Accuracy and noise performance of logarithmic amplifiers is very important for modern EW systems. This is especially true, when after digitizing the analog signal; a system engineer applies digital signal processing in an attempt to correct for analog errors. In this case, errors due to noise contributions have to be less than the desired accuracy prior to the digital error correction, otherwise, the error correction may actually degrade the overall accuracy instead of improving it.

Generally, the tradeoff between measurement time and accuracy is well known as a major tradeoff in measurement systems. Each case requires specific analysis and an appropriate solution. In nonlinear systems like logarithmic amplifiers, analysis could be complicated and the results of the analysis may not be apparent.

We will discuss noise performance of logarithmic amplifiers and their settling time performance. We will then present the most desirable approach for digital logarithmic amplifiers.

**TYPICAL BLOCK DIAGRAM OF EXTENDED RANGE DLVAs**

![Typical Block Diagram of Extended Range DLVAs](image)

**FIGURE 1**

**NOISE PERFORMANCE OF LOGARITHMIC AMPLIFIERS**

Before developing MITEQ’s first digital output log amplifier, we had a long and successful history in the design and manufacturing of low-noise successive detection logarithmic video amplifiers commonly known as SDLVAs. MITEQ also developed extended range detector logarithmic video amplifiers as shown in Figure 1.

Incorporating an A/D converter into the design of an extended range detector log video amplifier is considerably more difficult than it appears. The high noise content inherent in a DLVA and the non-linear nature of logarithmic signal processing makes this design considerably more complicated than most mixed signal designs.

A plot of the noise power versus RF input power for both SDLVAs and ERDLVAs are shown in Figure 2. As seen in the plot, the SDLVA exhibits a non-linear increase in signal-to-noise ratio as input power increases. On the other hand the carrier-to-noise ratio of the ERDLVA looks entirely different, increasing linearly for more than half the input dynamic range. This results in higher noise output levels, which inhibits accurate digitizing and error correction. In both cases the tangential sensitivity (TSS) is quite low, which is no surprise because TSS correlates with noise at low input power levels.

**TYPICAL NOISE POWER OF LOGARITHMIC AMPLIFIERS**

![Typical Noise Power of Logarithmic Amplifiers](image)

**FIGURE 2**
There are three major sources of noise that can effect a DLVAs performance. Two sources are the result of RF noise downconversion. The third source is the video amplifier. To simplify our discussion we will define continuous white RF noise as in triple independent sources located equidistant within the frequency spectrum with 1 MHz spacing. This approximation is illustrated in Figure 3.

The definitions are as follows:

- **Noise-Noise Term-Noise Downconversion or N&N**
  Resulting from the RF noise downconversion due to beating between the different noise components.

- **Noise-Carrier Term-Noise Downconversion or N&C**
  Resulting from the RF noise downconversion due to beating between the RF carrier and the noise components.

- **Video Amplifier Noise Contribution**

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**FIGURE 3**

![Noise Spectrum Approximation Diagram](image-url)
NOISE-CARRIER TERM-NOISE DOWNCONVERSION

Each arbitrarily taken noise component, with frequency $F_i$ beats with carrier $F_c$ and creates a baseband product with a frequency of $F_c - F_i$. The resulting RMS voltage will be proportional to the RMS voltages of the carrier and the RF noise component. Therefore the RMS voltage of one beating pair ($V_{oneBP}$) conversion is:

$$V_{oneBP} = 2\sqrt{2} \left( \frac{K_d}{R_v} \right) \cdot V_c \cdot V_{RF\text{noise}}$$

(1)

The power of this baseband component expressed in mW:

$$P_{oneBP} = 2 \left( \frac{K_d}{R_v} \right)^2 \cdot P_c \cdot P_{RF\text{noise}} = 2M^2 \cdot P_c \cdot P_{RF\text{noise}}$$

(2)

Where:

- $V_c$ is the RMS voltage of the carrier
- $P_c$ is the power of the carrier expressed in mW
- $V_{RF\text{noise}}$ is the RMS voltage of the noise within a 1 MHz bandwidth
- $K_d$ is the detector’s sensitivity expressed in mV/mW
- $R_v$ is the detector’s video resistance in ohms
- $M$ is the detector’s figure of merit, which is equal to $\sqrt{\frac{K_d^2}{R_v}}$
- $P_{RF\text{noise}}$ is noise power in mW within a 1 MHz bandwidth

Noise power in mW in a 1 MHz bandwidth itself depends on RF gain ($G_{RF}$) and noise figure ($NF_{RF}$).

$$P_{RF\text{noise}} = kT \cdot 10^{0.1(G_{RF} + NF_{RF})}$$

(3)

Total noise power density ($NPD$ in mW/MHz) in the baseband depends on the number of beating pairs ($N$) that will provide the equal baseband frequencies ($F_c - F_i$).

$$NPD = P_{oneBP} \cdot N$$

(4)

In the case of noise-carrier beating, $N$ is frequency independent and equals two because the same product could be obtained from the carrier beating with two noise components, first with frequencies lower than the carrier and second with frequencies higher than the carrier.

Combining Equations (2) and (4) gives us:

$$NPD = 4M^2 \cdot P_{RF\text{noise}} \cdot P_c$$

(5)

The total noise power in mW of the noise-carrier term ($P_{N&C}$) in the video bandwidth ($BW_{vid}$) in MHz is equal to:

$$P_{N&C} = NPD \cdot BW_{vid} = 4M^2 \cdot P_c \cdot P_{RF\text{noise}} \cdot BW_{vid}$$

(6)

$P_{N&C}$ does not depend on the RF bandwidth and is directly proportional to the following:

- The video bandwidth
- The RF noise figure as part of Equation (3)
- The input signal level
APPLICATION NOTES

NOISE-NOISE TERM-NOISE DOWNCONVERSION

Two arbitrarily taken noise components with frequencies $F_i$ and $F_j$ mix with each other to create two products, one with the RF frequency $F_i + F_j$ and the other with baseband frequency $F_{bb} = |F_i - F_j|$. For our discussion only the baseband product is of interest.

Equation (2) is applicable to the noise-noise downconversion with the substitution of $P_C$ by another $P_{RF\text{noise}}$.

So the power of the baseband product of the noise-noise beating pair is equal to:

$$P_{\text{oneBP}} = 2M^2 \cdot P^2_{RF\text{noise}}$$  \hspace{1cm} (7)

Where:

$P_{RF\text{noise}}$ is equal to the noise power in mW within a 1 MHz bandwidth, and expressed by Equation (3).

The number of beating pairs differs from the previous case significantly. In this case $N$ depends on the frequency of the baseband product $F_{bb}$ where $F_{bb} = |F_i - F_j|$. It is apparent that each component gives a 1 MHz product when beating with adjacent components. Please refer to our approximation which assumes 1 MHz spacing between independent noise sources. So where $F_{bb} = 1$ MHz

$$N_{@F_{bb} = 1 \text{ MHz}} = \frac{\text{BW}_{RF}}{1 \text{ MHz}}$$

For producing a 2 MHz product, each component has to mix with another one shifted by 2 MHz. In comparison with a 1 MHz product, we lose one beating pair.

$$N_{@F_{bb} = 2 \text{ MHz}} = \frac{\text{BW}_{RF}}{1 \text{ MHz}} - 1$$

And so on,

$$N_{@F_{bb} = 3 \text{ MHz}} = \frac{\text{BW}_{RF}}{1 \text{ MHz}} - 2$$

$$N_{(F_{bb})} = \frac{\text{BW}_{RF} - F_{bb}}{1 \text{ MHz}}$$  \hspace{1cm} (8)

Combining Equations (4), (7) and (8) we get a baseband noise power density, $NPD(F_{bb})$, expressed in mW/MHz.

$$NPD(F_{bb}) = 2(K_d^2/R_v) \cdot P^2_{RF\text{noise}} \cdot (\text{BW}_{RF} - F_{bb})$$  \hspace{1cm} (9)
NOISE-NOISE TERM-NOISE DOWNCONVERSION (CONT.)

Therefore the total noise power in mW in the video bandwidth \( (\text{BW}_{\text{vid}} \text{ in MHz}) \) can be derived from the following:

\[
P_{\text{N\&N}} = \sum_{\text{F}_{\text{bb}}} \text{NPD} (\text{F}_{\text{bb}}) = M^2 \cdot (P^2_{\text{RFnoise}} \cdot (2\text{BW}_{\text{RF}} \cdot \text{BW}_{\text{vid}} - \text{BW}_{\text{vid}}^2) = M^2 \cdot P^2_{\text{RFnoise}} \cdot \sqrt{\text{BW}_{\text{eff}}^2} \quad (10)
\]

Where effective bandwidth \( \text{BW}_{\text{eff}} \) is equal to:

\[
\text{BW}_{\text{eff}} = \sqrt{2\text{BW}_{\text{RF}} \cdot \text{BW}_{\text{vid}} - \text{BW}_{\text{vid}}^2} \quad (11)
\]

\( P_{\text{N\&N}} \) is therefore directly proportional to the following:

- The video bandwidth
- The RF noise figure as part of Equation (3)
- The RF bandwidth

\( P_{\text{N\&N}} \), on the other hand, does not depend on input signal level.

VIDEO AMPLIFIER NOISE CONTRIBUTION

The noise power, expressed in mW, of the video amplifier into the video bandwidth is equal to:

\[
P_{\text{vidAmpNoise}} = 4kT \cdot \text{F}_{\text{vid}} \cdot \text{BW}_{\text{vid}} \quad (12)
\]

Where the noise factor of the video amplifier \( (\text{F}_{\text{vid}}) \) is directly related to its noise figure as:

\[
\text{NF}_{\text{vid}} = 10\log (\text{F}_{\text{vid}}) \quad (13)
\]
If the video bandwidth is less than the RF bandwidth, the noise-carrier term is independent of RF bandwidth. The noise-noise term is a function of video bandwidth. These differences become clear when we look at the noise power density of the terms.

Noise power density of the noise-carrier term is flat because the number of beating pairs equals two and is independent of the baseband product frequency. Extending the RF noise bandwidth does not increase the NPD, instead it stretches the baseband noise spectrum, as seen in Figure 4. Therefore the portion of the baseband noise spectrum, which remains after video filtering (shown by gray bar) does not depend on the RF bandwidth.

As seen in Figure 5, in the case of noise-noise downconversion, the number of beating pairs does depend on the RF bandwidth, as well as the frequency of the baseband product. Therefore the portion of the baseband noise spectrum which remains after video filtering (shown by gray bar) increases proportionally with increasing RF bandwidth.

The difference between the N&N term and the N&C term is very important. It shows that by narrowing the RF bandwidth, one could improve TSS due to the reduction of the N&N term. At the same time it does not reduce noise at the medium to high input power levels (almost the entire input dynamic range of the log amplifier), where the N&C term is dominant.
When we understand each noise source, obtain expressions of their noise power density [Equations (5) and (9)] and their noise power in the video bandwidth [Equations (6), (10) and (12)], we are able to compare the contribution of these noise sources and understand which of the noise sources dominates under different conditions.

The N&C term is the only one which is directly proportional to the input power level of the carrier. This means that at low input power levels, near TSS, the N&C term is insignificant. Therefore, at low power levels, only the N&N [Equation (10)] and the video amplifiers noise [Equation (12)] make significant contributions to the output noise power.

Since N&C term is directly proportional to the input power, there must be a level \( P_{Ccr} \), where these terms have equal contribution with the N&C term. Above \( P_{Ccr} \), the N&C term will be dominant.

In comparing the three noise terms, the N&N terms contribution is very small (see NOTE 1) and may be disregarded for practicality. Therefore, \( P_{Ccr} \) occurs when the N&C term [Equation (6)] is equal to the video amplifiers noise term [Equation (12)] and can be calculated by the expression:

\[
P_{Ccr} = 30 - 2G_{RF} - NF_{RF} + NF_{vid} - 20\log M \tag{14}
\]

Critical input power is an important parameter, because noise dependence versus input power level changes significantly around this point.

To calculate the noise power at the output of the log amplifier, one has to go through a few transformations. We convert the noise power at the detector’s output [Equations (6), (10) and (12)] into RMS voltages. We then multiply these voltages with the log video transfer function then transform back into power.

Upon completion we get the noise power of the log amplifier’s output, in dBm, for each noise term.

For the video amplifier’s noise term, which dominates when input power is less than the critical point:

\[
P_{NoiseOut}^{VidAmp} = -95 + 10\log(Sl^2/R_{load}) + 10\log BW_{vid} + NF_{vid} - 20\log M - 2G_{RF} - 2P_C \tag{15}
\]

Where:

- \( Sl \) is the slope of the DLVA in mV/dB
- \( R_{load} \) is the video amplifier’s load resistor

The important feature of this equation is the rapid decline of noise power as a function of input power. This is due to the term of \(-2P_C\).

For the N&C term, which dominates when input powers are above the critical point:

\[
P_{NoiseOut}^{N&C} = -125 + 10\log(Sl^2/R_{load}) + 10\log BW_{vid} + NF_{RF} - P_C \tag{16}
\]

The important feature of this equation is a moderate decline of noise power as a function of input power. This is due to the term of \(-P_C\).

NOTE 1: This assumes that the RF system noise is insignificant. Usually the RF systems noise is greater than the video amplifier noise. Under this condition the N&C term and the N&N term will be most significant in the noise calculations.
Plots of the output noise power versus input power for the ERDLVA with two detector stages are shown in Figures 6A and 6B. Figure 6A presents the results taken on an ERDLVA with a center frequency of 500 MHz, with an input dynamic range of -70 to 0 dBm. Figure 6B presents the results taken of an ERDLVA with a center frequency of 5.6 GHz with an input dynamic range of -54 to +16 dBm.

Both examples demonstrate the measured output noise power correlate with results of our calculations. The important thing is, that for the high power stage, the noise versus input power degradation exhibits a constant slope of -2 while the low power stage changes its slope. It is -2 at low power levels and -1 at higher input power levels. When the input signal level is within the operating range of the low power stage, the noise from the high power stage remains at the same high level and dominates over the noise of the low power stage.

Now that we understand all the sources of noise let us analyze the design of a low-noise DLVA.

One can see from Figure 6A and 6B that the high power stage makes a significant contribution in the total output noise magnitude even for input powers below this level. This being the case, to design a low-noise DLVA, we must suppress the noise from the high power stage over the input dynamic range of the low power stage. By doing this we only utilize the low noise portion of the high power stage.

Figure 7 presents the measured results of the DLVAs noise performance under the following conditions:

• The performance of a standard ERDLVA.
• The noise performance of the low power stage when the high power stage is disconnected.
• The noise performance of the low power stage shifted into the range of the high power stage while the high power stage is disconnected.

The noise reduction then becomes apparent. The only drawback of this approach is a substantial reduction in input dynamic range of the analog output. However, we have developed a technique that would allow the restoration of the entire input dynamic range. This technique substantially improves processing speed and accuracy due to the aforementioned noise reduction.

We have applied this approach of noise reduction to an ERDLVA with a center frequency of 5.6 GHz and an input dynamic range of -54 to +16 dBm. The results are shown in Figure 8. The first two plots in Figure 8 present the log accuracy measured at the digital output over temperature, after digital error correction. The last graph presents the RMS error of the digital output, which is directly related to the noise of the analog portion. This RMS error was calculated on the basis of 400 samples for each 0.2 dB of input dynamic range. It becomes apparent from this plot that logarithmic error reduces to less than ±0.25 dB and that three times the RMS error is less than ±0.2 dB. Successful usage of digital error correction is possible only due to the substantial suppression of the RMS noise level. A different approach to suppress the noise is using additional filtering. This can be done using either analog or digital techniques. However, additional filtering means narrower video bandwidth, which leads to increased settling time thereby increasing the minimum pulse width, which can be accurately measured.

MITEQ has overcome this limitation and offers an ERDLVA, which measures pulses as narrow as 250 ns, with an accuracy of ±0.25 dB over 70 dB of input dynamic range.
FIGURE 6A

OUTPUT NOISE POWER OF A 500 MHz ERDLVA

FIGURE 6B

OUTPUT NOISE POWER OF A 5.6 GHz ERDLVA
APPLICATION NOTES

LOW-NOISE ERDLVA DESIGN TYPICAL TEST DATA (CONT.)

FIGURE 7
APPLICATION NOTES

DIGITAL ERDLVA DESIGN TYPICAL TEST DATA

MODEL NO. FBDL-5.4/5.9-70
SERIAL NO. 489150

Temperature @ 20°C
Absolute Accuracy, dB

Temperature @ 40°C
Absolute Accuracy, dB

Temperature Reading @ 40°C
Digital Reading

4 Sigma, Min., Max.

FIGURE 8