AN1994

Reviewing key areas when designing with the SA605

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Document information

Info	Content
Keywords	Gilbert cell mixer, Received Signal Strength Indicator (RSSI), Local Oscillator (LO), IF limiting amplifiers, quadrature detector, Frequency Shift Keying (FSK), Amplitude Shift Keying (ASK), Colpitts oscillator, AM rejection
Abstract	This application note provides key design considerations specifically for the SA605 and SA604A RF/IF devices. However, the concepts and techniques apply to all the RF/IF Building Blocks in the NXP SA6xx family. Many application issues are addressed in the ‘Questions and Answers’ section at the end of the application note.

Revision history

Rev	Date	Description
2.0	20140807	Application note; second release Modifications: • The format of this application note has been redesigned to comply with the new identity guidelines of NXP Semiconductors. • Legal texts have been adapted to the new company name where appropriate. • deleted application note outline • “SA602” changed to “SA602A” • “SA604” changed to “SA604A” • deleted section “Related Application Notes” • Table 3 “Application component list” updated • Added Section 5 “Abbreviations”
1.0	19971103	Application note; initial release

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1. Introduction

This application note addresses key information that is needed when designing with the SA605. Since the SA602A and the SA604A are closely related to the SA605, a brief overview of these chips will be helpful. Additionally, this application note will divide the SA605 into four main blocks where a brief theory of operation, important parameters, specifications, tables and graphs of performance will be given. A question and answer section is included at the end.

2. Background

2.1 History of the SA605

Before the SA605 was made, the SA602 (double-balanced mixer and oscillator) and the SA604 (FM IF system) existed. The combination of these two chips make up a high-performance low cost receiver. Soon after, the SA605 was created to be a one-chip solution, using a newer manufacturing process and design. Since the newer process and design in the SA605 proved to be better in performance and reliability, it was decided to make the SA602 and the SA604 under this new process. The SA602A and the SA604A were created. To assist the cost-conscious customer, Philips Semiconductors also offered an inexpensive line of the same RF products: the SA612, SA614, and SA615.

Because the newer process and design proved to be better in performance and reliability, the older chips are discontinued. Therefore, only the SA602A, SA612A, SA604A, SA614A, SA605 and SA615 are available.

Figure 1 shows a brief summary of the RF chips mentioned above. Under the newer process, minor changes were made to improve the performance. A designer, converting from the SA602 to the SA602A, should have no problem with a direct switch. However, switching from the SA604 to the SA604A, might require more attention. This will depend on how good the original design was in the system. In Section 4 "Questions and Answers", the SA604 and SA604A are discussed in greater detail. This will help the designer, who used the SA604 in their original design, to switch to the 'A' version. In general, a direct switch to the SA604A is simple.

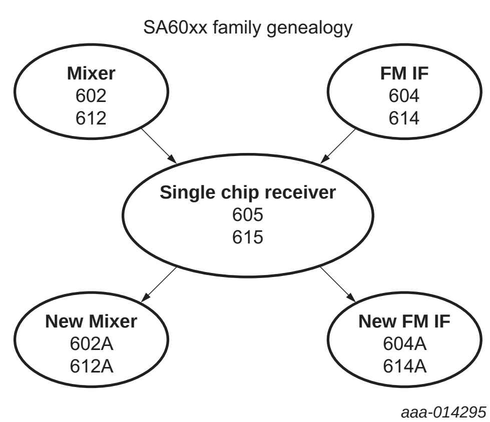


Fig 1. Overview of selected RF chips

3. Overview of the SA605

In Figure 2, the SA605 is broken up into four main areas: the mixer section, the IF section, the demodulator section, and the output section. The information contained in each of the four areas focuses on important data to assist you with the use of the SA605 in any receiver application.

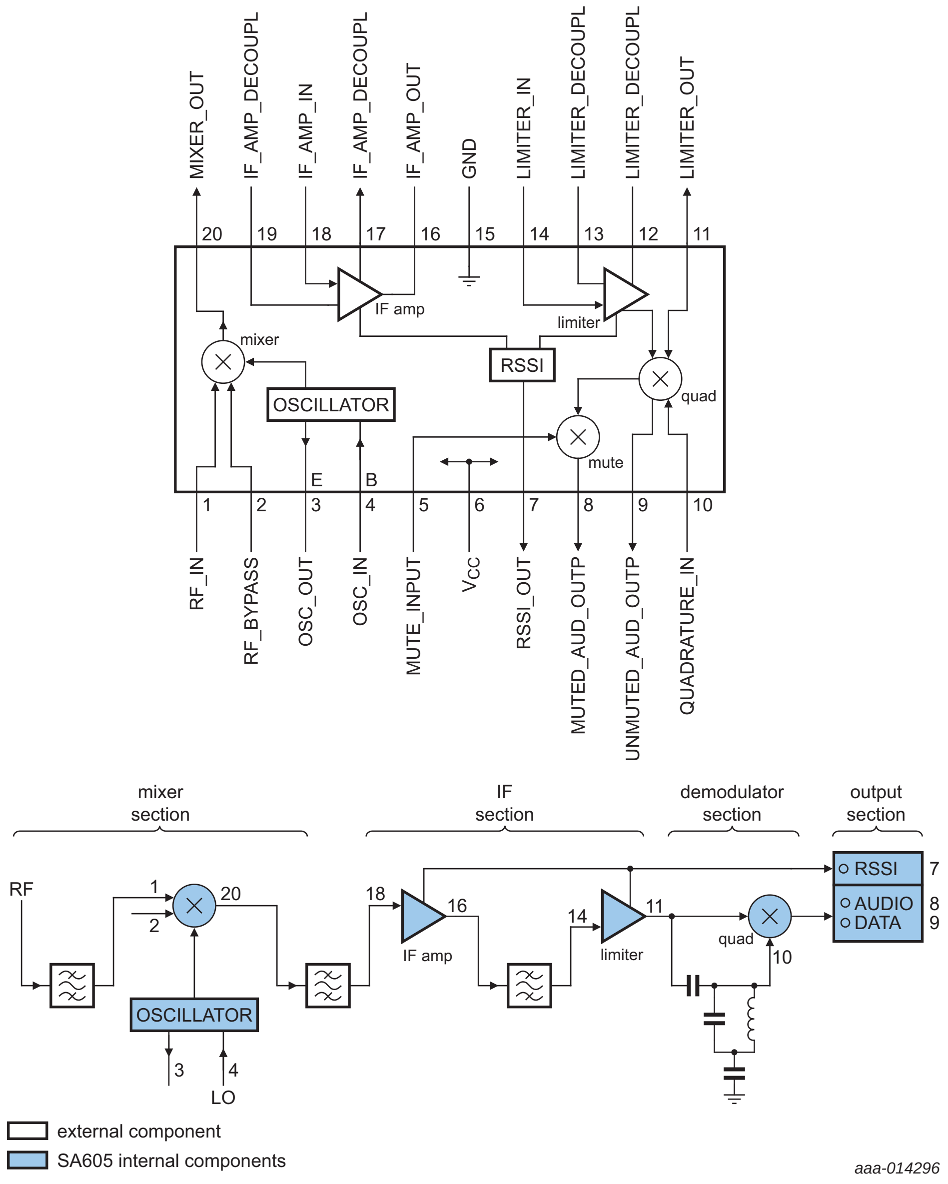
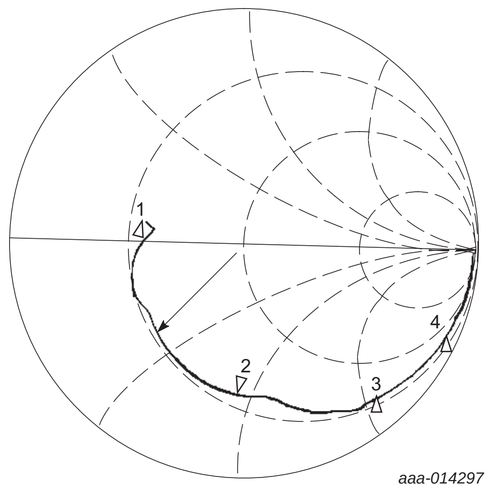


Fig 2. SA605 broken down into four areas

3.1 Mixer section

There are three areas of interest that should be addressed when working with the mixer section. The RF signal, LO signal and the output. The function of the mixer is to give the sum/difference of the RF and LO frequencies to get an IF frequency out. This mixing of frequencies is done by a Gilbert cell four-quadrant multiplier. The Gilbert cell is a differential amplifier (pins 1 and 2) that drives a balanced switching cell.

The RF input impedance of the mixer plays a vital role in determining the values of the matching network. Figure 3 shows the RF input impedance over a range of frequency. From this information, it can be determined that matching 50 Ω at 45 MHz requires matching to a 4.5 kΩ resistor in parallel with a 2.5 pF capacitor. An equivalent model can be seen in Figure 4 with its component values given for selected frequencies. Since there are many questions from the designer on how to match the RF input, an example is given in Section 3.1.1.1.



Marker 1: f = 1.5 GHz; 21 Ω || 10.6 nH

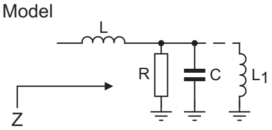
Marker 2: f = 900 MHz; 120 Ω || 3.25 pF

Marker 3: f = 500 MHz; 588 Ω || 2.75 pF

Marker 4: f = 250 MHz; 1785 Ω || 2.5 pF

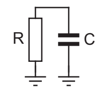
Fig 3. Smith chart of SA605 RF input impedance (pin 1 or pin 2)

- Model



- Inductor L_1 can be neglected until the frequency approaches 1 GHz (neglect 'C').

- Two element model



Frequency	R || C
10 MHz	5 kΩ || 2.5 pF
50 MHz	4651 Ω || 2.5 pF
100 MHz	3100 Ω || 2.5 pF
250 MHz	1785 Ω || 2.5 pF
500 MHz	588 Ω || 2.75 pF
750 MHz	175 Ω || 3.12 pF
900 MHz	120 Ω || 3.2 pF
1.1 GHz	48 Ω || 3.4 pF
1.56 GHz	21 Ω || 10.6 nH

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Fig 4. Equivalent model of RF input impedance

3.1.1 RF section of mixer

The mixer has two RF input pins (Pin 1 and Pin 2), allowing the user to choose between a balanced or unbalanced RF matching network. Table 1 shows the advantages and disadvantages for either type of matching. Obviously, the better the matching network, the better the sensitivity of the receiver.

Table 1. Comparing balanced versus unbalanced matching

SA605 or SA602	Matching	Advantages	Disadvantages
Pin 1 and Pin 2 (RF input)	Single-ended (unbalanced)	• very simple circuit • no sacrifice in third-order performance	• increase in second-order products
	Balanced	• reduce second-order products	• impedance match difficult to achieve

3.1.1.1 Example

Using a tapped-C network, match a 50 Ω source to the RF input of the SA605 at 45 MHz. Refer to Figure 5.

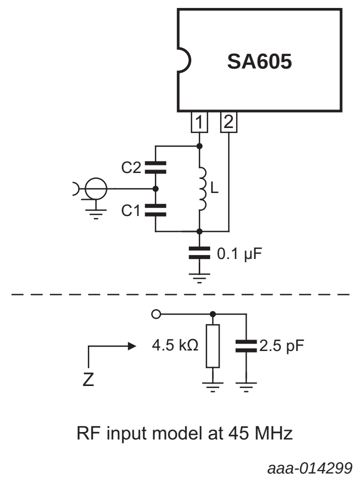


Fig 5. Tapped-C network

Step 1 Choose an inductor value and its 'Q' L = 0.22 μH Q_P = 50 (specified by the manufacturer).

Step 2 Find the reactance of the inductor.

$$X_P = 2\pi f L$$
$$= 2\pi(45\ MHz)(0.22\ \mu H)$$
$$\therefore X_P = 62.2\ \Omega$$

Step 3 Then,

$$R_P = Q_P X_P$$
$$= (50)(62.2)$$
$$\therefore X_P = 3.11\ k\Omega \text{ (the inductance resistance)}$$

Step 4

$$Q = (R_{total}/X_P)$$
$$= ((R_S' \| R_L \| R_P)/X_P)$$
$$\text{where } R_S' = R_L$$
$$= 4.5k \| 4.5k \| 3.11k/62.2$$
$$= 21.39$$
$$\therefore Q \cong 21 \text{ (the Q of the matching network)}$$

where:
R_S = source resistance
R_L = load resistance
R_S' = what the source resistance should look like to match R_L
R_P = inductance resistance

Step 5

$$\frac{C1}{C2} = \sqrt{\frac{R_S'}{R_S}} - 1 = 8.6$$

Step 6

$$C_T = \frac{1}{X_P\ \omega} = \frac{1}{(62.2)\ 2\pi\ 45\ MHz} = 56.86\ pF$$

Step 7

Using $C_T = \frac{C1C2}{C1 + C2}$, where C_T = 56.86 pF, $\frac{C1}{C2} = 8.6$

$$C_T = \frac{C1}{\frac{C1}{C2} + 1}$$

$\therefore C_1 = C_T\left(\frac{C1}{C2} + 1\right)$ and $C_2 = \frac{C1}{8.6}$,

thus:
C1 = 539 pF
C2 = 64 pF
L = 0.22 μH (value started with)

Step 8

Frequency check

$$\omega = \frac{1}{\sqrt{LC}}$$

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

f = 45 MHz (...so far so good)

Step 9

Taking care of the 2.5 pF capacitor that is present at the RF input at 45 MHz

$$\frac{C2_A}{C1_A} = \frac{64\ pF}{540\ pF}\ \text{(Equation 1)}$$

$$C_{TN} = \frac{C1_A C2_A}{C1_A + C2_A}\ \text{(Equation 2)}$$

where $C_{TN} = C_T - 2.5$ pF (recall value of C_T from Step 6).

Making use of Equations 1 and 2, the new values of C1 and C2 are:
$C1_A$ = 524 pF
$C2_A$ = 60.6 pF

Remark: At this frequency, the 2.5 pF capacitor could probably be ignored, since its value at 45 MHz has little effect on C1 and C2.

Step 10 Checking the bandwidth

$$Q = \frac{f}{BW}$$

$$BW = f_U - f_L$$

BW = bandwidth
f_U = upper 3 dB frequency
f_L = lower 3 dB frequency

Using the above formulas results in:
f_U = 46 MHz
f_L = 44 MHz
BW = 2 MHz

The above shows the calculations for a single-ended match to the SA605. For a balanced matching network, a transformer can be used. The same type of calculations will still apply once the input impedance of the SA605 is converted to the primary side of the transformer (see Figure 6). But before we transform the input impedance to the primary side, we must first find the new input impedance of the SA605 for a balanced configuration. Because we have a balanced input, the 4.5 kΩ transforms to 9 kΩ (4.5 kΩ + 4.5 kΩ = 9 kΩ), while the capacitor changes from 2.5 pF to 1.3 pF (2.5 pF in series with 2.5 pF is 1.3 pF). Notice that the resistor values double while the capacitor values are halved. Now the 9 kΩ resistor in parallel with the 1.3 pF capacitor must be transformed to the primary side of the transformer (see Figure 6).

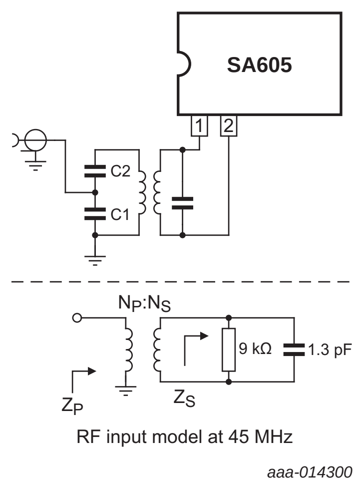


Fig 6. Using a transformer to achieve a balanced match

3.1.1.2 Procedure

Step 1

$$\frac{Z_P}{Z_S} = \left(\frac{N_P}{N_S}\right)^2$$

where:
Z_P = impedance of primary side
Z_S = impedance of secondary side
N_P = number of turns on primary side
N_S = number of turns on secondary side

Step 2

Recall,
$Z_S = R \,||\, X_C$
$Z_S = 9k \,||\, j\ 2.7k$
where:
R = 9k

$$X_C = \frac{1}{2\pi fC} = 2.7k \text{ at } f = 45\ MHz$$

Step 3

Assume 1 : N turns ratio for the transformer.

$$Z_P = \frac{Z_S}{N^2} = 2.25k \,\|\, j\ 680$$

(assuming N = 2)

Step 4

$$\therefore C = \frac{1}{2\pi fX_C} = 5.2\ pF$$

R = 2.25k
(These are the new values to match using the formulas in tapped-C.)

Step 5

Because the transformer has a magnetization inductance L_M, (inductance presented by the transformer), we can eliminate the inductor used in the previous example and tune the tapped-C network with the inductance presented by the transformer.

Let's assume L_M = 0.22 μH (Q = 50), therefore:
C1 = 381 pF
C2 = 66.8 pF
f_U = 46.7 MHz
f_L = 43.3 MHz
BW = 3.4 MHz

Taking the input capacitor into consideration:
C1 = 347 pF
C2 = 61 pF
L = 0.22 μH (Q = 50)

Because of leakage inductance, the transformer is far from ideal. All of these leakages affect the secondary voltage under load, which will seem like the indicated turns ratio is wrong. The above calculations show one method of impedance matching. The values calculated for C1 and C2 do not take into account board parasitic capacitance, and are, therefore, only theoretical values. There are many ways to configure and calculate matching networks. One alternative is a tapped-L configuration. But the ratio of the tapped-C network is easier to implement than ordering a special inductor. The calculations of these networks can be done on the Smith chart. Furthermore, there are many computer programs available which will help match the circuit for the designer.

3.1.2 Local oscillator section of mixer

The SA605 provides an NPN transistor for the local oscillator where only external components like capacitors, inductors, or resistors need to be added to achieve the LO frequency. The oscillator's transistor base and emitter (Pin 4 and Pin 3 respectively) are available to be configured in Colpitts, Butler or varactor controlled LC forms. Referring to Figure 7, the collector is internally connected directly to V_{CC}, while the emitter is connected through a 25 kΩ resistor to ground. Base bias is also internally supplied through an 18 kΩ resistor. A buffer/divider reduces the oscillator level by a factor of three before it is applied across the upper tree of the Gilbert cell. The divider de-sensitizes the mixer to oscillator level variations with temperature and voltage. A typical value for the LO input impedance is approximately 10 kΩ.

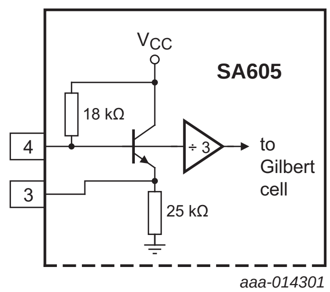


Fig 7. On-board NPN transistor for local oscillator

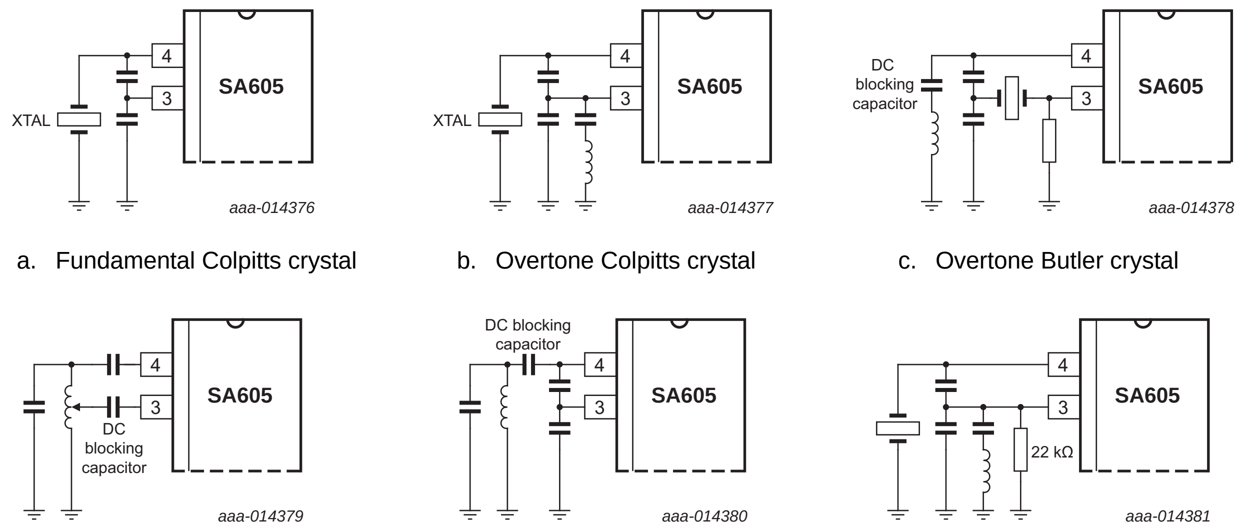
The highest LO frequency that can be achieved is approximately 300 MHz with a 200 mV (RMS) signal on the base (Pin 4). Although it is possible to exceed the 300 MHz LO frequency for the on-board oscillator, it is not really practical because the signal level drops too low for the Gilbert Cell. If an application requires a higher LO frequency, an external oscillator can be used with its 200 mV (RMS) signal injected at Pin 4 through a DC blocking capacitor. Table 2 can be used as a guideline to determine which configuration is best for the required LO frequency.

Table 2. LO configurations

LO	Suggested configuration using on-board oscillator
0 MHz to 30 MHz	Fundamental mode, use Colpitts
30 MHz to 70 MHz	Third overtone mode, use Colpitts
70 MHz to 90 MHz	Third to fifth overtone mode, use Colpitts with 22 kΩ resistor connected from the emitter pin to ground
90 MHz to 170 MHz	Use Butler, crystal in series mode, and a 22 kΩ resistor connected from the emitter pin to ground
170 MHz to 300 MHz	LC configuration

Because the Colpitts configuration is for parallel resonance mode, it is important to know, when ordering crystals, that the load capacitance of the SA605 is 10 pF. However, for the Butler configuration, the load capacitance is unimportant since the crystal will be in the series mode. Figure 8 shows the different types of LO configurations used with SA605.

If a person decides to use the Colpitts configuration in their design, they will probably find that most crystal manufacturers have their own set of standards of load capacitance. And in most cases, they are unwilling to build a special test jig for an individual's needs. If this occurs, the designer should tell them to go ahead with the design. But, the designer should also be ready to accept the crystal's frequency to be off by 200 Hz to 300 Hz from the specified frequency. Then a test jig provided by the designer and a second iteration will solve the problem.



a. Fundamental Colpitts crystal

b. Overtone Colpitts crystal

c. Overtone Butler crystal

d. Hartley L/C tank

e. Colpitts L/C tank

f. Third to fifth overtone Colpitts crystal

Fig 8. Oscillator configurations

3.1.3 Output of mixer

Once the RF an LO inputs have been properly connected, the output of the mixer supplies the IF frequency. Knowing that the mixer's output has an impedance of 1.5 kΩ, matching to an IF filter should be trivial.

3.1.4 Choosing the appropriate IF frequency

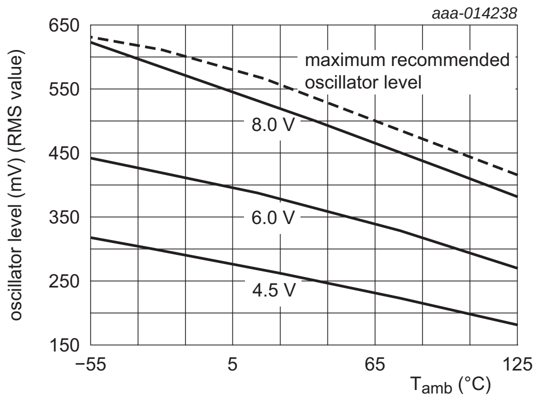
Some of the standard IF frequencies used in industry are 455 kHz, 10.7 MHz and 21.4 MHz. Selection of other IF frequencies is possible. However, this approach could be expensive because the filter manufacturer will probably have to build the odd IF filter from scratch.

There are several advantages and disadvantages in choosing a low or high IF frequency. Choosing a low IF frequency like 455 kHz can provide good stability, high sensitivity and gain. Unfortunately, it can also present a problem with the image frequency (assuming single conversion). To improve the image rejection problem, a higher IF frequency can be used. However, sensitivity is decreased and the gain of the IF section must be reduced to prevent oscillations.

If the design requires a low IF frequency and good image rejection, it is best to use the double conversion method. This method allows the best of both worlds. Additionally, it is much easier to work with a lower IF frequency because the layout will not be as critical and will be more forgiving in production. The only drawback to this method is that it will require another mixer and LO. But, a transistor can be used for the first mixer stage (which is an inexpensive approach) and the SA605 can be used for the second mixer stage. The SA602A can also be used for the first conversion stage if the transistor approach does not meet the design requirements.

If the design requires a high IF frequency, good layout and RF techniques must be exercised. If the layout is sound and instability still occurs, refer to Section 3.4.2 "RSSI output", which suggests solutions to these types of problems.

Another issue to consider when determining an IF frequency is the modulation. For example, a narrowband FM signal (30 kHz IF bandwidth) can be done with an IF of 455 kHz. But for a wideband FM signal (200 kHz bandwidth), a higher IF is required, such as 10.7 MHz or 21.4 MHz.



C_{BE} = 5.6 pF

Fig 9. SA605 application oscillator level

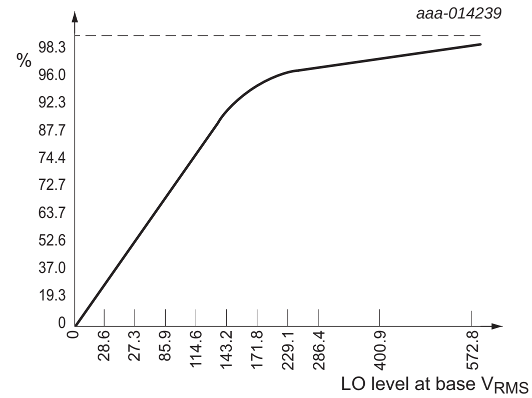
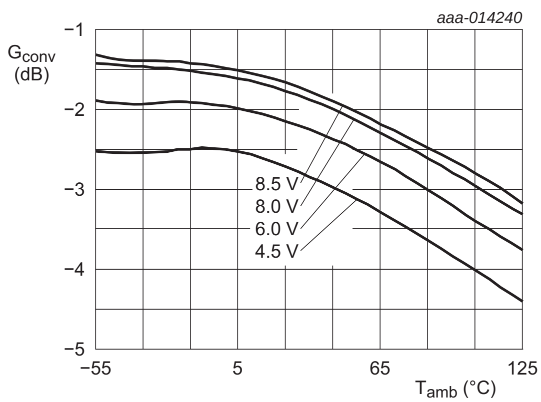
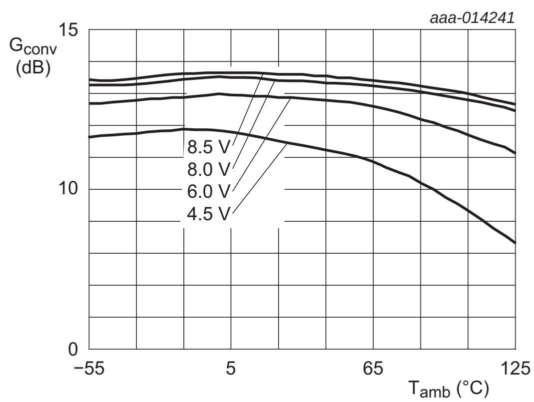


Fig 10. Mixer efficiency versus normalized LO level



50 Ω input; 1500 Ω output; external LO 220 mV

Fig 11. 50 Ω conversion gain

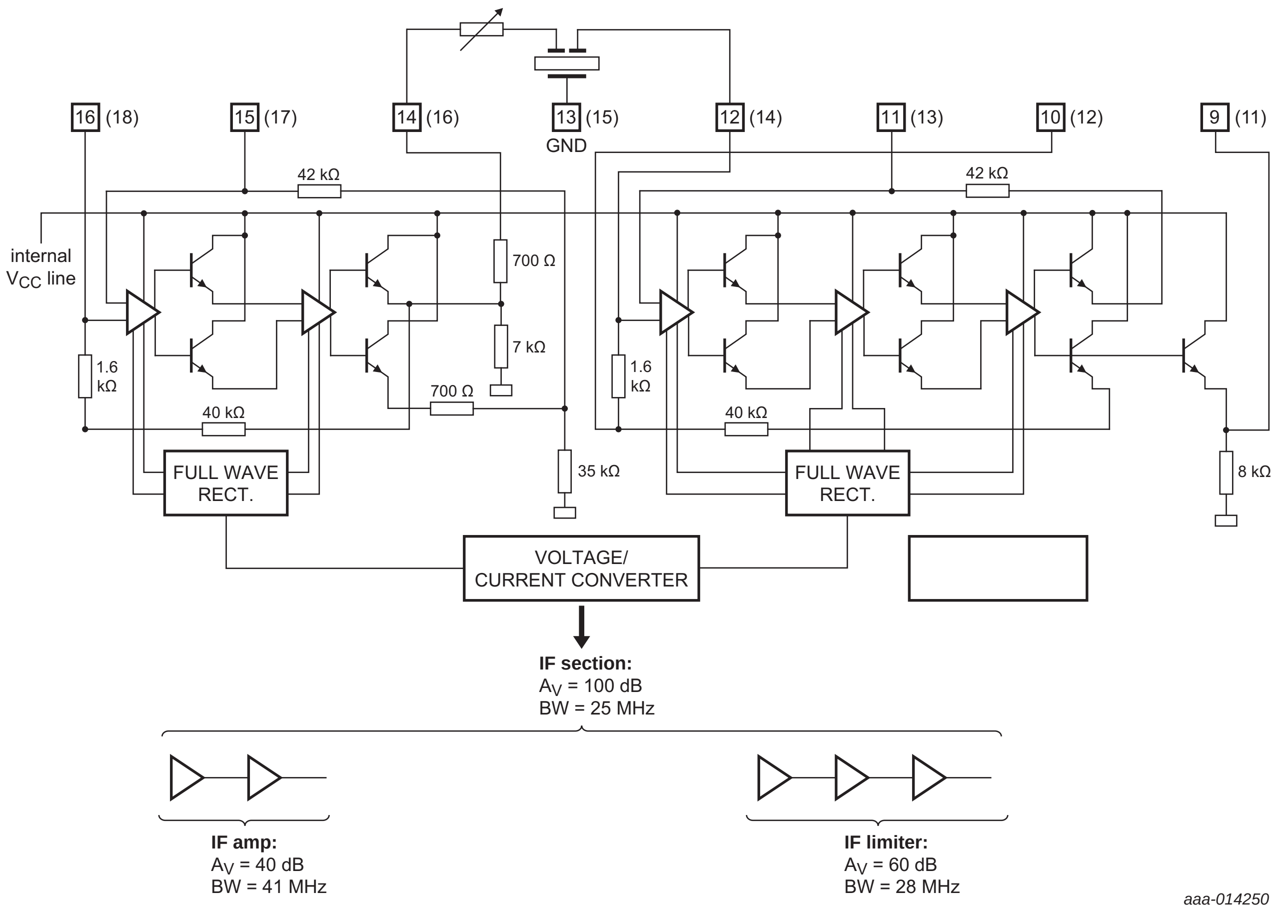


50 Ω to 1.5 kΩ, 14.5 dB matching step-up network

Fig 12. Single-ended matched input conversion gain

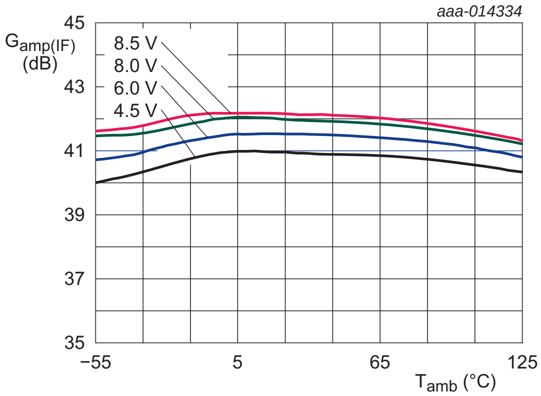
3.2 IF section

The IF section consists of an IF amplifier and IF limiter. With the amplifier and limiter working together, 100 dB of gain with a 25 MHz bandwidth can be achieved (see Figure 13). The linearity of the RSSI output is directly affected by the IF section and will be discussed in more detail later in this application note.



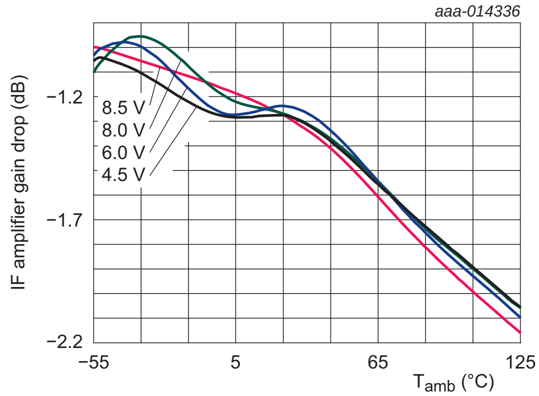
Pin numbers in parentheses are for SA605.

Fig 13. IF section of SA604A (SA605)



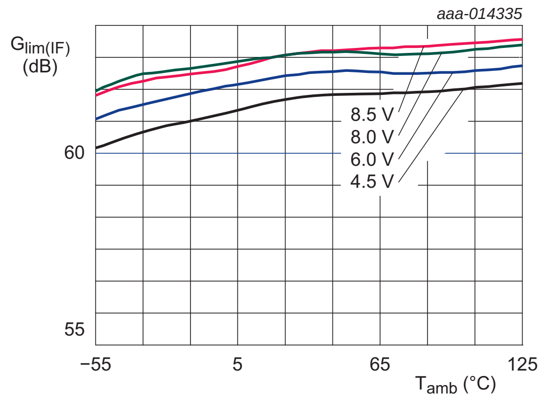
50 Ω source

Fig 14. IF amplifier gain



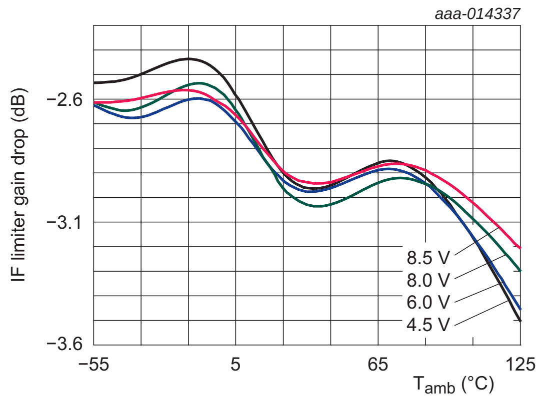
50 Ω source

Fig 15. IF amplifier gain drop, 20 MHz response



50 Ω source

Fig 16. IF limiter gain

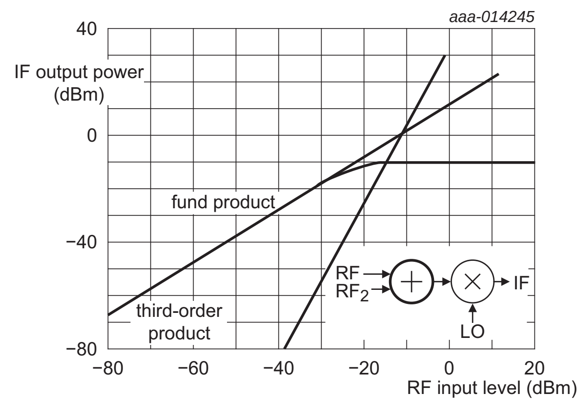


50 Ω source

Fig 17. IF limiter gain drop, 20 MHz response

3.2.1 IF amplifier

The IF amplifier is made up of two differential amplifiers with 40 dB of gain and a small signal bandwidth of 41 MHz (when driven by a 50 Ω source). The output is a low-impedance emitter follower with an output resistance of about 230 Ω, and an internal series build out of 700 Ω to give a total of 930 Ω. One can expect a 6 dB loss in each amplifier's input since both of the differential amplifiers are single-ended.



RF = 45 MHz; IF = 455 kHz; RF2 = 45.06 MHz; LO = 44.545 MHz

Fig 18. Third-order intercept and compression

The basic function of the IF amp is to boost the IF signal and to help handle impulse noise. The IF amp will not provide good limiting over a wide range of input signals, which is why the IF limiter is needed.

3.2.2 IF limiter

The IF limiter is made up of three differential amplifiers with a gain of 63 dB and a small signal AC bandwidth of 28 MHz. The outputs of the final differential stage are buffered to the internal quadrature detector. The IF limiter's output resistance is about 260 Ω with no internal build-out. The limiter's output signal (Pin 9 on SA604A, Pin 11 on SA605) will vary from a good approximation of a square wave at lower IF frequencies like 455 kHz, to a distorted sinusoid at higher IF frequencies, like 21.4 MHz.

The basic function of the IF limiter is to apply a tremendous amount of gain to the IF frequency such that the top and bottom of the waveform are clipped. This helps in reducing AM and noise presented upon reception.

3.2.3 Function of IF section

The main function of the IF section is to clean up the IF frequency from noise and amplitude modulation (AM) that might occur upon reception of the RF signal. If the IF section has too much gain, then one could run into instability problems. This is where crucial layout and insertion loss can help (also addressed later in this application note).

3.2.4 Important parameters for the IF section

3.2.4.1 Limiting

The audio output level of an FM receiver normally does not change with the RF level due to the limiting action. But as the RF signal level continues to decrease, the limiter will eventually run out of gain and the audio level will finally start to drop. The point where the IF section runs out of gain and the audio level decreases by 3 dB with the RF input is referred to as the –3 dB limiting point.

In the application test circuit, with a 5.1 kΩ interstage resistor, audio suppression is dominated by noise capture down to about the –120 dBm RF level at which point the phase detector efficiency begins to drop (see Section 3.2.4.4 "Interstage loss").

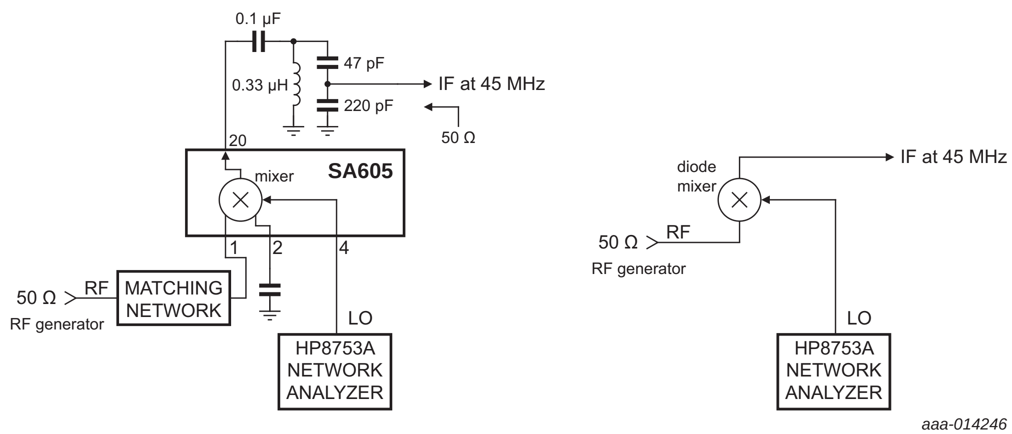


Fig 19. Test circuits for SA605 mixer versus diode mixer

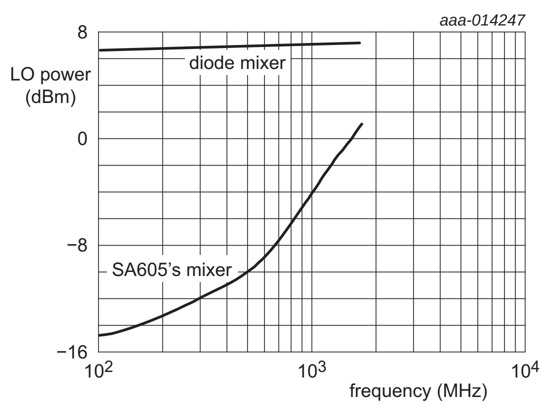


Fig 20. LO power requirements (matched input)

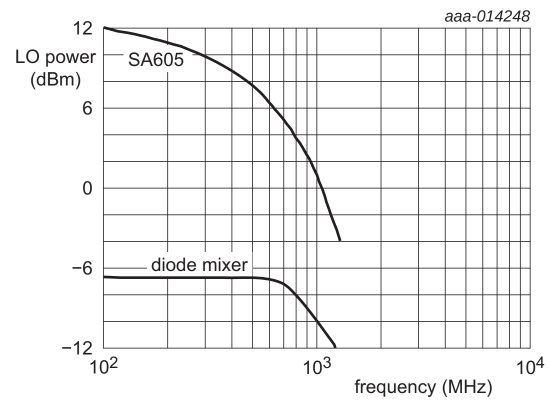


Fig 21. SA605 conversion gain versus diode mixer

The audio drop that occurs is a function of two types of limiting. The first type is as follows: As the input signal drops below a level which is sufficient to keep the phase detector compressed, the efficiency of the detector drops, resulting in premature audio attenuation. We will call this 'gain limiting'. The second type of limiting occurs when there is sufficient amount of gain without de-stabilizing regeneration (that is, keeping the phase detector fully limited), the audio level will eventually become suppressed as the noise captures the receiver. We will call this 'limiting due to noise capture'.

Figure 22 shows the 3 dB drop in audio at about 0.26 μV (RMS value), with a –118.7 dBm/50 Ω RF level for the SA605. Note that the level has not improved by the 11 dB gain supplied by the mixer/filter since noise capture is expected to slightly dominate here.

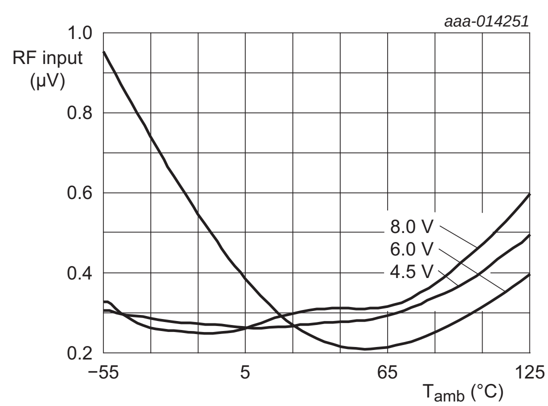
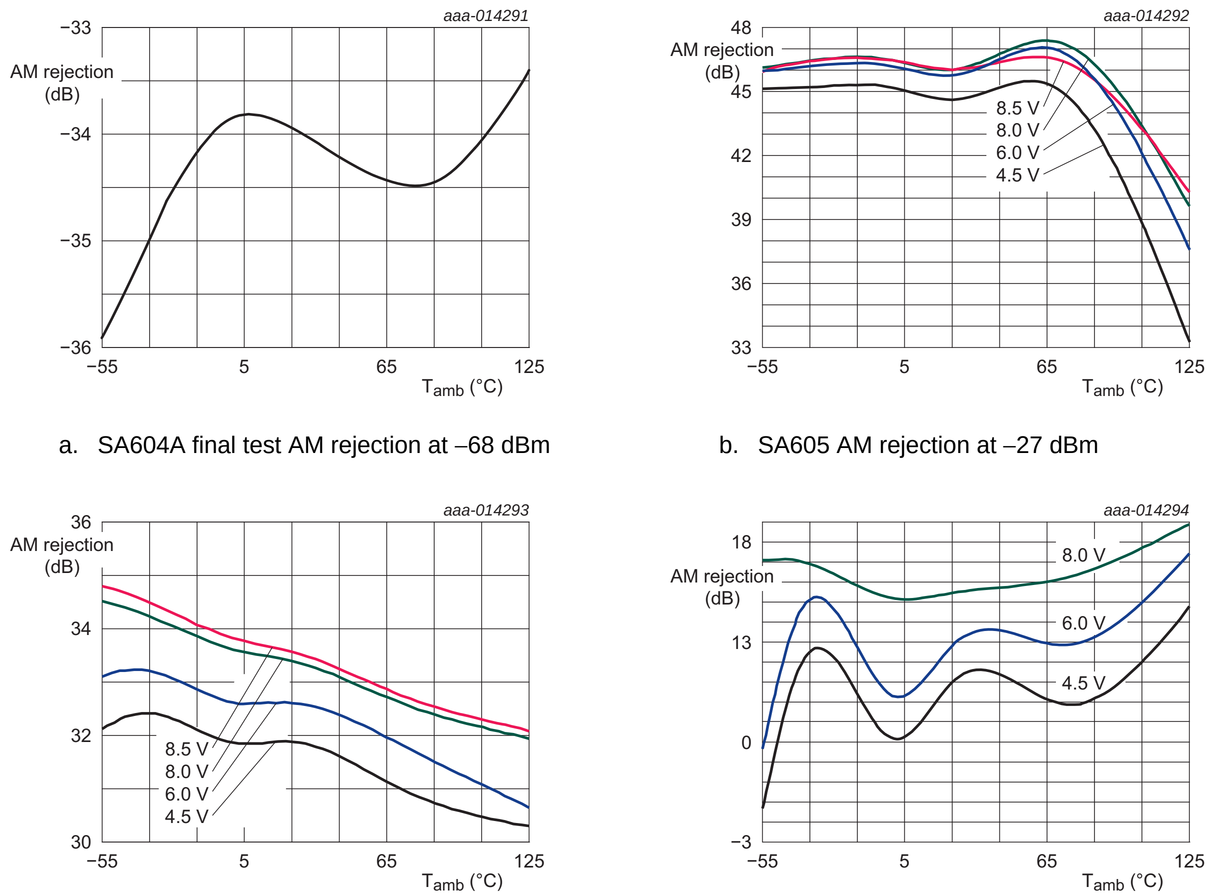


Fig 22. SA605 application board, –3 dB limiting (drop in audio)

3.2.4.2 AM rejection

The AM rejection provided by the SA605/604A is extremely good even for 80 % modulation indices as depicted in Figure 23a through Figure 23d. This performance results from the 370 mV peak signal levels set at the input of each IF amplifier and limiter stage. For this level of compression at the inputs, even better performance could be expected except that finite AM to PM conversion coefficients limit ultimate performance for high level inputs as indicated in Figure 23b.

Low level AM rejection performance degrades as each stage comes out of limiting. In particular as the quadrature phase detector input drops below 100 mV peak, all limiting will be lost and AM modulation will be present at the input of the quad detector (See Figure 23d).



a. SA604A final test AM rejection at –68 dBm

b. SA605 AM rejection at –27 dBm

c. SA605 AM rejection at –76 dBm

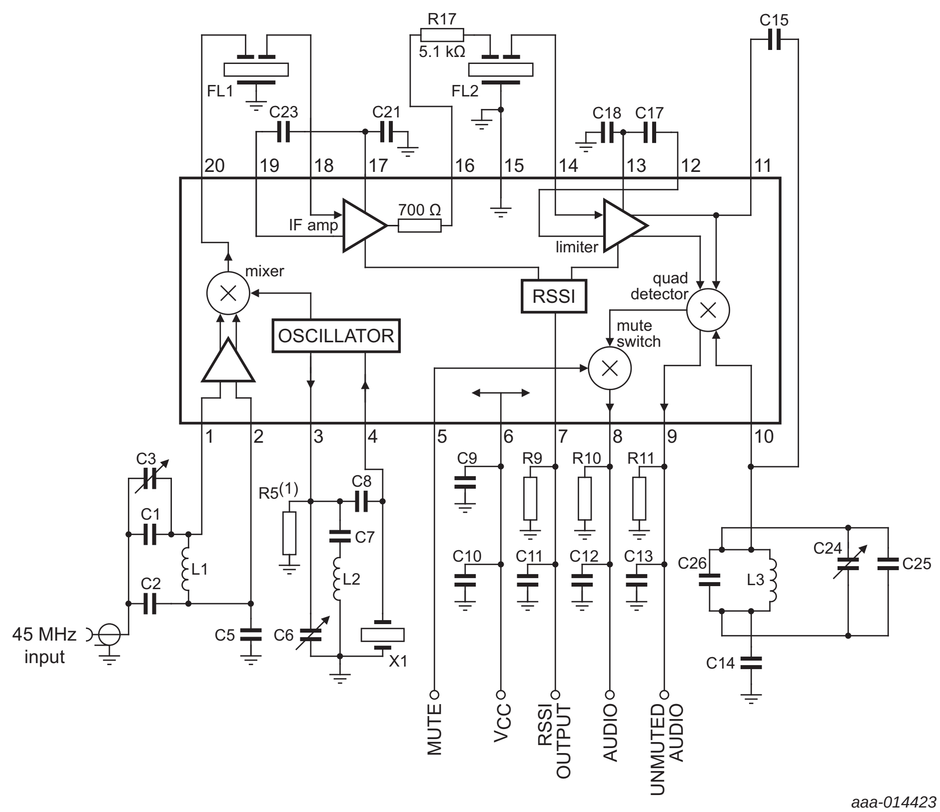
d. SA605 AM rejection at –118 dBm

Fig 23. AM rejection results at different input levels

3.2.4.3 AM to PM conversion

Although AM rejection should continue to improve above –95 dBm IF inputs, higher order effects, lumped under the term AM to PM conversion, limit the application rejection to about 40 dB. In fact this value is proportional to the maximum frequency deviation. That is lower deviations producing lower audio outputs result directly in lower AM rejection. This is consistent with the fact that the interfering audio signal produced by the AM/PM conversion process is independent of deviation within the IF bandwidth and depends to a first estimate on the level of AM modulation present. As an example reducing the maximum frequency deviation to 4 kHz from 8 kHz, will result in 34 dB AM rejection. If the AM modulation is reduced from 80 % to 40 %, the AM rejection for higher level IFs will go back to 40 dB as expected. AM to PM conversion is also not a function of the quad tank Q, since an increase in Q increases both the audio and spurious AM to PM converted signal equally.

As seen above, these relationships and the measured results on the application board (Figure 24) can be used to estimate high level IF AM rejection. For higher frequency IFs (such as 21.4 MHz), the limiter's output will start to deviate from a true square wave due to lack of bandwidth. This causes additional AM rejection degradation.



(1) R5 can be used to bias the oscillator transistor at a higher current for operation above 45 MHz. Recommended value is 22 kΩ, but should not be below 10 kΩ.

Fig 24. SA605 45 MHz application circuit

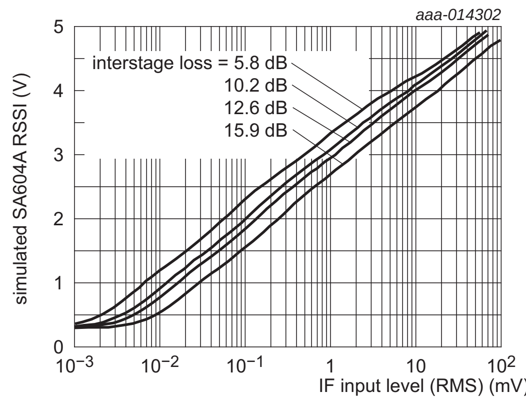
Table 3. Application component list

Component	Value	Description	Package	Part number
C1	33 pF	NPO ceramic	C0805K	445-127x-1-ND
C2	220 pF	NPO ceramic	C0805K	445-7484-6-ND
C3	5 pF to 30 pF	NPO ceramic; Murata TZC3P300A 110R00	TRIMCAP	490-1994-2-ND
C5	100 nF ± 10 %	100 nF ± 10 % monolithic ceramic	C0805K	311-1036-1-ND
C6	5 pF to 30 pF	NPO ceramic; Murata TZC3P300A 110R00	TRIMCAP	490-1994-2-ND
C7	1 nF	ceramic	C0805K	399-3293-1-ND
C8	10 pF	NPO ceramic	C0805K	490-1994-2-ND
C9	100 nF ± 10 %	monolithic ceramic	C0805K	311-1036-1-ND
C10	22 μF	tantalum	C1812	478-3117-1-ND
C11	100 nF ± 10 %	monolithic ceramic	C0805K	311-1036-1-ND
C12	15 nF ± 10 %	ceramic	C0805K	399-1161-1-ND
C13	150 pF ± 2 %	N1500 ceramic	C0805K	399-1125-1-ND
C14	100 nF ± 10 %	monolithic ceramic	C0805K	311-1036-1-ND
C15	10.0 pF	NPO ceramic	C0805K	311-1036-1-ND
C17	100 nF ± 10 %	monolithic ceramic	C0805K	311-1036-1-ND
C18	100 nF ± 10 %	monolithic ceramic	C0805K	311-1036-1-ND
C21	100 nF ± 10 %	monolithic ceramic	C0805K	311-1036-1-ND
C23	100 nF ± 10 %	monolithic ceramic	C0805K	311-1036-1-ND
C24	5 pF to 30 pF trim	NPO ceramic; Murata TZC3P300A 110R00	TRIMCAP	490-1994-2-ND
C25	470 pF	monolithic ceramic	C0805K	
C26	39 pF	monolithic ceramic	C0805K	
CN1		8-pin header	MA08-1	399-8083-10ND
CN2		BU-SMA-H	J502-ND-142-0701-881/886	520-142-0701-881
FL1, FL2		ceramic filter; Murata CFUKF455KB4X or equivalent	surface mount	CFUKF455KB4X-R0
L1	330 nH	Coilcraft 1008CS-331	WE-KI_1008_B	1008CS-331
L2	1.2 μH	fixed inductor Coilcraft 1008CS-122XKLC	WE-KI_1008_B	1008CS-122
L3	220 μH	fixed inductor	WE-GF_L	1812LS-224XJB
R9	100 kΩ ± 1 %	1/4 W metal film	R0603	311-100KCRCT-ND
R10[1]	100 kΩ ± 1 %	1/4 W metal film	C0805K	311-100KCRCT-ND
R11[1]	100 kΩ ± 1 %	1/4 W metal film	C0805K	311-100KCRCT-ND
R17	5.1 kΩ ± 5 %	1/4 W carbon composition	C0805K	311-5.10KCRDKR-ND
U1		SA605DK	TSSOP20	568-2087-5-nd
X1	44.545 MHz	resonant 3rd-overtone crystal	UM-1	49HC/11453

[1] Optional.

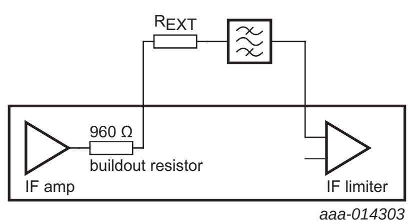
3.2.4.4 Interstage loss

Figure 25 plots the simulated IF RSSI magnitude response for various interstage attenuation. The optimum interstage loss is 12 dB. This has been chosen to allow the use of various types of filters, without upsetting the RSSI linearity. In most cases, the filter insertion loss is less than 12 dB from point A to point B. Therefore, some additional loss must be introduced externally. The easiest and simplest way is to use an external resistor in series with the internal build-out resistor (Pin 14 in the SA604A, Pin 16 in the SA605). Unfortunately, this method mismatches the filter which might be important depending on the design. To achieve the 12 dB insertion loss and good matching to the filter, an L-pad configuration can be used. Figure 26 shows the different setups.

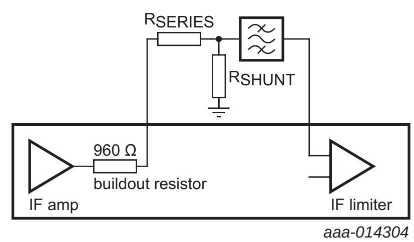


R_L = 91 kΩ

Fig 25. SA604's RSSI curve at different interstage losses



a. Series resistor configuration



b. L-pad configuration

Fig 26. Implementing the 12 dB insertion loss

Equation 1 is an example of how to calculate the resistors values for R_{EXT} in Figure 26a.

$$X_{dB} = 20log\frac{\sqrt{(960 + R_{EXT})R_{FLT}}}{960 + R_{EXT} + R_{FLT}} - FIL\ (dB) \tag{1}$$

where:

X = the insertion loss wanted in dB

R_{EXT} = the external resistor

R_{FLT} = the filter's input impedance

FIL = insertion loss of filter in dB

For the application board:

X = 12 dB

R_{FLT} = 1.5 kΩ

FIL = 3 dB

Therefore, using the Equation 1 gives: R_{EXT} = 5.1 kΩ.

Below are the design equations for calculating R_{SERIES} and R_{SHUNT} in Figure 26b:

$$R_{SERIES} = \left|960 - \frac{R_{FLT}}{2 \times 10^{\left(\frac{-X_{dB}}{20}\right)}}\right| \tag{2}$$

$$R_{SHUNT} = \frac{R_{FLT}}{1 - 2 \times 10^{\left(\frac{-X_{dB}}{20}\right)}} \tag{3}$$

In this case, let's assume FIL = 2 dB, therefore X_{dB} = +10, R_{FLT} = 1.5 kΩ. The results are:

R_{SERIES} = 1.41 kΩ

R_{SHUNT} = 4.08 kΩ

3.2.5 IF noise figure

The IF noise figure of the receiver may be expected to provide at best a 7.7 dB noise figure in a 1.5 kΩ environment from about 25 kHz to 100 MHz. From a 25 Ω source the noise figure can be expected to degrade to about 15.4 dB.

3.3 Demodulator section

Once the signal leaves the IF limiter, it must be demodulated so that the baseband signal can be separated from the IF signal. This is accomplished by the quadrature detector. The detector is made up of a phase comparator (internal to the SA605) and a quadrature tank (external to the SA605).

The phase comparator is a multiplier cell, similar to that of a mixer stage. Instead of mixing two different frequencies, it compares the phases of two signals of the same frequency. Because the phase comparator needs two input signals to extract the information, the IF limiter has a balanced output. One of the outputs is directly connected to the input of the phase comparator. The other signal from the limiter's output (Pin 1) is phase shifted 90° (through external components) and frequency selected by the quadrature tank. This signal is then connected to the other input of the phase comparator (Pin 10 of the SA605). The signal coming out of the quadrature detector (phase detector) is then low-passed filtered to get the baseband signal. A mathematical derivation of this can be seen in the SA604A data sheet (Ref. 2).

The quadrature tank plays an important role in the quality of the baseband signal. It determines the distortion and the audio output amplitude. If the 'Q' is high for the quadrature tank, the audio level will be high, but the distortion will also be high. If the 'Q' is low, the distortion will be low, but the audio level will become low. One can conclude that there is a trade-off.

3.4 Output section

The output section contains an RSSI, audio, and data (unmuted audio) outputs which can be found on Pins 7, 8, and 9, respectively, on the SA605. However, amplitude shift keying (ASK), frequency shift keying (FSK), and a squelch control can be implemented from these pins. Information on ASK and FSK can be found in application note *AN1993, "High sensitivity applications of low-power RF/IF integrated circuits"* (Ref. 3).

Although the squelch control can be implemented by using the RSSI output, it is not a good practice. A better way of implementing squelch control is by comparing the bandpassed audio signal to high frequency colored FM noise signal from the unmuted audio. When no baseband signal is present, the noise coming out of the unmuted audio output will be stronger, due to the nature of FM noise. Therefore, the output of the external comparator will go high (connected to Pin 5 of the SA605), which will mute the audio output. When a baseband signal is present, the bandpassed audio level will dominate and the audio output will now unmute the audio.

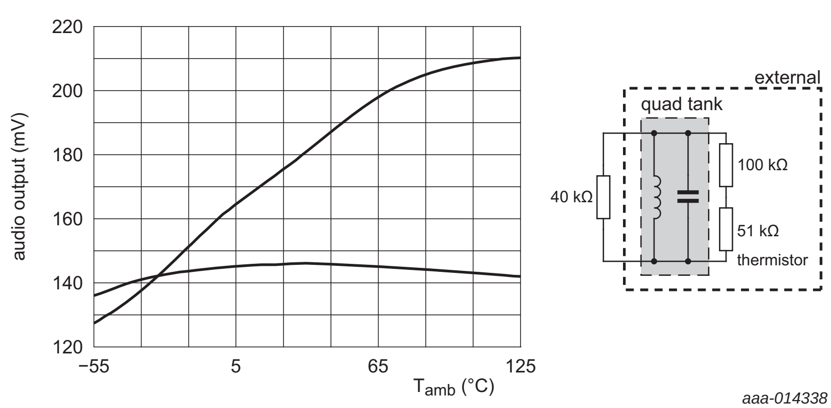
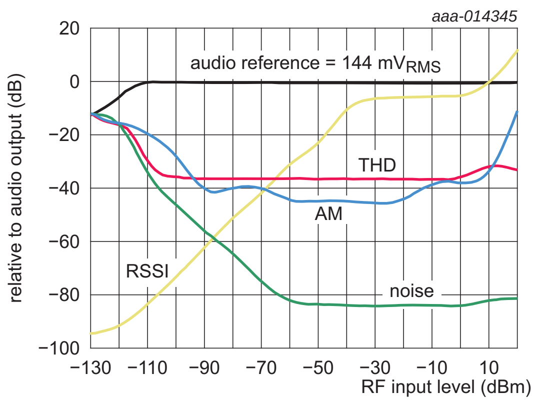
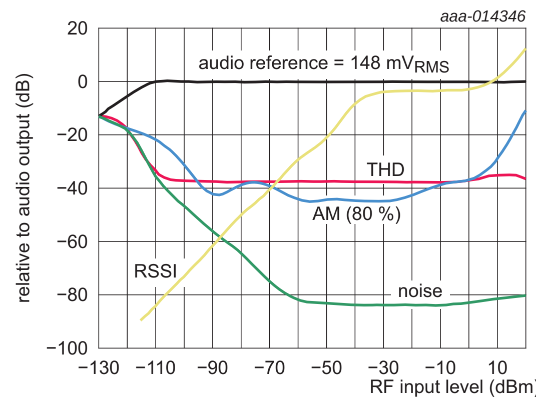


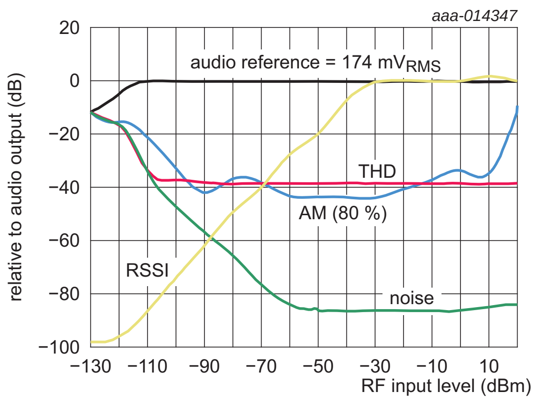
Fig 27. Audio output: compensated versus uncompensated



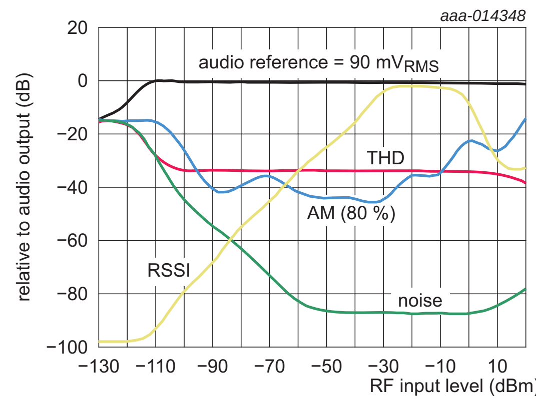
a. SA605 application board at –55 °C



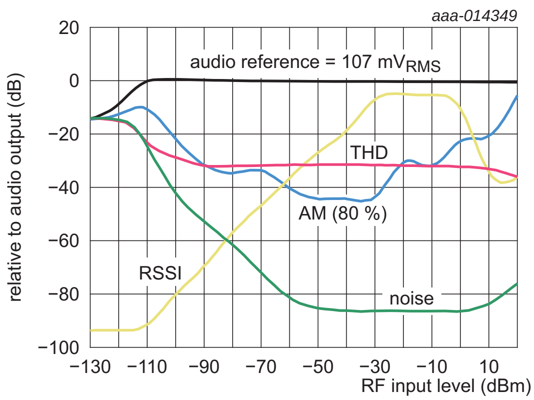
b. SA605 application board at –40 °C



c. SA605 application board at 25 °C



d. SA605 application board at 85 °C



e. SA605 application board at 125 °C

Fig 28. Performance of the SA605 application board at different temperatures (RF = 45 MHz; IF = 455 kHz; V_{CC} = 6 V)

3.4.1 Audio and unmuted audio (data)

The audio and unmuted audio outputs (Pin 8 and 9, respectively, on the SA605) will be discussed in this section because they are basically the same. The only difference between them is that the unmuted audio output is always ‘on’ while the audio output can either be turned ‘on’ or ‘off’. The unmuted audio output (data out) is for signaling tones in systems such as cellular radio. This allows the tones to be processed by the system but remain silent to the user. Since these tones contain information for cellular operation, the unmuted audio output can also be referred to as the ‘data’ output. Grounding Pin 5 on the SA605 mutes the audio on Pin 8 (connecting Pin 5 to V_{CC} unmutes it).

Both of these outputs are PNP current-to-voltage converters with a 55 kΩ nominal internal load. The nominal frequency response of the audio and data outputs are 300 kHz. However, this response can be increased with the addition of an external resistor (<58 kΩ) from the output pins to ground. This will affect the time constant and lower the audio’s output amplitude. This technique can be applied to SCA receivers and data transceivers (as mentioned in the SA604A data sheet, Ref. 2).

3.4.2 RSSI output

RSSI (Received Signal Strength Indicator) determines how well the received signal is being captured by providing a voltage level on its output. The higher the voltage, the stronger the signal.

The RSSI output is a current-to-voltage converter, similar to the audio outputs. However, a 91 kΩ external resistor is needed to get an output characteristic of 0.5 V for every 20 dB change in the input amplitude.

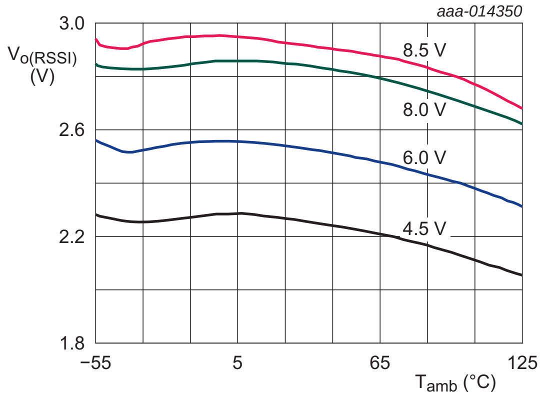
As mentioned earlier, the linearity of the RSSI curve depends on the 12 dB insertion loss between the IF amplifier and IF limiter. The reason the RSSI output is dependent on the IF section is because of the V/I converters. The amount of current in this section is monitored to produce the RSSI output signal. Thus, the IF amplifier’s rectifier is internally calibrated under the assumption that the loss is 12 dB.

Because unfiltered signals at the limiter inputs, spurious products, or regenerated signals will affect the RSSI curve, the RSSI is a good indicator in determining the stability of the board’s layout. With no signal applied to the front end of the SA605, the RSSI voltage level should read 250 mV (RMS value) or less to be a good layout. If the voltage output is higher, then this could indicate oscillations or regeneration in the design.

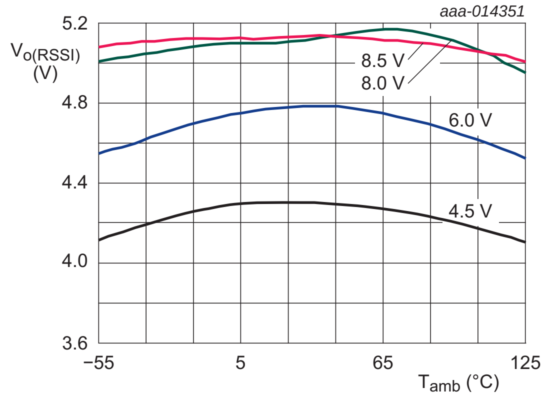
Referring to the SA604A data sheet, there are three primary ways to deal with regeneration:

- Minimize the feedback by gain stage isolation
- Lower the stage input impedances, thus increasing the feedback attenuation factor
- Reduce the gain

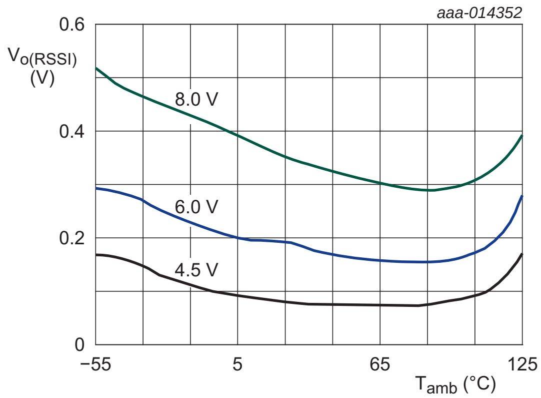
Gain reduction can be accomplished by adding attenuation between stages. More details on regeneration and stability considerations can be found in the SA604A data sheet (Ref. 2).



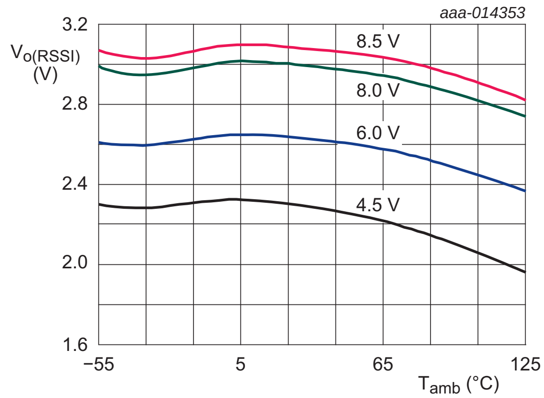
a. SA604A for –68 dBm, RSSI output



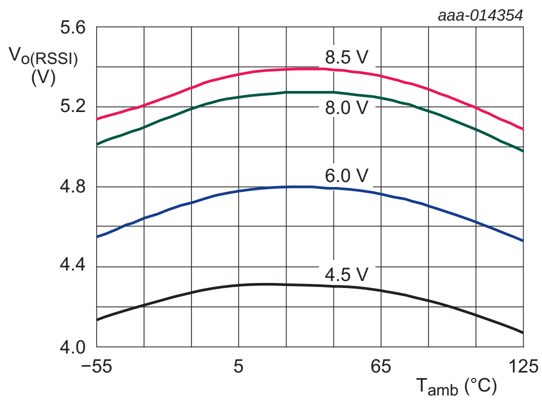
b. SA604A for –18 dBm, RSSI output



c. SA605 RSSI at –120 dBm

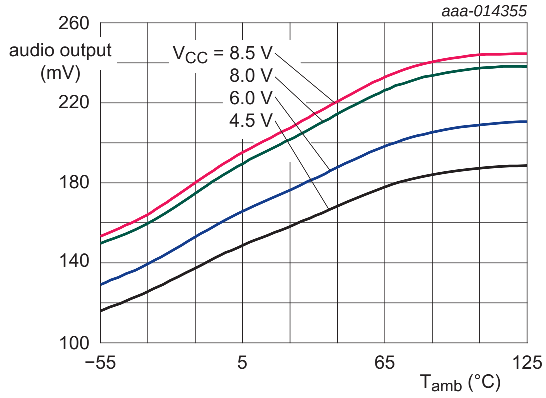


d. SA605 RSSI, –76 dBm



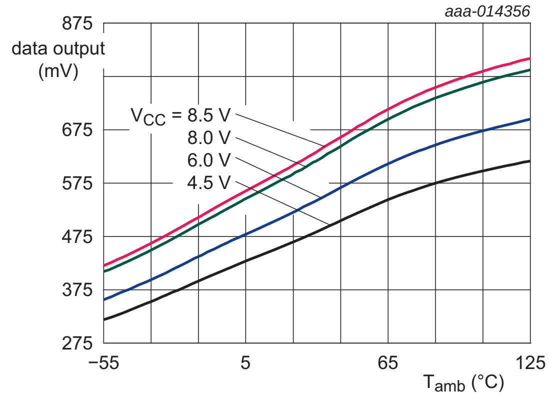
e. SA605 RSSI at –28 dBm

Fig 29. RSSI response for different inputs (50 Ω source; 91 kΩ load; IF = 455 kHz)



RF input = –76 dB; 100 kΩ load

Fig 30. Audio level versus temperature



RF input = –76 dB; 100 kΩ load

Fig 31. Data level versus temperature

4. Questions and Answers

Question: **Bypass.** How important is the effect of the power supply bypass on the receiver performance?

Answer: While careful layout is extremely critical, one of the single most neglected components is the power supply bypass in applications of SA604A or SA605. Although increasing the value of the tantalum capacitor can solve the problem, more careful testing shows that it is actually the capacitor's ESR (Equivalent Series Resistance) that needs to be checked. The simplest way of screening the bypass capacitor is to test the capacitor's dissipation factor at a low frequency (a very easy test, because most of the low frequency capacitance meters display both C, and Dissipation factor).

Question: **On-chip oscillator.** We cannot get the SA605 on-chip oscillator to work. What is the problem?

Answer: The on-board oscillator is just one transistor with a collector that is connected to the supply, an emitter that goes to ground through a 25 kΩ resistor, and a base that goes to the supply through an 18 kΩ resistor. The rest of the circuit is a buffer that follows the oscillator from the transistor base (this buffer does not affect the performance of the oscillator).

Fundamental mode Colpitts crystal oscillators are good up to 30 MHz and can be made by a crystal and two external capacitors. At higher frequencies, up to about 90 MHz, overtone crystal oscillators (Colpitts) can be made like the one in the cellular application circuit. At higher frequencies, up to about 170 MHz, Butler type oscillators (the crystal is in series mode) have been successfully demonstrated. Because of the 8 GHz peak f_T of the transistors, LC Colpitts oscillators have been shown to work up to 900 MHz. The problem encountered above 400 MHz is that the on-chip oscillator level is not sufficient for optimum conversion gain of the mixer. As a result, an external oscillator should be used at those frequencies.

Generally, about 220 mV (RMS value) is the oscillator level needed on Pin 4 for maximum conversion gain of the mixer. An external oscillator driving Pin 4 can be used throughout the band. Finally, since the SA605's oscillator is similar to the SA602, all of the available application notes on SA602 apply to this case (assuming the pin-out differences are taken into account by the user).

Below are a couple of points to help in the oscillator design. The oscillator transistor is biased around 250 μA which makes it very hard to probe the base and emitter without disturbing the oscillator (a high-impedance, low-capacitance active FET probe is desirable).

To solve these problems, an external 22 kΩ resistor (as low as 10 kΩ) can be used from Pin 3 to ground to double the bias current of the oscillator transistor. This external resistor is put there to ensure the start-up of the crystal in the 80 MHz range, and to increase the f_T of the transistor for above 300 MHz to 400 MHz operation. Additionally, this resistor is required for operations above 80 MHz to 90 MHz. When a 1 kΩ resistor from Pin 1 to ground is connected on the SA605, half of the mixer will shut off. This causes the mixer to act like an amplifier. As a result, Pin 20 (the mixer, now amplifier output) can be probed to measure the oscillator frequency. Furthermore, the signal at Pin 20 relates to the true oscillator level. This second resistor is just for optimizing the oscillator, of course. Without the 1 kΩ resistor, the signal at Pin 20 will be a LO feedthrough, which is very small and frequency dependent.

Finally, in some very early data sheets, the base and emitter pins of the oscillator were inadvertently interchanged. The base pin is Pin 4, and the emitter pin is Pin 3. Make sure that your circuit is connected correctly.

Question: **Sensitivity at higher input frequencies.** We cannot get good sensitivity like the 45 MHz case at input frequencies above 70 MHz. Do you have any information on sensitivity versus input frequency?

Answer: The noise figure and the gain of the mixer degrade by less than 0.5 dB, going from 50 MHz to 100 MHz. Therefore, this does not explain the poor degradation in sensitivity. If other problems such as layout, supply bypass, and so on are already accounted for, the source of the problem can be regeneration due to the 70 MHz oscillator. What is probably happening is that the oscillator signal is feeding through the IF, getting mixed with the 455 kHz signal, causing spurious regeneration. The solution is to reduce the overall gain to stop the regeneration.

This gain reduction can be done in a number of places. Two simple points are the attenuator network before the second filter and the LO level (see Figure 26). The second case will increase the mixer's noise figure which is not desirable. Therefore, increasing the interstage loss, despite minimal effect on the RSSI linearity, is the correct solution. As the interstage loss is increased, the regeneration problem is decreased, which improves sensitivity, despite lowering of the overall gain (the lowest RSSI level will keep decreasing as the regeneration problem is decreased). For an 81 MHz circuit it was found that increasing the interstage loss from 12 dB to about 17 dB produced the best results (–119 dBm sensitivity). Of course, adding any more interstage loss will start degrading sensitivity.

Conversely, dealing with the oscillator design, low LO levels could greatly reduce the mixer conversion gain and cause degradation of the sensitivity. For the 81 MHz example, a 22 kΩ parallel resistor from Pin 3 to ground is required for oscillator operation where a Colpitts oscillator like the one in the cellular application circuit is used. The LO level at Pin 4 should be around 220 mV (RMS value) for good operation. Lowering the LO level to approximately 150 mV (RMS value) may be a good way of achieving stability if increasing interstage attenuation is not acceptable. In that case, the 22 kΩ resistor can be made a thermistor to adjust the LO level versus temperature for maintaining sensitivity and ensuring crystal start-up versus temperature. At higher IF frequencies (above 30 MHz), the interstage gain reduction is not needed. The bandwidth of the IF section will lower the overall gain. So, the possibility of regeneration decreases.

Question: **Mixer noise figure.** How do you measure the mixer noise figure in SA605, and SA602?

Answer: We use the test circuit shown in the SA602 data sheet. The noise figure tester is the HP8970A. The noise source we use is the HP346B (ENR = 15.46 dB). Note that the output is tuned for 10.7 MHz. From that test circuit the NF-meter measures a gain of approximately 15 dB and 5.5 dB noise figure.

More noise figure data is available in the paper titled *“Gilbert-type Mixers vs. Diode Mixers”* presented at RF Expo ’89 in Santa Clara, California.

Question: What is the value of the series resistor before the IF filter in the SA605 or SA604A applications?

Answer: A value of 5.1 kΩ has been used by us in our demo board. This results in a maximally straight RSSI curve. A lower value of about 1 kΩ will match the filter better. A better solution is to use an L-pad, as discussed earlier in this application note.

Question: What is the low frequency input resistance of the SA605?

Answer: The data sheets indicated a worst-case absolute minimum of 1.5 kΩ. The typical value is 4.7 kΩ.

Question: What are BE-BC capacitors in the SA605 oscillator transistor?

Answer: The oscillator is a transistor with the collector connected to the supply and the emitter connected to the ground through a 25 kΩ resistor. The base goes to the supply through an 18 kΩ resistor. The junction capacitors are roughly about 24 fF (fempto Farads) for CJE (Base-emitter capacitors), and 44 fF for CJC (Collector-base capacitors). There is a 72 fF capacitor for CJS (Collector-substrate capacitor). This is all on the chip itself. It should be apparent that the parasitic packaging capacitors (1.5 pF to 2.5 pF) are the dominant values in the oscillator design.

Question: What are the differences between the SA604 and the SA604A? (See Table 4.)

Answer: The SA604A is an improved version of the SA604. The SA604 has been discontinued. Customers, who have been using the SA604 in the past, should have no trouble doing the conversion.

The main differences are that the small signal IF bandwidth is 25 MHz instead of 15 MHz, and the RSSI is internally temperature compensated. If external temperature compensation was used for the SA604, the designer can now cut cost with the SA604A. The designer can either get rid of these extra parts completely or replace the thermistor (if used in original temperature compensated design) with a fixed resistor.

Those using the SA604 at 455 kHz should not see any change in performance. For 10.7 MHz, a couple of dB improvement in performance will be observed. However, there may be a few cases where instability will occur after using SA604A. This will be the case if the PC-board design was marginal for the SA604 in the first place. This problem, however, can be cured by using a larger than 10 μF tantalum bypass capacitor on the supply line, and screening the capacitors for their ESR (equivalent series resistance) as mentioned earlier. The ESR at 455 kHz should be less than 0.2 Ω. Since ESR is a frequency-dependent value, the designer can correlate good performance with a low frequency dissipation factor, or ESR measurement, and screen the tantalum capacitors in production. There are some minor differences as well. The SA604A uses about 1 mA more current than the SA604. An emitter-follower has been added at the limiter output to present a lower and more stable output impedance at Pin 9. The DC voltage at the audio and data outputs is approximately 3 V instead of 2 V in the SA604, but that should not cause any problems. The recovered audio level, on the other hand, is slightly higher in the SA604A which should actually be desirable. Because of these changes, it is now possible to design 21.4 MHz IFs using the SA604A, which was not possible with the SA604.

The two chips are identical, otherwise. The customers are encouraged to switch to the SA604A because it is a more advanced bipolar process than the previous generation used in the SA604. As a result, we get much tighter specifications on the SA604A.

Table 4. Summary of differences for SA604/SA604A

Parameter	SA604	SA604A
RSSI	no temperature compensation	internally temperature-compensated
IF bandwidth	15 MHz	25 MHz
IF limiter output	no buffer	emitter-follower buffer output with 8 k in the emitter
Current drain	2.7 mA	3.7 mA

Question: How does the SA605 mixer compare with a typical double-balanced diode mixer?

Answer: Some data on the comparison of the conversion gain and LO power requirements are shown in this application note. These two parameters reveal the advantages in using the SA605 mixer.

The only drawback of the SA605 may seem to be its lower third-order intercept point in comparison to a diode mixer. But, this is inherent in the SA605 as a result of the low power consumption. If one compares the conversion gain of the SA605 with the conversion loss of a low cost diode mixer, it turns out that the third-order intercept point, referred to the output, is the same or better in the SA605. Another point to take into account is that a diode mixer cannot be used in the front end of a receiver without a preamp due to its poor noise figure. A third-order intercept analysis shows that the intercept point of the combination of the diode mixer and preamp will be degraded at least by the gain of the preamp. A preamp may not be needed with SA605 because of its superior noise figure.

For more detailed discussion of this topic please refer to the paper titled *"Gilbert-type Mixers vs. Diode Mixers"*.

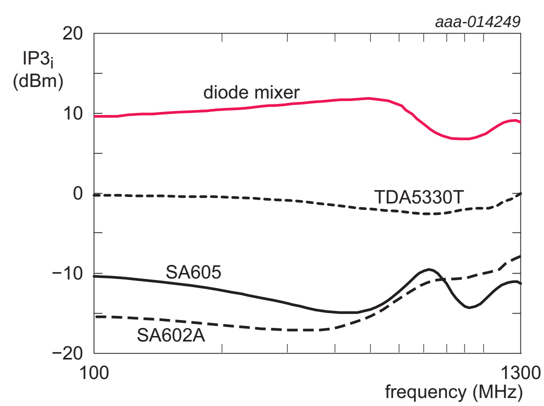


Fig 32. Comparing different types of mixers

Question: How can we use the SA605 for SCA FM reception?

Answer: The 10.7 MHz application circuit described in application note *AN1993* (Ref. 3) can be used in this case. The LO frequency should be changed and the RF front-end should be tuned to the FM broadcast range. The normal FM signal, coming out of Pin 8 of the SA605, could be expected to have about 1.5 μV (into 50 Ω) sensitivity for 20 dB S/N. This signal should be band-pass filtered and amplified to recover the SCA sub-carrier. The output of that should then go to a PLL SCA decoder, shown on the data sheet of Philips Semiconductors SA565 phase lock loop, to demodulate the base-band audio. The two outputs of the SA605 Pin 8 and Pin 9 can be used to receive SCA data as well as voice, or features such as simultaneous reception of both normal FM, and SCA. The RSSI output, with its 90 dB dynamic range, is useful for monitoring signal levels.

Question: What is the power consumption of the SA605 or SA604A versus temperature and V_{CC}?

Answer: The SA605 consumes about 5.6 mA of current at 6 V. This level is slightly temperature and voltage dependent as shown in Figure 33. Similar data for the SA604A is shown in Figure 34.

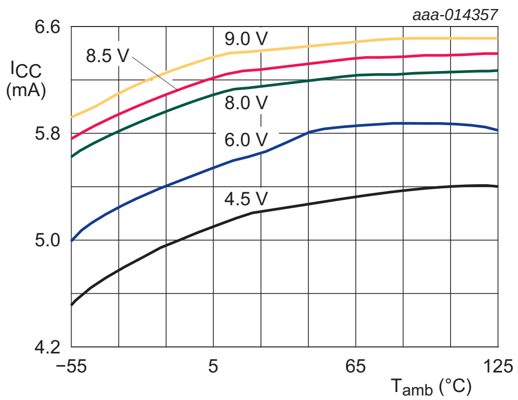


Fig 33. SA605 supply current versus temperature

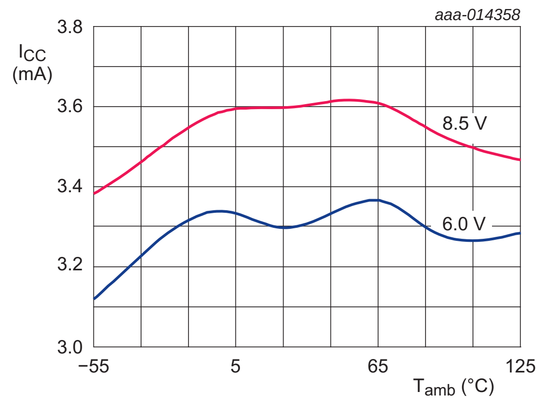


Fig 34. SA604A supply current versus temperature

Question: How can you minimize RF and LO feedthroughs?

Answer: The RF and LO feedthroughs are due to offset voltages at the input of the mixer's differential amplifiers and the imbalance of the parasitic capacitors. A circuit, such as the one shown in Figure 35, can be used to adjust the balance of the differential amplifiers. The circuit connected to Pin 1 and Pin 2 will minimize RF feedthrough, while the circuit shown connected to Pin 6 will adjust the LO feedthrough. The only limitation is that if the RF and LO frequencies are in the 100 MHz range or higher, these circuits will probably be effective for a narrow frequency range.

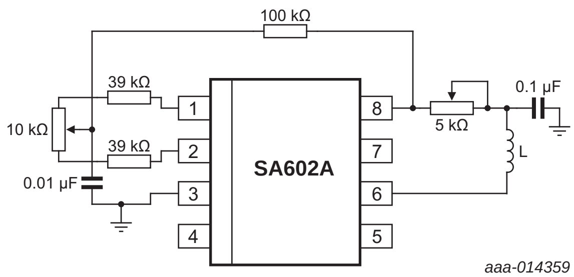


Fig 35. Minimizing RF and LO feedthrough

Question: **Distortion versus RF input level.** We get a good undistorted demodulated signal at low RF levels, but severe distortion at high RF levels. What is happening?

Answer: This problem usually occurs at 10.7 MHz or at higher IF. The IF filters have not been properly matched on both sides causing a sloping IF response. The resulting distortion can be minimized by adjusting the quad tank at the FM threshold where the IF is out of limiting. As the RF input increases, the IF stages will limit and make the IF response flat again. At this point, the effect of the bad setting of the quad tank will show itself as distortion. The solution is to always tune the quad tank for distortion at a medium RF level, to make sure that the IF is fully limited. Then, to avoid excessive distortion for low RF levels, one should make sure that the IF filters are properly matched.

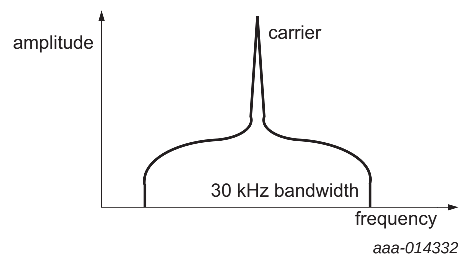
Question: **The most commonly asked questions:** “Why doesn’t the receiver sensitivity meet the specifications?”; “Why is the RSSI dynamic range much less than expected?”; “Why does the RSSI curve dip at 0.9 V and stay flat at 1 V as the RF input decreases?”; “Why does the audio output suddenly burst into oscillation, or output wideband noise as the RF input goes down, instead of dying down slowly?”; “When looking at the IF output with a spectrum analyzer, why do high amplitude spurs become visible near the edge of the IF band as the RF level drops?”

Answer: These are the most widely observed problems with the SA605. They are all symptoms of the same problem; instability. The instability is due to bad layout and grounding.

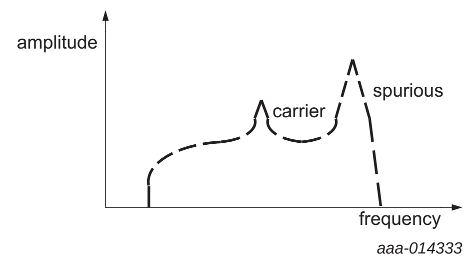
Regenerative instability occurs when the limiter’s output signals are radiated and picked up by the high impedance inputs of the mixer and IF amp. This signal is amplified by both the IF amp and limiter. Positive feedback causes the signal to grow until the signal at the limiter’s output becomes limited. Due to the nature of FM, this instability will dominate any low RF input levels and capture the receiver (see Figure 36).

Since the receiver behaves normally for high RF inputs, it misleads the designer into believing that the design is okay. Additionally the RSSI circuit cannot determine whether the signal being received is coming from the antenna or the result of regenerative instability. Therefore, RSSI will be a good instability indicator in this instance because the RSSI will stay at a high level when the received signal decreases. Looking at the IF spectrum (Pin 11 for SA605, Pin 9 for SA604A) with the RF carrier present (no modulation), the user will see a shape as shown in Figure 36. When regenerative instability occurs, the receiver does not seem to have the ultimate sensitivity of which it is capable.

Make sure that a double-sided layout with a good ground plane on both sides is used. This will have RF/IF loops on both sides of the board. Follow our layouts as faithfully as you can. The supply bypass should have a low ESR 10 μF to 15 μF tantalum capacitor as discussed earlier. The crystal package, the inductors, and the quad tank shields should be grounded. The RSSI output should be used as a progress monitor even if is not needed as an output. The lowest RSSI level should decrease as the circuit is made more stable. The overall gain should be reduced by lowering the input impedance of the IF amplifier and IF limiter, and adding attenuation after the IF amplifier, and before the second filter. A circuit that shows an RSSI of 250 mV or less with no RF input should be considered close to the limit of the performance of the device. If the RSSI still remains above 250 mV, the recommendations mentioned above should be revisited.



a. Carrier and in-band noise for high RF levels



b. For low RF levels, noise and spurious signals capture the receiver

Fig 36. Amplitude versus frequency

Question: Without the de-emphasis network at the audio output, the –3 dB bandwidth of the audio output is limited to only 4.5 kHz. The maximum frequency deviation is 8 kHz, and the IF bandwidth is 25 kHz. What is the problem?

Answer: What is limiting the audio bandwidth in this case is not the output circuit, but the IF filters. Remember that Carson's rule for FM IF bandwidth requires the IF bandwidth to be at least:

$$2(\text{maximum frequency deviation} + \text{audio frequency})$$

With a 25 kHz IF bandwidth and 8 kHz frequency deviation, the maximum frequency that can pass without distortion is approximately 4.5 kHz. 2(8 kHz + 4.5 kHz) is 25 kHz as expected.

Question: What are the equivalent RF input impedances for the SA606, given the equivalent circuit model of Figure 4?

Answer: The SA606 input impedance versus frequency is given in Table 5.

The input matching technique discussed in this application note is applicable for the SA606.

Table 5. Input impedance versus frequency

Frequency	R || C
10 MHz	7.5 kΩ || 2.4 pF
50 MHz	7.5 kΩ || 2.7 pF
100 MHz	6.25 kΩ || 2.7 pF
150 MHz	5.5 kΩ || 2.7 pF
250 MHz	5.0 kΩ || 2.8 pF

5. Abbreviations

Table 6. Abbreviations

Acronym	Description
AM	Amplitude Modulation
ASK	Amplitude Shift Keying
ENR	Excess Noise Ratio
ESR	Equivalent Series Resistance
FET	Field-Effect Transistor
FM	Frequency Modulation
FSK	Frequency Shift Keying
IF	Intermediate Frequency
LC	inductor-capacitor network
LO	Local Oscillator
NPN	bipolar transistor with N-type emitter and collector and a P-type base
PLL	Phase-Locked Loop
PM	Phase Modulation
PNP	bipolar transistor with P-type emitter and collector and a N-type base
RF	Radio Frequency
RMS	Root Mean Squared
RSSI	Received Signal Strength Indicator
SCA	Subsidiary Communications Authorization
THD	Total Harmonic Distortion

6. References

[1] **SA605, "High performance low power mixer FM IF system" —** Product data sheet; NXP Semiconductors; www.nxp.com/documents/data_sheet/SA605.pdf

[2] **SA604A, "High-performance low-power FM IF system" —** Product data sheet; NXP Semiconductors; www.nxp.com/documents/data_sheet/SA604A.pdf

[3] **AN1993, "High sensitivity applications of low-power RF/IF integrated circuits" —** Application note; NXP Semiconductors; www.nxp.com/documents/application_note/AN1993.pdf

[4] *"AN199, Designing with the SA604"* — Philips Semiconductors Linear Data Manual, 1987.

[5] *"AN1981, New Low Power Single Sideband Circuits"* — Philips Semiconductors Linear Data Manual, 1988.

[6] *"Applying the Oscillator of the SA602 in Low Power Mixer Applications"* — Philips Semiconductors Linear Data Manual, 1988.

[7] *"RF Circuit Design"* — Bowick. C., Indiana: Howard W. Sams & Company, 1982.

[8] *"The ARRL Handbook for the Radio Amateur"* — American Radio Relay League, 1986.

[9] *"Communications Receivers: Principles & Design"* — Rohde, U., Bucher, T.T.N., McGraw Hill, 1988.

[10] *"Gilbert-type Mixers vs. Diode Mixers"* — proceedings of R.F. Expo 1989, Fotowat, A., Murthi, E., pp. 409-413.

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8. Contents

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