SA900 I/Q transmit modulator for 1GHz applications AN1892

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INTRODUCTION

The SA900 (Figure 1) is a truly universal in-phase and quadrature (I/Q) radio transmitter that can perform many types of analog and digital modulation including AM, FM, SSB, QAM, BPSK, QPSK, FSK, etc. It is a highly integrated system which saves space and cost for the manufacturers producing cellular and wireless products. The device allows baseband signals to directly modulate the I/Q carriers, which are generated by internal phase shift network, in the 1GHz range, and to maintain good linearity required for linear modulation scheme (e.g., π/4-DQPSK). It contains an on-chip frequency divider, phase detector, and VCO, which can be built into a phase-locked loop (PLL) frequency synthesizer to create a transmit offset frequency. Its unique internal design allows frequency conversion without having an external image rejection filter for eliminating the sum term after mixing. The SA900 meets the specifications required by the IS-54, the industry standard for North America Digital Cellular (NADC) system. This application note reviews the basic concept of I/Q modulation and discusses the key points when designing the SA900 for an RF transmitter.

I/Q MODULATION

Any bandpass RF signals can be represented in polar form by

$$s(t) = A(t)\cos[\omega_c t + \phi(t)] \qquad \text{(EQ. 1)}$$

where A(t) is the signal envelope and ϕ(t) is the phase. By using the trigonometric identities, we can represent EQ. 1 in rectangular form by

$$s(t) = I(t)\cos[\omega_c t] - Q(t)\sin[\omega_c t], \qquad \text{(EQ. 2)}$$

$$I(t) = A(t)\cos[\phi(t)]$$

$$Q(t) = A(t)\sin[\phi(t)]$$

Since the baseband signals I(t) and Q(t) modulate two exactly 90° out-of-phase carriers $\cos(w_c t)$ and $-\sin(w_c t)$ respectively, we call the system implementing EQ. 2 an in-phase and quadrature (I/Q) modulator. Figure 2 shows the mathematics and hardware implementation of an I/Q modulator.

The local oscillator, usually a VCO within a PLL, generates the carrier and is split into two equal signals. One goes directly into a double-balanced mixer to form the I-channel and the other one goes into the other mixer via a 90° phase shifter (realized by passive elements) to provide the Q-channel. The baseband signals I(t) and Q(t), either analog or digital in nature, modulate the carrier to produce the I and Q components which are finally combined to form the desired RF transmitting signal. Since any RF signal can be represented in the I/Q form, any modulation scheme can be implemented by an I/Q modulator.

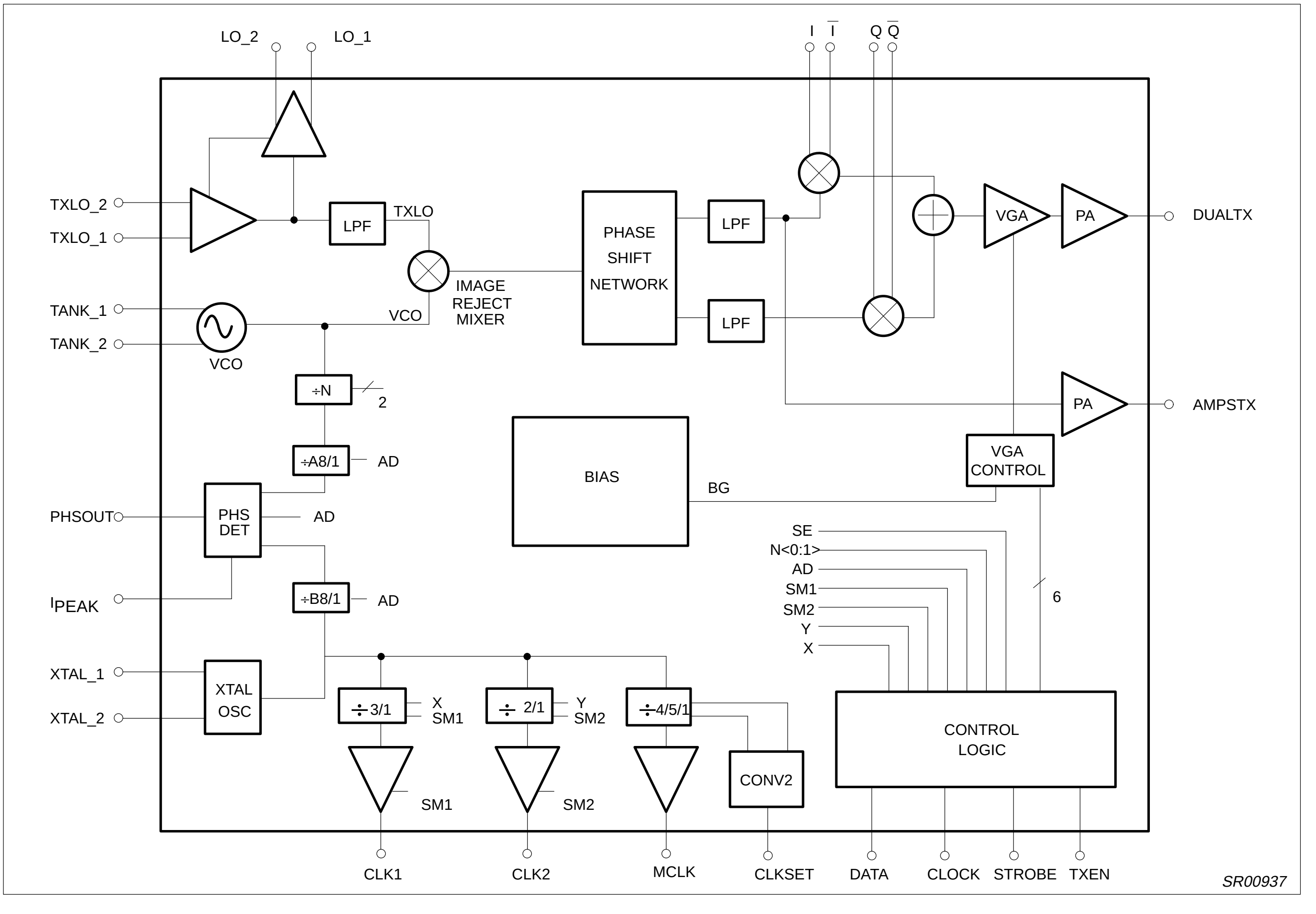

Figure 1. SA900 Transmit Modulator

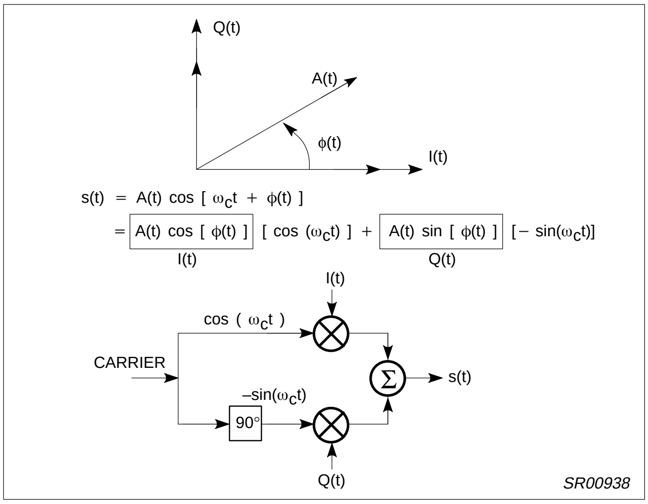

Figure 2. Mathematical Representation and Hardware Implementation of I/Q Modulator

Linear Digital Modulation

Linear digital modulation techniques depend on varying the phase and/or magnitude of an analog carrier according to some digital information: ones and zeros. This digital information can be the output of an analog-to-digital converter (e.g. voice codec), or it can be digital data in some standard formats (e.g. ASCII). The most popular digital signaling format is non return-to-zero (NRZ), where 1s and 0s are converted into signal with amplitude of 1 and -1, respectively, in a symbol duration. Since NRZ signal has infinite bandwidth, transmit filters have to be used to limit the spectral spreading. To ensure each NRZ symbol does not smear into its neighbors due to low-pass filtering and channel distortion causing inter-symbol interference (ISI), the frequency response of the low-pass filter has to satisfy Nyquist criteria. One example of this type of filter is the linear phase square-root-raised cosine filter. Together with the same type of filter for receive low-pass filtering, the signal is guaranteed ISI free in a Gaussian environment. One straight-forward technique of transmitting these bandlimited signals through communication channels would be applying it directly to the mixer of the I-channel to generate the RF signal. This is known as binary phase shift keying (BPSK), where the phase of the carrier is shifted 180° to transmit a data change from 0 to 1 or 1 to 0.

Quadrature or quaternary phase shift keying (QPSK) is a much more common type of modulation scheme used in mobile and satellite communications. It has four possible states (90° apart) and each of them represents two bits of data. Figure 3 shows the baseband generator for QPSK (without the differential phase encoder). NRZ data bits go through the serial-to-parallel converter (see Figure 4) and are mapped in accordance to some rules to generate I and Q values. The generic rule will be the values of I and Q components are 1 and 1 for the data bits “11” (45°) and -1 and -1 for the data bits “00” (-135°). These discrete signals have to be bandlimited by Nyquist low-pass filters to be ISI free.

A more sophisticated way of mapping results in π/4-DQPSK (D for differential encoding), which is chosen for North America Digital Cellular (IS-54), Personal Digital Cellular (PDC) in Japan, and Personal Handy Phone System (PHS) in Japan. In this scheme, consecutive pairs of bits are encoded into one of the four possible phases: π/4 for “11”, 3π/4 for “01”, -3π/4 for “00”, and -π/4 for “10”. However, unlike the previous case that “11” is always π/4 and “00” is always -3π/4, the encoded phases are the degrees that the carrier has to shift at each sampling instances. Thus, the information is contained in the phase difference (differential) instead of absolute phase for π/4-DQPSK.

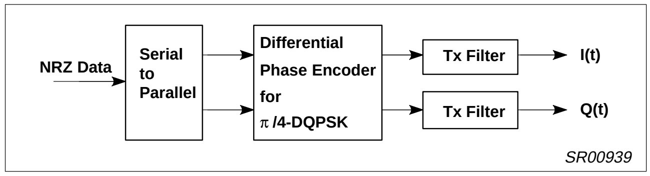

Figure 3. QPSK and π/4-DQPSK Baseband Generator

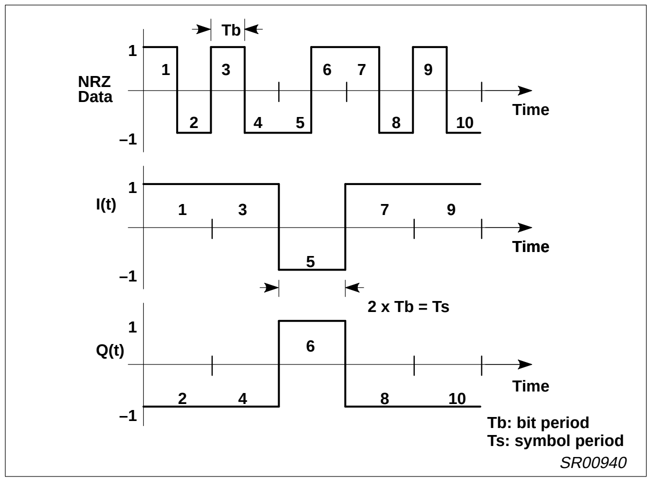

Figure 4. Serial-to-Parallel Conversion

A better way to tell the difference between QPSK and π/4-DQPSK is by looking at the signal constellation diagram, shown in Figure 5, which displays the possible values of I and Q vectors and change of states. Constellation diagram is also known as phase diagram because it shows the phase of the carrier at the sampling point. Notice that the phases of QPSK are assigned for every two bits of data; therefore, it can transmit twice as much information as BPSK in a given bandwidth, i.e., more bandwidth efficient. 8-PSK is another type of modulation used for high efficiency requirements. It maps three bits into 8 phases, 45° apart, in the constellation. More spectral efficient modulation can be created by mapping more bits into one phase at each sampling point. However, as you put more dots in the signal constellation, the signal susceptibility to noise is lower because the decision distance is shorter (dots are closer). Then, it requires higher carrier-to-noise (C/N) ratio to maintain the same bit error rate (BER).

One common misconception is that since π/4-DQPSK has 8 states in the constellation, it is just another type of 8-PSK. Notice that at every sampling instant, the carrier of π/4-DQPSK is only allowed to switch to one of the 4 possible states (see Figure 5). So, we still have two data bits which get encoded into 4 phases. Thus, it has the same spectral efficiency as QPSK for the same carrier power. The reason for using this modulation scheme is twofold. First, the envelope fluctuation, which causes spectral spreading due to nonlinearity of transmitter and amplifier, is reduced because the maximum phase shift is 135° instead of 180°. Second, the signal can be demodulated non-coherently which simplifies the receiver circuitry by eliminating the need for carrier recovery.

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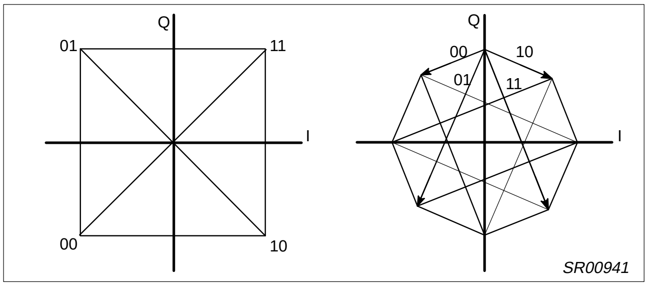


Figure 5. Signal Constellation of QPSK and π/4-DQPSK

Digital and Analog FM

Another family of digital modulation is categorized by frequency change of the carrier instead of phase and/or amplitude change. One of them is frequency shift keying (FSK), where the carrier switches between two frequencies. FSK is also known as digital FM because it can be generated by feeding the NRZ data stream into an analog VCO. FSK appears as a unit circle in the signal constellation because the RF signal envelope is constant and the phase is continuous. Baseband filtering is usually applied for FSK to limit the RF bandwidth of the signal so that more channels can fit into a given frequency band.

One common modulation of this type is known as Gaussian minimum shift keying (GMSK), which is used for GSM and some other wireless applications. GMSK can be generated by following its definition: bandlimit the NRZ data stream by a Gaussian low-pass filter, then modulate a VCO with modulation index (2 x frequency deviation/bit rate) set to 0.5. In other words, the single-sided frequency deviation is one fourth of the bit rate ($\Delta f = R/4$).

Another way of generating GMSK is by I/Q modulator. Referring back to EQ. 2, any RF signal can be split into I and Q components. Unlike the QPSK mentioned before, baseband I(t) and Q(t) are not discrete points for FM signals; rather, they are continuous functions of time. The way to produce FM is shown in Figure 6. We first store all the possible values of cos[ϕ(t)] and sin[ϕ(t)] in a ROM lookup table, which will be addressed by the incoming data to generate the I and Q samples. The output data from the ROM is then applied to D/A converters, after low-pass filtering for signal smoothing, to produce the analog baseband I and Q signals. This method guarantees the modulation index to be exactly 0.5, which is required for coherent detection of GMSK (e.g. GSM system). The same I/Q principle can also be applied to generating analog FM signals.

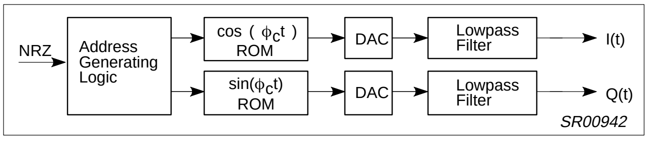


Figure 6. Digital FM (e.g. GMSK) Baseband I/Q Generator

Single sideband AM (SSB-AM)

AM signals can be divided into 3 types: the conventional AM, double sideband suppressed carrier AM (DSB-AM) and SSB-AM. The first type is not attractive because for 100% modulation, two-thirds of the transmit signal power appears in the carrier, which itself conveys no information at all. By using a balanced mixer (e.g. Gilbert cell), one can generate DSB-AM, where the carrier is totally suppressed and only the upper and lower sidebands are present. However, this is still not the best because the information is transmitted twice, once in each sideband. To further increase the efficiency of transmission, only one sideband is needed to deliver the information. The SSB-AM can be generated by an I/Q modulator with the baseband information feeding the modulator (by quadrature), as shown in Figure 7. This modulation technique can greatly reduce the bandwidth of the signal and allows more signals to be transmitted in a given frequency band. This topic is discussed in detail in Philips RF application note, #AN1981, "New low-power single sideband circuits".

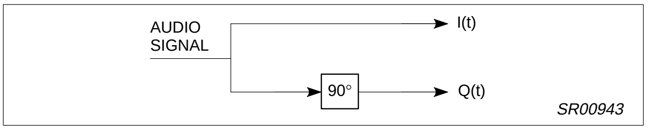


Figure 7. Baseband Processing for SSB-AM

SYSTEM ARCHITECTURE

There are usually two schemes, the dual conversion and direct conversion, used for implementing transmit modulators. Dual conversion is simpler to implement by modulating an oscillator at lower frequency and then up-converting to the carrier frequency. This scheme, however, is more expensive due to the need for additional filtering and more PC board space. By using only one mixer, direct conversion requires fewer components but is harder to implement.

The problems that direct conversion suffers are carrier leakage and modulated signal coupling. Poor RF isolation of the surface mount packages will allow the carrier to be present at the transmitter output thus making it difficult to have -40dBc carrier suppression. In addition to that, modulated RF signal would couple back to the oscillator (usually a VCO in a PLL synthesizer loop) and cause modulation distortion.

Based on the concept of dual conversion, the SA900 uses an image rejection mixer to eliminate the need for IF filtering and allow monolithic integration. The transmit carrier (LO) is down-converted by the frequency synthesized by the on-chip VCO, which operates from 90 to 140MHz. This LO is then modulated by the baseband I/Q signals to obtain a complex modulation scheme. The image (sum term) after mixing and LO is sufficiently suppressed by the image rejection mixer. Any residual amounts can be further suppressed by an external duplex filter.

Figures 8 and 9, respectively, show how the SA900 can be used in frequency division duplex (FDD) and time division duplex (TDD) transceivers. Notice that the LO for both systems is running at a frequency which is higher than the transmit frequency, thus minimizing carrier leakage. In the FDD system only one external VCO is required for generating both transmit and receive LO when using the SA900.

Figure 10 shows the IS-54 front-end chip set which consists of the SA601, SA7025, SA900, and SA637. This receiver architecture (SA637) supports a digital magnitude/phase baseband demodulator. An alternate configuration will be using the SA606 FM/IF receiver in conjunction with an external I/Q demodulator IC. The following table shows the possible configurations for the IS-54 handsets using the SA900 as transmitters.

Rx 1st IF	On-Chip VCO Frequency	On-Chip ÷N Value	Crystal Frequency
83.16MHz	128.16MHz	6	21.36MHz
71.64MHz	116.64MHz	6	19.44MHz
45MHz	90MHz	6	15MHz
84.6MHz	129.6MHz	9	14.4MHz

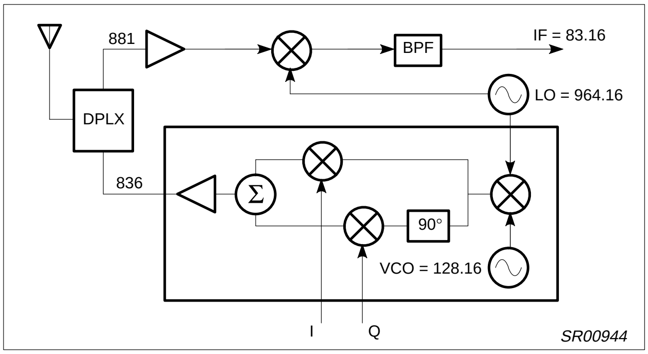


Figure 8. FDD System Using SA900

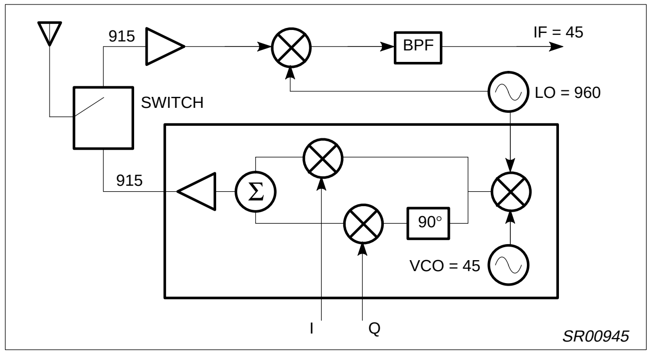


Figure 9. TDD System Using SA900

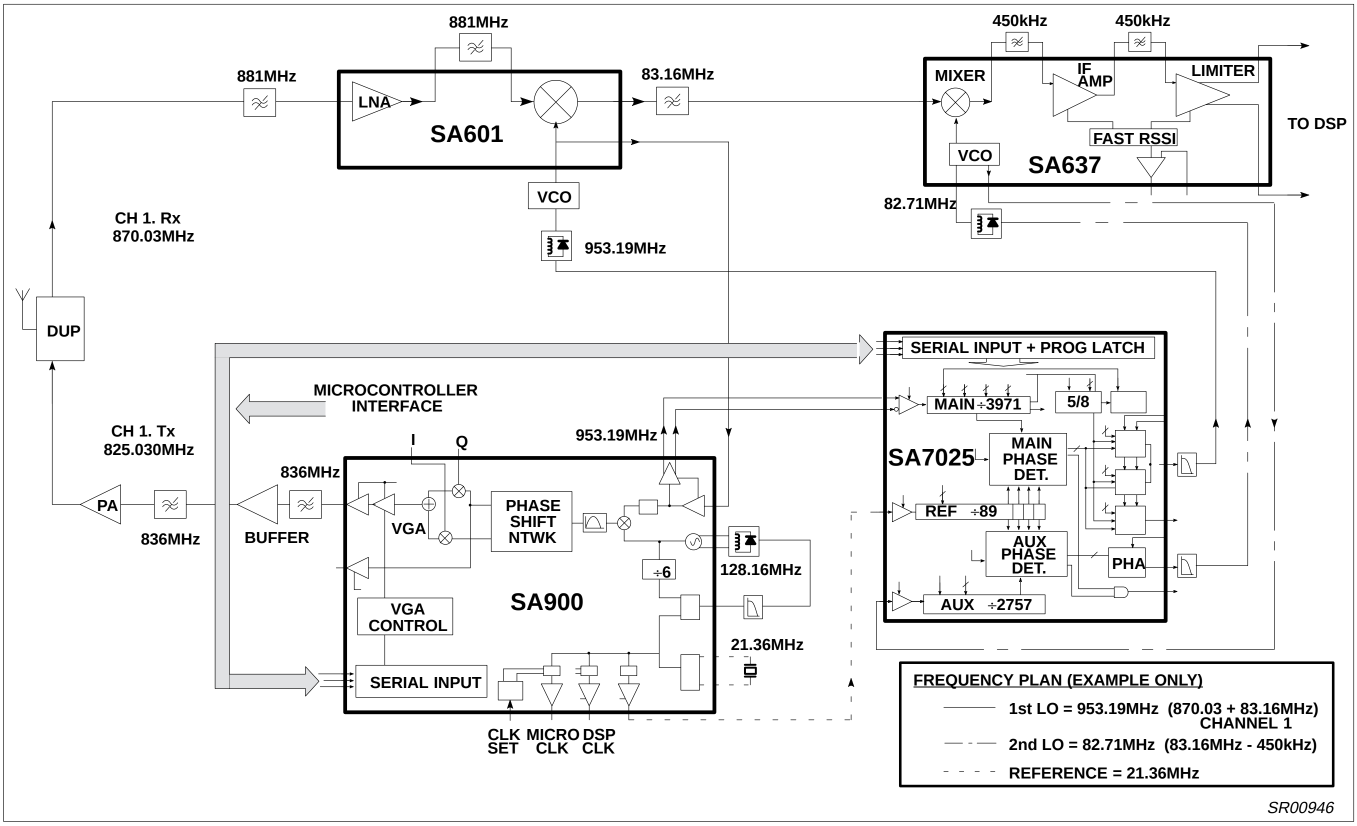


Figure 10. IS-54 front-end chip set from Philips

DESIGNING WITH THE SA900

Baseband I/Q Inputs

The baseband modulation inputs are designed to be driven differentially for the SA900 to operate at its best. The I and Q inputs should have a DC offset of V_{CC} /2, which is externally provided by common DSP chips. If all four inputs are biased from the same source, the device can tolerate ±0.5V_{DC} error; however, inaccuracy of DC bias between I1/I2 or Q1/Q2 causes reduced suppression of the carrier. Thus, it is important to have a well regulated DC supply for I and Q signal biasing. The bandwidth of the inputs is much higher than the specified 2MHz. Approximately 2dB of power loss will be experienced if the I and Q inputs are 50MHz.

The SA900 generates a minimum of 0dBm of power to a 50Ω load when the amplitude of the I and Q signals are 400mV_{P-P}. The output power will decrease by 6dB for every 50% decrease in I/Q amplitude. Single-ended I and Q sources can be used but are not recommended due to the degradation in carrier suppression (more than 10dB compared to differential). In addition, the entire noise performance of the device will suffer. V_{CC} /2 should be applied to I2 and Q2 pins if the part is driven single-endedly.

Transmit Local Oscillator

The transmit local oscillator path consists of a TXLO input buffer, LO output buffer, VCO, image rejection mixer and phase shift network. Together with a few external components, this section provides the I and Q carrier for modulation.

The TXLO inputs and LO outputs are designed to be used in an external PLL which synthesizes different frequencies for channel selection. The RF signal being generated is fed into TXLO inputs and then comes out of LO outputs to complete the system synthesizer loop. The TXLO inputs are differential in nature and have a VSWR of 2:1 with input impedance of 50Ω. Single-ended sources can be used by AC grounding the TXLO_2, as done on the demoboard. This signal should also be AC coupled into the TXLO_1. The frequency range for these inputs is from 900 to 1040MHz while the input power should be between -10 to -13dBm. The output level will be changed significantly if the input level is below -25dBm.

The output power of the LO buffered signal changes by about 2dB when the SA900 is in a different mode of operation. Typical values are -13.5dBm and -15.5dBm for DUAL mode and STANDBY mode, respectively.

The 90° phase shift network, realized by RC networks, is capable of operating over a wide frequency range. Even though their frequency characteristics are optimized for cellular band, the part can also be used in other applications in a different band. In such cases, designers have to test the part experimentally to find out the performance, such as sideband suppression, carrier suppression, and image rejection.

Crystal Oscillator

The crystal oscillator (XTAL_1 and XTAL_2 pins) is used to provide reference frequency between 10 and 45MHz for the phase detector and the three on-chip clocks. It can be configured as a crystal oscillator using external crystal and capacitors, or it can be driven by an external source. In the latter case, pin XTAL_2 can be left floating. Information regarding crystal oscillator design can be found in Philips RF application note, #AN 1982, "Apply the Oscillator of the NE602 in Low-Power Mixer Applications."

VCO

The VCO, together with the phase detector, the divider and external low-pass filter, can form a PLL for the transmit offset frequency. The image reject mixer down-converts the TXLO signal to the RF carrier by the amount of VCO frequency. Thus, the TXLO frequency should be the desired channel frequency plus the IF offset generated by the VCO. Notice that the part will not function if the VCO section is not used.

The VCO is designed for generating IF frequency between 90MHz and 140MHz. Together with an external varactor diode and resonator, it can be configured as an oscillator as shown in Figure 11.

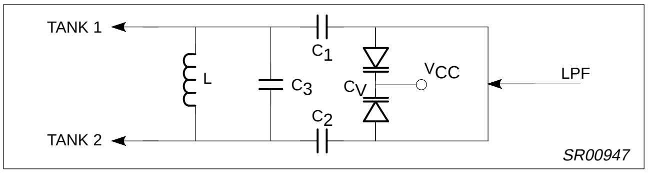


Figure 11. VCO Tank Circuit

The resonant frequency of such a circuit is

$$f_{VCO} = \frac{1}{2\pi\sqrt{LC_T}} \qquad \text{(EQ. 3)}$$

where $C_T = (C_1 // C_2 // C_V) + C_3$. C_V is a varactor diode of which capacitance changes linearly with the voltage across it.

Calculation:
$C_1 = C_2 = 33pF$
$C_3 = 5.6pF$
$C_V = 33.5pF$ @ 2.5V
$L = 100nH$

$C_T = 5.6 + (1/33 + 1/33 + 1/33.5)^{-1} = 16.7pF$

$$f_{VCO} = \frac{1}{2\pi\ (100e\text{-}9 \cdot 16.7e\text{-}12)^{0.5}} = 123MHz$$

On the demoboard, a 1:1 ratio RF transformer is also included to allow single-ended external source driving differential inputs when the VCO is not used.

When designing the VCO, careful PCB layout has to be made. Traces have to be short to avoid the parasitic capacitance and inductance which may cause unwanted oscillation. Referring to EQ. 3, there is a large combination of L and C_T values that will give the same resonant frequency. If undesired spurs are found in the design due to PCB layout, experimenting with a different set of LC values may sometimes solve the problem.

Output impedance matching

The equivalent output impedance at the DUALTX pin is approximately equal to 600Ω in parallel with 2pF at 830MHz. It has to be matched properly to generate maximum power into a 50Ω load (e.g. SAW filter). Figure 12 shows the recommended matching network. The shunt inductor (L1) is used to provide maximum swing at the output (short at DC) and also provide reactance to make the real impedance 50Ω looking into the matching network. The remaining negative reactance is canceled by the series inductor (L2). The values used on the demoboard can be used as a reference but may not be suitable if a different layout is implemented. The two shunt capacitors are included to bypass the high frequency RF signal, avoiding direct coupling into V_{CC}. The series AC coupling capacitor is used to maintain the proper bias for the output stage. Their values are big enough to be left out in impedance matching calculation.

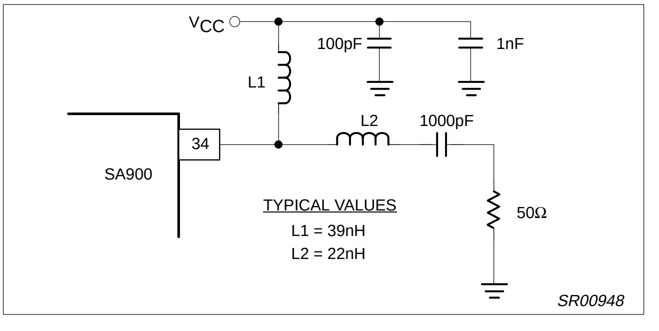
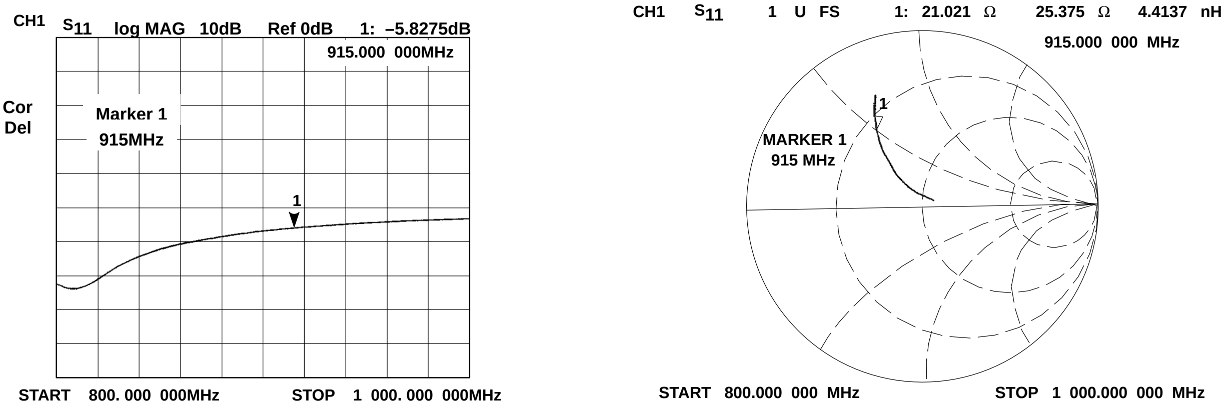
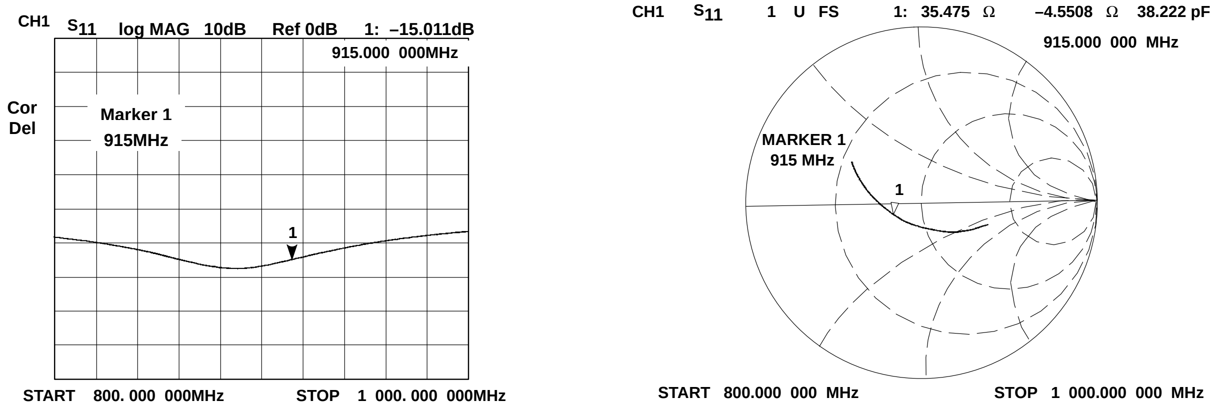


Figure 12. DUALTX Output Matching Network

Using a network analyzer to measure the S characteristic is necessary for obtaining optimum matching which generates maximum output power. Figure 13a-d shows how to match the output impedance to a 50Ω load at 915MHz. First, calibrate the network analyzer to the DUALTX SMA connector on the demoboard. Then, short the point where the series inductor is located and use the DELAY feature of the network analyzer to move the point of reference in the Smith Chart to the leftmost point. Now the network analyzer is calibrated to the beginning of the matching network, not just the SMA connector. The frequency response (Figure 13a) shows that the "dip" is around 830MHz, the frequency where the board was originally matched. The Smith Chart shows that it requires less inductance to bring the marker to the center of the chart (50Ω). By using a 15nH series inductor, the "dip" was moved closer to 915MHz (-15dB) and a better matching is achieved (Figure 13b).



a. S_{11} Response and Smith Chart when L2 = 22nH



b. S_{11} Response and Smith Chart when L2 = 15nH

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Figure 13.

On-Chip Clocks

The crystal oscillator is buffered to provide three external clock signals: CLK1, CLK2, and MCLK. Table 1 shows the divide ratio and the controlling mechanism:

Table 4.

CLK1	divide by 3	X (bit 18) = 1
	divide by 1	X (bit 18) = 0
CLK2	divide by 2	Y (bit 19) = 1
	divide by 1	Y (bit 19) = 0
MCLK	divide by 4	CLKSET pin = V_{CC}
	divide by 5	CLKSET pin = $V_{CC}/2$
	divide by 1	CLKSET pin grounded

CLK1 is usually used for the system synthesizer (e.g. SA7025) reference. Since MCLK is active all the time, it is ideal for providing the master clock for the microcontroller. When the device is in STANDBY mode, CLK1 and MCLK provide the clock signals necessary for receiving RF signals. CLK2 can also be used as a clock for digital signal processing (DSP) chip.

Modes of Operation

The SA900 is intended for either AMPS mode (analog cellular) or DUAL mode (digital cellular, IS-54) operation. When the device is running in AMPS mode, the I/Q modulator, variable gain amplifier (VGA) and phase shifter are disabled. The fixed gain amplifier is powered up during AMPS mode operation. However, since the divide ratio is too low (6, 7 or 8), the comparison frequency of the on-board PLL is too high, making it very difficult for the loop bandwidth to be less than 300Hz for analog FM modulation.

The device includes two power saving modes of operation which disable partial circuitry to reduce the power consumption of the overall chip. The SLEEP mode disables all the circuitry except the master clock (MCLK pin) of the SA900. The STANDBY mode shuts down everything except the TXLO buffer, MCLK, and CLK1, which allows the system synthesizer (e.g. SA7025) to continue running. These two power saving modes are common to both AMPS and DUAL mode operation. The SA900 draws 60mA in DUAL mode, reduced to 3mA and 8mA, respectively, in SLEEP and STANDBY modes.

TXEN pin is for hardware powering down the modulator and synthesizer. The falling edge of the signal disables the modulator and synthesizer while the rising edge enables the modulator. To power down the synthesizer using software, send a data word with SE bit set to ‘0’ (‘1’ for enable). The synthesizer will be disabled right after the strobe signal is transmitted. Either SE or TXEN going low will turn off the synthesizer. This operation is common to both AMPS and DUAL mode.

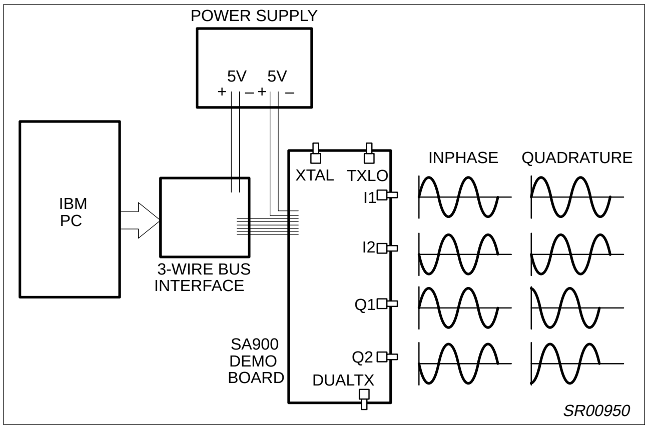


Figure 14. In-phase and Quadrature Modulation Test

PERFORMANCE OF THE SA900

Performance Criteria

Since the I/Q modulator is a universal transmitter, measuring only the frequency stability and modulation index of a generated FM signal would not be useful for other modulation schemes. Measurement parameters should be general enough so that they can represent the performance of modulators when applying different types of modulation and allow fair comparisons among different I/Q modulators. Based on this idea, two measurement techniques, in-phase modulation and quadrature modulation, are used for evaluating I/Q modulators.

The in-phase modulation relies on injecting two equal frequencies and phase signals at f_{mod} into the I and Q inputs. The result of this modulation is two sidebands appearing at f_{mod} offset from the carrier, with the carrier totally suppressed. This is also known as double-sideband (DSB) conversion. The quadrature modulation requires two equal frequencies (but 90° out-of-phase signals) being injected into the I and Q inputs. The result is a single-sideband suppressed carrier (SSB-SC) signal with either the upper or lower sideband at f_{mod} carrier offset being suppressed. This is also known as single-sideband (SSB) up-conversion. Figure 14 summarizes these two tests.

In a practical system, imperfection of an I/Q modulator is directly related to these two measurements. Sideband and carrier suppression from the quadrature modulation test will show the amount of gain imbalance, phase imbalance, and DC offset. On the other hand, intermodulation product suppression from the in-phase modulation test will show the linearity of an I/Q modulator. When making measurements, it is important to have well-balanced I and Q baseband modulating signals for measurement since the signal imperfection will translate into degradation in sideband and carrier suppression.

Performance Graphs

In making those measurements for the demoboard, the following parameters were used:

In-phase modulation:

PIN 43 I1=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=0°

PIN 42 I2=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=180°

PIN 41 Q1=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=0°

PIN 40 Q2=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=180°

Quadrature modulation:

PIN 43 I1=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=0°

PIN 42 I2=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=180°

PIN 41 Q1=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=90°

PIN 40 Q2=400mV$_{P-P}$, DC=V$_{CC}$/2 at 200kHz, Phase=270°

Figures 15a and 15b illustrate what the typical output spectrum would be if in-phase and quadrature modulation were applied to an I/Q modulator. Quadrature modulation will produce lower sideband (LSB) or upper sideband (USB) signal, depending on the phase angle between the I and Q signals. The SA900 was designed to have USB suppressed when the I signal is leading the Q signal. The undesired signals are carrier breakthrough and the harmonic products of the baseband modulating signals sitting at

$f_c \pm nf_{mod}$, where n is an integer ≥2.

Referring to Figure 15a, the output power is 1.3dBm (cable loss = 0.7dB) for the LSB while better than -38dBc of carrier, sideband, and harmonics suppression is measured. The USB better than -26dBc implies the residual AM of the transmit signal is better than 5%, a requirement of the IS-54 specification.

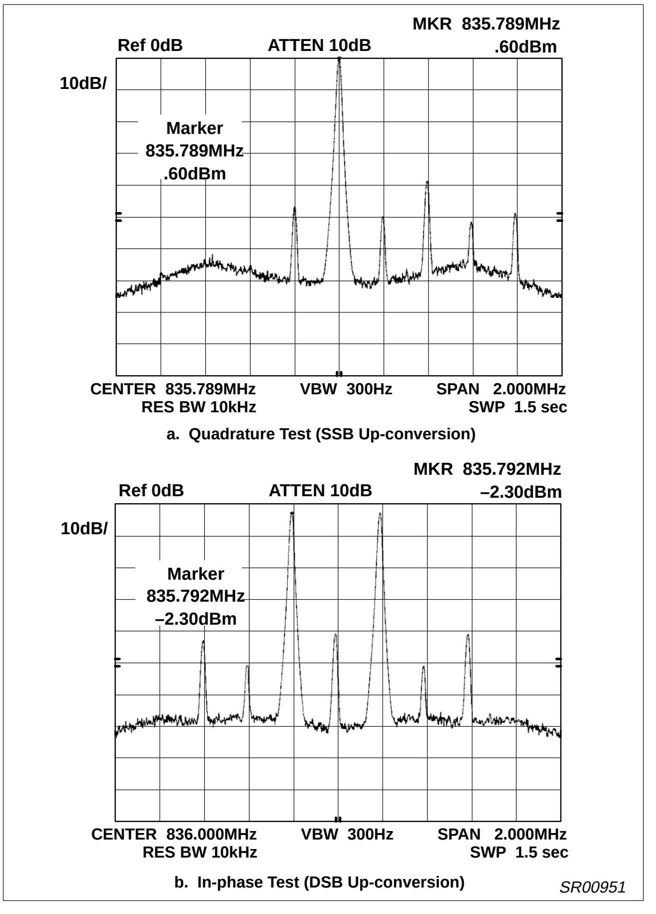


Figure 15.

In-phase modulation test will generate both LSB and USB. Beside these two tones, the carrier breakthrough and the harmonics, intermodulation (IM) products will all appear at the output. The odd IM products are dominant, and they satisfy the following rules:

Let $f_1 = f_c - f_{mod}$, $f_2 = f_c + f_{mod}$

3rd order IM:	$2f_1 - f_2 = f_c - 3f_{mod}$,	$2f_2 - f_1 = f_c + 3f_{mod}$
5th order IM:	$3f_1 - 2f_2 = f_c - 5f_{mod}$,	$3f_2 - 2f_1 = f_c + 5f_{mod}$
7th order IM:	$4f_1 - 3f_2 = f_c - 7f_{mod}$,	$4f_2 - 3f_1 = f_c + 7f_{mod}$

Referring to Figure 15b, both LSB and USB are -1.6dBm (cable loss = 0.7dB) in power, which is 3dB less than the measured power for the quadrature modulation test. The IM3 is better than -35dBc. Much higher order IM products are totally suppressed.

Amplitude and phase unbalance

Both amplitude and phase unbalance (error) of an I/Q modulator can be calculated directly from the SSB performance plots. Assume phase error equals ϕ radian and amplitude error equals K, the sideband suppression, X, in dBc can be expressed as follows (see APPENDIX for derivation):

$$\text{SSB suppression, } X(\text{dBc}) = 10\log\left(\frac{K^2 + 2\cdot K\cdot\cos(\phi) + 1}{K^2 - 2\cdot K\cdot\cos(\phi) + 1}\right) \qquad \text{(EQ. 4)}$$

Collecting the like terms and express ϕ in terms of K and X, it becomes:

$$\phi = \cos^{-1}\left(\frac{10^{x/10}\cdot K^2 + 10^{x/10} - 1 - K^2}{2\cdot K + 2\cdot K\cdot 10^{x/10}}\right) \qquad \text{(EQ. 5)}$$

For a given X, there will be a set of ϕ and K that satisfies EQ. 5. We can represent this relationship graphically, as shown in Figure 16. The contours show the phase and amplitude errors for SSB suppression, X, from -44 to -26dBc. When X equals -40dBc, phase error is less than 1.2° with a 0dB amplitude error. By the same token, the amplitude error is less than 0.2dB with a 0° phase error.

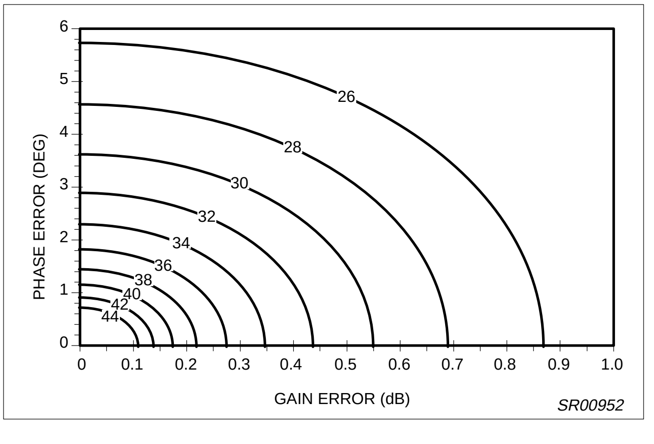
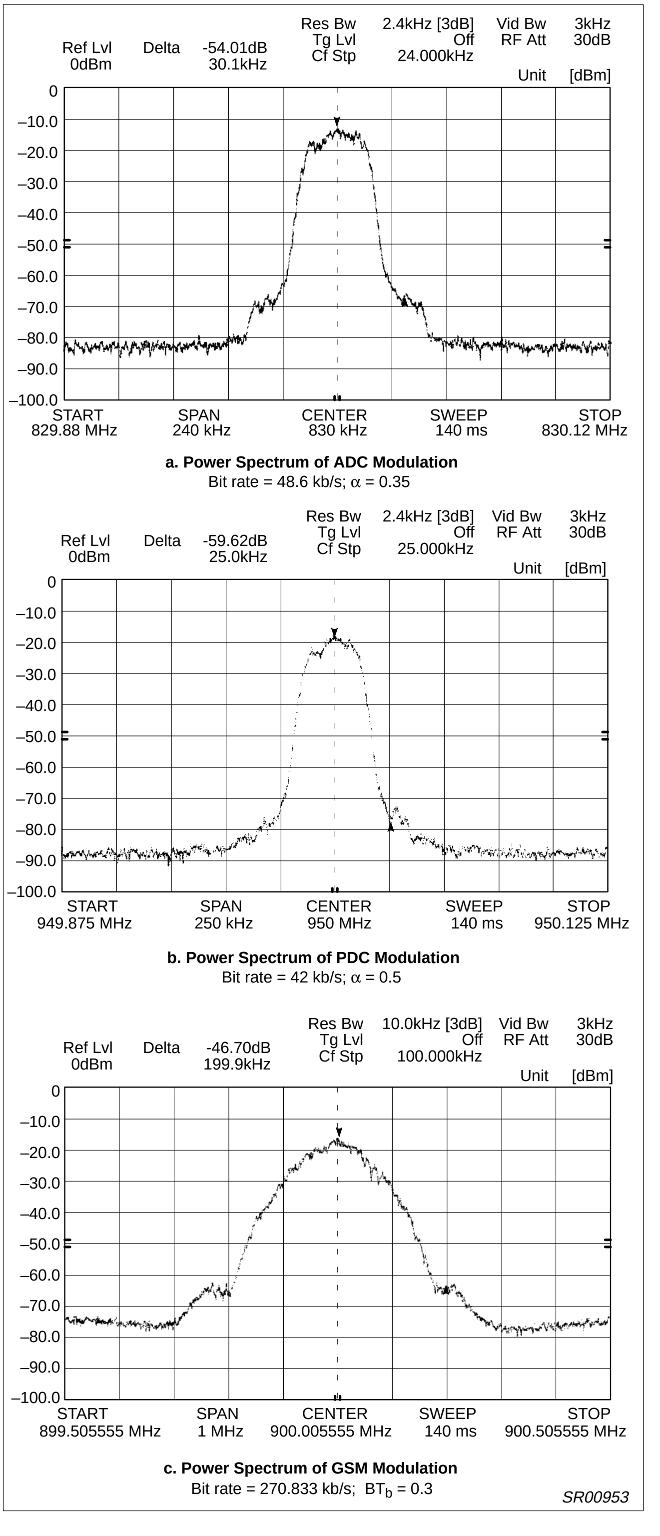


Figure 16. SSB Suppression Contours



a. Power Spectrum of ADC Modulation
Bit rate = 48.6 kb/s; α = 0.35

b. Power Spectrum of PDC Modulation
Bit rate = 42 kb/s; α = 0.5

c. Power Spectrum of GSM Modulation
Bit rate = 270.833 kb/s; $BT_b = 0.3$

Figure 17.

Spectral mask

To fully characterize the performance of an I/Q modulator, measurements of the power spectral density of various digital modulation schemes have to be made. Figures 17a and 17b show the measured spectral masks of IS-54 and PDC standards, which designate $\pi/4$-DQPSK as the modulation format.

GMSK is a digital modulation scheme widely used for wireless and mobile communications. Figure 17c shows the spectral mask of the modulation format required by GSM, the digital cellular standard in Europe. At 200kHz and 300kHz carrier offset, the power of the signal is suppressed by 46dB and 58dB, respectively, which is well within the GSM specification.

Power ON time

The power ON time for the SA900 is mainly determined by the loop bandwidth of the on-board PLL frequency synthesizer. It can be measured by using the HP 53310A Modulation Domain Analyzer set to the EXTERNAL TRIGGERED mode. The STROBE signal from 3-wire bus is used to trigger the equipment. Figure 18 shows that the part can be powered up and locked in about 62µs.

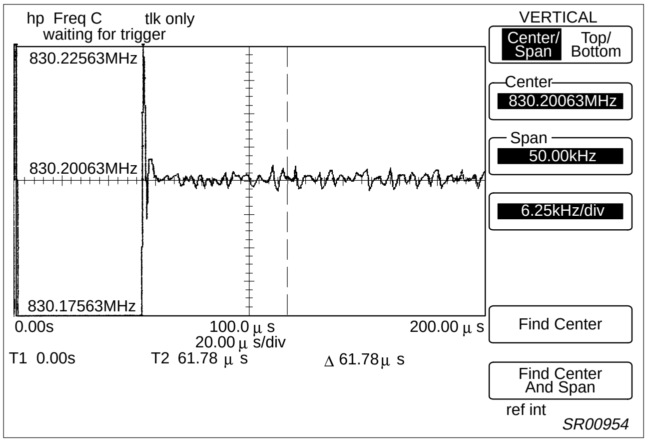


Figure 18. Power ON Time Measurement

ISM band application

The FCC has recently assigned three bands for ISM type of application. The one below 1GHz is from 902 to 928MHz. This band becomes very attractive because users are allowed, without having a license, to transmit up to 1 Watt of power when frequency hopping or direct sequence CDMA is used. The wide bandwidth nature of the SA900 fits well into this application. Figures 19a and 19b are the output spectrum of the SA900 showing how well the image reject mixer works. A common IF (45MHz) was chosen to be the offset frequency, and then injected externally into the VCO pins. The closest images are sitting at 45MHz apart and are better than -36dBc.

COMPONENTS FUNCTION

C241, C242, C243, C245, C246 - Supply bypassing capacitors

C247 - provides AC ground for TXLO2 pin

C249 - AC couples an external signal into TXLO1 pin

C253, C254, C255 - part of the LC tank circuit

C263 - AC couples an external signal into XTAL1 pin

C267, C269 - part of the PLL low-pass filter

C281, C287, C288 - AC coupling capacitors for the clocks

C301 - AC coupling capacitor for the DUALTX pin

C305, C306 - Bypass RF signal coming from DUALTX pin

C312, C313, C314 - AC coupling capacitors

C370, C371 - AC couples an external signal into TANK1 and TANK2 pins when on-board PLL is not used

L252 - part of the LC tank circuit

L304, L372 - matching network for the DUALTX output

R260 - termination resistor

R262 - current setting resistor for the charge pump

R264, R266 - part of the PLL low-pass filter

R274 - jumper

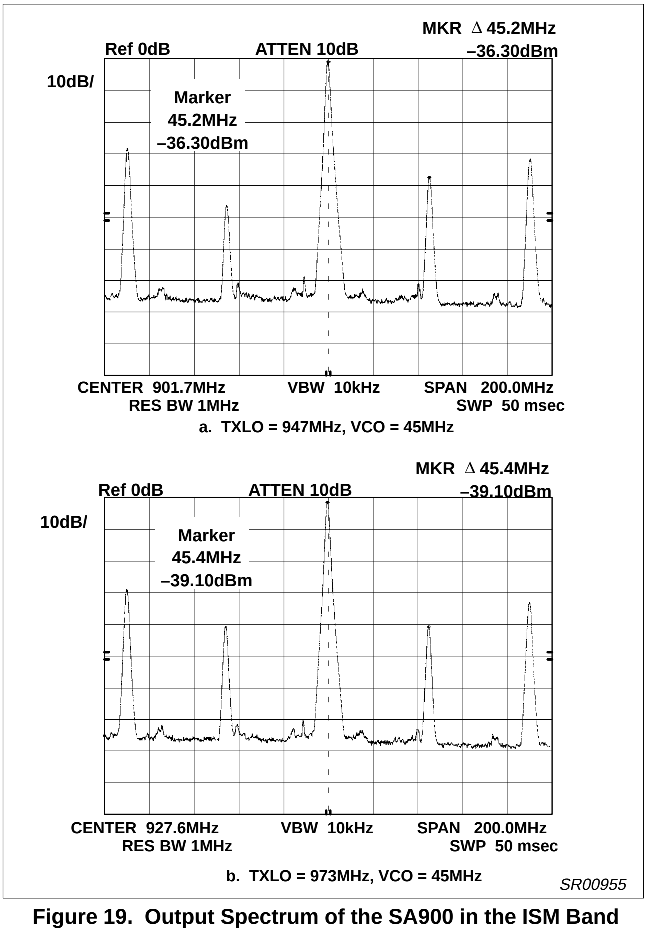


Figure 19. Output Spectrum of the SA900 in the ISM Band

R349, R350 - make up a voltage divider for selecting divide ratio of the MCLK

R352 - termination resistor

R333, 358 - isolation resistors between the LC tank and the PLL low-pass filter

FREQUENTLY ASKED QUESTIONS

Q. What is the bandwidth of the phase shifter for generating I/Q carriers?

A. The bandwidth is between 820 and 920MHz. The part is still functional below 820MHz and above 920MHz, but the carrier and sideband suppression are not guaranteed. In addition, the DUALTX output matching network needs to be optimized for a different frequency.

Q. Can I frequency modulate (FM) the on-board VCO to generate RF signal for AMPS system?

A. Since the divide ratio for the VCO is too low, it is very difficult to obtain the required loop bandwidth (<300Hz) to do AMPS modulation.

Q. What signals constitute the spurious output referred to under the DUALTX function of the AC electrical characteristics in the data sheet?

A. Those spurs could be N*TXLO, N*VCO, TXLO+VCO, N*XO, and TXLO ±N*VCO.

Q. Can external circuitry be added or modified to reduce the broadband noise floor below -136dBm/Hz?
A. Customers can put a bandpass SAW filter at the output of the TXLO to improve the broadband noise floor.

Q. Can the SA900 generate BPSK signal?
A. Yes, it can. Feed the baseband signal into I1 and I2 and leave Q1 and Q2 open or tie them to V_{CC} /2.

Q. What is the response of the image rejection filter?
A. It is actually a SSB mixer; not an image rejection filter.

Q. What happens if the VCO is not used?
A. There will not be any signal at the DUALTX and AMPS output if the VCO is not used.

REFERENCES:

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"Digital and Analog Communications Systems," Leon W. Couch II, Macmillan, 1990.

"Cellular System Dual-Mode Mobile Station-Base Station Compatibility Standard", IS-54-B, EIA/TIA, April 1992.

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"π/4-QPSK MODEMS for Satellite Sound/Data Broadcast Systems", Chia-Liang Liu and Kamilo Feher, IEEE Transactions on Broadcasting, March 1991, pp. 1-8.

"PCD5070 GSM Baseband Interface", Preliminary specification, Philips Semiconductors, September, 1992.

APPENDIX

Assume an imperfect I/Q modulator with gain error, K, and phase error, ϕ, modulated by quadrature I/Q signals (SSB up-conversion) wm. Then the signal, s(t), at the output of the I/Q modulator becomes,

$$s(t) = K\cos(\omega_c t + \phi)\cos(\omega_m t) - \sin(\omega_c t)\cos(\omega_m t + 90^o) \quad \text{(EQ. A.1)}$$

Using trigonometric identity and let $\omega c - \omega_m = A$ and $\omega_c + \omega_m = B$, we obtain,

$$s(t) = \frac{K}{2}\cos[At + \phi] + \frac{K}{2}\cos[Bt + \phi] + \frac{1}{2}\cos[At] - \frac{1}{2}\cos[Bt] \quad \text{(EQ. A.2)}$$

Assume the information is in LSB, i.e. A, and the spur is the USB, i.e., B, we have,

$$\text{Signal} = \frac{K}{2}\cos A\cos\phi + \frac{1}{2}\cos A - \frac{K}{2}\sin A\sin\phi \quad \text{(EQ. A.3)}$$

$$\text{Noise} = \frac{K}{2}\cos B\cos\phi + \frac{1}{2}\cos B - \frac{K}{2}\sin B\sin\phi \quad \text{(EQ. A.4)}$$

To find the power, we have to evaluate the envelope (amplitude) of these two signals. Recall that for any given bandpass signal in rectangular form,

Bandpass signal = $X\cos\omega t - Y\sin\omega t$,

the envelope is

Envelope = $(X^2 + Y^2)^{0.5}$

Therefore, from EQ. A.3 and A.4,

$$\text{Signal} = \left[\left(\frac{K}{2}\cos\phi + \frac{1}{2}\right)^2 + \left(\frac{K}{2}\sin\phi\right)^2\right]^{0.5} \quad \text{(EQ. A.5)}$$

$$\text{Noise} = \left[\left(\frac{K}{2}\cos\phi - \frac{1}{2}\right)^2 + \left(\frac{K}{2}\sin\phi\right)^2\right]^{0.5} \quad \text{(EQ. A.6)}$$

Finally, the S/N ratio can be found by taking 20 log the ratio of EQ. A.5 and A.6.

$$\frac{S}{N} = 10\log\left(\frac{K^2 + 2K\cos\phi + 1}{K^2 - 2K\cos\phi + 1}\right)$$

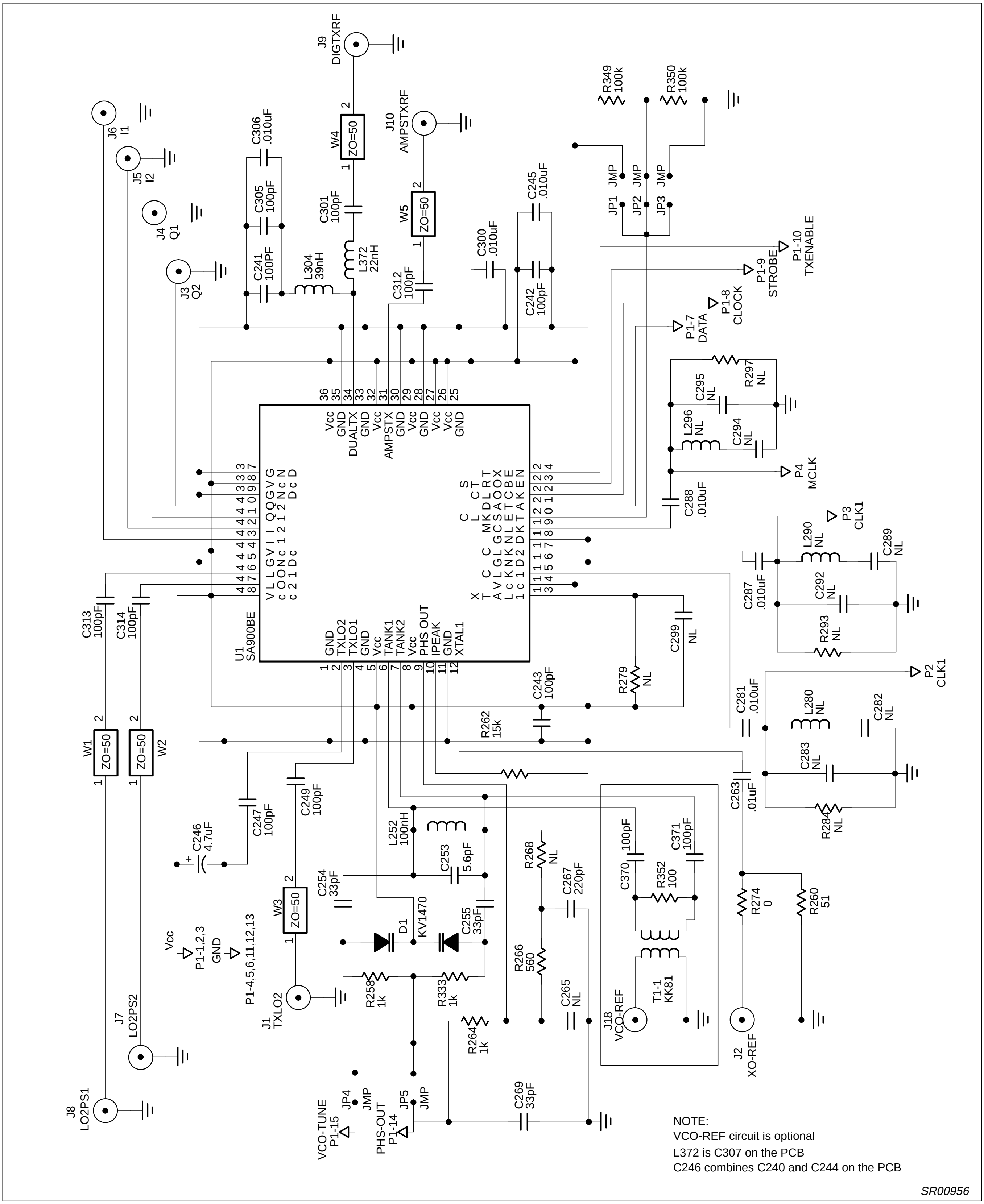
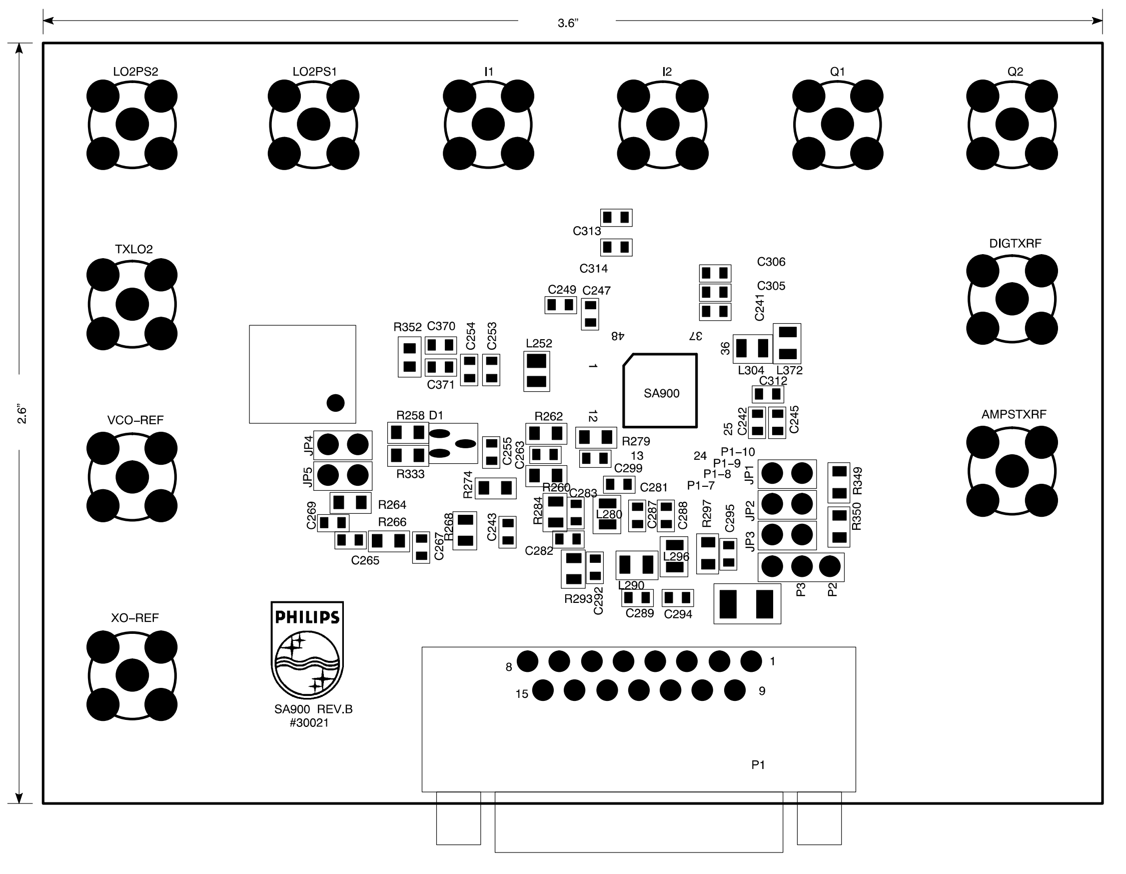


Figure 20. SA900 Demoboard



SR00957

Figure 21. SA900 Board Layout

Table 5. Customer Application Component List for SA900BE

Qty.	Part Value	Volt	Part Reference	Part Description	Vendor	Mfg	Part Number
Surface Mount Capacitors							
1	5.6pF	50V	C253	Cap. cer. 0603 NPO ±0.25pF	Garrett	Rohm	MCH185A5R6CK
3	33pF	50V	C254, C255, C269	Cap. cer. 0603 NPO ±5%	Garrett	Rohm	MCH185A330JK
13	100pF	50V	C241, C242, C243, C247, C249, C300, C301, C305, C312, C313, C314, C370, C371	Cap. cer. 0603 NPO ±5%	Garrett	Rohm	MCH185A101JK
1	1000pF	50V	C301	Cap. cer. 0603 X7R ±10%	Garrett	Rohm	MCH185C102KK
1	2200pF	50V	C267	Cap. cer. 0603 X7R ±10%	Garrett	Rohm	MCH185C222KK
6	0.01µF	25V	C245, C263, C281, C287, C288, C306	Cap. cer. 0603 X7R ±10%	Garrett	Philips	MCH182C103KK
1	4.7µF	10V	C246	Tant. chip cap. B 3528 ±10%	Garrett	Philips	49MC475B010KOAS
8	NL		C265, C282, C283, C289, C292, C294, C295, C299				
Surface Mount Resistors							
1	0Ω		R274	Res. chip 0603 1/16W ±5%	Garrett	Rohm	MCR03JW000E
1	51Ω		R260	Res. chip 0603 1/16W ±5%	Garrett	Rohm	MCR03JW510E
1	100Ω		R352	Res. chip 0603 1/16W ±5%	Garrett	Rohm	MCR03JW101E
1	560Ω		R266	Res. chip 0603 1/16W ±5%	Garrett	Rohm	MCR03JW561E
1	1KΩ		R258, R264, R333	Res. chip 0603 1/16W ±5%	Garrett	Rohm	MCR03JW102E
1	15KΩ		R262	Res. chip 0603 1/16W ±5%	Garrett	Rohm	MCR03JW153E
2	100KΩ		R349, R350	Res. chip 0603 1/16W ±5%	Garrett	Rohm	MCR03JW104E
Surface Mount Diodes							
1			D1	SMD Diode (Varactor)	Digikey	TOKO	KV1470TR00
Surface Mount Inductors							
1	0.022µH		L372	Inductor SM Mold/WW A	Garrett	J.W. Miller	PM20-R022M
1	0.039µH		L304	Inductor SM Mold/WW A	Garrett	J.W. Miller	PM20-R039M
1	0.10µH		L252	Inductor SM Mold/WW A	Garrett	J.W. Miller	PM20-R10M
Surface Mount Integrated Circuits							
1			U1	I/Q Transmit modulator	Philips	Philips	SA900BE
Miscellaneous							
1			K353	RF Transformer	Mini-Circuits	Mini-Circuits	T1-1 KK81
11			J1, J2, J3, J4, J5, J6, J7, J8 ,J9, J10, J18	SMA Right Angle Jack Receptacle	Newark	EF Johnson	142-0701-301
5			JP1, JP2, JP3, JP4, JP5	straight, dual row	Newark	IPI	929836-01-36-ND
1			P1	15 pins receptacle D-sub. conn.	Newark	Dupont	51F2456
1				Printed circuit board	Philips	Philips	SA900-30021
66 Total Parts							