LightSquared base station location

Emissions exceed FAA TSO C145/146/196 test level (-86.4 dBm)

>1-dB SNR degradation for Receiver #2 (-64 dBm)

Maximum interference level = -36.1 dBm

Figure 6-20. Initial LightSquared Deployment (2391 of 40000+ Towers) Aircraft at 500'

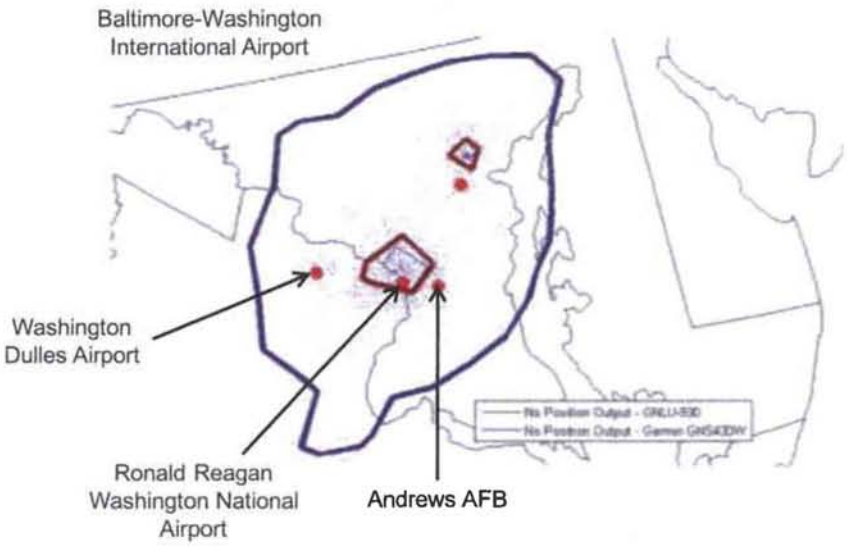

Figure 6-21. Initial LightSquared Deployment (2391 of 40000+ Towers) Aircraft at 500' (Zoom View above Baltimore-Washington)

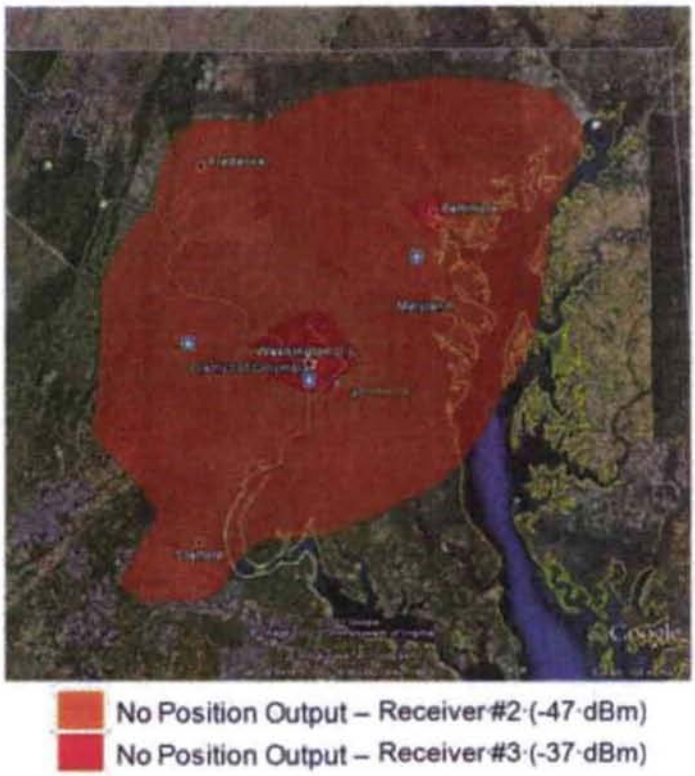

Figure 6-22. Initial LightSquared Deployment Aircraft at 500'

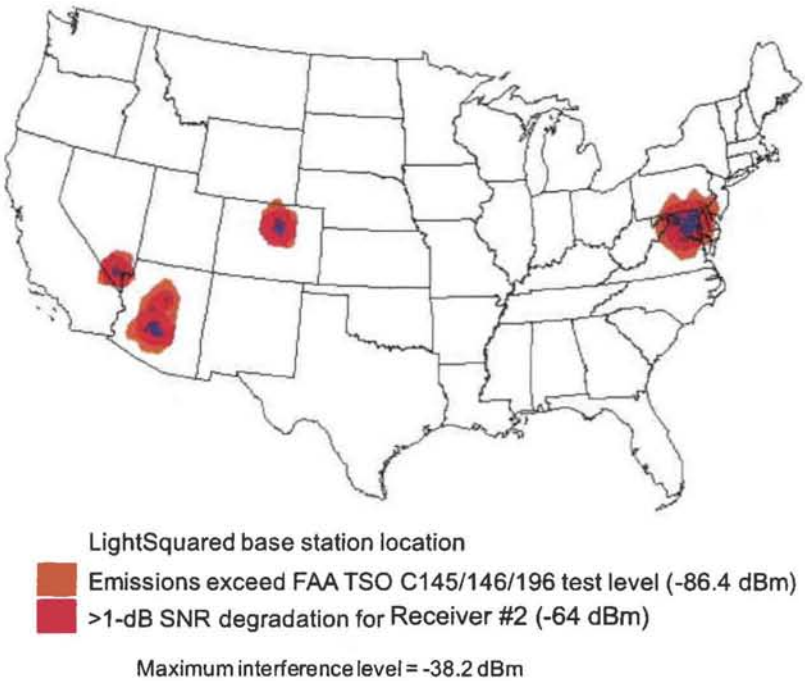

Figure 6-23. Initial LightSquared Deployment Aircraft at 1,000'

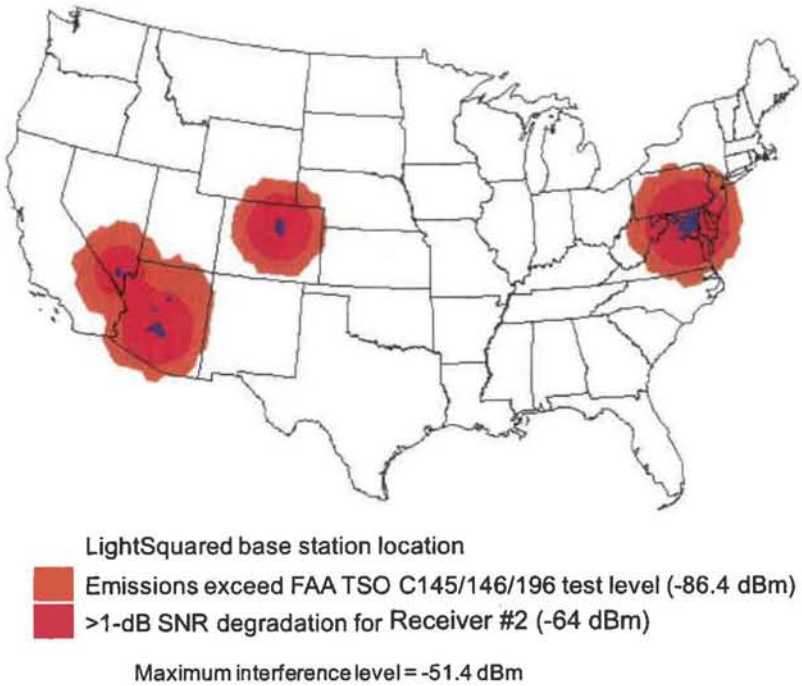

Figure 6-24. Initial LightSquared Deployment Aircraft at 10,000'

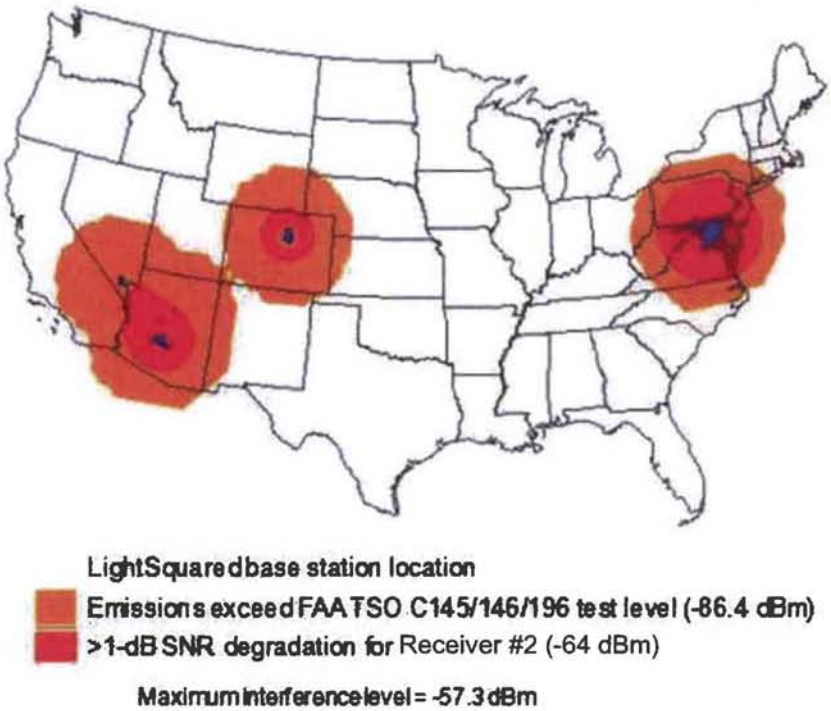

Figure 6-25. Initial LightSquared Deployment Aircraft at 20,000'

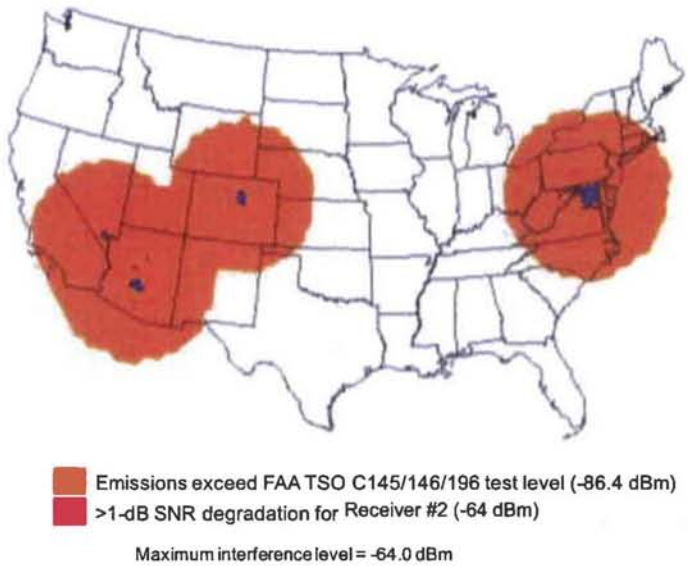

Figure 6-26. Initial LightSquared Deployment Aircraft at 40,000'

Intermodulation Product Simulation

Overview

This report provides an assessment of intermodulation products that may arise in some GPS receivers due to high-powered LightSquared ancillary terrestrial component (ATC) base station emissions driving low noise amplifiers (LNAs) within the receiver into saturation.

LNA Model

Consider the simple LNA system model shown in Figure 6-27. The LNA takes an input voltage, *x(t)*, which is typically the filtered output of a passive antenna element, and provides an output voltage, *y(t)*, with a nominal power gain of *G*. This note focuses on a typical airborne active antenna LNA that provides a nominal power gain of 34.5 dB[2].

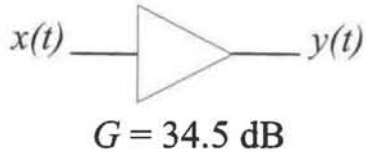

Figure 6-27. Low Noise Amplifier System Model

[2] Per RTCA DO-301, the overall active antenna amplifier subassembly must provide a minimum gain of 26.5 dB and a nominal gain of 29.5 dB. The nominal gain of 34.5 dB, used here for the LNA subcomponent, presumes 5 dB of losses for, e.g., pre-/post-selection filters and burnout protection circuitry.

For an ideal LNA, the input-output voltage characteristics may be described as:

$$y = a_1 x \tag{1}$$

where $a_1 = \sqrt{G}$ with G being the nominal power gain .

As is well-known, actual LNAs are only well-modeled by (1) for small input voltages. For larger input voltages, the output voltage saturates. A truncated Taylor series expansion is often used (see, e.g., [1]) as a more accurate model:

$$y = \sum_{i=1}^{N} a_i x^i \tag{2}$$

For example, Figure 6-28 shows the voltage input-output characteristics of an LNA modeled using equation (2) with $N = 5$ and the following coefficients:

$$\begin{aligned} a_1 &= 53.088 \text{ (unitless)} \\ a_2 &= 0 \\ a_3 &= -997490 / R \text{ (volts}^{-2}\text{)} \\ a_4 &= 0 \\ a_5 &= 6.5e9 / R^2 \text{ (volts}^{-4}\text{)} \end{aligned} \tag{3}$$

Where $R = 50\ \Omega$ is the resistance assumed to relate voltage to power. The a_1 coefficient was selected as $a_1 = \sqrt{G}$ to provide a nominal gain of 34.5 dB. The a_2 and a_4 coefficients were selected as zero to provide an odd-symmetric input-output voltage characteristic.

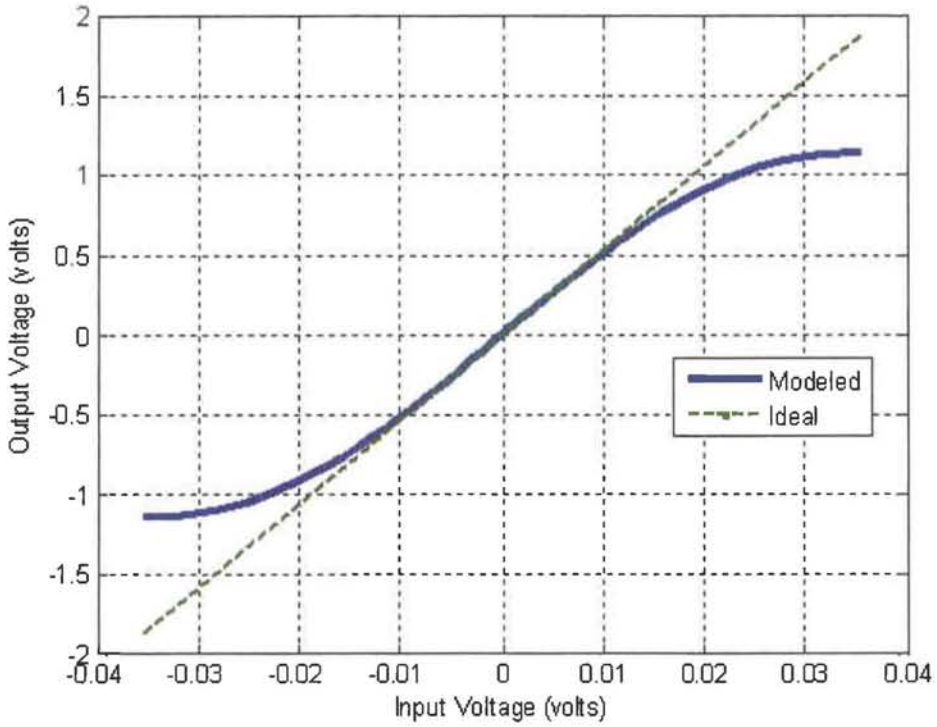

Figure 6-28. Input-Output Voltage Characteristic for Modeled Airborne Active Antenna LNA

a_3 was selected to achieve a representative *1-dB compression point* and *third-order intercept point*. The 1-dB compression point, P_1, is either the input (*input 1-dB compression point*) or output (*output 1-dB compression point*) power level at which the LNA provides 1-dB less gain than an ideal LNA with the same nominal gain value. Figure 6-29 shows the input-output power characteristics of the modeled LNA. Airborne antenna active subassemblies are required to have an input 1-dB compression point above -25 dBm within the passband. These subassemblies are defined to include protection circuitry and a preselector filter between the passive antenna output port and LNA input port. The modeled LNA is consistent with this requirement, providing an input 1-dB compression point of -22.2 dBm (see Figure 6-29), which would provide an input 1-dB compression point above -20 dBm for the overall active subassembly presuming combined insertion losses of greater than 2.2 dB for the preselector filter and protection circuit.

The concept of the third-order intercept point is explained, e.g., in [1]. A typical LNA has a third-order intercept point, P_3, which is 10 – 15 dB above its 1-dB compression points (provided that both are consistently referenced to either the input or output). The magnitude of the a_3 coefficient in equation (3) was selected using the formula [4]:

$$P_{3,output} = \frac{2a_1^3}{3|a_3|} \tag{4}$$

and a target $P_{3,input}$ value that was set to 10 dB above a target $P_{1,input}$ value of -24.5 dBm. The sign of a_3 was chosen to be negative, since this is required for gain suppression rather than gain

enhancement at higher input voltages. Using this selection process, the LNA model perfectly provides the target P_3 value. However, it is only fortuitous that the achieved P_1 value of -22.2 dBm is close to the target value of -24.5 dBm. The final coefficient, a_5, was selected to make the output voltage stay as close as possible to "flat" for high input voltage magnitudes.

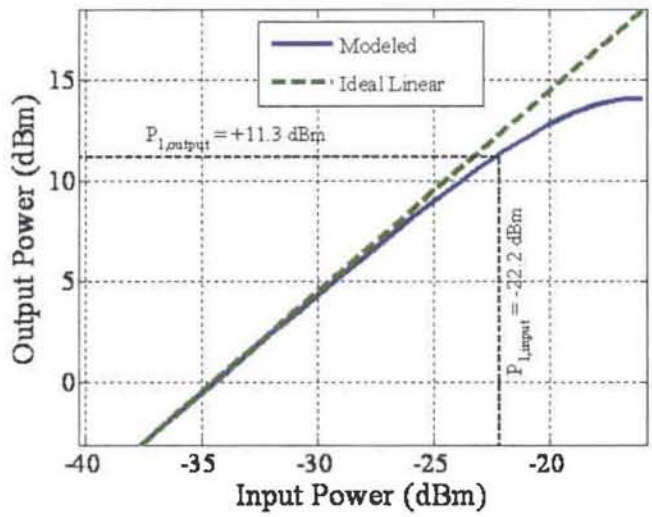

Figure 6-29. Modeled LNA Input-Output Power Characteristics

Response of LNA to LightSquared Emissions

Noting that the input-output characteristics are virtually unchanged for input power levels below -20 dBm by truncating the Taylor series to $N = 3$, here we focus on the simpler LNA input-output voltage model:

$$y = a_1 x + a_3 x^3 \tag{5}$$

with the input voltage presumed to be well-modeled as a Gaussian, wide sense stationary random process.

The output voltage autocorrelation may be determined as:

$$\begin{aligned} R_y(\tau) &\equiv E\left[y(t+\tau)y(t)\right] \\ &= E\left[\left(a_1 x(t+\tau) + a_3 x^3(t+\tau)\right)\left(a_1 x(t) + a_3 x^3(t)\right)\right] \\ &= \left(a_1^2 + 6a_1 a_3 \sigma_x^2 + 9a_3 \sigma_x^4\right) R_x(\tau) + 6a_3^2 R_x^3(\tau) \end{aligned} \tag{6}$$

where $R_x(\tau)$ is the input voltage autocorrelation and $\sigma_x^2 = R_x(0)$ is the variance of *x(t)* (i.e., input power aside from a possible scale factor for a non-unity resistance).

From equation (6), the following expression may be derived to relate the output power spectrum, $S_y(f)$, of the LNA to its input power spectrum, $S_x(f)$:

$$S_y(f) = \left(a_1^2 + 6a_1a_3\sigma_x^2 + 9a_3\sigma_x^4\right)S_x(f) + 6a_3^2 S_x(f) * S_x(f) * S_x(f) \qquad (7)$$

where * is the convolution operator.

Results with No Preselection Filtering

Figure 6-30 through Figure 6-32 show the LNA input and output power spectra for various LightSquared received power levels, presuming no filtering prior to the LNA. Phase 1 emissions are assumed, and these emissions are very simply modeled as perfectly rectangular 5-MHz blocks centered at 1528.8 MHz and 1552.7 MHz with total input power as indicated on each figure. An input noise floor of -201.5 dBW/Hz is also included[3].This truly would not fully present at the LNA input even for a receiver with this effective N_0 value, since some of the effective N_0 is due to the LNA noise figure.

In Figure 6-30, the LightSquared signal at the input of the LNA is at a power level of -45 dBm, which is much greater than the power from the thermal noise floor. This input power level is well below the LNA's 1-dB input compression point, and the LNA is operating very linearly. The output appears to be perfectly identical to the input, with the exception that the output power level is 34.5 dB (the nominal gain value) greater than the input.

In Figure 6-31 and Figure 6-32, the LightSquared input levels are increased to -35 dBm and -25 dBm, respectively. The LNA is being driven closer to its 1-dB compression point, and the output signal is seen to be increasingly distorted with a significant third-order intermodulation product clearly visible centered at 2 × 1552.7 – 1528.8 = 1576.6 MHz.

[3]This level would not truly be present at the LNA input even presuming that the overall front-end had precisely this effective N_0 value, since a good portion of the effective N_0 arises from the LNA noise figure. Nonetheless, ignoring rigor in this area does not materially affect the results presented here.

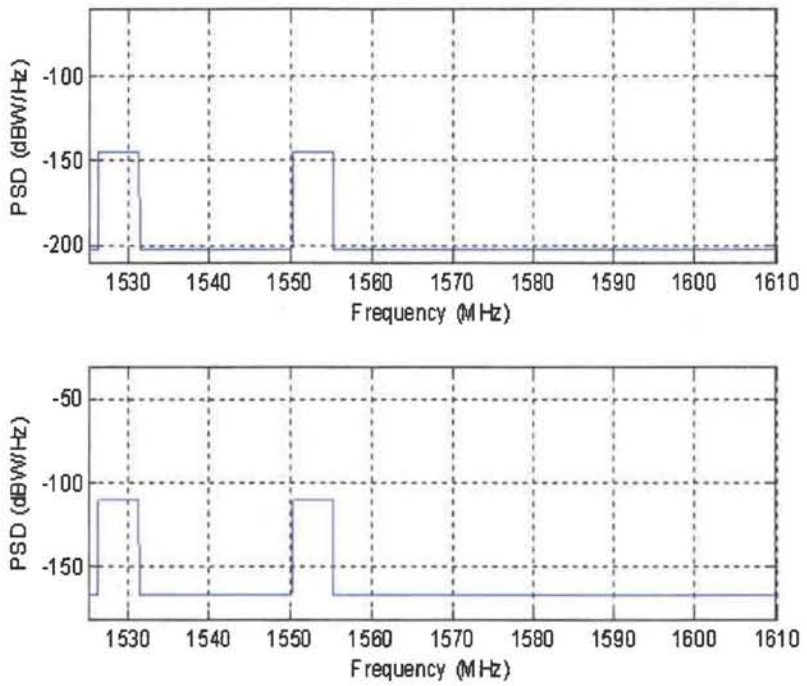

Figure 6-30. Input (top) and Output Power Spectrum of Modeled LNA, for Input Power of -45 dBm (LNA in Linear Region)

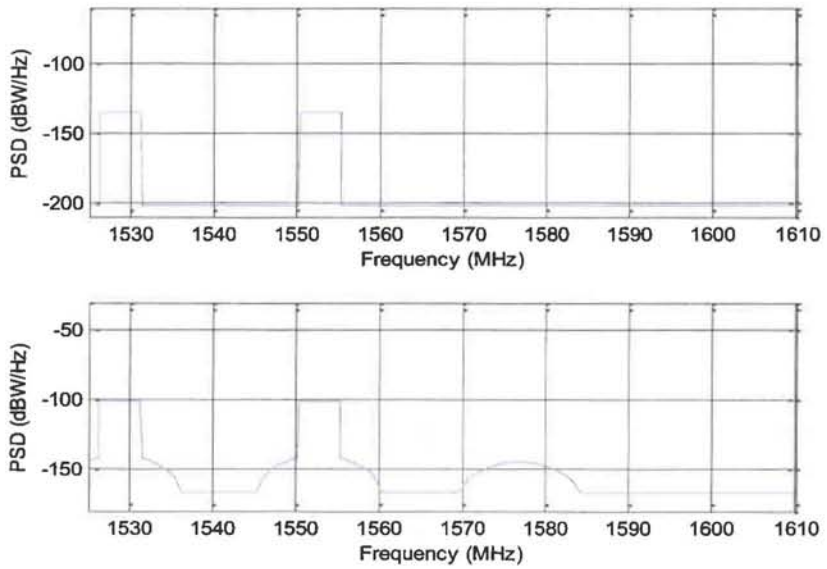

Figure 6-31. Input (top) and Output Power Spectrum of Modeled LNA, for Input Power of -35 dBm (LNA Entering into Compression)

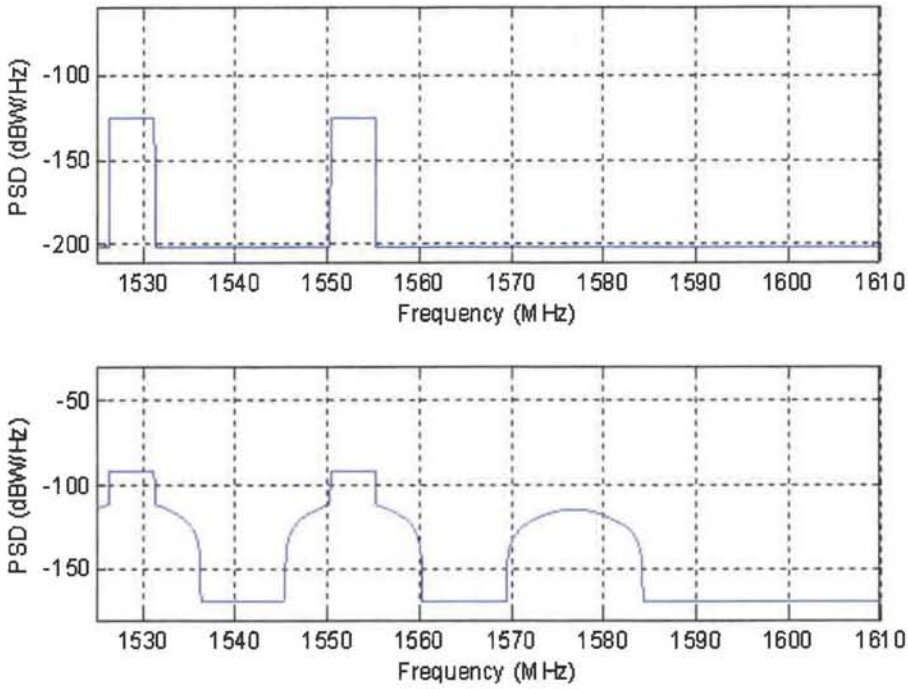

Figure 6-32. Input (top) and Output Power Spectrum of Modeled LNA, for Input Power of -25 dBm (LNA Nearly to 1-dB Compression Point)

Results with Preselection Filtering

Preselection filtering can significantly diminish the strength of the third-order intermodulation product. For example, a representative 3-pole ceramic filter provides the attenuation characteristics shown in Figure 6-33. This particular filter has an insertion loss of 2.2 dB, and for the results to follow an additional 0.5 dB insertion loss for protection circuitry is presumed to be present between the passive antenna output port and LNA input. With this configuration, Figure 6-34 through Figure 6-36 show input-output power spectra for the modeled LNA using the same remaining assumptions as were used for Figure 6-30 through Figure 6-32.

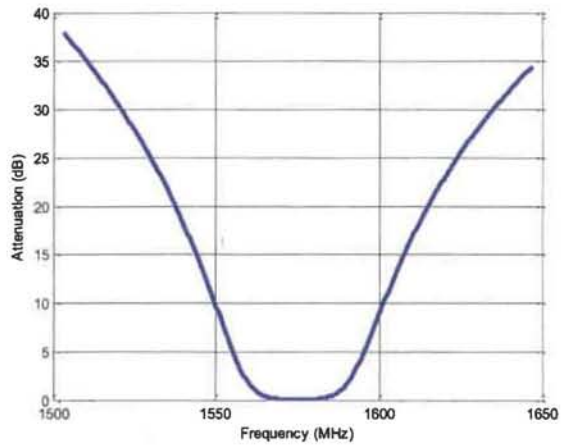

Figure 6-33. 3-Pole Ceramic Preselection Filter Attenuation (24 MHz 1-dB Bandwidth)

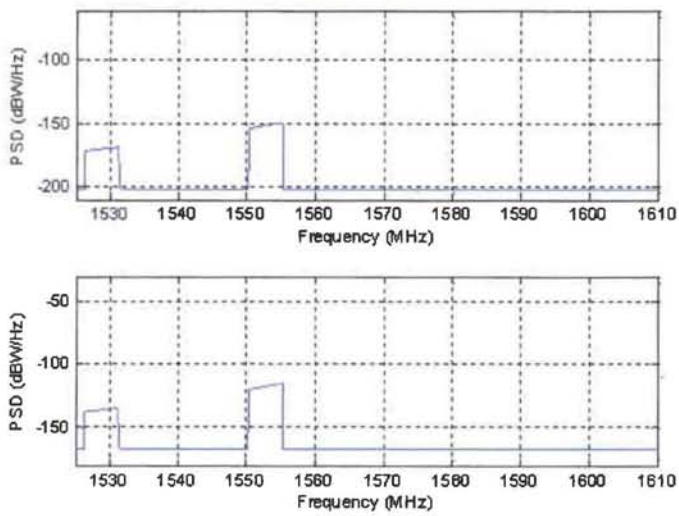

Figure 6-34. Input (top) and Output (bottom) Power Spectrum of Modeled LNA; LightSquared Signal Power at Antenna Output Port is -45 dBm, LNA Input Power is -57.5 dBm (LNA in Linear Region)

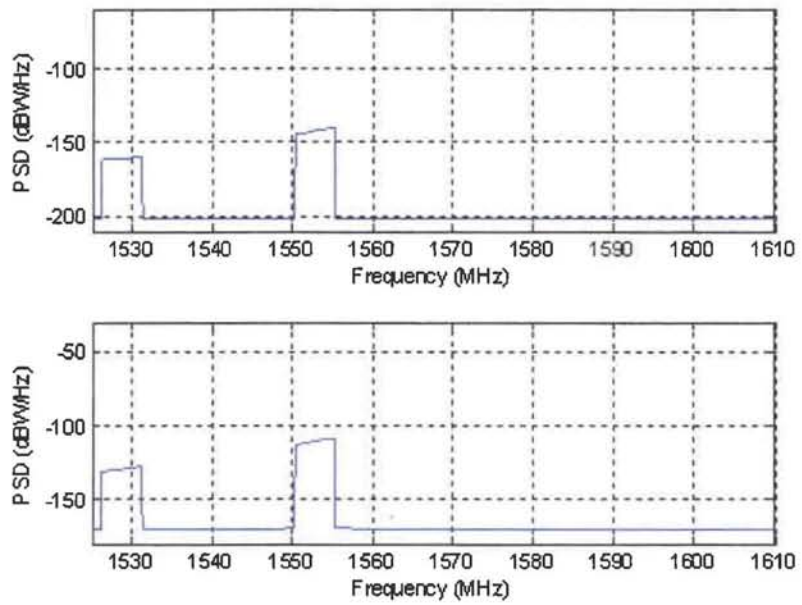

Figure 6-35. Input (top) and Output (bottom) Power Spectrum of Modeled LNA; LightSquared Signal Power at Antenna Output Port is -35 dBm, LNA Input Power is -47.6 dBm (LNA Lightly Entering into Compression)

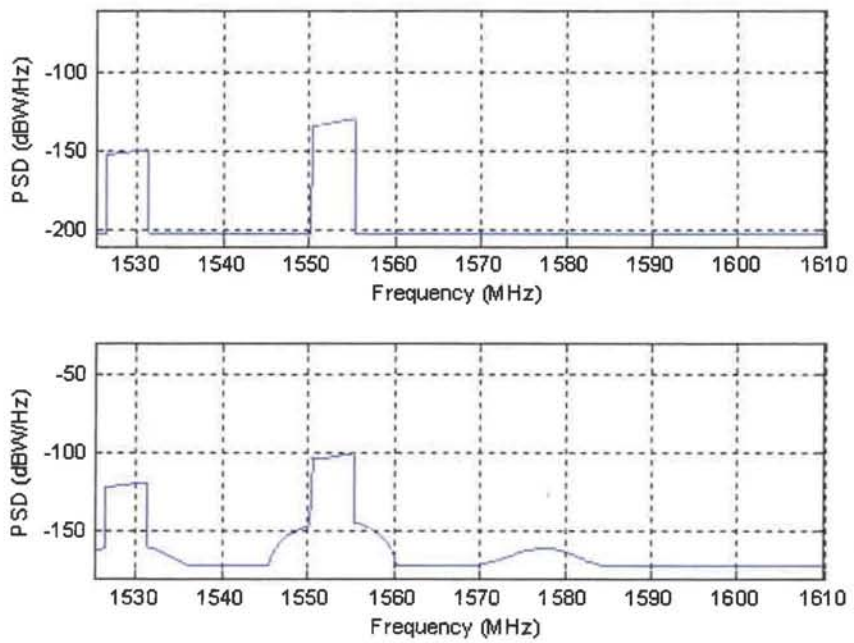

Figure 6-36. Input (top) and Output (bottom) Power Spectrum of Modeled LNA; LightSquared Signal Power at Antenna Output Port is -25 dBm, LNA Input Power is -37.6 dBm (LNA Entering Deeper into Compression)

Figure 6-37 shows the degradation to the effective N_0 for a C/A-code receiver due to the third-order intermodulation product, presuming that the remainder of the receiver front-end is perfectly linear and perfectly bandlimited to 20 MHz[4]. The plot was produced by computing the inner product of the LNA output power spectrum, referred back to the passive antenna port by scaling by the true LNA gain and assumed 2.7 dB insertion loss (for the preselector filter and protection circuitry), against the normalized power spectrum of the C/A-code. The inner product was computed over L1 +/-10 MHz under the presumption that the fundamental LightSquared emissions would be suppressed completely by later filtering within the receiver. A 0.5 dB degradation is seen at a LightSquared received power level of -31.5 dBm.

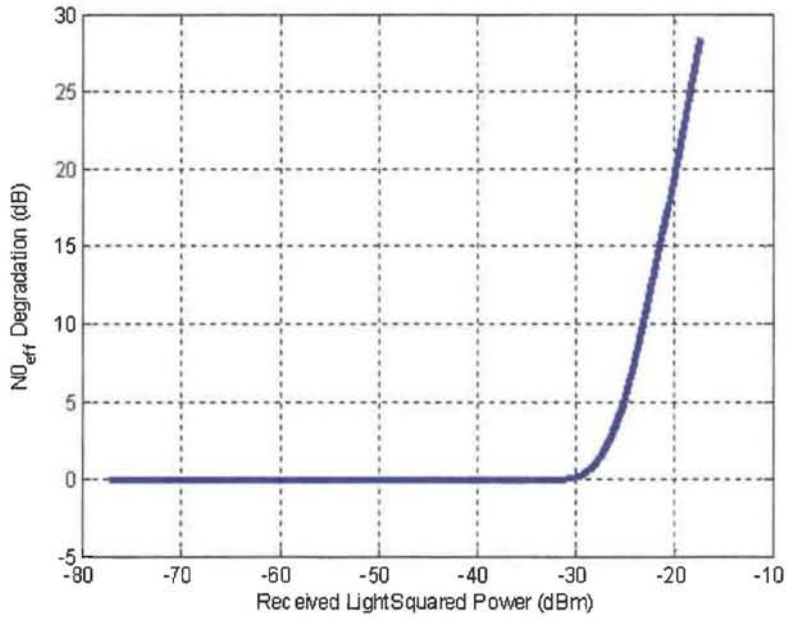

Figure 6-37. Degradation to Receiver Effective Noise Floor due to the Third Order Intermodulation Product

Comparison with Simulations and Measurements

Figure 6-36 looks very similar to a simulation result produced for the same Phase 1 emission scenario, but for a commercially available GPS receiver front-end module, see Figure 6-38 (from [2]). To produce this figure, the author of [2] created a time-domain simulation of the LightSquared Long Term Evolution (LTE) signals and fed these signals through a time-domain model of the front-end module.

[4]These are obviously big assumptions that are not likely to be true for most fielded receivers. In fact, it is the authors' view that saturation is far more likely in later receiver components such as LNAs or mixers within the receiver before appreciable attenuation of the fundamental LightSquared emissions is provided by distributed filtering (if at all).

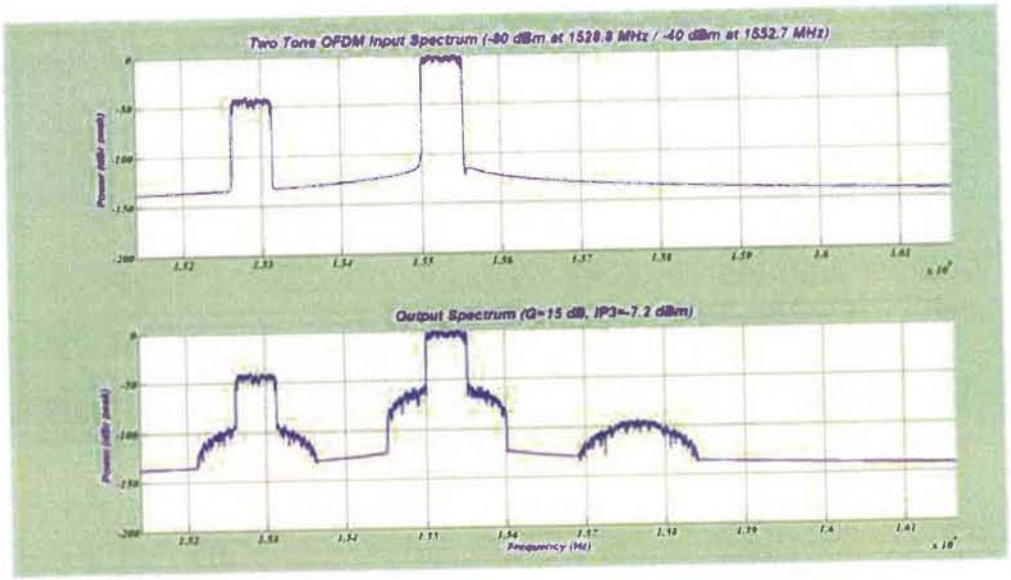

Figure 6-38. .Maxim MAX2741 Input (top) and Output (bottom) Power Spectra based Upon Simulations [2]

The model-predicted power spectra are also similar to those measured at the output of a GPS LNA from a live LightSquared base station. Figure 6-39 shows a measurement that was made by Jet Propulsion Laboratory (JPL) personnel at the Holloman Air Force Base Live Sky test event. A third-order intermodulation product is clearly visible in this spectrum analyzer screen-shot with some similar features as observed in the model results earlier in this note.

Figure 6-39. Measured Output Power Spectrum of LNA as Measured by JPL at the Holloman Live Sky Test Event

There is, however, one noticeable difference between the measured power spectra and the model results. The spectrum of the measured third-order intermodulation product is not as broad in bandwidth as that predicted by the model described in this note, or as was seen in the simulation results in [2]. This difference cannot be attributed to the fact that the LNA used for the measurement was somehow different than the one that was modeled in this note or within [2]. Note that regardless of the "tuning" of the LNA model described in this note (i.e., choice of particular Taylor series coefficients to match the actual LNA's input-output voltage characteristics), the third-order intermodulation products are always predicted to have a power spectrum shape that looks like one of the two fundamental emissions convolved with itself three times, i.e., a bandwidth that is three times larger than each fundamental emission. The live measurement intermodulation product, on the other hand, appears to have a power spectrum that is only as wide as each of the fundamental emissions near its peak.

One potential, speculative explanation is that the two carriers were produced using the same "dummy" data so that the baseband signal for each carrier was identical, or that there is some inherent characteristic of dual-carrier LTE signals that violates the assumptions made in deriving the model described in this note. This topic will be the subject of further study.

References

[1] Pozar, David M., *Microwave Engineering*, 3rd Edition, Hoboken, New Jersey: John Wiley & Sons, 2005.

[2] Scott, Logan, "Comments on LightSquared Ancillary Terrestrial Component (ATC) Interference Potential," GPS World Market Insights Webinar, 21 April 2011.

7. Subtask 7 - Work Plan, Test Planning, and Field Test Activities

Task Statement

Coordinate work plan, test planning, and field activities with the FAA, LightSquared, NTIA, and the EXCOM departments and agencies to measure emissions and determine representative technical and operational GPS receiver effects as a function of distance from a LightSquared terrestrial base station.

Overview

The LightSquared test strategy was designed to investigate possible incompatibilities between LightSquared ATC and GPS User Equipment. The NPEF performed various conductive tests as well as an Anechoic Chamber and Live Sky test to investigate these concerns. In general, the test team modulated the power of the LightSquared signal to replicate varying distances from a single ATC tower.

Conducted emissions tests were performed at Zeta, JPL, SPAWAR and 746th Test Squadron Laboratories. LightSquared was present for all unclassified tests and provided their ATC Emission Mask Filters for these events.

The Army Research Laboratory (ARL) EMVAF, located at WSMR, houses an Anechoic Chamber measuring 110' x 70' x 40'. This chamber was large enough to house the military and commercial user equipment, an antenna platform containing all UE antennas spaced so no antenna shielded any other from the GPS or LightSquared signal. The WSMR EMVAF Anechoic Chamber housed the Advanced Global Navigation Simulator (AGNS), GPS UE and antennas, and the LightSquared ATC Tower signal simulator and hardware. LightSquared was present for this test to comment on the setup and provide technical guidance.

Following anechoic chamber testing, the team transitioned test operations to an open air venue known as the "Balloon Pad" on the western portion of Holloman Air Force Base (HAFB). The 870 ft x 470 ft asphalt surface Pad provided sufficient level area for a test command center tent and even spatial distribution of the participating test organizations, their equipment and mobile diesel generators. This test was performed in close coordination with LightSquared and utilized actual LightSquared ATC equipment.

The approved details of the two tests are contained in the Anechoic Chamber Test Plan and the Live Sky Test Plan. These plans were coordinated with the NPEF, LightSquared, NTIA and FCC.

8. Subtask 8 – Identify & Evaluate Potential Mitigations for GPS Applications

Task Statement

Assess potential mitigation techniques and their expected effectiveness/costs for various representative GPS receivers in each of the selected scenarios. Assessments should include analysis, simulation, and prototype testing (as practical).

Overview

This task report addresses possible mitigation techniques applicable to GPS equipment and the GPS constellation. Four mitigations are considered: the use of additional filtering, adaptive antennas, GPS system changes, and operational solutions. The last section addresses costs associated with each of these mitigation approaches.

Additional Filtering

Fielded GPS receivers were, for the most part, designed with the expectation that the 1525 – 1559 MHz band would continue to be "quiet", as it has been occupied for many years primarily by low-level (as seen by GPS users on or near the Earth's surface) downlink signals for satellite communications. Although all known fielded GPS receivers include some filtering, the amount of attenuation provided by this filtering towards LightSquared emissions in the 1525 – 1559 MHz band is, in many cases, quite limited. This subsection considers the inclusion of additional filtering as a possible mitigation.

Filter Characteristics

Important filter characteristics for GPS applications are described in the following subsections. As will be discussed in the Filter Technologies section below, selecting from available filter technologies often involves making tradeoffs among these characteristics.

Type

Bandpass and low pass filters are most commonly used for GPS applications. As illustrated in Figure 8-1, ideal bandpass and low pass filters allow only a range of frequencies to pass through and completely suppress signals outside of this frequency range. For bandpass filters, the frequency range spans between two positive frequency values. Bandpass filters are used within GPS receivers or active antennas to pass the desired radiofrequency (RF) GPS signals before downconversion (e.g., at L1 ± 15 MHz for a receiver wishing to process all of the L1 signals), and may also be used at an intermediate frequency (IF) after downconversion. Low pass filters are often found after downconversion within receivers that utilize a low or no IF, or at times even with a high IF to suppress harmonics generated within the mixers used for downconversion.

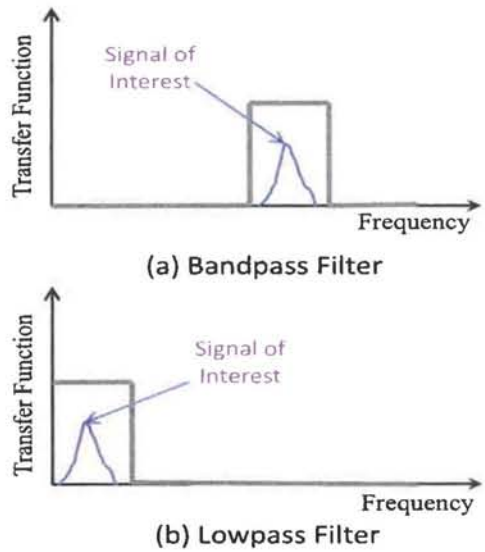

Figure 8-1. Ideal Bandpass and Low Pass Filters

Selectivity

Selectivity is the amount of attenuation that is provided by the filter towards undesired signals (i.e., those that fall outside of the filter passband). Although as illustrated in Figure 8-1 ideal filters completely suppress signals outside of the passband, as will be seen in typical characteristics to be presented throughout this section, realizable filters can only provide a finite amount of attenuation. Achievable attenuation in realizable filters typically is generally small close to the passband, and increases as frequency separation from the passband grows.

The passband is selected based upon the signals that are intended to be processed. Figure 8-2 depicts the current and future GPS L1 signals. Although 90% of the C/A-code power is contained within L1 ± 1 MHz, importantly, the ability of a GPS receiver to precisely range upon the C/A-code or any other GPS signal is enhanced tremendously in the presence of noise and multipath by additionally processing the sidelobes (see, e.g., [8-1]). Modern high-precision receivers generally utilize the full bandwidth of the signals transmitted by the GPS satellites (up to L1 ± 20 MHz for some of the operational satellites) and often use a passband that is even broader than this for reasons related to group delay (see discussion in the Group Delay section) and/or the desire to track signals that are or will be broadcast by other satellite navigation systems in this band (e.g., GLONASS, Galileo, COMPASS, satellite-based augmentation systems [SBAS], QZSS).

Selectivity is generally described using the magnitude of the transfer function, $|H(f)|$, of the filter, which is the ratio of the filter output spectrum to the spectrum of its input. Decibel (dB) units are typically used, with attenuation in dB equal to $20\log_{10}|H(f)|$. For RF filters, vendors often provide scattering matrix parameters (S-parameters) with the S_{21} parameter providing the transfer function.

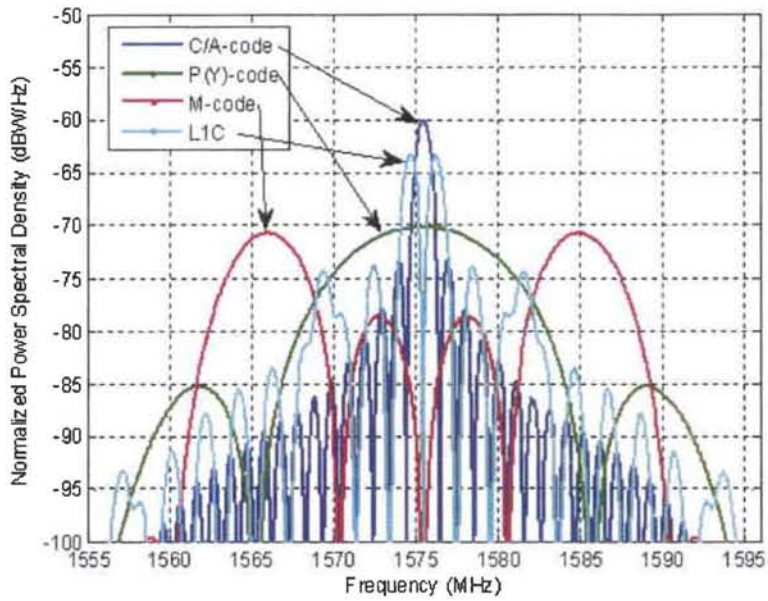

Figure 8-2. Power Spectra of GPS L1 Signals

Insertion Loss

Realizable analog filters will always provide some undesired attenuation of signals in the passband, which is referred to as *insertion loss*. Minimizing insertion loss is important, especially for any filter prior to the first significant gain stage within a GPS receiver's front-end since this filter characteristic impacts the receiver's noise floor. In a benign environment, GPS receivers see a noise floor that is due to:

- Undesired energy received from the antenna
- Undesired energy from sources internal to the receiver/antenna, e.g., due to thermal agitation of electrons within the antenna/receiver components.

Additional filtering will always increase the noise floor, and the extent to which this occurs can be quantified using the expressions:

$$N_0 = kT_{sys} \tag{8-1}$$

$$T_{sys} = T_s + T_0\left[L_1 - 1 + L_1\left[NF_1 - 1 + G_1^{-1}\left[L_2 - 1 + ...\right]\right]\right] \tag{8-2}$$

where N_0 is the noise density (in units of W/Hz) referenced to the output port of the passive antenna, k = 1.38E-23 J/K is Boltzmann's constant, and T_{sys} is the system temperature (in units of K). T_s is the source or antenna temperature (75 – 100 K for a typical GPS antenna that provides a broad gain pattern, i.e., the upper hemisphere), T_0 is 290 K, and the second term in the right-hand-side of equation 8-2 is the receiver temperature. Of importance to the present discussion,

the receiver temperature is influenced greatly by the first loss (with loss, in linear units, L_1) suffered between the output port of the passive antenna and the first gain stage (with gain, in linear units, G_1) in the receiver, as well as the noise figure, NF_1, of the first amplifier. Losses and noise figures of components further downstream in the receiver front-end is of lesser importance if the gain, G_1, is sufficiently high. Equation 8-2 is recursive, in that its form repeats for additional losses, gains and noise figures.

For the reasons described above, it is desirable for any filtering prior to the first low noise amplifier (LNA) within a GPS receiver front-end to have extremely low insertion loss. Insertion loss adds to the noise. Typical target design values can range from under 2.2 dB for aviation receivers with a clear view of the sky to less than 0.5 dB for some other applications. Filters with higher insertion losses can often be tolerated later within the RF/IF chain provided that the preceding net gain far outweighs the preceding net loss. So for instance, filters with insertion losses of up to 15 – 20 dB may be found at IF in numerous fielded receivers with little detrimental impact on receiver noise floor.

Group Delay

The phase response of a filter is also of great importance for many applications. Any phase response within the passband that is not linear with frequency will distort the desired signals. The derivative of the phase response with respect to frequency is referred to as the *group delay response* because this function of frequency describes how much time delay is incurred upon each frequency component of the desired signal.

For navigation and positioning applications, the absolute value of the group delay is not consequential since it does not affect position accuracy. For such applications, the group delay differential, which describes how much the group delay varies over the passband, is the critical characteristic. It is desirable to keep the group delay differential as small as possible over the passband to enable better positioning performance. This design goal is especially important for receivers that make measurements from more than one satellite navigation system within a band, see, e.g., [8-2]. Receivers used for time transfer, ionospheric mapping, and other science applications also require the group and phase delays to be stable over variations in temperature.

Group delay differential generally grows with increasing filter selectivity, and the frequencies where maximum group delay differential is typically seen are in the transition region between the passband and stopband. For these reasons, many fielded high-precision receivers use passbands that extend beyond the 1559 – 1610 MHz radionavigation satellite services (RNSS) band. It is also possible for some filter technologies to use specialized designs to provide delay compensation to minimize group delay differential. The use of narrow filters with sharp cutoffs, as well as some implementations of delay compensation, increases the variation of group delay and phase versus temperature.

Size, Weight, and Cost

Size, weight, and cost are of obvious importance to ensure that the filters can fit within the form factor and weight allowance for the receiver and antenna, and are additionally affordable.

Filter Technologies

As described in the following sections, a wide variety of filtering technologies are commonly used for GPS receiver applications. These technologies are described and typical and specified performance characteristics are presented. As a caveat on the typical performance characteristics, it is important to note that components can only be relied upon to meet specifications not "typical" values to allow for manufacturing and temperature variations. The relevance of specifications over typical performance is particularly true for filters with sharp cutoff transitions for which small variations can have very significant effects. The specifications allow for margin to account for such effects.

Dielectric Resonators

Dielectric resonators are a very popular technology for GPS RF and, occasionally, IF filters. These filters use a small disc or cube of low-loss high dielectric constant material as a microwave resonator to provide a low-cost, high-selectivity bandpass response [8-3]. They are also often referred to as *ceramic filters*, since ceramic is a common dielectric material used in their fabrication.

Figure 8-3 shows the selectivity of representative commercially available 3- and 4-pole dielectric resonator filters for GPS L1[5]. Both filters provide a minimum passband of 24 MHz as defined by the range of frequencies between 1-dB attenuation points. The maximum insertion losses of the 3-pole and 4-pole filters are 2.2 dB, and 3.0 dB, respectively. The 3-pole filter size is 0.80 × 0.55 × 0.27 inches, and the 4-pole filter size is 1.04 × 0.55 × 0.27 inches, both in a leadless surface mount package. A smaller form factor is available, with about one-half the volume for each, but at the cost of increased insertion loss (+0.7 dB for each) and poorer selectivity characteristics. Very little attenuation (4 – 6 dB) is provided for the upper LightSquared carrier, which extends to 1555.2 MHz. Greater attenuation (20 – 30 MHz) is provided for the lower LightSquared carrier, which extends to 1536 MHz. To increase the selectivity, more poles can be added (implemented by adding dielectric resonator sections in the filter construction, which increases filter size), but the insertion loss will also increase beyond an acceptable level for use before the first LNA (see the Insertion Loss section).

Importantly, the quality factor (Q) of dielectric resonators is not high enough to support smaller passbands centered at the GPS L1 frequency. Achievable 1-dB fractional bandwidths (ratio of 1-dB passband to carrier frequency) for this technology is typically in the range of 1.5 – 20% (24 MHz and up at L1), so receivers that use dielectric filters that desire smaller passbands will generally accomplish this through additional filtering at IF or baseband. This commonly implemented solution leaves LNAs and mixers utilized in the front-end up until the narrower bandwidth IF or baseband filters vulnerable to saturation from strong out-of-band signals, such as the planned LightSquared emissions.

[5] Specifically, these are part numbers 4DR35-1575/U24-1.9 and 3DR35-1575/U24-1.9 from K&L Microwave. Their S-parameters can be obtained by entering the part numbers into K&L Microwave's Filter Wizard at www.klfilterwizard.com.

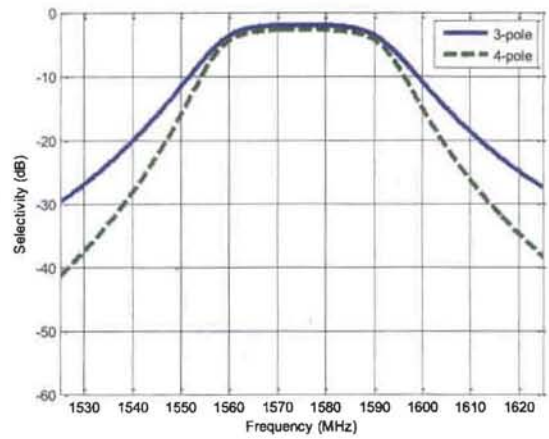

Figure 8-3. Selectivity of Representative 3- and 4-Pole Dielectric Resonator Filters with 24 MHz (1-dB Bandwidth) Passband

Even less selectivity is possible for receivers requiring wider passbands (e.g., for high-precision applications, or military equipment that is processing the wide-bandwidth P(Y)-code or M-code signals). Figure 8-4 shows representative selectivity for 3- and 4-pole dielectric filters with a 40 MHz (1-dB Bandwidth) passband. There is no attenuation provided at the upper LightSquared carrier frequency, and only a modest amount (10 – 20 dB) of attenuation for the lower carrier.

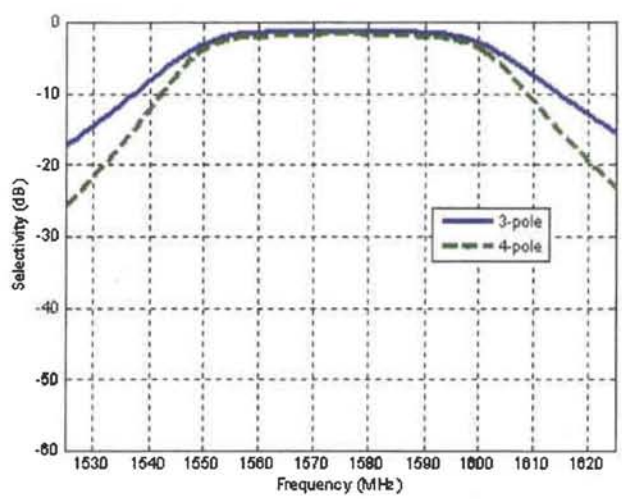

Figure 8-4. Selectivity of Representative 3- and 4-Pole Dielectric Resonator Filters with 40 MHz (1-dB Bandwidth) Passband

The group delay characteristics of the four dielectric resonator filters described in this section are shown in Figure 8-5. Over 24 MHz centered at L1, the differential group delays for the four filters in the order listed in the figure legend are 4.2 ns, 5.7 ns, 1.1 ns, and 1.5 ns. Note that

improved differential group delay performance is obtained as the passband bandwidth is widened.

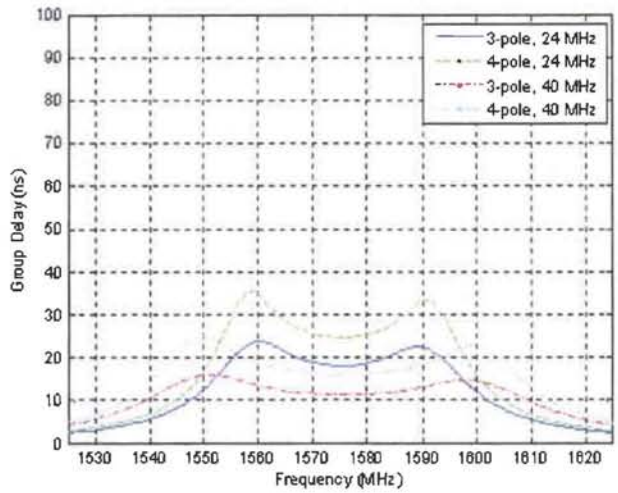

Figure 8-5. Group Delay Characteristics for Representative 3- and 4-Pole Dielectric Resonator Filters

Surface acoustic wave (SAW) filters are available for both RF and IF frequencies for GPS receivers. SAW filters use resonators that operate by converting the input electrical signal into an acoustic wave that propagates along the surface of a piezoelectric substrate [8-4]. They are inexpensive and typically much smaller than dielectric resonators, which makes them an extremely popular choice for applications where size is of utmost importance, such as GPS receivers integrated into cell phones and other mobile devices.

Low insertion loss SAW filters for use at the GPS L1 frequency are offered by many vendors. These filters are primarily marketed for low-bandwidth, low-precision C/A-code applications. They are not readily found with specified bandwidths wide enough for high precision GPS applications or for military receivers that process the wideband P(Y) or M-code signals. For instance, TriQuint Semiconductor offers 10 GPS L1 SAW filter models with advertised bandwidths ranging only from 2 – 2.4 MHz. The selectivities of two representative TriQuint RF SAW filter models are shown in Figure 8-6. Both models are only 1.4 × 1.2 × 0.46 mm in size, and have insertion losses over a passband defined as L1 ±1 MHz of less than 1.4 dB. Note that very little attenuation (< 5 dB) is provided at the upper LightSquared carrier frequency, and greater attenuation at the lower frequency (~20 - 30 dB). Importantly, the selectivity of SAW filters is sensitive to frequency, and the passband upper and lower limits may vary by up to ±5 MHz over the temperature range of -30°C to +85°C (-22°F to 185°F).

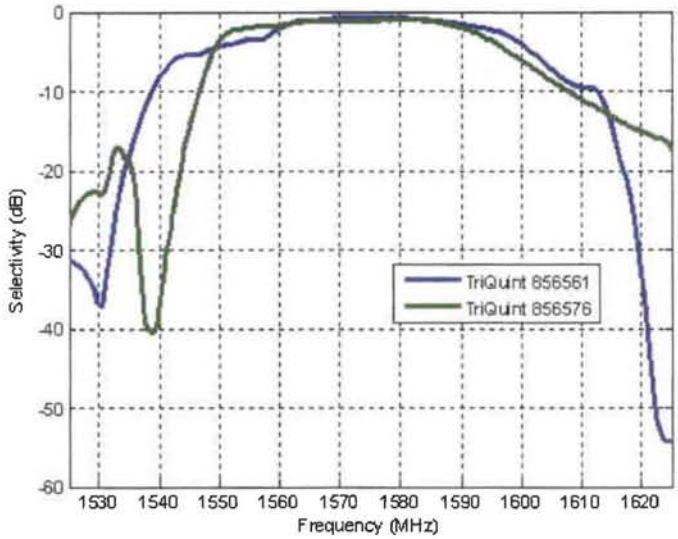

Figure 8-6. Selectivity of Two Representative GPS L1 RF SAW Filter Models

The group delay characteristics for the two representative GPS L1 RF SAW filters are shown in Figure 8-7. Although the differential group delay for these two examples is fairly small over L1 ±12 MHz, group delay performance is highly variable with temperature and manufacturing tolerances and thus generally not even specified for GPS RF SAW filters. For these reasons, in addition to the bandwidth limitation mentioned earlier, RF SAW filters are usually avoided for high-precision GPS applications.

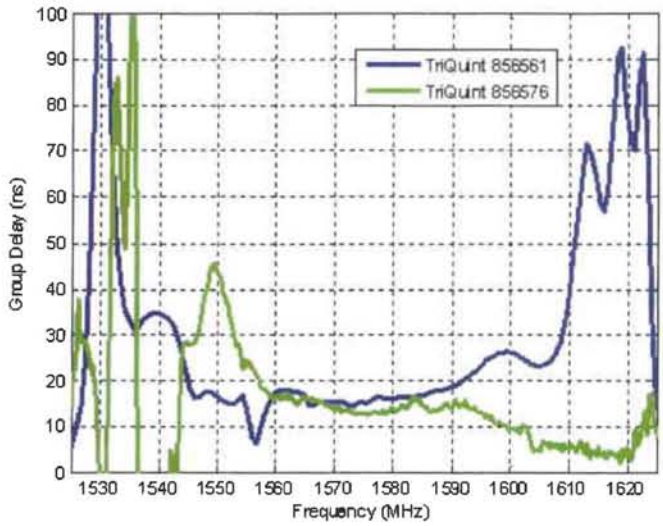

Figure 8-7. Group Delay Characteristics of Two Representative GPS L1 RF SAW Filter Models

SAW filters are also commonly used in GPS receivers at IF. SAW IF filters provide great selectivity, but at the drawback of high insertion loss (~10 – 20 dB). This tradeoff is generally desirable at IF because of the considerations discussed in Section 8.2.1.3. At typical GPS IF frequencies, which range from tens to hundreds of MHz, SAW filters are available in a wide variety of bandwidths ranging from 2 – 40 MHz. Figure 8-8 shows the selectivity characteristics of 3 representative GPS IF SAW filters for an IF frequency of 70 MHz. The filters have minimum specified 3-dB bandwidths of 40 MHz, 20 MHz, and 2 MHz. Their maximum specified insertion losses are 22 dB, 15 dB, and 8.25 dB, respectively, and their maximum specified group delays are 90 ns (over 36 MHz), 90 ns (over 18 MHz), and 340 ns (over 1.4 MHz). The large group delay differentials might preclude even the widest bandwidth IF SAW from use for high-precision applications.

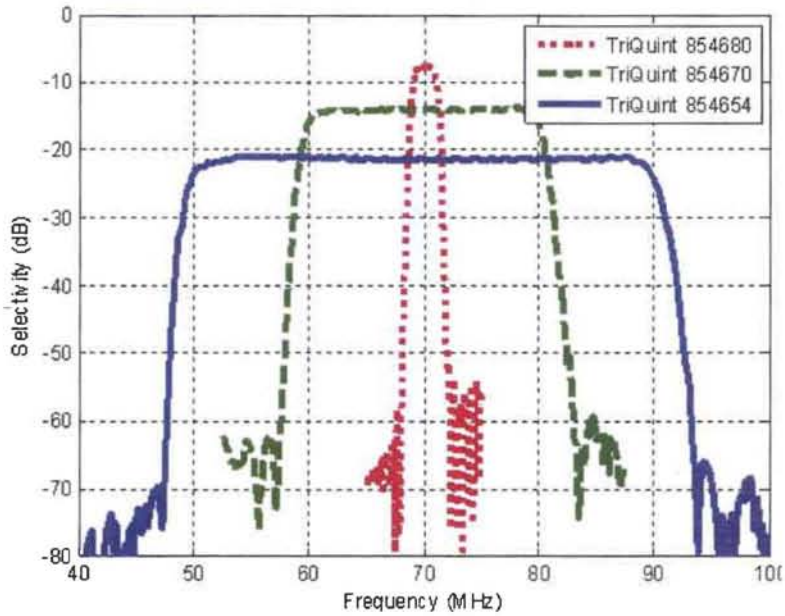

Figure 8-8. Selectivity of Three Representative GPS IF SAW Filters (70 MHz IF)

To achieve an IF frequency of 70 MHz, a GPS receiver must mix the incoming received L1 signal with a local oscillator at either 1505.42 MHz (low-side mix) or 1645.42 MHz (high-side mix). The LightSquared signals would end up, in the low-side mix case, at 20.6 – 30.6 MHz (lower carrier) and 39.8 – 49.8 MHz (upper carrier). All three of the IF SAW filters would suppress the LightSquared lower carrier signal by over 50 dB. The 2 MHz and 20 MHz filters would suppress the upper carrier signal by around this same amount, but the 40 MHz filter would provide very little suppression (relative to the 22 dB insertion loss that is also applied to the desired signal).