Application Note
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microelectronics group

Lucent Technologies
Bell Labs Innovations

W2020 RF Mixer

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Introduction

The W2020 RF mixer performs the first down-conversion of the received signal in the GSM transceiver to the 71 MHz IF stage. The RF mixer must have sufficient gain and a low noise figure to surpass the –102 dBm sensitivity requirement, while also having the dynamic range to handle the large signal levels at 1.6 MHz and 3 MHz offsets under blocking signal conditions. This performance must be met while working from a 2.7 V to 4.5 V supply.

Integrated Double-Balanced Mixer

The basic RF mixer circuit is shown in Figure 1. The input is an internally biased emitter-coupled pair, and although inherently differential, the application is to be driven from a single-ended RF SAW filter after the low-noise amplifier (LNA). Therefore, the input is driven single-ended with the opposite NPN base ac grounded. The outputs from the UHF local oscillator (LO) buffer amplifier drive into the Gilbert-Cell transistors. The LO signal level into the RF mixer is nominally 600 mV_{P-P} drive level within the IC. Integrated active mixer circuits have the advantage of relatively low-power LO drive levels compared to discrete diode mixers.

The differential output is taken from the collectors of Q3 and Q6. The RC filtering is for rejection of the RF input and UHF LO frequencies. The dc collector currents are nominally 3.6 mA per NPN. To have sufficient ac gain, the collector load resistors must be large; however, the dc voltage drop in the resistors would become prohibitive. Therefore, the dc collector current must be fed through off-chip inductors to keep the NPN transistors out of saturation. The inductors will allow the output signal to swing above the V_{CC} supply to have sufficient conversion power gain. The inductors can be treated as external chokes or else absorbed into the output matching network.

Also, since typical resistors in most IC processes have ±20% tolerances, external resistors of typical tolerance values such as 2% are used to set the gain for the mixer. This minimizes the gain variation that the system design must accommodate.

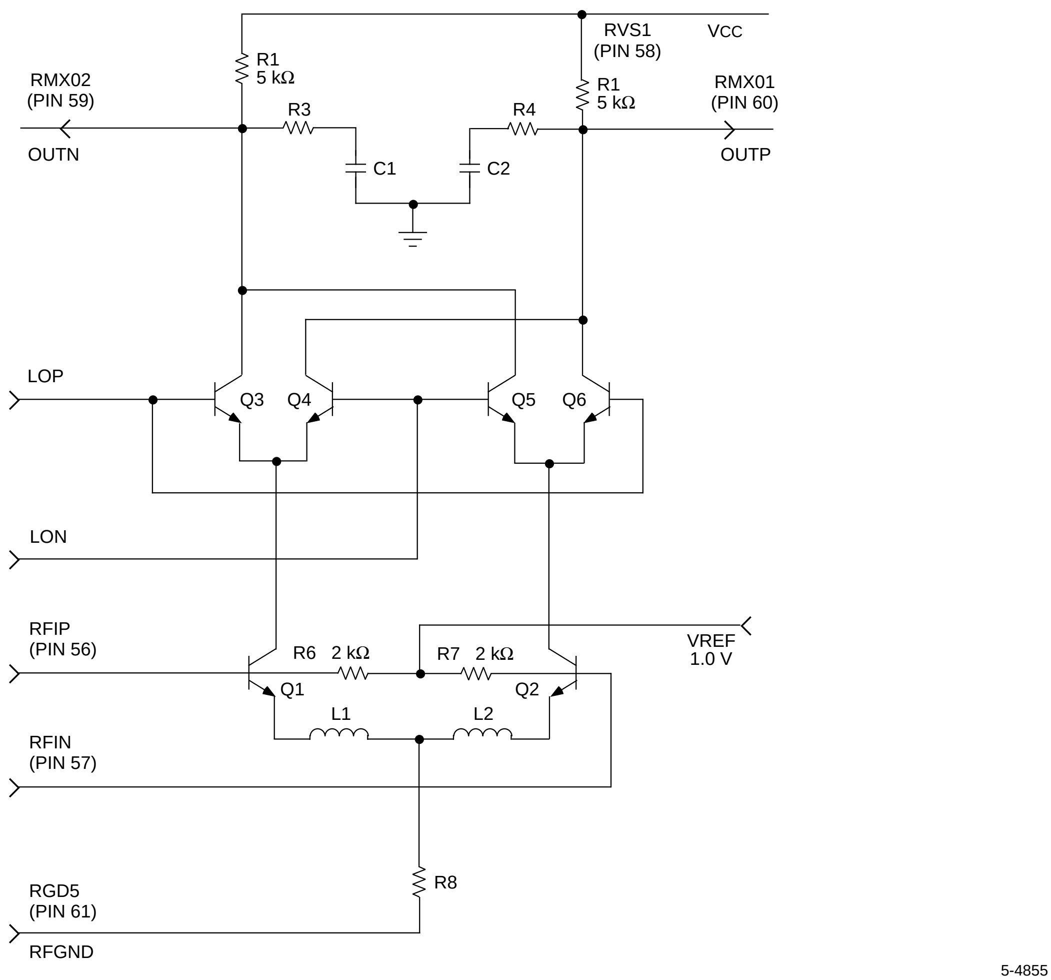


Figure 1. W2020 Double-Balanced RF Mixer Circuit

Input/Output Matching Network

The configuration presented for this application is for the mixer to be driven single-ended from a 50 Ω source with the opposite differential input ac grounded. The full differential output signal must be used to achieve the maximum conversion gain and optimum noise figure. This requires the output matching network to not only perform impedance matching, but also to be a low-loss accurate 71 MHz 0°/180° power combiner so that common-mode noise on the output can be rejected. The output signal is then input single-ended into the IF SAW filter. The matching networks are shown in Figure 2. The acceptable tolerance for the external components is 5% for the inductors, capacitors, and resistors. Preferably, 2% resistors would be used for R41 and R42 to set the gain in the RF mixer.

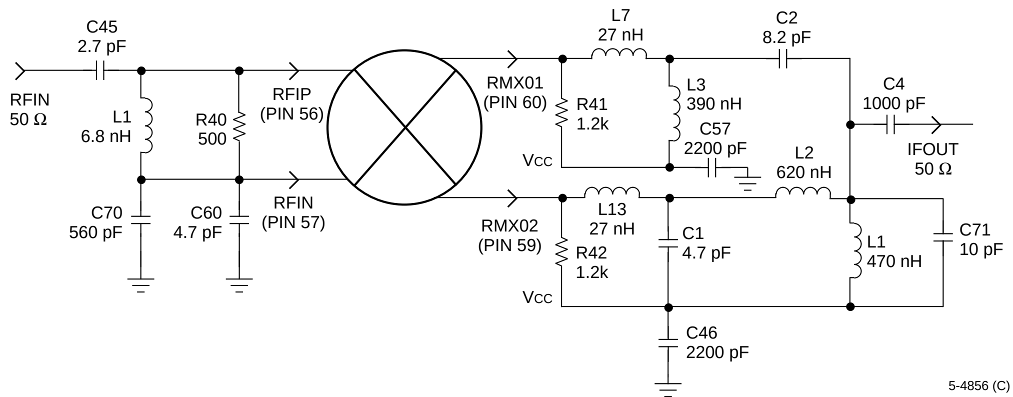


Figure 2. W2020 RF Mixer Input and Output Matching Networks

Single-Ended Input

The input matching network of Figure 2 is also documented in the *Evaluation Board Schematic SID-95RD2092, Sheet #2*, which is available with the evaluation board.

Noise Figure Considerations

As seen in Figure 1, the input is internally biased and would require a dc block in the input matching network. With the negative input ac grounded, the nominal input impedance at 947 MHz for the RF mixer is **Z_{IN} = 88 – j120. The matching network transforms this impedance down to 50** Ω. A high-pass network is chosen to provide a low impedance (shunt inductor) looking back toward the RF image SAW filter for the input transistor shot noise. Improved performance could be achieved by also using a series inductor (along with a different value series cap) to add a low-pass characteristic to the matching network to filter out noise at the RF image frequency UHF LO + 71 MHz IF. The mixer input impedance measured differentially is shown in Figure 23; the corresponding input S-parameters are shown in Table 3.

Another issue critical to achieving the best gain and noise figure performance is to have a very low-impedance ac ground on the opposite input of the emitter-coupled pair. A poor RF bypass on RFIN (pin 57 of Figure 1) will act as an attenuator to the RF input signal and will decrease the mixer conversion gain and directly increase the noise figure. To optimize the performance, a small value capacitance (such as C60 in Figure 3), which will tune out the PCB route and IC package parasitic inductances, works better than a larger value capacitor for RF bypassing. **This is critical for achieving the best noise figure for the mixer**. The capacitor must be placed close to the W2020 pin 57; otherwise, the capacitor value goes to zero in order to tune out the inductance at RF. On the evaluation board, the 4.7 pF capacitor C60 is connected through a via, but it is very close to the W2020 IC. The actual value for the capacitor is dependent on the terminal manufacturer's layout. The optimum capacitor value is found by peaking the conversion gain through the mixer. This also gives the best noise figure.

Input/Output Matching Network (continued)

Also important in the input matching network is to provide a larger (than C60) bypass capacitor on pin 57 to shunt IF noise at the input. On the evaluation board, a value of 560 pF with the PCB-layout inductance looks like a short at 71 MHz at the IC solder pad. This bypass cap has a secondary but noticeable effect on the mixer noise figure.

Compression Point Considerations

The mixer input matching network of Figure 3 also has a shunt 500 Ω resistor at the input. This is necessary to achieve the proper input compression point performance. Design analysis showed that a shunt value of 470 Ω gave optimum performance by improving the compression point by 2 dB, but since it also shunted higher-frequency shot noise of the input transistor, the noise figure degraded only by about 0.5 dB from the value obtained without the resistor. An external 500 Ω resistor is used to take into account the finite Q of the shunt inductor in the network, which effectively introduces a parallel loss in shunt with the discrete input resistor. **This resistor is critical to properly setting the mixer compression point.**

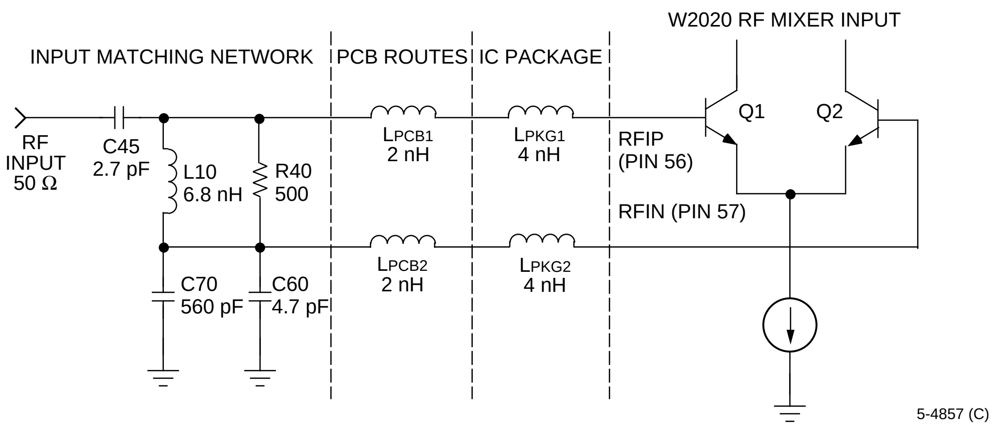


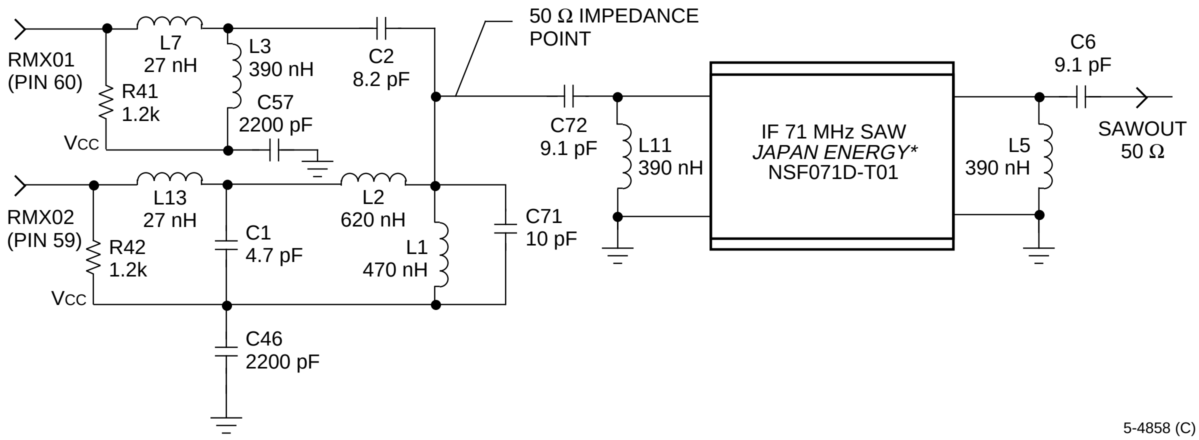
Figure 3. Mixer Input Matching Network and Parasitic Elements Consideration

Differential to Single-Ended Output

The output matching network is also documented in the *Evaluation Board Schematic SID-95RD2092, Sheet #2*, which is available with the evaluation board.

The matching network uses external 1.2 kΩ collector load resistors to nominally set the gain to about 8 dB. (See Figure 4.) The series 27 nH inductors in each output help to low-pass filter the UHF LO feedthrough to the IF output but do not affect the 71 MHz IF signal. The nominal single-ended output impedance for each output of the mixer at 71 MHz is **Z_{OUT} = 200 – j637**. The matching network transforms this impedance to 50 Ω, but also does a combined 180° phase shift of the originally balanced output signals so that they combine in phase at the single-ended 50 Ω output of the matching network. The mixer output impedance measured differentially is shown in Figure 24; the corresponding output S-parameters are shown in Table 4.

Input/Output Matching Network (continued)



* *Japan Energy* is a trademark of Japan Energy Corporation.

Figure 4. Balanced RF Mixer Output Matching Network to 50 Ω to Unbalanced IF SAW Input

The derivation of this matching network/power combiner was done with 2.5 kΩ load resistors, giving a nominal mixer gain of 10.7 dB, but the same equations applied and were used for the current network with 1.2 kΩ resistors. The derivation of the matching network is presented in the following section.

RF Mixer Output Matching Network to 50 Ω

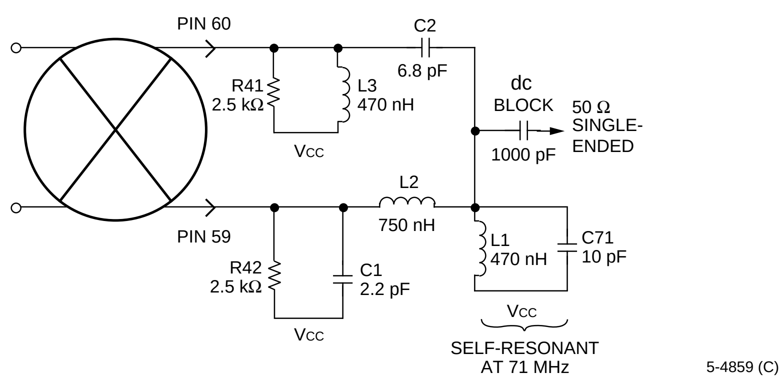


Figure 5. W2020 Output Circuit with 2.5 kΩ Resistors

Input/Output Matching Network (continued)

Theory

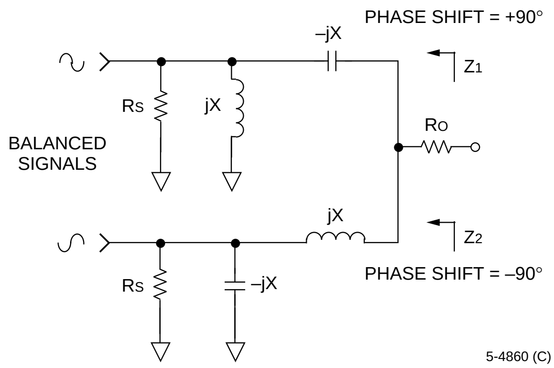


Figure 6. Ideal Circuit

Signals will combine at R_O in-phase when X = geometric mean of R_S and 2 • R_O:

$$X = \sqrt{2 \bullet R_S \bullet R_O} \tag{1}$$

From the ideal circuit in Figure 6:

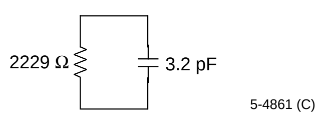
$$Z_1 = \frac{2 \bullet R_S \bullet R_O}{R_S + 2 \bullet R_O} - j\frac{2 \bullet R_O \bullet \sqrt{2 \bullet R_O \bullet R1}}{R_S + 2 \bullet R_O} \tag{2}$$

$$Z_2 = \frac{2 \bullet R_S \bullet R_O}{R_S + 2 \bullet R_O} + j\frac{2 \bullet R_O \bullet \sqrt{2 \bullet R_O \bullet R1}}{R_S + 2 \bullet R_O} \tag{3}$$

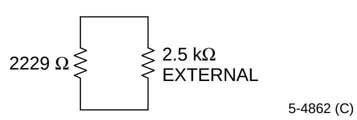
Example

W2020 Pin 59, Pin 60 at 71 MHz.

S_{22} = 200 – j637



R_S =



R_S = 1.178 kΩ

R_O = 50 Ω

Match each half-circuit to these impedance values:

$$Z_1 = 92.1 - j26.9 \tag{4}$$

$$Z_1 \parallel Z_2 = 50\ \Omega \tag{5}$$

$$Z_2 = 92.1 + j26.9 \tag{6}$$

Find the necessary reactance for the combined phase shift = 180°:

$$X = \sqrt{2 \bullet 50 \bullet 1178}\ \Omega = 343.3\ \Omega \tag{7}$$

Reactance = $X_L = X_C$

Input/Output Matching Network (continued)

Solve for the reactive components in the matching network:

$$L = \frac{X_L}{2\pi f} = \frac{343.3\ \Omega}{2\pi \bullet 71\ \text{MHz}} = 770\ \text{nH} \quad (8)$$

Use 750 nH standard value:

$$C = \frac{1}{2\pi f X_C} = \frac{1}{2\pi \bullet 71\ \text{MHz} \bullet 343.3\ \Omega} = 6.5\ \text{pF} \quad (9)$$

Use a 6.8 pF standard value. This gives actual values for series C2 and L2 components.

The shunt L and C (of Figure 6) both must have the shunt parasitic capacitance deembedded to obtain the actual external component to use. The total reactance of the shunt component and parasitic capacitance at its connecting node must equal the result X = 343.3 Ω found in (7). This is demonstrated in Figure 8.

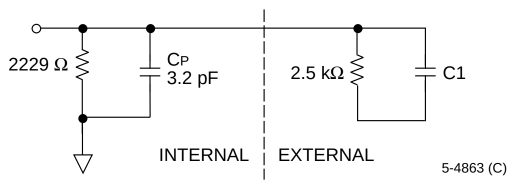


Figure 7. Pin 59 Shunt Capacitor C1

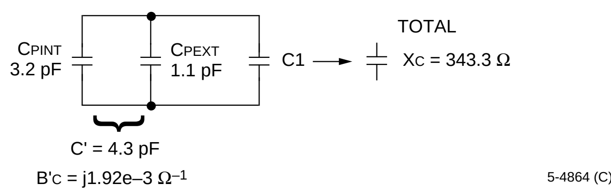


Figure 8. Deembedding of Shunt Parasitics to Find the Value of C1

$$C' = 4.3\ \text{pF} \quad (10)$$

$$B'_C = +j1.92e{-}3\ \Omega^{-1} \quad (11)$$

The 2.5 kΩ resistor and board parasitics give a total external parasitic capacitance of 1.1 pF. The value was determined from calculations once the proper value of C1 (which gave the desired Z_2 half-circuit impedance) was found experimentally. The total capacitive susceptance B_C is found from:

$$B_C = \frac{1}{X_C} = \frac{1}{343.3\ \Omega} = +j2.913e{-}3\ \Omega^{-1} \quad (12)$$

$$B_C = B'_C + B_{C1} \quad (13)$$

$$B_{C1} = B_C - B'_C = 0.993e{-}3\ \Omega^{-1} \quad (14)$$

The reactance for C1 is found by:

$$X_{C1} = \frac{1}{B_{C1}} = -j1.01e3\ \Omega \quad (15)$$

$$C1 = \frac{1}{2\pi \bullet 71\ \text{MHz} \bullet X_{C1}} = 2.2\ \text{pF} \quad (16)$$

Input/Output Matching Network (continued)

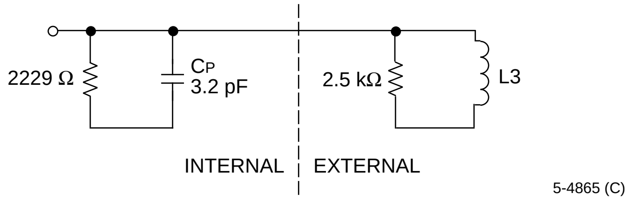


Figure 9. Pin 60 Shunt Inductor L3

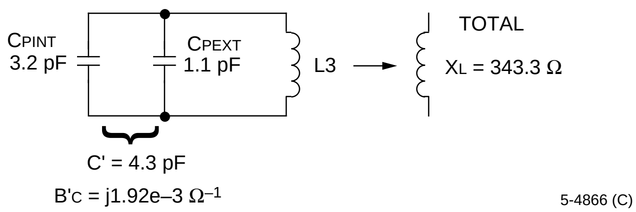


Figure 10. Deembedding of Shunt Parasitics to Find the Value of L3

The total inductive susceptance B_L is found from:

$$B_L = \frac{1}{X_L} = \frac{1}{343.3\ \Omega} = -j2.913e{-3}\ \Omega^{-1} \quad (17)$$

$$B_L = B'_C + B_{L3} \quad (18)$$

$$B_{L3} = B_L + B'_C = -j4.83e{-3}\ \Omega^{-1} \quad (19)$$

The reactance for L3 is found by:

$$X_{L3} = \frac{1}{B_{L3}} = j206.9\ \Omega \quad (20)$$

$$L_S = \frac{X_{L3}}{2\pi \bullet 71\ \text{MHz}} = 463.8\ \text{nH} \quad (21)$$

Use 470 nH standard value.

Measured Results

1. Pin 60 half-circuit Z_1 = 89 – j29, +22° phase shift.
2. Pin 59 half-circuit Z_2 = 87 + j21, –151° phase shift.
3. Combined output = 50 + j10, 173° phase shift.

Note: Since the starting impedance S_{22} of the mixer output is not purely resistive, the phase shift in each half-circuit is offset from 90°. The criterion is only that the combined phase shift be near 180°.

Effects of Phase Imbalance

As the combined phase shift for the total circuit departs from 180°, the mixer output power through the matching network will decrease. This is due to the signals combining slightly out of phase at the output of the mixer matching network. From the measured results for the output matching network, the small phase imbalance from the ideal 180° shift was 7°. (See Figure 11.)

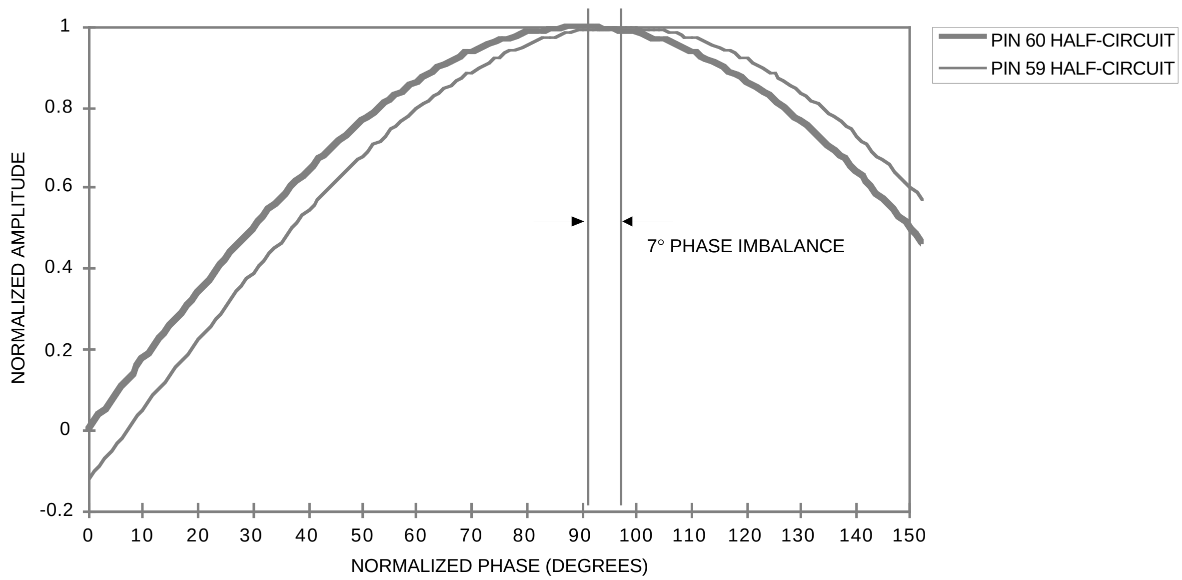


Figure 11. Phase Balance of Signals at Matching Network Output

Once the phase imbalance at the output of the matching network is known, the additional loss due to the non-ideal combined phase shift of the mixer output signals can be calculated:

$$\text{Loss} = 20 \log (\cos (0.5 * \text{Phase Imbalance})) = 20 \log (\cos(3.5°)) \quad (22)$$

$$= -0.016 \text{ dB}$$

For small values of phase imbalance, the loss will be insignificant but will increase greatly for larger phase imbalance (>20°) due to the logarithmic relationship.

PCB Layout

These notes make reference to the W2020 evaluation board *PCB Layout Plots SID-95RD2092, Sheets #4 & 7*, which are available with the evaluation board.

The critical aspects of location of external components in the mixer input matching network were discussed in the Single-Ended Input section and can be seen in the PCB layout plots. The top-level layout is shown in Figure 12. The via which connects the 4.7 pF C60 bypass capacitor to the W2020 RFIN (pin 57) is located close to the W2020 pad underneath the W2020 device. The routes to the differential input and output of the mixer are kept as similar as possible, especially at the RF input.

Since many of the components in the output matching network are referenced to VCC, it is important to have the supply well bypassed to ground so as not to introduce additional impedance into the matching network which would offset the phase shift in each output path and thereby degrade the power-combining effect of the output matching network.

PCB Layout (continued)

The IF SAW filter is positioned near the corner of the W2020 IC in such a way that the signal flow through the mixer output network is orthogonal to the signal flow through the IF strip input matching network. This is to achieve the best isolation of the 71 MHz IF signal across the SAW filter connection so as not to degrade the out-of-band rejection of the SAW filter. The evaluation board has a dual pad arrangement to allow either the *Japan Energy* NSF071D-T01 SAW filter or the GEC *Plessey*[*] DW9276 71 MHz SAW filter to be used.

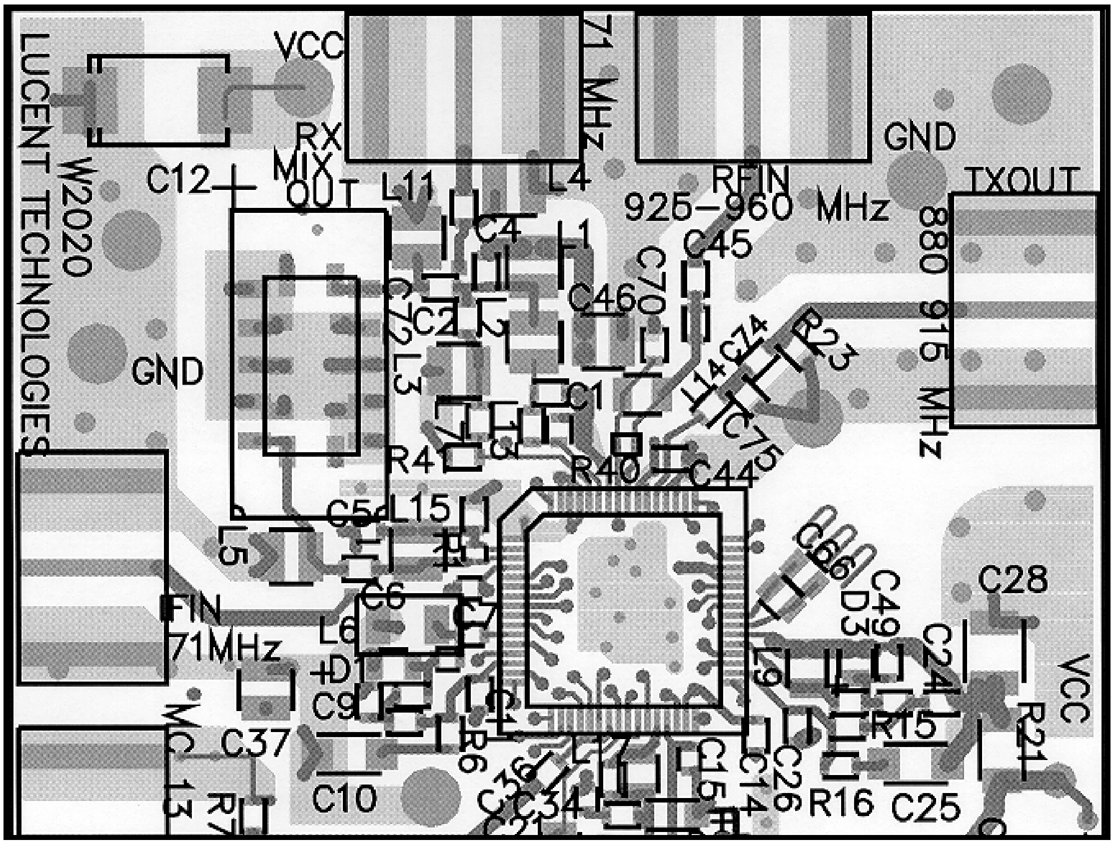


Figure 12. PCB Layout for RF Mixer Matching Network and IF SAW Filter

The substrate parameters for the mixer input and output PCB routes are shown in Table 1.

Table 1. RF Mixer PCB Route/Microstrip Substrate Parameters for Simulations

Parameter	MSUB EVB
Dielectric Constant	ER = 4.5
Loss Tangent	TAND = 0.03
Resistivity (relative to copper)	RHO = 1.42 (Gold)
Metal Thickness	T = 1.4 mils (0.0014 in.)
Roughness	RGH = 0.055
Units (for mils or thousandths of an inch)	0.0254
Height (above GND reference plane)	H = 34 mils (0.034 in.)

[*] *Plessey* is a registered trademark of The Plessey Company Limited.

RF Mixer Performance

The critical RF mixer performance parameters, typical results, and electrical specifications are listed in Table 2. The results are obtained with the evaluation board, and assume a shunt 500 Ω resistor at the input, 1.2 kΩ load resistors at the outputs, and 50 Ω input and output matching networks as described in this application note.

Table 2. Key RF Mixer Electrical Specifications and Typical Performance
(V_{CC} = 3.0 V ± 10%, T_A = 25 °C ± 3 °C)

Parameter	Min	Typical	Max	Unit
Conversion Power Gain	6.0	8.5	10	dB
Noise Figure (DSB)	—	7.8	9.0	dB
Input P-1dB Compression Point	–10	–8.0		dBm
Input Third-order Intercept Point (IP3)	—	0	—	dBm

Conversion Gain

Conversion gain is specified from the RF input to IF output including the 50 Ω matching networks, and it is true power gain. Figures 13—15 show the mixer gain versus RF input, V_{CC}, and temperature.

Measurement Method

The gain is measured by first applying a single-tone RF input of known power in dBm, in the EGSM RX band frequency range 925 MHz to 960 MHz. Then the 71 MHz IF output power level is measured with the increase being the conversion gain of the mixer. While a 50 Ω spectrum analyzer can be used, the *HP* 431* power meter was used to provide sufficient resolution for looking at small gain changes with temperature. Best results were obtained using a calibrated pad and 300 MHz low-pass filter on the mixer output into the power sensor.

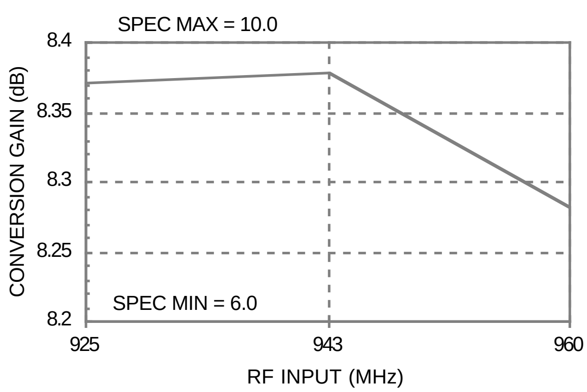


Figure 13. RF Mixer Gain vs. RF Input Frequency

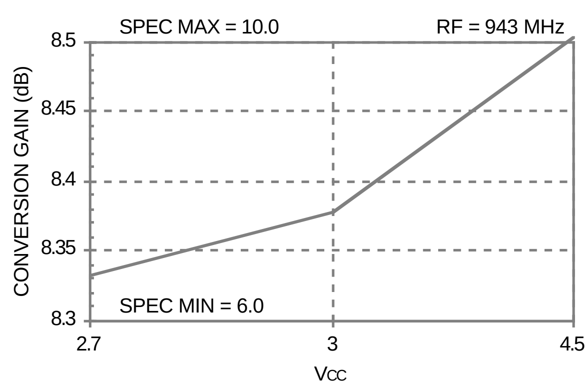


Figure 14. RF Mixer Gain vs. Vcc

* *HP* is a registered trademark of Hewlett-Packard Company.

RF Mixer Performance (continued)

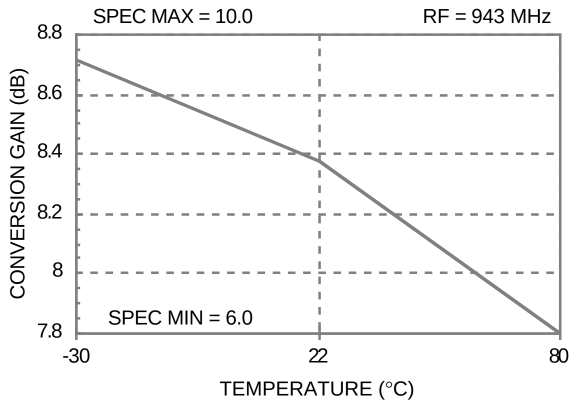


Figure 15. RF Mixer Gain vs. Temperature

Input P-1dB Compression Point

As previously mentioned, the RF mixer must have a large dynamic range to have acceptable noise figure for the receiver sensitivity, and it must not go into gain compression under blocking signal conditions. Since the maximum signal level at the mixer input is for the condition of single-tone blocking signals in GSM, the input P-1dB compression point is a more critical specification than the third-order intercept point (IP3) which relates to nonlinear intermodulation effects in the mixer.

Figures 16—19 show the mixer gain compression characteristics and also the input referred P-1dB compression point vs. RF input, V_{CC}, and temperature.

Measurement Method

In a similar method to making the gain measurements, the RF input level was increased gradually until the mixer went into 2 dB to 3 dB of gain compression. Power in and power out measurements were made to determine the 1 dB gain compression point.

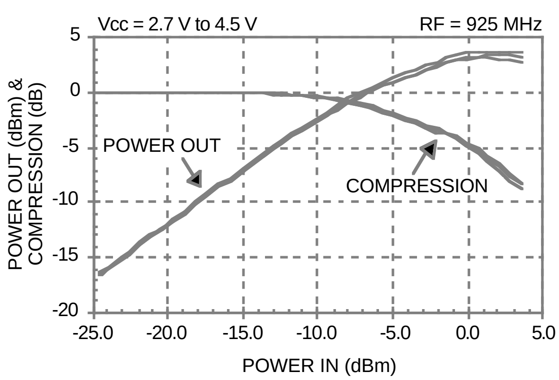


Figure 16. RF Mixer Gain Compression Characteristics

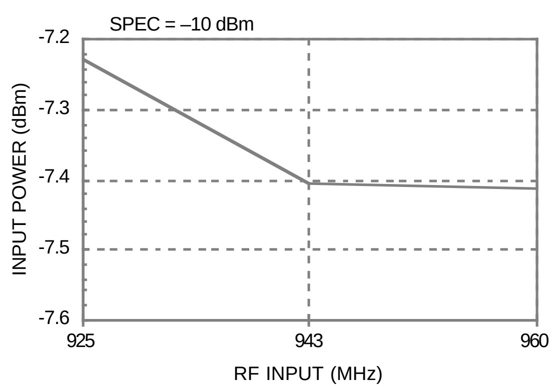


Figure 17. RF Mixer Input P-1dB Compression Point vs. RF Input Frequency

RF Mixer Performance (continued)

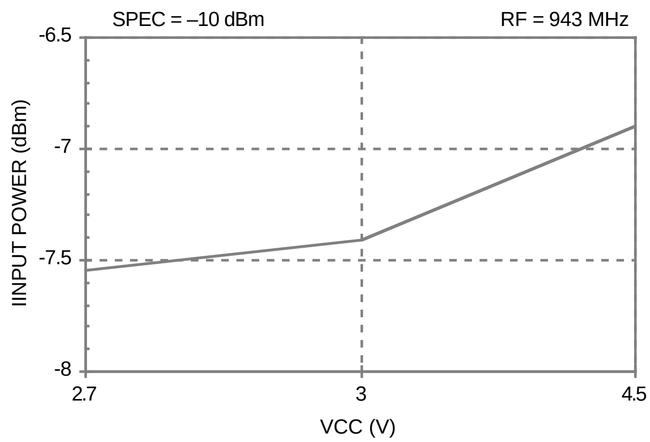


Figure 18. RF Mixer Input P-1dB Compression Point vs. Vcc

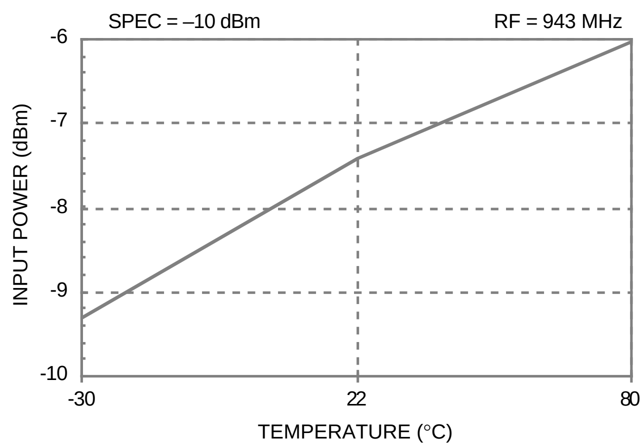


Figure 19. RF Mixer Input P-1dB Compression Point vs. Temperature

RF Mixer Performance (continued)

Noise Figure

Figures 20—22 show the mixer noise figure versus RF input, VCC, and temperature.

Measurement Method

The noise figure was measured using a broadband noise source at the RF input and an *HP* 8970B Noise Figure Meter. The meter was used in the 1.3 special function mode with the IF at 71 MHz, and it used a varying UHF LO. The evaluation board input matching network has 2.1 dB rejection of the RF image frequency. All measurements were corrected to give single sideband (SSB) readings in the meter.

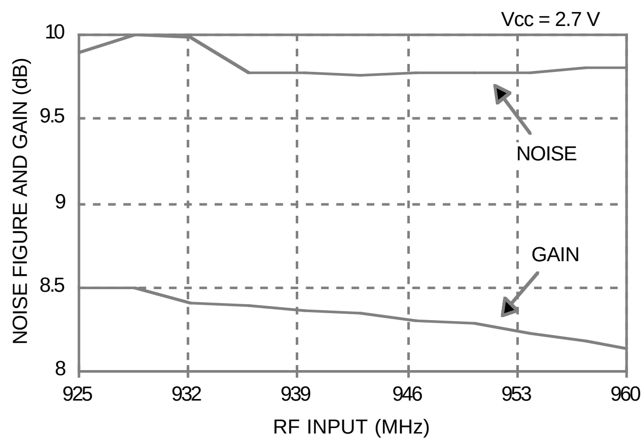


Figure 20. RF Mixer SSB Noise Figure vs. RF Input Frequency

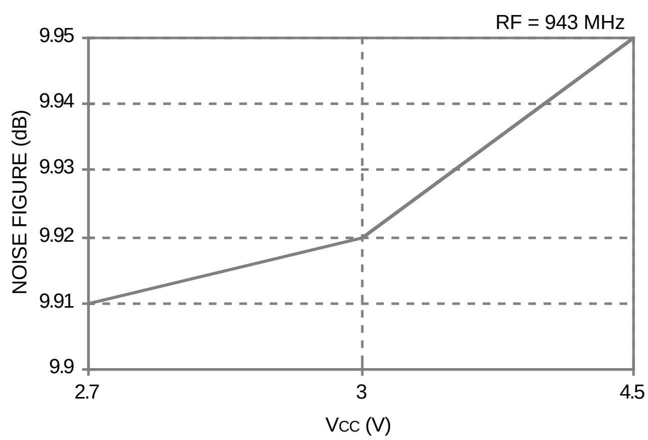


Figure 21. RF Mixer SSB Noise Figure vs. Vcc

RF Mixer Performance (continued)

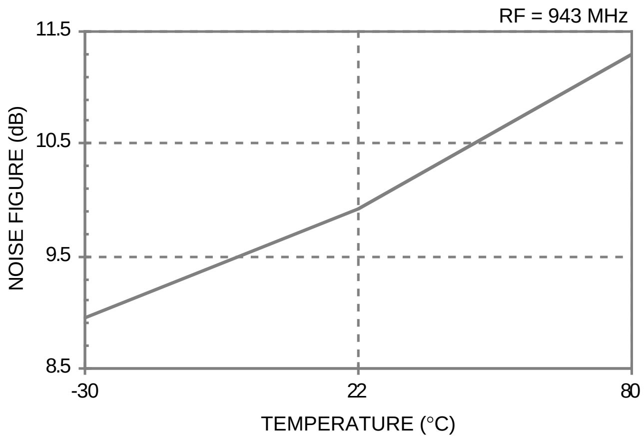


Figure 22. RF Mixer SSB Noise Figure vs. Temperature

S-Parameters

The RF mixer input and output S-parameters are measured differentially, although the mixer in the current application is driven single-ended at the input. The input S-parameters in Table 3 are measured with Port 1 connected to RFIP (pin 56) and Port 2 connected to RFIN (pin 57). The output S-parameters in Table 4 are measured with Port 1 connected to RMX02 (pin 59) and Port 2 connected to RMX01 (pin 60).

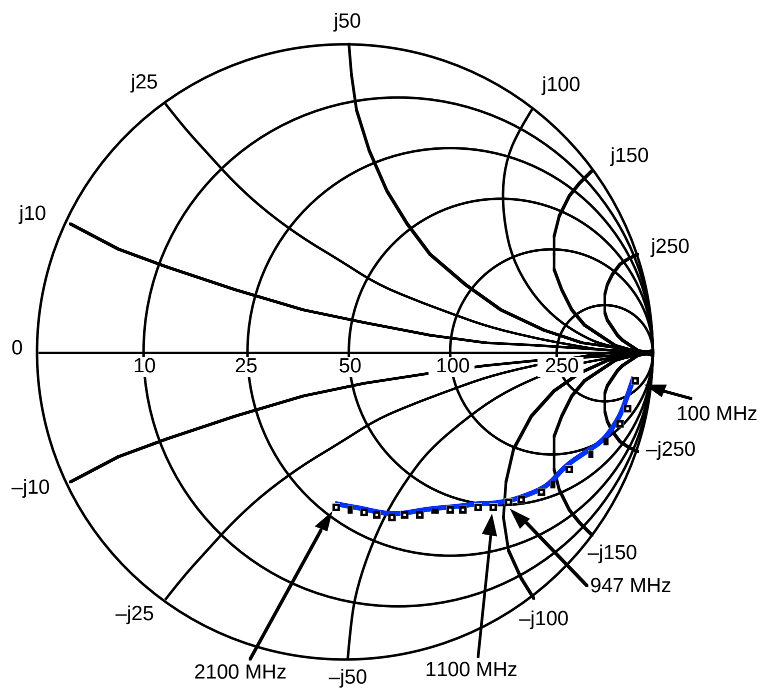


Figure 23. W2020 RF Mixer Input Impedance (S11)

RF Mixer Performance (continued)

Table 3. W2020 RF Mixer Differential Input S-Parameters

File: Rfmxdfin.s2p, W2020B RF Mixer Differential Input 25 °C, Vcc = 3.0 V								
Frequency (MHz)	S-Parameter							
	S11 M	S11 A	S21 M	S21 A	S12 M	S12 A	S22 M	S22 A
	S MA R	50						
100	0.9193	–6.102	0.0352	50.941	0.0389	53.687	0.9144	–6.267
200	0.8947	–11.579	0.0734	46.269	0.074	45.7	0.8932	–11.776
300	0.8642	–16.21	0.0973	34.866	0.0996	35.672	0.864	–16.661
400	0.8312	–20.388	0.1243	27.108	0.119	27.687	0.8301	–20.274
500	0.8048	–24.153	0.1411	21.135	0.1357	19.643	0.812	–23.559
600	0.7881	–27.578	0.1498	9.663	0.1461	9.519	0.8018	–27.992
700	0.7744	–32.091	0.1358	–5.442	0.1515	–2.964	0.7918	–32.058
800	0.7491	–36.953	0.1296	–15.777	0.138	–13.305	0.7647	–37.788
900	0.7157	–41.328	0.1212	–20.143	0.124	–18.842	0.7357	–43.087
947.5	0.7043	–43.39	0.0973	–21.197	0.1067	–21.543	0.7284	–44.645
1000	0.6888	–45.417	0.1019	–18.632	0.1062	–20.315	0.714	–48.074
1100	0.6594	–49.051	0.0858	–17.347	0.0944	–22.024	0.6933	–52.63
1200	0.6297	–52.912	0.0754	–18.79	0.0834	–24.307	0.672	–58.187
1300	0.6058	–55.898	0.0681	–13.605	0.0736	–25.957	0.6503	–63.864
1400	0.584	–58.992	0.0618	–8.483	0.0668	–26.463	0.6318	–71.614
1500	0.5765	–63.039	0.0599	–4.518	0.0592	–25.993	0.6083	–78.229
1600	0.5644	–67.681	0.0562	4.51	0.0459	–28.621	0.5834	–85.956
1700	0.5441	–72.805	0.0535	18.901	0.0284	–29.109	0.5645	–94.368
1800	0.5231	–77.395	0.0542	28.329	0.0125	9.095	0.5403	–103.962
1900	0.5094	–82.229	0.057	29.388	0.0158	84.827	0.5198	–112.656
2000	0.4965	–88.304	0.0714	33.063	0.0454	91.377	0.496	–121.878
2100	0.4769	–94.367	0.0816	34.904	0.0608	86.941	0.4858	–132.115

RF Mixer Performance (continued)

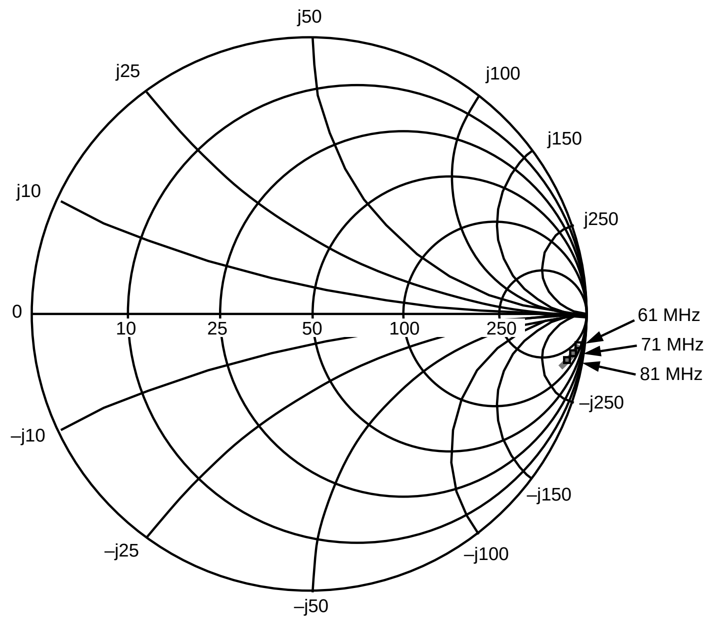


Figure 24. W2020 RF Mixer Output Impedance (S22)

Table 4. W2020 RF Mixer Differential Output S-Parameters

File: Ifmx20ot.s2p, W2020B RF Mixer Differential Output Port 1 = Pin 59, Port 2 = Pin 60, 25 °C, VCC = 3.0 V, Span = 20 MHz								
Frequency (MHz)	Parameter							
	S11 M	S11 A	S21 M	S21 A	S12 M	S12 A	S22 M	S22 A
	S MA R	50						
61	0.9633	–6.951	0.0121	85.816	0.0115	81.366	0.9631	–7.066
63	0.962	–7.162	0.0124	86.163	0.0118	81.033	0.9627	–7.31
66	0.9605	–7.487	0.0133	85.929	0.0125	82.126	0.9609	–7.616
67	0.9599	–7.583	0.0135	85.209	0.0126	81.021	0.9605	–7.748
68	0.9583	–7.699	0.0136	85.729	0.0127	81.101	0.9602	–7.886
69	0.9586	–7.84	0.0134	83.353	0.0144	80.286	0.9595	–7.977
70	0.9581	–7.917	0.0142	85.374	0.0133	81.717	0.9588	–8.113
70.9	0.957	–8.031	0.0141	85.437	0.0134	82.357	0.9584	–8.176
71	0.9568	–8.042	0.0145	85.734	0.0134	81.668	0.9587	–8.199
71.1	0.9565	–8.062	0.0144	85.686	0.0136	81.249	0.9589	–8.207
72	0.9561	–8.104	0.0149	85.065	0.0136	81.423	0.9579	–8.314
73	0.9564	–8.206	0.0148	83.964	0.0137	82.047	0.9574	–8.406
74	0.9558	–8.367	0.0148	84.378	0.0139	81.626	0.957	–8.561
75	0.9544	–8.436	0.0155	84.16	0.0147	81.265	0.9561	–8.645
76	0.9545	–8.55	0.0154	84.857	0.0144	81.968	0.9553	–8.745
79	0.952	–8.861	0.0159	85.002	0.015	80.911	0.9537	–9.045
81	0.9505	–9.068	0.0163	85.103	0.0154	81.799	0.9517	–9.266

Notes

Notes

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