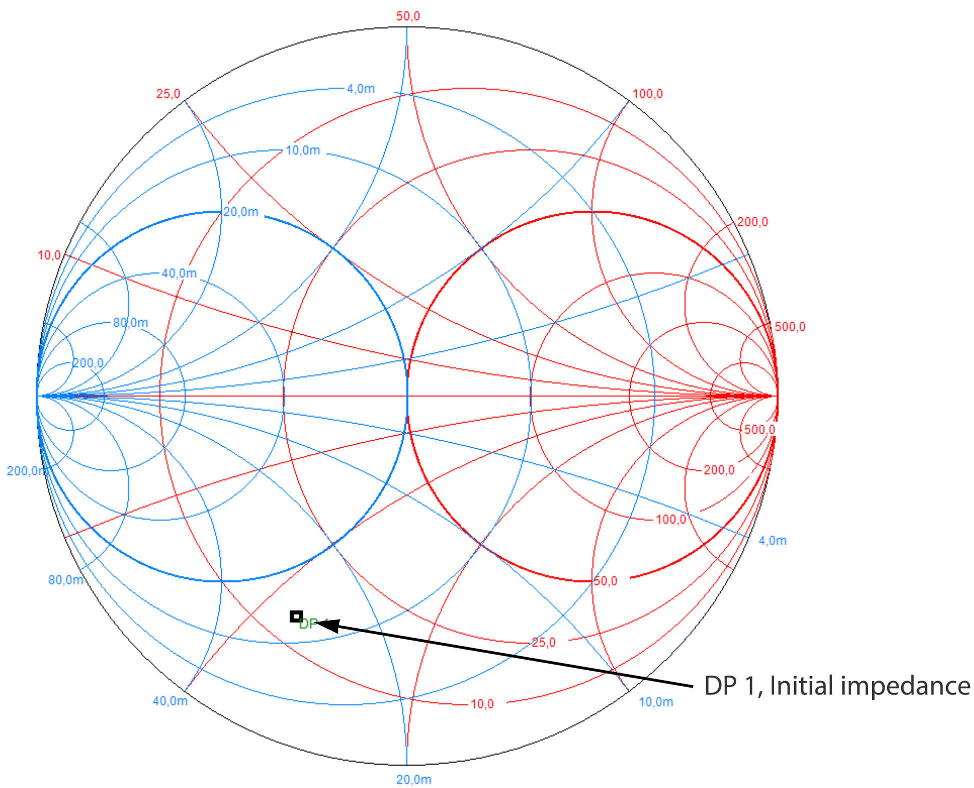


Figure 20 *Adding more than one component, initial impedance*

A shunt inductor moves the impedance up along the conductance circles in the admittance chart up to TP 2 in ***Figure 21***. TP 2 has an impedance of 1.0+j0.7 ohm normalized or 50+j35 ohm.

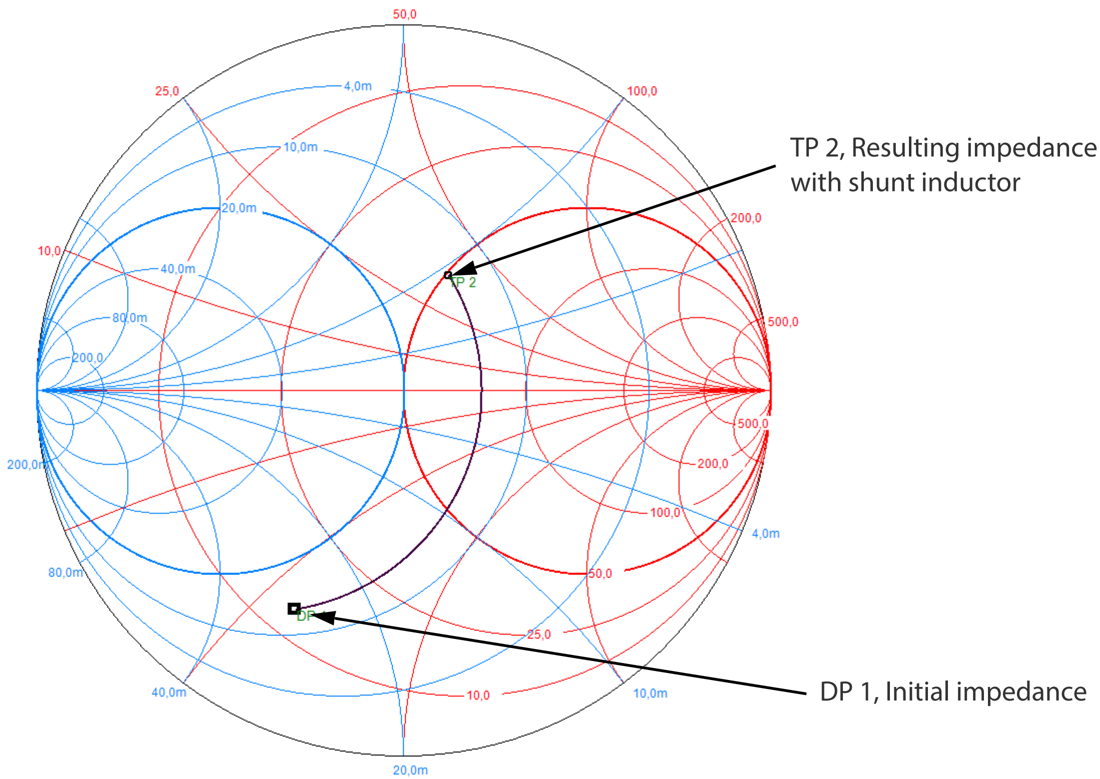


Figure 21 *Adding more than one component, first step: shunt inductor added*

Now a series capacitor moves the impedance down along the resistance circles in the impedance chart towards TP 3 in ***Figure 22***.

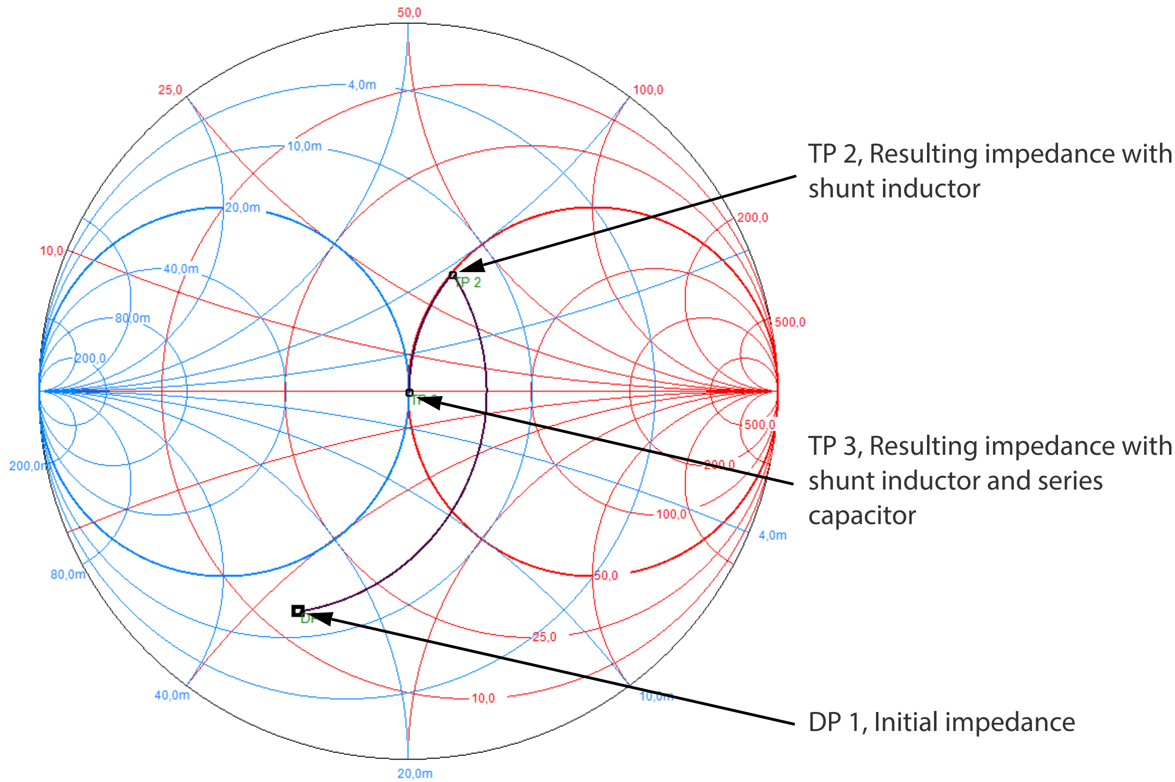


Figure 22 *Adding more than one component, second step: series capacitor added*

To find the value of the components, first find the admittance in DP 1 and TP 2 and calculate the difference in susceptance on the two points.

When moving along the susceptance circles, the conductance is constant. Read out the susceptance value of DP 1 and subtract it from the value found in TP 2. The values can be found by moving along the susceptance lines to the edge.

To find the values in the system impedance (50 ohm) instead of the normalized Smith chart, divide the values by the system impedance.

DP 1 reads 1.4 S in the normalized Smith chart. TP 2 reads -0.475 S.

$$\boxed{B_1 - B_2 = j\frac{1.4 - (-0.475)}{50} = j0.0375S}$$

Equation 15 *Find values in the system impedance*

The inductor value then becomes:

$$B_L = \frac{1}{j\omega L} \Rightarrow L = \frac{1}{j\omega B_L} = \frac{1}{j \cdot 2 \cdot \pi \cdot f \cdot B_L} = \frac{1}{j \cdot 2 \cdot \pi \cdot 2.4GHz \cdot 0.0375} = 1.8nH$$

Equation 16 *Inductor value calculation*

To calculate the capacitor value, use the reactance value at TP 2. The value is j0.75 ohm normalized or j36.5 ohm in the system impedance chart. To get to the 50 ohm point (origin), the capacitor must add -j36.5 ohm.

The capacitor value then becomes:

$$C = \frac{1}{2\pi f X_C} = \frac{1}{2 \cdot \pi \cdot 2.4GHz \cdot 36.5} = 1.8pF$$

Equation 17 *Capacitor value calculation*

The resulting pi-network has the following configuration:

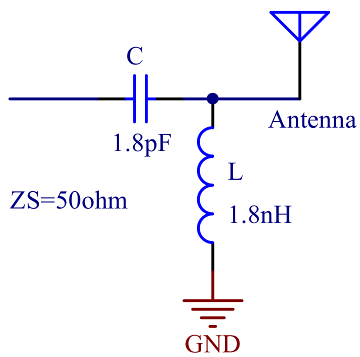


Figure 23 *Pi network configuration*

6 Non-ideal components

Unfortunately, there is no such thing as a perfect component. No component alters only the reactance or susceptance. All real components has parasitic effects caused by the physical housing, the terminals, and how it is mounted on the PCB.

6.1 Capacitors

A capacitor has an equivalent model as shown in ***Figure 24***.

- C is the actual capacitor value
- ESR is the equivalent series resistance
- L is the inductance in the leads or terminals.

In addition to these, there is always some stray capacitance to ground.

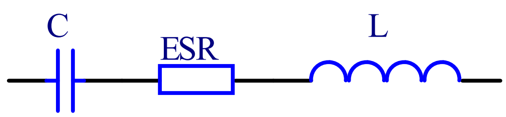


Figure 24 *Capacitor model*

At low frequencies, the capacitor's impedance is just like expected from the specified capacitor value, because the parasitic effects are relatively small. As the frequency increases, the parasitic effects in the component starts to influence the overall impedance. Smaller components have a greater parasitic influence than larger components.

At the self-resonance frequency, the capacitive and inductive impedances cancel each other out leaving only a resistive component. The self-resonance is given by:

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

Equation 18 *Self-resonance*

At frequencies above the self-resonance frequency, the capacitor behaves like an inductor.

6.2 Inductors

An inductor has an equivalent model as shown in ***Figure 25***.

- Rdc is equal to the DC resistance
- L is the inductance
- C is the capacitance between the terminals

The inductor is purely resistive at the self-resonance frequency.

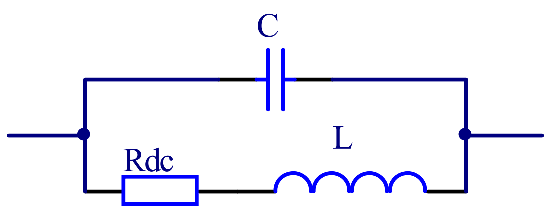


Figure 25 *Inductor model*

If the self-resonance frequency is known, the parasitic components can be calculated:

$$\boxed{C_P = \frac{1}{(2\pi f_0)^2 L_0}}$$

Equation 19 *Parasitic value calculation*

Or

$$\boxed{L_P = \frac{1}{(2\pi f_0)^2 C_0}}$$

Equation 20 *Parasitic value calculation*

This means that when the frequency increases, the real component value will start to deviate from the ideal value. This complicates the calculations of the component values.

A component is specified at a given frequency by the manufacturer. If used at higher frequencies, the component value starts to deviate. ***Figure 26*** shows how an inductor's value increases as the frequency increases towards the resonance frequency.Note that when the inductance starts to deviate, the component-to-component variation also increases, because the SFR of each component has a significant influence. It is therefore always best to choose inductance values that ensure that the inductance value is in its 'constant' range.

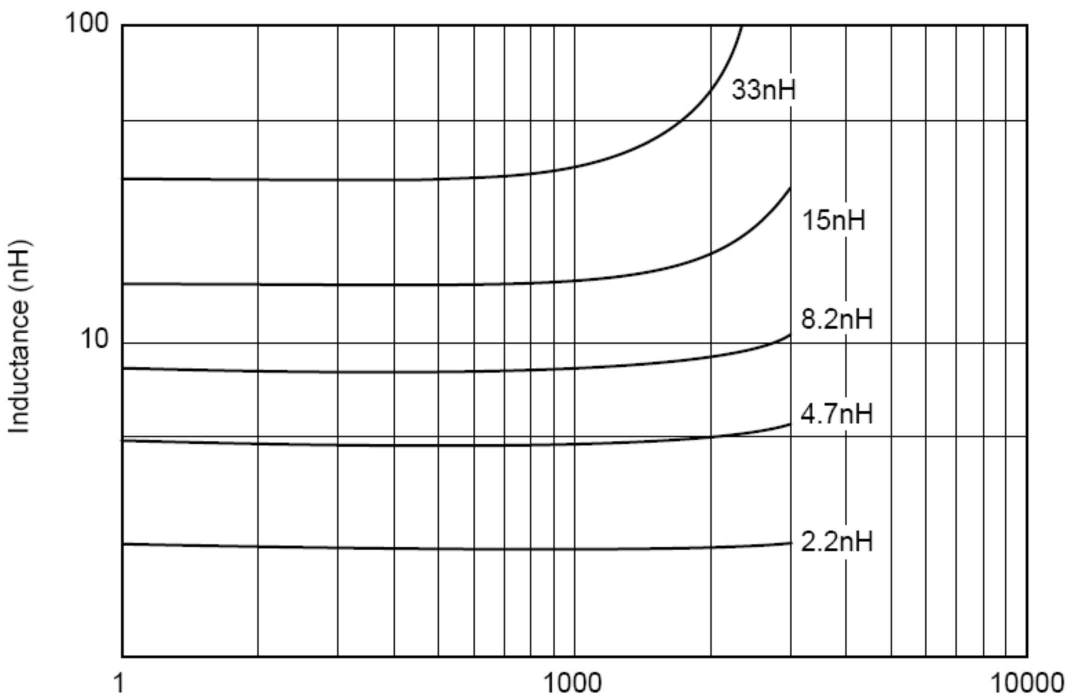


Figure 26 *Inductor values over frequency*

To speed up the calculations of accurate component values when tuning antennas at higher frequencies, RF simulation software can be used. But it is often better and faster to estimate the values using the Smith chart and then fine tune the values by measurements on the actual device.

7 Equipment needed to tune an antenna

For accurate impedance measurement, a Vector Network Analyzer (VNA) must be used. The VNA measures both amplitude and phase, so it will display complex impedance values. The VNA must always be calibrated before use, to compensate for cable length and loss.

Solder a short coaxial cable to the circuit that is to be analyzed. The length of the coaxial cable can be compensated on the VNA. It is important to use short cables and solder the screen to a good ground connection close to where the impedance is to be measured. The center conductor must also be kept as short as possible to avoid adding too much series inductance.

8 Examples

To better explain how the impedance matching is done, some examples are needed. ***Section 8.1*** shows how to match a chip antenna and ***section 8.2 on page 33*** shows how to match a ¼ wave PCB antenna.

8.1 Chip antenna

A chip antenna needs to be tuned with external components. Measure the initial impedance by shorting the series component with a 0 ohm resistor and removing any shunt components. The antenna must be isolated from the radio so it may be best to work on a board with the radio and matching network removed. ***Figure 27*** shows the impedance measured into the chip antenna.

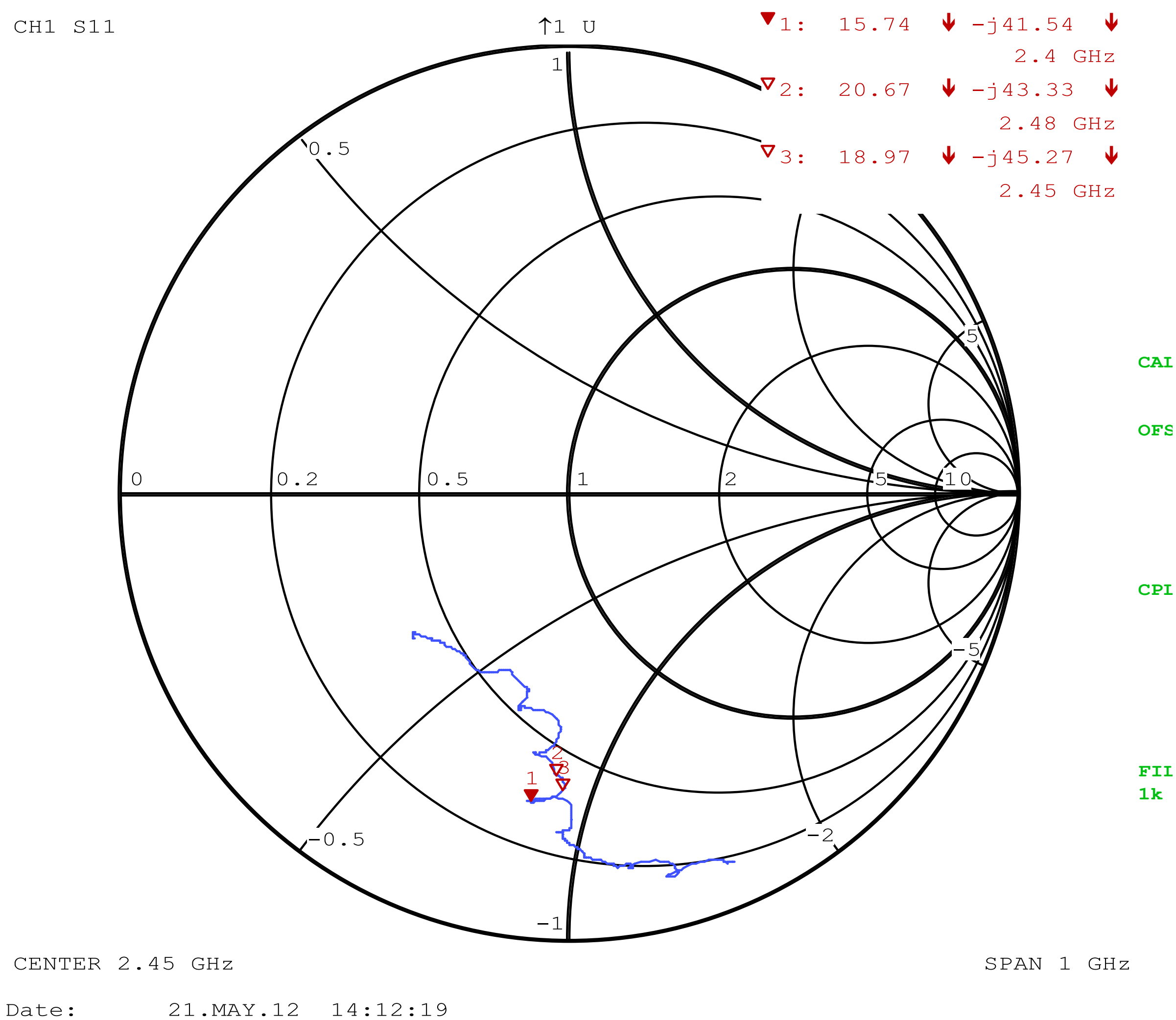


Figure 27 *Initial impedance*

From the Smith chart and the figures (forbidden regions), two topologies can be used:

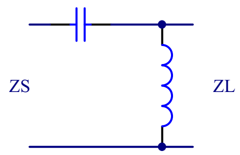


Figure 28 *Option (a)*

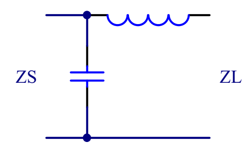


Figure 29 *Option (b)*

Calculated values for the two topologies gives the following schematics:

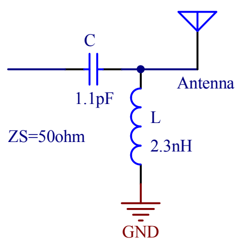


Figure 30 *Calculated values for (a)*

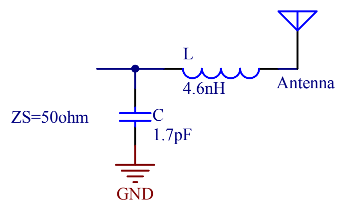


Figure 31 *Calculated values for (b)*

Both (a) and (b) have reasonable component values. Capacitors between 0.5 pF and 5.6 - 10 pF and inductors between 1.0 nH and 10 nH are preferable. The standard value closest to the calculated value must be selected.

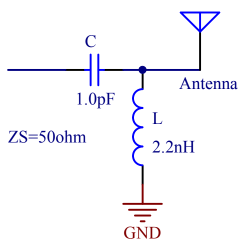


Figure 32 *Standard values for (a)*

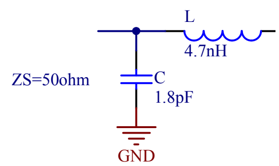


Figure 33 *Standard values for (b)*

This results in the measurements shown in ***Figure 34*** and ***Figure 35***.

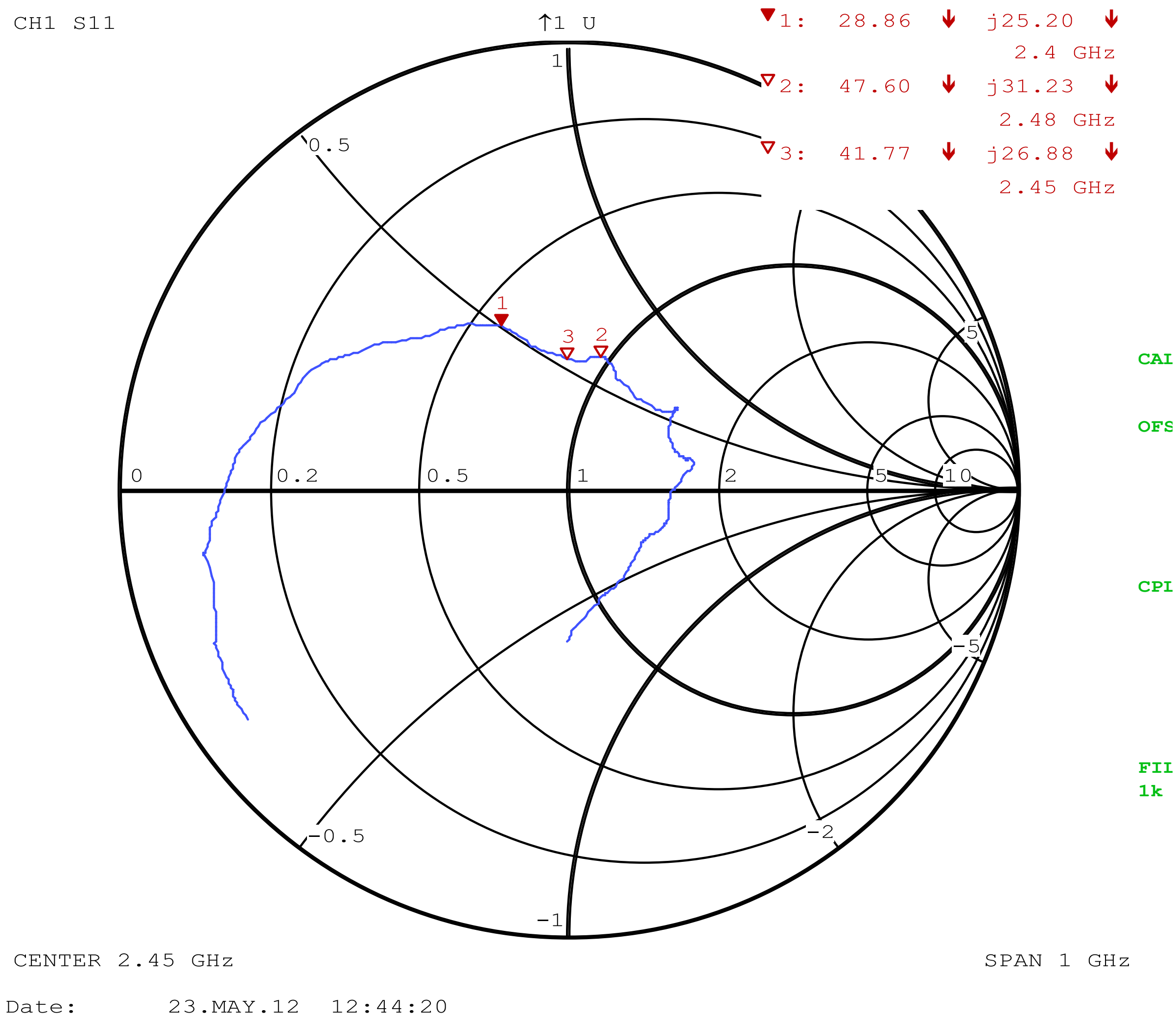


Figure 34 *First measurement option (a)*

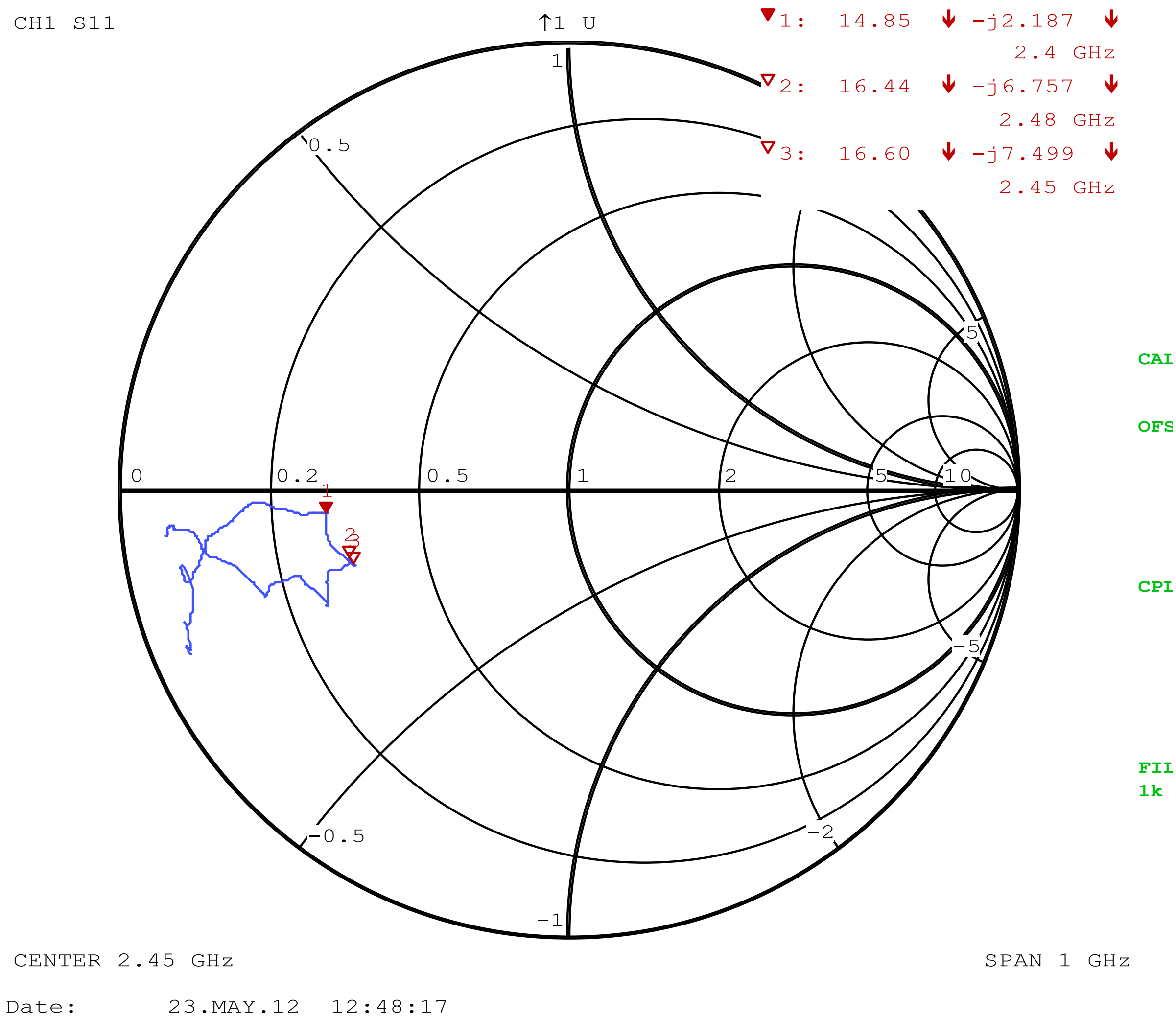


Figure 35 *First measurement option (b)*

The resulting impedance is not in origin so some adjustments have to be done. Capacitors must often be selected smaller than the calculated value because of pad to ground capacitance and capacitance between the pads. The 0 ohm resistor used to measure the initial capacitance equals a 1 nH inductor. This means in the option (b) case, the series inductor must be made larger to compensate for this. The values that give the best performance are given in ***Figure 36*** and ***Figure 37***.

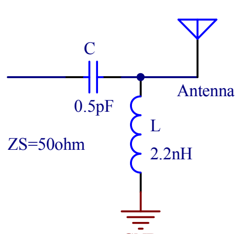


Figure 36 *Option (a)*

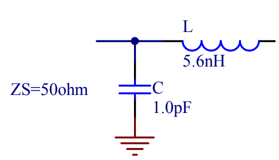


Figure 37 *Option (b)*

Impedance and SWR measurements can be found in ***Figure 38*** to ***Figure 41***. A normal aim is for a SWR equal to or below 2.0. Both option (a) and option (b) satisfy this requirement.

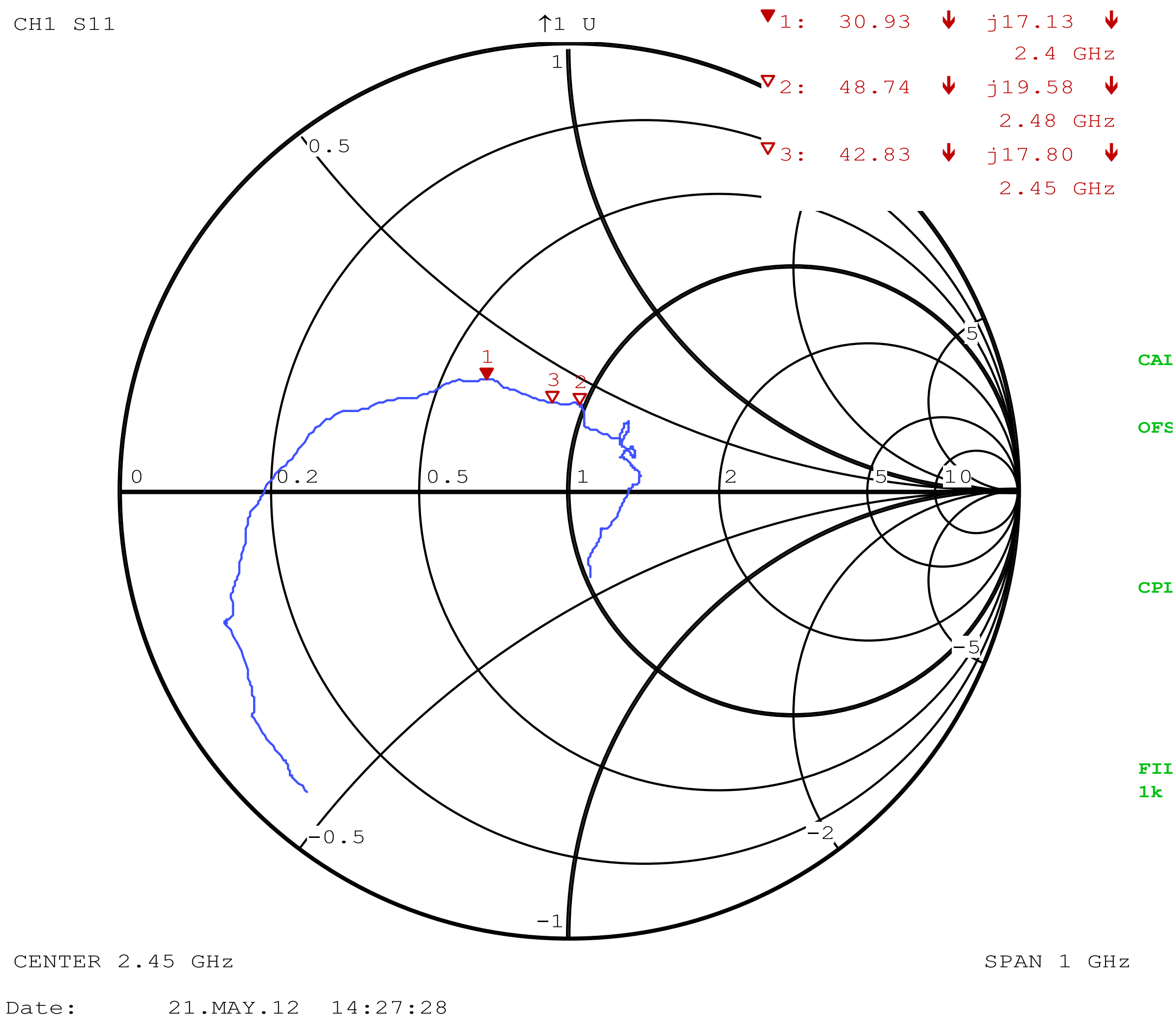


Figure 38 *Impedance after matching option (a)*

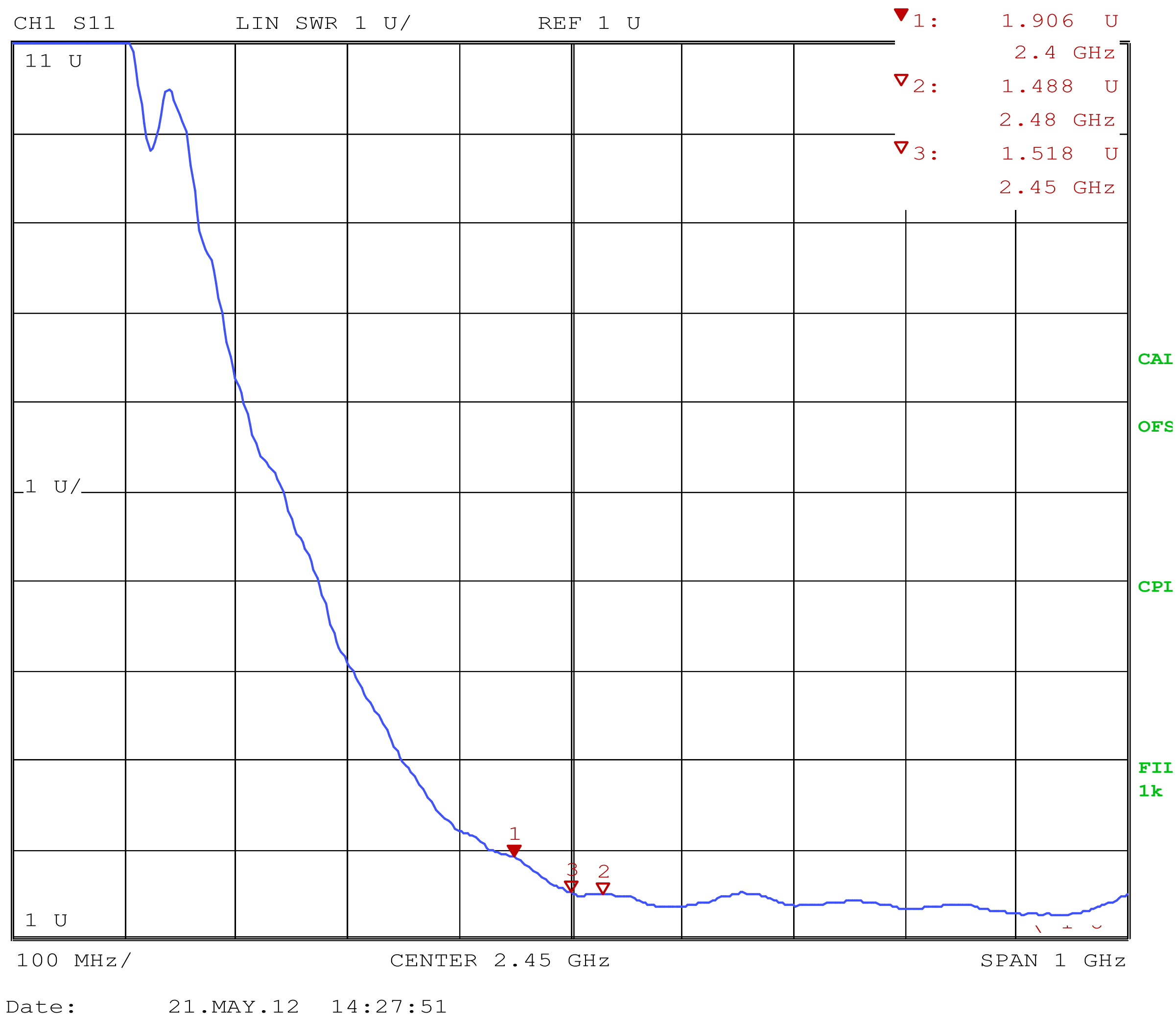


Figure 39 *SWR after matching option (a)*

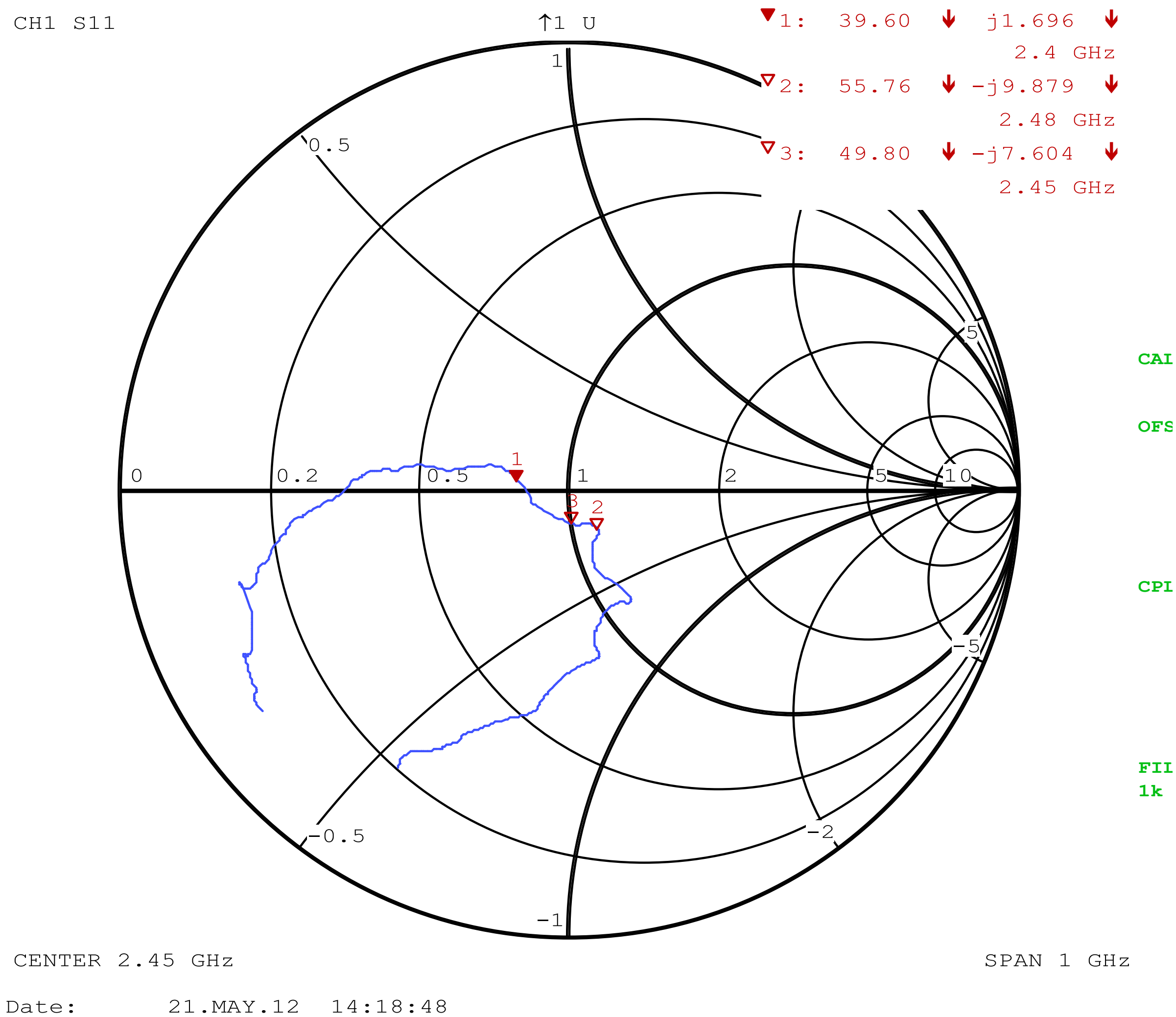


***Figure 40** Impedance after matching option (b)*

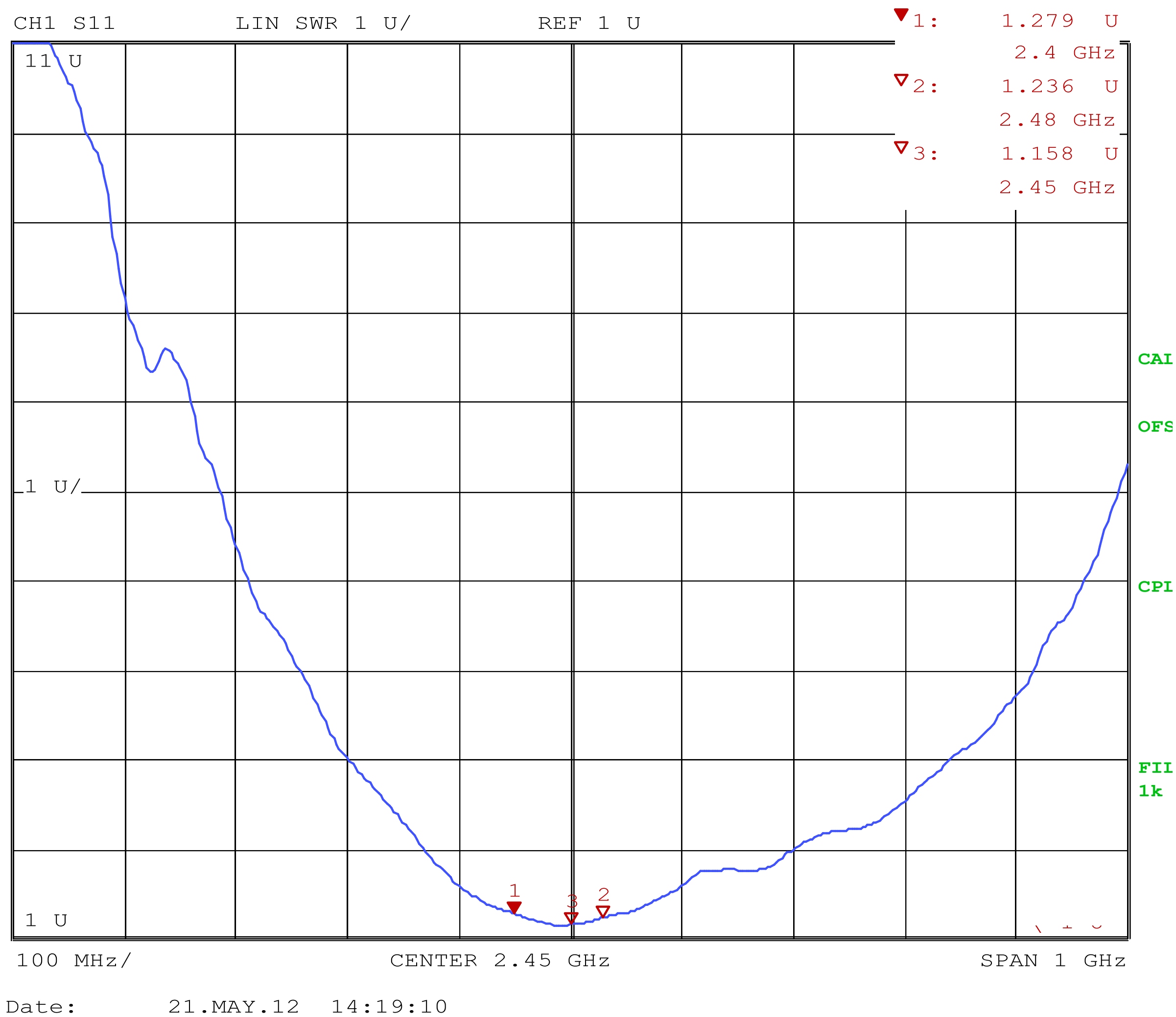


Figure 41 *SWR after matching option (b)*

8.2 PCB antenna

A PCB antenna is often made of a ¼ wavelength PCB track that extends out from the ground plane. If the antenna is laid parallel with the edge of the ground plane, its impedance will not be 50 ohm when it resonates. Theoretically, a ¼ wavelength antenna has an impedance of 73 ohm when it resonates.

Figure 42 shows the impedance measured when the antenna is isolated from the radio. In this case, the length of the antenna can be adjusted, and a single component is used to tune the antenna.

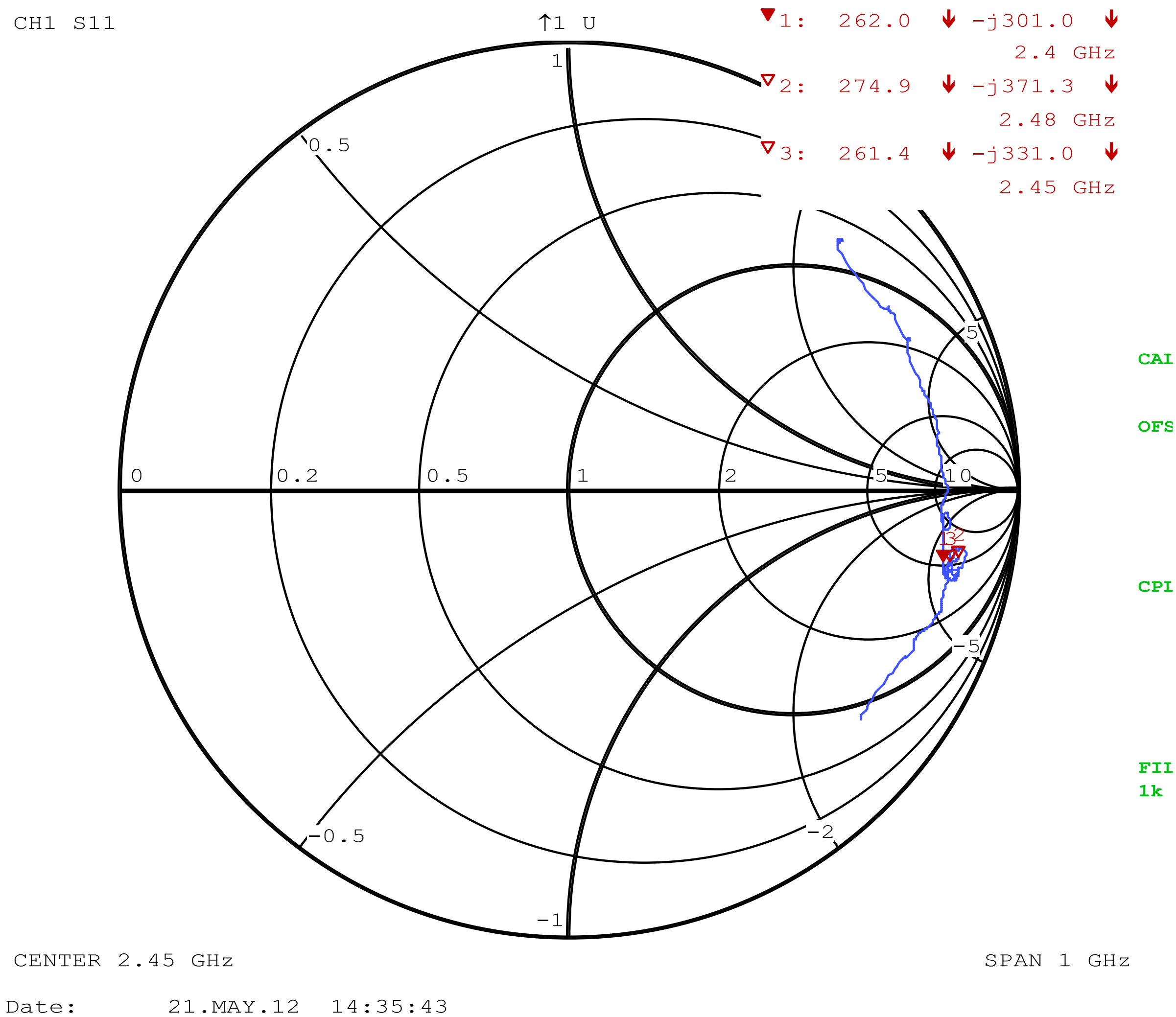


Figure 42 *Initial impedance*

If it is not possible to get the impedance exactly 50 ohm by adjusting the length of the antenna, a component must be used to pull the impedance to the 50 ohm point. It is preferable to use a shunt capacitor since a capacitor is cheaper than an inductor and because a shunt component can be removed without any impact.

When the length of the antenna is adjusted, it is important to completely remove the piece of track that is cut off since any remaining metal affects the antenna.

After the length is adjusted, and the impedance is getting close to the conductance circle that passes through the center of the Smith chart, see ***Figure 43***, a suitable capacitor is used to bring the impedance to the 50 ohm point, see ***Figure 44***. In this case, a 1.0 pF capacitor gives the best results. The exact value can be calculated, or it can be found by trial and error.

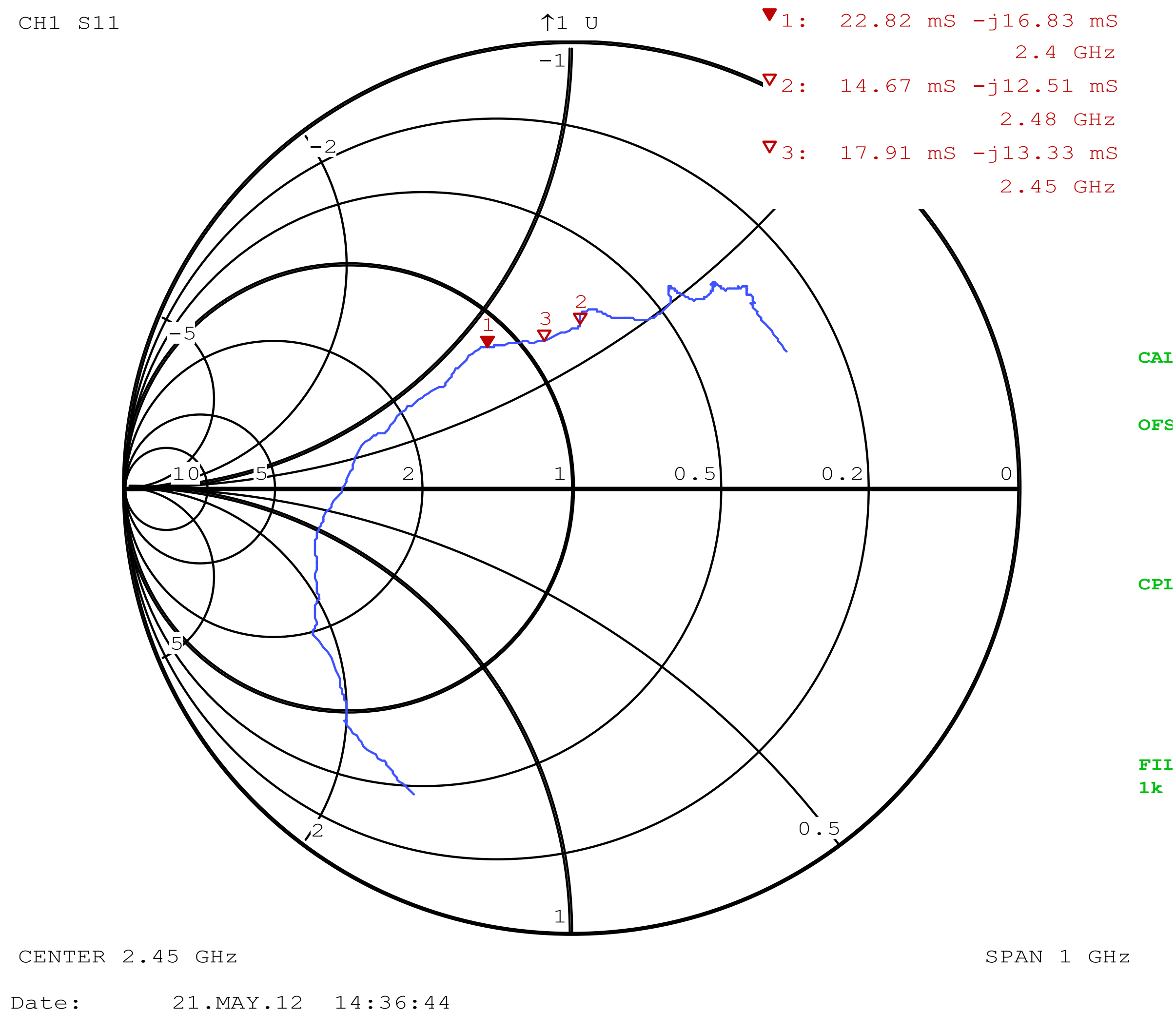


Figure 43 *Impedance after length adjustment*

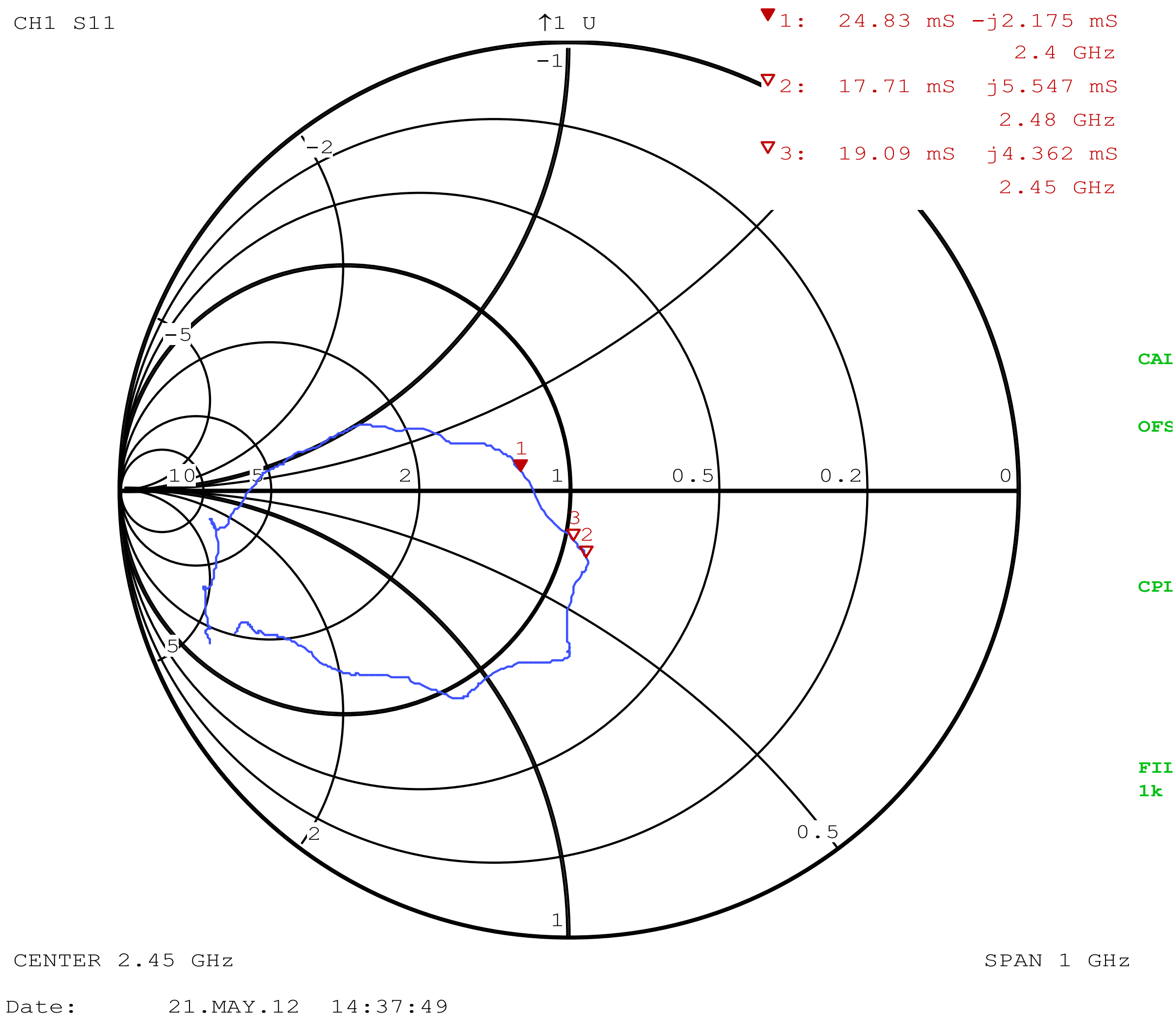


Figure 44 *Impedance after length adjustment and adding a shunt capacitor*

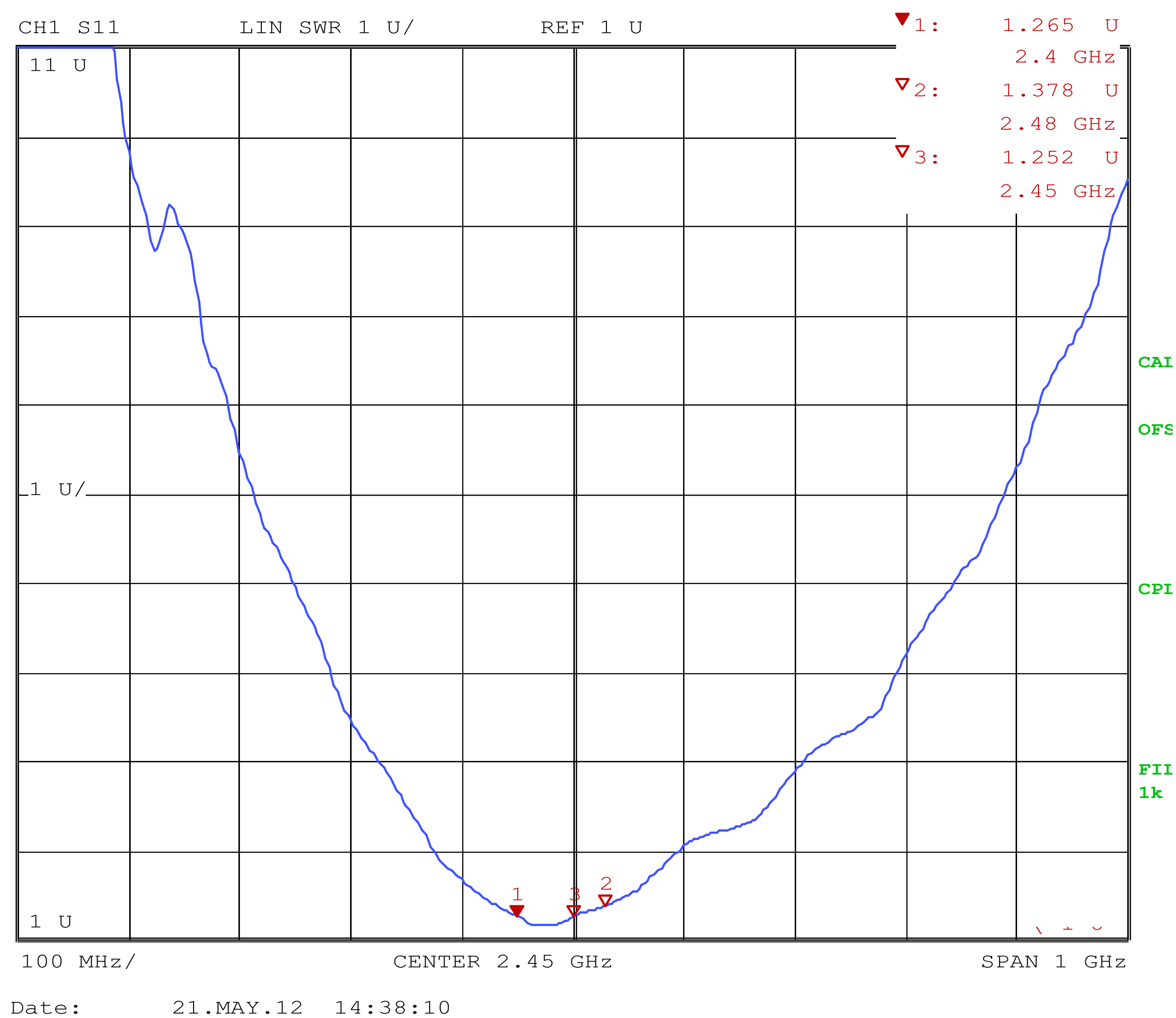


Figure 45 *SWR after length adjustment and adding a shunt capacitor*

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