THE USE OF OPTICAL TECHNIQUES FOR SIGNAL DISTRIBUTION AND BEAM FORMATION IN PHASED ARRAY ANTENNAS

by

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The Use of Optical Techniques for Signal Distribution and Beam Formation in Phased Array Antennas.

Abstract

The thesis examines the feasibility of using optics for signal distribution and beamforming in phased array antennas. The requirements for signal distribution in phased array antennas are identified. A computer model is developed to evaluate the performance of analogue optoelectronic distribution systems. Using this model, various optical distribution techniques which can be used in phased arrays are examined, with particular reference to a satellite-based application. A model for semiconductor laser amplifiers with intensity modulated inputs is presented, and the optimum use of amplifiers for signal distribution is considered. Finally the feasibility of using optical techniques to perform other tasks such as beamforming is examined, and a novel opto-electronic beamformer is experimentally demonstrated. Results show that optics offer an excellent alternative to microwave techniques for signal distribution and beamforming.
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\[
\frac{\left[1/\tau_{sp}\right] \cos \omega m t + \omega m \sin \omega m t}{\omega m^2 + \left[1/\tau_{sp}\right]^2}
\]

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CHAPTER 1

INTRODUCTION: HISTORY OF PHASED ARRAYS - THE NEED FOR OPTICAL DISTRIBUTION TECHNIQUES IN PHASED ARRAY ANTENNAS - GUIDE TO THE THESIS

1.1. A brief history of radars and phased arrays - the need for phased arrays in ground systems, battleships, air fighters and satellites

1.1.1. A brief history of radars in ground systems and the need of phased arrays in this area

The general idea of radars started in 1923 when two American scientists, Breit and Tuve used the method of transmission of radio waves to measure the height of the ionized gas in the ionosphere. As far as military devices are concerned, the idea started in the United Kingdom in the late thirties when Watson Watt demonstrated the detection of aircrafts using radio signals from a shortwave radio transmitter. The next step was to use pulsed transmissions, and by early 1939 there was a fully operational chain of radar stations along the east coast of England.

These first radars operated at metric wavelengths - of the order of 3 meters. They usually used two antennas one for transmission and one for reception of the transmitted pulses (of the order of 50 per second). To display the echo signals a cathode ray tube was used with the so called type - A display which showed the range and the amplitude. The direction of the object could be found by the pointing of the antennas. Usually no indication of the height was given. The main purpose of these radars was to detect aircrafts or ships before they attacked, as they approached the surveillance area and to get the defence fighters up.

The next generation of radars came with the invention of the cavity magnetron. A brilliant achievement at the University of Birmingham had led to the possibility of radar at a wavelength of ten centimetres. This made possible very narrow beams, and a performance from small aerials never obtained before. The new radars were highly accurate and with better resolution. Now radars could be used not only to locate the presence of the
aircraft but also to guide the aircraft against the targets. In general the invention of the radar is recognized as the most significant invention of the war. Churchill himself in his World War II Memoirs refers to the "decisive role played by radar" and deals with the whole subject in surprising detail.

There were of course many important developments in radar since then. However the most recent and potentially more far reaching is the concept of phased arrays. A phased array (see Fig 1.1) consists of many individual radiating elements situated on the same surface in an array configuration. Phased arrays synthesize their resultant beam by the vector addition of the wavefronts of all the individually radiating elements. Suppose all the signals from the elements of an array are radiated with the same phase. Then their amplitudes will add up constructively at all distant points in space along the normal to the array plane, forming a lobe (beam) in the angular dimension. At points away from the normal to the array, the signals from different radiating elements will travel different distances to reach these points, so they interfere destructively weakening or even totally suppressing one another. The direction of the beam can be controlled electronically by introducing a phase slope between successive elements of the array. The wavefronts of the phase-shifted signals in this case add constructively along a direction at an angle to the normal of the array thus forming a deflected beam.

The idea of phased array antennas with fixed (nonsteered) beams dates back to World War I [1.1]. During World War II most parties used phased arrays with mechanically controlled phase shifters [1.1]. In the fifties the mechanically controlled phase shifters were replaced by electronic ones, mainly with the introduction of the Ferrite Phase Shifters [1.1], [1.2]. Finally in the sixties digital phase shifters made their appearance [1.1], [1.3].

As reference [1.4] states, at that time about 11 ground-based and 8 airborne phased array projects started in the United States. However few of them ever became operational, mainly large ballistic missile detection radars. A typical such configuration (PAVE PAWS) [1.5], [1.4] consists of two 72.5 ft diameter apertures each containing 1792 active T/R modules along with 835 dummy elements for impedance matching. Each such module radiates about 320 W of average power at a frequency near 435 MHz.
Fig (1.1) A typical signal distribution system in a phased array antenna.
Beamforming is achieved by 4 bit digitally controlled phase shifters. As ref [1.4] states mainly two air defence phased array radars became operational at that time, mainly the US Army’s PATRIOT radar and the US Navy’s AN/SPY 1.

The advantages of phased array radars for ground applications over conventionally rotating radars are mainly the following:

a) When the number of elements in the array is big a very large amount of power can be radiated. This makes them ideal to be used for long distance detection of objects such as ballistic missiles etc.

b) Their use removes the straightjacket of constant dwell time dependent on the mechanical speed of the antenna, constant data rate, and constant distribution of power in elevation and azimuth. The electronically controlled pencil beam can be steered to a different direction much faster - in microsecond time intervals - as compared to seconds for mechanical antennas. This allows flexibility of pointing direction and dwell time. It also gives control over spatial power distribution and, very importantly, the decoupling of the surveillance and tracking functions. Some of the benefits arising from this are superior surveillance and tracking performance and more effective clutter suppression.

c) The array can emit multiple beams which can act as independent radars, performing different functions. So one beam can be used for general surveillance with another one for tracking of a particular target. Conceptually a surveillance radar and several dedicated tracking radars can be replaced by a single phased array radar.

d) The reliability of phased arrays greatly increases in comparison with high voltage conventional radars. Failure of a single element of the array does not present any considerable problems in the correct operation of the rest of the system. However some care has to be taken if an element malfunctions to switch it off automatically, otherwise it may introduce unwanted phase shifts which might affect the alignment of the whole beam.
e) Dynamic radiation pattern control. The sidelobe levels and null locations can be controlled to adapt to changing scenarios expected during operation.

All the above arguments clearly show that phased arrays offer much superior performance compared to conventional mechanically rotating radars. However their main disadvantage is that they are too expensive and complicated to fabricate. Their complex microwave distribution networks together with thousands of elements, phase shifters, control computers etc have made their costs prohibitive. Another disadvantage of the first generation of phased arrays was their non-optimal implementation with the microwave technology of that time leading to high sidelobes, slow beam steering times, low data rates etc. This is the reason why, despite the initial research efforts and enthusiasm, there are so few operational phased arrays. However developments and changes in microwave, digital and optical technology during recent years have changed the situation. It is now possible to integrate a whole T/R element on a single GaAs substrate [1.6]. Fibre optics offer the possibility of low-weight distribution networks and the advances in digital technology offer cheap and sophisticated central processors to control the array. This situation together with increasingly demanding requirements on military radars (targets of diminishing reflectivity multifunction operations etc) has led to a renewed activity and interest in phased array radars.

1.1.2. A brief history of radars for naval applications and the need for phased arrays in this area

In this area the history follows approximately that of the ground system radars. Initially ships were only equipped with one radar to perform all the tasks. However as weapon systems became more and more complicated, each ship started being equipped with many radars, each one performing a different task. So in a typical warship today, there is a central radar for general surveillance while there are many dedicated radars for tracking of individual targets. So with the number of radars per ship continuously increasing, the need has emerged to replace all these single function radars by a multi-function one which can perform at the same time all the
different tasks. For this purpose a phased array is the only possible candidate.

1.1.3. A brief history of radars in aircraft and the need for phased arrays in this area

The first appearance of aircraft radars was during the second world war and their antennas consisted of some dipole-like elements operating at the VHF frequency range. These could be accommodated at the outside walls of the aircraft and they did not introduce any fundamental aerodynamic problem [1.4], [1.5]. However this situation changed with the introduction of the cavity magnetron and its microwave frequencies. This new type of radars introduced many advantages over the previous one in airplanes. It could provide the pilot with a map-like picture of the ground over which he was flying. Coastlines, rivers, towns and even large bridges could now be seen. However the narrower beamwidths of this type of radars require a mechanically steerable parabolic antenna. This presents several problems in the aerodynamic design of the aircraft as this type of antenna and its pointing mechanisms require a specially designed space at the inside of the aircraft. So a system is required which can be readily mounted on the walls of the aircraft so as not to affect its aerodynamics. Furthermore electronic stabilization of the beam can be used to compensate for motion of the aircraft. So taking into consideration all the above, the need for lightweight phased array radars for aircrafts becomes evident.

1.1.4. The need for phased arrays in satellite systems

In the last two decades the field of satellite communications has seen a major expansion. Communication satellites are used as relays to receive and re-transmit telephone, television and other signals between different parts of the world. However, as more satellites are launched in space and the information capacity per satellite is increasing, the available frequency bands become saturated. To overcome this problem one solution is to use pencil microwave beams like the ones produced by phased array antennas. In this way the available frequency channels can be re-used by more than one satellite, or in some cases by the same satellite. Also for satellite
systems which are not geostationary it is highly desirable to have an electronically steerable beam to compensate for the movements of the satellite relative to the earth station.

Finally apart from the area of telecommunications, phased arrays are used in imaging applications on board satellites to make map-like pictures of the earth’s or other planets surface. Such applications require very high resolutions (typically 30 m X 30 m spot sizes) which can only be achieved with the use of synthetic aperture radars (SAR’s).

1.2. The need for optical distribution techniques in phased arrays

Every element of a phased array antenna needs to be interfaced to a large number of transmit and receive signals. So one of the problems encountered in the design and construction of such antennas is the distribution and organization of these signals to the often thousands of elements of the array. The use of microwave signal distribution techniques starts becoming impractical, especially as arrays move to the shorter wavelengths. Waveguide distribution systems, although they present low losses and good phase stability, are extremely heavy, occupy high volume and are not flexible. Coaxial distribution systems are lighter and flexible, but present high RF losses especially at high frequencies, together with bad phase stability which may depend on factors such as mechanical movements, temperature variations etc. One solution which has received much attention over the last few years to tackle this problem involves the use of a fibre optic system [1.9]. A typical distribution system for the analog signals of a phased array antenna can be seen in Fig (1.2). Optical fibres present a number of advantages for distribution of a microwave signals over coaxial cable or metallic waveguides These are mainly the following:

a) Their volume and mass are low (see Table 1.1).

b) Their insertion loss is small (see Table 1.1) and it is independent of the frequency of the modulating signal.

c) The pass band of the optical signal is wide so it is possible to use
Fig (1.2) A typical optical distribution network for the analogue signals of a phased array antenna.
wavelength division multiplexing of the various signals to be transmitted between the elements and the antenna processor.

d) They have a lower thermal expansion coefficient (see Table 1.1), and can operate at a higher ambient temperature.

e) They are insensitive to external RF interference and strong electromagnetic pulses.

f) Conversely because of the very short wavelength of light they do not emit any potentially interfering radiation, thus not suffering from crosstalk between adjacent cables.

<table>
<thead>
<tr>
<th></th>
<th>COAXIAL CABLE S A L F L E X - 2 a t 5 GHz</th>
<th>OPTICAL FIBRE 8/125 µm SMF</th>
<th>UNITS</th>
<th>RATIO</th>
</tr>
</thead>
<tbody>
<tr>
<td>1) MASS</td>
<td>47</td>
<td>0.8</td>
<td>g/m</td>
<td>59</td>
</tr>
<tr>
<td>2) TEMPERATURE</td>
<td>1.7</td>
<td>0.3</td>
<td>d e g / ° C</td>
<td>5.7</td>
</tr>
<tr>
<td>STABILITY</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3) CROS S ECTION</td>
<td>19.6</td>
<td>0.6</td>
<td>mm²</td>
<td>33</td>
</tr>
<tr>
<td>(AREA)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4) INSERTION</td>
<td>0.6</td>
<td>0.003</td>
<td>d B / m</td>
<td>200</td>
</tr>
<tr>
<td>LOSS</td>
<td></td>
<td></td>
<td></td>
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</tr>
</tbody>
</table>

Table 1.1. Relative characteristics between optical fibre and a typical coaxial cable.

(After W.I. Mc Millan, Marconi Space Systems Ltd.)

In Table 1.1 the relative characteristics of optical fibre and coaxial cable are tabulated with respect to mass, temperature stability, volume and insertion loss.

The main problems faced when using optical fibres for distribution of the above signals to the elements of the array and vice versa are the following:
Chapter 1: Introduction

a) High electrical to optical and optical to electrical conversion losses. Typical losses for an optical link are of the order of 40 dB. However progress is being made by designing and constructing low loss links with losses of the order of 0 dB for directly modulated links and gains of 10 dB for externally modulated ones [1.10], [1.11], [1.12].

b) Multichannel branching of the optical paths. Using conventional optical technology, the splitting losses limit the number of elements which can be fed from a single optical source to about 25 [1.13] if S/N ratios of the order of 110 dBc Hz are to be obtained. However this problem may be solved with the use of more efficient optical links [1.10], [1.11], [1.12] and/or optical amplifiers.

c) Especially when the phased array is used on board satellite systems, one of the problems to be kept in mind during the design of the optical network is the electrical power consumption of the optical sources, R.F. amplifiers and temperature stabilization circuits.

1.3. History of optics for phased array antennas

The use of optics for signal distribution in phased array antennas was proposed in 1981 at University College London by Professor Forrest [1.14]. He suggested that as the cost of T/R modules reduces due to monolithic integration, there will come a time when the array cost will be dominated by the cost of distributing and combining the R.F., I.F. and control signals to and from the elements of the array.

In one of his early papers on the subject, [1.15] Forrest et al. carried out design calculations and showed that for a 10 mW laser and a 20-way split, the power out of the photodiode was only 10 nW, or 50 dB less than the required 1 mW at the input to the T/R module. However as time passed, better optical components have emerged from the laboratories and these figures have been considerably improved. Optical links with gains have been demonstrated [1.16] and the use of optical amplifiers [1.17], [1.18], [1.19] has made possible the construction of optical distribution networks which do not suffer from branching losses.

In general the history of optics in phased arrays follows closely the
availability of optical components which were developed primarily for the telecommunications industry. Single mode optical fibres, high speed photodetectors, fast narrow linewidth lasers, optical switches and laser amplifiers are more or less transplanting technology from the telecommunications industry to the needs and specifications of microwave phased array antennas. However there are some exceptions from this rule:

In the area of optically controlled microwave oscillators a lot of pioneering work has been done by Forrest et al. in references such as [1.20]. Also the group from Drexel University [1.21], [1.22], [1.23] has shown that it is possible to lock an oscillator to a higher order harmonic of a laser diode thus substantially increasing the upper usable frequency limit for directly modulated semiconductor lasers.

In the area of optical to electronic signal conversion Gomes et al. [1.24] have demonstrated the use of opto-electronic mixers for optical signal recovery, which promises significant reductions in the circuit complexity of optically controlled T/R modules.

Finally a lot of pioneering work aimed at phased array antennas has been done in the area of optical beamforming techniques. References such as [1.25] and [1.26] use the coherent nature of light to demonstrate microwave phase shifters, and from this principle a whole range of coherent optical processors have emerged. The most important of them are those of Koepef [1.27], [1.28] which uses a miniature optical replica of the antenna pattern to construct the microwave beam, that of Dolfi [1.29] which combines a coherent and a switched true-time-delay optical processor and that of Birkmayer [1.30] which uses integrated optics to make a full-scale coherent processor.

1.4. Conclusions and guide to the thesis

The discussion in this chapter has shown that optically controlled phased array antennas will play a key role in future military and civil radar and communications systems. However several problems have yet to be solved in the optical distribution networks such as the high electro-optic and opto-electronic conversion losses, splitting losses etc. This thesis intends to tackle some of these problems and propose the best possible
solutions using the present status of technology. Also it will go beyond that and attempt to tackle the problem of performing other operations such as the beamforming in the optical domain.

In particular the thesis is organized as follows:

The second chapter presents the specifications and requirements of typical phased array antennas, and identifies the signals which have the potential to be distributed optically. The specifications and requirements are given for every signal in terms of S/N ratio, dynamic range, phase accuracy and nonlinear distortions. Finally the beamformer requirements are reviewed, mainly with respect to the need for true time delay beamforming as compared with frequency independent phase steering.

The next five chapters (3 -7) analyze and compare the various possible architectures in which signals can be distributed to and from the elements of the array.

In particular the third chapter presents the equations and design parameters of a computer program which is used to model analogue optical and optoelectronic links.

The fourth chapter analyzes and compares the architectures of various types of passive optical links - that is without the use of optical amplifiers.

The next three chapters (5 -7) investigate the possibility of using optical amplifiers in the distribution networks of phased arrays.

In particular the fifth chapter outlines and compares the general characteristics of the two main types of optical amplifiers (semiconductor and fibre) and identifies which signals in a phased array may benefit from the use of optical amplifiers and suggests some possible configurations.

The sixth chapter presents a computer model which evaluates the performance of semiconductor laser amplifiers under intensity modulated optical inputs. Results are presented, with particular reference to the problem of nonlinearities in the response of the amplifiers.

30
Finally the seventh chapter examines the use of optical amplifiers with reference to a particular system. This system is the forward link of a satellite-based array. Design curves predicting the performance of the link by varying its various parameters are plotted, and using these curves three possible networks structures are presented and their performances compared.

The next two chapters (8 and 9) examine whether other tasks such as the beamforming can be performed in the optical domain.

The eighth chapter presents a review of the various beamforming techniques which have been reported in the literature. Each technique is assessed and its characteristics compared with those of other solutions.

The ninth chapter presents the design and experimental construction of a novel optoelectronic beamformer. The theory behind the beamformer is analyzed, and the experimental results are presented with respect to phase accuracy, signal budget and noise measurements. The experimental performance of the beamformer is compared to the modelled one, and possible ways in which the performance can be improved are identified.

Finally the tenth chapter presents some conclusions including predictions about the use of optics in future systems.

1.4.1. Novel contributions of this thesis

It is normally expected that a PhD thesis should contain some novel contributions. In this thesis, the main novel contributions as compared to the existing literature are the following:

a) The examination in chapters 5 - 7 of the use of optical amplifiers for signal distribution in optically controlled phased array antennas. Although in the literature the use of optical amplifiers has been extensively examined for digital telecommunications links and cable television distribution systems, a recent literature survey has shown that there is no paper which examines the use of laser amplifiers in optically controlled
phased array antennas. The purpose of the above chapters is to cover this gap. The results show that the use of optical amplifiers can offer reduced electrical power consumption, improved S/N ratio and reduced circuit complexity.

b) The novel optoelectronic beamformer for which the design, constructional details and results are given in chapter 9. The results show a highly linear response and its structure offers the potential for integration with optical distribution networks thus offering a very attractive proposition in the area of opto-electronic beamformers.

1.5. References for Chapter 1


[1.25] Bone M.C., Jackson J.D. and Wolfson R.I. "Electro-optic phase
control of R.F. array antennas" Milicom '82 IEEE Military communications conference pp. 188-193 Session 5A paper No 5.


CHAPTER 2
TECHNICAL SPECIFICATIONS AND REQUIREMENTS OF OPTICALLY CONTROLLED PHASED ARRAYS

2.1. Introduction

Before discussing the structures and architectures of the optical distribution networks in phased array antennas it is vital to know the exact technical specifications and requirements of each type of antenna. This chapter will attempt to identify these specifications for each type of phased array antennas based on a review of the relevant literature. The chapter is organized in two sections:
First there is a discussion of the specifications of the signals which need to be distributed to and from the elements of the array. Then there is a discussion of the beamformer requirements with particular attention to the need for true-time delays or pseudo-phase shifting. Finally in APPENDIX A the relevant equations for each type of beamforming are derived and possible solutions are proposed.

2.2. Identification and specifications of the signals which need to be distributed to each element of the array

In a typical phased array the signals which are candidates for optical distribution are the following [2.1], [2.2], [2.3], [2.4], [2.5], [2.6], [2.7], [2.8]:

1) For the transmit function of the array:
   a) The R.F. reference signal
   b) The I.F. transmit signal
   c) The element control signals

2) For the receive function of the array.
   a) The heterodyne local oscillator signal
   b) The received I.F. signal
   c) The element monitoring signals
2.2.1. Transmit function

For the transmit signals of the array two architectures are possible (see Fig 2.1), generally characterized by the point at which the communications data (or frequency chirp) and the RF carrier are combined. In CPU-level designs, the communications or other data is impressed on the RF carrier before conversion to light. This type of architecture is the simplest conceptually, but is the most demanding on the optical requirements of the link such as bandwidth, dynamic range and noise figure. A more realizable architecture at present is that of combining the communications data with the RF carrier at the elements of the array. The greatly reduced high frequency RF bandwidths required, lead to relaxed system specifications but components for mixing RF and data must be incorporated in the array element design. In such a configuration - which will be the main one considered in this section - the signals candidates for optical distribution are the following:

2.2.1A. The R.F. reference signal

This is the highest frequency signal in the system. Its frequency is usually close to the radiated frequency from the array. In typical systems this frequency may lie anywhere between UHF (for example in ground based ballistic missile radars) and 100 GHz (such as for future satellite systems [2.6], [2.9]). This signal consists of a single frequency, although in military frequency agile radars [2.1] this frequency can appear anywhere within the system bandwidth.

In some designs the photodetected RF signal is amplified and fed directly to the mixer with the data signal. Another common approach is to use the detected RF signal to lock an oscillator in frequency and phase. This locking can be either at the fundamental frequency of the received signal [2.10] or at a sub-harmonic [2.11], [2.12], [2.13].

The distribution of the RF signal requires a tree-type network, with one microwave reference source modulating an optical carrier. This optical carrier is distributed to the elements of the array through equal length optical branches, so that phase coherence is maintained across the array.
Fig (2.1) Differences between a) CPU-level transmit architecture in which the I.F. signal is combined with the R.F. signal at the central processor of the array and b) element combining transmit architecture in which the I.F. and R.F. signals are distributed separately and combined at the array elements.

Fig (2.2) Differences between a) CPU-level receive architecture in which the received signal is combined with the L.O. at the central processor and b) element combining receive architecture in which the mixing of the received signal and the L.O. takes place at the elements of the array.
Chapter 2: Technical specifications and requirements

If phase compensating devices are not included in the elements of the array, the fibres have to be cut very accurately. Typical specifications in the literature vary between 1° [2.8] and 6.5° [2.1] (this corresponds for example to fibre lengths of 0.05mm and 0.33mm at 10 GHz). These are of course very difficult to achieve specifications so in practice there are usually phase compensating devices in the R/T modules. In this way the lengths of the fibres do not have to be equal and final adjustment of phase can take place after the array has been assembled.

The small instantaneous bandwidth means that distortions caused by nonlinearities in the distribution network are not important thus making it desirable to use high modulation depths on the optical carrier. However the presence of noise on the transmitter reference is important in that it is transferred, via the received signal to the signal processing unit. Thus regarding the signal to noise ratio of this signal, for Doppler radars an important feature of the radar system is its ability to distinguish small returns from moving targets in the presence of much larger returns from nearby obstacles (clutter). The difference between these two signals (apart from level) is the Doppler frequency shift present on the moving target. The transmitter noise which will appear on all target returns extends over a finite spectrum and hence noise returned from stationary clutter can mask a Doppler shifted target return unless the noise is of a sufficiently low level. According to ref [2.1] for a typical system with a peak clutter to signal ratio of 80 dB, the transmitter noise, 2 KHz from the carrier, must be reduced to at least 80 dB below the carrier in a 200 Hz bandwidth to allow accurate Doppler measurements to be performed on a 100 m/sec target. Thus the transmitter AM noise spectrum can be specified at -103 dBc/Hz at 2 KHz from the carrier. As the LO noise will have an identical effect, it would be more realistic to set both the transmitter and LO noise performance at -106 dBc/Hz maximum. Other authors place even more severe requirements on the S/N ratio. For example ref [2.8] argues that in modern radar systems a S/N ratio of 130 dBC Hz is required at 10 KHz offset from the carrier.

In other communications phased arrays not used as Doppler radars, for example in satellite based systems, the noise requirements are determined by other factors, such as the received S/N ratio and detailed calculations need to be performed for each particular system.
2.2.1B. The I.F. transmit signal

This signal needs only to be distributed in phased arrays where mixing of the RF and IF signals is done at the array element level. In some Doppler systems only an RF pulse needs to be transmitted. So the only information required from the IF signal is a pulse which can be time-multiplexed with other digital signals in the system. In other phased array systems such as those used in telecommunications, the IF signal will carry all the communications data. In contrast with the RF signal, where linearity of the distribution network is not important, here any nonlinearity will produce undesired harmonics and intermodulation products [2.14], [2.15]. Some of the third order intermodulation products will most probably be within the signal bandwidth [2.14] while the harmonics may limit the usable bandwidth of the signal and prevent frequency multiplexing of other signals into the channel [2.16]. However, in contrast with the received IF signal, the dynamic range here is small. So there should not be great problems in its distribution.

2.2.1C. Element control signals

A number of control signals need to be distributed to the array elements. This is the least demanding requirement in terms of device technology and can usually be satisfied with current digital optics technology. A typical configuration consists of a central processor from which all data is determined and transmitted. Such addressing scenarios usually require high serial bit rates to address a large number of elements. An alternative is to provide every element or sub-array with a separate processor. In this way the system intelligence is distributed, and the serial data bit rates significantly reduced.

The element control signals are usually the following:

i) Digital phase shifter settings
ii) Transmit/receive switch settings.
iii) RF polarization selection (horizontal or vertical).
iv) RF amplitude control.
v) Built in test.
vi) Timing reference.

In general these control signals do not require more than 20 bits for each element of the array per update time [2.1]. The update time is governed by the slew rate of the beam and according to [2.1] for a military radar this should not be less than 10 μsec. So in an array of say 1000 elements the data rates from the central processor should be 1 Gbit/sec. This is a relatively high bit rate, but implies a tree type distribution where all the elements receive data from the same centrally controlled serial bus. In a practical system the distribution will be probably divided into sub-arrays where every sub-array gets the data from a central processor in its own serial bus. Ref [2.1] also suggests that in certain systems it is possible to distribute control signals via the transmitter/LO network between periods of transmission and reception. Finally in systems where this is not possible, it may be possible to use frequency division multiplexing or wavelength division multiplexing to the transmitter/LO network.

2.2.2. Receive function

Again for the receive network of the array two architectures are possible characterized by the point at which the incoming signal is mixed with the L.O. signal (see Fig 2.2). The first, and most straight-forward conceptually, is to feed the incoming R.F. signals straight to the optical links, combine them together and mix them with the L.O. signal at the central processor. However this architecture is the most demanding - if not totally unrealizable - on the requirements of the optical link such as dynamic range, noise figure and bandwidth. A more realizable architecture (see Fig 2.2) is that of heterodyning the L.O. with the incoming signal at the array element level. This is the approach which will be mainly discussed here. The signals required for such an architecture to be distributed optically are:

2.2.2A. The heterodyne local oscillator signal
Chapter 2: Technical specifications and requirements

This is an RF signal of approximately the same frequency as the transmitter reference signal. It is used during the receive operation of the array and is heterodyned with the incoming signal. Their frequency difference produces the received IF signal.

The requirements of the optical network to distribute this signal to the elements are practically the same as those for the RF transmitter reference signal. In Doppler radars, and other systems where the transmission and reception of signals do not occur at the same time, time multiplexing can be used to combine the two signals on one optical carrier. If the designer has chosen to use two mixing stages at the array elements, then a second L.O. is required but of much lower frequency. This second L.O. can be frequency or wavelength division multiplexed with the first one.

2.2.2B. The received I.F. signal

This is probably the most challenging signal to be distributed optically. In Doppler radar systems, the return signal consists of many signals reflected from various moving and non-moving objects. Thus as the signals of interest may be only a small proportion in power compared to the other returning signals (clutter), a wide dynamic range is required from the receiver network. In [2.1] the dynamic range for the return IF signal is specified as 50 - 90 dB while in [2.8] it is specified as at least 70 dB over the instantaneous bandwidth.

The instantaneous bandwidth required mainly depends on the range resolution required by the radar.

Typical bandwidths encountered in a real signal are of the order of 5 MHz to 20 MHz [2.1]. So a dynamic range of 70 dB in a 10 MHz bandwidth is equivalent to a spurious-free signal to noise ratio of 140 dBc Hz. The nonlinear performance of the optical network is also important. Nonlinearities will distort the signal, producing undesired spectral components in the frequency domain which correspond to spurious targets in the time domain. These echoes will limit the Doppler and range accuracy, and they must hence be suppressed as much as possible.

In communications systems the problem of intermodulation performance is usually of a different nature, arising from strong signals from other systems within the receiver passband. In this case the specifications
depend on the particular system under examination.
The noise figure of the optical link must also be reduced to the lowest possible value. Typical degradations of the S/N ratio must be less than 0.5 dB [2.1]. Some articles [2.1], [2.2] argue that all of the array element signals must be available separately at the input to the signal processing unit, which means that every array element must have its own dedicated optical link. However if intensity modulation is used in the return links, the links can be combined together in an inverse tree structure in the optical domain. Then the output of this structure will be the required IF signal from the array. Amplitude and phase weighting of the individual IF signals can be performed at the array element level. More discussion about such possible architectures will be presented in chapter 4.
Finally [2.1] suggests 3 types of modulation which can be used to modulate the IF signal on the optical carrier; intensity modulation, digital modulation and frequency modulation of a microwave carrier. A comparison of the three techniques will be presented in chapter 4.

2.2.2C. Element monitoring signals

In addition to the I.F. signal which needs to be distributed from the array elements to the central processor there are also some digital signals which are needed to perform functions such as monitoring the status of the element and to check for element malfunctions so that it can be switched off. Also in some Doppler or telemetry systems there may be a need to send an indication to the central processor of the strength of the incoming signal when it reaches the array elements. The specifications for these links are more or less the same as those for the element control signals. The main differences are that the data rates do not usually need to be so high because they do not need to steer the beam in microsecond time intervals.
Thus these signals can usually be wavelength or time multiplexed with the return I.F. signals.

2.3. Beamformer requirements

If the bandwidth of the radiated signal from the array and the array size
are small then conventional phase shifters (i.e. phase shifters which introduce the same phase shift on the signal over the frequency range of interest) are acceptable. However when the bandwidth of the radiated signal and the antenna size are large then a phenomenon known to antenna designers as "squint" arises. This is illustrated in Fig (2.3). For a single frequency $f_o$, the beam can be steered to an angle $\phi$ by an introduction to the $n_{th}$ element of a phase shift $\theta$ given by:

$$\theta = \frac{2 \pi n d \sin \phi}{\lambda_o} \quad (2.1)$$

where $d$ is the distance between two consecutive elements, and $\lambda_o$ is the wavelength at $f_o$. For a conventional phase shifter though [2.17], [2.18] the phase versus frequency characteristic is constant. Thus at lower frequencies with longer wavelengths the beam angle will be larger to satisfy (2.1) and vice versa. So, as can be seen in Fig (2.3), the power of the beam in the desired direction has "dispersive" characteristics with frequency and the bandwidth of the antenna is limited. The problem manifests itself more as the array size, bandwidth and scan angle increase. Translating the same phenomenon from the frequency into the time domain, suppose there is an array like that of Fig (2.4) and it is desired to send a pulse modulated on an R.F. carrier. It can be seen from Fig (2.4) that in order for the pulse from element B to be on the same plane wavefront together with the pulse emitted from element A it needs to experience a time-delay of $c \Delta l \sin \phi$ where $\Delta l$ is the distance between elements A and B, $\phi$ is the angle of the beam and $c$ is the speed of light. When the radiation angle of the array changes to $-\phi$, the pulse from element A needs to experience a time delay $c \Delta l \sin \phi$ and so on. Thus in order to scan the beam at different angles, variable true time delays are needed in every element or sub-array of the phased array antenna.

In APPENDIX A an analysis is presented of the equations for conventional phase shifting and true-time delays. It is shown that true time delays may take place at the I.F. (probably at a sub-array level) while only conventional phase shifting is needed for every element on the R.F. single frequency signal before it is mixed with the I.F.. This has the following advantages in comparison with R.F. true time delay beamforming:
Fig (2.3) The need for true-time-delays in phased arrays (frequency domain)

Fig (2.4) The need for true-time-delays in phased arrays (time domain).
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i) It is easier and cheaper to build delay devices at lower frequencies with less loss and dispersion. For example surface acoustic wave (SAW) devices can operate at frequencies between 500 MHz and 5 GHz [2.19] which is usually the frequency range of the I.F. signal of a phased array.

ii) One degree of phase shift at R.F. is equivalent to one degree of phase shift at the I.F.. At I.F. it takes a much larger delay error to produce this degree than it takes at R.F.

iii) If the radiation frequency of the array is variable the delay devices need only operate over a bandwidth equal to the system’s instantaneous bandwidth and not to the R.F. tuning range.

iv) True time delays need only be introduced at sub-array level while only conventional phase shifting can be introduced at each element. This greatly reduces the number of the more expensive true-time delay phase shifters required.

2.4. Conclusions

In this chapter the requirements of the various signals which need to be distributed to and from the elements of optically controlled phased array antennas have been specified. All of them can be transmitted optically with the currently available technology, except for the IF return signal of Doppler radars which presents certain problems due to its high dynamic range. However it is expected that these difficulties can be overcome with advances in the optical technology or the use of alternative optical transmission techniques such as frequency or digital modulation.

Finally in this chapter and in Appendix A the basic beamformer requirements of phased arrays have been specified and a true time delay configuration has been proposed in which the true time delay is introduced at the IF level while only conventional phase shifting is introduced at the RF level.

2.5. References for Chapter 2

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CHAPTER 3

COMPUTER PROGRAM TO MODEL NOISE AND POWER BUDGETS ON OPTICAL LINKS

3.1. Introduction

Having established the specifications of phased array antennas, the next step is to decide whether these specifications can be satisfied using optical technology, determine the best architectures and make detailed calculations of the system budgets. To make these calculations, a computer model is required to analyze the performance of optical networks carrying RF and microwave signals. In this Chapter the equations and the design approach for this model are described in detail.

3.2. Design approach to the problem

To determine the power and noise performance of optical networks other designers such as Way [3.1], Forrest [3.2], and Bakhrakh [3.3] have combined the equations for the individual components of an optical link into a single equation. Then they have used this equation to calculate signal-to-noise (S/N) ratio, output power and other parameters. This approach, although the simplest in concept has the disadvantage that it does not give flexibility to the user. Also it is difficult to construct a single equation for a complex system with many inter-dependent parameters. Furthermore the task becomes even more complicated when the number of components in the link increase, for example to model the combination of an optical link with a following microwave link.

In this model a more distributed approach has been chosen. Within this approach every microwave or optical component is modelled separately, as a block. Then its output parameters are passed to the next component which processes them according to its describing equations and so on. In this way, by the combination of many such blocks, a complex system can be modelled easily. Furthermore the user has the flexibility to construct his own configuration by combining many such blocks together as in a real system. For example to model a simple optical link the user has to combine
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A microwave source, a laser, an optical attenuator and a photodiode. In each of these blocks, parameters such as the relative intensity noise (RIN) of the laser or the quantum efficiency of the photodiode can be specified by the user. In addition there is a routine in the system which automatically plots the effects of varying any of the parameters in the blocks against any of the output results. For example the output S/N ratio can be plotted as a function of the laser RIN, the optical attenuation, the photodiode efficiency or any other parameter of the link.

In the next part of the Chapter the equations used for every block will be described in detail. Also there will be a description of the user-interfaces to input and output data to and from the model, together with a description of the interfaces between components-blocks of the system.

3.3. Interfaces between blocks in the R.F. - microwave domain

In the RF - microwave domain the following variables need to be transferred from one component to the next (for example from a photodiode to an RF amplifier):

a) The frequency of the RF signal \( f_c \) (in Hz).
b) The RF power of the signal \( P_s \) (in Watts).
c) The noise power \( P_{nse} \) in a normalized bandwidth (in Watts/Hz).
d) The frequency offset \( f_{off} \) where the noise power is measured from the carrier (in Hz).
e) The input & output impedance of the component \( R_{in} \) and \( R_{out} \) (in \( \Omega \)). For the time being real impedances are assumed. In the future if necessary a component can be constructed to deal with complex impedances.

3.4. Interfaces between blocks in the optical domain

In the optical domain the following variables need to be transferred from one component to the next (for example from a laser to a photodiode):

a) The average optical power of the optical signal \( P_{oavg} \) (in Watts)
b) The frequency of the modulated signal on the optical carrier. (in Hz)
c) The rms value of the modulated component of the optical signal $P_{\text{oms}}$. For sinusoidally modulated optical signals the relation between $P_{\text{oms}}$ and the peak value of the modulated component of the optical signal $P_{\text{om}}$ is:

$$P_{\text{oms}} = \frac{P_{\text{om}}}{\sqrt{2}}$$

(3.1)

d) The noise equivalent optical power $P_{\text{opase}}$. In Sze [3.4] this is defined as the noise of an optical signal with an rms optical power $P_{\text{oms}}$ which when photodetected with a noiseless photodiode (no shot or thermal noise) gives a (S/N) = 1 in a 1 Hz bandwidth. Its units are Watts/Hz.

3.5. Equations used to calculate the transfer characteristics of a semiconductor laser

In a typical laser diode, for values of current $I$ larger than the threshold current $I_{\text{th}}$, and operating in the linear regime the output optical power $P_{\text{opav}}$ is:

$$P_{\text{opav}} = a (I - I_{\text{th}})$$

(3.2)

where $a$ is the electro-optic conversion constant of the laser and has units W/A. This equation is used in the model to calculate the average optical power out of the laser in terms of the bias current and the threshold current.

Laser diodes usually have internal impedances of the order of 3-5 $\Omega$ say $R_{B}$. There are mainly two ways to match this impedance to the 50 $\Omega$ transmission line:

a) Through the use of a 47 $\Omega$ resistance in series with them, say $R_{A}$. In this case the electrical power in the circuit is:

$$P_{\text{el}} = I^2 (R_{A} + R_{B})$$

(3.3)

Combining equations (3.2) and (3.3) the optical power out of the laser can
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be obtained in terms of the electrical power:

\[ P_{\text{opav}} = a \left[ \frac{P_{\text{e}1}}{(R_A + R_B)} - I_{\text{th}} \right] \quad (3.4) \]

For small depths of sinusoidal modulation and for modulation frequencies below the photon-electron resonance the current equation (3.4) becomes:

\[ P_{\text{oms}} = a \left[ \frac{P_s}{(R_A + R_B)} \right] \quad (3.5) \]

This equation is used to calculate the rms value of the modulated component of the optical signal from the R.F. electrical power into the laser.

b) The second way to match the laser to the 50 Ω transmission line is to use a reactive method. Various lumped element and transmission line techniques have been developed [3.5]. These are usually valid for narrow frequency bands. At low R.F. frequencies (up to about 1 GHz) a common method of reactive matching which is valid over relative wide bandwidths is to use R.F. transformers. At the frequencies where the laser is matched to the transmission line, the rms value of the modulated component of the optical signal is:

\[ P_{\text{oms}} = a \left[ \frac{\eta P_s}{R_B} \right] \quad (3.6) \]

where \( \eta \) is the R.F. power efficiency of the matching circuit (for example the transformer).

The equivalent optical noise power of the signal has its origins in two factors:
i) Noise carried by the electrical R.F. signal which is modulated on the optical carrier together with the signal. This can be entered into the model using equation (3.5):
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\[ P_{\text{op}} = a \left[ \frac{P_{\text{nse}}}{(R_A + R_B)} \right] \]  \hspace{1cm} (3.7)

ii) Noise produced by the laser itself. This has optical power of:

\[ P_{\text{RIN}} = \sqrt{\text{RIN} \cdot P_{\text{opav}}^2} \]  \hspace{1cm} (3.8)

where RIN is the relative intensity noise of the laser.

Assuming that the two noise sources are statistically independent (uncorrelated) the total noise equivalent optical power out of the laser \( P_{\text{nse}} \) due to the incoming electrical noise and the RIN is:

\[ P_{\text{opnse}} = \sqrt{P_{\text{op}}^2 + P_{\text{RIN}}^2} \]  \hspace{1cm} (3.9)

These are the equations used to calculate the transfer characteristics of a laser diode.

Typical component parameters for a semiconductor laser can be found in APPENDIX B.

3.6. Equations used to calculate the characteristics of an external modulator

A discussion about the characteristics of links using electro-optic modulators is presented in Section 4.2.2.

The model for the electro-optic modulator has necessarily two input interfaces. The first is optical, to interface the modulator to the input laser and the second is electrical to interface the modulator to the R.F. source.

The rms value of the modulated component of the optical signal out of the modulator is:

\[ P_{\text{oms(out)}} = \frac{\Gamma \cdot P_{\text{opav(in)}} \cdot a_{tt} \cdot \sqrt{P_{\text{R}} \cdot R_M}}{V_\pi} \]  \hspace{1cm} (3.10)
where \( V_\pi \) is the switch voltage of the modulator, \( a_\pi \) the optical attenuation coefficient, \( R_M \) the impedance of the modulator and \( \Gamma \) a factor given by:

\[
\Gamma = \frac{2 J_1 \left( \frac{m \pi}{2} \right)}{m} \tag{3.11}
\]

where \( m \) is the optical power modulation depth and \( J_x(y) \) is the Bessel function of \( y \) of order \( x \). For small modulation depths \( \Gamma \) becomes equal to \( \pi/2 \), while for large modulation depths \( \Gamma \) approaches unity. In the model the above approximations for small and large signal depths are used.

For large signal modulation depths, the nonlinear response of the modulator produces harmonics and intermodulation products. The factor \( \Gamma \) for the \( 2k-1 \) order harmonic \((k = 1, 2, 3, \ldots n)\) is given by:

\[
\Gamma_{2k-1} = \frac{2 J_{2k-1} \left( \frac{m \pi}{2} \right)}{m} \tag{3.12}
\]

This factor \( \Gamma_{2k-1} \) can be substituted into equation (3.10) to give the optical power contained in the \( 2k-1 \) order harmonic.

The magnitudes of the intermodulation products of an electro-optic modulator are calculated in reference [3.6]. In particular the ratio of the third order intermodulation product to the main carrier (when the signal is in the optical domain - optical dB) is given by:

\[
r = \frac{J_2 \left( \frac{m \pi}{2} \right)}{J_0 \left( \frac{m \pi}{2} \right)} \tag{3.13}
\]

where \( m \) is the modulation depth of each of the two closely spaced carriers used to calculate the third order intermodulation products.
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It should be noted that third order intermodulation products do not appear when the optical link is used to transmit a single carrier, for example in the calculations of Table (4.2).

Since the model does not calculate nonlinearities, it does not include the above equations for harmonics and third order intermodulation products. However these formulas can be used separately when it is believed that such nonlinearities may degrade the performance of the system.

The noise equivalent optical power out of the modulator has two sources:
*a) The incoming electrical noise from the R.F. source \( P_{nse(in)} \)
*b) The incoming noise on the input optical signal due to the RIN of the laser with noise equivalent optical power of \( P_{opnse(in)} \).

The optical noise component due to the incoming electrical noise has noise equivalent optical power of:

\[
P_{op} = \frac{\pi P_{opnse(in)} a_{ut} \sqrt{P_{nse(in)}} R_M}{2 V_\pi}
\]  

(3.14)

The optical noise component due to the RIN noise has noise equivalent optical power of:

\[
P_{op} = \frac{P_{opnse(in)} a_{ut}}{2}
\]  

(3.15)

Assuming again that the two noise sources are uncorrelated, the total noise equivalent optical power optical power out of the modulator is:

\[
P_{opnse(out)} = \sqrt{P_{op}^2 + P_{op}^2}
\]  

(3.16)

Typical component parameters for an external modulator can be found in APPENDIX B.

3.7. Equations used to calculate the transfer characteristics of a
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A semiconductor laser amplifier and a fibre amplifier.

Although a full discussion of the equations to model a laser amplifier will be presented in Chapter 6 an approximate analysis is used here for the purpose of integrating the model with other components and calculating parameters such as the signal to noise ratio for the whole optical link. Although these equations have been derived for semiconductor laser amplifiers, the principles from which they have been obtained also apply to fibre amplifiers. Thus only small modifications are needed to apply them to fibre amplifiers. These are indicated in the text.

As can be seen from Fig (7.2) the gain $G$ of semiconductor laser amplifiers versus input optical power is constant for low levels of input optical signal but decreases as the signal level becomes higher. The same applies to fibre amplifiers. Also the gain depends on the input current (or pump power) to the amplifier. To account for this behaviour in the model, the gain versus input optical power characteristics for a certain current (or pump power) have to be entered as parameters. In the model the user enters some points of the gain versus input optical power curve. Then the computer draws straight lines between these points and can calculate the gain $G$ for every input optical power to the amplifier. In semiconductor amplifiers the gain $G$ is the gain from facet to facet, not including coupling losses to and from the input and output optical fibres. Another important parameter in semiconductor amplifiers is the internal single pass gain of the amplifier $G_s$. To calculate it equation (4) of Mukai [3.7] is used which gives the gain as a function of the single pass gain:

$$G = \frac{(1 - R_1)(1 - R_2)G_s}{(1 - G_s \sqrt{\frac{R_1 R_2}{R_1 + R_2}})^2} \quad (3.17)$$

However, here the single pass gain as a function of the total gain is required:

$$G_s = \frac{1}{\sqrt{\frac{R_1 R_2}{R_1 + R_2}}} + \frac{(1 - R_1)(1 - R_2)}{G R_1 R_2}$$
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\[
\sqrt{\left[ \frac{1}{R_1 R_2} + \frac{(1 - R_1)(1 - R_2)}{G R_1 R_2} \right]^2 - \frac{1}{R_1 R_2}} \quad (3.18)
\]

Also in semiconductor amplifiers the coupling losses from the input fibre to the amplifier and from the amplifier to the output fibre are entered as parameters and the losses in optical power in and out of the amplifier are calculated accordingly.

In fibre amplifiers, since the reflectivities of the splices are approximately zero, the single pass gain \( G_s \) of the amplifier is approximately the same as the gain \( G \).

To calculate the noise produced by the amplifier [3.7] and [3.8] give the variance of the photon number out of the amplifier:

\[
\sigma_{\text{out}}^2 = G \langle n_{\text{in}} \rangle + (G - 1) n_{sp} B + 2 G (G - 1) n_{sp} \langle n_{\text{in}} \rangle \xi +
\]

\[
+ (G - 1)^2 n_{sp}^2 B \quad (3.19)
\]

where \( \langle n_{\text{in}} \rangle \) is the mean value of the photon number per second incident on the amplifier, \( n_{sp} \) is the population inversion parameter of the amplifying medium, \( \xi \) is the excess noise coefficient for the beat noise between signal and spontaneous emission and \( B \) is the optical bandwidth of the amplifier. The four terms on the right side of equation (3.19) represent amplified signal shot noise, spontaneous emission shot noise, beat noise between signal and spontaneous emission and beat noise between spontaneous emission components.

In equation (3.19) \( n_{sp} \) is the population inversion parameter given by:

\[
n_{sp} = \frac{n}{n - n_{tr}} \quad (3.20)
\]

where \( n \) is the average dynamic carrier density in the amplifier and \( n_{tr} \) is the transparency carrier density.

It can be proved that the relation between the average carrier density and the single pass gain to the amplifier \( G_s \) is:

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\[
\ln \left( \frac{G_s}{L} \right) + \alpha_{sc} + n_{tr} = \frac{n_{out}}{\Gamma / B} \quad (3.21)
\]

where \( \alpha_{sc} \) is the loss coefficient of the guided mode, \( \Gamma \) is the confinement factor and \( B \) is the gain constant of the amplifier.

Equation (3.19) calculates the variance \( \sigma_{out}^2 \) of the photons out of the amplifier. However in the model we are interested in the noise equivalent optical power produced by the amplifier \( P_{\text{opamp}} \). The relation between the two is:

\[
P_{\text{opamp}} = h \nu \sigma_{out}^2 \quad (3.22)
\]

Also in equation (3.19), \( \langle n_{in} \rangle \) the mean number of input photons per second to the amplifier is related to the input optical power to the front facet of the amplifier by:

\[
\langle n_{in} \rangle = \frac{P_{\text{opav(in)}}}{h \nu} \quad (3.19)
\]

So using (3.19), (3.22) and (3.23) it is possible to get an expression for the noise equivalent optical power produced by the amplifier:

\[
P_{\text{opamp}} = h \nu \sqrt{\frac{G P_{\text{opav(in)}}}{h \nu}} + (G - 1) n_{sp} B +
\]

\[
\frac{2 G (G - 1) n_{sp} P_{\text{opav(in)}} \xi}{h \nu} + (G - 1)^2 n_{sp}^2 B \quad (3.24)
\]

Finally to calculate the total noise equivalent optical power out of the amplifier \( P_{\text{opamp(out)}} \) the noise into the amplifier times the gain is added to the noise produced by the amplifier.
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\[ P_{\text{opnsel(out)}} = \sqrt{(GP_{\text{opnsel(in)}})^2 + P_{\text{opamp}}^2} \]  \hspace{1cm} (3.25)

where \( P_{\text{opnsel(in)}} \) is the input noise equivalent optical power to the amplifier. Again this equation is derived under the assumption that the two noise sources are uncorrelated.

To calculate the mean output optical power from the amplifier equation (1) of Mukai [3.7] is used:

\[ \langle n_{\text{out}} \rangle = G \langle n_{\text{in}} \rangle + (G - 1) n_{sp} B \]  \hspace{1cm} (3.26)

Combining this with (3.23) it becomes:

\[ P_{\text{opav(out)}} = G P_{\text{opav(in)}} + h f (G - 1) n_{sp} B \]  \hspace{1cm} (3.27)

The first term in the right side of the above equation represents the amplified optical power while the second one the optical power due to spontaneous emission.

Finally the output rms value of the modulated component of the optical signal \( P_{\text{orms(out)}} \) can be calculated from:

\[ P_{\text{orms(out)}} = G P_{\text{orms(in)}} \]  \hspace{1cm} (3.28)

Typical component parameters for a semiconductor laser amplifier can be found in APPENDIX B.

3.8. Equations used to calculate the transfer characteristics of a photodiode

The DC current out of a photodiode is given by:

\[ I_{dc} = \frac{\eta P_{\text{opav}} e}{hf} \]  \hspace{1cm} (3.29)

where \( P_{\text{opav}} \) is the mean optical power of the incident signal, \( \eta \) is the quantum efficiency, \( e \) is the electronic charge and \( hf \) is the energy of an
incident photon to the photodiode.

The electrical power $P_{el}$ due to the current to a load resistance $R_L$ following the photodiode is:

$$P_{el} = \left( \frac{\eta P_{opsv} e}{\hbar f} \right)^2 R_L$$ (3.30)

with the responsivity of the photodiode $R$ defined as:

$$R = \frac{\eta e}{\hbar f}$$ (3.31)

The RF electrical power out of the photodiode $P_s$ is given by:

$$P_s = \left( \frac{\eta P_{orms} e}{\hbar f} \right)^2 R_L$$ (3.32)

where $P_{orms}$ is the rms value of the modulated component of the optical signal. This formula applies for frequencies well within the 3 dB gain bandwidth of the photodiode. For higher frequencies an equivalent circuit of the photodiode has to be constructed and analyzed. A typical simplified approach is to take account of the capacitance of the depletion region plus any parasitic capacitance due to packaging. Then equation (3.32) becomes:

$$P_s = \left( \frac{\eta P_{orms} e}{\hbar f (1 + sR_L C)} \right)^2 R_L$$ (3.33)

Thus adding a pole at $\omega = 1/R_L C$ to the response of the photodiode.

To calculate the noise out of the photodiode, three sources of noise have to be taken into account:

i) The noise which is modulated on the optical carrier. This may have as origin the RIN of the laser, noise from the electrical source which modulates the laser etc. Using equation (3.32) the output electrical noise power of the photodiode due to this effect per unit bandwidth is:
\[ P_o = \left( \frac{\eta P_{opse} e}{hf} \right)^2 R_L \]  \hspace{1cm} (3.34)

where \( P_{opse} \) is the noise equivalent optical power as defined in Section (3.4) part (d).

ii) The shot noise of the photodiode which has power per unit bandwidth:

\[ P_{shot} = 2 e I_{dc} R_L \]  \hspace{1cm} (3.35)

where \( I_{dc} \) is the DC current out of the photodiode and is given by (3.29)

iii) The thermal noise which is generated at the load resistance of the photodiode and has mean square voltage given by:

\[ v^2 = 4kTBR_L \]  \hspace{1cm} (3.36)

However in the model, since the definition of the noise figure for the first amplifier presents certain problems (normally four parameters are needed to specify the noise figure of a transimpedance amplifier [3.9]) an effective input noise temperature \( T_a \) to the amplifier can be defined. The noise power per unit bandwidth measured at the output of the amplifier (at the frequency of interest) with the photodiode, bias circuit and matching circuit connected but without light shining at the photodiode (so the shot and RIN noise are effectively zero) is taken as \( GkT_a \) where \( k \) is Boltzman’s constant. So the equivalent input noise power per unit bandwidth at the amplifier (which is considered to be noiseless) is:

\[ P_{thermal} = kT_a \]  \hspace{1cm} (3.37)

Alternatively instead of defining the equivalent input noise temperature of a noiseless amplifier, the equivalent noise figure of a noisy amplifier can be defined as follows: If the amplifier was matched to an input source, then the input noise power to it per unit bandwidth would be \( KT \) where \( T \) is the ambient temperature (290°K). So the equivalent noise figure of the
amplifier can be defined as:

\[
F = \frac{S_{\text{in}}}{S_{\text{out}}} \frac{N_{\text{out}}}{kT} = \frac{T_a}{T} \quad (3.38)
\]

The total noise power \( P_{nse} \) out of the photodiode is:

\[
P_{nse} = P_o + P_{\text{shot}} + P_{\text{thermal}} \quad (3.39)
\]

The load resistance \( R_L \) of the photodiode is taken from the next component after the photodiode which can be an amplifier, a transmission line etc.

3.9. Equations used to calculate the transfer characteristics of an avalanche photodiode

The equations to model an avalanche photodiode are the same as those to model a standard p-i-n photodiode except for the following:

a) The responsivity is multiplied by the multiplication factor \( M \) i.e.

\[
R = \frac{M \eta e}{h f} \quad (3.40)
\]

b) The shot noise except for its increase by the multiplication factor (due to the current increase) is multiplied by \( 10^\xi \) where \( \xi \) is the excess noise factor of the photodiode (usually of the order of \( \xi = 0.2 \) to \( \xi = 0.9 \))

c) The bandwidth decreases with increasing multiplication factor. A complete analysis of this frequency dependence of the multiplication factor can be found in references such as [3.10] and is beyond the scope of this thesis. In general for high multiplication factors the gain-bandwidth product is constant. So in practice the 3 dB bandwidth can be found experimentally (or from the data sheet) for a given multiplication factor, and from that, and assuming a constant gain-bandwidth product, the frequency response can be calculated for every multiplication factor.

Finally the model includes some other components like optical attenuators, R.F. filters and R.F. amplifiers. Most of the equations for these components are straight-forward and do not need special discussion.

In an optical attenuator the parameters entered by the user are the total optical attenuation in dB and the number of elements the signal is divided into. So if the user enters an optical attenuation of 10 dB and the signal is split in 3 elements the total optical attenuation of the signal is:

\[ A = 10 + 10 \log_{10} 3 = 10 + 4.77 = 14.77 \text{ dB}. \]

In an R.F. filter the user enters some points from the attenuation versus frequency curve of the filter. Then the computer draws straight lines between them and calculates the attenuation at every particular frequency. The filter can be used in conjunction with other electrical components to model their non - constant frequency response. For example to model a laser which does not have a constant frequency response curve, the models of an R.F. filter and a laser can be cascaded in series.

In an R.F. amplifier the user enters the gain of the amplifier \( G \) and the noise figure \( F \). Then the signal power out of the amplifier is:

\[ P_{s(out)} = G \, P_{s(in)} \]  \hspace{1cm} (3.41)

and the noise power out is:

\[ P_{nse(out)} = G \, P_{nse(in)} + G \, (F - 1) \, k \, T \]  \hspace{1cm} (3.42)

where \( k \) is Boltzmann's constant and \( T \) is the absolute temperature.

3.11. A typical example

As a typical example of such a component-block and its interfaces with other components, the modelling block of a photodiode can be seen in Table
Chapter 3: Computer program to model optical links

3.1.
In the first column the first 4 lines represent the optical parameters required from the previous optical component. In the next 3 lines are the calculated average optical power of the optical signal $P_{opav}$, the rms value of the modulated component of the optical signal $P_{oms}$ and the noise equivalent optical power $P_{pase}$ in dBm and dBm/Hz. These are directly calculated from lines 2 and 4 above them using the standard formula:

$$P_{(dBm)} = 10 \log_{10}(1000*P_{(Watts)})$$ (3.43)

These lines give the user an indication of the signal and noise levels at each block in the system using the familiar dBm representation.

Under these 3 lines are the parameters of the particular photodiode used. These are determined by the user and they include the quantum efficiency of the photodiode, the wavelength of light and the temperature of the photodiode.

The first 4 lines of the second column are the output RF parameters of the photodiode. These will be passed to the next electrical block (which may be for example an RF amplifier). The next 3 lines below are the signal & noise powers $P_s$ & $P_nse$ in dBm and dBm/Hz and the S/N ratio in dBC Hz. These are again calculated from the first 3 lines and their purpose is to present the signals in the familiar dB representation.

Finally the last 12 lines of the second column are either fundamental constants such as Planck’s constant, electronic charge etc or other factors calculated from the photodiode parameters or the input signal. Examples of these are the temperature in K, shot noise etc.


Having presented the equations to model the individual components of optical links, some of the equations can be combined to give analytical expressions for the S/N ratio and R.F. link loss of directly and indirectly modulated optical links. The results are the following:

a) For directly modulated links
## PHOTODIODE

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<td>1.0E+09</td>
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<td>Total optical power in</td>
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<td></td>
<td>3.1E-03</td>
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<tr>
<td>Noise optical power in</td>
<td>$P_{opnse}$</td>
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<td>1.1E-09</td>
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<tr>
<td>Signal optical power in</td>
<td>$P_{om}$</td>
<td></td>
<td>6.7E-05</td>
</tr>
<tr>
<td>Total optical power in (dBm)</td>
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<td></td>
<td>4.9E+00</td>
</tr>
<tr>
<td>Noise optical power in (dBm)</td>
<td></td>
<td></td>
<td>5.9E+01</td>
</tr>
<tr>
<td>Signal opt. power in (dBm)</td>
<td></td>
<td></td>
<td>1.1E+01</td>
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</table>

<table>
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<th>Unit</th>
<th>Value</th>
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</thead>
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## PARAMETERS

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<td>Temperature (° celcius)</td>
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## CALCULATED & CONSTANTS

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<tr>
<td>Resp of photodio</td>
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<tr>
<td>Total current</td>
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<tr>
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</tr>
<tr>
<td>kT</td>
<td>4.0E-21</td>
</tr>
<tr>
<td>Shot noise</td>
<td>1.3E-20</td>
</tr>
</tbody>
</table>

Table 3.1. A typical component of the model.
Chapter 3: Computer program to model optical links

i) The S/N ratio is given by:

$$\frac{a^2 P_{R.F.in}}{(R_A + R_B)\left[a^2(I - I_{th})RIN + \frac{2hfa}{\eta \alpha_{TL}}(I - I_{th}) + \frac{kT(h\nu)^2}{(\eta \alpha_{TL} e)^2 R_L}\right]}$$

(3.44)

where $a$ is the electro-optic conversion efficiency of the laser, $P_{R.F.in}$ is the input R.F. power to the laser, $R_B$ is the internal impedance of the laser, $R_A$ is the matching impedance of the laser to the 50 $\Omega$ line, $(I-I_{th})$ is the bias current to the laser above threshold, $\eta$ is the quantum efficiency of the photodiode, $\alpha_{TL}$ is the optical attenuation of the link $e$ is the electron charge and $R_L$ is the load impedance of the photodiode.

ii) The R.F. link loss (from the input of the laser to the output of the photodiode) is given by:

$$\text{Loss}_{RF} = \frac{R_L (\eta a \alpha_{TL} e)^2}{(R_A + R_B) (h\nu)^2}$$

(3.45)

b) For indirectly modulated links and if small modulation depths are assumed.

i) The S/N ratio is given by:

$$\frac{\Gamma P_{opav(in)} a_{tt}}{V_\pi}\left[\frac{P_{opav(in)} a_{tt}}{2}\right]^2 RIN + \frac{hf P_{opav(in)} a_{tt}}{\eta \alpha_{TL}} + \frac{kT(h\nu)^2}{(\eta \alpha_{TL} e)^2 R_L}$$

(3.46)

where $P_{opav(in)}$ is the optical power to the modulator, $a_{tt}$ the optical attenuation of the modulator $P_s$, the R.F. power to the modulator $R_M$ the input R.F. impedance of the modulator, $V_\pi$ the half wave voltage of the modulator and $\Gamma$ a factor given in equation (3.11) which for small
modulation depths is $\pi/2$.

ii) The R.F. link loss (from the input of the modulator to the output of the photodiode) is given by:

$$\text{Loss}_{RF} = \frac{(\Gamma \eta \alpha_{IL} e^{P_{opav(n)}} a_{in})^2 R_L R_M}{(hf V_n)^2}$$  \hspace{1cm} (3.47)

It is not attempted in this thesis to obtain an analytical expression for the S/N ratio and gain of an optical link with one or more laser amplifiers. Although this is theoretically possible, in practice it is difficult as the equations to model an optical amplifier are complex and some parameters of the link are inter-dependent, for example the gain of the amplifier is a function of the input optical power to it.

3.13. Conclusions

In this Chapter a computer model has been presented which emulates passive or active optical links. To confirm its accuracy the model has been tested against experimental and theoretical results from other authors and the agreement has been of the order of 1-2 dB. So the model can be used to make predictions and as a helping CAD tool in designing analogue opto-electronic links. The modelling of such real systems using this computer program will be presented in Chapters 4, 7 and 9. Also the flexibility of the model allows for the future addition of other components and so it can be used in a wide range of microwave and optical applications.

3.14. References for Chapter 3


Chapter 3: Computer program to model optical links

on optical technology for microwave applications Arlington VA USA May 1984)


CHAPTER 4

ANALYSIS OF PASSIVE OPTICAL DISTRIBUTION NETWORKS AND COMPARISON WITH EXISTING TECHNIQUES

4.1. Introduction

Having analyzed the equations to model the power and noise parameters of optical links, these equations will be used to evaluate the theoretical performance of the various possible types of optical networks. All the parameters of the networks will be analyzed in detail and the effect of each of these parameters on the overall performance of the link will be discussed. Based on this analysis the currently available passive optical network architectures will be compared with one another and finally, using the current status of technology, the optimum solution will be proposed for each of the links involved in a typical phased array system. The links analyzed in this Chapter do not use optical amplifiers. An analysis of the feasibility and performance of optical distribution networks using optical amplifiers is presented in Chapters 5, 6 and 7.

The Chapter is organized as follows: First the various methods for the distribution of the R.F. reference and local oscillator signal are analyzed and compared. Then the various methods for the transmission of the return I.F. signal from the elements to the central processor of the array are discussed. As a specific example the discussion is focused on a satellite-based phased array currently being developed by Marconi Space Systems for the European Space Agency.

4.2. R.F. reference signal and local oscillator signal.

As discussed in Chapter 2, the distribution of these two signals presents more or less the same link requirements, thus the analysis will be common to both. For the transmitter four main techniques are possible to modulate the microwave signal onto the optical carrier:

a) Direct modulation of a laser diode.

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Chapter 4: Analysis of passive optical distribution techniques

b) The use of higher order harmonic modulation of a laser diode.
c) External modulation using an electro-optic modulator.
d) Heterodyning of two lasers to produce a microwave difference frequency.

In the receiver part of the link two techniques are possible: Either to amplify the signal and transmit it directly or use a microwave oscillator locked in frequency and phase to the reference signal. The two techniques will be compared, together with some other architectural details such as the choice of the photodiode.

4.2.1. Directly modulated links

In direct laser modulation the R.F. signal is superimposed onto the bias current of a laser resulting in an intensity modulation of the optical signal. In Chapter 3 the equations which describe the performance of directly modulated optical links have been presented. However these equations are rather complicated and it is difficult for the designer of an optical link to see how each parameter affects the overall performance of the link. It is thus desirable to present some typical design curves based on these equations. The example chosen is a proposed satellite-based link for a synthetic aperture radar (SAR) currently under research from Marconi Space Systems for the European Space Agency [4.1], [4.2], [4.3]. Its specifications and technical characteristics can be found in APPENDIX C.

In Fig (4.1) the S/N ratio of the optical link is plotted as a function of the total optical attenuation and the RIN of the laser. It can be seen from the curve that to achieve the typical S/N ratio of 110 dBc Hz required at the array elements (see Section 2.2.1A) the RIN of the laser is not a critical factor and lasers with RIN as high as -125 dB/Hz can be used without any degradation in performance. In Fig (4.2) the total attenuation of the optical link can be seen as a function of the number of elements the signal is divided into. The straight line represents the splitting losses if it is assumed that the optical splitters do not present any excess loss and that the optical connectors are lossless. The other curve above the straight line is more realistic and takes account of the excess losses of typical commercial optical splitters as used in [4.1] and [4.2].
Fig (4.1) Output S/N ratio of the link (after the two R.F. amplifiers) versus total optical attenuation for various levels of laser’s RIN.

Fig (4.2) Total optical attenuation of the link versus number of elements. The straight line curve represents the minimum theoretical value i.e. if it is assumed that there exist splitters and connectors with zero excess loss. The upper curve represents the value obtained assuming there are only the splitters and connectors used in the experimentally demonstrated link from Marconi analyzed in Appendix C.
Fig (4.3) shows how the contribution of the various noise sources affects the total S/N ratio curve of the link. These noise sources are:

a) The RIN which is the dominant factor for small values of optical attenuation. In this curve a value of RIN of -140 dB/Hz is assumed for the laser.

b) The thermal noise which is the dominant factor for large values of optical attenuation and

c) The shot noise of the photodiode.

In Fig (4.4) the S/N ratio and the R.F. loss of the link are plotted as a function of the load resistance of the photodiode. It can be seen that the sensitivity of the detector increases with higher load impedances. So the use of transimpedance amplifiers is recommended after the photodiode. However it should be kept in mind that at high input impedances and high frequencies of operation any parasitic capacitance (due to the photodiode, the tracks of the printed circuits etc) manifests itself much more and becomes the limiting factor in the performance of the link. So in optical high input impedance receiver designs, these capacitances become the limiting factor for the choice of the load impedance of the photodiode and should be eliminated if possible.

Two important parameters in the design of optical links are the noise produced by the laser (RIN) and the nonlinearities introduced by the laser and other link components. The next paragraphs will discuss these parameters.

Laser noise has been studied extensively by many authors [4.4], [4.5], [4.6], [4.7], [4.8]. It has its origin mainly in shot and recombination processes. For non-coherent links only optical intensity noise is important since phase noise manifests itself only in coherent applications.

References such as [4.9] derive analytic expressions based on the rate equations for the RIN of a laser versus frequency and a typical plot is presented in Fig (4.5). The point where the curve peaks is the relaxation frequency of the laser.

The second design problem with optical links is nonlinearity. Nonlinearity results in the generation of harmonics and intermodulation products. In a
Fig (4.3) Output S/N ratio of the link versus total optical attenuation. The relative contributions of the various noise sources in the system are shown. The RIN of the laser is -140 dB/Hz.

Fig (4.4) Output S/N ratio of the link and R.F. link loss (from the input of the laser to the load resistance of the photodiode) versus load resistance of photodiode.
Fig (4.5) Laser's RIN versus frequency of a typical laser diode (After [4.9]).

Fig (4.6) Plot of the second order harmonic (curve A), third order harmonic (curve B) and third order intermodulation product (curve C) versus modulation frequency of a typical laser diode. The modulation depth is 1%. (After [4.9]).
typical communications fibre optic link the nonlinearities are mainly produced in the directly modulated laser or electro-optic modulator but may also be produced in the R.F. amplifiers or by nonlinear effects in the optical fibre [4.10]. Various articles analyze the subject [4.11], [4.12], [4.13]. When the bandwidth of the links is small the third-order intermodulation products are the main cause of degradation of performance. This is because the harmonics are typically out-of-band while the third order intermodulation products are within the link passband. For directly modulated lasers the main cause of these nonlinearities is the laser photon-electron interaction. To model these nonlinearities the complete set of rate equations of the laser has to be solved. Based on the rate equations ref [4.9] derives expressions for the second and third order harmonic and the intermodulation products of a typical laser diode relative to the original carrier intensity and the results are plotted in Fig (4.6) for a modulation depth of 1 %.

Finally fibre nonlinearity is an important source of distortions in some optical links. However this manifests itself mainly in long multi-channel wavelength-division multiplexed single mode optical links. For a 30 metre single channel optical link, and at the optical powers of interest this nonlinearity is very unlikely to produce any effects [4.10].

Another important quantity when characterizing an optical link is its dynamic range. This gives a measure of the variation of the signal levels that can be carried on the link. Although there are many definitions for the dynamic range [4.14] usually it is defined as the ratio of the fundamental output to the biggest spurious response which affects the performance of the link when this spurious response equals the noise level. For example when the bandwidth is small, since the harmonics do not affect the performance of the link this spurious response is a third order intermodulation product. For the R.F. and L.O. links the dynamic range is usually not a significant problem as high levels of distortions can be tolerated.

4.2.2. Externally modulated links

Externally modulated links consist of an unmodulated laser which provides the optical carrier and an electro-optic modulator such as the one shown in
Chapter 4: Analysis of passive optical distribution techniques

Fig (4.7) which modulates a microwave signal onto the optical carrier. The electro-optic modulator (Mach-Zehnder interferometer) is usually a travelling wave device constructed on lithium niobate. As can be seen from Fig (4.7) the device consists of an optical waveguide over or near which a transmission line conductor is deposited which carries the microwave signal. The microwave signal generates fields which slightly change the refractive index of the optical waveguides. This in turn changes the relative phase of the two optical wavefronts in the two arms of the interferometer. By adding the two optical wavefronts at the end of the interferometer the resulting wave becomes intensity modulated at the microwave frequency.

The basic equations to model an electro-optic modulator are presented in Section 3.6. The modulator has a low-pass frequency response and the bandwidth is determined by the mismatch in the velocities between the microwave and optical signals. Because the modulation process involves no atomic transitions the modulator does not add any amplitude or phase noise to the signal. Also the intermodulation products are independent of frequency but depend on the nonlinearities of the transfer function of the modulator.

As the bandwidth is determined by the velocity mismatch between the microwave and the optical signals, various attempts have been made to reduce this mismatch. Lithium niobate is not very promising as the microwave and optical refractive indices have values of the order of 4.25 and 2.2 respectively [4.23]. However in GaAs and InP materials the two refractive indices are similar, and near zero velocity mismatch can be obtained [4.23].

Further discussion on structural details of electro-optic modulators is out of the scope of this thesis. Some articles which the reader can consult on electro-optic modulators are [4.15], [4.16], [4.17], [4.18], [4.19], [4.20], [4.21] [4.22] [4.23] and others.

In the following Table (Table 4.1) the typical parameters are given for some experimentally demonstrated electro-optic modulators.
Fig (4.7) Schematic diagram of an electro-optic modulator (After [4.22]).

Fig (4.8) Output S/N ratio of the link and R.F. link loss (from the input of the laser to the load resistance of the photodiode) versus input optical power to the modulator.
Chapter 4: Analysis of passive optical distribution techniques

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<th>3 dB FREQ. RANGE GHz</th>
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<th>$V_\pi$ Volt</th>
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<th>OPT. LOSS FACTOR dB</th>
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</table>

Table 4.1. Typical parameters for experimentally demonstrated electro-optic modulators.

For the reader to get a picture of the power budgets involved in a typical link using external modulation, the satellite-based link presented in APPENDIX C will be analyzed using an electro-optic modulator as the optical source. For this link and other single-tone systems for R.F. and L.O. signal distribution, the bandwidth is not a critical requirement so the modulator can be optimized for a single frequency. So the switching voltage $V_\pi$ does not need to be very high. In the model a $V_\pi$ of 2 V is assumed. With reference to Table 4.1 this is a reasonable assumption for a frequency of 5 GHz and a bandwidth of 300 MHz. Also the optical loss factor of this modulator is assumed to be 4 dB. Under these conditions to achieve an output S/N ratio of 113 dBC Hz an input optical power to the modulator of 3.2 mW is required. The R.F. power to the modulator required is 9.5 dBm which corresponds to an optical modulation depth of 0.95. Table 4.2 shows the modelled power and noise levels of the whole system. The input power to the modulator is less than that used in the directly modulated example but this strongly depends on the switch voltage $V_\pi$ of the device. Also the parameters of the link depend strongly on the input optical power to the modulator. Fig (4.8) shows the output S/N ratio and R.F. link loss of the link as a function of the input optical power to the modulator for various levels of the laser's RIN. From the curve it can be seen that in order to
<table>
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<th>SIGNAL POWER (dBm)</th>
<th>0.0</th>
<th>9.5</th>
<th>-3.7</th>
<th>-7.7</th>
<th>-11.2</th>
<th>-24.8</th>
<th>-57.0</th>
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<td>GAIN (dB)</td>
<td>9.5</td>
<td>4.0</td>
<td>3.5</td>
<td>13.5</td>
<td></td>
<td>24.0</td>
<td>33.0</td>
<td></td>
<td></td>
</tr>
<tr>
<td>NOISE POWER OUT (dBm/Hz)</td>
<td>-174.0</td>
<td>-158.9</td>
<td>-72.0</td>
<td>-76.0</td>
<td>-79.5</td>
<td>-93.0</td>
<td>-173.5</td>
<td>-146.1</td>
<td>-113.1</td>
</tr>
<tr>
<td>NOISE FIGURE (dB)</td>
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<td></td>
<td></td>
<td></td>
<td>3.7</td>
<td>5.0</td>
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<td>dB</td>
<td>dB</td>
<td>dB</td>
</tr>
<tr>
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<td>1.0</td>
<td>0.5</td>
<td>1.5</td>
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<table>
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<th>atten. coeff.</th>
<th>No of element</th>
<th>No of element resist.</th>
</tr>
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<tbody>
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<td></td>
<td>4.0</td>
<td>2.0</td>
<td>16.0</td>
</tr>
<tr>
<td></td>
<td>dB</td>
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<table>
<thead>
<tr>
<th>OTHER CHARACTERISTIC</th>
<th>laser power in</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>3.2 mW</td>
</tr>
</tbody>
</table>

Table (4.2) Power and noise levels of the forward link presented in APPENDIX C but using an electro-optic modulator instead of direct modulation.
Chapter 4: Analysis of passive optical distribution techniques

achieve high link performance, high input optical power is required. It also shows that using high input optical powers very low R.F. link losses can be obtained in comparison with directly modulated links. However the maximum amount of optical power that can be fed to the modulator is usually limited to between 10 mW and 100 mW by optical damage effects to the single mode waveguide. In Fig (4.9) the S/N ratio and R.F. link loss of the link are plotted as a function of the switching voltage $V_\pi$ of the modulator. It can be seen that in order to achieve efficient links modulators with low $V_\pi$ are required.

In general the main advantages which electro-optic modulators claim in comparison with directly modulated systems are the following:

a) The separation of the light generation and modulation functions which results in the optimization of each of these functions. For example CW lasers with very low values of RIN ($\approx -160$ dB/Hz) can be employed. However as can be seen from Fig 4.8, 4.1 and 4.3 the RIN of the laser mainly becomes of importance when it is required to make links with very high S/N ratio and the links have very low optical attenuation. For example Fig 4.1 shows that for the typical S/N specification of 110 dBc Hz lasers with RIN of as high as -120 dB/Hz can be used.

b) Using high input optical powers to the modulators very low R.F. link losses can be obtained in comparison to directly modulated links. For example [4.24] reports a net R.F. link gain of 11 dB with an input optical power to the modulator of 55 mW (17.5 dBm). This property allows more elements to be fed from a single optical source than in the case of directly modulated links. However the maximum optical power to the modulator is limited by damage effects to the waveguide. Also when optical amplifiers are used in the distribution networks this property becomes of limited importance.

c) The upper frequency-response limit of electro-optic modulators can be much higher than that of directly modulated lasers. For example devices which modulate the light up to 60 GHz have been demonstrated [4.25]. So electro-optic modulators is the most likely choice in links operating above $\approx 25$ GHz and below $\approx 60$ GHz.
Fig (4.9) Output S/N ratio of the link and R.F. link loss (from the input of the laser to the load resistance of the photodiode) versus modulator’s switching voltage $V_\pi$.

Fig (4.10) The use of the higher order harmonics of a laser diode to indirectly lock an oscillator.
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The main disadvantage of electro-optic modulators relative to directly modulated links is that they add extra complexity to the design and construction of the system as more components are required.

Based on the above arguments electro-optic modulators present an advantage and should be the choice instead of directly modulated lasers when:

a) Optical links with S/N ratios higher than $\approx 130$ dBC Hz are required
b) Modulation frequencies higher than $\approx 25$ GHz and up to $\approx 60$ GHz are required
c) Passive optical networks in which each optical source has to feed many T/R elements are required.

4.2.3. The use of higher order harmonic modulation of laser diodes.

The bandwidth of directly modulated optical links is currently limited by the upper frequency limit of laser diodes. This is currently below 25 GHz and it is not anticipated that it will exceed 25 GHz in the near future. Also the received signal from the photodiode is used in some architectures to lock an oscillator in frequency and phase. So the Microwave - Lightwave engineering group of Drexel University has proposed another solution to achieve high frequencies using directly modulated laser diodes [4.26], [4.27], [4.28], [4.29], [4.30]. In this scenario (see Fig 4.10) the transmitting laser is heavily modulated with a signal of frequency $f_o$. Because the intensity of this input R.F. signal is very high, the laser is driven into its nonlinear region and harmonics of frequencies $2f_o$, $3f_o$ etc are emitted alongside with the fundamental carrier. The frequencies of these harmonics can be above the upper modulation limit of the laser. At the receiver side of the optical link, a bandpass filter is used to select only one of these harmonics (for example $4f_o$ in Fig 4.10). This signal is usually used to indirectly lock an oscillator which works in an even higher frequency, for example $12f_o$. In this way links which operate at frequencies up to 40 GHz have been demonstrated [4.26] and the authors argue that frequencies up to 100 GHz can be achieved [4.31]. The characteristics of this architecture should be compared with those of indirectly modulated links as both are capable of transmitting at much
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higher frequencies than it is possible with directly modulated laser diodes. The main comparison points are the following:

a) This technique can be realized with relatively simple opto-electronic hardware while if indirect modulation is used, a special modulator will have to be developed, optimized for the frequency of interest.

b) The harmonic of interest is usually a small proportion of the fundamental carrier and the other harmonics of the diode. Thus the R.F. link losses of such links are high and a limited number of elements can be fed from a single laser diode if passive optical techniques are used. If optical amplifiers are used, these will saturate due to the fundamental power and they will exhibit lower gains and reduced electro-optic efficiencies. This is because they will have to amplify the fundamental carrier and the other harmonics of the signal together with the harmonic of interest.

So in generally it is recommended that for frequencies above the maximum modulation frequencies of lasers (= 25 GHz) electro-optic modulators are used except when the costs to develop such devices are not justified by the applications concerned.

4.2.4. The use of coherent techniques

Another promising technique for the generation of the R.F. and L.O. reference signals is to use the beating of two closely spaced optical carriers. A typical network can be seen in Fig 4.11. Here the optical frequencies of a master and a slave laser are tuned so that their difference is equal to the microwave reference signal. Coherent methods offer a number of advantages relative to direct modulation. These are the following:

a) High microwave frequencies of operation with good phase accuracies and large modulation depths can be achieved. In fact this is probably the only currently available technique with the potential to achieve reasonable modulation of light above ~ 60 GHz.

b) Using this technique opens more possibilities to perform the beamforming
Fig (4.11) The use of two heterodyned laser diodes to produce a microwave carrier.

Fig (4.12) Optical control of a microwave oscillator combining direct injection locking and a phased locked loop. (After [4.38]).
and other processing in the optical domain. In the literature several coherent optical beamforming techniques have been demonstrated (see Chapter 8, Sections 8.4 and 8.5).

However it is not technically easy to construct such optical transmitters mainly because of the following reasons:

a) The optical frequency of a laser depends on its temperature ($\approx 20 \text{ GHz/}^\circ\text{C}$ for semiconductor lasers) and thus the temperature of the lasers must be controlled to very narrow limits. With temperature stabilization the beat frequency linewidth can be reduced to a few MHz and other techniques such as phase locking are needed to reduce it even further.

b) The mechanical-optical-electronic setup is difficult to construct and is very sensitive to external factors such as vibrations, air currents and temperature fluctuations.

c) The linewidth of the lasers is a very critical factor and must be chosen to be as low as possible [4.32]. Thus either external cavity or distributed feedback (DFB) or Nd:YAG lasers are the most likely choices for use in such systems.

Whether such systems can be integrated in phased array antennas mainly depends on their future volume, cost, power consumption, reliability and sensitivity to external factors such as temperature and vibrations.

If optical amplifiers are used in the network, then only one optical source can be used for the whole array so its price and power consumption are not major constraints. However its reliability and sensitivity to external factors are very important since failure of the source results in failure of the whole array. If optical amplifiers are not used and especially for satellite-based applications these coherent sources should be in an integrated form.

In general in the future coherent sources may replace directly and indirectly modulated links provided they can be made inexpensive, small and reliable in the following cases:

a) When the frequency and phase stability required are beyond the
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capabilities of conventional sources and b) when coherent beamforming
techniques are used to steer the beam.

4.2.5. The use of oscillators locked to the received reference signals

At the elements of the array, after the photodetection of the R.F. or L.O.
reference microwave signal, two architectures are possible:

a) To amplify directly this signal and either send it to the antenna or use
it as the local oscillator. This architecture has the advantage that the
communication data or the frequency chirping in a radar can be introduced
at the central processor of the array. However this places severe
requirements on the optical link in terms of bandwidth, dynamic range and
noise figure. Furthermore high phase accuracy is required from the
microwave amplifiers, which may not be easy to achieve at high frequencies
and gains.

b) A second technique which appears to offer more advantages is to use the
received reference signal to lock an oscillator in frequency and phase.
This locking can be either at the fundamental frequency of the reference
signal or at a harmonic of it. Various papers have investigated the subject
and three possible approaches have been demonstrated. These are:
i) direct optical injection locking, ii) indirect optical injection locking
and iii) combination of one of the previous methods with phase lock loops.
In the direct optical injection locking scenario, the received optical
reference signal is coupled directly into the active region of the
oscillator. The technique has been demonstrated for IMPATT [4.33], [4.34]
[4.35] and FET [4.36], [4.37] oscillators for frequencies up to X-Band, but
in general the coupling to the active region of the devices has been poor
resulting in small locking ranges (about 0.1 % of the operating frequency.)
In indirect optical injection techniques, the optical signal is detected by
a photodiode, amplified and then electrically injected to the oscillator.
The advantages of this method relative to the previous one are mainly the
following: a) Higher coupling and quantum efficiencies of the photodiodes
as compared with the active regions of the oscillators. b) The signal is
amplified after photodetection and prior to injection to the oscillator and
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c) Availability of commercial components.
This method can be used to lock the oscillator either to the frequency of
the reference signal or to a harmonic of it [4.26]. Using this technique
together with utilization of the higher order harmonics of heavily
modulated laser diodes, locking an oscillator at 39 GHz while modulating
the laser at only 3.2 GHz [4.26] has been achieved. In ref [4.31] the
technique is proposed to achieve locking at frequencies up to \( \approx 100 \) GHz.

One of the problems of using either direct or indirect optical injection
locking is the phase difference between the locking signal and the
oscillator output. This depends on factors such as the oscillator
free-running frequency [4.38]. So in a phased array, variations in the
free-running frequency between oscillators can occur due to manufacturing
tolerances and temperature gradients, and this may introduce unacceptable
phase shifts between elements [4.38]. To solve this problem, techniques
which combine optical injection and phase lock loops have been demonstrated
[4.38]. A typical example of such a scheme can be seen in Fig 4.12. In this
setup the phase difference between the reference signal and the directly
injected oscillator are compared with a phase detector and this in turn
sends a signal through the loop filter to the oscillator to control its
phase. Using this technology, phase stability can be achieved over the
locking range of the oscillator.

In general the use of oscillators and T/R level mixing appears to be
necessary for arrays operating at high frequencies where the required phase
accuracy cannot be achieved with direct amplification of the signal. If the
array is required to operate reliably without phase shifts arising from
temperature gradients and manufacturing tolerances, some form of phase
stabilization technique must be used together with injection locking. It is
expected that a solution using harmonic injection locking combined with
phase stabilization techniques will be demonstrated.

4.3. I.F. return signal

The next signal which is a candidate for optical distribution in a typical
phased array system is the return I.F. signal. Its characteristics and link
requirements are discussed in Chapter 2 Section 2.2.2B.
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Three main ways have been proposed for its distribution. These are the following:

a) Direct modulation of a laser (or a LED) diode.
b) Frequency modulation of an oscillator with the signal and then imposing the resulting signal on an optical carrier.
c) Conversion of the signal to digital format and then intensity modulating the optical carrier.

The use of electro-optic modulators is of course another possibility. However for the case of the I.F. signal they are not considered to be suitable. This is because each of them requires a CW operated laser source and the whole structure of the system is more complicated than the directly modulated laser diode. In the forward link these disadvantages may be offset by the fact that one electro-optic modulator can feed many T/R modules or sub-arrays. In the case of the return link every T/R module or sub-array needs its own dedicated electro-optic modulator. This makes every T/R module more complicated and power-consuming so that with the current status of the technology it is not believed they present an effective solution except when the SNR requirements of the link are very critical.

4.3.1. Analysis of a link using direct modulation of a LASER or a LED.

This is the most straightforward and conceptually simple solution. As discussed in Chapter 2 the dynamic range of this signal for modern military radars should be about 70 dB in 10 MHz bandwidth. This specification can be translated into a spurious-free dynamic range of 140 dBC Hz. From the analysis of the forward link Fig 4.1 and Fig 4.2 it can be seen that for a point-to-point link (no splitting) and with a laser of RIN -140 dB/Hz a S/N ratio can be achieved of 135 dBC Hz. This is 5 dB short of the required specifications and the situation becomes much worse when the intermodulation products of the laser are taken into account which for a typical laser operating under these conditions are about -60 dB relative to the carrier. So it can be seen that for such radars the specifications cannot be fulfilled with direct modulation and so some other solution has to be sought. However for communications and other radar systems not
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requiring so high dynamic ranges, the intensity modulation presents the technically easiest solution. Two architectures are possible for the combination of the optical signals from the elements of the array. The first and most conceptually simple is to use a dedicated optical link with its own laser and photodiode for every element of the array (see Fig 4.13). However this solution apart that it requires a large number of receiver circuits, it also requires an R.F. combining network to combine the received signals in the microwave domain which makes the whole system relatively complex.

A simpler solution is to combine the element signals in the optical domain (see Fig 4.14). In this solution only one receiver is needed for the whole array. One potential problem in using this solution is the laser beating noise, that is the noise produced due to the heterodyne of any two lasers emitting at optical frequencies whose difference is within the bandwidth of the I.F. signal. To produce a mathematical model to analyze this problem is not easy as it would need to take into account the statistics of the central wavelengths of the lasers used in the return link, the linewidths of these lasers and other factors. In general the results can be from catastrophic (if only two very low linewidth lasers are present in the system and the frequency of the heterodyned signal is in the bandwidth of the I.F. signal) to negligible (if the system consists of many incoherent or low coherence optical sources). Thus it is recommended that a large number of multi-mode, wide linewidth lasers are used in the return link, or even better incoherent LED’s. This last solution has been chosen by some designers [4.1], [4.2], [4.3] but it is not very efficient in terms of electrical power consumption and complexity. This is because due to their low electro-optic conversion efficiencies the LED’s need high R.F. powers to drive them.

Using optical combination of the I.F. signals results in combination gain. Assuming the noise contributions from the optical channels are uncorrelated, the equations to model this are the following:

\[
P_{\text{rms(out)}} = \left(\frac{N}{L}\right) P_{\text{rms(in)}}
\]

and

\[
P_{\text{opnse(out)}} = \sqrt{\frac{N}{L}} P_{\text{opnse(in)}}
\]
Fig (4.13) The use of dedicated optical links and an R.F. combining network for the reception of the I.F. signal.

Fig (4.14) The use of an optical combining network for the reception of the I.F. signal.
where \( N \) is the number of channels, \( L \) is the combination loss of the coupler, \( P_{\text{oms(out)}} \) the rms value of the modulated component of the optical signal out of the combiner (for a full definition of this quantity see Section 3.4 part c), \( P_{\text{opnse(in)}} \) the rms value of the modulated component of the optical signal of each channel into the combiner, \( P_{\text{opnse(out)}} \) the noise equivalent optical power out of the combiner (for a full definition of this quantity see Section 3.4 part d) and \( P_{\text{opnse(in)}} \) the noise equivalent optical power from each channel into the combiner.

So the optical signal to noise ratio is improved by \( \sqrt{\frac{N}{L}} \) and the R.F. S/N ratio (not taking into account the thermal and shot noise in the photodiode) is improved by \( \frac{N}{L} \). This improvement in the S/N ratio can significantly increase the dynamic range but in practice the excess losses of the couplers tend to be of very high levels, similar to the splitting losses of the same devices [4.1], [4.2]. Two types of optical couplers exist, fusion and waveguide, but they both tend to have high combination losses (of the order 3 dB per two channels combining). So these combination losses tends to offset the combination gains unless more efficient couplers become available. Two solutions to this problem are possible:

a) To shine the incoming signals in parallel to the facet of the photodiode. Using this method would also automatically solve the problem of laser beating noise as the optical signals are not mixed in the optical domain. However this solution is limited by how many optical fibres can shine in parallel to the area of the photodiode. (The area of the photodiode is limited by the operating frequency)

b) To use an optical amplifier before the photodiode to boost the optical signal to the required level. As it will be shown in Chapters 5, 6 and 7, fibre amplifiers offer low noise and distortion-free performance at the frequencies of interest.

4.3.2. The use of frequency or digital modulation to transmit the return signal

Since the use of directly modulated lasers cannot satisfy the dynamic range requirements of modern military radars, ref [4.39] proposes two alternative solutions: Frequency modulation of a microwave carrier and digital
modulation of the IF signal.

4.3.2A. Frequency modulation of a microwave carrier

This technique consists of frequency modulating the required signal on an R.F. carrier using a voltage controlled oscillator and then intensity modulating this onto the optical carrier. The frequency of the R.F. carrier should be about 5 - 10 times that of the I.F. signal. This technique is used to avoid the problem of the thermal noise generated by the photodiode which limits the performance of an intensity modulated system requiring high dynamic range. Using frequency modulation the requirements are now much more easily satisfied as only a moderate frequency deviation is enough to satisfy the dynamic range requirements and the signal is much above the thermal noise limit.

A block diagram of such a solution can be found in Fig 4.15. Special attention should be given in the linearity of the voltage controlled oscillator (VCO) and the discriminator over the tuning range which may result in unwanted spectral components. The amount of frequency deviation with voltage is determined by two factors: a) The highest signal level must be within the linear tuning regime of the oscillator and discriminator and b) The lowest signal level must produce a frequency deviation above the FM noise of the oscillator.

In general the above specifications can be satisfied without any particular difficulty. However to construct a VCO and a discriminator for every element of the array is a difficult and expensive task. Also optical combining cannot be used and a microwave combining network is required after the signal has been converted into intensity modulation format. So it is doubtful whether optics offers any advantage over a purely microwave combining network situated after the receive elements.

4.3.2B. Digital modulation

This consists of the conversion of the return IF signal into digital format and then transmitting it through an optical link. After investigating this possibility ref [4.39] comes to the conclusion that in order to satisfy the dynamic range and bandwidth requirements of modern Doppler radars, better
Fig (4.15) The use of frequency modulation of a microwave carrier to transmit the I.F. signal.
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analog-to-digital converters are required. However this article was written in 1984 and today with the advance of integrated digital technology this solution can be realized though at the expense of power hungry and expensive analog-to-digital converters at every element of the array. [4.40], [4.41]. At the central processor of the array after reception of the digital signals two scenarios are possible:

a) The digital signals are converted into analogue format and then a microwave network is used to combine them. However this solution does not offer any advantages to justify its complexity and it is easier for a purely microwave combining network to be used after the array elements.
b) The signals are combined in the digital domain using one or more very fast digital processors. In such scenarios the phase shifts are introduced digitally by the central processor and not at the elements of the array. The advantage of this solution is that it is possible for the processor to introduce different phase slopes to the received signals simultaneously thus allowing the array to look at many directions with maximum gain simultaneously. This is very important for example in communication arrays which are required to receive the signals of many mobile terminals with maximum antenna gain simultaneously.

However a further discussion of digital beamforming techniques is out of the scope of this thesis and there are many papers in the literature to which the reader can refer [4.40], [4.41].

4.4. Conclusions

In this Chapter an examination and comparison of the passive optical techniques for the distribution of the microwave signals to and from the elements of the array has been carried out. The results show that for the distribution of the RF and the LO signals, direct modulation of a laser diode is the technically simplest solution.

Electro-optic modulators should be used when:
a) Optical links with S/N ratios higher than $\approx 130$ dBC Hz are required
b) Modulation frequencies higher than $\approx 25$ GHz and up to $\approx 60$ GHz are required
c) Passive optical networks in which each optical source has to feed many T/R elements are required.
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The technique of higher order harmonic modulation of laser diodes offers an advantage only when it is difficult and expensive to develop an electro-optic modulator to work at the frequency of interest.

Coherent techniques, although difficult to construct, offer the potential of very high frequencies of operation with good phase stability and large modulation depths. Also they can be used with coherent beamforming techniques.

At the array elements, oscillators locked to the base frequency or to a harmonic of the received signal offer increased phase accuracy together with the potential of higher frequencies of operation.

For the IF signal again direct modulation is the technically easiest solution but cannot satisfy the dynamic range of some military radars. As an alternative two other solutions are proposed, namely frequency modulation and digital modulation.

4.5. References for Chapter 4


[4.5] Yamamoto Y., Saito S., and Mukai T. "AM and FM quantum noise in

*References [4.1] and [4.2] are not in the public domain and permission has been obtained from Marconi Space Systems and the European Space Agency to use some of the information contained in them.
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1991


[4.38] Blanchflower I.D. and Seeds A.J., "Optical control of phase and frequency of microwave oscillators" IEE Colloquium on Optical Control and Generation of Microwave and Millimetre-Wave Signals, April 1989


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CHAPTER 5

THE USE OF OPTICAL AMPLIFIERS FOR SIGNAL DISTRIBUTION IN OPTICALLY CONTROLLED PHASED ARRAY ANTENNAS - A) GENERAL CHARACTERISTICS OF OPTICAL AMPLIFIERS AND NETWORK REQUIREMENTS

5.1. Introduction

One of the main limitations of the passive optical networks analyzed in Chapter 4 is the small number of elements (usually about 30) which can be fed from a single optical source. To overcome this problem most designers use many optical sources, each one of them feeding a small number of elements. [5.1], [5.2], [5.3] A microwave network is used to distribute the R.F. signals to these sources. However, as antenna applications move to higher frequencies the optical sources and other microwave components become more expensive. Also, new, more complicated sources such as laser heterodyne systems are being considered for use in phased arrays [5.4], with control circuits which are difficult and expensive to construct. So there is a desire to use a single optical source to feed all the elements of the array. This can be done if the output optical signal of this source can be amplified in the optical domain. This can be achieved using optical amplifiers.

Although in the literature the use of optical amplifiers has been extensively examined for digital telecommunications links [5.5], [5.6] and cable television distribution systems [5.7],[5.8] little attention has been given to the use of laser amplifiers in optically controlled phased array antennas. The object of this and the next two Chapters is to cover this gap and examine the feasibility of using laser amplifiers in the distribution networks of optically controlled phased array antennas.

Various types of optical amplifiers have been demonstrated. Of these the most important and most widely used are the semiconductor laser amplifier and the rare earth doped fibre amplifier. When this work was beginning semiconductor laser amplifiers were already a technology under intensive research, while fibre amplifiers were just beginning to emerge in some
laboratories. So in this thesis the research is mainly focused on semiconductor laser amplifiers and there is some discussion - mainly comparison - with the characteristics of fibre structures.

The work presented in this and the next two Chapters is organized as follows:
First the general characteristics of the two types of amplifiers will be discussed with reference to their use in optically controlled phased array antennas. The discussion will be with respect to their structure, optical transmission frequency, gain stability with respect to bias current, polarization and temperature, and their noise characteristics. Then requirements for optical amplification in a phased array will be considered.
To assess the suitability of laser amplifiers in optically controlled phased array antennas the following factors have to be taken into account:

a) Nonlinearities in the response of the amplifiers. These may produce harmonics and intermodulation products.
b) Noise performance: degradation the laser amplifiers introduce in the S/N ratio of the optical links.
c) The power budget of optical links using optical amplifiers.

The first of the above factors is discussed in Chapter 6 where a computer program is presented to model the performance of semiconductor amplifiers with intensity modulated signals. The remaining factors are discussed in Chapter 7, where the use of optical amplifiers is analyzed for the forward link of the phased array antenna whose main characteristics are presented in Appendix C and has been analyzed using only passive optical networks in Chapter 4.

5.2. General characteristics of semiconductor and fibre amplifiers

5.2.1. Structure of semiconductor laser amplifiers

The semiconductor laser amplifier has essentially the same structure as the laser oscillator, but anti-reflection coatings are applied to the laser
Chapter 5: General characteristics of optical amplifiers

facets to reduce their reflectivity (see Fig 5.1). In the literature semiconductor laser amplifiers are usually divided into two types: Fabry-Perot (FP) and travelling wave (TW). They differ in their facet reflectivity.

FP amplifiers have high facet reflectivities (0.01 - 0.3) and operate in resonant mode usually just below threshold. Their 3 dB gain bandwidths are small (typical value \( \approx 5 \) GHz) \([5.5]\) so to keep the gain constant the device temperature and current, and the wavelength of the input signal have to be tightly controlled. FP amplifiers also exhibit gain saturation at a lower input power level than TW amplifiers and have lower gain.

TW amplifiers have low facet reflectivities (<0.01) arising from anti-reflection coatings applied to their facets. TW amplifiers offer wide bandwidths (\( \approx 1000 \) GHz), while the gain is less affected by temperature. For the above reasons only TW amplifiers will be considered for applications in optically controlled phased arrays.

If a laser amplifier has exactly the same structure as a laser oscillator (i.e. the planes of the facets are at 90° to the waveguide) then very tight control of the anti-reflection coating process is needed to achieve low reflectivities \([5.9]\), \([5.10]\), \([5.11]\). For example to obtain a 30 dB internal gain with less than 3 dB gain ripple the facet reflectivities must be \(1 \times 10^{-4}\) or less, which requires extremely tight control on the refractive index and the thickness of the dielectric layers. Eisenstein et al for example \([5.10]\) have applied sputtered SiN films whose refractive indices were controlled by adjusting the partial pressure of nitrogen in plasma. This method of tight control of the evaporation process has the following disadvantages:

a) It is very expensive for mass production since the amplifiers have to be processed one at a time and the coating process must be controlled very accurately.

b) Since the wavelengths giving the minimum reflectivity for both facets must be adjusted to the operating wavelength of the TWA, the amplifier has effectively a lower operating bandwidth.

A better way to achieve low facet reflectivities is to make the facets of the amplifier at an angle to the waveguide \([5.12]\), \([5.13]\), \([5.14]\), \([5.15]\) (see Fig (5.2)). In this way the light reflected by the cleaved facets does
Fig (5.1) Principle of operation of a semiconductor laser amplifier.

\[ \theta_i = 7^\circ \quad \theta_o = 22^\circ \]

Fig (5.2) Structure of semiconductor laser amplifier with angled facets (After [5.12]).
not couple back into the waveguide [5.12]. This results in an effective facet reflectivity of about 0.02 without anti-reflection coating. The effective facet reflectivity is further reduced to $1 \times 10^{-4} - 8 \times 10^{-4}$ by the application of conventional 0.01 anti-reflection coatings to both facets [5.13], [5.14].

Light is usually coupled into and out of the amplifier's waveguide through lens-ended fibres. This requires precise alignment of the fibres and results in optical losses of the order of 3 dB - 5 dB per facet. This considerably reduces the effective gain of the amplifier.

5.2.2. Structure of fibre amplifiers

Fibre optical amplifiers consist of standard glass optical fibres whose core is doped with a rare earth element usually erbium or neodymium. Typical doping concentrations vary from a few tens to a few hundreds of parts per million [5.16], [5.7]. Fibre amplifiers work on the following principle: A pump signal of shorter wavelength than the amplified signal is injected into the fibre. This excites the rare earth ions into a higher energy level thus causing a population inversion. When the signal to be amplified enters the fibre, it stimulates the emission of photons of the same frequency and phase and so is amplified. The pump optical signal can be removed before the receiving photodiode (so that it does not induce extra shot noise) using an optical filter.

The optimum length of the fibre amplifier depends on the doping concentration, the available pump power, the efficiency, and the gain required [5.17], [5.18]. Typical values vary between 1 m and 100 m.

Since fibre amplifiers have essentially the same structure as standard optical fibres, they can be fusion spliced into systems with splicing losses as low as 0.1 dB. This means that unlike semiconductor laser amplifiers they do not have the problem of end-facet reflectivities or coupling losses.

5.2.3 Monolithic integration of semiconductor and rare earth doped amplifiers

Since semiconductor laser amplifiers are integrated structures, they can be
easily integrated into more complex opto-electronic circuits. In such structures, the problem of facet reflectivity is also automatically solved since there are no air interfaces between the facets of the amplifier and the input and output waveguides. However in the case of erbium amplifiers, the dopant concentrations cannot be very high, and thus such structures are not offered for monolithic integration. Recently some work has started at University College London, attempting to tackle the problem.

5.2.4. Wavelength of operation of semiconductor laser amplifiers

The central wavelength of operation of semiconductor laser amplifiers is similar to that of the corresponding laser if its facets were not anti-reflection coated. If the facet reflectivities of the amplifier are very small, the 3 dB bandwidth for the unsaturated amplifier depends on the gain spectrum of the medium and is proportional to $1/\Gamma L^2$ [5.5] thus decreasing as the length $L$ and the confinement factor $\Gamma$ increase. Typical values of bandwidth are of the order of 5 nm to 100 nm ($\approx 1$ THz - 15 THz). If the reflectivities of the facets are finite then ripples appear in the gain versus wavelength curve with a period of the order of 1 nm due to the Fabry-Perot resonances in the amplifier cavity. If the reflectivities are further increased then these effectively limit the bandwidth of the amplifier to very small values.

5.2.5. Wavelength of operation of fibre amplifiers

For fibre amplifiers two wavelengths are of importance: a) the absorption wavelength of the doping ions and thus the wavelength at which the pump source should be operated and b) the fluorescence wavelength of the doping ions, which is the wavelength of the input and output signal. A comprehensive study of the absorption and fluorescence spectra of all except one of the rare earth ions in $\text{SiO}_2-\text{GeO}_2-\text{P}_2\text{O}_5$ glass has been reported by Ainslie et al [5.19], and should be consulted for further information. From all the rare earth elements, erbium, praseodymium and neodymium are the most important since their fluorescence wavelengths are compatible with those used in telecommunications. Erbium ions have absorption wavelengths
of interest at 800 nm, 980 nm and 1480 nm. The fluorescent wavelength of interest is 1550 nm which coincides with the third telecommunications window. At this wavelength the bandwidth is about 4 nm (< 500 GHz). Praseodymium has an absorption band at 1450 nm and fluorescence bands at 1050 nm and 1550 nm. Finally neodymium has absorption bands at 800 nm and 900 nm and fluorescence bands at 1060 nm, 900 nm, and 1320 nm although high efficiency at this wavelength is hard to obtain. From this and the previous paragraph it can be seen that rare earth and especially erbium fibre amplifiers have narrower bandwidths. This has the advantage that it provides an effective filter which limits the accumulation of spontaneous emission in a chain of many amplifiers, but makes the wavelength tolerances of the source lasers in a microwave distribution system stricter and the use of incoherent wavelength division multiplexing techniques more difficult.

5.2.6. Polarization dependence of gain of optical amplifiers

In erbium doped amplifiers, since the structure has cylindrical symmetry the basic gain is the same for both input polarizations. However, in semiconductor laser amplifiers the gain is different for the TE and TM polarization modes. This is mainly because the confinement factors $\Gamma$ are not the same for the two polarization directions. Fig 5.3 shows the optical gain versus input current for the two polarization modes of a commercially available laser amplifier (BT&D SOA1100/SOA3100). It can be seen that for a typical bias current of 100 mA the gain difference between the two modes is about 4 dB. This difference creates a problem in the use of semiconductor laser amplifiers in optically controlled phased arrays as it introduces an uncertainty in the R.F. power budget of 8 dB. There are 3 possible ways to overcome this:

a) To calculate for the "worst case" when designing the power budget of the system and then use automatic gain control (AGC) amplifiers in the optical receivers. In this way the final S/N ratio characteristics of the system will be better than the calculated results.

b) To construct polarization insensitive semiconductor laser amplifier
Fig (5.3) Fibre-to-fibre gain versus bias current characteristics for the two polarization modes of a commercial semiconductor laser amplifier (BT&D SOA1100/SOA3100).
(After the data sheet of BT&D SOA1100/SOA3100).

Fig (5.4) Gain versus current to the pump laser diode of the fibre amplifier presented in reference [5.35]. The voltage across the laser diode is about 2 V.
structures. Various ideas have been reported in the literature [5.20], [5.21], [5.22], [5.23], [5.24]. The most important of these are:
i) The design and construction of a semiconductor laser amplifier chip structure with equal confinement of the two modes and thus reduced polarization sensitivity [5.22].
ii) To cascade two amplifiers in series or in parallel with their polarization angles differing by 90°. In the parallel configuration the gain is limited by the losses of the optical splitters and couplers while in the series configuration the main problem is the optical coupling between the two amplifier cavities. To solve this problem, reference [5.20] inserts a polarization insensitive isolator between the two amplifiers.

c) An alternative method to control dynamically the polarization and temperature characteristics of semiconductor laser amplifiers using an all-electronic method has been proposed and experimentally demonstrated by Ellis et al [5.25], [5.26], [5.27]. The method uses the product of the voltage and the current at the terminals of the optical amplifier to detect the optical power inside the amplifier. Then using a standard control loop the optical power can be kept approximately constant against variations in gain due to the input signal, signal polarization and optical amplifier temperature.

5.2.7. Temperature dependence of gain of semiconductor laser amplifiers and fibre amplifiers.

The gain of semiconductor laser amplifiers also depends on the temperature. For TW amplifiers the main factors which contribute to this are a reduction in the material gain constant B and an increase in the transparency carrier density n₀ when the temperature increases. According to [5.5] for a typical 500 µm length amplifier operating in the wavelength band of 1.5 µm the gain increases with decreasing temperature by about 0.6 dB/K. In addition to the increase in gain when the temperature falls there is also an increase and a wavelength shift in the passband ripple of the amplifier if there is some residual reflectivity. So for phased arrays which need to operate reliably over a range of temperatures (the satellite-based phased array described in Appendix C needs to operate between -15°C and +50°C.)
this temperature dependence has to be compensated for.

One way to do this is to use a Peltier cooler and a feedback circuit. The commercial amplifier SOA3100 from BT&D is packaged with a temperature control mechanism containing the Peltier cooler and thermistor. This way of directly controlling the temperature is good for applications in which electrical power is not expensive to supply (for example in ground based phased arrays etc.) However for space applications where electrical power is expensive this method cannot be used or can be used only if a few amplifiers are present in the array.

An alternative method of controlling the temperature and polarization characteristics of semiconductor laser amplifiers using an all-electronic method has been proposed and demonstrated by Ellis et al [5.25], [5.26], [5.27] and has been briefly described in the previous Section.

For fibre amplifiers, although the characteristics of the amplifiers do not considerably vary with temperature, the output power of the pump source usually does. Again, for applications where electrical power is expensive, this has to be compensated for electronically, without the use of Peltier coolers.

5.2.8. Noise characteristics of optical amplifiers

The equations describing the various noise contributions in semiconductor laser amplifiers have been presented in Chapter 3. These equations have been derived from the mean and variance in the photon number at an amplifier output described by the photon master equation [5.28] and assume a flat frequency response over any reasonable communications bandwidth. Recently a RIN-like behaviour of semiconductor laser amplifiers has been reported [5.29] but the mechanisms and equations for this have not yet been fully studied in the literature. So in this analysis only the conventional noise equations will be used. As can be seen from Chapter 3 there are four main noise contributions to the total noise of the amplifier; the beat noise between signal and spontaneous emission components, the amplified signal shot noise (quantum noise), the beat noise between spontaneous emission components and the spontaneous emission shot noise.

The noise figure $F$ of an optical amplifier (semiconductor or fibre) is defined as the degradation in the $S/N$ ratio before and after amplification

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of a quantum noise limited optical signal. If the signal-spontaneous beat noise is predominant over the spontaneous-spontaneous beat noise then using equation (3.20) of Chapter 3 the noise figure $F$ can be reduced to the simple expression:

$$F = 2 \, n_{sp} \, \xi$$  \hspace{1cm} (5.1)

where $n_{sp}$ is the population inversion parameter of the medium and $\xi$ the excess noise coefficient for the beat noise between signal and spontaneous emission. The best theoretical noise figure which can be achieved for an optical amplifier is 3 dB under complete population inversion of the amplifying material i.e. when $n_{sp} \, \xi = 1$. This 3 dB arises from quantum mechanical considerations, because the uncertainty product for the simultaneous measurement of the input and output quantum limited signals of a noiseless amplifier is twice that of the separate measurement of any of the two signals [5.30].

Erbium doped fibre amplifiers have very low noise figures of the order of 3.2 dB for a co-propagating pump signal and 5.5 dB for a counter-propagating pump signal. [5.31], [5.32]. For semiconductor amplifiers noise figures varying from 5.2 dB for TWA’s [5.33] to 13 dB for FP structures [5.34] have been reported.

5.2.9. Electrical power consumption and efficiency of optical amplifiers

When this Chapter was first written (1990), the optimum semiconductor laser amplifiers had about five times better electrical power efficiencies than the optimum fibre amplifiers. Typical pump power requirements to achieve comparable gains of about 15 dB from fibre-to-fibre were about 70 mA for semiconductor laser amplifiers and 100 mW of optical power for rare earth amplifiers [5.30]. Typical laser diodes needed about 500 mA to produce 100 mW of optical power. However with the design of more efficient fibre structures and better semiconductor laser diodes this situation has changed. In particular reference [5.35] reports the design and construction of a highly efficient fibre amplifier pumped at 980 nm. Fig (5.4) shows the gain characteristics against the driving current to the pump laser diode. In this configuration the voltage to the laser diode was about 2 V. From
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the curve it can be seen that erbium amplifiers can achieve similar electrical power efficiencies as semiconductor laser amplifiers.

5.3. Signals which may require optical amplification in a typical phased array

Having discussed the general characteristics of the two main types of amplifiers, the next step is to determine which signals need optical amplification in an optically controlled phased array antenna. The candidate signals are the following:

5.3.1. R.F. and L.O. signals

These are attractive candidates for optical amplification for two reasons:
i) Since these signals need to be distributed to every element of the array, if a passive optical network is used the intensity of the optical carriers will be reduced by splitting losses. Due to the limitations from thermal noise, only a limited number of elements (usually = 30) can be fed from a single optical source. However if optical amplifiers are used, the splitting losses will be compensated by the gains of the amplifiers and so it may be possible to feed all the elements of the array from a single optical source.

ii) The frequency of these signals is usually in the microwave region. Thus electrical amplifiers and electrical-optical converters (laser diodes, electro-optic modulators or heterodyne systems) are difficult and expensive to construct. In contrast the gain of optical amplifiers is not dependent on the modulation frequency. In fact, as can be seen from Chapter 6, semiconductor laser amplifiers perform better in terms of nonlinearities at high modulation frequencies, since at these frequencies the gain of the amplifier is not modulated by variations of the optical signal.

5.3.2 The return I.F. signal

In this case it is very unlikely that optical amplifiers will offer any advantage. If a separate optical link with its own dedicated laser and
photodiode is used for the transmission of every element signal as suggested in [5.36] then clearly optical amplifiers are not required. This is because the link distances are small and there are no losses due to optical splitters or combiners. If the return signals are combined in the optical domain as discussed in Section 4.3.1 of Chapter 4, then again optical amplifiers are not required since the signal gain increases due to the combination gain of the element signals.

5.3.3. The I.F. transmit signal

If this signal exists in the distribution system, then as with the R.F. and L.O. signals it needs to be distributed to all the elements of the array. So again optical splitting losses will be present, and optical amplifiers will be required if the signal from a single source is to be distributed to all the elements of the array. Since the bandwidth of this signal is usually less than 1 GHz, if semiconductor laser amplifiers are used nonlinearities in the response of the amplifiers may introduce harmonic or intermodulation distortion as discussed in Chapter 5. To avoid this, the input optical signals to the semiconductor laser amplifiers must be of sufficiently low level that the amplifiers do not operate in their saturation region or the modulation indices of the optical signals must be small. However if fibre amplifiers are used in the distribution network then the response will be linear for all frequencies of interest.

5.3.4. The element control signal

This signal, like the R.F. and L.O. has to be distributed from a central point to all the elements of the array. So some designers [5.1], [5.2], [5.3] choose to modulate it onto a carrier of the same frequency as the R.F. and the L.O.. In this case if transmit/receive time requirements permit, the element control signal can be time multiplexed with the R.F. and/or the L.O., and the same photodiode, R.F. amplifiers and filters can be used for its reception at the remote interface. So in this case the same optical amplifiers will be used to amplify this signal as for the R.F. and L.O..

In some military systems where short beam dwell times are required, the bit
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rates are too high for the element control signal to be time multiplexed with the R.F./L.O.. In some cases it is even impossible for one laser to provide this signal to all the array elements. In these circumstances dedicated digital optical networks will have to be used for the transmission of this signal and if a single laser has to feed more than \( \sim \) 30 elements, optical amplifiers may have to be used. The use of optical amplifiers for digital systems has been extensively studied in the literature ([5.5], [5.6], [5.16] and their references), and no further discussion will be included in this thesis.

5.3.5. The element monitoring signals

For these signals as discussed in Section 2.2.2.C of Chapter 2 the bit rates are usually low.

If separate optical links are used for every element to transmit the return IF signal, then the element monitoring signals can be time-multiplexed with it. In this case no optical amplifiers are required. If an optical combining network is used for the return IF signal [5.1], then as discussed in Section 4.3.1 of Chapter 4, there will be some combining losses due to the optical combiners. In this case since only one element may transmit its signal at a time, these combining losses are not offset by the combining gains as in the case of the return IF signal. In the solution of [5.1] it is shown that the element monitoring signals can withstand the losses of a 512 element combining network with the bit error rates at the central photodiode being of an acceptable level. If lasers are used instead of the LED’s proposed in [5.1] (see Section 4.3.1 of Chapter 4) then this number of 512 can be further increased to say 10,000 due to the higher electro-optic conversion efficiencies of the lasers. So it may finally be possible to use a passive combining network only. In the case that the required specifications cannot be achieved with the use of a single passive combining network two solutions can be considered:

i) The use of one optical amplification stage which will also help to improve the S/N ratio performance of the return IF signal.

ii) The use of more than one optical combining networks and the use of a microwave combining network to combine the analogue IF outputs while the
digital outputs, can be sent to the central processor through another
digital network.

5.4. Conclusions

From the above discussions it can be seen that both semiconductor and rare
earth fibre amplifiers are suitable for use in optically controlled phased
array antennas. When this chapter was first written, fibre amplifiers
suffered from higher electrical power consumptions due to the non-optimum
implementations of the time. However the problem has now been solved, and
fibre amplifiers appear to offer superior performance than semiconductor
laser amplifiers for all systems, except from those requiring
opto-electronic integration. For space applications some other potential
advantages of semiconductor amplifiers are their small size, low mass and
perhaps higher reliability. Finally the signals which will benefit from the
use of optical amplifiers are mainly the signals which need to be
distributed from the central processor to the elements of the array. These
are the high frequency R.F. and L.O. signals, the transmit I.F. signal (if
it exists) and the element control signals.

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use some of the information contained in them.
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CHAPTER 6

THE USE OF OPTICAL AMPLIFIERS FOR SIGNAL DISTRIBUTION IN OPTICALLY CONTROLLED PHASED ARRAY ANTENNAS - B) COMPUTER PROGRAM TO MODEL SEMICONDUCTOR LASER AMPLIFIERS UNDER INTENSITY MODULATED SIGNALS

6.1. Introduction

Having discussed the general characteristics of optical amplifiers and established the need for them in the distribution networks of optically controlled phased array antennas the next stage is to make a computer program to model the behaviour of semiconductor laser amplifiers with intensity modulated optical input signals. In the literature, whilst much work has been done on the modelling of semiconductor laser amplifiers, [6.1] - [6.6], little attention has been given so far to performance with RF intensity modulated inputs [6.7].

The chapter is organized as follows. First a description of the model is presented including its structure, the equations used, and the improvements of equations and algorithms over other similar models. Then a brief analysis of the rate equations is presented to predict the performance of an amplifier as the modulation frequency of the input optical signal varies. Finally the results of the program are presented and compared with the results of the rate equations analysis and with experimental measurements reported in the literature.

Although the primary purpose of the program is to handle sinusoidally modulated signals, it is a general dynamic amplifier model which with little change can be used to predict the performance of semiconductor laser amplifiers under other input conditions such as pulse and square waves.

6.2. Description of the model

6.2.1. Static part of the program

The static part of the program is the part where the power of the input optical signal to the amplifier is constant with time. Fig (6.1) shows a
Fig (6.1) Block diagram for the static part of the program.
flow diagram for the static part of the program.
The optical signal within the amplifier is represented by a normalized field $f$ defined by:

$$f = \frac{n_g P}{h v c A} \exp i(2\pi v t - \beta z)$$  \hspace{1cm} (6.1)

where $P$ is the optical power, $v$ is the optical frequency, $n_g$ is the group refractive index (defined in Appendix D), $A$ is the cross sectional area of the input signal, $c$ is the speed of light in vacuo and $h$ is Planck’s constant. The propagation constant $\beta$ is given by:

$$\beta = \frac{2 \pi n_e v}{c}$$  \hspace{1cm} (6.2)

where $n_e$ is the effective refractive index of the gain medium (defined in Appendix D).

Fig (6.2) shows the signal amplitudes at the facets of the amplifier. The (+) and (-) superscripts represent optical signals travelling in the positive and negative directions respectively. The numbers in parentheses represent the spatial position of the signal if the left facet of the amplifier is assumed to be the origin. The cavity is of length $L$.

In the program equation (6.1) is used to calculate the normalized signal amplitude $a^2(L)$ of the incoming signal to the amplifier from the optical power of this signal.

The signal amplitudes inside the amplifier obey the following differential equation:

$$\frac{\partial^2 f(z,t)}{\partial z^2} = \left[ \frac{n_e + \frac{i n_e [g(z,t) - a_{sc}]}{2\beta}}{c^2} \right]^2 \frac{\partial^2 f(z,t)}{\partial t^2}$$  \hspace{1cm} (6.3)

Where $g(z,t)$ is the optical gain coefficient and $a_{sc}$ is the optical absorption coefficient. The above equation is derived from Maxwell’s equations but the refractive index is complex to account of the gain $g(z,t)$ and attenuation $a_{sc}$ of the material. From this equation and for the static
Fig (6.2) Input and output signals to a laser amplifier.

$$\Delta \lambda_{1/2} = \frac{\lambda_0}{Q - \frac{1}{4 Q}} = 8.3 \text{nm}$$

$$\lambda_0 = 830 \text{nm}$$

Fig (6.3) Normalized Lorentzian response for the gain curve of the modelled laser.
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case the following equations can be derived [6.3]:

\[
\frac{d f^+(z)}{d z} = i\beta f^+(z) + 0.5(g(x)-a_{sc})f^+(z) \tag{6.4}
\]

\[
\frac{d f^-(z)}{d z} = i\beta f^-(z) - 0.5(g(x)-a_{sc})f^-(z) \tag{6.5}
\]

whose solutions are:

\[
f^+(z) = f^+(0)\exp(-i\beta z)\exp\{0.5\int_0^z [g(x)-a_{sc}]dx\} \tag{6.6}
\]

\[
f^-(z) = f^-(L)\exp[-i\beta(L-z)]\exp\{0.5\int_z^L [g(x)-a_{sc}]dx\} \tag{6.7}
\]

The amplifier cavity is divided into spatial mesh points with separation \(\Delta L\). The signal amplitudes of the forward and backward signal amplitudes waves \(f^+(z)\) and \(f^-(z)\) are calculated at every mesh point using equations (6.6) and (6.7).

To calculate \(f^+(0)\) and \(f^-(L)\) Marcuse [6.3] has shown that in the static case:

\[
f^+(0) = \frac{(1 - R_1)^{0.5} a^+(0)}{1 - (R_1R_2)^{0.5}G_s \exp(-2i\beta L)} \tag{6.8}
\]

and:

\[
f^-(L) = \frac{[G_s R_2 (1-R_1)]^{0.5} a^+(0) \exp(-i\beta L)}{1 - (R_1R_2)^{0.5}G_s \exp(-2i\beta L)} \tag{6.9}
\]

Where \(a^+(0)\) is the normalized field applied to the amplifier (calculated using equation (6.1) from the input power) and \(R_1\) and \(R_2\) are the mirror reflectivities of the laser. In the above equations \(G_s\) is given by:

\[
G_s = \exp\{\int_0^L [g(x)-a_{sc}]dx\} \tag{6.10}
\]

The next step is to calculate the gains for every cavity mesh point. To
calculate the peak gains, the equation

\[ g_p(z) = \Gamma B(n_n - n_{tr}) \]  \hspace{1cm} (6.11)

is used, where \( B \) is the gain constant of the amplifier, \( \Gamma \) is the confinement factor and \( n_{tr} \) the transparency carrier density. From that the gain for the specific input frequency of the signal is calculated using the Lorentzian approximation [6.8] illustrated in Fig (6.3):

\[ g = \frac{g_p \omega_o^2}{((\omega - \omega_o)^2 + (\omega_o/2Q)^2)4Q^2} \]  \hspace{1cm} (6.12)

This equation is valid close to the peak of the gain curve around which most of the laser power is produced. In the equation \( \omega \) is the optical frequency of the signal, \( \omega_o \) the optical frequency where the amplifier gain is maximum and \( Q \) the quality factor of this Lorentzian response.

Fig (6.1) shows the scheme of iteration for the equations. Initial values are required for the amplifier gain and the carrier density. In the steady state:

\[ R_1 R_2 G_p \leq 1. \]  \hspace{1cm} (6.13)

where \( G_p \) is the round trip power gain. Also for finite current injection \( g > 0 \) since some population inversion then exists. Thus, averaging the above equations, a useful starting condition is:

\[ g = \frac{\ln (1/R_1 R_2)}{4 L} + \frac{a_{sc}}{2} \]  \hspace{1cm} (6.14)

The initial carrier density is set to give a single pass gain of unity

\[ n = n_{tr} + \frac{a_{sc}}{\Gamma B} \]  \hspace{1cm} (6.15)

At each iteration the gain is checked to ensure that the single pass gain,
neglecting material attenuation, is greater than unity and that the round trip gain is within the limits for stability:

\[
\frac{1}{R_1 R_2} > \exp\{2\int_0^1 [g(x) - a_{sc}']dx\} \quad (6.16)
\]

So that

\[
\frac{1}{R_1 R_2} > \exp\{2\int_0^1 [g(x) - a_{sc}']dx\} > \exp(-2L_{sc}) \quad (6.17)
\]

If this condition is not met, the gain at every cavity point is set to the average of the gain of the previous iteration and the gain of this iteration. Then iteration continues until the total gain falls within the correct limits and convergence is obtained. For cases where the carrier density oscillates between values for successive iterations or diverges an averaging procedure is used to ensure convergence.

The carrier density at each mesh point is found from the rate equation:

\[
\frac{dn(z,t)}{dt} = \frac{J}{(ed)} - \frac{n(z,t)}{\tau_{sp}} - \frac{(c/n_e)}{S_t} \quad (6.18)
\]

Where \(J\) is the current density, \(e\) is the electronic charge, \(d\) is the active layer thickness and \(S_t\) is the total photon density due to the signal and spontaneous emission.

For the static case \(dn/dt = 0\) giving:

\[
n(z) = \frac{\tau_{sp}}{J/(ed)} - \frac{(c/n_e)}{\sum S_t} \quad (6.19)
\]

In the above equation the term \(\sum S_t\) is calculated by adding the contribution of photon density times gain due to spontaneous emission and due to the signal:

\[
\sum S_t g = \sum S_{sp} g_{av} + (|f^+|^2 + |f^-|^2) g \quad (6.20)
\]

The next step is to calculate the photon density gain product \((S_{sp} g)\) due to spontaneous emission. Marcuse [6.3] calculates the spontaneous photon density at every point in the cavity. However this would be very
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computation intensive, especially in a time varying program. As the spatial position of spontaneous photon density is not important for our purpose, the approximation for the photon density due to spontaneous emission of equation (6.19) of Adams [6.1] is used i.e.

$$\Sigma S_{sp} g_{av} = \left[ \left( e^{g_{av}L} - 1 \right) (1 - R_2) (1 + R_1) e^{g_{av}L} \right] + \left( 1 - R_1 \right) (1 + R_2) e^{g_{av}L} \right] \frac{R_{sp} n \beta_{(eff)} - 2}{c} \frac{g_{av} L}{g_{av} L (1 - R_1 R_2 e^{2g_{av}L})}$$

(6.21)

where the summation is done over all the modes of the device through the use of an effective spontaneous emission factor $\beta_{(eff)}$ which is the sum of the spontaneous emission factors of all the modes of the device. In this model the effective spontaneous emission factor is entered as a constant.

The gain $g_{av}$ is calculated as the average of all the gains in the cavity:

$$g_{av} = \frac{1}{L} \int_0^L g(z) \, dz \quad \text{where} \quad g(z) = g[n(z)] \quad (6.22)$$

The $R_{sp}$ factor, is considered to be:

$$R_{sp} = \frac{n}{\tau_{sp}} \quad (6.23)$$

where $n$ is the carrier density and $\tau_{sp}$ is the electron lifetime due to spontaneous photon emission).

At the end of the program the output signal amplitude $c^*(L)$ (see Fig (6.2)) can be calculated from the formula:

$$c^*(L) = (1 - R_2)^{1/2} \, f^*(L) \quad (6.24)$$

In the above equation $(1 - R_2)^{1/2}$ represents the transmission coefficient of the signal amplitude. Finally from $c^*(L)$ the output optical power of the amplifier can be calculated using the equation:
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\[ P = \frac{|e^*(L)|^2 h v_c A}{n_g} \]  \hspace{1cm} (6.25)

which is derived from equation (6.1).

6.2.2. Dynamic part of the program

The dynamic part of the program is the part in which the input optical power signal is no longer constant but varies with time. In this work we are particularly interested in sinusoidal modulation of the input power. Two representations of the input signal are used:

a) An intensity modulated input signal with equation:

\[ P = P_o (1 + \delta \cos 2\pi f_m t) \]  \hspace{1cm} (6.26)

where \( P_o \) is the average optical power, \( \delta \) is the modulation index and \( f_m \) is the modulation frequency.

or

b) A heterodyned signal produced from the combination of two (or more) optical sources with field equations:

\[ f_1 = K_1 \exp(i \nu_1 t) \]  \hspace{1cm} (6.27)

\[ f_2 = K_2 \exp(i \nu_2 t) \]  \hspace{1cm} (6.28)

where \( \nu_1 \) and \( \nu_2 \) are the optical frequencies whose difference is in the microwave range and \( K_1 \) and \( K_2 \) are the peak normalized field amplitudes of the signals. It can be shown that such a signal produced by two such sources of equal power is equivalent to an amplitude modulated signal whose modulation index \( \delta \) is 1 (100% modulated) and whose modulation frequency is the difference between the optical frequencies of the two signals.

In Fig (6.4) a block diagram for the dynamic part of the program is presented. In this part the same basic equations are used as for the static part but modified to account for the fact that \( \frac{d}{dt} \) is no longer zero.

In the case of a modulated input signal equation (6.26) is used as the
Fig (6.4) Block diagram for the dynamic part of the program.
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input equation of the signal and from this the normalized field amplitude of the input signal can be calculated using equation (6.1):

\[ a^\dagger(0,t) = \exp i2\pi vt \frac{n_g P^o (1 + \delta\cos2\pi f_m t)}{h\nu A} \]  

(6.29)

This equation is used to calculate the input signal amplitude in the case of a directly modulated input optical signal.

In the case of two (or more) signals heterodyned at their difference frequency, equation (6.1) is used to calculate the input signal amplitude \( a^\dagger(z,t) \) to the amplifier for every optical component of the signal. Then these input signal amplitudes are added vectorially to produce the resulting signal.

The next step is to calculate the signal amplitudes at every point in the cavity. When the modulation frequency is small (the period of the modulation signal is much bigger than the round trip time of light inside the amplifier) then equations (6.6) to (6.10) are used to calculate the field amplitude at every point in the cavity.

Also the program calculates the timeperiod of the modulation signal:

\[ T_m = \frac{1}{f_m} \]  

(6.30)

and divides it into a number of timesteps which can be defined by the user. Equations (6.6) to (6.10) are used at every timestep to calculate the field amplitudes.

In the case of a heterodyned input signal with the period of the heterodyne frequency being much greater than the round trip time of light inside the amplifier, the same equations are used as in the direct modulated case.

In the case of a modulated or heterodyned input signal where the period of the modulation or heterodyne frequency is of the order of or smaller than the round trip time of light inside the amplifier, equations (6.6) to (6.10) cannot be used any more. This is because to derive equations (6.8) and (6.9) it is assumed that during the round trip time of the signal inside the amplifier the amplitude of the input signal remains the same. So equation (6.3) is solved directly and the solutions are:
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\[ f^+(z+\Delta L, t+\Delta t) = f^+(z, t) \exp\left(\text{i}[-\beta \Delta L + \omega \Delta t]\right) \exp\left\{0.5 \int_{z}^{z+\Delta L} \left[g_s(x, t) - a_{sc}\right] \, dx\right\} \] (6.31)

and

\[ f^-(z-\Delta L, t+\Delta t) = f^-(z, t) \exp\left(\text{i}[-\beta \Delta L + \omega \Delta t]\right) \exp\left\{0.5 \int_{z}^{z-\Delta L} \left[g_s(x, t) - a_{sc}\right] \, dx\right\} \] (6.32)

The following boundary conditions should be satisfied at the ends of the cavity:

\[ f^+(0, t) = t_1 a^+(0, t) + r_1 f^-(0, t) \] (6.33)

\[ f(L, t) = r_2 f^+(L, t) \] (6.34)

And for output signals from the amplifier:

\[ a^+(0, t) = r_1 a^+(0, t) + t_1 f^-(0, t) \] (6.35)

\[ c^+(L, t) = t_2 f^+(L, t) \] (6.36)

where:

\[ r_1 = R_1^{1/2} \] (6.37)

\[ r_2 = R_2^{1/2} \] (6.38)

\[ t_1 = (1 - R_1)^{1/2} \] (6.39)

\[ t_2 = (1 - R_2)^{1/2} \] (6.40)

are the reflection and transmission coefficients.

The period of the modulated or heterodyned signal is again divided into a number of time steps. Equations (6.31) - (6.40) are applied at each time step for all mesh points in the cavity.

Both the number of time steps for one period of the signal and the number of mesh points in the cavity are specified initially by the user. This has the advantage of flexibility but presents a problem in the structure of the
program. This is that if $\Delta L$ is the distance between two consecutive points in the cavity and $\Delta t$ is the time difference between timesteps then $\Delta t$ is not going to be equal to $v \Delta L$ where $v$ is the velocity of light in the medium. This implies that in a time $\Delta t$ the light has travelled a distance $k \Delta L$ where $k$ is not necessarily an integer.

Other designers of such computer programs [6.4] - [6.6] have avoided this problem altogether by making $\Delta t = \Delta L/v$.

In this computer program the integer and fractional part of $k$ are taken and defined as $\epsilon$ and $\eta$ respectively. Then the light signal can be propagated $\epsilon$ times along the cavity using equations (6.31) to (6.36) as illustrated in Fig (6.5). After that the light signal is propagated back in time by an amount $(1-\eta)$ and then moves forward by $\Delta L$ using equations (6.31) to (6.36). The equations to move the signal backwards in time by an amount $\Delta \tau$ are:

$$f^+(t-\Delta \tau)=f^+(t)\exp(-i\omega \Delta \tau)/\exp\{0.5\int_{z}^{z+\Delta l}[g(x,t)-a_{sc}]dx\} \quad (6.41)$$

and $$f^-(t-\Delta \tau)=f^-(t)\exp(-i\omega \Delta \tau)/\exp\{0.5\int_{z}^{z-\Delta l}[g(x,t)-a_{sc}]dx\} \quad (6.42)$$

Where $Dl = (1-\eta)\Delta \tau$

thus completing the equations for the calculation of the signal amplitudes.

For the calculation of the new carrier density the static rate equation (6.19) cannot be used any more since now $dn/dt \neq 0$. Instead in the rate equation:

$$dn(z,t)/dt = J/(ed) - n(z,t)/\tau_{sp} - (c/n_e)\sum S_{i,g} \quad (6.18)$$

for each time step $dn(z,t)/dt$ becomes $(n^{(m)}_m - n^{(m-1)}_{m-1})/\Delta t$, where $\Delta t$ is the time difference between the two points in time and $n^{(m)}_m$ and $n^{(m-1)}_{m-1}$ are the values of the carrier density at these two points in time. Also the carrier density $n$ in equation (6.18) is taken as the average of the $n^{(m)}_m$ and the $n^{(m-1)}_{m-1}$ i.e. $n = (n^{(m)}_m + n^{(m-1)}_{m-1})/2$.

With these substitutions the rate equation becomes:
Fig (6.5) Graphical representation of method of translating the signal in space and time along the cavity when both the number of model time steps and the number of model points along the cavity have been specified by the user.
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\[
\frac{\left[ n_{(m-1)} \left( \frac{1}{\Delta t} - \frac{1}{2\tau_s} \right) \right] + \frac{J}{ed} - \frac{c}{n_e \sum_s S_s g}}{\left( \frac{1}{\Delta t} + \frac{1}{2\tau_s} \right)}
\]

(6.43)

To reduce computing time, equation (6.21) is used to calculate spontaneous emission. Finally, as in the static part, equation (6.20) is used to calculate photon density times gain and (6.11) and (6.12) to calculate gain. To calculate the output optical signal power from the amplifier, \( c^*(L) \) from equation (6.24) is substituted into the equation (6.25).

6.3. Input & output of data to the program

Because of the different types of input and output data required, three versions of the program have been constructed. The first is a static program (STAT), accepting only an unmodulated optical input signal. The second version (MODUL) accepts a sinusoidally modulated optical input signal, the depth and frequency of modulation being determined by the user. The third version (HETER) accepts two or more unmodulated optical signals of slightly different wavelengths, thus modelling inputs from wavelength division multiplex optical systems.

a) Input of data.

The input of data to the 3 versions of the program (STAT, MODUL and HETER) consists of 2 parts which are stored in 2 separate files: i) The parameters of the laser. A set of numerical values for these parameters for the HLP 1400 GaAs/AlGaAs semiconductor laser are given in Appendix D. ii) The operating conditions for the laser. In the static (STAT) version these consist of the current to the laser, the input optical power and the wavelength of the optical signal. In the modulation (MODUL) version these consist of the average input optical power, the microwave modulation frequency and the modulation depth of the input optical signal, the centre wavelength of the input optical signal and the current to the laser. In the heterodyne (HETER) version these consist of the optical powers and
wavelengths of the two or more input optical signals and the current to the laser.

b) Output of data.

A commercially available spreadsheet - Lotus 123 is used for output. In the STAT version of the program, a file with the spatial variation of the signal amplitude inside the amplifier is imported into Lotus-123 together with the input data files. From these the spatial variation of optical power inside the amplifier together with the output optical power is calculated and displayed graphically. In the MODUL and HETER versions of the program the variation of output power with time is calculated and displayed graphically.

Another advantage of using a spreadsheet to output the results, is the flexibility that it gives the user to make his own calculations and calculate other parameters from the data already existing in the spreadsheet.

6.4. Analytic treatment of the carrier density rate equation for an intensity modulated signal

Before the results of the computer program are presented, an analytic treatment of the effects of an intensity modulated input signal on amplifier operation will be given. This will then be compared with the results from the program. Consider the rate equation:

\[
\frac{dn}{dt} = J/(ed) - n/\tau_p - (c/n_e) \sum S_i g
\]  

(6.18)

For a sinusoidally modulated input giving a photon density of the form \( a(1 + \delta \cos \omega \ t) \) the rate equation becomes:

\[
\frac{dn}{dt} = J/(ed) - n/\tau_p - (c/n_e) \ a(1 + \delta \cos \omega \ t) g
\]  

(6.44)

Solving this equation analytically gives:
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\[ n = \tau_{sp} \left( \frac{J}{ed} - \frac{cga}{n_e} \right) - \frac{cga}{n_e} \left( \frac{[1/\tau_{sp}]}{\omega_m^2 + [1/\tau_{sp}]^2} \cos \omega_m t + \omega_m \sin \omega_m t \right) \]

+ K \exp(-t/\tau_{sp}) \quad (6.45)

Where K is a constant to be determined by the initial conditions of the problem, but since the term \( K \exp(-t/\tau_{sp}) \) is just a decaying exponential it can be ignored in the case of the steady state solution for an intensity modulated signal. Considering the first term of the above equation it can be seen that this is the solution of the carrier density for a constant input optical signal. The second term describes the response to the intensity modulated input. For \( \omega_m >> 1/\tau_{sp} \) the term tends to zero, so that at high modulation frequencies the carrier density and hence the amplification only depend on the average value of the input signal. For \( \omega_m \rightarrow 0 \) the solution becomes:

\[ n = \tau_{sp} \left[ \frac{J}{ed} - \frac{cga(1 + \delta \cos \omega_m t)}{n_e} \right] \quad (6.46) \]

showing that the carrier density and the amplification are modulated at every time instant to the values that they would take if a constant signal was fed into the amplifier with a value equal to the value of the input signal at this time instant. This modulation of the amplification at low modulation frequencies leads to considerable distortion of the input signal.

Fig (6.6) plots the amplitude and the phase of the frequency dependent term:

\[ \frac{[1/\tau_{sp}]}{\omega_m^2 + [1/\tau_{sp}]^2} \cos \omega_m t + \omega_m \sin \omega_m t \quad (6.47) \]

for a typical value of \( \tau_{sp} \) for GaAs of 4 ns. The term is seen to be significant below a frequency of 1 GHz in good agreement with other experimental and theoretical results [6.9], [6.10], [6.11], [6.12]. A more detailed comparison between the results of the program, the above results
Fig (6.6) Plots of the amplitude and the phase of the term:

\[
\frac{\left[\frac{1}{\tau}\right] \cos \omega t + \omega_m \sin \omega t}{\omega_m^2 + \left[\frac{1}{\tau}\right]^2}
\]
and results from other papers are given in section 9.

6.5. Program performance

The program is written in FORTRAN and all the arithmetic operations are performed with double precision. Due to the boundary conditions ((6.17) and the conditions for oscillation and divergence) the static program converges quickly. Operating on an 8 MHz PC compatible computer the static program requires an average time of less than half a minute to reach convergence. The dynamic part of the program is usually slower depending on the number of time steps and the modulation frequency. Typical times vary from about two minutes to ten minutes.

Ten mesh points were used for most of the calculations but this number can be defined by the user. The run time increases proportionately with the number of mesh points and the number used in the calculations was a compromise between run time and accuracy.

6.6. Static results

The laser selected for modelling studies was a commercially available GaAs/AlGaAs CSP device (Hitachi HLP 1400). A set of typical static results is briefly presented. This includes the amplification of the amplifier against the input power of the signal for a travelling wave amplifier with $R_1 = R_2 = 0.01$, Fig (6.7) and the amplification of the amplifier against the wavelength of the optical signal for a Fabry-Perot amplifier with $R_1 = R_2 = 0.3$, Fig (6.8). The curves are in good agreement with the experimental results of [6.13] where a 300 μm AlGaAs amplifier operating at 843 nm is examined. In particular the gain curves have approximately the same 3 dB saturation points (agreement of the input optical powers within about 3 dB.) and the tuning curves have the same shape and periodicity (agreement in the distance between two modes of about 10%).

6.7. Dynamic results

The dynamic model outputs the values of the input signal at the one end of the amplifier and the output signal at the other end to a file. At low
Fig (6.7) Travelling wave amplifier amplification as a function of input power for 3 different values of current. Mirror reflectivities: 0.01.

Fig (6.8) Amplification against wavelength for a Fabry-Perot laser amplifier. (Under these conditions the amplifier becomes unstable at about 57.5 mA). The current to the amplifier is 57 mA, the mirror reflectivities are 0.3 and the input power $1 \times 10^6$ W.
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modulation frequencies an amplification versus time curve can be constructed by dividing the output signal over the input at each point in time. However at higher frequencies this is not possible since the output signal has a phase offset from the input signal resulting from the propagation time delay in the amplifier cavity. A solution is to find the single pass amplification from the equation:

\[ G_s = \exp\{\int_0^1 g_e(x) - a_{sc} \, dx\} \quad (6.10) \]

And this represents the average amplification within the amplifier cavity. Fig (6.9) shows the amplification variation over one cycle of the modulation frequency for various modulation frequencies (with the input signal 100% sinusoidally modulated). At present the model does not include chirp effects; if the reflection coefficient at the facets is significant additional distortion would result from intensity dependent modulation of the cavity resonance.

Fig (6.10) presents the output optical signal powers corresponding to the gain curves of Fig (6.9). The curves for 1GHz and 10 GHz modulation frequencies are nearly sinusoidal while the ones for 10 and 100 MHz are heavily distorted. Fig (6.11) shows the corresponding spectra for the signals of Fig (6.10). The harmonic content is seen to decrease for high modulation frequencies. In this example the input signals and the amplifier current were chosen to emphasize these nonlinearities. If the modulation depth is small or the input signal power small these nonlinearities are greatly reduced in accordance with the predictions of Fig (6.7).

Another interesting point which these results illustrate is the amount of cross talk between two separate frequency channels passing through the same amplifier. Even in the case of very small signal modulation depth the two optical carrier frequencies form effectively a heterodyned modulation frequency with an 100% modulation depth. If the optical frequencies of the two channels are closer to one another than say 5 GHz the amplification is modulated by the frequency difference, as shown in Fig (6.9) and distortion occurs. However for higher separations this amplification modulation becomes negligible. This can impose a limit on the minimum channel separation.
Fig (6.9) Amplification variation with time for one period of the signal at various modulation frequencies. Optical input signal power: $3 \times 10^{-5}$ W, 100% sinusoidally modulated. Amplifier bias current: 82 mA. Mirror reflectivities: 0.01.

Fig (6.10) Modelled output signal powers for conditions of Fig (6.9).
Fig (6.11) Fourier spectra of the output signals of Fig (6.10).

Fig (6.12) Peak to peak variation of the carrier density as a function of frequency from a) the computer model b) the analytical model of section 5. The parameters of the amplifier and the input optical signal are the same with those of Fig (6.9).
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6.8. Comparison of the dynamic results of the program with the analytical results of section 5 and other results.

In Fig (6.12) the peak to peak carrier density variation is plotted versus the modulation frequency of the input signal. The upper curve represents the results from the computer model with the input signal and other amplifier parameters being the same as those of Fig (6.9). The lower curve shows the corresponding results from the analytical method of section 5. These use the average values of gain and photon density obtained from the computer model. As can be seen the shapes of the two curves are the same while the carrier density variation obtained from the computer model is higher than the analytical one. This is believed to be due to the assumption that the gain $g$ is not a function of the carrier density $n$ when deriving the analytical model.

In Fig (6.13) the normalized second harmonic divided by the fundamental carrier is plotted as a function of the frequency of separation of two optical signals passing through the amplifier. The amplifier operates at 1300 nm and the spontaneous lifetime $\tau_p$ is 280 ps. On the same curve the analytical and experimental results of Darce et al [6.9], [6.11] are also plotted. Excellent agreement can be seen between the results, confirming the accuracy of the analytical model.

6.9. The use of optical amplifiers for signal distribution in phased array systems

The modelling work carried out was primarily directed towards the use of semiconductor laser amplifiers in phased array signal distribution systems. The most important signal to be distributed optically in phased array systems is the high frequency carrier (R.F.) signal. This is usually in the microwave range with a frequency greater than 1 GHz. The modulation depth of the optical signal must be high. In the case of direct modulation of the laser this is because high conversion efficiency from the electrical to the optical domain and vice versa is desired. In the case of the heterodyne of 2 lasers with equal optical power [6.14] analysis shows that this is equivalent to sinusoidal modulation with 100% depth.

According to the results of this program under the above operating
Fig (6.13) Normalized second harmonic divided by the fundamental carrier as a function of the frequency of separation of two optical signals passing through the amplifier. The solid line shows the results from the computer model, the dashed lines the analytical results from [6.9] and the squares the experimental results from [6.11]. The spontaneous lifetime $\tau_{sp}$ and the wavelength of operation of the amplifier are 280 ps and 1300 nm in accordance with [6.11].
conditions semiconductor laser amplifiers present a suitable solution for the distribution of this signal. For distribution of other lower frequency signals such as the intermediate frequency (IF) the results suggest that low modulation depths must be used or else severe distortion will occur.

6.10. Conclusions

A new computer model has been presented to model the dynamic behaviour of a laser amplifier. The main novel points which distinguish it from other models are:
Equations (6.14) and (6.15) which give a very good starting point for gains and carrier density in the iteration.
Equation (6.17) which makes the static program converge quickly.
Equation (6.43) which is an equation to model the carrier density under dynamic conditions.
Equations (6.31) and (6.32) which represent the signal propagation equations under dynamic conditions.
Equations (6.41) and (6.42) which are the equations to propagate the signal backwards so that the number of cavity mesh points and time steps do not need to be interrelated.

The results for an intensity modulated signal show that the amplification variation decreases with increasing frequency and predict the existence of nonlinearities at low modulation frequencies and also cross talk when the channel separation frequencies are small.
For signal distribution in phased array systems the use of semiconductor laser amplifiers is recommended for the high frequency carrier signal but not for the I.F. signals if direct intensity modulation is to be used. An alternative solution is to use erbium doped fibre amplifiers; experimental results suggest that the frequency below which distortions occur is less than 100 KHz [6.15], probably around $\approx 0.5$ KHz due to the very long flourescence lifetime ($\tau_{sp} = 15$ms).
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6.11. References for Chapter 6


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CHAPTER 7

THE USE OF OPTICAL AMPLIFIERS FOR SIGNAL DISTRIBUTION IN OPTICALLY CONTROLLED PHASED ARRAY ANTENNAS - C) THEORETICAL ANALYSIS OF THE USE OF SEMICONDUCTOR LASER AMPLIFIERS IN THE FORWARD LINK OF A SATELLITE BASED PHASED ARRAY

7.1. Introduction

Having discussed the general characteristics of the two types of optical amplifiers in Chapter 5 and the performance of semiconductor laser amplifiers under intensity modulated inputs in Chapter 6, the use of semiconductor laser amplifiers will be analyzed with reference to a typical system. The analysis of a system using fibre amplifiers is similar except that their noise figures are usually lower. The system to be analyzed is the forward link of a satellite based phased array developed without the use of optical amplifiers by Nawaz et al [7.1], [7.2], [7.3] and analyzed in Appendix C and Chapter 4. The analysis of other types of arrays is usually similar with the main differences usually arising from individual system specifications such as the number of elements and the operating frequency. To perform the analysis a computer model is used whose equations have been presented in Chapter 3

7.2. Characteristics of the optical amplifier used in the analysis

The amplifier chosen for the analysis is a polarization insensitive structure constructed by British Telecom Laboratories [7.4]. It is not yet commercially available, though it is expected that it soon will be for both 1500 nm and for 1300 nm wavelength operation. The commercially available amplifier SOA1100/SOA3100 was not chosen mainly due to its polarization dependent gain characteristics which introduce uncertainties in the calculations. Fig (7.1) shows the internal gain versus current characteristics of the amplifier. Assuming a reasonable coupling loss of \( \approx 5 \) dB to each fibre, the fibre-to-fibre gain of the structure should be
Fig (7.1) Internal gain versus current for the polarization insensitive amplifier used in the analysis. (After [7.4]).

Fig (7.2) Internal gain versus input optical power for the polarization insensitive amplifier used in the analysis. (After [7.4]).
about 10 dB less than the values shown in Fig (7.1). Fig (7.2) shows the internal gain versus facet output power for the above structure for two values of current. In the analysis the upper curve will be used, for which the current is about 115 mA. The model parameters of the amplifiers, since they are not given in [7.4] are taken from an amplifier of similar structure analyzed by Lowery [7.5]. These parameters are:

\[
\begin{align*}
L &= \text{cavity length} = 190 \mu\text{m} \\
ne &= \text{effective index} = 3.4 \\
a_{sc} &= \text{loss coefficient of the guided mode} = 1500 \text{ m}^{-1} \\
tr &= \text{transparency carrier density} = 9 \times 10^{23} \text{ m}^{-3} \\
\Gamma &= \text{confinement factor} = 0.3 \\
B &= \text{gain constant} = 2.7 \times 10^{20} \text{ m}^2 \\
Q &= \text{gain curve Q factor} = 22 \\
\text{Eff}_{ia} &= \text{input coupling efficiency} = 5 \text{ dB} \\
\text{Eff}_{out} &= \text{output coupling efficiency} = 5 \text{ dB}
\end{align*}
\]

7.3. General analysis and design curves of the optical link

Fig (7.3) shows the relative contributions of the various noise sources in an optical link containing one laser amplifier situated in the middle of the link. For low values of attenuation less than 18 dB the laser RIN noise is the dominant source. For medium values of attenuation between 18 and 30 dB the internal amplifier noise dominates. Finally for values of attenuation over 30 dB the link becomes thermal noise limited.

The internal amplifier noise has its origins in 5 sources [7.6]. Fig (7.4) shows the relative contribution of these sources in the total S/N ratio of the link. For all values of optical attenuation the most important contribution is the signal-spontaneous emission noise. After that, for values of attenuation less than 27 dB comes the quantum noise and for high values of attenuation the beat noise between spontaneous emission components. Finally the least important contribution is the shot-spontaneous emission noise.

Fig (7.5) shows the S/N ratio versus total optical attenuation of the link (not including the gain due to the amplifiers using a) no optical amplifiers b) one optical amplifier and c) two optical amplifiers. In these
Fig (7.3) Modelled S/N ratio of the link using one semiconductor laser amplifier versus passive optical attenuation. The various noise contributions can be seen.

Fig (7.4) Modelled S/N ratio of the link using one semiconductor laser amplifier versus passive optical attenuation. The various noise contributions due to the laser amplifier can be seen.
Fig (7.5) Modelled S/N ratio versus passive optical attenuation for the link using a) no semiconductor laser amplifiers b) one semiconductor laser amplifier and c) two semiconductor laser amplifiers.

Fig (7.6) R.F. link loss (from the input to the laser to the load resistance of the photodiode) versus passive optical attenuation for links containing one to four optical amplifiers.
curves the attenuation between successive optical components was kept the same. For example in the curve with two amplifiers, if the total passive optical attenuation is $A$, this attenuation is distributed between successive optical components as follows:

\[
\text{LASER ATTEN. 1st AMPLIF. = } A/3 \quad \text{ATTEN. 2nd AMPLIF. = } A/3 \quad \text{ATTEN. PHOTOD. = } A/3
\]

It can be seen that using either one or two amplifier stages all the 512 elements of the array can be fed.

Fig (7.6) shows the R.F. link loss (from the input of the laser to the load resistance of the photodiode) versus the total passive optical link attenuation for links containing zero to four optical amplifiers. Again in these curves the optical attenuation between successive optical components was kept the same. It can be seen that for low values of optical attenuation and for more than two amplifiers net R.F. link gains can be obtained. Also for low values of attenuation the extra R.F. gain obtained by placing an additional optical amplifier in the link is reduced as more amplifiers enter the link. This is due to the gain saturation of the amplifiers at high input optical powers. However as the attenuation of the optical link increases the curves diverge from one another increasing the gain contributions from individual amplifier stages.

One of the problems in designing optical networks with laser amplifiers is to determine where to put the amplifiers inside the network in order to achieve the maximum advantage in terms of S/N ratio and R.F. link loss. For example assume there is an optical link with a certain optical loss and it is required to incorporate an amplifier stage. Where should this amplifier stage be inserted to bring the maximum advantage? The problem has been analyzed using the computer model described in Chapter 3 and Fig (7.7) and Fig (7.8) show the output S/N ratio and R.F. link loss as a function of the percentage of the optical attenuation in dB of the link before the optical amplifier. The four curves represent four different total optical link attenuations. It can be seen that the optimum S/N ratio can be obtained if the amplifier is placed somewhere in the middle of the link. This is because when the amplifier is placed in the beginning of the link it
Fig (7.7) Output S/N ratio as a function of the position of the amplifier inside the optical link. (% attenuation of the link in dB before the amplifier) for various values of total link attenuation.

Fig (7.8) R.F. link loss (from the input to the laser to the load resistance of the photodiode) as a function of the position of the amplifier inside the optical link. (% attenuation of the link in dB before the amplifier) for various values of total link attenuation.
Chapter 7: The use of optical amplifiers in a typical system

operates in its saturation region and so its gain is less than its maximum gain. When it is placed at the end of the link its internal noise contributions become more important as the input optical signal is weaker. The lowest RF link loss can be obtained when the amplifier works in the unsaturated regime i.e. somewhere between the middle and the end of the link.

When the link has two amplifiers, the problem of where they should be put to obtain the optimum performance becomes more complex. Fig (7.9) and Fig (7.10) show the output S/N ratio and R.F. link loss as a function of the relative positions of the amplifiers inside the link (percentage attenuation in dB before each amplifier). For example if the total attenuation of the link is 100 dB and the configuration is the following:

```
LASER 40 dB 1st 25 dB 2nd 35 dB PHOTOD.
       ATTN. AMPLIF. ATTN. AMPLIF. ATTN.
```

then the percentage attenuation in dB before the 1st amplifier is 40 and the percentage attenuation in dB before the 2nd amplifier is 40 + 25 = 65. The total attenuation of the link modeled in Fig (7.9) and Fig (7.10) is 35.5 dB i.e. the laser-to-element attenuation of a 512 element array distribution system like the one demonstrated by Pescod et al [7.1], [7.2], [7.3] using:

a) 2 X 16 way splitters (2 X 12 dB splitting loss + 2 X 1.5 dB excess loss)
b) 1 X 2 way splitter (3 dB splitting loss + 0.5 dB excess loss)
c) 5 optical connectors (5 X 1 dB loss)

From Fig 7.9 it can be seen that the lowest S/N ratio of 128 dBC Hz is obtained when the percentage attenuation in dB before the first amplifier is 24 and before the second amplifier 62 i.e. the configuration of the link is:

```
LASER 8.5 dB 1st 13.5 dB 2nd 13.5 dB PHOTOD.
       ATTN. AMPLIF. ATTN. AMPLIF. ATTN.
```

However from Fig (7.10) in this configuration the total R.F. link loss is
Fig (7.9) S/N ratio (constant contours) of a link with 2 amplifiers as a function of the position of the amplifiers inside the link (% passive attenuation in dB before each amplifier). The total optical link loss is 35.6 dB.

Fig (7.10) R.F. link loss (constant contours) of a link with 2 amplifiers as a function of the position of the amplifiers inside the link (% passive attenuation in dB before each amplifier). The total optical link loss is 35.6 dB.
about 45 dB which is 17 dB worse than the minimum possible attenuation of 28 dB. To achieve this minimum possible attenuation of 28 dB with the best possible S/N ratio (121 dBC Hz) the percentage attenuation in dB before the first amplifier should be 50 and after the second amplifier 0, i.e. the second amplifier is used as a preamplifier to the photodiode. The configuration is:

```
LASER 17.75 dB 1st ATTEN. AMPLIF. ATTEN. AMPLIF. 2nd PHOTOD.
```

The above link configurations would be the optimum in terms of S/N ratio and R.F. link loss if the link consisted of only one laser and one photodiode. In the case of a tree-type distribution network with one laser and many photodiodes there is a third important factor which has to be taken into account when designing the link and that is the total number of amplifiers required for each network implementation. For example in Fig (7.10) if the two amplifiers are placed in series just after the laser (percentage attenuation in dB before each amplifier = 0), the total number of amplifiers required in a 512 element distribution system is only 2. However if they are placed in series before the photodiode (percentage attenuation in dB before each amplifier = 100) the total number of amplifiers required is $2 \times 512 = 1024$. Fig (7.11) shows the number of amplifiers required if a single amplifying stage is to be inserted in a tree-type network and the amplifiers are at a certain level of attenuation away from the laser. This curve assumes a 2 dB loss (due to connectors) and a 0.72 dB excess loss per bifurcation (These values were chosen to match the characteristics of the satellite-based phased array analyzed in Appendix C).

Fig (7.12) presents the same information as Fig (7.11) but in a form compatible with Fig (7.9) and Fig (7.10). i.e. the $\log_{10}$ of the total number of amplifiers required for the 512 element array for each possible position of the two amplifier stages inside the distribution network. So figures (7.7), (7.8) and (7.11) are the complete design curves for a one-amplifier stage tree-type network and figures (7.9), (7.10) and (7.12) are the complete set of design curves for a two-amplifier stage network.
Fig (7.11) Number of amplifiers required if an amplifier stage is to be inserted at a certain point of a tree-type network, versus total attenuation before this amplifier stage.

Fig (7.12) $\log_{10}$ of total number of amplifiers required for each tree-type network implementation of fig (7.9) and (7.10).
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with 512 elements and a total attenuation of 35.5 dB.

7.4. Three possible network structures for a 512 element array using semiconductor laser amplifiers

Taking into account the above design curves and the technical specifications of the satellite-based phased array described in Chapter 4, three possible network structures have been designed which use semiconductor laser amplifiers.

The first uses one amplifier stage and has the following structure:

In Table 7.1 the signal and noise levels for this network can be seen. Apart from the optical amplifiers, the other components are the same as those used in the proposed distribution network by ref [7.1], [7.2], [7.3]. As can be seen above and in Table 7.1 this solution uses 16 optical amplifiers and all the 512 elements of the array can be fed by using only one laser transmitter as compared with 16 in the Marconi solution. The output S/N ratio is 117 dBC Hz which exceeds the required S/N ratio specifications. The power consumption of the network is about the same as that proposed by Marconi due to the fact that the 16 lasers are effectively replaced by 16 optical amplifiers which have about the same total power consumption. The main advantage of this network as compared to the Marconi solution is that it eliminates the need for a microwave network to distribute the 5 GHz signal to the 16 lasers.

The second proposed network has the following structure:

The signal and noise levels for this network can be seen in Table 7.2. As can be seen the network uses two amplifier stages and again feeds all the
<table>
<thead>
<tr>
<th>SIGNAL</th>
<th>FIRST</th>
<th>ORTEL</th>
<th>2</th>
<th>CORNING</th>
<th>SEMICOND.</th>
<th>CORNING</th>
<th>CORNING</th>
<th>LASERTRON</th>
<th>FIRST</th>
<th>SECOND</th>
</tr>
</thead>
<tbody>
<tr>
<td>GENERATOR</td>
<td>R.F.</td>
<td>1510B</td>
<td>CONNECTORS</td>
<td>1:16</td>
<td>LASER</td>
<td>1:2</td>
<td>1:16</td>
<td>ODE-075C</td>
<td>R.F.</td>
<td>R.F.</td>
</tr>
<tr>
<td>AMPLIFIER</td>
<td>LASER</td>
<td>SPLITTER</td>
<td>AMPLIFIER</td>
<td>SPLITTER</td>
<td>SPLITTER</td>
<td>SPLITTER</td>
<td>PHOTOD.</td>
<td>AMPLIFIER</td>
<td>AMPLIFIER</td>
<td></td>
</tr>
</tbody>
</table>

**ELECTRICAL/OPTICAL SIGNAL POWER (dBm)**

|       | 0.0  | 10.0 | -2.8 | -4.8 | -19.3 | -3.3 | -7.8 | -22.4 | -52.3 | -28.3 | 2.8 |

**ELECTRICAL/OPTICAL GAIN (dB)**

|       | 10.0 | -2.0 | -14.5 | 16.0 | -4.5 | -14.5 | 24.0 | 31.1 |

**NOISE POWER OUT (dBm/Hz)**

|       | -174.0 | -158.4 | -70.3 | -72.3 | -86.8 | -66.4 | -70.9 | -85.4 | -172.1 | -145.4 | -114.3 |

**S/N RATIO (dBc Hz)**

|       | 174.0 | 168.4 |       |       |       |       | 119.8 | 117.1 | 117.1 |
| (dBc 60 MHz) | 96.2 | 90.6 |       |       |       |       |        | 42.0 | 39.3 | 39.3 |

**NOISE FIGURE (dB)**

|       | 5.6 | 4.7 | 3.7 | 5.0 |

**OTHER CHARACTERISTIC**

<table>
<thead>
<tr>
<th>conversion efficiency</th>
<th>loss</th>
<th>excess loss</th>
<th>excess loss</th>
<th>excess loss</th>
<th>responsivity</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.0374</td>
<td>1.0</td>
<td>2.5</td>
<td>1.5</td>
<td>2.5</td>
<td>0.75</td>
</tr>
<tr>
<td>W/A</td>
<td>dB</td>
<td>dB</td>
<td>dB</td>
<td>dB</td>
<td>A/W</td>
</tr>
</tbody>
</table>

**OTHER CHARACTERISTIC**

<table>
<thead>
<tr>
<th>bias current</th>
<th>No of elements</th>
<th>No of element</th>
<th>No of element</th>
<th>load resist.</th>
</tr>
</thead>
<tbody>
<tr>
<td>40.0</td>
<td>16.0</td>
<td>2.0</td>
<td>16.0</td>
<td>315.5</td>
</tr>
<tr>
<td>nA</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
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</table>

**OTHER CHARACTERISTIC**

<table>
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<tr>
<th>threshold current</th>
<th>15.0</th>
</tr>
</thead>
<tbody>
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<td>------------------</td>
<td>----------------------</td>
</tr>
<tr>
<td>ELECTRICAL/OPTICAL SIGNAL POWER (dBm)</td>
<td>0.0</td>
</tr>
<tr>
<td>ELECTRICAL/OPTICAL GAIN (dB)</td>
<td>10.0</td>
</tr>
<tr>
<td>NOISE POWER OUT (dBm/Hz)</td>
<td>-174.0</td>
</tr>
<tr>
<td>S/N RATIO (dBc Hz)</td>
<td>174.0</td>
</tr>
<tr>
<td>(dBc 60 MHz)</td>
<td>56.0</td>
</tr>
<tr>
<td>NOISE FIGURE (dB)</td>
<td>5.6</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>OTHER CHARACTERISTIC</th>
<th>conversion efficiency</th>
<th>loss</th>
<th>excess</th>
<th>loss</th>
<th>excess</th>
<th>loss</th>
<th>excess</th>
<th>responsivity</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0.0374 W/R</td>
<td>1.0 dB</td>
<td>2.5</td>
<td>1.5</td>
<td>2.5</td>
<td>0.75</td>
<td>A/W</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>OTHER CHARACTERISTIC</th>
<th>bias current</th>
<th>No of elements</th>
<th>No of elements</th>
<th>No of element load resist.</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>40.0 mA</td>
<td>16.0</td>
<td>2.0</td>
<td>315.5 OHM</td>
</tr>
</tbody>
</table>

Table (7.2) Modelled signal and noise levels of configuration 2.
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512 elements of the array. The total number of amplifiers required for the network is $16 + 2 \times 16 = 48$. Since the optical signal level is higher at the receive photodiodes, only two amplifier chips are needed in the second R.F. amplifier stage as compared with six in the Marconi solution. The power consumption of the complete 512 element forward link is shown in Table 7.3

Table 7.3. Power consumption calculation of the 512 element forward link for the second solution:

<table>
<thead>
<tr>
<th>Components</th>
<th>Amplifier plus control transistor</th>
<th>First R.F. amplifier stage</th>
<th>2 Plessey P35-4100 amplifier chips</th>
</tr>
</thead>
<tbody>
<tr>
<td>Power consumption per component</td>
<td>115 mA X 5 V = 575 mW</td>
<td>900 mW</td>
<td>2 X 330 mW = 660 mW</td>
</tr>
<tr>
<td>Number of components in a 512 element array</td>
<td>48</td>
<td>512</td>
<td>512</td>
</tr>
<tr>
<td>Total power consumption of the array</td>
<td>827 W</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

*In the above calculation the assumption is being made that the electronic circuit to control the current to the amplifier in order to compensate for variations in temperature, consumes a small proportion of the total current. Also it is assumed that the voltage required by the amplifier and its control transistor is 5 V.*

This figure of 827 W should be compared with 1508 W in the Marconi solution.

From Table 7.2 it can be seen that this solution also offers improved S/N ratio of 125 dBC Hz as compared with 115 dBC Hz in the Marconi solution.

The third proposed network has the following structure:

```
LASER 1:16  LASER 1:16  LASER 1:2  PHOTOD.
SPLIT. AMPLIF. SPLIT. AMPLIF. SPLIT.        
[ ]→[ ]→[ ]→[ ]→[ ]→[ ]→[ ]→[ ]
```

The signal and noise levels for this network can be seen in Table 7.4. The
<table>
<thead>
<tr>
<th>SIGNAL GENERATOR</th>
<th>FIRST R.F. AMPLIFIER</th>
<th>ORTEL 1510B LASER CONNECTORS 1:16</th>
<th>CORNING LASER 1:16</th>
<th>SEMICOND. CORNING LASER 1:2</th>
<th>PHOTOD. AMPLIFIER</th>
<th>FIRST ODE-075C R.F. SPLITTER AMPLIFIER SPLITTER AMPLIFIER SPLITTER</th>
</tr>
</thead>
<tbody>
<tr>
<td>ELECTRICAL/OPTICAL SIGNAL POWER (dBm)</td>
<td>0.0</td>
<td>10.0</td>
<td>-2.8</td>
<td>-4.8</td>
<td>-19.3</td>
<td>-3.3</td>
</tr>
<tr>
<td>ELECTRICAL/OPTICAL GAIN (dB)</td>
<td>10.0</td>
<td>-2.0</td>
<td>-14.5</td>
<td>16.0</td>
<td>-14.5</td>
<td>15.3</td>
</tr>
<tr>
<td>NOISE POWER OUT (dBm/Hz)</td>
<td>-174.0</td>
<td>-158.4</td>
<td>-70.3</td>
<td>-72.3</td>
<td>-86.8</td>
<td>-66.4</td>
</tr>
<tr>
<td>S/N RATIO (dBc Hz)</td>
<td>174.0</td>
<td>168.4</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>S/N RATIO (dBc 60 MHz)</td>
<td>96.2</td>
<td>90.6</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>NOISE FIGURE (dB)</td>
<td>5.6</td>
<td>4.7</td>
<td>4.6</td>
<td>3.7</td>
<td></td>
<td></td>
</tr>
<tr>
<td>OTHER CHARACTERISTIC</td>
<td>conversion efficiency</td>
<td>loss each</td>
<td>excess loss</td>
<td>excess loss</td>
<td>excess loss</td>
<td>responsivity</td>
</tr>
<tr>
<td></td>
<td>0.0374</td>
<td>1.0</td>
<td>2.5</td>
<td>2.5</td>
<td>1.5</td>
<td>0.75</td>
</tr>
<tr>
<td>OTHER CHARACTERISTIC</td>
<td>bias current</td>
<td>No of elements</td>
<td>No of elements</td>
<td>No of element</td>
<td>load resist.</td>
<td></td>
</tr>
<tr>
<td></td>
<td>40.0</td>
<td>16.0</td>
<td>16.0</td>
<td>2.0</td>
<td>315.5</td>
<td></td>
</tr>
<tr>
<td>OTHER CHARACTERISTIC</td>
<td>threshold current</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>15.0</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Table (7.4) Modelled signal and noise levels of configuration 3.
structure again uses two amplifier stages but in the second amplifier stage each amplifier feeds only two elements. The total number of amplifiers required for this network is \(16 + 256 = 272\). At the remote interface only the first R.F. amplifier stage (transimpedance amplifier) is required. Making similar calculations as in Table 7.3 the total power consumption of this network is 617 W. The S/N ratio is about the same with that of the previous network i.e. 124 dBC Hz

A Table summarizing the results of the 3 proposed networks using optical amplifiers and the passive solution from Marconi can be seen below (see Table 7.5)

**Table 7.5**

The main features of the 3 proposed networks

<table>
<thead>
<tr>
<th>Network no</th>
<th>No of lasers used</th>
<th>No of optical amplifiers used</th>
<th>No of Plessey no.35-4110 R.F. amplifier chips used after first amplifier stage</th>
<th>Total power consumption (W)</th>
<th>S/N ratio (dBC Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Marconi</td>
<td>16</td>
<td>0</td>
<td>6</td>
<td>1508</td>
<td>115</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>16</td>
<td>6</td>
<td>1496</td>
<td>117</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>48</td>
<td>2</td>
<td>827</td>
<td>125</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>272</td>
<td>0</td>
<td>617</td>
<td>124</td>
</tr>
</tbody>
</table>

In the above calculations the assumption has been made that it is possible to compensate for temperature variations in the optical amplifiers using electronic feedback circuits similar to those demonstrated by Ellis et al. in references [7.7], [7.8] and [7.9]. However, although this method has been demonstrated for undersea applications, it still has to be demonstrated for space applications, and suitable electronic control circuits have to be developed for this purpose. In the case the method does not sufficiently satisfy the requirements for space application, an additional budget is included here of the additional power required for temperature control.

A typical Peltier cooler which can be used for semiconductor optical amplifiers [7.10] requires a current of 1.1 A at a voltage of 0.48 V. If the voltage drop at the control transistor is 1.4 V and an efficient control circuit design is assumed, then each optical amplifier consumes...
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about 2 W. Table 7.6 shows the power budget for the three networks presented in table 7.5 if temperature control is included.

Table 7.6
Power budget for the three networks if temperature control proves necessary in the optical amplifiers.

<table>
<thead>
<tr>
<th>Network no</th>
<th>No of lasers used</th>
<th>No of optical amplifiers used</th>
<th>Total power consumption (no temperature control) (W)</th>
<th>Additional power consumption for temperature control (W)</th>
<th>Total power consumption (temperature control) (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>16</td>
<td>1496</td>
<td>32</td>
<td>1528</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>48</td>
<td>827</td>
<td>96</td>
<td>923</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>272</td>
<td>617</td>
<td>544</td>
<td>1161</td>
</tr>
</tbody>
</table>

Thus while the use of temperature compensation seriously degrades the power consumption performance of network number 3, it has little effect on the performance of network number 2 which appears to be the best candidate for use in a real application.

7.5. Conclusions

The above analysis shows that the use of optical amplifiers in the forward links of optically controlled phased array antennas can offer significant advantages in terms of electrical power consumption, S/N ratio and simplicity of construction. In particular a single optical source can be used to feed all the elements of the network, opening the way to the use of expensive and difficult to construct optical sources such as those consisting of two lasers heterodyned at their difference optical frequency (see Section 4.2.2). In terms of electrical power consumption, optical amplifiers can replace electrical ones with an advantage, especially if efficient electronic feedback control circuits can be constructed to compensate for the variation of the output optical power of the amplifiers with factors such as temperature and polarization sensitivity. In terms of signal to noise ratio, the requirements of phased arrays can be met, and the internal noise sources of the amplifiers do not significantly affect.
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the performance of the system.
To obtain the optimum performance out of the amplifiers, it is recommended
that they are placed as close as possible to the front end of the optical
link (towards the optical source) but not as close so as to reach
saturation. However in general, to design an optimized optical network, it
is recommended that a computer program similar to the one in Chapter 3 is
used, and the possible solutions compared. Finally, for the time being, the
price of optical amplifiers presents a limitation, especially for systems
at the lower end of the market, but this is expected to be solved in the
near future, given large volume applications.

7.6. References for Chapter 7

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126-132 Apr 1988

* References [7.2] and [7.3] are not in the public domain and permission has
been obtained from Marconi Space Systems and the European Space Agency to
use some of the information contained in them.
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CHAPTER 8

REVIEW OF OPTICAL BEAMFORMING TECHNIQUES FOR PHASED ARRAYS

8.1. Introduction

Having analyzed the problem of optically distributing the various signals to the elements of a phased array, the next step is to examine whether other functions of the phased array, such as beamforming, can be done in the optical domain. Traditionally beamforming and steering are performed in the microwave domain by individually controlled microwave phase shifters and variable gain amplifiers in the signal path of every array element or sub-array. In the literature various ideas have been proposed to perform beamforming in the optical domain. These can be divided into two main categories, those which can offer true-time delays and those which cannot. In general all the true-time delay beamformers can also be used for non-true-time delay phase steering if they can offer enough resolution at the operating frequency.

Optical beamforming techniques can be sub-divided into the following main categories:

a) Optical control of microwave phase shifters
b) The use of optical switches to change an optical path length.
c) Coherent phase shifters which can extend to
d) Coherent optical processors
e) Acousto-optic techniques.
f) The use of a piezo-electric crystal to stretch a fibre.

In this Chapter a review of the various beamformers which have been proposed in the literature will be presented and their relative advantages and disadvantages compared.

8.2. Optical control of microwave phase shifters
Traditionally microwave phase shifters operate by controlling the phase and amplitude of the microwave signal at each element of the array. Optical control of the bias circuits of the amplifiers of such standard microwave phase shifters has been proposed [8.1], [8.2]. Fig (8.1) shows the arrangement used. Two LED's produce optical signals with intensities proportional to their drive currents. These optical signals illuminate two optical detectors which after suitable amplification control the gain and phase of standard microwave phase shifters. However the method is of little practical use. It is usually much easier to use a digital optical network to control the gain and phase of each phase shifter. The method described above has the disadvantages that it needs two optical links, it is dependent on uncertainties of how much optical power couples to each element, it can feed only a few elements, due to the optical splitting and excess losses involved, and the optical power of the LEDs is highly dependent on factors such as temperature.

8.3. The use of optical switches to change the optical path length

Switching of microwave delay lines has been used extensively to provide true-time delays in phased array antennas using diode phase shifters [8.3] or configurations such as the Blass matrix [8.4]. However these have significant shortcomings including large size, mass and phase stability with temperature and vibrations, especially if long delays are required. Optical delay techniques consist of switching the modulated optical signal through different lengths of optical waveguide to obtain the required delay. A typical optical fibre switch configuration is shown in Fig (8.2). Here a LiNbO$_3$ optical switch array composed of six 2x2 directional coupler switches together with three fibre delay lines of lengths corresponding to $\Delta t$, $2\Delta t$ and $4\Delta t$ make up a three bit switchable fibre optical delay network. Theoretically if all the switches are the same only two voltages are required to operate them. However in practice the switches suffer high optical losses and each switch has its own optimum switching voltages. Thus either the control system becomes very complicated or the performance of the switch array is degraded. Typical values for the losses of such switches are of the order of 7 dB for a Ti:LiNbO$_3$ 8x8 matrix [8.5].
Fig (8.1) Optical control of a microwave phase shifter. (After [8.1])

Fig (8.2) A typical switching optical delay line phase shifter. (After [8.6])
extinction ratios of integrated optical switches are typically in the range
of 15 to 20 dB. Recently a switch has been demonstrated with an extinction
ratio of 27 dB and this parameter is expected to exceed 40 dB [8.5] before
long.

Another switching configuration which has received some attention is shown
in Fig (8.3). The idea was first proposed in [8.6] and experimentally
demonstrated in [8.7] and [8.8]. The idea is similar to the Rotman
microwave beamformer and an optical fibre of appropriate length goes from
every array element to every beam port. Thus by electrically switching to
the photodiode for the appropriate beam port, a certain beam direction can
be selected. The technique can also be used to transmit a signal, by
changing the beam port photodiodes to lasers and the array element lasers
to photodiodes.

The system has the disadvantages that it needs \( N \times M \) optical fibres, an \( M \)
way optical divider in each array element and an \( N \) way combiner in each
beam port. Thus the system is only suitable for arrays with small number of
elements (or sub-arrays) and a limited number of beam positions.

In general various architectures are possible to arrange the switches and
the delay lines in a switching network. References [8.9] and [8.10] analyze
the main possible configurations and present the optimum ones in terms of
simplicity and performance.

Finally some integrated versions of switched delay lines have been
presented [8.11], [8.12]. The most interesting of these consists of a
"folded" optical waveguide integrated onto the optical substrate (see Fig
8.4). The switching elements consist of integrated gratings which can be
made either transmissive or reflective at the required wavelength by
electrically changing the refractive index. These gratings can offer about
\( \approx 100 \% \) reflective efficiencies and thus by using the two dimensions of the
substrate relatively long delays can be achieved.

8.4. Coherent phase shifters

The basic principle behind coherent phase shifters and processors is the
following: If two optical beams whose optical frequencies differ by a
microwave frequency are combined together and then photodetected, the
microwave difference appears at the output of the photodetector. If the
Fig (8.3) An N element X M beam electro-optic beamformer. (After [8.8])

Fig (8.4) The integrated "folded" waveguide technique presented in ref [8.12]. (After [8.12])
optical phase of one of the two beams changes, then this change transfers to the phase of the photodetected microwave signal. Changing the phase of an optical signal is usually much easier than changing that of a microwave one, as sufficient phase change can be obtained by changing the refractive index of an optical waveguide using an electric field. The two optical beams can be produced from two separate lasers heterodyned together or from a single laser with an optical frequency shifter.

Coherent beamforming using the first technique has been demonstrated in ref [8.13] (see Fig 8.5). In this technique the two (or more) optical carriers $\omega_{LO}$ and $\omega_i$ enter an integrated network of power splitters and amplifiers and the phase and amplitude of each of the outputs are controlled by coherent phase shifters and optical attenuators. Then each pair of signals is combined together and transmitted through an optical fibre to an element of the array. The receive operation is similar.

A second method of producing the two optical carriers is to use an optical frequency shifter. Experimental demonstrations of such techniques are presented in ref. [8.14], (bulk optics) and [8.15] (integrated). Also some general discussion about coherent phase shifters can be found in [8.16] and [8.17].

The main disadvantage of coherent optical phase shifters arises due to the temperature sensitivity of the phase of the two optical carriers. These cannot be transmitted through long distances in separate optical fibres without their relative optical phase difference being affected by temperature gradients or vibrations. Thus power splitting, amplification, phase shifting and mixing of the optical signals must take place at the central processor of the array and not at the array elements. This will result in relatively complex optical processors if the number of the array elements is big. However integrated solutions to this problem have been demonstrated [8.13].

The advantage of coherent phase shifters is that they can be made physically small, offering lower mass and electrical power consumption than other solutions.

8.5. Coherent optical processors

The idea behind coherent processors is similar to that of coherent phase
Fig (8.5) Coherent beamforming using two or more heterodyned lasers.
(After [8.13])
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shifters but the processing (modulation of the optical carrier and introduction of phase shifts) is applied to a single laser beam and then the beam enters into a bundle of fibres which distribute the optical signals into the elements of the array. A typical configuration is shown in Fig (8.6) where the beam of the laser is divided into two parts. The first passes through a spatial light modulator to control its spatial optical phase and the second through an optical frequency shifter to introduce a frequency offset to the optical carrier equal to the microwave frequency of the array. The two beams are then heterodyned together giving a resulting beam intensity modulated at their frequency difference, while the phase of the modulated microwave at every point in the cross section of the beam is proportional to the optical phase introduced by the spatial light modulator. Techniques based on this principle are described in [8.18], [8.19], while the technique described in [8.20] and [8.21] amplifies the frequency shifted beam by mode-locking a slave laser.

Another coherent optical processor is proposed by Koepf in ref [8.22] and an experimental demonstration is presented by the same author in ref [8.23]. This is based on the following principle: The far-field cross section of a microwave antenna beam can be described as a two-dimensional Fourier transform of the antenna illumination. In the reverse direction, the illumination is obtained by a Fourier transform of the antenna far-field distribution. In the optical processor demonstrated by Koepf [8.22], [8.23] the two dimensional Fourier transform is performed in real time by far-field diffraction of coherent light. Beam formation is therefore possible by coherent illumination of a mask that represents a scaled-down replica of the desired antenna beam, and by down-converting the spatial information of the optical far field into the spectral region where the antenna emits. This is shown schematically in Fig (8.7). The down-conversion is possible with an array of photodetectors, one detector per element in the array antenna. A more detailed diagram of the configuration proposed by Koepf is shown in Fig (8.8). A laser that emits two light beams with the desired frequency offset is used as a source. One beam is directed through an electro-optic modulator and focused on a pinhole, which represents a circular antenna spot beam. The spatial amplitude and phase information is obtained in the focal plane of the FT lens. The second beam passes through two electro-optic crystals that are
Fig (8.6) Basic configuration of a coherent beamformer. (After [8.18])

Fig (8.7) Principle of operation of Koepl's beamformer. (After [8.22])
Fig (8.8) Configuration of Koepf's beamformer. (After [8.22])

Fig (8.9) The use of coherent techniques together with switched true time delays in bulk optics. (After [8.25])
used to steer the beam in its azimuth and elevation. The beam is then expanded to form a plane wave and is combined with the first beam. A bundle of optical fibres is used to sample the superposition of the fields and to distribute the optical signals to the elements of the array.

One of the main disadvantages of the coherent phase shifters and optical processors presented so far is that they do not provide true-time delays. Two solutions to the problem have been presented:
The first presented by Dolfi [8.24], [8.25] combines coherent phase shifting with switched delays. In this configuration using a high power laser and a Bragg cell, (see Fig 8.9) two beams are produced with their optical frequencies differing by the microwave frequency applied to the Bragg cell. These beams are polarized with their polarization angles differing by 90° and they travel along the same path. After being expanded through a beam expander (BE) the combined beam intercepts \( M_o \) a liquid crystal SLM of \( p \times p \) pixels. This SLM provides an analogue control of the microwave signal phase since it changes the relative optical phase of the cross polarized components of the dual frequency beam. After this non true time-delay phase shifter, the two polarizations are recombed on a 45° oriented polarizer and intercept a set of spatial light modulators SLM\(_i\), polarizing beam splitters PBS\(_i\) and prisms P\(_i\). They provide the parallel control of the time-delays to the antenna as follows: each SLM consists of an array of \( p \times p \) pixels and the beam polarization is rotated by 0° or 90° according to the applied voltage on each pixel. Each PBS\(_i\) consists of two polarizing beam splitters joined side by side. When the polarization is horizontal, the PBS\(_i\) is perfectly transparent and the light intersects the next SLM. When the polarization is vertical, then the first polarizing beam splitter of PBS\(_i\) is perfectly reflecting and deflects the beam towards a prism P\(_i\). P\(_i\) acts as a corner cube and the beam reflects off the second polarizing beam splitter of (PBS\(_i\)) towards the next SLM. The collimated beam travels through all the SLM and PBS and is focused by an array of microlenses \( L \) to an array of \( p \times p \) photodiodes (PDA) or can be entered into a bundle of fibres to feed the elements of the array. The method has been demonstrated experimentally [8.25].
The second coherent true-time delay optical processor has been suggested and experimentally demonstrated by Toughlian and Zmuda [8.26].
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technique works as follows: (see Fig 8.10)
An optical splitter splits the beam of a laser into two paths. The one is
used as a reference beam and the other passes through an optical frequency
shifter which shifts the optical frequency by an amount equal to the input
microwave signal. The resulting optical beam hits a diffraction grating and
its optical frequency components become spatially dispersed as shown in Fig
(8.10). A cylindrical lens is used to stop the spread and image the
spatially separated frequency components on a reference plane in front of a
mirror which can be both translated and rotated. So as can be seen from Fig
(8.10) each optical frequency component experiences a slightly different
time delay due to the angular position of the mirror. This time delay can
be varied by rotation and translation of the mirror. Since the heterodyne
process in coherent phase shifters translates the optical phase into the
output microwave signal, after combination with the reference beam, a
microwave phase shift equal to the optical phase shift results. This
microwave phase shift varies linearly with frequency and in effect provides
true time delay.
The technique can use the parallel nature of light to make true time delay
optical processors and the authors suggest several realizable
configurations.
Both of these true-time delay coherent optical processors can offer compact
size, good phase resolution and compatibility with fibre signal
distribution systems. Their main problem is that it is difficult to realise
the precise mechanical construction required for these processors.

8.6. The use of acousto-optic techniques

A problem of using variable optical or microwave paths in achieving
true-time delays is that since the speed of light is very high, the lengths
of the delay lines required to achieve the required delays are relatively
large. Thus it is difficult to integrate these structures on a single chip.
Another promising solution is to use the speed of sound as the mean of
delay. Since the acoustic speed is much lower than the speed of light the
lengths involved to achieve the same delays are much smaller. Integrated
SAW delay lines (not using optics) offer a compact and cheap solution to
achieve true-time delays [8.27], [8.28]. The upper frequency which SAW
Fig (8.10) The principles of operation of a true-time delay coherent phase shifter with optical frequency separation. (After [8.26])

Fig (8.11) A typical acousto-optic beamformer. (After [8.30])
components can reach is $\approx 10$ GHz [8.27] although as shown in Chapter 2, true-time delay phase shifters can be used in the I.F. domain while only non-true-time delay phase shifters [8.29] are required in the R.F. domain. Thus for the frequencies involved in the I.F. domain, SAW phase shifters can often satisfy the frequency and bandwidth requirements. However further examination of purely electro-acoustic phase shifters is beyond the scope of this thesis. The acousto-optic true-time delay beamformers [8.30], [8.31], [8.32] are mainly based on the following principle (see Fig 8.11):

The input signals from the elements of the phased array antenna are applied to a column of acoustic transducers which emit acoustic waves to an acousto-optic crystal. Each transducer is driven independently and the acoustic waves are spatially separated into rows as shown. Typically the row to row isolation is 30 dB [8.30]. With attention to any particular row the electrical drive signal is converted into a spatially travelling wave in the crystal. A strong laser beam is applied to the acousto-optic crystal in a direction perpendicular to the direction of travel of the acoustic waves. The crystal is followed by a matrix of pixels which can be turned either on (allow the light to pass) or off (stop the light). The light passing from each pixel is modulated by the acoustic wave through the process of acousto-optic diffraction. So in a certain row the light passing through every pixel carries a time-delayed version of the original signal applied to the transducer of this row. So depending on which of the pixels in a certain row is on a different time delay can be applied to the signal. After the acousto-optic crystal there is a lens which focuses all the beams from the pixels to a photodetector. The output from the photodetector is the combination of all the element signals.

The array of pixels can be replaced by other types of masks. For example [8.32] replaces the array of pixels by a mask with a narrow slit. By mechanically rotating the mask, the direction of the input beam to the array can be rotated.

Since the light coming out of the array of pixels is very weak (only one pixel in a row is on at a time), [8.30] proposes a way to amplify it by coherently mixing it with a strong reference beam using a photo-refractive crystal.

The use of acousto-optic techniques offers a compact solution to introduce true-time delays. However it is usually easier to keep the signal in the
electro-acoustic domain (i.e. use switched SAW delay lines or other electro-acoustic components) than to introduce the extra complexity of optics especially when the optical components required for the acousto-optic beamformers are not the same as those used for the signal distribution in the array.

8.7. The use of a piezo-electric crystal to stretch a fibre.

The use of mechanical stretching of the fibre to achieve true time delay has been suggested and experimentally demonstrated in references [8.33] and [8.34]. It uses a cylinder made out of piezo-electric material around which the optical fibre is wrapped. When a voltage is applied to the piezo-electric crystal it expands thus stretching the fibre. This expansion of the fibre and its subsequent contraction after the PZT voltage is removed introduces a variable true-time delay for the microwave modulated signal travelling through the fibre.

The concept has been experimentally demonstrated [8.33] using a 9.6 cm diameter PZT crystal with 75 turns of fibre wrapped around it. A 4.2 KV voltage is applied to this crystal and for a signal of 10 GHz modulated on the input optical carrier a phase shift of 20° is obtained. The time constant for the optical fibre to relax to its original position is a few minutes and good repeatability is obtained.

The disadvantages of using this method for true-time delays are the following:

a) Since the excess length obtained by stretching the fibre is only a very small portion of its original length, very large lengths of fibre are needed to produce significant true-time delays.
b) Very high voltages are required to drive the PZT crystal.
c) The fibre needs some minutes to relax to its original position after being stretched. This means that the idea cannot be used when fast beam steering is required.

8.8. Conclusions

In this Chapter the principles of operation of the main types of optical
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and optoelectronic beamformers which have been reported in the literature have been presented. The techniques which can provide true-time delays are optical switched delay lines, coherent optical processors with frequency shifting or with switched delays, acousto-optic techniques and fibre stretching by mechanical means. From these techniques the optical switched delay lines usually require large amounts of optical fibre, so they can be mainly used to provide true-time delays in large sub-arrays in conjunction with the use of non-true time delay phase shifters at each element of the sub-array.

Acousto-optic techniques, although they can be made compact and lightweight, utilizing the parallel processing capability of light, have the major disadvantage that they are not compatible with fibre optic distribution techniques.

The most promising method of all the true-time delay techniques appear to be the use of coherent techniques with optical frequency shifting or with switched delays. These techniques can utilize the parallel nature of light and provide compact true-time delay optical processors. Again as with most coherent optical processors the construction tolerances and optical source requirements may be difficult to meet but this problem should be solved in the future.

From the non-true time delay beamformers, coherent optical processors appear to offer the best solution for future systems as they fully utilize the parallel processing potential of light, can be made compact and lightweight for applications such as satellites and they are compatible with fibre optical distribution techniques. Their main drawbacks are their critical optical source requirements and precision of construction, but these can be solved in the future.

8.9. References for Chapter 8


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[8.5] Pan J.J. "Fiber optics for wideband extra high frequency (EHF) phased arrays." Proc SPIE vol 886 pp. 60-63 (Conf. on optoelectronic signal processing for phased array antennas Los Angeles CA USA Jan 1988)


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CHAPTER 9

A NOVEL OPTICAL BEAM-FORMER FOR MICROWAVE PHASED ARRAY ANTENNAS

9.1. Introduction

The review of the optical beamforming techniques in the previous Chapter, has shown that the most important techniques presented in the literature, either require a critical coherent optical source and an accurate mechanical design, or a large number of low-loss optical switches. Thus their implementation usually requires high opto-electronic technology which is not always easy to achieve.

In this Chapter, the theory and experimental construction of a novel opto-electronic beamformer will be presented which can be implemented using relatively simple optoelectronic and microwave technology.

9.2 Theory of the beamformer

The theory behind the beamformer is the following (see Fig (9.1)): Suppose the radiated frequency for the array is $F$. This signal is distributed to the elements of the array through a dispersive fibre network. In this network successive fibre elements are made longer than preceding ones so as to introduce an extra phase shift of $2k\pi$ to the microwave modulated optical signal at the boresight frequency, where $k$ is an integer. Clearly these $2k\pi$ phase shifts would not affect the boresight alignment of the phase front. However, any frequency shift in the frequency $F$ say $\Delta F$, would produce a phase slope of:

$$\Delta \phi = \frac{2k \pi n \Delta F}{F}$$  \hspace{1cm} (9.1)

across the array where $n$ is the number of elements. For an array of length $D$ the beam would be steered by:

$$\theta = \sin^{-1} \left( \frac{k\Gamma n \Delta F}{F D} \right)$$  \hspace{1cm} (9.2)
Fig (9.1) The working principles behind a variable frequency beamformer.
where $\Gamma$ is the wavelength at the centre frequency.

There are many operational radar systems which use this principle for beamsteering. However if the system application does not allow the frequency to change with scan angle, an alternative arrangement (fig 9.2) can be used: Suppose the radiated frequency for the array is $f$ and $F$ is another frequency which controls the inter-element phase change. These two signals have phase $(2\pi Ft)$ and $(2\pi ft)$. The $(2\pi Ft)$ signal is distributed through a dispersive link (link A) and two signals with a phase difference of say $\phi$ given by (1) reach mixers C and D with phase of $(2\pi Ft+\phi)$ and $(2\pi Ft)$ respectively. The original signals are mixed at mixer A. At the output of mixer A two signals arise with phase $\{2\pi(f-F)t\}$ and $\{2\pi(f+F)t\}$. After filter B only the $\{2\pi(f-F)t\}$ component remains. This is distributed through a non dispersive link (link B) and two signals $\{2\pi(f-F)t\}$ reach mixers C and D. At the output of mixer C two signals arise one with phase $\{2\pi(f)t+\phi\}$ and one with phase $\{2\pi(f+2F)t+\phi\}$. The second one is removed by filter E and only the $(2\pi ft+\phi)$ component remains. Similarly at the output of filter F a $(2\pi ft)$ component remains. Thus by changing the control frequency $F$ a phase slope $\phi$ given by (1) is introduced to the radiated frequency $f$ of the array. It can be seen that these signals are of frequency $f$ with a phase difference $\phi$ between them.

This technique has previously been demonstrated using microwave components by Huggins [9.1] and Davies [9.2]. This experiment demonstrates this method using optical signal distribution. That is links A and B use single mode optical fibre instead of coaxial cable, the microwave signals being modulated onto optical carriers. The use of single mode optical fibre offers the advantages of low weight and volume, good temperature stability and good phase stability with environmental factors such as shocks and vibrations. Also since the transmission line lengths usually required in such systems are large, at high frequencies the use of optical techniques can offer improved attenuation.

9.3. Description of the experiment

To make an experimental demonstration of the principle described in the previous paragraph, the following setup was used (see Fig (9.3)): Only two
Fig (9.2) Block diagram of the beamformer.
Fig (9.3) Detailed block diagram of the beamformer.
elements were used and their phase relative to one another was compared using a network analyzer. The frequencies used were 1500 MHz for f and 200 MHz for F. For the first part of the experiment only the 200 MHz link was constructed, and then a 1300 MHz link was added to the whole system. The setup works as follows: Two frequency generators produce the 1500 MHz and 200 MHz frequencies. The 200 MHz frequency is split into two parts one of which modulates the laser (laser 1) and the other is mixed with the 1500 MHz frequency to produce two sidebands one at 1300 MHz and one at 1700 MHz. A filter (filter 1) removes the 1700 MHz sideband and the 1300 MHz one is distributed through a non dispersive network (via laser 2) to the two elements. There it is mixed with the 200 MHz frequency which comes out of the photodiode and amplifier of the dispersive link. This mixing results in two sidebands, one at 1500 MHz and one at 1100 MHz. The 1100 MHz sideband is filtered with a bandpass filter and so only the 1500 MHz sideband remains whose phase is compared with that of the other element using a network analyzer.

In the next paragraphs the design and construction of every individual part in the experimental setup will be described.


Two optical links have been constructed for this experiment, to transmit the 200 MHz and 1.3 GHz signals. These consist of a laser driver circuit, a laser, a single mode optical fibre, (which in the real case splits into 2 branches through a fibre coupler to feed the two elements of the experiment) a photodiode and a photodiode amplifier.

The laser used was a single mode Hitachi HLP 1400 together with a suitable bias and modulation circuit. This laser has an effective resistance of about 3 Ω. So a resistance of 47 Ω was put in series with it to match it to the 50 Ω impedance of the input signal transmission line. The use of LC or transformer impedance matching circuits was avoided because they introduce phase shifts when the frequency changes. In real systems matching circuits can be used which improve the performance of the links [9.3], [9.4]. However attention to their frequency - phase characteristics is needed when designing them for this application.

The laser was coupled to a single mode optical fibre without the use of
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lenses to collimate and focus the beam to the fibre. In a production system the use of lens-ended fibre [9.5] would give better coupling efficiency (typical value about 5 dB optical loss). The photodetector used was a Telefunken S 213 PIN photodiode with the optical fibre shining directly on its window so that again the use of focusing optics was avoided. The RF loss of the links (from the laser input to the output of the photodiode can be seen in Fig (9.4). As can be seen there is a loss of 52 dB in the 200 MHz link and 57 dB in the 1.3 GHz link. To compensate for these losses amplifiers were used in both links after the photodiodes. The amplifier for the 200 MHz link consisted of 6 cascaded stages of the silicon integrated circuit RS 560 in two RF screened boxes while the one for the 1.3 GHz link consisted of 5 cascaded stages of the Avantek MSA 0135-21 silicon integrated circuit in two RF shielded boxes. To stop the amplifiers from oscillating, sheets of microwave absorbing material were placed between the amplifier PCB and the metallic top of the box. Ring diode mixers were used to mix the signals, the first being an ANZAC MD-525-4 and the second (in the remote interface) an ANZAC MD-113. The microwave filters were realized with microstrip technology on duroid. Filter 1 was a 7th order lowpass and filter 2 a 3rd order bandpass using coupled lines. Finally a 1.5 GHz bandstop filter was added before the last mixer to remove the unwanted 1.5 GHz signal from the first mixer and filter 1. Leakage of this signal through laser 2 and the optical link to the final mixer and filter would cause a periodic variation of the output amplitude.

9.5. Spectrum analyzer plots of the signals in the system

Before the phase versus frequency results are presented, the experimental spectrum analyzer plots of the various signals inside the system will be presented.

Fig (9.5) and (9.6) show the outputs of the 200 MHz and 1.5 GHz signal generators respectively. Fig (9.7) shows the output of the first mixer which is used to mix the 1.5 GHz and 200 MHz signals. The 1.3 GHz & 1.7 GHz mixing products can be seen together with the unwanted 1.5 GHz local oscillator signal. Fig (9.8) shows the signal after it has passed through filter 1 (lowpass filter) to remove the 1.7 GHz mixing product. As can be seen the level of the 1.5 GHz and 1.7 GHz signals are about 33 dB below the
Fig (9.4) Experimental R.F. link loss (from the input of the laser to the load resistance of the photodiode) versus frequency of the two optical links (no splitting losses present).

Fig (9.5) The 200 MHz input signal to the beamformer.
Fig (9.6) The 1.5 GHz input signal to the beamformer.

Fig (9.7) The signal output of the first mixer to combine the 1.5 GHz and the 200 MHz frequencies.
Fig (9.8) The output signal of filter 1. It can be seen that the 1.7 GHz sideband is suppressed approximately 30 dB relative to the 1.3 GHz one.

Fig (9.9) The output signal of the 200 MHz optical link.
Fig (9.10) The output signal of the 1.3 GHz optical link. The parasitic 1.5 GHz signal can be clearly seen.

Fig (9.11) The output of the bandcut filter before it goes to the final mixer. This should be compared with Fig (9.10) The parasitic 1.5 GHz signal has now disappeared.
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1.3 GHz signal. Fig (9.9) shows the output, after photodetection and amplification, of the 200 MHz link. The signal is of power 8 dBm to drive the LO of the second mixer while the harmonic at 400 MHz is about 20 dB below the signal. Fig (9.10) shows the output, after photodetection and amplification, of the 1300 MHz link. This signal passes through the 1.5 GHz bandcut filter and the output response is shown in Fig (9.11). Note that the 1.5 GHz parasitic signal of Fig (9.10) is suppressed by more than 20 dB in Fig (9.11). Finally the 1.5 GHz output signal of the system is shown in Fig (9.12).

9.6. Theoretical and experimental characteristics of phase versus frequency of the beamformer

Rearranging equation (1) the theoretical formula for the phase change versus frequency and extra length of fibre between the two elements is:

\[
\Delta \Phi = \frac{2 \pi \Delta F \Delta l}{c} n
\]  

(9.3)

Where \( \Delta \Phi \) is the phase change in radians, \( \Delta F \) is the frequency shift of the frequency F, \( \Delta l \) is the extra length of fibre between the two elements, \( n \) is the refractive index of the fibre and \( c \) is the speed of light in vacuo.

From this formula it can be seen that when the length \( \Delta l \) of the fibre between the two elements is big, then with a small change in frequency a large change in phase can be achieved. This is desirable so that differences in phase and amplitude characteristics of the amplifiers and filters versus frequency do not significantly affect the response of the system. In real systems of course the differences between lengths of fibre between successive elements cannot be too big due to practical reasons. As a compromise, typical lengths in a practical system can be of the order of 1 to 10 metres which require frequency shifts of 187.5 MHz and 18.75 MHz respectively to produce a 360° phase shift.

However for a two element experimental demonstration like in this case a large additional length does not present any serious problems in the construction and helps demonstrate the principle, since by sweeping the frequency over a small range many 360° phase circles can be obtained. For
Fig (9.12) The final 1.5 GHz output signal of the beamformer.

Fig (9.13) The final phase versus frequency response of the system. The graph includes 2 complete 360° phase circles. Experimental points were taken every 50 KHz on the frequency axis.
these reasons 150 m of fibre were placed as the additional length $\Delta l$ in the 200 MHz dispersive optical link. The results of the experiment are shown in Fig (9.13) for two complete 360° cycles. As can be seen from the curve the phase versus frequency characteristic is linear within about 1 % which is due to experimental error. The frequency difference for a complete 360° rotation of the phase is about 1.25 MHz. For this experiment by substituting in the above formula $\Delta l = 150 \text{ m}$, $n = 1.6$ for a typical glass fibre and $\Delta \Phi = 2\pi$ for a complete 360° cycle, a value of $\Delta \Phi = 1.25 \text{ MHz}$ is obtained. This is exactly in agreement with the experimental observation of Fig (9.13).

9.7 Noise analysis of the beamformer

Two types of noise are significant for this type of beamformer: phase noise and intensity noise. Phase noise is significant in the sense that any residual frequency modulation in the control frequency $F$ will produce a modulation in the output phase of the beamformer. From equation (3) the rms phase error in the output phase is given by:

$$\Delta \Phi_{\text{rms}} = \frac{2 \pi \Delta F_{\text{rms}} \Delta l n}{c} \quad (9.4)$$

where $\Delta F_{\text{rms}}$ is the rms frequency error of the control frequency $F$. Thus for a typical $\Delta F_{\text{rms}}$ of 20 Hz and a fibre length difference between two elements of 20 metres the rms phase error will be $\Delta F_{\text{rms}} = 0.00077^\circ$ which is clearly not significant.

To model the intensity noise performance of the beamformer the various parameters of the components of the 1.3 GHz optical link have been entered into the computer model of Chapter 3. Since balanced and double balanced mixers are designed to reject noise coming from the L.O., the noise contribution of the 200 MHz link which is used as the L.O. in the final mixer has not been taken into account. Instead the mixer is modelled as an attenuator for the 1.3 GHz signal.

The parameters of the components of the 1.3 GHz link are the following:
Fig (9.14) The frequency response of the laser HLP1400 and its bias circuit for various bias currents (measured with a very high frequency photodiode).

Fig (9.15). Modelled and experimental S/N ratio versus total optical attenuation of the beamformer (The attenuation was measured from the output facet of the laser to the input facet of the photodiode).
Chapter 9: A novel optical beamformer

For simplicity the input to the link has been considered to be the -15 dBm 1300 MHz signal from the first mixer. This signal passes through the low pass filter which has a measured ≈ 4.9 dB attenuation at the operating frequency. Then it passes through a 24 dB gain amplifier with a noise figure of 3.5 dB to feed laser 2. The frequency response of this laser and its bias circuit can be seen in fig (9.14) for various values of input current. This response was measured using a photodiode with a frequency cutoff point well above that of the laser. The low frequency response is mainly due to the coupling capacitor of the bias circuit while the high frequency response is mainly due to the laser itself. As can be seen from fig (9.14) at a frequency of 1300 MHz (log frequency = 9.11) there is an attenuation of ≈ 10 dB relative to the maximum frequency response. In the noise measurement experiment the input current to the laser was 57 mA while its threshold current was ≈ 48 mA. Its conversion efficiency was measured to be ≈ 1.41 W/A while from fig (6) of [9.6] the RIN of the laser for this current and frequency was estimated to be about -130 dB/Hz at this bias current and frequency. The optical attenuation was measured to be about 15 dB optical loss from the output facet of the laser to the input facet of the photodiode. Of this, 12 dB was attributed to coupling losses and excess losses due to the splitter, and the remaining 3 dB to splitting losses (since 2 elements were fed from a single laser). The photodiode at this frequency and bias (-30V) has a quantum efficiency of ≈ 0.65. The load impedance of this photodiode is the input impedance of the R.F. amplifier (50 Ω). The first amplifier box has a measured gain of 27 dB at 1300 MHz. From the data sheet of the integrated circuit used the noise figure is about 5 dB when the input impedance is 50 Ω. When the input impedance is the photodiode, the output noise level at the frequency of interest was measured to be approximately the same (0.1 dB difference). So this noise figure can be used both when the input impedance to the amplifier is 50 Ω and when it is the photodiode. The second amplifier box has a gain of 22 dB at 1300 MHz and again a noise figure of 5 dB. From the data sheet of the ANZAC MD113 mixer its conversion loss at 1300 MHz is about 7.5 dB while the measured loss of the 1.5 GHz bandpass filter at 1.5 GHz is about 4 dB.

Using these parameters the modelled signal and noise levels for the system can be seen in Table (9.1). The measured results can be seen in Table (9.2). These results were taken using a HP 8569B spectrum analyzer and the
<table>
<thead>
<tr>
<th></th>
<th>FIRST</th>
<th>R.F.</th>
<th>HITACHI</th>
<th>TOTAL</th>
<th>TELEFUNKEN</th>
<th>FIRST</th>
<th>SECOND</th>
<th>SECOND</th>
<th>BANDPASS</th>
</tr>
</thead>
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<td>MIXER</td>
<td>OUTPUT</td>
<td>AMPLIFIER</td>
<td>HLP-1400</td>
<td>OUTPUT</td>
<td>LASER</td>
<td>LOSS</td>
<td>PHOTOD.</td>
<td>BOX</td>
<td>BOX</td>
</tr>
</tbody>
</table>

**ELECTRICAL/OPTICAL SIGNAL POWER**

| OUT (dB) | -15.0 | 4.1 | -4.9 | -14.0 | -63.3 | -36.3 | -14.3 | -21.8 | -25.8 |

**ELECTRICAL/OPTICAL GAIN (dB)**

| (1300 MHz) (1300 MHz) (1300 MHz) (1500 MHz) | 24.0 | -15.0 | 27.0 | 22.0 | -7.5 | -4.0 |

**ELECTRICAL/OPTICAL NOISE POWER**

| OUT (dB/Hz) | -174.0 | -146.5 | -64.0 | -79.0 | -173.1 | -141.7 | -119.7 | -127.2 | -131.2 |

**S/N RATIO (dBc Hz)**

| 159.0 | 150.6 | 109.8 | 105.4 | 105.4 | 105.4 | 105.4 |

**OTHER CHARACTERISTIC**

| noise conversion coupling quantum noise noise |
| figure efficiency loss efficiency figure figure |
| 3.5 | 0.141 | 12.0 | 0.65 | 5.0 | 5.0 |

**OTHER CHARACTERISTIC**

| bias splitting load |
| current loss resist. |
| 57.0 | 3.0 | 50.0 |

**OTHER CHARACTERISTIC**

| threshold current |
| resist. |
| 48.0 | 0Hm |

Table (9.1) Modelled signal and noise levels in the beamformer.
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<thead>
<tr>
<th></th>
<th>FIRST OUTPUT</th>
<th>R.F. AMPLIFIER</th>
<th>HITACHI</th>
<th>TOTAL TELEFUNKEN</th>
<th>FIRST SECON SECON</th>
<th>SECOND FILTER</th>
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<td><strong>ELECTRICAL/OPTICAL</strong></td>
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<td>SIGNAL POWER</td>
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<tr>
<td>OUT (dBm)</td>
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<td>-26.0</td>
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<tr>
<td>GAIN (dB)</td>
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<tr>
<td>NOISE POWER</td>
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<td>104.8</td>
<td>106.0</td>
<td>97.0</td>
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<td>loss efficiency</td>
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<td>mA</td>
</tr>
<tr>
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<td></td>
<td></td>
<td></td>
<td></td>
<td>mA</td>
</tr>
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</table>

Table (9.2) Experimental signal and noise levels in the beamformer.
Chapter 9: A novel optical beamformer

averaging noise function was used to measure the noise level. The noise level was measured at a distance of about 100 KHz from the carrier. 2 dB was added to the measured noise level to account for various factors such as for example that the noise bandwidth of the spectrum analyzer is slightly different from its signal resolution bandwidth [9.7].

Comparing the signal levels between Tables (9.1) and (9.2) it can be seen that the experimental and modelled results are about the same. Comparing the signal-to-noise ratios in the two Tables the results are the same except for the following:

a) The S/N ratio of the output of the first mixer to the first amplifier. The experimental S/N ratio appears to be lower than the modelled one (132 dBC Hz versus 159 dBC Hz) but this difference is mainly due to the noise introduced by the spectrum analyzer.

b) The modelled S/N ratio of the output 1500 MHz signal is 8 dB higher than the experimental one (105 dBC Hz versus 97 dBC Hz). Part of this is due to the noise coming from the 200 MHz L.O. input of the mixer. In the model the mixer is modelled as an attenuator with its loss being its conversion loss at that frequency. To experimentally confirm this the L.O. input to the mixer from the 200 MHz link was replaced with a 200 MHz signal of the same level from a signal generator. Using this configuration the output S/N ratio increased from 97 dBC Hz to 102 dBC Hz i.e. only 3 dB below the modelled value of 105 dBC Hz.

Also to investigate whether this additional noise had its origin in the 200 MHz or the 1500 MHz band of the L.O. link, a 1.5 GHz bandcut filter was inserted between the output of the 200 MHz L.O. link and the final mixer. However with the insertion of this filter no improvement in the noise level was recorded. So it appears that this noise indeed has its origin in the 200 MHz band of the L.O. link.

In fig (9.15) the modelled and experimental S/N ratio of the output signal is shown versus the total optical attenuation. In the experiment the optical attenuation was measured by placing an optical power meter i) at the output facet of the laser and ii) at the output end of the fibre of one branch of the 1.3 GHz optical link. The optical attenuation could be controlled by slightly moving the xyz positioner used to couple the input end of the fibre to the laser. As can be seen from fig (9.15) the experimental and modelled curves have about the same slope and the S/N
Fig (9.16) Modelled S/N ratio out of the photodiode versus total optical attenuation showing the various noise contributions.

Fig (9.17) Modelled S/N ratio and R.F. link loss versus photodiode load resistance.
ratio is mainly thermal noise limited. This is better shown in fig (9.16) where the S/N ratio of the signal out of the photodiode versus total optical attenuation can be seen, showing the various noise contributions due to RIN, thermal and shot noise.

The experimental and modelled values of S/N ratio for the link is not high enough to satisfy the requirements of a typical phased array radar where usually S/N ratios of the order of 110 dBC Hz are required (see Chapter 2). However various parameters can be improved in the link. First the optical attenuation can be improved with the use of lensed fibres to couple the light from the laser into the optical fibre - or in a practical system with the use of pigtailed lasers. This could give a reduction in the optical attenuation of about 7 dB which from fig (9.15) would give an improvement in the S/N ratio of about 11 dB thus increasing the output S/N ratio from 97 dBC Hz to 108 dBC Hz. Also a transimpedance amplifier could be used at the output of the photodiode to increase its sensitivity. Fig (9.17) shows the modelled output S/N ratio and R.F. link loss (from the input of the laser to the output of the photodiode) of the 1.3 GHz link against the load resistance of the photodiode. It can be seen from this curve that if the effective load resistance of the photodiode can be increased from 50 Ω to 300 Ω then the S/N ratio of the link can be increased by about 7 dB. Thus it can be seen that with reduced optical attenuation and a transimpedance amplifier at the output of the photodiode, the S/N ratio can exceed 113 dBC Hz which is sufficient for most applications. Finally the input R.F. power to the laser can be increased, the noise figure of the first amplifier can be reduced and optical amplifiers can be used to amplify the signals in the distribution path. Using these methods the output S/N ratio can potentially exceed 120 dBC Hz.

9.8. Suitability of the phase shifter for real systems - advantages and disadvantages.

The microwave version of this system (Huggins phase shifter [9.1], [9.2]) had several disadvantages, leading to its having been superseded by ferrite phase shifters. The main disadvantages are:

a) Since the total transmission line lengths are usually big, they present
the disadvantages of high attenuation (especially at high frequencies), large weight and volume. Furthermore they have decreased temperature stability and also bad phase stability due to other environmental factors such as shocks, vibrations etc.

b) Since an amplifier is required at every element, it cannot be used in passive phased arrays

c) It does not offer true-time delay phase shifting.
The use of optical techniques overcomes most of these disadvantages as follows:

1) The networks can be realized in single mode optical fibre offering extremely low loss and dispersion Also it is much lighter in weight and occupies less volume than coaxial cable or waveguide (see Table 1.1).

2) Fused tapered fibre couplers can be used for power division within the network giving wide bandwidth and excess losses below 0.2 dB per bifurcation [9.8].

3) The use of single mode fibres for network realization offers improved phase stability relative to coaxial cable or waveguide [9.9].

4) The networks can form part of an optical signal distribution system for the other array signals (received signal, control and telemetry) by using frequency or wavelength division multiplexing.

5) The microwave method (Huggins phase shifter) requires a power amplifier at the end of the last mixer in each element of the array. In contrast ferrite phase shifters can introduce phase shifts in relatively high power signals which can be radiated from the array without any further amplification.

However in optically controlled phased array systems, an amplifier is needed in any case to amplify the signal out of the photodiode before it can be radiated. Also with modern microwave circuit technologies these amplifiers can occupy a very small volume and can be relatively cheap.
Mixers can also occupy a low volume, the only problem probably is the size of the microstrip filters which will be however of the order of size of the radiating antenna.

The disadvantages of the system are as follows:

1) It does not offer true-time delay phase shifting

2) It does not offer independent control of phase at each element. This means that only pencil-like beams can be created and not other beam shapes which may be required for other operations.

9.9. Conclusions

A novel beam-forming technique for optically controlled phased array antennas which offers independent specification of output frequency and phase-slope has been demonstrated. By simultaneously sweeping the frequency sources f and F frequency independent beam steering can be obtained. The networks used are fully compatible with general optical signal distribution requirements. It is possible to extend the technique to two dimensional arrays with mono-pulse tracking and other requirements [9.10]. The technique can be applied to antennas operating at frequencies up to 30 GHz using directly modulated lasers or external modulators [9.11], [9.12]. For higher frequencies the sources could be replaced with an optical heterodyne system [9.13]. A significant reduction in the element complexity could be achieved by replacing the photodetectors and mixers with opto-electronic mixers [9.14].

9.10. References for Chapter 9


* Reference [9.1] is not available in the public domain.


CHAPTER 10

CONCLUSIONS

10.1 Main results

This thesis had two main objectives. The first was to examine whether optical amplifiers can be used with advantage in optically controlled phased arrays and to set design guidelines for their integration into future systems. The second was to design and build a novel opto-electronic beamformer and characterise it. Both objectives were fulfilled bringing new contributions to the field.

10.1.1 The use of optical amplifiers in phased arrays

Until now, the literature on optically controlled phased arrays has only examined passive network distribution architectures, that is architectures which do not amplify the signal in the optical domain. Such architectures impose many limitations on the design, especially when high microwave frequencies are involved. In references [10.1] - [10.5] and in Chapters 5, 6 and 7 of this thesis a thorough analysis of the use of optical amplifiers in phased arrays has been presented for the first time. In particular two computer models which model networks with optical amplifiers have been presented. The first in Chapter 3 models the noise performance and power budgets of active or passive optical networks. The second in Chapter 6 and reference [10.1] is a rate equation based model for optical amplifiers which examines their nonlinear performance. This second model has also introduced a number of novel equations to the field of semiconductor laser amplifier modelling.

10.1.2 Experimental demonstration of a novel opto-electronic beamformer

The second achievement of this thesis has been the design, experimental demonstration and characterization of a novel opto-electronic beamforming technique. From the review of optical beamforming techniques in Chapter 8 it was seen that the most important techniques presented in the literature,
either require a critical coherent optical source and an accurate mechanical or integrated optical design, or a large number of low-loss optical switches. Thus their implementation usually requires high opto-electronic technology which is not always easy to achieve. The beamformer presented in Chapter 9 and in references [10.6] - [10.7], uses relatively simple opto-electronic and microwave technology and provides a solution for optically controlled phased array antennas in situations where it is too expensive to develop a more complicated technology.

10.1.3 Review of the literature on optically controlled phased array antennas

In addition to the work described above, a critical review of the literature on optically controlled phased array antennas was undertaken. Although this does not represent a new achievement, it provides a very useful reference for newcomers to the field and informs those who have previously worked in the area about the latest technological developments. It is also worth mentioning that this review, apart from being presented here for academic purposes, has also been used by the personnel of the European Space Agency (under contract No 112897) to select the optimum architecture for future state-of-the-art design of optically controlled phased arrays.

10.2 The future of optics in phased array antennas

Since 1981 when Professor Forrest of University College London first proposed the use of optics in phased array antennas, an increased number of papers has been published every year, establishing optics as a signal distribution and beamforming alternative to microwave techniques. System designers are considering the use of optics for future systems, however there is still a reluctance among microwave designers to accept a novel technology which they are not very familiar with, unless it can offer some significant advantages compared with existing techniques. The main drawback of optics is that it needs to convert the signal to a different domain - the optical domain - introducing an extra level of complexity. However, once the signal is inside the optical fibre, then its advantages are
manifested which are in terms of phase stability, mass, dispersion and attenuation. These advantages become important as the transmission frequency and the lengths of the distribution or delay lines increase.

So in the near future the first area where optics are likely to penetrate into the world of phased arrays is in constructing switched true-time delay devices. These will be used at the sub-array level, together with conventional non-true-time delay phase shifters at every element of the array. Since these devices are going to be used in sub-arrays, only a few of them are required for the whole array, so their price and difficulty of construction is not a primary concern. Recently an experimental phased array using this architecture has been constructed [10.8]. Since in such architectures the signals will be converted to the optical domain, the next logical step is to use optics for the distribution of the signals from the central array processor to the sub-arrays.

Whether the distribution of the signals from the sub-arrays to the elements of the array will be in the optical or the microwave domain depends on several factors:

a) The price and efficiency of the optical receivers required at each array element

b) The price and efficiency of the optical amplifiers required to compensate the splitting losses between the elements of the sub-array.

c) Whether non-true-time delay phase shifting can be easily and efficiently performed in the optical domain for every element of the array.

With reference to factor (a), integrated optical receivers can easily be constructed in the T/R modules so this presents no major problem. For space applications the power consumption of the RF amplifiers after the photodiode may be a critical factor.

With reference to factor (b), the calculations of Chapters 5, 6 and 7 of this thesis have for the first time demonstrated that efficient optical distribution networks using optical amplifiers can be constructed to satisfy the requirements of phased arrays.

With reference to factor (c) it is doubtful whether there is currently any element-based optical non-true time-delay phase shifter to match the simplicity of its microwave counterpart.

However it is believed that technical advances in the field will overcome these problems giving way to efficient all-optical forward signal
distribution and beamforming networks.
For the IF return links, again optics can be used with advantage to perform
switched true-time delays and hence transmission of signals from the
sub-arrays to the central processor. Whether optics can be beneficially
used for signal transmission from the element level into the sub-array
level depends mainly on the efficiency of the optical transmitters, and
again it is believed that problems such as power consumption and laser mode
beating noise can be solved without great difficulty in the near future.
In the more distant future, arrays which are based on compact optical
processors such as those presented by Koepl [10.9], Dolfi [10.10] and
perhaps Zmuda [10.11] can become a reality provided their mechanical design
and alignment requirements can be reliably met. These processors utilize
the parallel processing capability of light thus offering the potential for
lower beamforming cost-per-element than the distributed microwave
techniques. Finally integrated coherent beamforming techniques such as that
of Birkmayer [10.12] have also the potential to offer lower beamforming
cost-per-element, as integrated optical technology reaches a sufficiently
mature stage to allow for the realization of these techniques at low cost.

10.3 Suggestions for future work

Having completed this thesis, in the next paragraphs some suggestions will
be presented for future work which can be done in the area of optically
controlled phased array antennas. These are as follows:

10.3.1 Temperature compensation using electronic feedback techniques

Although the use of optics can offer mass savings in space-based arrays,
the major optical components, especially lasers and optical amplifiers
usually need some form of temperature compensation. Designers of optical
systems tend to provide this compensation using Peltier coolers and
heaters, a solution which results in excessive power consumption making the
use of optical techniques unattractive. So some future research is needed
on methods of electronically compensating for temperature induced optical
power variations at the optical networks using electronic feedback
circuits. These circuits should be designed with the aim of minimizing the
Chapter 10: Conclusions

power consumption. For example, in the field of optical amplifiers, the method of Ellis et al [10.13], [10.14], [10.15] should be reviewed and based on this review, efficient control circuits tailored for the requirements of phased arrays should be designed.

10.3.2 Design of lasers for the I.F. return links.

For the lasers used in the I.F. return links of phased arrays, there are two requirements which need to be fulfilled: The first is to present low threshold currents and good electro-optic conversion efficiencies. The second is that their spectral characteristics are such that they do not present any problems with mode-beating when signal combining is done in the optical domain.

So the first task of a future research effort should be to produce a mathematical model to analyze the problem of mode-beating and establish the spectral requirements for the optical sources. Then based on these results, the second task is to design or identify suitable lasers which can fulfil these requirements.

10.4 References for Chapter 10


[10.3] Zaglanikis C.D. and Seeds A.J. "The use of optical amplifiers for signal distribution in optically controlled phased array antennas" Invited paper from Hungarian Communications Journal, Special Issue on 'Microwave Optoelectronics'


[10.12] Birkmayer W. and Wale M.J. "Optical BFN for telecom satellites and/or SAR applications: analysis and results" Proceedings of European
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Space Agency workshop on advanced beamforming networks for space applications, Noordwijk, The Netherlands, 26-28 November 1991 pp. 3.7.1 - 2.7.11 (ESA publication number ESA-WPP-030)


APPENDIX A

EQUATIONS FOR TRUE AND NON-TRUE TIME DELAY
R.F. AND I.F. BEAMFORMING

In this appendix the equations for non-true time delay and true time delay beamformers are analyzed. It is shown that true time delays may take place at the I.F. (probably at a sub-array level) while only conventional phase shifting is needed for every element in the R.F. single tone signal before it is mixed with the I.F..

A.1. Equations for conventional (non true-time delay) beamforming

Suppose an array like that of Fig (2.4) radiates a single tone signal with equation \( \cos(\omega t) \) at an angle \( \phi \). The non true time delay phase shifters of the 2 elements introduce phase shifts \( \theta_A \) and \( \theta_B \) respectively to the output signals of the two elements. So at the plane wavefront of Fig (2.4) the two radiated signals have equations:

\[
S_A = \cos(\omega t + \theta_A + \beta \Delta l \sin \phi) \quad (A.1)
\]

\[
S_B = \cos(\omega t + \theta_B) \quad (A.2)
\]

where \( \beta = \omega/c \).

For them to have equal phase at the plane wavefront of Fig (2.4):

\[
\omega t + \theta_A + \beta \Delta l \sin \phi = \omega t + \theta_B \quad (A.3)
\]

i.e. \( \theta_A - \theta_B = \omega \Delta l \sin \phi/c \) \quad (A.4)

Thus the phase difference introduced by the phase shifter between the two signals is proportional to the frequency \( \omega \). If the frequency \( \omega \) changes, the phase difference has to change in order to steer the beam to the same angle.
Appendix A: Beamforming equations

A.2. Equations for true-time delay beamforming.

In the case of true-time delay phase shifting in equations (A.1) and (A.2) the angles introduced by the phase shifters, \( \theta_A \) and \( \theta_B \), are now proportional to the frequency \( \omega \). So they are replaced by \( K_A \omega \) and \( K_B \omega \) where \( K_A \) and \( K_B \) are constants. So equation (A.4) becomes:

\[
K_A \omega - K_B \omega = \omega \Delta l \sin \phi / c \quad (A.5)
\]

i.e. \( K_A - K_B = \Delta l \sin \phi / c \quad (A.6) \)

So in this case the beam scan angle is independent of the frequency \( \omega \).

A.3. Equations for true-time delay beamforming at I.F. level combined with conventional phase shifting in the R.F. level.

A third solution which is equivalent to the introduction of true-time delays in the array radiation signal is the introduction of true time delays in the I.F. signals together with non-true time delays in the R.F. signal. This solution is only possible of course when the mixing of the I.F. and R.F. take place at the array element level.

Suppose the I.F. signals of the elements have equations:

\[
I_A = \cos(\omega r t + K_A \omega t) \quad (A.7)
\]

\[
I_B = \cos(\omega t + K_B \omega t) \quad (A.8)
\]

where \( K_A \omega \) and \( K_B \omega \) are the true-time delays introduced at the I.F. level \( K_A \) and \( K_B \) being variables set by the controlling computer. Also suppose the single-tone R.F. signals have equations:

\[
R_A = \cos(\omega r t + \theta_A) \quad (A.9)
\]

\[
R_B = \cos(\omega r t + \theta_B) \quad (A.10)
\]

where \( \theta_A \) and \( \theta_B \) are the non-true time delays introduced by conventional phase shifters at the R.F. level. Mixing the two signals one of the two
mixing products has the form:

\[ P_A = \cos(\omega_R + \omega_I)t + K_A \omega_I + \theta_A \]  \hspace{1cm} (A.11)

\[ P_B = \cos(\omega_R + \omega_I)t + K_B \omega_I + \theta_B \]  \hspace{1cm} (A.12)

At the plane wavefront of Fig (2.4) the two signals have equations

\[ P_A = \cos(\omega_R + \omega_I)t + K_A \omega_I + \theta_A + \frac{\omega_R + \omega_I}{c \Delta l \sin \phi} \]  \hspace{1cm} (A.13)

\[ P_B = \cos(\omega_R + \omega_I)t + K_B \omega_I + \theta_B \]  \hspace{1cm} (A.14)

So to cancel out the factor \( \frac{\omega_R + \omega_I}{c \Delta l \sin \phi} \):

\[ \frac{\omega_R + \omega_I}{c \Delta l \sin \phi} = \theta_B - \theta_A + (K_B - K_A)\omega_I \]  \hspace{1cm} (A.15)

The above equation can be divided into two equations which have to be simultaneously satisfied:

i) \[ \frac{\omega_I}{c \Delta l \sin \phi} = (K_B - K_A)\omega_I \]  \hspace{1cm} (A.16)

or \[ \frac{1}{c \Delta l \sin \phi} = K_B - K_A \]  \hspace{1cm} (A.17)

which is readily satisfied if the array's computer sets \( K_B - K_A \)
equal to \[ \frac{1}{c \Delta l \sin \phi} \]

ii) \[ \frac{\omega_R}{c \Delta l \sin \phi} = \theta_B - \theta_A \]  \hspace{1cm} (A.18)

but here \( \omega_R \) is a single tone signal and the equation can be satisfied simply if the array's computer sets \( \theta_B - \theta_A \) equal to \[ \frac{\omega_R}{c \Delta l \sin \phi} \) so it can be seen that if the phase shifting is done before the final mixing,
conventional phase shifters are sufficient for the R.F. level while true-time delay ones are needed for the I.F. level. Furthermore since $\omega_R >> \omega_I$ the bandwidth of $\omega_I$ equation (A.15) suggests that a true-time delay phase shifter is needed for every sub-array only and not for every element of the array. The maximum dimensions of the sub-array which can be handled from a single true-time delay phase shifter depends on the specifications of the particular system under consideration.
APPENDIX B

TYPICAL OPTO-ELECTRONIC COMPONENT PARAMETERS
FOR THE MODEL OF CHAPTER 3.

B.1. PIN photodiode

\[ \eta = \text{quantum efficiency of photodiode} = 0.4 \]
\[ \lambda = \text{wavelength of light} = 830 \text{ nm} \]
\[ T = \text{temperature} = 16 \, ^\circ\text{Celsius} \]

B.2. Semiconductor laser

\[ \text{RIN} = \text{relative intensity noise} = -140 \, \text{dB/Hz} \]
\[ I = \text{input current} = 90 \, \text{mA} \]
\[ I_{\text{th}} = \text{threshold current} = 60 \, \text{mA} \]
\[ R = \text{laser’s internal impedance} = 3 \, \Omega \]
\[ R = \text{matching impedance} = 47 \, \Omega \]
\[ a = \text{laser’s conversion constant} = 0.15 \, \text{W/A} \]
\[ T = \text{temperature} = 16 \, \text{degrees Celsius} \]

B.3. Electro-optic modulator

\[ V = \text{half wave voltage} = 2 \, \text{Volts} \]
\[ R = \text{modulator’s input impedance} = 50 \, \Omega \]
\[ R + R = \text{circuit’s input impedance} = 50 \, \Omega \]
\[ a_{\text{nt}} = \text{optical power loss coefficient} = 4 \, \text{dB} \]
\[ T = \text{temperature} = 16 \, \text{degrees Celsius} \]

B.4. Semiconductor laser amplifier

\[ \text{input coupling efficiency} = 5 \, \text{dB} \]
\[ \text{output coupl. efficiency} = 5 \, \text{dB} \]
\[ R = \text{input reflectivity} = 0.01 \]
\[ R = \text{output reflectivity} = 0.01 \]
\[ m = \text{effective number of modes} = 6 \]
Appendix B: Typical component parameters for the model of chapter 3

L = cavity length = 300 $\mu$m
\( n_e \) = effective index = 3.5
\( \alpha_c \) = loss coefficient = 1500 m$^{-1}$
\( n_tr \) = transparency carrier density = \( 1.50 \times 10^{24} \) m$^3$
\( \Gamma \) = confinement factor = 0.5
B = gain constant = \( 3 \times 10^{-20} \) m$^2$
Q = gain curve q factor = 100
\( \lambda \) = wavelength of light = 830 nm


<table>
<thead>
<tr>
<th>OPT. POWER (dBm)</th>
<th>GAIN (dB)</th>
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<tbody>
<tr>
<td>-50</td>
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<tr>
<td>-40</td>
<td>23.7</td>
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<tr>
<td>-30</td>
<td>22.5</td>
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<td>-20</td>
<td>17.7</td>
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<tr>
<td>-10</td>
<td>11.2</td>
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APPENDIX C

DETAILS OF THE OPTICAL LINK USED AS EXAMPLE IN THE DISCUSSION

The optical link used as an example in the discussions of Chapters 4 and 7 is the proposed forward link of a satellite-based Synthetic Aperture Radar (SAR) used to take radar images of the earth’s surface. The design has been developed by Marconi Space Systems for the European Space Agency. The specifications for the link are that a 5 - 5.3 GHz R.F. signal needs to be distributed to 512 T/R modules. The transmit signal consists of a chirped 5.3 GHz signal and the local oscillator is a continuous single tone of 5 GHz. The signal to noise (S/N) ratio of the output signal should be at least 113 dBc Hz. The input and output signal levels should both be 0 dBm to provide a transparent system. The basic characteristics of the proposed and experimentally demonstrated link [C.1], [C.2], [C.3] are the following:

One pigtailed laser (Ortel 1510B) operating at 1300 nm feeds 32 T/R modules. The conversion efficiency of the laser a is 3.7 W/A. The threshold current of the laser is 15 mA and its RIN is less than -140 dB/Hz for the frequency range of interest. The laser is biased at 40 mA and the maximum R.F. signal into the laser is 11.3 dBm. Two optical splitters are used in the link, one to divide the signal into 2 ways and one to divide it into 16 ways (2 X 16 = 32). Apart from the splitting losses the splitters present excess losses of 0.5 dB optical loss for the 2 way and 1.5 dB for the 16 way unit. Finally there are 4 optical connectors with about 1 dB optical loss each. On the other side of the link a pigtailed PIN photodiode is used which has a responsivity of 0.7 A/W. This corresponds (if the coupling losses are considered negligible) to a quantum efficiency of 0.716. The photodiode is reactively matched to the first transimpedance R.F. amplifier resulting in an improvement of 8 dB in the signal as compared with its performance using a 50 Ω load. The photodiode is followed by two R.F.
Appendix C: Details of the link used as example in the discussion

amplifiers. The first has a gain of 24 dB and a noise figure of 3.65 dB\(^1\). The second amplifier has a variable gain of 41 dB maximum and a noise figure of 5 dB. With these parameters the authors experimentally obtain a noise power of -115 dBm/Hz which results in a S/N ratio of 115 dBC Hz. The above link parameters have been entered into the computer model whose equations were described in Chapter 3. In the model to obtain the required output signal power level of 0 dBm the gain of the second (variable gain) amplifier had to be set to 31 dB. Also for the load resistance of the photodiode to give a gain of 8 dB as compared to the 50 \(\Omega\) load, in the model this had to be set to 316 \(\Omega\). Under these conditions the model gave an output S/N ratio of 115 dBC Hz which exactly agrees with the experimental result from the authors. The R.F. and optical signal and noise levels predicted by the computer model for every component in the link can be seen in Table (C.1).

References for Appendix C


\(^1\) As the noise figure of a transimpedance amplifier depends on the impedance of the input source to it, normally four parameters are needed to characterize it [C.4]. In this analysis it will be assumed that the authors have measured the noise figure at the frequency of interest using a source of impedance equal to the effective impedance of the input circuit to the amplifier (photodiode plus matching network).
<table>
<thead>
<tr>
<th></th>
<th>SIGNAL GENERATOR</th>
<th>FIRST R.F. AMPLIFIER</th>
<th>ORTEL LASER</th>
<th>4 CORNING CONNECTORS</th>
<th>1:2 SPLITTER</th>
<th>CORNING 1:16 SPLITTER</th>
<th>LASERTRON ODE-07SC</th>
<th>R.F. PHOTO. AMPLIFIER</th>
<th>SECOND R.F. AMPLIFIER</th>
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Table (C.1) Modelled power and noise levels for the forward link
Appendix C: Details of the link used as example in the discussion

Signal distribution and interface system demonstrator manufacture", technology definition of an advanced SAR study, GEC Marconi Research Centre, Y/256/170, December 1989.\(^2\)


\(^2\)References [C.2] and [C.3] are not in the public domain and permission has been obtained from Marconi Space Systems and the European Space Agency to use some of the information contained in them.
APPENDIX D

NUMERICAL VALUES FOR THE PARAMETERS USED IN THE MODELLING OF THE HITACHI HLP-1400 LASER (CHAPTER 6)

\[ \lambda = \text{free space wavelength} = 830 \text{ nm} \]
\[ L = \text{cavity length} = 300 \ \mu\text{m} \]
\[ n_{tr} = \text{transparency carrier density} = 1.5 \times 10^{24} \ \text{m}^{-3} \]
\[ a_{sc} = \text{loss coefficient of the guided mode} = n_g/(c\tau_{ph}) = 1500 \ \text{m}^{-1} \]
\[ n_g = \text{group index} = 4 \ \text{*1} \]
\[ n_e = \text{effective index} = 3.5 \ \text{*2} \]
\[ c = \text{light speed} = 3 \times 10^8 \ \text{m/sec} \]
\[ \Gamma = \text{confinement factor} = 0.5 \]
\[ B = \text{gain constant} = 3 \times 10^{-20} \ \text{m}^2 \]
\[ \tau_{sp} = \text{electron lifetime due to spontaneous photon emission} = 4 \ \text{nsec} \]
\[ e = \text{electron charge} = 1.6 \times 10^{-19} \ \text{C} \]
\[ d = \text{active layer thickness} = 0.1 \ \mu\text{m} \]
\[ W = \text{active region width} = 3 \ \mu\text{m} \]
\[ \omega_o = \text{peak angular frequency of the amplifier} \]
\[ Q = \text{gain curve Q factor} = 100 \]

*1 The group index gives the group velocity of the light inside the optical waveguide. So it determines the longitudinal mode spacing around the lasing frequency.
*2 The effective index gives the phase velocity of light at the laser frequency. So it is a measure of the absolute frequency of oscillation of the device.
APPENDIX E

PUBLICATIONS


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