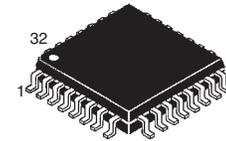


**Legacy Device:** Motorola MC13158

The ML13158 is a wideband IF subsystem that is designed for high performance data and analog applications. The ML13158 has an on-board grounded collector VCO transistor that may be used with a fundamental or overtone crystal in single channel operation or with a PLL in multi-channel operation. The mixer is useful to 500 MHz and may be used in a balanced differential or a single ended configuration. The IF amplifier is split to accommodate two low cost cascaded filters. RSSI output is derived by summing the output of both IF sections. A precision data shaper has an Off function to shut the output off to save current. An enable control is provided to power down the IC for power management in battery operated applications.

Applications include DECT, wideband wireless data links for personal and portable laptop computers and other battery operated radio systems which utilize GFSK, FSK or FM modulation.

- Designed for DECT Applications
- 1.8 to 6.0 Vdc Operating Voltage
- Low Power Consumption in Active and Standby Mode
- Greater than 600 kHz Detector Bandwidth
- Data Slicer with Special Off Function
- Enable Function for Power Down of Battery Operated Systems
- RSSI Dynamic Range of 80 dB Minimum
- Low External Component Count
- Operating Temperature Range –  $T_A = -40$  to  $+85^\circ\text{C}$

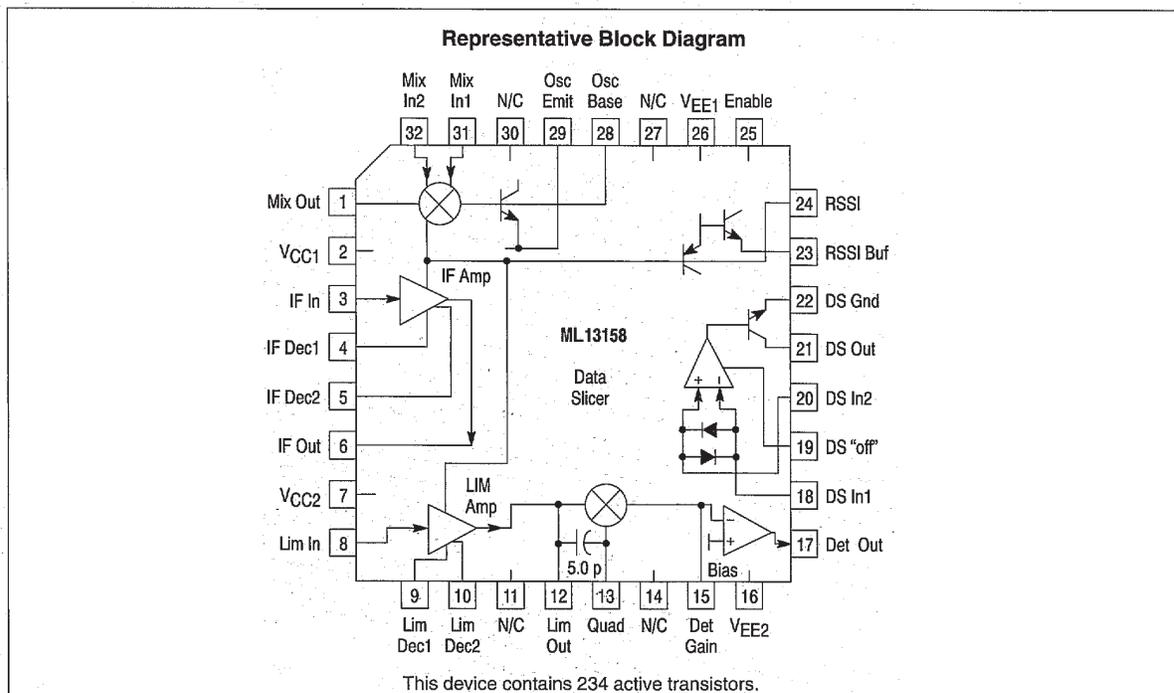


**QFP 32 = 8P**  
PLASTIC QFP PACKAGE  
CASE 873

**CROSS REFERENCE/ORDERING INFORMATION**

PACKAGE	MOTOROLA	LANSDALE
QFP 32	MC13158FTB	ML13158-8P

**Note:** Lansdale lead free (Pb) product, as it becomes available, will be identified by a part number prefix change from **ML** to **MLE**.



**MAXIMUM RATINGS**

Rating	Pin	Symbol	Value	Unit
Power Supply Voltage	16, 26	$V_{S(max)}$	6.5	Vdc
Junction Temperature		$T_{JMAX}$	+150	°C
Storage Temperature Range		$T_{stg}$	-65 to +150	°C

**NOTE:** 1. Devices should not be operated at or outside these values. The "Recommended Operating Limits" provide for actual device operation.

**RECOMMENDED OPERATING CONDITIONS** ( $V_{CC} = V_2 = V_7$ ;  $V_{EE} = V_{16} = V_{22} = V_{26}$ ;  $V_S = V_{CC} - V_{EE}$ )

Rating	Pin	Symbol	Value	Unit
Power Supply Voltage $T_A = 25^\circ\text{C}$ $-40^\circ\text{C} \leq T_A \leq 85^\circ\text{C}$	2, 7 16, 26	$V_S$	2.0 to 6.0	Vdc
Input Frequency	31, 32	$F_{in}$	10 to 500	MHz
Ambient Temperature Range		$T_A$	-40 to +85	°C
Input Signal Level	31, 32	$V_{in}$	200	mVrms

**DC ELECTRICAL CHARACTERISTICS** ( $T_A = 25^\circ\text{C}$ ;  $V_S = 3.0$  Vdc; No Input Signal; See Figure 1.)

Characteristic	Condition	Pin	Symbol	Min	Typ	Max	Unit
Total Drain Current	$V_S = 2.0$ Vdc $V_S = 3.0$ Vdc $V_S = 6.0$ Vdc See Figure 2	16, 26	$I_{TOTAL}$	2.5 3.5 3.5	5.5 5.7 6.0	8.5 8.5 9.5	mA

**DATA SLICER** (Input Voltage Referenced to  $V_{EE}$ ;  $V_S = 3.0$  Vdc; No Input Signal)

Output Current; $V_{18}$ LO; Data Slicer Enabled (DS "on")	$V_{19} = V_{EE}$ $V_{18} < V_{20}$ $V_{20} = V_S/2$ See Figure 3	21	$I_{21}$	2.0	5.9	-	mA
Output Current; $V_{18}$ HI; Data Slicer Enabled (DS "on")	$V_{19} = V_{EE}$ $V_{18} > V_{20}$ $V_{20} = V_S/2$ See Figure 4	21	$I_{21}$	-	0.1	1.0	$\mu\text{A}$
Output Current; Data Slicer Disabled (DS "off")	$V_{19} = V_{CC}$ $V_{20} = V_S/2$	21	$I_{21}$	-	0.1	1.0	$\mu\text{A}$

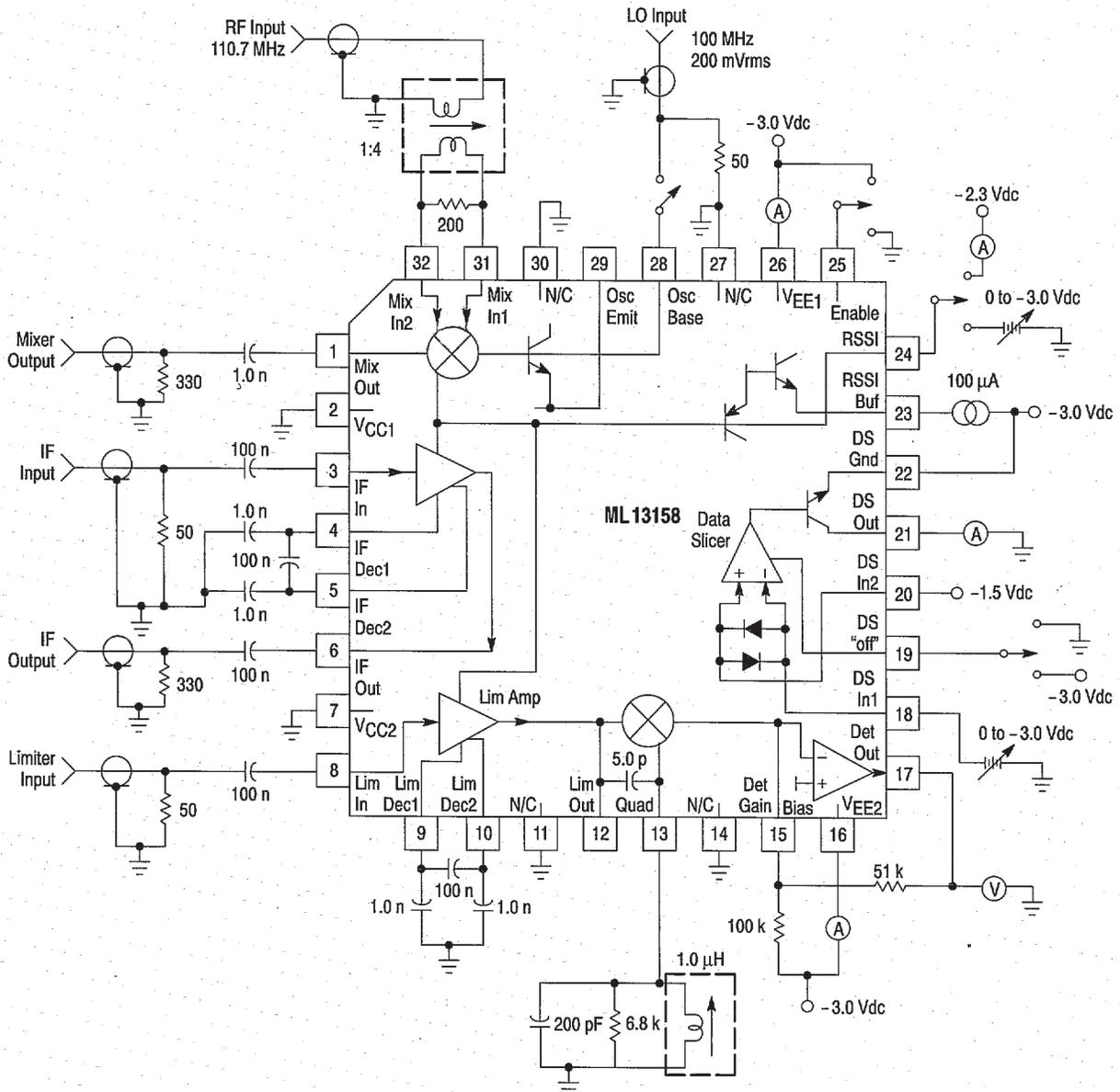
**AC ELECTRICAL CHARACTERISTICS** ( $T_A = 25^\circ\text{C}$ ;  $V_S = 3.0$  Vdc;  $f_{RF} = 110.7$  MHz;  $f_{LO} = 100$  MHz; See Figure 1 unless otherwise specified.)

Characteristic	Condition	Pin	Symbol	Min	Typ	Max	Unit
<b>MIXER</b>							
Mixer Conversion Gain	$V_{in} = 1.0$ mVrms See Figure 5	31, 32, 1		-	22	-	dB
Noise Figure	Input Matched	31, 32, 1	NF	-	14	-	dB
Mixer Input Impedance	Single-Ended See Figure 15	31, 32	$R_p$ $C_p$	- -	865 1.6	- -	$\Omega$ pF
Mixer Output Impedance		1		-	330	-	$\Omega$

**AC ELECTRICAL CHARACTERISTICS (continued)** ( $T_A = 25^\circ\text{C}$ ;  $V_S = 3.0\text{ Vdc}$ ;  $f_{RF} = 110.7\text{ MHz}$ ;  $f_{LO} = 100\text{ MHz}$ ; See Figure 1 unless otherwise specified.)

Characteristic	Condition	Pin	Symbol	Min	Typ	Max	Unit
<b>MIXER</b>							
<b>IF AMPLIFIER SECTION</b>							
IF RSSI Slope	See Figure 8	23		0.15	0.3	0.4	$\mu\text{A/dB}$
IF Gain	$f = 10.7\text{ MHz}$ See Figure 7	3, 6		-	36	-	dB
Input Impedance		3		-	330	-	$\Omega$
Output Impedance		6		-	330	-	$\Omega$
<b>LIMITING AMPLIFIER SECTION</b>							
Limiter RSSI Slope	See Figure 9	23		0.15	0.3	0.4	$\mu\text{A/dB}$
Limiter Gain	$f = 10.7\text{ MHz}$	8, 12		-	70	-	dB
Input Impedance		8		-	330	-	$\Omega$

Figure 1. Test Circuit



Typical Performance Over Temperature

(per Figure 1)

Figure 2. Total Supply Current versus Ambient Temperature, Supply Voltage

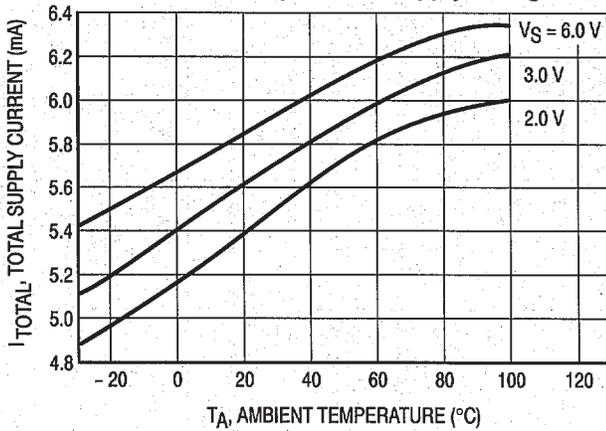


Figure 3. Data Slicer On Output Current versus Ambient Temperature

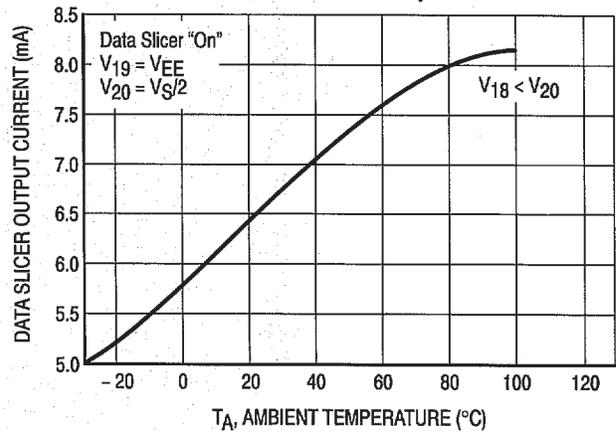


Figure 4. Data Slicer On Output Current versus Ambient Temperature

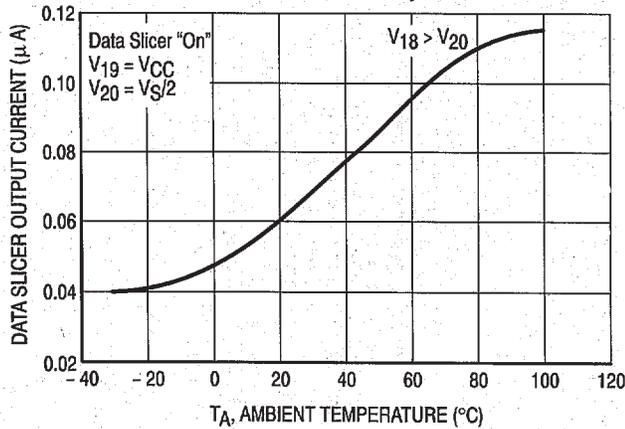


Figure 5. Normalized Mixer Gain versus Ambient Temperature

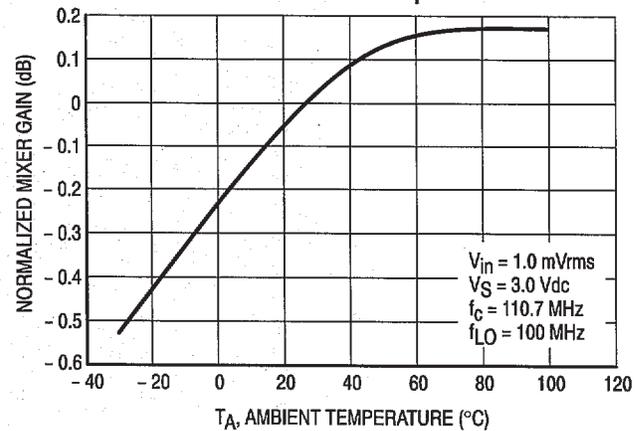


Figure 6. Mixer RSSI Output Current versus Ambient Temperature, Mixer Input Level

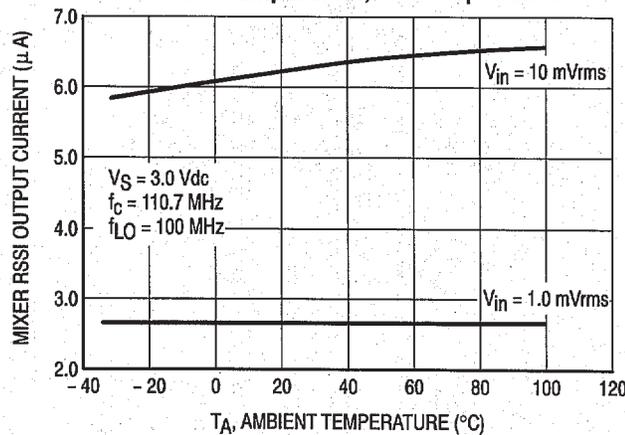
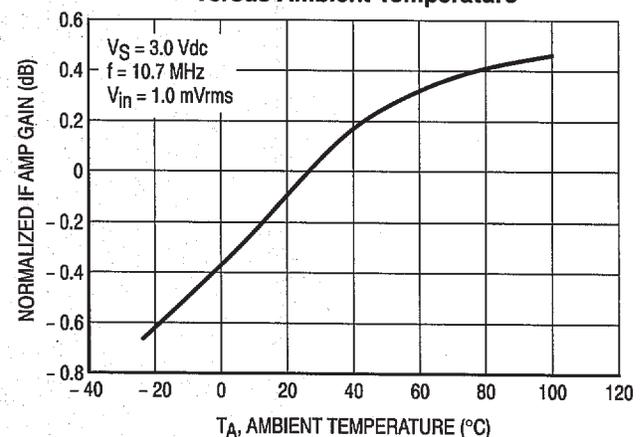


Figure 7. Normalized IF Amp Gain versus Ambient Temperature



Typical Performance Over Temperature

(per Figure 1)

Figure 8. IF Amp RSSI Output Current versus Ambient Temperature, IF Input Level

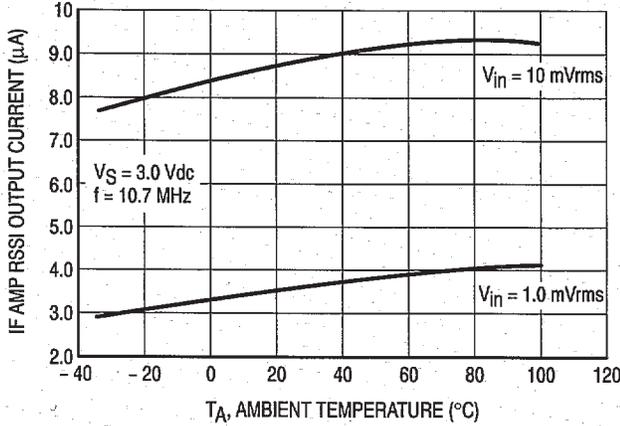


Figure 9. Limiter Amp RSSI Output Current versus Ambient Temperature, Input Signal Level

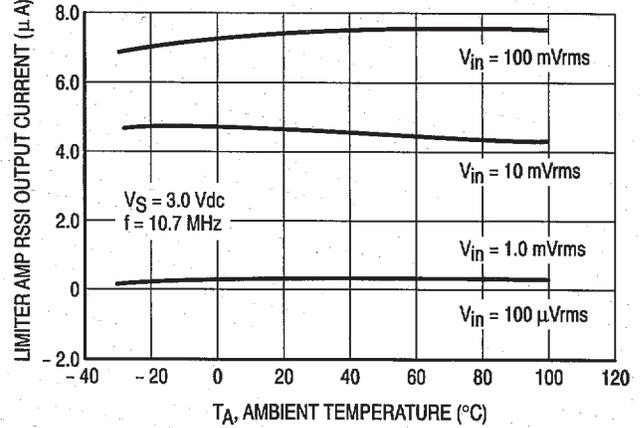


Figure 10. Total RSSI Output Current versus Ambient Temperature (No Signal)

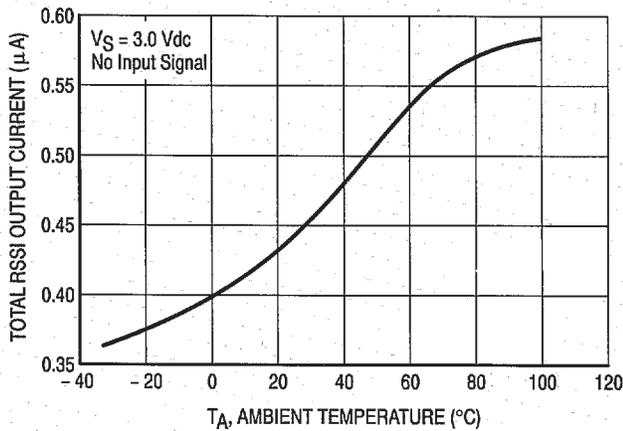
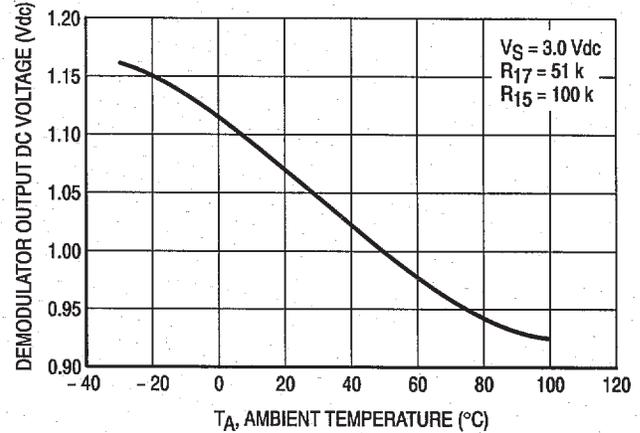


Figure 11. Demodulator DC Voltage versus Ambient Temperature



SYSTEM LEVEL AC ELECTRICAL CHARACTERISTICS ( $T_A = 25^\circ\text{C}$ ;  $V_S = 3.0 \text{ Vdc}$ ;  $f_{RF} = 112 \text{ MHz}$ ;  $f_{LO} = 122.7 \text{ MHz}$ )

Characteristic	Condition	Notes	Symbol	Typ	Unit
12 dB SINAD Sensitivity: Narrowband Application	$f_{RF} = 112 \text{ MHz}$ $f_{mod} = 1.0 \text{ kHz}$ $f_{dev} = \pm 125 \text{ kHz}$ SINAD Curve	1			dBm
Without Preamp	Figure 25			-101	
With Preamp	Figure 26			-113	
Third Order Intercept Point	$f_{RF1} = 112 \text{ MHz}$ $f_{RF2} = 112.1 \text{ MHz}$ $V_S = 3.5 \text{ Vdc}$ Figure 28	2	IIP3	-32	dBm
1.0 dB Comp. Point			1.0 dB C.Pt.	-39	

NOTES: 1. Test Circuit & Test Set per Figure 24.  
2. Test Circuit & Test Set per Figure 27.

## CIRCUIT DESCRIPTION

### GENERAL

The ML13158 is a low power single conversion wideband FM receiver incorporating a split IF. This device can be used as the backend in digital FM systems such as Digital European Cordless Telephone (DECT) and wide band data links with data rates up to 2.0 Mbps. It contains a mixer, oscillator, Received Signal Strength Indicator (RSSI), IF amplifier, limiting IF, quadrature detector, power down or enable function, and a data slicer with output off function. Further details are covered in the Pin Function Description which shows the equivalent internal circuit and external circuit requirements.

### CURRENT REGULATION

Temperature compensating voltage independent current regulators which are controlled by the enable pin (Pin 25) where "low" powers up and "high" powers down the entire circuit.

### MIXER

The mixer is a double-balanced four quadrant multiplier and is designed to work up to 500 MHz. It can be used in differential or in single-ended mode by connecting the other input to the positive supply rail. The linear gain of the mixer is approximately 22 dB at 100 mVrms LO drive level. The mixer gain and noise figure have been emphasized at the expense of intermodulation performance. RSSI measurements are added in the mixer to extend the range to higher signal levels. The single-ended parallel equivalent input impedance of the mixer is  $R_p \sim 1.0 \text{ k}\Omega$  and  $C_p \sim 2.0 \text{ pF}$ . The buffered output of the mixer is internally loaded resulting in an output impedance of 330  $\Omega$ .

### LOCAL OSCILLATOR

The on-chip transistor operates with crystal and LC resonant elements up to 220 MHz. Series resonant, overtone crystals are used to achieve excellent local oscillator stability. Third overtone crystals are used through about 65 to 70 MHz. Operation from 70 MHz up to 180 MHz is feasible using the on-chip transistor with a 5th or 7th overtone crystal. To enhance operation using an overtone crystal, the internal transistor bias is increased by adding an external resistor from Pin 29 to  $V_{EE}$ , however, with an external resistor, the oscillator stays on during power down. Typically, -10 dBm of local oscillator drive is needed to adequately drive the mixer. With an external oscillator source, the IC can be operated up to 500 MHz.

### RSSI

The Received Signal Strength Indicator (RSSI) output is a current proportional to the log of the received signal amplitude. The RSSI current output is derived by summing the currents from the mixer, IF and limiting amplifier stages. An increase in RSSI dynamic range, particularly at higher input signal levels is achieved. The RSSI circuit is designed to provide typically 85 dB of dynamic range with temperature compensation.

Linearity of the RSSI is optimized by using external ceramic bandpass filters which have an insertion loss of 4.0 dB and 330  $\Omega$  source and load impedance. For higher data rates used in DECT and related applications, LC bandpass filtering is necessary to acquire the desired bandpass response; however, the RSSI linearity will require the same insertion loss.

### RSSI BUFFER

The RSSI output current creates a voltage across an external resistor. A unity voltage-gain amplifier is used to buffer this voltage.

The output of this buffer has an active pull-up but no pull-down, so it can also be used as a peak detector. The negative slew rate is determined by external capacitance and resistance to the negative supply.

### IF AMPLIFIER

The first IF amplifier section is composed of three differential stages with the second and third stages contributing to the RSSI. This section has internal DC feedback and external input decoupling for improved symmetry and stability. The total gain of the IF amplifier block is approximately 40 dB at 10.7 MHz.

The fixed internal input impedance is 330  $\Omega$ . When using ceramic filters requiring source and loss impedances of 330  $\Omega$ , no external matching is necessary. Overall RSSI linearity is dependent on having total midband attenuation of 10 dB (4.0 dB insertion loss plus 6.0 dB impedance matching loss) for the filter. The output of the IF amplifier is buffered and the impedance is 330  $\Omega$ .

### LIMITER

The limiter section is similar to the IF amplifier section except that five differential stages are used. The fixed internal input impedance is 330  $\Omega$ . The total gain of the limiting amplifier section is approximately 70 dB. This IF limiting amplifier section internally drives the quadrature detector section and it is also brought out on Pin 12.

### QUADRATURE DETECTOR

The quadrature detector is a doubly balanced four quadrant multiplier with an internal 5.0 pF quadrature capacitor between Pins 12 and 13. An external capacitor may be added between these pins to increase the IF signal to the external parallel RLC resonant circuit that provides the 90 degree phase shift and drives the quadrature detector. A single pin (Pin 13) provides for the external LC parallel resonant network and the internal connection to the quadrature detector.

Internal low pass filter capacitors have been selected to control the bandwidth of the detector. The recovered signal is brought out by the inverting amplifier buffer. An external feedback resistor from the output (Pin 17) to the output amplitude; it is combined with another external resistor from the input to the negative supply (Pin 15) controls the output amplitude; it is combined with another external resistor from the input to the negative supply (Pin 16) to set the output DC level. For a resistor ratio of 1, the DC level at the detector output is  $2.0 V_{BE}$  (see Figure 12). A small capacitor  $C_{17}$  across the first resistor (from Pin 17 to 15) can be used to reduce the bandwidth.

### DATA SLICER

The data slicer is a comparator that is designed to square up the data signal. Across the data slicer inputs (Pins 18 and 20) are back to back diodes.

The recovered data signal from the quadrature detector can be DC coupled to the data slicer DS IN1 (Pin 18). In the application circuit shown in Figure 1 it will be centered at  $2.0 V_{BE}$  and allowed to swing  $\pm V_{BE}$ . A capacitor is placed from DS IN2 (Pin 20) to  $V_{EE}$ . The size of this capacitor and the nature of the data slicer shapes up the recovered signal. The time constant is short for large peak to peak voltage swings or when there is a change in DC level at the detector output. For small signal or for continuous bits of the same polarity which drift close to the threshold voltage, the time

constant is longer.

A unique feature of the data slicer is that inverting switching stages in the comparator are supplied through the emitter pin of the output transistor (Pin 22 – DS Gnd) to V<sub>EE</sub> rather than internally to V<sub>EE</sub>. This is provided in order to reduce switching feedback to the front end. A control pin is provided to shut the data slicer output off (DS “off” – Pin 19). With DS “off” pin at V<sub>CC</sub> the data slicer output is

shut off by shutting down the base drive to the output transistor. When a channel is being monitored to make an RSSI measurement, but not to collect data, the data output may be shut off to save current.

**PIN FUNCTION DESCRIPTION**

Pin	Symbol	Internal Equivalent Circuit	Description/External Circuit Requirements	
1	Mix Out		<p><b>Mixer Output</b> The mixer output impedance is 330 Ω; it matches to 10.7 MHz ceramic filters with 330 Ω input impedance.</p> <p><b>Supply Voltage (V<sub>CC1</sub>)</b> This pin is the V<sub>CC</sub> pin for the Mixer, Local Oscillator, and IF Amplifier. The operating supply voltage range is from 1.8 Vdc to 5.0 Vdc. In the PCB layout, the V<sub>CC</sub> trace must be kept as wide as possible to minimize inductive reactances along the trace; it is best to have it completely fill around the surface mount components and traces on the circuit side of the PCB.</p>	
2	V <sub>CC1</sub>			
3	IF In		<p><b>IF Input</b> The input impedance at Pin 3 is 330 Ω. It matches the 330 Ω load impedance of a 10.7 MHz ceramic filter. Thus, no external matching is required.</p> <p><b>IF DEC1 &amp; DEC2</b> IF decoupling pins. Decoupling capacitors should be placed directly at the pins to enhance stability. Two capacitors are decoupled to the RF ground V<sub>CC1</sub>; one is placed between DEC1 &amp; DEC2.</p>	
4	IF Dec1			
5	IF Dec2			
6	IF Out		<p><b>IF Output</b> The output impedance is 330 Ω; it matches the 330 input resistance of a 10.7 MHz ceramic filter.</p>	

PIN FUNCTION DESCRIPTION

Pin	Symbol	Internal Equivalent Circuit	Description/External Circuit Requirements
7 8 9 10	VCC2 Lim In Lim Dec1 Lim Dec2		<p><b>Supply Voltage (VCC2)</b> This pin is VCC supply for the Limiter, Quadrature Detector, data slicer and RSSI buffer circuits. In the application PC board this pin is tied to a common VCC trace with VCC1.</p> <p><b>Limiter Input</b> The limiter input impedance is 330 <math>\Omega</math>.</p> <p><b>Limiter Decoupling</b> Decoupling capacitors are placed directly at these pins and to VCC (RF ground). Use the same procedure as in the IF decoupling.</p>
11, 14, 27 & 28	N/C		<p><b>No Connects</b> There is no internal connection to these pins; however it is recommended that these pins be connected externally to VCC (RF ground).</p>
12 13	Lim Out Quad		<p><b>Limiter Output</b> The output impedance is low. The limiter drives a quadrature detector circuit with in-phase and quadrature phase signals.</p> <p><b>Quadrature Detector Circuit</b> The quadrature detector is a doubly balanced four-quadrant multiplier with an internal 5.0 pF capacitor between Pins 12 and 13. An external capacitor may be added to increase the IF signal to Pin 13. The quadrature detector pin is provided to connect the external RLC parallel resonant network which provides the 90 degree phase shift and drives the quadrature detector.</p>
15 17 16	Det Gain Det Out VEE2		<p><b>Detector Buffer Amplifier</b> This is an inverting amplifier. An external feedback resistor from Pin 17 to 15, (the inverting input) controls the output amplitude; another resistor from Pin 15 to the negative supply (Pin 16) sets the DC output level. A 1:1 resistor ratio sets the output DC level at two VBE with respect to VEE. A small capacitor from Pin 17 to 15 can be used to set the bandwidth.</p> <p><b>Supply Ground (VEE2)</b> In the PCB layout, the ground pins (also applies to Pin 26) should be connected directly to chassis ground. Decoupling capacitors to VCC should be placed directly at the ground pins.</p>

PIN FUNCTION DESCRIPTION

Pin	Symbol	Internal Equivalent Circuit	Description/External Circuit Requirements	
19	DS "off"		<p><b>Data Slicer Off</b> The data output may be shut off to save current by placing DS "off" (Pin 19) at VCC.</p> <p><b>Data Slicer Output</b> In the application example a 10 kΩ pull-up resistor is connected to the collector of the output transistor at Pin 21.</p> <p><b>Data Slicer Ground</b> All the inverting switching stages in the comparator are supplied through the emitter pin of the output transistor (Pin 22) to ground rather than internally to VEE in order to reduce switching feedback to the front end.</p>	
21	DS Out			<p><b>Data Slicer Inputs</b> The data slicer has differential inputs with back to back diodes across them. The recovered signal is DC coupled to DS IN1 (Pin 18) at nominally <math>V_{18}</math> with respect to VEE; thus, it will maintain <math>V_{18} \pm V_{BE}</math> at Pin 18. DS IN2 (Pin 20) is AC coupled to VEE. The choice of coupling capacitor is dependent on the nature of the data signal. For small signal or continuous bits of the same polarity, the response time is relatively large. On the other hand, for large peak to peak voltage swings or when the DC level at the detector output changes, the response time is short. See the discussion in the application section for external circuit design details.</p>
22	DS Gnd			<p><b>RSSI Buffer</b> A unity gain amplifier is used to buffer the voltage at Pin 24 to 23. The output of the unity gain buffer (Pin 23) has an active pull up but no pull down. An external resistor is placed from Pin 23 to VEE to provide the pull down.</p> <p><b>RSSI</b> The RSSI output current creates a voltage drop across an external resistor from Pin 24 to VEE. The maximum RSSI current is 26 μA; thus, the maximum RSSI voltage using a 100 kΩ resistor is approximately 2.6 Vdc. Figure 22 shows the RSSI Output Voltage versus Input Signal Level in the application circuit.</p> <p>The negative slew rate is determined by an external capacitor and resistor to VEE (negative supply). The RSSI rise and fall times for various RF input signal levels and R24 values without the capacitor, C24 are displayed in Figure 24. This is the maximum response time of the RSSI.</p>
23	RSSI Buf			
24	RSSI			

PIN FUNCTION DESCRIPTION

Pin	Symbol	Internal Equivalent Circuit	Description/External Circuit Requirements
25	Enable		<p><b>Enable</b> The IC regulators are enabled by placing this pin at VEE.</p>
26	VEE1		<p><b>VCC and VEE ESD Protection</b> ESD protection diodes exist between the VCC and VEE pins. It is important to note that significant differences in potential (<math>&gt; 0.5 V_{BE}</math>) between the two VCC pins or between the VEE pins can cause these structures to start to conduct, thus compromising isolation between the supply busses. VCC1 &amp; VCC2 should be maintained at the same DC potential, as should VEE1 &amp; VEE2.</p>
28	Osc Base		<p><b>Oscillator Base</b> This pin is connected to the base lead of the common collector transistor. Since there is no internal bias resistor to the base, VCC is applied through an external choke or coil.</p> <p><b>Oscillator Emitter</b> This pin is connected to the emitter lead; the emitter is connected internally to a current source of about 200 <math>\mu</math>A. Additional emitter current may be obtained by connecting an external resistor to VEE; <math>I_E = V_{29}/R_{29}</math>.</p> <p>Details of circuits using overtone crystal and LC varactor controlled oscillators are discussed in the application section.</p>
31	Mix In1		<p><b>Mixer Inputs</b> The parallel equivalent differential input impedance of the mixer is approximately 2.0 k<math>\Omega</math> in parallel with 1.0 pF. This equates to a single ended input impedance of 1.0 k<math>\Omega</math> in parallel with 2.0 pF.</p> <p>The application circuit utilizes a SAW filter having a differential output that requires a 2.0 k<math>\Omega</math>    2.0 pF load. Therefore, little matching is required between the SAW filter and the mixer inputs. This and alternative circuits are discussed in more detail in the application section.</p>

Legacy Applications Information

APPLICATIONS INFORMATION

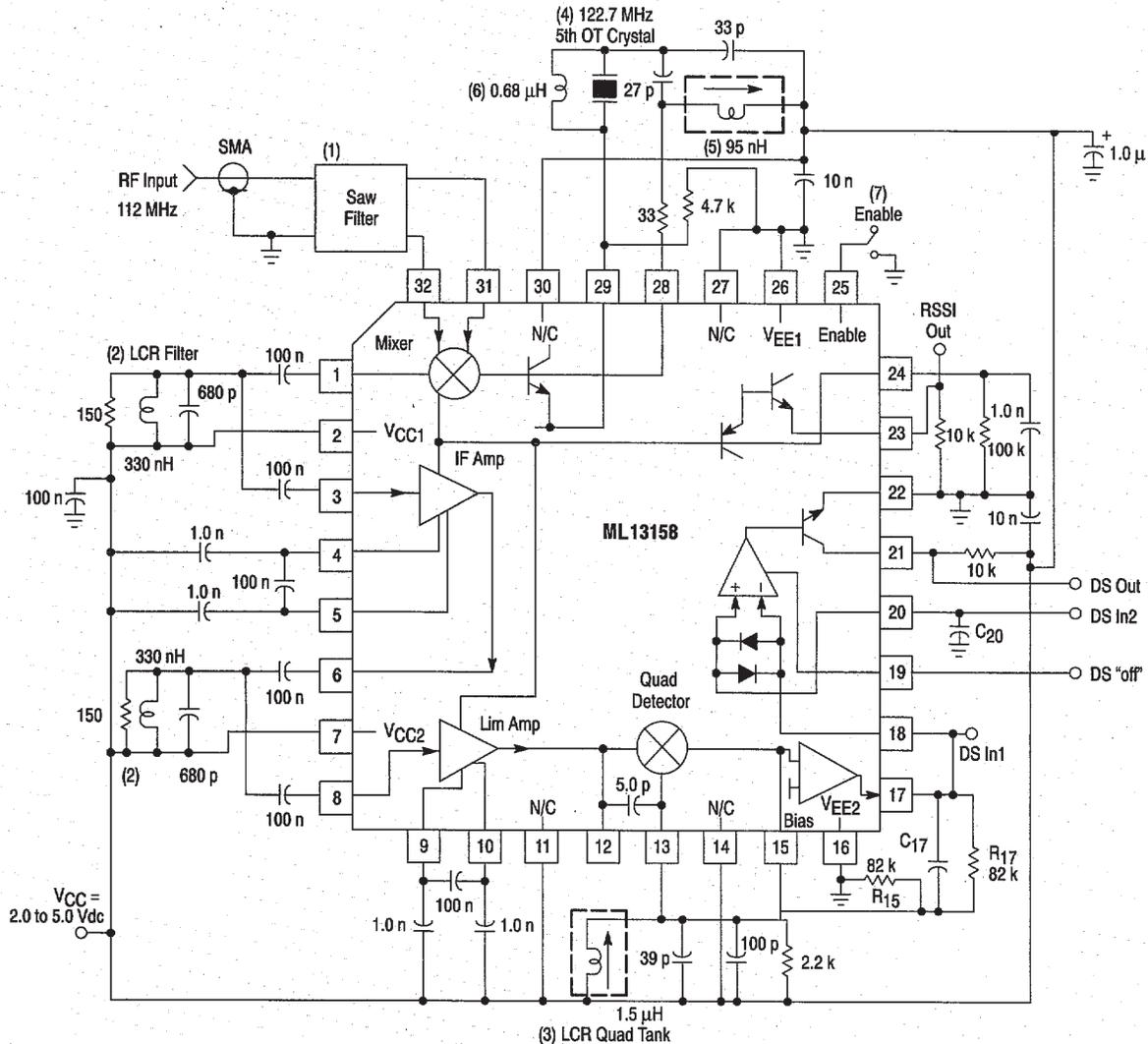
EVALUATION PC BOARD

The evaluation PCB is very versatile and is intended to be used across the entire useful frequency range of this device. The center section of the board provides an area for attaching all SMT components to the circuit side and radial leaded components to the component ground side (see Figures 29 and 30). Additionally, the peripheral area surrounding the RF core provides pads to add supporting and interface circuitry as a particular application requires. This evaluation board will be discussed and referenced in this section.

COMPONENT SELECTION

The evaluation PC board is designed to accommodate specific components, while also being versatile enough to use components from various manufacturers and coil types. Figures 13 and 14 show the placement for the components specified in the application circuit (Figure 12). The application circuit schematic specifies particular components that were used to achieve the results shown in the typical curves and tables but alternate components should give similar results.

Figure 12. Application Circuit

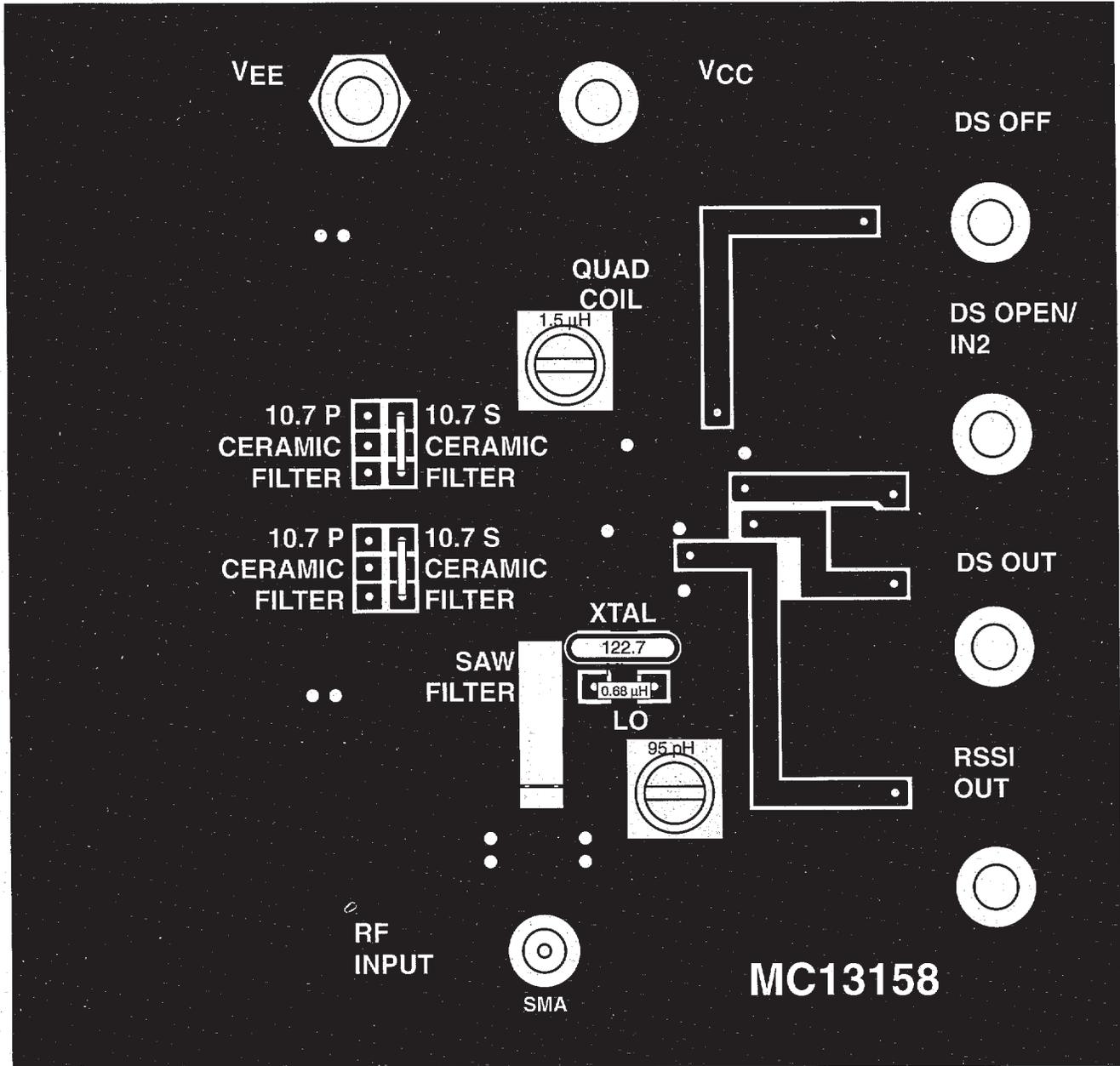


- NOTES:**
1. Saw Filter – Siemens part number Y6970M(5 pin SIP plastic package).
  2. An LCR filter reduces the broadband noise in the IF; ceramic filters may be used for data rates under 500 kHz. 4.0 dB insertion loss filters optimize the linearity of RSSI.
  3. The quadrature tank components are chosen to optimize linearity of the recovered signal while maintaining adequate recovered signal level. 1.5 μH 7.0 mm variable shielded inductor: Toko part # 292SNS-T1373Z. The shunt resistor is approximately equal to  $Q(2\pi fL)$ , where  $Q \sim 18$  (3.0 dB BW = 600 kHz).
  4. The local oscillator circuit utilizes a 122.7 MHz, 5th overtone, series resonant crystal specified with a frequency tolerance of 25 PPM, ESR of 120 Ω max. The oscillator configuration is an emitter coupled butler.
  5. The 95 nH (Nominal) inductor is a 7.0 mm variable shielded inductor: Coilcraft part # 150-04J08S or equivalent.
  6. 0.68 μH axial lead chokes (molded inductor): Coilcraft part # 90-11.
  7. To enable the IC, Pin 25 is taken to VEE. The external pull down resistor at Pin 29 could be linked to the enable function; otherwise if it is taken to VEE as shown, it will keep the oscillator biased at about 500 μA depending on the VCC level.
  8. The other resistors and capacitors are surface mount components.



Legacy Applications Information

Figure 14. Ground Side Component Placement



## Legacy Applications Information

### INPUT MATCHING/COMPONENTS

It is desirable to use a SAW filter before the mixer to provide additional selectivity and adjacent channel rejection. In a wideband system the primary sensitivity of the receiver backend may be achieved before the last mixer. Bandpass filtering in the limiting IF is costly and difficult to achieve for bandwidths greater than 280 kHz.

The SAW filter should be selected to easily interface with the mixer differential input impedance of approximately 2.0 k $\Omega$  in parallel with 1.0 pF. The PC board is dedicated to the Siemens SAW filter (part number Y6970M); the part is designed for DECT at 112 MHz 1st IF frequency. It is designed for a load impedance of 2.0

k $\Omega$  in parallel with 2.0 pF; thus, no or little input matching is required between the SAW filter and the mixer.

The Siemens SAW filter has an insertion loss of typically 10 dB and a 3.0 dB bandwidth of 1.0 MHz. The relatively high insertion loss significantly contributes to the system noise and a filter having lower insertion loss would be desirable. In existing low loss SAW filters, the required load impedance is 50  $\Omega$ ; thus, interface matching between the filter and the mixer will be required. Figure 15 is a table of the single-ended mixer input impedance. A careful noise analysis is necessary to determine the secondary contribution to system noise.

**Figure 15. Mixer Input Impedance**  
(Single-ended)

f (MHz)	R <sub>s</sub> ( $\Omega$ )	X <sub>s</sub> ( $\Omega$ )	R <sub>p</sub> ( $\Omega$ )	X <sub>p</sub> ( $\Omega$ )	C <sub>p</sub> (pF)
50	930	-350	1060	-2820	1.1
100	480	-430	865	-966	1.6
150	270	-400	860	-580	1.8
200	170	-320	770	-410	1.9
250	130	-270	690	-330	1.85
300	110	-250	680	-300	1.8
400	71	-190	580	-220	1.8
500	63	-140	370	-170	1.9
600	49	-110	300	-130	2.0

### SYSTEM NOISE CONSIDERATIONS

The system block diagram in Figure 16 shows the cascaded noise stages contributing to the system noise; it represents the application circuit in Figure 12 and a low noise preamp using a MRF941 transistor (see Figure 17). The preamp is designed for a conjugately matched input and output at 2.0 Vdc. V<sub>CE</sub> and 3.0 mAdc I<sub>C</sub>. S-parameters at 2.0 V, 3.0 mA and 100 MHz are:

$$\begin{aligned} S_{11} &= 0.86, -20 \\ S_{21} &= 9.0, 164 \\ S_{12} &= 0.02, 79 \\ S_{22} &= 0.96, -12 \end{aligned}$$

The bias network and V<sub>CE</sub> at 2.0 V and I<sub>C</sub> at 3.0 mA for V<sub>CC</sub> = 3.0 to 3.5 Vdc. The preamp operates with 18 dB gain and 2.7 dB noise figure.

In the cascaded noise analysis the system noise equation is:

$$F_{\text{system}} = F_1 + [(F_2 - 1)/G_1] + [(F_3 - 1)]/[(G_1)(G_2)]$$

where:

- F1 = the Noise Factor of the Preamp
- G1 = the Gain of the Preamp
- F2 = the Noise factor of the SAW filter
- G2 = the Gain of the SAW filter

F3 = the Noise factor of the Mixer

Note: the proceeding terms are defined as linear relationships and are related to the log form for gain and noise figure by the following:

$$F = \log^{-1}[(\text{NF in dB})/10] \text{ and similarly}$$

$$G = \log^{-1}[(\text{NF in dB})/10]$$

The noise figure and gain measured in dB are shown in the system block diagram. The mixer noise figure is typically 14 dB and the SAW filter adds typically 10 dB insertion loss. Addition of a low noise preamp having a 18 dB gain and 2.7 dB noise figure not only improves the system noise figure but it increases the reverse isolation from the local oscillator to the antenna input at the receiver. Calculating in terms of gain and noise factor yields the following:

$$F_1 = 1.86; G_1 = 63.1$$

$$F_2 = 10; G_2 = 0.1$$

$$F_3 = 25.12$$

Thus, substituting in the equation for system noise factor:

$$F_{\text{system}} = 5.82; \text{NF}_{\text{system}} = 7.7 \text{ dB}$$

## Legacy Applications Information

Figure 16. System Block Diagram for Noise Analysis

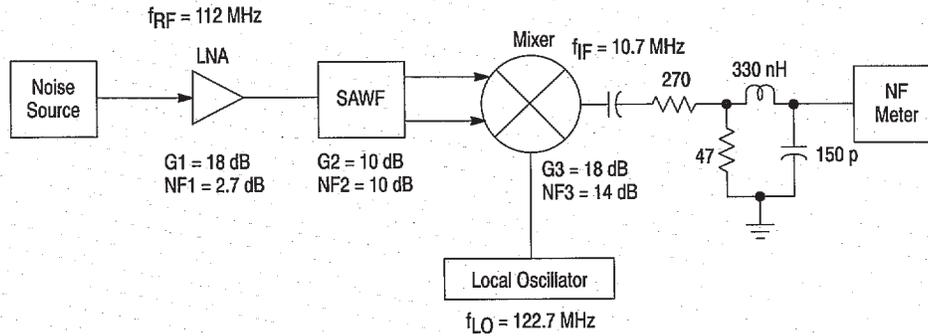
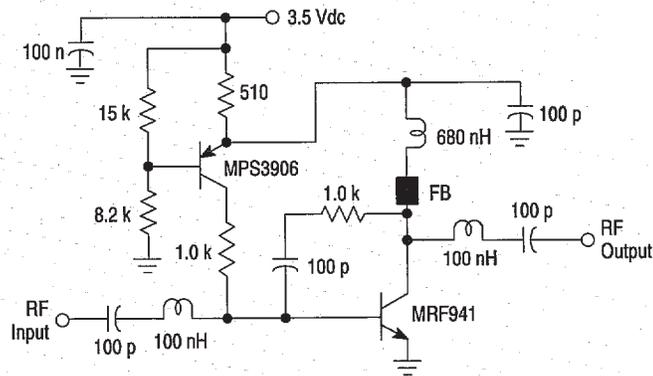


Figure 17. 112 MHz LNA



## LOCAL OSCILLATORS

## VHF APPLICATIONS

The on-chip grounded collector transistor may be used for HF and VHF local oscillator with higher order overtone crystals. The local oscillator in the application circuit (Figure 12) shows a 5th overtone oscillator at 122.7 MHz. This circuit uses a Butler overtone oscillator configuration. The amplifier is an emitter follower. The crystal is driven from the emitter and is coupled to the high impedance base through a capacitive tap network. Operation at the desired overtone frequency is ensured by the parallel resonant circuit formed by the variable inductor and the tap capacitor and parasitic capacitances of the on-chip transistor and PC board. The variable inductor specified in the schematic could be replaced with a high tolerance, high Q ceramic or air wound surface mount component if the other components have tight enough tolerances. A variable inductor provides an adjustment for gain and frequency of the resonant tank ensuring lock up and start-up of the crystal oscillator. The overtone crystal is chosen with ESR of typically 80  $\Omega$  and 120  $\Omega$  maximum; if the resistive loss in the crystal is too high the performance of oscillator may be impacted by lower gain margins.

A series LC network to ground (which is  $V_{CC}$ ) is comprised of the inductance of the base lead of the on-chip transistor and PC board

traces and tap capacitors. Parasitic oscillations often occur in the 200 to 800 MHz range. A small resistor is placed in series with the base (Pin 28) to cancel the negative resistance associated with this undesired mode of oscillation. Since the base input impedance is so large a small resistor in the range of 27 to 68  $\Omega$  has very little effect on the desired Butler mode of oscillation.

The crystal parallel capacitance,  $C_O$ , provides a feedback path that is low enough in reactance at frequencies of 5th overtones or higher to cause trouble.  $C_O$  has little effect near resonance because of the low impedance of the crystal motional arm ( $R_M-L_M-C_M$ ). As the tunable inductor which forms the resonant tank with the tap capacitors is tuned "off" the crystal resonant frequency it may be difficult to tell if the oscillation is under crystal control. Frequency jumps may occur as the inductor is tuned. In order to eliminate this behavior an inductor,  $L_O$ , is placed in parallel with the crystal.  $L_O$  is chosen to be resonant with the crystal parallel capacitance,  $C_O$ , at the desired operation frequency. The inductor provides a feedback path at frequencies well below resonance; however, the parallel tank network of the tap capacitors and tunable inductor prevent oscillation at these frequencies.

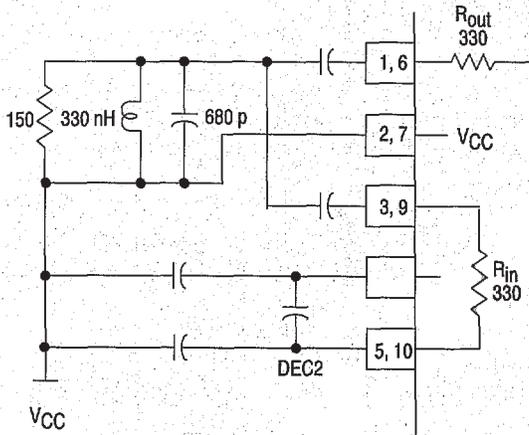
Legacy Applications Information

**IF FILTERING/MATCHING**

In wideband data systems the IF bandpass needed is greater than can be found in low cost ceramic filters operating at 10.7 MHz. It is necessary to bandpass limit with LC networks or series-parallel ceramic filter networks. Murata offers a series-parallel resonator pair (part number KMFC545) with a 3.0 dB band width of ±325 kHz and a maximum insertion loss of 5.0 dB. The application PC board is laid out to accommodate this filter pair (a filter pair is used at both locations of the split IF). However, even using a series parallel ceramic filter network yields only a maximum bandpass of 650 kHz. In some applications a wider band IF bandpass is necessary.

A simple LC network yields a bandpass wider than the SAW filter but it does reduce an appreciable amount of wideband IF noise. In the application circuit an LC network is specified using surface mount components. The parallel LC components are placed from the outputs of the mixer and IF amplifier to the VCC trace; internal 330Ω loads are connected from the mixer and IF amplifier outputs DEC2 (Pin 5 and 10 respectively). This loads the outputs with the optimal load impedance but creates a low insertion loss filter. An external shunt resistor may be used to widen the bandpass and to acquire the 10 dB composite loss necessary to linearize the RSSI output. The equivalent circuit is shown in Figure 18.

Figure 18. IF LCR Filter Equivalent



The following equations satisfy the 12 dB loss (1:4 resistive ratio):  

$$\frac{(R_{ext})(330)}{(R_{ext} + 330)} = \text{Requivalent}$$

$$\frac{\text{Requivalent}}{\text{Requivalent} + 330} = 1/4$$
 Solve for Requivalent:  

$$4(\text{Requivalent}) = \text{Requivalent} + 330$$

$$3(\text{Requivalent}) = 330$$

$$\text{Requivalent} = 110$$

Substitute for Requivalent and solve for Rext  

$$330(R_{ext}) = 110 (R_{ext}) + (330)(110)$$

$$R_{ext} = (330)(110)/220$$

$$R_{ext} = 165 \Omega$$

The IF is 10.7 Mhz although any IF between 10 to 20 MHz could be used. The value of the coil is lowered from that used in the quadrature circuit because the unloaded Q must be maintained in a surface mount component. A standard value component having an unloaded Q = 100 at 10.7 MHz is 330 nH; therefore the capacitor is 669 pF. Standard values have been chosen for these components;

$R_{ext} = 150 \Omega$   
 $C = 680 \text{ pF}$   
 $L = 330 \text{ nH}$

Computation of the loaded Q of the is LCR network is  

$$Q = \text{Requivalent}/X_L$$
 where  $X_L = 2\pi f l$  and Requivalent is 103 Ω  
 Thus,  $Q = 4.65$

The total system loss is  

$$20 \log (103/433) = -12.5 \text{ dB}$$

**QUADRATURE DETECTOR**

The quadrature detector is coupled to the IF with an internal 5.0 pF capacitor between Pins 12 and 13. For wideband data applications, the drive to the detector can be increased with an additional external capacitor between these pins; thus, the recovered signal level output is increased for a given bandwidth.

The wideband performance of the detector is controlled by the loaded Q of the LC tank circuit. The following equation defines the components which set the detector circuit's bandwidth:

$$Q = R_T/X_L \tag{1}$$

where  $R_T$  is the equivalent shunt resistance across the LC Tank  
 $X_L$  is the reactance of the quadrature inductor at the IF frequency ( $X_L = 2\pi f l$ ).

The inductor and capacitor are chosen to form a resonant LC tank with the PCB and parasitic device capacitance at the desired IF center frequency as predicted by

$$f_c = [2\pi (LC_p)^{1/2}]^{-1} \tag{2}$$

where  $L$  is the parallel tank inductor  $C_p$  is the equivalent parallel capacitance of the parallel resonant tank circuit.

The following is a design example for a wideband detector at 10.7 MHz and a loaded Q of 18. The loaded Q of the quadrature detector is chosen somewhat less than the Q of the IF bandpass. For an IF frequency of 10.7 MHz and an IF bandpass of 600 kHz, the IF bandpass Q is approximately 6.4.

Legacy Applications Information

**EXAMPLE:**

Let the external  $C_{ext} = 139 \text{ pF}$ . (The minimum value here should be much greater than the internal device and PCB parasitic capacitance,  $C_{int} = 3.0 \text{ pF}$ ). Thus,  $C_p = C_{int} + C_{ext} = 142 \text{ pF}$ .

Rewrite equation (2) and solve for L

$$L = (0.159)^2 / (C_p f_c^2)$$

$L = 1.56 \text{ } \mu\text{H}$ ; Thus, a standard value is chosen:

$L = 1.56 \text{ } \mu\text{H}$  (tunable shielded inductor)

The value of the total damping resistor to obtain the required loaded Q of 18 can be calculated by rearranging equation (1):

$$R_T = Q(2\pi f l)$$

$$R_T = 18(2\pi)(10.7)(1.5) = 1815 \Omega$$

The internal resistance,  $R_{int}$  at the quadrature tank Pin 13 is approximately  $13 \text{ k}\Omega$  and is considered in determining the external resistance,  $R_{ext}$  which is calculated from

$$R_{ext} = ((R_T)(R_{int})) / (R_{int} - R_T)$$

$R_{ext} = 2110$ ; Thus choose the standard value

$R_{ext} = 2.2 \text{ k}\Omega$

It is important to set the DC level of the detector output at Pin 17 to center the peak to peak swing of the recovered signal. In the equivalent internal circuit shown in the Pin Function Description, the reference voltage at the positive terminal of the inverting, op amp buffer amplifier is set at  $1.0 V_{BE}$ . The detector DC level,  $V_{17}$  is determined by the following equation:

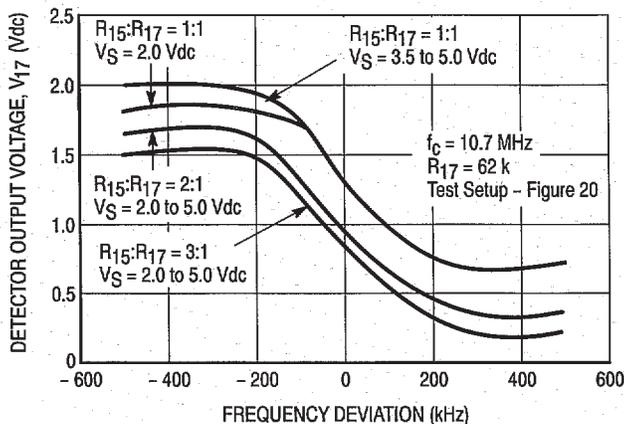
$$V_{17} = [(R_{15}/R_{17}) + 1] / (R_{15}/R_{17}) V_{BE}$$

Thus for a 1:1 ratio of  $R_{15}/R_{17}$ ,  $V_{17} = 2.0 V_{BE} = 1.4 \text{ Vdc}$ .

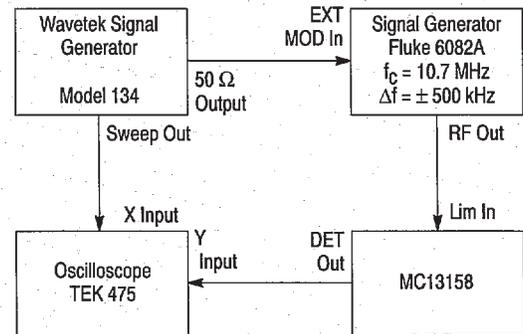
Similarly for a 2:1,  $V_{17} = 1.5 V_{BE} = 1.05 \text{ Vdc}$ ; and for 3:1,  $V_{17} = 1.33 V_{BE} = 0.93 \text{ Vdc}$ .

Figure 19 shows the detector "S-Curves", in which the resistor ratio is varied while maintaining a constant gain ( $R_{17}$  is held at  $62 \text{ k}\Omega$ ).  $R_{15}$  is  $62 \text{ k}\Omega$  for a 1:1 ratio; while  $R_{15} = 120 \text{ k}\Omega$  and  $180 \text{ k}\Omega$  to produce the 2:1 and 3:1 ratio. The IF signal into the detector is swept  $\pm 500 \text{ kHz}$  about the  $10.7 \text{ MHz}$ , IF center frequency. The resulting curve show how the resistor ratio and the supply voltage effects the symmetry of the "S-curve" (Figure 21 Test Setup). For the 3:1 and 2:1 ratio, symmetry is maintained with  $V_S$  from 2.0 to 5.0 Vdc; however, for the 1:1 ratio, symmetry is lost at 2.0 Vdc.

**Figure 19. Detector Output Voltage versus Frequency Deviation**



**Figure 20. Demodulator "S-Curve" Test Setup**



**DATA SLICER CIRCUIT**

$C_{20}$  at the input of the data slicer is chosen to maintain a time constant long enough to hold the charge on the capacitor for the longest strings of bits at the same polarity. For a data rate of  $576 \text{ kHz}$  a bit stream of 15 bits at the same polarity would equate to an apparent data rate of approximately  $77 \text{ kbps}$  or  $38 \text{ kHz}$ . The time constant would be approximately  $26 \text{ } \mu\text{s}$ . The following expression equates the time constant  $t$ , to the external components;

$$t = 2\pi (R_{18})(C_{20})$$

Solve for  $C_{20}$ :

$$C_{20} = t / 2\pi (R_{18})$$

where the effective resistance  $R_{18}$  is a complex function of the demodulator feedback resistance and data slicer input circuit. In the data input network the back to back diodes form a charge and discharge path for the capacitor at Pin 20; however, the diodes create a non-linear response. This resistance is loaded by the  $\beta$ , beta of the detector output transistor;  $\beta = 100$  is a typical value (see Figure 21). Thus, the apparent value of the resistance at Pin 18 (DS IN1) is approximately equal to:

$$R_{18} \sim R_{17} / 100$$

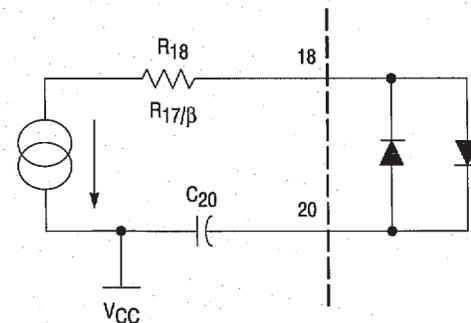
where  $R_{17}$  is  $82 \text{ k}\Omega$ , the feedback resistor from Pin 17 to 15.

Therefore, substituting for  $R_{18}$  and solving for  $C_{20}$ :

$$C_{20} = 15.9 (t) / R_{17} = 5.04 \text{ nF}$$

The closest standard value is  $4.7 \text{ nF}$ .

**Figure 21. Data Slicer Equivalent Input Circuit**



Legacy Applications Information

RSSI

In Figure 22, the RSSI versus RF Input Level shows the linear response of RSSI over a 65 dB range but it has extended capability over 80 dB from -80 dBm to +10 dBm. The RSSI is measured in the application circuit (Figure 12) in which a SAW filter is used before the mixer; thus, the overall sensitivity is traded off for the sake of selectivity. The curves are shown for three filters having different bandwidths:

- 1) LCR Filter with 2.3 MHz 3.0 dBm BW (Circuit and Component Placement is shown in Figure 12)
- 2) Series-Parallel Ceramic Filter with 650 kHz 3.0 dB BW (Murata Part # KMFC - 545)
- 3) Ceramic Filter with 280 kHz 3.0 dB BW.

Figure 22. RSSI Output Voltage versus Signal Input Level

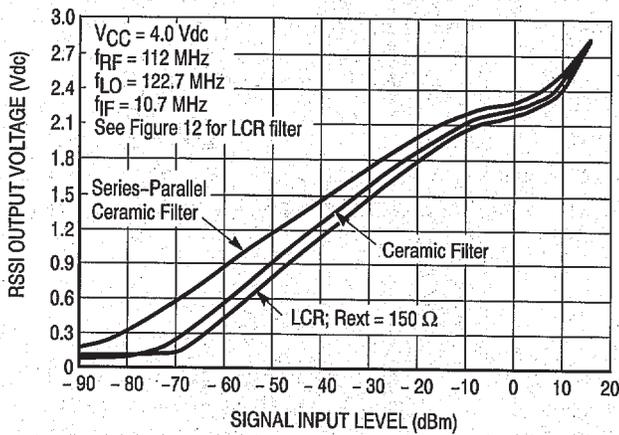
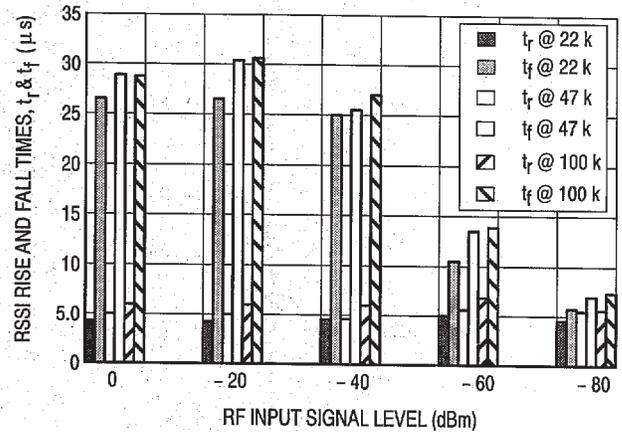


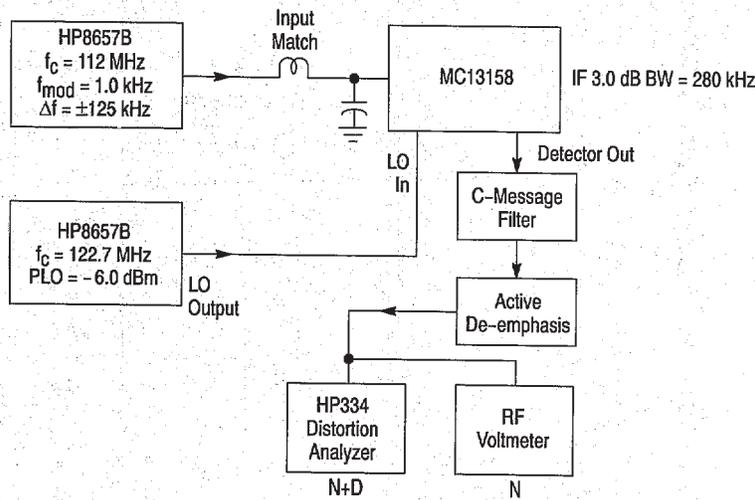
Figure 23. RSSI Output Rise and Fall Times versus RF Input Signal Level



SINAD PERFORMANCE

Figure 24 shows a test setup for a narrowband demodulator output response in which a C-message filter and an active de-emphasis filter is used following the demodulator. The input is matched using a 1:4 impedance transformer. The SINAD performance is shown in Figure 25 with no preamp and in Figure 26 with a preamp (Preamp-Figure 16). The 12 dB SINAD sensitivity is -101 dBm with no preamp and -113 dBm with the preamp.

Figure 24. Test Setup for Narrowband SINAD



Legacy Applications Information

Figure 25. S+N+D, N+D, N versus Input Signal Level (without preamp)

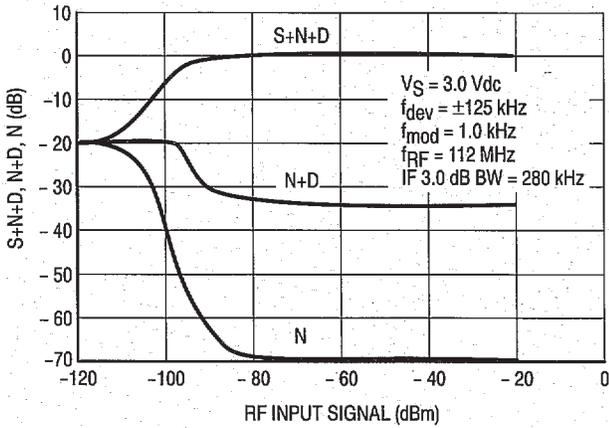


Figure 26. S+N+D, N+D, N versus Input Signal Level (with preamp)

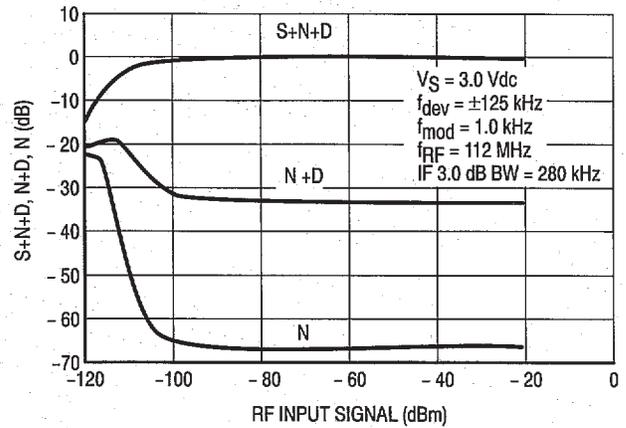


Figure 27. Input IP3, 1.0 dB Compression Pt. Test Setup

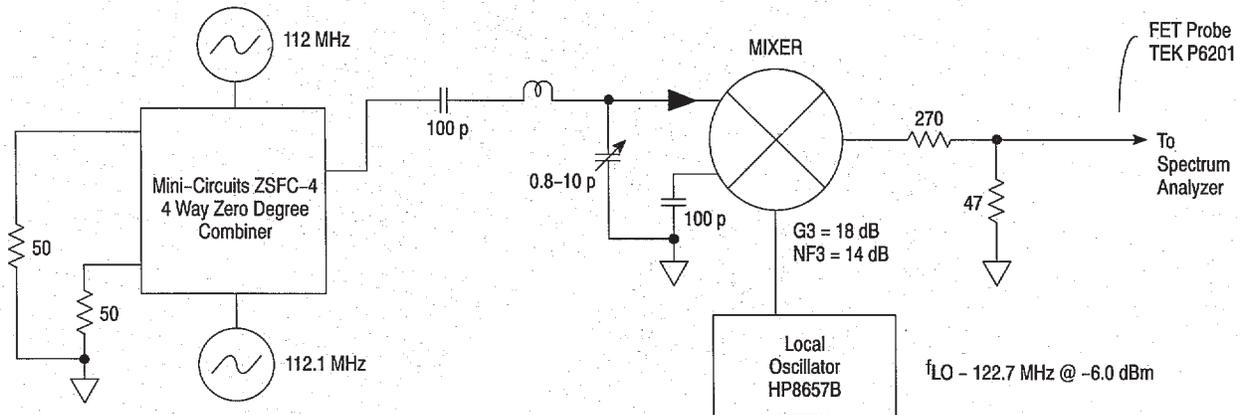
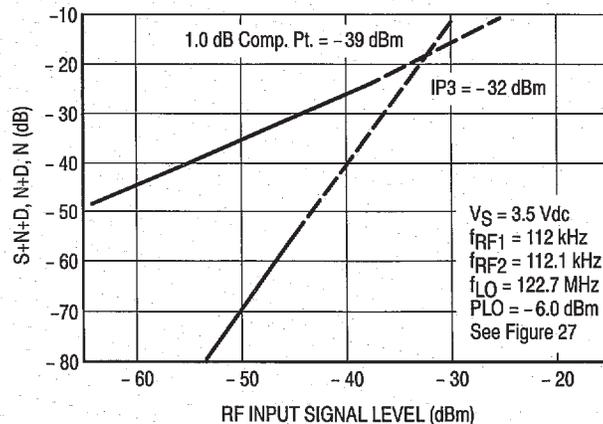
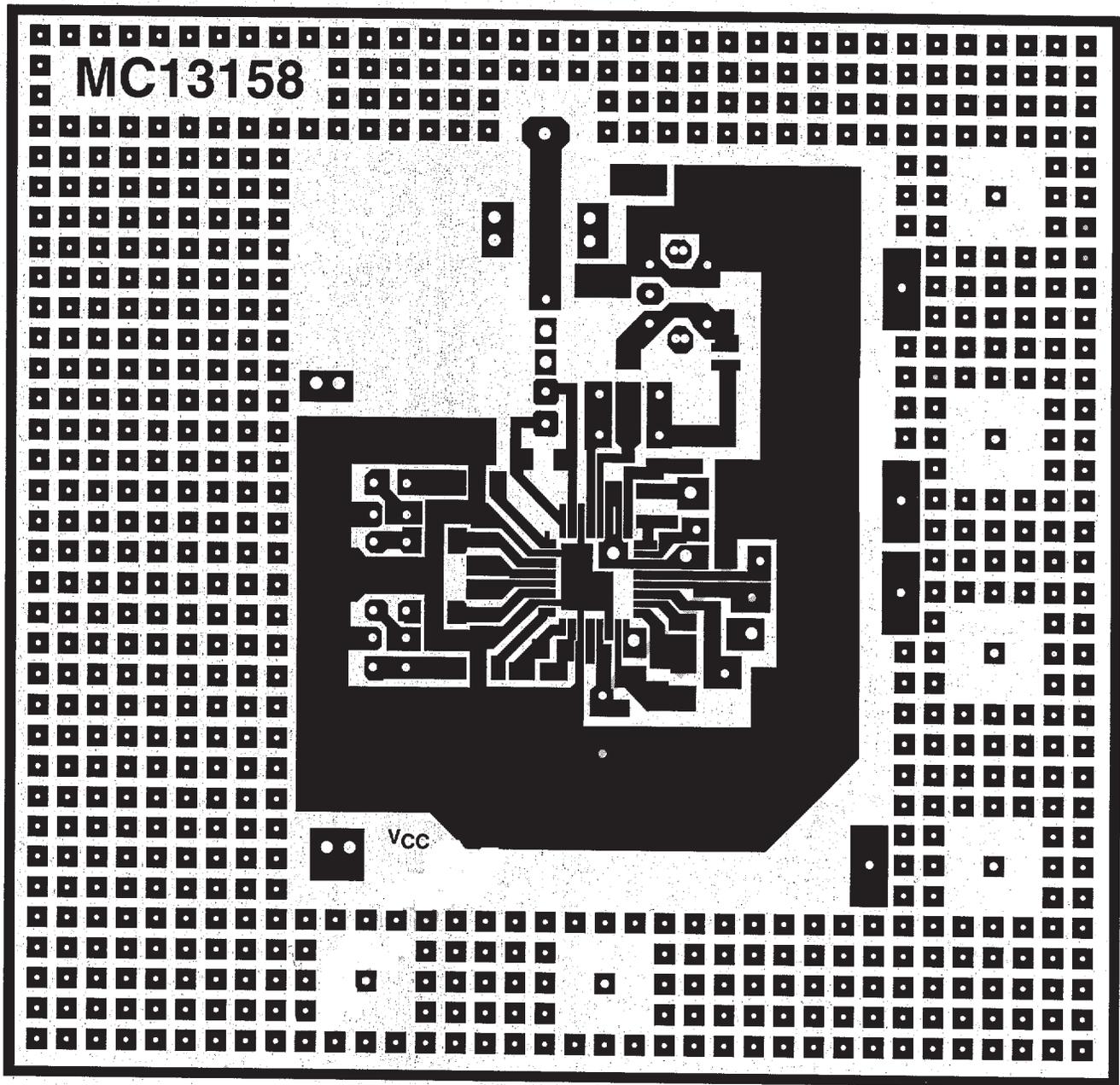


Figure 28. 1.0 dB Compression Pt. and Input Third Order Intercept



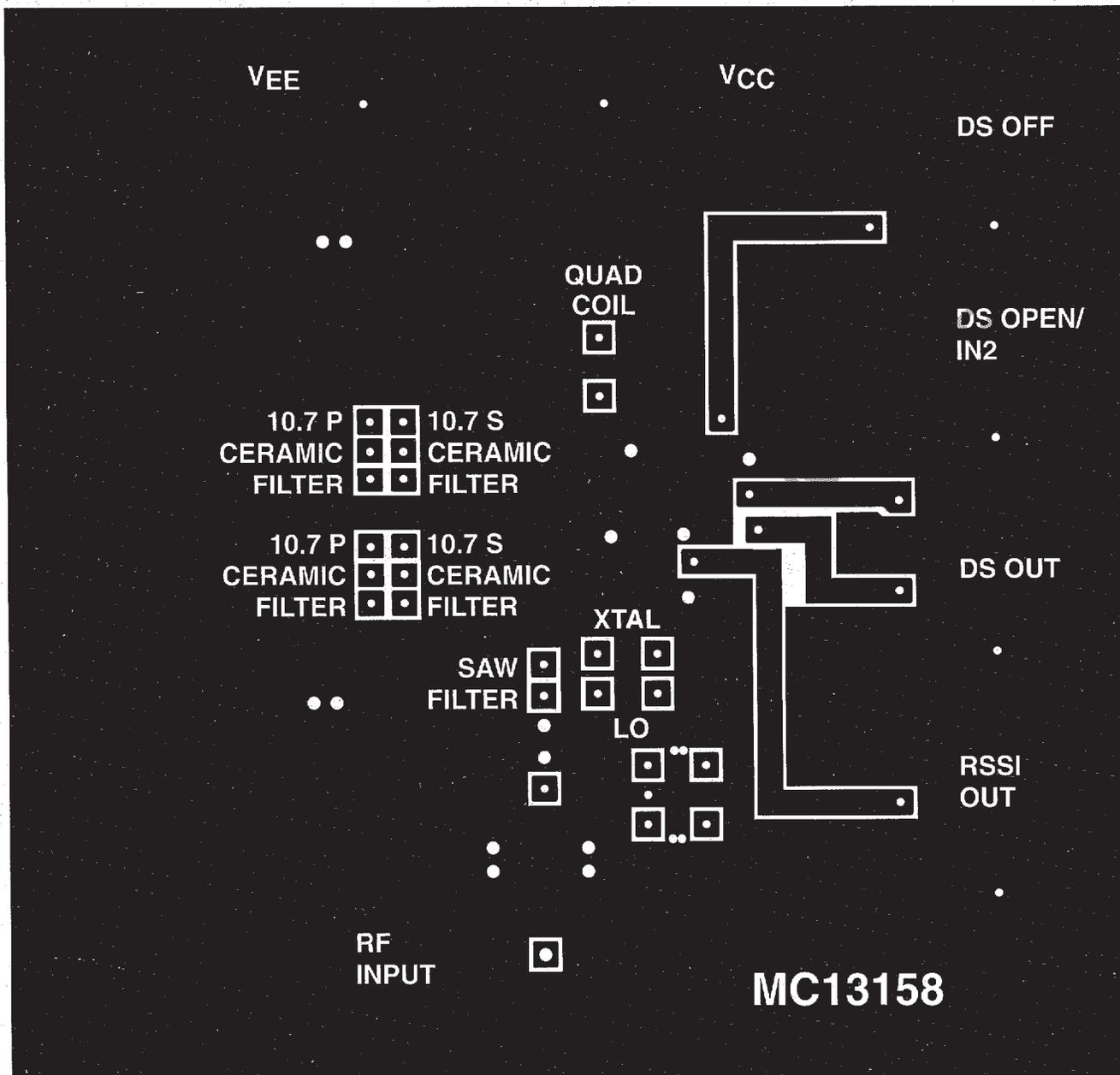
Legacy Applications Information

Figure 29. Circuit Side View



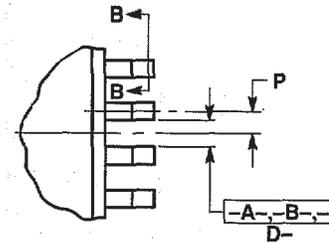
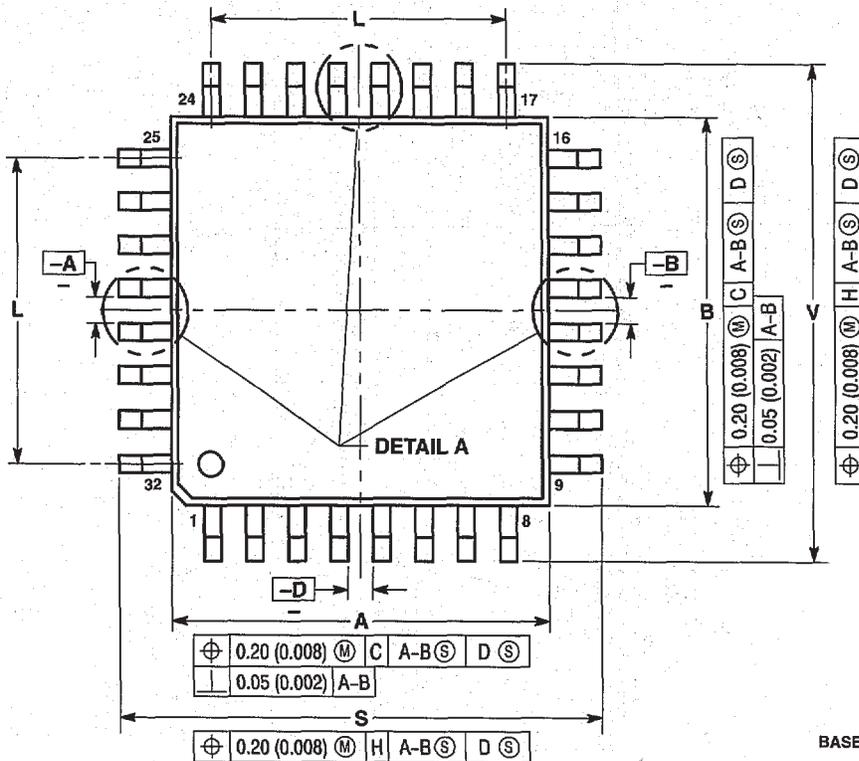
Legacy Applications Information

Figure 30. Ground Side View

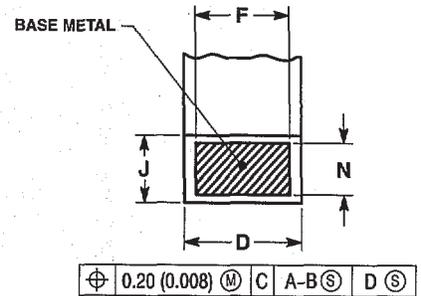


OUTLINE DIMENSIONS

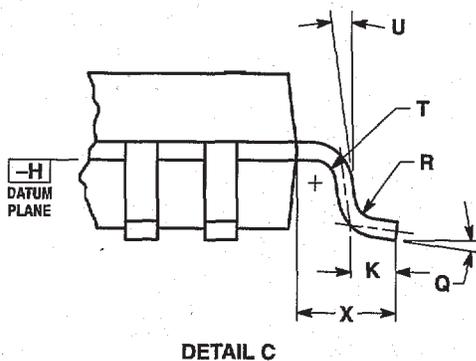
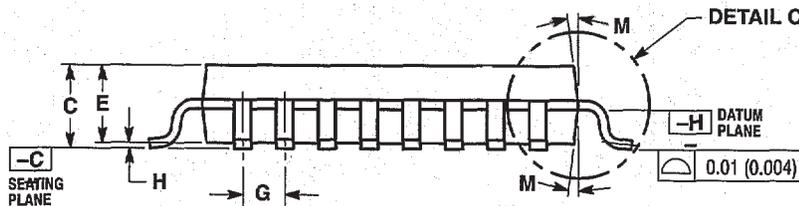
QFP 32 = 8P  
 PLASTIC QFP PACKAGE  
 (ML13158-8P)  
 CASE 873-01



DETAIL A



SECTION B-B  
 VIEW ROTATED 90° CLOCKWISE



DETAIL C

NOTES:

1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: MILLIMETER.
3. DATUM PLANE -H- IS LOCATED AT BOTTOM OF LEAD AND IS COINCIDENT WITH THE LEAD WHERE THE LEAD EXITS THE PLASTIC BODY AT THE BOTTOM OF THE PARTING LINE.
4. DATUMS -A-, -B- AND -D- TO BE DETERMINED AT DATUM PLANE -H-.
5. DIMENSIONS S AND V TO BE DETERMINED AT SEATING PLANE -C-.
6. DIMENSIONS A AND B DO NOT INCLUDE MOLD PROTRUSION. ALLOWABLE PROTRUSION IS 0.25 (0.010) PER SIDE. DIMENSIONS A AND B DO INCLUDE MOLD MISMATCH AND ARE DETERMINED AT DATUM PLANE -H-.
7. DIMENSION D DOES NOT INCLUDE DAMBAR PROTRUSION. ALLOWABLE DAMBAR PROTRUSION SHALL BE 0.08 (0.003) TOTAL IN EXCESS OF THE D DIMENSION AT MAXIMUM MATERIAL CONDITION. DAMBAR CANNOT BE LOCATED ON THE LOWER RADIUS OR THE FOOT.

DIM	MILLIMETERS		INCHES	
	MIN	MAX	MIN	MAX
A	6.95	7.10	0.274	0.280
B	6.95	7.10	0.274	0.280
C	1.40	1.60	0.055	0.063
D	0.273	0.373	0.010	0.015
E	1.30	1.50	0.051	0.059
F	0.273	-	0.010	-
G	0.80 BSC	-	0.031 BSC	-
H	-	0.20	-	0.008
J	0.119	0.197	0.005	0.008
K	0.33	0.57	0.013	0.022
L	5.6 REF	-	0.220 REF	-
M	6°	8°	6°	8°
N	0.119	0.135	0.005	0.005
P	0.40 BSC	-	0.016 BSC	-
Q	5°	10°	5°	10°
R	0.15	0.25	0.006	0.010
S	8.85	9.15	0.348	0.360
T	0.15	0.25	0.006	0.010
U	5°	11°	5°	11°
V	8.85	9.15	0.348	0.360
X	1.0 REF	-	0.039 REF	-

Lansdale Semiconductor reserves the right to make changes without further notice to any products herein to improve reliability, function or design. Lansdale does not assume any liability arising out of the application or use of any product or circuit described herein; neither does it convey any license under its patent rights nor the rights of others. "Typical" parameters which may be provided in Lansdale data sheets and/or specifications can vary in different applications, and actual performance may vary over time. All operating parameters, including "Typicals" must be validated for each customer application by the customer's technical experts. Lansdale Semiconductor is a registered trademark of Lansdale Semiconductor, Inc.