Design Considerations of a
Submarine Laser Communications System

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After almost twenty years of research and technology development, it now appears as though the Navy is ready to initiate the development of an optical communication system between satellites and submarines. This system will operate in the blue/green region of the spectrum. A great deal of the past work has been to develop the critical components that will allow operations through clouds and water to submarines at depth, and to determine some of the key environment quantities that will be encountered. In this paper we will show how the theoretical efforts in the area of optical communications must be merged with the practical limitations that will occur under operational conditions. We will go through a sample design of such a system taking into account the theory, the components, the environment and sample requirements.

1.0 LASER COMMUNICATIONS IN
THE OPTICAL SCATTER CHANNEL

To achieve optimum system performance in any proposed optical communications link, the amount and form of the anticipated received background, and information-carrying radiance distributions must be known [1, 2]. Ideally the incoming information signal would be deterministic and only the various noise sources incident on the communications receiver would be stochastic in nature. Under these circumstances there would be no need to write this paper. Classical detection theory [3, 10] would be more than sufficient to optimally configure the desired communication system. However in real-life, all optical transmission channels affect an information exchange in a non-deterministic fashion to some extent and this must be taken into account in the system design to assure reliable communications can occur with

reasonable link availability. The result is receiver configurations far more complex and computationally-intensive than classical theory implies. This paper provides the necessary tools for designing communications systems in the optical scatter channel.

Diffraction-limited laser communications systems designed for the atmospheric and marine channels are inherently line-of-sight links, and any interaction with their particulate constituents will result in large system losses in the tens and hundreds of decibels [20, 21, 22, 23, 24, 27]. The impact of these large losses is an on-line availability for a single receiver in any proposed system of at most 40-70% in the atmospheric scatter channel, and only over short distances in the marine environment [22,25,26]. Fortunately much of the lost signal energy is potentially recoverable through proper receiver design, and this will increase both link availability and transmitter/receiver separations in these channels. Direct detection communication systems benefit the most since they only require a growth in the basic receiver design parameters to benefit in link performance [22,25,26,27]. Heterodyne communication systems response the least as the inherent phase and temporal coherence of the initial laser radiation is totally lost after a few particulate scatterings. The conclusion is that direct detection communications are the most desirable in the atmospheric and marine optical scatter channels. This is the design emphasis which will be pursued in the remaining parts of this section.

1.0.1 GENERAL SYSTEM CONSIDERATIONS
FOR DIRECT DETECTION LINKS

Gagliardi and Karp have shown that M-ary pulse-position-modulation (PPM) is a near optimal coding for background-limited optical communications [10, 11], and it also appears to be the best approach for communicating in the optical scatter channel [22]. In this subsection

we will review this powerful coding technique and show how one performs a first order design of such a direct detection system.

Figure 1.1 illustrates synchronous PPM and its important relations.

SLOT TIME = ΔT
 DEAD TIME = $N\Delta T$
 INFORMATION SLOTS = M
 AVERAGE PULSE INTERVAL = $(M+N)\Delta T$
 BITS PER PULSE = $\log_2 M$
 AVERAGE DATA RATE = $\log_2 M / (M+N)\Delta T$

Figure 1.1 Pulse-position-modulation (PPM) and its important relationships

In a basic time period called the interpulse interval, one determines in which of M slots of width ΔT seconds a laser pulse resides. For this discussion, we will ignore pulses overlapping more than one slot, i.e. pulse broadening by the optical scatter channel is assumed minimal for the present. The difference between the interpulse interval and $M\Delta T$, the frame time, is called the deadtime and is denoted $N\Delta T$. In most situations, N is not necessarily an integer. This latter time is usually associated with the time required for the laser to recharge or for some other such system consideration. The value of M determines the number of bits transmitted per pulse. Specifically M equals 2^k , where k is the number of bits sent per pulse; conversely, k equals $\log_2 M$. Figure 1.2 depicts the binary word (001) in an octary PPM scheme.

Figure 1.2 The binary word (001) in 8-ary PPM.

It is important to note that this type of coding assumes absolute time synchronization and has an average data rate equal to

$$\log_2 M$$

$$(N+M)\Delta T$$

A variation of the above is asynchronous PPM, which generally goes by the name of differential pulse-interval modulation (DPIM) [22]. Figure 1.3 shows DPIM and its important relations.

SLOT TIME = ΔT
 DEAD TIME = $N\Delta T$
 INFORMATION SLOTS = M
 AVERAGE PULSE INTERVAL = $(N+M/2)\Delta T$
 BITS PER PULSE = $\log_2 M$
 AVERAGE DATA RATE = $\log_2 M / (N+M/2)\Delta T$

Figure 1.3 Differential pulse-interval-modulation (DPIM) and its important relationships.

In this scheme, an initial pulse is sent and the time difference between it and a subsequent pulse, minus $N\Delta T$, determines which of M slots contain the signal. This later pulse is then used as the time reference for the next pulse and so on (See figure 1.3). The interesting aspect of this coding scheme is that the average frame time is $M\Delta T/2$. This implies that

the average data rate of

$$\frac{\log_2 M}{(N+M/2)\Delta T}$$

It is apparent that DPIM conveys information faster on the average than synchronous PPM. Unfortunately this higher data rate has to be traded off against the fact that DPIM requires one extra pulse and a probability of bit error P_E equal to one half of that needed for PPM for equivalent quality of service [22]. This is sometimes worth the price, but only a detailed look at the application will say. Once a coding scheme has been selected, the optimum value of M (or equivalently k) must be chosen. In general, the primary objective of any communication system is to convey as much information as feasible in the shortest possible time with the fewest anticipated errors. This first aspect of information transfer translates into optimizing the average data rate of the system; the second into the designing the communication link for the largest possible signal-to-noise ratio that the technology, and available funds and schedule can support. Let us first look at how one determines M .

To maximize information transfer, one needs to determine the optimum average data rate R . However this selection is not arbitrary, but requires that

$$\frac{\log_2 M}{(N+M)\Delta T} \text{ for PPM} \quad (1.1a)$$

$R =$

$$\frac{\log_2 M}{(N+M/2)\Delta T} \text{ for DPIM} \quad (1.1b)$$

be optimized with $N\Delta T$ given a k constrained to be an integer value. Mathematically, this can be done as follows. Let

$$\begin{aligned} T_D &= N\Delta T \\ \Delta T &= \alpha T_D \text{ for PPM} \\ &= 2\alpha T_D \text{ for DPIM} \end{aligned}$$

where we have redefined the communication slotwidth in terms of the laser deadtime. Here α is some arbitrary constant to be

determined. The pulse repetition rate R_p of the system can then be rewritten as

$$R_p = \frac{1}{T_D(1+\alpha M)} \quad (1.2)$$

$$R_{pmax} = \frac{1}{T_D} \text{ for } \Delta T = 0 (= \alpha)$$

This gives an average data rate equal to

$$\begin{aligned} R &= R_p \log_2 M \\ &= R_{pmax} \left[\frac{\log_2 M}{1 + \alpha M} \right] \end{aligned} \quad (1.3)$$

or in terms of the maximum data rate

$$R/R_{pmax} = \frac{\log_2 M}{1 + \alpha M} \equiv \beta < 1 \quad (1.4)$$

To maximize R/R_{pmax} , eq (1.4) is differentiated and set to zero. This yields

$$(1 + \alpha M) \log_2 e = \alpha M \log_2 M$$

Manipulating this expression around, we obtain the following relationship

$$\frac{1}{\alpha} \log_2 e = M \log_2 (M/e) \quad (1.5)$$

Figure 1.4 shows α as a function of word length in bits per pulse (word lengths of 0 and 1 give negative values for α , which are unrealizable conditions, and are omitted).

Figure 1.4

Figure 1.5 gives the maximum normalized average data rate of β as a function of word length in bits per pulse.

Figure 1.5

It is apparent from this last figure that the normalized data rate moves to its asymptotic maximum value (equal to the number of bits per pulse sent) with a κ equal to 6. This means that one could potentially send as much information as one wanted by adjusting the slotwidth to the appropriate interval. Unfortunately, neither the optical scatter channel nor existing electro-optical and electronic technologies allow an arbitrary slotwidth to be employed. As might be expected, the former is really the limiting factor. The presence of clouds within any communications link could increase the initial source pulsewidth in the tens of nanoseconds to the tens of microseconds through particulate multiple scattering. Hence one might be forced to create a realtime adaptive communications link, fix the slotwidth to one or more sizes which will encompass a major portion of

the possible clear sky/cloud scenarios for the link, or perform some other modification to the initial system design to permit "optimal" communication. The determination of the best average data rate can be accomplished using graphical techniques.

The best word length for the non-adaptive, channel-limited communications link is found by maximizing the average data rate R given by eq (1.1) where both NAT and ΔT are given and κ is still constrained to be an integer value. As in the last discussion, NAT is defined by the operating characteristics of the source and the ratio of R/κ can not exceed the pulse repetition frequency (prf) of this source. However in this case ΔT is set by the optical scatter channel and the type of communication receiver approach used; in particular, its value(s) are predetermined. As an example of the required analysis, let us compare the performance of two candidate systems with equivalent source prf's and a slotwidth of 7 microseconds. The difference of the two systems comes from the inherent deadtime of the source. In figures 1.6 and 1.7 we have plotted average data rate (ADR) R versus the number of bits sent per pulse for a HgBr or down-converted XeCl laser [12,14] and a frequency-doubled Nd:YAG laser [15], respectively, assuming PPM communications.

Figure 1.6

Figure 1.7

In these figures, we have utilized eq (1.1) and the relation

$$R = (\log_2 M) (\text{prf}) \quad (1.6)$$

under various permissible prfs for each laser technology. Figure 1.6 shows source deadtime of 5 milliseconds for the excimer technology and figure 1.7 gives a 18.1 millisecond deadtime for the Nd:YAG technology [12-15]. Both plots assume a common prf of 30 Hz. Referring to figure 1.6, the intersection of the 7 microsecond slotwidth curve and the 30 Hz prf line is at k equal 8 bits per pulse (b/p). As k must be an integer, the actual number of bits per pulse which would be used in the system would be 12 bits per pulse in order not to exceed a maximum 30 Hz prf. Repeating the same analysis for figure 1.7, we find an optimum k equal to 11.088 b/p which again implies the same number of bits per pulse (12 b/p) as we found for the excimer technology. However this does not imply equivalent data rates. The average data rates for these two 12 b/p-communication systems would be 356.4 and 256.6 bits/second (b/s), respectively. In other words, the smaller deadtime of the excimer technology moves more information through the link than the Nd:YAG technology, even though both have maximum prf limits of 30 Hz. Thus two conclusions can be drawn from the above discussion:

- (1) For a fixed deadtime and a large slotwidth and prf variation range, use the largest number of bits per second the system will allow. This implies the smallest slotwidth in its possible range and the largest prf.
- (2) For fixed slotwidth, deadtime and prf, use the largest number of bits per

second which places the system closest to its prf limit.

Although these two axioms appear rather straightforward to apply, one must be careful in their application. Specifically if a high prf laser system is employed whose average power is modest, the resulting energy per pulse may be too small to provide any reasonable quality of service. This point introduces the other major design factor which must be considered concurrently with this data rate optimization; namely, transferring information with the fewest possible errors.

With the "optimum" value of k identified using the above analysis, the next step is to determine if enough signal energy is available to communicate with some fixed information quality or fidelity. The standard measure of communication is the system's probability of bit error P_E , as cited in our discussion on PPM and DPIM coding. When one defines a method of pulse detection and coding, an equation for P_E is then derivable. This in turn allows a required signal-to-noise ratio (SNR) to be defined for a specific level of bit error probability. Given this knowledge, the next step is to establish whether or not the channel can sustain an SNR at least to this required minimum level.

The pulse detection techniques available for PPM or DPIM communications are quite numerous and more a function of the channel effects and a system designer prejudice. For example, a PPM system can employ a threshold filter, threshold integrator, "greatest-of" integrator or other such technique to establish the presence of an incoming signal [3, 10, 11, 16]. A similar set of detection schemes exist for DPIM [10, 11, 16]. For the purposes of this discussion, we will assume a "greatest of" integrator for our pulse detection technique and PPM coding. In this situation, the received energy in each slot is integrated over the slottime and the slot yielding the largest value is designated as the "correct word" position for that interpulse interval. This is a "soft" decision technique meaning that all of the information in an interpulse interval is obtained before a decision is made. This contrast a "hard" decision scheme where the first slot to yield a detected energy level obeying

some predetermined criteria (e.g. the received energy level exceeds some fixed threshold) is selected as the correct word position. The expected mean power-signal-to-noise ratio for a post-detection integrator can be written as

$$SNR \approx \frac{0.22 (\eta e/h\nu) E_R^2}{c P_B t_m}$$

assuming background-limited operation [16]. In this equation, we have

$(\eta e/h\nu) \equiv$ Responsivity of the detector

$c \equiv 1.6 \times 10^{-19}$ coulombs

$E_R \equiv$ Received energy at the detector

$P_B \equiv$ Total Received background Noise Power

$t_m \equiv$ Time of peak for the incoming signal = $\Delta t/2.45$

$\Delta t \equiv$ Temporal bandwidth of the incoming signal

The probability of a bit error for a "greatest of" detection system is equal to

$$P_E = 1 - \frac{1 + \alpha}{\sqrt{2\pi - \alpha}} \int \exp\{-(x - SNR)^2/2\} (1 - \text{erfc}[x])^{M-1} dx \quad (1.7)$$

where

$$\text{erfc}[x] = \frac{1}{\sqrt{2\pi}} \int_x^\infty \exp\{-t^2/2\} dt \quad (1.8)$$

which reduces to

$$P_E \approx \frac{M-1}{\sqrt{\pi SNR}} \exp\{-SNR/4\} \quad (1.9)$$

for a $SNR \gg 1$ [5]. Figure 1.8 depicts the probability of bit error for this type of detection method as a function of SNR and various values of K . For example if a system is expected to operate with an $K = 12$ b/p and a bit error probability of 6.6×10^{-6} , then the required minimum SNR will be 19 dB.

Figure 1.8

One would now evaluate eq (1.7) for the worst communication scenario envisioned for this system to determine if this minimum SNR is achievable with a given set of electro-optical/electronic technology, or perform a trade-off study to determine a reasonable set of technology to do the job. This type of analysis can be performed for a DPIM system. However unlike the PPM approach which can operate with either a hard or soft decision format, DPIM can only operate with hard decision pulse detection formats and requires a larger minimum SNR level of equivalent fidelity performance, as noted earlier.

To put all the previous discussion in perspective, we shall perform a first order link budget design of an low data rate (hundreds of bits per second) air/space-to-subsurface communication system to illustrate how the material contained in this chapter comes into play in the design process. For this example, let us assume a 0.55 wavelength laser with a 5 millisecond dead time and an unspecified (as yet) prf, and an optical receiver [21] with a one meter aperture, and a one steradian field-of-view (in water).

1.1 AN AIR/SPACE-TO-SUBSURFACE COMMUNICATIONS SYSTEM DESIGN

It is clear from the previous discussion that the design of any reliable communication system is strongly dependent on the environment it must operate in. Thus one must approach this problem with a strong knowledge of the communications channel and all its signal degrading effects, as well as classical communication theory. There is no one design which will work for all situations and requirements. Each design must be approached as a new and unique challenge

which the engineer must overcome. Only through creative thought will the challenge be met. To illustrate this point we will walk the reader through the basic flow of an air/space-to-subsurface communication system design. Although somewhat cursory, it will give the reader a general feel of the fundamental considerations one must go through in planning a system for use in the optical scatter channel. For this discussion, let us assume a high flying aircraft- or satellite-based laser transmitter performing a step/stare pattern on a portion of the ocean surface directly overhead of a submerged optical receiver pointed upward. This implies that the incident spot on the ocean will be large enough to assume a planewave-like nature; i.e. the signal irradiance on the ocean surface is essentially uniform in magnitude. This assumption will simplify the equations used in our analysis while still providing the basic flow in performing a system design without loss in generality.

The optical scatter channel will affect performance of any communication system through the following three mechanisms:

- (1) the received peak power will be effectively reduced by the temporal and angular spreading induced by the particulates and air/sea interface within the channel,
- (2) the initial signal energy level will be attenuated by the particulate absorption and scattering experienced in traversing the channel, and
- (3) the wider system slotwidths required to accommodate the multipath time spreading of the incident signal allows more background power to be detected.

Each of the above directly reduce the received signal-to-noise ratio; but it is the first two mechanisms which will do the most damage. These properties will become clearer shortly during our design analysis.

In the optical scatter channel it is the atmospheric portion which can spread an initial signal the greatest, both in time and angle; specifically those situations when clouds, fog or rain are present within the link. Of these three entities, clouds occur the most frequently in nature, and they usually

have the most impact because of their physical size and particle density. Hence they will be the limiting factor in the availability and ultimate performance of the link. The bottom of a cloud or cloud layer resembles a lambertian radiator and the anticipated pulse spreading induced by these entities will range from tenths of microseconds to 200 microseconds. A wide field-of-view optical receiver can minimize this loss to a couple of dB. However the time spreading of the initial laser signal can not be accommodated so easily. As most laser transmitters generate pulses in the tens to few hundreds of nanoseconds, the reduction of peak power by this channel (proportional to the ratio of initial to received pulsewidths to first order) can be tremendous. The reason is that clouds have very little absorption, and this allows their associated poly-dispersion phase function to generate a large amount of reverberation, creating multipath in the incident light distribution. The rms scatter angles for clouds is in the 20-40 degree range. The question now is what value of cloud parameter do we design to? For this example we are interested in communicating with a submerged receiver. Figure 1.9 shows the cumulative distribution for optical thickness. For a 100% availability of the optical link, the communication system must be capable of transmitting through clouds of optical thickness of 200 or less.

Figure 1.9

Figure 1.10 depicts the anticipated multipath time spread Δt_c from a cloud layer, for a range of optical thickness up to 200.

Figure 1.10

Table 1 gives the relationship between optical and physical thickness used to produce this last figure. Here we have assumed $\langle \cos \Theta \rangle = 0.875$, $W_0 = 0.9999$ and the relation

$$\langle \Theta^2 \rangle \approx 2 [1 - \langle \cos \Theta \rangle] \quad (1.10)$$

(18, 19).

OPTICAL THICKNESS t	EXTINCTION LENGTH, b (m)
$0. \leq t < 11.2$	350
$11.2 \leq t < 18.5$	65
$18.5 \leq t < 27.5$	50
$27.5 \leq t < 36.5$	45
$36.5 \leq t < 44.0$	20
$44.0 \leq t < 63.0$	55
$63.0 \leq t < 97.0$	15
$97.0 \geq t$	25

Table 1 Relationship between cloud optical thickness range and its associated volume extinction coefficient.

The nonlinear relationship shown in figure 1.10 is a manifestation of the fact that all cloud types have characteristic optical thickness and extinction lengths ranges. What it implies is that a low optical thickness cloud layer does not necessarily mean small multipath time spreading, hence it may give a misleading availability projection if one is not careful. From this graph we find that an optical thickness of 200, which implies a physical thickness of 5 km, gives an anticipated time spreading of 104 microseconds and is the limiting condition for this example.

The pulse detection method assumed for this communication link will be a threshold, low-pass filter scheme [16].

The resulting signal-to-noise ratio in this case is given by

$$\text{SNR} \approx \left[\frac{(\eta e / h \nu) E_R^2}{e P_B} \right] [X_0 / t_e] \quad (1.11)$$

where

$$t_e \equiv \text{estimated time-of-peak for the incoming signal} \\ = \Delta t_{ce} / 2.45$$

$\Delta t_{ce} \equiv$ estimated pulsewidth of the incoming signal

$(X_0 / t_e) \equiv$ Post-Detection Gain Factor

$$= \frac{\mu_0 t_e}{2} (T / t_m) \exp \{-2\mu_0 T\}, \quad t_m = RC \quad (1.12a)$$

$$= \frac{2\mu_0 t_e}{(1 - \mu_0 t_m)} [e^{-\mu_0 T - T/t_m} (1 + \frac{T[1 - \mu_0 t_m]}{t_m})^2], \quad t_m \neq RC \quad (1.12b)$$

$$T = 2.5 t_e \equiv \text{observation time}$$

$$\mu_0 = 1/RC = 1/2 t_e$$

$RC \equiv$ Low-pass filter time constant

and the other parameters have been defined previously. Figure 1.11 illustrates the ratio of X_0 / t_e as a function of the actual received pulsewidth Δt_e for various estimated signal pulsewidths, Δt_{ce} . The quantity X_0 / t_e describes the dependence of the signal-to-noise ratio on the received signal pulsewidth for a given design.

Figure 1.11

Two things are clear from this graph. One, it is apparent in figure 1.11 that the shorter and closer the estimated and actual signal pulsewidths are, the larger X_0 / t_e . This is not too surprising a

result since the received peak power is inversely proportional to the pulsewidth of the incoming signal, and its value degrades for actual and design pulsewidth mismatches. However the relative response of the low-pass filter for pulsewidth mismatches appears to degrade faster for small slotwidths than with the large slotwidths. This is just an artifact of the normalization factor t_e . X_0 is the same for each of those plots and it is its ratio with t_e which gives it the different response curves. These curves bring up the second point; namely that large slotwidths allow more background noise power to be detected. The relative differences in peaks illustrate the impact on signal-to-noise ratio for this effect very clearly. On the otherhand, a large slotwidth low-pass filter degrades gracefully with the actual signal, on the order of 3-4 dB over a nominal two decade range. This insensitivity will become advantageous in our design example shortly. In any event this figure has shown that system performance can be significantly reduced by the peak power being distributed over a larger time interval by particulate multiple scattering and because of the normalization factor t_e , also by the increase in background noise power received. These are two of the major properties of the optical scatter channel noted above. The other property, channel attenuation reduction of initial signal energy levels, will become more apparent shortly.

The probability of pulse error for threshold detection of a PPM coded system is given by

$$P_{PPM} = 1 - Q_d [1 - Q_0]^{M-1} \quad (1.13)$$

with

$$Q_d \equiv \text{"Detection" probability} \\ = \text{erfc} [d_0 - \text{SNR}] \quad (1.14)$$

$$Q_0 \equiv \text{"False Alarm" probability} \\ = \text{erfc} [d_0] \\ d_0 = 1/2 \sqrt{\text{SNR} + \ln(M-1)} / \sqrt{\text{SNR}} \quad (1.15)$$

(3, 5). For DPIM coding, we have

$$P_{DPIM} = 1 - (1 - P_{PPM})^2 \quad (1.16a)$$

$$\approx 2 P_{PPM}, P_{PPM} \ll 1 \quad (1.16b)$$

The probability of bit error, P_E , can be

written as

$$P_E = \frac{M}{2(M-1)} P_P$$

where the appropriate P_P has been inserted for the coding scheme assumed. As one is generally interested in maximizing information transfer, we shall assume DPIM coding for the remainder of this example. This assumption also eliminates a system requirement of extremely accurate timing clocks and other related aspects of absolute timing of pulse receipt. Figure 1.12 shows the probability of bit error as a function of SNR for k between 6 and 16 b/p₀ for the purposes of our discussion, let us also assume a required probability of PPM pulse error less than or equal to 10^{-2} for our requisite quality of service.

To optimize the performance of our communication system, we need to decide a basic design philosophy. One approach is to optimize the system's data rate (ADR) with minimal impact to the required SNR under the worst case background conditions. This would imply a communications scenario where the cited cloud conditions exist with the sun illuminating this layer from directly overhead. The multipath time spreading under the worst cloud conditions was found to be 104.3 microseconds and this implies that

$$\Delta t_{ce} = 104.0 \mu s$$

which yields

$$T = 1.02 t_e = 106.3 \mu s$$

and

$$RC = 0.82 t_c = 85 \mu s$$

Figure 1.13 is a graph of the ADR for DPIM communications versus the number of bits sent per pulse for selected slotwidths and prfs, and a source deadtime of 5 milliseconds.

Figure 1.13

Referring to this figure it is evident that the values of κ which optimize the average data rate for prfs equal to 100, 75, 50 and 30 Hz are 7, 8, 9 and 10, respectively, given 106 microsecond slotwidths. Their corresponding data rates are 594, 431, 280 and 169 b/s, respectively. A trade-off is now possible between spot-dwell time, message length and prf, or spot-dwell time and message length for a fixed prf source. Given these communication aspects have been resolved, the next part of the analysis involves determining if the prospective link can support the minimum signal-to-noise ratio over the assumed availability of the channel, 100%.

The probability of bit error for DPIM is given by

$$P_{E,DPIM} = \frac{M}{M-1} P_{P,PPM} = 10^{-2} \left(\frac{M}{M-1} \right) \quad (1.17)$$

Table 2 gives the signal-to-noise ratio SNR for $\kappa = 7, 8, 9$ and 10 for values of P_E near its assumed required value.

κ	PE	SNR REQ
7	1.17×10^{-2}	16.3 dB
8	1.14×10^{-2}	16.5 dB
9	1.13×10^{-2}	16.7 dB
10	1.11×10^{-2}	16.8 dB

TABLE 2 Required SNR for various values of κ and a $PE \approx 10^{-2}$

By adding 4 dB to these numbers, we are able to provide the necessary margin for estimated/received pulsewidth mismatch and have a single design which will operate over the entire anticipated environmental climatology range. For convenience, let the required SNR be 21 dB for the rest of

this discussion. For the sun directly overhead and illuminating a cloud layer of $\tau=200$ and $z=5$ km, the specific forms of E_R and P_0 given in eq (1.11) are given to first order by

$$E_R = E_P \gamma_R \left(\frac{d_o^2}{d_s^2} \right) T_a T_c T_{a/s} T_w f_{sq} (\Theta_{FOV}) \quad (1.18)$$

and

$$P_B = H_{su} B_{opt} \gamma_R \left(\frac{\pi d_o^2}{4} \right) T_a T_c T_{a/s} T_w f_{sq} (\Theta_{FOV}) \quad (1.19)$$

respectively, where

E_P = Optical energy per laser pulse (Joules)

H_{su} = Exoatmospheric solar irradiance (watt/micrometer-meter**2)

B_{opt} = Spectral bandwidth of the receiver

γ_R = Transmittance of the receiver optics

d_o = Diameter of the receiver clear aperture (meters)

d_s = Emerging source diameter from the cloud layer
 $= (d_1^2 + 4.7z^2 \tau^{-0.16})^{1/2} \quad (1.20)$

T_a = Atmospheric transmittance to top of cloud
 $= \exp \{-0.357\} \quad (1.21)$

T_c = Cloud transmittance

$T_{a/s}$ = Air/sea Interface transmittance

T_w = Irradiance transmittance to depth
 $= \exp \{-kD\} \quad (1.22)$

k = Diffuse attenuation coefficient (inverse meters)

D = Receiver depth (meters)

f_{sq} = Finite FOV receiver collection factor

In the above we have ignored the additional spatial and temporal spreading from the radiance propagation from the bottom of the cloud layer to the top of the air/sea interface. For large diameter beams, this is a rather small effect and can be omitted without loss of generality. The waters off Hawaii are expected to be a IB water-type. Since our laser has a center wavelength of 0.55 micrometers, the diffuse attenuation coefficient in this case will be around 0.07 inverse-meters down to depths and

Wavelength (nm)	Jerlov Type									
	IB	IB-II		II	II-III		III			
425	0.042	0.051	0.061	0.071	0.081	0.100	0.120	0.140	0.160	0.180
430	0.040	0.049	0.059	0.069	0.079	0.097	0.115	0.135	0.155	0.175
440	0.038	0.047	0.056	0.064	0.073	0.091	0.109	0.127	0.145	0.163
450	0.036	0.044	0.052	0.060	0.068	0.085	0.102	0.118	0.134	0.150
460	0.035	0.042	0.050	0.057	0.065	0.081	0.097	0.112	0.127	0.142
470	0.033	0.040	0.048	0.055	0.063	0.077	0.092	0.106	0.120	0.134
475	0.033	0.040	0.048	0.055	0.062	0.072	0.082	0.099	0.111	0.123
480	0.035	0.041	0.048	0.056	0.064	0.077	0.090	0.103	0.116	0.128
490	0.038	0.045	0.053	0.060	0.067	0.079	0.091	0.103	0.115	0.127
500	0.042	0.049	0.056	0.063	0.070	0.081	0.092	0.103	0.114	0.125
510	0.047	0.053	0.060	0.066	0.072	0.082	0.093	0.104	0.115	0.125
520	0.052	0.057	0.063	0.069	0.075	0.085	0.095	0.105	0.115	0.125
525	0.054	0.059	0.065	0.070	0.076	0.086	0.096	0.106	0.116	0.125

Let us now see how channel attenuation affects system performance.

$$\begin{aligned}
10 \log [\text{SNR}] &= 10 \log \left[\frac{(\eta e / h \nu) E_p^2 \gamma_p^2}{e H_{\text{su}} B_{\text{opt}} (\pi d_0^2 / 4) \gamma_R t_e'} \right] \\
+ 10 \log \left[\frac{d_0^2}{d_s^2} \right] &+ 10 \log \left[\frac{X_0}{t_e} \right] + 10 \log [T_A] \\
+ 10 \log [T_C] &+ 10 \log [T_{\text{a/s}}] + 10 \log [T_w] + 10 \log [f_{\text{sq}}]
\end{aligned}
\quad (1.22)$$
$$\frac{(\eta e/h\nu) E_p^2 \gamma_R^2}{e H_{su} B_{opt} (\pi d_o^2 / 4) \gamma_R t_c'} = 1 \times 10^{19}$$

inherent gain of our communication system. Since we are assuming that link availability is totally dominated by a cloud layer existence, the atmospheric and air/sea interface transmittances, finite FOV receiver collection factors can be considered fixed channel losses. We will use -1.55 dB, -2.05 and -3.0 dB, respectively. These factors reduce the inherent system gain to 183.4 dB. Figure 1.15 illustrates the signal-to-noise ratio at depth for various incident spot sizes on the top of the worst case cloud layer for Jerlov IB water at $\lambda=0.55$ micrometers.

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or change and/or improve the assumed system design. For example if the time availability specification was reduced to a 90%, the resulting signal-to-noise ratio at depth plot is shown in figure 1.16. Here the worst case cloud layer has an optical thickness of 100 and a physical thickness of 2.5 km. It is apparent from this figure that the reduced time availability allows a broader range of spot diameter and receiver depths to be used. Other possible system performance solutions include choosing another wavelength laser to reduce the water loss and redesigning the system to be adaptive in nature.

Figure 1.16

For example, bad clouds do not exist all the time and one could design a system which utilizes remote sensed climatology data on cloud optical thickness to configure the optimal information transfer to filter bank receiver system. A change in operating wavelength chosen to penetrate the water better will reduce the slopes of the graphs in figure 1.15 and 1.16 to reflect an even broader range of spot diameter and receiver depths to trade-off. The conclusions one can draw are twofold: (1) These last two figure clearly demonstrate that absorption (from the marine channel) and scattering (from both the atmospheric and marine channels) attenuate the initial signal energy quite dramatically, as noted in the optical scatter channel above; and (2) System design is a complex trade-off between channel limitations, performance goals, and state-of-the-art technology, totally dependent on the creativity and intuition of the engineer involved.

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