AN923.2: EFR32 Series 2 sub-GHz Matching Guide

The EFR32 Series 2 family of RFICs includes chip variants that provide 2.4 GHz and/or sub-GHz operation. This application note provides a description of the RF matching network design principles including single-, dual- and wideband frequency operations for those EFR32 Series 2 chip variants that provide sub-GHz operation.

Note: This document does ***not*** address the matching procedure for the 2.4 GHz RF path. The 2.4 GHz matching procedure is described in application note, *AN930.2: EFR32 Series 2 2.4 GHz Matching Guide*. For information on PCB layout requirements for proper operation, refer to application note, *AN928.2: EFR32 Series 2 Layout Design Guide*.

KEY FEATURES

- Provides an overview of RF matching procedure
- Specifically discusses design procedures for the PA and LNA impedance transformation networks
- Single-, dual- and wideband matching networks
- Shows how to apply the information provided
- Recommended and tuned matching network component values with measurement data provided

Table of Contents

1. Device Compatibility

This application note supports the following EFR32 Wireless Gecko Series 2 devices:

- EFR32FG23
- EFR32ZG23
- EFR32FG25
- EFR32FG28
- EFR32ZG28
- EFR32SG28

2. Introduction

The EFR32 Series 2 family of RFICs includes chip variants that provide 2.4 GHz and/or sub-GHz operation. This document provides a description of the RF matching circuit design principles for those EFR32 Series 2 chip variants that provide sub-GHz operation, e.g., EFR32xG23, EFR32xG25, and EFR32xG28.

This document provides technical details and description of EFR32 Series 2 sub-GHz matching solutions applied on the publicly available Silicon Labs reference radio boards. The matching networks discussed in this document are targetted for single-band applications and use SMD0201 discrete components in the RF path. Additionally, these matching networks utilize the so-called TX-RX direct-tie topology, so both TX and RX paths are routed to the same and single antenna port without the need of use an external RF switch. Futhermore, besides the single-band matching network examples, this application note discusses the design details and provides matching circuit examples for applications covering dual- and wideband frequencies.

The layout design of the matching circuit is critical to achieve the targeted TX power, RX sensitivity, and power efficiency. Silicon Labs suggests copying the RF part of the reference PCB design, or if it is not possible, applying the layout design rules and guidelines described in *AN928.2: EFR32 Series 2 Layout Design Guide*.

The matching effort strives to simultaneously achieve these goals:

- Support tying together TX and RX signal paths, external to the RFIC without an RF switch.
- Provide the desired nominal TX output power level (measured at the connector to the antenna, load).
- Obtain this nominal TX output power at the nominal supply voltage.
- Achieve desired PA output linearity, e.g., good EVM performance, when using the EFR32xG25's high linear PA for SUN OFDM modulations
- Provide optimal RX Sensitivity.
- Minimize current consumption (i.e., maximize power efficiency).
- Comply with regulatory specifications for spurious emissions (e.g., especially TX harmonics).

The matching procedures outlined in this document will help achieve the goals listed above.

Table 4.1 Summary of EFR32xG23 Matching Component Values vs. Frequency on page 30 is provided for users more interested in quickly obtaining matching component values for EFR32xG23 than in the methodology used to develop the matching network.

Table 4.2 Summary of EFR32xG28 Matching Component Values vs. Frequency on page 31 is provided for users more interested in quickly obtaining matching component values for EFR32xG28 than in the methodology used to develop the matching network.

Table 4.3 Summary of Matching RF Conducted Performance on page 31 summarizes the conducted RF performance of the for EFR32xG23 matches used in Silicon Labs reference radio board designs.

5. Recommended Matching Network for EFR32xG25 summarizes the recommended matching network solutions available on Silicon Labs' radio board designs for the EFR32xG25 parts.

3. Matching Procedure

3.1 Overview of EFR32xG23

The sub-GHz LNA and PA circuits in the EFR32xG23 parts are single-ended and are not tied together inside the chip. As a result, SUBG_O0 or SUBG_O1 for the TX output, and SUBG_I0 or SUBG_I1 for the RX input pins are required on the RFIC to provide access to the LNA and PA circuits. Additionally, two separate TX and two separate RX RF pins are available at chip pins. These four pins are adjacently located, as shown, and highlighted in the figure below, regardless of the chip variant. The RF matching circuit must provide for connecting the TX and RX signal paths together, external to the RFIC.

The sub-GHz antenna interface consists of two single-ended input pins (SUBG_I0 and SUBG_I1) that interface directly to two LNAs and two single-ended output pins that interface directly to two +14 dBm or +20 dBm PA (SUBG_O0 and SUBG_O1). Integrated switches select either SUBG_O0 or SUBG_O1, SUBG_I0 or SUBG_I1 to be the active paths. EFR32xG23 parts are available either with the +14 dBm (LPA) or +20 dBm (HPA) Class-D PAs, these are internally bonded options so different OPN part numbers are used for the chip variants with different maximum TX power levels.

Using the two TX and two RX ports available separately, a single-BOM, true dual-frequency band matching network can easily be achieved by populating the separately-optimized, single-band matching network component values in separate RF paths.

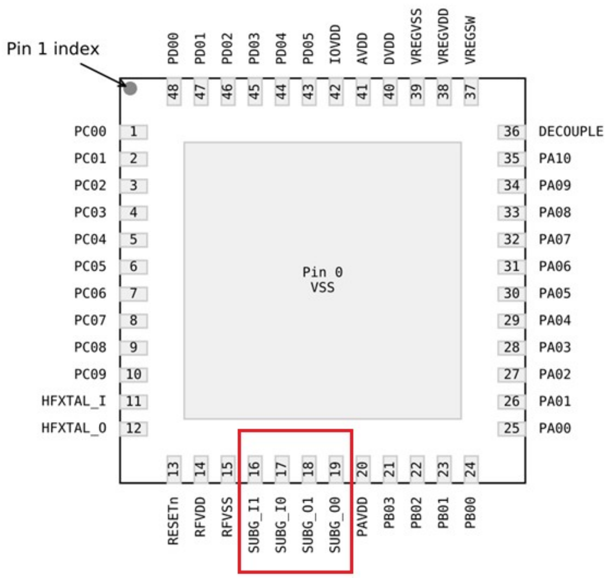

Figure 3.1. Example Pin Locations of sub-GHz TX and RX Functions

3.2 Overview of EFR32xG25

The EFR32xG25 family features two radio transceivers. Using a hardware modem and high efficiency PA to support proprietary wireless protocols, proprietary FSK, SUN FSK, and proprietary O-QPSK. Using a highly configurable software defined modem and high linearity PA, the radio supports SUN O-QPSK, and SUN OFDM.

The Sub-GHz antenna is attached to an RF matching network which interfaces to a single-ended input pin (SUBG_I) that interface directly to the internal LNA input, and two differential output pins (SUBG_ON and SUBG_OP). The SUBG_ON and SUBG_OP pins interface internally to both +16 dBm differential PAs (the constant envelop high efficiency FSK Class-D PA, and the high linearity OFDM Class-A PA). The matching networks change based on the operating frequency band. The external components, RF matching networks are shown in later sections.

The linear PA and the constant envelope PA are output on the same external differential pins, and can be used only one at a time. See a simplified radio block diagram in the figure below.

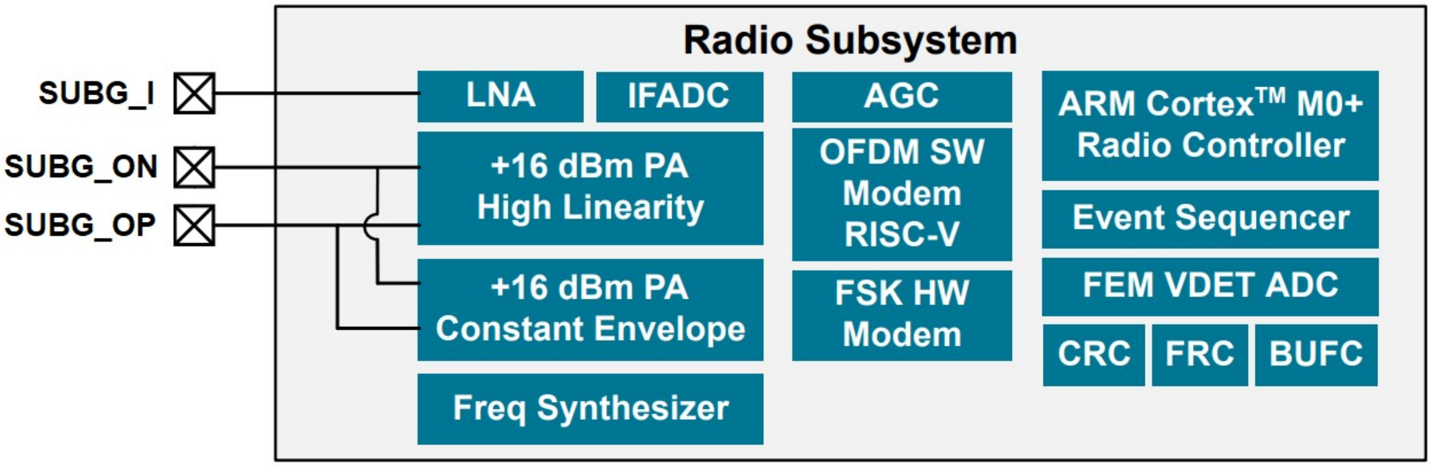

Figure 3.2. EFR32xG25 Radio Block Diagram

3.3 Overview of EFR32xG28

The sub-GHz and 2.4 GHz LNA and PA circuits in the EFR32xG28 parts are single-ended and are not tied together inside the chip. Two separate TX and two separate RX RF pins are available at chip pins. These four pins are adjacently located, as shown, and highlighted in the figure below, regardless of the chip variant. The RF matching circuit must provide for connecting the TX and RX signal paths together, external to the RFIC. Based on the selected OPN, sub-GHz-only and dual-band (sub-GHz and 2.4 GHz) variants are available. The sub-GHz-only variant provides 2-2 sub-GHz TX/RX ports similarly as EFR32xG23, while the dual-band variant provides 1-1 sub-GHz TX/RX and 1-1 2.4 GHz TX/RX ports (shown in the figure below).

The sub-GHz-only variant's antenna interface consists of two single-ended input pins (SUBG_I0 and SUBG_I1) that interface directly to two LNAs and two single-ended output pins that interface directly to two +14 dBm or +20 dBm PA (SUBG_O0 and SUBG_O1). Integrated switches select either SUBG_O0 or SUBG_O1, SUBG_I0 or SUBG_I1 to be the active paths. EFR32xG28 parts are available either with the +14 dBm (LPA) or +20 dBm (HPA) sub-GHz Class-D Pas. These are internally bonded options so different OPN part numbers are used for the chip variants with different maximum TX power levels.

The dual-band variant's antenna interface consists of a sub-GHz and a 2.4 GHz single-ended input pin (SUBG_I1, RF2G4_I0) that interface directly to sub-GHz and 2.4 GHz LNAs, respectively, and a sub-GHz and a 2.4 GHz single-ended output pin that interface directly to the sub-GHz +14 dBm or +20 dBm PA (SUBG_O1) and to the 2.4 GHz +10 dBm PA (RF2G4_O0), respectively. EFR32xG28 parts are available either with the +14 dBm (LPA) or +20 dBm (HPA) sub-GHz Class-D Pas. These are internally bonded options so different OPN part numbers are used for the chip variants with different maximum TX power levels. The 2.4 GHz PA is a +10 dBm Class-D PA and its matching procedure is discussed in application note, AN930.2: EFR32 Series 2 2.4 GHz Matching Guide.

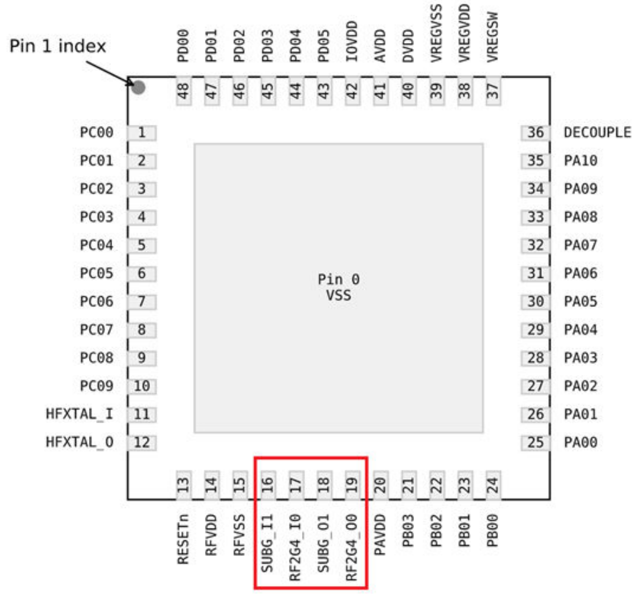

Figure 3.3. Example Pin Locations of TX and RX Functions of Dual-Band Variants

3.4 PA Supply Voltage Conditions

The EFR32xG23/28 parts include an on-chip DCDC converter, so the design is also optimized for battery-powered applications. The separate PAVDD pin supplies the internal PAs and is recommended to feed the LPA (+14 dBm PA) from the DCDC output (VREGSW) for the output power levels equal or below +14 dBm. For +20 dBm TX power, and above +14 dBm, connect the PAVDD pin needs to the main battery voltage (e.g., +3.3 V) to supply the HPA (+20 dBm PA). The SUBG_O0 and SUBG_O1 TX ports must not be tied to a dc supply voltage, and furthermore the transmit RF path must include a series dc-blocking capacitor. The SUBG_I0 and SUBG_I1 RX ports should not be connected to a dc supply voltage, but the LNA pins include an on-chip dc-blocking capacitor.

There is an on-chip pre-regulator block for the EFR32xG23/28 power amplifiers. Linear regulator with an input of PAVDD and output of PA blocks. Regulates to a target V_{DD} voltage when PAVDD > V_{DD}. Follows supply with a ~30 mV offset when PAVDD ≤ V_{DD} (and PA power will trail off also). For the +14 dBm LPA the target V_{DD} voltage is 1.65 V, while for the +20 dBm HPA it is 3.0 V (with the exception of 3.15 V in the 868-915 MHz frequency bands).

The EFR32xG25 parts need PAVDD voltage >= 3.45 V, nominally 3.6 V +/- 4 %, to keep the output linearity of the linear OFDM PA. The regulated internal voltage to the FSK PA is 3.0 V.

3.5 EFR32xG23/28 Class-D PA and Optimum PA Load Impedance

Different EFR32xG23/28 OPN parts are available for the maximum available TX power levels of +20 or +14 dBm (bond-out options). The +20 dBm PA is the high-power PA, i.e., HPA, while the +14 dBm PA is referred as low-power PA, i.e., LPA. The different power amplifiers require an optimal different load impedance for the maximum TX power level.

The EFR32xG23/28 has a complementary Class-D low-power PA and high-power PA available with the following generic features:

- Single-ended output, which ensures simple RF matching topology.
- Two transistors are working as on/off switches during operation between GND and V_{DD}. TX matching RF path needs a series dc-blocking capacitor.
- High power efficiency, i.e., low current consumption, due to the switched-PA behavior. Square-wave voltage with shifted sinusoidal current waveforms on the PA output.

The AC voltage swing is well controlled in both on and off switching cycles. Therefore, the AC over-voltage stress is not a concern because the maximum AC voltage swing cannot exceed V_{DD}. The PA output swings between its regulated supply (V_{DD}) and ground. As a result, the swing is limited to a maximum of 3 V for HPA and 1.65 V for LPA, irrespective of PAVDD pin supply, antenna VSWR (different load conditions at the SMA port) and other design parameters. So, APC mechanism (monitoring and possibly backing-off the power for long-term reliability) is not required for the PA device on the EFR32xG23/28 parts.

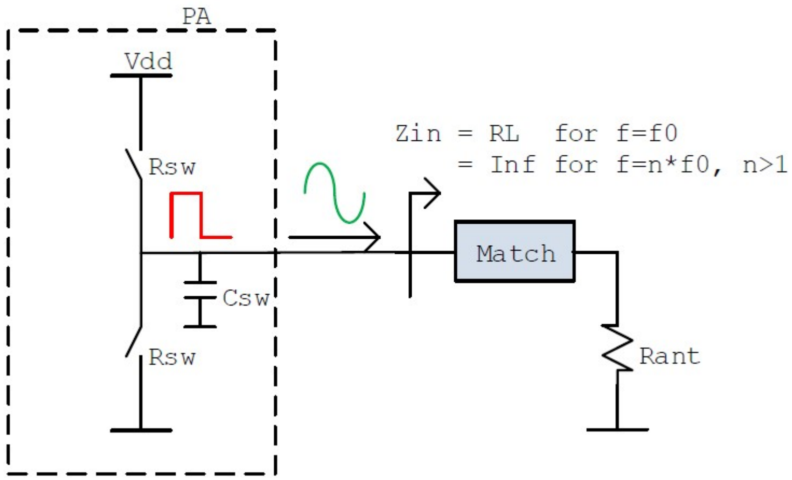

Figure 3.4. Class-D PA Structure

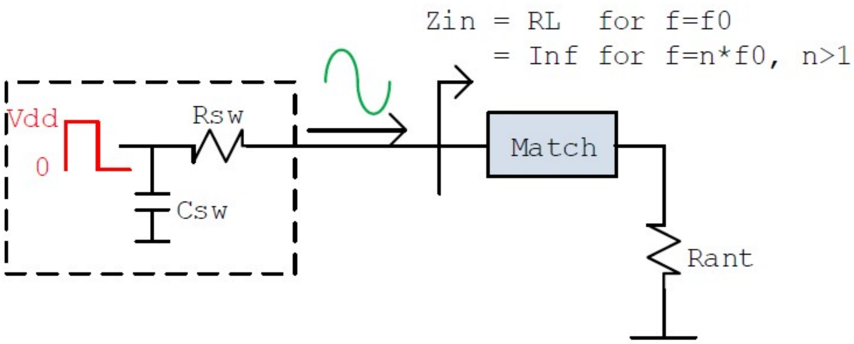

Figure 3.5. Simplified Class-D PA Structure

The PA can be modeled as a parallel RC network and the PA on-off impedance values are given in the table below.

Table 3.1. PA Impedance

PA	Mode	R_{SW}	C_{SW}
LPA	ON	2.2 Ω	2 pF
	OFF	25 kΩ	1 pF
HPA	ON	1.2 Ω	4 pF
	OFF	17 kΩ	2 pF

As the table shows, the active mode PA impedance is low, while the off-mode PA impedance is large. This behavior comes from the complementary Class-D switched PA operation.

An optimal PA matching network shows R_L, as optimum PA load impedance, at the fundamental frequency, while a high-impedance load is provided to the PA at all TX harmonics, as also shown in the figures above. Naturally, it is difficult the full-fill the high-impedance harmonic load requirement, but it is recommended to provide the possible highest load impedance at the 3rd harmonic for the PA, at least.

The optimum PA load impedance at the fundamental frequency, R_L, is determined by the regulated V_{DD} PA supply voltage and desired TX output power level. The following equation helps derive the required R_L impedance.

$$P_{OUT} = \frac{1}{2}\left(\frac{2}{\pi}V_{DD}\right)^2 \frac{R_L}{(R_{SW}+R_L)^2}$$

The bonding wire inductance can be estimated to L_{BWTX} = 1.2 nH for each TX pin. The optimum PA load impedance at the PA, before the bonding wires and matching network, can be estimated as shown in the table below for different desired TX power levels.

Table 3.2. Optimum PA Load Impedance at the PA

Parameters	20 dBm	17 dBm	14 dBm	10 dBm	0 dBm
PAVDD [V]	3.3	3.3	1.8	1.8	1.8
Regulated PA V_{DD} [V]	3.0	3.0	1.65	1.65	1.65
R_{SW} [Ω]	1.2	3.0	2.2	6.0	60
Calculated R_L [Ω]	10	18	10	22	200
Recommended R_L [Ω]	7	15	9	20	200

Power ramping is done digitally by controlling PA units, e.g., (N_{HPA}), as follows:

$$(N_{HPA}):\ R_{SW} = 1.2\Omega\ \frac{511}{N_{HPA}}$$

The regulated V_{DD} voltage for the HPA is 3.15 V in the 868 and 915 MHz frequency bands for +20 dBm TX power. Calculated load impedance assumes 1.5-2 dB of losses, so the TX power is given at the antenna port. The optimum PA load impedance value (here, R_L = Z_{IN}) in the table above is given at the PA before the bonding wire as shown in the figure below. For a custom matching network design, Silicon Labs recommend starting from these conditions given in the table above. The recommended PA load impedance ensures some additional margin to compensate all switching, internal, external losses, and technical spreading.

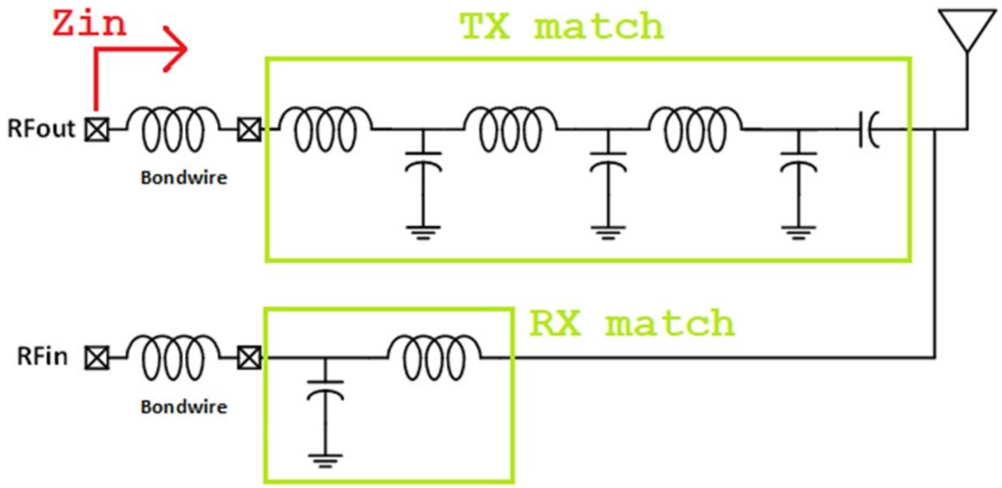

Figure 3.6. Optimum PA Load Impedance Given Before Bonding Wire

The table below shows the optimum input impedance of TX matching networks (plus the bonding wires) at different frequencies, supply voltage and TX power requirements, available on Silicon Labs reference radio board designs. In other words, Z_{IN} impedances in the table below are fine-tuned, compared to R_L values in Table 3.2 Optimum PA Load Impedance at the PA on page 10, for the available reference radio board designs.

The low-pass filter design should be handled together with the TX matching network design. However, a front-end matching designer could design the TX matching network separately from the LPF design, e.g., use a maximum of 2-element TX match plus a symmetrical 50-to-50 Ω LPF. However, with this approach the harmonic impedance load is typically lower than desired, so re-tuning of the TX matching network is required which also affects the low-pass filter design. As a result, Silicon Labs chose a more-element TX matching LPF where both TX matching and low-pass filtering functions are handled simultaneously and the 50 Ω point appears only at the end of the TX matching LPF, i.e., at the antenna port only. This approach lets the matching designer to set the required harmonic load impedance and overall Q, bandwidth of the matching network.

Table 3.3. Optimum PA Load Impedance at the PA on Available Reference Designs

Frequency	TXP	PA	PAVDD Pin Supply Voltage	Regulated Internal V_{DD}	Z_{IN} at Fundamental Frequency	Z_{IN} at 3rd Harmonic Frequency
915 MHz	14 dBm	LPA	1.8 V	1.65 V	7.5 + j3 Ω	144j Ω
868 MHz	14 dBm	LPA	1.8 V	1.65 V	9.0 - j1 Ω	N/A
915 MHz	10 dBm	LPA	1.8 V	1.65 V	20 + j4 Ω	210j Ω
868 MHz	10 dBm	LPA	1.8 V	1.65 V	23 + j0 Ω	N/A
915 MHz	20 dBm	HPA	3.3 V	3.15 V	4.8 + j1 Ω	338j Ω
868 MHz	20 dBm	HPA	3.3 V	3.15 V	5.7 - j2.5 Ω	N/A
490 MHz	17 dBm	HPA	3.3 V	3.0 V	15.3 + j2 Ω	193j Ω
434 MHz	10 dBm	LPA	1.8 V	1.65 V	21.3 + j0 Ω	246j Ω
315 MHz	10 dBm	LPA	1.8 V	1.65 V	18.0 + j0 Ω	160j Ω
169 MHz	20 dBm	HPA	3.3 V	3.0 V	6.9 + j1 Ω	116j Ω

The optimum PA load impedance just before the bonding wire is estimated as the optimum input impedance of the TX matching network toward the matching network from the PA die just before the bonding wire and all other on-chip parasitics are included; however, the on-chip bonding wire, leadframe, and all external PCB and SMD parasitics should be accounted for during the matching network design. These impedance values given in the table above are based on the EFR32xG23 reference radio board designs available. Due to the different chip package and minor PCB layout differences (e.g., because of the dual-band layout requirements), these impedance values may slightly differ for designs with EFR32xG28 parts.

3.6 EFR32xG25 Optimum PA Load Impedance

The EFR32xG25 family contains of two +16 dBm differential PAs. The high-efficiency FSK PA is a Class-D design similary as for EFR32xG23 but with differential TX ouputs. The linear OFDM PA is a Class-A design with also differential PA outputs. These two power amplifiers are internally routed to the same differential TX ouput pins but can be used only one of them at a time.

The optimum differential PA load impedance at the chip pins at the fundamental frequency is:

- **Z_{IN} = 14.0 - j*5.6 Ω at 915 MHz**
- **Z_{IN} = 13.2 + j*2.1 Ω at 868 MHz**
- **Z_{IN} = 13.6 + j*2.3 Ω at 470 MHz**

These PA load impedance values are optimal for +14/16 dBm TX power output. Furthermore, Silicon Labs reference designs use one common TX matching network simulateneously for both FSK and OFDM PAs, and the performance is more optimized for the linear OFDM PA in order to keep the output linearity and therefore EVM specifications for SUN OFDM modulations. For OFDM PA performance (linearity), a more negative imaginary part is more favorable, however this is at the cost of FSK PA's efficiency performance. For FSK-only designs, the optimum differential FSK PA load impedance at the chip pins using the impedance values above is around **Z_{IN} + j*3 Ω**.

The PA off-mode impedance is around 17 kΩ with a parallel 10 pF capacitance.

3.7 EFR32xG2x LNA Impedance

The LNA is a high impedance design where a complex conjugate match is basically not possible at the given frequencies using SMD components. However, the matching design goal is to provide the possible highest voltage gain for the receiver. The LNA design is the same for both EFR32xG23/28 and EFR32xG25 parts, it can also be modeled as a parallel RC network and values are given in the table below.

Table 3.4. LNA impedance

LNA	Mode	R_{LNA}	C_{LNA}
LNA	RX	from 10 to 6 kΩ	1.2 pF
	TX	10 Ω	1.2 pF

The LNA impedance is high in RX active mode (from 10 kΩ to 6 kΩ across the entire sub-GHz frequency range), while there is an internal switch to GND which turns on during TX mode to protect the LNA. The C_{LNA} capacitance in the table includes all chip parasitics. The bonding wire inductance can be estimated to series L_{BWRX} = 1.5 nH.

Because of the high value of equivalent parallel resistance (6 kΩ at 915 MHz), Silicon Labs makes no attempt to construct a true complex conjugate match at the RX interface. Such an extreme impedance transformation ratio (e.g., ANT = 50 Ω to RX = 6000 Ω) would require an extremely high-Q, narrowband, and difficult-to-tune matching network. The LNA circuitry acts more as a voltage amplifier than a power amplifier. Therefore, less emphasis is placed on maximum power transfer to the LNA input and more emphasis on the passive voltage gain of the RX matching network.

EFR32 Series 2 sub-GHz chip variants include single-ended LNA with simple external RX matching circuit.

3.8 Matching Topology Used on Silicon Labs Reference Designs

Silicon Labs reference radio boards use the so-called TX-RX direct-tie matching topology, which means the TX and RX paths are directly connected to each other to interface to a single-ended 50 Ω antenna without the need of any external RF switch in the RF-FE path.

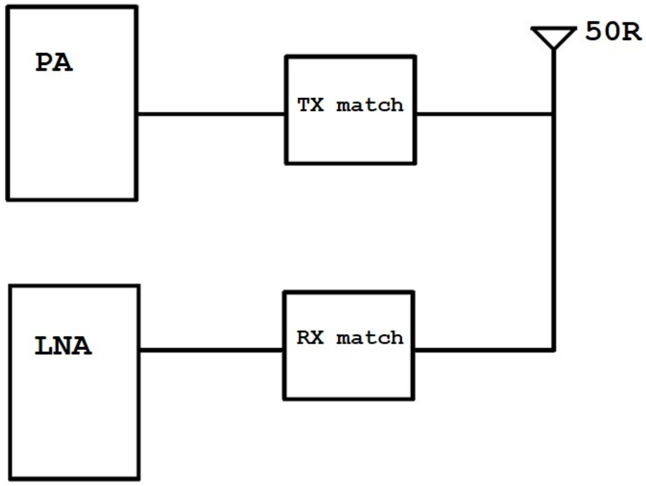

Figure 3.7. TX-RX Direct-tie Matching Network Topology

The TX matching network includes the low-pass filtering function as well to suppress the TX harmonics, while the RX matching network does the possible highest voltage gain to the LNA by transforming the impedance up while the LNA capacitance is resonated out.

In transmit mode, the LNA port is shorted to the GND by an internal switch (with a series 10 Ω resistance) to protect the LNA. The input of the RX match needs to show high impedance under these conditions to deliver the power to the 50 Ω load, antenna. Additionally, the TX match transforms the impedance between the optimal PA load impedance and 50 Ω antenna, while the input impedance of the TX match at the harmonics (typically, at least at the 3rd harmonic) should be high to enhance the harmonic suppression.

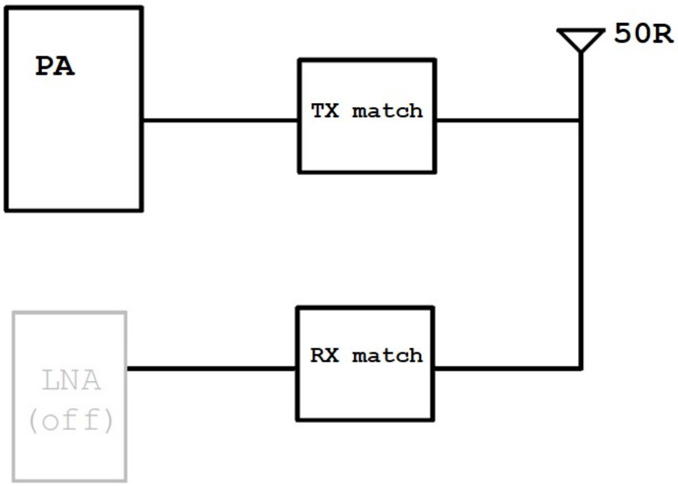

Figure 3.8. Effective Matching Circuit in TX Mode

In receive mode, the PA operates in off mode, where the PA impedance is high. Under these conditions, the output of the TX match should show high impedance to transfer the received power towards the LNA.

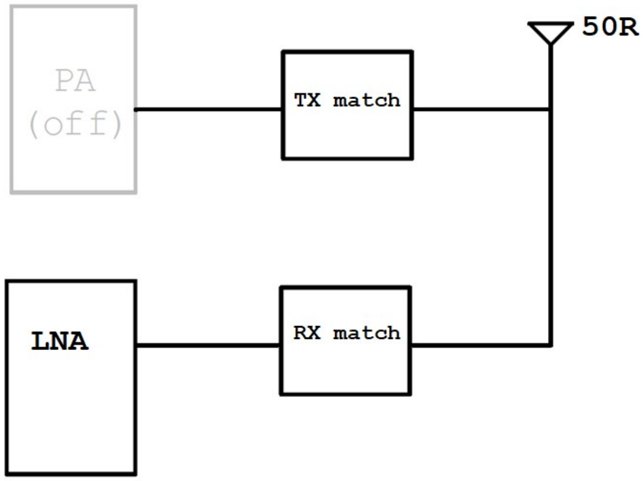

Figure 3.9. Effective Matching Circuit in RX Mode

3.9 Separating RF Paths and Inserting SAW Filter

Silicon Labs provides matching network solutions in the TX-RX direct-tie matching topology. However, separating the RF paths also gives tuned and optimal matching solutions for separate TX-RX RF path matches, e.g., for designs with RF switch or FEM. Inserting a SAW filter, for any reason, into the design is also possible, where Silicon Labs recommends placing the SAW filter in the RX path only and separate the TX and RX paths with an RF switch.

UPG2214TB-A, CG2179M2-C4, AS179-92LF, AS213-92LF, SKY13323-378LF are the possible RF switch candidates.

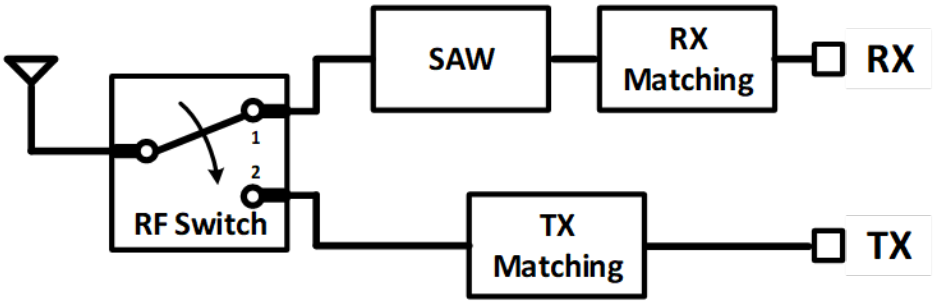

Figure 3.10. Inserting SAW Filter

3.10 Matching Network Design Details

3.10.1 EFR32xG23/28 TX Matching Network Design

The TX matching network must meet the following requirements:

- Provide the optimum PA load impedance at the fundamental frequency.
- Provide high impedance load at the harmonics, focused for the 3rd harmonic. Recommended to start the TX match with a series inductor to increase the load impedance at the harmonics (i.e., minimize harmonic current).
- In PA off mode the output impedance of the TX match should be high-Z for proper direct-tie RX operation.
- Suppress the harmonics to comply with world-wide emission regulatory limits.

The optimum load impedance must be provided to the PA when the antenna port is terminated by 50 Ω. The impedance matching and low-pass filtering functions can be combined in the TX matching network design, i.e., the task is to design a low-pass filter between the optimum PA load and 50 Ω impedances. The Q of the TX matching network can be adjusted by the component values and number of matching components, while the harmonic suppression can also be enhanced by increasing the order of the matching LPF. The TX pin must be dc-blocked, so the TX matching network must contain of a series dc-blocking capacitor as well.

The target for the input impedance of the TX match at the 3rd harmonic is 250 Ω from 3.3V, 200 Ω form 2.5V and 150 Ω from 1.8V with any topology, to comply with the radiated harmonic emission limits with comfortable margins. For instance, for 915 MHz +14 dBm match, Silicon Labs use a 4-element LCLC ladder match to provide 150 Ω at the 3rd harmonic. However, this typically requires a higher inductor value at the TX pin, which introduces bigger match loss and degraded wideband performance. For 915 MHz +20 dBm match, an LCLC ladder match cannot be used to get 250 Ω at the 3rd harmonic. So, a parallel LC needs to be used in the place of the first series inductor of the LCLC match, which also gives lower match loss and better wideband performance. However, care must be taken about the higher harmonics.

Radiation sources can be the antenna; RF traces especially closer to the TX pin, e.g., PAVDD pin, where it is recommended to place a series ferrite or inductor close to the chip pin to reduce the leaked harmonic current and thus minimize the unwanted emissions. GPIO lines can also be unwanted radiation sources because the PA is switched single-ended, the return currents of harmonics come back to the chip through PA ground which can couple to other ground domains within the chip including GPIOs.

To achieve a TX match, design a ladder LC network, starting with a series inductor close to the TX pin, and therefore the high-impedance load can be ensured for the harmonics. To enhance the suppression of a specific harmonic (e.g., 3rd harmonic), a parallel LC network can also be placed in the series RF path close to the TX pin, as the first element of the ladder LC TX matching network. This is required for +20 dBm operation in the 868-915 MHz frequency band to comply with the regulatory limits. The high impedance load at the harmonics minimizes the harmonic currents coming out of the PA and therefore limiting the radiated emissions.

The TX matching procedure of the +14 dBm 868-915 MHz design is shown as an example in this section.

Step 1: Selecting optimum PA load impedance: The optimum PA load impedance values are listed in the Table 3.2 Optimum PA Load Impedance at the PA on page 10 and Table 3.3 Optimum PA Load Impedance at the PA on Available Reference Designs on page 11, which provide the required matching network impedance (with bonding wires) seen by the PA. For an 868-915 MHz design with one BOM, the $Z_{IN} = R_L = 9 + j0\ \Omega$ is selected for simplicity. This impedance must be seen by the PA before the bonding wire + matching network at the 898 MHz center frequency (for the band between 868 and 928 MHz).

Step 2: Matching and filtering network design to 50 Ω: The selected optimum PA load impedance needs to be matched to 50 Ω together with a low-pass filter design. The PA bonding wire inductance and external PCB trace inductance is estimated to 1.2 and 0.75 nH, respectively. The match starts with a series L and then a parallel C is required to get to the 50 Ω point as the figure shows below.

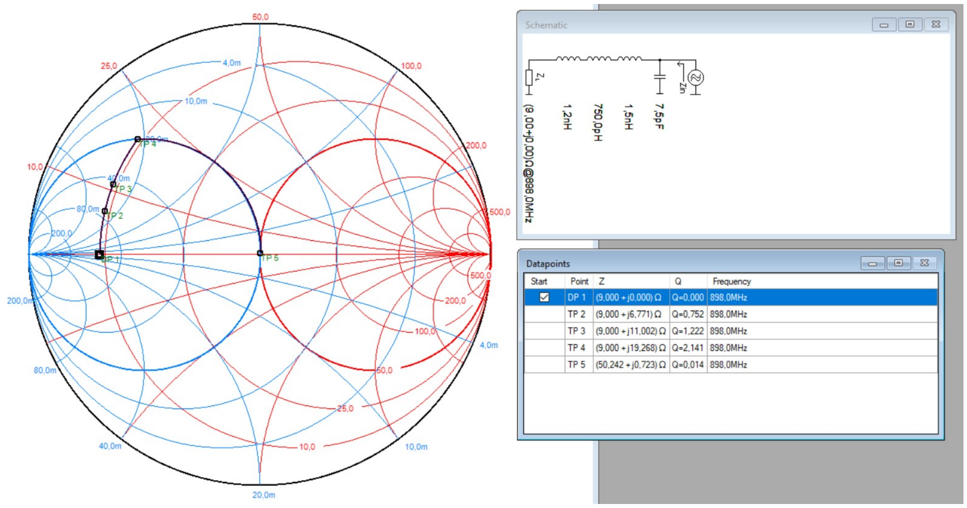

Figure 3.11. 2-Element TX Match on Smith Chart

A symmetrical 50-to-50 Ω low-pass filter can be added after the 2-element TX match (series 1.5 nH, shunt 7.5 pF), shown above. This solution provides the required impedance transformation at the fundamental frequency but the input impedance of this TX match at the 3rd harmonic does not appear to be high enough, as shown in the following figures.

The figure below shows the simulated TX matching circuit including the bonding wire, PCB trace inductances. The 2-element TX match calculated above is combined with a 5.1 pF – 6.8 nH – 5.1 pF symmetrical, 3-element low-pass CLC π-filter.

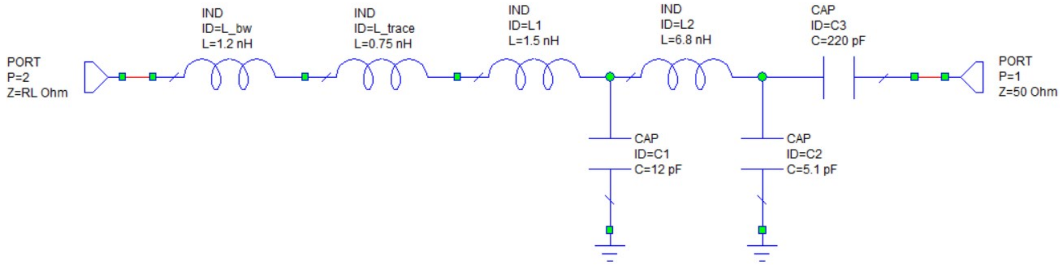

Figure 3.12. 2-Element TX Match Plus Symmetrical LPF

To have higher impedance load at the 3rd harmonic, the first series inductor, L1, value must be increased which also requires re-tuning the TX match together with the LPF. Further increasing the L1 inductance does not allow to do the TX match by two components only, so the entire 4-element LCLC needs to be the part of both TX matching and low-pass filtering functions. See an example for this match with an increased L1 inductor value on the figure below.

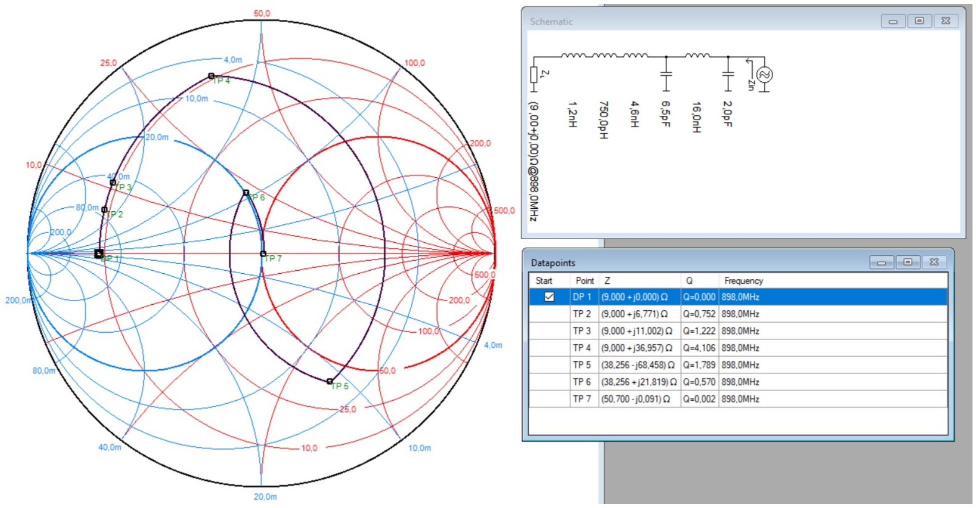

Figure 3.13. 4-Element TX Matching LPF on Smith Chart

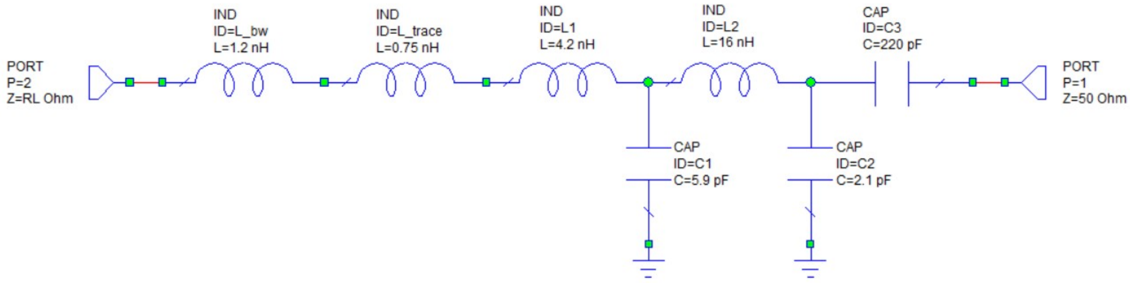

Figure 3.14. 4-Element TX Matching LPF

Both versions of TX matching networks does the required impedance transformation between the PA and antenna ports at the fundamental frequency, but the TX match variant with the increased L1 inductor value provides higher impedance load at the 3rd harmonic, which is also a TX matching design goal. See the input impedances of the TX matches for comparison in the figure below. The simulations include SMD and PCB parasitics. The higher impedance load at the harmonic (3^{rd}) helps reduce the harmonic current generated and leaked out, and therefore can help provide lower radiated harmonic emissions.

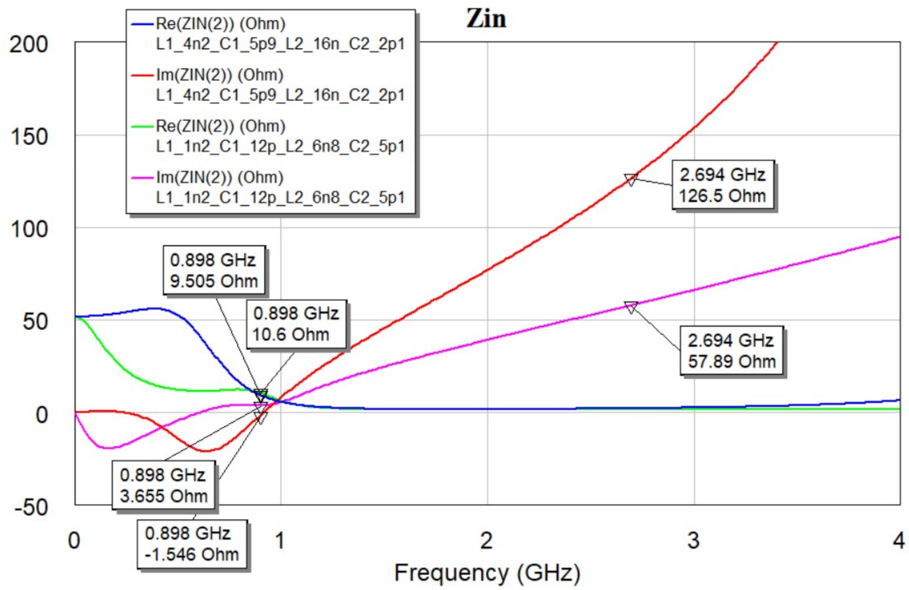

Figure 3.15. Input Impedances of TX LCLC Matching Variants

The two TX matching options have different Q and bandwidth. The TX match with the lower first inductance has a larger bandwidth. As a result, designers can choose and trade-off between harmonic suppression and TX bandwidth requirements. Silicon Labs reference designs follow the higher Q, lower bandwidth solution which provides better harmonic suppression because the BW criteria of the targeted applications is also well satisfied (-10 dB BW > 100 MHz). In other words, Silicon Labs recommend utilizing larger inductor values in the TX matching network (i.e., higher-Q match) until the BW requirement is still satisfied under all conditions. See the comparison of S parameters of these matching solutions in the figure below.

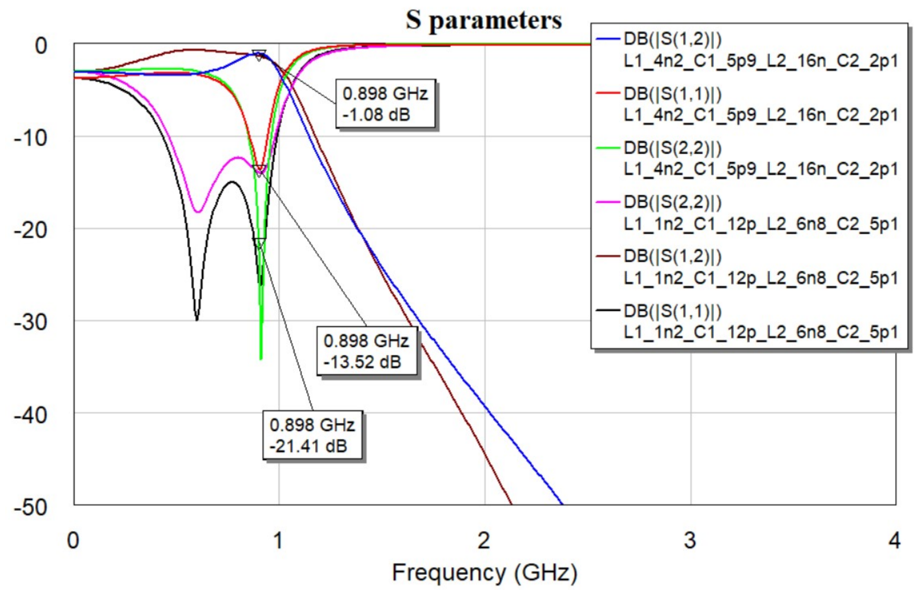

Figure 3.16. S Parameters of TX LCLC Matching Variants

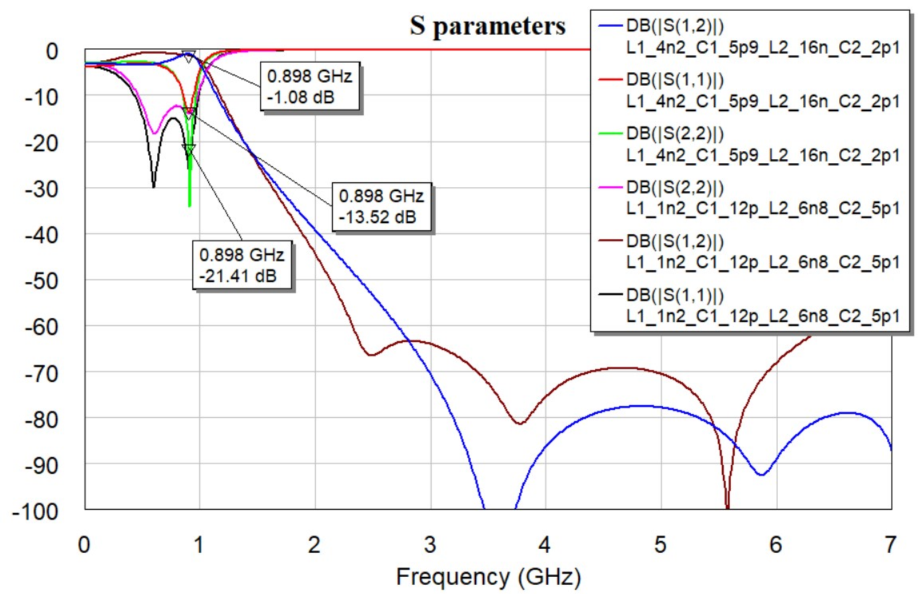

Figure 3.17. S Parameters of TX LCLC Matching Variants, Increased Frequency Span

The input impedance of the TX match can also be plotted on Smith Chart, where you can observe that the TX match with the increased L1 inductor approaches the high-Z point closer at the 3rd harmonic frequency.

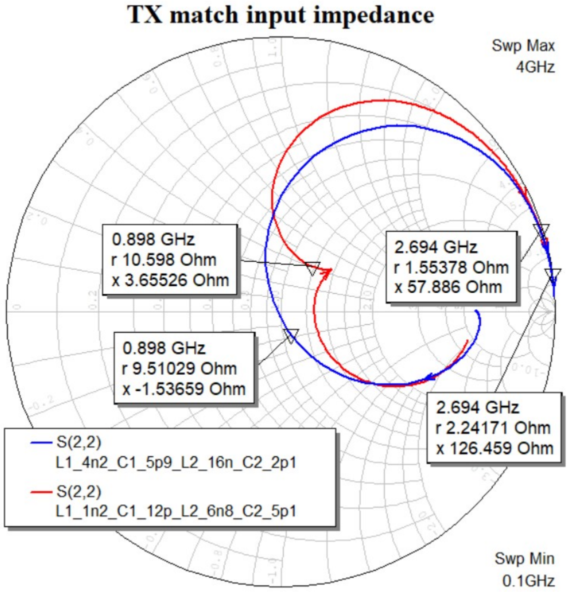

Figure 3.18. Input Impedances of TX LCLC Matching variants on Smith Chart

The simplified simulation structure of the TX match including the bonding wires, PCB and SMD parasitics with equivalent discrete circuit model is shown in the figure below (at the antenna port [1] the series dc-blocking capacitor is not shown here).

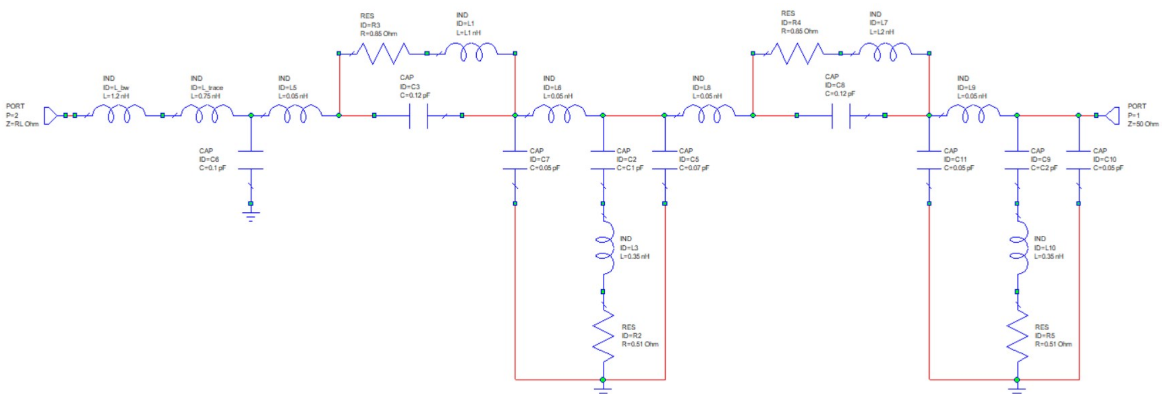

Figure 3.19. Simplified Simulation Model of TX LCLC Matching Network with Parasitics

Step 3: Checking the output impedance in PA off mode: This is an important step for direct-tie TX-RX matching configurations, where the TX and RX paths are directly connected to each other without the use of an external RF switch. In PA off mode, the output impedance of the TX matching network must show high impedance at the fundamental frequency for proper direct-tie RX operation. In the PA off mode, its capacitance needs to be included in the simulations as well.

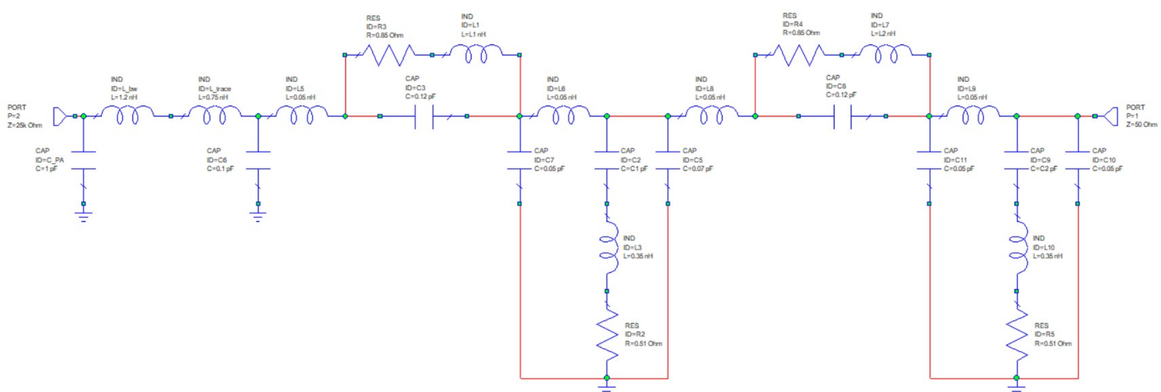

Figure 3.20. Simplified Simulation Model of TX LCLC Matching Network with Parasitics, PA-off Mode

The output impedance of the TX match in PA off mode is shown in the figure below for both matching solutions, where you can observe that the match with larger inductor values provides higher impedance load. The higher-Z output impedance yields better RX performance in the TX-RX direct-tie matching configuration.

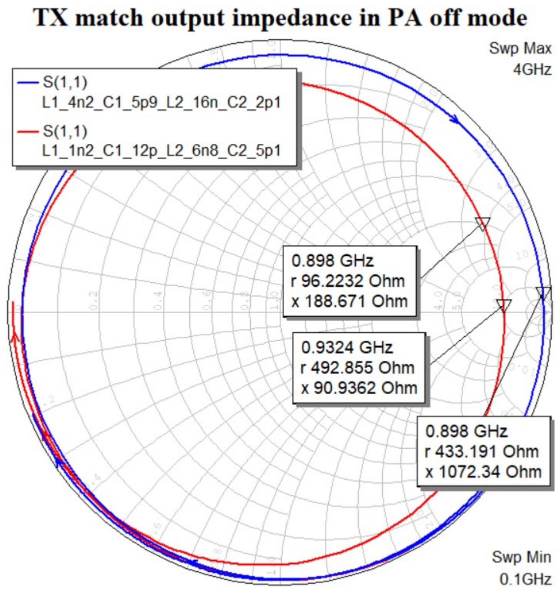

Figure 3.21. Output Impedances of TX LCLC Matching Variants in PA-off Mode

3.10.2 RX Matching Network Design for EFR32xG23/28 and EFR32xG25

The RX matching network must meet the following requirements:

- Provide the optimum NF and possible highest voltage gain at the fundamental frequency for the LNA.
- In LNA off mode the input impedance of the RX match should be high-Z for proper direct-tie TX operation.

The LNA impedance is in the range of a parallel resistance > 6 kΩ with an internal shunt 1.15 - 1.2 pF capacitance. As a result, complex conjugate match cannot be done at the given sub-GHz frequencies using SMD components. The RX matching network rather needs to provide the possible highest impedance transformation (while keeping Q < 6) to the RX pin while the on-chip LNA capacitance is resonated out. The RX matching network therefore can be a single series inductor which resonates with the on-chip LNA capacitor. This provides a very simple matching circuit. Silicon Labs reference radio board designs follow this first-order match with a single series inductor in the RX match, where the only exception is the 169 MHz design where an additional shunt capacitor is also used at the RX pin to limit the Q of the matching circuit which provides better BW and lower NF (due to the extremely large inductor's own loss, noise becomes worse than the matching improvements at 169 MHz when Q > 6).

The LNA capacitance can be modeled with a ~1.2 pF shunt capacitor (with all on-chip parasitics included). This value does not significantly change over the sub-GHz frequencies, which also means that the single series RX matching inductor must be selected to resonate with the given LNA capacitance at the desired fundamental RF frequency. This results that the Q of the matching network decreases and the BW of the match increases with frequency.

Silicon Labs reference radio boards are available in TX-RX direct-tie matching configuration, which means that a longer trace must also be used in the RX path to connect the RX match to the TX match output which utilizes more components and therefore occupies a bigger area of the layout. Silicon Labs radio boards use a relatively long, inductive trace for the direct-tie connection where the inductance of the additional series PCB trace can be estimated to 4 nH in total. Furthermore, the LNA bonding wire inductance is around 1.5 nH.

For example, in the 868-915 MHz RX matching network design, the LNA capacitance of 1.15 - 1.2 pF resonates at the center frequency of 898 MHz with a series ~24 nH inductor, as shown in the Smith chart figure below, while the resulted LNA load impedance by the RX matching network ends up around 420 Ω (at lower frequencies this impedance gets higher with the fixed LNA capacitance but increased series matching inductor value). The RX matching inductor needs to be adjusted back by the bonding wire and PCB trace inductances which results in 18-19 nH. Silicon Labs radio boards utilize 18 nH in the 868 and 915 MHz bands.

The LNA design has an on-chip dc-blocking capacitor in the signal path before the LNA and a large pull-down resistor that provides a dc path between RX pin and ground.

Therefore, the RX matching network design consists of a single series inductor, which needs to be selected to resonate with the on-chip LNA capacitance of ~1.2 pF. Additional shunt capacitor at the RX pin can be recommended when the center frequency is below 300 MHz. In the high-frequency bands, an additional shunt inductor could be used to resonate out some of the LNA capacitance and therefore achieve a higher matching network Q, but the tradeoff is the shunt inductor own loss, noise, and the cost of an additional inductor. Silicon Labs decided not going with this approach because the benefit didn't appear to be substantial. In the low-bands, a shunt capacitor is used to limit the inductor size, beyond a certain Q (not recommended to go beyond Q > 6, which means the suggested LNA load impedance is < 1.85 kΩ), since the noise of the inductor starts dominating the RX noise floor, so an optimal Q (~5) should be chosen in these cases.

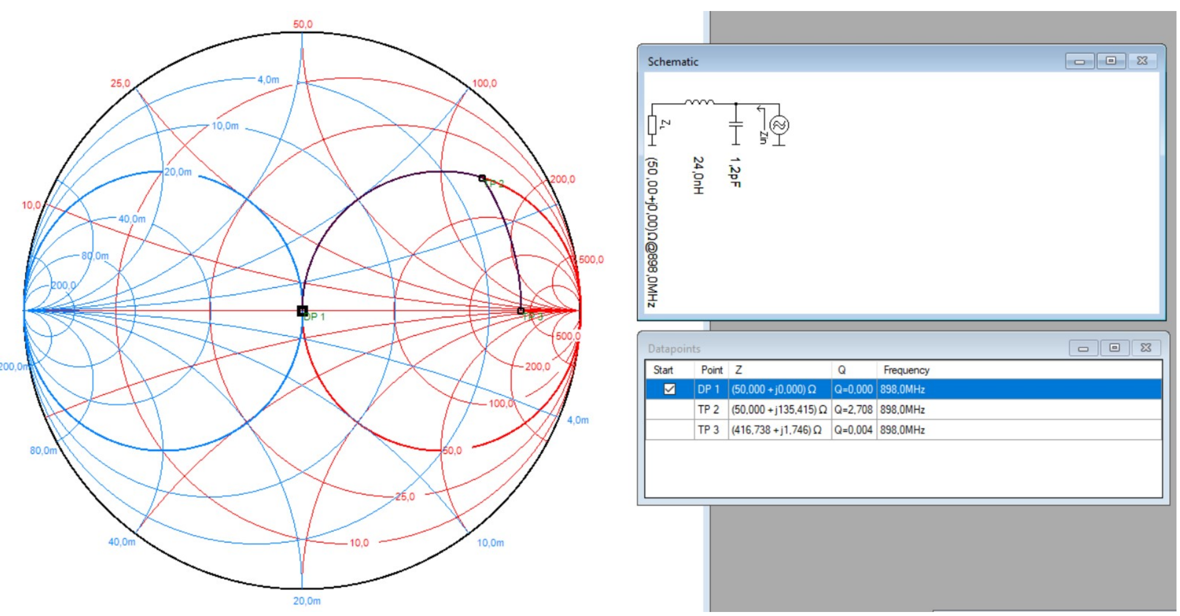

Figure 3.22. Single-Element RX Match on Smith Chart

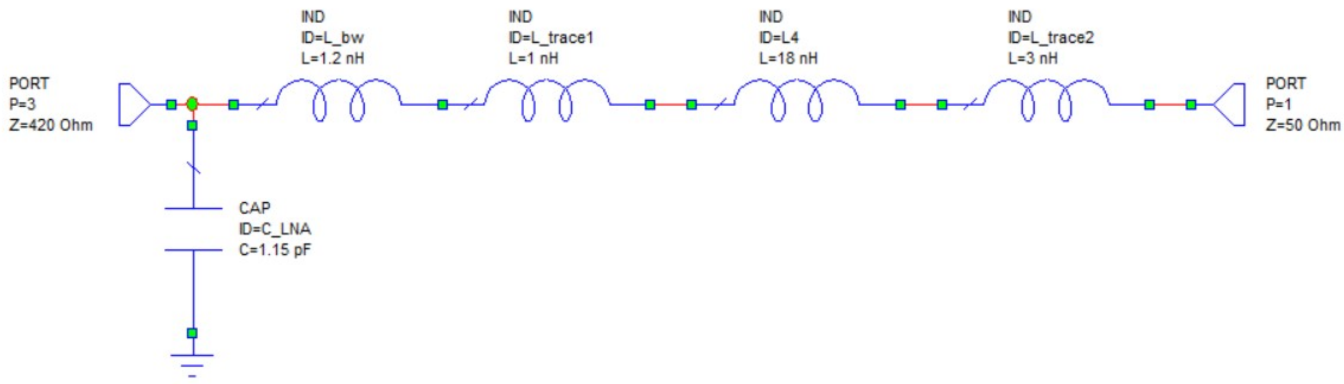

Figure 3.23. RX Match with LNA Capacitance, Bonding Wire and PCB Trace Inductances

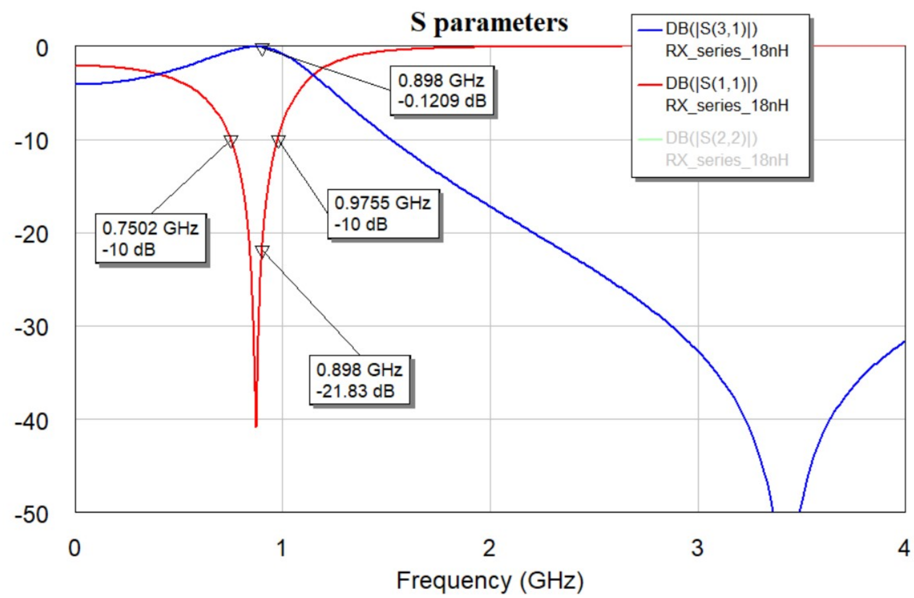

Figure 3.24. S Parameters of RX Match with 18 nH Single Inductor, Plus Bonding and Trace Inductances, LNA load = 420 Ω

The RX matching in TX mode must show a higher impedance at the fundamental frequency for proper TX operation. In TX mode, the LNA port is shunted to GND by an internal switch with 10 Ω series resistance.

In this case, a large-valued RX inductor is connected parallel to the GND (through a 10 Ω resistor) between the antenna port and output of the TX matching LPF. The large shunt inductor with absolute impedance > 150 Ω does not have considerable effects on the TX performance.

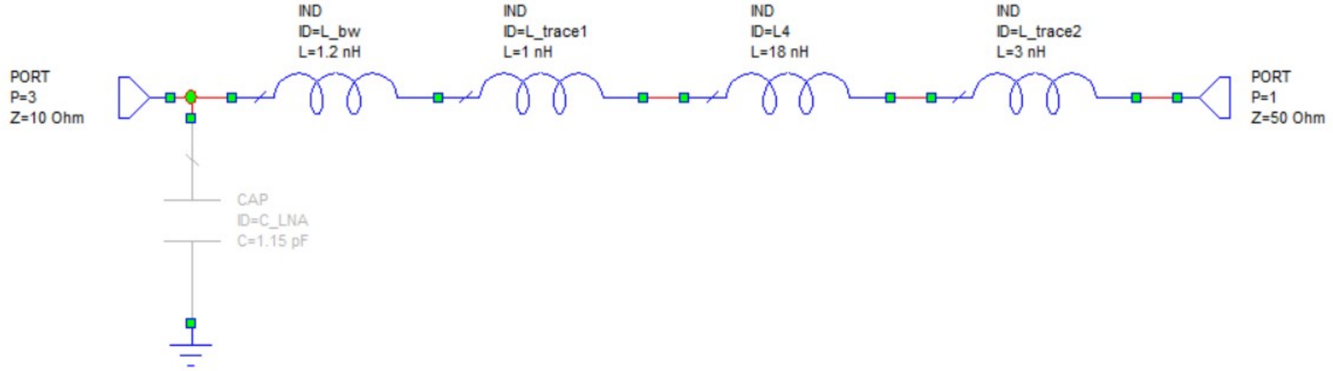

Figure 3.25. RX Match with Bonding Wire and PCB Trace Inductances in TX Mode

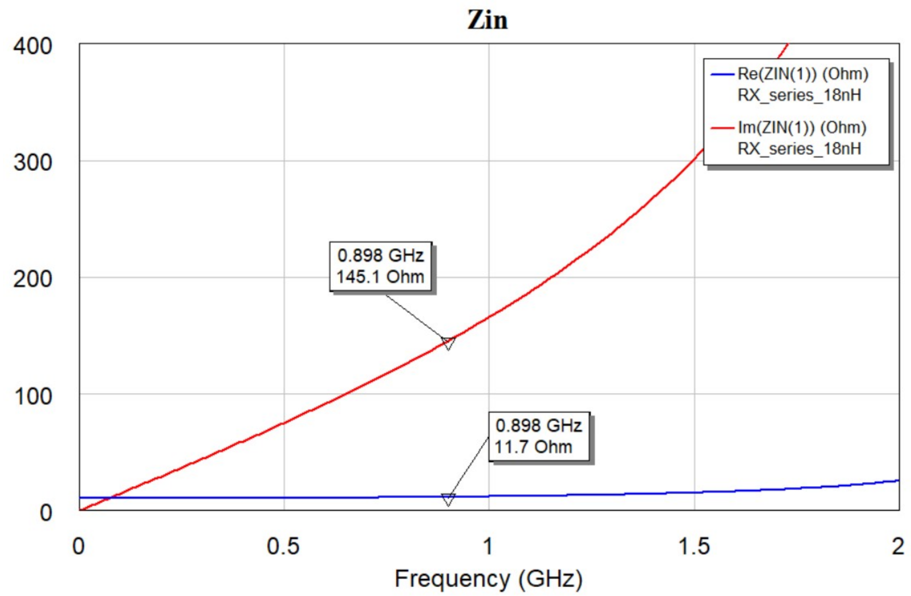

Figure 3.26. Input Impedance of RX Match of Single Inductor Plus Parasitics in TX Mode

3.10.3 TX-RX Direct-Tie Matching Network Design – Putting TX and RX Together (Example with EFR32xG23)

The separate TX and RX matching circuits are designed to show high TX output and high RX input impedances when the radio is operating in RX and TX modes, respectively. In other words, if these matching circuit paths are directly connected to each other, no performance degradation is expected in either TX or RX mode. At the same time, it is worth checking these in simulations, as this section of the document shows.

TX performance can be compensated by the last-connected TX matching shunt capacitor, if needed, to resonate out the additional shunt inductor load at the TX matching output, i.e., to achieve higher RX matching network input impedance in TX mode.

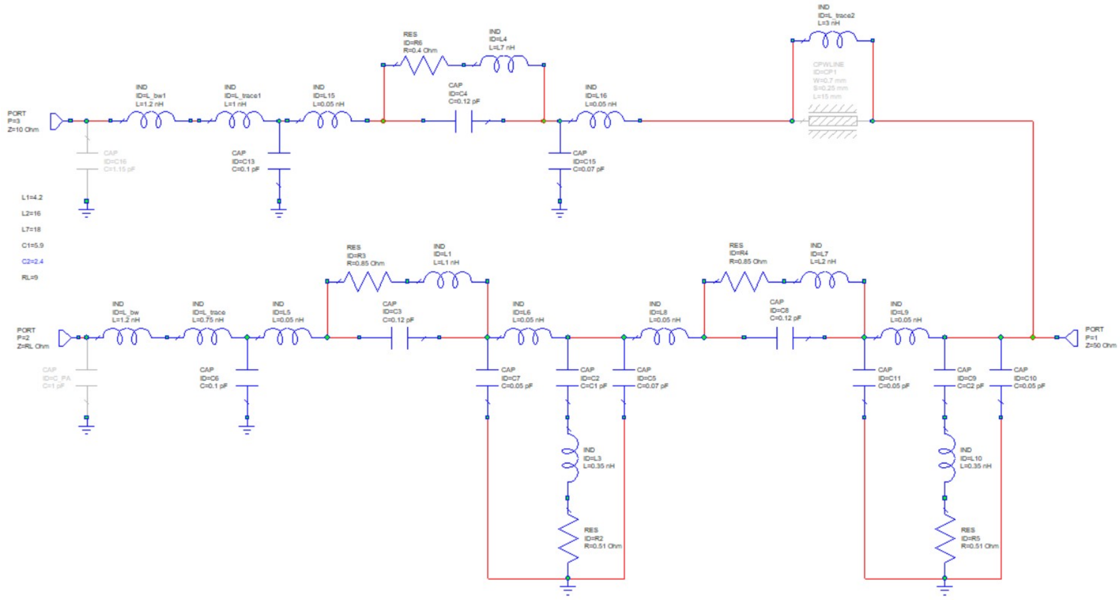

Figure 3.27. Simplified Simulation Model of TX-RX Direct-Tie Matching Network with Parasitics, TX Mode

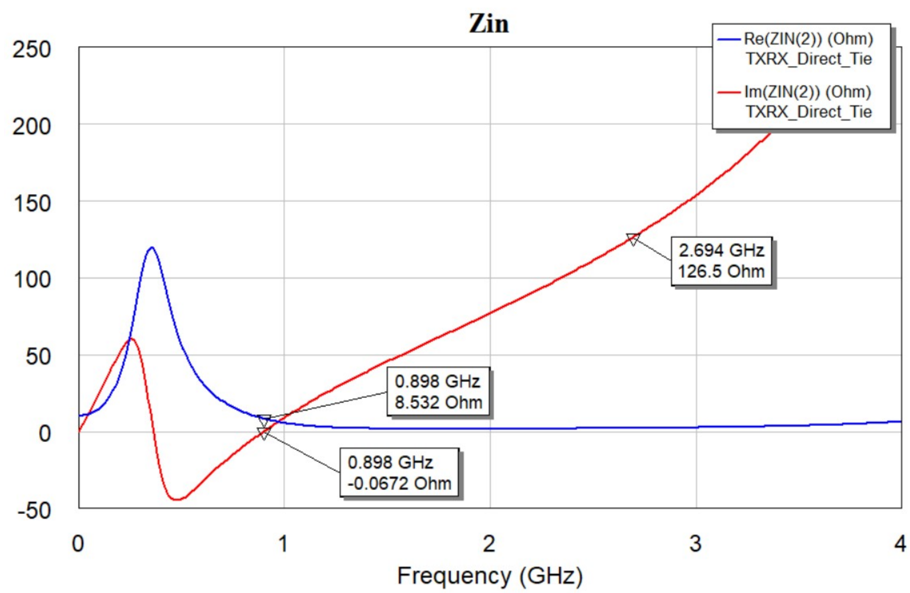

Figure 3.28. TX Match Input Impedance of TX-RX Direct-Tie Matching Network

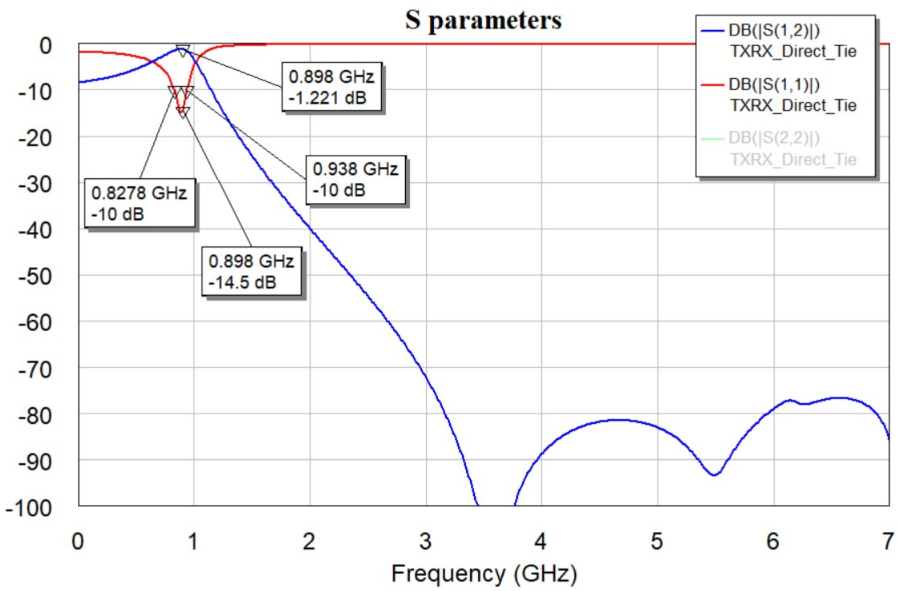

Figure 3.29. Transmit Mode S Parameters of TX-RX Direct-Tie Matching Network

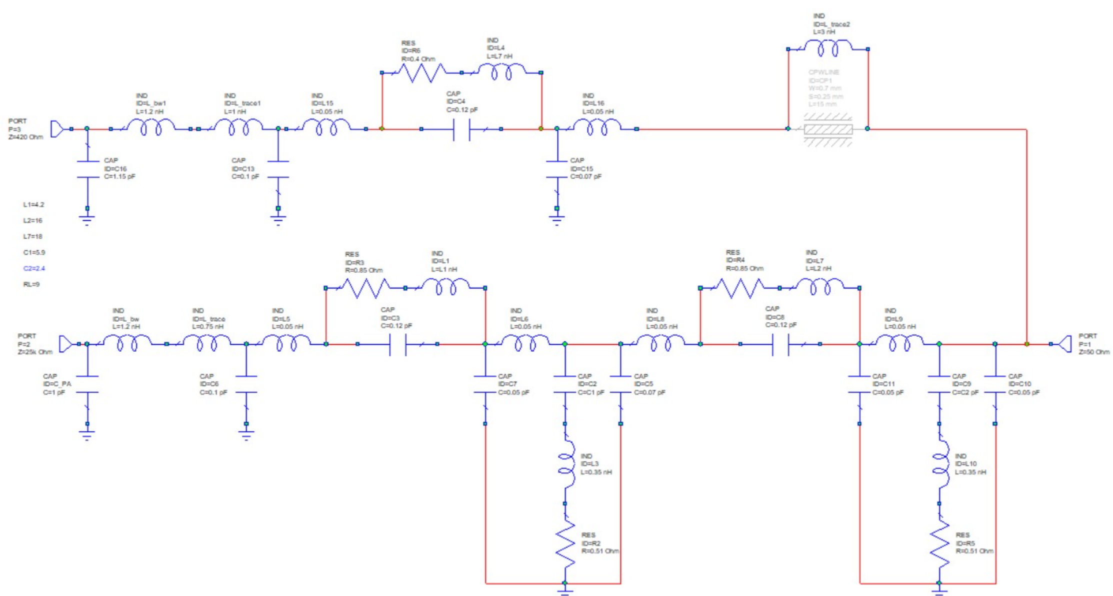

Figure 3.30. Simplified Simulation Model of TX-RX Direct-Tie Matching Network with Parasitics, RX Mode

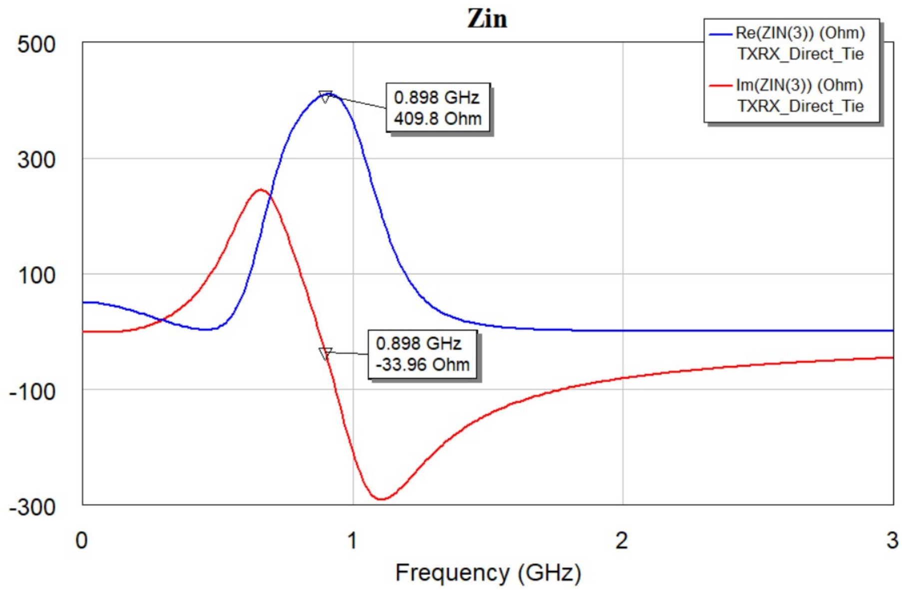

Figure 3.31. LNA Load Impedance of TX-RX Direct-Tie Matching Network, RX Mode

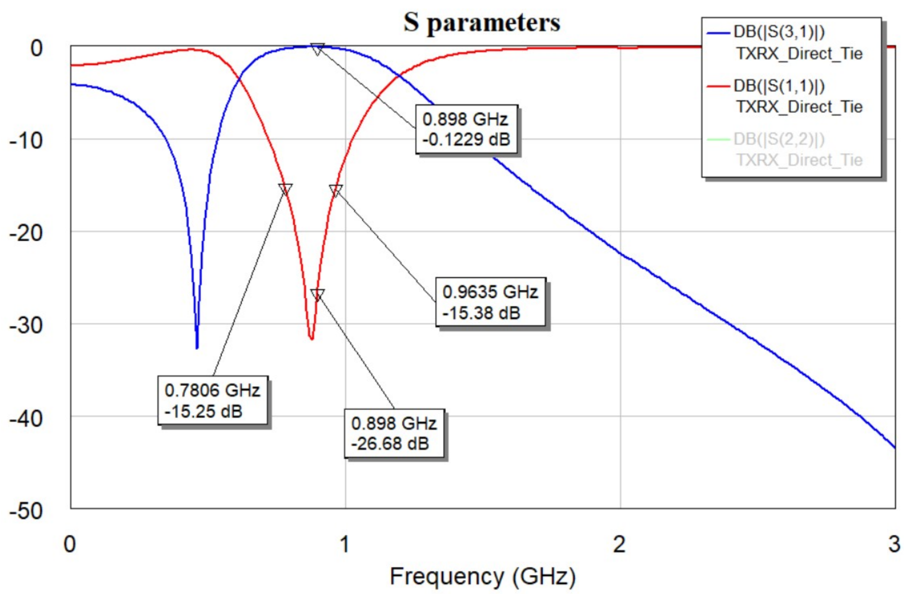

Figure 3.32. Receive Mode S Parameters of TX-RX Direct-Tie Matching Network

The simulation results shown in the plots above predict good RF performance on both TX and RX side.

Note: For more accurate simulation results, in addition to the circuit simulations of the matching network including the discrete parasitics model of PCB and SMD components, a full EM simulation can also be performed using the given PCB details and layout together with S-parameters files or Spice models of the SMD elements from component vendors.

3.10.4 EFR32xG25 TX/RX Matching Network Design

The matching network design procedure is similar to the design considerations and steps as presented previously for EFR32xG23, with the only expection that a balun function in the TX path is also required for the EFR32xG25 parts due to the differential TX outputs.

The on-chip RX front-end design, i.e., LNA, is identical between the EFR32xG23 and EFR32xG25 devices, so the RX matching network and its design principles are also exactly the same between them.

The EFR32xG25 TX matching network is built up from the following sections:

- Differential-to-differential PA matching network between the optimum PA load impedance and 50 Ω. Component designators from the figure of matching example below: LN1, LP1, CD1, CN1, CP1
- Differential-to-single-ended 1:1 impedance transformer balun which can be an external ceramic balun or can be implemented by discrete SMD components as well. From the figure below: CN3, LN2, LP2, CP3
- Low-pass filter between the single-ended port of the balun and antenna port: CM1-5, LM1, LM2

Additionally, the TX path requires dc blocking, therefore CN2 and CP2 from the example below are also applied in the matching network.

The RX matching network consists of a single series inductor, LR1, while it is recommenedd to also utilize a shunt capacitor at the RX pin to reduce the RX matching Q factor below ~ 300 MHz.

In a full discrete TX matching network design (i.e., without a ceramic balun) the first two functions of impedance matching and balun can also be combined into one discrete matching balun circuitry in order to further reduce the overall BOM of the matching network.

Silicon Labs provides the following matching network design examples:

- Radio board with full discrete matching design for 470 MHz
- Radio board with ceramic balun design for the 868 MHz band
- Radio board with ceramic balun design for the 915 MHz band
- Minimal BOM full discrete matching design for 868-915 MHz

The high-band solutions (868/915 MHz) with the external ceramic balun do have slightly better EVM margin at a given TX power level. See the comparison performance data in 5. Recommended Matching Network for EFR32xG25.

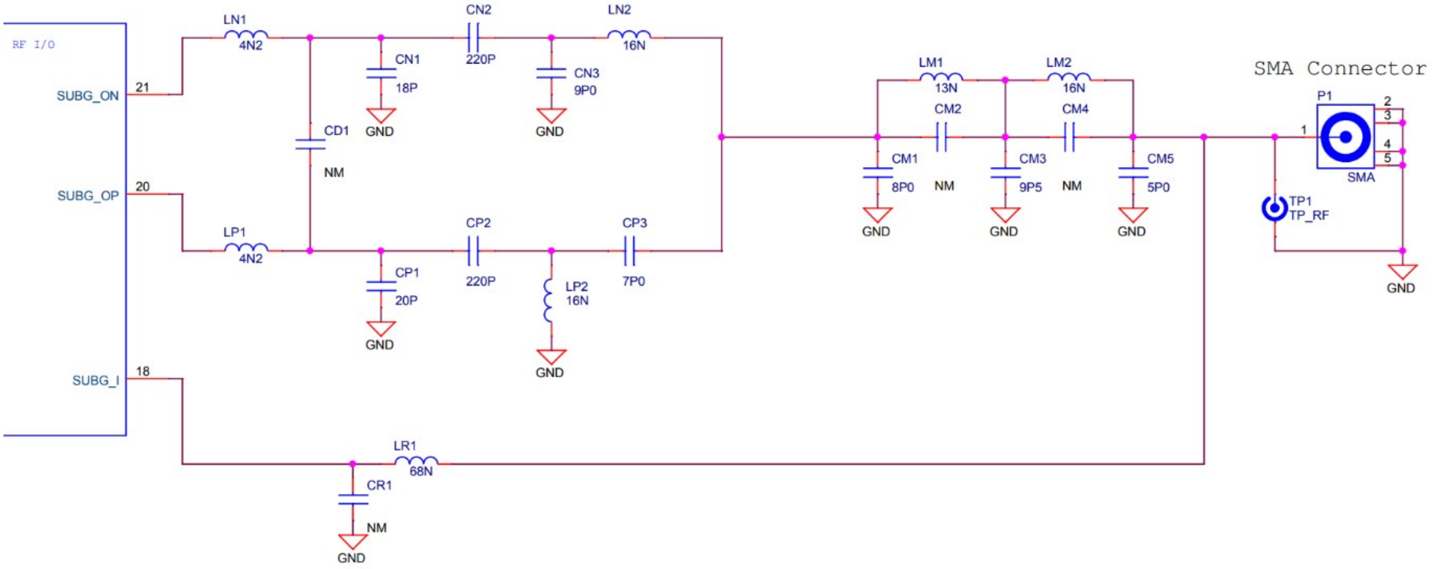

Figure 3.33. EFR32xG25 RF Discrete Matching Network Example (470 MHz)

The following charts show what TX/RX simulation results need to be considered to make sure the matching design is correct. The design example is gone through the 470 MHz radio board match.

S parameters used during the simulations:

- Port 1: Single-ended 50 Ω SMA port, TX mode
- Port 2: Differential TX ports
- Port 4: Single-ended 50 Ω SMA port, RX mode
- Port 6: Single-ended RX port

In TX mode, the LNA port is internally switched to the GND by a 10 Ω resistor and the PA needs to be matched to the optimal PA load impedance. In RX mode, the PA presents the PA off-mode impedance of 17 kΩ with a shunt 10 pF capacitor, and the LNA capacitance needs to be resonated out while transforming the RX impedance from 50 Ω up to ~400 – 1600 Ω depending on the frequency.

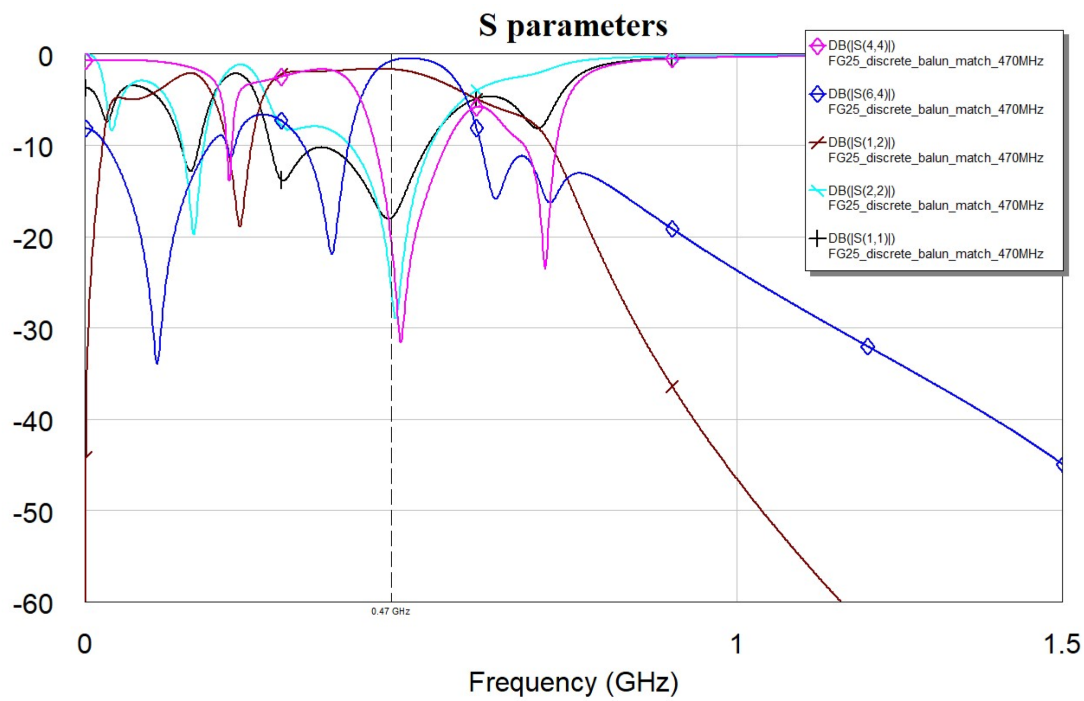

Figure 3.34. Simulated TX/RX S-Parameters, Discrete 470 MHz Match

The phase difference (ideally, 180°) is also simulated between the TX ports to check the balun function.

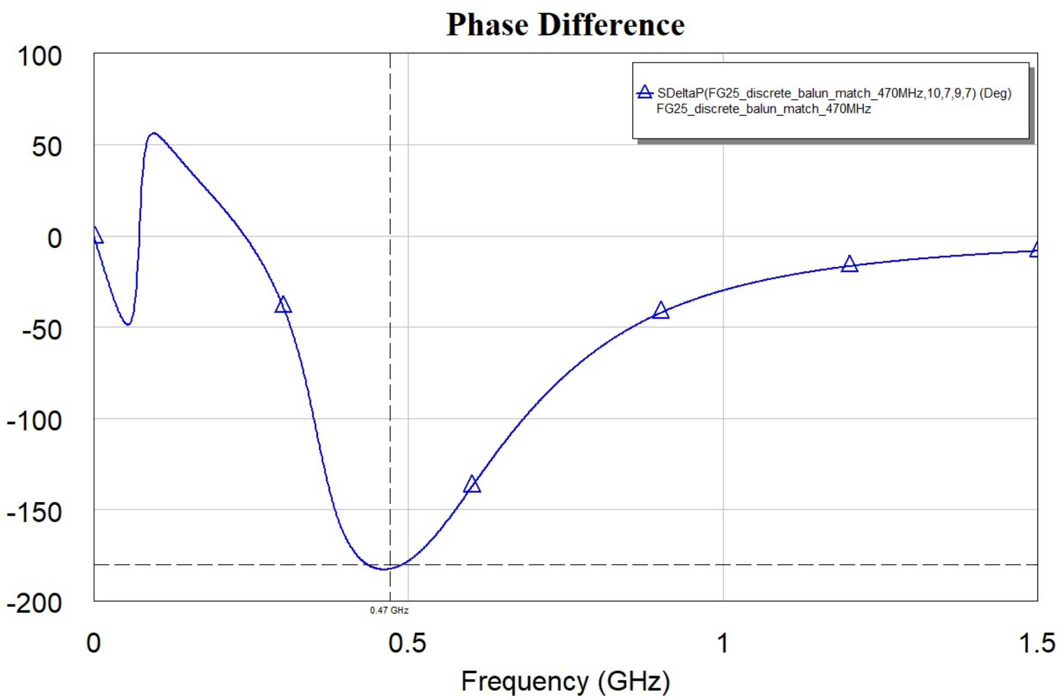

Figure 3.35. Simulated TX Phase Difference between the TX Ports, Discrete 470 MHz Match

The input impedance of the TX matching network must show the optimum PA load impedance at the TX ports, as the simulation result shows in the figure below.

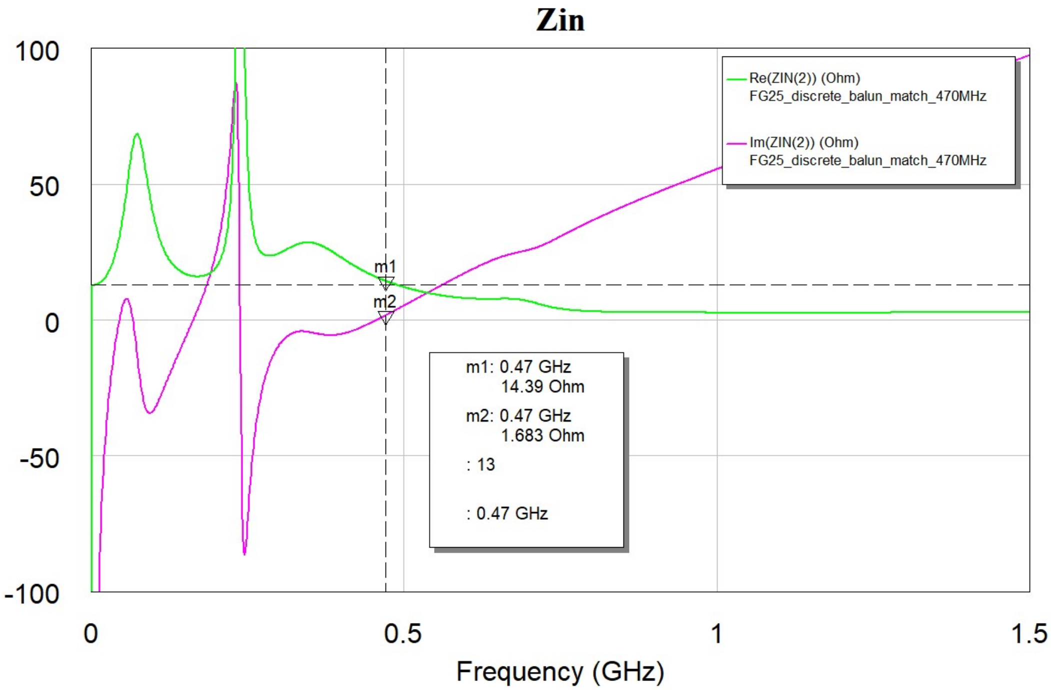

Figure 3.36. Simulated TX Input Impedance at the TX Ports, Discrete 470 MHz Match

4. Recommended Matching Network for EFR32xG23/28

Silicon Labs reference radio boards use SMD0201 components for the matching network design, on a 4-layer PCB. The overall RF performance of the matching network also depends on the layout drawing. As a result, it is recommended to follow the layout guidelines summarized in *AN928.2: EFR32 Series 2 Layout Design Guide*.

The typical RF impedance matching network topology for any sub-GHz applications is shown in the figure below, where all components do not need to be populated, but different BOM options are required for different frequencies and output power level requirements. For example, C4 is only required at 169 MHz to limit the Q of the RX matching network, while C6 is only required for +20 dBm TX power in the 868 and 915 MHz frequency bands to limit the 3rd harmonic current and therefore ensure radiated harmonic compliance with comfortable margins. For lower power levels, neither L3 and C3 components are required. The SUBG_O0 and SUBG_O1, SUBG_I0 and SUBG_I1 ports are identical. Therefore, use the same RF matching network for both RF pair ports. Series dc-blocking capacitor of C5 in the TX path is required, while dc decoupling in the RX path is not needed. The unused RF ports should be left floating, i.e., with an open load.

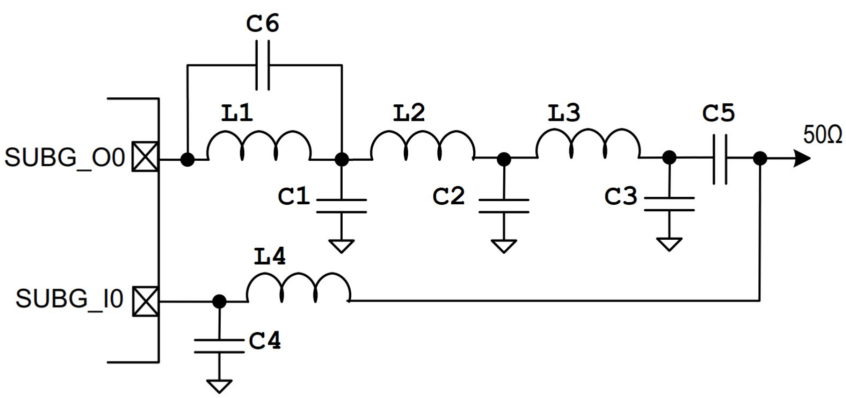

Figure 4.1. Typical RF Matching Network Circuit

The summary of the single-band matching network component values with conducted measurement data is provided in the tables below.

Table 4.1. Summary of EFR32xG23 Matching Component Values vs. Frequency

Freq. Band	P_{OUT}	PAVDD	L1	L2	L3	L4	C1	C2	C3	C4	C5	C6
169 MHz	10 dBm	1.8 V	22 nH	47 nH	51 nH	220 nH	33 pF	27 pF	15 pF	2.6 pF	220 pF	N.M.
	20 dBm	3.3 V	12 nH	40 nH	51 nH	220 nH	62 pF	27 pF	16 pF	2.6 pF	220 pF	N.M.[1]
315 MHz	10 dBm	1.8 V	22 nH	47 nH	0R	150 nH	15 pF	9.1 pF	N.M.	N.M.	220 pF	N.M.
	20 dBm	3.3 V	9.1 nH	39 nH	36 nH	150 nH	22 pF	8.2 pF	N.M.	N.M.	220 pF	N.M.
434 MHz	10 dBm	1.8 V	20 nH	39 nH	0R	82 nH	8.5 pF	4.6 pF	N.M.	N.M.	220 pF	N.M.
	14 dBm	1.8 V	10 nH	33 nH	0R	82 nH	15 pF	5.6 pF	N.M.	N.M.	220 pF	N.M.
	20 dBm	3.3 V	10 nH	33 nH	0R	82 nH	15 pF	5.6 pF	N.M.	N.M.	220 pF	N.M.
470 MHz[2]	17 dBm	3.3 V	5.1 nH	9.1 nH	30 nH	68 nH	N.M.	10 pF	7.6 pF	N.M.	220 pF	N.M.
	20 dBm	3.3 V	7.5 nH	0R	30 nH	68 nH	N.M.	13 pF	4.3 pF	N.M.	220 pF	N.M.
650 MHz	20 dBm	3.3 V	7 nH	0R	11 nH	30 nH	N.M.	16 pF	5.6 pF	N.M.	220 pF	N.M.
780 MHz	20 dBm	3.3 V	4 nH	15 nH	0R	24 nH	8.7 pF	3 pF	N.M.	N.M.	220 pF	N.M.
868 MHz	20 dBm	3.3 V	1.5 nH	1.3 nH	13 nH	18 nH	N.M.	7.2 pF	1.3 pF	N.M.	220 pF	1.9 pF
915 MHz												

Freq. Band	P_{OUT}	PAVDD	L1	L2	L3	L4	C1	C2	C3	C4	C5	C6
868 MHz	14 dBm	1.8 V	4.2 nH	16 nH	0R	18 nH	5.9 pF	2.1 pF	N.M.	N.M.	220 pF	N.M.
915 MHz												

Note:
1. If the 169 MHz 20 dBm matching is used with P_{OUT} reduced to 10-15 dBm, the 2nd harmonic emission might become marginal for the -36 dBm ETSI limit. Using C6=15 pF can resolve the issue but at the expense of 0.7-0.8 dB lower fundamental power.
2. The 470 MHz match is tuned for the 470-510 MHz frequency band.

Table 4.2. Summary of EFR32xG28 Matching Component Values vs. Frequency

Freq. Band	P_{OUT}	PAVDD	L1	L2	L3	L4	C1	C2	C3	C4	C5	C6
169 MHz	10 dBm	1.8 V	22 nH	47 nH	51 nH	220 nH	33 pF	27 pF	15 pF	2.6 pF	220 pF	N.M.
	20 dBm	3.3 V	12 nH	40 nH	51 nH	220 nH	62 pF	27 pF	16 pF	2.6 pF	220 pF	N.M.
315 MHz	10 dBm	1.8 V	22 nH	47 nH	0R	150 nH	15 pF	9.2 pF	N.M.	N.M.	220 pF	N.M.
434 MHz	10 dBm	1.8 V	20 nH	39 nH	0R	82 nH	8.5 pF	4.6 pF	N.M.	N.M.	220 pF	N.M.
	14 dBm	1.8 V	10 nH	33 nH	0R	82 nH	15 pF	5.6 pF	N.M.	N.M.	220 pF	N.M.
470 MHz	17 dBm	3.3 V	5.1 nH	9.1 nH	30 nH	68 nH	N.M.	10 pF	7.6 pF	N.M.	220 pF	N.M.
	20 dBm	3.3 V	1.5 nH	3.1 nH	20 nH	68 nH	N.M.	20 pF	7.1 pF	N.M.	220 pF	N.M.
868 MHz	20 dBm	3.3 V	1.5 nH	1.3 nH	13 nH	18 nH	N.M.	7.6 pF	1.3 pF	N.M.	220 pF	2.0 pF
915 MHz												
868 MHz	14 dBm	1.8 V	4.2 nH	16 nH	0R	18 nH	6.3 pF	2.1 pF	N.M.	N.M.	220 pF	N.M.
915 MHz												
868 MHz	10 dBm	1.8 V	5.6 nH	13 nH	0R	18 nH	5.1 pF	3.1 pF	N.M.	N.M.	220 pF	N.M.
915 MHz												

Table 4.3. Summary of Matching RF Conducted Performance

Freq. Band	P_{OUT_GOAL}	PAVDD	Raw Power Setting	P_{FUND}	P_{2nd}	P_{3rd}	I_{TX}	RX Sens. BER < 0.1%	2-GFSK data rate/freq. deviation
169 MHz	10 dBm	1.8 V	63	10.5 dBm	-44.3 dBm	-59.5 dBm	12.4 mA	-124.2 dBm	2.4 kbps / 1.2 kHz
	20 dBm	3.3 V	240	20.4 dBm	-36.1 dBm	-52.1 dBm	82.4 mA	-124.2 dBm	
300 MHz[1]	20 dBm	3.3 V	240	19.9 dBm	-42.4 dBm	-52.5 dBm	71.1 mA	-110.6 dBm	100 kbps / 50 kHz
315 MHz	10 dBm	1.8 V	63	10.4 dBm	-51.2 dBm	-47.3 dBm	14.0 mA	-114.5 dBm	38.4 kbps / 20 kHz
	20 dBm	3.3 V	240	20.1 dBm	-44.7 dBm	-55.4 dBm	71.3 mA	-110.4 dBm	100 kbps / 50 kHz

Freq. Band	P_{OUT_GOAL}	PAVDD	Raw Power Setting	P_{FUND}	P_{2nd}	P_{3rd}	I_{TX}	RX Sens. BER < 0.1%	2-GFSK data rate/freq. deviation
434 MHz	10 dBm	1.8 V	63	10.4 dBm	-48.9 dBm	-49.9 dBm	13.4 mA	-110.8 dBm	100 kbps / 50 kHz
	14 dBm	1.8 V	187	14.0 dBm	-50.0 dBm	-46.0 dBm	26.4 mA	-110.4 dBm	
	20 dBm	3.3 V	240	19.8 dBm	-43.7 dBm	-41.5 dBm	76.8 mA	-110.6 dBm	
470 MHz	17 dBm	3.3 V	170	17.2 dBm	-46.6 dBm	-44.3 dBm	41.5 mA	-120.5 dBm	10 kbps / 5 kHz
470 MHz	20 dBm	3.3 V	240	19.7 dBm	-49.2 dBm	-38.9 dBm	73.7 mA	-120.6 dBm	
490 MHz	17 dBm	3.3 V	135	17.4 dBm	-48.1 dBm	-49.2 dBm	44.2 mA	-120.5 dBm	
490 MHz	20 dBm	3.3 V	240	20.2 dBm	-52.0 dBm	-40.4 dBm	84.2 mA	-120.4 dBm	
510 MHz	20 dBm	3.3 V	240	19.7 dBm	-42.0 dBm	-41.5 dBm	82.1 mA	-120.4 dBm	
650 MHz	20 dBm	3.3 V	240	20.6 dBm	-41.0 dBm	-42.0 dBm	87.7 mA	- 110.5 dBm	100 kbps / 50 kHz
780 MHz[2]	20 dBm	3.3 V	240	20.0 dBm	-42.8 dBm	-45.9 dBm	82.3 mA	-110.0 dBm	100 kbps / 50 kHz
868 MHz	20 dBm	3.3 V	240	20.2 dBm	-38.3 dBm	-46.1 dBm	85.8 mA	-113.9 dBm	38.4 kbps / 20 kHz
915 MHz	20 dBm	3.3 V	240	20.3 dBm	-39.6 dBm	-50.6 dBm	90.0 mA	-97.1 dBm	2 Mbps / 500 kHz
868 MHz	14 dBm	1.8 V	240	14.3 dBm	-47.2 dBm	-48.0 dBm	26.5 mA	-113.9 dBm	38.4 kbps / 20 kHz
915 MHz	14 dBm	1.8 V	240	14.6 dBm	-52.2 dBm	-53.9 dBm	27.8 mA	-97.1 dBm	2 Mbps / 500 kHz

Note:

1. At 300 MHz, the 315 MHz matching network is used with C4=0.2 pF mounted.
2. At 780 MHz, with L2=13 nH change, Pfund=19.8 dBm with 2 dB lower 2nd harmonic and TX current = 73.1 mA. Other measured RF parameters are the same.

HPA is used above +14 dBm TX power level, while LPA is used equal or below +14 dBm TXP. PAVDD is the dc supply voltage level at the PAVDD pin which is further regulated to supply the PA within the part. SMD0201 components are used in the RF matching network. Radio boards are available with inductors from the LQP03HQ series to provide the best RF performance.

Measurement results shown in the table above are taken on designs using EFR32xG23 parts.

5. Recommended Matching Network for EFR32xG25

Silicon Labs reference radio boards use SMD0201 components for the matching network design, on a 4-layer PCB. The overall RF performance of the matching network also depends on the layout drawing. As a result, it is recommended to follow the layout guidelines summarized in https://www.silabs.com/documents/public/application-notes/an928.2-efr32-series2-layout-design-guide.pdf.

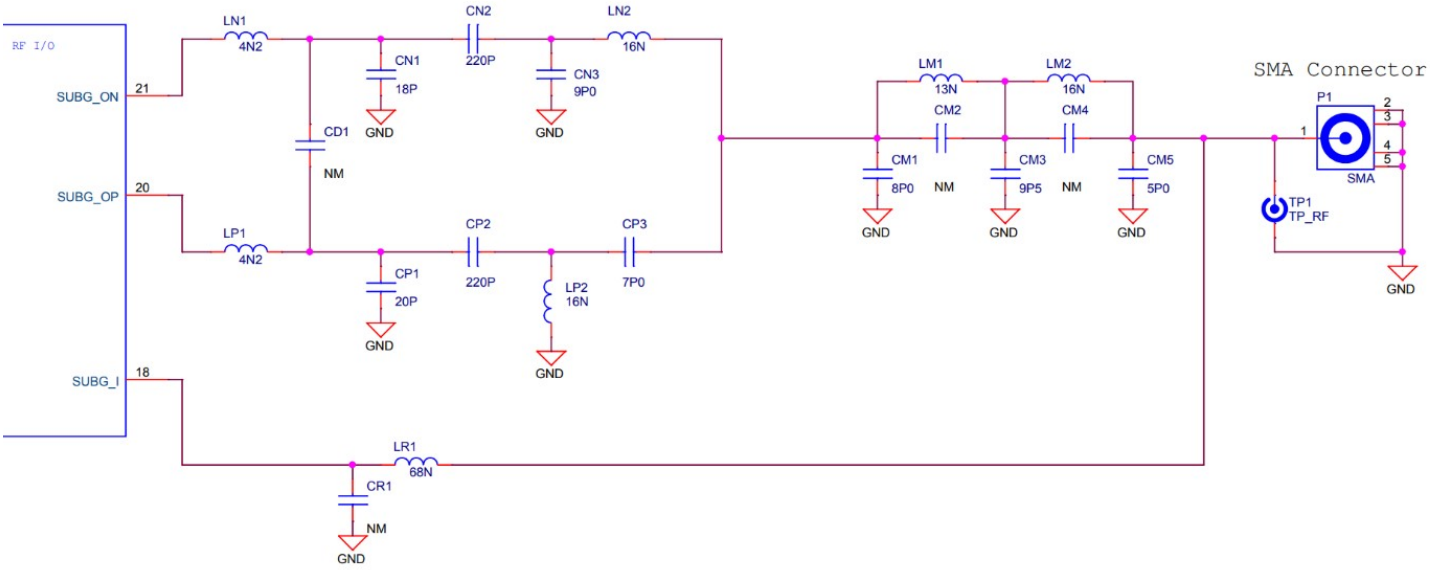

Figure 5.1. Recommended Full Discrete Match for 470 MHz

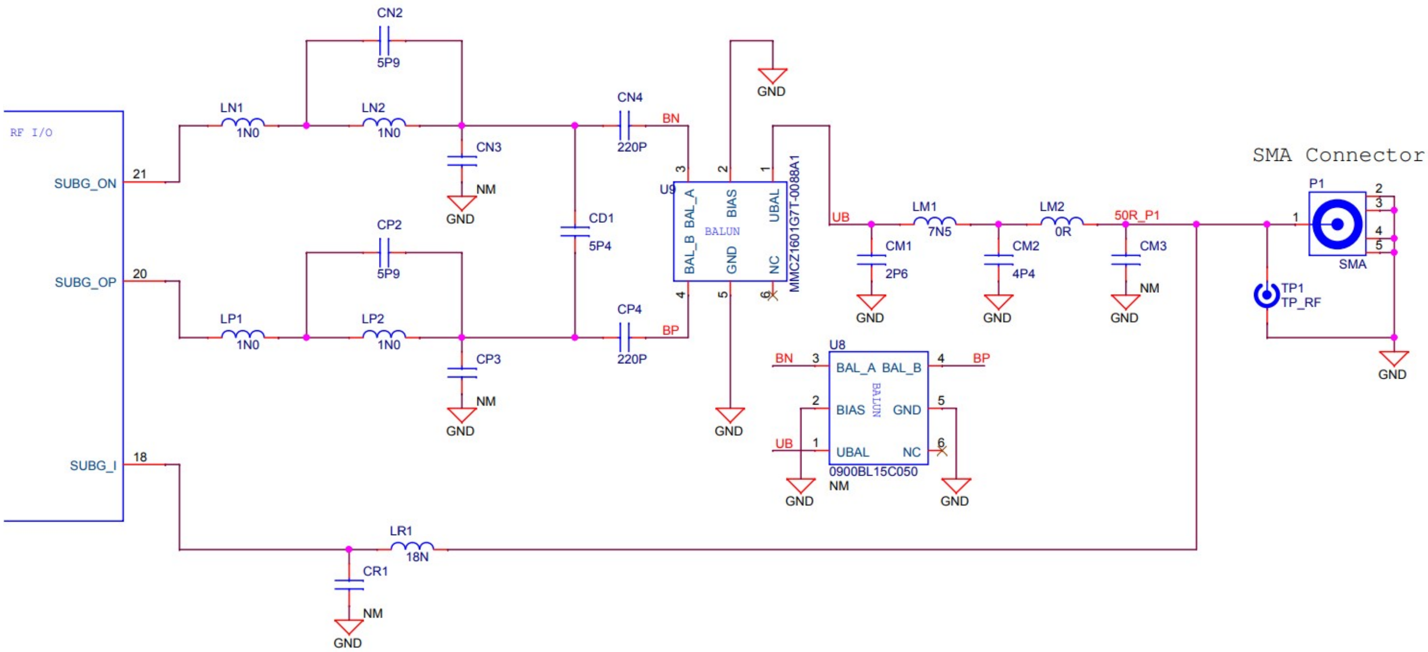

Figure 5.2. Recommended Ceramic Balun Match for 868 MHz

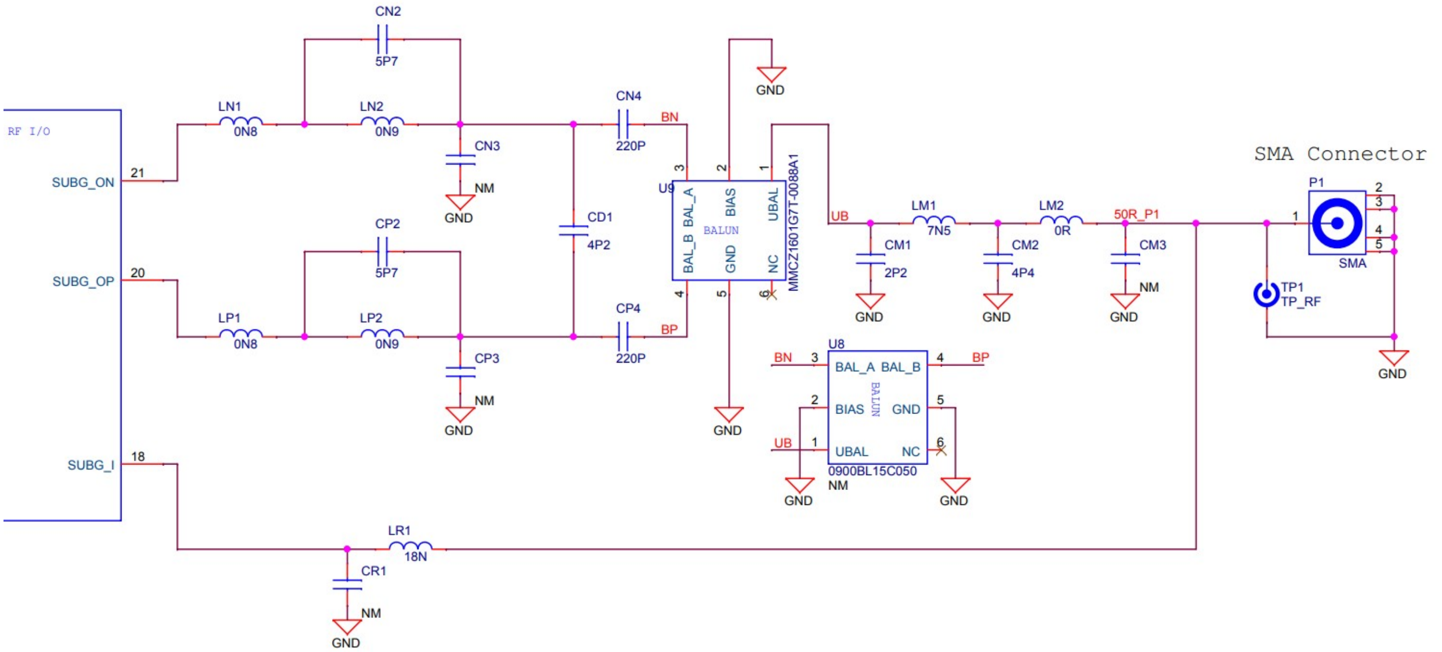

Figure 5.3. Recommended Ceramic Balun Match for 915 MHz

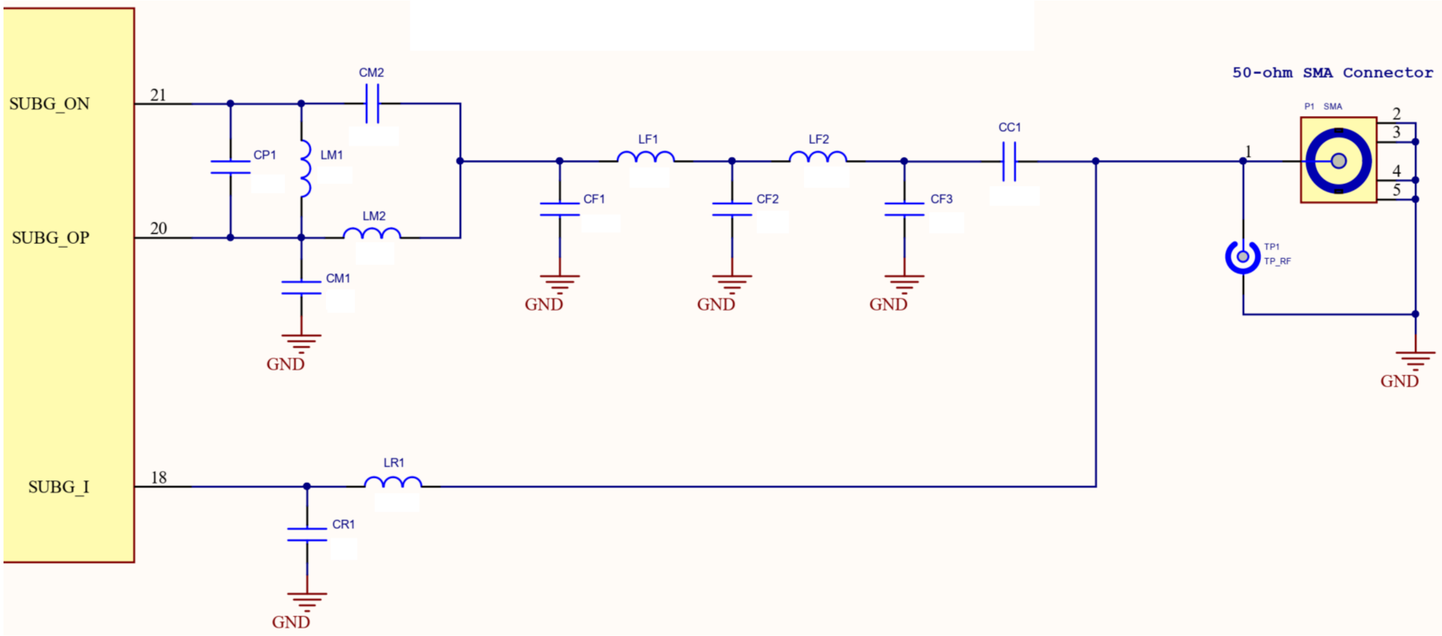

Figure 5.4. Recommended Minimal-BOM Full Discrete Matching Balun Schematic

Table 5.1. Recommended Component Values for the Minimal-BOM Full Discrete Matching Balun for the 863 – 928 MHz Frequency Band

Name	LM1	LM2	CM1	CM2	LF1-2	CF1	CF2	CF3	CP1	LR1
#1	22 nH	6.2 nH	11 pF	4.3 pF	6.8 nH	4.7 pF	10 pF	5.1 pF	1.5 pF	20 nH
#2	22 nH	7.5 nH	9.1 pF	4.3 pF	6.8 nH	4.7 pF	10 pF	5.1 pF	1.5 pF	20 nH
Note 1: CC1 = 100 pF; CR1 = N.M. (not mounted).										

		FSK PA CW with full power		OFDM PA CW with full power		OFDM 868 MHz MCS6 OPT4		OFDM 915 MHz MCS6 OPT1		OFDM 923 MHz MCS6 OPT2	
		868 MHz	915 MHz	868 MHz	915 MHz	Channel Power [dBm]	EVM [dB]	Channel Power [dBm]	EVM [dB]	Channel Power [dBm]	EVM [dB]
#1	Fund. [dBm]	14.3	14.5	17.5	18.2	17.3	-20.5	17.2	-20.6	17.2	-21.1
	H2 [dBm]	-47.0	-54.8	-50.8	-57.6	16.5	-23.0	16.1	-23.3	16.4	-23.9
	H3 [dBm]	-47.7	-49.3	-60	-60	13.4	-31.1	13.7	-32.4	13.3	-32.5
	H4 [dBm]	-60	-60	-60	-60	-5.5	-34.9	-6.1	-37.3	-5.7	-37.8
	H5 [dBm]	-53.1	-60	-60	-60	N/A	N/A	N/A	N/A	N/A	N/A
	TX current [mA]	80.2	80.2	218.1	220.8	N/A	N/A	N/A	N/A	N/A	N/A
	RX sens. [dBm]	-109.9	-109.9	N/A	N/A	N/A	N/A	N/A	N/A	N/A	N/A
#2	Fund. [dBm]	15.1	15.4	17.5	18.1	17.1	-19.8	17.0	-19.9	17.1	-20.6
	H2 [dBm]	-44.6	-48.5	-49.7	-55.7	16.4	-22.1	16.1	-22.4	16.3	-23.1
	H3 [dBm]	-47.4	-48.9	-60	-60	13.4	-30.2	13.5	-31.5	13.2	-31.8
	H4 [dBm]	-60	-60	-60	-60	-5.4	-35.2	-6.4	-37.2	-5.8	-38.1
	H5 [dBm]	-55.6	-60	-60	-60	N/A	N/A	N/A	N/A	N/A	N/A
	TX current [mA]	78.1	78.2	214.3	216.8	N/A	N/A	N/A	N/A	N/A	N/A
	RX sens. [dBm]	-109.7	-109.8	N/A	N/A	N/A	N/A	N/A	N/A	N/A	N/A
BRD4270B	Fund. [dBm]	14.9	15.4	17.3	18.5	16.4	-21.8	17.0	-20.2	17.4	-20.4
	H2 [dBm]	-34.1	-39.2	-60	-53.3	16.1	-22.6	16.3	-23.3	16.3	-23.5
	H3 [dBm]	-44.5	-54.0	-60	-60	14.4	-27.2	15.0	-25.8	15.5	-25.9
	H4 [dBm]	-60	-60	-60	-60	-6.0	-33.6	-5.9	-36.9	-5.4	-37.8
	H5 [dBm]	-60	-60	-60	-60	N/A	N/A	N/A	N/A	N/A	N/A
	TX current [mA]	78.1	77.5	209.9	214.5	N/A	N/A	N/A	N/A	N/A	N/A
	RX sens. [dBm]	-109.6	-109.6	N/A	N/A	N/A	N/A	N/A	N/A	N/A	N/A

Note 1: TX current consumption is measured at the 5 V net in CW mode with full power
Note 2: FSK sensitivity measured with 2-FSK, 100 kbps data rate, 50 kHz deviation at BER < 0.1%

Figure 5.5. Performance Numbers of the Minimal-BOM Full Discrete 863 – 928 MHz Matching Balun in Comparison with the 915 MHz Radio Board Match with Ceramic Balun

The design package of the reference board with the minimal-BOM full discrete matching network is available under the Silicon Labs hardware design examples github page: https://github.com/SiliconLabs/hardware_design_examples.

For the performance numbers of the 470 MHz full discrete, 868 MHz and 915 MHz ceramic balun matching networks available on the Silicon Labs radio boards, refer to the datasheet specifications.

6. Multi-Band Matching Network Design for EFR32xG23/28

This section of the application note provides design examples dedicated to applications that require dual-, multi- or wideband frequency coverage in the sub-GHz region. Design methodology with measurement results is presented.

The matching network naming follows a similar approach as was applied for Series 1 devices and documented in the application note of AN1180: EFR32 Series 1 sub-GHz Discrete Matching Solutions.

- Wideband match: a frequency band coverage with large BW (e.g., 315–434 MHz).
- Dual-band match: 2 far-placed frequency bands with narrow BW of each (e.g., 434 and 868–915 MHz).
- Multi-band match: 2-far placed frequency bands with a narrow and a large BW coverage (e.g., 315–434 and 868–928 MHz).
- Dual-wideband match: 2 far-placed frequency bands with large BW coverage on both (e.g., 300–510 and 780–928 MHz).

6.1 Wideband Matching Network

A wideband matching network can cover frequency bands that fall close to each other, i.e., covers one frequency band with relatively large bandwidth (e.g., 315–434 MHz).

6.1.1 Design Example for 315–510 MHz RX-only Wideband Match

This section drives the reader through the design steps for a wideband matching network.

Step 1: Determine the theoretical BW limit and required number of matching components

Bode and Fano's VSWR versus decrement factor (δ) curves help determine the required number of matching elements based on the load conditions.

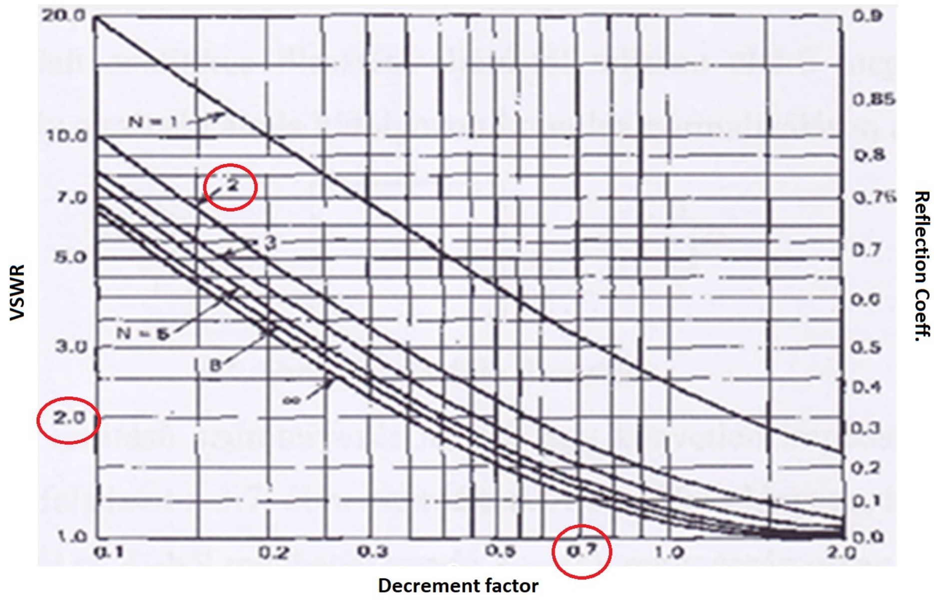

Figure 6.1. Fano's VSWR vs. Decrement Factor Curves

The parameter of N is the order of the matching network. The decrement factor is inversely proportional with the relative BW and the load Q as the following equation shows.

$$\delta = \frac{Q_{BW}}{Q_L} = \frac{\sqrt{\omega_1 \omega_2}}{\omega_2 - \omega_1} \frac{1}{Q_L} = \frac{\omega_0}{\omega_2 - \omega_1} \frac{1}{Q_L} = \frac{\omega_0}{2\pi * BW_f * Q_{RC}}$$

Q_{BW}: -3 dB bandwidth definition of a resonant circuit (~ f_c/BW_{3dB}).

Q_L or Q_{RC}: stored energy definition of a load (~ $2\pi f_c$ * "energy stored/power loss") (e.g., for a parallel RC load: $Q_{RC}=\omega_c RC$).

From the figure above, the decrement factor of 0.7 yields S11 better than -10 dB (VSWR between 1.5 and 2) with using a 2-element match.

Design example (e.g., LNA match) with R = 1.2 kΩ and C = 1.1 pF (with parallel RC model) at f_c = 410 MHz:

$$Q_{RC} = 2\pi f_c RC = 2\pi * 410 * 10^6 * 1200 * 1.1 * 10^{-12} = 3.4$$

$$BW_{theory} = \frac{f_c}{\delta * Q_{RC}} = \frac{410 * 10^6}{0.7 * 3.4} = 172 \quad MHz$$

Targeted BW in the sub-GHz low-band is **~200 MHz** to cover frequencies from 315 up to 510 MHz, so it is recommended to increase the number of matching elements.

Decrement factor of 0.5 yields S11 better than -10 dB with a **4-element** match and the theoretical BW is larger than 200 MHz as the calculation shows below.

$$BW_{theory} = \frac{f_c}{\delta * Q_{RC}} = \frac{410 * 10^6}{0.5 * 3.4} = 241 \quad MHz$$

Step 2: Calculate the matching network component values using Youla wideband match prototypes

A 2nd-order single-ended Youla wideband match prototype (given between equal generator [antenna model] and load [LNA model] real part impedances) is shown in the figure below.

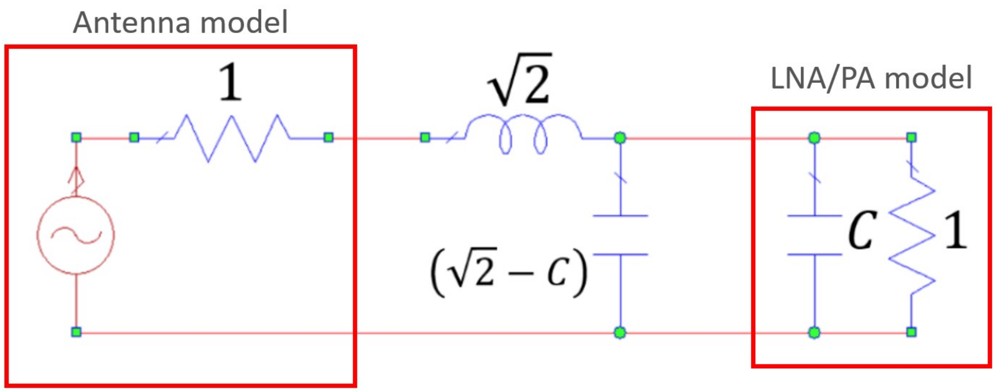

Figure 6.2. 2nd-order Youla Wideband Match Prototype

The relative units can be calculated as follows:

$$R = R_e;\ \omega = \omega_e;\ L_e = \frac{R_e}{\omega_e};\ C_e = \frac{1}{R_e\omega_e}$$

Design example (e.g., LNA match) with R = 1.2 kΩ and C = 1.1 pF at f_c = 410 MHz:

$$L_e = \frac{R_e}{\omega_e} = \frac{1200}{2\pi * 410 * 10^6} = 466 \quad nH$$

$$L_{ser1} = L_e\sqrt{2} = 659 \quad nH$$

$$C_e = \frac{1}{R_e\omega_e} = \frac{1}{1200 * 2\pi * 410 * 10^6} = 0.32 \quad pF$$

$$C_{par} = \left(\sqrt{2} - \frac{C}{C_e}\right) * C_e = C_e\sqrt{2} - C = 0.32 * \sqrt{2} - 1.1 = -0.64 \quad pF$$

These above give the max flat S11 curve between equal 1.2 kΩ and 1.2 kΩ impedances and the -3 dB point at f_c, as the simulation also proves it below.

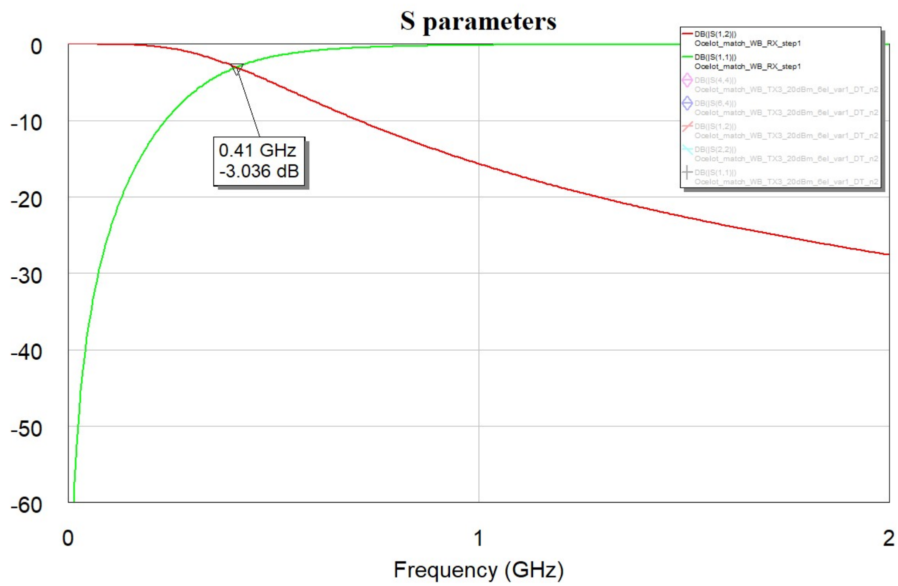

Figure 6.3. Max Flat S11 Curve Based on Youla Prototype

2nd-order prototype for non-equal generator and load impedance real parts are not analytically synthesized yet, instead approximately calculated as shown in the next steps.

Step 3: Estimate the matching network component values between non-equal real-part impedances

To have the match between 1.2 kΩ and R_{ANT}=50 Ω, let's have two stages of 2nd-order Youla prototypes in series, where the internal impedance point is the geometric mean of R=1.2 kΩ and R_{ANT}=50 Ω to minimize overall matching Q, and therefore maximize BW. This is needed because a 2-element only match would not satisfy the BW requirement as calculated in step 1 above.

$$R_{INT} = \sqrt{R * R_{ANT}} = 245 \quad \Omega$$

Design between R = 1.2 kΩ with C = 1.1 pF and R_{INT} = 245 Ω at f_c = 410 MHz:

$$L_{ser1} = L_e\sqrt{2} = 659 \quad nH$$

$$C_{par} = \left(\sqrt{2} - \frac{C}{C_e}\right) * C_e = C_e\sqrt{2} - C = 0.32 * \sqrt{2} - 1.1 = -0.64 \quad pF$$

Add L_x in parallel with the series inductor of L_{ser1} which matches between R = 1.2 kΩ and R_{INT} = 245 Ω.

$$L_x = \frac{\sqrt{R * R_{INT}}}{2\pi f_c} = \frac{\sqrt{1200 * 245}}{2\pi * 410 * 10^6} = 210.5 \quad nH$$

$$L_{ser} = \frac{L_{ser1} L_x}{L_{ser1} + L_x} = \frac{659 * 210.5}{659 + 210.5} = 159.5 \quad nH$$

Keep C_{par} = -0.64 pF as calculated between equal real-part impedances.

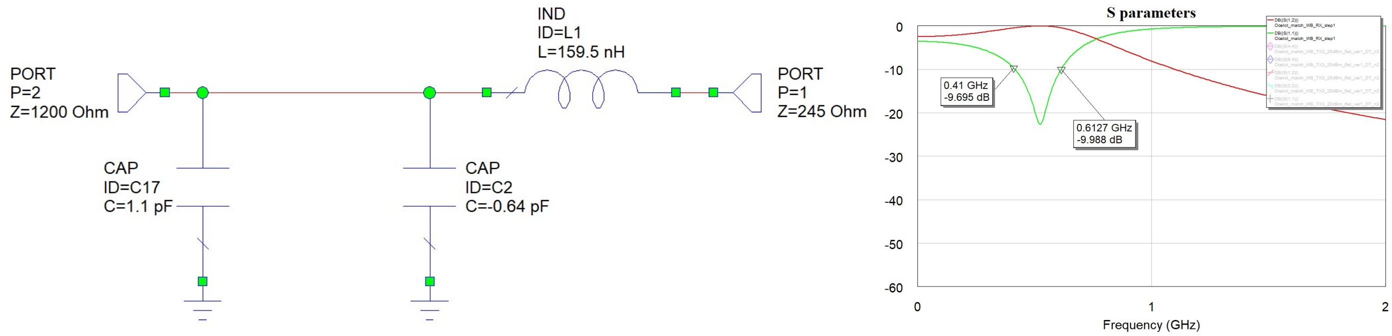

Figure 6.4. Estimated Wideband Match between Non-Equal Real-Part Impedances

These above give a good approximation for the wideband matching network component values but with a frequency offset observed. In order to compensate the frequency offset, the recommended matching component values should be re-calculated at the starting frequency of 300 MHz after the introduction of Lx.

1st stage matching design between R = 1.2 kΩ with C = 1.1 pF and R_{INT} = 245 Ω at f_s = 300 MHz:

L_{ser} = 218 nH

C_{par} = -0.47 pF

2nd stage match between R_{INT} = 245 Ω and R_{ANT} = 50 Ω at f_s = 300 MHz:

L_{ser} = 44.5 nH

C_{par} = 3.06 pF

Putting these together will provide the desired frequency characteristic response as shown below.

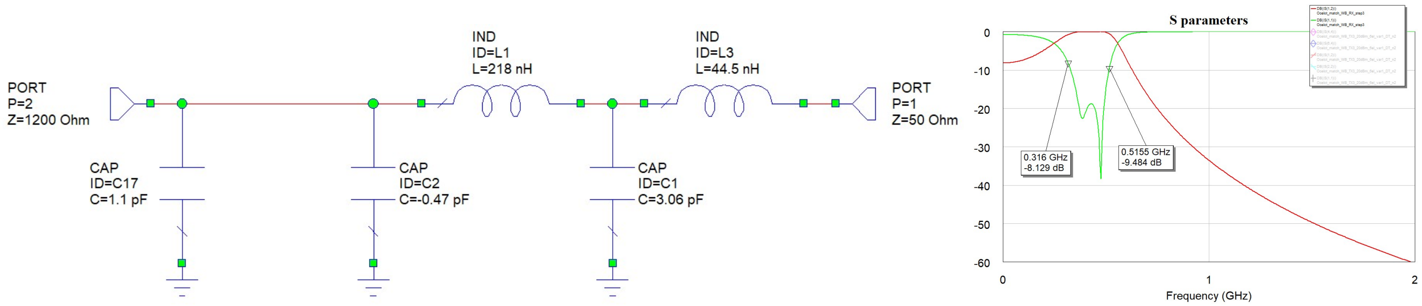

Figure 6.5. Re-Calculated Wideband Match between Non-Equal Real-Part Impedances

Step 4: Eliminating non-realizable matching elements

Unfortunately, C_{par} = -0.47 pF does not give a realizable solution.

Replacing C_{par} = -0.47 pF with a shunt inductor of 321 nH (resonates with 0.47 pF at the center 410 MHz) is a BW-reducing step due to the different frequency characteristic of an inductor. It is shown in the figure below on the left-hand side.

With tuned component values (shunt 570 nH → series 171 nH → shunt 3.26 pF → series 46.3 nH) between R = 1.2 kΩ with C = 1.1 pF and R_{ANT} = 50 Ω, this effect can even be compensated as shown in the figure below on the right-hand side.

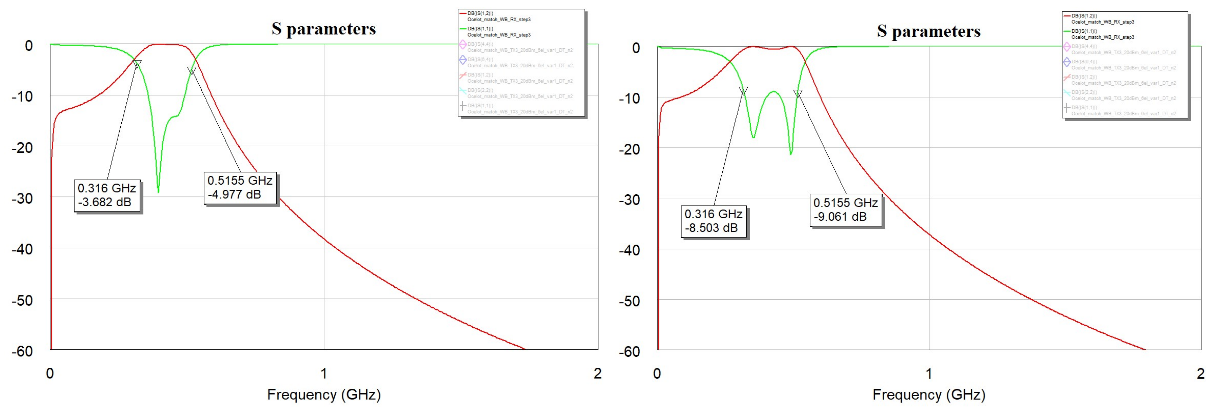

Figure 6.6. Frequency Response after Replacing Non-Realizable Components

Step 5: Tune the matching network including parasitics

The final matching network solution includes the following:

- Eliminating the first shunt inductor (previously, negative capacitance)
- Introducing chip (e.g., bonding wire, stray capacitance), external SMD and PCB parasitics
  - cause drop in real part of input impedance achievable by the matching circuit
  - propose to do the impedance match for R = 800 Ω, instead of 1.2 kΩ
    - this adds only about $1\text{-}S_{11}^2$=-0.18 dB mis-match loss
    - re-calculate the component values (ideal) with R = 800 Ω

Calculations with R = 800 Ω using ideal components:

1st stage matching network between R = 800 Ω with C = 1.1 pF and R_{INT} = 200 Ω at f_s = 300 MHz

L_{ser} = 156.8 nH

C_{par} = -0.16 pF, it would be replaced by a shunt 942 nH inductor (it can basically be eliminated without big impact)

2nd stage match between R_{INT} = 200 Ω and R_{ANT} = 50 Ω at f_s = 300 MHz

L_{ser} = 39.2 nH

C_{par} = 3.75 pF

Putting these above together and tuning with parasitics (not shown in the schematic) included:

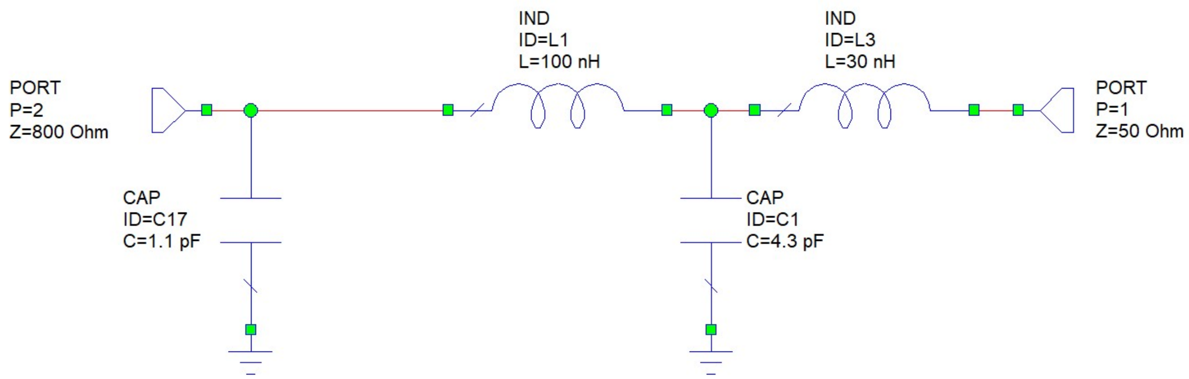

Figure 6.7. Final Wideband RX Matching Network

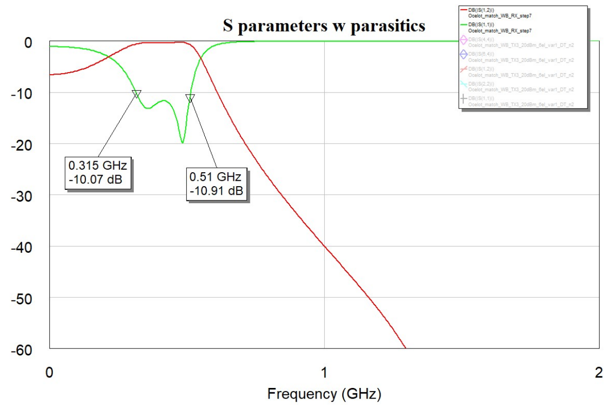

Figure 6.8. Simulation Results of Final Wideband RX Match

6.1.2 Smith-Chart with Constant Q Curves Representation

The BW of the matching network is determined by the max Q of the impedance points of trajectory.

Examples below show a 2-element and a 4-element match between the same impedances. A 4-element match can stay within a lower-valued constant Q curve which provides a larger impedance bandwidth.

This section is intended to provide a graphical representation of matching procedure for a wideband matching network.

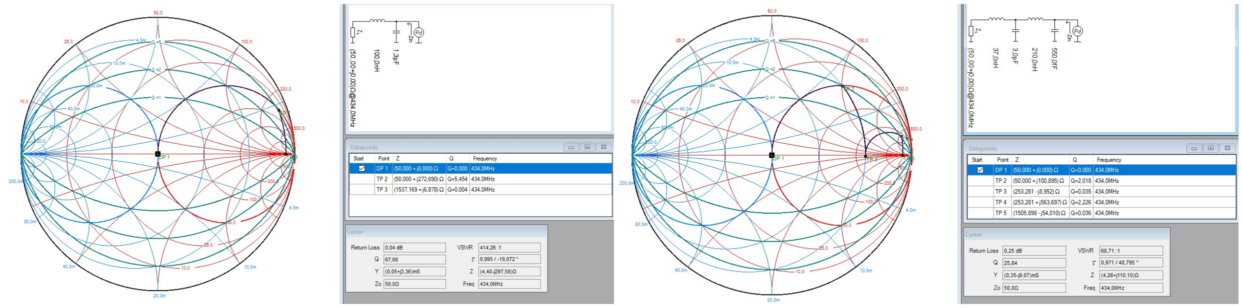

Figure 6.9. Smith-Chart Representation for a Wideband Match

6.1.3 Summary of RX Matching Networks

Covering frequency bands between 315–510 MHz needs the 3-element T-matching network structure as designed above:

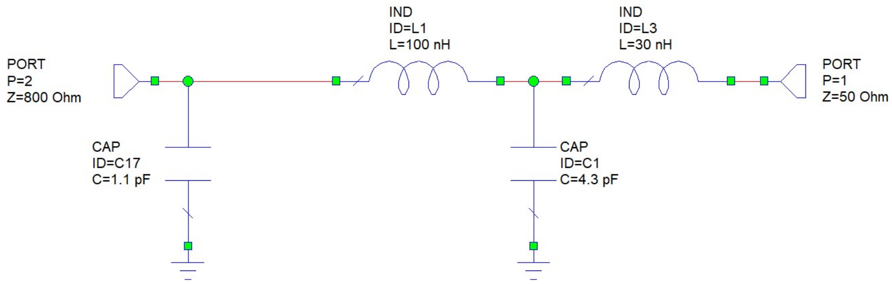

Figure 6.10. RX-Only Wideband Match for the 315–510 MHz Band

A single series inductor match satisfies the following frequency bands with acceptable RX performance:

- 315–434 MHz
- 434–510 MHz
- 868–928 MHz
- 780–928 MHz

6.1.4 TX-Only Wideband Matching Design Steps

The PA match does not require the Youla matching technique for a wideband operation since the matching Q is relatively smaller. So, simple matching designs can satisfy the BW requirements on the TX path.

The required PA load impedance values depend on the PAVDD and TXP as shown in Table 3.3 Optimum PA Load Impedance at the PA on Available Reference Designs on page 11. The PA can be modeled as a parallel RC (as well as the LNA).

Design example between $R=7\ \Omega$ *with* $C=4$ *pF and* $R_{ANT} = 50\ \Omega$ *at* $f_c = 400$ *MHz*

- +20 dBm designs have the biggest Q, therefore the smallest BW

2-element simple match:

- $Q_{RC} = 2\pi f_c RC = 2\pi * 400 * 10^6 * 7 * 4 * 10^{-12} = 0.07$
- $R_s = \dfrac{R}{1 + Q^2_{RC}} = 6.97\ \ \Omega$
- $C_s = C * \dfrac{1 + Q^2_{RC}}{Q^2_{RC}} = 811.7\ \ pF \rightarrow X_s = \dfrac{1}{2\pi f_c * C_s} = 0.49\ \ \Omega$
- $Q_{match} = \sqrt{\dfrac{R_{ANT}}{R_S} - 1} = 2.49$
- $L_{ser} = \dfrac{R_s * Q_{match}}{2\pi f_c} + \dfrac{X_s}{2\pi f_c} = 7.1\ \ nH$
- $C_{par} = \dfrac{Q_{match}}{2\pi f_c * R_{ANT}} = 19.8\ \ pF$

4-element match with mid-impedance point at $R_{INT} = \sqrt{50 * 7} = 18.7\ \ \Omega$:

- L_{ser1} = 3.8 nH; C_{par1} = 27.6 pF; L_{ser2} = 9.6 nH; C_{par2} = 10.3 pF

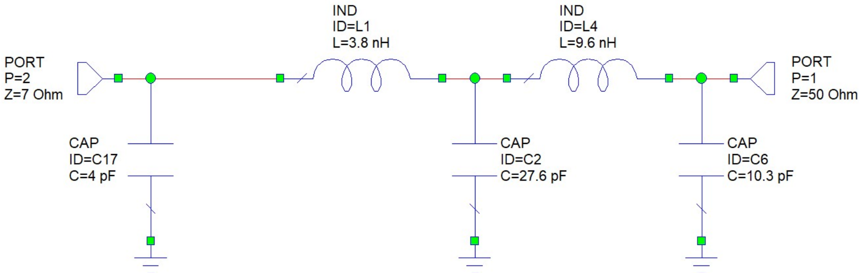

Figure 6.11. 4-Element PA Match

- 4-element match provides large impedance BW as shown below.
- The Q of the matching network can even be increased to cover a 200 MHz BW, i.e., from 315 up to 510 MHz, for instance.
- An increased Q of the match can ensure better harmonic suppression.

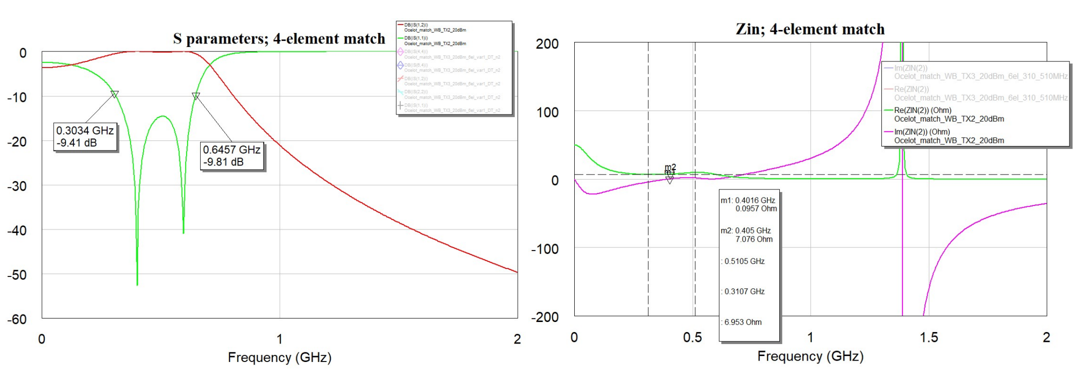

Figure 6.12. Simulation Results of a 4-element PA Match

Design recommendations:

TX matching design goal is to maximize the harmonic suppression performance, which typically yields to use the highest-Q (until a reasonable limit, e.g., component technical spreading is well tolerated) for the matching network which still satisfies the impedance BW criteria.

Design example between *R=7 Ω with C=4 pF and* R_{ANT} *= 50 Ω at* f_c*= 400 MHz* with a 4-element matching for 220 MHz BW:

- Determine R_{INT}
  - $Q_{RC} = 2\pi f_c RC = 2\pi * 400 * 10^6 * 7 * 4 * 10^{-12} = 0.07$
  - $R_s = \dfrac{R}{1 + Q^2_{RC}} = 6.97 \quad \Omega$
  - $Q_{match} = \sqrt{\dfrac{R_{INT}}{R_S} - 1} = \dfrac{f_c}{BW_{theory}} \rightarrow R_{INT} = \dfrac{R_s f_c^2}{BW^2_{theory}} + R_s = 30 \quad \Omega$
- Two-stage match between R = 7 Ω with C = 4 pF and R_{INT} = 30 Ω, and between R_{INT} = 30 Ω and R_{ANT} = 50 Ω
  - L_{ser1} = 5.2 nH; C_{par1} = 24.1 pF; L_{ser2} = 9.8 nH; C_{par2} = 6.5 pF

To further increase the harmonic suppression, the 2nd-stage match can be a 3-element PI filter as well and/or can be suggested to design a 6-element L-C ladder matching circuit (i.e., increase the order of the LPF function).

6.1.5 Simulations of TX-only Wideband Matching Networks

There are two TX wideband matching circuits designed for the following frequency bands:

- 315–434 MHz +14 dBm match: low power FCC, ETSI bands
- 434–510 MHz +20 dBm match: high power ETSI, China bands

To achieve the desired TX harmonic suppression performance a 6-element ladder L-C matching network has been designed for these targeted frequency bands. The matching network schematic shown in the figures below does not include the parasitics, while the simulation results do.

315 – 434 MHz +14 dBm match:

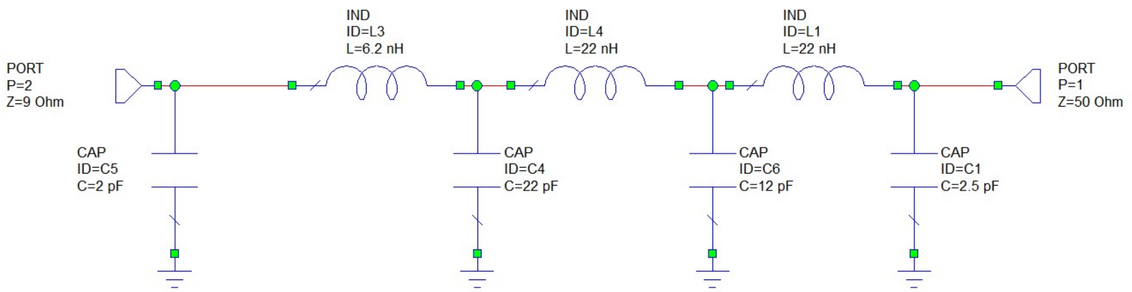

Figure 6.13. 6-Element PA Match for 315-434 MHz +14 dBm

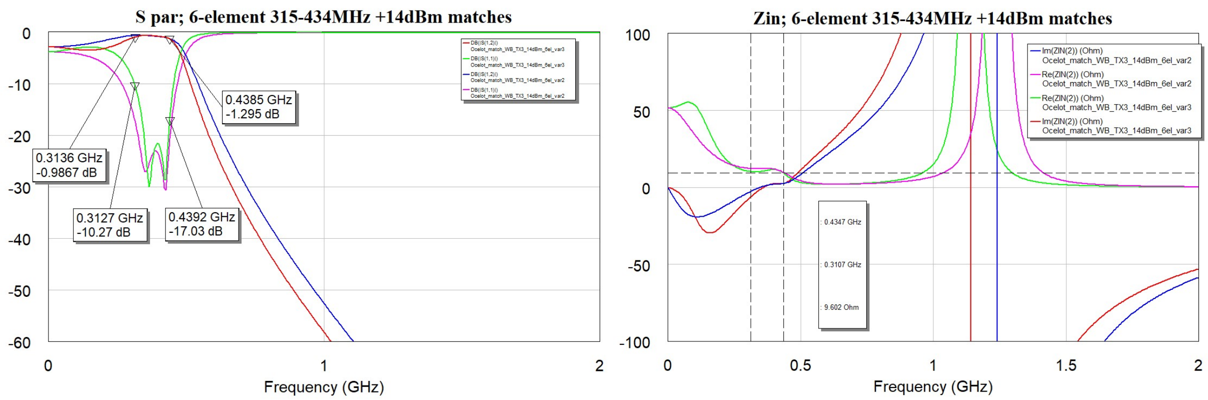

Figure 6.14. Simulation Results of the PA Match for 315-434 MHz +14 dBm

434 – 510 MHz +20 dBm match:

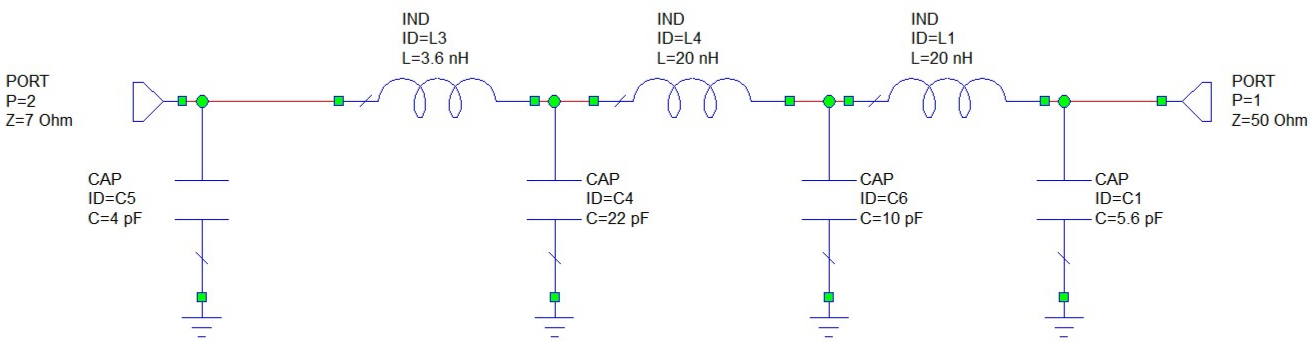

Figure 6.15. 6-Element PA Match for 434-510 MHz +20 dBm

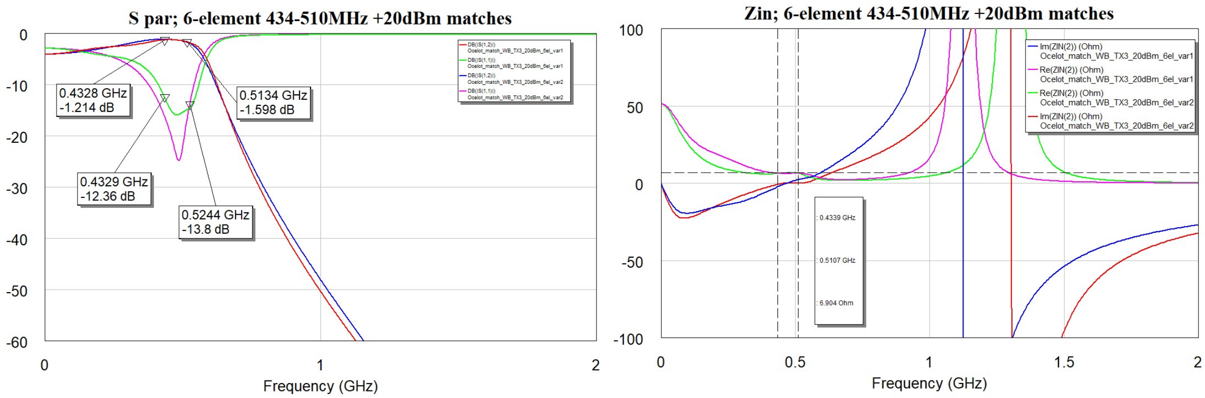

Figure 6.16. Simulation Results of the PA Match for 434-510 MHz +20 dBm

6.1.6 Direct-Tie TRX Wideband Matching Network Considerations

Direct-tying the TX and RX paths together is typically a bandwidth-reducing step, and the followings should be considered.

- TX match output impedance must be high-Z in PA off-mode
  - PA off-mode impedance
    - HPA: 17 kΩ with shunt 2 pF
    - LPA: 25 kΩ with shunt 1 pF
- RX match input impedance must be high-Z in LNA off-mode
  - LNA off-mode impedance: 10 Ω with shunt 1.1 – 1.2 pF
- RX match limitations in wideband:
  - 3-element wideband T-match (between 315 – 510 MHz frequencies) has a low-impedance resonance within the band
  - Wideband direct-tie match is only possible when using a single-inductor RX match, e.g., in 315 – 434 MHz, 434 – 510 MHz, 780 – 928 MHz. Not possible in the full 315 – 510 MHz frequency band.
- TX match compromises:
  - Bandwidth versus harmonic suppression
  - In direct-tie matching configuration, RX sensitivity versus harmonic suppression as well

<u>Example of 434 – 510 MHz +20 dBm TRX direct-tie match:</u>

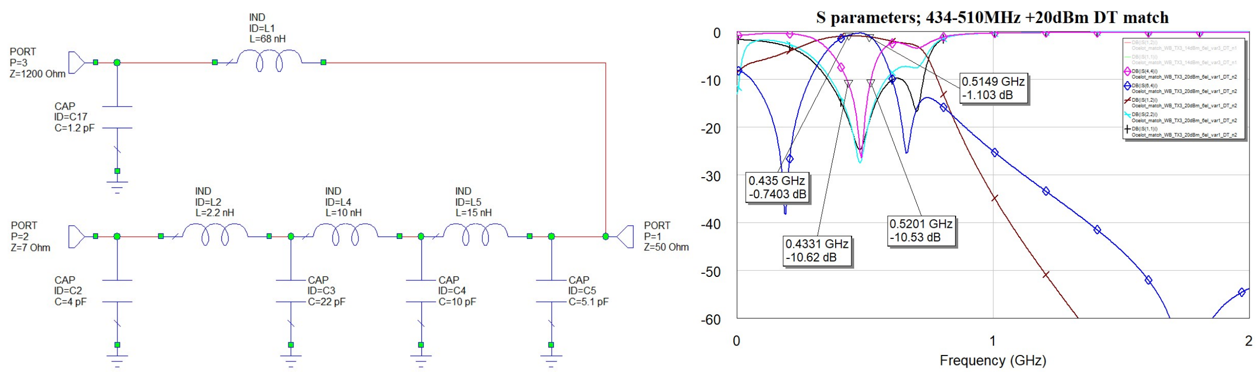

Figure 6.17. TRX Direct-Tie Match for 434-510 MHz +20 dBm

<u>Example of 315 – 434 MHz +14 dBm TRX direct-tie match:</u>

- 2nd harmonic trap is required in the TX path to have the desired attenuation
- L5=16 nH: sensitive for harmonic suppression (brown curve) and unwanted resonance in RX mode (blue curve)
  - If its value is increased it improves harmonic suppression, but the unwanted notch moves closer or into the RX pass-band

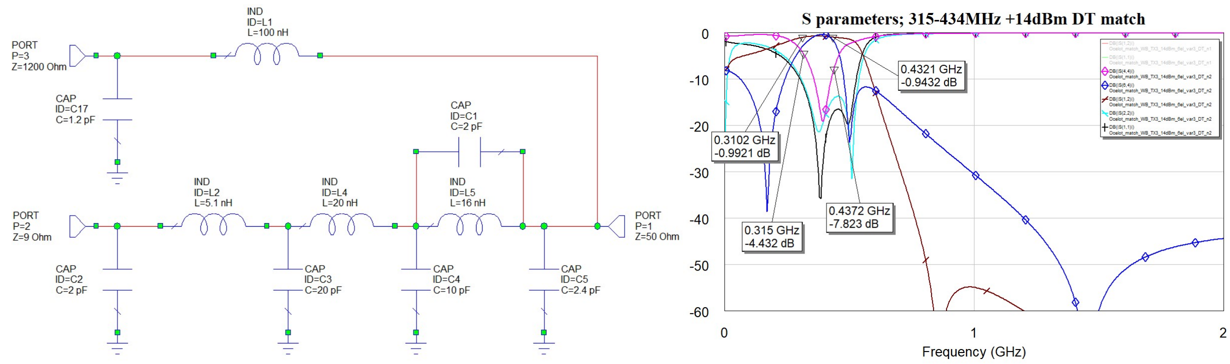

Figure 6.18. TRX Direct-Tie Match for 315-434 MHz +14 dBm

Example of 315 – 510 MHz +20 dBm TRX direct-tie match:

- 2nd harmonic suppression of 315 MHz band (at 630 MHz) is challenging while 510 MHz is still a pass-band.
- L3 – C1 has a low-impedance resonance in the RX path in TX mode (brown curve). It shunts out TX power in TX-RX direct-tie configuration. This effect cannot be tuned back while keeping the desired RX wideband performance.
- RX-only match was already presented above in this document. TX-only match might be possible as shown below (depending on the harmonic suppression requirements).

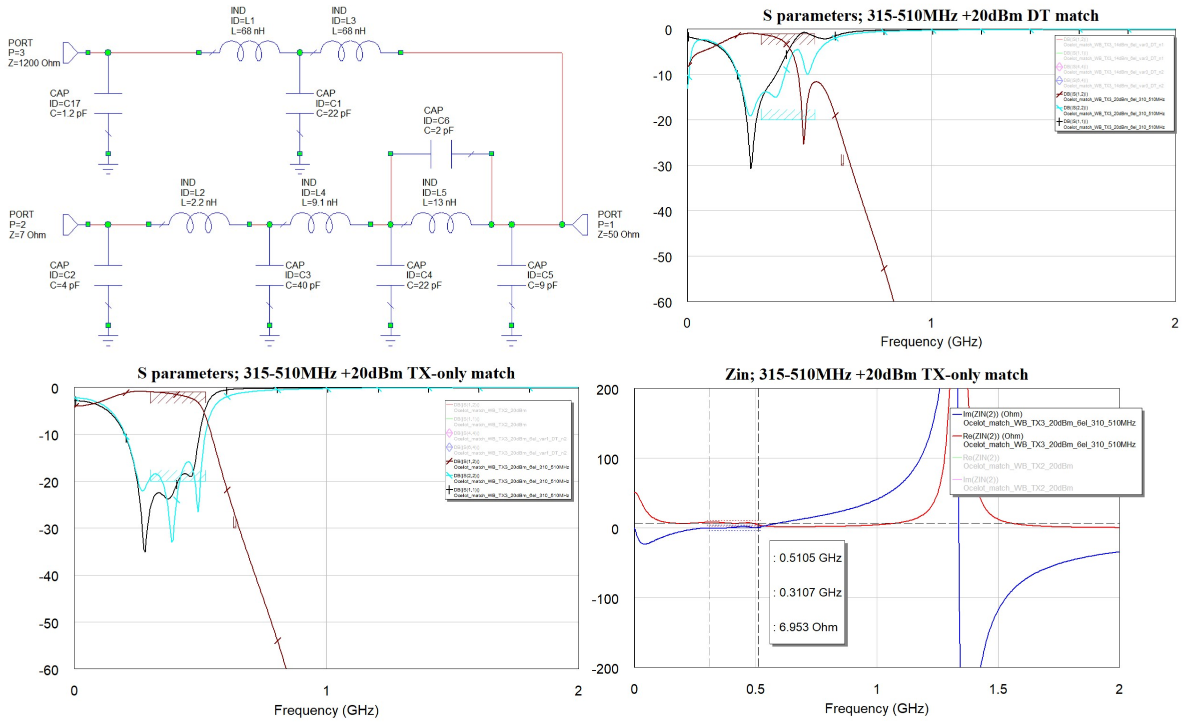

Figure 6.19. TRX Direct-Tie Match for 315-510 MHz +20 dBm

6.2 Dual-Band Matching Network

A dual-band matching network can cover frequency bands that fall far from each other, i.e., covers far-placed frequency bands with relatively small bandwidths (e.g., 434 and 868-928 MHz).

For a dual-band solution, inductors and/or capacitors with properly decreasing values with frequency are required. This can be done by the 3-element resonators. See section “5. The 3-Element Resonators” in the application note AN1180: EFR32 Series 1 sub-GHz Discrete Matching Solutions for a detailed description with equations.

Dual-band matches designed for EFR32FG23 on a single RF port:

- Dual-band RX match for 434–868 MHz, 434–868/915 MHz, 315–915 MHz and 315–868/915 MHz.
- Dual-band TX match does not make sense on a single RF port.
  - Needs separation for LPF functions → use separate RF port or with external switch.

6.2.1 Design Example of RX Dual-Band Match for 434–868/915 MHz

- Single-band RX match is a single series inductor.
  - It basically resonates with the LNA capacitance.
    - Assuming C_{LNA} = 1.2 pF → 112 nH at 434 MHz; 26 nH at 900 MHz.
  - Radio boards use 82 nH at 434 MHz and 18 nH at 868/915 MHz.
    - Slightly lower-Q match, compensate long PCB trace parasitics in the RX path.
  - Calculate the component values of a 3-element resonator based on the equations shown in section “5. The 3-Element Resonators” in AN1180: EFR32 Series 1 sub-GHz Discrete Matching Solutions.
  - Select the inductor type 2 where PCB trace inductance can be compensated easier (by one component value change).
  - Calculated values for 112 nH at 434 MHz and 26 nH at 868/915 (900) MHz:
  - L_1 = 61.5 nH, L_3 = 29.1 nH, C_3 = 1.95 pF

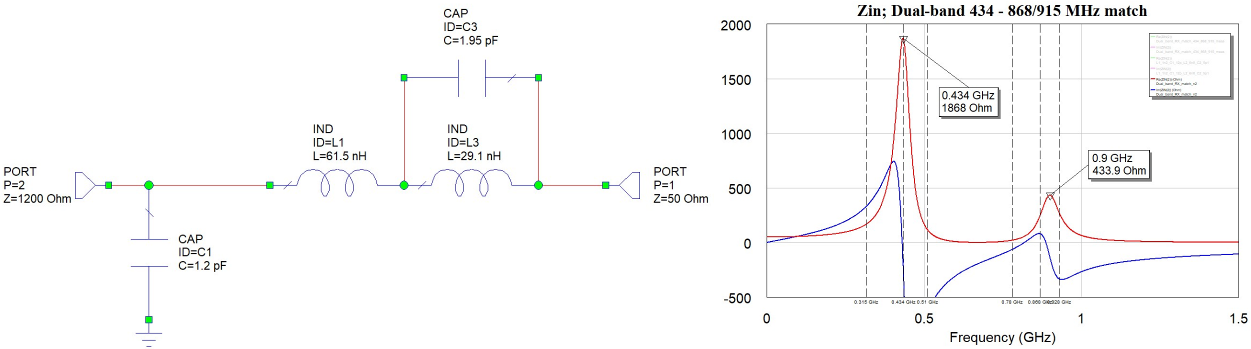

Figure 6.20. 3-Element Resonator as Dual-and 434–868/915 MHz RX Match

6.2.2 Simulation Results with PCB and SMD Parasitics

- Re-calculate the 3-element resonator component values and use slightly lower-Q match to be able to cover the 868–915 MHz band
  - Calculated for 82 nH at 434 MHz and 18 nH at 900 MHz
    - L_1 = 44.4 nH, L_3 = 21.7 nH, C_3 = 2.63 pF
  - Final values with all parasitics and for a high-band with wider bandwidth
    - L_1 = 30 nH, L_3 = 24 nH, C_3 = 3.3 pF
- Simulation results for 315 – 868/915 MHz bands:
  - L_1 = 24 nH, L_3 = 47 nH, C_3 = 3.3 pF

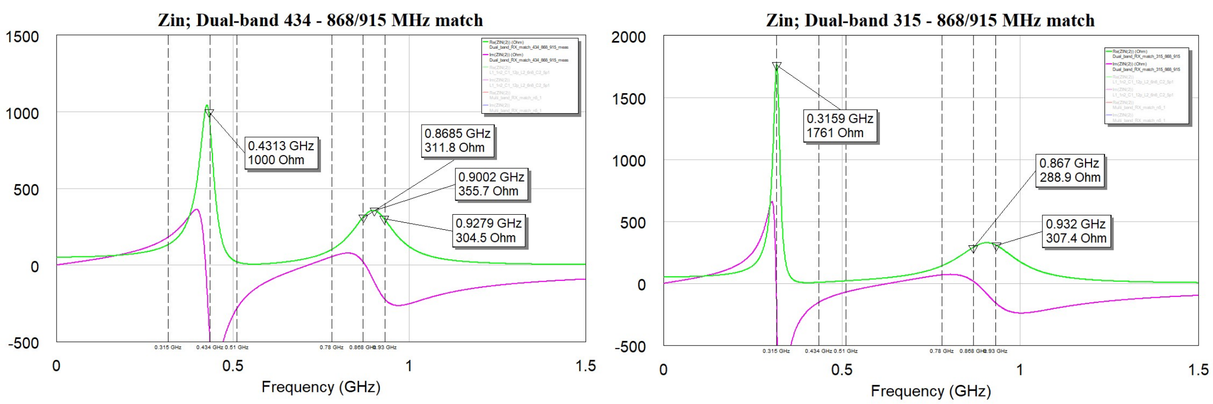

Figure 6.21. Simulation Results of Dual-Band RX Matches

6.3 Multi-Band and Dual-Wideband Matching Network

A multi-band matching network can cover 2-far placed frequency bands with a narrow and a large BW coverage (e.g., 315–434 and 868–928 MHz). Similarly, a dual-wideband match can cover 2 far-placed frequency bands with large BW coverage on both (e.g., 300–510 and 780–928 MHz).

For a multiband or dual-wideband solution the design methods of the wideband and dual-band matching designs need to be combined.

6.3.1 Design Steps

- Design the separate wideband (and/or single-band) matching networks.
- Apply 3-element resonators to have one matching network structure covering the separately designed wideband matching circuits.
- Optimize the BOM of the resulted network.
  - To keep the multi-band or dual-wideband operation, at least, one 3-element resonator should be kept in the matching design.
- Tune and optimize the component values in circuit simulations.

Design example for a RX match in the 315–510 & 780–928 MHz and in the 315–434 & 868–928 MHz frequency bands:

- Utilize the 3-element T-wideband match (ideal, 122 nH – 4.3 pF – 31 nH) for 315–510 MHz (center at 410 MHz) and the single series RX matching inductor (ideal, 26 nH) for 780–928 MHz (center at 860 MHz).

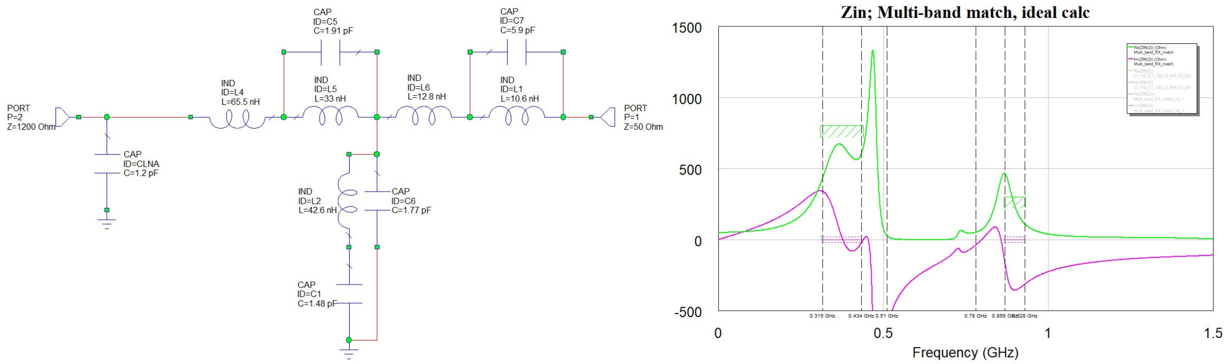

Figure 6.22. Simulations of Multi-Band RX Match

6.3.2 BOM Optimization

Step 1

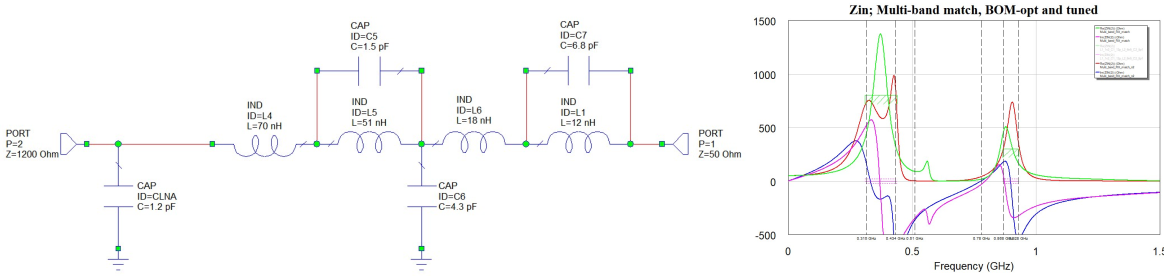

Figure 6.23. Simulations of Multi-Band RX Match with Reduced BOM

Step 2 (Minimal BOM)

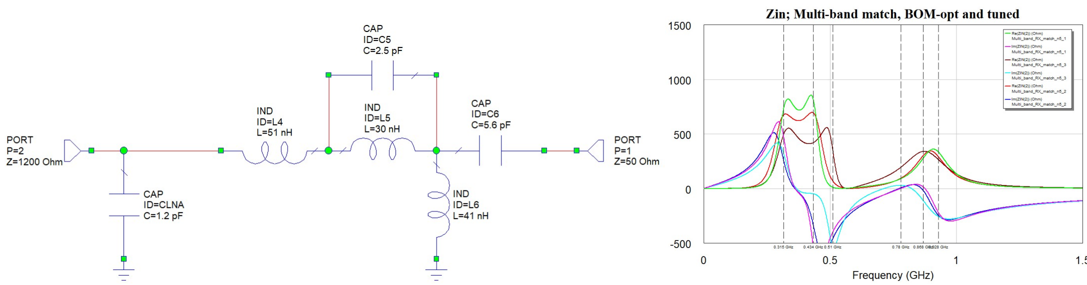

Figure 6.24. Simulations of Multi-Band RX Match with Minimal BOM

6.3.3 Simulations with Parasitics

- 5-element multi-band matching network with minimal BOM has been selected.
  - Simulation results shown with several different tuned component values to cover the targeted frequency bands.
- Targeted frequency bands:
  - 315–434 and 868–928 MHz
  - 315–510 and 780–928 MHz

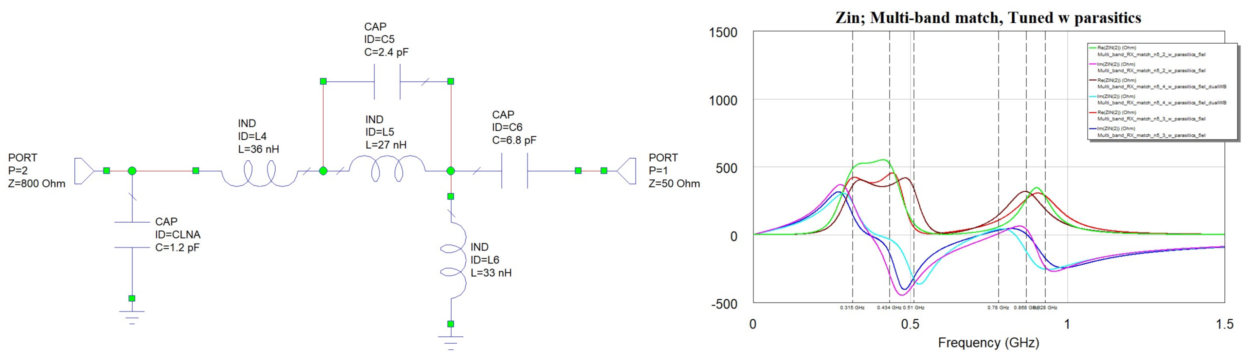

Figure 6.25. Simulations of Multi-Band Minimal-BOM RX Match with Parasitics

6.4 Measurement Results

6.4.1 Common Schematic Structure

The schematic reference with its component designators as shown in the figure below is used to summarize all conducted measurement results in this section. The RF boards used for these measurements follow the radio board layout design concept and documented in AN928.2: EFR32 Series 2 Layout Design Guide.

The measurements were taken on EFR32xG23 parts, but these matching network solutions can also be recommended for designs with EFR32xG28 parts.

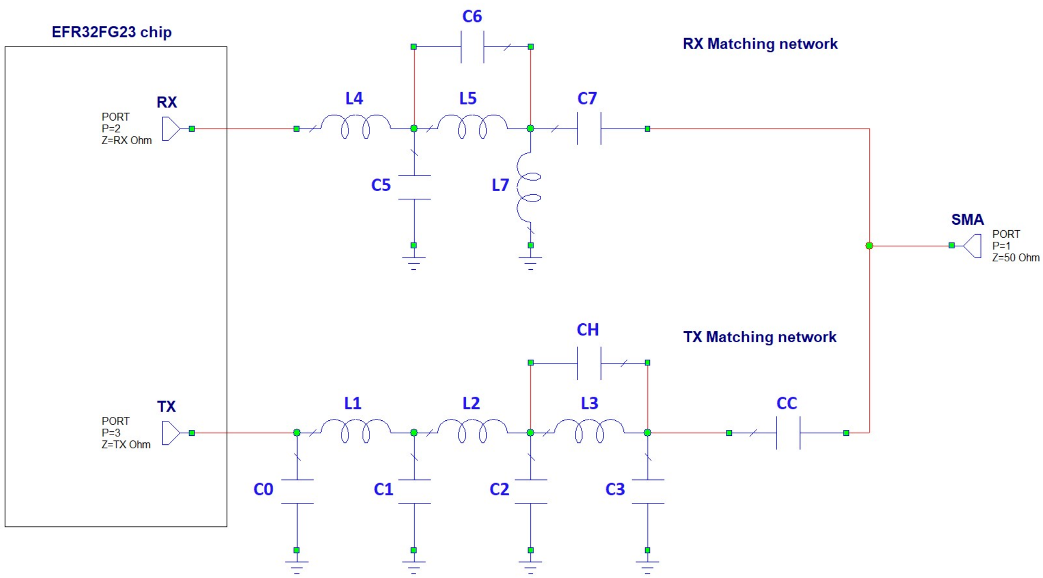

Figure 6.26. Common Schematic Reference

RX sensitivity test conditions: 2-GFSK modulation with 100 kbps data rate and 50 kHz frequency deviation (BT = 0.5). BER limit is at 0.1%.

6.4.2 Wideband RX-only Matching Network

Table 6.1. Measurement Results of Low-Band WB RX Match

Component Values						Sens. [dBm] at Freq. [MHz]						
L4 [nH]	**L5 [nH]**	**C5 [pF]**	**C6 [pF]**	**L7 [nH]**	**C7 [pF]**	**285**	**315**	**390**	**434**	**470**	**490**	**510**
68	0R	NM	NM	NM	0R	NA	-104.8	-108.6	-109.7	-110.2	-110.3	-110.2
82	0R	NM	NM	NM	0R	NA	-108.0	-109.8	-110.6	-110.2	-109.7	-109.1
100	0R	NM	NM	NM	0R	NA	-109.0	-110.3	-109.8	-108.5	-107.5	-106.6
100	30	4.3	NM	NM	0R	NA	-110.1	-110.2	-109.8	-109.8	-109.8	-109.6
120	0R	NM	NM	NM	0R	-108.6	-109.3	-110.2	-108.7	NA	NA	NA
120	33	4.7	NM	NM	0R	-110.0	-110.4	-110.0	-109.8	NA	NA	NA

Table 6.2. Measurement Results of High-Band WB RX Match

Component Values						Sens. [dBm] at Freq. [MHz]					
L4 [nH]	**L5 [nH]**	**C5 [pF]**	**C6 [pF]**	**L7 [nH]**	**C7 [pF]**	**780**	**830**	**868**	**902**	**915**	**928**
18	0R	NM	NM	NM	0R	-109.5	-110.0	-110.2	-110.1	-110.2	-110.1
20	0R	NM	NM	NM	0R	-109.8	-110.1	-110.2	-110.1	-110.0	-109.8

6.4.3 Wideband TX-only Matching Network

Table 6.3. Measurement Results of 434-510 MHz +14 dBm WB TX Match

Component values									TXP [dBm] at freq. [MHz]				
L1 [nH]	L2 [nH]	L3 [nH]	C1 [pF]	C2 [pF]	C3 [pF]	C0 [pF]	CH [pF]	CC [pF]		434	470	490	510
6.2	18	0R	18	6	NM	NM	NM	220	Fund. [dBm]	13.9	14.2	14.3	14.0
									H2 [dBm]	-46.7	-43.5	-45.8	-48.1
									H3 [dBm]	-41.3	-44.8	-46.5	-47.8
									H4 [dBm]	-60.0	-60.0	-60.0	-60.0
									H5 [dBm]	-60.0	-60.0	-60.0	-60.0
									Current [mA]	20.9	23.7	25.2	25.1

Table 6.4. Measurement Results of 315-434 MHz +14 dBm WB TX Match

Component values									TXP [dBm] at freq. [MHz]			
L1 [nH]	L2 [nH]	L3 [nH]	C1 [pF]	C2 [pF]	C3 [pF]	C0 [pF]	CH [pF]	CC [pF]		315	390	434
6.2	22	22	22	12	2.4	NM	NM	220	Fund. [dBm]	14.1	14.2	14.2
									H2 [dBm]	-41.1	-57.3	-60.0
									H3 [dBm]	-49.6	-60.0	-60.0
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-60.0	-60.0	-60.0
									Current [mA]	21.0	22.7	24.2
7.5	30	30	20	10	2.4	NM	NM	220	Fund. [dBm]	14.1	14.0	13.9
									H2 [dBm]	-48.6	-60.0	-60.0
									H3 [dBm]	-54.2	-60.0	-60.0
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-59.8	-60.0	-60.0
									Current [mA]	22.3	22.6	24.2
6.8	30	30	20	10	1.5	NM	NM	220	Fund. [dBm]	14.0	14.0	14.3
									H2 [dBm]	-47.9	-60.0	-60.0
									H3 [dBm]	-52.1	-60.0	-60.0
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-60.0	-60.0	-60.0
									Current [mA]	22.0	22.6	26.7

Table 6.5. Measurement Results of 285-434 MHz +20 dBm WB TX Match

Component values									TXP [dBm] at freq. [MHz]				
L1 [nH]	L2 [nH]	L3 [nH]	C1 [pF]	C2 [pF]	C3 [pF]	C0 [pF]	CH [pF]	CC [pF]		285	315	390	434
2.4	15	16	36	15	4.3	NM	2.7	270	Fund. [dBm]	20.0	20.4	20.5	20.0
									H2 [dBm]	-27.2	-35.2	-40.0	-40.6
									H3 [dBm]	-41.9	-44.2	-44.5	-44.1
									H4 [dBm]	-50.0	-50.0	-50.0	-50.0
									H5 [dBm]	-47.5	-48.9	-50.0	-50.0
									Current [mA]	63.5	70.0	77.8	71.2
4.3	20	15	30	13	4.3	NM	3.6	270	Fund. [dBm]	19.5	19.9	20.1	19.3
									H2 [dBm]	-35.9	-40.6	-40.2	-41.8
									H3 [dBm]	-42.1	-42.9	-45.3	-47.9
									H4 [dBm]	-50.0	-50.0	-50.0	-50.0
									H5 [dBm]	-48.5	-50.0	-50.0	-50.0
									Current [mA]	54.0	60.5	73.0	65.2

Table 6.6. Measurement Results of 434-510 MHz +20 dBm WB TX Match

Component values									TXP [dBm] at freq. [MHz]				
L1 [nH]	L2 [nH]	L3 [nH]	C1 [pF]	C2 [pF]	C3 [pF]	C0 [pF]	CH [pF]	CC [pF]		434	470	490	510
3.3	20	20	22	10	3.3	NM	NM	220	Fund. [dBm]	20.4	20.5	20.4	20.2
									H2 [dBm]	-53.1	-54.8	-57.5	-55.1
									H3 [dBm]	-52.5	-50.4	-48.6	-48.1
									H4 [dBm]	-60.0	-60.0	-60.0	-60.0
									H5 [dBm]	-56.2	-56.9	-56.6	-55.9
									Current [mA]	79.9	84.6	84.0	83.7
2	16	22	27	11	2.7	NM	NM	220	Fund. [dBm]	20.8	20.5	20.3	20.1
									H2 [dBm]	-51.4	-54.0	-57.9	-55.9
									H3 [dBm]	-52.5	-49.9	-47.9	-47.4
									H4 [dBm]	-60.0	-60.0	-60.0	-60.0
									H5 [dBm]	-55.7	-55.1	-56.4	-56.0
									Current [mA]	88.0	84.7	81.9	82.9

6.4.4 Wideband TRX Direct-tie Matching Network

Table 6.7. Measurement Results of 315-434 MHz +14 dBm WB TRX Match

Component values									TXP and RX Sens. [dBm] at freq. [MHz]			
L1 [nH]	L2 [nH]	L3 [nH]	C1 [pF]	C2 [pF]	C3 [pF]	CH [pF]	L4 [nH]	CC [pF]		315	390	434
5.1	20	16	22	10	2.4	2	100	220	Fund. [dBm]	14.2	14.9	14.3
									H2 [dBm]	-41.7	-60.0	-60.0
									H3 [dBm]	-47.5	-48.3	-49.4
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-55.2	-58.3	-59.7
									TX Current [mA]	22.2	25.8	24.7
									RX Sens. [dBm]	-109.2	-110.2	-109.2
5.1	20	16	22	10	2.4	2.4	100	220	Fund. [dBm]	14.2	14.9	14.1
									H2 [dBm]	-44.7	-60.0	-60.0
									H3 [dBm]	-44.7	-47.5	-48.8
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-55.3	-58.3	-58.7
									TX Current [mA]	22.5	25.8	23.6
									RX Sens. [dBm]	-109.2	-110.1	-109.1
5.1	20	16	22	10	2.7	2.4	100	220	Fund. [dBm]	14.1	14.9	14.1
									H2 [dBm]	-45.0	-60.0	-60.0
									H3 [dBm]	-44.9	-47.4	-49.1
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-55.5	-58.2	-59.5
									TX Current [mA]	22.2	26.1	24.0
									RX Sens. [dBm]	-109.2	-110.2	-109.1
Note: L5=C7=0R; C0=C5=C6=L7=N.M.												

Table 6.8. Measurement Results of 315-390 MHz +20 dBm WB TRX Match

Component values									TXP [dBm] at freq. [MHz]			
L1 [nH]	L2 [nH]	L3 [nH]	C1 [pF]	C2 [pF]	C3 [pF]	CH [pF]	L4 [nH]	CC [pF]		315	345	390
5.1	22	15	30	11	3.9	2.2	120	270	Fund. [dBm]	20.4	20.7	20.4
									H2 [dBm]	-42.1	-45.9	-49.5
									H3 [dBm]	-45.0	-45.3	-44.0
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-53.6	-55.6	-57.4
									Current [mA]	72.0	80.0	82.0
									RX Sens. [dBm]	-109.7	-109.9	-109.6
5.6	24	15	30	11	3.9	2.2	120	270	Fund. [dBm]	20.4	20.7	20.0
									H2 [dBm]	-44.0	-50.2	-53.8
									H3 [dBm]	-46.2	-45.1	-45.3
									H4 [dBm]	-60.0	-60.0	-60.0
									H5 [dBm]	-54.2	-55.9	-57.8
									Current [mA]	73.0	81.0	78.0
									RX Sens. [dBm]	-109.6	-109.8	-109.5
Note: L5=C7=0R; C0=C5=C6=L7=N.M.												

Table 6.9. Measurement Results of 434-510 MHz +20 dBm WB TRX Match

Component values								TXP and RX Sens. [dBm] at freq. [MHz]				
L1 [nH]	**L2 [nH]**	**L3 [nH]**	**C1 [pF]**	**C2 [pF]**	**C3 [pF]**	**L4 [nH]**	**CC [pF]**		**434**	**470**	**490**	**510**
2	12	18	22	10	4.3	68	220	Fund. [dBm]	20.5	20.7	20.8	20.6
								H2 [dBm]	-43.9	-53.1	-54.8	-60.0
								H3 [dBm]	-43.4	-43.2	-43.4	-44.0
								H4 [dBm]	-60.0	-60.0	-60.0	-60.0
								H5 [dBm]	-54.9	-55.7	-55.2	-53.0
								TX Current [mA]	79.8	87.9	89.8	87.6
								RX Sens. [dBm]	-109.9	-110.0	-109.9	-109.8
2	12	18	22	10	3.9	68	220	Fund. [dBm]	20.5	20.7	20.7	20.5
								H2 [dBm]	-43.4	-52.4	-54.5	-60.0
								H3 [dBm]	-43.2	-42.7	-42.9	-43.5
								H4 [dBm]	-60.0	-60.0	-60.0	-60.0
								H5 [dBm]	-54.8	-55.6	-55.3	-53.2
								TX Current [mA]	80.2	86.7	87.8	85.0
								RX Sens. [dBm]	-109.9	-110.0	-110.0	-109.8
Note: L5=C7=0R; C0=CH=C5=C6=L7=N.M.												

Table 6.10. Measurement Results of 780-928 MHz +20 dBm WB TRX match

Component values								TXP [dBm] at freq. [MHz]			
L1 [nH]	L2 [nH]	L3 [nH]	C1 [pF]	C2 [pF]	C0 [pF]	L4 [nH]	CC [pF]		780	868	915
1.8	16	7.5	8.9	3.6	NM	20	100	Fund. [dBm]	20.1	20.8	19.9
								H2 [dBm]	-74.6	-54.1	-51.8
								H3 [dBm]	-42.5	-41.3	-40.7
								H4 [dBm]	-67.4	-73.5	-72.0
								H5 [dBm]	-62.0	-56.5	-48.1
								Current [mA]	92.2	98.3	91.9
								RX Sens. [dBm]	-110.5	-110.4	-109.8
2.0	13	8.2	8.2	3.9	7.5	20	100	Fund. [dBm]	19.9	20.5	20.0
								H2 [dBm]	-57.5	-60.7	-58.9
								H3 [dBm]	-42.5	-44.2	-43.3
								H4 [dBm]	-69.6	-74.3	-75.1
								H5 [dBm]	-71.9	-59.7	-53.0
								Current [mA]	88.9	94.1	93.9
								RX Sens. [dBm]	-110.5	-110.5	-110.0
Note: L5=C7=0R; CH=C3=C5=C6=L7=N.M.											

6.4.5 Dual-band RX-only Matching Network

Table 6.11. Measurement Results of Dual-Band RX Matches

Component Values			Sens. [dBm] at Freq. [MHz]					
L4 [nH]	L5 [nH]	C6 [pF]	315	434	868	902	915	928
30	24	3.3	NA	-110.6	-109.5	-109.8	-109.8	-109.8
30	22	3.6	NA	-110.6	-109.6	-109.8	-109.8	-109.6
33	24	3	NA	-110.9	-109.5	NA	-109.5	-109.2
33	24	3.3	NA	-110.6	-109.9	NA	-109.2	NA
27	47	3.3	-110.0	NA	-109.7	-109.7	-109.8	-109.9
27	43	3.6	-109.9	NA	-109.8	-109.9	-109.9	-109.8
30	56	2.4	-110.3	NA	-109.4	-109.7	-109.7	-109.7
Note: C7=0R; C5=L7=N.M.								

6.4.6 Multi-band and Dual-wideband RX-only Matching Network

Table 6.12. Measurement Results of Multi-band and Dual-wideband RX Matches

Component Values					Sens. [dBm] at Freq. [MHz]											
L4 [nH]	**L5 [nH]**	**C6 [pF]**	**L7 [nH]**	**C7 [pF]**	**300**	**315**	**390**	**434**	**470**	**490**	**510**	**780**	**868**	**902**	**915**	**928**
39	27	2.4	33	6.8	-109.5	-110.1	-110.7	-110.4	-110.0	-109.0	-108.0	NA	-108.7	-109.1	-109.0	-108.7
33	20	3.9	36	6.2	-109.6	-110.1	-110.3	-110.1	-109.8	-108.6	-106.2	-107.7	-109.6	-109.5	-109.5	-109.2
33	15	4.7	33	6	-108.3	-109.5	-110.3	-110.1	-110.1	-110.2	-109.6	-107.8	-109.6	-109.4	-109.2	-108.9
33	15	5.1	33	6	-108.7	-109.4	-110.2	-110.1	-110.1	-109.8	-108.5	-108.6	-109.7	-109.3	-109.2	-108.6
33	13	5.1	33	6	-108.6	-109.5	-110.2	-110.0	-110.1	-110.3	-110.0	-107.9	-109.6	-109.3	-109.2	-108.6
Note: C5=N.M.																

7. Revision History

Revision 0.8

July 2023

- Added EFR32xG28 sub-GHz matching guide and related considerations.

Revision 0.7

June 2023

- Updated EFR32xG25 minimal-BOM full discrete matching network recommendation for the 868/915 MHz frequency bands.

Revision 0.6

January 2023

- Added EFR32xG23 recommended matching network component values for +20 dBm in the 285-434, 315-390 and 780-928 MHz frequency bands.

Revision 0.5

December 2022

- Added EFR32xG25 matching guide and related considerations.

Revision 0.4

June, 2022

- Added recommended matching network component values for +20 dBm at 315 and 780 MHz.
- Added dual-band and wideband matching network design examples.
- Added 6. Multi-Band Matching Network Design for EFR32xG23/28.

Revision 0.3

January 2022

- Added recommended matching network components values for +20/14 dBm at 434 and for +20 dBm at 650 MHz.

Revision 0.2

September 2021

- Added recommended matching network component values for +20 dBm at 470-510 MHz.

Revision 0.1

September 2021

- Initial release.

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