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Specifying and Measuring Aggregate Broadband Noise of CATV Modulators and Upconverters

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Specifying and Measuring Aggregate Broadband Noise of CATV Modulators and Upconverters

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Abstract

The proliferation of an increased number of frequency agile digital channel transmitters adversely impacts the level of noise aggregation at the headend and hubs. As operators expand and carry as much as 50 digital channels on their downstream lineup, the use of QAM transmitters with excessive broadband noise can result in a few dB loss of noise margin at the subscriber tap, which threatens the reliability of 256 QAM services. It is therefore important to be able to specify, predict and measure the level of aggregate noise produced by a plurality of combined modulators or upconverters.

This paper discusses the practical limitations and the subtle assumptions that must be incorporated into the specifications and measurements of Aggregate Broadband Noise of CATV upconverters and digital QAM transmitters.

For *K* channel headend systems, the types of Aggregate Broadband Noise measures discussed in the paper are based on the following three transmitter selection regimes: (a) *K Unit Sample Aggregate Broadband Composite Noise*; (b) *K Unit Ensemble Aggregate Broadband Composite Noise* and (c) *Single-Unit Self-Aggregate Broadband Composite Noise*. It is shown that while the first two methods come intuitively closer to headend combining realities, they require actual measurements of many transmitters in order to make a single pronouncement. In contrast, the third method yields an approximate upper bound that provides a measure attributable to (and measured on) a *single* transmitter device, thereby permitting a comparison with set limits or with other competitive devices.

This paper presents the considerations that govern the accuracy and measurement dynamic range required for such Aggregate Broadband Noise measures. Novel methods involving the separation of the Aggregate Broadband Noise measure into its composites, namely, modulated distortion noise components and other broadband noise components are discussed. Optimal ways in which such measurement techniques have been automated to enable rapid and efficient evaluation of transmitter devices are also presented along with examples of typical results.

1 Introduction

The proliferation of frequency-agile modulators and upconverters used in CATV headends and hubs presents new realities and challenges with respect to channel combining practices and the maintenance of satisfactory broadband noise levels at the output of such channel combining networks. Fixed frequency modulators and upconverters employ an output channel filter that filters out any spurious signals and noise that are out-of-band, thereby enabling the combining of all such channels in the lineup without incurring excessive output noise floors. In contrast, frequency-agile transmitters do not have narrow band filters at their output and consequently they can produce appreciable levels of broadband noise [1]. To the extent that many such agile frequency CATV transmitters are used in combination, their composite noise can aggregate unfavorably.

In a companion paper [2], it is shown that under currently prevailing headend and HFC signal transport practices, mass deployment of certain classes of QAM transmitters and upconverters used (or proposed to be used) for Video On Demand (VOD), will, in the aggregate, produce excessive noise accumulation from adjacent channel modulated distortion terms as well as broadband noise. Specifically, it is shown that as operators expand and carry 50 digital channels on their downstream lineup using these QAM transmitters, a loss of 2.4 dB C/(N+I) at the subscriber tap compared to the levels that can be realized with transmitters that meet the DOCSIS RF requirements. Because many subscriber installations do not have the benefit of any margin, this loss can render the 256 QAM reception unreliable.

Thus, the broadband noise specifications constitute major criteria for accepting or rejecting a specific frequency-agile CATV transmitter for headend applications. This measure plays an important quantitative role in assessing the impact on digital QAM downstream channel performance, as it is a significant component in an overall degradation budget. The buildup and use of such budget to predict satisfactory QAM channel performance is described in Reference [2].

The contribution of broadband noise over various frequencies in such frequency agile transmitters depends on the channel being tuned to, the type of filtering used and other design approaches that impact the broadband noise floor levels. Hence, the full characterization of the broadband noise density of such transmitters would require a spectral mask over the full frequency band for every tuned channel case. This exhaustive approach can prove to be prohibitively cumbersome and impractical.

An alternative practice usually adopted is to specify the aggregate noise level on each channel (or its maximum over the entire band) which results from combining a full consecutive channel line-up of identical transmitters, each tuned to distinct channels. Note that this aggregate noise level includes both thermal noise contributions and modulated distortion contributions from each identical transmitter. We shall concentrate here on the output of digital transmitters carrying Quadrature Amplitude Modulated ("QAM") signals, wherein the interference and degradation effects of the modulated distortion terms are practically indistinguishable from those of thermal noise if that noise has the same spectral density function. It is therefore the spectral density of the total noise and distortion that is of interest, which shall be hereinafter defined as "*Composite Noise"*. For a given transmitter j, denote the composite noise spectral density on frequency f by $\mathbf{S}_j(f,n)$ where n is the channel being tuned to. Designating the j^{th} transmitter to be tuned to the channel equal to its index ($n=j$), we obtain the result that the aggregate broadband composite noise density $\mathbf{a}(f)$ from all combined transmitters is given by

$$(1) \qquad \mathbf{a}(f) = \sum_n \mathbf{S}_n(f,n).$$

The aggregate composite noise power integrated over a measured channel M is given by integrating the aggregate composite density $\mathbf{a}(f)$ over the frequencies f that belong to the 6 MHz frequency interval corresponding to the measured channel M, (which we denote here by $f \in M$):

$$(2) \qquad \mathbf{A}(M) = \int_{f \in M} \mathbf{a}(f)\,df = \int_{f \in M} \sum_n \mathbf{S}_n(f,n)\,df = \sum_n \int_{f \in M} \mathbf{S}_n(f,n)\,df = \sum_n \mathbf{D}_n(M,n),$$

where $\mathbf{D}_n(M,n)$ in the right hand side of Equation (2) is a discrete random variable depending on the specific choice of transmitter n and expressing the integrated composite noise measured on channel M emanating from such transmitter tuned to a channel corresponding to its index n. Thus, the aggregate composite noise $\mathbf{A}(M)$ measured on channel M is a random variable that depends on the *collection* of sampled transmitters when combined and not on the performance of *a single* transmitter. To specify aggregate broadband composite noise one can construct the following measures:

- **(a)** <u>*K Unit Sample Aggregate Broadband Composite Noise*</u>. By obtaining *one* sample headend having K transmitters tuned to K distinct channels and combining them to yield a single sample measure of $\mathbf{A}(M)$.
- **(b)** <u>*K Unit Ensemble Aggregate Broadband Composite Noise*</u>. Repeat method (1) above over an *ensemble* Ω (multiple number) of 'headends', each consisting of different K transmitters. Forming such an ensemble may also include channel shuffling the same collection of K transmitters. The result is an ensemble of random variables $\mathbf{A}_\Omega(M)$. One can than observe the statistics over Ω and obtain a single measure such as the average or, more appropriately, the worst-case value given by $U(M) = \max_\Omega \{A_\Omega(M)\}$ for every M.
- **(c)** <u>*Single-Unit Self-Aggregate Broadband Composite Noise*</u>. Using a single unit j, obtaining the broadband composite noise measures $\mathbf{D}_j(M,n)$ on each channel M with the *same* unit tuned to channel n that assumes values running over all K channels and summing such individual contributions:

$$(3) \qquad \mathbf{A}_j(M) = \sum_{n=1}^{K} \mathbf{D}_j(M,n).$$

While the first two methods come closer to headend combining realities, they require actual measurements of many transmitters in order to make a single pronouncement. In contrast, the third method in (**c**) above yields a result that provides a measure attributable to (and measured on) a *single* transmitter device, thereby permitting a comparison with set limits or with other competitive devices. In this paper, we refer to this aggregate noise measure as the '*self-aggregate composite noise*' of the transmitter and for convenience we drop the transmitter index j from the expression in Equation (3).

This paper describes a measurement method for the determination of the self-aggregate composite noise of a transmitter. The paper also relates this self-aggregate metric (which is an attribute of a single transmitter), to the aggregate noise performance of a headend comprised of multiple (different) transmitters, each transmitter of which manifests its own unique self-aggregate composite noise performance, but all of which satisfy a common self-aggregate specification limit. To that end, a qualitative presentation of the value of self-aggregate composite noise specifications to the prediction of headend performance is presented in Section 3.

2 Measurement Considerations and Description

2.1 Overview

In section 1, self-aggregate composite noise is described as the aggregate noise level on each channel of a headend that is comprised of a full consecutive channel line-up of identical transmitters, each tuned to a distinct channel.
A measurement approach for obtaining a measure of self-aggregate composite noise would proceed as follows:

1. Tune device under test (DUT) to the first channel (n=1).
2. For all channels (including the channel to which we are tuned), measure total composite noise power residing within the channel bandwidth. Figure 2-1 represents the results of such a measurement. We can interpret Figure 2-1 as an illustration of the sequence $\mathbf{D}(n,M)$ for every measured channel M consisting of the elements { $\mathbf{D}(n,1)$, $\mathbf{D}(n,2)$, … , $\mathbf{D}(n,K)$ } where the value of n corresponds to channel 13 and where $\mathbf{D}(n,M)$ represents the composite noise power measured on channel index M when the device under test is tuned to channel index n.
3. Repeat steps 1 and 2 above for all channels n up to K.
4. Compute the aggregate distortion plus noise as follows: At the conclusion of step (3) we will have obtained K sequences similar to that shown in Figure 2-1. Next, obtain the self-aggregate composite noise sequence $\mathbf{A}(M)$ in accordance with Equation (3) above.

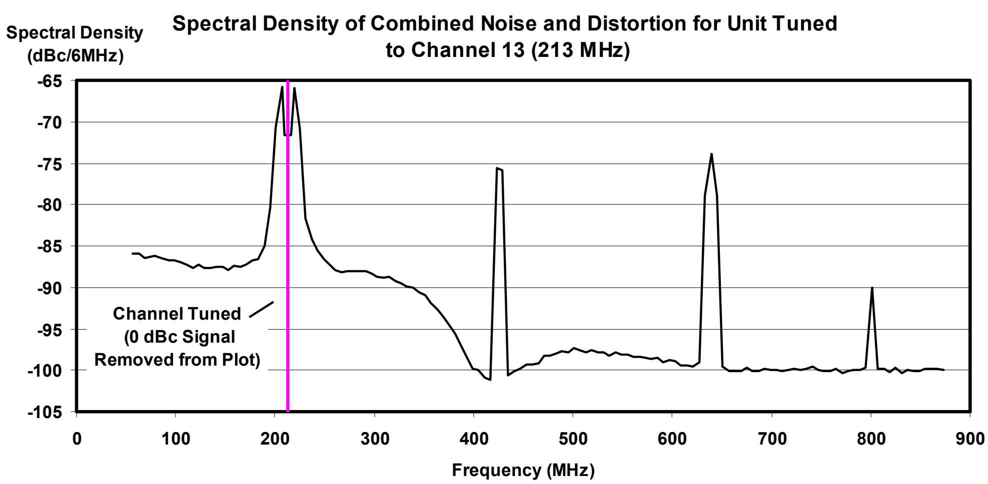

Figure 2-1. Composite noise contribution of single upconverter tuned to 213 MHz. Spectral density is referred to the QAM signal level.

The brute force measurement approach described above envisions simultaneous measurements of distortion plus noise. In fact, practical considerations result in a measurement process that separates noise measurements from distortion measurements.

Figure 2-1 illustrates the noise and distortion contributions of a single channel to the aggregate composite noise performance. We observe that the composite noise contributions of a channel consist of a noise profile plus several distinct distortion terms.

4 distinct distortion components are present in Figure 2-1:

- **Spectral regrowth** around 213 MHz above and below the signal.
- **2nd harmonic** around 426 MHz.
- **3rd harmonic** around 639 MHz
- **Mixer cross term** around 800 MHz.

With the exception of the mixer cross term, these types of distortion terms will also be generated by a broadband measurement device such as a spectrum analyzer when it is fed by the transmitter's signal. For CW distortion terms, the effects of spectrum analyzer distortion can typically be mitigated by increasing analyzer attenuation and reducing analyzer resolution bandwidth. For the modulated distortion terms generated by a digital signal however, even the best analyzers do not have sufficient dynamic range to reliably distinguish DUT distortion from analyzer distortion, and there is a requirement to filter the main carrier at the spectrum analyzer input (See explanation in text box below).

Requirement for measurement filters

We see from Figure 2-1 that we have a requirement to make second harmonic measurements of terms that are on the order of –75 dBc. The Rohde & Schwarz FSEA-30 is a high dynamic range spectrum analyzer that has a second harmonic intercept (*SHI*) of +50 dBm and a specified noise floor of –155 dBm/Hz (in practice we have observed analyzer noise density measurements as low as –160 dBm/Hz). Suppose we require that the analyzer distortion be at least 20 dB below the DUT distortion and that the analyzer noise floor be at least 10 dB below the DUT distortion. We thus require that the analyzer distortion be less than – 95 dBc. We compute the maximum power allowed at the analyzer input as follows:

$$P_{\text{analyzer}} < SHI - 95 = -45 \text{ dBm}$$

The value of – 45 dBm represents 57 dB of attenuation relative to a typical DUT output of +12 dBm.

In order to satisfy the requirement that the analyzer noise floor be at least 10 dB below the DUT distortion floor, we require that the distortion spectral density be greater than –145 dBm/Hz. If we assume that modulated distortion power is evenly spread over the 6 MHz channel bandwidth (which we will show is not exactly the case), we see that the DUT distortion at the analyzer input must be greater than –77 dBm in order to satisfy the 10 dB above noise floor requirement.

For a typical DUT output of +12 dBm applied directly to the analyzer input, – 75 dBc DUT distortion corresponds to distortion power of – 63 dBm, leaving us only 14 dB of carrier attenuation before we encounter the – 77 dBm limit identified above. Thus, we conclude that the 52 dB carrier attenuation requirement must be applied selectively to the main carrier and not the distortion term; i.e. there is a requirement for filtering between the DUT and the spectrum analyzer. Similar considerations can be applied to other distortion terms, showing that filtering is required.

Referring to Figure 2-1 again, the noise profile includes terms ranging from –70 dBc to –100 dBc. Moreover, the noise terms of primary interest (those with the most power) are in the vicinity of the carrier. The requirement to attenuate (filter) the main carrier for distortion measurements is incompatible with the requirement to measure low-level noise terms in the vicinity of the carrier. Thus, we must measure aggregate noise separately from aggregate distortion.

The aggregate noise measurements and aggregate distortion measurements are described separately in sections 2.2 and 2.3.

2.2 Aggregate Noise Measurement

2.2.1 Description of Measurement Configuration and Calibration Considerations

Figure 2-2 illustrates the aggregate noise test configuration. As noted in section 2.1, the DUT output is connected directly to the analyzer input (though a matching pad) in order to maximize dynamic range.

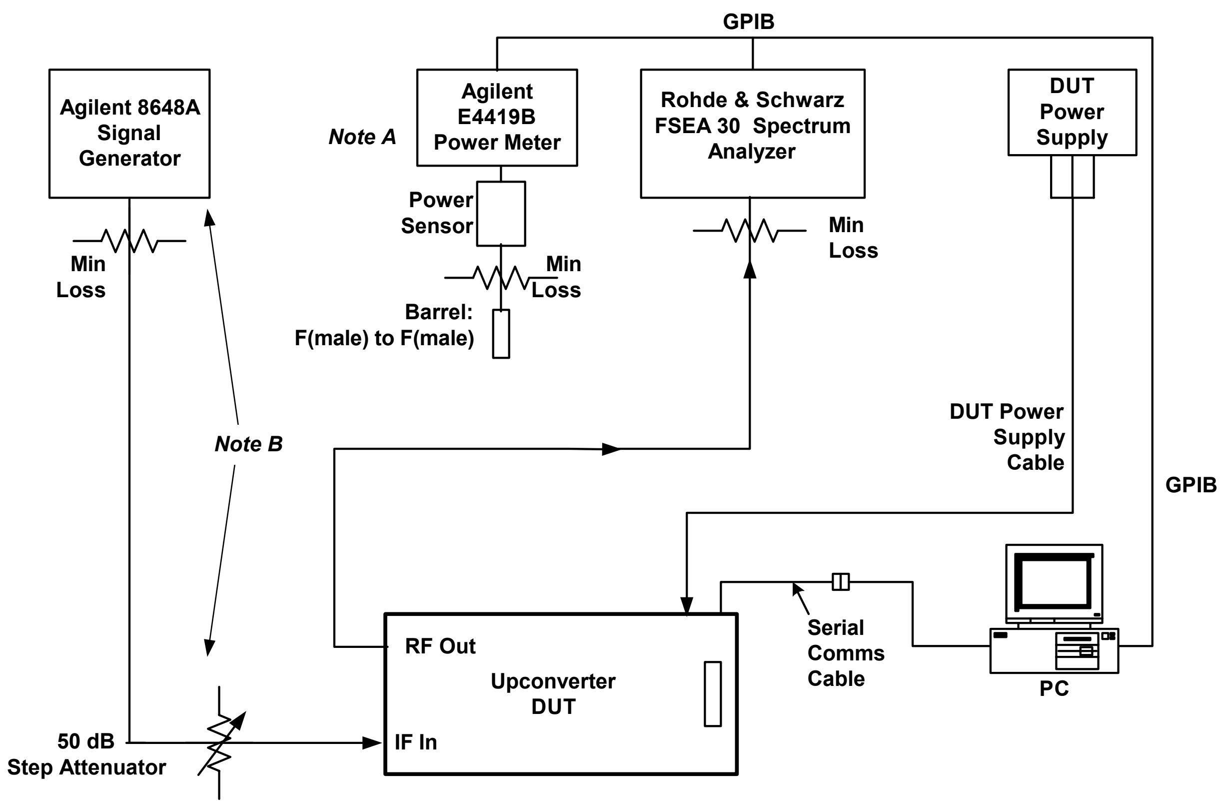

Figure 2-2. Aggregate noise test configuration. ***Note A*:** Power meter used during spectrum analyzer calibration and for obtaining DUT reference power measurements**.** ***Note B*:** Signal generator and 50 dB attenuator used during calibration only. For noise measurements, the DUT IF input is terminated.

Measurement of DUT noise presented several instrumentation challenges. These challenges are addressed below.

1. *Distinguishing DUT noise from spurious signals (both DUT spurious and environmental spurious)*. The extremely low levels of the DUT noise and the fact that spurious signals might exist in the measurement channel bandwidth rendered integrated noise density measurements impractical. Thus, we chose to perform the measurements in zero-span mode at a location within the channel bandwidth where DUT spurs would not be present. A resolution bandwidth setting of 300 kHz was used, which represented a compromise between spurious mitigation and measurement speed.
2. *The spectrum analyzer noise floor* represents another source of measurement error, particularly because many of the DUT noise measurements are close to the analyzer noise floor. Thus, the analyzer noise floor is measured independently and all spectrum analyzer measurements are corrected for the analyzer noise floor contribution.
3. *Absolute accuracy errors* due to cable and spectrum analyzer characteristics.

In order to improve measurement accuracy, we elected to calibrate the spectrum analyzer (including the cable between the DUT and the spectrum analyzer) against a power meter. This calibration presented a challenge in that the signal levels seen by the analyzer during noise measurements (typically less than –70 dBm) are not measurable by a power meter. Therefore a special calibration procedure was devised. This calibration is based on comparison of the results of two measurements of the DUT output power at all frequencies of interest: one is conducted using a power meter at the nominal output level of the DUT; the second is conducted using the spectrum analyzer when the output level of the DUT is reduced by 50 dB.

The calibration process consisted of the following steps.

a. Apply a CW IF input signal to the DUT and measure the power level at the RF output of the DUT using a power meter. Perform this measurement for all channel tunings of the DUT.
b. Attenuate the CW IF input signal by 50 dB and connect the DUT output to the spectrum analyzer as shown in Figure 2-2. Perform power measurements on the spectrum analyzer under the same conditions that DUT noise measurements will subsequently be made (see step (1)). Also, note that the DUT settings (RF level setting in particular) must be identical to the settings used in calibration step (a). Finally, the spectrum analyzer readings obtained in this step must be corrected for the error introduced by the analyzer noise floor.
c. Measure the exact attenuation of the 50 dB attenuator at the IF frequency.

The DUT was chosen as the calibration source specifically because it allows us to perform the 50 dB attenuation at the IF frequency only, and hence the attenuator needs to be calibrated at one frequency only. An alternative calibration method could have used a tunable signal generator, and power meter measurements of the signal generator output would have been compared with spectrum analyzer measurements of the signal generator output attenuated by 50 dB. This technique however, would require broadband characterization of the attenuator.

A second advantage of using the DUT as the calibration source is that it implicitly takes into account possible broadband impedance mismatches between the DUT and the measurement system.

Measurements (a)-(c) are used to obtain an analyzer power correction factor (as a function of measurement frequency) that is applied to all DUT noise measurements made using the spectrum analyzer.

The correction and calibration steps identified in steps (2) and (3) (with the exception of the attenuator calibration) were carried out under computer control. Calibration files were obtained and subsequently used during the processing of the DUT noise measurements.

2.2.2 Description of Measurement Process

The process of measuring aggregate noise included the following steps.

1. Perform measurement system calibration as described in section 2.2.1.
2. Perform reference power measurements at the DUT output on all channels. These measurements are obtained using the power meter with a nominal IF input signal applied to the DUT and the DUT set at a nominal RF level. These measurements serve as the reference power measurements for the subsequent DUT noise measurements, which are expressed in dBc relative to DUT carrier power.
3. Terminate the DUT IF input and connect the DUT output directly to the spectrum analyzer as shown in Figure 2-2. Perform noise measurements as described below:

    a. Tune DUT to the first channel and set DUT RF level to the same level used when reference power is obtained in step (2) above.
    b. Configure the analyzer to make the DUT noise measurement as described in section 2.2.1. Step through the RF band and perform a noise measurement on every channel frequency. Note that the DUT noise measurements are obtained in 'noise marker' mode and are in fact spectral density measurements (dBm/Hz). These measurements are subsequently 'integrated' by the test software to obtain a noise power measurement over the 6 MHz channel bandwidth.
    c. Repeat steps (a) and (b) for all channel tunings.

Steps (2) and (3) above are performed under computer control and calibration corrections are performed automatically. The outcome of these measurements is a two-dimensional array where each row corresponds to a channel setting of the DUT and each column corresponds to a channel on which DUT noise spectral density is measured. A particular cell (*n,M*) in the array represents the DUT noise (in dBc) measured on channel index *M* when the DUT is tuned to channel index *n*.

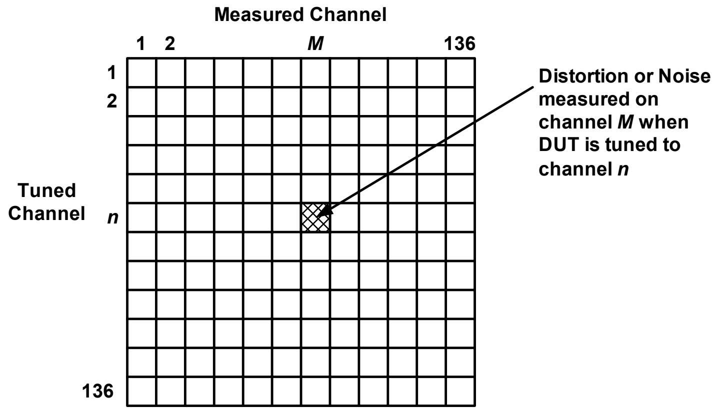

Figure 2-3. Format of noise and distortion measurements. For each channel tuned to, one makes noise measurements for all other channels.

The data in the array depicted in Figure 2-3 can be processed in several ways. Data from particular rows can be plotted as a function of measured channel frequency, thus representing the noise generated by the DUT across the RF band as a function of a particular channel tuning. Figure 2-4 depicts several such plots. We see that individual channel tunings are distinguished by a narrow noise pedestal centered at the tuned channel resting upon a broader pedestal associated with broadband filters.

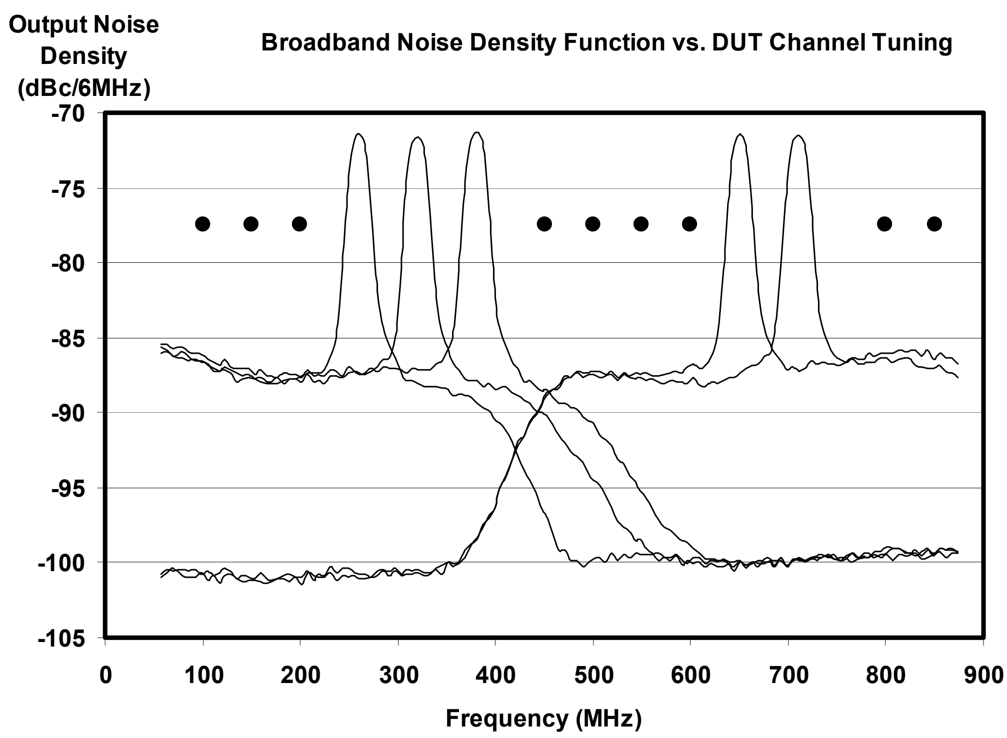

Figure 2-4. Broadband noise as a function of upconverter channel tuning. Noise density is referred to QAM signal level when turned on.

Alternatively, the individual columns of the array can be summed to obtain a single plot of self-aggregate noise as a function of measured channel frequency. Figure 2-7 includes such a plot.

2.3 Aggregate Distortion Measurement

2.3.1 Differences Between Aggregate Noise Measurement and Aggregate Distortion Measurement

As noted in section 2.1, the necessity to filter the main carrier at the spectrum analyzer input in order to make distortion measurements leads us to perform distortion measurements using a different test setup than that used for noise measurements. Moreover, the predictable location of distortion terms relative to the main carrier permits us to perform a much smaller number of measurements for each DUT channel tuning than were required for aggregate noise measurements.

Referring once again to Figure 2-1, we see that the distortion terms resulting from a particular DUT channel tuning may include Spectral regrowth, 2^{nd} harmonic, 3^{rd} harmonic and Mixer cross term

Thus, for each channel tuning, it appears that we need to make no more than 4 measurements, and that our measurement scheme would proceed as follows:

1. Tune device under test to first channel.
2. Measure the distortion power of the known distortion terms associated with each channel tuning.
3. Repeat steps 1 and 2 for all channel tunings.
4. Compute the aggregate distortion for each measured channel by summing up the appropriate distortion measurements (i.e. only those measurements that fall on the channel of interest) obtained during the previous steps.

In fact, there are several subtleties associated with the scheme described above: First, the filtering requirements vary as a function of both the channel to which we are tuned and the channel on which we are measuring. Thus, there is a conceptual addition to step (2) in which an appropriate channel filter is selected. Second, the power of the various modulated distortion terms is not restricted to a single channel. Thus, in step (2), measurements need to be made on multiple channels in order to fully capture all of the distortion power. We address the filtering considerations first.

<u>Spectral regrowth filtering:</u>

Spectral regrowth is also often referred to as Adjacent Channel Leakage Power and the details of making such measurements have recently been covered in an application note [3]. In order to determine whether or not there was a requirement to perform individual channel filtering in support of spectral regrowth measurements, we manually determined spectrum analyzer regrowth contribution as a function of signal level and extrapolated the regrowth contribution we could expect at the level used during measurements.

For the Rohde & Schwarz FSEA-30 analyzer, the analyzer noise in a 6 MHz bandwidth is approximately –90 dBm. Thus, for carrier levels of –20 dBm, we can make regrowth measurements (assuming we correct for the analyzer noise floor) on the order of –67 or –68 dBc. We found that at carrier levels below –10 dBm, we did not see any measurable improvement in regrowth performance, implying that the analyzer was not limiting our ability to make the regrowth measurements. Thus, we concluded that carrier filtering was not required for regrowth measurements.

Subsequently, we performed spectral regrowth measurements on selected channels using channel-specific filters and did not discern any significant difference relative to the measurements made without filtering.

<u>2^{nd} and 3^{rd} harmonic filtering:</u>

Table 2-1 summarizes the characteristics of the harmonic filters required for these measurements:

<u>Mixer cross term:</u>

The mixer cross term is a distortion term generated by the DUT on a frequency equal to F_0-*f*, where F_0 is a frequency dependent on internal DUT frequency design and *f* is the frequency tuned by the DUT. This term is not generated by the spectrum analyzer and therefore there is not a requirement to attenuate the signal at the spectrum analyzer input when performing mixer cross term measurements.

Tuned Channel Range (MHz)	Harmonic Range: Min (2nd H) – Max (3rd H) (MHz)	Filter Type	Stop Band (MHz)
57 – 85	114 – 255	HPF	< 91
93 – 159	186 – 477	HPF	< 174
165 – 231	330 – 693	HPF	< 300
237 - 435	474 – 870	BPF	229 - 462

Table 2-1 Harmonic measurement filter characteristics

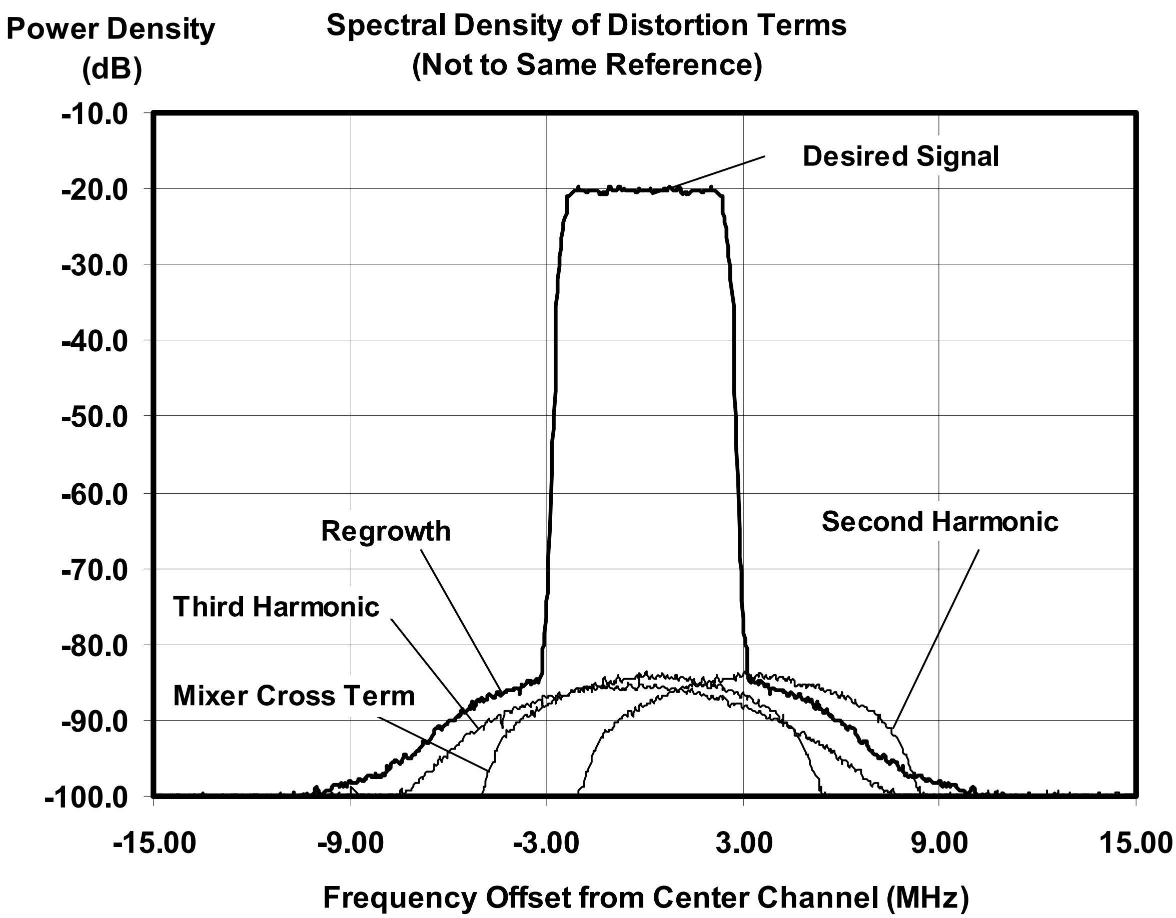

Figure 2-5. Desired signal and distortion terms shown with respect to channel boundaries (-3 MHz, +3 MHz) with Standard CATV plan tuning. All terms shown do not necessarily have the same power reference.

The distortion terms created under conditions of digital modulation occupy multiple channels. Figure 2-5 illustrates the various modulated distortion terms. This work was done on the Standard CATV frequency plan for which the specifics reported below apply. The 2nd harmonic term is centered at a channel boundary and the power of the term is evenly divided between the two channels on either side. The modulated 3rd harmonic term is centered within a channel and the power is spread across three channels. The modulated spectral regrowth term occupies the adjacent channels relative to the carrier with virtually all of the signal power resident in the first and second adjacent channels. The modulated mixer cross term is a second order-like non-linearity term, with a spectral shape similar to that of the second harmonic term, but centered within the measurement channel. Unlike 2nd harmonic, in which measurements were made on two channels, the mixer cross term distortion measurement was made on one (worst-case) channel. Table 2-2 summarizes the characteristics of the particular distortion measurements.

Term	Number of terms measured	Center of measured channels relative to main carrier frequency f (in MHz)
Spectral Regrowth	4	f-12, f-6, f+6, f+12
2nd Harmonic	2	2f-3, 2f+3
3rd Harmonic	3	3f-6, 3f, 3f+6
Mixer Cross Term	1	$F_0 - f$

Table 2-2 Summary of distortion measurements for operation on the Standard frequency plan.

The measurements summarized above specifically exclude other potential sources of composite noise, such as higher order harmonics and discrete (unmodulated) spurs. In the case of higher order modulated harmonics, the combination of inherently low distortion levels and the high degree of spectral spreading (resulting in lower distortion power densities) render the harmonics immeasurable.

In contrast, discrete spurs, which are readily measurable due to their relatively high spectral density, are typically specified separately from modulated terms and noise. For example, the DOCSIS specifications for both adjacent channel and other channel spurious and noise explicitly exclude up to 3 discrete spurs [4]. Thus, we concluded that it was not appropriate to include the discrete spurs in the self-aggregate measurements.

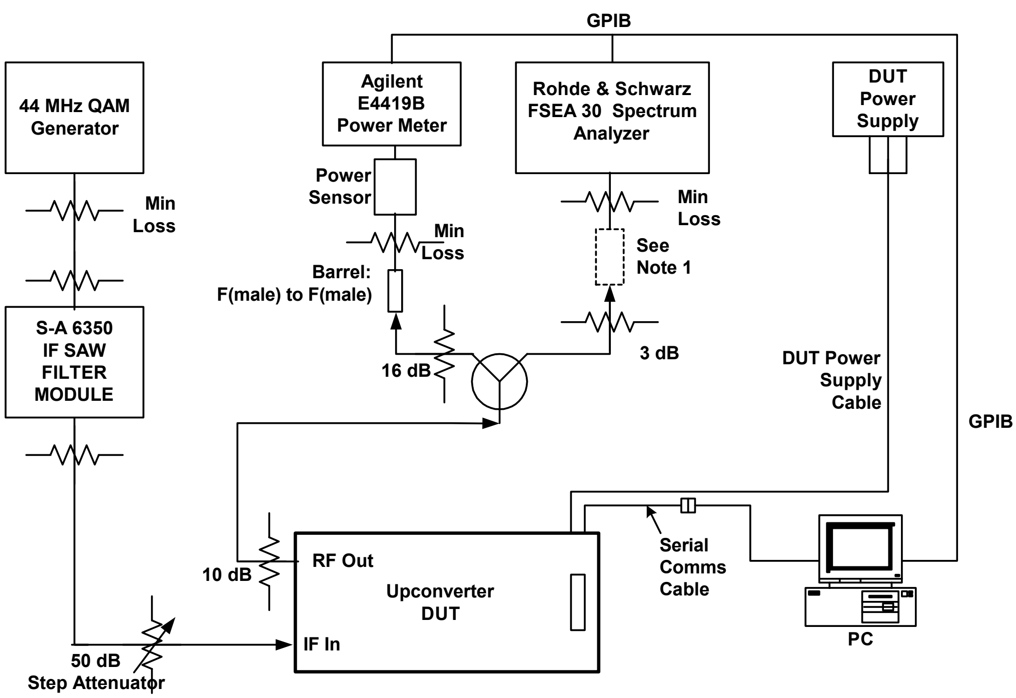

Figure 2-6. Modulated distortion test configuration. ***Note 1*****:** This in-line device is either a pad, a coaxial filter or direct connection as specified in Table 2-3.

Figure 2-6 illustrates the modulated distortion test configuration. This configuration differs in several ways from the test configuration used for aggregate noise measurements:

1. Aggregate noise measurements were obtained under conditions of no input signal. Aggregate distortion measurements specifically require a modulated IF source. The test configuration includes an Agilent ESG generator programmed to generate a 256 QAM signal at a symbol rate of 5.360537 Msps. The ESG output is subsequently filtered to reduce sidebands generated by the ESG.
2. As discussed earlier, harmonic filtering is implemented at the spectrum analyzer input. This functionality is represented by the block labeled 'Coaxial Filter or Pad'.

3. Aggregate noise measurements required a direct connect between the DUT and the analyzer in order to maximize sensitivity. The fact that distortion terms of interest are typically greater than –70 dBc (in order to have any measurable impact on aggregate composite noise) and that carrier filtering is implemented at the spectrum analyzer input allows us to tolerate signal attenuation in the path between the DUT and the spectrum analyzer. The benefit of this configuration is that power meter measurements can be made at the same time that spectrum analyzer measurements are made. Thus, the aggregate noise test step of performing reference carrier power measurements prior to the aggregate noise measurements is eliminated.

Note that the presence of the signal path from the DUT to the power meter introduces a requirement to calibrate the path (path loss as a function of frequency). This particular calibration step was not required for aggregate noise measurements as all power meter measurements were made directly at the DUT output.

In addition to the differences in test configuration, the aggregate distortion measurements differed from the aggregate noise measurements with respect to the particular manner in which distortion was measured. Specifically, the distortion measurements were performed as integrated power measurements over a channel bandwidth, whereas the noise measurements were performed as spot noise measurements that were computationally integrated based on knowledge of the channel bandwidth.

As noted in section 2.2, due to the relatively low levels of the noise, there was concern that an integration technique would inadvertently capture spurs of the same order of magnitude as the noise. Moreover, the noise spectral density is relatively flat over the channel bandwidth, and a spot measurement technique is a reliable and fast surrogate for integrated measurements under these conditions. As seen in Figure 2-5, however, the distortion measurements show significant departure from flat spectra within a channel. We could have devised correction factors for the known distortion shaping, but were concerned about the introduction of measurement errors if the distortion did not behave in an ideal fashion. In fact, we did observe spectra that were distorted relative to the ideal cases illustrated above.

The fact that distortion measurements were performed as integrated power measurements led to several changes in the spectrum analyzer calibration method described for aggregate noise. These differences are summarized below:

1. Noise power measurements performed in support of calibration of the spectrum analyzer noise floor were performed as integrated power measurements over a 6 MHz channel bandwidth.
2. The calibration of the spectrum analyzer was performed using the modulated signal described above and the measurement made at the spectrum analyzer was an integrated power measurement over the 6 MHz channel bandwidth. As in the case of the aggregate noise calibration, a power measurement performed directly at the DUT output (on a particular channel) is compared to a power measurement made at the spectrum analyzer on the same channel in order to derive a spectrum analyzer calibration factor.
3. It is important to note that the calibration of the spectrum analyzer against the power meter is in fact a calibration of the complete signal path between the DUT and the analyzer (including the analyzer). In the case of aggregate noise, only a single signal path (a cable) was used. For aggregate distortion measurements, multiple paths (distinguished by filter selection) exist between the DUT and the spectrum analyzer. Thus, there is a requirement to perform spectrum analyzer/signal path calibration for each signal path.

A final calibration consideration is the fact that any DUT distortion measurement also includes DUT noise. Thus, measurements of aggregate distortion must be preceded by measurements of aggregate noise, and the individual measurements used in the determination of aggregate noise must also be used to correct the individual distortion measurements that will subsequently be used in the determination of aggregate distortion.

2.3.2 Description of Measurement Process

The process of obtaining aggregate distortion measurements for a particular DUT includes the following steps:

1. Using the techniques described in section 2.2 for aggregate noise, obtain individual channel noise measurements. The output of this process is a 136 × 136 array that includes the DUT noise measured on all channels for all possible channel tunings (see Figure 2-3).
2. Perform 'measurement-independent' calibration of the test configuration. This includes:
    a. Calibration of the signal path between the DUT and the power meter.

   b. Calibration of the spectrum analyzer noise floor.
3. Perform 'measurement-specific' calibration of the test configuration. This explicitly includes calibration of the various signal paths between the DUT and the spectrum analyzer. The concept of multiple signal paths is represented in Figure 2-6 by the dotted block labeled 'Coaxial Filter or Pad'. A particular coaxial filter or pad is installed at this location as a function of measurement type. The relationship between measurements and filter/pad type is summarized in Table 2-3 below.

Measurement Type	Tuned Channel Range (MHz)	Filter/Pad Type
Harmonics	57 – 85	91 MHz HPF
Harmonics	93 – 159	174 MHz HPF
Harmonics	165 – 231	300 MHz HPF
Harmonics	237 - 435	229 MHz – 462 MHz BPF
Spectral Regrowth	All	10 dB Pad
Mixer Cross Term	All	Direct Connection

Table 2-3 Relationship between measurement type and filter/pad type used

4. Perform distortion measurements. This consists of the following steps:
   a. Establish the signal path (i.e. install the appropriate filter/pad) associated with the measurement (see Table 2-3). This step (a) is implemented manually.
   b. For each channel in the Tuned Channel Range:
      i. Measure the carrier power.
      ii. Measure the integrated distortion power on the channels where the distortion falls. Table 2-2 summarizes the particular distortion measurements that are made.
      iii. Calculate distortion in dBc.

   Step 4(b) is implemented under computer control. The calibration corrections identified in steps 2 and 3 above are performed automatically by the test software.

5. Process the data obtained in step 4 and determine aggregate distortion as a function of frequency.

The measurement data obtained in step (4) is retained in a series of 136 × 136 arrays, one for each measurement type. The array structure is illustrated in Figure 2-3. Specifically, an array element represents the distortion measured on channel M when the DUT is tuned to channel n.

11 unique arrays were created. One array included the DUT noise measurements described in step (1) above. The remaining 10 arrays are associated with the 10 unique measurement types identified in the right-most column of Table 2-2 (2nd harmonic measured at frequency $2f$–3 MHz, 2nd harmonic measured at frequency $2f$+3 MHz, etc.).

Processing of these arrays included the following steps:

a. *Correction of the distortion arrays for DUT noise*. Each distortion array underwent a correction process in which the distortion measurement was corrected for DUT noise. Thus, in each distortion array, the distortion measurement in every cell (i,j) was reduced by the noise measurement in the associated cell (i,j) of the noise array.
b. *Summation of the corrected distortion arrays to produce an aggregate distortion array*. Recalling the terminology introduced in section 2.1, the aggregate distortion array consists of array elements $\mathbf{D}_D(n,M)$, each of which represents the aggregate distortion measured on channel index M when the DUT is tuned to channel index n. Each aggregate distortion array element $\mathbf{D}_D(n,M)$ is computed as follows:

   (4) $$\mathbf{D}_D(n,M) = \sum_{i=1}^{10} \mathbf{D}(n,M)_i$$, where $\mathbf{D}(n,M)_i$ represents the cell (n,M) in the i^{th} of 10 distortion measurement arrays.

c. *Calculation of the aggregate distortion sequence*. Obtain $\mathbf{A}_D(M)$ as follows:

(5) $\mathbf{A}_D(M) = \sum_{n=1}^{K} \mathbf{D}_D(M,n)$; using the values of $\mathbf{D}_D(n,M)$ obtained in Equation (4) above.

Note that the calculations described in steps (b) and (c) above have applicability beyond the determination of the aggregate distortion sequence $\mathbf{A}_D(M)$. For example, the 'summing over channel columns' step described mathematically in step (c) can be applied to the noise array to develop the aggregate noise sequence $\mathbf{A}_N(M)$. Also, individual measurement arrays can be added as described in step (b) to develop aggregate arrays for individual distortion types. For example, the 4 spectral regrowth arrays can be added to arrive at an aggregate regrowth array.

Finally, all 11 arrays (the noise array plus the 10 'noise-corrected' distortion arrays) can be added together to develop a single aggregate noise+distortion array $\mathbf{A}_{N+D}(M)$=$\mathbf{A}(M)$. Figure 2-7 illustrates plots of aggregate noise, aggregate distortion, and aggregate noise+distortion, arriving at the aggregate composite noise levels that are the subject of this paper.

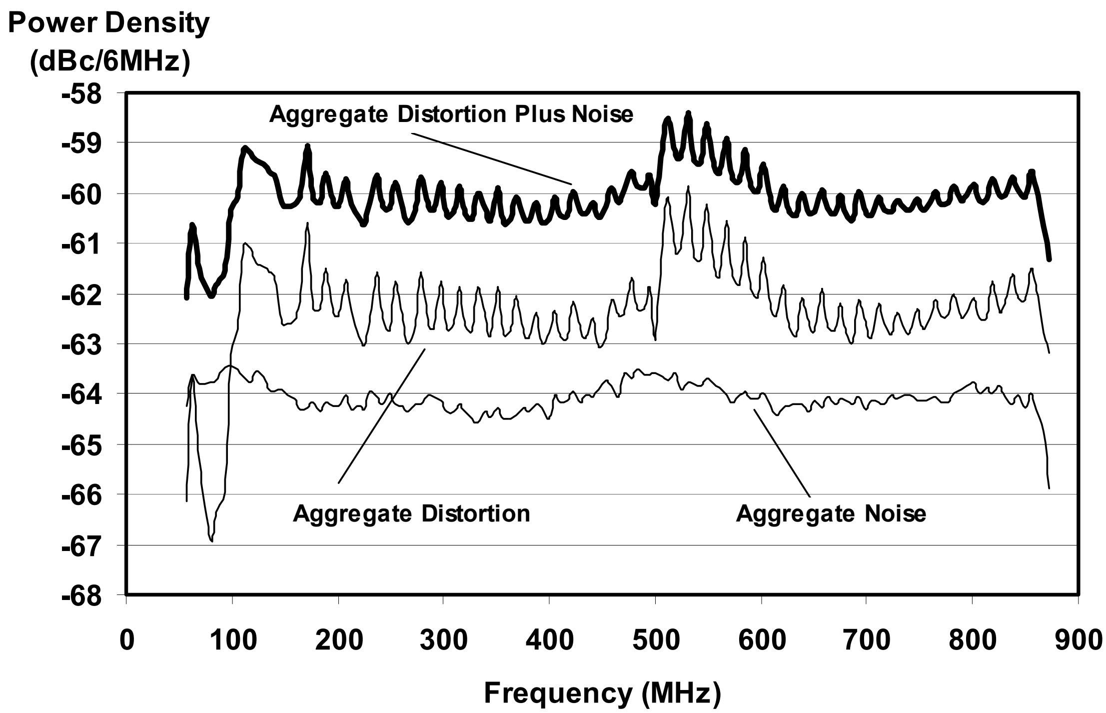

Figure 2-7. Plots of typical 136 channel self-aggregate broadband composite noise with its thermal noise and distortion components shown separately. Aggregate levels are referenced to the desired QAM signal level. Note that units compliant with DOCSIS RF specifications can aggregate under the same conditions up to a level of – 49.3 dBc per 6MHz[2], a level that is 9 dB higher than the highest level shown above.

Several characteristics of the plots are noteworthy:

1. Although aggregate distortion makes a slightly stronger contribution to aggregate composite noise than does aggregate thermal noise, their level in this upconverter device is comparable, generally within 1.5 dB. This is a significant indicator for an overall near optimal upconverter design, as it can be shown that in a system which maximizes dynamic range, the noise and distortion contributions are equal.
2. The relatively better aggregate distortion performance between 50 and 100 MHz is due to at least 2 factors:
    a. Lack of 2nd harmonic contribution in this region.
    b. Improved spectral regrowth performance due to non-uniform channel spacing.
3. The 18 MHz periodicity in the aggregate distortion plot is due to the 3rd harmonic characteristics (see Figure 2-5). Every 3rd channel represents the peak of the modulated 3rd harmonic.

4. The relatively better aggregate distortion performance in the small regions centered about 159 MHz and 225 MHz are in fact due to measurement errors associated with treatment of the harmonics generated when tuned to the offset channels 5 and 6 (centered at 79 MHz and 85 MHz respectively).

It should be emphasized that the aggregate composite noise calculated in Figure 2-7 is the level that would have been seen if 136 identical upconverters of the type that was measured were to be combined. In practice, fewer channels are likely to be combined for digital services in the foreseeable future. However, this measure provides a good guideline and a comparison yardstick. The worst channel result of –58.5 dBc shown for a typical upconverter we evaluated is approximately better by 9 dB than that which may result from an aggregation of 136 DOCSIS compliant transmitters [2].

2.3.3 Other Error Correction Considerations

In section 2.3.2, two potential sources of distortion calculation error are identified:

1. All modulated distortion measurements include a DUT noise component that must be subtracted out.
2. Distortion components that result when the DUT is tuned to channels 5 and 6 must be treated differently as these components do not fall at the measurement locations illustrated in Figure 2-5. For example, the center of the 2nd harmonic when the DUT is tuned to channel 5 will sit at 158 MHz, which is not a channel boundary. Thus, care must be taken when assigning the resulting distortion measurement to a particular measured channel.

An additional source of calculation error is due to the fact that under certain circumstances we capture multiple distortion terms in a single measurement, despite our intention to capture a single term. In the particular case of the upconverter we were evaluating, the mixer cross term (of the form $F_0 - f$ MHz, where F_0 is a constant and f is the frequency to which the DUT is tuned) would occasionally coincide with other distortion terms. These coincident terms are clearly deterministic and the data reduction process can be programmed to avoid double counting of terms. Moreover, approximate values of the individual terms can be extrapolated based on unambiguous measured values obtained on nearby channels.

3 Relationship of Single Unit Self-Aggregate Composite noise Performance to Real Headend Aggregate Noise Performance

The self-aggregate composite noise performance of a particular transmitter describes the aggregate performance of a headend comprised of multiple copies of that particular transmitter. The self-aggregate specification of a transmitter differs from traditional transmitter specifications in that it bounds the transmitter performance in aggregate, whereas traditional specifications bound individual channel performance of a single transmitter.

For example, DOCSIS specifies adjacent channel and other channel spurious and noise levels. These specifications apply to every tuned channel of the transmitter. Using these specifications, an operator can bound the aggregate performance of a headend comprised of individual (different) transmitters, each of which satisfies DOCSIS specifications on all channels. Thus, a specific virtue of individual channel bounds is that headend aggregate performance can be readily bounded.

What statements can be made about the aggregate performance of a headend comprised of transmitters that satisfy a particular self-aggregate bound, but for which individual channel spurious and noise performance is not (explicitly) specified?

We have demonstrated in this paper that the self-aggregate performance of a transmitter is derived from individual channel spurious and noise characteristics such as those depicted in Figure 2-1. Moreover, the aggregate performance of a headend is clearly a function of the individual channel spurious and noise performance of its various transmitters.

If individual channel performance is highly correlated among different transmitters (i.e. if all transmitters when tuned to a particular channel *n* manifest similar spurious plus noise performance), which would be the case for transmitters produced by the same manufacturer (since the individual channel performance is a manifestation of a common transmitter design), then we would expect that the self-aggregate performance of a single transmitter would be highly predictive of the aggregate performance of a headend comprised of multiple transmitters.

Alternatively, if the individual channel performance of the same transmitter tuned to different channels is highly correlated, then we might expect that the aggregate performance of a sample headend based on worst-case selection would be tightly related to the self-aggregate performance of the worst transmitter if that transmitter is a part of such sample headend.

For example, Figure 2-4 depicts broadband noise measurements for a particular transmitter tuned to several different channels. We observe a high degree of correlation in the broadband noise profile for nearby channel tunings. This is not a surprising result, as the noise profile is a function of a small number of transmitter blocks, most of which are engaged across multiple channel tunings. If this transmitter was a particularly bad transmitter within a sample headend, then we might expect that its self-aggregate performance would provide an aggregate noise bound for the sample headend in which it is installed.

Further study of the statistics of ensemble and individual channel spurious plus noise characteristics of transmitters is required before definitive statements can be made about the degree to which the self-aggregate performance of individual transmitters is predictive of an upper bound for the aggregate headend performance over an ensemble of headends. It is argued, however, that it is a good approximation and thus provides a useful comparative figure of merit that we encourage the industry to use.

4 Summary

This paper intends to provide insight into the derivation and measurement of transmitter self-aggregate composite noise, which the authors believe will become important criteria for cable operator's selection of QAM transmitters. A rationale is provided for separate treatment of broadband noise measurements and broadband modulated distortion measurements, and specific methods for accomplishing each measurement type are described. Techniques for combining these terms into a single self-aggregate composite noise measure are presented. Finally, this self-aggregate measure, which is an attribute of a single transmitter, is related to the performance of a real headend consisting of multiple different transmitters. A transmitter that has a self-aggregate composite noise value for 136 channels that exceeds – 49 dBc would most likely not support further digital channel expansion and a collection of such transmitters can render 256 QAM links unreliable in normal HFC systems. Cable operators are encouraged to require their QAM transmitter vendors to provide information on their product's aggregate noise levels for the proper number of channels.

5 References

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