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Indoor Optical Wireless Receiver – Theory And Design

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Abstract: This paper presents a systematic approach to the design of receiver for indoor optical wireless communication. In particular, it is concerned with how one can properly chooses the front-end preamplifier and biasing circuitry for the photodetector. A comparison between differences types of amplifier, and bandwidth enhancement technique is also discussed. For most photodetector applications, large values of R_L and C_D would severely restrict bandwidth. It is shown that a proper front-end design incorporates a transimpedance preamplifier which tends to integrate the detector output. Such a design provides significant reduction in photodiode capacitance and increase bandwidth when compared to a design which does not integrate initially. Two novel techniques, using bandwidth adjustment for better service quality with a bootstrapped transimpedance amplifier and bootstrapped composite transimpedance amplifier is presented. A controllable capacitance is introduced at the output of the second stage of the amplifier. These technique permits a bandwidth adjustment from 52Hz to 233MHz for a capacitance range of 10uF to 1pF, while the composite amplifier bandwidth adjustment of the circuit can be controlled in the frequency range of 6MHz to 60MHz for a capacitance range of 50pF to 1nF.

Keywords: photodetector, boostrap transimpedance preamplifier, composite transimpedance amplifier

1. Introduction

The purpose of this paper is to provide insight into the research effort to date in optical wireless receiver both in academic and industrial contexts. The idea of using the optical medium for wireless communications is not new, having been proposed as a means for indoor communications almost two decades ago.[4][6][7] However, recent years have seen an increasing interest in the potential for free space optical systems to provide portable data communications. One of the main aspects for reconsidering the use of an optical carrier in the wireless context is the demand for greater transmission bandwidths. The radio frequency spectrum is already exceedingly congested and frequency allocations of sufficient bandwidths are extremely hard to obtain. Proponents of optical wireless systems argue that the optical medium is the only cost-effective way to provide high bit-rate mobile services to volume markets. In most proposed infrared optical wireless communication [4-10], it is well known that the signal level in an optical wireless receiver is weakest at the front end. Weakest front end means that it is too noisy, too slow or both. The two parameters are not unrelated; it is easy to have a fast front end by preparing to sacrifice signal to noise ratio. This is where the system signal-to-noise ratio is determined and system performance level established.

2. Photodetector Input-Output Relationships

Semiconductor junctions that convert photon energy of light into an electrical signal by releasing and accelerating current-conducting carriers, ultimately to produce a baseband voltage for regeneration is called a photodiode.[1] In order to appreciate its performance in practical optical wireless systems, we have to characterize the photodiode from three points of view : the physical viewpoint, the circuit viewpoint and the statistical viewpoint.

The physical viewpoint

The presence of incident optical power, entering a semiconductor device produces thermal agitation that release holes-electron pairs generated at various points within the diode as illustrated in Fig 1.[2] These carriers drift toward opposite ends of the device under the influence of the applied field. When a carrier passes through the high-field region, it may gain sufficient energy to generate one or more new pairs of holes and electrons through collision ionization. These new pairs will in turn generate additional pairs by the same mechanism. Carriers accumulate at opposite ends of the diode, thereby reducing the potential across the device until they are removed by the biasing and other circuitry in parallel with the diode as shown in Fig 2. The chances that a carrier will generate a new pair when passing through the high-field region depends upon the type of

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carrier, the material out of which the diode is constructed and the voltage across the device. The depth and extent of the junction determines the location of the depletion region and the light wavelengths that produce an efficient response. For a given photodiode and a given wavelength, a photodiode responsivity expresses the resulting efficiency through[2]

$$I_p = r_Ø Ø_e$$

where $r_Ø$ – diode's flux responsivity and $Ø_e$ – radiant flux energy in watts

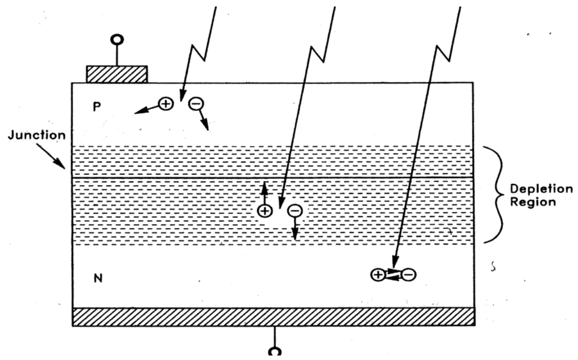


Fig. 1 Photodetector

The circuit viewpoint

From the discussion above and more detailed investigation, modeling the characteristics of the photodiode with discreet circuit components permits analysis of application circuits. Fig 2 shows the resulting model with an ideal diode, a current source and parasitic elements. C_d is the junction capacitance of the diode across which voltage accumulates when charges produced within the device separate under the influence of the bias field. The current generator $i_p(t)$ represents the production of charges by optical and thermal generation and collision ionization in the diode high-field region. Resistance R_D represents the diode's dark resistance, which is the resistance of the zero-biased diode junction. In order to use the photodiode efficiently we must design a circuit which will respond to the current $i_p(t)$ with as little distortion and added noise as possible. In order to derive information from the circuit responding to $i_p(t)$ we must understand the statistical relationship between $i_p(t)$ (the equivalent current generator) and the incident optical power p(t).[1-2]

The statistical viewpoint

In Fig 2, the current source $i_p(t)$ can be considered to be a sequence of impulses corresponding to electrons generated within the photodiode due to optical or thermal excitation or collision ionization. From various physical studies it has been concluded that for cases of current interest, the electron production process can be modeled as shown in Fig 3. Let the optical power falling upon the photon counter be p(t). In response to this power and due to thermal effects, the photon counter of Fig 3 produces electrons at average rate λ(t) per second where

$$\lambda(t) = [(\eta/h\Omega)p(t)] + \lambda_o \qquad (1)$$

where η – photon counter quantum efficiency

hΩ – energy pf a photon

λ_o – dark current "counts" per second

λ(t) is only the average rate at which electrons are produced. p(t) the number of electronic produced in any interval is statistically independent of the number produced in any other disjoint interval.[1]

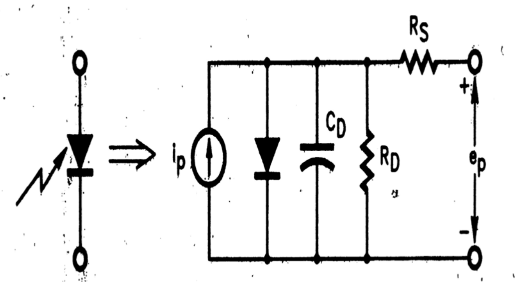


Fig.2 Photodiode circuit model

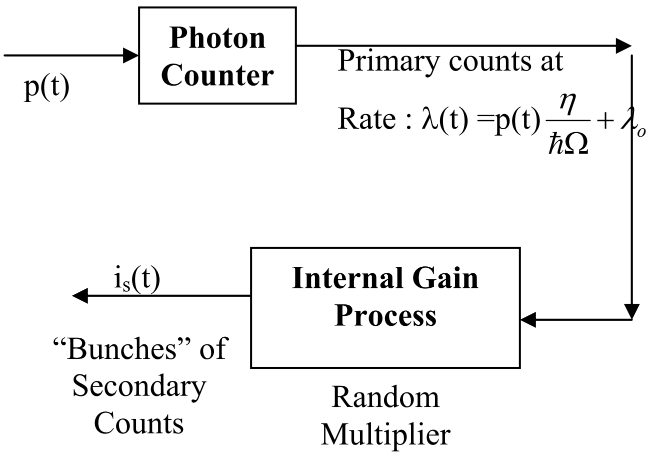


Fig.3 Model of $i_s(t)$ generation process

Each of the primary impulses produced by the photon counter enters a random multiplier where, corresponding to collision ionization it is replaced by g contiguous "secondary" impulses. The number g is governed by the statistics of the internal gain mechanism of the photodiode. Each primary impulse is "multiplied" in this manner by a value g which is statistically independent of the value g assigned to other primaries. Thus the current leaving the photodiode consists of bunches of electrons. For applications interest here, it will be assumed that all electrons in a bunch exit the photodiode at the time when the primary is produced. This shows that the duration of the photodiode response to a single primary hole-electron pair is very short compared to the response times of circuitry to be used with the photodiode.

Photodiode Variations

Two variations of the basic photodiode improve the diode's response. Physical study shows that PIN photodiodes increase the spectral bandwidth or range of light frequencies that produce an efficient photo response. Avalanche photodiodes increase the magnitude of the output current and the response speed by permitting diode bias at the verge of breakdown. For application interest, PIN photodiode is preferably in optical wireless communication system.

Noise Contributions : As a photodiode amplifier, the current to voltage converter exhibits a complex noise behavior. The major sources of noise in front ends are listed in Table 1, where e_N and i_N are rms values of random fluctuations.[6],[9]

Table 1 Noise source

Source	Type	Formula	Dominates When :
Photocurrent	Shot noise	$i_{NSHOT} = (2eI_d)^{1/2}$	Bright light, large load resistor
Load resistor	Johnson noise	$I_{Nth} = (4kT/R)^{1/2}$	Dim light, small R
Amplifier	Input current noise	$i_N = \sqrt{\frac{4kT}{R}}$	Ideally, never
	Input voltage noise	$e_N = \sqrt{4kTR}$	Dim light, large RC or a fast noisy amp

3. Theoretical Insight of Transimpedance Amplifier

This section of the paper explained an example of how to design a front-end amplifier for a visible or near IR photodiode and how to get improvement in bandwidth without a big sacrifice of SNR.

Given a detector (Fig 2) whose output is a current, the easiest way to form a voltage from it is to have a load resistor (R_L). The output full swing appears across the detector capacitance C_d, rolls off starting at $f_{RC} = \frac{1}{2\pi R_L C_D}$. The signal voltage $V_o(f) = \frac{i_p(f) R_L}{1 + j2\pi R_L C_D f}$. Reducing R_L will reduce the RC product and speed up the system, while increasing R_L increase the bandwidth and dynamic range of the system. Another key idea is to reduce the swing across C_D, by making the detector work into a virtual ground using a transimpedance amplifier shown in Fig 4.[6],[9] The inverting input of A_1 draws no current, feedback forces the voltage there to be close to zero at all times. A_1 senses the voltage across C_D and wiggles other end of R_F to zero it out. Provided that A_1 has high loop gain A_{VOL}, the swing across C_D is greatly reduced and the bandwidth greatly improved. The amplifier input adds a significant amount of its own capacitance C_{in}. For a typical transimpedance topology using active devices with load feedback resistance R_f , the transimpedance gain A_z can be approximated by :

$$A_z \approx \frac{-R_f}{1 + j\omega \frac{R_f (C_D + C_{in})}{A_{VO}}} \qquad (2)$$

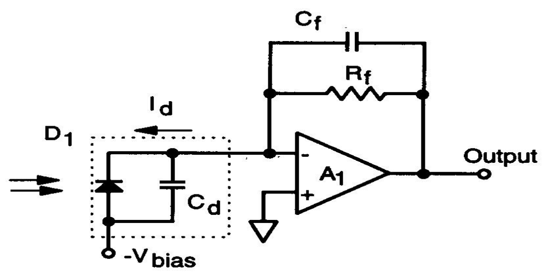
where A_{vo} is the open loop voltage gain of the amplifier and ω is the angular frequency.

The transimpedance rolls off depends on the magnitude of the impedance of the feedback elements. Therefore the transimpedance amplifier bandwidth is calculated with the following equation :[6],[9]

$$f_{3dB} \approx \sqrt{\frac{f_{RC} f_T}{2}} \qquad (3)$$

where f_T is the unity-gain crossover frequency

Equation 2 shows that the upper 3dB cut-off frequency of the preamplifier is a function of the capacitance from the detector, feedback resistor and open loop voltage gain. In other words, a large detector means a large C_d. Hence, in order to achieve large bandwidths either the value of R_f is reduced or A_{vo} is increased. Unfortunately, increasing A_{vo} will jeopardize amplifier stability and reducing R_f will increase thermal noise into the system.



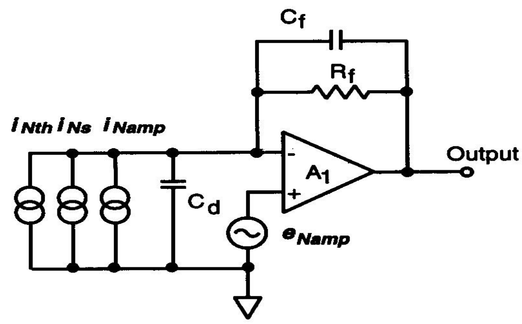


Fig. 4 Transimpedance amplifier circuit and noise model

Noise in transimpedance amplifier : From Fig 4, it is obvious that all the current sources are in parallel : i_d, i_{NSHOT}, i_{Nth} and i_{amp}. The Johnson noise i_{Nth} appears across R_f but since the op amp output impedance is low and the currents add linearly, the other end of i_{Nth} is at ground for noise purpose. The rolloff in the frequency response does not degrade the signal to current noise ratio. The amplifier's voltage noise, e_{Namp} is treated differently because A_1 is a differential amplifier. Therefore e_{Namp} is multiplied by A_1's noninverting gain.[6],[9]

$$A_{Vcl} = \frac{A_{Vol}}{1 + \frac{A_{Vol}}{1 + jwC_D R_f}} \quad (4)$$

If e_{Namp} is very low or if we are not trying to get a huge bandwidth improvement through $(f_T * f_{RC})^{1/2}$ mechanism, this rising noise contributions will not limit. Otherwise, it will dominate the noise starting at :

$$f_{3dB} = \frac{1}{2\pi e_{Namp} C_D} \sqrt{2eI_D + i_{Namp}^2 + \frac{4kT}{R_L}} \quad (5)$$

In order that op amp do not dominate the noise according to Hobbs[9], we should choose it by the following rules :

$i_{Namp} < 0.5 i_{Nth}$ (Noise of R_f dominates i_{Namp})
$e_{Namp} < 0.5\ R_f\ i_{Nth}$ (e_{Namp} should be the same in flatband)
$e_{Namp} < 0.5 i_{Nth} / (2\pi f_{3dB} (C_D + C_{in})$ (Noise peak should not dominate anywhere in the frequency band)
$f_T > 2 f_{3dB}^2 / f_{RC}$ (The amplifier has to raise enough bandwidth)
$f_T < 10 f_{3dB}^2 / f_{RC}$ (Speed too fast risks trouble with ringing and oscillation)

The transimpedance amplifier does not improve the SNR of the photodiode it just changes the frequency response. There's nothing inherent or inescapable about noise peak in system, it comes from a poor choice of circuit topology that can be amended.

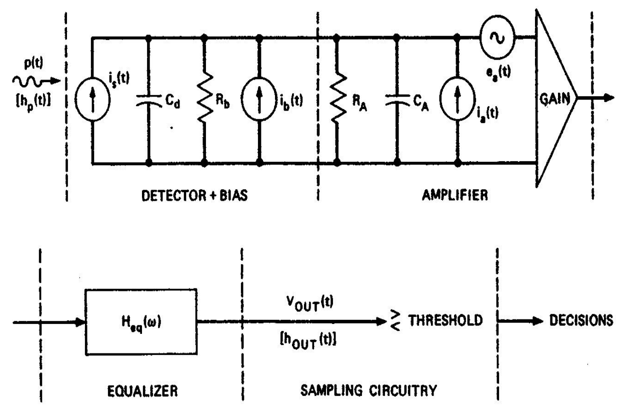


Fig. 5 High impedance amplifier with equalizer

High impedance amplifier theory : Fig 5 shows a typical receiver in schematic form consisting of a photodiode, an amplifier and an equalizer. This amplifier is modeled as an ideal high-gain impedance amplifier with an equivalent shunt capacitance and resistance at the input, in addition with two noise sources referred to the input. The noise sources for this particular part will be assumed to be white, Gaussian and uncorrelated. It is also assumed that amplifier gain is sufficiently high, that the noises introduced by the equalizer are negligible.
The average detector output current, $i_S(t)$ is given by :[1]

$$i_S(t) = \frac{\eta e p(t)}{\hbar \Omega} + e\lambda_o \quad (6)$$

where e = electron charge, λ_o = dark current electrons/second, $\frac{\eta p(t)}{\hbar \Omega}$ = average optical primary electrons/second.

Therefore the average voltage (neglecting dc components) at equalizer output, where " * " indicates convolution and A is an arbitrary constant :

$$v_{out}(t) = \frac{A \eta e p(t)}{\hbar \Omega} * h_{fe}(t) * h_{eq}(t) \quad (7)$$

$h_{fe}(t) = F\{ \frac{1}{\frac{1}{R_T} + j\omega(C_d + C_A)} \}$ is the amplifier input circuit current impulse response $R_T = [\frac{1}{R_b} + \frac{1}{R_A}]^{-1}$ is the total detector parallel load resistance $h_{eq}(t)$ is the equalizer impulse response

High impedance amplifier tends to give a degraded frequency response as the bandwidth relationship $\frac{1}{2\pi R_b (C_d + C_A)} \geq B$ is not maintained for wideband operation[1]. The detector output is effectively integrated over a large time constant and must be restored by differentiation. This is performed with the correct equalization. Therefore the high impedance front end gives a better improvement in sensitivity over the low impedance front end design, but eventually creates a heavy demand for equalization and has problem of limited dynamic range. The limited dynamic range is because of the attenuation from the low frequency signal components by the equalization process which causes the amplifier to saturate at high level signals. If the amplifier is saturated before equalization has occurred the signal will be heavily distorted, thus reducing the dynamic range which is dependent upon the amount of integration and subsequent equalization employed.

4. Bootstrap Transimpedance Amplifier

The majority of optical wireless receivers proposed [3][5][7][8][10][11] or demonstrated to date have employed transimpedance amplifier, due to the fact this configuration largely overcomes the drawbacks of high impedance front end by utilizing a low noise, high input impedance amplifier with positive or negative feedback. Bandwidth enhancement techniques are required that do not affect low frequency behavior, so that both high dc gain and large bandwidth can be obtained. Several of these techniques have been proposed in the literature in order to maximize the gain-bandwidth product of an amplifier stage.

One of the method is a transimpedance amplifier is built up with two capacitive coupled voltage dividers (R_1-R_2, R_3-R_4) instead of a single feedback resistor[10]. The basic concept of this network is its different behavior at low and high frequencies. At low frequencies the transimpedance of the transimpedance amplifier is approximately the sum of R_1 and R_2. For higher frequencies the parasitic capacitance of resistor R_2 has no effect owing to the low resistance of R_4, which has no effect on the bandwidth of the transimpedance amplifier. Only parasitic capacitance of resistor R_1 has an influence on the bandwidth of the circuit at a nine times higher frequency than before.
Cascading and capacitance neutralization compensate the input Miller capacitance thus expanding the bandwidth if the input pole is dominant. Unfortunately these techniques are less effective in low power amplifiers that use high speed bipolar transistors, if the output pole becomes dominant.

Another well known method to increase bandwidth of amplifiers uses peaking capacitors[8] or inductors. This method usually places inductors or capacitors in a strategic location of the amplifier circuit, resulting in a resonance with parasitic capacitances, which broadens the bandwidth of the amplifier. Although inductive peaking do increase the amplifier bandwidth, stray capacitances of the inductor often causes bandwidth degradation rather than an improvement. Capacitive peaking design is preferable, but this technique is extremely sensitive to process variations and could cause large peaking.

Therefore, a much improved version of a front-end, incorporated within a transimpedance amplifier, is shown in Fig 6, the topology being known as the bootstrapped transimpedance amplifier (BTA)[11]. The BTA is an attractive design as it reduces the effective detector capacitance, C_d, seen by the signal. The output of the emitter follower stage is feedback to the photodetector by a bootstrapping capacitor, C_3. Fig 7 proposes a combination of bootstrapped technique with a capacitive peaking technique by placing a capacitor, C_4, in series with the emitter resistor in the second gain stage, with a feedback resistor R_8 [12]. By varying the capacitor, the bandwidth of the circuit can be controlled. The output results of Fig 6 and Fig 7 are shown in Fig 8 and Fig 9. In each case the amplifier output is taken from the collector at Q_3. Assuming that the gain stages and the emitter follower can be approximated by a simplified hybrid-π model, as shown in Figure 10, $R_1, R_2 >> R_7$ and we consider frequencies where C_1, C_2 and C_3 are short circuits, the transimpedance gain, A_z for the circuit is approximated by equation (8):

A_z =

$$\frac{V_o}{I_p} = \frac{g_{m3}R_6[2r_3 + R_7(1-A)]}{(A+1)[1+r_3 g_{m3}] + j\omega C_4[r_3(1+g_{m3}R_7)+R_7]} \qquad (8)$$

where A is the voltage gain of the first stage amplifier

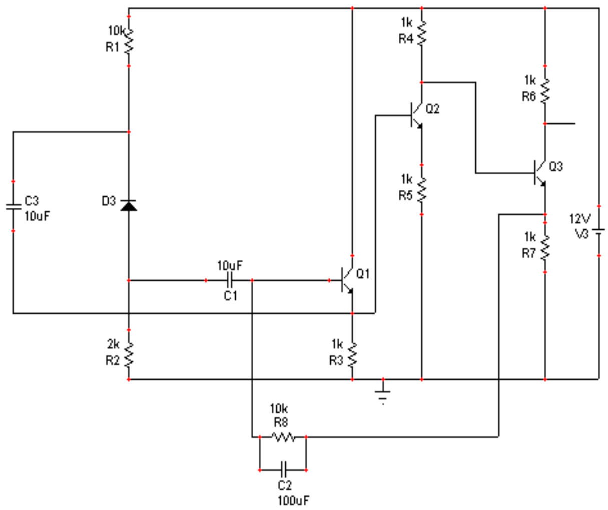


Fig. 6 Bootstrapped transimpedance amplifier circuit

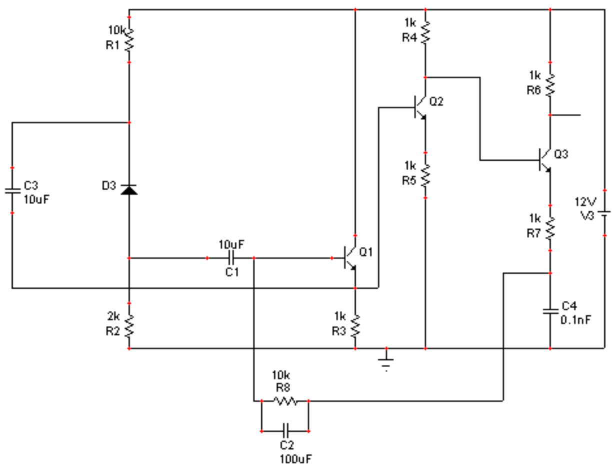


Fig. 7 Bootstrapped with capacitor peaking transimpedance amplifier circuit

Equation (8) shows that the receiver bandwidth is determined not only by the R_8C_d time constant but by a complex function of g_{m3}, r_3, R_7 , R_6 and C_4. The circuit effectively operates in between the low and the high frequency range. Thus, the modified circuit shows that varying capacitor C4, thus modifying the second stage gain can vary the bandwidth. The lower the value of C_4, the higher the bandwidth becomes. This technique permits a bandwidth adjustment from 350MHz to 2.5MHz for a capacitance range of 100µF to 100pF.

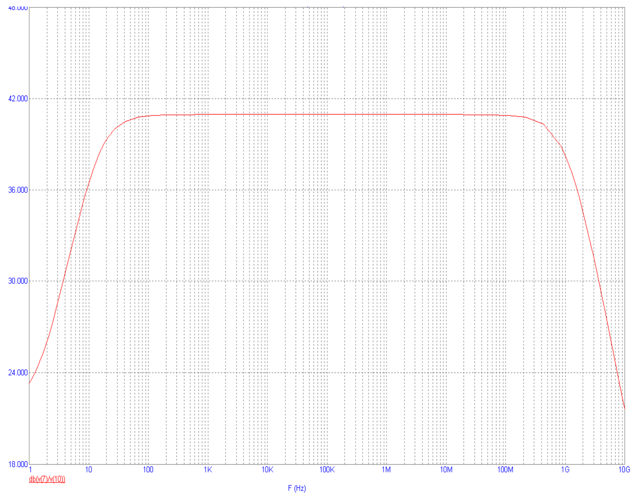


Fig. 8 Frequency response bootstrapped transimpedance amplifier

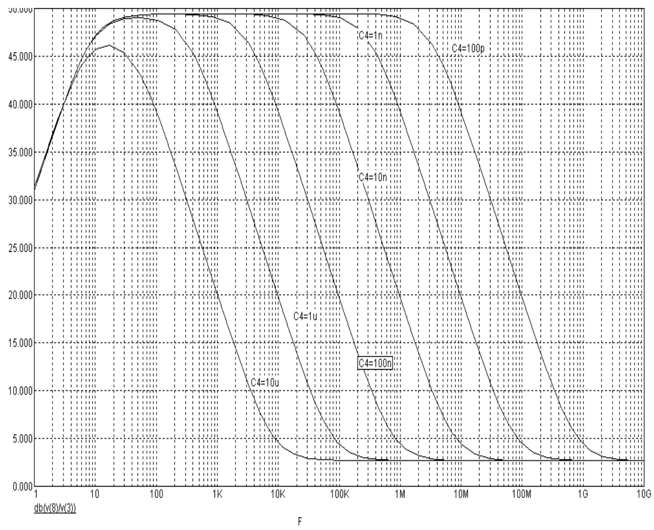


Fig. 9 Frequency response bootstrapped with capacitor peaking transimpedance amplifier

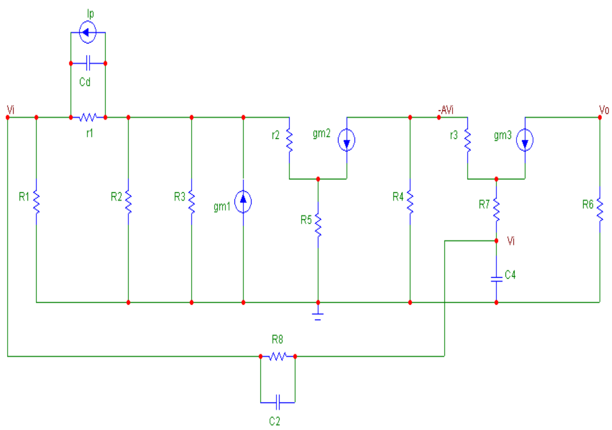


Fig.10 Model hybrid-pi of bootstrapped with capacitor peaking transimpedance amplifier circuit

A better proposed topology using a composite amplifier provides high bandwidth as shown in Fig 11. The bootstrapped transimpedance amplifier is connected in series with a voltage feedback amplifier and a RC filter. By varying the capacitor, C_6 between 50pF to 1nF the bandwidth of the circuit can be controlled in the frequency range of 6MHz to 60MHz as shown in Fig 12. There is a trade-off between gain and bandwidth compared to Fig 9. As the bandwidth is increased the gain of the circuit is reduced. If we assume that the gain stages, the emitter follower and second stage amplifier can be approximated by a simplified hybrid-π model, as shown in Figure 13. The transimpedance gain, A_{z1} for the circuit can be approximated by equation (9) considering frequencies where C_1, C_2 and C_3 are short circuits :

$$A_{z1} = \frac{V_{o2}}{I_p} = \frac{R_{19} R_{i1} R_6 A_1 [R_7 + A R_7 R_6 - r_3]}{r_3 (R_{19} R_{i1} + R_{19} + A_1 R_{i1})(1 + j\omega C_6 R_{20})} \quad (9)$$

where A is the voltage gain of the first stage amplifier and A_1 is the voltage gain of the second stage amplifier.

Bootstrapping transimpedance amplifier effectively allows for a higher transimpedance gain and a lower R_f thermal noise contribution. Composite configuration of transimpedance amplifier effectively allows for a higher bandwidth with a trade-off of 10dB – 20dB gain, while maintain its noise contribution.

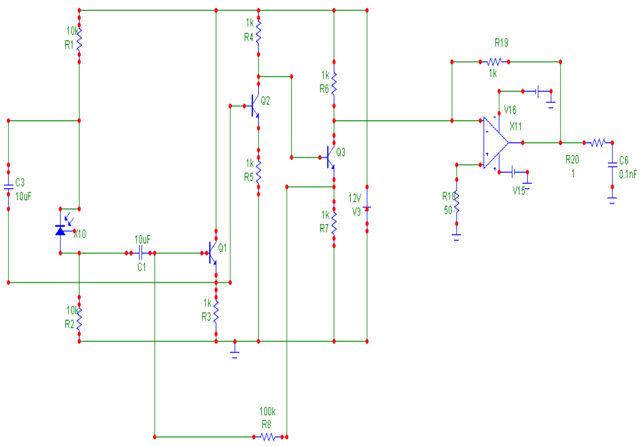


Fig. 11 Composite transimpedance amplifier

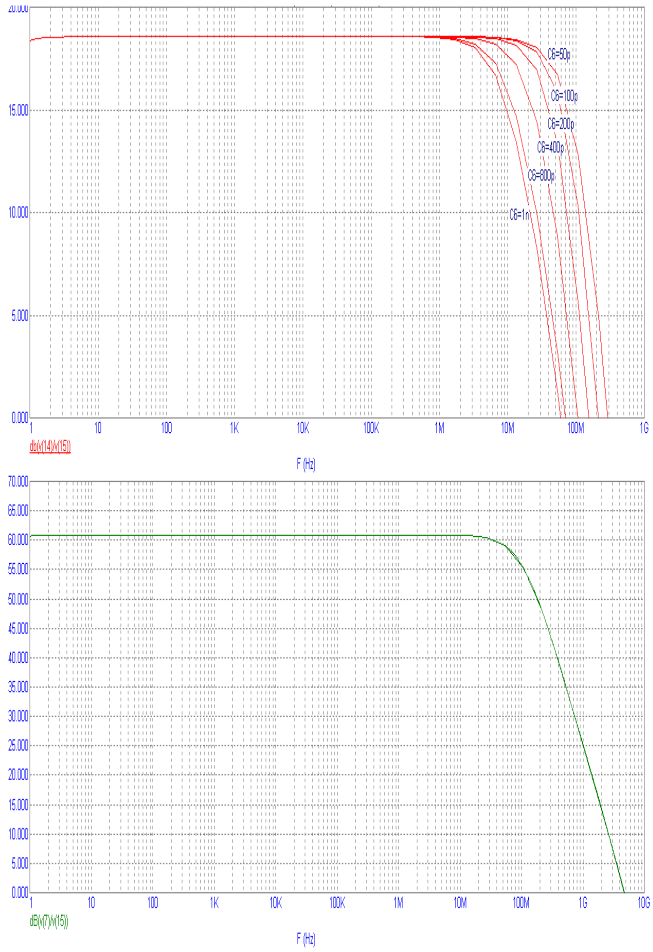


Fig. 12 Frequency response composite transimpedance amplifier

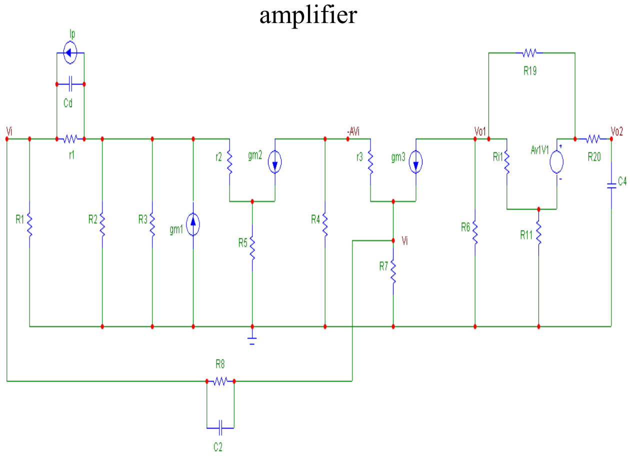


Fig. 13 Model hybrid-pi of composite transimpedance amplifier

5. Transimpedance Amplifier Implementation

In application wise, base on the above simulation design it is suggested that the composite amplifier receiver to be use to accommodate the high frequency range bandwidth adjustment, where a variable switching capacitor circuit could be use to vary the capacitance from the range of 50pF to 1nF. The propose switching circuit which is design to be incorporated together with the receiver will then automatically select the right capacitance value based on the input received signal quality and adjust the receiver to the required bandwidth.

6. Summary

Receiver design is particularly challenging because not only dynamic range and bandwidth criteria has to be met, but also significant problems of high ambient light levels noise has to be dealt with. This paper has provided an insight issues associated with the front-end design of a wireless infrared communication. It has highlighted the significant maxims on choosing the biasing circuitry. To summarize, high impedance amplifier with a large R_f diminish the effects of thermal noise. However the receiver bandwidth is then usually smaller then the signal bandwidth, which require an equalization stage following the preamplifier as discussed. The circuit will be tricky, as the equalizer effect reduces the overall dynamic range of the receiver. The transimpedance front ends resolve these issues by using a large feedback resistor and an inverting amplifier, which boots the bandwidth without thermal noise and dynamic range problems. In addition, infrared application dominant source of noise is due to background radiation, not thermal and circuit noise, which makes the sensitivity of the transimpedance front end more attractive. Therefore, in conclusion, infrared optical wireless communication had yet to fully exploit all the potential benefits offered by the medium. There is still a great deal of work to be done in the adaptation and optimization of coverage areas, especially with unknown receiver orientation

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