

Introduction



The PRISM1KIT-EVAL wireless LAN PC card is a complete wireless high speed modem utilizing the Harris PRISM™ Direct Sequence Spread Spectrum (DSSS) wireless transceiver chip set.

The card is packaged in an open frame PCMCIA Type II extended coverset for evaluation access and a connector output for use in the laboratory. Evaluation kits include two cards, Microsoft® Windows® 95 software and documentation to get your WLAN evaluation started quickly. The PRISM1KIT-EVAL is not FCC approved as an intentional radiator and is intended for use with cabled connections only (30dB antenna port to antenna port attenuation recommended). An FCC experimental license is required while transmitting over the air with unapproved equipment. Please refer to the WLANKITPR1-KIT for an FCC-approved radio kit for wireless LAN evaluations. This application note details the RF and analog design of these cards. The physical layer (PHY) sections of these PC Cards are described in detail. The medium access control (MAC) section of the PC Cards will be described in detail in the pending AMD application note titled "Wireless LAN DSSS PC Card Reference Design" [1].

Figure 1 shows a block diagram of the reference radio design. This radio has been designed to conform to the draft IEEE 802.11 specification but does not include the antenna diversity selection.

The specifications of the PC Card Wireless LAN are as follows:

Ordering Information

PART NUMBER	DESCRIPTION	CARDS PER KIT
PRISM1KIT-EVAL	Evaluation Kit	2

General Specifications

- Targeted Standard IEEE 802.11 (Draft)
- Data Rate 1Mbps DBPSK
2Mbps DQPSK
- Range 400ft Indoor (Typ) (Note 1)
3700ft Outdoor (Typ) (Note 1)
- Frequency Range 2412MHz to 2484MHz
- Step Size 1MHz
- IF Frequency 280MHz
- IF Bandwidth 17MHz
- RX/TX Switching Speed 2μs (Typ)
- Operating Voltage 4.5V_{DC} - 5.5V_{DC}

- Standby Current 190mA at 1μs Recovery (Note 4)
70mA at 25μs Recovery (Note 4)
60mA at 2ms Recovery (Note 4)
30mA at 15ms Recovery (Note 4)
- Operating Temperature Range 0°C to 70°C (Note 2)
- Storage Temperature Range -55°C to 125°C
- Mechanical . Type II PCMCIA Card, with Antenna Extension
- Antenna Interface SMA, 50Ω

Receive Specifications

- Sensitivity -91dBm (Typ), 1Mbps, 8E-2 FER (Note 3)
-88dBm (Typ), 2Mbps, 8E-2 FER (Note 3)
- Input Third Order Intercept Point -20dBm (Typ)
- Image Rejection 65dB (Typ)
- IF Rejection 80dB (Typ)
- Adjacent Channel Rejection 63dB (Typ)
at 25MHz Offset
- Supply Current 287mA (Typ), 2Mbps,
100% Duty Cycle

Transmit Specifications

- Output Power +17.5dBm (Typ)
- Transmit Spectral Mask. . . . -32dBc (Typ) at First Side-Lobe
- Supply Current 488mA (Typ), 2Mbps, 100% Duty Cycle

NOTES:

- Using M/A-COM AND-C-107 omnidirectional antenna.
- AM79C930 limited to 0°C to 70°C.
- FER = Frame Error Rate or Packet Error Rate.
- Recovery times do not include MAC recovery.

Receive Processing

Referring to the block diagram in Figure 1, the schematic on our web site [2], and the bill of materials in Appendix A, a single antenna is used. Up to two antennas are supported in the HFA3824A [3] Baseband Processor to implement diversity, countering the adverse effects of multipath fading. As space is at a premium in a PC Card environment, only one antenna is used. In an actual system implementation, if one can achieve diversity in at least one end of a link, such as at the access point where it is possible to achieve physical separation between diversity antennas, multipath performance will be improved.

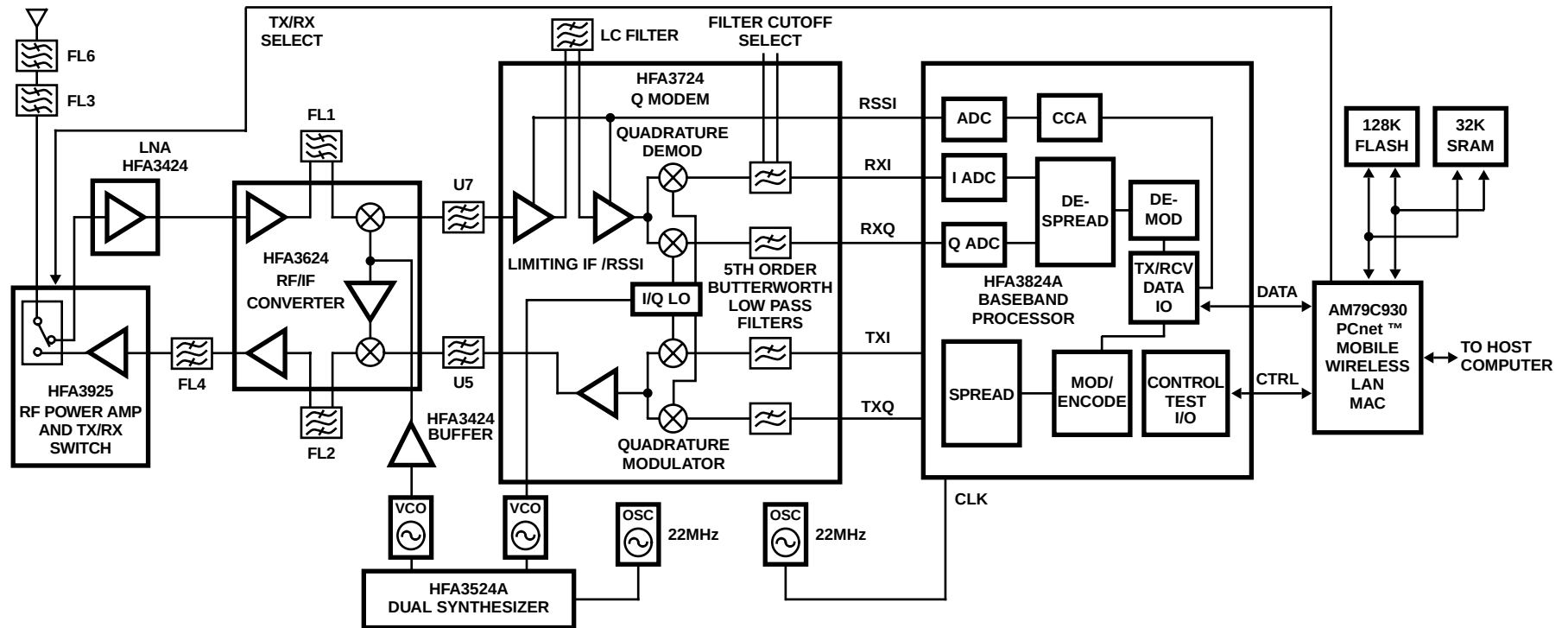


FIGURE 1. PRISM PC CARD BLOCK DIAGRAM

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From the antenna, the received input is applied to FL3 and FL6, a Toko TDF2A-2450T-10 two pole dielectric bandpass filters, which are used to provide image rejection for the receiver. The IF frequency is 280MHz, and low-side injection is used, thereby placing the received image 560MHz below the tuned channel. FL3 and FL6 also provide protection for the RF front-end from out of band interfering signals.

The T/R switch is integrated in the HFA3925 [4] RF Power Amplifier (RFPA). The HFA3925 RFPA operates from the unregulated 5V PC Card supply.

Following the T/R switch, the HFA3424 [5] Low Noise Amplifier (LNA) is used to set the receiver noise figure. The HFA3424 LNA operates from a regulated 3.5V supply. A logic-level PMOS switch, RF1K49093 [9], is used to control the drain supply voltage to the HFA3424 LNA, and implement a power down mode when transmitting.

A trade-off between noise figure and input intercept point exists in any receiver, to balance these conflicting requirements in the PRISM radio, an attenuator follows the HFA3424 LNA, and it may be mounted. To improve input intercept point, this attenuation may be increased. The cascaded front-end noise figure, input intercept point, and gain distribution analysis are shown in Table 1.

Next, the signal enters the HFA3624 [6] RF/IF Converter LNA section, which aids in setting receiver NF. FL1 is used to suppress image noise generated in both the HFA3424 LNA and the HFA3624 LNA, and is a Murata LFJ30-03B2442B084 two pole monolithic LC bandpass filter. Only modest attenuation at the image frequency is required. The insertion loss is not critical, since at this point in the receiver, component loss or NF is offset by the preceding gain stages. All sections of the HFA3624 RF/IF Converter operate from a regulated 3.5V supply.

Down-conversion from the 2.4GHz - 2.5GHz band is performed in the HFA3624 RF/IF Converter mixer section. As previously mentioned, the IF center frequency is 280MHz, and low-side local oscillator (LO) injection is used. A discrete LC matching network is used at the mixer output to differentially combine the IF outputs, as well as

match impedance to a 50Ω environment. A trimmer capacitor is used as part of the narrow-band matching network. An alternative, broadband matching network is described in the HFA3624 RF/IF Converter application note, and does not require any tunable elements. A direct impedance match to the IF filter, U7 could be implemented if desired. The 50Ω environment was chosen to allow ease in measurement of portions of the radio with external test equipment. An analysis of the mixer spurious responses is shown in Appendix B. There are no crossing spurious responses, therefore only tuned responses are shown.

The IF receive filter, U7, is a Toyocom TQS-432 SAW bandpass filter. The center frequency is 280MHz, the 3dB bandwidth is 17MHz, and the differential group delay is less than 100ns. Insertion loss is typically 6dB, making it ideal for single-conversion systems. The impedance of the SAW is 270Ω, in parallel with 5pF, and a series 33nH inductor is used to match the filter input to 50Ω. The SAW output is matched directly to the IF input of the HFA3724 Quadrature IF Modulator/Demodulator, using a shunt 33nH inductor and 261Ω resistor. This presents a 250Ω source impedance to the limiter input, thereby optimizing the limiter's NF. Measured performance of the SAW filter is shown in Appendix C.

In the receive mode, the HFA3724 Quadrature IF Modulator/Demodulator provides two limiting amplifiers, a quadrature baseband demodulator, and two baseband low pass filters. All sections of the HFA3724 operate from a regulated 3.5V supply. The first limiting amplifier establishes the NF of the IF strip at approximately 7dB. A discrete one pole LC differential filter is placed between the two limiters to restrict the noise bandwidth of the first limiter. As both limiters exhibit a broadband response, with over 400MHz bandwidth, a noise bandwidth reduction filter is appropriate to ensure that the second limiter is fully limiting on the front-end noise within the signal bandwidth, as opposed to the broadband noise generated by the first limiter. This filter has a center frequency of 280MHz, and a 3dB bandwidth of 50MHz. It consists of a fixed 10nH inductor and a fixed 20pF capacitor, as described in the HFA3724 data sheet.

TABLE 1. PRISM CASCADED FRONT-END ANALYSIS

STAGE	G	GC	F	FC	IP30	IP30C	IP3IC
FL6 RF FILTER	-2.0	-2.0	2.0	2.0	100.0	100.0	102.0
FL3 RF FILTER	-2.0	-4.0	2.0	4.0	100.0	95.9	99.9
HFA3925 T/R SWITCH	-1.2	-5.2	1.2	5.2	34.0	34.0	39.2
HFA3424 LNA	13.0	7.8	2.0	7.2	11.1	11.0	3.3
HFA3624 LNA	15.6	23.4	3.8	7.4	15.0	14.7	-8.7
FL1 RF FILTER	-3.0	20.4	3.0	7.4	100.0	11.7	-8.7
HFA3624 MIXER	3.0	23.4	12.0	7.5	4.0	3.6	-19.8
U7 IF FILTER	-10.0	13.4	10.0	7.5	100.0	-6.4	-19.8
HFA3724 [7] IF STRIP	0.0	13.4	7.0	7.7	100.0	-6.4	-19.8

Cascaded Gain = 13.4dB Cascaded NF = 7.7dB Cascaded Input IP3 = -19.8dBm

NOTE: G (individual stage gain, dB), GC (cumulative gain, dB), F (NF, dB), FC (cumulative NF, dB), IP30 (individual stage output IP3, dBm), IP30C (cumulative output IP3, dBm), IP3IC (cumulative input IP3, dBm).

The gain distribution/limiter noise analysis is shown in Appendix F. If the alternative HFA3624 broadband matching network is used, the HFA3624 mixer conversion gain will be higher, which will help ensure that the second limiter is fully limited on front-end noise.

At the output of the limiters, a 200mV_{P-P} differential signal level is maintained under all input conditions. This limited signal is then mixed in quadrature to baseband in the HFA3724 Quadrature IF Modulator/Demodulator. The LO needed for the quadrature mixing is applied at twice the IF frequency, or 560MHz. A divide by two circuit then provides an accurate quadrature LO for the mixers. The baseband outputs of the quadrature mixers are AC coupled off-chip to the integrated fifth order Butterworth filters. The output levels of the low pass filters are nominally 500mV_{P-P} single-ended, and are intended to be AC coupled to the HFA3824A Baseband Processor. The AC coupling time constant is approximately 25 times longer than the symbol period, and is implemented with 0.01μF series capacitors. These coupling capacitors must be taken into account, however, when estimating the time it takes to power up or awaken from sleep mode.

At the input to the HFA3824A Baseband Processor, the quadrature signals are analog to digital converted in wideband 3 bit converters. The sample rate is 22MSPS, which results in two samples per chip. A 22MHz Fox F4106 crystal oscillator is used to provide the main clock for the HFA3824A. The signals are spread spectrum with no DC term, so it is feasible to AC couple the signals to the ADCs and avoid DC bias offsets. The signal at this point has been limited to a constant IF amplitude and then passed through two separate mixer and low pass filter paths. The component variations in these two paths can introduce offsets in amplitude and phase and can also use up some of the headroom in the ADCs. The maximum amplitude variation is 2dB and the maximum phase balance variation is 4 degrees. Since the signal is limited, the IF signals will have low peak to average ratios even with noise as an input. The I and Q signals will have sinusoidal properties with PSK modulation imposed. It is their combined vector magnitude that is limited, not their individual amplitudes. To optimize the demodulator's performance, the ADCs are operated at the point where they are at full scale on either I or Q one third of the time. To maintain this operating point in the face of component variations, there is an optional active adjustment of the ADC reference voltage by feedback. This avoids the necessity of allowing extra headroom for the variation. The adjustment circuit is very slow and averages the energy from the two channels over both packet and noise conditions.

The HFA3824A Baseband Processor correlates the PN spreading to remove it and to uncover the differential BPSK or QPSK data. The processor initially uses differential detection to identify and lock onto the signal. It then makes measurements of the carrier and symbol timing phase and frequency and uses these to initialize tracking loops for fast

acquisition. Once demodulating and tracking, the processor uses coherent demodulation for best performance. Since this radio uses a spread spectrum signal with 10.4dB of processing gain ($10 \log 11$), the signal to noise ratio (SNR) in the chip rate bandwidth is approximately 0dB when the demodulator is at the desired bit error rate in BPSK. The radio operates with about 2.5dB of implementation loss relative to theoretical performance and achieves a sensitivity of -91dBm in the BPSK mode of operation.

The HFA3824A Baseband Processor provides differential decoding and descrambling of the data to prepare it for the Media Access Controller (MAC). The MAC is an AMD AM79C930 PCnet™ -Mobile controller. All packet signals have a preamble followed by a header containing a start frame delimiter (SFD), other signal related data and a cyclic redundancy check (CRC). The MAC processes the header data to locate the SFD, determine the mode and length of the incoming message and to check the CRC. The MAC then processes the packet data and sends it on through the PC Card interface to the host computer. The MAC checks the packet data CRC to determine the data purity. If corrupted data is received, a retransmission is requested by the MAC which handles the physical layer link protocols.

Transmit Processing

Data from the host computer is sent to the MAC via the PC Card interface. Prior to any communications, however, the MAC sends a Request to Send (RTS) packet to the other end of the link and receives a Clear to Send (CTS) packet. The MAC then formats the payload data packet (MPDU) by appending it to a preamble and header and sends it on to the HFA3824A Baseband Processor which clocks it in. The HFA3824A Baseband Processor scrambles the packet and differentially encodes it before applying the spread spectrum modulation. The data can be either DBPSK or DQPSK modulated at 1MSPS and is a baseband quadrature signal with I and Q components. The BPSK spreading is an 11 chip Barker sequence that is clocked at 11MHz and is modulated with the I and Q data components. These are then output to the HFA3724 as CMOS logic signals. Following the RTS/CTS/MPDU is an acknowledge (ACK) packet by the receiving side of the link.

Transmit quadrature single-bit digital inputs are applied to the HFA3724 Quadrature IF Modulator/Demodulator from the HFA3824A Baseband Processor. These inputs are attenuated by 1/7 and DC coupled to the fifth order Butterworth low pass filters, which are used to provide shaping of the phase shift keyed (PSK) signal. The required transmit spectral mask, at the antenna, is -30dBc at the first side-lobe relative to the main-lobe. An unfiltered PSK waveform would have the first side-lobe suppressed only -13dBc. The fifth-order filters are tuned to an approximate 7.7MHz cutoff, using a 909Ω fixed tuning resistor external to the HFA3724.

In the PC Card wireless LAN, the goal is to control the regrowth of the sidelobes, with the HFA3925 RFPA dominating the regrowth. This will result in maximum transmitted power available. To achieve this goal, once the PSK waveform is filtered at baseband, all remaining transmit elements are operated at a 6dB back-off from compression, except for the HFA3925 RFPA, which is operated at less back-off.

The low pass filters provide initial shaping of the PSK waveform. Final shaping is provided by a transmit IF filter, U5, a Toyocom TQS-432 SAW bandpass filter. The low pass filter outputs are off-chip AC coupled to the quadrature up-converter in the HFA3724. As in the receive mode, the baseband AC coupling time constant is approximately 25 times longer than the symbol period, and is implemented with 0.01 μ F series capacitors. The same twice IF frequency LO used previously is also used in this up-conversion. The IF output of the HFA3724 is reactively matched to U5, with a 250 Ω resistive load presented to the HFA3724. A shunt 33nH inductor, in parallel with a 316 Ω resistor, is used to provide this match, negate the effects of board and component capacitance, and provide a DC return to V_{CC} to prevent saturation in the IF output stage of the HFA3724.

The output of U5 is terminated in a 200 Ω potentiometer that is used for transmit gain control. A shunt 27nH inductor is used to negate the effects of parasitic board and component shunt capacitance, as well as match the SAW output to the potentiometer. This potentiometer has its center wiper connected to the HFA3624 RF/IF Converter transmit IF input, which has an input resistance of approximately 3k Ω . By varying the potentiometer, the gain of the transmit chain is controlled, allowing for precise control of the signal back-off at the HFA3925 RFPA. Therefore, this potentiometer is adjusted to achieve the desired compromise between transmit output power and the mainlobe to sidelobe ratio of the output PSK waveform, typically -32dBc to -35dBc, at an output power of +17.5dBm.

Upconversion to the 2.4GHz - 2.5GHz band is performed in the HFA3624 RF/IF Converter transmit mixer. The mixer output is filtered with FL2, a Murata LFJ30-03B2442B084 two pole monolithic LC bandpass filter. This filter suppresses the LO feedthrough from the mixer, and selects the upper sideband. The transmit buffer in the HFA3624 RF/IF Converter amplifies the selected sideband, easing the requirement for HFA3925 RFPA gain.

FL4, a Toko TDF2A-2450T-10 two pole dielectric bandpass filter, is used to further suppress both the transmit LO leakage and the undesired sideband.

The HFA3925 RFPA amplifies the transmit signal to a level of approximately +21.5dBm, as measured at the T/R switch output. This represents a back-off from 1dB compression of approximately 3dB. Transmit side-lobe performance is approximately -32dBc to -35dBc with this level of back-off. Allowing for approximately 4dB of loss between the T/R

switch output and the antenna connector results in a final output power of +17.5dBm.

The HFA3925 RFPA is the only physical layer component that operates directly from the 5V PC Card supply. To supply the needed negative gate bias to the HFA3925 RFPA, a ICL7660SIBA [8] charge pump is used. A second potentiometer is used to adjust the drain current on the third stage of the HFA3925 to a quiescent operating current of 90mA, as measured through a one ohm sense resistor. A base-emitter junction is used as part of the gate bias network to provide temperature compensation, and all three gates are driven from one source to reduce the impact of process variation on pinch-off voltage. The nominal quiescent drain bias currents are 20mA for stage one, 53mA for stage two, and 90mA for stage three.

A second logic-level PMOS switch, RF1K49093, is used to control the drain supply voltage to the HFA3925 RFPA, and implement a power down mode when receiving. A 2N2222 NPN transistor is used to level shift the 3.5V logic level from the AM79C30 MAC to drive the 5V PMOS switch gates, as well as the 5V HFA3925 RFPA T/R control gate. The T/R V_{DD} pin is connected to the three PA V_{DD} pins, and is powered down in the receive mode by the PMOS switch. In this manner the T/R control pin transfer characteristic is less dependent on its voltage, with the receive state being valid for T/R control voltages as low as 3V. If the T/R V_{DD} pin was connected to a supply in both transmit and receive modes, the T/R control voltage would have to be within a few hundred millivolts of the supply to obtain similar performance.

Following the T/R switch, FL3 and FL6 are reused in the transmit mode to attenuate harmonics generated in the HFA3925 RFPA, as well as provide additional suppression of the LO. As the loss of FL3 and FL6 is approximately 4dB, the amount of transmit power available at the antenna is approximately +17.5dBm.

As the transmit chain is operated linearly, any gain flatness from the HFA3624 and HFA3925, as well as from FL2, FL3, FL4, and FL6, will result in the transmit output power varying across the operating channels. To reduce the amount of variability, three 1pF capacitors are used as coupling elements to provide a form of simple equalization. Care must be exercised to ensure that the filter rejection is still acceptable in meeting the requirements of FCC 15.247. If desired, more complicated equalization could be used to maintain an improved 50 Ω environment for all passband frequencies. Using the simple equalization, the transmit output varies approximately 2.5dB across the band.

Synthesizer Section

The dual frequency synthesizer section uses the HFA3524A [10] Synthesizer and two voltage controlled oscillators to provide a tunable 2132MHz to 2204MHz first

LO, and a fixed 560MHz second LO. Both feedback loops use a 1MHz reference frequency that is derived from a second 22MHz Fox F4106 crystal oscillator. Two separate crystal oscillators were used for the HFA3824A and HFA3524A to maintain a high quality, low spurious reference for the synthesizer. Sharing a common 22MHz oscillator is possible if care is taken to isolate the HFA3824A from the HFA3524A. Both feedback loops are third order and were designed to have loop bandwidths of 10kHz, and phase margins of 50 degrees. The feedback loop analysis is included for both loops in Appendix D. Measured phase noise performance and calculated RMS phase jitter is included in Appendix E. All components in the synthesizer section operate from a regulated 3.5V supply.

The tunable 2132MHz to 2204MHz first LO oscillator is a Motorola™ KXN1332A VCO. To ensure operation at low tuning voltages, a start-up circuit was added to force the tuning voltage from the HFA3524A Synthesizer RF charge pump to a high state for a short period (~1ms) following HFA3524A programming. A 2N2907 PNP transistor was used to implement this function, and the AM79C930 MAC device provides the control signal. The output level of the first LO to the HFA3624 RF/IF Converter is attenuated to approximately -3dBm. An active buffer using an additional HFA3424 is used to provide additional isolation between the VCO and the HFA3624 LO input.

The fixed 560MHz second LO oscillator is a discrete design, using a Phillips BFR505 transistor and a Siemens BBY51 varactor, as described in the HFA3524A Synthesizer evaluation board documentation. The output level of the second LO to the HFA3724 Quadrature IF Modulator/Demodulator is attenuated to approximately -6dBm and a three pole low pass filter is included to preserve the duty cycle of the output. High even order components in the second LO can result in offsets from a 50% duty cycle, and will degrade the quadrature phase accuracy of the HFA3724. A transconductance network is used at the HFA3724 LO input to convert the second LO voltage into a current, as recommended in the HFA3724 data sheet. As the HFA3524A Synthesizer auxiliary IF input covers the 560MHz range, the internal divide-by-two LO buffer output of the HFA3724 is disabled, as recommended in the HFA3724 data sheet.

Regulator Section

Linear voltage regulators are used to provide filtering and isolation from the 5V PC Card input supply. An additional advantage of using voltage regulators is a savings in overall supply current, as all of the components that are regulated consume less current at a 3.5V operating point, as opposed to a 5V operating point. The 3.5V operating point was chosen specifically to support the AM79C930 MAC controller. At the time of publication 3.5V was the lowest nominal operating voltage approved by AMD for 40MHz AM79C930 operation.

The only components operating directly from the 5V supply are the HFA3925 RFPA, in order to maximize RF output power, and the PC Card interface sections of the AM79C930 MAC controller.

A total of three regulators, 3.5V Toko TK11235MTL, are used in the PC Card wireless LAN. One regulator supplies voltage to the HFA3824A Baseband Processor and portions of the AM79C930, as well as the HFA3424 LNA and HFA3624 RF/IF Converter. A second regulator supplies voltage to the synthesizer. The third regulator supplies voltage to the HFA3724 Quadrature IF Modulator/Demodulator.

PCB Layout Guidelines

Although the actual PCB layout is proprietary, some of the techniques utilized are worthy of discussion [11]. As there are many RF, IF, analog, and digital circuits in close proximity, isolation is of prime concern. All RF and IF circuits utilize coplanar waveguide with ground transmission line techniques to allow for easy integration of varied line widths and component pin widths, and to provide a low dispersion, high isolation environment. A Radio Schematic is available on the internet at <http://www.semi.harris.com/prism/lanref.htm>.

The outside two planes of each side of the PCB are dedicated to RF and IF signal processing, and form two pairs of coplanar waveguide with ground circuits. As the two sides of the PCB contain circuitry that must be isolated from each other, blind via techniques are used, and the only places that the two sides share common ground or signal connections are when signals are passed between them - mainly when LO1 and LO2 need to pass from the synthesizer side to the RF/IF transceiver side.

In general, the RF and IF circuit layouts need to be as short and direct as possible to avoid costly shielding. This is especially critical in the receive IF stages where spurious signal coupling can easily occur, resulting in poor sensitivity or high packet error rates.

Appendix A Reference Radio Bill Of Materials

TABLE 2. REFERENCE RADIO BILL OF MATERIALS (Rev. A1, 3-26-98)

LINE ITEM	PART #/CUST. #	VENDOR	VENDOR P/N	COMMODITY	TYPE	MECH. DES. MOIST.	REFERENCE DESCRIPTION	QTY	COMMENTS
1	00000003206C	Murata, AVX, Kemet, TDK	GRM39X7R104K016AD 0603YC104KAT2A C0603C104K5RAC C1608X7R1C104KT	Capacitor	Ceramic	SMT0603	C21, C26, C27, C34, C36, C37, C38, C40, C41, C42, C45, C46, C53, C55, C57, C58, C66, C70, C71, C74, C79, C80, C82, C99, C101, C102, C106, C119, C120, C121, C124-C130, C132, C136	39	
2	00000003209C	Motorola, Siemens	MMBT2907ALT1 SMBT2907A	Transistor	BJT-PNP	SMTSOT23	Q4	1	
3	00000004945C	Toko	LL1608-F15NK	Inductor	Power	SMT	L7, L9, L14, L23	4	
4	00000006661C	Rohm	MCH212C683KP	Capacitor	Ceramic	SMT0805	C97	1	
5	00000009762C	Harris	ICL7660SIBA-T	Linear	Converter	SMTSOIC N	U12	1	Super Voltage Converter
6	00000009897C	Siemens	BBY51-E6327	Diode-Rect	Varactor	SMTSOT23	CR3	1	CT = 5.3PF at 1V, CT = 3.1PF at 4V
7	00000009898C	Dialight	597-3401-407	Optic	LED Single	SMT	DS2	1	SMT, 7" Reel, 2500 Pieces/Reel
8	00000009899C	Dialight	597-3111-407	Optic	LED Single	SMT	DS3	1	SMT, 7" Reel, 2500 Pieces/Reel
9	00000009900C	Harris	HFA3724IN	Wireless/Rf	Mod/Demod	TQFP Y-6	U1	1	400MHz Quadrature IF Modulator/Demod.
10	00000009901C	Dialight	597-3311-407	Optic	LED Single	SMT	DS1	1	SMT, 7" Reel, 2500 Pieces/Reel
11	00000009902C	Harris	HFA3624IA96	Wireless/Rf	Prescaler	SMTSSOP N	U4	1	2.4GHz RF to IF Converter
12	00000009903C	Philips	BFR 505	Transistor	BJT-NPN	SMTSOT23	Q2	1	9GHz Wideband Transistor, High Gain
13	00000009904C	Toko	TK11235AMTL	Linear	Volt Reg	SMTSOT23 N	U13, U16, U21	3	3.5V Voltage Regulator with On/Off Switch
14	00000009907C	AMD	AM79C930	Controller		TQFP Y-4	U3	1	IC LAN Media Access Controller
15	00000009935C	Toko	LL1608-F33NK	Inductor	Power	SMT0805	L2, L4-L5	3	Chip Inductor, DCR = 0.65Ω
16	00000009936C	AVX, TDK	04025A5R6CAT2A C1005C0G1H5R6CT	Capacitor	Ceramic	SMT0402	C78, C122	2	
17	00000009937C	AVX, TDK	04025A4R7CAT2A C1005COG1H4R7CT	Capacitor	Ceramic	SMT0402	C93	1	
18	00000009938C	AVX, TDK	04025C501JAT2A C1005X7R1H501JT	Capacitor	Ceramic	SMT0402	C32, C154	2	

TABLE 2. REFERENCE RADIO BILL OF MATERIALS (Rev. A1, 3-26-98) (Continued)

LINE ITEM	PART #/CUST. #	VENDOR	VENDOR P/N	COMMODITY	TYPE	MECH. DES. MOIST.	REFERENCE DESCRIPTION	QTY	COMMENTS
19	00000009943C	AVX, TDK	04025A470JAT2A C1005COG1H470JT	Capacitor	Ceramic	SMT0402	C13, C18, C28, C60	4	
20	00000009944C	AVX, TDK	04025A100CAT2A C1005COG1H100CT	Capacitor	Ceramic	SMT0402	C31	1	
21	00000009945C	AVX, TDK	04025A101JAT2A C1005COG1H101JT	Capacitor	Ceramic	SMT0402	C6, C9, C11, C12, C14, C15, C22, C30, C43, C47, C48, C65, C69, C72, C90, C92, C100, C104, C109, C110, C118	21	
22	00000009947C	AVX, TDK	0402YC103KAT*A C1005X7R1C103KT	Capacitor	Ceramic	SMT0402	C1, C2, C3, C4, C5, C7, C8, C10, C16, C29, C35, C76, C94, C96, C147, C153	16	* = 1, 2, M OR N
23	00000009948C	AVX, TDK	04025A220JAT2A C1005COG1H220JT	Capacitor	Ceramic	SMT0402	C23, C24, C52, C67, C98, C143	6	
24	00000009949C	AVX, TDK	04023C102KAT2A C1005X7R1C102KT	Capacitor	Ceramic	SMT0402	C17, C19, C25, C33, C39, C50, C63, C64, C75, C95, C113, C138, C140, C141	14	
25	00000009950C	Voltronics	JZ060	Capacitor	Trim	SMT	C139	1	
26	00000009974C	Panasonic, Rohm	ERJ2GEJ911X MCR01MZSJ911	Resistor	Film	SMT0402	R31	1	
27	00000009975C	Panasonic, Rohm	ERJ2GEJ432X MCR01MZSJ432	Resistor	Film	SMT0402	R63, R64	2	
28	00000009976C	Panasonic, Rohm	ERJ2GEJ221X MCR01MZSJ221	Resistor	Film	SMT0402	R2, R33	2	
29	00000009977C	Toko	TDF2A-2450T-10	Filter	Dielectric	SMT	FL3-FL4,FL6	3	
30	00000009978C	Panasonic, Rohm	ERJ2RKF2610X MCR01MZSF2610	Resistor	Film	SMT0402	R5	1	
31	00000009979C	Panasonic, Rohm	ERJ2GEJ561X MCR01MZSJ561	Resistor	Film	SMT0402	R9	1	
32	00000009981C	Panasonic, Rohm	ERJ2GEJ560X MCR01MZSJ560	Resistor	Film	SMT0402	R23, R87	2	
33	00000009982C	Panasonic, Rohm	ERJ2GEJ472X MCR01MZSJ472	Resistor	Film	SMT0402	R38, R76	2	
34	00000009983C	Panasonic	ERJ2GEJ153X	Resistor	Film	SMT0402	R46, R77	2	
35	00000009983C	Rohm	MCR01MZSJ153	Resistor	Film	SMT0402	R99-R104	6	
36	00000009984C	Panasonic, Rohm	ERJ2GEJ104X MCR01M2SJ104	Resistor	Film	SMT0402	R27	1	

TABLE 2. REFERENCE RADIO BILL OF MATERIALS (Rev. A1, 3-26-98) (Continued)

LINE ITEM	PART #/CUST. #	VENDOR	VENDOR P/N	COMMODITY	TYPE	MECH. DES. MOIST.	REFERENCE DESCRIPTION	QTY	COMMENTS
37	00000009986C	Panasonic, Rohm	ERJ2GEJ152X MCR01MZSJ152	Resistor	Film	SMT0402	R35, R50	2	
38	00000009989C	Panasonic, Rohm	ERJ2GEJ100X MCR01M2SJ100	Resistor	Film	SMT0402	R39, R85	2	
39	00000009990C	Panasonic, Rohm	ERJ2GEJ101X MCR01M2SJ101	Resistor	Film	SMT0402	R45, R78	2	
40	00000009991C	Panasonic, Rohm, KOA	ERJ2GEJ103X MCR01MZSJ103 RM73B1ET103J	Resistor	Film	SMT0402	R41	1	
41	00000009992C	Panasonic, Rohm	ERJ2GEJ750X MCR01MZSJ750	Resistor	Film	SMT0402	R51	1	
42	00000009994C	Panasonic, Rohm	ERJ2GEJ681X MCR01MZSJ681	Resistor	Film	SMT0402	R42	1	
43	00000010004C	Panasonic, Rohm	ERJ2GEJ392X MCR01MZSJ392	Resistor	Film	SMT0402	R13	1	Resistor 3.9K 5% 0402 1/16W
44	00000010005C	Panasonic, Rohm	ERJ2GEJ822X MCR01MZSJ822	Resistor	Film	SMT0402	R22	1	Resistor 8.2K 5% 1/16W 0402
45	00000010006C	Panasonic, Rohm	ERJ2GEJ912X MCR01MZSJ912	Resistor	Film	SMT0402	R1	1	Resistor 5% 9.1K 1/16W 0402
46	00000010007C	Panasonic, Rohm	ERJ2GEJ331X MCR01MZSJ331	Resistor	Film	SMT0402	R21, R24	2	Resistor 330Ω 5% 0402 1/16W
47	00000010008C	Panasonic, Rohm	ERJ2GEJ102X MCR01MZSJ102	Resistor	Film	SMT0402	R4	1	Resistor 1K 5% 1/16W 0402
48	00000010009C	Panasonic, Rohm	ERJ2GE0R00X MCR01MZSJ0R00	Resistor	Film	SMT0402	R7, R20, R48, R49, R55, R56, R59, R60, R61, R62, R66, R69, R97	13	Resistor 0Ω Jumper 0402
49	00000010010C	Panasonic, Rohm	ERJ3GSYJ681V MCR03EZHMJ681	Resistor	Film	SMT0603	R58, R65	2	
50	00000010011C	Panasonic, Rohm, KO, Q Dale	ERJ3EKF3160V MCR03EZHMF3160 RK73H1JTF3160 CRCW06033160FRT1	Resistor	Film	SMT0603	R17	1	
51	00000010012C	Rohm, Panasonic	MCR10EZHMJW331 ERJGEYJ331V	Resistor	Film	SMT0805	R32	1	
52	00000010013C	Philips	9C06031A1R0F	Resistor	Film	SMT0603	R3	1	This was changed from the SMT0505
53	00000010014C	Harris	HFA3524AIA96	Wireless/RF	Synth	TSOP II N	U22	1	2.5GHz/600MHz Dual Freq Synth.

TABLE 2. REFERENCE RADIO BILL OF MATERIALS (Rev. A1, 3-26-98) (Continued)

LINE ITEM	PART #/CUST. #	VENDOR	VENDOR P/N	COMMODITY	TYPE	MECH. DES. MOIST.	REFERENCE DESCRIPTION	QTY	COMMENTS
54	00000010021C	Harris	HFA3824AVI	Wireless/RF	BBP	TQFP Y-2	U6	1	Baseband processor
55	00000010085C	NIC	NCB1206B320TR	Inductor	Ferr Chip	SMT	L1, L8, L12, L19, L21, L22, L24	7	
56	00000010169C	Harris	HFA3424IB96	Wireless/RF	LNA	SMTSOIC N	U19, U24	2	RF component, 2.5GHz Low Noise Amplifier
57	00000010170C	Harris	HFA3925IA96	Wireless/RF	PA	SMTSSOP N	AR1	1	2.4-2.5GHz 25MW RF Power Amplifier
58	00000010171C	Murata	LFJ30-03B2442B084	Filter	EMI	SMT	FL1-FL2	2	Tape and Reel 2.442GHz, BW = 84MHz
59	00000010172C	Panasonic, Rohm	ERJ2RKF4990X MCR01MZSF4990	Resistor	Film	SMT0402	R10	1	Resistor, 499Ω 1% 1/16W 0402
60	00000010214C	Statek	CX-6V-SM2-32.768K-	Freq Control	Crystal	SMT	U11	1	Nickel tin plated termination
61	00000011540C	Harris	RF1K4909396	Transistor	FET P-CH	SMTSO8	U9, U25	2	Dual MOSFET, 8 Pin DIL SMT package
62	00000011729C	Panasonic, Rohm	ERJ2GEJ751X MCR01MZSJ751	Resistor	Film	SMT0402	R26, R52	2	Resistor, 750Ω 5% 0402 1/16W
63	00000011734C	Panasonic	EVM1SSW50B22	Resistor	Pots	SMT	R19	1	
64	00000011758C None	AVX, TDK	04025C222JAT2A C1005X7R1H222JT	Capacitor	Ceramic	SMT0402	C142	1	
65	00000011759C None	AVX, NIC	04023A101FAT2A NMC0402NPO101F025T	Capacitor	Ceramic	SMT0402	C44, C49, C54, C61, C111, C155	6	
66	00000011760C None	AVX, TDK	04023A200FAT*A C1005COG1E200CT	Capacitor	Ceramic	SMT0402	C62	1	* = 1, 2, M or N for AVX P/N
67	00000011762C None	AVX, TDK, NIC	04023A150FAT2A C1005COG1E150FT NMC0402NPO150F25TR	Capacitor	Ceramic	SMT0402	C73, C77	2	
68	00000011836C N/A	AVX	L0603100GFWTR	Inductor	Power	SMT0603	L3	1	RF IND. L at 450Ω, DCR = 0.13Ω
69	00000011837C N/A	AVX	L0603120GFWTR	Inductor	Power	SMT0603	L17	1	RF IND. L at 450Ω, DCR = 0.20Ω
70	00000011866C	Panasonic, Rohm	ERJ3GSYK106V MCR03EZHU106	Resistor	Film	SMT0603	R57	1	
71	00000011869C	Panasonic, Rohm	ERJ2GEJ430X MCR01MZSJ430	Resistor	Film	SMT0402	R36	1	
72	00000011870C	Panasonic, Rohm	ERJ2GEJ240X MCR01MZSJ240	Resistor	Film	SMT0402	R47, R53	2	Resistor, 24Ω 5% 0402

TABLE 2. REFERENCE RADIO BILL OF MATERIALS (Rev. A1, 3-26-98) (Continued)

LINE ITEM	PART #/CUST. #	VENDOR	VENDOR P/N	COMMODITY	TYPE	MECH. DES. MOIST.	REFERENCE DESCRIPTION	QTY	COMMENTS
73	00000012189C	Samsung, Mitsubishi	KM62256CLTG-5L M5M5256CVP-55LL	Memory	SRAM	TSOP I Y-3	U8	1	
74	00000012987C	Fox, Statek	F4106-22.000M CXO- M10N22M(25PPM)	Freq Control Oscillator		SMT	U20, U23	2	
75	00000012990C	Panasonic, Rohm	ERJ2RKF9090X MCR01MZSF9090	Resistor	Film	SMT0402	R34	1	Resistor, 909Ω 1% 0402 1/16W
76	00000012992C	Panasonic, Rohm	ERJ2RKF3011X MCR01MZSF3011	Resistor	Film	SMT0402	R14, R25	2	Resistor, 3.01K 1% 1/16W
77	00000013250C	Panasonic, Rohm	ERJ2RKF7150X MCR01MZSF7150	Resistor	Film	SMT0402	R43, R84	2	Resistor, 715Ω 1% 1/16W 0402
78	00000013251C	Panasonic, Rohm	ERJ2GEJ511X MCR01MZSJ511	Resistor	Film	SMT0402	R82	1	
79	00000013303C			Sheet Metal	Bottom Shield		Bottom Shield	1	
80	00000013319C None	AVX, TDK	04025A5R0CAT*A C1005COG1H5R0CT	Capacitor	Ceramic	SMT0402	C123, C144, C151	3	* = 1 or 2
81	00000013325C None	AVX, TDK	0402YC562JAT*A C1005X7R1C562JT	Capacitor	Ceramic	SMT0402	C91	1	* = 1 or 2
82	00000013326C None	AVX	0805YC563JAT*A	Capacitor	Ceramic	SMT0805	C115	1	* = 1 or 2
83	00000013327C	AVX, Kemet	TAJS475K6R3R T491S475K006AS	Capacitor	Tantalum	SMTS	C51, C56, C83, C85, C133, C134	6	Low Profile - 3216L
84	00000013329C	AVX	TAJT475K010R	Capacitor	Tantalum	SMTT	C84, C135	2	
85	00000013330C	AVX	TAJT106K010R	Capacitor	Tantalum	SMTT	C81	1	3528L - Low Profile
86	00000013345C	Panasonic, Rohm	ERJ2GEJ394X MCR01MZSJ394	Resistor	Film	SMT0402	R79	1	
87	00000013875C	AVX, Kemet, Murata	06035C682KAT*A C0603C682K5RAC GRM39X7R682K050%#	Capacitor	Ceramic	SMT0603	C89	1	* = 1 or 2 or M: % = A or B: # = D or L
88	00000014977C	AVX, Murata	04025A1R0CAT*A GRM36COG1R0C050%#	Capacitor	Ceramic	SMT0402	C59, C131, C137	3	* = 1 or 2 or M, % = A or B, # = D or L
89	00000015327C	Statek Fox	CXO-M-10N-40MHz (100PPM) F3355-40MHz	Freq Control Oscillator		SMT	U10	1	
90	00000016302C	Panasonic, Rohm	ERJ2GEJ680X MCR01MZSJ680	Resistor	Film	SMT0402	R86	1	
91	00000016708C	Murata, TDK AVX	GRM39COG2R0C050A# C1608COG1H020CT 06035A2R0CAT*A	Capacitor	Ceramic	SMT0603	C116	1	* = 1, 2, M or N; # = D or L

TABLE 2. REFERENCE RADIO BILL OF MATERIALS (Rev. A1, 3-26-98) (Continued)

LINE ITEM	PART #/CUST. #	VENDOR	VENDOR P/N	COMMODITY	TYPE	MECH. DES. MOIST.	REFERENCE DESCRIPTION	QTY	COMMENTS
92	00000016709C	Murata TDK AVX	GRM36COG8R0C050A# C1005COG1H080CT 04025A8R0CAT*A	Capacitor	Ceramic	SMT0402	C68	1	* = 1, 2, M or N; # = D or L
93	00000016710C	Toko	LL2012-F27NK	Inductor	Fixed	SMT0805	L6	1	Chip Inductor, DCR = 0.55Ω
94	00000016812C	Panasonic	EVM1SSX50B53	Resistor	Potentiometer	SMT	R75	1	5K SMP Pot, Type X 2000 per Reel
95	00000017117C	Toyocom	TQS-432E-7R	Filter	Saw	SMT	U5, U7	2	Tape and Reel
96	00000017587C			Label Outline Drawing	General				
97	00000018647C	ITT-Cannon		Coverset (top and bottom)					
98	00000018864C			Sheet Metal	Top Shield		Top shield	1	
99	00000019393C	Motorola	KXN1332A	Freq Control	Oscillator	SMT	U18	1	Voltage Controlled, 2132-2204MHz range
100	00000019584C			PCB			PRISM1BRD	1	Raw PC Board
101	00000019598C			Coppertape top cut out					
102	00000019599C			Coppertape bottom cut out					
103	00000025454C	Panasonic, Rohm, KOA	ERJ2GEJ202X MCR01MZSJ202 RM73B1ET202J	Resistor	Film	SMT0402	R37	1	
104	00000027262C	KOA, Rohm	RM73B1ET132J MCR01MZSJ132	Resistor	Film	SMT0402	R44	1	
105	00000027369C	AVX, Murata TDK	04023C122JAT*A GRM36X7R122J025A# C1005X7R1E122JT	Capacitor	Ceramic	SMT0402	C86	1	* = 1, 2, M or N; # = D or L
106	00000032228C	AMD	Device Driver	Device Driver	Software		April '98 Software Win95 Driver	1	
107	00000032227C	AMD	Firmware Version 1.6 or later		Firmware		Firmware Version 1.6 April '98 or later	0	
108	000000932607C	ITT-Cannon		Connector 68-pin PC card side	Terminal		P1	1	
109	0000013H9703fP	Toko	LL1608-F12NK	Inductor	Fixed	SMT0603	L15-L16	2	(±10%)
110	000000941687	Motorola	MMBT2222ALT1	Transistor	BJT-NPN	SMTSOT23	Q3, Q5	2	
111	000000943218	AMD	AM29F010-55EC	Memory	Flash	TSOP I Y-3	U2	1	32 Pin Standard TSOP

Appendix B Receive Mixer Spurious Analysis

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PRISM Reference Radio, 08/29/1996, 16:08:48

Receive Crossover Responses in Receive Band (2400MHz to 2500MHz)

LO Level = +7dBm

<u>IF FREQ</u>	<u>LO FREQ</u>	<u>RCV FREQ</u>	<u>SPUR FREQ</u>	<u>M</u>	<u>N</u>	<u>MXR</u>
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No spurs recorded.

PRISM Reference Radio, 08/29/1996, 16:08:50

Spurious Responses in 0MHz to 5000MHz Band (-90dB Minimum)

LO Level = +7dBm

Fixed Bandpass Preselector

Band Edges are 2400MHz and 2500MHz

4 Section Butterworth - Corner Attenuation = 1dB

Ultimate Preselector Rejection = 80dB

<u>IF FREQ</u>	<u>LO FREQ</u>	<u>RCV FREQ</u>	<u>SPUR FREQ</u>	<u>M</u>	<u>N</u>	<u>MXR</u>	<u>FLT</u>	<u>TOT</u>
280.0000	2120.0000	2400.0000	2400.0000	+1	-1	0	1	1
280.0000	2120.0000	2400.0000	1840.0000	-1	+1	0	80	80
280.0000	2120.0000	2400.0000	2488.0000	-5	+6	90	0	90
280.0000	2120.0000	2400.0000	2426.6667	-6	+7	90	0	90
280.0000	2120.0000	2400.0000	2462.8571	+7	-8	90	0	90
280.0000	2130.0000	2410.0000	2410.0000	+1	-1	0	0	0
280.0000	2130.0000	2410.0000	1850.0000	-1	+1	0	80	80
280.0000	2130.0000	2410.0000	2438.3333	-6	+7	90	0	90
280.0000	2130.0000	2410.0000	2474.2857	+7	-8	90	0	90
280.0000	2140.0000	2420.0000	2420.0000	+1	-1	0	0	0
280.0000	2140.0000	2420.0000	1860.0000	-1	+1	0	80	80
280.0000	2140.0000	2420.0000	2450.0000	-6	+7	90	0	90
280.0000	2140.0000	2420.0000	2485.7143	+7	-8	90	0	90
280.0000	2140.0000	2420.0000	2405.7143	-7	+8	90	0	90
280.0000	2150.0000	2430.0000	2430.0000	+1	-1	0	0	0
280.0000	2150.0000	2430.0000	1870.0000	-1	+1	0	80	80
280.0000	2150.0000	2430.0000	2461.6667	-6	+7	90	0	90
280.0000	2150.0000	2430.0000	2417.1429	-7	+8	90	0	90
280.0000	2160.0000	2440.0000	2440.0000	+1	-1	0	0	0
280.0000	2160.0000	2440.0000	1880.0000	-1	+1	0	80	80
280.0000	2160.0000	2440.0000	2473.3333	-6	+7	90	0	90
280.0000	2160.0000	2440.0000	2428.5714	-7	+8	90	0	90
280.0000	2170.0000	2450.0000	2450.0000	+1	-1	0	0	0
280.0000	2170.0000	2450.0000	1890.0000	-1	+1	0	80	80
280.0000	2170.0000	2450.0000	2485.0006	-6	+7	90	0	90
280.0000	2170.0000	2450.0000	2440.0000	-7	+8	90	0	90

Application Note 9624

<u>IF FREQ</u>	<u>LO FREQ</u>	<u>RCV FREQ</u>	<u>SPUR FREQ</u>	<u>M</u>	<u>N</u>	<u>MXR</u>	<u>FLT</u>	<u>TOT</u>
280.0000	2180.0000	2460.0000	2460.0000	+1	-1	0	0	0
280.0000	2180.0000	2460.0000	1900.0000	-1	+1	0	80	80
280.0000	2180.0000	2460.0000	2273.3333	+3	-3	50	39	89
280.0000	2180.0000	2460.0000	2451.4286	-7	+8	90	0	90
280.0000	2190.0000	2470.0000	2470.0000	+1	-1	0	0	0
280.0000	2190.0000	2470.0000	1910.0000	-1	+1	0	80	80
280.0000	2190.0000	2470.0000	2283.3333	+3	-3	50	37	87
280.0000	2190.0000	2470.0000	2462.8571	-7	+8	90	0	90
280.0000	2200.0000	2480.0000	2480.0000	+1	-1	0	0	0
280.0000	2200.0000	2480.0000	1920.0000	-1	+1	0	80	80
280.0000	2200.0000	2480.0000	2293.3333	+3	-3	50	35	85
280.0000	2200.0000	2480.0000	2474.2857	-7	+8	90	0	90
280.0000	2210.0000	2490.0000	2490.0000	+1	-1	0	0	0
280.0000	2210.0000	2490.0000	1930.0000	-1	+1	0	80	80
280.0000	2210.0000	2490.0000	2303.3333	+3	-3	50	32	82
280.0000	2210.0000	2490.0000	2485.7143	-7	+8	90	0	90
280.0000	2220.0000	2500.0000	2500.0000	+1	-1	0	1	1
280.0000	2220.0000	2500.0000	1940.0000	-1	+1	0	79	79
280.0000	2220.0000	2500.0000	2360.0000	+2	-2	74	15	89
280.0000	2220.0000	2500.0000	2313.3333	+3	-3	50	30	80

TABLE 3. RECEIVE SPUR TABLE - LO POWER = 7dBm

	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
(M) 15	99															
14	99	99														
13	90	99	90													
12	99	99	99	99												
11	90	99	90	95	90											
10	99	99	99	99	99	99										
9	90	95	90	90	90	99	90									
8	99	95	99	95	99	95	99	95								
7	90	90	90	90	90	87	90	90	90							
6	90	90	90	90	90	90	90	90	90	90						
5	90	80	90	71	90	68	90	65	88	65	85					
4	86	90	86	88	88	85	86	85	90	85	85	85				
3	67	64	69	50	77	47	74	44	74	47	75	44	70			
2	73	73	74	70	71	64	69	64	69	62	74	62	72	60		
1	24	0	35	13	40	24	45	28	49	33	53	42	60	47	63	
0	---	26	35	39	50	41	53	49	51	42	62	51	60	47	77	50

Appendix C IF SAW Filter Measured Data

Toyocom TQS-432E-7R, VHF-Range Wideband Low Loss SAW Filter

SPECIFICATIONS (TENTATIVE)	
Reference Frequency (f_0):	280MHz
Passband:	$f_0 \pm 8.5\text{MHz}$
Maximum Insertion Loss in Passband:	10dB Max
Attenuation:	$f_0 \pm 38.8\text{MHz}$ at 50dB
Terminating Impedance:	$270\Omega // -5\text{pF}$
Operating Temperature Range:	-10°C to 50°C
Package:	SS-752A

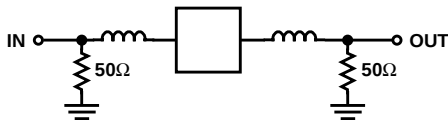


FIGURE 3.

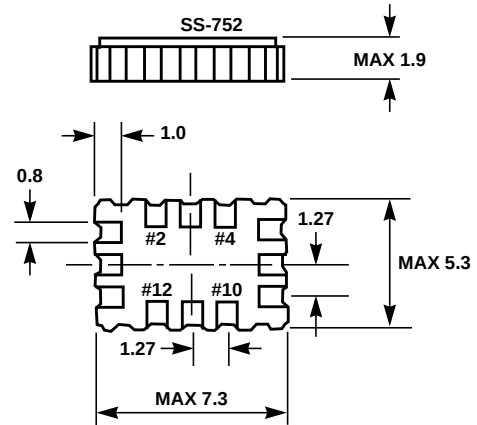


FIGURE 2.

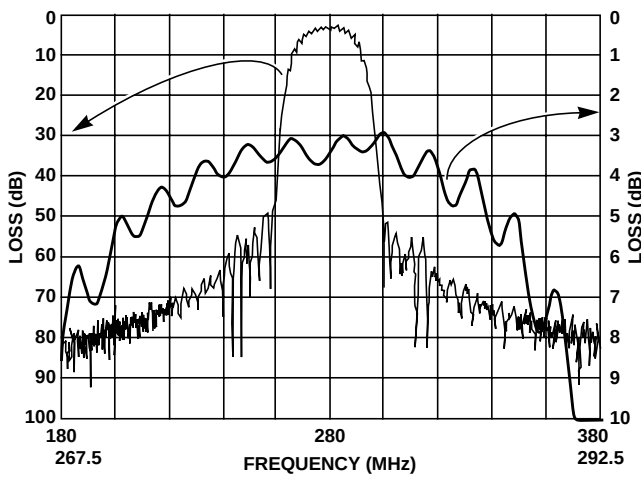


FIGURE 4. FREQUENCY RESPONSE

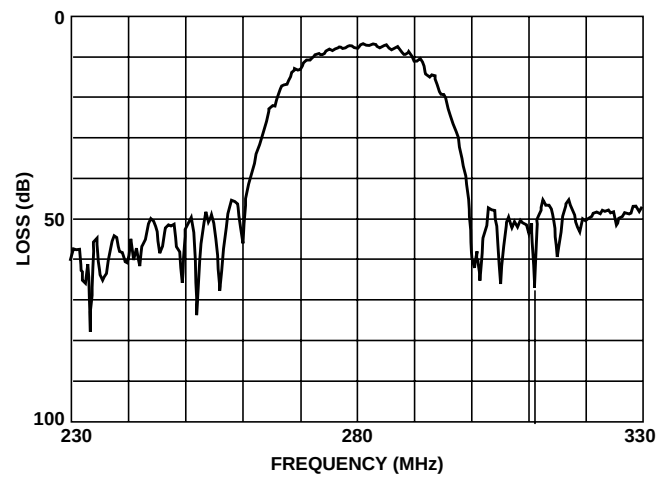


FIGURE 5. FREQUENCY RESPONSE

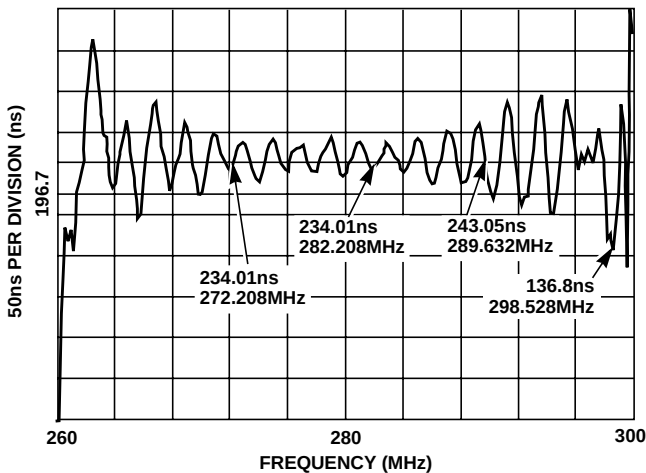


FIGURE 6. GROUP DELAY

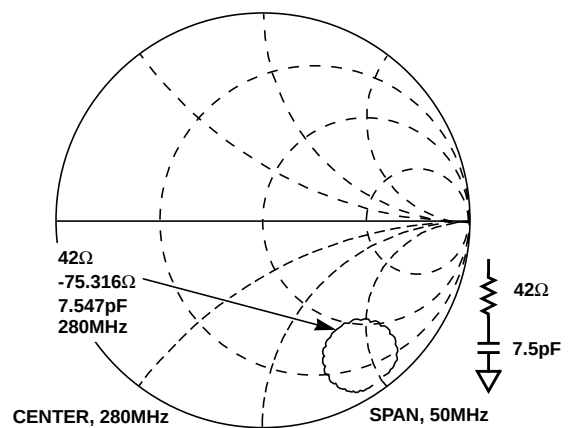


FIGURE 7. INPUT IMPEDANCE CHARACTERISTIC

Appendix D Synthesizer Loop Analysis

A linear control system model of the phase feedback for a PLL in the locked state is shown in Figure 8. The open loop gain is the product of the phase detector gain (K_{PD}), the VCO gain (K_{VCO}/s), and the loop filter gain $Z(s)$ divided by the gain of the feedback counter modulus (N). The passive loop filter configuration used is displayed in Figure 9.

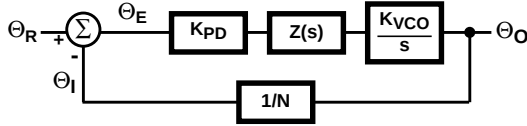


FIGURE 8. PLL LINEAR MODEL

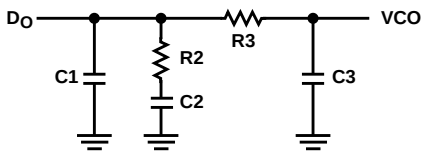


FIGURE 9. PASSIVE LOOP FILTER

This analysis calculates the loop filter components for the PRISM HFA3524A Dual PLL.

A. Schaldenbrand and R.D. Schultz, 5/8/96

Where:

f_{loop} , 2132MHz to 2204MHz,

f_{loop} is the desired bandwidth of the PLL, in Hz,

f_{ref} is the reference frequency of the loop, in Hz,

ϕ is the desired phase margin of the PLL, in degrees,

K_{vco} is the VCO Tuning Voltage Constant in Hz/V,

K_{pd} is the Phase Detector Pump Constant in A/rad,

N is the divider ratio,

1.424 is a constant to account for impact of $R3/C3$ pole.

$$f_{loop} := 10 \cdot 10^3 \quad \omega_p := 2 \cdot \pi \cdot f_{loop} \quad \omega_p = 6.283 \cdot 10^4$$

$$f_{ref} := 1 \cdot 10^6 \quad \omega_r := 2 \cdot \pi \cdot f_{ref} \quad \omega_r = 6.283 \cdot 10^6$$

$$\phi := 50^\circ \quad \phi_r := \left(\pi \cdot \frac{\phi}{180^\circ} \right)$$

$$K_{vco} := 53 \cdot 10^6 \quad K_{vco} := 2 \cdot \pi \cdot K_{vco} \quad K_{vco} = 3.33 \cdot 10^8$$

$$K_{pd} := 0.004 \quad K_{pd} := \frac{K_{pd}}{2 \cdot \pi} \quad K_{pd} = 6.366 \cdot 10^{-4}$$

$$N := 2168$$

$$T3 := \frac{1}{10 \cdot \omega_p} \quad \omega_p := 1.424 \cdot \omega_p$$

$$T1 := \frac{\sec(\phi_r) - \tan(\phi_r)}{\omega_p} \quad T1 = 4.068 \cdot 10^{-6}$$

$$\omega_c := \frac{\tan(\phi_r) \cdot (T1 + T3)}{[T1 + T3^2 + T1 \cdot T3]} \cdot \left[\sqrt{1 + \frac{(T1 + T3)^2 + T1 \cdot T3}{(\tan(\phi_r) \cdot (T1 + T3))^2}} - 1 \right]$$

$$T2 := \left[\frac{1}{\omega_c^2 \cdot (T1 + T3)} \right] \quad f_c := \frac{\omega_c}{2 \cdot \pi}$$

$$C1 := \left(\frac{T1 \cdot K_{pd} \cdot K_{vco}}{N \cdot T2 \cdot \omega_c^2} \right) \cdot \sqrt{\frac{1 + (\omega_c \cdot T2)^2}{[1 + (\omega_c \cdot T1)^2] \cdot [1 + (\omega_c \cdot T3)^2]}}$$

$$C2 := C1 \cdot \left[\left(\frac{T2}{T1} \right) - 1 \right]$$

$$R2 := \frac{T2}{C2}$$

Loop Filter Component Values

$$T1 = 4.068 \cdot 10^{-6}$$

$$T2 = 4.472 \cdot 10^{-5}$$

$$T3 = 1.592 \cdot 10^{-6}$$

$$f_c = 1 \cdot 10^4$$

$$C1 = 6.47 \cdot 10^{-9}$$

$$C1 := 0.068 \cdot 10^{-6}$$

$$C2 = 6.47 \cdot 10^{-8}$$

$$C2 := 0.68 \cdot 10^{-6}$$

$$R2 = 691.064$$

$$R2 := 680$$

Rules of thumb for selecting the values of the reference frequency suppression filter are:

1. Choose $C3 < C1/10$

$$C3 := \frac{C1}{10} \quad R3 := \frac{T3}{C3}$$

$$R3 = 234.051 \quad C3 = 6.8 \cdot 10^{-9}$$

2. Choose $R3 < 2 \cdot R2$

$$R3 := 2 \cdot R2 \quad R3 = 1.36 \cdot 10^3 \quad R3 = 1300$$

$$C3 := \frac{T3}{R3} \quad C3 = 1.224 \cdot 10^{-9} \quad C3 := 1200 \cdot 10^{-12}$$

$$k = 10..70 \quad kp(k) := 10 \left(\frac{k}{10} \right)$$

$$\omega(k) := 2j \cdot \pi \cdot kp(k)$$

$$GH(k) := K_{pd} \cdot K_{vco} \cdot \frac{1 + \omega(k) \cdot T2}{(C1 \cdot N \cdot \omega(k) \cdot \overline{\omega(k)}) \cdot (1 + \omega(k) \cdot T1)} \cdot \frac{T1}{T2} \cdot \left[\frac{1}{(1 + \omega(k) \cdot T3)} \right]$$

$$MAGdb_GH(k) := 20 \cdot \text{Log}(GH(k) \cdot \overline{GH(k)})$$

$$PHI_GH(k) := \text{atan}\left(\frac{\text{Im}(GH(k))}{\text{Re}(GH(k))}\right)$$

$$f_GH(k) := \left(\frac{180 \cdot PHI_GH(k)}{\pi} \right) - 180$$

$$\text{MAG_GH}(k) := (\text{GH}(k) \cdot \overline{\text{GH}(k)}) \quad \text{Phim}(k) := 180 + f_GH(k)$$

Plots of the magnitude and phase of GH(k) are shown in Figures 10 and 11.

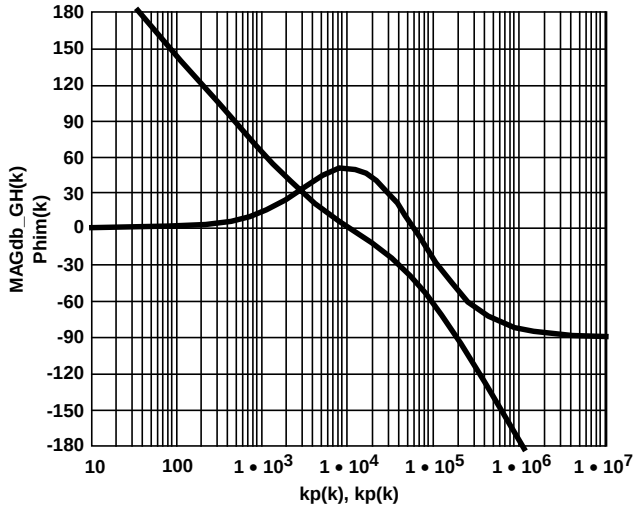


FIGURE 10.

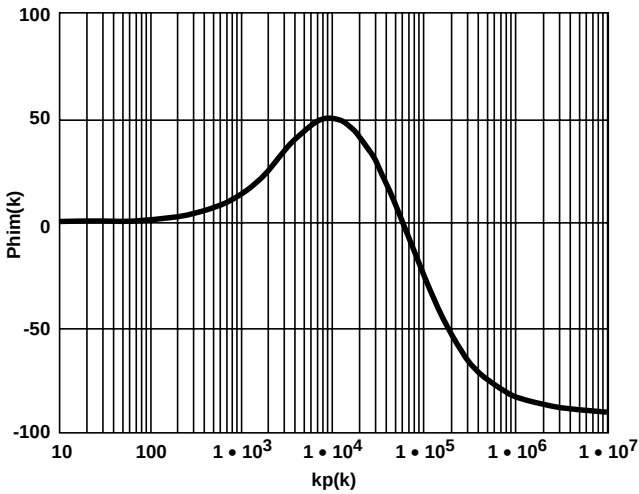


FIGURE 11.

This analysis calculates the loop filter components for the PRISM HFA3524A Dual PLL.

A. Schaldenbrand and R.D. Schultz, 5/8/96

Where:

LO2, 560MHz,

f_{loop} is the desired bandwidth of the PLL, in Hz,

f_{ref} is the reference frequency of the loop, in Hz,

ϕ is the desired phase margin of the PLL, in degrees,

Kvco is the VCO Tuning Voltage Constant in Hz/V,

Kpd is the Phase Detector Pump Constant in A/rad,

N is the divider ratio,

1.424 is a constant to account for impact of R3/C3 pole.

$$f_{loop} := 10 \cdot 10^3 \quad \omega_p := 2 \cdot \pi \cdot f_{loop} \quad \omega_p = 6.283 \cdot 10^4$$

$$f_{ref} := 1 \cdot 10^6 \quad \omega_r := 2 \cdot \pi \cdot f_{ref} \quad \omega_r = 6.283 \cdot 10^6$$

$$\phi := 50^\circ \quad \phi_r := \left(\pi \cdot \frac{\phi}{180^\circ} \right)$$

$$Kvco := 13 \cdot 10^6 \quad Kvco := 2 \cdot \pi \cdot Kvco \quad Kvco = 8.168 \cdot 10^7$$

$$Kpd := 0.004 \quad Kpd := \frac{Kpd}{2 \cdot \pi} \quad Kpd = 6.366 \cdot 10^{-4}$$

$$N := 560$$

$$T3 := \frac{1}{10 \cdot \omega_p} \quad \omega_p := 1.424 \cdot \omega_p$$

$$T1 := \frac{\sec(\phi_r) - \tan(\phi_r)}{\omega_p} \quad T1 = 4.068 \cdot 10^{-6}$$

$$\omega_c := \frac{\tan(\phi_r) \cdot (T1 + T3)}{[T1 + T3]^2 + T1 \cdot T3} \cdot \left[1 + \frac{[(T1 + T3)^2 + T1 \cdot T3]}{[(\tan(\phi_r) \cdot (T1 + T3))^2]} - 1 \right]$$

$$T2 := \left[\frac{1}{\omega_c^2 \cdot (T1 + T3)} \right] \quad f_c := \frac{\omega_c}{2 \cdot \pi}$$

$$C1 := \left(\frac{T1 \cdot Kpd \cdot Kvco}{N \cdot T2 \cdot \omega_c^2} \right) \cdot \sqrt{\frac{1 + (\omega_c \cdot T2)^2}{[1 + (\omega_c \cdot T1)^2] \cdot [1 + (\omega_c \cdot T3)^2]}}$$

$$C2 := C1 \cdot \left[\left(\frac{T2}{T1} \right) - 1 \right]$$

$$R2 := \frac{T2}{C2}$$

Loop Filter Component Values

$$T1 = 4.068 \cdot 10^{-6}$$

$$T2 = 4.472 \cdot 10^{-5}$$

$$T3 = 1.592 \cdot 10^{-6}$$

$$f_c = 1 \cdot 10^4$$

$$C1 = 6.149 \cdot 10^{-9}$$

$$C2 = 6.146 \cdot 10^{-8}$$

$$R2 = 727.746$$

$$C1 := 5600 \cdot 10^{-12}$$

$$C2 := 0.056 \cdot 10^{-6}$$

$$R2 := 750$$

Rules of thumb for selecting the values of the reference frequency suppression filter are:

1. Choose $C3 < C1/10$

$$C3 : = \frac{C1}{10} \quad R3 : = \frac{T3}{C3}$$

$$R3 = 2.842 \cdot 10^3 \quad C3 = 5.6 \cdot 10^{-10}$$

2. Choose $R3 > 2 \cdot R2$

$$R3 : = 2 \cdot R2 \quad R3 = 1.5 \cdot 10^3 \quad R3 = 1500$$

$$C3 : = \frac{T3}{R3} \quad C3 = 1.061 \cdot 10^{-9} \quad C3 : = 1000 \cdot 10^{-12}$$

$$k = 10..70 \quad kp(k) : = 10^{\left(\frac{k}{10}\right)}$$

$$\omega(k) : = 2j \cdot \pi \cdot kp(k)$$

$$GH(k) : = Kpd \cdot Kvco \cdot \frac{1 + \omega(k) \cdot T2}{(C1 \cdot N \cdot \omega(k) \cdot \overline{\omega(k)}) \cdot (1 + \omega(k) \cdot T1)} \cdot \frac{T1}{T2} \cdot \left[\frac{1}{(1 + \omega(k) \cdot T3)} \right]$$

$$MAGdb_GH(k) : = 20 \cdot \text{Log}(GH(k) \cdot \overline{GH(k)})$$

$$PHI_GH(k) : = \text{atan}\left(\frac{\text{Im}(GH(k))}{\text{Re}(GH(k))}\right)$$

$$f_GH(k) : = \left(\frac{180 \cdot PHI_GH(k)}{\pi} \right) - 180$$

$$MAG_GH(k) : = (GH(k) \cdot \overline{GH(k)}) \quad \text{Phim}(k) : = 180 + f_GH(k)$$

Plots of the magnitude and phase of $GH(k)$ are shown in Figures 12 and 13.

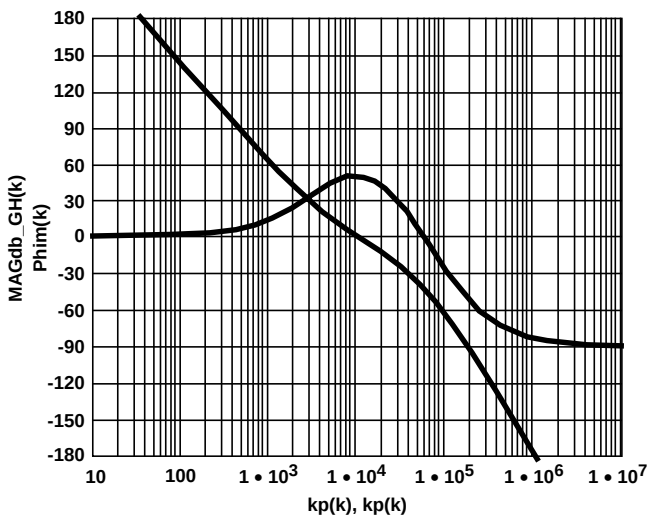


FIGURE 12.

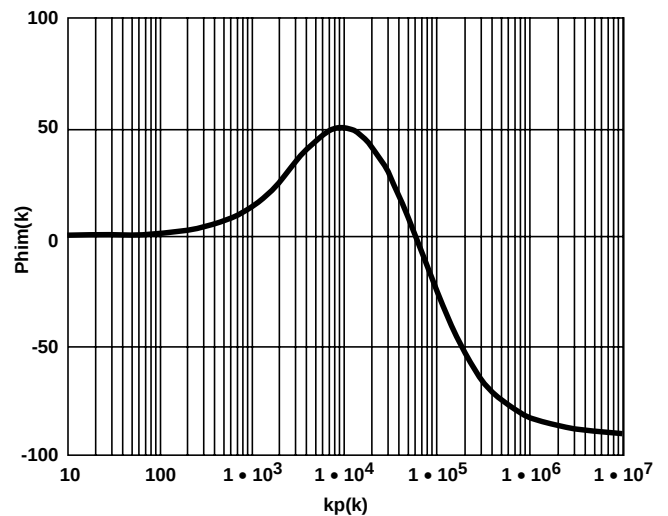


FIGURE 13.

Appendix E Synthesizer Phase Noise/Jitter Analysis

TABLE 4. PRISM 1 LO PHASE NOISE ANALYSIS

PRISM 1 LO PHASE NOISE ANALYSIS				
3/22/96 RDS				
2.1GHz LO MEASURED PERFORMANCE				
FREQ OFFSET	dBc/Hz	NOTES	PHASE JITTER (RADIAN S SQUARED)	PHASE JITTER (RMS DEGREES)
10K	-76.6		2.382E-04	0.884
20K	-81.8		1.243E-04	0.639
50K	-92.4		8.781E-05	0.537
200K	-97.6		1.782E-05	0.242
300K	-103.7		5.511E-06	0.135
400K	-107.6		7.919E-06	0.161
750K	-111.2			
Total			4.816E-04	1.257
560MHz LO MEASURED PERFORMANCE				
10K	-83.5		4.051E-05	0.365
20K	-90.5		1.768E-05	0.241
50K	-102.1		3.865E-06	0.113
100K	-106		1.608E-05	0.230
>200K	-111			
Total			7.814E-05	0.506
"Assume worst case, RMS measured 2.1GHz and 560MHz LO"				
			PHASE JITTER (RADIAN S SQUARED)	PHASE JITTER (RMS DEGREES)
2.1GHz LO			4.816E-04	1.257
560MHz LO			7.814E-05	0.506
Total			5.597E-04	1.355

Appendix F Gain Distribution/IF Limiting Analysis

The IF stage, including the limiters, is of a differential design to improve noise rejection and stability for these high gain stages. The RF front end, on the other hand, is single ended to reduce complexity. A receive chain block diagram is shown in Figure 14.

The minimum limiter 1 and 2 voltage gains are 39dB at 2.7V and 400MHz. The radio design uses the part at a less extreme operating point of 3.5V and 280MHz where the minimum performance is 42dB. The limiter 1 and 2 output limiting voltage is 200mV_{P-P} into a differential 500Ω load. Using 3dB loss in the limiter Bandpass Filter (BPF), the limiter chain (LIM1, BPF, LIM2) cascaded voltage gain is 81dB, with typical performance above 90dB. With a 200mV_{P-P} output, the input limiting voltage is 17.8μV_{P-P} or -98dBm at 250Ω source impedance.

This is calculated as follows: since the source and load impedances are different (250Ω vs 500Ω) the input signal is calculated in terms of voltage. Remember that the limiter is a voltage gain device and so gain is independent of source impedance. Substitute the result of Equation 1 into Equation 2 and calculate V_{IN(P-P)}.

$$\text{gain} = 10^{\left(\frac{\text{GAIN (dB)}}{20}\right)} \quad (\text{EQ. 1})$$

$$V_{\text{OUT}} = \text{gain} \cdot V_{\text{IN}} \quad (\text{EQ. 2})$$

Calculate the input power with a 250Ω impedance by using Equation 3 to get V_{RMS} and then substitute into Equation 4 with R = 250Ω to get power in Watts. Equation 3 assumes a sinewave crest factor for the $\sqrt{2}$ term. Power in Watts is converted to dBm with Equation 5 to get -98dBm input power at 250Ω source impedance.

$$V_{\text{RMS}} = \frac{V_{\text{P-P}}}{2 \cdot \sqrt{2}} \quad (\text{EQ. 3})$$

$$\text{Pwr(Watts)} = \frac{V_{\text{RMS}}^2}{R} \quad (\text{EQ. 4})$$

$$\text{Pwr(dBm)} = 10\log(\text{Pwr(Watts)} \cdot 10^3) \quad (\text{EQ. 5})$$

The limiters have a noise bandwidth of over 500MHz and so the cascaded limiters will fully limit on their own noise, if no BPF is used between the stages. The thermal noise voltage delivered from the 250Ω source to the limiters in a 500MHz band is -87dBm, as calculated from Equation 6. This thermal noise adds to the limiter noise figure (NF) of 7dB resulting in an equivalent input noise power of -80dBm, which is significantly higher than the -98dBm required for limiting.

$$P(\text{Watts})_A = kT\Delta f \quad (\text{EQ. 6})$$

Where:

$$P(\text{Watts})_A = \text{Available Noise Power}$$

$$k = 1.38042 \times 10^{-23}$$

Boltzmann's Constant

$$T = 300 \text{ Degree Kelvin}$$

$$\Delta f = 500\text{MHz}$$

The RF front end 3dB bandwidth is 17MHz, with an estimated noise bandwidth of 20MHz, as defined by the IF SAW filter. This makes the available thermal noise at the limiter input -101dBm and with the 7dB limiter noise figure, is an equivalent -94dBm.

If the limiter BPF was also 20MHz, the front end would only need to supply 7dB of noise floor gain to overcome the limiter noise figure. This would result in a receiver that limits in a 20MHz bandwidth from front end noise with no margin. The 20MHz limiter BPF would require a second SAW filter and therefore is not cost effective or practical.

The alternative chosen to be implemented is a simple one pole LC BPF with a bandwidth wide enough so that the variability of fixed components do not result in the filter being off frequency. The filter selected has a 3dB bandwidth of 50MHz, and an estimated noise bandwidth of 100MHz. Using this method, the front end gain must be increased to compensate for excess limiter bandwidth.

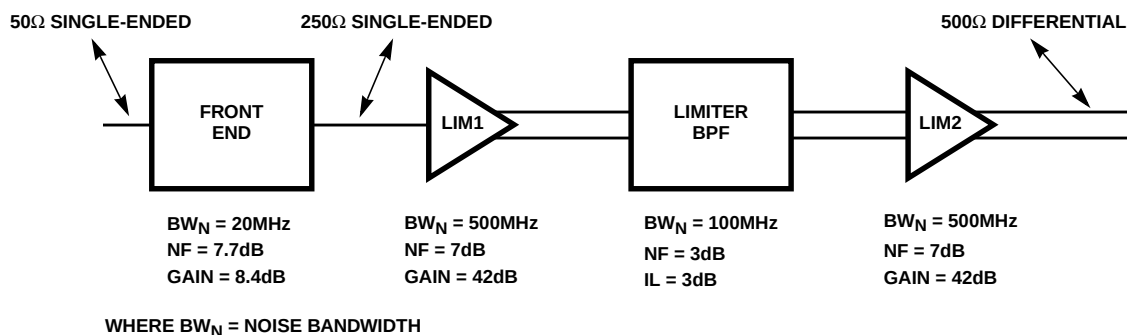


FIGURE 14. RECEIVE CHAIN GAIN DISTRIBUTION DIAGRAM

It is desired that the second limiter to be fully limited on front end noise, as opposed to noise generated in the first limiter. This requires that the front end noise floor must be greater than the sum of the following; the limiter NF of 7dB, the ratio of limiter BPF noise bandwidth to front end IF SAW bandwidth ($10\log(100\text{MHz}/20\text{MHz})$ or 7dB), and the amount of limiting margin (6dB for -1dB limiting). The front end output noise floor must therefore be greater than 20dB.

The limiting margin was measured on the HFA3724 IF Mod/Demod, with good agreement to a theoretical estimate based upon the hyperbolic tangent response of a bipolar differential limiter. The measurement means that if a non-desired jamming signal, noise in this case, is 6dB below the level of the desired signal, the desired output will be reduced 1dB.

The front end noise floor was previously calculated for a 20MHz bandwidth as -101dBm, if the required gain is now added, this noise floor becomes -81dBm as shown in Equation 7. Now verify that this level is larger than the minimum signal power required to limit the limiter chain and guarantee that the radio limits on front end noise (-81dBm > -98dBm).

$$(20\text{MHz kTB}) + \text{NF}_{\text{LIM}} + \text{BPF}_{\text{Noise}} + \text{Lim}_{\text{Margin}} = \text{Front End Noise} \quad (\text{EQ. 7})$$

i.e.

$$-101\text{dBm} + 7\text{dB} + 7\text{dB} + 6\text{dB} = -81\text{dBm}$$

Where;

(20MHz kTB) is Available Thermal Noise

NF_{LIM} is Noise Figure of Limiter Chain

$\text{BPF}_{\text{Noise}}$ is Added Noise due to Limiter BPF Bandwidth wider than Front End

$\text{Lim}_{\text{Margin}}$ is Front End Margin required to guarantee limiter jammer rejection

The actual front end gain is 13.4dB and NF is 7.7dB for a front end output noise floor increase of 21.1dB typical, which is better than desired 20dB. This results in limiting, when no signal is present, on mainly front end noise, but also some limiter broadband noise. This limiter output, although still fully limited, when band filtered will have a slight drop in baseband quadrature output voltage to the HFA3824A ADCs as compared to totally limiting on front end noise due to some of the limiter noise being out of band. The net impact to the system is a small reduction in sensitivity due to reduced ADC signal to full scale. If desired, additional front end gain, or a reduced bandwidth limiter BPF could be used to gain back performance.

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Harris AnswerFAX (407) 724-7800.

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