Low probability of detection underwater acoustic communications for mobile platforms

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***Abstract* - Mobile underwater platforms, such as autonomous underwater vehicles (AUVs), use underwater acoustic communications to form a network for command and control. Direct sequence spread spectrum signaling can be used for multi-access communications. It also provides high processing gain for communications using a low source level. Probability of detection by an intruder is minimized due to the decreased signal-to-noise ratio outside the operation area. A simple receiver algorithm is presented for DSSS communications between mobile platforms in the presence of multipaths. This paper analyzes the required source level for a given operating area and the corresponding counter detection range by a intruder.**

I. INTRODUCTION

Three methods are known to provide underwater acoustic communications with a low probability of detection (LPD). The first method uses time-reversal to focus the sound at the intended receiver with a signal level up to 10logN dB higher than the adjacent locations, where N is the number of sources used to transmit the sound. The signal level in the surrounding area is assumed to be low and difficult to detect. The disadvantage of this method is that it requires a large aperture source array which is not practical, and the focusing depends on transmission of a high level probe signal which can be easily detected although it may be infrequently transmitted. The second method uses frequency-hopping frequency-shift-keying (FH FSK) signals. The signal bit occupies only one (or a few) frequency bin, which hops from frequency to frequency, so that the average signal level over the transmission band is low. The drawback is that the information bearing frequency bin has a high signal-to-noise SNR. Once the hopping pattern has been recognized, the message can be detected and likely intercepted as well. The most commonly used LPD communications use the direct sequence spread spectrum approach (DSSS). DSSS uses a code sequence to spread the symbols at the transmitter and a de-spreader (a correlator or a matched filter) at the receiver to recover the transmitted symbols. At a certain range from the source and beyond, the signal will be weaker than the noise so that it will not likely be detected by a commonly used energy detector. The intended receiver, knowing the transmitted code sequence, can, on the other hand, pick up the signal using a coherent processor, i.e., the de-spreader, which compresses the signal and boosts it above the noise level. This form of communications can be carried out using a low level source, depending on how much processing gain is derived from the matched filter. Furthermore, DSSS uses a pseudo-random signal that is noise like and hence less likely to be recognized and intercepted. The drawback is that the data rate is reduced by the length of the spreading code which is not a problem for autonomous underwater vehicles (AUVs) operating in a network, in which short messages need to be transmitted frequently for command and control of the vehicles. The DSSS method minimizes the probability of detection and interception of these messages by an unfriendly intruder. It provides multi-access communications between the vehicles whereby different vehicles use different spreading codes which are orthogonal to each other. It uses less energy by transmitting a low level signal.

Communications using DSSS signaling in an underwater acoustic channel faces several technical challenges. One is the multipath arrivals, which create severe inter-chip and inter-symbol interferences [1,2]. Decision feedback equalizer (DFE) and Rake receiver have been adapted for DSSS communications. To achieve precise symbol synchronization and channel equalization, high SNR signals are required. To achieve a minimal BER, multiple receivers are required. Recently, Yang et al have demonstrated two new approaches for low input-SNR DSSS communications between fixed nodes without requiring a multichannel DFE [1,2]. These methods have been demonstrated with at sea data using a single receiver, where minimal (<1%) bit error rate (BER) were achieved for in-band SNR as low as -11 to -14 dB.

DSSS signaling method uses code "orthogonality" to minimize interference between symbols. In the presence of multipath, the matched filter yields the channel impulse response modulated by the symbol sequence. The symbol phase is often path dependent and changing rapidly with time. At-sea data showed that the symbol phase error estimated from the matched filter output is often larger than allowed (i.e., the phase difference between the symbols), resulting in unacceptable bit errors even with high input SNR. However, the peak of the correlation of the matched filter outputs can be used to determine the

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relative (differential) phase between two consecutive symbols with high reliability. While this suggests that a phase-locked loop (PLL) can be used to track symbols with binary phases, it is much simpler to use differential phase shift keying (DPSK) symbols in practice. This is one of the methods proposed in [1,2], which works well with a single receiver. This method requires only "coarse" synchronization at the symbol level and is applicable to low input-SNR communications between fixed nodes. The other method estimates the impulse response from a pilot signal and convolves it with the matched filter output to estimate the symbol phase.

The estimation of differential phase between symbols by correlating the matched-filter outputs works well when the channel coherence time is longer than two symbol periods (two code sequences), i.e., when the channel impulse response hasn't changed much between the two symbol time frames. This assumption is supported by fixed source/receiver data as previously shown in [1,2] but not so by data from a moving source as shown below. The problem is that the differential phase between the symbols fluctuates rapidly with time for a moving source. The phase error (between the estimated and true differential phase) is often larger than ±90°, resulting a high BER even for DPSK signals.

Similar to [1,2], this paper develops a method for moving source/receiver by exploring the use of the code sequence of DSSS without requiring DFE as mentioned above. Using a modified code sequence, a pair of (coherent) energy detector is used to detect symbol transition (i.e., whether the next symbol is the same as the current). By not measuring the (differential) symbol phase, this method is insensitive to the rapidly fluctuating symbol phase problem and is quite effective with the moving source data. This method is demonstrated for low SNR communications involving mobile platforms using at-sea data. Considering cooperative AUVs working within a network as an example, one can then study the probability of detection (P_D) of the communication signals by a hostile interceptor as a function of range to the source. The result is expressed in terms of a counter detection range for a given probability of false alarm (P_{FA}).

II. LOW PROBABILITY OF DETECTION COMMUNICATIONS

Acoustic communications with an input signal much weaker than the ambient noise (e.g., -10 dB SNR in the signal band) are difficult to detect by an un-alerted, unfriendly listener, who has no prior knowledge about the signal. If the signals are like noise, they are difficult to decode without a prior knowledge of the structure of the signal. Communications of this type are often called covert communications when the probability of interception and detection by an unfriendly interceptor is low. The probability of interception and detection is generally a function of the SNR at the receiver. The probability of detection increases as the interceptor moves close to the source.

For low SNR communications, the transmitted symbols must be brought above the noise by signal processing so that they can be decoded. Thus an essential element of the communication method is the processing gain (PG). DSSS signal provides a PG which equals theoretically to the time-bandwidth product of the signal, or the length of the spreading code. The PG enables communications to a friendly receiver and prevents detection/interception by a hostile interceptor.

One notes that to avoid detection, one would not want to use a (loud) probe signal since it has a high chance being detected. Second, training data are undesirable, since the data rate has been slowed by the spreading code and training data take up an unnecessary overhead. Because the low data rate, one is interested more in the uncoded data rate aided with a simple error correction code. Third, to transmit a decent size message, a long packet or many packets will need to be sent. This requires an algorithm that is not limited by the short channel coherence time. Fourth, at low input SNR, precise symbol synchronization (at the chip sampling rate) is difficult. The receiver algorithm must be able to work with coarse synchronization. Fifth, the receiver algorithm needs to be robust under complex environmental condition. Sixth, from the operationally point of view, the receiver algorithm needs to be computationally simple to minimize the processing load. Lastly, it is highly desirable that the receiver algorithm works with a single receiver, as (widely-spaced) multiple sources/receivers are operationally unavailable.

III. COHERENT ENERGY-DETECTOR

In a multipath environment, the signal phase is influenced by interference between the multipaths. Medium fluctuation can cause symbol phase to change with time. For a moving source, the individual paths encounter, in addition, a different phase change (ray path-length change) due to source changing range. A higher rate of phase fluctuation is thus expected for a moving source than for a fixed source. The moving source data show that, after proper Doppler compensation, the relative (differential) phase between adjacent symbols is often larger than the phase difference between symbols even with a high input SNR. This causes incorrect symbol identification and bit errors using the method mentioned above.

For DSSS signaling involving a mobile platform, a new method, which is insensitive to the phase fluctuation, is described below which follows the same principle as the original matched filter described above. The difference is that it uses a different set of code sequences.

Let C be the original code sequence, and C_1 be the first half of the code sequence, and C_2 be the second half of the code sequence, i.e., $C \equiv [C_1, C_2]$. A new pair of

transition detector is proposed which uses the following sequence as the matched filter: $C_P = [C_2, C_1]$, and $C_N = [C_2, -C_1]$.

How this method works is shown by the schematic in Fig. 1 for DPSK signals. The data are first synchronized using the original code sequence and divided into overlapping blocks covering the symbol time period. Doppler shift is estimated for each symbol using a wideband ambiguity function by correlating the Doppler shift code sequence with the data. The peak of the ambiguity surface determines the Doppler shift as a function of the symbol number. The data is then corrected for Doppler shift including signal dilation/compression. The Doppler corrected data are divided into new blocks shifted by half of the original block length. The pair of energy detectors is applied to each block of data to decide whether the two symbols covered by the block are of the same kind or different kind as described above. If the adjacent two symbols are of the same kind, the matched filter using C_P will yield the impulse response as the original matched filter will, with a peak value M times larger than the sidelobe level, where M is the length of the code sequence. The matched filter using C_N, will yield a much smaller value since C_N does not match the data. On the other hand, if the two adjacent symbols are of the opposite kind, then the reverse will be true, since in this case, C_N will match the data whereas C_P does not. Thus by comparing the (total) energy of the two matched filter outputs one can determine whether the two adjacent symbols are of the same kind or not. This determines the DPSK symbol sequence.

This method is checked first by using the fixed source data as a test case. One finds a slightly degraded but still satisfactory performance compared with the first method mentioned above. The method is then applied to moving source data collected during the same (TREX04) experiment. The result is shown in Fig. 2 for the moving source data which is to be compared with the fixed source data using the same method.

IV. P_D AND COUNTER DETECTION RANGE

To motive the discussion in this section, one could consider a potential scenario involving a group of AUVs operating cooperatively with each other. Low source level will be used for DSSS signaling among the AUVS to minimize the detection of the communications signals by an interceptor assumed initially outside the operating area of the AUVs. Form the system point of view, the acoustic source level must be high enough to enable communications between neighboring AUVs (with minimum bit errors), and yet low enough to avoid detection by an interceptor at a distance away. The question is at what range will the detection by the interceptor become unavoidable.

The probability of detection (P_D) depends on the construction of the detector; the more the detector knows about the signal, the more features the detector can use to improve the detection. If the signal waveform is known, one can employ a coherent detector, e.g., a matched filter, which often yields a high P_D given the processing gain derivable from the signal. Noise-like signals are difficult to detect using a coherent detector as the signal is random and difficult to replicate compared with an impulse like signal. This reason favors the DSSS signaling as compared with, for example, frequency-hopping frequency-shift-keying signaling, which possess a certain hopping pattern that can be constructed from the data.

If the signal waveform is not known, one is left with an energy detector. The most common energy detector is the spectrogram method, which detects a sudden increase in the SNR as a function of frequency and time. This normally requires a fairly good SNR (say 5 dB) over the signal frequency band and is most effectively done by a human being who can interactively adjust the Fourier transform window to match the signal. A computer-aided detector needs to know the signal bandwidth and signal duration, as mismatch in bandwidth and duration between the signal and detector can significantly decrease the detection performance. The mismatch in bandwidth is self evident, as incorrect bandwidth results in loss of part of the signal energy and addition of more noise energy. Likewise, the signal duration also influences the detector performance. If the integration time is much longer than the signal duration, it effectively reduces the SNR as more noise is included in the detector output. Without knowing the signal duration, a narrowband detector is often used to detect narrowband energy and a broadband pulse detector (i.e., short integration time) is commonly used to detector a burst energy. If an alarm has been sounded based on the detector outputs, the data can be reprocessed using variable integration times to improve the detection performance.

The P_D for an energy detector depends very much on the input SNR. Normally, a detector is first alerted for a potential signal of interest, when the probability of an alert exceeds the recognition differential for the signal of interest; the probability of alert is also a function of input SNR. Pseudo-random noise-like DSSS signals with an input level lower than the ambient noise (e.g., SNR ~ -8 dB within the signal band) are difficult to detect by a narrowband or broadband energy detector assuming an un-alerted listener. As input SNR increases (as when source-receiver ranges decrease), the probability that it will be detected increases. From a system point of view, it is useful to define a counter-detection range such as the range below which the P_D will exceed 50% assuming a given P_{FA} of, say, 1%. However, there is no one value for the counter detection range as it depends on the signal level, signal transmission loss (TL), noise level and signal fading statistics which vary from ocean to ocean. A quantitative analysis of the P_D will be carried out below to illustrate the point.

The analysis involves two receivers. One is the intended (friendly) receiver and the other is the unfriendly

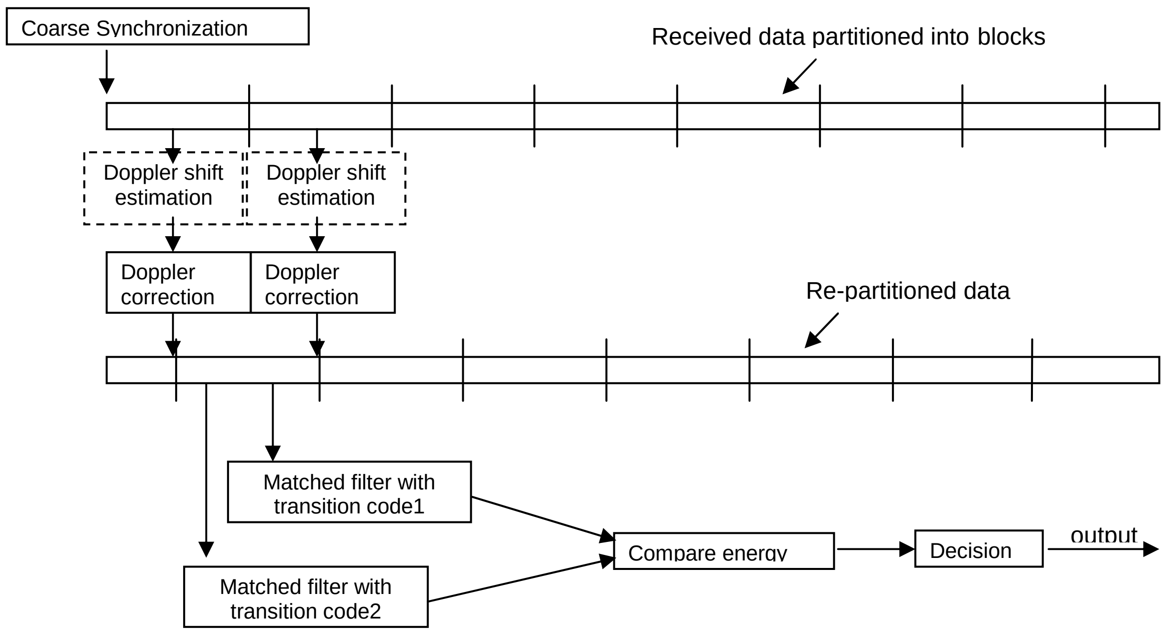

Figure 1. Block diagram of the transition detector method.

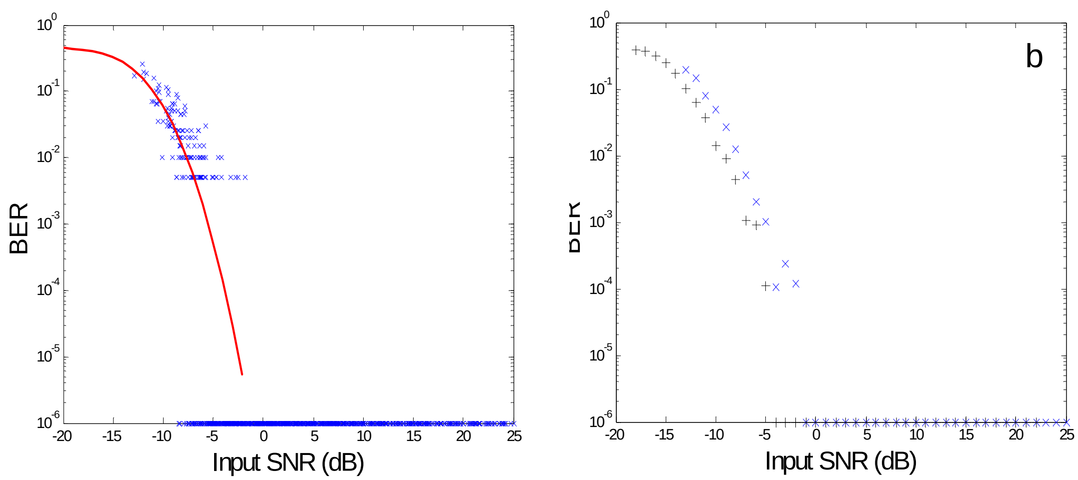

Figure 2. BER (a) and average BER (b) as a function of input-SNR using the transition detector method applied to the moving source data. 'x' and the fixed source data '+' of Ref. 2.

receiver, the interceptor. In general, the source level of the transmitter should be set high enough such that the message is received (at the intended receiver) with a high enough SNR for decoding the message, yet low enough to minimize the P_D by an interceptor. The analysis involves not only the P_D but also probability of false alarm (P_{FA}) as a function of the source-interceptor range for a given source level.

To model signal (amplitude) fading, we consider three cases: non-fading, Rayleigh-fading and lognormal-fading cases, from which one calculates P_D as a function of P_{FA} for a given input SNR, the so called receiver operation characteristic (ROC) curve, as shown in Fig. 3. For the lognormal fading, we use μ_s = -1.6449, and σ^2_s = 0.0304 as measured from at-sea data. Note in Fig. 3 that the P_D versus P_{FA} relation based on the log-normal statistics is much closer to the non-fading case than the Rayleigh fading case.

The above results can also be displayed in a different manner, for example, showing the P_D as a function of SNR for a given P_{FA}. As the input SNR is a function of source-receiver range for a given source level, one can calculate the P_D as a function of source-receiver range for a given P_{FA}. Counter detection range will be defined in this paper

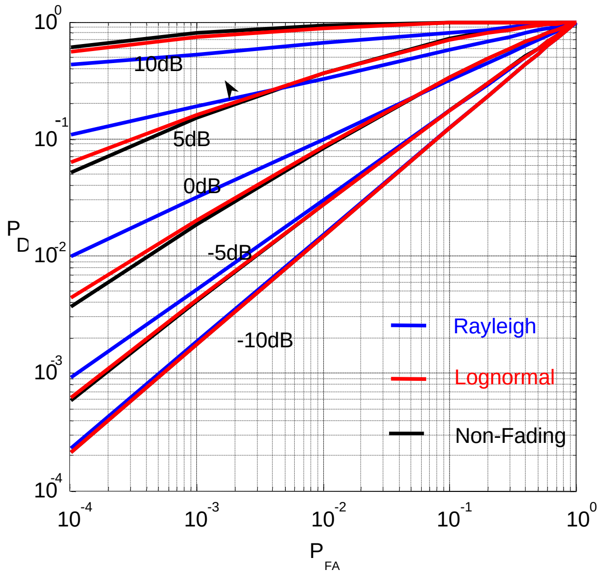

Figure 3. Receiver operation curves showing PD as a function of PFA for various SNR and signal fading distributions.

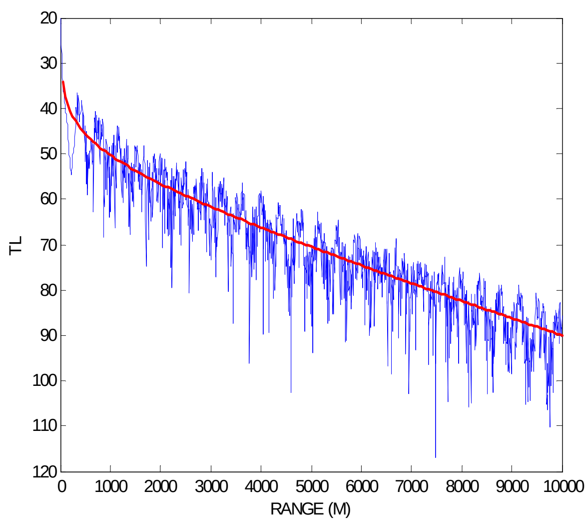

Figure 4. Transmission loss as a function of range using the RAM PE model. The smooth line is the empirical model calculation based on sound attenuation by sea water.

as the range below which P_D > 0.5 for P_{FA} = 0.01. To model the SNR at the receiver, we calculate the transmission loss (TL) for the TREX04 environment using the RAM PE model.[14] The result is shown in Fig. 4 for a 17 kHz signal. Superimposed is a sonar-equation-based calculation which assumes spherical spreading to a range of 50m and cylindrical spreading beyond 50m. For absorption by the water medium we use the following expression due to Thorp,[15]

$$\alpha = \left(\frac{0.1 f^2}{1 + f^2} + \frac{40 f^2}{4100 + f^2} + 2.75\text{x}10^{-4} f^2 + 0.003 \right) \text{x}0.875,$$

where α is the absorption loss dB/km, f is the acoustic frequency in kHz, and 0.875 is an depth correction factor determined by fitting the RAM calculation; the original formula is for a depth of 1000 m. To determine the SNR at the receiver, we note that at 17 kHz, the noise is dominated by wind generated noise, with a spectral level given by[16]

$$10 \log N(f) = 50 + 7.5 \text{w}^{1/2} + 20 \log f - 40 \log(f + 0.4),$$

where w is the wind speed in m/s. Assuming w = 10 m/s, one find NL = 49 dB. Given the TL and noise level, one can calculate the SNR as a function of range for a give source level (SL) from which one can calculate the P_D as a function of range for a given P_{FA} = 0.01. The result is shown in Fig. 5 for three source levels (SLs): 143, 155 and 164 dB, as needed for communications to an intended (friendly) receiver at a range of ~2, 4, and 7 km respectively. Figures 5a and 5b show the P_D as a function of range for Rayleigh signal fading and lognormal signal fading statistics respectively. One finds that the counter-detection range (for $P_D \geq 0.5$) is approximately 1.3, 3.6 and 5.8 km for Rayleigh signal fading and approximately 1.4, 3.8 and 6.1 km for lognormal signal fading statistics. Naturally, the higher the SL, the greater the counter-detection range. As expected, the counter-detection range for an energy detector is shorter than the communication range to a friendly receiver using a matched filter so that the communication signal will not likely be detected by an interceptor located outside the communication range (or the operation area). But as the interceptor approaches one of the transmitting nodes, detection by the interceptor is unavoidable. The hope is that the interceptor remains un-alerted, such as when the signal is noise like.

Note that at P_D = 0.5 the differences in the counter-detection ranges are relatively small between the two signal fading scenarios. However, if the detection ranges were defined at a higher P_D, e.g., P_D = 0.9, the differences between the two cases would be significant. The difference is that the P_D increases faster with decreasing range for the lognormal than for Rayleigh distribution.

V. SUMMARY

In this paper, a simple receiver algorithm is presented for DSSS signals from a moving source using only matched filters. This method is demonstrated with at-sea data with low ($<10^{-2}$) BER for input SNR as low as -8 dB using only a single receiver. The probability of detection and counter detection range are analyzed assuming an operating range of 2 km between mobile nodes. One finds that signal is unlikely to be detected unless the intruders get inside the operating area to a range ≤ 1.3 km from the transmitting node.

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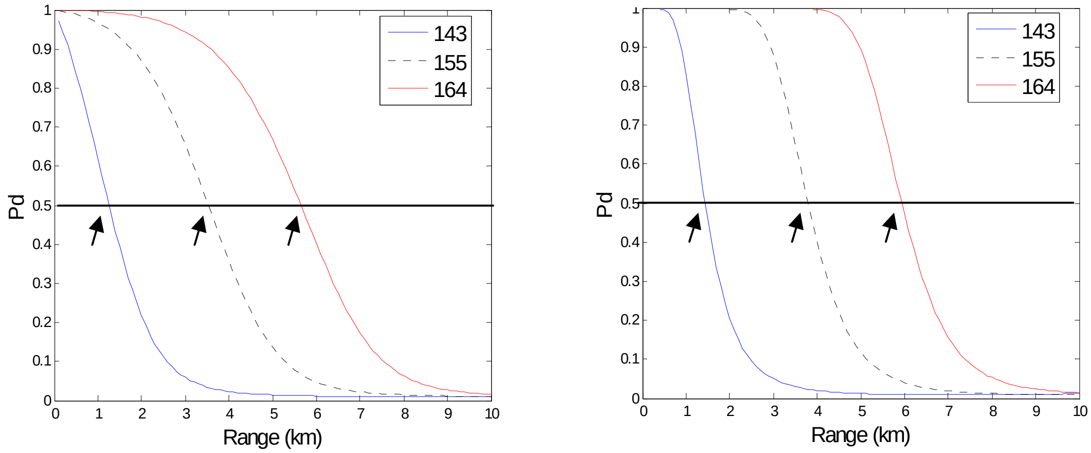

Figure 5. P_D as a function of source-receiver range for P_{FA} = 0.01 for three assumed source levels. Left figure assumes Rayleigh signal fading and right figure assumes lognormal signal fading.

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