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Noise Analysis Adjustable Bootstrap Transimpedance and Voltage Feedback Receiver Amplifier

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Abstract: An understanding of the noise is required for a receiver's performance to be precisely characterized. The amount of noise present in a receiver will be the primary factor that determines the receiver's sensitivity. In this study, the definition of noise in infrared communication is discussed to demonstrate the effects of receiver noise on performance for adjustable bootstrap transimpedance and voltage feedback receiver amplifier. The noise sources that are commonly found in an optical wireless receiver are then discussed, including noises that are of optical as well as of electrical origin. The equivalent circuit model that will describe the noise performance of the bootstrap transimpedance and voltage feedback amplifier circuit shows, that the input noise current density is flat 380 pV/$\sqrt{\text{Hz}}$ from 1 Hz to 10 GHz and starts to increase. The output noise density, shows a flatness of 1.1nV/$\sqrt{\text{Hz}}$ from 1 Hz to 80 MHZ when it starts descending according to the capacitor value which adjust the bandwidth. Results also showed that the output noise density remains constant during the bandwidth adjustment process.

Key words: Ambient noise, thermal noise, shot noise

INTRODUCTION

Noise, defined in the broadest practical terms, is any signal present in the receiver other than the desired signal or any unwanted disturbance that masks, corrupts, reduces the information content of or interferes with the desired signal. In the optical wireless communication environment, it is known as "ambient noise". The sources of noise available in a receiver circuit are divided into two classes (Alexander, 1997):

- Intrinsic noise sources arising from fundamental physical effects in optoelectronic and electronic devices used to construct the receiver
- Coupled noise sources arising from interactions between receiver circuitry and the surrounding environment

In addition, noise in a receiver can be described as either additive or signal-dependent. Additive noise is a source of noise that is present whether there is a signal at the receiver or not, while a signal-dependent noise is one that is observed only when there is a signal present at the receiver. Figure 1 illustrates a simple model for an optical receiver and the major contributors to the noise present in the receiver. The received signal and any optical background that may be present, are photodetected and then amplified in a linear signal path.

The implications of each noise source will be addressed, shown right to left in Fig. 1. Receiver electronic noise consists of three primary components: thermal noise, electronic shot-noise and 1/*f* noise. Thermal noise is the type most often associated with receivers. Also known as Johnson noise, it is a result of thermally-induced random fluctuations in the charge carriers in a resistance. The power spectral-density for thermal noise is white for all frequencies within a defined bandwidth and, since thermal noise inherently results from the accumulated effect of large quantity of individual charge motions, it exhibits Gaussian statistics. Nyquist showed that the open circuit RMS voltage produced by a resistance R is as follows (Nyquist, 1928; Baker, 2005):

$$V_n = \sqrt{4kTBR} \text{ (volt rms)} \tag{1}$$

where, k is Boltzmann's constant, T is absolute temperature in Kelvin and B is the observation bandwidth in Hz.

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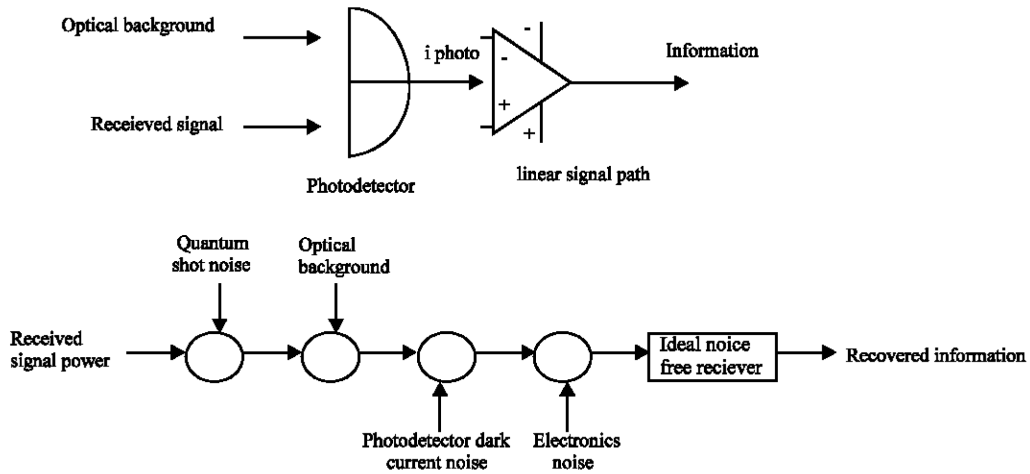

Fig. 1: Simple receiver model and noise sources in the receiver (Alexander, 1997)

Electronic shot noise is that associated with the passage of carriers across a potential barrier. This means that any photocurrent in the photodiode will have electronic shot-noise associated with it. Based strictly on the characteristics of the noise, it is impossible to distinguish between the quantum shot noise arising from photons being detected with a photodiode and the electronic shot noise arising from the photocurrent flowing through the junction in the photodiode. Therefore, the total electronic shot noise associated with a current, I_{DC}, flowing through a potential barrier is given as follows (Baker, 2005):

$$I_{shot} = \sqrt{2qI_{DC}B} \text{ (amps)} \tag{2}$$

where, q is the electronic charge and B is the bandwidth

$1/f$ noise has been observed as low frequency fluctuations in the resistance of a semiconductor. In resistors it is called "excess noise" and has been a concern in optical receiver design when the receiver is required to have a low frequency cut-off that is less than a few tens of MHZ. The amount of $1/f$ noise present depends on the choice of transistors used in the first stages of the receiver. A Si bipolar transistor's $1/f$ noise is noticeable in the tens of kHz region.

The noise in an optical wireless receiver can be influenced by additional non-signal related sources of optical energy that fall on the photodetector, namely quantum shot noise and optical background noise. Quantum shot noise is the result of the discreteness of photon arrivals. It is due to background light sources, such as sunlight, fluorescent lamp light and incandescent lamp light, as well as the signal-dependent source. Since the background light striking the photodetector is normally much stronger than the signal light, the dependency of noise on the input signal may be neglected and the photon noise can be considered to be additive white Gaussian noise (Carruthers, 2002). The amount of background radiation collected by a free-space optical receiver is dependent on the receiver's field-of-view, as well as its optical bandwidth.

LITERATURE REVIEW

Noise model of a transimpedance amplifier: Amplification of low-level signals is a critical function of any receiver, due to the noise sources mention earlier. The overall noise performance of an amplifier is related to the noise characteristics of the individual devices and components that form the amplifier circuit. Two established techniques have been used as indicators of the noise performance of an amplifier. The first, known as the noise figure, NF, is simply the noise factor, F, expressed in terms of dB:

$$\text{where, } F = \frac{SNR_{input}}{SNR_{output}} \quad \text{and } NF = 10\log_{10}(F) \text{ (dB)} \tag{3}$$

The first formula for the overall noise factor for a three stage amplifier combination is given by:

$$F_{Total} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} \tag{4}$$

where, F_1 and G_1 are noise factor and power gain of first stage amplifier.

F_2 and G_2 are noise factor and power gain of second stage amplifier etc and all amplifiers are assumed to have the same bandwidth.

Equation 4 shows that if the gain of the first stage is large, the noise performance of a cascade of subsystems will be dominated by the noise performance of the first-stage. In an optical wireless receiver, the photodetector and the first stages of the amplifier are known as the receiver front-end. This will generally be the principal factor in determining the overall noise performance and sensitivity of the receiver. Unfortunately, noise figure has some drawbacks when used to describe noise performance of amplifiers intended for an optical wireless receiver. It is defined using a resistive source, but a photodiode is dominated by capacitance, so the magnitude of the source impedance varies with frequency in the first stage amplifier in the receiver. Therefore the second technique that overcomes the noise figure drawbacks is to model an amplifier using equivalent noise sources, $V_n(\omega)$ and $I_n(\omega)$, as illustrated in Fig. 2 for a receiver consisting of a signal source such as a photodetector and an amplifier.

Assuming only amplifier noise, using the noise analysis principles approach discussed by Motchenbacher and Fitchen (1973) and the principle of superposition in linear circuits, the noise power at the amplifier output is given by:

$$V_{no-1}^2(\omega) = I_n^2(\omega)\frac{Z_f(\omega)^2}{[1+\frac{1}{A(\omega)}+\frac{Z_f(\omega)Z_s(\omega)}{A(\omega)}\}]^2} \tag{5}$$

Replacing $V_n(\omega)$ and $V_{n\text{-}r}(\omega)$ as short circuits, where $Z_f = R_f/C_f$

Then, the output noise for voltage noise is given by:

$$V_{no-2}^2(\omega) = V_n^2(\omega)\frac{[1+Z_f^2(\omega)Z_s^2(\omega)]}{[1+\frac{1}{A(\omega)}+\frac{Z_f(\omega)Z_s(\omega)}{A(\omega)}]^2} \tag{6}$$

Replacing $I_n(\omega)$ as an open circuit and $V_{n\text{-}r}(\omega)$ as a short circuit

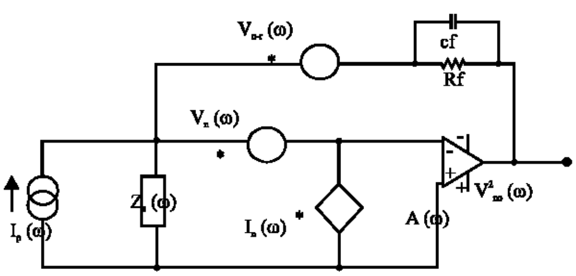

Fig. 2: Noise model of amplifier

$$V_{no-3}^2(\omega) = V_{n-r}^2(\omega)\frac{1}{[1+\frac{1}{A(\omega)}+\frac{Z_f(\omega)Z_s(\omega)}{A(\omega)}]^2} \tag{7}$$

Replacing $I_n(\omega)$ as an open circuit and $V_n(\omega)$ as a short circuit

The transfer function:

$$\frac{V_{no}(\omega)}{I_p(\omega)} = \frac{-Z_f(\omega)}{1+\frac{1}{A(\omega)}+\frac{Z_f(\omega)Z_s(\omega)}{A(\omega)}]} \tag{8}$$

The total output source and amplifier noise is:

$$I_{neq}^2(\omega) = \frac{V_{no-1}^2(\omega)+V_{no-2}^2(\omega)+V_{n0-3}^2(\omega)}{[\frac{V_{no}(\omega)}{I_p(\omega)}]^2} \tag{9}$$

Substituting (5) and (6) into (8), the simplified equivalent input current-noise:

$$I_{neq}^2(\omega) = I_n^2(\omega)+V_n^2(\omega)\{\frac{1}{Z_f^2(\omega)}+Z_s^2(\omega)\}+V_{n-r}^2(\omega)\{\frac{1}{Z_f^2(\omega)}\} \tag{10}$$

Since a photodetector is a capacitive current source, $Z_s(\omega) = j\omega C_d$, setting $Z_f(\omega) = R_f$:

$$I_{neq}^2(\omega) = I_n^2(\omega)+V_n^2(\omega)((\frac{1}{R_f^2}+(j\omega C_d R_f)^2)+V_{n-r}^2(\omega)(\frac{1}{R_f^2}) \tag{11}$$

Equation 11 shows that the influence of voltage-noise increases with frequency. The detector capacitance acts as high pass filter to the voltage noise sources of the amplifier. At low frequencies the contribution of voltage noise to the overall current flowing is small, due to the large impedance of the capacitor. At high frequencies, the amounts of circulating current due to voltage noise increase because the capacitor's impedance decreases. The implications of noises for the front-end of an optical wireless receiver can be summarised in Fig. 3.

Assume an optical signal and background noise impinge on the photodetector inducing a current in the external load resistor. If P_t is the average optical power, then I_p the photocurrent, is given by:

$$I_p(\omega) = \frac{\eta q P_t(\omega)}{h\nu} \tag{12}$$

where, q is the electron charge, η is the quantum efficiency, h is Planck's constant and ν is the optical frequency of the light.

Therefore, the expressions for the current sources in the models illustrated in Fig. 2, assuming that a PIN photodetector is being used in the receiver, are as follows:

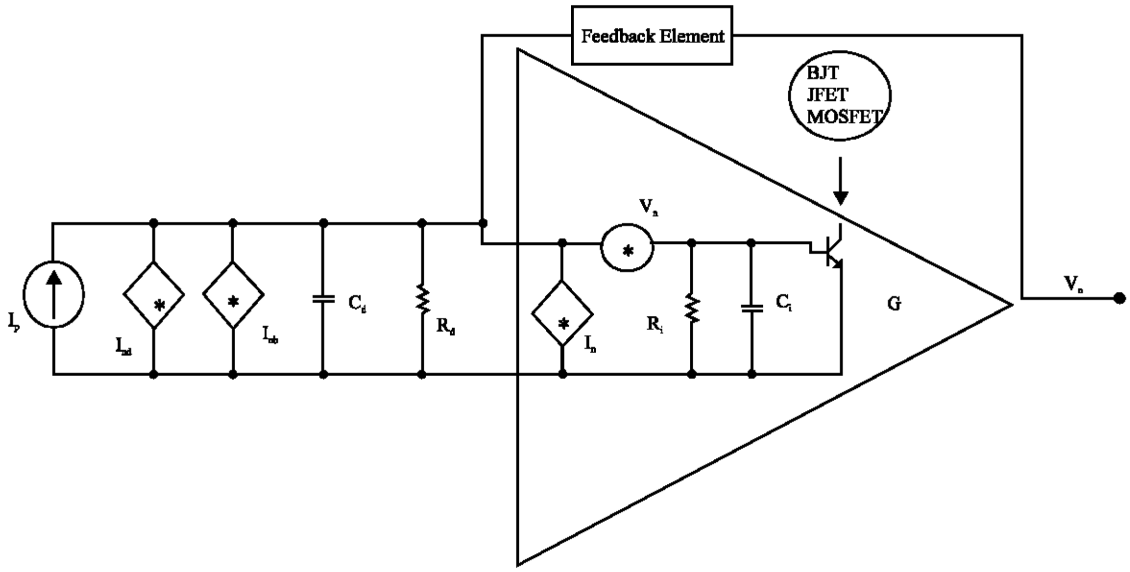

Fig. 3: An equivalent noise model of input stage of preamplifier, where I_p is the photocurrent, I_{nd} is the detector noise, I_{nb} is the background noise, C_d, R_d are capacitance and resistance of a detector, I_n, V_n are current noise and voltage noise of a preamplifier, R_i, C_i are input resistance and input capacitance of a preamplifier, G is the voltage gain of a preamplifier (Bielecki *et al*., 2003).

$$I^2_{Total}(\omega) = I^2_{neq} + 2qI_{dk} + I^2_{shot} + I^2_{nb} \tag{13}$$

$$I_{dk} = I_{dm} + I_b$$

Where:

- I_{dk} : is photodiode dark current
- I_{dm} : is unmultiplied dark current
- I_b : is DC photocurrent due to optical background
- I_{nb} : is optical background noise
- I_{neq} : is equivalent input current-noise from the receiver amplifier including photodiode thermal noise

If the general expression for $I_{Total}(\omega)$ is converted to frequency f in Hz, the equivalent input current-noise can be expressed as a power series:

$$I^2_{Total}(f) = I^2_{shot} + \sum_{j=-b}^{a} x_j f^j \tag{14}$$

where, I_{shot} is the spectral density of the unmultiplied quantum shot-noise associated with the signal and the coefficients x_j may, depending on the details of the receiver and system implementation, be proportional to received signal power, optical background, etc. The $1/f$ noise has been excluded in the above expression, since the receivers are used at high frequencies (Park *et al*., 1988; Su *et al*., 1983). In practical cases, Eq. 14 can be limited to (Muoi, 1987):

$$I^2_{Total}(f) = I^2_{shot} + x + x_1 f + x_2 f^2 \tag{15}$$

From Eq. 15, a figure of merit for an optical wireless receiver can be defined and used to describe the noise performance of the system, by assuming the amount of signal related quantum shot noise is essentially constant:

$$I^2_{Total}(f) = I^2_{shot} + I^2_{rcvr}(f) \tag{16}$$

I^2_{shot} is unmultiplied quantum shot-noise associated with the signal and $I^2_{rcvr}(f)$ is the equivalent input current noise due to all other noise sources present in the receiver.

Parameter k(f) is indicative of the difference between the receiver's total equivalent input current noise density and the noise-density due to the quantum shot noise of the signal as:

$$\kappa(f) = \frac{I^2_{Total}(f)}{I^2_{shot}} = \frac{I^2_{shot} + I^2_{rcvr}(f)}{I^2_{shot}} \tag{17}$$

Using Eq. 10 to describe the amplifier noise and Eq. 13 to describe photodetector noise, the signal photocurrent and dark current, then the equivalent input current-noise at the receiver's input is given by:

$$I^2_{Total}(\omega) = I^2_n(\omega) + V^2_n(\omega)\{[\frac{1}{Z_f^2} + Z_s^2(\omega)]\} + V^2_{n-r}(\omega)\{\frac{1}{Z_f^2}\} + I^2_{th-Rd} + 2qI_{dk} + I^2_{shot} + I^2_{nb} \quad (18)$$

where, $Z_s(\omega) = j\omega C_d$, $Z_f(\omega) = R_f$, neglecting C_f:

$$I^2_{th-Rd} = \frac{4kT}{R_d}$$

$$V^2_{n-r}(\omega) = 4kTR_f$$

Assuming that the receiver is illuminated by a signal with a constant power level P_t and ignoring any optical background, the equivalent input current noise density is:

$$I^2_{Total}(\omega) = I_n^2 + V_n^2\{\frac{1}{R_f^2} + [j\omega C_d R_f]^2\} + \frac{4kT}{R_f} + \frac{4kT}{R_d} + 2qI_{dk} + I^2_{shot} \quad (19)$$

$$I^2_{Total}(f) = I^2_{shot} + x_0 + x_2 f^2$$

Where:

$$x_0 = I_n^2 + \frac{V_n^2}{R_f^2} + \frac{4kT}{R_f} + \frac{4kT}{R_d} + 2qI_{dk}$$

$$x_2 = V_n^2\{4\pi^2 C_d^2 R_f^2\}$$

From Eq. 16, the degradation from the quantum shot-noise limited current density is then given by:

$$\kappa = 10\log\frac{I^2_{shot} + x_0 + x_2 f^2}{I^2_{shot}}\ \text{dB} \quad (20)$$

The total equivalent input current noise is:

$$I^2_{Total} = I^2_{shot}\int_0^{\infty}|H_t(f)|^2\,df + x_0\int_0^{\infty}|H_t(f)|^2\,df + x_2\int_0^{\infty}|H_t(f)|^2 f^2\,df \quad (21)$$

where, $H_t(f)$ is the amplitude normalized frequency dependent portion of the overall receiver transimpedance, given by $H_t(f) = 1$ for $0<f<B$, 0 for $f > B$.

METHODOLOGY

In this section, the discussion is focussed on the noise analysis for adjustable bootstrap transimpedance and voltage feedback amplifier circuits. The analysis is facilitated by the consideration of the input equivalent noise voltage and noise current. Bipolar Junction Transistors (BJTs) and Field-effect Transistors (FET) are one of the key building blocks of electronic amplifiers. The noise sources present in both the components have been extensively studied (Motchenbacher and Fitchen, 1973; Muoi, 1984; Smith and Personick, 1982; Sze, 1981; Abdullah *et al*., 2004; Abdullah and Green, 2010). The small-signal model for the FET and the BJT are shown in Fig. 4 and 5, respectively. To simplify the analysis of the FET, direct effects from the gate-drain feedback capacitor will be ignored. This is usually valid, since in most devices the gate source capacitance is approximately ten times the gate drain capacitance (Alexander, 1997).

For the BJT, the base spreading resistance, r_x, accounts for any resistance between the base terminal contact and the actual active base region of the device. The base current and collector current shot noises are accounted for by two noise current generators i_{bBJT} and i_{cBJT}. The resistances r_μ, r_μ and r_0 are dynamic resistances, they do not dissipate energy and do not contribute thermal noise. In this analysis, secondary effects that are introduced by the internal feedback r_μ and C_μ will be ignored for simplicity. C_π is composed of two contributions from the space charge in the emitter junction and the diffusion capacitance of the emitter junction that increases with emitter current.

The bootstrapped transimpedance amplifier is connected in series with a voltage feedback amplifier, the LMH6624. An RC filter is used as a termination as shown

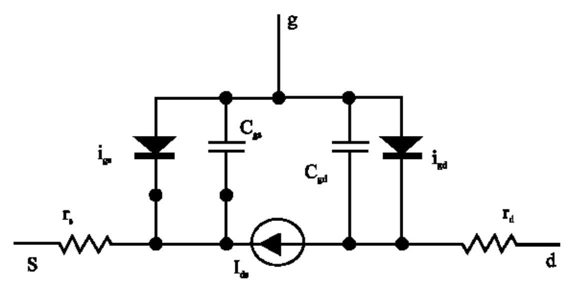

Fig. 4: FET small-signal model

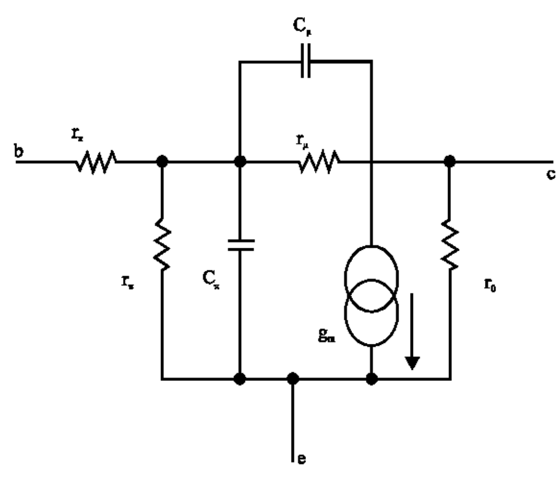

Fig. 5: BJT small-signal hybrid-π model

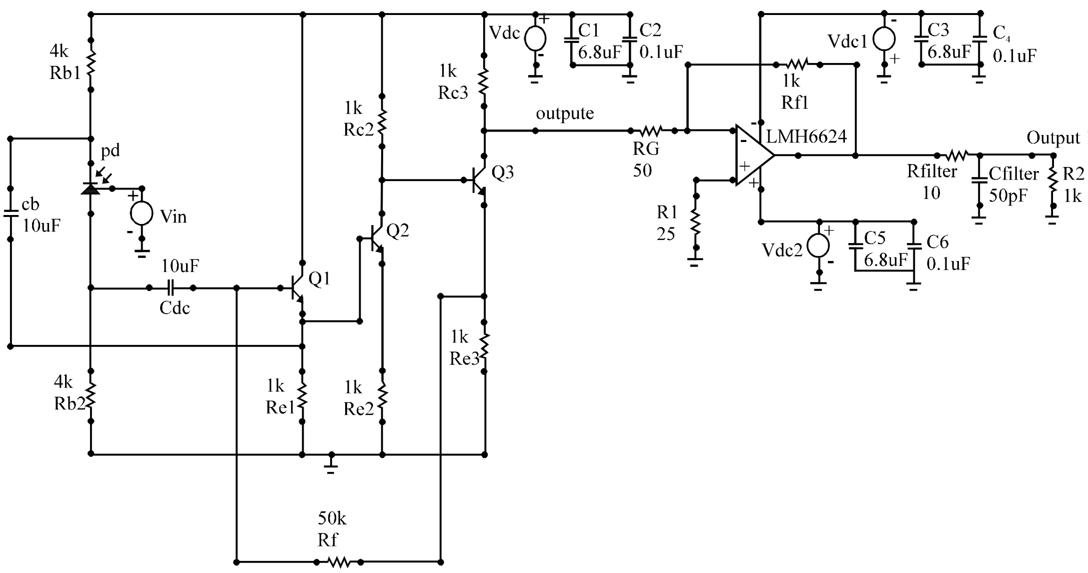

Fig. 6: Composite bootstrap transimpedance amplifier with VFA

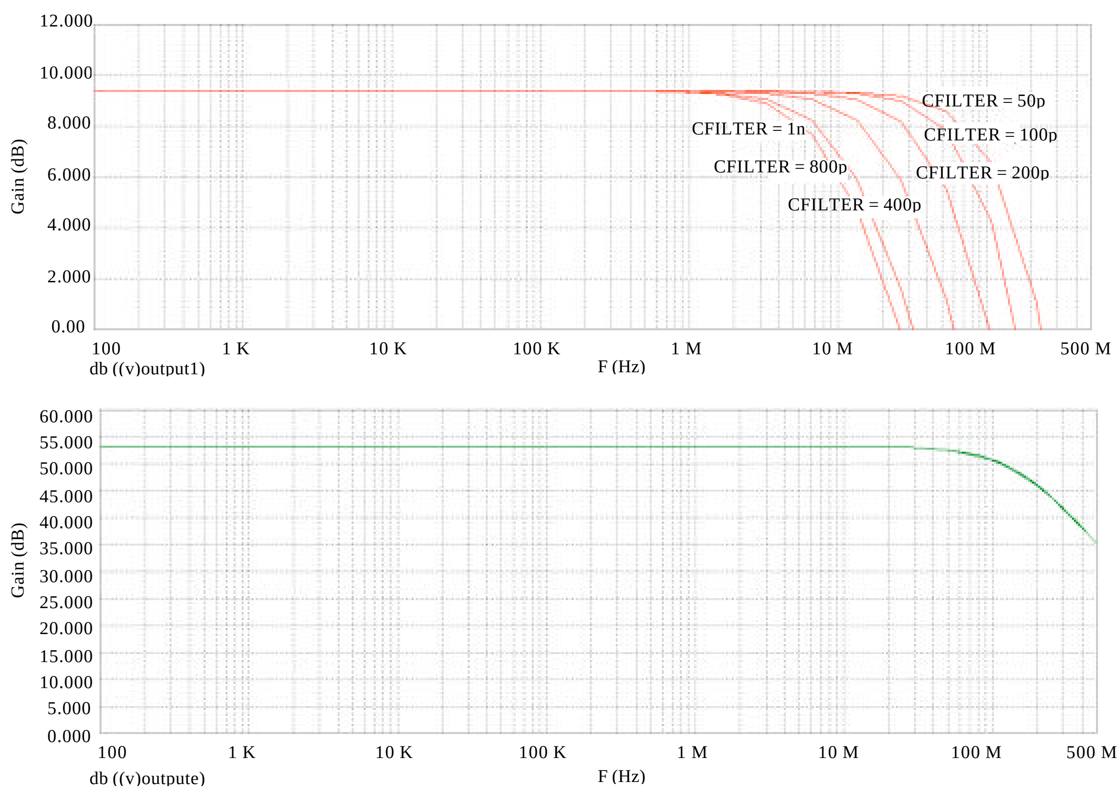

Fig. 7: Frequency response composite transimpedance amplifier

in Fig. 6. The first stage of the BTA produces a cut-off frequency of 94.5 MHZ, with a gain of 51.5 dB. By varying the capacitor, C_{filter}, of the RC filter between 50pF to 1nF, the bandwidth of the composite circuit can be controlled. The cut-off frequency obtained by this RC filter on the composite amplifier is from 9.5 to 103.5 MHZ, as shown on the frequency response plot of Fig. 7. The overall gain is reduced to 12.3 dB. There is a trade-off between gain and bandwidth, as the bandwidth is increased the gain of the circuit is reduced.

Assuming that the gain stages and the emitter follower can be approximated by the simplified amplifier model as shown in Fig. 8., R_{b1}, R_{b2}>>R_{e1} and frequencies are considered where C_{dc} and C_b are short circuits, than the transimpedance gain, A_z, for the circuit is approximated using the following assumption where A is the voltage gain of the first stage amplifier and A_1 is the voltage gain of the second stage amplifier:

$$A_z = \frac{V_{output}}{I_p} = \frac{AR_f R_{e3}}{R_{e3}(A-1) + AR_f - j\omega C_d R_{e3} R_f} \tag{22}$$

$$\frac{V_{output1}}{V_{output}} = -A_1 \frac{\frac{1}{j\omega C_{filter}}}{R_{filter} + \frac{1}{j\omega C_{filter}}} \cong \frac{-A_1}{j\omega R_{filter} C_{filter} + 1} \tag{23}$$

From (22) and (23):

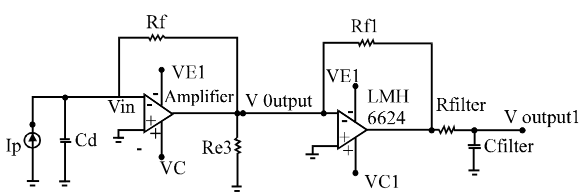

Fig. 8: Simplified model of Figure 6

$$\frac{V_{output1}}{I_p} = \frac{-AA_1 R_f R_{e3}}{1 + R_{e3}(A-1) + AR_f - j\omega C_d R_{e3} R_f + j\omega R_{filter} C_{filter}} \tag{24}$$

Equation 24 shows that the feedback resistor plays an important part in determining the gain of the circuit. As the feedback resistor, R_f increases, the gain of the overall system increases but, as the second stage amplification feedback resistor, $R_{f1,}$ increases, the overall system gain reduces.

ANALYSIS AND RESULTS

The combination of a bootstrap transimpedance amplifier with the feedback impedance (R_f in parallel with the parasitic C_f) has been referred back to the input, such that the input impedance is equal to $R_f//C_d//C_f$. The gain set for the voltage feedback amplifier is:

$$\frac{-R_f}{R_G} = \frac{-1000}{50} = -20$$

And the calculated noise figure for the voltage feedback amplifier will be 2.6, where, R_f = 1k Ω, R_1=25 Ω and R_G=50 Ω. Therefore, the input equivalent noise current density, is:

$$I^2_{nTotal}c(f) = I^2_{shot} + 2qI_{dk} + \frac{4kT}{R_d}10^{\frac{NF_{LMH6624}}{10}} f + \frac{2q}{\beta} I_{cBJT} + 4kTr_x\left[\frac{1 + (2\pi(C_d + C_f)R_f)^2}{R_f^2}\right]f^2 + 2q\frac{V_t^2}{I_{cBJT}}\left[\frac{1 + (2\pi(C_\pi + C_d + C_f)R_T)^2}{(R_T)^2}\right]f^2 \tag{25}$$

where, R_T=$R_f//r_\pi$

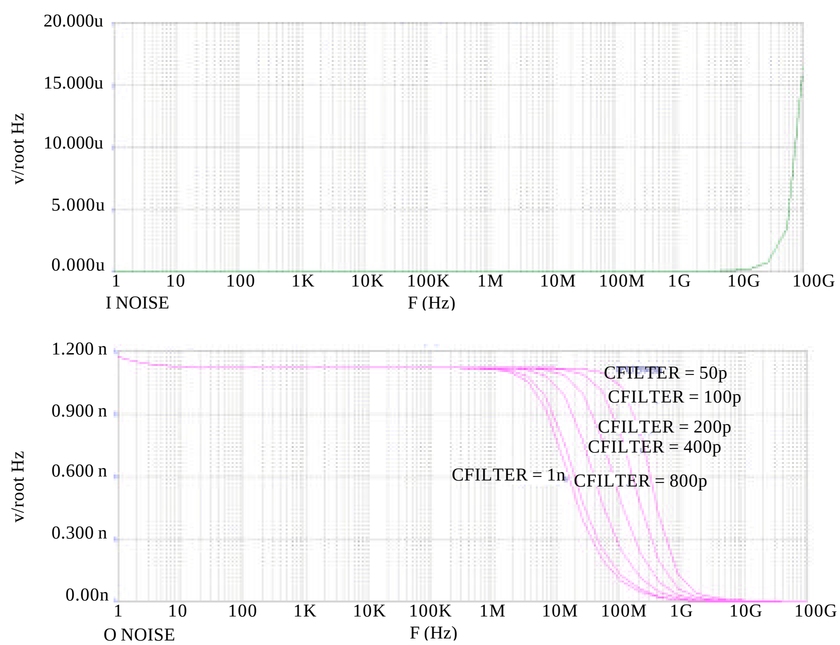

Fig. 9: Input and Output noise density for bootstrap transimpedance amplifier with voltage feedback amplifier

Figure 9 shows the simulated input and output noise density for bootstrap transimpedance amplifier and voltage feedback amplifier. The input noise current density shows a flat 380 pV/$\sqrt{\text{Hz}}$ from 1 Hz to 10 GHz and starts to increase. In simulation the output noise density, shows a flatness of 1.1 nV/$\sqrt{\text{Hz}}$ from 1 Hz to 80 MHZ when it starts descending according to the capacitor value which adjust the bandwidth. The simulated results also showed that the output noise density remains constant during the bandwidth adjustment process.

CONCLUSION

This study has discussed the noise performance analysis and simulation for transimpedance amplifier using the composite amplifier techniques. Results showed that the bootstrapped transimpedance with voltage feedback amplifier exhibits an identical output noise density.

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