Abstract

This Thesis is concerned with optoelectronic mixing in heterojunction bipolar transistors (HBTs). Optoelectronic mixing is classified into the electrically pumped and optically pumped categories. In electrically pumped mixing, the local oscillator (LO) source is applied electrically to an optoelectronic mixer which photodetects and converts a radio frequency (RF) modulated optical signal to an electrical signal at an intermediate frequency (IF). In optically pumped mixing, the LO is applied optically to an optoelectronic mixer which converts an electrical input signal at RF to another electrical output signal at IF. Optoelectronic mixing is required in many microwave over fibre and fibre-radio systems, and the conventional approach is to employ a separate photodiode for optical-to-electrical conversion and a microwave mixer for frequency conversion. In this work, optoelectronic mixing has been carried out using a two-terminal edge-coupled InP/InGaAs HBT for the first time and a three-terminal normal-incidence InP/InGaAs HBT. This single HBT approach to optoelectronic mixing offers a simpler alternative to the conventional method.

The two HBT optoelectronic mixers have been experimentally characterised in terms of the mixing conversion gain, frequency response for the three-terminal HBT, signal-to-noise ratio and spurious-free dynamic range, obtained from two-tone third-order intermodulation distortion measurement using a two-laser approach. The distortion characteristics for HBT optoelectronic mixers are reported for the first time.

Harmonic-balance techniques have been applied for the first time to model the conversion gain characteristics of the two electrically pumped HBT optoelectronic mixers. A large-signal equivalent circuit in the T-topology has been used for the three-terminal HBT. However a quasi-static model has been used for the two-terminal HBT because of the absence of an electrical base terminal which would otherwise allow the internal transistor parameters to be determined through DC and s-parameter measurements and the construction of a more physical HBT model.

A number of fibre-radio system architectures employing HBT optoelectronic mixers are discussed.
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Chin Pang LIU
To my mother FAN Kam-Fung 樊金鳳

and sister LIU Tak-Ching 廖德晶
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Chapter One

Introduction

Section 1-1: Introduction to Optoelectronic Mixing

The microwave mixer, built using semiconductor devices exhibiting nonlinear characteristics, is one of the most important elements in radio communication systems and is described in most microwave circuit design textbooks. In general the function of a microwave mixer is to convert the frequency of an input signal to a different frequency at the mixer output. Such a frequency conversion of the signal is often necessary because, for example, modulation and demodulation can be more easily carried out at low frequencies while the actual radio frequency used may be different due to the required spectrum being available in a different frequency band. Figure 1-1 shows a symbol for a microwave mixer.

![Mixer symbol](image)

Figure 1-1: A symbol representing a microwave mixer. RF: radio frequency, IF: intermediate frequency, LO: local oscillator.

A mixer is a three-port device consisting of an RF, an IF and an LO port. The RF port is usually used for signal input while the IF is for signal output and both the RF and IF signals have much lower power levels than the LO input. The LO port is connected to a constant (frequency and amplitude) large-signal (~10 dBm) microwave source (the local oscillator) which is required to pump the mixer. Under this large LO condition and assuming that the RF is much smaller in amplitude, the mixer then mixes the LO with the RF and the output IF will consist of components at frequencies given by

\[ f_{IF} = |\pm f_{RF} \pm k f_{LO}| \]

Eqn 1-1
where $f_{IF}$, $f_{RF}$ and $f_{LO}$ are the frequencies of the IF, the RF and the LO, respectively, and $k$ is any integer. For fundamental mixing, which is used throughout this work, $k$ is equal to 1. $k \geq 2$ denotes harmonic mixing. For a given $f_{IF}$, the conversion gain of a microwave mixer is defined as the ratio of the output IF power at $f_{IF}$ delivered to the load to the available input RF power at $f_{RF}$, i.e.

$$G_{con.} = \frac{P_{IF}}{P_{RF}}$$  \hspace{1cm} \text{Eqn 1-2}

When both the input and output ports of the mixer are conjugately matched to the signal source and output load, respectively, then the available input RF power will be equal to the RF power delivered to the mixer input, and the mixer output available IF power will be equal to the power delivered to the load at $f_{IF}$.

Commercially available microwave mixers are usually specified in terms of the conversion loss, which is the reciprocal of the conversion gain defined above, because those mixers are built using diodes and therefore their output power is always smaller than the input power. When active transistors are employed for mixer constructions, a conversion gain becomes possible.

Unlike the microwave mixer which has three electrical ports, an optoelectronic mixer, in the context of this Thesis, is a mixer which has one electrical output, one electrical input and one optical input. According to this description, there are therefore two categories of optoelectronic mixers, depending on whether the LO is input to the mixer optically or electrically. When the LO is given by the intensity modulation of an optical source and the RF input is electrical, the optoelectronic mixer is described as optically pumped. If on the other hand the LO is supplied by a normal electrical microwave source while the RF is delivered optically on an optical carrier, the optoelectronic mixer is described as electrically pumped. In both cases, the output of the optoelectronic mixer is still electrical. Figure 1-2 shows the symbols for these two kinds of optoelectronic mixers.

Before proceeding further, it is instructive to clarify the terminology of the terms IF and RF. Throughout the computer modelling work (Chapter 3) and the experimental work (Chapters 4 and 5) in this Thesis, the term RF is used to refer to the input port of the optoelectronic mixer while the term IF refers to the corresponding output port, regardless
of whether the mixer is down-converting or up-converting. In the literature, however, it is common to find that IF is used to mean the mixer input port while the RF the output port when the mixer is up-converting where the frequency of the output mixed product (RF) is higher than that of the input signal (IF). In order to avoid confusion, whenever the terms RF and IF are used in this Thesis, it will be made clear which port each of the terms is referring to.

Therefore the function of an electrically pumped mixer is to detect and convert the intensity modulation of the incident light modulated at \( f_{RF} \) into an electrical output at \( f_{IF} \). The function of an optically pumped mixer is similar to that of a microwave mixer, except that the LO is supplied optically by the intensity modulation of the incident light.

A microwave mixer can be configured as an optoelectronic mixer by adding a separate photodetector to convert the optical signal into an electrical signal and this represents the conventional approach to optoelectronic mixing as shown in Figure 1-3.
In practice, matching may be required in the electrically pumped configuration after the PD output in order to increase the RF power delivered to the electrical input of the mixer, thus increasing the conversion gain. Also matching and amplification may be used in the optically pumped configuration after the PD so that sufficient electrical LO power is available to drive the mixer.

In this work, optoelectronic mixing is accomplished by using single heterojunction bipolar transistors (HBTs) which can be optically controlled. An HBT, because of its two internal pn junctions, is inherently a nonlinear device and possesses the nonlinearity required in the mixing processing. Two types of InP/InGaAs HBTs have been investigated for such an application. The first type is a two-terminal, edge-coupled HBT while the second type is a three-terminal, normal-incidence HBT. Full details of the two types of HBTs will be given in Chapters 4 and 5, respectively. The use of single HBT optoelectronic mixers offers a number of advantages. Since a single HBT is now used instead of a separate photodiode and a separate microwave mixer, system simplicity and cost savings in equipment are possible. In the electrically pumped HBT optoelectronic mixer, the photodetected RF signal is amplified and mixed with the LO simultaneously by the HBT, thus providing a gain advantage over the conventional microwave mixer-photodiode approach (Figure 1-3(a)). In the optically pumped HBT optoelectronic mixer, electrical connection of the LO and the consequent matching and parasitics problems are avoided while the electrically applied RF is also amplified by the HBT, giving a similar gain advantage. The topologies of the two types of HBTs configured as electrically pumped optoelectronic mixers are illustrated in Figure 1-4.
The corresponding optically pumped optoelectronic mixer topologies are identical to those in Figure 1-4 except that the roles of the RF and LO are reversed.

As can be seen in Figure 1-4, the three-terminal HBT is a natural candidate for being an electrically pumped optoelectronic mixer because the LO can be applied to the base terminal while the IF output can be extracted from the collector separately when the base is illuminated with the RF modulated light. The two-terminal HBT mixer being without an electrical base connection, however, has only the emitter available for both the LO injection and the IF output. A diplexer is therefore required to channel the LO and IF signals.

The organisation of this Chapter is as follows. In Section 1-2, formal definitions for the optoelectronic mixing conversion gains will be given. In Section 1-3 the applications of optoelectronic mixers in microwave subcarrier multiplexed and fibre-ratio systems will be described. In Section 1-4, previous work on optoelectronic mixing involving other semiconductor devices and techniques will be examined. In Sections 1-5 and 1-6, the reason for the use of a heterojunction and the photodetection process in the HBT will be briefly reviewed. From Section 1-7 to Section 1-9, the current status of HBT direct photodetectors, optoelectronic mixers as well as optically controlled oscillators will be summarised. Section 1-10 will list the aims and a number of novel achievements of this Thesis. Finally a conclusion and the detailed structure of the remainder of this Thesis will be given in Section 1-11.

Section 1-2: Definitions for the Optoelectronic Mixing Conversion Gains and Other Figures-of-Merit Relating to Photodetectors

The conversion gain for an optically pumped optoelectronic mixer is defined the same way as that for a microwave mixer and is given by Eqn 1-2. The conversion gain for an electrically pumped optoelectronic mixer, however, requires further consideration since the RF input is optical but the IF output is electrical.
To obtain a meaningful conversion gain definition for an electrically pumped optoelectronic mixer, the RF modulated optical input is first converted to an equivalent electrical input power given by

$$P_{RF\_Elec} = \left( \frac{mq\lambda P_{opt}}{hc} \right)^2 \frac{R_{Load}}{2}$$  \hspace{1cm} \text{Eqn 1-3}$$

where \( m \) is the intensity modulation depth of the incoming optical beam, \( q \) is the electron charge, \( P_{opt} \) is the mean optical power, \( h \) is the Planck's constant, \( c \) is the speed of light, \( \lambda \) is the wavelength and \( R_{Load} \) is the load impedance which is almost always equal to 50 \( \Omega \) in practice. \( P_{RF\_Elec} \) is thus equal to the ac electrical power delivered by a 100 \% quantum efficient photodiode under the same illumination to a directly attached load \( R_{Load} \). Note that \( P_{RF\_Elec} \propto P_{opt}^2 \) since electrical current is directly proportional to optical power.

The system conversion gain for an electrically pumped optoelectronic mixer is then defined as

$$G_{sys} = \frac{P_{IF}}{P_{RF\_Elec}}$$  \hspace{1cm} \text{Eqn 1-4}$$

where \( P_{IF} \) is the mixer electrical IF output power delivered to the load. When applying Eqn 1-4, it is important that the impedance with which \( P_{IF} \) is measured is equal to the impedance \( R_{Load} \) with which \( P_{RF\_Elec} \) is calculated. In practice both impedances are almost always equal to 50 \( \Omega \).

\( G_{sys} \) is a useful definition in the system sense because it includes the external quantum efficiency of an electrically pumped optoelectronic mixer and this factor has to be taken into account when the electrically pumped optoelectronic mixer is employed in practical situations. A similar intrinsic conversion gain \( G_{int} \) can also be defined by excluding the effect of the external quantum efficiency as

$$G_{int} = \frac{1}{\eta_{ext}^2} \frac{P_{IF}}{P_{RF\_Elec}}$$

$$= \frac{1}{\eta_{ext}^2} G_{sys}$$  \hspace{1cm} \text{Eqn 1-5}$$
where $\eta_{\text{ext}}$ is the external quantum efficiency and is given as the number of carriers per unit time in the primary photocurrent divided by the number of incident photons per unit time. $G_{\text{int}}$ is a useful figure of merit when the internal mixing mechanism of the electrically pumped optoelectronic mixer is the subject of the investigation. Both $G_{\text{int}}$ and $G_{\text{sys}}$ are affected by the transistor internal current gain and internal mixing efficiency. However $G_{\text{int}}$ is independent of the transistor external quantum efficiency since this factor is removed in Eqn 1-5.

When the HBT is used as a direct photodetector, a system signal gain can be defined as

$$G_{\text{sig}} = \frac{P_{\text{out}}}{P_{\text{RF-Elec}}}$$  \hspace{1cm} \text{Eqn 1-6}

where $P_{\text{out}}$ is the HBT electrical output signal power delivered to the load.

The responsivity $R$ is defined as $I_{\text{ph}}/P_{\text{opt}}$ where $I_{\text{ph}}$ is the total photocurrent due to the incident optical power $P_{\text{opt}}$.

The optical gain $G_{\text{opt}}$ is given by the number of carriers in the total photocurrent divided by the number of incident photons, and can be expressed as

$$G_{\text{opt}} = \left(\frac{hc}{q\lambda}\right)\left(\frac{I_{\text{ph}}}{P_{\text{opt}}}ight)$$  \hspace{1cm} \text{Eqn 1-7}

\section*{Section 1-3: HBT Optoelectronic Mixer Applications}

\section*{Section 1-3a: Microwave Subcarrier Multiplexed Systems}

In order to deliver a number of channels down a single transmission medium, e.g. an optical fibre or a microwave cable, some form of multiplexing is necessary. In digital transmission systems, time-division multiplexing is commonly used. In analogue transmission systems, however, multiplexing is more conveniently carried out in the frequency domain. Figure 1-5 illustrates how a number of baseband signals are
frequency-multiplexed and transmitted over a single optical fibre in a simplified microwave subcarrier multiplexed (SCM) system [1,2,3,4,5,6].

![Simplified schematic diagram illustrating a microwave subcarrier multiplexed system using fibre-optic technology.](image)

In Figure 1-5, each baseband signal (BB 1, BB 2, etc.) is up-converted to its respective frequency ($f_{sc1}$, $f_{sc2}$, etc.) by the subcarrier modulator. The up-converted signals are in different frequency bands and can therefore be combined by a microwave power combiner forming a microwave subcarrier multiplexed composite signal. The spectra of the signals before and after subcarrier multiplexing are shown in Figure 1-6. The composite signal then directly modulates a laser diode for transmission over a single optical fibre.

![Subcarrier multiplexing of a number of baseband signals (BB 1, BB2, etc.).](image)

Figure 1-6: Subcarrier multiplexing of a number of baseband signals (BB 1, BB2, etc.). (a) individual baseband signals before multiplexing, (b) the single composite signal consisting of up-converted baseband channels in different frequency bands after multiplexing.
optical fibre.

At the receiver end, the light carrying the microwave subcarrier multiplexed composite signal is photodetected by a photodiode. A low noise amplifier (LNA) then amplifies the detected signal level in order to overcome the losses and noise figures of the subsequent mixer chain and other electronics. To select the desired channel for demodulation by the final receiver, the frequency of LO 1 is such that mixer 1 frequency translates the desired photodetected subcarrier channel to fall within the frequency range of the bandpass filter. The bandpass filter filters out the noise outside its bandwidth and prevents other subcarrier channels from breaking through. The filtered subcarrier channel can then be down-converted to recover the baseband signal by mixer 2. The lowpass filter is used to filter out the up-converted output from mixer 2 and other signals outside the baseband which might otherwise saturate the receiver. Figure 1-5 only shows a simplified system. In practice, an IF amplifier is usually used after the bandpass filter to strengthen the IF signal level.

One important advantage of the SCM system is that most existing microwave components and techniques can be reused and this point can be better illustrated by comparing Figure 1-5 with Figure 1-7 which shows a simplified frequency division multiplexed system using a conventional microwave cable as the transmission line. It can be readily seen that many components are used in both systems, making the transfer from conventional cables to fibre optic technology less costly. SCM networks find their main applications in the CATV industry [7].

![Diagram](Figure 1-7: A simplified frequency division multiplexed system using conventional microwave cables.)
In the simple SCM system diagram shown in Figure 1-5, the optical-electrical conversion at the receiver side was performed by a photodiode. The photodiode was then followed by an LNA in order to increase the photodetected signal strength before the desired subcarrier channel is frequency converted to the IF by Mixer 1. By operating an HBT as an electrically pumped optoelectronic mixer as illustrated in Figure 1-4, the photodetection, amplification and frequency conversion can be performed simultaneously by a single HBT, offering system simplicity and other advantages as discussed in Section 1-1. Figure 1-8 illustrates how an electrically pumped HBT optoelectronic mixer can be substituted for the photodiode, LNA and Mixer 1 required in a conventional SCM system in Figure 1-5.

The output IF in Figure 1-8 can be above or below the laser subcarrier modulation frequency (RF). The choice of IF depends on a number of factors. An output IF below the input RF (i.e. down-conversion) can usually allow the optoelectronic mixer to have a better conversion efficiency. Cheaper IF amplifiers, if required, are also available in the low frequency range. However the use of a very low IF increases the unwanted reception of the image signal. Also it has been observed that HBT optoelectronic mixers can have a much higher noise output at low frequencies because of the higher gain in this frequency region. On the other hand, when the output IF is above the input RF (i.e. up-conversion), the desired signal and the image signal are separated by a wider frequency range and therefore image breakthrough can be easily filtered out. If the conversion efficiency is comparable to that in the down-conversion, a better signal-to-noise ratio for up-conversion should be possible because the HBT has a lower output noise at high
frequencies. However, the choice of a higher IF also implies that the IF amplifier, if required, might be more expensive and the design of the bandpass filter might be more difficult.

Section 1-3b: Microwave/Millimetre-Wave Fibre-Radio Systems

Fibre-radio systems are characterised by having both a fibre-optic link and a free-space radio path. The use of a free-space radio path as the final drop to the end-users provides the flexibility that the end-users do not have to be fixed in location. Such systems are important in a number of applications, including mobile communications, wireless local area networks (LANs), wireless local loops [8], etc. Figure 1-9 shows an example of a simple mobile telephone system in the fibre-radio architecture, consisting of a central station, one antenna site or base station and some mobile users.

![Figure 1-9: An illustration of a fibre-radio system for mobile communications.](image)

A low-loss, high-capacity fibre-optic link connects the central station and one of its base stations. Since the central station and the base stations are separated by some distance, optical fibres are preferred to conventional coaxial cables in such an application because of the loss consideration. For example a 1 km long, good-quality single-mode fibre has an optical loss of only 0.2 dB at 1550 nm [9], which is equivalent to an electrical loss of 0.4 dB, while a common RG-405 semi-rigid microwave cable of the same length would have a loss of 2430 dB at 10 GHz [10]!

The base station performs optical-electrical conversion of the signal in the down-link direction (central station to base station) and electrical-optical conversion in the up-link
direction (base station to central station). The use of the free-space radio path enables the end-users to be mobile.

In wireless LANs and wireless local loops, the final free-space radio path allows fast and economical deployments of the user receiving equipment. In antenna remoting, the antenna site, connected by a fibre-optic link, can be situated far away from the control centre to improve satellite visibility and reduce interference with terrestrial systems.

The demand for high data rate services is increasing and in order to deliver such services over a radio link, a wider radio frequency spectrum is required. As a result the radio link should use higher frequency carriers because of the spectrum congestion at low frequencies (<20 GHz). In the European Union's RACE II Mobile Broad System (MBS) project, for example, radio carriers in the 40 GHz and 60 GHz bands are chosen to deliver broadband (up to 155 Mbits/s) services to mobile users. The use of the 60 GHz band also attracts a significant atmospheric attenuation (around 17 dB/km [11]) because of the oxygen absorption peak and therefore allows better frequency reuse in mobile cellular systems. Delivery of microwave/millimetre-wave signals using optical fibres between the central station and the base stations, and free-space radio paths between the base stations and the end-users may be termed microwave/millimetre-wave fibre-radio system [12,13,14].

At present, the highest modulation bandwidth for a semiconductor laser reported to date is around 30 GHz [15] and therefore direct modulation of the laser beyond this limit is not yet practical. The bandwidths of most commercially available, packaged lasers are even lower. Although external modulators can be used and fast external modulators with a 3-dB electrical bandwidth of 75 GHz have been demonstrated in research laboratories [16], they usually require excessive RF drive power (+18 to +24 dBm) to operate. Fibre dispersion further limits the link performance at such high frequencies. If only the low-frequency IF input signal needs to be transmitted over the fibre and at the end of the fibre-optic link, this IF signal is photodetected and up-converted to millimetre-wave frequencies before being radiated into free-space and delivered to the end-users, then the losses due to the optoelectronic components and fibre dispersion can be greatly reduced. An HBT electrically pumped up-converting optoelectronic mixer (or simply up-
converter) at a base station is particularly suited for such an application and is illustrated in Figure 1-10.

![Figure 1-10](image_url)

Figure 1-10: A millimetre-wave fibre-radio system employing an HBT optoelectronic up-converter at a base station. The RF is in the millimetre-wave region and is higher than the IF.

The HBT optoelectronic up-converter again reduces the complexity of the base station by simultaneously performing photodetection, amplification and mixing which would otherwise be carried out by separate components.

There are other commonly used photodetectors and one of them is the photoconductor. Photoconductors can provide gain but suffer from excessive shot noise due to the dark current. Also the frequency response of the photoconductor is inversely proportional to the gain. On the other hand, photodiodes such as the PIN photodiode (50 GHz bandwidth, 0.75 A/W responsivity [17]), MSM photodiode (105 GHz, 0.1 A/W [18]), Schottky-barrier photodiode (60 GHz, 0.2 A/W [19]) have very wide bandwidth with no internal gain. Avalanche photodiodes have very high gain-bandwidth product (160 GHz gain-bandwidth product with a 3 dB bandwidth greater than 20 GHz at 1.55 μm [20]) due to the avalanche multiplication. However such a multiplication process is probabilistic and an excess noise factor (ratio of the total noise to the multiplied shot noise) of greater than 5 for a multiplication factor of 100 can be incurred by a silicon avalanche photodiode [21]. Moreover the reverse bias voltage required for the avalanche to occur ranges from around 20V to 160V [22], rendering them unsuitable for low-power optoelectronic applications.

Popular phototransistors being considered for fibre-radio applications include metal semiconductor field effect transistors (MESFETs) [23,24], high electron mobility
transistors (HEMTs) [23,24] and heterojunction phototransistors (HPTs) or heterojunction bipolar transistors (HBTs) [25] which is the subject of this Thesis. The main advantage of phototransistors over other types of photodetectors is that they can function as a photodetector as well as a high gain electrical amplifier in a single device without requiring high bias voltages, as opposed to the avalanche photodiode.

In the next section, a brief review of the work on optoelectronic mixing performed using other optoelectronic devices (i.e. not HBTs) will be given first.

**Section 1-4: Optoelectronic Mixing Performed with Other Optoelectronic Devices.**

In the early optoelectronic mixing research, the avalanche photodiode was a popular choice. Since the avalanche multiplication factor is a function of the reverse bias voltage, by modulating this bias voltage with an LO, the avalanche gain applied to the photodetected signal is varied which provides the necessary mixing mechanism.

An electrically pumped avalanche photodiode optoelectronic mixer was reported by Davis and Kulczyk [26] as early as in 1970 in which an avalanche photodiode (type unspecified) was pumped by an LO at 100 MHz. The light source was a He-Ne laser (wavelength unspecified), externally modulated at 70 MHz with 1–10 % modulation depth and the output electrical IF was measured at 30 MHz. Since the incident optical power, the gain of the amplifier used and the electrical output power were not specified, neither the system conversion gain nor the intrinsic conversion gain defined in Section 1-2 can be calculated. In their paper, the conversion loss \( L \) was defined as \( \frac{i_{OF}^2}{i_{RF}^2} \) where \( i_{OF}^2 \) and \( i_{RF}^2 \) are the squares of the mixer output currents at IF (30 MHz) and laser modulation frequency RF (70 MHz), respectively. With a 115 V DC bias voltage, the conversion loss \( L \) was reduced from around 16 dB with 7 V peak LO voltage, to around 2.5 dB with 29 V peak LO voltage. The diode breakdown voltage was 142 V.

Another electrically pumped avalanche photodiode optoelectronic mixer was reported by Seeds and Lenoir [22]. The silicon avalanche photodiode was not only used as a fundamental mixer but also a harmonic mixer (up to the 5\(^{th}\) harmonic, see Eqn 1-1). The
applied LO frequency was 100 MHz and the modulation frequency of the GaAs/GaAlAs laser (780 nm wavelength) was such that for each harmonic mixing, the IF was 10 MHz. However it was not specified whether the laser modulation frequency was above or below the corresponding LO harmonic frequency. Since the optical modulation depth and the electrical output power were not given, the system conversion gain, as defined in Section 1-2, cannot be calculated. In their paper, the conversion loss $L$ was defined as the ratio of the square of the signal current, with the device used as a photodetector, to that of the IF current when it is used as a mixer. Such a definition is equivalent to the reciprocal of the intrinsic conversion gain definition given in Eqn 1-5, provided that the avalanche multiplication factor remains equal to 1 when the device is used as a photodetector. The conversion loss $L$ was a decreasing function of the applied LO power for all five harmonic mixings, with the higher order harmonic mixing conversion losses decreasing more rapidly. With a 160 V reverse bias voltage giving a 10.5 avalanche multiplication factor, an optical illumination giving a 10 $\mu$A unmultiplied photocurrent and +14 dBm LO power, the conversion losses were around 6, 13, 20, 28 and 37 dB for the fundamental, $2^{nd}$, $3^{rd}$, $4^{th}$ and $5^{th}$ harmonic mixing, respectively.

The main drawback of the use of avalanche diodes, as shown in previous two examples, is the very high reverse bias voltage required. In [26] the reverse bias voltage used was from 97 V to 133 V while in [22] it was 160 V.

Gomes and Seeds [27] demonstrated an optically pumped tunnelling metal-semiconductor contact optoelectronic mixer. The semiconductor used was heavily n-doped ($2 \times 10^{18}$ cm$^{-3}$) GaAs and the metal was gold. Two kinds of optical sources were used in turn. In the first experiment, a GaAs semiconductor laser, emitting at 797 nm wavelength, was directly modulated by a 76 MHz LO. The average incident optical power was 3 mW with a generated average photocurrent of 120 $\mu$A. With the output IF set at 2 MHz, a minimum conversion loss of 18 dB was obtained with a forward bias voltage of around 0.14 V. In the second experiment, a frequency-doubled mode-locked Nd: YAG laser (532 nm wavelength) was used as the optical LO with a pulse repetition rate of 76 MHz. With 3.4 mW incident optical power, a minimum conversion loss of 15 dB was recorded at around 0 V bias.
Lam and MacDonald [28] investigated the frequency response of an electrically pumped GaAs photoconductive detector optoelectronic mixer. The 100 μm x 100 μm GaAs device had a 0.3 μm active-layer n-type doped to 10^{16} cm^{-3}, and had interdigitated contacts with 10 μm finger spacing. The laser modulation frequency and the LO frequency were swept simultaneously from 1 to 4.5 GHz and from 0.5 to 4 GHz, respectively, giving a constant output IF of 500 MHz. With the (Mitsubishi) MIT-5308 laser biased at 34 mA and modulated with 0 dBm, and the LO power set at 10 dBm, the highest 3 dB bandwidth was 4.8 GHz in the LO frequency. Since the actual output power was not given, neither the system nor intrinsic conversion gains can be calculated.

Ogawa and Kamiya [29] used InGaAs PIN photodiodes (responsivity of 0.58 A/W and 3-dB bandwidth of 10 GHz) as both electrically and optically pumped optoelectronic mixers in two fibre-optic transmission configurations. In the electrically pumped experiment, the 10 GHz LO was generated at the base station by photodetecting with a separate photodiode the 5th harmonic of a laser diode directly modulated at 2 GHz at the central station. The input signal to be up-converted by the optoelectronic mixer was a 0.9 GHz intensity modulated light source. With an applied LO power of 4 dBm and input signal power of 13 dBm at the central station, the up-converted signal power at the base station was measured to be around -30 dBm. The optical power levels reaching the base station were not specified.

In the optically pumped experiment, the LO was produced by harmonic distortion of the laser diode, which was now directly modulated at 4 GHz. The input signal to the optoelectronic mixer was obtained by photodetecting with a separate photodiode a 0.9 GHz intensity modulated light source. With 10 dBm LO power applied to the harmonic generating laser and 8.3 dBm signal power applied to another laser, the measured power levels of the up-converted signals were around -30 dBm at 4 ± 0.9 GHz and 8 ± 0.9 GHz, and -37 dBm at 12 ± 0.9 GHz. The optical power levels reaching the base station were not specified.

Liu et al. [30,31,32,33] performed electrically pumped optoelectronic mixing with a GaAs Metal-Semiconductor-Metal (MSM) photodiode (PD) integrated with a two-stage transimpedance MESFET amplifier. The MSM PD had an active area of 100 μm x 100
Chapter 1

\(\mu m\) with a 1 \(\mu m\) finger width and a 3 \(\mu m\) spacing. The DC responsivity, which is the dc photocurrent divided by the incident mean optical power, was 0.2 A/W at 10 V bias voltage. The two-stage MESFET amplifier had a transimpedance of 48 dBΩ. In the mixing experiment, an AlGaAs laser diode emitting at 780 nm wavelength was directly modulated at 60 MHz. The optical power incident on the MSM PD was 0.4 mW, but the modulation depth was not specified. The DC bias voltage of the MSM PD was modulated by a 17 dBm, 100 MHz LO so that the responsivity of the MSM PD became time-varying and provided the necessary mixing mechanism. After a 20 dB gain amplifier, the output IF signal at 40 MHz was measured and had a power of around -37 dBm. The conversion loss, defined in their papers as the ratio of the square of the current, with device used as a photodetector, to that of the IF current when it is used as a mixer, was about 14 dB.

Later Liu et al. performed a slightly different experiment in which the laser diode was directly modulated by two signal tones, one at 150 MHz and the other at 160 MHz. The two tones had the same power level of -10 dBm. The incident optical power on the MSM PD was 0.4 mW. The applied LO power and frequency were 0 dBm and 1.65 GHz, respectively. After being bandpass filtered at around 1.8 GHz, amplified by a 50 dB gain amplifier, radiated by a 2 dBi discone antenna and transmitted over a distance of 2 m, the up-converted signals were received by another discone antenna followed by two 25 dB gain amplifiers. Measurements at a position just before the radiating antenna showed that the two up-converted signals were -20 dBm with the third-order intermodulation 44 dB lower, limited by the laser nonlinearity. Measurements after the radio path and the two 25 dB gain amplifiers at the receiver side showed similar power levels for both the signals and the intermodulation products.

Other microwave transistors have also been used in optoelectronic mixing.

Malone et al [34] reported an electrically pumped MESFET optoelectronic mixer. The commercial MESFET had four 75 \(\mu m\) wide gate fingers, a gate length of 0.8 \(\mu m\) and a dopant concentration of 3\(\times10^{17}\) cm\(^3\). The 850 nm wavelength laser source had 1.8 mW output optical power and around 100 % intensity modulation depth. The light was conveyed to the MESFET gate on a multimode fibre. The electrical gate terminal was
also pumped by a 5 dBm LO source. The output IF signal, which was fixed at 50 MHz, was taken from the MESFET drain and measured with a spectrum analyser. With $V_{gs} = -0.5 \text{ V}$ and $V_{ds} = 2 \text{ V}$, the laser modulation frequency was swept from 50 MHz to 3 GHz while the LO was swept from 100 MHz to 3.05 GHz, giving a constant 50 MHz IF. The IF output power decreased from -26 dBm at 50 MHz RF, to -44 dBm at 3 GHz RF. Assuming all the laser output optical power was incident on the MESFET, these would correspond to system conversion gains of -12 dB and -30 dB, respectively. The IF output of the optoelectronic mixer was also 5 to 10 dB lower than the directly detected laser modulation signal in the above frequency range.

An electrically pumped HEMT image-rejection optoelectronic up-converter was reported by Kamitsuna and Ogawa [35] and Figure 1-11 illustrates its configuration.

![Figure 1-11: An electrically pumped HEMT image-rejection optoelectronic up-converter](image)

The optoelectronic mixer consisted of two MMIC HEMTs whose two gates (100 $\mu$m wide) were driven by an LO through an in-phase power divider. By directly modulating the two laser with two 110 MHz IF input signals which were 90 degrees out of phase, and combining the outputs of the two HEMTs through a 90-degree power combiner, image rejection was achieved. In their experiment, $V_{ds} = 2 \text{ V}$, $V_{gs}$ was set near the pinch-off voltage, the LO power was set at 4 dBm, the IF power applied to the 90-degree power divider was 16 dBm and the average optical power incident on each HEMT was around 0.2 mW with the modulation depth unspecified. When the LO frequency was swept from 28.5 GHz to 31.5 GHz, the power level of the desired upper sideband of the up-converted signal remained between -44 and -41 dBm in the frequency range $f_{LO} + 110 \text{ MHz}$. The
power level of the lower sideband, which was to be rejected, was at least 15 dB below the corresponding upper sideband, thus confirming the proper operation of the image-rejection optoelectronic up-converter. With $f_{\text{LO}}$ fixed at 30 GHz while sweeping the IF from 70 to 170 MHz, an image-rejection of better than 20 dB was obtained.

In the optoelectronic mixing techniques discussed so far, the optical intensity modulation has been first photodetected and then mixed with the applied electrical LO power. It is, however, also possible to perform the necessary frequency conversion purely in the optical domain with an external modulator driven by an LO and then use a photodiode or any direct photodetector to convert the intensity modulated optical signal into an electrical output. Such an approach using external modulators is particularly useful in millimetre-wave fibre-radio link. Park et al [36] reported using a LiNbO₃ Mach-Zehnder external optical amplitude modulator (EOM), driven by a 38 GHz LO and DC biased at the half-intensity point, to up-convert the intensity modulation of a 1310 nm wavelength laser which had been directly modulated with a 150 MHz digitally modulated carrier. Figure 1-12 illustrates their experimental arrangement.

![Figure 1-12: Optoelectronic mixing experimental arrangement using a Mach-Zehnder modulator [36]. PD: photodiode.](image)

After transmission through a 5m long single-mode fibre, around 1 mW optical power was detected with a high-speed photodiode and the output was an electrical spectrum with a carrier at 38 GHz and two sidebands 150 MHz above and below this carrier. The conversion loss, defined in their paper as the ratio of the detected power at 150 MHz to that of the detected up-converted power at 38 GHz ± 150 MHz, was 18 dB. The actual measured electrical powers were not given. In such a configuration, the modulation bandwidth around the 150 MHz carrier is only limited by the laser, but not the external
modulator because the external modulator simply up-converts the laser modulation to the millimetre-wave region.

Although not strictly an optoelectronic mixer according to the descriptions given in Section 1-1, Fetterman et al. [37] used a two-stage GaAs FET amplifier to detect the beat signal of two coherent light sources. The two light sources were a visible tunable CW ring dye laser and a stabilised He-Ne laser (632.8 nm wavelength) and simultaneously incident on the FET amplifier. The wavelength of the tunable laser was adjusted to give a beat signal frequency of 32 GHz at which the FET amplifier had a gain of 16 dB. The output beat signal was -60 dBm with a signal-to-noise ratio of 12 to 15 dB in a 1 MHz resolution bandwidth.

Later working with Fetterman and Chew, Ni et al. [38] performed a similar experiment with a GaAs FET (0.3 μm gate length and 0.2 μm drain-source distance) in which a beat signal of 62 GHz was generated by coherently mixing the optical outputs of a stabilised He-Ne laser and a tunable CW ring dye laser. The FET gate-source port was attached to a mm-wave printed-circuit twin dipole antenna. To measure the mm-wave beat signal indirectly, the 62 GHz signal was down-converted to a 2 GHz IF by applying an LO which was fed to the FET through a horn and the twin dipole antenna. A 37 dB gain amplifier was used to amplify the output IF. With 50 mW and 20 mW LO powers applied to the horn, the output IF power were -52 dBm and -66 dBm, respectively. The optical power densities incident on a separate control FET were 1 kW/cm² from the He-Ne laser and 40 kW/cm² from the dye laser.

In the next Section, the concept of the heterojunction will be described. The use of such a heterojunction in an HBT results in superior high-frequency performance compared to the homojunction bipolar transistor (BJT) and the reasons for this will also be qualitatively explained.

Section 1-5: The Heterojunction in Bipolar Transistors

A heterojunction, which is one formed between two dissimilar semiconductors, was first suggested by Shockley [39] and the use of such a heterojunction in the bipolar transistor
was later proposed by Kroemer [40] in which the base-emitter heterojunction was formed by a wide bandgap emitter and a narrow bandgap base, and the resultant device is called the HBT. Since then the use of the wide bandgap emitter has been extended to phototransistors (i.e. heterojunction phototransistor or HPT) and an early theoretical analysis of the HPT was reported by Morizumi et al. [41]. Further theoretical analysis of the HPT was also carried out by other researchers including Milano et al. [42], Chand et al. [43] and Campbell [44].

A number of material systems can be used to form heterojunctions, including Al\(_{0.3}\)Ga\(_{0.7}\)As/GaAs (1.82eV/1.42eV), GaAs/Ge (1.42eV/0.66eV), InP/In\(_{0.53}\)Ga\(_{0.47}\)As (1.35eV/0.75eV), Al\(_{0.48}\)In\(_{0.52}\)As/InP (1.46eV/1.35eV), Ga\(_{0.52}\)In\(_{0.48}\)P/GaAs (1.89eV/1.42eV) and Si/Si\(_{0.8}\)Ge\(_{0.2}\) (1.12eV/0.92eV). However, in order for the HBT to be compatible with long wavelength (1300 nm and 1550 nm) optical communication systems, the InP/In\(_{0.53}\)Ga\(_{0.47}\)As (or InP/InGaAs for short) material system is chosen in which the base and the collector are made of InGaAs and the emitter made of InP.

The use of a wide bandgap emitter in bipolar transistors offers a number of advantages. Assuming no avalanche multiplication, the common base current gain of a bipolar transistor, be it heterojunction or homojunction, is given by [45]

\[
\alpha_0 = \gamma \alpha_T
\]

where \(\gamma\) is the emitter efficiency and \(\alpha_T\) is the base transport factor, both of which are smaller than unity. The common-emitter current gain can be expressed in terms of \(\alpha_0\) [45]

\[
h_F = \frac{\alpha_0}{1-\alpha_0}
\]

Therefore in order to achieve a high common-emitter current gain, \(\alpha_0\) should be made as close to unity as possible. In a BJT, \(\alpha_0\) is made close to unity by doping the base very lightly relative to the emitter so that the emitter efficiency can be increased. However having a very lightly doped base has a number of undesirable effects. Since the base doping is so low, the base sheet resistance is correspondingly higher, thus reducing the

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1 Energy bandgap
transistor power gain at high frequencies. To keep the base resistance below an acceptable value, the base therefore cannot be made too thin which results in a longer base transit time and lower high-frequency current and power gains. To improve the high-frequency performance of a bipolar transistor, it is therefore important to have a very low-resistance and thin base without compromising the emitter efficiency. A base-emitter heterojunction can eliminate this constraint.

Figure 1-13 shows the band energy diagram of the base-emitter heterojunction of an idealised Npn\(^2\) type HBT.

![Band diagram of an idealised base-emitter heterojunction.](image)

Figure 1-13: Band diagram of an idealised base-emitter heterojunction. \(E_c\): emitter energy bandgap, \(E_C\): bottom of the emitter covalent band energy level, \(E_{v'}\): top of the emitter valence band energy level, \(E_b\): base energy bandgap, \(E_{CB}\): bottom of the base covalent band energy level, \(E_{vb}\): top of the base valence band energy level, \(E_{barrier-c}\): covalent band energy barrier for electrons, \(E_{barrier-v}\): valence band energy barrier for holes. \(e^+\): electrons, \(h^+\): holes.

The energy bandgap difference in the emitter and the base is such that in an Npn HBT, holes injected from the base are blocked from entering the emitter because of a higher valence band energy barrier. On the other hand, the electrons injected from the emitter experience only a small conduction band energy barrier and can therefore enter the base.

\(^2\) It is customary to use upper case letters to denote the doping type of the wide bandgap materials and lower case letters for that of the narrow bandgap materials.
already. It is this imbalance in the energy barriers faced by the electrons and the holes travelling in opposite directions at the junction that makes the emitter efficiency independent of the relative doping levels in the base and the emitter. As a result, the base of an HBT can then be highly doped to reduce the base resistance and a much thinner base can be made. It should be noted that the reasons for use of the heterojunction in an HBT is not to increase the emitter efficiency, but rather to afford the device designers the freedom to tailor the device doping levels to improve the performance without compromising the emitter efficiency and hence the gain of the HBT. Impressive HBT results have been reported. For example Oka et al\[46\] reported an InP/InGaAs HBT with a current gain cutoff frequency ($f_T$) of 209 GHz and a maximum oscillation frequency ($f_{max}$) of 138 GHz. The base was only 50 nm thick and p-type doped with carbon at $2 \times 10^{19}$ cm$^-3$.

### Section 1-6: Photodetection in HBTs

To understand how an HBT can function as a phototransistor, it is necessary to identify which parts of the HBT are photo-sensitive at a specific wavelength. Consider an idealised Npn type InP/InGaAs two-terminal HBT under a constant illumination at 1550 nm wavelength via an optical window in the emitter contact as shown in Figure 1-14.

![Figure 1-14: An idealised two-terminal Npn type InP/InGaAs HBT under optical illumination at 1550 nm wavelength.](image)

The 1550 nm wavelength corresponds to a photon energy of 0.8 eV. In such an InP/InGaAs HBT, the InP emitter is not able to absorb any photon because its bandgap
energy (1.35 eV) is larger than the photon energy. The InGaAs base and the InGaAs collector have a bandgap energy of 0.75 eV which is slightly smaller than the incident photon energy. Therefore only the base and the collector are the photo-sensitive parts at this wavelength. Like a normal homojunction bipolar transistor (BJT), the HBT is usually operated with the base-collector junction reverse biased and the base-emitter forward biased and this mode of operation is called the forward-active mode. The reverse biased base-collector junction resembles a pn junction photodiode. In an Npn HBT, photons are absorbed in the base-collector junction depletion region, the neutral base and collector regions. A number of electron-hole pairs are generated due to a number of photons absorbed and the ratio is determined by the internal quantum efficiency $\eta_{\text{int}}$. Electron-hole pairs generated in the base-collector junction depletion region are separated by the electric field with the electrons drifting towards the collector and the holes towards the base constituting a photocurrent. Electron-hole pairs generated in the neutral base and collector regions where there is very weak electric field face two fates. If they are generated within a diffusion length from the depletion region, they can diffuse into the depletion region and get separated by the electric field, constituting a photocurrent. Otherwise they will recombine after their recombination life-times have elapsed contributing no current. It should be noted that diffusion is a rather slow process and in high-speed applications, it is important to ensure that most photons are absorbed in the depletion region where electrons and holes traverse the depletion region at their respective saturation velocities due to the strong electric field. Therefore a lightly doped collector is usually employed so that the base-collector junction depletion region can extend further into the collector.

The photocurrent discussed in the last paragraph is of primary nature and therefore it is called the primary photogenerated current or simply primary photocurrent. The concept of the primary photocurrent or $I_{\text{prim}}$ can be used to devise the equivalent circuit representation for an HBT under illumination shown in Figure 1-15.

The primary photocurrent generation is represented by an illuminated photodiode connected between the collector and the base of a normal HBT. In the absence of an externally supplied base current, as in the two-terminal HBT, $I_{\text{prim}}$ acts as the sole base bias current controlled by the light and is amplified by the HBT by normal transistor
action. In the presence of an externally supplied base current in the three-terminal HBT, $I_{\text{prim}}$ adds to the externally supplied base bias current. However when $V_{BE}$ of the HBT is fixed, there will be no photocurrent gain because it is $V_{BE}$ which controls the injection of electrons from the emitter.

**Section 1-7: HBTs as Direct Photodetectors**

In order to enhance the performance of the HBT as a direct photodetector, optoelectronic mixer and an optically controlled oscillator, it is fundamentally important to improve the basic HBT parameters such as the quantum efficiency, light coupling loss, current gain and frequency response. Some important work in these areas will be reviewed below.

**Section 1-7a: Experimental Investigations of Two-Terminal HBTs by Campbell and Co-Workers**

Campbell and co-workers published a number of papers on two-terminal, collector-up, back-illuminated HBTs in the early 1980s. In [47,48], they described the fabrication and characterisation of a two-terminal InP/InGaAs HBT. The light was shone through the InP emitter. The optical gain achieved was about 40 from 0.95 $\mu$m to 1.4 $\mu$m wavelength at an optical power of 1 nW. The shorter wavelength limit was due to the absorption in the InP emitter region and the longer wavelength limit was due to the bandgap energy of the InGaAs base and collector being greater than the incident photon energy. The optical gain increased to 1000 at 5 $\mu$W optical power. This dependence of gain on the optical
power or the resultant collector current is a result of the base-emitter junction defect current (or recombination current). At very small optical powers (< 100 nW), however, the gain became small (around 40), but relatively independent of the optical power because the base-collector junction leakage current, which provided a quiescent current for the base-emitter junction, was larger than the photogenerated current. Campbell et al. in [49] showed that the sensitivity of an optical receiver utilising their HBTs mentioned above could be as good as or exceed that of a hybrid PIN photodiode/field effect transistor-preamplifier combination.

Campbell et al. [50] reported another two-terminal, collector-up, back-illuminated InP/InGaAs HBT. Using a He-Ne laser of 1.154 μm wavelength, the common-emitter current gain was 100 at as low as 20 nW optical power and 600 at 2.5 μW optical power. The cutoff frequency was 100 MHz at $I_C = 10 \mu A$ and 1.7 GHz at $I_C = 500 \mu A$. The dependence of the cutoff frequency on $I_C$ was due to the emitter dynamic resistance, $r_e$, which is governed by

$$\text{Eqn 1-10} \quad r_e = \frac{kT}{qI_C}$$

where $k$ is the Boltzmann constant, $T$ is the temperature in Kelvin and $q$ is the electron charge.

In 1983 Campbell et al. [51] reported a two-terminal, avalanche InP/InGaAs HBT with a very high gain. Similar to previous reported HBTs, the device was back-illuminated through the transparent InP substrate. The avalanche multiplication took place in the base-collector junction. If the transistor has a common-emitter current gain $h_{fe}$ and the base-collector junction has an avalanche multiplication factor $M$, then the collector current is given by

$$\text{Eqn 1-11} \quad I_C = \frac{(h_{fe} + 1)M}{1 - (M - 1)h_{fe}} [I_{\text{prim}} + I_{\text{co}}]$$

where $I_{\text{prim}}$ and $I_{\text{co}}$ are the primary photocurrent and the base-collector leakage current, respectively. It is noted that the breakdown voltage of the avalanche phototransistor is smaller than the breakdown voltage of the base-collector junction alone due to the gain, $h_{fe}$, of the transistor. An extremely high optical gain of 25000 was achieved when the
phototransistor was operated near its breakdown voltage where \( 1 - (M - 1)h_R = 0 \). To operate the HBT near its breakdown voltage, the bias voltage has to be very stable.

The HBTs reported by Campbell to date are all two-terminal HBTs. While this configuration has the advantage of eliminating the base contact capacitance, the optical gain is usually very small at low optical input powers due to the base-emitter junction recombination current. The optical gain at low optical powers can be improved by providing a quiescent emitter current by electrical means, i.e., having an electrical base, or by optical means. Campbell et al. [52] in 1982 integrated a light emitting diode (LED) with their HBT so that optically coupled dc bias current was provided. The optical gain was increased from 30 to 60 when the LED was driven by a 6 mA current.

A summary of the work carried out by Campbell and co-workers along with some general theories of the HBT can be found in [53].

**Section 1-7b: Resonant Cavity (Wavelength Selective)**

The external quantum efficiency of a thin collector HBT can be improved by placing the HBT structure in a resonant cavity. Dodabalapur et al. [54] have demonstrated a high optical gain of about 2500 from their two-terminal InGaAlAs/InGaAs/InAlAs HBT with the resonant cavity wavelength at 1.3 \( \mu \text{m} \). Their resonant cavity consisted of a gold emitter contact serving as the top reflector and an InAlAs/InGaAlAs quarter-wave stack (QWS) as the bottom reflector. Unlu et al. [55] have also used a resonant cavity enhanced two-terminal AlGaAs/GaAs HBT with an intermediate InGaAs layer in the collector to achieve an external quantum efficiency of 43% at 0.9 \( \mu \text{m} \) wavelength. A 43% external quantum efficiency has also been achieved by Huang et al. [56] with a two-terminal HBT at 860 nm wavelength. In their resonant cavity the top reflector was the GaAs/semiconductor interface, with a reflectivity of about 30%, while the bottom mirror was 13 pairs of QWS, giving a reflectivity of about 90%. Use of the resonant cavity makes the HBT wavelength selective which may be useful in wavelength division multiplexing. Some authors, however, have investigated wide spectral bandwidth, resonant cavity enhanced HBTs. Sjolund et al. [57] have incorporated an InGaAs/GaAs multi-quantum well structure in the collector region of a two-terminal HBT as an
absorber in such a way that the integrated absorption is insensitive to the standing wave
distribution in the resonant cavity. This has resulted in a responsivity of 400 A/W with
100 μW incident optical power at 929 nm wavelength with a full width at half maximum
(FWHM) spectral bandwidth of about 9 nm.

Section 1-7c: Edge-Coupled HBTs

In conventional HBTs, the optical input is normal to the plane of the epilayer. The length
for optical absorption is the total thickness of the base and the collector in such a
configuration. However the base is invariably kept very thin to maximise the base
transport factor and minimise the base transit time, and the base and collector areas are
kept small for minimum capacitance. As a result, coupling light into the HBT becomes
very difficult and inefficient. Wake et al. [58] demonstrated a two-terminal InP/InGaAs
HBT in which the input light was edge-coupled, in parallel with the epilayer, to the base-
collector depletion region. The device area was 5 x 10 μm² with the narrow side serving
as the optical input. Therefore the length for optical absorption was 10 μm. Their HBT
showed a responsivity of 175 A/W and a DC current gain greater than 270. The HBT
also showed a common-collector unity gain frequency beyond 30 GHz when compared to
an edge-coupled photodiode mounted in an identical package [59]. In this work an
electrically pumped optoelectronic mixer has been built using such a two-terminal HBT
provided by Wake et al. [58] and the experimental results will be given in Chapter 4.

Margnin et al. [60] reported both two-terminal and three-terminal edge-coupled
InP/InGaAs HBTs whose structures were very similar to those by Van de Casteele et al.
[81] which will be described in Section 1-8. The width of the emitter mesa was 4 μm
while the length of the waveguide varied from 4 μm to 20 μm, depending on the cleaving
operation. The paper compared the frequency responses of the two types of HBTs.
While the three-terminal HPT exhibited flatter frequency response with a photocurrent
gain-bandwidth product of 40 GHz, the two-terminal HBT had a gain which was over 10
dB higher than the three-terminal counterpart below 1 GHz. Also the two-terminal HBT
frequency response rolled off more steeply with a gain-bandwidth product of only 25
GHz. Although no explanation was given for the observed phenomena, it is believed that
the higher low-frequency gain of the two-terminal HPT was due to the absence of an
electrical base connection to which the photogenerated signal in the base-collector junction may leak, as in the case of a three-terminal HBT. The higher gain-bandwidth product observed in the three-terminal HBT is believed to be due to the higher quiescent emitter current (invoked by a 400 μA base current) which in turn lowered the emitter dynamic resistance (Eqn 1-10).

**Section 1-7d: Indium Tin Oxide (ITO) Emitter Contact HBTs**

Coupling of light into HBTs is usually done through the transparent substrate or the emitter where an optical window is located in the emitter metallisation. However the emitter contact area must be kept small because it would otherwise increase the base-emitter capacitance and reduce high frequency response. As a result, the optical window is very small leading to difficulties in optical alignment. A number of researchers have fabricated HBTs using the transparent Indium Tin Oxide (ITO) as the emitter contact in an attempt to improve the light coupling efficiency. However reported results of ITO HBTs are not very encouraging to date. Bashar *et al.* [61,62] reported both three-terminal AlGaAs/GaAs and InP/InGaAs HBTs with transparent ITO emitter contacts. A responsivity of 2.5 A/W at 630 nm for the AlGaAs/GaAs HBT was achieved. For the InP/InGaAs HBT, the responsivities were 5.4 A/W at 780 nm and 30 A/W at 1310 nm. For both types of HBTs, the emitter/base and base/collector areas were $7.8 \times 10^{-5}$ cm$^2$ and $3.7 \times 10^{-4}$ cm$^2$, respectively. The common-emitter current gains were 16 for the AlGaAs/GaAs HBT and 28 for the InP/InGaAs HBT. The low current gain might have been due to the high base-emitter recombination current and small base transport factor rather than the use of the ITO emitter contact. A plot of $\log I_C$ vs. $V_{BE}$ with $V_{BC} = 0$ V in their paper suggests that the ITO emitter contact might have had high resistance because the plot starts to deviate from a straight line at around $I_C = 0.1$ mA with a greater-than-unity ideality factor. Their work also shows that the InP/InGaAs HBT has very low turn-on voltage while the AlGaAs/GaAs has a turn-on voltage of about 0.7 V.

Li *et al.* [63] have also fabricated three-terminal GaAs/Al$_{0.25}$Ga$_{0.75}$As HBTs with ITO emitter contacts. The emitter area was $2 \times 10 \mu$m$^2$. They obtained a current gain of around 10 to 20, a current gain cutoff frequency $f_T = 18$ GHz, a maximum oscillation frequency $f_{max} = 20$ GHz at $I_C = 10$ mA and $V_{CE} = 3$ V. A 6 GHz optically injection
locked oscillator using such an HBT has also been built by their group [85] and will be described in Section 1-9.

**Section 1-7e: Use of an Optical Reflector**

Another way of improving the light coupling efficiency is to use a reflector so that light is reflected after passing the base and collector once, thus effectively doubling the length for optical absorption. Fukano *et al.* [64] demonstrated a three-terminal InP/InGaAs HBT with a nonalloyed electrode metal serving as the emitter metallisation and as an optical reflector. Light at 1.3 μm or 1.5 μm entered the InGaAs collector and base through the semi-insulating InP substrate. The photons which were not absorbed by the collector and the base in the first pass got reflected back by the emitter metallisation, making a second pass. The measured external quantum efficiencies at 1.3 μm and 1.55 μm were 37% and 21%, respectively, and these figures, according to the authors, are about 50% larger than those HBTs when the light is illuminated from the top window. The emitter and base areas were 3 x 3 μm² and 6 x 8 μm². At 1.3 μm wavelength, a DC optical gain of greater than 70 was achieved. When the base was terminated in a 50 Ω resistor, the unity optical gain frequencies were 22 GHz and 14 GHz at λ = 1.3 μm and λ = 1.55 μm, respectively.

**Section 1-7f: Use of a Guard-Ring Structure**

Twynam *et al.* [65] have used a guard-ring in their GaInP/GaAs HBT to reduce the surface leakage current at the emitter periphery. The guard-ring is a metal contact which is located near the emitter periphery and totally surrounds the emitter contact. When a positive bias is applied to the guard-ring, with respect to the emitter contact, the emitter-base voltage at the perimeter of the junction is reduced and the surface leakage current is suppressed. Optical access was through a top emitter window. The static current gain was 400 and optical gain at 870 nm was greater than 2000.

**Section 1-7g: Three-Terminal and Two-Terminal HBT Comparisons**

In HBTs, the base-emitter junction is formed by two different semiconductors. The quality of the heterojunction might not be as good as that of a homojunction due to the lattice constant mismatch between the two materials and, therefore, a number of
recombination centres exist at the heterojunction interface. These recombination centres trap the photogenerated holes which effectively wastes some of the photogenerated current and results in a reduced optical gain, especially at low incident optical power. The resultant low emitter current at low optical powers further decreases the frequency response of the HBT because the emitter dynamic resistance is increased (Eqn 1-10). The current which is wasted to fill these recombination centres is termed recombination current or defect current. At higher incident optical power, a larger collector current flows and a relatively small portion of this current is used to compensate for the recombination current and therefore the optical gain is not affected significantly. However large optical power is not always available.

One way of improving the performance of a phototransistor is to have an electrical base to supply a base current so that even without any light, a quiescent collector is still flowing which can compensate for the above-mentioned recombination current. This was first suggested by Fritzsche et al. [66] and re-addressed by Chandrasekhar et al. [67]. Chandrasekhar et al. [67] reported and compared experimental results of an emitter-up, InP/InGaAs HBT with and without a base terminal. The backside of the device was polished for optical access through the semi-insulating InP substrate. The base-emitter junction area and the base-collector junction area were 12 x 12 $\mu$m$^2$ and 20 x 33 $\mu$m$^2$, respectively. At 1 $\mu$W incident optical power, the DC optical gain was increased from 40 for the open-base HBT to 190 for the three-terminal configuration when an external base current of 11 $\mu$A was supplied. By intensity modulating the 1$\mu$W incident optical power, the frequency response was also measured. The optical gain 3-dB bandwidth was increased from 3.5 MHz in the open-base HBT to 52 MHz for the three-terminal configuration when an external base current of 15 $\mu$A was supplied. Although performed at relatively low frequencies, these results clearly show the advantages of having an electrical base connection for supplying a base current.

**Section 1-7h: Punchthrough HBTs**

It is now apparent that three-terminal HBTs with an electrical base have better gain at low optical power than their two-terminal counterparts due to the base-emitter junction recombination current being compensated by the externally supplied base current.
However the applied base current will introduce shot noise which is then amplified by the transistor small signal gain. Therefore the three-terminal HBT is only useful when the total thermal noise of the load and the subsequent pre-amplifier is dominant.

Y. Wang et al. [68] have proposed a punchthrough HBT in which the base is lightly doped so that under normal operation, the base is completely depleted. When the depletion width reaches the base-emitter heterojunction, a quiescent collector current flows which compensates the base-emitter recombination current. The quiescent collector current is an exponential function of $V_{CE}$. As a result the optical gain should still be high even at low optical power. According to their simulation, the punchthrough HBT should have a better signal-to-noise ratio than the conventional three-terminal HPT on the assumption that the shot noise associated with the punchthrough quiescent collector current does not get amplified by the HBT gain. However no experimental result was provided to substantiate this claim. A two-terminal, emitter-up HBT was fabricated with the following doping profiles: $n$-type GaAs sub-collector (600 nm, $3 \times 10^{18}$ cm$^{-3}$), $n$-type GaAs collector (500 nm, $1 \times 10^{16}$ cm$^{-3}$), $p$-type GaAs base (500 nm, $3 \times 10^{16}$ cm$^{-3}$), $n$-type AlGaAs emitter (200 nm, $2 \times 10^{18}$ cm$^{-3}$) and $n$-type GaAs cap layer (100 nm, $5 \times 10^{18}$ cm$^{-3}$). The reported optical gain was about 1240 at wavelengths from 790 nm to 850 nm with 0.5 µW incident optical power.

Section 1-71: Travelling-Wave HBT with Polyimide Waveguides

Prakash et al [69,70,71] reported a two-terminal AlInAs/InGaAs travelling-wave HBT integrated with polyimide waveguides. To fabricate the HBT as a travelling-wave transistor, the metal pads of the HBT were coplanar transmission lines with the centre line contacting the emitter and the ground plane contacting the collector. The centre electrode width was 20 µm and separation between the centre electrode and the ground plane was 15.11 µm providing a 50 Ω characteristic impedance. The $10\times10$ µm$^2$ polyimide waveguide was defined on top of the emitter in the gap between the centre electrode and the ground plane. When intensity modulated light was coupled to the polyimide waveguide horizontally and leaked into the HBT active region along the length of the device (because the refractive index of the semiconductor is higher than that of the polyimide), a microwave signal was generated on the transmission line as light was
absorbed along the length of the HBT. 20, 200 and 2000 μm long HBTs were fabricated and had DC optical gains of 2.3, 6.17 and 13.5, respectively. This HBT is potentially very fast and limited only by the velocity mismatch between the optical and electrical waves, and impedance mismatch with external circuits. As a demonstration of their HBT high-frequency performance, a 2-mm long HBT was used to photodetect the 20 GHz beat signal of two tunable diode pumped YAG lasers at 1.3 μm. The incident optical power was 2 mW and a 25 dB S/N in 10 kHz resolution bandwidth was obtained. This HBT can also handle very high optical power without optical gain saturation because the absorption of the incident light is distributed along the entire length of the device. To demonstrate this, the authors repeated the above heterodyning experiment with a beat signal at 60 GHz using a 200 μm long HBT. The incident optical power was adjusted so that the resultant collector current varied from 0.3 mA to 50 mA. Within this current range, no power saturation was observed. However, the output signal powers were only given in arbitrary units, therefore no knowledge of the HBT responsivity at high frequencies can be derived.

**Section 1-8: HBT Optoelectronic Mixers Reported by Other Researchers**

It should be noted that all HBT optoelectronic mixers reported by other researchers to date have only used three-terminal HBTs. In fact the only two-terminal HBT optoelectronic mixer reported is that of this work which will be presented in Chapter 4.

Urey *et al.* [72] were the first to demonstrate an HBT optoelectronic mixer. In their work, the signal-to-noise ratio (S/N) of an HBT, a HEMT and a JFET down-converting optoelectronic mixer were compared. These devices were all InGaAs based. The HBT had a 5 x 5 μm² transparent window in the 8 x 20 μm² emitter. The base, collector and sub-collector were 0.1 μm, 0.3 μm and 0.8 μm thick, respectively. Both electrically pumped and optically pumped mixing experiments were conducted. In electrically pumped mixing, the base of the HBT (and the gates of the HEMT and JFET) was driven by a microwave signal source acting as the LO and the input signal was the intensity modulated light. In the optically pumped mixing, the LO was used to modulate a DFB laser close to full modulation and optically amplified to 15 mW mean optical power.
before being focused on the HBT and other transistors. The electrical input signal was applied to the base (or the gate of other transistors).

In a 3 MHz resolution bandwidth, the HBT showed the highest S/Ns of 35.7 dB and 28.9 dB among the transistors in the electrically pumped mixing (RF = 10 GHz / IF = 600 MHz) and optically pumped mixing experiments (LO = 6 GHz / IF = 600 MHz), respectively. Using the information given in that reference, the system conversion gains of the HBT, HEMT and JFET for electrically pumped mixing have been calculated to be -21 dB, -26 dB and -29 dB, respectively. In optically pumped mixing, the conversion gains of the HBT, HEMT and JFET were 17.9 dB, 6.9 dB and 4 dB, respectively. These results suggest that the HBT is the preferred optoelectronic device in such an application to the HEMT and the JFET.

It was mentioned in Section 1-3b that the HBT, configured as an up-converting optoelectronic mixer, can be applicable in microwave/millimetre-wave fibre-radio systems. Suematsu et al. [73,74] have used a GaAs/AlGaAs HBT as an optoelectronic up-converter in delivering millimetre-wave signals over fibre. The HBT studied was not specially designed as a phototransistor but as an MMIC component. When light of 0.83 μm wavelength was incident normally on the HBT, the photons were absorbed in the base and the collector through the gaps among the emitter, base, and collector electrodes. The photodiode formed by the base-collector junction of the HBT had an external quantum efficiency of 30% to 40%. Three different fibre optic link configurations were considered in their papers. In link 1 an external modulator was used to intensity modulate a laser beam with an RF input signal at millimetre-wave frequencies at the transmitter and an HBT or a PIN photodiode was employed to detect and regenerate the RF signal at the receiver. In link 2 the external modulator intensity modulated the laser beam merely with the low-frequency IF input signal and the HBT, whose base was driven by the LO source at millimetre-wave frequencies, was used to up-convert the IF modulated optical input signal to an RF output signal at millimetre-wave frequencies. In link 3 the laser beam was again intensity modulated by the external modulator at the same input IF, but the HBT was simply used as a direct photodetector which was then followed by a double balanced diode mixer to perform the up-conversion. It was observed that the S/N for link 2 was 10 dB and 21 dB higher than that for link 1 when the HBT and the p-i-n photodiode...
were used, respectively. The reason why link 2 performed better than link 1 is because certain components in the link are more efficient at the lower IF than at the higher RF. For example, the external modulator used had a 3-dB bandwidth of only 5 GHz. On the other hand, link 2 had a S/N which was 2 dB lower than that of link 3 because, according to their papers, the HBT, when operated as a linear device, had a higher $S_{21}$ than when being pumped. The intrinsic conversion gain of the HBT in link 2 was -4 dB at IF = 3.2 GHz / RF = 30 GHz and -14 dB at IF = 3.2 GHz / RF = 50.8 GHz, respectively.

More recently, another HBT optoelectronic mixer in the up-converting configuration has been reported by Gonzalez et al [75,76]. The HBT was of the InP/InGaAs material system and was top illuminated through a 56 $\mu$m$^2$ optical window with an external quantum efficiency of 26%. In the mixing experiment, the LO was fixed at 30 GHz and -4.5 dBm, and was applied to the base of the HBT. The optical source at 1550 nm wavelength was intensity modulated by an IF input signal from 200 MHz to 2.5 GHz with 50% modulation depth. The average incident optical power on the HBT was 840 $\mu$W. The up-converted signals were measured at $f_{LO} \pm f_{IF}$. The two up-converted signals had similar power level which only varied within ±3 dB over the frequency range. Biased at $V_{CE} = 1.6$ V, $V_{BE} = 0.65$ V and $I_C = 780 \mu$A, the system conversion gain was around -16 dB from $f_{IF}$ to 30 GHz ± $f_{IF}$.

An interesting configuration was reported by Sawada et al. [77,78] of a self-oscillating optoelectronic up-converter using an HBT with a 2 $\mu$m x 20 $\mu$m emitter area. The LO signal at 9.567 GHz was generated by the HBT oscillator itself whose frequency was determined by a dielectric resonator connected to the HBT base. It was not mentioned in their paper whether the LO frequency would be affected by different incident optical powers, resulting in an optically tuned oscillator. Insufficient information was provided for the system and intrinsic conversion gain calculations. Also no structural detail of the HBT used was given.

Their second configuration was a balanced self-oscillating HBT optoelectronic up-converter. The advantage is that it suppressed the LO power at the output by over 40 dB.
Similar to earlier work reported by Fetterman et al. [37] described in Section 1-4, Scott et al. [79,80] reported using an HBT to generate a 59.5 GHz signal when illuminated by two laser beams. The HBT had an AlInAs emitter (180 nm thick), a GaInAs base (80 nm), a GaInAs collector (270 nm) and a GaInAs sub-collector (800 nm). To provide optical access, an $8 \times 8 \, \mu m^2$ emitter window was included with an external quantum efficiency of 22%. The first laser emitted at 632.8 nm wavelength and the wavelength of the second laser was tuned from 600-640 nm in order to generate the desired beat frequency. The incident optical power from each of the lasers on the HBT was 0.6 mW. The achieved S/N was 45 dB in a 1 MHz resolution bandwidth.

Van de Casteele et al. [81] produced an electrical mixed product by photodetecting and mixing the intensity modulations of two incident optical beams in a two-terminal edge-coupled HBT. The layer structure of their HBT is shown in Figure 1-16.

![Figure 1-16: HBT layer structure by Van de Casteele et al.](image)

An 6 $\mu m \times 115 \, \mu m$ HBT was cleaved for measurements. Two optical sources, a 2 GHz directly modulated laser beam of 1.33 $\mu m$ wavelength and a 2 kHz square wave modulated laser beam of 1.319 $\mu m$ wavelength, were coupled to a singlemode optical fibre via a 3 dB coupler and then delivered to the HBT with average incident optical powers of 15 $\mu W$ and 50 $\mu W$. The mixed output was the two sidebands 10 dB below and at $\pm 2 \, kHz$ offset from the 2 GHz, -40 dBm output signal. The mixing was caused by the gain nonlinearities of the HBT at low input optical power rather than the beat frequency of the two optical sources as in the work by Scott et al. [79,80].
Section 1-9: Optically Controlled HBT Oscillators

As demonstrated by Sawada et al [77,78], a self-oscillating HBT can function equally well as an optoelectronic up-converter. The HBT photodetects the intensity modulated optical input and up-converts it by the LO frequency which is generated by the HBT itself. The frequency of the up-converted signal can be varied by tuning the HBT oscillation frequency. There are two ways to change the HBT oscillation frequency and up-conversion frequency optically when the HBT also is simultaneously used as an optoelectronic mixer or a microwave mixer. The first is called optical tuning [23] where the frequency of a free-running HBT oscillator can be changed by the level of the incident optical power. By changing the incident average optical power, the oscillation frequency can be varied. The other method is called optical injection locking [23]. In this method an optical source is intensity modulated at a frequency which in turn is used to lock the HBT oscillation frequency. When the locking frequency is close enough to the HBT free-running frequency, the HBT oscillator will be locked to the laser modulation frequency. By changing the locking frequency, the HBT oscillation frequency can also be changed and follow the locking frequency, and therefore the up-conversion frequency can be controlled remotely through an optical fibre. Optical injection locking of a MESFET self-oscillating microwave mixer has been demonstrated by Callaghan et al [82]. An injection locking range of 8.5 MHz at 6.3 GHz was achieved and a conversion gain of 1-2 dB for an IF between 50 and 200 MHz was obtained.

It is of interest to review the current status of the work on optically controlled HBT oscillators due to their application in optoelectronic mixing.

Bangert et al. [83] were the first to report an optically controlled HBT oscillator. Two standard GaInP/GaAs HBTs, connected in a differential amplifier configuration, were used to form a simple LC oscillator. One of the HBTs was illuminated with an 840 nm optical source above the 10 x 100 \( \mu \text{m}^2 \) emitter finger. In the optical tuning experiment, they achieved a frequency shift from 506 MHz to 481 MHz as the incident optical power was changed from 0 mW to 1 mW, an optical tuning range, \( \Delta f \), of around 25 MHz. In the injection locking experiment the average optical power used was 950 \( \mu \text{W} \) and a...
locking signal power of 7.9 mW was provided by a sweep generator to the laser. An optical injection locking range, $\Delta f_i$, of 15 MHz was achieved with the free-running oscillation frequency, $f_0$, at 500 MHz ($\Delta f_i / f_0 = 3\%$).

Karakucuk et al. [84] later reported a 2.6 GHz direct optically injection locked GaAs/AlGaAs HBT oscillator built on a 1 mm thick epoxy substrate using microstrip lines for impedance matching. A 30-period GaAs/AlGaAs multiquantum-well (MQW) was included between the base and the collector, and used to increase the breakdown voltage to 15 V and provided spectral tunability although none of the mentioned features were used in the experiments. The circular emitter contact had a 4 $\mu$m diameter hole for optical access to the MQW base-collector depletion region where absorption occurred.

With 40 mW (16 dBm) modulation power applied to the AlGaAs laser diode ($\lambda = 830$ nm), an optical injection locking range of around 6 MHz was achieved with the free-running oscillation frequency at 2.65 GHz, thus $\Delta f_i / f_0 = 0.2\%$. A 5 MHz optical tuning range was also obtained by changing the incident optical power. In both experiments the incident optical power was not specified. Karakucuk et al. [85] reported in their next paper another optically injection locked GaAs/AlGaAs HBT oscillator built on a 15 mil thick Duroid substrate using microstrip lines for impedance matching. In that work the emitter contact of the HBT was made of transparent Indium Tin Oxide (ITO) for improved optical coupling into the HBT. Light was focused onto the transparent ITO emitter contact of the HBT. With approximately 26 dBm modulation power applied to the AlGaAs laser diode ($\lambda = 830$ nm), an optical injection locking range of 2.5 MHz was obtained with the oscillator free-running frequency at 6 GHz ($\Delta f_i / f_0 = 0.04\%$). This small locking range was mainly due to the small optically generated RF power in the HBT which in turn was limited by small laser modulation depth. This RF power was reported to be 30 dB lower than the oscillator output. The optical tuning range was 25 MHz. No information about the incident optical power was given.

Freeman et al. [86] extended the frequency of oscillation with an optically controlled InAlAs/InGaAs HBT MMIC oscillator to the 14 GHz band with an integrated optical waveguide structure. The oscillator was constructed in a simple feedback configuration consisting of a planar rectangular inductors and thin film capacitors for the reactive feedback elements. The input optical signal of 1.55 $\mu$m wavelength was first guided
within an integrated optical waveguide running in parallel with and underneath the HBT sub-collector layer. The waveguide had no upper cladding vertically below the HBT so that the light bent upward and reached the fully depleted collector. The measured external quantum efficiency was 30%. The optical tuning experiment showed that a shift of -100 MHz from the free-running oscillation frequency of 13.9 GHz was achieved when the input optical power was changed from 0 μW to 200 μW. An optical injection locking range of only 0.5 MHz was recorded ($\Delta f / f_0 = 3.6 \times 10^{-3}$%) due to a very weak optically generated microwave locking signal power of -61 dBm in the HBT, compared to a free-running oscillator output power of -4 dBm.

It can be seen from the review above that optically injection locked oscillators operated at higher frequencies tend to have smaller fractional injection locking range and the reason is believed to be due to the low light coupling efficiency and inefficient direct modulation of the laser diode at such frequencies.

**Section 1-10: Aims and Novel Achievements of the Thesis**

The aims of this *Thesis* are as follows:

i. To investigate the performance of a two-terminal edge-coupled and a three-terminal normal incidence InGaAs/InP HBT optoelectronic mixer in terms of conversion gains, frequency responses, noise performance and intermodulation distortion characteristics;

ii. To compare the experimental results with other reported work;

iii. To model the two optoelectronic mixers in the electrically pumped configurations using harmonic-balance techniques in order to predict and understand better the conversion gain characteristics;

iv. To consider a number of different fibre-radio architectures employing an HBT optoelectronic mixer.
This *Thesis* contains a number of novel achievements, including

i. The first two-terminal edge-coupled InP/InGaAs HBT optoelectronic mixer, optoelectronic mixing conversion gain characteristics, noise performance characteristics and intermodulation distortion characteristics;

ii. Highest reported optoelectronic system conversion gain (defined in Section 1-2) of +7 dB (RF = 2.41 GHz, LO = 2.91 GHz, IF = 500 MHz) using the two-terminal HBT achieved for any single-transistor optoelectronic mixer;

iii. The first two-terminal and first three-terminal HBT electrically pumped optoelectronic mixer computer models using harmonic-balance techniques;

iv. The first experimental results on the intermodulation distortion characteristics of a three-terminal HBT optoelectronic mixer using a two-laser approach.

### Section 1-11: Conclusion and Organisation of the Thesis

This *Chapter* began with a brief introduction to optoelectronic mixing, describing and distinguishing between electrically pumped and optically pumped optoelectronic mixers, and between the conventional and the HBT approaches to these tasks. To facilitate comparisons with other reported work and characterise the performance of the HBT optoelectronic mixers in this *Thesis*, formal definitions for the system and intrinsic conversion gains were given which will be used throughout this *Thesis*. The applications of optoelectronic mixers in SCM and fibre-radio systems were described. The reason for the use of the heterojunction and the photodetection process in an HBT were qualitatively explained. Also a comprehensive review of prior important work on optoelectronic mixing using different semiconductor devices and techniques, HBT direct photodetectors, optoelectronic mixers and optically controlled oscillators was included.

A summary of some of the most important prior results on HBT optoelectronic mixers, optically controlled oscillators and direct photodetectors is given in Table 1-1. Although excellent results have been reported for HBT direct photodetectors, the achieved
optoelectronic mixing system conversion gains have been low (< 0 dB). There is therefore scope for further improvement in this area.

The organisation of this Thesis is as follows. Chapter 2 will examine the theoretical basis of the optical control of the HBT and other performance limitation issues. Chapter 3 will present the computer modelling and simulation results for the two-terminal and three-terminal HBT electrically pumped optoelectronic mixers using harmonic-balance techniques. Chapter 4 will present the experimental results for the two-terminal HBT optoelectronic mixer and Chapter 5 will present the experimental results for the three-terminal HBT optoelectronic mixer. In Chapter 6 a number of system architectures using HBT optoelectronic mixers in fibre-radio systems will be discussed. A conclusion and suggestions for further work will be given in Chapter 7. Finally the Appendix will give a list of publications resulting from this work.
Table 1-1: Summary of some of the most important prior results on HBT optoelectronic mixers, optically controlled oscillators and direct photodetectors.

<table>
<thead>
<tr>
<th>Configuration</th>
<th>Parameter</th>
<th>Value</th>
<th>Remark</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optoelectronic mixers</td>
<td>Optoelectronic mixing conversion</td>
<td>-21 dB system conversion gain with 35.7 dB S/N in 3 MHz res. B. W.</td>
<td>Down-conversion from 10 GHz to 600 MHz</td>
<td>Urey et al. [72]</td>
</tr>
<tr>
<td></td>
<td>gain</td>
<td>-23.04 dB system conversion gain with 30.8 dB S/N in 3 MHz res. B. W.</td>
<td>Down-conversion from 18 GHz to 600 MHz</td>
<td>Urey et al. [72]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>-4 dB intrinsic conversion gain</td>
<td>Up-conversion from 3.2 GHz to 30 GHz</td>
<td>Suematsu et al. [73,74]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>-14 dB intrinsic conversion gain</td>
<td>Up-conversion from 3.2 GHz to 30.8 GHz</td>
<td>Suematsu et al. [73,74]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>-16 dB system conversion gain</td>
<td>Up-conversion from IF to 30 GHz ± IF where IF is from 200 MHz to 2.5 GHz</td>
<td>Gonzalez et al [75,76]</td>
</tr>
<tr>
<td>Optical beat signal detection</td>
<td>mm-wave generation frequency</td>
<td>59.5 GHz with 45 dB S/N in 1 MHz res. B. W.</td>
<td>Mixing of two laser beams</td>
<td>Scott et al. [79,80]</td>
</tr>
<tr>
<td>Optically controlled oscillators</td>
<td>Free-running freq.: f₀, injection locking range: Δf₀, tuning range: Δf₀</td>
<td>f₀=500 MHz, Δf₀=15 MHz, Δf₀/Δf₀=3%</td>
<td>Differential amplifier configuration with 2 HBTs</td>
<td>Bangert et al. [83]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>f₀=2.65 GHz, Δf₀=6 MHz, Δf₀/Δf₀=0.2%</td>
<td>Multiquantum-well used between base and collector</td>
<td>Karakucuk et al. [84]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>f₀=6 GHz, Δf₀=2.5 MHz, Δf₀/Δf₀=0.04%</td>
<td>Transparent ITO emitter contact</td>
<td>Karakucuk et al. [85]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>f₀=13.9 GHz, Δf₀=0.5 MHz, Δf₀=100 MHz, Δf₀/Δf₀=3.6x10⁻³%</td>
<td>Integrated optical waveguide below the HBT</td>
<td>Freeman et al. [86]</td>
</tr>
<tr>
<td>HBT direct photodetectors</td>
<td>Optical gain &amp; responsivity</td>
<td>2500, 2620 A/W</td>
<td>Resonant cavity at 1.3 μm</td>
<td>Dodabalpur et al. [54]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2300, 1600 A/W</td>
<td>Guard ring structure, 860 nm</td>
<td>Twynam et al. [65]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>534, 400 A/W at 100 μW</td>
<td>Resonant cavity at 929 nm</td>
<td>Sjolund et al. [57]</td>
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<td></td>
<td></td>
<td>1000, 927 A/W at 5 μW</td>
<td>Back-illuminated InGaAs HBT</td>
<td>Campbell et al. [47,48]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1240, 850 A/W at 0.5 μW</td>
<td>Punchthrough HBT, 850 nm</td>
<td>Wang et al. [68]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>25000, 23185 A/W at 2.3 μW</td>
<td>Avalanche phototransistor at 1.15 μm, Highly stabilised bias voltage required.</td>
<td>Campbell et al. [51]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>147, 175 A/W at 0.16 mW to 0.18 mW</td>
<td>2-T edge-coupled HBT at 1.48 μm</td>
<td>Wake et al. [58]</td>
</tr>
<tr>
<td>External quantum efficiency</td>
<td></td>
<td>43 %</td>
<td>Resonant cavity at 0.9 μm</td>
<td>Unlu et al. [55]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>43 %</td>
<td>Resonant cavity at 860 nm</td>
<td>Huang et al. [56]</td>
</tr>
<tr>
<td>Common-emitter current gain</td>
<td></td>
<td>100 at 20 nW and 600 at 2.5 μW</td>
<td>2-T HBT</td>
<td>Campbell et al. [50]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>260 at 40 nW</td>
<td>Insertion of a low doped layer into emitter</td>
<td>Leu et al. [87,88,89]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>270 at 0.16 mW to 0.18 mW</td>
<td>2-T edge-coupled HBT</td>
<td>Wake et al. [58]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>400</td>
<td>Guard ring structure, 860 nm</td>
<td>Twynam et al. [65]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1400 at 20 mA Iᵣ</td>
<td>First 3-T HBT</td>
<td>Fritzscache et al. [66]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>5000</td>
<td>Resonant cavity at 1.3 μm</td>
<td>Dodabalpur et al. [54]</td>
</tr>
<tr>
<td>Unity optical gain frequency</td>
<td></td>
<td>22 GHz at 1.3 μm, 14 GHz at 1.55 μm</td>
<td>Reflector. Base terminated in 50 Ω</td>
<td>Fukano et al. [64]</td>
</tr>
<tr>
<td></td>
<td></td>
<td>10 GHz at Iᵣ=15 μA, 1 μW, 1.3 μm</td>
<td>3-T HBT</td>
<td>Chandrasekhar et al. [67]</td>
</tr>
<tr>
<td>Common collector unity gain frequency</td>
<td></td>
<td>30 GHz compared to a PIN photodiode in identical package</td>
<td>2-T edge-coupled HBT at 1.3 μm, Iᵣ=8 mA</td>
<td>Wake et al. [58]</td>
</tr>
</tbody>
</table>

*Resolution bandwidth.
References


[19] New Focus GaAs/Schottky photodiode, model 1002.

Chapter 1


Chapter 1


[57] O. Sjolund, M. Ghisoni and A. Larsson, "High gain resonant cavity enhanced InGaAs/AlGaAs heterojunction phototransistor resonant at 930 nm," *Electronics Letters*, vol. 31, no. 11, pp. 917-918, May 1995.


Chapter 1


Chapter Two

Optical Control of Heterojunction Bipolar Transistors
and Optoelectronic Mixing

Section 2-1: Introduction

The analysis of the optical control of heterojunction bipolar transistors (HBTs) can be divided into two separate parts. The first part involves analysing the way the primary photocurrent is generated when the HBT is under optical illumination. The second part is to analyse how the HBT responds to the primary photocurrent so generated. Therefore it is possible to use the equivalent circuit in Figure 2-1 to analyse an illuminated HBT.

![Figure 2-1: An equivalent circuit representation of an HBT under illumination. PD: photodiode.](image)

In Figure 2-1, the PD models the illuminated base-collector junction of the HBT.

In this Chapter, the general theory of the optical control of HBTs will be described. Specifically in Section 2-2, the optical absorption processes in both the normal incidence and edge-coupled HBTs will be examined, resulting in the primary photocurrent expressions for the two types of structures. Then the discussion of the HBT will be carried out purely in the electrical domain. In Section 2-3 the static operation of the bipolar transistor will be summarised. In Section 2-4, some commonly used figures of
merit for the bipolar transistor in DC operation will be derived from the transistor current-voltage equations, showing how the transistor structural parameters affect the DC operation performance. In Section 2-5 the use of a heterojunction in an HBT will be examined in detail. In Section 2-6, the effect of the base-emitter junction space-charge recombination current on the transistor common-emitter current gain will be explained and illustrated with some measured results. In Section 2-7, the four time constants which affect a transistor current gain cutoff frequency \( f_T \) will be discussed in turn, leading to a simplified derivation of the equation for \( f_T \). In Section 2-8, the mechanisms responsible for the mixing in HBTs will be described. Finally a conclusion is given in Section 2-9.

**Section 2-2: Optical Absorption in HBTs**

Both three-terminal normal-incidence and two-terminal edge-coupled InP/InGaAs HBTs have been used in optoelectronic mixing in this work. In the following two sub-sections, the optical absorption processes in these two types of HBTs will be considered and their respective primary photocurrents derived.

**Section 2-2a: Optical Absorption in a Three-Terminal InP/InGaAs Normal Incidence HBT**

The cross-section of the three-terminal InP/InGaAs normal-incidence HBT is shown in Figure 2-2. Details of the device fabrication will be given in Chapter 5.

![Figure 2-2: Cross-section of the normal-incidence HBT. Emitter area is 4 x 11 \( \mu \text{m}^2 \) and the base area is 9 x 23 \( \mu \text{m}^2 \). Optical window area on the base mesa is 5 x 6 \( \mu \text{m}^2 \).](image)
In Figure 2-2 the light is incident on the optical window of the HBT located on the base. The direction of the light is normal to the epi-layers of the device. The base and collector are made of InGaAs which has an energy bandgap of 0.75 eV. The wavelength used in the experiments was around 1550 nm, corresponding to 0.80 eV photon energy. Therefore as the light travels through the base and the collector, the photon flux density (no. of photons per unit area per second) starts to decrease exponentially because of the absorption by the base and the collector. Taking the positive y-axis as the downward direction as shown in Figure 2-3,

\[ \Phi(y) = \Phi_0 e^{-\alpha y} \]  \hspace{1cm} \text{Eqn 2-1}

where \( \alpha \) is the absorption coefficient which is approximately 10,000 cm\(^{-1}\) or 1 \( \mu m\)^{-1} for InGaAs at around 1550 nm wavelength [2], \( \Phi_0 \) is the photon flux density at the edge of the base (\( y = 0 \)) just below the air-base interface and is given by

\[ \Phi_0 = \frac{\lambda P_{ap}(1 - R_{ref})}{A_{ap}hc} \]  \hspace{1cm} \text{Eqn 2-2}
where $R_{\text{ref}}$ is the air-base reflectance, $A_{\text{opt}}$ is area of the optical window on the base mesa, $h$ is the Planck's constant, $c$ is the speed of light, $\lambda$ is the wavelength of light and $P_{\text{opt}}$ is the amount of optical power incident on and within the optical window. Assuming each photon absorbed generates an electron and a hole, the electron-hole pair generation rate per unit volume as a function of $y$ is then given by

$$G(y) = \alpha \Phi_0 e^{-\alpha y} \quad \text{Eqn 2-3}$$

As seen in Figure 2-3, photon absorption and hence electron-hole pair generation occur in three regions, which are the neutral base, the base-collection depletion and the neutral collector regions. As a result, the primary photocurrent $I_{\text{prim}}$ is given by the sum of three different current components, depending on where the electron-hole pairs are generated.

The electron-hole pairs generated in the base-collector depletion region are immediately separated by the electric field, forming the drift current component $I_{\text{dr}}$ of $I_{\text{prim}}$. Assuming no recombination takes place in the reverse-biased base-collector junction depletion region and all the photogenerated electrons and holes are swept out of the depletion region, $I_{\text{dr}}$ is then given by

$$I_{\text{dr}} = q A_{\text{opt}} \int_{w_a}^{w_a + w_{\text{bc}}} G(y) \cdot dy \quad \text{Eqn 2-4}$$

$$= q A_{\text{opt}} \Phi_0 \exp(-\alpha W_B)[1 - \exp(-\alpha W_{\text{bc}})] \quad \text{Eqn 2-5}$$

where $q$ is the electron charge.

In normal operation, the base-collector junction is reverse biased. Therefore as in any reverse biased pn junction, the minority carrier concentrations at the edges of that junction are approximately zero. That is

$$n_b(y = W_B) = 0 \quad \text{Eqn 2-6}$$

$$p_c(y = W_B + W_{\text{bc}}) = 0 \quad \text{Eqn 2-7}$$

where $n_b$ and $p_c$ are the electron and hole concentrations in the base and collector, respectively. The above two boundary conditions give rise to minority concentration gradients which cause the electrons in the neutral base and holes in the neutral collector to diffuse into the depletion region.
In the steady state, the minority electron concentration in the neutral base is governed by the following diffusion equation

\[ \frac{d^2 n_b(y)}{dy^2} - \frac{n_b(y) - n_{ob}}{L_{nb}^2} + \frac{G(y) A_{opt}}{D_{nb} A} = 0 \]  

Eqn 2-8

where \( L_{nb} \) is the electron diffusion length, \( D_{nb} \) is the electron diffusion coefficient, \( A \) is the base and collector area which is larger than \( A_{opt} \), \( n_b(y) \) is the electron concentration at position \( y \) and \( n_{ob} \) is the electron concentration at thermal-equilibrium. The subscripts 'b' on these variables state that these are quantities in the base. In practice, measuring the primary photocurrent in an HBT requires that the base-emitter junction voltage \( V_{BE} \) be set at 0 V so that no electron is injected from the emitter side. Therefore for \( V_{BE} = 0 \) V the electron concentration at \( y = 0 \) is equal to its thermal-equilibrium value, i.e.,

\[ n_b(y = 0) = n_{ob} \]  

Eqn 2-9

With the boundary conditions given in Eqn 2-6 and Eqn 2-9, the electron concentration profile in the neutral base is found to be

\[ n_b(y) = C \frac{\sinh[(W_b - y)/L_{nb}]}{\sinh[W_b/L_{nb}]} + \left[ C \exp(-\alpha W_b) - n_{ob} \right] \frac{\sinh[y/L_{nb}]}{\sinh[W_b/L_{nb}]} - C \exp(-\alpha y) + n_{ob} \]  

Eqn 2-10

where \( C = \frac{A_{opt}}{A} \frac{\Phi_0 aL_{nb}^2}{D_{nb} (\alpha^2 L_{nb}^2 - 1)} \).

The diffusion current due to the electrons diffusing from the neutral base to the base-collector depletion region is given by

\[ I_{nb}(y = W_b) = -qAD_{nb} \frac{dn_b(y)}{dy} \bigg|_{y=W_b} \]  

Eqn 2-11

or

\[ I_{nb}(y = W_b) = -\frac{qAD_{nb}}{L_{nb}} \left[ C\alpha L_{nb} \exp(-\alpha W_b) - C \text{csch} \left( \frac{W_b}{L_{nb}} \right) \right] \]

\[ + \left[ C \exp(-\alpha W_b) - n_{ob} \right] \text{coth} \left( \frac{W_b}{L_{nb}} \right) \]  

Eqn 2-12
Note that the base width $W_g$ is very small compared to both the electron diffusion length $L_{nb} \approx 2 \mu$m in InGaAs [3]) and the reciprocal of the absorption coefficient $\alpha^{-1}$, and also that $\alpha^2 L_{nb}^2 > 1$, therefore Eqn 2-12 can be simplified to

$$I_{nb}(y = W_g) = qA_{opt} \Phi_0 [1 - \exp(-\alpha W_g)] + \frac{qAD_{nb}n_{nb}}{W_g}$$  \hspace{1cm} \text{Eqn 2-13}$$

Examining Eqn 2-13 reveals that because of the absence of $\Phi_0$, the second term is the HBT dark current due to the thermally generated electrons in the base with $V_{BE} = 0$ V and the first term is photogenerated. Comparing Eqn 2-13 with Eqn 2-5 also shows that for a thin-base normal incidence HBT, all the photogenerated electrons in the base diffuse to the base-collector depletion and therefore contribute to the total primary photocurrent. One can argue that such a simple equation (Eqn 2-13) could have been promptly obtained, without lengthy analysis, simply by assuming that all the photogenerated carriers in a very thin base must be able to diffuse to the base-collector depletion region and contribute to the primary photocurrent. However such an assumption, although correct for a top-illuminated HBT, is wrong for the edge-coupled HBT where the direction of incident photons is perpendicular to that of the current flow and even for a thin-base edge-coupled HBT, only half of the photogenerated carriers in the neutral base can diffuse to the base-collector depletion, as will be demonstrated in Section 2-2b.

The treatment for the photogenerated holes in the neutral collector is similar to that for the electrons in the neutral base. The diffusion equation for the holes in the neutral collector is

$$\frac{d^2 p_c(y)}{dy^2} - \frac{p_c(y) - p_{eq}}{L_{pc}^2} + \frac{G(y) A_{opt}}{D_{pc} A} = 0$$ \hspace{1cm} \text{Eqn 2-14}$$

where $L_{pc}$ is the hole diffusion length, $D_{pc}$ is the hole diffusion coefficient, $p_c(y)$ is the hole concentration at position $y$ and $p_{eq}$ is the hole concentration at thermal-equilibrium. Assume that the thickness of the neutral collector is much greater than both the hole diffusion length and the reciprocal of the absorption coefficient $\alpha^{-1}$ such that at large $y$, the hole concentration is equal to its thermal-equilibrium value, i.e.,

$$p_c(y = \infty) = p_{eq}$$ \hspace{1cm} \text{Eqn 2-15}$$

Using Eqn 2-15 and Eqn 2-7 as the boundary conditions, the hole concentration profile is found to be
The hole diffusion current in the neutral collector at the base-collector junction can be evaluated as

\[ I_{pc}(y = W_B + W_{BC}) = qAD_{pc} \frac{dp_c(y)}{dy} \bigg|_{y=W_B+W_{BC}} \quad \text{Eqn 2-17} \]

or

\[ I_{pc}(y = W_B + W_{BC}) = qA_{opt} \Phi_0 \left\{ \frac{\alpha L_{pc}}{\alpha L_{pc} + 1} \exp(-\alpha(W_B + W_{BC})) \right\} + \frac{qAD_{pc}p_{oc}}{L_{pc}} \quad \text{Eqn 2-18} \]

Because of the absence of \( \Phi_0 \), the second term in Eqn 2-18 is the HBT dark current due to the thermally generated holes in the neutral collector and the first term is photogenerated.

Finally the collector current is equal to the total current crossing the base-collector junction which is given by the sum of \( I_{dr} \) in Eqn 2-5, \( I_{nb}(y = W_B) \) in Eqn 2-13 and \( I_{pc}(y = W_B + W_{BC}) \) in Eqn 2-18. Therefore for \( V_{BE} = 0 \) V, the collector current is

\[
I_{CE}(V_{BE} = 0 \text{ V}) = I_{dr} + I_{nb}(y = W_B) + I_{pc}(y = W_B + W_{BC})
= qA_{opt} \Phi_0 \exp(-\alpha W_B) [1 - \exp(-\alpha W_{BC})]
+ qA_{opt} \Phi_0 [1 - \exp(-\alpha W_B)]
+ qA_{opt} \Phi_0 \left\{ \frac{\alpha L_{pc}}{\alpha L_{pc} + 1} \exp(-\alpha W_B - W_{BC}) \right\}
+ qA \left( \frac{D_{nb}n_{oh}}{W_B} + \frac{D_{pc}p_{oc}}{L_{pc}} \right) \quad \text{Eqn 2-19}
\]

which can be further simplified to
In Eqn 2-20 the primary photocurrent can therefore be identified as

\[
I_{\text{prim}} = q A_{\text{opt}} \Phi_0 \left\{ 1 - \frac{\exp \left[ -\alpha (W_B + W_{BC}) \right]}{\alpha L_{pc} + 1} \right\} + qA \left( \frac{D_{ab} n_{ab}}{W_B} + \frac{D_{pc} P_{oc}}{I_{pc}} \right)
\]

Eqn 2-21

which represents the magnitude of the primary photocurrent flowing through the photodiode in Figure 2-1 for a normal incidence HBT. \( I_{\text{dark}} \) is the HBT dark current for a short-circuited base-emitter junction.

Having determined the primary photocurrent, the analysis of the HBT under illumination can proceed, treating the HBT purely as an electrical transistor with the optical illumination accounted for by including a primary photocurrent generator between the base and the collector.

The next section will consider the two-terminal edge-coupled HBT.

Section 2-2b: Optical Absorption in a Two-Terminal InP/InGaAs Edge-Coupled HBT

The cross-section of the two-terminal edge-coupled InP/InGaAs HBT is shown in Figure 2-4.
In this edge-coupled HBT, light is incident from one side in parallel with the device epitaxial layer. The photon flux density decreases exponentially in the $x$ direction (Figure 2-5) due to the absorption in the base and collector regions. Along the $y$ direction in which the current flows, the photon flux density is constant. Assuming that light is uniformly incident on the base and collector, and exits the device after a length of $d$, which is the device lateral dimension, the average electron-hole pair generation rate per unit volume along the $y$ direction is given by

$$G = \frac{\Phi_0 (1 - e^{-\alpha d})}{d} \quad \text{Eqn 2-22}$$

This electron-hole pair generation rate per unit volume is constant in both the base and the collector for a given point along the $x$ axis.

The drift current due to the photogenerated electron-hole pairs in the base-collector depletion region is given by

$$I_{dr} = qA \int_{w_{b}}^{W_{b} + W_{BC}} G \cdot dy \quad \text{Eqn 2-23}$$

Substituting the $y$-independent electron-hole pair generation rate per unit volume from Eqn 2-22 into Eqn 2-23 gives

$$I_{dr} = \frac{qA \Phi_0 W_{BC}}{d} (1 - \exp(-\alpha d)) \quad \text{Eqn 2-24}$$
To find the electron diffusion current component of the primary photocurrent in the neutral base, the following electron diffusion equation is solved to obtain the electron concentration profile \( n_b(y) \).

\[
\frac{d^2 n_b(y)}{dy^2} - \frac{n_b(y) - n_{ob}}{L_{nb}^2} + \frac{G}{D_{nb}} = 0
\]  
Eqn 2-25

Similar to the three-terminal HBT case, the boundary conditions are

\[
n_b(y = 0) = n_{ob} \quad \text{Eqn 2-26}
\]

\[
n_b(y = W_B) = 0 \quad \text{Eqn 2-27}
\]

The solution to Eqn 2-25 is therefore

\[
n_b(y) = \frac{-\left( n_{ob} + \frac{L_{nb}^2 G}{D_{nb}} \right)}{\sinh \left( \frac{W_B}{L_{nb}} \right)} \sinh \left( \frac{y}{L_{nb}} \right) + \frac{L_{nb}^2 G}{D_{nb} \sinh \left( \frac{W_B}{L_{nb}} \right)} \sinh \left[ \frac{y - W_B}{L_{nb}} \right] + n_{ob} + \frac{L_{nb}^2 G}{D_{nb}}
\]

Eqn 2-28

The electron current diffusing into the base-collector depletion region can be obtained by evaluating

\[
I_{nb}(y = W_B) = -qAD_{nb} \frac{dn_b(y)}{dy} \bigg|_{y=W_B}
\]

Eqn 2-29

which, after substituting Eqn 2-28, gives

\[
I_{nb}(y = W_B) = \frac{qAD_{nb} n_{ob} \coth \left( \frac{W_B}{L_{nb}} \right)}{L_{nb}} + qA W_B G \left[ \frac{\cosh \left( \frac{W_B}{L_{nb}} \right) - 1}{\left( \frac{W_B}{L_{nb}} \right) \sinh \left( \frac{W_B}{L_{nb}} \right)} \right]
\]

Eqn 2-30

The first term on the right in Eqn 2-30 is the HBT dark current due to the thermally generated electrons in the neutral base. The second term in Eqn 2-30 is due to the photogenerated electrons diffusing into the depletion. For a very thin base in which \( W_B / L_{nb} \approx 0 \), the term within the square brackets approaches 1/2. Therefore

\[
I_{nb}(y = W_B) \approx \frac{qAD_{nb} n_{ob}}{W_B} + \frac{qA W_B G}{2}
\]

Eqn 2-31

\[
\approx \frac{qAD_{nb} n_{ob}}{W_B} + \frac{qA \Phi \chi W_B}{2d} [1 - \exp(-\alpha d)]
\]

Eqn 2-32
Eqn 2-31 reveals that for a thin-base edge-coupled HBT in which the direction of the incident light is perpendicular to the current flow, only half of the photogenerated electrons in the neutral base diffuse into the base-collector depletion region and contribute to the total primary photocurrent. This reduction in photogenerated electron diffusion current compared to that for a normal incidence HBT is due to the fact that in a normal incident HBT, there are two factors which cause the electrons to diffuse to the depletion region. The first is the zero electron concentration condition \( n_b(y=W_b)=0 \) imposed by the reverse biased base-collector junction. The second is the exponential decrease in the electron-hole pair generation rate as the photons are absorbed along the \( y \) direction. The decreasing electron-hole pair generation gives rise to an additional concentration gradient and encourages the electrons to diffuse towards the base-collector depletion. Although the first factor mentioned applies to both types of transistors, the second factor does not help electrons diffuse towards the depletion region in the edge-coupled HBT because the concentration gradient due to the exponentially decreasing electron-hole pair generation rate occurs along the \( x \) direction which is orthogonal to the direction of the HBT current flow.

The hole concentration profile in the neutral collector region can be obtained by solving the following diffusion equation

\[
\frac{d^2 p_c(y)}{dy^2} - \frac{p_c(y) - p_{uc}}{L_{pc}^2} + \frac{G}{D_{pc}} = 0
\]

Eqn 2-33

The boundary conditions are

\[
p_c(y=W_a+W_{bc})=0 \tag{Eqn 2-34}
\]

\[
p_c(y=\infty)=\text{finite} \tag{Eqn 2-35}
\]

The boundary condition in Eqn 2-35 needs some explanation. It can be seen by inspection that the solution to Eqn 2-33 is of the form

\[
p_c(y) - p_{uc} = E_1 \exp\left(\frac{y}{L_{pc}}\right) + E_2 \exp\left(-\frac{y}{L_{pc}}\right) + E_3
\]

Eqn 2-36

where \( E_1, E_2 \) and \( E_3 \) are constants to be determined. After substituting Eqn 2-36 back Eqn 2-33, \( E_3 \) is found to be
Although it is not apparent at this stage what value \( p_c(y = \infty) \) will take, it is obvious that \( p_c(y = \infty) \) cannot be equal to infinity, therefore \( E_1 \) in Eqn 2-36 must be zero. The remaining constant \( E_2 \) can be determined from Eqn 2-34. The solution to Eqn 2-33 is found to be

\[
p_c(y) = -\left( p_{oc} + \frac{L_{pc}^2 G}{D_{pc}} \right) \exp\left[ \frac{(W_B + W_{BC} - y) - L_{pc}}{L_{pc}} \right] + \left( p_{oc} + \frac{L_{pc}^2 G}{D_{pc}} \right)
\]

Eqn 2-38

Now from Eqn 2-38, Eqn 2-35 can be rewritten as

\[
p_c(y = \infty) = p_{oc} + \frac{L_{pc}^2 G}{D_{pc}}
\]

Eqn 2-39

The current due to the holes in the neutral collector diffusing into the base-collector depletion region can be obtained by evaluating

\[
I_{pc}(y = W_B + W_{BC}) = qAD_{pc} \frac{dp_c(y)}{dy} \bigg|_{y=W_B+W_{BC}}
\]

Eqn 2-40

which, after substituting Eqn 2-38, gives

\[
I_{pc}(y = W_B + W_{BC}) = \frac{qAD_{pc} p_{oc}}{L_{pc}} + qAL_{pc} G
\]

\[
= \frac{qAD_{pc} p_{oc}}{L_{pc}} + \frac{qA\Phi_0 L_{pc}}{d} [1 - \exp(-\alpha d)]
\]

Eqn 2-41

Eqn 2-42

Eqn 2-41 shows that only those holes photogenerated within a distance \( L_{pc} \) from the depletion can contribute to the primary photocurrent. The first term in Eqn 2-42 is the dark current due to the thermally generated holes in the neutral collector.

The collector current is equal to the total current crossing the base-collector junction which is given by the sum of \( I_{dr} \) in Eqn 2-24, \( I_{nb}(y = W_B) \) in Eqn 2-32 and \( I_{pc}(y = W_B + W_{BC}) \) in Eqn 2-42. Therefore for \( V_{BE} = 0 \) V, the collector current is
\[ I_c(V_{BE} = 0 \text{V}) = I_{dr} + I_{nb}(y = W_b) + I_{pc}(y = W_b + W_{BC}) \]

\[ = \frac{qA\Phi_0 W_{BC}}{d}(1 - \exp(-\alpha d)) + \frac{qA\Phi_0 W_B}{2d}(1 - \exp(-\alpha d)) + \frac{qA\Phi_0 L_{pe}}{d}(1 - \exp(-\alpha d)) \]

\[ + \left( \frac{qA D_{nh} n_{sh}}{W_b} + \frac{qA D_{pe} P_{soc}}{L_{pc}} \right) \]

Eqn 2-43

which further simplifies to

\[ I_c(V_{BE} = 0 \text{V}) = \frac{qA\Phi_0}{d} \left( W_{BC} + \frac{W_B}{2} + L_{pe} \right)(1 - \exp(-\alpha d)) + \frac{qA}{W_b} \left( \frac{D_{nh} n_{sh}}{2} + \frac{D_{pc} P_{soc}}{L_{pc}} \right) \]

\[ = I_{\text{prim}} + I_{\text{dark}} \]

Eqn 2-44

In Eqn 2-44, the primary photocurrent can therefore be identified as

\[ I_{\text{prim}} = \frac{qA\Phi_0}{d} \left( W_{BC} + \frac{W_B}{2} + L_{pe} \right)(1 - \exp(-\alpha d)) \]

Eqn 2-45

which represents the magnitude of the primary photocurrent flowing through the photodiode in Figure 2-1 for an edge-coupled HBT. \( I_{\text{dark}} \) is the HBT dark current for a short-circuited base-emitter junction.

In summary, the optical absorption and electron-hole pair generation processes in both normal incidence and edge-coupled HBTs have been considered and the primary photocurrents derived in this section. In a three-terminal HBT, the primary photocurrent can be experimentally measured by short-circuiting the base-emitter junction and reverse biasing the base-collector junction, and the primary photocurrent is then equal to the collector current for a given incident optical power. In a two-terminal HBT, however, the base-emitter junction cannot be short-circuited because of the absence of an electrical base terminal and the photocurrent measured is due to both the primary photocurrent and the transistor current gain. An indirect way of measuring the primary photocurrent in a two-terminal edge-coupled HBT is to fabricate an edge-coupled pn photodiode whose structure is identical to the base-collector section of the two-terminal HBT. The photocurrent of this pn photodiode illuminated with the same optical power can then be taken as an approximation for the primary photocurrent for the two-terminal HBT.
The concept of the primary photocurrent has been used to reduce the problem of fully analysing a phototransistor into analysing a normal electrical transistor with a photodiode connected between the base and the collector. Since the primary photocurrent has been considered, the following sections will proceed to consider the HBT as a purely electrical device.

**Section 2-3: Summary of Static Bipolar Transistor Operation**

The HBT can be regarded as a successor to the mature BJT technology. To appreciate the advantages of having a wide bandgap emitter in an HBT, it is useful to review the basic operation principle of a BJT first.

A BJT, in its simplest form, consists of two pn junctions made of the same semiconductor material. In the npn-type BJT, the two junctions share a very thin p-type region known as the base. The other two n-type regions are the emitter and the collector. Figure 2-6 shows an idealised one-dimensional BJT structure of the npn-type. This structure can be used to explain the basic principles of BJTs.
In most analogue applications, the BJT is operated with the base-emitter junction forward-biased while the base-collector junction is reverse-biased, and this is known as the forward-active mode of operation. With the base-emitter junction forward-biased, a large current flows across this junction which consists of the flow of electrons injected from the emitter into the base and the flow of holes back injected from the base into the emitter as shown in Figure 2-6. To achieve a current gain, the number of holes back injected into the emitter should be much smaller than the number of electrons injected into the base. This is achieved, in a homojunction BJT, by doping the emitter much more heavily relative to the base so that at the base-emitter junction, the electron current component is much larger than the hole current component. The electron concentration in the base is now above its thermal-equilibrium value because of the injection of the electrons from the emitter. As a result the injected electrons and the majority holes in the base tend to recombine in order to restore equilibrium. To conserve this large injected electron current and hence the transistor current gain, all bipolar transistors are designed to have very thin bases so that most of the injected electrons from the emitter can diffuse through the base and reach the reverse-biased base-collector junction with only little recombination with the majority holes. Once arrived at the edge of the base-collector junction, the electrons are then swept across the depletion region by the electric field. Those electrons that have reached the edge of the base-collector junction form the major part of the collector current. Since the base-collector junction is also reverse-biased, a small reverse saturation current flows across that junction and forms the remaining part of the collector current which is not represented in Figure 2-6.

It is important to note that those holes that have been back injected into the emitter and those that have recombined with the traversing electrons in the base have to be replenished by providing an external base current. In a well-designed BJT where the base is both very thin and lightly doped relative to the emitter, this base current is very small. When a BJT is used as a current amplifier, this small base current acts as the input current and the amplified output current can be taken from the collector or the emitter.

The operation of a BJT has been described qualitatively in the section. In the next section, the current-voltage characteristics will be expressed analytically so that the common-base current gain, the emitter efficiency, the base transport factor and the
common-emitter current gain can be related to the BJT structural parameters and doping levels.

Section 2-4: Current Flows in an Ideal Transistor and Figures of Merit for Static Operation

To understand how the terminal behaviour of the BJT is related to its structure, it is useful to describe the emitter and collector currents analytically in terms of the BJT structural parameters and junction voltages. The assumptions used in the derivation for an ideal BJT are

(a) no generation-recombination currents exist in the two junction depletion regions;
(b) there are no series resistances in the BJT;
(c) each of the three regions of the BJT is uniformly doped. Graded-base BJT is not being considered;
(d) the injection level is low, i.e. in the base the injected minority electron concentration is small compared to the majority hole concentration there;
(e) the thicknesses of the emitter and collector are long compared to the minority carrier diffusion lengths;
(f) the current flows are governed by the drift-diffusion model.

Detailed derivation can be found in many semiconductor device textbooks (e.g. [2]) and the results are quoted below without derivation. The emitter and collector currents of a one-dimensional idealised npn BJT shown in Figure 2-6 are

\[
I_E = Aq \left[ \frac{D_{nb} n_{ob}}{L_{nb}} \coth \left( \frac{W_B}{L_{nb}} \right) + \frac{D_{pe} P_{oe}}{L_{pe}} \right] \exp \left( \frac{qV_{BE}}{kT} \right) - 1 \right] - \frac{AqD_{nb} n_{ob}}{L_{nb} \sinh \left( \frac{W_B}{L_{nb}} \right)} \exp \left( \frac{qV_{BC}}{kT} \right) - 1 \right]
\]

Eqn 2-46

and
where \( k \) is Boltzmann constant, \( T \) is the temperature in Kelvin, \( A \) is the device area, \( D_{pe} \) is the hole diffusion coefficient in the emitter, \( p_{oe} \) is the thermal-equilibrium hole concentration in the emitter and \( L_{pe} \) is the hole diffusion length in the emitter. \( V_{BE} \) and \( V_{BC} \) are the base-emitter and base-collector junction voltages, respectively. Eqn 2-46 and Eqn 2-47 show that in the forward-active mode, both the emitter and collector currents are an exponential function of \( V_{BE} \). For \( V_{BC} = 0 \, \text{V} \), Eqn 2-46 and Eqn 2-47 also reveal that the emitter current consists of a hole component identified by the term containing \( p_{oe} \), and an electron component identified by the term containing \( n_{ab} \). The collector current, however, consists of an electron component only, identified by the term containing \( n_{ab} \). The base current is simply given by

\[
I_B = I_E - I_C
\]

Eqn 2-48

By examining Eqn 2-46 and Eqn 2-47 further, it can be seen that what make a BJT different from two separate pn junction diodes connected back-to-back are the terms \( \sinh\left(\frac{W_B}{L_{nb}}\right) \) and \( \coth\left(\frac{W_B}{L_{nb}}\right) \). In a BJT the base is so thin that most of the minority carriers injected from one junction can diffuse through the base and reach the other junction. If the base width was very thick, i.e. \( W_B/L_{nb} \rightarrow \infty \), then \( \coth\left(\frac{W_B}{L_{nb}}\right) \rightarrow 1 \), \( \sinh\left(\frac{W_B}{L_{nb}}\right) \rightarrow \infty \), and Eqn 2-46 and Eqn 2-47 would simply reduce to those for two isolated pn junction diodes with no transistor action at all.

Some commonly used figures of merit for static operation can also be derived from Eqn 2-46 and Eqn 2-47. Assuming the BJT is operated with \( V_{BE} > 0 \, \text{V} \) and \( V_{BC} = 0 \, \text{V} \), the common-base current gain \( \alpha_o \) for an npn BJT is defined as

\[
\alpha_o = \frac{\partial I_C}{\partial I_E} = \frac{\partial I_{nb}(0)}{\partial I_E} \frac{\partial I_{nb}(W_B)}{\partial I_E} \frac{\partial I_C}{\partial I_{nb}(W_B)}
\]

Eqn 2-49
where $I_{nb}(0)$ and $I_{nb}(W_B)$ are the injected electron currents at the emitter side of the base and at the collector side of the base, respectively. The first term on the right hand side (r.h.s.) of Eqn 2-49 is the emitter efficiency $\gamma$ which is the ratio of the incremental electron current injected from the emitter to the total incremental emitter current:

$$
\gamma = \frac{\partial I_{nb}(0)/\partial V_{BE}}{\partial I_{E}/\partial V_{BE}} = \frac{Aq D_{nb} n_{ob} \coth \left( \frac{W_B}{L_{nb}} \right)}{L_{nb}} Aq \left[ \frac{D_{nb} n_{ob} \coth \left( \frac{W_B}{L_{nb}} \right)}{L_{nb}} + \frac{D_{pe} p_{oe}}{L_{pe}} \right]
$$

$$
= \frac{1}{1 + \frac{D_{pe} p_{oe}}{D_{nb} n_{ob}} \frac{L_{nb}}{L_{pe}} \tanh \left( \frac{W_B}{L_{nb}} \right)}
$$

Eqn 2-50

From the mass-action law [1] and assuming all donors and acceptors are ionised, then $p_{oe} = n_i^2 / N_{De}$ and $n_{ob} = n_i^2 / N_{Ab}$ where $n_i$ is the intrinsic carrier concentration, $N_{De}$ and $N_{Ab}$ are the emitter donor and base acceptor concentrations, respectively.

Since also $W_B << L_{nb}$, the emitter efficiency can be approximated as

$$
\gamma = \frac{1}{1 + \frac{D_{pe} N_{Ab} W_B}{D_{nb} N_{De} L_{pe}}}
$$

Eqn 2-51

Eqn 2-51 clearly shows that to maintain the emitter efficiency of a BJT close to unity, the emitter must be more heavily doped than the base.

In a BJT, the intrinsic carrier concentration $n_i$ is the same for the emitter, the base and the collector. In an HBT in which the emitter is made of a wide energy bandgap material, the intrinsic carrier concentration for the emitter is different from those for the base and the collector. As a result, the mass-action law has to be applied to the emitter and the base separately, using different intrinsic carrier concentrations. The effect of this on the emitter efficiency will be considered in the next section.
The second term on the r.h.s. of Eqn 2-49 is the base transport factor $\alpha_T$ which is the ratio of the incremental injected electron current reaching the collector to the incremental electron current injected from the emitter. From Eqn 2-46 and Eqn 2-47

$$\alpha_T = \frac{\partial I_{nb}(W_B)}{\partial V_{BE}} / \frac{\partial I_{nb}(0)}{\partial V_{BE}} = \frac{AqD_{nb}n_{nb}}{L_{nb}} \sinh\left(W_B / L_{nb}\right) \left[ \frac{AqD_{nb}n_{nb}}{L_{nb}} \coth\left(W_B / L_{nb}\right) \right]$$

Eqn 2-52

Eqn 2-52 shows that the base transport factor can be made very close to unity if the base width is much smaller than the electron diffusion length in the base so that most of the electrons injected from the emitter can diffuse to the collector.

The third term on the r.h.s. of Eqn 2-49 is the collector multiplication factor $M$ which is given by

$$M = \frac{\partial I_C}{\partial I_{nb}(W_B)}$$

Eqn 2-53

For a base-collector junction voltage well below the avalanche breakdown voltage, $M \equiv 1$ and the common-base current gain becomes

$$\alpha_o \equiv \gamma \alpha_T$$

Eqn 2-54

Finally the common-emitter current gain $h_{fe}$, defined as the ratio of the incremental collector current to the incremental base current, can be derived from $\alpha_o$, $\gamma$, and $\alpha_T$:

$$h_{fe} = \frac{\partial I_C}{\partial I_B} = \frac{\partial I_C}{\partial I_E - \partial I_C} = \frac{\partial I_C / \partial I_E}{1 - \partial I_C / \partial I_E}$$

$$= \frac{\alpha_o}{1 - \alpha_o} \equiv \frac{\gamma \alpha_T}{1 - \gamma \alpha_T}$$

Eqn 2-55

It can therefore be seen in Eqn 2-55 that a large common-emitter current gain can only be obtained if both the emitter efficiency and the base transport factor are made close to unity.

In the next section, the effect of a wide energy bandgap emitter heterojunction on the transistor emitter efficiency will be considered.
Section 2-5: The Heterojunction

It was explained qualitatively in Section 1-4 that the heterojunction in an HBT can maintain the emitter efficiency close to unity even though the base doping level is substantially higher than that of the emitter. This is in direct contrast to the BJT case in which a high emitter efficiency is obtained by doping the emitter more heavily than the base. Having a highly doped and very thin base makes the HBT superior to the BJT in high frequency operations which was also briefly explained in Section 1-4. In this section, the effect of the bandgap difference between the emitter and the base on the emitter efficiency will be considered analytically. The equation derived will confirm that the emitter efficiency is relatively independent of the base and emitter doping levels in an HBT.

Recall that in Section 2-4, the emitter efficiency expression before the mass-action law was applied was given by Eqn 2-50. For a heterojunction, the intrinsic carrier concentrations for the two semiconductors are different. Therefore for a base-emitter heterojunction and assuming all donors and acceptors are ionised, the mass-action law equations for the base and emitter are

\[ P_{oe} N_{De} = n_{ie}^2 \]  \hspace{1cm} Eqn 2-56

\[ n_{ob} N_{Ab} = n_{ib}^2 \]  \hspace{1cm} Eqn 2-57

where \( n_{ie} \) and \( n_{ib} \) are the emitter and base intrinsic carrier concentrations, respectively.

Substituting Eqn 2-56 and Eqn 2-57 into Eqn 2-50 results in

\[ \gamma = \frac{1}{1 + \frac{D_{pe}}{D_{nb}} \frac{N_{Ab}}{N_{De}} \frac{L_{nb}}{L_{pe}} \left( \frac{n_{ie}}{n_{ib}} \right)^2 \tanh \left( \frac{W_B}{L_{nb}} \right) } \]   \hspace{1cm} Eqn 2-58

In Eqn 2-58 the term \( \left( \frac{n_{ie}}{n_{ib}} \right)^2 \) is directly controlled and determined by the band structures of the two semiconductors forming the junction shown in Figure 2-7.
Figure 2-7: Energy band diagram of an InP/InGaAs heterojunction.

Invoking the mass-action law once again gives

\[ n_{e}^2 = n_{e}p_{e} \]  
Eqn 2-59

where

\[ n_{e} = \frac{2}{\hbar^3} \left( 2\pi m_{ne}^* kT \right)^{3/2} \exp \left[ -\frac{(E_{ce} - E_F)}{kT} \right] \]  
Eqn 2-60

\[ p_{e} = \frac{2}{\hbar^3} \left( 2\pi m_{pe}^* kT \right)^{3/2} \exp \left[ -\frac{(E_F - E_{ve})}{kT} \right] \]  
Eqn 2-61

and

\[ n_{b}^2 = n_{b}p_{b} \]  
Eqn 2-62

where

\[ n_{e} = \frac{2}{\hbar^3} \left( 2\pi m_{ne}^* kT \right)^{3/2} \exp \left[ -\frac{(E_{cb} - E_F)}{kT} \right] \]  
Eqn 2-63

\[ p_{b} = \frac{2}{\hbar^3} \left( 2\pi m_{pb}^* kT \right)^{3/2} \exp \left[ -\frac{(E_{F} - E_{vb})}{kT} \right] \]  
Eqn 2-64

where \( n_{e} \) and \( p_{e} \) are the electron and hole concentrations in the emitter, \( n_{b} \) and \( p_{b} \) are the electron and hole concentrations in the base, \( m_{ne}^* \) and \( m_{pe}^* \) are the density-of-state effective masses for electrons and holes in the emitter, \( m_{nb}^* \) and \( m_{pb}^* \) are the density-of-state effective masses for electrons and holes in the base, \( E_{F} \) is the Fermi energy level, \( E_{ce} \) and \( E_{vb} \) are the energy levels at the top of the valence bands in the emitter and base,
respectively, and $E_{Ce}$ and $E_{Cb}$ are the energy levels at the bottom of the covalent bands in the emitter and base, respectively.

Using Eqn 2-59 to Eqn 2-64, the term $(n_{ie}/n_{ib})^2$ can be expressed as

$$\left(\frac{n_{ie}}{n_{ib}}\right)^2 = \left(\frac{m_{pe}^* m_{ne}^*}{m_{pb}^* m_{nb}^*}\right)^2 \exp\left[\frac{-\frac{(E_{Ce} - E_{Ve})}{kT}}{\frac{-\frac{(E_{Cb} - E_{Vb})}{kT}}{kT}}\right]$$  \hspace{1cm} \text{Eqn 2-65}

where $\Delta E_g = E_e - E_b$ is the energy difference between the emitter semiconductor bandgap $E_e$ and base semiconductor bandgap $E_b$. Substituting Eqn 2-65 into Eqn 2-58, the emitter efficiency for an HBT is obtained

$$\gamma = \frac{1}{1 + \frac{D_{pe} \cdot N_{th} \cdot L_{nb}}{D_{nb} \cdot N_{De} \cdot L_{pe}} \left(\frac{m_{pe}^* m_{ne}^*}{m_{pb}^* m_{nb}^*}\right)^3 \tanh\left(\frac{W_b}{L_{nb}}\right) \exp\left[\frac{-\Delta E_g}{kT}\right]}$$  \hspace{1cm} \text{Eqn 2-66}

For a very thin base where $W_b << L_{nb}$, Eqn 2-66 can also be approximated as

$$\gamma = \frac{1}{1 + \frac{D_{pe} \cdot N_{th} \cdot W_b}{D_{nb} \cdot N_{De} \cdot L_{pe}} \left(\frac{m_{pe}^* m_{ne}^*}{m_{pb}^* m_{nb}^*}\right)^3 \exp\left[\frac{-\Delta E_g}{kT}\right]}$$  \hspace{1cm} \text{Eqn 2-67}

Comparing Eqn 2-67 with Eqn 2-51 shows that the emitter efficiency equation for an HBT has two extra terms in the denominator, one containing the density-of-state effective masses and one containing an exponential function of the bandgap energy difference. The bandgap energy difference between InP and InGaAs is 0.6 eV and the exponential term in the denominator of Eqn 2-67 is therefore equal to $85 \times 10^{-12}$. As the following calculation will show, this extremely small number maintains an emitter efficiency close to unity and a high common-emitter current gain despite having a more heavily doped base. In the three-terminal HBT shown in Figure 2-2, the base width is $W_b = 0.05 \mu m$ and using the following material parameters from reference [3],

$L_{pe} = 2.4 \mu m$

$$\frac{qD_{pe}}{kT} = 150 \text{ cm}^2/\text{V} \cdot \text{sec}, \quad \frac{qD_{nb}}{kT} = 5000 \text{ cm}^2/\text{V} \cdot \text{sec}$$
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\[ N_{De} = 5 \times 10^{17} \text{ cm}^{-3}, \quad N_{Ab} = 10^{19} \text{ cm}^{-3} \]

\[ m_{pe}^* = 0.56 \quad m_0, \quad m_{ph}^* = 0.50 \quad m_0 \]

\[ m_{ne}^* = 0.08 \quad m_0, \quad m_{nb}^* = 0.041 \quad m_0 \]

\[ \Delta E_g = 0.6 \text{ eV} \]

where \( m_0 \) is the electron rest mass, the calculated emitter efficiency for an InP/InGaAs HBT at \( T = 300 \text{ K} \) is then greater than \( 1 - 10^{-12} \) which, assuming a base transport factor of 0.99, gives a common-emitter current gain of 99. On the contrary if there was no bandgap energy difference, an otherwise homojunction transistor would have an emitter efficiency of 0.954 and a common-emitter current gain of only 17.

Because the emitter efficiency is assured, device designers now have the freedom to set the base and emitter doping levels independently without affecting the emitter efficiency.

Section 2-6: Base-Emitter Junction Space-Charge Recombination Current

In Section 2-3, it was explained that in a BJT, an external base current is required to replenish those holes that are back-injected into the emitter and those that recombine with the traversing electrons in the neutral base. In homojunction silicon BJT, the back-injection hole current is the dominant base current component. In an HBT, however, the base-emitter heterojunction effectively eliminates this back-injection hole current. What was not considered in Section 2-3 and not shown in Figure 2-6 was the third current component that the base current needs to supply and this is the base-emitter junction space-charge recombination current \( I_{scr} \) which is also known as the defect current or simply the junction recombination current.

Whenever equilibrium is disturbed, mechanisms exist and attempt to restore it. Under forward-bias condition, the electron and hole concentrations in the base-emitter junction depletion (space-charge) region are above their equilibrium values. As a result, these two kinds of carriers will try to recombine in order to restore their respective equilibrium concentrations. The physical process responsible for the recombination is the capture of carriers through recombination centres in the forbidden energy bandgap. The recombination centres are localised states due to lattice imperfections as well as impurity
atoms, and act as intermediate steps for an electron from the valence band first to reside in one of these states and then to be transferred to the conduction band or vice versa.

The junction recombination current in the base-emitter depletion region can be calculated from

\[ I_{scr} = qA \int_{-W_{BE}}^{0} R(y) \cdot dy \]  
Eqn 2-68

where \( R(y) \) is the rate of recombination by the recombination centres and \( W_{BE} \) is the width of the base-emitter junction depletion region. Liu [4] showed that \( I_{scr} \) can be described by a diode equation with an ideality factor close to 2, i.e.,

\[ I_{scr} \equiv K \exp \left( \frac{qV_{BE}}{2kT} \right) \]  
Eqn 2-69

where \( K \) is a constant. Therefore an extra component \( I_{scr} \) is added to the emitter and base currents in addition to those predicted by Eqn 2-46 to Eqn 2-48. \( I_{scr} \) dominates the base current at low \( V_{BE} \) and therefore the base current will show a higher ideality factor. As \( V_{BE} \) is increased, the base current ideality factor decreases because the recombination current in the neutral base region becomes more dominant than \( I_{scr} \) and has the same ideality factor as that of the collector current which is close to 1.

On the other hand, for a given \( V_{BE} \) the collector current is not affected by \( I_{scr} \) because the electrons, injected from the emitter constituting \( I_{scr} \), disappear in the base-emitter depletion region by recombining with their hole counterparts injected from the base without travelling to the collector. The net result is a common-emitter current gain which varies with \( V_{BE} \) because the base and the collector currents do not vary with \( V_{BE} \) with the same ideality factor.

To illustrate the effect of \( I_{scr} \), Figure 2-8 shows a measured Gummel plot (\( I_B \) and \( I_C \) versus \( V_{BE} \) with \( V_{BC} = 0 \) V) for the three-terminal InP/InGaAs HBT. The discontinuities in the measured data points from \( V_{BE} = 0.9 \) V onwards are due to the oscillation of the semiconductor parameter analyser.
It can be seen that while the collector current varies with an ideality factor of 1.29 for most part of the Gummel plot, the ideality factor of the base current varies. For very small $V_{BE}$, both the base and collector currents are due to device leakage and can be modelled satisfactorily by a $3.5 \times 10^9 \, \Omega$ parallel resistor connected between the base and the emitter, and between the collector and the emitter.

The effect of $I_{nr}$ on the common-emitter current gain $h_{fe}$ can be illustrated by transforming Figure 2-8 into a plot of $h_{fe}$ versus $I_C$ which is shown in Figure 2-9. As can be seen, the common-emitter current gain was initially below 1 for $I_C < 50 \, \text{nA}$. As $I_C$ is increased, $h_{fe}$ also increases and when $I_C = 10 \, \text{mA}$, $h_{fe}$ is almost 200. So the effect of $I_{nr}$ is to reduce the common-emitter current gain when the transistor is operated at low current.

It should also be noted that recombinations do not only occur in the junction depletion region, but also at the surface of the extrinsic base and the emitter side-walls [5]. The combined effect is a reduced $h_{fe}$ at low currents.
Section 2-7: High-Frequency Performance Limitations

The theory of the HBT at DC has been described in previous sections. It should be noted that the performance, such as the common-emitter current gain, of an HBT and a BJT can be comparable when the frequency of operation is low. At much higher frequencies in the GHz region, the HBT is far more superior to the BJT because of a very thin and heavily doped base. Nevertheless, there are limitations on the high frequency performance of the HBT and these factors will be addressed in this section.

The high-frequency limitation for a bipolar transistor is most often measured and specified in terms of its common-emitter current gain cutoff frequency $f_T$ (also known as the common-emitter gain-bandwidth product or simply current gain cutoff frequency) and maximum oscillation frequency $f_{max}$. $f_T$ is the frequency at which the common-emitter, short-circuit current gain $h_{fe}$ (Eqn 2-55) has fallen to unity. $f_{max}$ is the frequency at which the unilateral power gain of the transistor has fallen to unity.

$f_T$ is commonly approximated [2] as
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\[ f_T = \frac{1}{2\pi \tau_{ec}} \quad \text{Eqn 2-70} \]

where \( \tau_{ec} \) is the total delay time the carriers encounter as they travel from the emitter to the collector in a bipolar transistor. \( \tau_{ec} \) consists of four other delay times:

\[ \tau_{ec} = \tau_e + \tau_{bt} + \tau_{cd} + \tau_c \quad \text{Eqn 2-71} \]

The rigorous derivations of Eqn 2-70 and Eqn 2-71 are lengthy and involve devising a transistor’s \( y \) parameters in both common-base and common-emitter configurations from the device physical parameters and conversions from \( y \) to \( z \) and \( h \) parameters [4]. An alternative but less formal approach to deriving Eqn 2-70 and Eqn 2-71 will be presented in Section 2-7d. The four time constants contained in Eqn 2-71 will first be considered.

**Section 2-7a: Emitter Capacitance Charging Time \( \tau_e \) and Collector Capacitance Charging Time \( \tau_c \)**

When the voltage across a pn junction is varied, the width of the depletion is also varied and majority carriers move into, for increasing forward bias, or out of, for increasing reverse bias, the depletion region to reflect the new terminal voltage. It is the movements of these majority carriers in response to the variation in the junction voltage that give rise to the junction capacitance. The base-collector homojunction capacitance is given by

\[ C_{bc} = \frac{AE_s}{W_{bc}} \quad \text{Eqn 2-72} \]

where \( E_s \) is the semiconductor permittivity in the depletion region. The base-emitter heterojunction capacitance is given by [6]

\[ C_{be} = A \left[ \frac{qN_{De}N_{Ab}E_bE_e}{2(E_sN_{De} + E_bN_{Ab})(V_{bi} - V)} \right]^{1/2} \quad \text{Eqn 2-73} \]

where \( E_b \) and \( E_e \) are the base and emitter permittivities, respectively, \( V_{bi} \) is the junction built-in voltage and \( V \) is the applied voltage (+ve for forward bias). From Eqn 2-73 it can be seen that if one side of a junction is heavily doped, the junction capacitance will be mostly determined by the doping level of the lightly doped side. In an HBT, the doping level of the emitter is almost always at least an order of magnitude lower than the base doping and therefore a lower emitter doping level can reduce the depletion capacitance.
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since $C_{bc} \approx N_{bc}^{\frac{1}{2}}$ [7]. When considering the $f_T$ of a bipolar transistor, the so-called base-emitter diffusion capacitance, which is due to the movements of the minority carriers, is ignored because it decreases with frequency [2]. At $f_T$, the diffusion capacitance will be insignificant compared to the junction depletion capacitance given by Eqn 2-73.

The effect of these junction capacitances is to increase the base reactive current when the junction voltage is time-varying, hence reducing the common-emitter current gain at high frequencies. Figure 2-10 shows a hybrid-$\pi$ equivalent circuit model for a bipolar transistor in which the only frequency dependent components are assumed to be the impedances of $C_{be}$ and $C_{bc}$. The frequency dependence of the base transport factor and the effect of the collector delay time will be dealt with in Section 2-7b and Section 2-7c, respectively.

![Figure 2-10: A common-emitter hybrid-$\pi$ equivalent circuit model for a bipolar transistor [8].](image)

In Figure 2-10, $R_C$ and $R_E$ are the collector and emitter series resistances, respectively, and $\beta$ is the transistor DC common-emitter small-signal current gain. After some straightforward, albeit tedious, algebraic manipulation, it can be shown that the frequency-dependent common-emitter current gain with the output AC shorted, as in Figure 2-10, is given by

$$h_{fe}(\omega) = \frac{i_{out}}{i_{in}} \bigg|_{v_{ce}=0} = \frac{\beta - j\omega R_C C_{bc} - j\omega R_E C_{bc}(1 + \beta + j\omega \tau C_{bc})}{1 + j\omega \tau (C_{be} + C_{bc}) + j\omega C_{bc}(R_E + R_C)(1 + \beta + j\omega \tau C_{bc})}$$

Eqn 2-74

where $\omega = 2\pi f$ and at frequencies of interest ($\leq f_T$)
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\[ h_{fe}(\omega) = \frac{i_{out}}{i_{in}} \bigg|_{v_{ce}=0} = \frac{1}{j\omega R_c (C_{be} + C_{bc}) / \beta + C_{bc} (R_E + R_C)} \]  

Eqn 2-75

It is also noted that

\[ \frac{r_e}{\beta} = \frac{1}{g_m} = \frac{kT}{qI_c} \]  

Eqn 2-76

where \( r_e \) is the emitter dynamic resistance and \( g_m \) is the DC transconductance of the transistor. Substituting Eqn 2-76 into Eqn 2-75 gives

\[ h_{fe}(\omega) = \frac{i_{out}}{i_{in}} \bigg|_{v_{ce}=0} = \frac{1}{j\omega \left( \frac{kT}{qI_c} \right) (C_{be} + C_{bc}) + C_{bc} (R_E + R_C)} \]  

Eqn 2-77

The emitter charging time \( \tau_e \) is the time required to change the base voltage by charging up the base-emitter junction and base-collector junction capacitances through the emitter dynamic resistance and can be identified in Eqn 2-77 as

\[ \tau_e = \frac{kT}{qI_c} (C_{be} + C_{bc}) \]  

Eqn 2-78

Similarly the collector capacitance charging time can be identified in Eqn 2-77 as

\[ \tau_c = (R_C + R_E) C_{bc} \]  

Eqn 2-79

Defining

\[ f_e = \frac{1}{2\pi \left( \frac{kT}{qI_c} \right) (C_{be} + C_{bc})} = \frac{1}{2\pi \tau_e} \]  

Eqn 2-80

and

\[ f_c = \frac{1}{2\pi (R_C + R_E) C_{bc}} = \frac{1}{2\pi \tau_c} \]  

Eqn 2-81

and substituting Eqn 2-80 and Eqn 2-81 into Eqn 2-77 gives
\[ h_{fe}(f) = \left. \frac{i_{out}}{i_{in}} \right|_{f_c = 0} \equiv \frac{1}{j\omega \left( \frac{1}{f_e} + \frac{1}{f_c} \right)} \quad \text{Eqn 2-82} \]

Therefore at frequency equal to \( 1/[2\pi(\tau_e + \tau_c)] \), the magnitude of this base-emitter and base-collector capacitance limited \( h_{fe} \) will become unity.

**Section 2-7b: Base Transit Time \( \tau_{bl} \)**

When the minority electrons are injected from the emitter, it will take them some time to travel across the base and reach the collector. During the time spent in the base, some electrons will be lost due to recombination with the majority holes and therefore the electron current at the collector side will be smaller than the injected electron current at the emitter side. The ratio of these two currents depends on the frequency at which the transistor is operated and can be obtained by first solving the time-dependent continuity equation for the minority electrons in an electric-field free base:

\[ \frac{\partial n_b(y,t)}{\partial t} = D_{nb} \frac{\partial^2 n_b(y,t)}{\partial y^2} - \frac{n_b(y,t) - n_{nb}}{\tau_{nb}} \quad \text{Eqn 2-83} \]

Assume the electron concentration in the base has the following form

\[ n_b(y,t) = n_{b,dc}(y) + n_{b,ac}(y) \cdot \exp(j \omega t) \quad \text{Eqn 2-84} \]

where \( n_{b,dc}(y) \) is the electron concentration profile in the base under DC bias and \( n_{b,ac}(y) \) is amplitude of the electron concentration fluctuation due to the small-signal excitation. The boundary conditions for solving Eqn 2-83 are \( n_{b,ac}(y) = n_{b,ac}(0) \) at the emitter side \( (y = 0) \) and \( n_{b,ac}(y) = 0 \) at the collector side \( (y = W_b) \). Once \( n_{b,ac}(y) \) has been determined, the frequency dependent base transport factor can be evaluated according to

\[ \alpha_r(\omega) = \left( \frac{dn_{b,ac}(y)}{dy} \right)_{y=W_b} \left( \frac{dn_{b,ac}(y)}{dy} \right)_{y=0}^{-1} \quad \text{Eqn 2-85} \]

which, after some algebraic manipulations, yields
where $\tau_{bt} = \frac{W_b^2}{2D_{nb}}$ is the base transit time \cite{1}. Assuming the high-frequency performance of an HBT was limited only by the base transport factor, the common-emitter current gain could be written as

$$h_{fe}(\omega) = \frac{\alpha_{r}(\omega)}{1 - \alpha_{r}(\omega)} = \frac{1}{\frac{1}{\alpha_{r}(\omega)} - 1}$$

$$= \frac{1}{j\omega\tau_{bt}} = \frac{1}{j2\pi f_{bt}}$$

Eqn 2-87

Clearly at frequency

$$f_{bt} = \frac{1}{2\pi\tau_{bt}}$$

Eqn 2-88

the magnitude of the common-emitter current gain would become unity.

**Section 2-7c: Collector Delay Time $\tau_{cd}$**

After diffusing across the base, the injected electrons enter the base-collector depletion region where a strong electric field is present due to the reverse bias. The electrons then drift across the depletion region at their saturation velocity $v_{sat}$. The moving electrons anywhere in this high electric field region instantly induce currents \cite{9} on both ends of the base-collector depletion region and therefore the junction current is equal to the average of all induced currents by the moving electrons in the depletion region. However if the depletion width is large and/or the electrons drift at a low saturation velocity, there will be a large phase difference between the currents induced by the electrons at one end and that at the other end. The phase difference also increases with frequency. The consequence is that as this phase difference between induced currents increases, the junction current will decrease because of the average of the induced currents which are progressively out of phase. Now consider that the electron current entering the base-collector depletion region at $y = W_b$ is sinusoidally time-varying, the mobile electron concentration in the base-collector depletion region is therefore a function of time and
location. Ignoring the constant mobile electron concentration term, the time-varying mobile electron concentration component in the depletion region between $y = W_B$ and $y = W_{BC}$ can be written as

$$n(y, t) = n(W_B) \exp \left[ j \omega \left( t - \frac{y - W_B}{v_{sat}} \right) \right] \quad \text{Eqn 2-89}$$

The junction current due to the average of all the induced currents is then given by

$$i_{jn}(t) = \int_{W_B}^{W_g+W_{BC}} \left( -qA v_{sat} n(y, t) \right) dy$$

$$= i_{ng}(t) \cdot \exp(-j \omega \tau_{cd}) \cdot \left( \frac{\sin \omega \tau_{cd}}{\omega \tau_{cd}} \right) \quad \text{Eqn 2-90}$$

where $i_{ng}(t) = -qA v_{sat} n(W_B) \exp(j \omega t)$ is the electron current entering the depletion region at $y = W_B$ and

$$\tau_{cd} = \frac{W_{BC}}{2v_{sat}} \quad \text{Eqn 2-91}$$

is the collector delay time and, under the assumption that the electrons travel at constant saturation velocity $v_{sat}$, is half the time the electrons take to travel across the base-collector depletion. As can be seen in Eqn 2-90, the induced current is delayed by $\tau_{cd}$ and scaled by $\sin \omega \tau_{cd}/\omega \tau_{cd}$ whose magnitude decreases with frequency for $0 < \omega \tau_{cd} \leq 2\pi$. The induced current on the collector side of the depletion region forms the collector current, i.e. $i_c = i_{jn}$. The difference between $i_{jn}$ and $i_{ng}$ is provided by the base external circuit. Suppose the high-frequency performance was limited solely by this collector delay time, then the collector-delay-time-limited common-emitter current gain would be written as

$$h_{fe}^{(0)}(\omega) = \frac{i_{jn}}{i_{ng}-i_{jn}} = \frac{\exp(-j \omega \tau_{cd}) \left( \frac{\sin \omega \tau_{cd}}{\omega \tau_{cd}} \right)}{1 - \exp(-j \omega \tau_{cd}) \left( \frac{\sin \omega \tau_{cd}}{\omega \tau_{cd}} \right)} \quad \text{Eqn 2-92}$$

At frequency $f_{cd} = 1/2\pi \tau_{cd}$, Eqn 2-92 becomes
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\[ h_{fe}(2\pi f_c) = \frac{e^{-j\sin 1}}{1 - e^{-j\sin 1}} = 0.94 \]  \hspace{1cm} \text{Eqn 2-93}

and the magnitude of this collector-delay-time-limited \( h_{fe} \) is then roughly equal to 1.

**Section 2-7d: Derivation of Current Gain Cutoff Frequency \( f_T \)**

In Sections 2-7a to 2-7c, the effects of the four time constants contained in Eqn 2-71 were examined individually. In this section, their combined effects on the common-emitter current gain will be considered, resulting in the derivations of Eqn 2-70 and Eqn 2-71 for \( f_T \).

From previous discussions, it can be thought that the effects associated with the four time constants contained in Eqn 2-71 are to reduce the common-base current gain of a bipolar transistor. Therefore when these four time constants are considered simultaneously, the transistor common-base current gain can be expressed as

\[
\alpha_o(f) = \frac{\alpha_o(0)}{1 + j \frac{f}{f_e} + j \frac{f}{f_{bt}} + j \frac{f}{f_{cd}} + j \frac{f}{f_c}}
\]  \hspace{1cm} \text{Eqn 2-94}

In arriving at Eqn 2-94, the frequency response associated with the collector delay time effect (Eqn 2-90) was approximated to have a -20 dB/decade roll-off. For frequencies much below \( f_e, f_{bt}, f_{cd} \) and \( f_c \), Eqn 2-94 can be approximated as

\[
\alpha_o(f) \approx \frac{\alpha_o(0)}{1 + j \frac{f}{f_e} + j \frac{f}{f_{bt}} + j \frac{f}{f_{cd}} + j \frac{f}{f_c}}
\]  \hspace{1cm} \text{Eqn 2-95}

Substituting Eqn 2-95 into Eqn 2-55 gives the common-emitter current gain as

\[
h_{fe}(f) = \frac{\alpha_o(0)}{1 + j \frac{f}{f_e} + j \frac{f}{f_{bt}} + j \frac{f}{f_{cd}} + j \frac{f}{f_c}} - \alpha_o(0)
\]

\[
\approx \frac{1}{j 2\pi (\tau_c + \tau_{bt} + \tau_{cd} + \tau_c)f} = \frac{1}{j 2\pi \tau_{ce} f}
\]  \hspace{1cm} \text{Eqn 2-96}
since $a_c(0)=1$. Therefore at frequency $f_T$ given by Eqn 2-70, the magnitude of the common-emitter current gain becomes unity and this completes the proofs for Eqn 2-70 and Eqn 2-71.

The maximum oscillation frequency $f_{\text{max}}$ can also be expressed in terms of $f_T$, the total base resistance $R_b$ and the base-collector junction capacitance $C_{bc}$ as [4]

$$f_{\text{max}} = \sqrt{\frac{f_T}{8\pi R_b C_{bc}}}$$

Eqn 2-97

Unlike a BJT, the base of an HBT can be much more heavily doped without affecting the emitter efficiency. As a result $R_b$ is low, so that high $f_T$ and $f_{\text{max}}$ can be simultaneously achieved.

**Section 2-8: Optoelectronic Mixing Mechanisms in HBTs**

The mixing mechanisms in an HBT arise from the two internal pn junctions since these two junctions exhibit nonlinear exponential current-voltage characteristics. For example when two voltages $v_1$ and $v_2$ at frequencies $f_1$ and $f_2$, respectively, appear across the forward-biased base-emitter junction, the junction will conduct a current which contains components at frequencies $|nf_1 \pm mf_2|$ where $n$ and $m$ are any integers. These current components will in turn cause other voltage components at the same frequencies to appear across the junction through the linear parts of the circuit, such as any resistance and capacitance, and the mixing continues. This description of the mixing mechanisms is very general and often computer modelling for a mixer is closely linked to such a description.

Another way of describing the mixing process in an HBT is through the time-varying small-signal current gain. In electrically pumped optoelectronic mixing, a large LO is applied to the mixer and unlike a small-signal excitation, this LO changes the HBT bias conditions considerably and periodically. It was shown in Figure 2-9 that due to the base-emitter junction space-charge recombination current, the common-emitter current gain of the three-terminal InP/InGaAs HBT is a function of the collector current which in turn is
controlled by $V_{BE}$. Also the emitter dynamic resistance, which affects the common-emitter current gain at high frequencies, is inversely proportional to the collector current. By applying an LO across the base-emitter terminal of a three-terminal HBT, the common-emitter current gain changes at the same frequency as that of the LO. Therefore when an intensity modulated light is incident on the HBT, a small AC primary photocurrent $i_{\text{prim}}(t)$ is generated and experiences a time-varying small-signal current gain $g(t)$ and the output current is given by

$$i_{\text{out}}(t) = i_{\text{prim}}(t) \cdot g(t)$$  \hspace{1cm} \text{Eqn 2-98}$$

Suppose

$$i_{\text{prim}} = mI_{\text{prim}} \cos(2\pi f_{RF} t)$$  \hspace{1cm} \text{Eqn 2-99}$$

and

$$g(t) = G_0 + G_1 \cos(2\pi f_{LO} t)$$  \hspace{1cm} \text{Eqn 2-100}$$

where $m$ is the light modulation index, $I_{\text{prim}}$ is the DC primary photocurrent, $G_0$ and $G_1$ are the Fourier coefficients of $g(t)$, $f_{RF}$ and $f_{LO}$ are the frequencies of the light modulation and the LO, respectively. Substituting Eqn 2-99 and Eqn 2-100 into Eqn 2-98 then yields

$$i_{\text{out}}(t) = mI_{\text{prim}} \left[ G_0 \cos(2\pi f_{RF} t) + \frac{G_1}{2} \left[ \cos(2\pi (f_{LO} - f_{RF}) t) + \cos(2\pi (f_{LO} + f_{RF}) t) \right] \right]$$  \hspace{1cm} \text{Eqn 2-101}$$

Eqn 2-101 clearly shows how the time-varying small-signal current gain $g(t)$ produces the mixed products at frequencies $|f_{LO} \pm f_{RF}|$.

In the two-terminal electrically pumped HBT optoelectronic mixer, the LO is applied across the collector and the emitter. In this case, the base-emitter junction voltage $V_{BE}$ is not directly controlled by the LO. Suppose this two-terminal HBT is optically biased by a constant optical source. If $V_{CE} = 0$ V, a small negative collector current will flow because the turn-on voltage of the InGaAs base-collector homojunction is lower than that of the
InP/InGaAs heterojunction [10], and the accumulated holes in the base forward-bias the base-collector junction in preference to the base-emitter junction. Therefore, a positive offset voltage has to be applied across $V_{CE}$ to reverse-bias the base-collector, blocking the holes from leaving the base, and hence raising $V_{BE}$. A large positive collector current can then flow.

When an LO is applied across the collector and emitter, it increases $V_{CE}$ in one half of the cycle and decreases $V_{CE}$ in the other half. When $V_{CE}$ is reduced to below the offset voltage, the base-collector junction is forward-biased (transistor in saturation mode) and the accumulated holes start to leave the base through the base-collector junction. As a result, $V_{BE}$ and $I_C$ are lowered and the small-signal current gain is reduced correspondingly in this part of the LO cycle. The net result is a periodically changing small-signal current gain $g(t)$ whose frequency is the same as the LO frequency. When the optical source is intensity modulated, an AC primary photocurrent $i_{prim}$ is then generated in the HBT and experiences this time-varying small-signal current gain $g(t)$ which, in this two-terminal HBT, is equal to $(1 + h_{fe}(t))$. As explained before, $h_{fe}(t)$ is collector current dependent which is modulated by the LO. This situation is similar to what has already been described for the three-terminal HBT earlier and the mixing mechanisms should therefore be obvious.

A qualitative explanation for the mixing mechanisms in the HBT has been given in this section. Computer modelling for the two electrically pumped HBT optoelectronic mixers have been performed using harmonic-balance techniques and will be presented in Chapter 3. For the three-terminal HBT, a large-signal equivalent circuit T-model is used while a quas-static model is employed for the two-terminal HBT.

**Section 2-9: Conclusion**

The idea of optically controlling HBTs has been introduced using an equivalent circuit model in Figure 2-1 in which the optical to electrical conversion is modelled by a photodiode connected between the base and the collector. Once the primary photocurrent
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of this photodiode has been obtained, the HBT can then be treated as a normal electrical transistor.

The optical absorption processes in a top-illuminated HBT and an edge-coupled HBT have been considered, resulting in the expressions for the corresponding primary photocurrents.

The static operation of the bipolar transistor has been briefly reviewed. Commonly used figures of merit in DC operation, such as the emitter efficiency, the base transport factor and the common-emitter current, have been expressed in terms of transistor structural and doping parameters.

The use of a wide bandgap emitter forming a base-emitter heterojunction has been shown to maintain the emitter efficiency close to unity regardless of the relative doping levels in the base and the emitter. With the emitter efficiency now assured, the base of the transistor can be heavily doped and made very thin, thus reducing the base transit time without significantly increasing the base resistance. The emitter can also be lightly doped, reducing the base-emitter depletion capacitance. As a result both $f_T$ and $f_{max}$ of the HBT are higher than those of the BJT.

It has been demonstrated using the measured Gummel plot for the three-terminal HBT that the base-emitter junction space-charge recombination current reduces the common-emitter current gain at low collect current.

The four time constants which directly affect $f_T$ have also been considered and a derivation of the expression (see Eqn 2-70 and Eqn 2-71) for $f_T$ in terms of these time constant has been given.

Finally the optoelectronic mixing mechanisms in HBTs have been explained in terms of the time-varying small-signal current gain which is modulated by the applied LO.

The purpose of this theory Chapter has been to provide an insight into the operation and limitations of HBTs. The equations derived should be useful in estimating the basic HBT
performance, but will not be used in the computer modelling in the next Chapter where
two large-signal HBT models will be developed and used in the harmonic-balance
simulations for the two electrically pumped HBT optoelectronic mixers.
References


Chapter Three

Optoelectronic Mixing Simulation using
Harmonic-Balance Techniques

Section 3-1: Introduction

This Chapter is concerned with the use of harmonic-balance techniques in simulating the two HBT electrically pumped optoelectronic mixers investigated in this work. Since mixing involves driving the HBT with a large local oscillator (LO) signal, a circuit model describing the HBT behaviour under such a large input is required and the circuit is commonly simulated in the time-domain as in SPICE. Fourier transforming the output waveform will give the frequency spectrum from which the conversion gain of the mixer is determined. There are, however, a few difficulties in such a time-domain approach. Firstly, the circuit might contain an input and output impedance matching network, signal filters and frequency-dependent transistor parameters such as the base transport factor. These linear elements can be easily described in the frequency-domain but it would be rather complicated to do so in the time-domain. Secondly, since only the steady-state solution of the mixing is of interest in the current investigation, the number of LO cycles the simulator takes to reach a steady state can be very high. One usually has to examine manually the waveforms and decide when a steady-state solution has been reached. Finally and most importantly, mixing involves not only the LO frequency but also the intermediate frequency (IF) and other up- and down-converted and image frequencies. The minimum time step in the time-domain analysis is set by the inverse of the sampling frequency which, as required by the Nyquist theorem, is at least twice the highest frequency component in the circuit being simulated. For example if the LO frequency is 1 GHz and ten harmonics of the LO are required, the minimum time step will be 50 ps. On the other hand, the time span over which the simulation has to be carried out is dictated by the inverse of the frequency resolution required. For example if the IF is 100 MHz, the simulation will have to span at least 10 ns and the corresponding number of simulation points will be 201. If the IF is reduced to 100 kHz, the number of simulation
points required will increase drastically to 200001. Therefore the computation time of such a time-domain analysis is affected significantly by the frequency relationship between the IF and LO.

The approach in this work to simulating electrically pumped optoelectronic mixing involves two steps. The first step is to employ the harmonic-balance techniques to analyse the mixer subjected to the large LO excitation alone without considering the small photogenerated radio frequency (RF) signal. The results of such an analysis will be the LO induced time-varying voltage waveform at each node of the mixer circuit model. The second step is to analyse the mixer in a quasilinear fashion in which the sole input to the mixer is the photogenerated RF signal without considering the LO any more. Frequency generation at the IF and other image frequencies are dealt with in the analysis using the concept of the conversion-matrix which is determined from the time-varying small-signal conductance of the nonlinear element in the model.

Harmonic-balance allows the linear parts of the model to be described and simulated in the frequency-domain and the nonlinear parts in the time-domain, and forward and inverse Fourier transforms are used to bridge the two parts. Therefore the first problem of having to describe the responses of the linear elements in the time-domain is avoided. Also harmonic-balance attempts to find the steady-state solution from the outset and so no time is wasted on the transient, which is the second problem mentioned in the first paragraph. Since the mixer is analysed separately for the LO and photogenerated RF signal excitations, therefore the number of simulation points and hence the computation time will be independent of the frequency relationship between the LO, IF and RF.

A very limited amount of work has been published on the modelling of HBT optoelectronic mixers. Betser et al [1] reported modelling a three-terminal HBT electrically pumped optoelectronic mixer using SPICE. The equivalent circuit used was a standard SPICE bipolar transistor (BJT) model with a current source connected between the base and the collector to account for the primary photocurrent generation. This model suffers from a number of weaknesses. First, the use of a standard SPICE BJT model implicitly assumes that the base current is dominated by the back-injection hole current from the base to the emitter, which has the same ideality factor as the injection current from the emitter to the base, and hence the collector current. As has been explained in
Section 2-6, this is not accurate because of the use of the heterojunction and the base current in the present InP/InGaAs HBT is dominated by the base-emitter junction recombination current, having a different ideality factor from the collector current. The difference in the ideality factors of the base and collector currents contributes a significant mixing mechanism because the transistor current gain becomes a function of $V_{BE}$ (or $I_C$) which in turn is modulated by the LO. If the base current has the same ideality factor as the collector current, as is assumed in the SPICE BJT model in [1], current gain variation with $V_{BE}$ can only occur when the frequency is higher than the inverse of the time constant of the base-emitter capacitance and dynamic resistance. In other words, the assumption of equal ideality factor implies that at low frequencies when the impedance of the base-emitter capacitance is high compared to the base-emitter dynamic resistance, mixing becomes less efficient because the transistor current gain does not vary much with $V_{BE}$.

Another weakness of the SPICE model is that the frequency response of the transistor is modelled by lumped capacitors only and the base transport factor frequency dependency is ignored. This can be understood because when SPICE is required to model the large-signal response of a transistor, it is carried out purely in the time-domain and the frequency dependent base transport factor is more difficult to model in the time-domain.

The model in [1] did, however, take the base-collector junction into account so that the behaviour of the HBT when driven into saturation can be modelled more accurately. The present model to be discussed in Section 3-5 for the three-terminal HBT assumes that the base-collector junction is always reverse biased because of a constant 1.5 V collector-emitter bias. The response of the HBT in the saturation mode can be taken into account when a full base-collector junction model is included in the present model.

Harmonic-balance techniques have been popular in large-signal microwave circuit analysis [2,3] and have recently been employed in modelling laser dynamics [4,5]. This work, however, has extended its application to modelling optoelectronic mixing in HBTs.

The order of this Chapter is as follows. In Section 3-2, the concept of harmonic-balance and its application in finding the steady-state large-signal behaviour of the HBT when
driven by an LO will be introduced and the algorithm employed for solving the harmonic-
balance problems will be described in Section 3-3. The conversion matrix, which relates
the currents and voltages at the RF, IF and other image frequencies in the nonlinear
elements, will then be explained in Section 3-4. The simulation implementations and
results for the three-terminal and two-terminal HBT electrically pumped optoelectronic
mixers will then be presented in Section 3-5 and 3-6, respectively. Finally a conclusion
will be given in Section 3-7.

Section 3-2: Fundamentals of Harmonic-Balance Techniques

Any large-signal equivalent circuit can be thought of as consisting of two kinds of
elements, one linear, such as ideal resistors and capacitors, and the other nonlinear, such
as pn junctions. In the small-signal analysis, the nonlinear elements are linearised around
the DC bias point so that their differential resistances or conductances can be found and
treated as if they were linear. Linear resistors are specified by real constants and reactive
elements by their frequency-dependent complex phasor representations. The result is a
network consisting solely of linear components whose response to a small-signal
excitation as a function of frequency can be easily found analytically using matrix algebra
or numerically requiring only one iteration at each frequency.

In conventional large-signal analysis, the nonlinear components must be modelled in the
time-domain. At the same time, the reactive elements can no longer be described by their
complex phasors but must be specified with their time-dependent differential equations.
Simulation is carried out for a number of LO cycles until a steady state is reached which
is usually controlled by the reactive elements. The situation is different if harmonic-
balance techniques are used.

The concept of harmonic-balance can be illustrated with Figure 3-1.
This model consists of a linear and a nonlinear subcircuit. The nonlinear subcircuit is described in the time-domain by

\[ I(t) = f(V(t)) \]  
\[ \text{Eqn 3-1} \]

where \( V(t) \) is the instantaneous terminal voltage across both the linear and nonlinear subcircuits and \( I(t) \) is the corresponding current flowing into nonlinear subcircuit. The LO is embedded in the linear subcircuit as an AC voltage source at frequency \( f_{LO} \). Since the only excitation is the LO in the model at this stage, both \( V(t) \) and \( I(t) \) are periodic with the period equal to \( f_{LO}^{-1} \). \( I_{non} \) is an \( n \)-by-1 matrix whose elements are the coefficients of the Fourier series representation of \( I(t) \) and \( n \) is the number of harmonics (including DC) of the LO considered in the model. \( n = 1 \) designates DC, \( n = 2 \) designates \( f_{LO} \) and so on. A large \( n \) will give a better Fourier series representation of \( I(t) \) but the computation time will have to be increased accordingly because of a smaller time step required. Similarly \( V \) is also an \( n \)-by-1 matrix whose elements are the coefficients of the Fourier series representation of \( V(t) \).

\( I_{lin} \) is an \( n \)-by-1 matrix representing the current at each harmonic frequency flowing into the linear subcircuit and is related to \( V \) by

\[ (I_{lin})_n = g(V_n, (n-1)f_{LO}) \]  
\[ \text{Eqn 3-2} \]

where \((I_{lin})_n \) and \( V_n \) are the \( n \)th elements of \( I_{lin} \) and \( V \), respectively, and the linear function \( g \) gives the current component \((I_{lin})_n \) in response to the voltage \( V_n \) at frequency \((n-1)f_{LO} \). \( g \) can be found analytically using matrix algebra. Alternatively if the expression for \( g \) is too complicated because of the high number of nodes in the linear
subcircuit, \( (I_{\text{lin}})_n \), can be directly evaluated using numerical methods, such as the Newton's method \([6]\) requiring only one iteration at each harmonic.

The aim of the harmonic-balance techniques is to find \( V \) so that the currents calculated from the nonlinear subcircuit and the linear subcircuit are balanced. Mathematically harmonic-balance aims to determine \( V \) such that

\[
I_{\text{lin}} + I_{\text{non}} = 0 \quad \text{Eqn 3-3}
\]

When the condition in Eqn 3-3 is met, the corresponding \( V \) must be the solution. In practice, since only a finite number of harmonics are used, the aim of the harmonic-balance becomes to minimise the error function \( e \)

\[
e = (I_{\text{lin}} + I_{\text{non}})^T (I_{\text{lin}} + I_{\text{non}}) \quad \text{Eqn 3-4}
\]

where the superscript \( T \) denotes the transpose of the matrix concerned and \(*\) denotes complex conjugate.

**Section 3-3: Solution Algorithms for Harmonic-Balance Problems**

A number of algorithms \([2]\) can be used to obtain solutions to harmonic-balance problems of the form given by Eqn 3-4. The success of these algorithms depends on how close the initial guess solution is to the real solution, how nonlinearly the system being considered behaves and how stable and computation intensive the method is. A popular choice is the iterative Newton's method whose application in solving one-dimensional nonlinear problems is well-known \([6]\). In harmonic-balance problems, however, the number of dimensions the Newton's method is required to solve increases to \( n \times m \) where \( m \) is the number of nodes in the model, and the number of partial derivatives that need to be generated for each iteration will be \((n \times m)^2\). Generating this many partial derivatives is very computation intensive. It was also empirically found that this method does not always converge because of the very nonlinear pn-junctions included in the model. However it should not be confused with the application of the Newton's method to solving the linear part of the problem as mentioned previously in which the method always converges and produces a solution with only one iteration at each harmonic.
Start Figure 3-2: Splitting method flow chart for solving harmonic-balance problems.

Initial $V$

Inverse Fourier transform $V$ to find $V(t)$

Determine $I(t)$ from nonlinear subcircuit $I(t) = f(V(t))$

Fourier transform $I(t)$ to give $I_{\text{non}}$

Equate $I_{\text{lin}} = -I_{\text{non}}$

Find $V_{\text{temp}}$ from $I_{\text{lin}}$ and linear subcircuit

$|V_{\text{temp}} - V| < \text{Tolerance?}$

No

Yes

Inverse Fourier transform $V$ to find $V(t)$

Output $V(t)$ as solution

Equate $V = V_{\text{new}}$

Find new $V$

$V_{\text{new}} = (1-s)V + sV_{\text{temp}}$

Splitting Method

Stop
The algorithm employed in this work is illustrated with a flowchart in Figure 3-2.

As with any iterative algorithm, this method starts the simulation by guessing the matrix \( V \) which, recalling from Figure 3-1, is the frequency-domain representation of the voltage across both the linear and nonlinear subcircuits. The algorithm first simulates the nonlinear subcircuit in the time-domain by inverse Fourier transforming \( V \) to \( V(t) \) from which the time-varying current \( I(t) \) can be determined. Once \( I(t) \) is found, \( I(t) \) is Fourier transformed to the frequency-domain \( I_{\text{non}} \). According to Figure 3-1 and Eqn 3-3, \( I_{\text{lin}} \) is equated to \(-I_{\text{non}}\). The algorithm next simulates the response of the linear subcircuit to \( I_{\text{lin}} \) in the frequency-domain and calculates a new matrix \( V \), temporarily named \( V_{\text{temp}} \).

If the initial matrix \( V \) was the solution, satisfying both the linear and nonlinear subcircuits, \( V_{\text{temp}} \) will be the same as the starting \( V \). A decision is therefore made based on whether \(|V_{\text{temp}} - V|\) is smaller than a specified tolerance. If this condition is met, then \( V \) must be a solution and will then be inverse Fourier transformed to find \( V(t) \). If the condition is not met a new matrix \( V \) needs to be generated and fed back to the beginning of the algorithm and the procedures will start again.

The success of this iterative algorithm depends critically on how the new \( V \), temporarily named \( V_{\text{new}} \), is generated. The assumption used in generating \( V_{\text{new}} \) is that the true solution \( V \) must be located somewhere between the initial \( V \) and \( V_{\text{temp}} \) which has been arrived at after the simulator has gone round the whole circuit model once. Therefore, using a linear approximation in the frequency-domain

\[
V_{\text{new}} = (1-s)V + sV_{\text{temp}}
\]

Eqn 3-5

where \( 0 < s < 1 \). The way of generating \( V_{\text{new}} \) using Eqn 3-5 is called the splitting method, similar to the one by Hicks and Khan [7]. The coefficient \( s \) determines how much \( V_{\text{new}} \) should differ from the initial \( V \). For weakly nonlinear circuits, \( s \) can be increased to accelerate the simulation. For highly nonlinear circuits, \( s \) must be reduced to ensure convergence. It was found empirically that for small LO powers, e.g. -12 dBm,
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$s = 0.5$ or smaller gives successful results, while for large LO powers, e.g. 0 dBm, $s = 0.05$ or smaller should be used.

It was also found empirically that the quality of the initial guess $V$ is not critical at all, and in fact in the simulations, the initial $V$ was simply replaced by a zero matrix or by the solution of a previous simulation evaluated at a smaller bias voltage.

The result of this large-signal harmonic-balance simulation is the time-varying voltage waveform at each node of the circuit model. The voltage waveforms across the nonlinear elements are particularly important since they allow the conversion matrices to be determined which describe the mixing of a large LO with a small RF signal in a quasilinear fashion. The conversion matrix will be the subject of the following section.

**Section 3-4: The Conversion Matrix**

In the last section, the circuit model was analysed with only the LO excitation. In this section the response of the circuit model to a small RF signal when simultaneously driven by the LO is considered.

Mixed products are produced by the nonlinear elements in a mixer when driven by more than one tone. Consider Figure 3-3 in which a nonlinear element is represented by a block model and its terminal behaviour by $I = f(V)$.

![Figure 3-3: A nonlinear circuit block model.](image)

$V$ and $I$ are large-signal time-varying quantities caused by the LO. Suppose a small-signal time-varying voltage $v$, induced by the RF, is superimposed on $V$, the current...
flowing into the nonlinear element will also have a small-signal current component \( i \) superimposed on \( I \):

\[
I + i = f(V + v)
\]

Eqn 3-6

The right-hand-side (r.h.s.) of Eqn 3-6 can be expanded in a Taylor's series as

\[
f(V + v) = f(V) + \frac{df(V)}{dV} v + \frac{1}{2} \frac{d^2 f(V)}{dV^2} v^2 + \ldots
\]

Eqn 3-7

Since \( v \) is assumed small, \( v^2 \) and other higher order terms are ignored, leaving only

\[
f(V + v) = f(V) + \left. \frac{df(V)}{dV} \right|_V v
\]

Eqn 3-8

and

\[
I + i = f(V) + \left. \frac{df(V)}{dV} \right|_V v
\]

Eqn 3-9

or

\[
i = \left. \frac{df(V)}{dV} \right|_V v
\]

Eqn 3-10

\[
\left. \frac{df(V)}{dV} \right|_V \text{ in Eqn 3-10 is the differential or small-signal conductance } g \text{ of the nonlinear element at } V. \text{ Since } V \text{ is a time-varying quantity having the same frequency as the LO, } g \text{ is also time-varying having the same frequency. Eqn 3-10 can be written as}
\]

\[
i(t) = g(t)v(t)
\]

Eqn 3-11

where the time dependencies of the three quantities are shown explicitly.

\( g(t) \) can be written in a complex Fourier series as

\[
g(t) = \sum_{k=-\infty}^{\infty} G_k e^{j\omega_{LO}t}
\]

Eqn 3-12

where \( \omega_{LO} = 2\pi f_{LO} \). \( i(t) \) and \( v(t) \) represent the mixed products with components at frequencies \( \pm f_{IF} \pm kf_{LO} \) where \( f_{IF} \) is the intermediate frequency and \( k \) is any integer.

Writing \( i(t) \) and \( v(t) \) in Fourier series gives
\[ i(t) = I_0 e^{j\omega_0 t} + e^{j\omega_0 t} \sum_{k=1}^{\text{upper}} \left( I_k^{\text{upper}} e^{jk\omega_{ab} t} + I_k^{\text{lower}} e^{-jk\omega_{ab} t} \right) + \]
\[ I_0 e^{-j\omega_0 t} + e^{-j\omega_0 t} \sum_{k=1}^{\text{lower}} \left( I_k^{\text{upper}} e^{jk\omega_{ab} t} + I_k^{\text{lower}} e^{-jk\omega_{ab} t} \right) \]  
Eqn 3-13

and

\[ v(t) = V_0 e^{j\omega_0 t} + e^{j\omega_0 t} \sum_{k=1}^{\text{upper}} \left( V_k^{\text{upper}} e^{jk\omega_{ab} t} + V_k^{\text{lower}} e^{-jk\omega_{ab} t} \right) + \]
\[ V_0 e^{-j\omega_0 t} + e^{-j\omega_0 t} \sum_{k=1}^{\text{lower}} \left( V_k^{\text{upper}} e^{jk\omega_{ab} t} + V_k^{\text{lower}} e^{-jk\omega_{ab} t} \right) \]  
Eqn 3-14

where \( \omega_{EF} = 2\pi f_E \) and the superscripts 'upper' and 'lower' designate the mixed products in the upper and lower sidebands, respectively. Note that the second terms in both Eqn 3-13 and Eqn 3-14 are simply complex conjugates of their respective first terms. Also if the second term of Eqn 3-14 is removed, the corresponding second term in Eqn 3-13 also vanishes according to Eqn 3-11 and Eqn 3-12. Therefore after removing the two second terms from both Eqn 3-13 and Eqn 3-14, Eqn 3-11 can be rewritten as

\[ i'(t) = g(t)v'(t) \]  
Eqn 3-15

where

\[ i'(t) = I_0 e^{j\omega_0 t} + e^{j\omega_0 t} \sum_{k=1}^{\text{upper}} \left( I_k^{\text{upper}} e^{jk\omega_{ab} t} + I_k^{\text{lower}} e^{-jk\omega_{ab} t} \right) \]  
Eqn 3-16

and

\[ v'(t) = V_0 e^{j\omega_0 t} + e^{j\omega_0 t} \sum_{k=1}^{\text{upper}} \left( V_k^{\text{upper}} e^{jk\omega_{ab} t} + V_k^{\text{lower}} e^{-jk\omega_{ab} t} \right) \]  
Eqn 3-17

It should be noted that \( i'(t) \) and \( v'(t) \) contain only half of the full Fourier series representations of \( i(t) \) and \( v(t) \), respectively. However no information is lost due to the use of \( i'(t) \) and \( v'(t) \) because \( i(t) \) and \( v(t) \) can be readily restored, if required, from

\[ i(t) = i'(t) + (i'(t))^* \]  
Eqn 3-18

and

\[ v(t) = v'(t) + (v'(t))^* \]  
Eqn 3-19

It is more convenient to write Eqn 3-16 and Eqn 3-17 as
Substituting Eqn 3-20, Eqn 3-21 and Eqn 3-12 into Eqn 3-15 and collecting terms at similar frequencies reveals that the relationship between the coefficients \( G_k \), \( I_k \) and \( V_k \) can be written in a simple matrix equation. To illustrate this, it is assumed that frequency components for \( |k| > 2 \) are small and can be ignored. Therefore,

The square matrix in Eqn 3-25 is the small-signal conductance conversion matrix of the nonlinear element which relates signals at different mixing frequencies. For example if \( v(t) \) is caused by the RF excitation at \( f_{RF} = f_{IF} + f_{LO} \), \( V_1 \) in Eqn 3-25 will be nonzero and will cause current components to appear at \( f_{RF} \) given by \( I_1 \), at the down-converted frequency \( f_{IF} \) given by \( I_0 \), and at the up-converted frequency \( f_{IF} + 2f_{LO} \) given by \( I_2 \). Mixed products, such as \( I_2 \) and \( I_{-1} \), through higher order mixing are also covered by the conversion matrix. In general, signal components at any frequency given by \( (\omega_k/2\pi) \) can be caused by any other signal component at any frequency given by \( (\omega_k/2\pi) \), but
higher order mixing is usually much weaker than fundamental mixing and so only a small number of terms, such as those in Eqn 3-25, need to be considered.

The elements of the above conversion matrix are simply the Fourier series coefficients of the time-varying small-signal conductance $g(t)$ which can be found readily once the time-varying voltage across the nonlinear element has been determined from the previous harmonic-balance analysis.

The use of the conversion matrix has allowed frequency mixing to be described by a linear matrix operation (Eqn 3-25). With the nonlinear element now represented by a conversion matrix, the overall mixer response can be simulated readily using highly efficient linear methods.

The harmonic-balance techniques and the conversion matrix method have been used to simulate a three-terminal normal-incidence HBT and a two-terminal edge-coupled HBT electrically pumped optoelectronic mixer. The implementations and simulation results will be presented in the following two sections.

Section 3-5: Implementation and Simulation Results for a Three-Terminal HBT Optoelectronic Mixer

The modelling of the HBT optoelectronic mixer consists of a large-signal simulation using harmonic-balance and the determination of the conversion gain using the conversion matrix method. These two steps will be described in the following two subsections.

Section 3-5a: Large-Signal Simulation Using Harmonic-Balance

To apply the simulation techniques discussed so far in this Chapter, the first step in modelling an HBT optoelectronic mixer is to have a large-signal equivalent circuit to be used in the harmonic-balance simulation. Figure 3-4 shows such an equivalent circuit for the HBT in the T-topology.
The component values are given in Table 3-1 and were obtained by a combination of device geometry estimations, parameter extractions and optimisations [8,9] from the DC and s-parameter measurements, as well as from the equivalent circuit of the bias-T manual. The value of $R_c$ is seen to be a bit too low. In practice, $R_c$ should have a value of around 3 $\Omega$ to 4 $\Omega$. The discrepancy in the value of $R_c$ could have resulted from the optimisation process, giving non-unique component values.

$V_{BE}$ and $V_{CE}$ in Figure 3-4 represent the HBT base-emitter voltage and collector-emitter voltage, respectively. $I_{\text{prim\_DC}}$ represents the average primary photocurrent $I_{\text{prim}}$ of 15 $\mu$A.

All voltages, currents and impedances printed in bold characters in Figure 3-4 are $n$-by-1 column matrices whose first elements ($n=1$) represent their DC values, second elements ($n=2$) represent their values at the first harmonic of $f_{LO}$ and so on. As a result $V_{BE}$,
Table 3-1: Component values used in Figure 3-4

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Iprim_DC</td>
<td>15μA</td>
</tr>
<tr>
<td>L_bias</td>
<td>2μH</td>
</tr>
<tr>
<td>C_bias</td>
<td>800pF</td>
</tr>
<tr>
<td>Lb, Lc, Le</td>
<td>0.39 nH</td>
</tr>
<tr>
<td>Rb</td>
<td>19Ω</td>
</tr>
<tr>
<td>Re</td>
<td>2.1Ω</td>
</tr>
<tr>
<td>Rc</td>
<td>0.001Ω</td>
</tr>
<tr>
<td>Rcb</td>
<td>12000Ω</td>
</tr>
<tr>
<td>Cbe</td>
<td>0.1nF</td>
</tr>
<tr>
<td>Cbc</td>
<td>80fF</td>
</tr>
<tr>
<td>α(f)</td>
<td>α0(1-j2πft)</td>
</tr>
<tr>
<td>α0</td>
<td>0.9983</td>
</tr>
<tr>
<td>τd</td>
<td>1.7ps</td>
</tr>
</tbody>
</table>

VCE and Iprim_DC have nonzero elements only when \( n = 1 \) while V_LO, representing the LO source, only has a nonzero element when \( n = 2 \). All other node voltages are described by a single \( n \)-by-\( m \) matrix \( V \) where \( m \) is the number of nodes and is equal to 5 in the current model. The element \( V_{n,m} \) of \( V \) is therefore the Fourier coefficient of the voltage at frequency \( (n-1)f_{LO} \) at node \( m \). It should be apparent that these matrices are frequency-domain quantities.

The two nonlinear elements in Figure 3-4 are the diodes D_REC and D_F. D_REC accounts for the base-emitter junction recombination current while D_F accounts for the emitter injection current. The IV characteristics of the D_REC and D_F can determined from the HBT static Gummel plot in Figure 3-5. The microwave behaviour of the two diodes is modelled by the base-emitter junction capacitance Cbe shown in Figure 3-4.

The discontinuities in the measured data points from \( V_{BE} = 0.9\,V \) onwards are due to the oscillation of the semiconductor parameter analyser. The collector current exhibits a constant ideality factor in the Gummel plot for \( V_{BE} > 0.3\,V \) and is characteristic of an injected current. The base current, on the other hand, consists mainly of a base-emitter junction recombination current and a current needed to compensate for the loss of electrons injected from the emitter as they traverse the base. The base-emitter junction
recombination current is dominant for $0.3V \leq V_{BE} \leq 0.8V$ whose ideality factor in the
Gummel plot is higher than that of the collector current. For higher $V_{BE}$, the base current
is expected to be dominated by the compensation current and has the same ideality factor
as that of the collector current. For very small $V_{BE}$, both the base and collector currents
are due to device leakage and can be modelled satisfactorily by a $3.5 \times 10^8 \ \Omega$ resistor
connected between the base and the emitter, and between the collector and the emitter.
The IV characteristics of D_REC can be determined from the base current for
$0.3V \leq V_{BE} \leq 0.8V$ in the Gummel plot. The IV characteristics of D_F can be
approximated to that of the collector current. Both the D_F and D_REC IV
characteristics are curve fitted so that they can be implemented analytically in the
simulator.

A portion of the injected emitter current $IE$ reaches the collector and is described by the
current source matrix $ICB$ whose $n^{th}$ element is given by

$$ICB_n = \alpha((n-1)f_{LO}) \cdot IE_n$$

Eqn 3-26

where $\alpha(f)$ is the frequency-dependent base transport factor and $IE_n$ is the $n^{th}$ element
of $IE$. The DC value of the base transport factor $\alpha_0$ can be determined from the
common-emitter current gain derived from Figure 3-5 at higher $V_{BE}$ where the base
current is dominated by compensation current.
In conventional large-signal analysis [10], the current source ICB is specified in the time-domain and is determined from the convolution integral of the product of \( IE(t) \), which is the time-domain equivalence of \( IE \), and the unit impulse response, \( h(t) \), of the base transport factor. The determination of \( h(t) \) is less straightforward compared to that of \( \alpha(f) \). The evaluation of the convolution integral also increases the computation time. In this work, advantage has been taken of the fact that circuit elements can be specified and simulated in either time-domain or frequency-domain, whichever is more convenient. Therefore the base transport factor has been specified in the model as a function of frequency and the otherwise required convolution integration has been replaced by a few simple multiplications given by Eqn 3-26.

Comparing Figure 3-4 with Figure 3-1 shows that the two diodes form the nonlinear subcircuit and the rest of the Figure 3-4 forms the linear subcircuit. The objective of the present simulation using harmonic-balance is to find the correct time-varying voltage waveform across the two diodes so that the conversion matrices can be determined. The simulation first determines \( V_{\text{non}}(t) \) by inverse Fourier transforming \( V_{\text{non}} \). Assuming 5 LO harmonics (DC counts one) are used, \( V_{\text{non}} \) is then given by

\[
V_{\text{non}} = \begin{bmatrix}
V_{1,4} - V_{1,3} \\
V_{2,4} - V_{2,3} \\
V_{3,4} - V_{3,3} \\
V_{4,4} - V_{4,3} \\
V_{5,4} - V_{5,3}
\end{bmatrix}_{\text{Initial}}
\]

Eqn 3-27

\( V_{\text{non}} \) in Eqn 3-27 are initial guess values. \( I_{\text{non}}(t) \) is then calculated from \( V_{\text{non}}(t) \) and the IV characteristics of the diodes D_F and D_REC. The next step is to Fourier transform \( I_{\text{non}}(t) \) and find \( I_{\text{non}} \). Equating \( I_{\text{lin}} = -I_{\text{non}} \) a new set of node voltages described by \( V \) can then be obtained from the linear subcircuit analytically using matrix algebra, or numerically using Newton's method. The latter was chosen because the large number of components in the equivalent circuit makes the matrix algebra approach more complicated.
The first step in implementing the Newton's method is to construct a current error function for each of the five nodes in the model evaluated at each harmonic, including DC. For each node, the current error function is the sum of all diverging currents from that node calculated from the neighbouring node voltages, current sources and the impedances in between. To illustrate this, the current error function for node 1 at frequency \((n-1)f_{LO}\) is

\[
I_{err,1} = \frac{(V_{n,1} - V_{n,4})}{Z_{b_n}} + \frac{(V_{n,1} - V_{-LO_n})}{50 + Z_{C_{bias}}_{n}} + \frac{(V_{n,1} - V_{BE_n})}{Z_{L_{bias}}_{n}}
\]

Eqn 3-28

where \(V_{n,1}, V_{n,4}, V_{-LO_n}, V_{BE_n}, Z_{b_n}, Z_{C_{bias}}_{n}, \) and \(Z_{L_{bias}}_{n}\) are the \(n^{th}\) elements of their respective matrices. As can be seen, the error functions at each frequency are functions of the node voltages. Obviously when all node voltages represent the correct solution, the sum of all diverging currents from any node must be zero, and so must the current error functions. After the error functions for all the nodes have been formulated, a matrix containing those error functions is constructed:

\[
\textbf{I}_{err}(V_{n,1}, V_{n,2}, V_{n,3}, V_{n,4}, V_{n,5}) = \begin{bmatrix}
I_{err,1} \\
I_{err,2} \\
I_{err,3} \\
I_{err,4} \\
I_{err,5}
\end{bmatrix}
\]

Eqn 3-29

The round brackets in Eqn 3-29 emphasize that \(\textbf{I}_{err}\) is a function of all other node voltages.

The second step in the implementation of the Newton's method is to construct a Jacobian matrix [6] for the error functions. The Jacobian matrix at frequency \((n-1)f_{LO}\) is given by
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which can be easily obtained with just a few lines of code. The Jacobian matrix of a multi-dimensional function is analogous to the gradient of a curve in the one-dimensional case.

The third step is to determine a set of node voltages which will satisfy \( I_{\text{lin}} = -I_{\text{non}} \) and make all the error functions zero. This set of node voltages are given by a very simple operation:

\[
\begin{bmatrix}
V_{n,1} \\
V_{n,2} \\
V_{n,3} \\
V_{n,4} \\
V_{n,5,\text{temp}}
\end{bmatrix}
= -\text{Jacob}^{-1}_n \text{err}_n (0,0,0,0,0)
\]

Eqn 3-31

where \( \text{Jacob}^{-1}_n \) is the matrix inverse of \( \text{Jacob}_n \). Although the proof of Eqn 3-31 is not difficult, it is not readily available in the textbooks. However an analogy can be drawn with a one-dimension linear function such as \( f(x) = mx + c \). This function is zero when \( x = -m^{-1}c = -m^{-1}f(0) \) which has the same form as Eqn 3-31. The operation in Eqn 3-31 is carried out for all values of \( n \).

The final step is to generate a new set of node voltages to be used in the next iteration. These voltages at frequency \( (n-1)f_{LO} \) are given by
where $s$ is the coefficient for the splitting method described in Section 3-3. The iteration continues until

$$\sum_{m=1}^{\text{No. of nodes}} \sum_{n=1}^{\text{No. of harmonics}} \left| V_{m,n}^{\text{initial}} - V_{m,n}^{\text{temp}} \right| < \text{Tolerance}$$

Eqn 3-33

In the simulation the tolerance used was set at $1 \times 10^{-9}$ and the number of harmonics including DC was 10 instead of 5 which has been assumed in the above paragraphs for simplicity. When Eqn 3-33 is met, the node voltages can then be used to determine the final $V_{\text{non}}(t)$ and the harmonic-balance part of the simulation is considered completed.

**Section 3-5b: Determination of Conversion Gain using Conversion Matrices**

After $V_{\text{non}}(t)$ is obtained, the small-signal conductance conversion matrices for the two diodes, $D_F$ and $D_{REC}$, can be calculated as explained in Section 3-4. Once the conversion matrices are known, the determination of the conversion gain of the HBT model is a purely linear operation and no longer requires the knowledge of the LO and other bias voltages and currents.

The equivalent circuit in Figure 3-4 needs to be modified and is shown in Figure 3-6.

Several changes have been made. The bias voltage and LO sources have been removed from Figure 3-6. Also the diodes $D_F$ and $D_{REC}$ have been replaced by two conductances characterised by the conversion matrices $G_F$ and $G_{REC}$, respectively. New notations with first letters in small case have been used for all the voltage, current and impedance matrices to coincide with the small-signal nature of the present analysis.
Let $f_{RF}$, $f_{LO}$ and $f_{IF}$ denote the laser modulation frequency (RF), LO frequency and IF frequency, respectively, and $f_{RF} = f_{IF} + f_{LO}$. Therefore rewriting Eqn 3-22, the corresponding mixing frequency sequence is given by

$$f_k = f_{RF} + kf_{LO}$$  \hspace{1cm} \text{Eqn 3-34}$$

where $-\infty \leq k \leq \infty$. Since fundamental mixing components are stronger than higher order mixing components, it can be assumed that $|k| \leq 2$ so that the number of frequencies considered is only five. All voltages and currents are therefore specified using 5-by-1 column matrices and constructed as explained in Section 3-4, in particular Eqn 3-22 and Eqn 3-25. The first elements of these column matrices correspond to their values at frequency $f_{-2}$ defined in Eqn 3-34, the second elements correspond to their values at frequency $f_{-1}$ and so on.
The only excitation in this equivalent circuit in Figure 3-6 is the primary photogenerated RF current given by the matrix \( i_{\text{prim RF}} \). Let \( i_{\text{prim}} \) be the AC peak primary photocurrent generated in the HBT which is related to the DC primary photocurrent through the light modulation index. \( i_{\text{prim RF}} \) is formulated as

\[
    i_{\text{prim RF}} = \begin{bmatrix}
        0 \\
        0 \\
        0 \\
        j \cdot i_{\text{prim}} / 2 \\
        0
    \end{bmatrix}
\]  
Eqn 3-35

The AC peak primary photogenerated RF current appears as the 4\(^{th}\) element of the matrix \( i_{\text{prim RF}} \) because \( f_{\text{RF}} = f_{\text{if}} + f_{\text{LO}} \) which corresponds to \( k=1 \) and appears as the 4\(^{th}\) element in the current column matrix (see Eqn 3-25). The factor \( j/2 \) is included because, as implied in Eqn 3-16 and Eqn 3-17, the current matrices and the voltage matrices only represent half of their full Fourier series.

Impedances of the linear elements and the base transport factor are represented by 5-by-5 diagonal matrices because they do not produce mixed products, as opposed to the conversion matrices. The first elements on the diagonals of these matrices correspond to their values at frequency \( f_{\pi} \) defined in Eqn 3-34, the second elements correspond to their values at frequency \( f_{-1} \) and so on.

Since there are five nodes in the equivalent circuit in Figure 3-6 and each node requires a current error function matrix, therefore five current error function matrices are formulated. Suppose the five node voltage matrices are \( \mathbf{v}_1, \mathbf{v}_2 \) and so on, the current error function matrix at node 4, for example, is given by
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\[
\text{ierr}_4 = \begin{bmatrix} 
\text{ierr}_{-2,4} \\
\text{ierr}_{-1,4} \\
\text{ierr}_{0,4} \\
\text{ierr}_{1,4} \\
\text{ierr}_{2,4} 
\end{bmatrix}
= (z_{BE}^{-1} + G_{REC} + G_F)(v_4 - v_3) + z_{BE}^{-1}(v_4 - v_1) + z_{CB}^{-1}(v_4 - v_5) - \alpha G_F(v_4 - v_3) - iprim_RF
\]

Eqn 3-36

After combining all five 5-by-1 current error function matrices into one 25-by-1 column matrix, whose first five elements are those of \( \text{ierr}_1 \) and so on, a Jacobian can be calculated and all the node voltages can be readily found using Eqn 3-31.

Finally the down- and up-converted signal powers delivered to the output 50 \( \Omega \) load can be calculated from the third and the fifth elements of \( v_2 \). Simulation results are presented in the next section.

Section 3-5c: Simulation Results

Computer simulations have been carried out for the three-terminal HBT electrically pumped optoelectronic mixer using the techniques discussed. The simulations were implemented in Matlab using the parameters given in Table 3-2. The number of iterations required to achieve a solution at each bias point for a tolerance of \( 1 \times 10^{-9} \) (Eqn 3-33) and with the initial values all set to zeros, was from 33 for low LO powers and small \( V_{BE} \), to over 1000 for high LO powers and large \( V_{BE} \). Running on a Cyrix P166 computer, similar to a Pentium 166, the corresponding simulation time at each bias point ranged from 3 seconds to 4 minutes.

The system conversion gains, which were defined in Section 1-2, for both down- and up-conversions, were computed and shown in Figure 3-7 and Figure 3-8, respectively, as a function of the base-emitter voltage and the LO power. The up-conversion frequency is 6.9 GHz, and the down-conversion frequency is 100 MHz IF.
For both down- and up-conversions at low LO powers, the system conversion gains initially increase with $V_{BE}$, reach a maximum and then decrease again. These characteristics can be understood by noting that the conversion gains depend both on how strongly dependent the current gain of the HBT on $V_{BE}$ (or $I_C$) is and the magnitude of the current gain itself. It is recalled from Section 2.6 that at small $V_{BE}$, the common-emitter current gain of HBT increases with $V_{BE}$ due to the effects of the base-emitter space-charge recombination current and is therefore strongly dependent on $V_{BE}$. Thus the HBT is very nonlinear and the conversion gains increase with $V_{BE}$ at low bias. At higher $V_{BE}$, the current gain becomes less dependent on $V_{BE}$, making the HBT more linear and mixing less efficient, and hence the conversion gains decrease with $V_{BE}$. An optimum $V_{BE}$ therefore exists for the highest conversion gain at small LO powers and is observed in Figure 3-7 and Figure 3-8. At higher LO powers, the voltage swing of $V_{BE}$ by the LO is so great that within one cycle of the LO, both the linear and nonlinear operating points of the HBT are involved and the conversion gains are less dependent on $V_{BE}$ because this
Figure 3-7: Simulated HBT system down-conversion gain. LO: 3.4 GHz, RF: 3.5 GHz, down-converted frequency: 100 MHz, VCE: 1.5 V, mean incident optical power: 70μW, modulation depth: 90 %, P_{RF,Elec}: -38 dBm, external quantum efficiency: 17.6 %.

Figure 3-8: Simulated HBT system up-conversion gain. LO: 3.4 GHz, RF: 3.5 GHz, up-converted frequency: 6.9 GHz, VCE: 1.5 V, mean incident optical power: 70μW, modulation depth: 90 %, P_{RF,Elec}: -38 dBm, external quantum efficiency: 17.6 %.

Voltage swing by the LO is now large compared to the range of $V_{BE}$ varied. The difference between the down- and up-conversion characteristics are due to the different frequency responses at the two down- and up-conversion frequencies.
Detailed comparisons between the simulation and experimental results will be presented in Chapter 5.

Section 3-6: Implementation and Simulation Results for a Two-Terminal Edge-Coupled HBT Optoelectronic Mixer

The main difference between the simulation implementation for the previous three-terminal HBT and for the two-terminal HBT is in the equivalent circuit model used. In fact, the model used for the two-terminal HBT is not a physical model but an empirical one which describes the terminal quasi-static behaviour of the HBT. The use of such a model rather than a more physics based one is mainly due to the lack of an electrical base terminal which would otherwise allow the transistor internal parameters to be measured and used in an equivalent circuit model similar to Figure 3-4. Apart from this, the implementation of the simulation for this two-terminal HBT is similar to the one described in the previous section.

To devise a model which describes the quasi-static HBT terminal properties, the static IV characteristics under a constant optical bias were first measured and are shown in Figure 3-9.

![Static IV characteristics of the two-terminal HBT under constant optical bias. Incident optical power: 0.3 mW. Markers are experimental data points while the curve is fitted.](Image)
These experimental data are then curve fitted so that the HBT IV can be specified analytically in the simulator. The range of the bias voltages used in the measurement was limited to \(-1V \leq V_{CE} \leq 1V\) to avoid device breakdown. The fitting equation, however, also predicts the HBT IV beyond this range, although this prediction has not been verified experimentally because the two-terminal HBT can be easily damaged even if \(V_{CE}\) is only slightly higher than 1 V.

The circuit model used for the harmonic-balance simulation is shown in Figure 3-10.

![Figure 3-10: A harmonic-balance model for the two-terminal HBT.](image)

It can be seen that the HBT as a whole is modelled as a single nonlinear element. The terminal behaviour of the HBT under the LO excitation at frequency \(f_{LO}\) is described simply by the transistor DC characteristics in Figure 3-9, and therefore Figure 3-10 is a quasi-static model. The LO, driving the HBT emitter at a few GHz, will see a different input impedance from that predicted from Figure 3-9. As a result, this method of modelling the HBT using its static terminal characteristics is accurate for frequencies below a few MHz and less accurate in the microwave region.

A large-signal harmonic-balance simulation is then carried out as already described in detail in the three-terminal HBT case. The result of this part of the simulation is the time-
varying voltage waveform at node 1 which allows the conversion matrix of the HBT to be determined.

The construction of the conversion matrix for this two-terminal HBT is different from that for the previous three-terminal HBT. This is due to the fact that the primary photocurrent is not explicitly represented as a current source in Figure 3-10 but generated from within the nonlinear HBT quasi-static model. So if the HBT symbol in Figure 3-10 was simply replaced by a conductance characterised by a small-signal conductance conversion matrix, as has been done for the two diodes in the three-terminal HBT equivalent circuit, there would be no current source for the primary photogenerated RF current.

The way to overcome this problem is to describe the two-terminal HBT in two separate stages. The first stage describes the photodetection and primary photocurrent generation while the second stage consists of a voltage-dependent current-controlled current source which effectively describes the photocurrent gain of the HBT. This idea is illustrated in Figure 3-11.

![Figure 3-11: Equivalent circuit for two-terminal HBT conversion gain calculation.](image)

The current-controlled current source is specified by a small-signal photocurrent gain conversion matrix $G$ at the laser modulation frequency $RF$. To construct $G$, the HBT small-signal photocurrent gain was measured and is shown in Figure 3-12.
Again curve fitting is used to implement these characteristics analytically in the simulator. Since the HBT is pumped by a large LO, the small-signal photocurrent gain is therefore modulated at the LO frequency, and can be described by a Fourier series (Eqn 3-12). The small-signal photocurrent gain conversion matrix is then constructed using the Fourier series coefficients according to Eqn 3-25.

The system conversion gains of the two-terminal HBT optoelectronic mixer can be readily found using the techniques described in Section 3-5b. However, it can be expected that the simulated down- and up-conversion gains using the models in Figure 3-10 and Figure 3-11 will be virtually identical because it is the frequency dependent parameters in the equivalent circuits that affect the down- and up-converted signals differently. In Figure 3-10 and Figure 3-11, those parameters are the impedances of the inductor and capacitor of the bias network. In the present frequency range of interest, the inductor and the capacitor can be treated as an open and a short circuit, respectively, and therefore both down- and up-conversions are the same.

The simulation parameters used are shown in Table 3-3.
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<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_{RF}$</td>
<td>3.025 GHz</td>
</tr>
<tr>
<td>$f_{ID}$</td>
<td>100 MHz</td>
</tr>
<tr>
<td>$f_{LO}$</td>
<td>2.925 GHz</td>
</tr>
<tr>
<td>Mean incident optical power</td>
<td>0.3 mW</td>
</tr>
<tr>
<td>Light modulation index</td>
<td>10 %</td>
</tr>
<tr>
<td>$i_{prem}$</td>
<td>37.5 mA</td>
</tr>
<tr>
<td>No. of harmonics including DC in harmonic balance</td>
<td>10</td>
</tr>
<tr>
<td>No. of frequencies in conversion matrix calculation</td>
<td>5</td>
</tr>
<tr>
<td>No. of LO cycles simulated</td>
<td>5</td>
</tr>
<tr>
<td>No. of simulation points per iteration</td>
<td>512</td>
</tr>
<tr>
<td>Splitting coefficient $s$</td>
<td>from 0.01 to 0.5</td>
</tr>
<tr>
<td>Tolerance</td>
<td>$1 \times 10^{-9}$</td>
</tr>
</tbody>
</table>

Table 3-3: Parameters used in the modelling of the two-terminal HBT electrically pumped optoelectronic mixer

The simulated system conversion gain for both down- and up-conversions is shown in Figure 3-13.

![Figure 3-13: Simulated system conversion gain for both down- and up-conversions of the two-terminal edge-coupled HBT optoelectronic mixer. Average incident optical power: 0.3 mW, laser modulation frequency (RF): 3.025 GHz, LO freq.: 2.925 GHz, down-conversion freq.: 100 MHz, up-conversion freq.: 5.95 GHz.](image)
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The up-conversion frequency is 5.95 GHz and the down-conversion frequency is 100 MHz IF. In general, the system conversion gain initially increase with $V_{CE}$, reach a maximum and then decrease with $V_{CE}$. The reason for these characteristics can be understood by referring to Figure 3-12. At low $V_{CE}$, the small-signal current gain of the two-terminal HBT is strongly dependent on and increases with $V_{CE}$, and therefore the system conversion gains increase correspondingly. At higher $V_{CE}$, the small-signal current gain stays large but becomes less dependent on Therefore the mixing process is less efficient and the system conversion gains decrease at very high $V_{CE}$. For low LO powers, it can be seen that the system conversion gains actually decrease with $V_{CE}$ initially before increasing again. This characteristic can be explained in terms of the HBT terminal impedance at around 0.1 V $V_{CE}$. Referring to Figure 3-9 which shows the two-terminal HBT IV characteristic under constant optical illumination. At around $V_{CE} = 0.1$ V, the gradient of the IV curve is high compared to other parts of the plot. Therefore the HBT exhibits a very low input impedance seen by the LO and as a result, the voltage swing developed by the LO across the HBT collector-emitter terminal is small. Consequently, the system conversion gains are also small at this bias point.

These simulation results will be compared with experimental results for the two-terminal HBT Chapter 4.

Section 3-7: Conclusion

A novel use of the harmonic-balance techniques and the concept of conversion matrices have been described and applied to the modelling of two HBT electrically pumped optoelectronic mixers. The three-terminal HBT was represented by a T-topology equivalent circuit and the only nonlinear elements included in the model were the two base-emitter junctions mainly for simplicity reasons. On the other hand, the two-terminal HBT was represented by a quasi-static model which describes the device terminal behaviour. The use of the quasi-static model was due to the lack of an electrical base terminal which would otherwise allow the transistor internal parameters to be measured and used for the construction of a more physical equivalent circuit model.
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The simulation results for the two mixers have been presented and commented on. The accuracy of the two HBT mixer models will be revealed in the following two chapters where the experimental results will be presented and comparisons made.
References


Chapter Four

Experimental Results for a Two-Terminal Edge-Coupled HBT Electrically Pumped Optoelectronic Mixer

Section 4-1: Introduction

Although early work on heterojunctionphototransistors was concentrated on the two-terminal type, i.e., no electrical base connection (see Chapter 1), to date all HBT-based optoelectronic mixers reported by other researchers have only used three-terminal HBTs. One of the reasons is that with a third terminal, the base of the HBT in electrically pumped mixing can be driven by the LO while the output signals can be separately obtained from the collector with the emitter being the signal ground. Therefore, electrically pumped optoelectronic mixing experiments can be carried out in a very simple configuration. If, on the other hand, electrically pumped optoelectronic mixing is to be carried out with a two-terminal HBT, either the emitter or the collector terminal can be used for simultaneous LO injection and output signal measurement. Consequently, frequency separation in the form of a filter needs to be provided at that terminal.

In this Chapter, a novel two-terminal edge-coupled HBT electrically pumped optoelectronic mixer [1] will be described and experimental results presented. The performance of this optoelectronic mixer will be compared with others reported in the literature and with the simulation results of Chapter 3.

The HBT samples were provided by D. Wake of British Telecom Laboratories (BT Labs). Although the full structural and fabrication details have been published by Wake et al [2] elsewhere, the relevant information about the device will be included here for completeness.

The order of this Chapter is as follows. In Section 4-2, the fabrication and the structure of the HBT as well as the experimental arrangement will be described. In Section 4-3, static IV characteristics of the HBT illuminated with different optical powers will be
given. In Section 4-4, the HBT frequency response to intensity modulated light in direct photodetection will be presented. In Section 4-5, the experiments on electrically pumped optoelectronic mixing will be described and the results presented and compared with those from the harmonic-balance modelling of Chapter 3. In Section 4-6, the noise performance characteristics will be given. In Section 4-7, the two-tone third-order intermodulation distortion characteristics measured using a two-laser approach will be described and the results will be used to calculate the spurious-free dynamic range. Finally a conclusion will be given in Section 4-8.

Section 4-2: Device Structure and Experimental Arrangement

The cross-sectional view of the HBT structure is shown in Figure 4-1. The HBT was grown by MOVPE on an $N^+$ substrate and consisted of a collector (InGaAs, $n = 10^{16} \text{ cm}^{-3}$, $t = 0.4 \mu\text{m}$), an electrically floating base (InGaAs, $p = 10^{19} \text{ cm}^{-3}$, $t = 0.1 \mu\text{m}$) and an emitter (InP, $n = 5 \times 10^{17} \text{ cm}^{-3}$, $t = 0.15 \mu\text{m}$) where $n$ and $p$ are the $n$- and $p$-type doping concentrations, respectively, and $t$ is the thickness. The $p$-type dopant was zinc and the $n$-type dopant was sulphur. A thin spacer layer (InGaAs, $n = 10^{16} \text{ cm}^{-3}$, $t = 0.02 \mu\text{m}$) between the base and emitter was used to reduce the effects of out-diffusion of the base dopant and a highly doped emitter contact layer (InGaAs, $n = 10^{19} \text{ cm}^{-3}$, $t = 0.05 \mu\text{m}$) was also included. The finished device area was $5 \times 10^2 \mu\text{m}^2$ with the narrow dimension of the device illuminated.

![Figure 4-1: Cross-section of the two-terminal edge-coupled HBT.](image)

In this HBT, light access is from the side and in parallel with the epilayers of the device, as compared to other conventional top-illuminated or back-illuminated phototransistors.
with the direction of light normal to the device epilayers. Light of 1550 nm wavelength is absorbed in the base and the collector. The primary photocurrent serves as the base bias as in a normal bipolar transistor.

Since the lateral dimension of this HBT is 10 \( \mu \text{m} \) long, the length available for optical absorption is much greater than in the conventional, top-illuminated or back-illuminated HBTs, where the length available for the optical absorption is the total thickness of the base and the collector, usually less than 1 \( \mu \text{m} \). Since the absorption coefficient for InGaAs at 1550 nm wavelength is around 1 \( \mu \text{m}^{-1} \) [3], Eqn 2-1 shows that almost all photons coupled into the base and the collector of this edge-coupled HBT can be absorbed and the external quantum efficiency of this HBT is mainly limited by coupling loss.

The HBT was originally delivered to UCL on a Kyocera laser mount. To perform measurements with the HBT, the Kyocera mount was subsequently secured in custom-designed brass mount and a short 50 \( \Omega \) characteristic impedance microstrip line was used to connect the Kyocera mount emitter contact to an external SMA connector. The collector was connected, through the Kyocera mount and the brass, to the signal ground. External bias and AC signal coupling were accomplished using an HP 11612A bias-T having a frequency range from 45 MHz to 26.5 GHz. The whole HBT package was then put on an \( x-y-z \) micro-positioner so that the HBT could be easily aligned with the laser beam.

Figure 4-2 shows the basic experimental arrangement for all measurements, which used free-space optics.

The light sources in Figure 4-2 are two very low noise GEC-Marconi LD6804 60mW lasers emitting at around 1550 nm wavelength and biased using HP 11612A bias-Ts. The two lasers were operated simultaneously only for the two-tone third-order intermodulation distortion product measurements in which the wavelengths of the two lasers were sufficiently offset so that the beat signal (above 20 GHz) did not interfere with the measurements at lower frequencies. For other optoelectronic measurements, only laser 1 was required. The collimating and focusing lenses were Newport F-L40B
Unlike a normal three-terminal electrical bipolar transistor, the only available electrical terminal from this HBT is the emitter (with the collector grounded). If the HBT is used simply as a direct photodetector, the signal output can be measured from the emitter directly. However when this two-terminal HBT is operated as an electrically pumped optoelectronic mixer, both the injection of the LO source and measurement of the mixer output signals can only be done through the single emitter terminal. As was mentioned in the introduction in this Chapter, a filter is required to separate the LO and other outputs from the HBT. A diplexer was therefore designed and fabricated to accomplish this task. The diplexer design was based on a microstrip coupled-line structure and consisted of a bandpass section and a bandstop section. The LO source signal was injected to the emitter of the HBT via the bandpass section and the HBT output signals were retrieved.
from the emitter via the bandstop section. Therefore it is important that both the bandpass and bandstop centre frequencies are the same so that the LO is sufficiently isolated from bandstop port. The original design centre frequency was 3 GHz. The diplexer was synthesised using procedures similar to those found in [4]. The final dimensions of each coupled-line section were calculated assuming TEM-type propagation [5]. The diplexer was built on a 1.3 mm thick RT/Duroid-6006 board with a quoted dielectric constant of 6.15. The physical layout of the actual diplexer is shown in Figure 4-3.

![Diagram of diplexer layout](image)

**Figure 4-3: Layout of the microstrip coupled-line diplexer.**

The finished diplexer required some post-fabrication lengthening of the two open-stubs in the bandstop section so that all the resonance frequencies were closely matched. A network analyser was used to measure its bandpass and bandstop frequency responses which are shown in Figure 4-4.

![Frequency response chart](image)

**Figure 4-4: Frequency response of the diplexer.**
The measured centre frequency of the diplexer was 2.925 GHz rather than 3 GHz as originally expected. Possible reasons for this discrepancy are the small tolerance variation in the quoted dielectric constant of the RT-Duroid board used and the TEM-type propagation assumption. The diplexer bandpass section had about 2.5 dB loss at 2.925 GHz.

After all the components shown in Figure 4-2 were put in place, the optoelectronic characteristics of the HBT were measured and will be presented in the following sections.

**Section 4-3: Static Photodetecting Characteristics**

Figure 4-5 shows the HBT static IV characteristics for different mean incident optical power levels.

![Figure 4-5: Two-terminal HBT static IV characteristics for different mean incident optical power levels. \( P_{\text{opt}} \): mean incident optical power.](image)

The mean incident optical power was measured in front of the HBT focusing lens using a large area optical power meter. It can be seen that the characteristics in Figure 4-5 resemble those of a conventional bipolar transistor with the incident optical power acting as the base bias current. Indeed for an ideal two-terminal bipolar phototransistor, no current should flow without optical illumination except some small leakage currents. The photons absorbed in the base and the collector regions generate electron-hole pairs. Those electron-hole pairs generated in the base-collector depletion region are
immediately separated by the junction electric field. The electrons and holes generated in
the neutral base and collector, respectively, diffuse into the base-collector depletion
region. Since the movement of these charged carriers constitute a primary photocurrent
and is analogous to having a current source connected between the transistor base and
collector, the incident optical power thus generates the base bias current.

Figure 4-5 can be re-plotted in terms of the HBT static responsivity shown in Figure 4-6.

![Figure 4-6: Two-terminal HBT static responsivity for different mean incident optical power levels.](image)

For a 100 % quantum efficient photodiode at 1550 nm wavelength, the responsivity is
1.25 A/W. By dividing the responsivity in Figure 4-6 by 1.25 A/W, the photocurrent gain
of the HBT can be determined. For example, at \( P_{opt} = 0.18 \text{ mW} \) and \( V_{CE} = 1 \text{ V} \), the static
photocurrent gain was about 84.

It is seen that the responsivity, and hence the photocurrent gain, of the HBT was a
function of the incident optical power. The dependence of the responsivity on the
incident optical power is a result of the base-emitter junction recombination current as
explained in Chapter 2. For example at \( V_{CE} = 1 \text{ V} \), the responsivity and the photocurrent
gain increased from 60 A/W and 48, respectively, at \( P_{opt} = 0.06 \text{ mW} \) to 105 A/W and 84,
respectively, at \( P_{opt} = 0.18 \text{ mW} \).
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It should be noted that the incident optical power was not equal to the optical power absorbed because of the coupling loss and the internal quantum efficiency of the HBT. In order to measure the external quantum efficiency, which includes the effects of the coupling loss and the internal quantum efficiency, it would be necessary to set the base-emitter junction voltage to 0 V so that the HBT would simply act as a p-i-n photodiode without any transistor gain. The external quantum efficiency would then be calculated from the measured primary photocurrent and the incident optical power. However no electrical base terminal is available for setting $V_{BE} = 0$ V in the present two-terminal HBT and the current measured was always the amplified version of the primary photocurrent. Since the internal current gain of the HBT is not known exactly, it was not possible to measure directly its external quantum efficiency. An indirect way of determining the external quantum efficiency is to use an edge-coupled p-i-n photodiode whose structure is similar to the base-collector junction of the HBT. Since the external quantum efficiency of the edge-coupled p-i-n photodiode can be measured readily, the measured external quantum efficiency of the p-i-n photodiode can be treated as an estimated value of that of the HBT. With a lens ended single-mode fibre with a 10 μm radius, Wake et al. [2] have estimated the external quantum efficiency of the HBT to be around 50 % using the described procedure.

Section 4-4: Frequency Response to Intensity Modulated Light in Direct Detection

To measure the frequency response of the HBT direct photodetector to intensity modulated light, the diplexer was not required and was removed from the experimental arrangement in Figure 4-2. Laser 1 was directly modulated with a microwave signal swept from 45 MHz to 6 GHz by the output port of a network analyser. However the modulation response of the laser was not flat and therefore had to be measured first against a wideband photodetector. The wideband photodetector used was a New Focus 1014 InGaAs Schottky photodiode with a stated 3 dB bandwidth of 45 GHz. Assuming the frequency response of this New Focus 1014 is flat from 45 MHz to 6 GHz, the modulation depth of the output intensity of laser 1 was measured. Knowing the mean incident optical power on the HBT and the corresponding modulation depth allowed the
equivalent electrical input power $P_{RF\_Elec}$ (defined in Section 1-2) to be determined. Keeping the laser bias and modulation conditions unchanged, the output power of the HBT was measured by the input port of the network analyser. The frequency response of the HBT in terms of the system signal gain (also defined in Section 1-2) is shown in Figure 4-7 for different collector currents controlled by varying the mean incident optical power.

![Figure 4-7: HBT frequency response to intensity modulated light. $V_{CE} = 1$ V.](image)

At low frequencies, the system signal gain increased with increasing collector current because of the base-emitter junction recombination current as explained in Chapter 2. The 3 dB bandwidth also increased with the collector current because of the reduced emitter dynamic resistance. The 3 dB electrical bandwidth was about 280 MHz at 15 mA collector current and 1V $V_{CE}$.

**Section 4-5: Electrically Pumped Optoelectronic Mixing Conversion Gain Measurements and Comparison with the Simulation Results**

This section describes the electrically pumped optoelectronic mixing experiments with the two-terminal edge-coupled HBT. The experimental results will be compared with the
simulation results of Chapter 3 and other HBT optoelectronic mixers reported in the literature.

To operate the HBT as an electrically pumped optoelectronic mixer, the emitter of the HBT was pumped by an LO through the bandpass section of the diplexer. The LO frequency was set at 2.91 GHz, within the bandwidth of the diplexer bandpass section. The LO power was varied from 0 dBm to 10 dBm. Laser 1 was directly modulated with an RF signal at 2.41 GHz. The mean incident optical power on the HBT was 0.14 mW and the light modulation depth was monitored by the LSA. The HBT photodetected the intensity modulated light and mixed it with the LO. The down-converted signal (IF) power at 500 MHz was measured through the diplexer bandstop section with a spectrum analyser. With 0.14 mW mean incident optical power and 51% light modulation depth, the equivalent electrical input power $P_{RF,Elec}$ was calculated to be $-37$ dBm. Figure 4-8 shows the measured system conversion gain (defined in Section 1-2) versus $V_{CE}$ in down-conversion for different LO powers.

![Figure 4-8: HBT optoelectronic mixer system down-conversion gain.](image)

In general, the system conversion gain was an increasing function of the applied LO power. The highest system conversion gain was +7 dB at $V_{CE} = 0.6$ V and $LO = 10$ dBm.
This figure is significant because it shows that such a single two-terminal HBT optoelectronic mixer would have a 7 dB gain advantage over an optoelectronic mixer built using a hypothetical 100 % quantum efficient photodetector and an ideal microwave mixer with 0 dB conversion loss in a 50 Ω impedance system. With reference to Table 1-1, this two-terminal HBT has also achieved the highest system conversion gain ever reported for a single HBT optoelectronic mixer although the frequencies used were different. Furthermore the LO power level used to achieve this 7 dB gain is also comparable to that required to drive most commercially available microwave double-balanced mixers.

Optoelectronic mixing has also been carried out with a second HBT sample, following the failure of the original one, and it is this HBT that was modelled in Chapter 3. In this second optoelectronic mixing experiment, the LO was set at 2.925 GHz and laser 1 was directly modulated at 3.025 GHz. Both down-conversion product at 100 MHz and up-conversion product at 5.95 GHz were measured through the bandstop section of the diplexer with a spectrum analyser. With 0.3 mW mean incident optical power and 10 % modulation depth, $P_{RF,elec}$ was calculated to be $-44.5$ dBm.

Figure 4-9 and Figure 4-10 show the system down- and up-conversion gains versus $V_{CE}$ for different LO powers.

![Figure 4-9: System down-conversion gain versus $V_{CE}$ for different LO powers. RF = 3.025 GHz, LO = 2.925 GHz, IF = 100 MHz, $P_{opt} = 0.3$ mW, $P_{RF,elec}$ = -44.5 dBm.](image)
For \( \text{LO} = 10 \, \text{dBm} \), the maximum system down- and up-conversion gains using this second HBT were \(-1\, \text{dB} \) at \( V_{CE} = 0.6 \, \text{V} \) and \( V_{CE} = 0.8 \, \text{V} \), respectively. The system conversion gains of this second transistor were several dBs smaller than that of the previous one mainly because of a lower responsivity. For example, the static responsivity of the second HBT was \( 37.5 \, \text{A/W} \) at \( V_{CE} = 1 \, \text{V} \) and \( P_{op} = 0.18 \, \text{mW} \). However under the same bias conditions, the first HBT responsivity was \( 105 \, \text{A/W} \). The lower responsivity could in turn be due to a higher base-emitter junction recombination current, although this could not be verified experimentally without an electrical base contact for constructing a Gummel Plot.

In Section 3-6, this two-terminal HBT electrically pumped optoelectronic mixer was modelled using a quasi-static model to describe the transistor terminal behaviour under LO excitation and optical illumination. The mixing mechanism was attributed to the small-signal photocurrent gain dependence on the HBT terminal voltage \( V_{CE} \). Such a photocurrent gain dependence was shown in Figure 3-12 and is re-plotted in Figure 4-11.

According to the assumption that the mixing mechanism is the dependence of the small-signal photocurrent gain on \( V_{CE} \), then the mixing should be most efficient where the slope of Figure 4-11 is the steepest. The steepest slope occurs in the range \( 0.6 \, \text{V} < V_{CE} < 0.9 \, \text{V} \).
Figure 4-11: HBT small-signal photocurrent gain dependence on $V_{CE}$. Gain measured at 3.025 GHz, $P_{opt} = 0.3$ mW.

and corresponds well to where the highest system down- and up-conversion gains were measured.

Another interesting feature seen in both the down- and up-conversion gain characteristics is that for low LO power levels, the conversion gains initially dipped to a minimum as $V_{CE}$ was increased from 0 V before the conversion gains increase again. Such a phenomenon can be explained in terms of the HBT terminal impedance. Figure 4-12 shows a plot of the second HBT emitter current versus $V_{CE}$ which was used to describe the HBT terminal behaviour in the model in Chapter 3.

Figure 4-12: HBT emitter current versus $V_{CE}$, $P_{opt} = 0.3$ mW.
It can be seen that as $V_{ce}$ is increased from 0 V, the slope of the plot increases and reaches a maximum at around 0.2 V which, in turn, represents a minimum terminal impedance presented by the HBT. As a result of this small terminal impedance, the voltage swing caused by the LO at the emitter and thus the conversion gains were reduced at around $V_{ce} = 0.2$ V. Figure 4-13 shows the peak AC voltage swing at the emitter at the LO frequency calculated from the large-signal harmonic-balance model in Chapter 3. Indeed a minimum voltage swing at around $V_{ce} = 0.2$ V is predicted by the model.

![Peak ac voltage swing at emitter at LO frequency determined by the model in Chapter 3. LO = 0 dBm.](image)

The measured optoelectronic mixing system conversion gain characteristics shown in Figure 4-9 and Figure 4-10 are compared with the simulation results from Chapter 3 in Figure 4-14 and Figure 4-15.
It can be seen that the simulation results show good qualitative agreement with the experimental ones. In particular, the model was able to predict that both down- and up-conversion gains would initially decrease as $V_{ce}$ was increased from 0 V, and reached a maximum in the range $0.6 \text{ V} < V_{ce} < 0.9 \text{ V}$.
However, because of the assumption in the model that the mixing mechanism is the small-signal photocurrent gain dependence on $V_{ce}$, the model did not distinguish between down-conversion and up-conversion, and therefore the simulation results were virtually identical in both processes. Clearly some differences were observed in the measured system down- and up-conversion gains, especially for $V_{ce} > 0.8$ V where the measured system down-conversion gain started to increase with $V_{ce}$ while the measured system up-conversion did not. This difficulty is due to the high-level assumption used in the model for generating the mixed products. Physically, the mixed products were generated by the nonlinear base-emitter and base-collector junctions of the HBT. Since the down-conversion and up-conversion products were a few GHz apart, the base-emitter and base-collector capacitances would affect the phases and amplitudes of the currents at the down- and up-conversion frequencies differently, and hence different down- and up-conversion characteristics would result. The model of Chapter 3, however, did not take this factor into account.

The discrepancies between the experimental and simulation results are partly due to the reason given in the last paragraph, and partly due to the quasi-static nature of the model. The terminal behaviour of the HBT was modelled by incorporating the transistor static IV characteristics under optical illumination. Clearly when the LO was at a few GHz, the terminal impedance presented by the HBT to the LO source would be different from what is predicted simply from the inverse of the slope of the plot in Figure 4-12 because of the HBT junction capacitances. As a result the voltage swing caused by the LO at the emitter would not be accurately predicted by the quasi-static model and hence further discrepancies between the experimental and simulation were caused.

It is recalled that the reason for using a quasi-static model in the simulation was the lack of an electrical base connection in this two-terminal HBT. Such a connection would allow a 2-port s-parameter and Gummel plot measurements to be carried out so that the HBT internal parameters could be extracted and a more physical equivalent circuit model could be constructed. The use of such a quasi-static model has, however, provided a simple way of predicting the conversion characteristics of the two-terminal HBT electrically pumped optoelectronic mixer.
**Section 4-6: HBT Noise Performance Measurements**

In this section, measurements of the noise performance of the two-terminal HBT direct photodetector and the electrically pumped optoelectronic mixer will be described and results presented. One important issue in such noise measurements on photodetectors is the noise originating from the optical source used. The noise performance of a laser is characterised by its relative intensity noise (RIN). Suppose a 100 % quantum efficient photodiode is illuminated by a laser beam which exhibits an optical power fluctuation. The photodiode is attached to a load $R_{\text{load}}$, as shown in Figure 4-16, which is at 0 K temperature so that the thermal noise does not need to be considered.

![Figure 4-16: A simple model illustrating the concept of the laser relative intensity noise. PD: 100 % quantum efficient photodiode.](image)

The optical power fluctuation will translate to an electrical noise power delivered to the attached load after photodetection. The ratio of the total noise power spectrum density $S_{\text{noise}}$ to the mean DC power $P_{\text{DC}}$ delivered to the load can be written as

$$\frac{S_{\text{noise}}}{P_{\text{DC}}} = \frac{(i_{\text{noise}})^2}{I_{\text{DC}}^2} = \frac{(i_{\text{RIN}})^2}{I_{\text{DC}}^2} + \frac{(i_{\text{shot}})^2}{I_{\text{DC}}^2}$$

Eqn 4-1

where $I_{\text{DC}}$ is the DC photocurrent, $(i_{\text{noise}})^2$, $(i_{\text{RIN}})^2$ and $(i_{\text{shot}})^2$ are the mean square spectral densities of the total noise current, the noise current due to the laser RIN and the shot noise current, respectively. The RIN is defined as

$$\text{RIN} = \frac{(i_{\text{RIN}})^2}{I_{\text{DC}}^2}$$

Eqn 4-2

and substituting the following well-known equation for the shot noise
into Eqn 4-1, the RIN can be expressed as

\[
\text{RIN} = \frac{S_{\text{noise}}}{P_{\text{DC}}} - \frac{2q}{I_{\text{DC}}}
\]

Eqn 4-4 shows that the definition for RIN excludes the contribution from shot noise.

A laser having a very high RIN makes measurements of the noise generated by the photodetector itself unreliable or even impossible. Previously a rather high RIN laser (RIN = \(-135\) dB/Hz at 110 MHz and \(-130\) dB/Hz at 2.81 GHz) was used as the optical source in the noise measurements of the HBT [6]. It was observed and confirmed by calculations that the output noise powers of the HBT both as a direct photodetector and an optoelectronic mixer were mainly due to the high laser RIN and therefore the noise generated by the HBT itself could not be measured accurately.

In the present measurements, the optical source used (laser 1 in Figure 4-2) was a very low RIN GEC-Marconi LD6804 strained layer quantum well DFB laser diode. The RIN of the laser was first measured with an HP70000/HP70810B LSA from 50 MHz to 25 GHz and is shown in Figure 4-17.

![Figure 4-17: Laser 1 relative intensity noise (RIN) at 140 mA bias current. Total output optical power \(\equiv 26\) mW.](image)
The laser had a RIN of better than $-160$ dB/Hz below 8 GHz which is at least 25 to 30 dB less noisy than the previous laser [6]. Such a low RIN value was obtained with a 140 mA bias current with the total output optical power from the laser approximately at 26 mW. Because of the large optical power, optical attenuators were used to reduce and adjust the amount of incident optical power on the HBT.

To improve the sensitivity of the system in these noise measurements, a preamplifier was used and inserted before the spectrum analyser.

With laser 1 biased at 140 mW unmodulated, the output noise power spectral densities (PSDs) spectra of the HBT direct photodetector were measured for different mean incident optical powers $P_{\text{opt}}$ and are shown in Figure 4-18 over the frequency range 45 MHz to 10 GHz.

![Figure 4-18: Spectra of the noise PSDs of the HBT direct photodetector for different mean incident optical powers. $V_{CE} = 1$ V, a 27 dB gain, 2.5 dB noise figure Miteq preamplifier was used and de-embedded.](image)

The peaks seen in Figure 4-18 were due to ambient electromagnetic pick-up by the unshielded microstrip transmission line connecting the HBT and the SMA connector amplified by the preamplifier. In general the HBT output noise PSD decreased with frequency and was, therefore, closely related to its frequency response. To determine whether the laser RIN was a significant factor in the measured HBT output noise PSDs,
suppose that the laser beam illuminates a 100% quantum efficient photodiode directly connected to a 50 Ω load. It is further assumed that the RIN is $-155$ dB/Hz below 10 GHz and the mean incident optical power on the photodiode is 0.3 mW. The noise PSD delivered to the 50 Ω load due to the laser RIN alone is calculated to be $-177$ dBm(1Hz).

So if the HBT had not generated noise of its own and the spectra in Figure 4-18 had been entirely due to the detection and amplification of the laser RIN, it would have required the HBT to have a system signal gain of 47 dB at very low frequencies to produce an output noise PSD of $-130$ dBm(1Hz). The HBT would also have had to provide a gain of 17 dB at 6 GHz in order to output a noise PSD of $-160$ dBm(1Hz). According to Figure 4-7, the HBT certainly did not have such a high gain and therefore the measured noise PSDs were not laser RIN limited but generated by the HBT itself.

Figure 4-19 shows the $V_{ce}$ dependence of the measured noise PSD at 3.025 GHz in direct photodetection. Also shown in Figure 4-19 is the corresponding output signal power under the same bias conditions when the laser was modulated at 3.025 GHz with 10% modulation depth.

![Graph](image)

**Figure 4-19:** Measured noise PSD and signal power in direct photodetection at 3.025 GHz. $P_{opt} = 0.3$ mW, 10% mod. depth, $P_{RE,loc} = -44.5$ dBm, a 20 dB gain, 8 dB noise figure NewFocus preamplifier was used and de-embedded.

Figure 4-19 can also be re-plotted in terms of the signal-to-noise ratio (S/N) and is shown in Figure 4-20.
The S/N increased with $V_{CE}$ at low bias voltages because in that region, the signal power increased faster than the noise PSD. For $V_{CE} \geq 0.7$ V, the S/N stayed at about 115 dB (1 Hz). If the mean incident optical power is kept constant while the modulation depth is increased, it is expected that the S/N will increase. For example, if 100 % modulation depth is used with 0.3 mW mean incident optical power on the HBT, the S/N for $V_{CE} \geq 0.7$ V should increase to 135 dB (1 Hz).

Similar noise measurements were also made with the HBT configured as down- and up-converting optoelectronic mixers. Figure 4-21 and Figure 4-22 show the measured noise PSDs at 100 MHz (down-conversion) and 5.95 GHz (up-conversion), respectively, when the HBT optoelectronic mixer was subjected to a 10 dBm LO source at 2.925 GHz. Also shown are the corresponding down-and up-conversion signal powers when the laser was modulated at 3.025 GHz with 10 % modulation depth.
Figure 4-21: Measured noise PSD and down-conversion signal power at 100 MHz versus $V_{ce}$, $P_{opt} = 0.3 \text{ mW}, 10 \%$ mod. depth at 3.025 GHz, $P_{RE, REC} = -44.5 \text{ dBm}$, LO = 10 dBm at 2.925 GHz, a 20 dB gain, 8 dB noise figure NewFocus preamplifier was used and de-embedded.

Figure 4-22: Measured noise PSD and up-conversion signal power at 5.95 GHz versus $V_{ce}$, $P_{opt} = 0.3 \text{ mW}, 10 \%$ mod. depth at 3.025 GHz, $P_{RE, REC} = -44.5 \text{ dBm}$, LO = 10 dBm at 2.925 GHz, a 20 dB gain, 8 dB noise figure NewFocus preamplifier was used and de-embedded.

In general, the measured noise PSD was much lower at the up-conversion frequency of 5.95 GHz than at 100 MHz because as Figure 4-7 and Figure 4-18 show, the HBT had very high gain and output noise at 100 MHz than at 5.95 GHz. The system signal gain at 100 MHz when the HBT was simultaneously driven by the LO was also measured and is
shown in Figure 4-23 together with the noise PSD at 100 MHz from the down-conversion measurement.

Figure 4-23: System signal gain at 100 MHz when the HBT was simultaneously driven by the LO and output noise PSD at 100 MHz in the down-conversion.

Figure 4-23 shows that the down-conversion noise and the system signal gain at 100 MHz had very similar dependence on \( V_{CE} \) and suggests that the measured down-conversion noise was mainly due to the direct amplification at 100 MHz by the HBT of some constant noise sources, such as the base thermal noise and the base-collector junction shot noise due to the primary photocurrent.

The measured up-conversion noise PSD at 5.95 GHz shown in Figure 4-22 was, in general, slightly higher than that in direct detection (Figure 4-18) under similar bias conditions and it is believed to be due mainly to the up-conversion of the HBT noise from 3.025 GHz.

The signal-to-noise ratios for the down- and up-conversions are shown in Figure 4-24.

The S/N for the up-conversion at high \( V_{CE} \) was significantly greater than for the down-conversion and it is due to the very relatively high output noise generated by the HBT at low frequencies.
The results so far have identified a problem with the transistor optoelectronic mixer approach. It can be seen in Figure 4-18 that the active HBT has a much higher output noise at low frequencies than at high frequencies. Since mixing takes place within the HBT which also generates such large noise output, it is therefore impossible to filter out the HBT output noise without simultaneously attenuating the mixed products. The situation is particularly severe in down-conversion because the input signal is frequency translated to a relatively noisier frequency band. However, the situation is somewhat alleviated in the up-conversion process. In the conventional multi-component approach in which a double-balanced mixer is used, a high-gain narrow-band low-noise amplifier can be used to increase the photodetected signal strength. Further filtering after the amplifier can also be provided by a bandpass filter so that noise at the image and down- and up-conversion frequencies can be attenuated before mixing takes place. Compared to a transistor, a double-balanced mixer has a lower and flatter output noise spectrum, and if the down-and up-conversion losses of the mixer are comparable, then the signal-to-noise ratios for these two processes should be similar, unlike those seen in Figure 4-24.

In the next section, the distortion characteristics of the HBT optoelectronic mixer will be examined.
Section 4-7: Two-Tone Third-Order Intermodulation Distortion Product and Dynamic Range Measurements

Although mixing is as a whole a nonlinear process, when an input signal is down- and up-converted to different frequencies, this frequency translation is required to introduce minimal distortion to the input signal. In other words, while the mixing between a large LO signal and some other small input signals is desired, mixing among the small input signals should be avoided in order not to generate intermodulation distortion products.

For narrow-band operations in which the bandwidth is smaller than an octave, the third-order products are of interest since they are close to the signals themselves and therefore are difficult to filter out. Second harmonics fall out of the band and can usually be filtered out easily. Suppose the input to a mixer consists of two fundamental frequencies at $f_1$ and $f_2$, and the mixer is pumped by an LO at $f_{LO}$. Ideally the up-converted output should only contain the two fundamental frequencies at $(f_1 + f_{LO})$ and $(f_2 + f_{LO})$ but in practice, third-order products at $(2f_1 - f_2 + f_{LO})$ and $(2f_2 - f_1 + f_{LO})$ are also produced.

A similar situation applies to down-conversion and direct detection.

Measurements of the third-order intermodulation product powers as a function of the input intensity modulation depth together with the knowledge of the output noise PSDs allow the spurious-free dynamic range (SFDR) to be determined.

The simplest way to generate a test signal would be to modulate a single laser diode directly with two microwave sources at slightly different frequencies. However, the laser diode also generates intermodulation distortion products of its own and the measured distortion products from the HBT can be erroneous if the laser diode is not significantly more linear than the HBT itself. The approach selected here was to use two separately modulated laser diodes at two slightly different modulation frequencies. Since each laser received only one of the two tones from its own microwave generator, intermodulation before detection by the HBT was avoided and the distortion characteristics of the HBT could be more accurately measured.
The experimental arrangement is as shown in Figure 4-2. Laser 1 and laser 2 were respectively modulated at $CF - 10 \text{kHz}$ and $CF + 10 \text{kHz}$ where $CF = 3.025 \text{GHz}$. The two laser beams were combined in a beam splitter cube and focused on the HBT. The modulation depth of each of the two beams was varied by adjusting the electrical modulating signal power to the laser and was monitored with a lightwave signal analyser. The mean incident optical power on the HBT from each laser was set at 0.15 mW so that total mean optical power the HBT received was 0.3 mW throughout, which was the same optical bias condition used in the previous noise measurements. The preamplifier used previously was also taken out to prevent introducing additional distortion to the system.

The intermodulation distortion characteristics in direct detection shown in Figure 4-25 were obtained by measuring the fundamental signals at $CF \pm 10 \text{kHz}$ and third order distortion products at $CF \pm 30 \text{kHz}$ as a function of the intensity modulation index per laser beam. It should be noted that a value of 0 on the x-axis denotes 100% modulation depth and any value greater than 0 is imaginary.

![Figure 4-25: Two-tone third order intermodulation distortion characteristics in direct detection as a function of intensity modulation index. $V_{CE} = 1 \text{ V}$, $P_{opt} = 0.3 \text{ mW}$ (0.15 mW from each laser). Lasers modulated at 3.025 GHz $\pm 10 \text{ kHz}$.

The distortion products increased 3 times as fast with the logarithmic of modulation index as the fundamental signals. For $V_{CE} = 1 \text{ V}$, from Figure 4-19 the HBT output noise PSD
was -145 dBm (1 Hz), which is included in Figure 4-25. The SFDR at the output, defined as the difference between the output fundamental signal power and the output third order product power when equal to the output noise PSD, was 95 dB (1 Hz).

Distortion characteristics in down- and up-conversions were also measured and are shown in Figure 4-26 and Figure 4-27, respectively.

![Figure 4-26: Two-tone third order intermodulation distortion characteristics in down-conversion as a function of intensity. $V_{CE} = 0.6 \, V$, $P_{opt} = 0.3 \, mW$ (0.15 mW from each laser). Lasers modulated at 3.025 GHz ± 10 kHz. LO = 10 dBm at 2.925 GHz. Fundamentals measured at 100 MHz ± 10 kHz and intermod. measured at 100 MHz ± 30 kHz.]

The SFDR for the down-conversion was 78 dB (1 Hz) while the SFDR for the up-conversion was 90 dB (1 Hz). The smaller down-conversion SFDR was mainly due to the higher output noise floor in the 100 MHz region compared to that in the 5.95 GHz region. The results in Figure 4-26 and Figure 4-27 therefore suggest that the HBT has higher performance as an up-converter than as a down-converter. Compared to performing optoelectronic mixing with an HBT direct photodetector and a microwave mixer, the penalty in SFDR for using a single HBT up-converter should be less than 5 dB because the HBT SFDR in direct detection was only 5 dB better than in the up-conversion.
Figure 4-27: Two-tone third order intermodulation distortion characteristics in up-conversion as a function of intensity. $V_{CE} = 0.6 \text{ V, } P_{opt} = 0.3 \text{ mW (0.15 mW from each laser).}$ Lasers modulated at 3.025 GHz ± 10 kHz. LO = 10 dBm at 2.925 GHz. Fundamentals measured at 5.95 GHz ± 10 kHz and intermod. measured at 5.95 GHz ± 30 kHz.

**Section 4-8: Conclusion**

A novel two-terminal edge-coupled HBT optoelectronic mixer has been built and characterised. The obtained system conversion gain was +7 dB when a 2.41 GHz intensity modulated light was detected and down-converted to a 500 MHz electrical signal. The result suggests that such a single two-terminal HBT optoelectronic mixer would have a 7 dB gain advantage over an optoelectronic mixer built using a hypothetical 100 % quantum efficient photodetector and an ideal microwave mixer with 0 dB conversion loss in a 50 Ω impedance system. Compared with Table 1-1 of *Chapter 1*, this HBT has also achieved the highest reported system conversion gain, although the frequencies used were different. The LO drive power used to achieve such a high system conversion gain was 10 dBm and is comparable to that required for most commercially available double-balanced mixers.

The measured system gain characteristics have been compared with the simulation results of the harmonic-balance model of *Chapter 3*. The simulation results have shown good qualitative agreement with the experimental results. The discrepancies between the
simulation and measured results were attributed to the high-level assumption that the mixing mechanism of the HBT mixer was the terminal voltage dependence of small-signal photocurrent gain, and the quasi-static nature of the model. Nevertheless, the model has provided a simple way of predicting the conversion characteristics of the HBT electrically pumped optoelectronic mixer.

Signal-to-noise ratio and two-tone third-order intermodulation distortion characteristics have been measured. The results of these two measurements allowed the spurious-free-dynamic ranges to be determined. The measured SFDRs for direct detection at 3.025 GHz, down-conversion from 3.025 GHz to 100 MHz, and up-conversion from 3.025 GHz to 5.95 GHz were 95, 78 and 90 dB(1 Hz), respectively. These results suggest that the HBT has higher performance as an up-converter than as a down-converter. Compared to performing optoelectronic mixing with an HBT direct photodetector and a microwave mixer, the penalty in SFDR for using a single HBT up-converter should be less than 5 dB because the HBT SFDR in direct detection was only 5 dB better than in the up-conversion.

In the next Chapter, a three-terminal normal-incidence HBT optoelectronic mixer will be described.
References


Chapter Five

Experimental Results for a Three-Terminal Normal-Incidence HBT Optoelectronic Mixer

Section 5-1: Introduction

This Chapter is concerned with the experimental results for a three-terminal normal-incidence HBT optoelectronic mixer. The HBTs were fabricated by project collaborators at the Technion in Israel and delivered to University College London (UCL) in die form. Packaging and experimental characterisations of the devices were performed at UCL. The experimental results presented here are therefore those of the packaged HBTs.

This Chapter is structured as follows. The HBT fabrication and structure, and the experimental arrangement will first be described in Section 5-2. The experimental results which follow will include, in Section 5-3, the static photo-responsivity to establish the device external quantum efficiency, in Section 5-4 the frequency response in direct photodetection, in Section 5-5 the electrically pumped optoelectronic conversion gains and comparisons with the simulation results from Chapter 3, in Section 5-6 the frequency response in electrically pumped optoelectronic mixer, in Section 5-7 noise performance and in Section 5-8 the two-tone third-other intermodulation distortion product measurements. The spurious-free dynamic range of the HBT electrically pumped optoelectronic mixer will be calculated using the results obtained. In Section 5-9 the performance of the HBT used as an optically pumped optoelectronic mixer in which the LO power is delivered optically will also be presented. Finally a conclusion is given in Section 5-10.

Section 5-2: Device Fabrication, Structure and Experimental Arrangement
The HBT epitaxial layers were grown on a semi-insulating InP substrate by a compact metal-organic molecular beam epitaxy system [1]. The emitter Ti/Pt/Au metal served as a mask for wet etching of the emitter mesa. Self-aligned non-alloyed Pt/Ti/Pt/Au contacts were evaporated on the base and collector mesas. A $5 \times 6 \mu m^2$ opening in the base metallisation served as an optical window. Ti/Au pads were evaporated after polyimide passivation and a curing process at 300°C for 1 hour. The polyimide layer covering the optical window was 450 nm thick. The cross-section of the HBT with doping levels is shown in Figure 5-1.

The HBT was secured in the middle of a custom-designed brass mount with 1 cm long, 50 Ω characteristic impedance microstrip transmission lines on both sides. The base and the collector were connected to the transmission lines using single bond wires estimated to be no longer than 0.5 mm each. The emitter was similarly connected to a neighbouring brass step serving as the signal ground. SMA connectors were then attached to both ends of the mount so that the HBT could be accessed electrically using conventional microwave cables. To enable optical access to the device, the mount was placed on an $x$-$y$-$z$ micro-positioner so that the HBT optical window could be aligned with the laser beam.

The basic experimental arrangement used in the static IV, direct photodetection and optoelectronic mixing measurements is shown in Figure 5-2.
Figure 5-2: Basic experimental arrangement for HBT measurements.

The light sources in Figure 5-2 are two very low noise GEC-Marconi LD6804 60mW lasers emitting at around 1550 nm wavelength. The two lasers were operated simultaneously for the two-tone third-order intermodulation distortion product measurements in which the wavelengths of the two lasers were sufficiently offset so that the beat signal (above 20 GHz) did not interfere with the measurements at lower frequencies. For other optoelectronic measurements, only laser 1 was required. The two beams were combined in free space with a beam splitter cube. One of the emergent beams from the splitter was focused onto the HBT and the other was coupled to a fibre and monitored with an LSA. Optical isolators were employed to ensure that reflections of optical energy back to the lasers were sufficiently suppressed.

Both lasers and the HBT base and collector terminals were biased via HP 11612A Bias-Ts having a frequency range of 45 MHz to 26.5 GHz. In the electrically pumped mixing experiments, the HBT base was driven by a large-signal local oscillator (LO). In the optically pumped mixing experiments, the base was driven by a small-signal RF signal source with the LO delivered to the HBT optically. The output signal of the HBT was taken from the collector and measured with a spectrum analyser.
**Section 5-3: External Quantum Efficiency Measurement**

The availability of an electrical base contact in a three-terminal HBT enables the external quantum efficiency $\eta_{ext}$ (defined in Section 1-2) to be measured directly because the current gain of the HBT can be switched off by setting $V_{BE} = 0 \text{ V}$ and the measured current is the primary photocurrent generated in the base and the collector. Such a measurement is not possible in a two-terminal HBT because no electrical base connection is available to set $V_{BE} = 0 \text{ V}$ and the measured current is always an amplified version of the primary photocurrent.

To obtain the external quantum efficiency of the present HBT, the responsivity (defined in Section 1-2) was first measured. Figure 5-3 shows the measured primary photocurrent (the collector current with $V_{BE} = 0 \text{ V}$) versus the mean incident optical power.

![Figure 5-3: Primary photocurrent vs. mean incident optical power. $V_{BE} = 0 \text{ V}$ and $V_{CE} = 1 \text{ V}$. Wavelength = 1550 nm.](image)

As expected from the theory, the measured primary photocurrent generated in the base and collector without any transistor gain was directly proportional to the mean incident optical power. The slope in Figure 5-3 is 0.22 A/W and therefore was the responsivity of the HBT. Since the responsivity of a 100% quantum efficient photodetector at 1550 nm wavelength would be 1.25 A/W, the external quantum efficiency of the HBT was
therefore calculated to be 17.6%. Such a low external quantum efficiency can be attributed to three factors. The first one is the optical reflections at the air-polyimide and polyimide-InGaAs base interfaces. The combined optical reflectance is given by \[ R_{\text{refl}} = \left( \frac{Z_{\text{poly}}(y = 0) - \eta_{\text{air}}}{Z_{\text{poly}}(y = 0) + \eta_{\text{air}}} \right)^2 \] \[ \text{Eqn 5-1} \]

where \( \eta_{\text{air}} \) is the characteristic impedance of air and is equal to 377 \( \Omega \). \( Z_{\text{poly}}(y = 0) \) is the load impedance in the polyimide at the air-polyimide interface and is given by

\[ Z_{\text{poly}}(y = 0) = n_{\text{poly}} \left[ \eta_{\text{base}} \cos(\beta_{\text{poly}} d) + j n_{\text{poly}} \sin(\beta_{\text{poly}} d) \right] \] \[ \text{Eqn 5-2} \]

where \( n_{\text{poly}} \) and \( n_{\text{base}} \) are the characteristic impedances of polyimide and the InGaAs base, and are given by \( 377/n_{\text{poly}} \) and \( 377/n_{\text{base}} \), respectively, where \( n_{\text{poly}} \) and \( n_{\text{base}} \) are the respective refractive indices. \( n_{\text{base}} = 3.74 \) (calculated from the relative permittivity from [3]) and \( n_{\text{poly}} \) is estimated to be around 1.6. \( \beta_{\text{poly}} \) is the wave number of polyimide and, at 1550 nm wavelength, equal to \( 6.50 \times 10^6 \) rad/m. \( d \) is the thickness of the polyimide layer and is measured to be 450 nm. Substituting all these figures into Eqn 5-1 and Eqn 5-2 gives

\[ R_{\text{refl}} = 0.32 \quad \text{Eqn 5-3} \]

The second factor is the ratio of the smaller optical window area to the larger spot size of the focused beam which is approximated to be 8 \( \mu \text{m} \) in diameter. Therefore this ratio is equal to

\[ k_{\text{area}} = \frac{5 \times 6 \mu\text{m}^2}{\pi \times 4^2 \mu\text{m}^2} = 0.6 \quad \text{Eqn 5-4} \]

The third factor limiting the external quantum efficiency is the percentage of the photons absorbed in the base and the collector. Using Eqn 2-1 and assuming there is no absorption in polyimide, the proportion of the photons absorbed relative to the number of photons illuminating the base and the collector is

\[ k_{\text{abs}} = 1 - \exp[-\alpha(W_B + W_{BC} + W_C)] \quad \text{Eqn 5-5} \]
where \( W_b + W_{bc} + W_c = 800 \text{ nm} \) is the total thickness of the base and the collector, and 
\( \alpha = 1 \mu m^{-1} \) is the absorption coefficient for InGaAs at 1550 nm wavelength [5]. 
Substituting these values into Eqn 5-5 gives  
\[
k_{\text{abs}} = 0.55 \tag{Eqn 5-6}
\]
Assuming every photon absorbed generates an electron and a hole, and all the electron-hole pairs so generated contribute to the primary photocurrent, then the external quantum efficiency is given by  
\[
\eta_{\text{ext}} = (1 - R_{\text{refl}}) \cdot k_{\text{area}} \cdot k_{\text{abs}} = 0.22 \text{ or } 22 \% \tag{Eqn 5-7}
\]
Therefore the measured external efficiency of 17.5 % is not too far from what would be expected according to the above calculation. Such a low external quantum efficiency is, however, not uncommon. Fukano et al [4] reported a substrate illuminated InP-InGaAs HBT employing a nonalloyed electrode metal as a reflector to double the optical absorption length. The total base and collector thickness, including an InGaAs spacer, was 400 nm and thus an effective optical absorption length of 800 nm, similar to the total thickness of the base and collector of the HBT here. The base area of the HBT in [4] was 6x8 \( \mu m^2 \) and the external quantum efficiency at 1550 nm wavelength was 21 %. This slightly higher external quantum efficiency was obtained with an HBT with a larger base area than the optical window of the HBT here.

Since, from Section 1-2, \( G_{\text{int}} = G_{\text{sys}} / \eta_{\text{ext}}^2 \), \( G_{\text{int}} \) will be 15 dB higher than \( G_{\text{sys}} \).

**Section 5-4: Frequency Response in Direct Photodetection**

One of the simplest applications of a phototransistor is the direct detection of intensity modulated light and the phototransistor is commonly characterised by its frequency response.
To measure the frequency response of the HBT to intensity modulated light, laser 1 was directly modulated with an RF signal swept from 10 MHz to 6 GHz. However, the frequency response of laser 1 was not flat in this frequency range and had to be measured first against a wideband photodetector. The wideband photodetector used was the New Focus 1014 InGaAs Schottky photodiode with a stated 3 dB bandwidth of 45 GHz. Assuming the frequency response of this New Focus 1014 is flat from 10 MHz to 6 GHz, the modulation depth of the emitted light from laser 1 was measured. From the DC primary photocurrent generated in the HBT, responsivity in Figure 5-3 and the modulation depth of the laser beam, the equivalent electrical input power $P_{RF_elec}$ (see Section 1-2) was calculated. After the output power of the HBT was measured, the system signal gain $G_{sig}$ was obtained and is shown in Figure 5-4.

![Figure 5-4: HBT frequency response in direct photodetection. $V_{CE} = 1$V, mean incident optical power = 0.13 mW, base terminated in 50 Ω.](image)

When $V_{BE} = 0$ V, the HBT was not active and $G_{sig}$ was that of the HBT base-collector photodiode and flat within ±2 dB in the frequency range.

When the HBT was switched on, $G_{sig}$ increased considerably. $G_{sig}$ at low frequencies was dependent on the collector current, especially at low currents, because of the base-emitter
recombination current. $G_{\text{sig}}$ started to roll off from a few hundred MHz, depending on the collector current. All top four curves show a similar system signal gain-bandwidth product of around 3.5 GHz.

**Section 5-5: Electrically Pumped Optoelectronic Mixing Conversion Gain Measurements and Comparison with the Simulation Results**

Electrically pumped optoelectronic mixing is conventionally carried out by first converting RF modulated light into an electrical signal with a photodetector. The RF signal is subsequently mixed in a microwave mixer with an LO to produce mixed products at both down- and up-conversion frequencies. In this Chapter electrically pumped optoelectronic mixing is carried out in a single HBT in which the base-collector junction acts as a photodiode while the current gain dependence on $V_{BE}$, which is modulated by an LO, provides the necessary mixing mechanism.

In this section, measurements of the HBT electrically pumped optoelectronic mixing conversion gains will be described and results compared with those from the computer simulation described in Chapter 3. The experimental procedures are as follows. Laser 1, biased at 41 mA, was directly modulated with a 3.5 GHz RF signal. The intensity modulation depth of the light emitted by laser 1 was measured by a lightwave signal analyser to be 90%. Optical attenuators were used to reduce the mean incident optical power $P_{\text{opt}}$ on the HBT to 70 $\mu$W, giving a primary photocurrent of 15 $\mu$A and $P_{RF, \text{Elec}} = -38$ dBm. The HBT base was driven by a 3.4 GHz LO whose power was varied. The down- and up-converted signals at 100 MHz and 6.9 GHz, respectively, were taken from the collector and measured with a spectrum analyser. $V_{CE}$ was fixed at 1.5 V while $V_{BE}$ was varied.

Figure 5-5 and Figure 5-6 show the measured system down- and up-conversion gains, respectively.
Figure 5-5: Measured system down-conversion gain versus $V_{BE}$ as a function of LO powers. $P_{opt} = 70 \, \mu W$. $V_{CE} = 1.5 \, V$, laser modulation frequency (RF) = 3.5 GHz, LO frequency = 3.4 GHz, down-conversion frequency = 100 MHz.

Figure 5-6: Measured system up-conversion gain versus $V_{BE}$ as a function of LO powers. $P_{opt} = 70 \, \mu W$. $V_{CE} = 1.5 \, V$, laser modulation frequency (RF) = 3.5 GHz, LO frequency = 3.4 GHz, up-conversion frequency = 6.9 GHz.
For low LO powers ($\leq -8 \text{ dBm}$), both the down- and up-conversion gains initially increased with $V_{BE}$. A conversion gain maximum was then reached at $V_{BE} = 0.82 \text{ V}$ for the down-conversion and at $V_{BE} = 0.80 \text{ V}$ for the up-conversion, respectively, before decreasing. It is also seen that for low LO powers, both conversion gains were strongly dependent on $V_{BE}$ because the voltage swing by the LO was small relative to how much $V_{BE}$ was varied. On the other hand, for large LO powers, the dependence of the conversion gains on $V_{BE}$ was weaker because the voltage swing by the LO was now large compared to the range of $V_{BE}$ varied.

It is quite difficult to explain qualitatively all the characteristics shown in Figure 5-5 and Figure 5-6 except for very small LO powers. The conversion gains depend both on how strongly dependent the current gain of the HBT on $V_{BE}$ is and the magnitude of the current gain itself. It is well known that at small $V_{BE}$, the current gain of a bipolar transistor increases with $V_{BE}$ due to the effects of the base-emitter recombination current [5] and is therefore strongly dependent on $V_{BE}$. Thus the HBT is very nonlinear and the conversion gains increase with $V_{BE}$ at low bias. At higher $V_{BE}$, the current gain becomes less dependent on $V_{BE}$, making the HBT more linear and mixing less efficient, and hence the conversion gains decrease with $V_{BE}$. An optimum $V_{BE}$ therefore exists for the highest conversion gain at small LO powers and is observed in Figure 5-5 and Figure 5-6.

At higher LO powers, the situation is more complicated since the voltage swing of $V_{BE}$ by the LO is so large that within one cycle of the LO, both the linear and nonlinear operating points of the HBT are involved.

Also the system up-conversion gain increased again at high $V_{BE}$ while the system down-conversion gain did not. The reason for this is not fully understood but two suggested explanations will be discussed later in this section.

The highest system down-conversion gain was around $-4 \text{ dB}$ while the highest system up-conversion gain was around $-11 \text{ dB}$. In general the down-conversion gain was a few dB higher than the up-conversion gain because of both the intrinsic parameters, such as
the base-emitter and base-collector capacitances, base transit time effect and the collector depletion delay time effect, and the parasitics such as the bond-wire inductances.

In Chapter 3, the HBT electrically pumped optoelectronic mixer was modelled using harmonic-balance and conversion matrices. The measured and simulated system down- and up-conversion gains are compared in Figure 5-7 and Figure 5-8, respectively, for low LO powers.

Figure 5-7: Comparison of measured (markers) and simulated (lines) system down-conversion gain for low LO powers.

Figure 5-8: Comparison of measured (markers) and simulated (lines) system up-conversion gain for low LO powers.
In both system conversion gains, the simulated results are in good agreement with the measured ones at low $V_{BE}$. In addition, the model correctly predicted that both conversion gains would initially increase with $V_{BE}$, then reach a maximum and finally decrease with $V_{BE}$. The model, however, failed to predict that the up-conversion gain would increase again at higher $V_{BE}$. The model also correctly predicted that the optimum $V_{BE}$ value for maximum up-conversion gain would be lower than that for maximum down-conversion gain, although the actual calculated optimum $V_{BE}$ values were not in good agreement with the measured ones.

At higher $V_{BE}$, the simulated results do not agree well with the measured ones. The reason for this could be the following. At high $V_{BE}$, the amplified AC current due to the LO developed a large AC voltage at the collector through the output 50 Ω load. Such a large AC voltage at the collector momentarily forward biased the base-collector junction, driving the HBT more into saturation and giving rise to an extra mixing mechanism. As $V_{BE}$ was increased, the reverse-bias voltage of the base-collector junction was reduced and the junction was forced to conduct for a longer time within each LO cycle. In the model described in Chapter 3, however, the base-collector junction was assumed to be reverse-biased at all times no matter what $V_{BE}$ and LO levels were. As a result, the model is not particularly accurate at high bias points and for large LO powers. The deficiency of the model is more obvious when the LO power levels are high. The simulated and measured system down- and up-conversion gains for large LO are compared in Figure 5-9 and Figure 5-10, respectively.

It can be seen that the discrepancies between the measured and simulated system conversion gains at large LO powers are greater than those at small LO powers. The simulated system down-conversion gain showed relatively better agreement with the measured one. In particular the model for down-conversion predicted that for large LO powers, the system conversion gain would be less strongly dependent on $V_{BE}$. 
To make use of the extra mixing mechanism provided by the momentarily forward-biased base-collector junction, the system down- and up-conversion gains were also measured with $V_{ce}$ now reduced to 1 V, and are shown in Figure 5-11 and Figure 5-12.
Figure 5-11: Measured system down-conversion gain versus $V_{BE}$ as a function of LO powers. $P_{opt} = 0.3$ mW, $V_{CE} = 1.0$ V, laser modulation frequency (RF) = 3 GHz, LO frequency = 2.9 GHz, down-conversion frequency = 100 MHz.

Figure 5-12: Measured system up-conversion gain versus $V_{BE}$ as a function of LO powers. $P_{opt} = 0.3$ mW, $V_{CE}$ = LO V, laser modulation frequency (RF) = 3 GHz, LO frequency = 2.9 GHz, up-conversion frequency = 5.9 GHz.

It can be seen that lowering $V_{CE}$ had a considerable effect on the patterns of the system down- and up-conversion gains at high $V_{BE}$ and large LO powers. Especially in Figure
5-12, the system up-conversion gain was enhanced significantly at higher $V_{be}^{\ast}$. One possible reason for the different patterns seen in the down- and up-conversion gains at high $V_{be}^\ast$ could be the following. The two nonlinear elements in the HBT are the base-emitter junction and the base-collector junction, both of which produce down- and up-converted currents. However the relative phase relationship between the down- and up-converted currents produced by the base-emitter junction might be different from that by the base-collector junction. For example, let, say, both the down- and up-converted currents by the base-emitter junction have positive phase. At the same time, assume that the down-converted current by the base-collector junction has negative phase while the up-converted current has positive phase. When these currents add up at the collector, the down-converted currents add destructively while the up-converted currents add constructively. As a result, the system up-conversion gain is enhanced but the system down-conversion gain is suppressed.

Another possible reason is due to the extra phase shift added by the base-transport factor of the HBT. After the down- and up-converted currents are generated at the base-emitter junction, they travel across the base and reach the collector. Since these currents are several GHz apart, the base transport factor will add a large phase shift to the up-converted current compared to that added to the down-converted current. Therefore when these currents arrive at the collector and add up with the down- and up-converted currents generated by the base-collector junction, the down-converted currents will add destructively because of the opposite polarities of the two junctions. If, however, the base-transport factor adds enough phase shift (up to 180 degrees), to the up-converted current, it will add constructively to that generated by the base-collector junction. As a consequence, the system up-conversion gain is enhanced. The exact reason for this phenomenon could be a mixture of the two possible ones just discussed and some others not yet discovered. A more detailed simulation should help understand this phenomenon better.

In summary, the measured system down- and up-conversion gains of the HBT have been presented and compare favourably with other published results on three-terminal HBT optoelectronic mixers summarised in Table 1-1 in Chapter 1. The measured system down- and up-conversion gains were compared with those from the HBT harmonic-
balance model. The simulation results show good agreement with the measured ones at small $V_{BE}$ and for low LO power levels. For $V_{BE} \leq 0.8$ V and LO $\leq -8$ dBm, the agreement was within 4 dB for down-conversion, and was within 7 dB for up-conversion. When the HBT was driven by large LO powers, the model becomes inaccurate. For $V_{BE} = 0.9$ V and LO $\geq -2$ dBm, the discrepancy was around 10 dB for down-conversion, and around 20 dB for up-conversion. The reason for the discrepancies could be due to the fact that at such large LO levels, the base-collector junction was momentarily forward-biased, giving rise to an extra mixing mechanism. However, the HBT model assumes that the base-collector junction is always reverse-biased. Therefore, the accuracy of the model is expected to improve when the base-collector junction is represented fully by a large-signal diode model.

Section 5-6: Frequency Response in Electrically Pumped Optoelectronic Mixing

Unlike the two-terminal HBT optoelectronic mixer described in Chapter 4 in which a change of the LO frequency would require a new diplexer, the frequency response of the three-terminal HBT electrically pumped optoelectronic mixer can be more conveniently measured because the LO frequency is not restricted by any external filter. In this section, the measurements of the frequency responses in both down- and up-conversions will be described and results presented.

The experimental procedures are as follows. In the down-conversion frequency response measurement, the down-converted frequency $f_{IF}$ was fixed at 100 MHz while the laser modulation frequency $f_{RF}$ and the LO frequency $f_{LO}$ were swept according to

$$f_{RF} = f_{LO} + 100 \text{ MHz} \quad \text{Eqn 5-8}$$

Laser 1 was biased at 50 mA and directly modulated by a -10 dBm RF signal. The modulation depth of the emitted light was measured and used to calculate $P_{RF, Elec}$ at each $f_{RF}$. The average incident optical power on the HBT was 0.3 mW. $V_{BE}$ and the LO power level were fixed at 0.82 V and -10 dBm, respectively. The system down-conversion gain of the HBT versus $f_{RF}$ for a fixed 100 MHz $f_{IF}$ is shown in Figure 5-13.
In general, the system down-conversion gain decreased with the laser modulation frequency because, as shown in Figure 5-4, the packaged HBT photoresponse started to roll off from just a few hundred MHz.

In the measurement of the up-conversion frequency response, the laser modulation frequency was fixed at 100 MHz, the LO frequency was swept from 3 GHz to 20 GHz and the corresponding up-converted signal was measured from 3.10 GHz to 20.10 GHz. The average incident optical power on the HBT was 0.3 mW and the modulation depth of the light was set at 10 %. The system up-conversion gain versus the LO frequency is shown in Figure 5-14 for two sets of $V_{BE}$ and LO drive levels.

As expected, the two system up-conversion gains decreased with increasing LO frequency. However the roll-off of the system up-conversion gain was not as steep as that of the system down-conversion gain in which both the laser modulation and LO frequencies were swept. Also in the down-conversion, the LO power was only -10 dBm which was not large enough to saturate the HBT. At low LO frequencies, the first curve showed a system up-conversion gain of around 6 dB. Even at 20 GHz LO frequency (20.1 GHz up-converted frequency), the system up-conversion loss was only 10 dB,
indicating that the HBT could be a compact and high-performance optoelectronic up-converter for mm-wave fibre-radio links.

**Section 5-7: HBT Noise Performance**

In this section, the measurements of the noise performance of the HBT in electrically pumped optoelectronic down- and up-conversions as well as in direct photodetection are described and results presented. One of the problems in measuring the noise performance of optoelectronic devices is the extra noise contribution from the optical source, characterised by its relative intensity noise (RIN). A previous noise performance characterisation of a two-terminal HBT optoelectronic mixer [6] was carried out using a high RIN laser (-130 dB/Hz at 2.81 GHz). Due to the fact that two-terminal HBT required a relatively higher incident optical power for biasing and the laser had a high RIN value, the results obtained were mostly laser RIN limited.

The optical source (laser 1) used in measuring the noise characteristics of the present three-terminal HBT had a RIN of less than -160 dB/Hz below 7.5 GHz when biased at 140 mA. A plot of the laser RIN versus frequency up to 25 GHz was given in Figure 4-16 in Chapter 4. On the other hand, the three-terminal HBT is less susceptible to the
optical noise because the HBT can be biased electrically by injecting a base current without requiring a high incident optical power.

To improve the sensitivity of the equipment in this noise measurement, a 27 dB gain Miteq low-noise amplifier was employed and inserted between the spectrum analyser and the collector of the HBT. The noise figure of this amplifier is around 2.5 dB above 500 MHz and therefore measurement of noise power spectral density as low as -171 dBm(1Hz) should be possible.

In the optoelectronic mixing measurements, the base of the HBT was pumped by an LO. However, after amplification by the HBT, the measured LO power at the HBT collector was so large that the Miteq amplifier was driven into saturation. To rectify this problem, two bandstop filters based on microstrip were designed and fabricated. The first bandstop filter was used to suppress the LO fundamental at 3.4 GHz and was based on a 3rd order Butterworth commensurate-line filter design. One of the criteria in the design was that the filter should have a bandwidth narrow enough not to affect the passages of the down-and the up-converted signals, but wide enough so that the exact centre frequency of the filter was not critical. The filter layout and measured frequency response are shown in Figure 5-15.

The second filter was required to suppress the second harmonic of the LO at 6.8 GHz. A very narrow bandwidth filter was needed because of the neighbouring up-converted

![Figure 5-15: Layout and frequency response of a 3.4 GHz commensurate-line microstrip bandstop filter.](image_url)
signal at 6.9 GHz. Therefore the filter was designed in microstrip coupled-line configuration. The filter layout and measured frequency response are shown in Figure 5-16.

![Layout and frequency response of a 6.8 GHz microstrip coupled-line bandstop filter.](image)

Some post-fabrication lengthening of the two vertical open stubs was required so that their resonance frequency matched that of the two coupled-line sections.

The third and higher harmonics of the LO were found to be small enough not to saturate the amplifier. These two bandstop filters were then placed in series between the HBT collector and the Miteq amplifier before the noise measurement in the mixing process was carried out.

After all the components were put in place, the noise performance characterisation of the HBT optoelectronic mixer was conducted as follows. The LO was changed slightly to 3.382 GHz so that the second harmonic of the LO frequency matched the centre frequency of the second narrow bandwidth bandstop filter. The LO power level was set at -4 dBm so that the HBT showed good system down- and up-conversion gains. Laser 1 was biased at 140 mA unmodulated so that its RIN was better than -160 dB/Hz below 7.5 GHz which was within the frequency band of interest. Optical attenuators were used to reduce the incident optical power on the HBT to 70 μW. If the laser had been modulated at 3.682 GHz, the down- and up-converted signals would have appeared at 300 MHz and 7.064 GHz, respectively. Assuming the modulation frequency was 3.682 GHz, the output
down- and up-converted noise power spectral densities (PSDs) at 300 MHz and 7.064 GHz, respectively, were measured as a function of $V_{BE}$ with $V_{CE}$ fixed at 1.5 V.

After the noise PSDs were measured, laser 1 was biased at 41 mA and modulated at 3.682 GHz, giving an intensity modulation depth of 90%. Optical attenuation was reduced so that the average incident optical power on the HBT remained at 70 µW. The down- and up-converted signal outputs were measured again for the same bias conditions. Figure 5-17 and Figure 5-18 show the measured noise PSDs and output signal powers for down- and up-conversions, respectively.

![Figure 5-17: Noise power spectral density and signal output at 300 MHz versus $V_{BE}$ in down-conversion. $V_{CE} = 1.5$ V, LO = -4 dBm at 3.382 GHz. Mean incident optical power = 70 µW. When measuring noise, laser was run in CW with RIN < -160 dB/Hz. When measuring signal, laser was modulated at 3.682 GHz with 90% intensity modulation depth, giving $P_{RE,LEC} = -38$ dBm.](image)

In general the output noise PSD in the down-conversion was higher than that in the up-conversion and this can be better understood by examining the output noise spectrum of the HBT in Figure 5-19 when no LO drive and illumination were present.
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Figure 5-18: Noise power spectral density and signal output at 7.064 GHz versus $V_{BE}$ in up-conversion. $V_{CE} = 1.5$ V, LO = -4 dBm at 3.382 GHz. Mean incident optical power = 70 μW. When measuring noise, laser was run in CW with RIN < -160 dB/Hz. When measuring signal, laser was modulated at 3.682 GHz with 90% intensity modulation depth, giving $P_{RE,Elec} = -38$ dBm.

The peaks seen in Figure 5-19 were due to ambient electromagnetic pick-up by some unshielded microstrip transmission lines in the setup amplified by the HBT and the Miteq amplifier. It can be seen that when the HBT was sufficiently biased, say 20 mA collector current, the noise level at the down-converted frequency was around 20 dB higher than that at the up-conversion frequency. In Figure 5-17 and Figure 5-18 when $V_{BE} = 0.9$ V, the collector current was also around 20 mA and the measured noise PSD in the down-

Figure 5-19: HBT output noise PSD spectrum. No LO and optical illumination were present. $V_{CE} = 1.5$ V.

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conversion was approximately 19 dB larger than that in the up-conversion which agrees well with what is predicted from Figure 5-19. It is also noted that the measured noise PSD for down-conversion increased faster with bias than it did for up-conversion and is again in agreement with Figure 5-19.

Assuming the laser RIN was -160 dB/Hz and noting that the mean incident optical power was 70 µW, it would therefore require the electrically pumped HBT optoelectronic mixer to have a system conversion gain of 34 dB to achieve a RIN contributed output noise PSD of -160 dBm (1 Hz). Obviously the present optoelectronic mixer cannot provide such a high system conversion gain and therefore the measured noise PSD was not laser RIN limited. On the other hand the primary photocurrent was only 15 µA compared to the externally applied base current of around 100 µA for a 20 mA collector current, the optical shot noise was therefore not as significant as the shot noise due to the externally applied base current.

In general the output noise PSD measured in down-conversion (i.e. at low frequencies) could be attributed to the result of the amplified base thermal and base shot noise because of the frequency dependence seen in Figure 5-19. In up-conversion, however, the output noise PSD at such high frequency could be significantly attributed to the collector shot noise. For example, for $V_{be} = 0.9$ V and $I_c = 20$ mA, the calculated shot noise PSD in a 50 Ω load is -155 dBm (1 Hz). The measured noise PSD for up-conversion with the same bias condition was only 3 dB lower than this value and could be due to low-pass filtering by the base-collector junction capacitance as well as by the bond-wire inductance in the package.

The highest system down- and up-conversion gains calculated from Figure 5-17 and Figure 5-18 were 0 dB and -10 dB, respectively, and were 4 dB and 1 dB higher than the previously measured values shown in Figure 5-5 and Figure 5-6. The reason could be the insertion of the two bandstop filters in these noise measurements which changed the impedance seen by the HBT collector, and as a result the base-collector junction was subjected to a different AC voltage, affecting the mixing process. The results shown in Figure 5-5 and Figure 5-6 were obtained without using any filters.
The signal-to-noise ratios for both down- and up-conversions can be obtained from Figure 5-17 and Figure 5-18, and are shown in Figure 5-20 and Figure 5-21, respectively.

![Figure 5-20: Signal-to-noise ratio versus $V_{BE}$ for the down-conversion obtained from Figure 5-17.](image)

![Figure 5-21: Signal-to-noise ratio versus $V_{BE}$ for the up-conversion obtained from Figure 5-18.](image)

The highest signal-to-noise ratios occurred at $V_{BE} = 0.6$ V in down-conversion and at $V_{BE} = 0.7$ V in up-conversion, and were comparable at around 112 dB (1 Hz). However, highest system conversion gains occurred at $V_{BE} = 0.8$ V in down-conversion and
$V_{BE} = 0.72\, \text{V}$ in up-conversion which were different from the $V_{BE}$ values for highest signal-to-noise ratios.

Similar measurements were carried out with the HBT configured as a direct photodetector. The corresponding results are shown in Figure 5-22 and Figure 5-23.

Figure 5-22: Noise power spectral density and signal output at 3.682 GHz versus $I_C$ in direct photodetection. $V_{CE} = 1.5\, \text{V}$. Mean incident optical power = 70 μW. When measuring noise, laser was run in CW with RIN < -160 dB/Hz. When measuring signal, laser was modulated at 3.682 GHz with 90 % intensity modulation depth, giving $P_{BE,\text{Elec}} = -38\, \text{dBm}$.

Figure 5-23: Signal-to-noise ratio versus collector current for direct photodetection obtain from Figure 5-22.
In general the maximum signal-to-noise ratio for direct detection was around 9 dB better than for down- and up-conversions because 1) direct detection was more efficient while both down- and up-conversions were lossy, 2) in down-conversion the photodetected signal at 3.682 GHz was frequency translated to a frequency region (300 MHz) where the HBT output noise, according to Figure 5-19, was almost 16 dB higher. Even in up-conversion, where the photodetected signal was frequency translated to 7.064 GHz where the HBT output noise was about 5 dB lower than at 3.682 GHz, the correspondingly higher conversion loss still made the direct detection signal-to-noise ratio at 3.682 GHz better. Suppose the HBT was simply used as a direct photodetector at 3.682 GHz and an external microwave mixer having a typical 10 dB single sideband noise figure and 10 dB conversion loss was employed to perform the down- and up-conversions. Also suppose a bandpass filter was used between the HBT direct photodetector and the mixer to suppress the noise in other frequency regions apart from the laser modulation frequency, then for $I_C = 20 \text{ mA}$, the signal-to-noise ratio would be 117 dB (1 Hz) in this HBT/external mixer configuration which is equal to that for the HBT direct photodetector. Compared to this HBT/external mixer configuration, the HBT electrically pumped optoelectronic mixer signal-to-noise ratio penalties for both down- and up-conversions are therefore 5 dB.

In summary, the signal-to-noise ratios for the mixing and direct photodetection were limited by the electrical noise generated in the HBT. To improve the signal-to-noise ratios, the external quantum efficiency of the HBT must be improved. The corresponding increase in the shot noise due to a higher primary photocurrent can be counterbalanced by reducing the base bias, thus maintaining a similar collector current.

Although signal-to-noise ratio is a useful figure of merit for a transistor performance, the spurious-free dynamic range (SFDR) is also vital for many applications. To determine the SFDR, knowledge of the noise level is required and this can be obtained from Figure 5-17, Figure 5-18 and Figure 5-22 for down-conversion, up-conversion and direct photodetection, respectively. SFDRs for these three modes of operation will be reported in the next section.
Section 5-8: Two-Tone Third-Order Intermodulation Distortion Product and Dynamic Range Measurements

Although mixing is a nonlinear process, when an input signal is down- and up-converted to different frequencies, this frequency translation is required to introduce minimal distortion to the input signal. In other words, while the mixing between a large LO signal and some other small input signals is required, mixing among the small input signals should be avoided in order not to generate intermodulation distortion products.

For narrow-band operations in which the bandwidth is smaller than an octave, the third-order products are of interests since they are close to the signals themselves and therefore are difficult to filter out. Second harmonics fall out of the band and can usually be filtered out easily. Suppose the input to a mixer consists of two fundamental frequencies at \( f_1 \) and \( f_2 \), and the mixer is pumped by an LO at \( f_{LO} \). Ideally the up-converted output should only contain the two fundamental frequencies at \( (f_1 + f_{LO}) \) and \( (f_2 + f_{LO}) \) but in practice, third-order products at \( (2f_1 - f_2 + f_{LO}) \) and \( (2f_2 - f_1 + f_{LO}) \) are also produced. A similar situation applies to down-conversion.

Measurements of the third-order intermodulation product power levels together with the noise floor given in the last section allow the SFDR to be determined. The experimental procedures were as follows. Laser 1 and laser 2 were separately modulated at \( CF - 5 \) kHz and \( CF + 5 \) kHz, respectively, where \( CF = 3.682 \) GHz. The two laser beams were then combined in free space in a beam splitter cube and focused on the HBT. The two-tone test signal was then generated in the HBT after detecting the combined laser beam. Equal amounts of incident optical power from each laser on the HBT were maintained by monitoring the primary photocurrent generated in the HBT due to each laser. By varying the amount of optical power incident on the HBT through optical attenuation and slight misalignments, the fundamental output signals and third-order intermodulation distortion products were measured. The equivalent electrical input power was used as the independent variable in this measurement.
The measured fundamental signals and the third-order intermodulation (IM) distortion products as a function of the equivalent electrical input power $P_{RF_{Elec}}$ in down- and up-conversion are shown in Figure 5-24, Figure 5-25, respectively.

Figure 5-24: Third-order intermodulation distortion product characteristics in down-conversion. LO = 3.382 GHz at -4 dBm. CF = 300 MHz, $dF = 5$ kHz. Only distortion products at CF-$3dF$ were measured. $V_{BE} = 0.75$ V, $V_{CE} = 1.5$ V.

Figure 5-25: Third-order intermodulation distortion product characteristics in up-conversion. LO = 3.382 GHz at -4 dBm. CF = 7.064 GHz, $dF = 5$ kHz. Only distortion products at CF-$3dF$ were measured. $V_{BE} = 0.75$ V, $V_{CE} = 1.5$ V.
Also shown in these two figures are the corresponding SFDRs and the third-order intercept points (IP3). Down- and up-conversions showed similar SFDRs of 94 dB (1 Hz) and 96 dB (1 Hz), respectively. Microwave subcarrier multiplexed (SCM) links connecting a satellite earth station with its antennas require a typical SFDR of 90 to 96 dB (1 Hz) [7] and therefore the up-converting HBT electrically pumped optoelectronic mixer should find applications in such a system. The IP3 was +2 dBm for down-conversion and -14 dBm for up-conversion. It is mainly the difference in the system conversion gain for the two processes which leads to a much lower IP3 for up-conversion than for down-conversion.

The third-order IM product characteristics and SFDR of the HBT configured as a direct photodetector were also measured and are shown in Figure 5-26.

Figure 5-26: Third-order intermodulation distortion product characteristics in direct photodetection. $CF = 3.682$ GHz, $dF = 5$ kHz. Only distortion products at $CF-3dF$ were measured. $V_{CE} = 1.5$ V, $I_c = 16$ mA.

The SFDR was 109 dB (1 Hz). The IP3 was +13 dBm. Both the IP3 and SFDR for direct detection were higher than those for down- and up-conversions because of the linear operation of the HBT. If electrically pumped optoelectronic mixing was carried out with the HBT direct photodetector and an external microwave mixer, and a bandpass filter was used to suppress the noise outside the laser modulation frequency, then the SFDR should remain at 109 dB (1 Hz). The SFDR penalties for the HBT optoelectronic mixer
compared to this HBT/external mixer configuration will be 15 dB (down-conversion) and 13 dB (up-conversion). However, the present HBT optoelectronic mixer requires only a low -4 dBm LO drive, is simpler and therefore remains an attractive option.

Section 5-9: Optically Pumped Mixing

So far the HBT optoelectronic mixers described have been of the electrically pumped type in which the LO is supplied electrically while the input signal is supplied optically. Another obvious type is the optically pumped optoelectronic mixer in which the light carries the LO power and the input signal to be down- or up-converted is applied electrically to the base of the HBT. The conversion gain characteristics of the HBT as an optically pumped mixer have been investigated and will be described in this section.

The main difference between performing electrically pumped and optically pumped mixing was the change in the incident optical power and electrical power applied to the HBT base. The conversion gain for the optically pumped HBT optoelectronic mixer is defined in Section 1-2 as the output down- or up-converted electrical signal power measured at the collector to the electrical input available power applied to the base. The experimental procedures were as follows. Laser 1 was biased at 47 mA and directly modulated by a 10 dBm LO source at 3.5 GHz. The modulation depth of the light was measured to be 90 %. The beam splitter and all optical attenuators were removed to maximise the optical power shone on the HBT. The measured primary photocurrent was 0.383 mA, indicating an incident optical power of 1.8 mW. With $V_{BE} = 0 \text{ V}$, $V_{CE} = 1 \text{ V}$, $I_C = 0.383 \text{ mA}$ and both base and collector terminated in 50 $\Omega$ loads, the directly detected LO signal, without transistor amplification, was -25 dBm measured at the collector with a spectrum analyser.

An input signal of -40 dBm at 3.4 GHz was applied to the HBT base. The down- and up-converted signals were measured from the collector at 100 MHz and 6.9 GHz, respectively. The optically pumped optoelectronic down- and up-conversion gains as a function of $V_{BE}$ are shown in Figure 5-27.
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Figure 5-27: Optically pumped optoelectronic mixing down- and up-conversion gain versus $V_{BE}$. $V_{CE} = 1.5$ V, mean incident optical power $= 1.8$ mW, modulation depth $= 90\%$ at 3.5 GHz (LO), input signal freq. $= 3.4$ GHz.

The highest down- and up-conversion gains were -4 dB and -10 dB, respectively. The characteristics shown in Figure 5-27 are very similar to those for the electrically pumped mixing. In particular Figure 5-27 shows that the optimum $V_{BE}$ for the highest down-conversion gain was higher than that for the highest up-conversion gain. Also shown is the increase in the up-conversion gain at high $V_{BE}$. The conversion gains were also comparable with those for electrically pumped mixing.

Apart from this work, only Urey et al [8] have reported optically pumped mixing experiments with an HBT. In their paper, the light source at 1550 nm wavelength delivered the LO power at 6 GHz with almost full modulation. The mean incident optical power on the HBT was boosted up to 15 mW by optical amplification. The RF and the output frequency were not specified. From the input and output powers, a conversion gain of 18 dB is calculated. This high conversion gain is probably due to the very large optical power used. The reason why optical amplification was not used in this work was to avoid damaging the HBTs.

The compression characteristics of the present optically pumped HBT mixer were also measured and are shown in Figure 5-28 for $V_{BE} = 0.78$ V. The 1-dB gain compression point was at around -15 dBm at the input. In electrically pumped mixing, however, no
gain compression point was reached because of the very small photogenerated input signal power.

![Optically pumped mixing gain compression characteristics.](image)

**Figure 5-28:** Optically pumped mixing gain compression characteristics. \( V_{BE} = 0.78 \text{ V}, \ V_{CE} = 1.5 \text{ V}, \) mean incident optical power \( = 1.8 \text{ mW}, \) modulation depth \( = 90 \% \) at 3.5 GHz (LO), input signal freq. \( = 3.4 \text{ GHz}. \)

**Section 5-10: Conclusion**

Experimental results for a three-terminal HBT optoelectronic mixer have been presented. The highest system down-conversion gain of 0 dB (conversion from 3.682 GHz to 300 MHz, Figure 5-17) and up-conversion gain of -10 dB (conversion from 3.682 GHz to 7.064 GHz, Figure 5-18) were obtained. Also an input signal at 100 MHz was up-converted by the current HBT to 20.1 GHz with only 10 dB system conversion loss and requiring an LO power of as low as -3 dBm, suggesting that the HBT can be a compact and highly efficient optoelectronic up-converter for mm-wave fibre-radio links. The measured noise performance in terms of signal-to-noise ratio has been limited by the low external quantum efficiency (17.6 \%) and should improve if lensed fibre is used to launch light onto the HBT or the HBT is designed as an edge-coupled phototransistor. Good SFDRs of 109 dB(1Hz), 94 dB(1Hz) and 96 dB(1Hz) were also obtained for direct photodetection, down-conversion and up-conversion, respectively. In addition to electrically pumped mixing, optically pumped mixing was also performed. The obtained optically pumped mixing down- and up-conversion gains of -4 dB and -10 dB,
respectively, were comparable to those for electrically pumped mixing, and should improve considerably if optical amplification is used.

In the next Chapter, a number of possible system architectures involving the three-terminal HBT optoelectronic mixer will be discussed.
References


Chapter Six

HBT Optoelectronic Mixer System Architectures in Fibre-Radio

Section 6-1: Introduction

In Chapter 1, the role of the HBT optoelectronic mixers in fibre-radio systems was introduced. The system diagrams included only the down-link path in which the data transfer was from the central station to the base station. In practice, for applications such as mobile telephones and wireless LANs, the base station also receives radio signal from the mobile or stationary equipment and this radio signal also needs to be delivered back to the central station for further processing. As a result the connection between the central and base stations should be a duplex link.

In this Chapter a number of system architectures employing an HBT optoelectronic mixer will be considered. The optoelectronic mixer used in the discussion will be the three-terminal HBT electrically pumped type. This three-terminal HBT allows a self-oscillating HBT mixer to be designed. Other competing architectures employing different optoelectronic devices, topologies and modulation methods will also be discussed.

Section 6-2: System Architectures Involving HBTs

In this section, a number of architectures involving HBTs for a simple fibre-radio system consisting of a central station and one of the base stations are considered. Such a fibre-optic network can be employed in systems like mobile cellular networks, wireless local loops [1] and wireless LANs (e.g. HIPERLAN [2]). To allow duplex operation, both the central station and the base station should be able to receive and transmit data over optical fibre to and from each other at the same time. In addition, the base station is required to convert the data on the optical carrier to an electrical signal which is then radiated from the attached antenna and received by the mobile or stationary equipment. The antenna also receives radio signals from the same mobile or stationary equipment and the signals
so received will need to be delivered back to the central station over an optical fibre for further processing.

The straightforward approach is the direct-modulation-direct-detection (DMDD) scheme depicted in Figure 6-1.

![Figure 6-1: A direct-modulation-direct-detection system. PD: photodiode, PA: power amplifier, LNA: low noise amplifier.](image)

In the down-link (central to base station) a laser diode at the central station is directly modulated with an RF signal carrying the information to be delivered to the base station and eventually the mobile or stationary equipment. The output of the laser is then coupled into a fibre. At the base station a photodiode then converts the received light back into the electrical domain. After amplification, the signal is radiated from the antennas. The situation is similar in the up-link in which the signal received by the antenna is first amplified and then used to modulate directly a laser diode at the base station. After transport over the fibre, a photodiode at the central station converts the optical signal into the electrical domain. This approach is simple to implement and works well at low frequencies (a few GHz). In fact it is possible to use low-cost LEDs and multimode fibres in short-range applications (e.g. within a building).

However the demand for high data rate services is increasing and in order to deliver such services over a radio link, a wider bandwidth is required. As a result the radio link should use higher frequency carriers because of the spectrum congestion at low frequencies (<20 GHz). In the European Union's RACE II Mobile Broad System (MBS) project, for
example, radio carriers in the 40 GHz and 60 GHz bands are chosen to deliver broadband (up to 155 Mbits/s) services to mobile users. The highest modulation bandwidth for a semiconductor laser reported to date is around 30 GHz [3] and therefore direct modulation of the laser beyond this limit is not yet practical. The bandwidths of most commercially available, packaged lasers are even lower. Although external modulators can be used and fast external modulators with a 3-dB electrical bandwidth of 75 GHz have been demonstrated in research laboratories [4], they usually require large RF drive power (+18 to +24 dBm) to operate.

One of the alternatives is to deliver the data in the low IF band over fibres in both downlink and up-link directions. At the base station, the incoming optical signal is detected and up-converted to the required frequency band while the received radio signal is down-converted to the low IF. In this approach the lasers in both central and base stations need not have high-speed modulation characteristics and external modulators will not be necessary. Figure 6-2 illustrates the idea using an HBT electrically pumped optoelectronic up-converter.

![Figure 6-2: A simple fibre-radio system employing an HBT electrically pumped optoelectronic up-converter driven by a dielectric resonator oscillator.](image)

The laser diode at the central station can be a low-cost DFB laser diode which is directly modulated by an IF from a few MHz to a few GHz. At the base station, the HBT, pumped by a local dielectric resonator oscillator (DRO), photodetects and up-converts the IF to the required frequency. Other types of LOs can be substituted for the DRO depending on the radio frequency and the costs of the LOs. The cost of a DRO, which is found in most satellite TV receiver dishes working in the 10 GHz band, is so low that
some satellite broadcast companies are now able to give away the receivers for free. Some representative temperature stability figures are 5 ppm/°C maximum for frequencies below 16 GHz and 10 ppm/°C maximum for frequency below 30 GHz [5]. The circulator is used to separate the incoming and outgoing radio signals and can be substituted with a diplexer. In the up-link the received radio signal is mixed down to the low IF band by a microwave mixer before directly modulating a laser diode. Two fibres are used to carry the respective down-link and up-link optical signals.

Recall from Chapter 5 that the three-terminal HBT electrically pumped optoelectronic mixer achieved a system conversion gain of -10 dB when up-converting an input signal from 3.682 GHz to 7.064 GHz. If one uses a commercially available New Focus 1014 fast photodiode, having a responsivity of 10 V/W into a 50 Ω load, and a Watkins-Johnson WJ-M89C double-balanced external mixer, having a conversion loss of 7.5 dB, to perform the same up-conversion, then the system conversion gain will be -23.5 dB. Therefore the three-terminal HBT up-converter provides 13.5 dB advantage over that achieved by using this commercial combination for the down-link link budget. Adding an extra amplifier to this combination will reduce this advantage, but it will also increase the component count. However the 45 GHz bandwidth of the New Focus photodiode and the 18 GHz bandwidth of the Watkins-Johnson mixer allow a more uniform frequency response.

Using two optical fibres to carry the down-link and up-link signals separately between each central and base station eases the system complexity and can be considered economical if both the central and other base stations are in proximity (e.g. within a building). There are other situations where using two fibres for each base station becomes expensive. For example if the central station is situated at a railway terminal and all the base stations are distributed along an intercity railway line in order to serve the train passengers, then the total length of the fibre required will increase significantly due mainly to the more remotely situated base stations. Using wavelength division multiplexing (WDM) in which two different optical carriers of different wavelengths are used to carry both down-link and up-link information on a single fibre for each base station, the total fibre length is then halved. Figure 6-3 illustrates one such system in
which the down-link laser wavelength is 1.55 \( \mu m \) and the up-link laser wavelength is 1.3 \( \mu m \). WDM couplers are used to combine and separate the two optical carriers.

As has been demonstrated by Sawada et al. [6,7], a self-oscillating HBT can function as an optoelectronic mixer without requiring a separate LO. There are broadly two types of oscillators: fixed frequency and tunable oscillators.

For the fixed frequency type, it is popular to use dielectric resonators because of their high quality-factors Q (3000 to 10,000) and high temperature stability (better than \( \pm 10 \) ppm/°C). Figure 6-4 shows a fixed frequency self-oscillating HBT optoelectronic up-converter employing a dielectric resonator at the base station.
In Figure 6-4, the emitter series capacitor provides the required feedback to make the HBT potentially unstable. The HBT base is then connected to a dielectric resonator which, at the resonance frequency, presents the HBT an input reflection coefficient that is within the unstable region and causes the HBT to oscillate. Because of the oscillation of the HBT, the photodetected IF signal can therefore be up-converted. The LO required by the down-converter in the up-link can also be derived from the HBT output using a diplexer. Since the HBT oscillation output power is usually smaller than 0 dBm, an amplifier may be necessary to increase the HBT output to a sufficient level (around +10 dBm) to drive the down-converting microwave mixer.

Tunable frequency oscillators are required in those situations in which the base station LO frequency needs to be remotely controlled in order to change the transmitted radio frequency. If this is the case, it may be useful to design the HBT as an optically injection-locked self-oscillating mixer (OILSOM). Figure 6-5 illustrates this idea.

The HBT is in the common-base configuration where the inductor attached to the base is used to increase both the input (emitter) and output (collector) reflection coefficients. The emitter capacitance is chosen to present to the HBT an input reflection coefficient within the unstable region and cause the HBT to oscillate. The LO modulated light transported from the central station on an optical fibre then illuminates the HBT. If the HBT self-oscillation frequency and the LO frequency are close enough, the HBT frequency will follow that of the LO within a limited frequency range and hence can be
controlled remotely from the central station. The incident optical signal carrying the IF is first detected by a photodiode. The electrical IF is then fed to the HBT base. Since the HBT oscillator is operated nonlinearly, the IF will be up-converted by the HBT. An optically injection-locked self-oscillating mixer which is based on MMIC MESFETs has been reported by Callaghan et al. [8]. The oscillation frequency was $f_0 = 6.3 \text{ GHz}$ with a locking range of $8.5 \text{ MHz}$ (0.13 %). An electrical conversion gain of about 1 to 2 dB was obtained when an electrical RF signal at either $f_0 - f_{IF}$ or $f_0 + f_{IF}$ was down-converted to an electrical IF at $f_{IF}$ where $f_{IF}$ is from 50 MHz to 200 MHz.

A number of different architectures involving the HBT optoelectronic mixers and HBT OILSOM have been described with applications in fibre-radio duplex systems. There are many other different approaches to the same problems and some of these will be discussed in the next sections.

**Section 6-3: Other Approaches to Fibre-Radio Systems**

If there was no frequency limitation on optoelectronic devices and no dispersion from optical fibres, then the DMDD architectures shown in Figure 6-1 would offer the simplest and cheapest solution to microwave/millimetre wave fibre-radio systems. Since the highest direct modulation bandwidth of semiconductor lasers is around 30 GHz, alternatives must be found if the operation frequency is to be increased. The HBT optoelectronic mixers (Figure 6-2 to Figure 6-4) and HBT optically injection-locked self-oscillating mixer (Figure 6-5) discussed in the previous section are some of those alternatives in which only the low frequency IF signal is transported over the fibre and frequency conversions to and from the higher microwave or millimetre-wave (mm-wave) band take place at the base station.

One of the key issues in mm-wave fibre-radio systems is the generation of the mm-wave carrier which is required by both the up- and down-converters at the base station. It is generally considered that generating the mm-wave carrier at each base station is more expensive than delivering it from the central station because each base station will need its own mm-wave source. Delivering the mm-wave carrier from the central station also
allows the radio frequency radiated from each base station to be controlled centrally in one place.

Many techniques have been devised to generate mm-wave carriers. These include optical heterodyning of two lasers [9,10], optical heterodyning of a single dual-mode laser [11], optical phase-lock loop involving one master and one slave laser [12], optical injection phase-lock loop, optical feed-forward optical field modulation [13], external modulators directly driven by a mm-wave source and frequency doubling using double-sideband suppressed carrier (DSB-SC) in an external modulator driven at half the desired mm-wave frequency [14]. The generated mm-wave carrier, transported on a separate optical fibre, can be used to injection lock the HBT OILSOM shown in Figure 6-5, or can first be photodetected, amplified and used to drive a microwave up-converting mixer directly. The laser at the central station is still modulated by a low IF.

At the central station it is also possible to first up-convert the IF to the mm-wave band which is then transported together with the mm-wave carrier over a single optical fibre. At the base station, a high-speed photodiode is then required to detect the incident light. A diplexer is used to separate the mm-wave carrier and the signal (see Figure 6-4 and Figure 6-5). The mm-wave signal, after further amplification, is radiated from the antenna. The mm-wave carrier, recovered by the diplexer, is amplified and used to drive the microwave down-converter. A simple implementation of this approach is to first mix the IF signal with a mm-wave carrier in a microwave mixer at the central station. The mixer output, which contains the mm-wave carrier and two sidebands centred round the mm-wave carrier representing the up-conversion of the IF signal, is then used to drive an external Mach-Zehnder modulator whose optical input is provided by a free-running DFB laser.

There are three drawbacks with the approach just described. First a high-speed photodiode is required at the base station to detect the mm-wave signal and the carrier. Since a high-speed photodiode usually has a low responsivity because of its small area, the detected mm-wave signal and carrier powers will be weak. Also the Mach-Zehnder modulator is inherently a nonlinear device because of its sinusoidal transfer characteristics, the resultant high intermodulation distortion can limit the dynamic range when the modulator is heavily driven. Finally the output light from the modulator
contains optical sidebands on either side of the optical carrier for each input modulating frequency. Now consider only the mm-wave carrier modulation and its two resultant optical sidebands. At the receiver, these two optical sidebands beat with the optical carrier and generate two beat signals which, when added constructively, produce a single signal component which is the original mm-wave carrier. The relative phase shifts induced by fibre chromatic dispersion between the upper and lower sidebands and the optical carrier, however, cause the two beat signals to be out of phase and lead to power degradation of the detected signals and will limit the fibre transmission distance [15]. When the phase difference becomes $\pi$, complete signal cancellations occur. The same situation applies to the other optical sidebands resulting from the up-conversion of the IF signal.

In order to overcome the intermodulation distortion problem of driving a Mach-Zehnder modulator with a composite signal, Park et al. [16] suggested that a Mach-Zehnder modulator be used to up-convert optically the intensity modulation of a low frequency, commercially available linearised laser diode which is directly modulated with the low IF signal. The intermodulation distortion characteristics depend mainly on the laser diode because the external modulator simply multiplies the laser modulation by the mm-wave carrier. Since also the external modulator is driven by the mm-wave carrier, only narrow band matching between the mm-wave source and the external modulator is required. However this approach still suffers from fibre chromatic dispersion and requires a high-speed photodiode at the base station.

In order to reduce the fibre-dispersion when transmitting mm-wave modulated optical signals over fibres, optical single sideband (SSB) modulation has been used. Yonenaga et al. [17] optically suppressed one of the two resultant modulation sidebands to obtain the optical SSB modulation. Smith et al. [18] instead employed a dual-electrode Mach-Zehnder modulator biased at quadrature and driven with signals at the RF electrodes which were $\pi/2$ out of phase, and showed that using this optical SSB modulation to distribute 2-20 GHz signals over 79.6 km of single-mode fibre, only a 1.5 dB degradation in RF power due to fibre dispersion was experimentally observed. These optical SSB modulation techniques, however, still require a high-speed photodiode at the base station when the modulation frequency is high.
In all the architectures discussed so far, the tasks of photodetection and optical intensity modulation at the base station have been accomplished separately using a photodetector and a directly modulated laser diode or an external modulator. Recently interest in electroabsorption modulator (EAM) transceivers has increased where an EAM is used both as a photodetector as well as an external intensity modulator [19]. Figure 6-6 illustrates an EAM in a loop-back configuration functioning as a photodetector and an intensity modulator at a base station.

In Figure 6-6 the down-link IF signal directly modulates a laser diode whose optical output is then coupled into a fibre. At the base station, the EAM photodetects this down-link optical signal. Since a loop-back configuration is used, there will be residual light coming out of the other optical port of the EAM and this light is coupled into the up-link fibre. Therefore when the EAM is simultaneously driven with the up-link IF signal, the residual light will carry the up-link IF signal back to the central station. Since the same optical carrier is used in both directions, it is important that different down-link and up-link frequencies are used. The main advantage of the EAM transceiver is the reduced component count in the base station by performing both photodetection and intensity modulation with a single device.

The use of unbiased EAM transceivers has received particular attention since they require no electrical power from the base station. Furthermore if the radio frequency is only a
few GHz and no frequency conversion is necessary, then in short-range applications (e.g. in an office) a completely passive, lightweight and cheap base station can be built requiring no amplifiers, oscillator and hence no power supply. D. Wake et a. [20] demonstrated an optically-fed radio access point which consisted only of an antenna and an unpowered EAM transceiver. A wireless Ethernet connection at 3 Mbits/s operating in the 2.4 GHz band between the radio access point and a mobile laptop computer fitted with a radio modem was successfully demonstrated in an 6x3 m\(^2\) office.

**Section 6-4: Discussions and Conclusion**

In this Chapter, a number of architectures for fibre-radio systems have been described and discussed. In particular in Section 6-2, the HBT electrically pumped optoelectronic up-converter allows a low frequency IF signal (instead of the mm-wave signal) to be transported over the fibre and up-converts the detected IF to a higher frequency band. Therefore a low-frequency DFB laser can be directly modulated without requiring an external modulator which usually requires over +15 dBm drive power. Since also the optical intensity is modulated at low frequency, fibre chromatic dispersion is insignificant.

Comparing to using a commercially available New Focus 1014 fast photodiode, having a responsivity of 10 V/W into a 50 Ω load, and a Watkins-Johnson WJ-M89C double-balanced external mixer, having a conversion loss of 7.5 dB, to perform up-conversion from 3.682 GHz to 7.064 GHz, the three-terminal HBT electrically pumped optoelectronic mixer will provide 13.5 dB advantage in the down-link link budget. Although adding an extra amplifier after the New Focus photodiode will reduce this advantage, it will increase the component count.

The LO source required by the HBT optoelectronic up-converter can be provided by a low-cost but highly stable DRO or the LO signal can be generated by a self-oscillating HBT optoelectronic mixer as depicted in Figure 6-4. If the LO frequency needs to be remotely controlled, an optically injection-locked self-oscillating HBT mixer can be designed.
Generating the LO at the base station is considered expensive because each base station will need its own LO source. A number of schemes involving optical heterodyning and external modulators to generate and deliver the mm-wave LO have been briefly mentioned in Section 6-3. Delivering the mm-wave signal and carrier over fibre to the base station requires a high-speed photodiode at the base station and such a photodiode usually has a low responsivity because of its small size. Fibre chromatic dispersion can be a problem if optical SSB modulation is not employed. Optically filtering of one of the sidebands requires complicated optics while the dual-electrode Mach-Zehnder method requires large RF drive (+10 dBm at each electrode).

EAM transceivers perform both photodetection and intensity modulation in one device. However the performance can be of concern because the optimum bias conditions for the EAM to act as a photodetector are in general different from those when the EAM is used as a modulator. Unbiased EAMs offer the ultimate design simplicity, but their performance has only been experimentally investigated at low frequencies (2.4 GHz) in a small office (6×3 m^2).

There is no one single architecture which satisfies all the performance requirements and provides the cheapest and simplest solution. The exact adaptation of any of the architectures discussed in this Chapter depends on the operation frequency, the fibre link, linearity requirement, cost and others. From the previous two experimental Chapters, it is observed that although the single HBT optoelectronic mixers might not outperform other hybrid architectures, their simplicity is the factor which makes them attractive for those systems which do not require state-of-the-art performance but do require low-cost design. The performance of a self-oscillating and an OILSOM HBT optoelectronic mixer will require further experimental investigation.
References


Chapter Seven

Summary and Conclusion

Section 7-1: Summary

Section 7-1a: Summary of Chapter 1

In Chapter 1, the subject of optoelectronic mixing was introduced by describing and distinguishing between the electrically pumped and optically pumped optoelectronic mixers. In electrically pumped mixing, the large-signal local oscillator (LO) source is applied to the optoelectronic mixer electrically while the input is an RF modulated optical source and output signal is an electrical signal at the intermediate frequency (IF). In optically pumped mixing, the LO is supplied to the optoelectronic mixer optically while both the RF input and the IF output are electrical. The conventional approach to optoelectronic mixing is to employ a separate photodiode to convert the RF or LO modulated optical beam into an electrical signal which is then mixed in a microwave mixer with the electrical LO or RF, depending on whether it is electrically pumped or optically pumped mixing, respectively. In this work, optoelectronic mixing has been carried out using two single HBTs. The two HBTs are a two-terminal edge-coupled InP/InGaAs HBT provided by the BT Labs and a three-terminal normal-incidence InP/InGaAs HBT provided by the project collaborators at the Technion in Israel.

In order to characterise and compare the achieved performance with other published results, definitions for the system and intrinsic conversion gains for electrically pumped optoelectronic mixers were developed. The conversion gain of an optically pumped optoelectronic mixer is defined identically to that for a normal microwave mixer because both the input and output signals are electrical.

The applications of HBT optoelectronic mixers as photodetecting and frequency converting devices were described using examples of the subcarrier multiplexed and fibre-radio systems. Previous work on optoelectronic mixing using different devices was
summarised and the current status of HBT optoelectronic mixers, direct photodetectors and optically controlled oscillators were reviewed in detail and summarised in Table 1-1.

So far all other reported HBT optoelectronic mixers have used only three-terminal HBTs and the system conversion gains have also been small (<0 dB). In factor the only two-terminal HBT optoelectronic mixer reported has been that of this work [1]. It can be seen in Table 1-1 that although reported results for HBT direct photodetectors have been encouraging, results from HBT optoelectronic mixers have not produced good system conversion gains.

**Section 7-1b: Summary of Chapter 2**

In Chapter 2, the principles of the optical control of HBTs were explained. Firstly the photodetection process in an HBT can be separately and equivalently represented by an illuminated PN photodiode connected between the base and the collector terminals. Secondly, the response of the HBT to the optical illumination can be determined from the HBT electrical response to the primary photocurrent generated by the PN photodiode. Therefore once the primary photocurrent has been determined from the incident photon flux density and the HBT structure, the photo-HBT can then be analysed purely as an electrical device. The optical absorption processes in the three-terminal normal-incidence and the two-terminal edge-coupled InP/InGaAs HBTs were considered and the expressions for the two primary photocurrents in these two HBTs were derived.

Since the principles of the HBT and BJT are closely related, the basic operation of the bipolar transistor was summarised in Section 2-3. From the BJT current-voltage equations given in Section 2-4, the static common-base current gain, emitter efficiency, base transport factor and the common-emitter current gain were related analytically to the transistor structural and doping parameters. It was shown that a high common-emitter current gain requires that both the base transport factor and the emitter efficiency be close to unity. In a BJT, a high emitter efficiency is obtained by doping the emitter more heavily than the base and as a result, the base resistance is high and cannot be made very thin which also lowers the base transport factor. The heterojunction introduced in Section 2-5 shows that an HBT having a wide bandgap emitter and a narrow bandgap base can maintain the emitter efficiency close to unity despite having a more heavily doped base
relative to the emitter. As a consequence, the heavily doped base having low resistance can be made very thin which not only increases the base transport factor but also reduces the base transit time which is important for high frequency operations. However a heterojunction nevertheless introduces a number of localised states in the depletion region because of the inevitable lattice mismatch between the two constituent semiconductors. As a result, a space-charge recombination current exists in the base-emitter heterojunction which, as explained in Section 2-6, not only reduces an HBT common-emitter current gain at low collector currents, but also makes the common-emitter current gain vary with the collector current. Since the collector current is in turn controlled by the base-emitter voltage, if an LO is applied across an HBT base-emitter terminal, the current gain will become time-varying at the LO frequency which will provide a mixing mechanism when the HBT simultaneously detects an intensity modulated light.

In Section 2-7, the HBT frequency performance limitations were considered. These limitations are the emitter charging time, the base transit time, the collector delay time and the collector capacitance charging time. These four time constants were individually examined and used to provide an alternative and simplified derivation for the current cutoff frequency expression.

In Section 2-8, the optoelectronic mixing mechanisms in HBT were briefly explained in terms of the transistor small-signal gain dependent on the collector current which is in turn modulated by the applied LO.

**Section 7-1c: Summary of Chapter 3**

In Chapter 3, the concept of harmonic-balance and the conversion matrix was introduced, and the modelling and simulation results for the two HBT electrically pumped optoelectronic mixers using such techniques presented. It was explained in Section 3-1 that modelling an electrically pumped mixer involves two steps. The first step is to use harmonic-balance to model the mixer response to a large-signal LO excitation without considering the small-signal photogenerated RF signal. The result in this part of the analysis is the LO induced time-varying voltage waveform at each node of the mixer circuit model. The second step is to analyse the mixer in a quasilinear fashion in which the sole input to the mixer is the photogenerated RF signal without considering the LO
any more. Frequency generation at the IF and other image frequencies are dealt with in the analysis using conversion matrices which are determined from the time-varying small-signal conductances of the nonlinear elements in the model.

Using harmonic-balance techniques to find the steady-state large-signal response of a circuit has a number of advantages compared to pure time-domain analysis. Firstly harmonic-balance allows the linear parts of the model to be described and simulated in the frequency-domain and the nonlinear parts in the time-domain, and forward and inverse Fourier transforms are used to bridge the two parts. Therefore the problem of having to describe the responses of the linear, frequency-dependent elements in the time-domain is avoided. Secondly harmonic-balance attempts to find the steady-state solution from the outset and so no time is wasted on the transient. Since the mixer is analysed separately for the LO and photogenerated RF signal excitations, therefore the number of simulation points and hence the computation time will be independent of the frequency relationship between the LO, IF and RF.

Using a simple model consisting of a linear subcircuit and a nonlinear subcircuit directly connected together, it was explained in Section 3-2 that the aim of the harmonic-balance analysis is to find the voltage which appears across both the linear and nonlinear subcircuits so that the sum of the current going into the linear subcircuit and that going into the nonlinear subcircuit is zero, and hence balanced. The algorithm used for solving the harmonic-balance problems was the splitting method and was explained in Section 3-3. Once this voltage is found, the next step is to construct a conversion matrix for each nonlinear element for calculating the conversion gain for the mixer which was explained in Section 3-4.

In Sections 3-5 and 3-6, the implementations and simulation results for the three-terminal and the two-terminal HBT electrically pumped optoelectronic mixers were presented, respectively. In the three-terminal HBT simulation, the HBT model used was the T-topology equivalent circuit whose parameters were obtained from a combination of device geometry estimations, parameter extractions and optimisations from the DC and s-parameter measurements, as well as from the equivalent circuit of the bias-T manual. In the two-terminal HBT simulation, however, the HBT model used was a quasi-static one which describes the device terminal behaviour. The use of a quasi-static model was due
to the lack of an electrical base terminal which would otherwise allow the transistor internal parameters to be measured and used for the construction of a more physical equivalent circuit model. The simulation results were commented on and compared with the experimental results in Chapters 4 and 5.

Section 7-1d: Summary of Chapter 4

In Chapter 4, a novel two-terminal edge-coupled HBT electrically pumped optoelectronic mixer was described and experimental results presented. With the collector grounded, this two-terminal HBT only had an electrical emitter terminal available and a diplexer was required to separate the applied LO and other outputs from the HBT.

The measured static responsivity increased with the amount of the incident optical power. For example at $V_{CE} = 1\,\text{V}$, the responsivity and the photocurrent gain increased from 60 A/W and 48, respectively, at $P_{opt} = 0.06\,\text{mW}$ to 105 A/W and 84, respectively, at $P_{opt} = 0.18\,\text{mW}$. The dependence of the responsivity and the photocurrent gain on the incident optical were due to the base-emitter junction recombination current which was explained in Section 2-6. In direct photodetection, the low frequency system signal gain increased with increasing collector current because of the base-emitter junction recombination current. The 3 dB bandwidth also increased with the collector current because of the reduced emitter dynamic resistance. The 3 dB electrical bandwidth was 280 MHz and the unity system signal gain frequency was 6 GHz for $P_{opt} = 0.3\,\text{mW}$, $V_{CE} = 1\,\text{V}$ and $I_C = 15\,\text{mA}$.

In the first electrically pumped experiment, with the LO at 2.91 GHz and 10 dBm, 0.14 mW mean incident optical power at 1550 nm wavelength, the laser modulation frequency (RF) at 2.41 GHz with 10 % modulation depth, output IF at 500 MHz, a maximum system conversion gain of 7 dB was achieved with $V_{CE} = 0.6\,\text{V}$. This 7 dB gain is significant because it states that such a single two-terminal HBT optoelectronic mixer would have a 7 dB gain advantage over an electrically pumped optoelectronic mixer built using a hypothetical 100 % quantum efficient photodetector and an ideal microwave mixer with 0 dB conversion loss in a 50 $\Omega$ impedance system. With reference to Table 1-
1, this two-terminal HBT has also achieved the highest system conversion gain ever reported for a single HBT optoelectronic mixer although the frequencies used were different. Furthermore the LO power level used to achieve this 7 dB gain is also comparable to those required to drive most commercially available microwave double-balanced mixers. Unfortunately this excellent HBT sample subsequently failed and a second HBT was used for further experiments and it is this second HBT sample that was modelled in Chapter 3. For RF = 3.025 GHz, LO = 2.925 GHz, $P_{opt} = 0.3 \text{ mW}$, down-conversion frequency at 100 MHz and up-conversion frequency at 5.95 GHz, the maximum system conversion gains for both down- and up-conversions were around -1 dB. The system conversion gains of this second transistor were several dB smaller than that of the first one mainly because of a lower responsivity. For example, the static responsivity of the second HBT was $37.5 \text{ A/V}$ at $V_{CE} = 1 \text{ V}$ and $P_{opt} = 0.18 \text{ mW}$. However under the same bias conditions, the first HBT responsivity was $105 \text{ A/W}$. The lower responsivity could in turn be due to a higher base-emitter junction recombination current, although this could not be verified experimentally without an electrical base contact for constructing a Gummel Plot.

The simulation results from Chapter 3 were compared with the experimental system conversion gain characteristics and they were in good qualitative agreement. In particular, the model correctly predicted that both down- and up-conversion gains would initially decreased as $V_{CE}$ was increased from 0 V, and reached a maximum in the range $0.6 \text{ V} < V_{CE} < 0.9 \text{ V}$. However, because of the assumption in the model that the mixing mechanism is the small-signal photocurrent gain dependence on $V_{CE}$, the model did not distinguish between down-conversion and up-conversion, and therefore the simulation results were virtually identical in both processes. Clearly some differences were observed in the measured system down- and up-conversion gains, especially for $V_{CE} > 0.8 \text{ V}$ where the measured system down-conversion gain started to increase with $V_{CE}$ while the measured system up-conversion did not. This difficulty was due to the high-level assumption used in the model for generating the mixed products. Physically, the mixed products were generated by the nonlinear base-emitter and base-collector junctions of the HBT. Since the down-conversion and up-conversion products were a few GHz apart, the base-emitter and base-collector capacitances would affect the phases and amplitudes of
the currents at the down- and up-conversion frequencies differently, and hence different
down- and up-conversion characteristics would result. The model of Chapter 3, however,
did not take this factor into account. The discrepancies between the experimental and
simulation results were also due to the quasi-static nature of the model. The terminal
behaviour of the HBT was modelled by incorporating the transistor static IV
characteristics under optical illumination. Clearly when the LO was at a few GHz, the
terminal impedance presented by the HBT to the LO source would be different from what
is predicted simply from the inverse of the slope of the plot in Figure 4-12 because of the
HBT junction capacitances. As a result the voltage swing caused by the LO at the emitter
would not be accurately predicted by the quasi-static model and hence further
discrepancies between the experimental and simulation were caused.

The noise performance of the HBT as a direct photodetector and as an electrically
pumped optoelectronic mixer was also measured using a very low RIN laser (RIN < -160
dB/Hz below 8 GHz). The output electrical noise spectrum of the directly detecting HBT
under different constant optical illuminations showed similar characteristics as the
frequency response of the HBT and therefore suggested that the output noise was actually
generated from within the HBT and amplified by the HBT gain. A simple calculation
further verified that the measured output noise was not laser RIN limited. A maximum
signal-to-noise ratio (S/N) of 115 dB (1 Hz) was achieved for direct photodetection at
3.025 GHz with 0.3 mW mean incident optical power and 10 % modulation depth. Under
the same optical illumination, the S/N was 86 dB (1 Hz) for down-conversion and 103 dB
(1 Hz) for up-conversion with $V_{CE} = 0.6$ V. The S/N for down-conversion was much
smaller than that for up-conversion because the HBT down-converted the photodetected
RF signal to a low frequency region where the HBT exhibited an output noise power
which was over 20 dB higher than at the up-conversion frequency.

The intermodulation distortion characteristics of the HBT were measured using a two-
laser approach in which the two lasers were separately modulated at two frequencies and
combined in free-space before being focused onto the HBT. This two-laser approach
ensures that the intermodulation distortion products were generated by the HBT but not
the laser. The spurious-free dynamic range (SFDR) was 95 dB (1 Hz) in direction
photodetection, 78 dB (1 Hz) in down-conversion and 90 dB (1 Hz) in up-conversion.
The much smaller SFDR in down-conversion compared to the up-conversion was again due to the conversion of the photodetected RF signal to a much noisier region at the down-conversion frequency. Compared to performing optoelectronic mixing with the HBT direct photodetector and an external microwave mixer, the penalty in SFDR for using a single HBT up-converter is expected to be less than 5 dB.

**Section 7-1e: Summary of Chapter 5**

Chapter 5 was concerned with the experimental results for a three-terminal normal-incidence HBT optoelectronic mixer. Details of the HBT fabrication and structure were given in Section 5-2. The light was normally incident on the HBT via a 5×6 μm² optical window located on the base mesa. The InGaAs base and collector were 50 nm and 750 nm thick, respectively, and the measured static responsivity for $V_{BE} = 0$ V was 0.22 A/W at 1550 nm wavelength. This number translates to a measured external quantum efficiency of 17.6 %. Based on a simple calculation taking into account the reflections at the air-polyimide and polyimide-base interfaces, the photon absorption in the base and collector, and the small spot size of the focused beam compared to the HBT optical window area, the expected external quantum efficiency was calculated to be 22 % which is not far from the measured 17.6 %. The frequency response of the HBT to intensity modulated light in direct photodetection was also measured. For $I_c = 20$ mA and $V_{CE} = 1$ V, the 3 dB electrical bandwidth was around 500 MHz and the unity system signal gain frequency was 3.5 GHz.

In the electrically pumped optoelectronic mixing experiments, the HBT base was driven by a 3.4 GHz LO while detecting a 3.5 GHz RF modulated light with 70 μW mean optical power and 90 % modulation depth. With $V_{CE} = 1.5$ V, the down-converted signal power at 100 MHz and the up-converted signal power at 6.9 GHz were measured as a function of $V_{BE}$ and the LO power. The highest system conversion was -4 dB in down-conversion ($V_{BE} = 0.82$ V and LO = -4 dBm) and -11 dB in up-conversion ($V_{BE} = 0.68$ V and LO = -3 dBm). The measured down- and up-conversion gain characteristics were compared with the simulation results, and for $V_{BE} \leq 0.80$ V and LO $\leq -8$ dBm, agreement within 4 dB was observed for down-conversion and 7 dB for up-conversion. When the HBT was
driven by larger LO powers, the model became inaccurate. For $V_{BE} = 0.9$ V and LO $\geq -2$ dBm, the discrepancy was around 10 dB for down-conversion, and around 20 dB for up-conversion. The reason for the discrepancies could be due to the fact that at such large LO levels, the base-collector junction was momentarily forward-biased, giving rise to an extra mixing mechanism. However, the HBT model assumed that the base-collector junction was always reverse-biased. Therefore, the accuracy of the model is expected to improve when the base-collector junction is represented fully by a large-signal diode model. To verify the significance of the momentarily forward-biased base-collector junction, electrically pumped optoelectronic mixing was also carried out with $V_{CE}$ reduced to 1 V. For a -3 dBm LO at 2.9 GHz, RF = 3 GHz, it was observed that the system up-conversion gain increased to -7.5 dB at $V_{BE} = 0.92$ V.

The frequency responses in both down- and up-conversions were measured and results presented in Section 5-6. In down-conversion, the intermediate frequency was fixed at 100 MHz while both the laser modulation frequency ($f_{RF}$) and the LO frequency ($f_{LO}$) were swept according to

$$f_{RF} = f_{LO} + 100 \text{ MHz}$$

The mean incident optical power was 0.3 mW, $V_{BE} = 0.82$ V and LO = -10 dBm. The system conversion gain decreased from around -6 dB at $f_{RF} = 3.1$ GHz to -24 dB at $f_{RF} = 9.6$ GHz. Higher $f_{RF}$ could not be used because of the limited laser direct modulation bandwidth.

In the measurement of the up-conversion frequency response, the laser modulation frequency was fixed at 100 MHz, the LO frequency was swept from 3 GHz to 20 GHz and the corresponding up-converted signal was measured from 3.10 GHz to 20.10 GHz. For $V_{BE} = 0.92$ V and LO = -3 dBm, the system up-conversion gain decreased from +6 dB at $f_{LO} = 3$ GHz to -10 dB at $f_{LO} = 20$ GHz. Such a low system conversion loss (10 dB) in the 20 GHz region indicated that the HBT could be a compact and high-performance optoelectronic up-converter for mm-wave fibre-radio applications.

The noise performance of the HBT was described in Section 5-7. Two microstrip filters were inserted after the HBT collector to prevent the low-noise amplifier from being
saturated by the LO leakage. After the insertion of the two filters, the maximum system down- and up-conversion gains, with -4 dBm LO at 3.382 GHz, RF = 3.682 GHz and $V_{ce} = 1\text{V}$, were measured to be 0 dB and -10 dB, respectively, which were 4 dB and 1 dB higher than the previously measured values. The reason could be the insertion of the two bandstop filters in these measurements which changed the impedance seen by the HBT collector, and as a result the base-collector junction was subjected to a different AC voltage, affecting the mixing process. Previous results were obtained without using any filters. With 70 $\mu$W mean incident optical power and 90 % modulation depth, the highest signal-to-noise ratios occurred at $V_{BE} = 0.6 \text{V}$ in down-conversion and at $V_{BE} = 0.7 \text{V}$ in up-conversion, and were comparable at around 112 dB (1 Hz). The signal-to-noise ratio for direction photodetection at 3.682 GHz was 117 dB (1 Hz) when $I_c = 20 \text{mA}$.

The third-order intermodulation distortion characteristics were measured using a two-laser approach. With LO = -4 dBm at 3.382 GHz, the two laser modulation frequencies centred around 3.682 GHz and separated by 10 kHz, the SFDR was 94 dB (1 Hz) for down-conversion and 96 dB (1 Hz) for up-conversion. Microwave subcarrier multiplexed (SCM) links connecting a satellite earth station and its antennas require a typical SFDR of 90 to 96 dB (1 Hz) [2] and therefore the up-converting HBT electrically pumped optoelectronic mixer should find applications in such a system. The SFDR for direct photodetection was 109 dB (1 Hz).

Optically pumped optoelectronic mixing has also been carried out with the three-terminal HBT. The HBT was optically pumped with a 1.8 mW mean optical power modulated at 3.5 GHz with 90 % modulation depth. An electrical RF signal at 3.4 GHz was applied to the base, and the down- and up-converted signals were measured from the collector at 100 MHz and 6.9 GHz, respectively. The maximum down- and up-conversion gains were -4 dB and -10 dB, respectively. The 1-dB down- and up-conversion gain compression points were -15 dBm at the input.

*Section 7-1f: Summary of Chapter 6*

In Chapter 6, the system aspect of the HBT optoelectronic mixer was considered in terms of a number of fibre-radio system architectures. Although the direct-modulation-direct-
detection offers the simplest architecture, the upper frequency limit of directly modulating a laser (30 GHz) and high power requirement by the external modulator (+18 dBm to +24 dBm) confines its applications to the low frequency region (a few GHz). The HBT electrically pumped optoelectronic up-converter allows a low frequency IF signal (instead of the mm-wave signal) to be transported over the fibre which is then detected and up-converted by the HBT to a higher frequency band at the base station. Therefore the laser only needs to be modulated at the low IF without requiring an external modulator which usually requires over +15 dBm drive power. Since also the optical intensity is modulated at low frequencies, fibre chromatic dispersion is insignificant. The LO source required by the HBT optoelectronic up-converter can be provided by a low-cost but highly stable DRO. Alternatively a self-oscillating HBT optoelectronic mixer can be designed. Furthermore if the LO frequency needs to be remotely controlled, an optically injection-locked self-oscillating HBT mixer can be designed.

Generating the LO at the base station is considered expensive because each base station will need its own LO source. A number of schemes involving optical heterodyning and external modulators to generate and deliver the mm-wave LO were briefly mentioned in Section 6-3. Delivering the mm-wave signal and carrier over fibre to the base station requires a high-speed photodiode at the base station and such a photodiode usually has a low responsivity because of its small size. Fibre chromatic dispersion can be a problem if optical SSB modulation is not employed. Optically filtering of one of the sidebands requires complicated optics while the dual-electrode Mach-Zehnder method requires large RF drive (+10 dBm at each electrode).

EAM transceivers perform both photodetection and intensity modulation in one single structure. However the performance can be of concern because the optimum bias conditions for the EAM to act as a photodetector are in general different from those when the EAM is used as a modulator. Unbiased EAMs offer the ultimate design simplicity, but their performance has only been experimentally investigated at low frequencies (2.4 GHz) in a small office (6×3 m²).

There is no one single architecture which satisfies all the performance requirements and provides the cheapest and simplest solution. The exact adaptation of any of the
architectures discussed in Chapter 6 depends on the operation frequency, the required link gain, linearity requirement and cost. From the experimental evidence, it is expected that a signal HBT electrically pumped optoelectronic up-converter, for simplicity, and an HBT direct photodetector-external mixer, for performance, will be useful in mm-wave over fibre systems. The performance of a self-oscillating and an OILSOM HBT optoelectronic mixer will require further experimental investigation.

Section 7-2: Conclusion and Future Work Suggestions

In conclusion, the work described in this Thesis has resulted in a number of novel achievements, including

i. The first two-terminal edge-coupled InP/InGaAs HBT electrically pumped optoelectronic mixer, measurements of the optoelectronic mixing conversion gain characteristics, noise performance characteristics and intermodulation distortion characteristics;

ii. Highest reported optoelectronic mixing system conversion gain (defined in Section 1-2) of +7 dB (RF = 2.41 GHz, LO = 2.91 GHz, IF = 500 MHz) using the two-terminal HBT;

iii. The first two-terminal and first three-terminal HBT electrically pumped optoelectronic mixer computer models using harmonic-balance techniques;

iv. The first experimental results on the intermodulation distortion characteristics of a three-terminal InP/InGaAs HBT optoelectronic mixer using a two-laser approach;

Table 7-1 compares the performance of the two HBT electrically pumped optoelectronic mixers with other reported results. It can be seen that the results achieved as a result of this work are competitive with others.

Compared to the conventional approach to optoelectronic mixing which involves a separate photodiode and an external microwave mixer, the two HBT optoelectronic mixers not only provide a much simpler alternative but also exhibit a gain advantage.
Table 7-1: Comparison of the results for the two HBT electrically pumped optoelectronic mixers (OEMs) in this Thesis with those reported by other authors.

<table>
<thead>
<tr>
<th>Configuration</th>
<th>Result</th>
<th>Condition</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>2-T HBT OEM (1st sample)</td>
<td>( G_{\text{op}} = 7 \text{ dB (2.41 GHz to 500 MHz)} )</td>
<td>( V_{CE} = 0.6 \text{ V}, P_{\text{opt}} = 0.14 \text{ mW}, \text{LO = 10 dBm} )</td>
<td>Chapter 4</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -1 \text{ db (3.025 GHz to 100 MHz)} )</td>
<td>( V_{CE} = 0.6 \text{ V}, P_{\text{opt}} = 0.3 \text{ mW}, \text{LO = 10 dBm} )</td>
<td>Chapter 4</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -1 \text{ db (3.025 GHz to 5.95 GHz)} )</td>
<td>( V_{CE} = 0.8 \text{ V}, P_{\text{opt}} = 0.3 \text{ mW}, \text{LO = 10 dBm} )</td>
<td>Chapter 4</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -10 \text{ db (100 MHz to 20.1 GHz)} )</td>
<td>( V_{CE} = 1.5 \text{ V}, V_{BE} = 0.82 \text{ V}, P_{\text{opt}} = 70 \mu\text{W}, \text{LO = -4 dBm} )</td>
<td>Chapter 5</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -11 \text{ db (3.5 GHz to 6.9 GHz)} )</td>
<td>( V_{CE} = 1.5 \text{ V}, V_{BE} = 0.68 \text{ V}, P_{\text{opt}} = 70 \mu\text{W}, \text{LO = -3 dBm} )</td>
<td>Chapter 5</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -4 \text{ db (3.05 GHz to 100 MHz)} )</td>
<td>( V_{CE} = 1.5 \text{ V}, V_{BE} = 0.92 \text{ V}, P_{\text{opt}} = 0.3 \text{ mW}, \text{LO = -3 dBm} )</td>
<td>Chapter 5</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -4 \text{ db (3.05 GHz to 5.95 GHz)} )</td>
<td>( V_{CE} = 1.5 \text{ V}, V_{BE} = 0.75 \text{ V}, \text{LO = -4 dBm} )</td>
<td>Chapter 5</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -14 \text{ db (3.2 GHz to 30.8 MHz)} )</td>
<td>( V_{CE} = 4 \text{ V}, V_{BE} = 1.3 \text{ V}, P_{\text{opt}} = 1 \text{ mW}, I_{C} = 2.6 \text{ mA}, )</td>
<td>Suematsu et al. [4,5]</td>
</tr>
<tr>
<td>2-T HBT OEM (2nd sample)</td>
<td>( G_{\text{op}} = -16 \text{ db (200 MHz to 30.2 GHz)} )</td>
<td>( V_{CE} = 1.6 \text{ V}, V_{BE} = 0.65 \text{ V}, P_{\text{opt}} = 840 \mu\text{W}, I_{C} = 780 \mu\text{A}, \text{LO = -4.5 dBm at 30 GHz} )</td>
<td>Gonzalez et al [6,7]</td>
</tr>
</tbody>
</table>

The need to interconnect different components, such as the LO, the microwave mixer and the photodiode, is also avoided. The measured system conversion gains for the two electrically pumped HBT optoelectronic mixers further confirm their potentials in microwave over fibre systems. A system up-conversion gain of -10 dB (up-conversion from 100 MHz to 20.1 GHz) from the three-terminal HBT also indicates that the HBT can be a compact and high-performance optoelectronic up-converter for mm-wave fibre-radio links.

The three-terminal HBT offers greater flexibility than the two-terminal HBT because the availability of a base terminal means that the HBT base can be electrically pumped by the LO while the mixed products can be obtained from the collector separately. In the two-terminal HBT, only the emitter terminal is available and a diplexer is required to separate the LO and other output signals at different frequencies. If a change of the LO frequency is required, a new diplexer is then needed for the two-terminal HBT. However this is not necessary for the three-terminal device.
Pumping the three-terminal HBT at the base also requires a much lower LO power level compared to the two-terminal HBT which is pumped by the LO across the collector-emitter port.

As far as the noise performance is concerned, the experimental results have shown that such a single transistor approach to optoelectronic mixing favours the up-converting configuration over the down-converting configuration because the transistor generates a much higher noise output at low frequencies. Therefore when the photodetected RF signal is down-converted to a lower frequency, the signal will then appear in a very noisy frequency region. Since the transistor output noise level decreases with frequency, the up-converted signal will then appear in a less noisy frequency band. Therefore if both the down- and up-conversion gains are comparable, the signal-to-noise ratio will be higher for up-conversion than for down-conversion. The spurious-free dynamic range is similarly affected.

It must be stressed that the reason to use such a single HBT optoelectronic mixer is not to provide a far better performance, in terms of gain, noise, frequency response and dynamic range, over the discrete component approach to optoelectronic mixing, but rather to provide a much simpler alternative. In fact, using discrete components to accomplish optoelectronic mixing should result in higher performance because each component can be individually tailored for its function and separately biased to achieve optimum performance.

To improve the performance of the HBT optoelectronic mixer, it is important to understand better the mechanisms responsible for the mixing process. To do that, an accurate computer model for the HBT optoelectronic mixer is very useful. The harmonic-balance model in Chapter 3 provided the first step in modelling the two HBT mixers and there is still room for improving the accuracy of the computer model.

Due to limited time available for this work, the three-terminal HBT equivalent circuit used in the harmonic-balance model in Chapter 3 was kept as simple as possible in which the only nonlinear element was the base-emitter junction, and the base-collector junction was assumed reverse-biased at all times. As the experimental results showed in Chapter 5, the model became inaccurate when the three-terminal HBT was subjected to a large LO
excitation. By including a full large-signal base-collector junction equivalent circuit in the three-terminal HBT model, it is expected that accuracy of the simulation will improve at large LO powers and $V_{BE}$. Also in reality, the nonlinear elements in an HBT which contribute to the mixing mechanisms are not limited to the static IV characteristics of the two junctions. It is well known that the base-emitter and base-collector junction capacitances depend on the junction voltages which affect the depletion widths. Also the base-emitter diffusion capacitance, which was not taken into account in Chapter 2 in the consideration of the HBT's $f_T$ because it decreases with frequency, has a collector current dependence. When a large LO is applied to an HBT, both the junction voltages and the collector current will vary with the LO and these factors should be included in the model. If the HBT is to detect light modulated at very high frequencies (>10 GHz), it is also important to take into account the frequency response of the base-collector photodiode.

The modelling of the two-terminal HBT is far more difficult because of the lack of an electrical base terminal which would otherwise allow the equivalent circuit parameters to be measured and a more physical equivalent circuit representation constructed. Therefore more research into developing a two-terminal phototransistor large-signal model is important. However as a first step, one should be able to construct a simple two-terminal HBT model using the equations quoted or derived in Chapter 2 along with those given in a recent publication on HBTs [8].

Although the two HBT optoelectronic mixers have a gain advantage over using a separate photodiode and a microwave mixer, the achieved system conversion gain might still not be enough in practical systems because of the coupling loss. Rather than using a separate amplifier to increase the overall gain, it is suggested that the HBT optoelectronic mixer be integrated with a second amplifier stage monolithically. However, as was experienced in the three-terminal HBT noise measurement, the LO leakage signal from the HBT may saturate any directly connected amplifier when no filter is used. However it is difficult to provide any filtering on chip. Increased LO isolation can be obtained if, instead of using a single-HBT optoelectronic mixer, the HBT mixing stage is designed in a balanced configuration, such as a Gilbert cell. Figure 7-1 illustrates an HBT Gilbert cell optoelectronic mixer.
The constant current source connected to the base of transistor Q1 is used to ensure that when Q1 is illuminated, the average collector currents of Q1 and Q2 remain balanced. Being a balanced mixer, the LO leakage from the HBT mixing stage should be much smaller than the single-HBT configuration.

To improve the linearity of the HBT optoelectronic mixer, it would be interesting to investigate designing the HBT as a travelling-wave photodetector [9,10,11]. Since the absorption of light is distributed along the entire length of the HBT, the HBT should not be easily saturated and hence an improvement in linearity should be possible.

Finally it is of importance to deploy an HBT optoelectronic mixer in a practical fibre-radio system and evaluate the complete link performance. This is the only way to confirm and demonstrate the ultimate feasibility of the HBT optoelectronic mixer in practical situations.
References


Appendix

Publication List


