New Potential of
Low-Frequency Radionavigation
in the 21st Century
COVER ILLUSTRATION:

3D representation of H-field measurements of local propagation phenomena around the Sunshine Skyway Bridge, Tampa Bay, Florida, USA
New Potential of

Low-Frequency Radionavigation

in the 21st Century

Proefschrift

ter verkrijging van de graad van doctor
aan de Technische Universiteit Delft,
op gezag van de Rector Magnificus prof. dr. ir. J.T. Fokkema,
voorzitter van het College van Promoties,
in het openbaar te verdedigen op
dinsdag 28 november 2006 om 10:00 uur

door
Wouter Johan PELGRUM
elektrotechnisch ingenieur
geboren te Bussum.
Dit proefschrift is goedgekeurd door de promotor:
Prof. dr. ir. L.P. Ligthart

Samenstelling promotiecommissie:

Rector Magnificus, voorzitter
Prof. dr. ir. L.P. Ligthart, Technische Universiteit Delft, promotor
Prof. dr. ir. D. van Willigen, Technische Universiteit Delft
Prof. dr. ir. F. van Graas, Ohio University
Prof. dr. J.D. Last, University College of North Wales
Prof. dr. B. Forssell, Norwegian University of Science and Technology
Prof. dr. A.G. Tijhuis, Technische Universiteit Eindhoven
Prof. ir. J.A. Spaans, Koninklijk Instituut voor de Marine
Prof. ir. P. Hoogenboom, Technische Universiteit Delft, reservelid


Copyright © 2006 by W.J. Pelgrum

All rights reserved. No part of this material may be reproduced or utilized in any form or by any means, electronic or mechanical, including photocopying, recording or by any other information storage and retrieval system, without permission from the author.
Summary

GPS (Global Positioning System) has enabled accurate, affordable, and almost ubiquitous positioning and timing. This has not only resulted in its wide spread usage, increased popularity, and numerous new applications, but it has also resulted in an increased dependency of the Global Navigation Satellite Systems. The ever-improving performance of GPS has long fueled the thought that GPS — and GPS alone — was to be the designated future of radio positioning and timing. The 2001 Volpe study, and later the 2004 proposed ERNP (European Radio Navigation Plan), stated otherwise: although very accurate, GPS and other satellite navigation systems are not considered reliable enough to be used as the sole-means for safety, environmental, and/or economically critical applications. Those critical applications need a backup system with dissimilar failure modes. The solution suggested by the Volpe-report and by the ERNP-proposal is — perhaps rather surprisingly — an old and almost forgotten radionavigation system: Loran-C. This system with its high-energy, low-frequency pulses is largely dissimilar to GPS. The combination of Loran-C and GPS, therefore, has the potential to be far more robust than either system individually. However, the “official” performance of the 1958 Loran-C system is no match for the stringent requirements of most modern applications. Fortunately, this Loran performance reflects the capabilities of outdated technology rather than the foundations of low-frequency radionavigation. The following question now arises:

What are the fundamental limits of low-frequency radionavigation and how do they affect potential applications?

Loran-C is currently the only operational and publicly available low-frequency radionavigation system with regional coverage. This dissertation, therefore, primarily focuses on Loran-C although most results will also be applicable to other low-frequency radionavigation systems. Chapter 2 introduces the system details of Loran-C.

The search for the fundamental limits begins by identifying the potential error sources. Chapter 3 contains a thorough system analysis, starting with the transmitter, and covering propagation, antenna, receiver algorithms, and concluding with calculated position and time.

Low-frequency ground waves experience delays as a function of ground conductivity, topography, seasons, and weather. These propagation delays can cause significant position errors if left uncompensated. Chapter 4 discusses the use of a differential reference station to compensate for the temporal fluctuations in propagation. A spatial correction map further reduces the propagation related positioning errors. The resulting positioning accuracy is potentially sufficient for such situations as the stringent 20 meter, 95% accuracy requirement of the maritime Harbor Entrance and Approach procedure.
Chapter 5 pays special attention to the H-field antenna; the chapter discusses error sources such as noise, E-field susceptibility, tuning, and cross talk thoroughly, as well as novel mitigation techniques and their successful implementation.

Chapter 6 discusses various measurement campaigns that bring the presented theory into practice. Throughout the Ph.D. research, the author of this work developed a highly accurate measurement system. The Reeuwijk measurements show the first step with precise dual-difference measurements; both the temporal domain and the spatial domain reveal local propagation effects. The land-mobile measurement campaign in Boston expands the measurement setup further. Simultaneous measurement of both E-field and H-field took place there allowing unprecedented analysis of re-radiation and an assessment of the applicability of low-frequency radionavigation in a land-mobile environment. The introduction of differential corrections and H-field antenna calibration for the Tampa Bay campaign resulted in an unprecedented measurement performance. Chapter 6 also shows the effect of bridges on positioning performance quantitatively, and presents the successful results of a unique re-radiation detection algorithm. This algorithm enables detection of local disturbances allowing a timely warning of potential erroneous position information. Finally, Chapter 6 shows the achievement of a positioning performance of better than 10 meters with 95% confidence during a realistic “Harbor Entrance and Approach” (HEA) scenario.

This dissertation concludes with an assessment of the potential of low-frequency radionavigation, based on the results of the author's Ph.D. research combined with his personal views.

Wouter J. Pelgrum
Delft, November 2006
## Contents

Summary ..................................................................................................................................... v

1 Introduction ....................................................................................................................... 1
   1.1 GNSS: almost perfect ................................................................................................. 2
   1.2 LF radionavigation in the 21st century ....................................................................... 4
   1.3 Research question .................................................................................................... 6
   1.4 Outline of dissertation ............................................................................................. 6
   1.5 References ............................................................................................................. 7

2 Loran System Description............................................................................................... 9
   2.1 Loran history .......................................................................................................... 10
   2.2 eLoran ..................................................................................................................... 11
   2.3 Loran System Characteristics ................................................................................ 11
      2.3.1 Loran-C system performance .............................................................................. 12
      2.3.2 eLoran system performance .............................................................................. 12
      2.3.3 Coverage ........................................................................................................... 13
   2.4 Loran-C Signal in Space definition ......................................................................... 14
      2.4.1 Pulse shape ....................................................................................................... 14
      2.4.2 System timing ................................................................................................... 16
      2.4.3 Positioning ....................................................................................................... 17
      2.4.4 Data communication ....................................................................................... 17
   2.5 Transmitter .............................................................................................................. 18
   2.6 Propagation ............................................................................................................. 20
   2.7 Noise and interference ............................................................................................ 21
   2.8 Receiver ................................................................................................................ 22
   2.9 References ............................................................................................................. 23

3 Loran-C Error Model .................................................................................................... 25
   3.1 From pseudorange to position .............................................................................. 26
      3.1.1 Geometry .......................................................................................................... 26
      3.1.2 Repeatable accuracy ....................................................................................... 27
      3.1.3 Absolute accuracy ........................................................................................... 27
   3.2 Overview of error model ....................................................................................... 28
   3.3 Transmitter timing ................................................................................................. 28
      3.3.1 SAM control ................................................................................................... 28
      3.3.2 TOT control .................................................................................................. 30
      3.3.3 UTC ............................................................................................................... 34
      3.3.4 Internal timing control .................................................................................. 35
   3.4 LF propagation ...................................................................................................... 36
      3.4.1 Ground wave versus sky wave ....................................................................... 36
      3.4.2 Spatial variations ............................................................................................. 41
5.2.4 \( R_{\text{loss}} \) ........................................................................................................... 127
5.3 Topology .................................................................................................................. 128
  5.3.1 Resonance configuration ..................................................................................... 128
  5.3.2 Wideband configuration ...................................................................................... 129
  5.3.3 Wideband versus resonance .............................................................................. 130
5.4 Noise ....................................................................................................................... 131
  5.4.1 Desired antenna noise performance ................................................................... 131
  5.4.2 Receiver noise analysis for resonance topology .................................................. 132
  5.4.3 Equivalent receiver-noise field-strength ............................................................. 135
  5.4.4 Orthogonal design ............................................................................................. 136
  5.4.5 H-field sensor implementation .......................................................................... 137
5.5 E-field susceptibility ............................................................................................... 142
  5.5.1 E-field susceptibility measurements ................................................................. 143
  5.5.2 E-field susceptibility compensation ................................................................. 146
  5.5.3 Balancing .......................................................................................................... 146
  5.5.4 Shielding ......................................................................................................... 148
  5.5.5 Rod construction .............................................................................................. 149
5.6 Tuning errors in a resonant H-field antenna ............................................................ 150
  5.6.1 Temperature coefficient of tuning capacitance .................................................... 150
  5.6.2 Temperature coefficient of antenna inductance .................................................. 151
  5.6.3 Temperature coefficient of the active part .......................................................... 153
  5.6.4 Measurement and calibration of tuning errors ..................................................... 153
  5.6.5 Common-mode versus differential-mode TOA errors ....................................... 155
5.7 Obtaining an omni-directional radiation pattern ....................................................... 156
  5.7.1 Quadrature addition .......................................................................................... 157
  5.7.2 Linear addition .................................................................................................. 160
5.8 Cross talk ................................................................................................................... 161
  5.8.1 Characterization of cross talk ............................................................................ 163
  5.8.2 Laboratory measurements of cross-talk parameters ........................................... 163
  5.8.3 Compensation for cross talk ............................................................................ 168
  5.8.4 Field measurements and compensation of cross-talk parameters ....................... 169
5.9 Discussion ............................................................................................................... 175
5.10 References .............................................................................................................. 176
6 LF Propagation Measurements ..................................................................................... 179
  6.1 Introduction .............................................................................................................. 180
  6.2 Reeuwijk measurements ....................................................................................... 180
    6.2.1 Spatial repeatability re-radiation measurements .............................................. 181
    6.2.2 Temporal stability re-radiation measurements ............................................... 187
  6.3 Boston highway measurements ............................................................................. 190
    6.3.1 Measurement setup .......................................................................................... 191
    6.3.2 Cloverleaf measurements ............................................................................... 193
    6.3.3 Measurements 495 south ............................................................................... 203
    6.3.4 Discussion .................................................................................................... 211
  6.4 Tampa Bay measurements ..................................................................................... 213
# Table of Contents

6.4.1 Measurement setup ............................................................................................................... 214
6.4.2 Reference station analysis .............................................................................................. 221
6.4.3 Measurements Tampa Bay area ...................................................................................... 223
6.4.4 Creation of ASF correction maps ...................................................................................... 225
6.4.5 E-field versus H-field in the far-field ............................................................................... 230
6.4.6 Re-radiation ...................................................................................................................... 231
6.4.7 HEA dLoran showcase ................................................................................................... 243
6.4.8 Discussion ...................................................................................................................... 246
6.5 General conclusions and recommendations .................................................................... 248
6.6 References ...................................................................................................................... 250

7 Conclusions, Discussion, and Recommendations .......................................................... 251
7.1 Conclusions ...................................................................................................................... 252
7.1.1 Loran-C System ............................................................................................................. 252
7.1.2 Transmitter .................................................................................................................. 253
7.1.3 Receiver ....................................................................................................................... 253
7.1.4 Interference .................................................................................................................. 254
7.1.5 Far-field propagation ................................................................................................. 255
7.1.6 Near-field propagation ............................................................................................... 255
7.1.7 In summary .................................................................................................................. 256
7.2 Discussion ....................................................................................................................... 256
7.2.1 Applications ............................................................................................................... 256
7.2.2 Is Loran-C the optimal LF radionavigation system? .................................................... 262
7.3 Recommendations ......................................................................................................... 265
7.4 References ..................................................................................................................... 267

List of Abbreviations and Acronyms ...................................................................................... 269

Appendix A RNP Definitions .................................................................................................. 273
Appendix B Applications and Their Requirements .................................................................. 277
Appendix C Timeline .............................................................................................................. 285
Appendix D Related Publications and Presentations by the Author ..................................... 287
Curriculum Vitae .................................................................................................................... 291
Samenvatting (Summary in Dutch) .......................................................................................... 293
Acknowledgements ............................................................................................................... 295
1 Introduction

A GNSS-independent backup is required that can provide *absolute* Position Velocity and Time (PVT), that is capable of bridging prolonged outages and that preferably can be used across different modalities (aviation, maritime and terrestrial). The cross-modal backup solution suggested by the Volpe study and by the proposed ERNP is, perhaps rather surprisingly, an old and almost forgotten radionavigation system: Loran-C.
Maritime and aviation users have been well acquainted with various radionavigation systems for a long time. These systems facilitate the users in safe and efficient travel by determining the position, velocity, and time from knowledge of the propagation of electromagnetic radio waves.

Traditionally, the required receivers were expensive, cumbersome and required a skilled operator. This made radionavigation the exclusive domain of professionals. The introduction of Global Navigation Satellite Systems (GNSS), in particular the Global Positioning System (GPS), has changed all that. Navigation has never been easier since GPS became fully operational in 1995. The system offers 3D positioning and time with global coverage; its use is straightforward, and can be considered “plug-and-play” in most situations. Furthermore, since the discontinuation of Selective Availability (SA) and the introduction of Space Based Augmentation Systems (SBAS), the performance of even consumer grade GPS units is extraordinary. Meter level positioning accuracy and better than 100 ns absolute timing is achieved using GPS OEM receivers costing less than $100.

The availability of such affordable high performance navigation sensors has made an enormous growth in navigation applications possible, both in a professional and in a consumer context. Possibly the most visible example of such an application is car navigation: affordable, compact, and easy to use systems providing door-to-door navigation for the general public. But the list continues: tracking packages, stolen goods, criminals, localizing emergency calls made from a cell phone (E911/E112) and also, for example, providing precise time and frequency for e.g. telecommunication networks and power grid synchronization. GPS has made the availability of Position Velocity and Time (PVT) affordable and almost ubiquitous. Because of that, PVT is often taken for granted and relied upon as such.

1.1 GNSS: almost perfect

Its ever-increasing performance made GNSS the “holy grail” of navigation. The United States Federal Radio Navigation Plan (FRP) of 1994 suggested gradual movement to sole-means satellite navigation in the year 2000 and hence abandonment most other existing aids to navigation.

Not everyone shared the same enthusiasm towards the increasingly critical dependency upon GPS in modern society. The President’s Commission on Critical Infrastructure Protection (PCCIP) concluded in 1997 that: “Global Positioning System (GPS) services and applications are

---

1 Global Navigation Satellite Systems (GNSS) consist among others of the American GPS, the Russian GLONASS and the future European Galileo system.

2 Until May 2000, the accuracy of the Standard Positioning Service (SPS) of GPS was intentionally degraded to 100 m horizontally (95%) under the so called Selective Availability (SA) program, for reasons of US national security.

3 Space Based Augmentation Systems (SBAS), such as the American WAAS, the European EGNOS and the Japanese MSAS, use a network of ground stations to determine differential corrections to the GPS pseudoranges, as well as signal-quality indicators. These are relayed to the user through geostationary satellites and used to improve the accuracy and integrity of his position solution.
susceptible to various types of interference. The effects of these vulnerabilities on civilian transportation applications should be studied in detail" [4][5].

The PCCIP findings were supported by the US Federal Aviation Administration (FAA) at the 1998 International Civil Aviation Organization (ICAO) meeting expressing the opinion that GPS “would not be approved for sole use navigation, and would need a backup” [6]. Also the 1999 US-FRP no longer advocated GPS as the sole-means solution to all navigation and revoked the earlier proposed termination of other navigation systems. Furthermore, the Volpe National Transportation Center⁴ executed the by the PCCIP suggested GPS vulnerability study and reported: “Backups for positioning and precision timing are necessary for all GPS applications involving the potential for life-threatening situations or major economic or environmental impact” [3]

Additionally, outside the US the same reserves against sole-means satellite navigation have been expressed, for example in the proposed European Radio Navigation Plan (ERNP): “The vulnerability assessment concluded that there is currently a strong reliance on GPS, and that fewer than 40 of the 137 applications analyzed would remain operational following the loss of GPS and its augmentations” [2].

These conclusions can be disillusioning: many crucial applications in current society critically depend upon GPS, a system that appears to be vulnerable. Therefore, sole-means reliance upon GPS is not an option for safety, environmental or economically critical applications; those applications require a backup system.

Fortunately, Europe is developing its own satellite navigation system: Galileo. This system will be operated independently from GPS and its slightly different design potentially even allows for superior performance. However, although the combined availability of GPS and Galileo will benefit the user, it will only modestly mitigate the vulnerability issue. Because Galileo is also a satellite navigation system, it possesses many of the same failure modes as GPS, which are rooted in the use of ultra high frequency, low power signals. Significant reduction of the chance of service disruption caused by signal blocking, intentional or unintentional interference requires a backup system with dissimilar failure modes.

The Inertial Reference System (IRS) is an excellent example of complete dissimilarity with satellite navigation. This system measures linear accelerations and angular rates in three dimensions. Given the starting position and orientation, the user can calculate his displacement by double integration of the accelerations. Due to the nature of the sensor, any measurement error has a cumulative effect on the calculated displacement. This limits the maximum coasting time through GNSS outages, especially when using low cost inertial sensors. Nevertheless, integration with inertial sensors shows significant potential for improving the robustness of satellite navigation. The latter is clearly expressed by the vast amount of research being conducted in this field. However, inertial systems are not a solution

---

⁴ The Volpe National Transportation Systems Center (RSPA/Volpe Center) conducted a vulnerability analysis of GPS and identified the potential impact to aviation, maritime transportation, railroads, and Intelligent Transportation Systems (ITS). The final report, Vulnerability Assessment of the Transportation Infrastructure Relying on the Global Positioning System was published on August 29, 2001.
for prolonged GNSS outages and they also cannot initialize without a start position and orientation, for example during a (prolonged) GNSS-denied situation.

An analog reasoning can be made for a GPS-independent time and frequency backup. Stable oscillators can coast through GNSS outages, the maximum coasting time obviously determined by the quality of the oscillator and the application’s requirements. A Rubidium oscillator performs significantly better than an oven controlled crystal oscillator but is also more expensive. Cesium technology allows for ultra high stability but its price is accordingly and mostly far out of reach for the majority of applications. Additionally, without help from a radio system, an oscillator can only provide frequency upon startup, not absolute time.

Concluding, a GNSS-independent backup is required that can provide absolute Position Velocity and Time (PVT), that is capable of bridging prolonged outages and that preferably can be used across different modalities (aviation, maritime and terrestrial). The cross-modal backup solution suggested by the Volpe study and by the proposed ERNP is, perhaps rather surprisingly, an old and almost forgotten radionavigation system: Loran-C.

### 1.2 LF radionavigation in the 21st century

The Volpe-report states that the low-frequency radionavigation system Loran-C is the only available potential cross-modal radionavigation backup to GPS [3]. This conclusion is endorsed by the proposed ERNP:

“A key issue is the vulnerability of satellite navigation services to signal jamming and unintentional interference. Mitigation strategies include the use of complementary services that have been allocated different parts of the spectrum, and the use of complementary non-radionavigation sensors or systems. (…) The stability and robustness of the current EU radionavigation service environment would be improved by the availability of GALILEO, EGNOS, and Loran-C services. (…) Loran-C services may mitigate system, service and user vulnerabilities that are currently dependent on GPS and its augmentations.” [2]

Low-frequency radionavigation systems are the ground-based predecessors of GNSS. In the pre-satellite era, high frequencies could not be used for radionavigation while they are limited to line-of-sight. Low frequencies however can propagate via ground waves, thereby providing stable signal reception far beyond the radio horizon. Unfortunately, the low end of the spectrum is cursed with extremely high atmospheric noise levels generated by thunderstorms all over the world. To overcome this nuisance, the Low Frequency radionavigation systems utilize extremely high transmitter powers. The difference in frequency and generated signal powers makes low-frequency radionavigation largely dissimilar and therefore complementary to GNSS. The combination of low-frequency radionavigation with GNSS has the potential to be significantly more robust than either system on its own.

Since the Second World War, many Low Frequency radionavigation systems have been in service, e.g. Loran-A (1950 kHz), Loran-C and Chayka (90-110 kHz), Decca (70-130 kHz) and Omega (10-14 kHz). The Loran-C system and its Russian counterpart Chayka are currently the only systems with regional coverage that are still in service. In this dissertation, these two systems are therefore used as representatives of low-frequency radionavigation. In most cases,
conclusions based on Loran-C are just as valid for other (hypothetical) LF radionavigation systems operating in the same frequency band and having similar sky-wave mitigation capabilities.

**Table 1-1: Dissimilarities between GNSS and Loran-C/ Chayka [7]**

<table>
<thead>
<tr>
<th>Property</th>
<th>GNSS</th>
<th>Loran-C / Chayka</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>Ultra High Frequency (1-2 GHz range)</td>
<td>Low Frequency (90-110 kHz)</td>
</tr>
<tr>
<td>Transmitters</td>
<td>Space based</td>
<td>Terrestrial</td>
</tr>
<tr>
<td>Transmitter power</td>
<td>Low (50-200W)</td>
<td>High (250 kW – 1 MW)</td>
</tr>
<tr>
<td>Signal structure</td>
<td>CDMA</td>
<td>Pulsed TDMA/CDMA</td>
</tr>
<tr>
<td>Signal Propagation</td>
<td>Line-of-sight</td>
<td>Ground wave / sky wave</td>
</tr>
</tbody>
</table>

The Loran-C system first became operational in 1958. Obviously, its original accuracy specification of a quarter nautical mile (463 m) is no match for the requirements of most contemporary applications. However, it is to be expected that an upgraded Loran-C system, “eLoran”, utilizing the latest technology will show significant performance improvements. Based on this potential, various reports have identified multiple applications for which an upgraded Loran-C can function as a backup for GNSS [3][2][1]:

For aviation, Loran-C is already approved as an en route supplementary air navigation system. Recent studies under direction of the FAA have concluded that a future eLoran is most likely also capable to meet the requirements of the more demanding “non-precision approach”-procedure [1].

The traditional maritime usage of Loran-C in open waters can potentially be extended into the harbor areas. Recent studies and measurements show promising results of eLoran meeting the most challenging accuracy demand of 20 m 95% of the “Harbor Entrance and Approach”-procedure. Even the future IMO (International Maritime Organization) requirement of 10 m 95% appears to be in range.

On land, eLoran has the prospect of e.g. aiding vehicle navigation and providing challenging track-and-trace capabilities.

Finally, eLoran has the capability to backup the time and frequency provision, which currently heavily relies upon GPS. The excellent Loran-C ground-wave stability allows for performances very close to that of satellite systems, potentially even when operated indoors.

These examples, which are further detailed in Appendix B, show the unique potential of eLoran as a cross-modal backup for GNSS.
1.3 Research question

In recent years, significant funding has been made available from US congress, NELS\(^5\) and industry for the upgrade process from “legacy” Loran-C to the new eLoran. This modernization process consists of the upgrading of the transmitter infrastructure, the development of modern integrated GPS-eLoran receivers and research of the capabilities of eLoran and its applicability as a GNSS backup. The results so far are promising: the already achieved increase in performance is impressive and robust integrated GNSS-eLoran receivers have been successfully demonstrated.

On the other hand, it has also become clear that low-frequency radionavigation will never achieve the performance of modern satellite navigation. The utilization of modern transmitter technology, state-of-the-art signal processing and elaborate computer modeling cannot fully mitigate all the nuisances of low-frequency radionavigation. For example, there will always be unknown aspects of the spatial and temporal variations of the far-field ground-wave propagation. Local objects often cause unpredictable near-field propagation phenomena and the presence of noise and interference in the low end of the radio spectrum is seldom insignificant.

This leads us to the research question of this dissertation:

What are the fundamental limits of low-frequency radionavigation and how do they affect potential applications?

1.4 Outline of dissertation

Loran-C is currently the only operational and publicly available Low Frequency radionavigation system with regional coverage. This dissertation therefore mainly focuses on Loran-C although most results will also be applicable to other low-frequency radionavigation systems. Chapter 2 introduces the system details of Loran-C.

The search for the fundamental limits is started by first identifying the potential error sources. Chapter 3 contains a thorough system analysis, starting with the transmitter, and covering propagation, antenna, receiver algorithms, and concluding with calculated position and time.

Low-frequency ground waves experience delays as a function of ground conductivity, topography, seasons, and weather. These propagation delays can cause significant position errors if left uncompensated. Chapter 4 discusses the use of a differential reference station to compensate for the temporal fluctuations in propagation. A spatial correction map further reduces the propagation-related positioning errors. The resulting positioning accuracy is

---

\(^5\) From 1995 till 2005, the Western European Loran-C transmitters were operated under the auspices of the Northwest European Loran System (NELS) organization.
potentially sufficient for such situations as the stringent 20 meter, 95% accuracy requirement of the maritime Harbor Entrance and Approach procedure.

Chapter 5 pays special attention to the H-field antenna; the chapter discusses error sources such as noise, E-field susceptibility, tuning, and cross talk thoroughly, as well as novel mitigation techniques and their successful implementation.

Chapter 6 discusses various measurement campaigns that bring the presented theory into practice. Throughout the Ph.D. research, the author of this work developed a highly accurate measurement system. The Reeuwijk measurements show the first step with precise dual-difference measurements; both the temporal domain and the spatial domain reveal local propagation effects. The land-mobile measurement campaign in Boston expands the measurement setup further. Simultaneous measurement of both E-field and H-field took place there allowing unprecedented analysis of re-radiation and an assessment of the applicability of low-frequency radionavigation in a land-mobile environment. The introduction of differential corrections and H-field antenna calibration for the Tampa Bay campaign resulted in an unprecedented measurement performance. Chapter 6 also shows the effect of bridges on positioning performance quantitatively, and presents the successful results of a unique re-radiation detection algorithm. This algorithm enables detection of local disturbances allowing a timely warning of potential erroneous position information. Finally, Chapter 6 shows the achievement of a positioning performance of better than 10 meters, 95% during a realistic “Harbor Entrance and Approach” scenario.

The dissertation concludes with an assessment of the potential of low-frequency radionavigation, based on the results of the author's Ph.D. research combined with his personal views.

1.5 References


This Chapter gives a brief summary of the Loran-C system to familiarize the reader with the general system concepts and vocabulary. For a more detailed discussion, the reader is referred to the Chapters 3, 4, and 5 and to the references.
2.1 Loran history

Loran-A

The Americans developed Loran-A when it became clear that they could not avoid involvement in WWII. The distances required to fly over the Pacific were enormous. The limited range of the WWII aircraft forced frequent refueling on the various specially built airfields on small islands and atolls. Celestial navigation was only possible during good weather conditions and required well-trained celestial navigators, who were scarce during the war. Loran-A was developed to overcome these navigation problems. It used 40 µs pulses transmitted at 1950 kHz by 200 kW transmitters and the system became operational in early 1943. The system, which was based on the UK “Gee” system, had an accuracy that varied according to location, time of day, weather and relative geometry of the transmitting stations. The error for navigating the 1400 miles to Japan from Tinian was about 28 miles [1].

Loran-B

Loran-B added phase matching to the Loran-A system to increase its accuracy. To accomplish this, the Loran-A transmitters were redesigned to transmit phase coherent pulses. Measurements were made on the pulse carrier, after cycle identification from the pulse envelope. This improved accuracy to about half a cycle, which equals 0.25 µs (75 m). System testing took place between 1948 and 1955 [1].

Loran-C

Loran-A and Loran-B did not provide enough accuracy over longer distances, which motivated the search for a more suitable frequency band providing better ground-wave stability. The USCG conducted successful tests with 100 µs long phase coherent pulses at 100 kHz in April 1956; these tests led to the development of the Loran-C system. The first Loran-C transmitters were built in 1957 in the Mediterranean and in the North-Eastern Atlantic, and became operational in 1958. Quickly many other transmitters followed, in the Pacific and elsewhere in the world. Among others, Loran-C has played an important role in navigating the US atomic submarines in the polar region during the cold war [27].

The current Loran-C constellation consists of transmitters in the USA, Canada, Europe, Asia, the Far East, and Russia (Chayka). The resulting coverage is discussed in Section 2.3.3. Loran-C provides 2D positioning as well as timing services for both civil and military air, land, and marine users and is approved as an en route supplemental air navigation system for both Instrument Flight Rules (IFR) and Visual Flight Rules (VFR) operations.

Chayka

Chayka is the Russian counterpart of the (originally American) Loran-C. With the exception of a slightly different waveform, Chayka is largely similar to Loran-C. Modern receivers are often capable of tracking both Chayka and Loran-C signals. A combined position solution, however, is hampered while the systems are not currently mutually time-synchronized.
2.2 eLoran

The growth in the number of Loran-C users has been reversed in favor of satellite navigation. GPS promised — and delivered — unprecedented performance for a very good price. It offers accurate 3-dimensional positioning as well as time with global coverage. The success of GPS led the USA to decide in their 1994 Federal Radionavigation Plan [4] to terminate Loran-C in the year 2000. As a result, Loran-C receiver manufacturers were reluctant to invest in further developments and (potential) Loran-C users decided against the purchase of new equipment.

However, the 1994 FRP has been superseded by the Volpe report [5] and the 2001 FRP [6], which underlines the vulnerability of GPS and the potential of Loran-C as a backup system to overcome this inadequacy. This change in American radionavigation policy led to a significant financial injection by US congress in the Loran-C system and to a renewed interest of industry and users. Additionally, Europe has acknowledged the future potential of Loran-C, which is clearly outlined in the proposed European Radio Navigation Plan (ERNP) [22].

With the renewed interest in Loran-C, the name of the system has also been changed to “eLoran” where the “e” stands for enhanced. As part of the US “Loran Recapitalization Program,” all of the US tube-type transmitters have been replaced by solid-state types and the Time and Frequency Equipment (TFE) has been upgraded towards Time-Of-Transmission (TOT) control (see Section 2.5). Furthermore, various receiver manufacturers have developed new DSP-based receivers using the modern signal processing techniques combined with H-field antennas (see Section 2.8). Finally, extensive research has been done on the potential capabilities of the eLoran system.

In 2004, it was concluded that eLoran is indeed capable of successfully mitigating the impact of a GPS outage on GPS position, navigation, and timing applications [9]. Examples of the eLoran capabilities are as follows:
- The ability to obtain absolute accuracies of 20 meters for Harbor Entrance and Approach (HEA)
- The provision of an independent, highly accurate source of Coordinated Universal Time (UTC)
- The function as a navigation source for Non Precision Approach according to the RNP 0.3 specifications.

This dissertation is the author’s contribution to the upgrade process from Loran-C towards eLoran.

2.3 Loran System Characteristics

At the time of this writing, Loran is in transition from legacy Loran-C to modern eLoran. Performance specifications based on the old Loran-C will, therefore, be too conservative; whereas, eLoran specifications are speculative because currently all system enhancements are not fully operational.
2.3.1 Loran-C system performance

The US Federal Radionavigation Systems of 2001 [7] specifies the official Loran-C System Performance as:

<table>
<thead>
<tr>
<th>Measure</th>
<th>Loran-C</th>
</tr>
</thead>
<tbody>
<tr>
<td>Absolute accuracy</td>
<td>0.25 NM (460 m) 95%</td>
</tr>
<tr>
<td>Repeatable accuracy</td>
<td>18-90 m 95%</td>
</tr>
<tr>
<td>Availability</td>
<td>99.6%</td>
</tr>
<tr>
<td>Coverage</td>
<td>Regional (Continental USA and coastal waters, North-West Europe, Western part of Russia, selected areas in Asia)</td>
</tr>
<tr>
<td>Fix reliability</td>
<td>99.7% (triad reliability)</td>
</tr>
<tr>
<td>Fix rate</td>
<td>10-20 fix/sec</td>
</tr>
<tr>
<td>Fix dimensions</td>
<td>2D + Time</td>
</tr>
<tr>
<td>System capacity</td>
<td>Unlimited</td>
</tr>
<tr>
<td>Ambiguity potential</td>
<td>Yes, easily resolved</td>
</tr>
</tbody>
</table>

These numbers are based on the original Loran-C systems infrastructure and legacy receiver technology, and are therefore very conservative. Note that the fix rate is based on the number of pulse groups that can be received per second from a given transmitter. The actual independent positioning update rate lies significantly lower because some noise reduction by integration is required.

2.3.2 eLoran system performance

<table>
<thead>
<tr>
<th>Current Definition of Capability (US FRP)</th>
<th>Accuracy</th>
<th>Availability</th>
<th>Integrity</th>
<th>Continuity</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0.25 nm</td>
<td>0.997</td>
<td>10 sec alarm</td>
<td>0.997</td>
</tr>
<tr>
<td>(463 m)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>FAA NPA (RNP 0.3)* Requirements</th>
<th>Accuracy</th>
<th>Availability</th>
<th>Integrity</th>
<th>Continuity</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0.016 nm</td>
<td>0.999-0.9999</td>
<td>10 sec alarm / 556 meter error (1·10⁻⁵)</td>
<td>0.999-0.9999 over 150 sec</td>
</tr>
<tr>
<td>(317 m)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>US Coast Guard HEA Requirements</th>
<th>Accuracy</th>
<th>Availability</th>
<th>Integrity</th>
<th>Continuity</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>0.004-0.01 nm</td>
<td>0.997-0.999</td>
<td>10 sec alarm / 20-50 m error (3·10⁻⁶)</td>
<td>0.9985-0.9997 over 3 hours</td>
</tr>
<tr>
<td>(8-20 m)</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

*Non-Precision Approach Required Navigation Performance
As of the time of this writing, no official eLoran specification has been written. However, a 2004 study done for the US Federal Aviation Administration [9] states that eLoran is capable of meeting RNP 0.3 for non-precision approach (NPA), the Harbor Entrance and Approach (HEA) requirements for maritime usage, and the Stratum-1 frequency specifications. These requirements are further described in Appendix B. It is most likely that the future eLoran system performance specification will be a superset of these requirements, as shown in Table 2-2.

### 2.3.3 Coverage

Figure 2-1 shows the regional coverage of the Loran-C and the Russian Chayka system. The Continental USA and its coastal waters, North-West Europe, Western part of Russia, and selected areas in Asia are covered by 79 transmitters (66 Loran and 13 Chayka). Note that the coverage as displayed in Figure 2-1 is based on the Loran-C 0.25 NM accuracy specification and single chain (triad) operation such as described in reference [6]. The eLoran coverage, in which the Non Precision Approach (NPA) RNP 0.3 and/or the Harbor Entrance and Approach (HEA) requirements can be met, for example, will also depend on the availability of future infrastructure such as integrity monitors and dLoran reference stations. Reference [9] contains assessments of the future eLoran HEA and NPA coverage for the continental US.

**Figure 2-1: Loran-C and Chayka positioning (solid lines) and data (dotted lines) coverage (Illustration courtesy of Megapulse [2])**

Where positioning requires at least three transmitters, data reception only requires one. Furthermore, data demodulation allows use of the full sky-wave potential, which significantly increases the effective range of a transmitter. Therefore, the Loran-C data broadcast coverage (Figure 2-1, dotted lines) expands well beyond the positioning coverage (Figure 2-1, solid lines).


2.4 Loran-C Signal in Space definition

This Section contains a brief overview of the Loran-C signal specification. For the official definition, the reader is referred to reference [13].

2.4.1 Pulse shape

The Loran-C transmitter transmits a series of 250 μs long pulses with a carrier frequency of 100 kHz. Figure 2-2 depicts a single Loran-C pulse.

Mathematically, the Loran-C pulse is described using:

\[ v(t) = A \cdot \left( \frac{t}{t_p} \right)^2 \cdot \exp\left( 2 - 2 \frac{t}{t_p} \right) \cdot \cos(\omega t + PC) \]  

(2-1)

where \( A \) is the pulse's maximum amplitude, \( t \) the time in seconds, \( t_p \) the time the pulse reaches its maximum (at 65 μs), \( \omega \) the angular frequency of 2π*100,000 rad/s and PC the phase code of 0 or π radians.

The Loran-C transmitter cuts the tail of the pulse at 250 μs with a so-called tail-biter. The Russian Chayka transmitters do not have such a system resulting in a “ringing” character at the end of the pulse. The shape of the Loran-C pulse forces 99% of the transmitted Loran-C energy in the frequency band between 90 and 110 kHz.

The Loran-C receiver tracks the phase of the pulse at the zero crossing around 30 μs typically [24]. At this point, the Time Of Arrival (TOA) is predominantly determined by the groundwave because the skywave delay is usually longer [18]. The correct zero crossing for tracking is found by comparing the envelope with a reference shape. This “shape-matching” is normally done by analyzing the envelope’s Half-Cycle-Peak-Ratio or HCPR:
\[ HCPR = \frac{\text{env}(t + 2.5 \mu s)}{\text{env}(t - 2.5 \mu s)} \]  

(2-2)

where \( \text{env} \) depicts the received Loran-C envelope and \( t \) the time in seconds. Figure 2-3 shows a commonly used cycle identification process:

Stored in memory, the receiver has a target value for the HCPR, which is a function of the antenna response and the filtering applied in the receiver. The target HCPR can be determined mathematically or by measurement. Next, the negative-to-positive zero crossing closest to the HCPR-match is designated as the measured Time Of Arrival (TOA) of the pulse. The discrepancy between the HCPR (envelope) match and this zero crossing (phase match) is called the ECD, Envelope-to-Cycle Discrepancy. Propagation effects cause a changing group delay and distortion of the pulse. This results in an ECD decreasing with distance. Usually, a Loran-C transmitter transmits the pulses with an ECD offset of about +2.5 µs such that an ECD of approximately 0 µs is expected at the border of the coverage area. This way, the chance of selecting the wrong cycle is minimized, especially at larger distances where the SNR is expected to be minimal. If, nevertheless, a wrong cycle is selected, the TOA will be off by approximately 10 µs resulting in a pseudorange error of about 3 km. Therefore, preventing such a “cycle slip” is crucial for the correct functioning of the system.

\[ \text{30 µs standard tracking point is the 3rd negative-to-positive zero crossing of a Loran-C pulse with a positive (0) Phase Code. For a pulse with a negative (π) Phase Code, it is obviously the 3rd positive-to-negative zero crossing.} \]
2.4.2 System timing

Loran-C transmitters are grouped together in chains of 3 to 6 stations. Each chain consists of a Master station, designated with the letter M, and up to 5 Secondary stations, designated with V,W,X,Y, and Z respectively. Every station transmits a group of eight pulses (nine for the Master station) at a specified interval. Individual pulses are separated 1 ms apart, except for the ninth Master pulse, which is at 2 ms after the eighth. Within a chain, the transmission time of each secondary station is specified as an offset, known as the Emission Delay (ED), from the transmission of the master station. The amount of time between transmissions of the pulse groups of a single transmitter, known as the Group Repetition Interval (GRI), is unique to each chain. Hence, chains are designated by their GRI, which is done in multiples of ten microseconds, for example, the Sylt GRI, GRI “7499”, is 74,990 µs long. The GRI is chosen such that interference of neighboring chains (Cross Rate Interference or CRI) and Continuous Wave Interference (CWI) is minimized. The GRI also needs to be long enough to prevent overlapping reception of two transmitters from the same chain within the coverage area of the chain. The minimum Loran-C GRI length is 40,000 µs, the maximum 99,990 µs.

Some transmitters are part of two chains; therefore, they are called “dual rated.” To avoid confusion, a physical transmitter is commonly called a “stick”; whereas, a single rate of a (dual-rated) transmitter is often called a “station.” Therefore, a stick can be formed by one or two stations. Because it is not possible for the transmitters to transmit two pulses simultaneously, the transmitter prioritizes one rate while “blanking” the other.

Table 2-3 shows the Loran-C Phase Code (PC), which is designed to cancel long delay skywaves, reduce the effect of Cross Rate Interference (CRI), and to assist in distinguishing between master and secondary stations. The phase code is implemented by reversing the carrier phase of the pulse in a predetermined pattern and is repeated every PCI (Pulse Code Interval), which consists of 2 GRIs.
2.4.3 Positioning

In the absence of absolute time, a Loran-C receiver needs phase measurements of minimal three transmitters to calculate a 2D position. Traditionally, this is done by measuring the difference in Time Of Arrival (TOA) of the signals of a master and multiple secondaries. These Time Differences (TDs) result in hyperbolic Lines Of Positioning (LOPs). The intersection between two or more LOPs gives the position, usually plotted on special Loran-C charts.

Modern Loran-C receivers use the pseudo-rho-rho principle: the TOAs are converted into pseudoranges and are used directly in an iterative position calculation in which the clock error automatically cancels. The latter technique also allows the use of the pseudoranges from multiple chains by the so-called “all-in-view” receivers. Pseudo-rho-rho positioning, in principle, also allows tight integration on a pseudorange level between (for example) Loran-C and GPS.

2.4.4 Data communication

The high-power low-frequency Loran infrastructure is a convenient platform to broadcast low-bitrate data over a large area. In contrast to high-frequency satellite signals, the coverage is not limited to line-of-sight. This advantage becomes especially clear at higher latitudes and in urban or mountainous areas with poor visibility towards geo-stationary satellites. In recent years various forms of Loran-C data communication have been demonstrated:

Eurofix [12] uses a three-state pulse position modulation (PPM) on the last six of eight pulses of a GRI. A Eurofix message is 30 GRIs long, containing effectively 56 data bits. Depending on the GRI, this results in a data rate between 18.7 and 46.7 bits/sec. Eurofix has been standardized by the ITU in 2001 [3], and demonstrated successful demodulation at ranges of up to 2000 km [14]. Eurofix is currently implemented on four of the NELS stations (Sylt, Lessay, Værlandet, and Bø) and on three Saudi stations (Afif, Ash Shaykh Humayd, and Al Muwassam).

The High Speed Loran-C Data Channel project (2000-2003) demonstrated the capability to broadcast the full 250 bits per second WAAS message by using intra-pulse frequency modulation [15]. After the rise time of the pulse (at 65 µs), the carrier frequency of the Loran-C pulses is hereby slightly changed, resulting in a changing phase throughout the tail of the pulse. Because, in principle, the tail of the pulse is not used for navigation, this should have

<table>
<thead>
<tr>
<th>Table 2-3: Phase code for master and secondary, “+” indicates 0 radians phase code, “−” indicates π radians phase code.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Master</td>
</tr>
<tr>
<td>GRI A</td>
</tr>
<tr>
<td>GRI B</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Master</th>
<th>Secondary</th>
</tr>
</thead>
<tbody>
<tr>
<td>GRI A</td>
<td>+ + - - + - + - · +</td>
</tr>
<tr>
<td>GRI B</td>
<td>+ - - + + + + · -</td>
</tr>
</tbody>
</table>
minimal impact on positioning performance. This modulation scheme, however, requires significant changes at the transmitter and it is doubtful whether it leaves legacy receivers unaffected [12].

Ninth-pulse modulation uses an extra data-only pulse, which is modulated using a 32-state PPM resulting in one five-bit word per GRI. The messages are 24 GRI long, and after subtracting error detection and correction bits, this results in an effective data rate of 45 bits per 24 GRI, or between 18.8 and 47.0 bits/sec. The additional 9th pulse is not used for tracking, thereby minimizing potential influence on positioning and timing performance. The 9th pulse modulation is part of the US Loran-C Recapitalization Program [9] and is on air on several transmitters in the USA since July 2005.

In recent years, various applications of Loran-C data broadcasts have been identified. Some examples:
– **GNSS augmentation**: the Loran-C data link provides differential corrections and integrity messages for (for example) GPS thereby increasing its accuracy and integrity.
– **Loran-C station identification**: normally a Loran-C station is identified via its GRI and Emission Delay. This allows a secondary station to be identified only after the master station of the chain has been found. A station-ID message allows a secondary to be identified independently of the master station.
– **Loran-C Time-Of-Transmission**: a data message containing the exact Time Of Transmission (TOT) allows a receiver to calculate precise time relative to UTC.
– **Loran-C integrity information**: integrity messages can inform the user about the transmitter health and warn for propagation anomalies such as early skywaves.
– **Loran-C temporal corrections**: temporal variations of propagation and transmitter timing can be monitored at a reference station and transmitted to the user via a Loran-C data link. This enables a significant improvement in timing and positioning performance (See also Section 4.3).

### 2.5 Transmitter

Loran-C transmitters have effective radiated peak output powers ranging from 10 kW (Okhotsk, Russia) to 1.6 MW (George, WA, USA). The European NELS transmitters range from 250 to 400 kW. These gigantic powers are needed to shout down the atmospheric noise present in the low-frequency band. Early transmitters used tube-type amplifiers, but most have been replaced by solid-state technology.

Figure 2-5-left shows a modern transmitter consisting of a number of redundant solid-state half-cycle generators from Megapulse, the leading manufacturer of Loran-C transmitters.

Efficient radiation of low-frequency radiowaves requires extremely tall transmitting antennas. Figure 2-6 shows the 200 m tall antenna of the German Sylt station. The antenna is brought into resonance at 100 kHz by using a series tuning network and a parallel capacitance. The parallel capacitance is called “top-loading” and is formed between the wire-“umbrella” at the peak of the mast and a number of radial-wires buried in the ground at the base of the mast.
There are two methods of controlling the transmitter timing: via System Area Monitor (SAM) control or via Time Of Transmission (TOT) control (Section 3.3). SAM control uses a monitoring station (SAM site), mostly located near an important shipping route. At this monitoring station, the Time Differences (TDs) between the secondaries and the master of a chain are measured and controlled towards a fixed value.

This method does not require two-way time-transfer systems and provides optimal positioning repeatability close to the SAM site. Unfortunately, at some distance from the SAM-site the Loran-C performance can degrade significantly. With TOT control, the Time Of Transmission of all stations is directly synchronized with Coordinated Universal Time (UTC).

The advantages of TOT control over SAM control are among others a better accuracy for cross chain and master independent use (for all-in-view eLoran receivers) and superior performance of Loran-C UTC receivers. All European Loran stations use TOT control as do the US transmitters, since the installation of the
new Time and Frequency Equipment (TFE), (an example is shown in Figure 2-5-right; operational since 2005).

2.6 Propagation

Loran-C LF radio waves propagate in two ways: as a ground wave along the earth’s surface, and as a sky wave reflected from the ionosphere (Section 3.4.1).

The Time of Arrival of the sky wave is a function of the height of the ionosphere, which is unknown and continuously changing. Therefore, the sky wave cannot be used for accurate absolute positioning. Typically, the sky wave arrives more than 30 µs later than the ground wave [18], which makes the standard tracking point, 30 µs after the start of the Loran-C pulse, skywave-free under normal circumstances. Unfortunately, rare events produced by extreme solar activity can cause so-called “early-skywave” events. Under these conditions, the sky-wave delay can be significantly shorter and can potentially cause tracking errors. A special 9th pulse Loran data message is suggested within the eLoran signal specification to warn the receiver not to use these sky-wave contaminated signals [9].

The Loran-C ground wave is very stable and allows for repeatable positioning accuracies in the meter-range [23]. The actual propagation time of the Loran-C ground wave is a function of the surface impedance and topography and is described using the following three factors (Section 3.4.2):

First, the Primary Factor (PF) compensates for the signal’s propagation time through the atmosphere rather than through free space.

The Secondary Factor or SF is a correction to compensate for propagation over an all-seawater path compared to the primary factor. This is not linearly related to distance, but can be calculated once an approximate path length is known.

Finally, the Additional Secondary Factor or ASF accounts for additional far-field propagation delays as compared to the PF and SF\(^2\). The ASFs are caused by the conductivity of earth’s landmasses, which causes the signals to propagate much slower than over salt seawater. Additionally, the signals are forced to travel over mountains and into valleys making the propagation path considerably longer. If left uncompensated, the ASFs often form by far the largest error source in absolute Loran-C positioning. ASF compensation can be done by mathematically modeling the effects using available ground conductivity [19] and topography databases [20] (Section 3.4.3). If extreme accuracy is required, for example during a Harbor Entrance and Approach (HEA) procedure, modeled ASFs are not accurate enough [9]. For those applications, the ASFs can be measured during dedicated measurement campaigns (Section 6.4.4). Finally, an integrated GPS-Loran receiver is capable of determining the ASFs by

\(^1\) Traditionally a more strict definition for ASF has been used, which only accounted for the additional propagation delays over land caused by the land’s conductivity, not its topography. The ASF definition used in this dissertation accounts both for the influence of conductivity and topography.
itself when both GPS and Loran are available [21]. Once GPS fails, Loran-C can take over using the last recorded ASF values.

On a micro-level, the LF radio wave propagation can be severely disturbed by the presence of local objects (Section 3.5). These objects cause re-radiation, which can very locally influence the E-field and H-field significantly, resulting in positioning errors ranging from tens to hundreds of meters. Low-frequency re-radiation can be compared with, but is not similar to, GPS multipath.

### 2.7 Noise and interference

Loran-C suffers from noise and interference internal and external to the receiving system:

The internal noise is dominated by the noise produced by the antenna. The 3 km wavelength of the Loran-C signal makes the receiving antenna infinitesimal, which forces a careful design with respect to the antenna noise. The latter is especially the case for H-field antennas (Section 5.4).

The external noise can be divided into non-local and local noise. Atmospheric noise is the most dominant non-local external noise source. Electromagnetic lightning discharges in thunderstorms all over the world create enormous energies in the LF radio band, and skywave propagation allows these signals to travel over extreme distances. The level of atmospheric noise is highly dependent upon the location on earth, the time of day, and the season of the year. Resulting from the high transmitter power, the received Loran-C signal strength usually drowns the more Gaussian part of the atmospheric noise to the background. The impulse part of the atmospheric noise can be dealt with by time-domain filtering (Section 3.6.1).

Another form of non-local external noise is Carrier Wave Interference or CWI. The frequency band of 90-110 kHz is strictly reserved for radionavigation [3] and since the termination of the DECCA Navigator system only used by the Loran-C system. Therefore, LF radio broadcasts are forced outside the “Loran-band,” but unfortunately can still be harmful. For example, the DCF77 time transmitter in Mainflingen (Germany), with its 77.5 kHz carrier frequency, is regarded as off-band interference, but can still increase the receiver noise level significantly if insufficiently suppressed. CWI can be dealt with by averaging, bandpass-filtering the desired (Loran) signal and by notch-filtering the CWI. Modern signal processing techniques allow for adjustable filtering, which continuously adapts to the momentary interference situation (Section 3.6.2).

Finally, Cross Rate Interference (CRI) is formed by Loran-C itself. All Loran-C transmitters use the same frequency and are identified and separated in the time- and code-domain. Transmitted signals from different chains, however, continuously overlap. Even transmitters too distant to be used for positioning can still cause, via skywave-propagation, a distinct increase in noise level. The PCI-timing structure of the Loran-C signal causes the CRI-level to decrease after integration. Unfortunately, due to the sensitivity of modern linear Loran-C receivers and to deficiencies in the Loran-C phase code, the PCI-integration is usually not an adequate remedy. Modern eLoran receivers use techniques such as estimate-and-subtract or detect-and-drop to further reduce CRI (Section 3.6.3).
Overcoming noise from local sources can be challenging, especially in land-mobile applications. Modern electronics such as switched power regulators and computer systems can cause significant interference at close distance. Additionally, for example, traffic detection loops and power distribution transformers tend to radiate significant energy in the Loran-C band, thereby interfering with nearby receivers. The Loran-C receiver needs to employ multi-domain interference cancellation techniques (e.g. in the time, frequency, amplitude, and spatial domain) to overcome these nuisances (Section 3.6.4).

2.8 Receiver

The majority of the traditional Loran-C receivers, also called “legacy receivers,” use an E-field “whip” antenna combined with a hard-limiter receiver. Given the 3 km wavelength of the Loran-C signal, the receiving E-field antenna can be considered an infinitesimal monopole. This active antenna provides an omni-directional radiation pattern and has a relative good noise performance. Unfortunately, it is also very susceptible to precipitation static caused by charged raindrops falling on the antenna. This p-static effect can dramatically decrease the performance of the Loran-C receiver, especially in airborne situations. Most legacy Loran-C receivers process the antenna signals by analog bandpass and notch filtering followed by a hard-limiting 1-bit Analog to Digital Converter (ADC) and a digital microprocessor. Hard limiting of the signals facilitates the suppression of impulse noise: the receiver is relative robust against noise from nearby thunderstorms and from cross-rating Loran-C chains. Continuous Wave Interference (CWI), however, poses more problems for the hard-limiting legacy receiver and the limited accuracy makes this type of receiver less suitable to fulfill the contemporary stringent demands.

Figure 2-7: Reelektronika ‘LORADD’ integrated eLoran-GPS receiver with associated Loran H-field/GPS antenna (Photo courtesy of Reelektronika B.V. [21])
eLoran introduces the next generation Loran-C receivers. Problems with p-static have been overcome by using an H-field antenna [25]. The H-field signals also penetrate better into deep urban canyons and even inside cargo containers allowing, for example, covert track-and-trace applications [26]. On the other hand, the H-field antenna requires a much more complex receiver design (Chapter 5) and often gives larger positioning error in the presence of re-radiation (Section 6.3 and Section 6.4.6). The hard-limiter receiver is replaced by a linear receiver using a high resolution ADC connected to a Digital Signal Processor (DSP). This DSP provides sufficient processing power to perform all the bandpass and notch filtering digitally. Furthermore, modern signal processing techniques allow meter-range accuracies and acquisition within seconds after a “cold start.” eLoran receivers are all-in-view: all transmitters in the region can contribute to the position solution. Integration with satellite navigation facilitates a seamless transition from GPS towards Loran when this is appropriate. Figure 2-7 shows a modern integrated GPS-eLoran receiver with associated Loran H-field / GPS antenna [21].

2.9 References

[2] Picture courtesy of Megapulse, North Billerica, MA, USA
[11] “Radionavigation”, D. van Willigen, lecture notes Et4-022, Delft University of Technologies
NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY


[17] Picture courtesy of Prof. D. van Willigen


Contemporary applications impose demanding performance requirements, often under stressful conditions. An assessment of the achievable performance of an LF radionavigation system requires knowledge of all significant error sources. This Chapter discusses these errors by first detailing the transmitter, followed by propagation, noise, interference, and finally the receiver. Most of the error sources presented are generally applicable to LF radionavigation, whereas some errors, such as Cross Rate Interference (CRI, see Section 3.6.3), are solely related to the Loran-C system.
3.1 From pseudorange to position

The accuracy requirements of an application such as Harbor Entrance and Approach (HEA) are commonly denoted as a 95- or 99-percentile position error. This means that 95 respectively 99 percent of the position solutions must be within a certain distance of the absolute position. This Chapter discusses primarily pseudorange errors, which must be converted into positioning errors to enable comparison with the application’s requirements. Transformation of pseudorange errors to a position error is a function of the following parameters:

- the biases on the pseudoranges
- the noise on the pseudoranges
- the geometry formed by the pseudoranges

3.1.1 Geometry

The GPS community has introduced Dilution of Precision (DOP) as an indication of the relation between pseudorange errors and the resulting position error. There are five different DOPs: PDOP for 3D position, HDOP for horizontal position, VDOP for vertical position, TDOP for the clock bias, and GDOP for 3D position + clock bias. Because Loran provides only 2D position (+time) we are interested in the HDOP:

$$\text{horizontal error (RMS)} = \sigma \cdot \text{HDOP}$$

where HDOP is the Horizontal Dilution of Position and $\sigma$ is the standard deviation of the pseudorange measurements. The DOP-definition can be directly derived from a given geometry, and provides quick insight into the relation between noise on the pseudorange measurements and the resulting position error. However, The DOP definition assumes a zero-mean error and equal standard deviation for all the pseudoranges used in the positioning solution. DOP also assumes that all the pseudorange errors are uncorrelated. The DOP value is
very useful in analyzing the effect of geometry on the positioning accuracy, but is not the sole parameter of interest.

The various error sources in the Loran-C error model can introduce (zero-mean) noise on the range measurement, a range bias, or both. The noise primarily influences the repeatable accuracy, whereas the range biases worsen the absolute accuracy.

3.1.2 Repeatable accuracy

The repeatable accuracy denotes the temporal stability of the position measurement. It determines how well a user can get back to a certain location or what the “position scatter” is when the user stays at the same location. Loran-C’s repeatable accuracy has always been very important to fishermen as this enables them to very accurately retrace their favorite fish spots.

The repeatable accuracy of Loran-C is relatively good: according to the original specifications, between 18-90 meters; however, with modern receivers, better than 10 meters is often achievable. Noise and interference are the main parameters determining the repeatable accuracy. Additionally, range biases in an over-determined position solution can have some influence on repeatable accuracy. The latter is possible as the range biases in an over-determined position solution cause range residuals. If the weight factors of the pseudoranges change, due to such things as changes in SNR, then the influence of these range residuals on the position solution will change as well, and hence, the repeatable accuracy is influenced. The inclusion or exclusion of a station from the position solution is obviously an example of (an extreme) change in weight factor.

3.1.3 Absolute accuracy

A radionavigation receiver measures the time of arrival of various signals, which are translated into (pseudo)ranges by using of a propagation model. An error in the propagation model leads to a range bias resulting in a deterioration of the absolute positioning accuracy.

The Additional Secondary Factor (see Section 3.4.2) is one of the phenomena that, when not properly accounted for in the propagation model, can give rise to discrepancies between the modeled propagation delay and the physical propagation delay¹. When ASFs are not, or not sufficiently, accounted for, the error in the derived absolute position can easily exceed hundreds of meters as can be seen in Figure 3-2.

Figure 3-2 clearly shows the difference between Loran-C absolute accuracy and repeatable accuracy. The repeatable accuracy is already very promising, but to meet Harbor Entrance and Approach (HEA) requirements, for example, the absolute accuracy must be almost as good. Hence, all error sources potentially creating a pseudorange bias need careful attention.

¹ Often, the Additional Secondary Factor is being wrongfully accused of being an error when in fact it is not, although it is the result of physical phenomena. On the other hand, omitting compensation for ASFs is erroneous and is the result of insufficient understanding of physics rather than physics itself.
3.2 Overview of error model

Figure 3-3 shows most of the error sources deteriorating the Loran-C (pseudo-)range measurements. Each of these error sources will be discussed in detail in the following Sections.

3.3 Transmitter timing

Because a transmitter timing error directly translates to a pseudorange error, it is crucial to maintain highly reliable and accurate time keeping at the transmitter sites. Each modern Loran-C transmitter is equipped with two or three cesium clocks, which provide excellent frequency stability. For a positioning system, however, it is necessary to have not only an accurate frequency, but also precise time. The time control of Loran-C transmitters is established by either SAM control or TOT control.

3.3.1 SAM control

With System Area Monitor (SAM) control, only the master station of a chain is synchronized to a time standard such as UTC. At the SAM site, on a specific location within the coverage of a chain, the time differences between the master station and the secondary stations are monitored and steered towards a constant value. In this fashion, the time-differences, and hence, the Loran-C position solution near the SAM site show excellent repeatability. The SAM
Figure 3-3: Graphical representation of all identified error sources deteriorating the Loran-C (pseudo) range measurements and, as a function of geometry, the positioning and timing accuracy.
sites are placed at the locations where the highest possible accuracy is needed most, such as close to an important harbor.

![Figure 3-4: Measurements taken in Tampa Bay, Florida show the influence of SAM control on station timing. Malone (7980M) is synchronized to UTC; the other stations are SAM controlled to achieve a constant TD at the SAM sites in New Orleans (LA), Mayport (FL) or Destin (FL). The timing steps introduced by the SAM sites actually increase the TD variations, and hence, degrade positioning accuracy as observed in Tampa Bay.](image)

SAM-controlled station timing is only optimal close to the SAM site. At other locations, the effect is less optimal or, as shown in Figure 3-4, SAM control can even degrade the pseudorange, and hence, positioning accuracy significantly at locations far from the SAM site. The TOA-bias introduced by SAM control is rather arbitrary and can be several hundred nanoseconds large. The seasonal and diurnal variation can exceed 100 ns.

SAM control is currently used in Asia, Saudi Arabia, and the US. The US Loran-C stations are being migrated from SAM control to TOT control as part of the Loran Recapitalization Program. The European Loran-C stations already use TOT control since 1995. In the remaining discussion of the Loran Error Budget, TOT control is assumed to be the transmitter timing control method of choice for eLoran, the modernized version of Loran-C.

### 3.3.2 TOT control

With Time of Transmission Control or TOT control, the timing of each Loran-C transmitter is synchronized to a common time-standard such as UTC-USNO. This synchronization can be achieved using the following methods:

- synchronization directly to UTC-GPS by using a GPS timing receiver
- time transfer via GPS common view
- two-way time transfer via geostationary satellite
- two-way time transfer via Loran-C
3.3.2.1 GPS timing

A stationary GPS receiver installed at a surveyed position needs only a single satellite to determine time. Typically, 6 to 12 GPS satellites are received simultaneously. The over-determined solution can be used for TRAIM, Time Receiver Autonomous Integrity Monitoring, to increase the accuracy and the integrity of the timing solution. The timing accuracy is further enhanced through very long averaging by using the highly accurate cesium clocks at the Loran transmitter sites. However, the timing solution is influenced by ionospheric and tropospheric delays and by satellite clock and orbit errors. The timing uncertainty after averaging over 20 hours is usually better than 20 ns, 2σ [3].

GPS common view time transfer

For positioning, it is important that all the transmitters are synchronized to the same common clock. An offset of this common clock to UTC, for example, will have no effect on the positioning accuracy. With GPS common view time transfer, two GPS receivers measure time with respect to the same GPS satellites. This way, satellite clock and orbit errors cancel in the time difference. Obviously, both receivers must be able to track identical satellites, which limits the distance between the two receivers. The accuracy of common-view time transfers after averaging over 20 hours is typically in the 1 to 10 ns range [3].

GPS timing is the easiest and cheapest way of distributing precise time to the Loran-C transmitters, but it implies a dependency upon the availability of GPS. Although the cesium clocks can coast for days during a GPS-outage while maintaining acceptable timing performance, it is politically a touchy point to have a system that is supposed to be a backup for GPS, but still depends upon GPS for its optimal operation. Furthermore, some scenarios described in the Volpe report [57], for example, assume prolonged GPS-outages of weeks. Such a long period becomes challenging for coasting, even when using cesium clocks.

3.3.2.2 Two-way time transfer using a geostationary satellite

Another way of distributing a common clock between the Loran-C stations is by means of two-way time transfer. Figure 3-5 shows this concept [1]:

...
Each ground station starts its Time Interval Counter (TIC) with a pulse from its own clock (typically a 1PPS), and at the same time, it transmits this pulse to the other station via a geostationary satellite. The TICs are stopped when the pulse from the remote station is received. The following timing information is recorded by the TICs:

\[
\begin{align*}
TIC(A) &= A - B + d_{TA} + d_{RA} + d_{SBA} + d_{SA} + d_{AS} + S_A \\
TIC(B) &= B - A + d_{TB} + d_{SB} + d_{SAB} + d_{SB} + d_{RB} + S_B
\end{align*}
\]  

(3-2)

where \(TIC(A)\) and \(TIC(B)\) are the Time Interval Counter readings, \(A\) and \(B\) the clock times and \(d_{xx}\) the propagation and equipment delays. \(S_A\) and \(S_B\) are the Sagnac corrections compensating for the rotation of the earth. Note that \(S_A = -S_B\) and \(S_A\) is positive if \(B\) is east of \(A\). The value of \(S_A\) is given by \(2\omega Ar / c^2\) where \(\omega\) is the angular velocity of the earth, \(c\) the speed of light and \(Ar\) the area defined by the projections onto the equatorial plane by the line segments connecting the satellite and the earth’s center to the two earth stations. The time difference between clocks \(A\) and \(B\) can now be determined by the following:

\[
A - B = \frac{1}{2} \left[ \begin{array}{c} (TIC(A) - TIC(B)) \\
+ (d_{TA} - d_{RA}) - (d_{TB} - d_{RB}) \\
+ (d_{SA} - d_{AS}) - (d_{SB} - d_{SB}) \\
+ (d_{SAB} - d_{SBA}) \\
- 2\omega Ar / c^2 \end{array} \right]
\]  

(3-3)

Most of the path delays will cancel as \((d_{SA} - d_{AS}) \approx (d_{SB} - d_{SB})\), but not entirely, because the propagation delays are not exactly the same if the up and down link frequencies are different.
The major source of inaccuracy and instability of the two-way time transfer is caused by the transmitting and receiving equipment, because there is no reason for \((d_{TA} - d_{RA})\) to be exactly the same as \((d_{TB} - d_{RB})\) seeing that they are caused by different pieces of equipment.

A high SNR and low multipath transmission-reception is achieved by using a relatively large dish antenna.

Two-way time-transfer measurement accuracies better than 1 ns can be achieved with high-end equipment. Accuracies in the order of 5 ns can be obtained at relatively low cost. The current accuracy and cost of this time-transfer method combined with the cesium clocks at the Loran-C transmitters opens the possibility of creating a robust and extremely accurate time-scale, totally independent of GPS [2].

### 3.3.2.3 Two-way time transfer using Loran-C

The stations of NELS (Northwest European Loran System) are TOT controlled using two-way Loran-C measurements as schematically outlined in Figure 3-6. The Loran-C transmitters from the NELS chain (Lessay, Sylt, Soustons, Værlandet, Bø, Berlevag, and Jan Mayen) are controlled at the French Naval Base in Brest. Brest maintains its own time-standard UTC\textsubscript{BREST}, which is derived from UTC\textsubscript{PARIS}.

![Diagram](image)

**Figure 3-6: NELS TOT synchronisation to UTC**

UTC is distributed through the NELS chain from station to station by measuring the time of arrival of the signals from adjacent transmitters relative to clock at each station. It is impractical or even impossible to accurately measure the Loran transmission very close to the
transmitter itself. Therefore, an H-field loop antenna, with the antenna null pointed towards
the transmitter antenna, is used for reception of the adjacent station. A Loran simulator is used
to generate a replica of the transmitted signal, which is transmitted directly into the receiving
loop antenna. This method leads to the following measurements:

\[
TD_{Rx\text{Lessay}} = T_{\text{Lessay}} + d_{\text{Sim\text{Lessay}}} - \left( T_{\text{Sylt}} + d_{\text{Tx\text{Sylt}}} + d_{\text{prop\text{Sylt\text{Lessay}}}} \right) + Err_{\text{Rx\text{Lessay}}}
\]

\[
TD_{Rx\text{Sylt}} = T_{\text{Sylt}} + d_{\text{Sim\text{Sylt}}} - \left( T_{\text{Lessay}} + d_{\text{Tx\text{Lessay}}} + d_{\text{prop\text{Lessay\text{Sylt}}} \right) + Err_{\text{Rx\text{Lessay}}}
\]

where \( TD_{RX} \) depicts the time difference measured by the receiver, \( d_{\text{sim}} \) the simulator delay, \( d_{\text{prop}} \) the propagation delay and \( Err_{RX} \) the receiver error. The time-difference measurements
measured by the receivers at each station are sent to the Brest control center via a permanent
data link. In Brest, the timing relations are calculated by subtracting the TD-measurements:

\[
T_{\text{Lessay}} - T_{\text{Sylt}} = \frac{1}{2} \begin{bmatrix}
TD_{Rx\text{Lessay}} - TD_{Rx\text{Sylt}} \\
d_{\text{Sim\text{Sylt}}} - d_{\text{Sim\text{Lessay}}} \\
d_{\text{Tx\text{Lessay}}} - d_{\text{Tx\text{Sylt}}} \\
Err_{\text{Rx\text{Sylt}}} - Err_{\text{Rx\text{Lessay}}} \\
d_{\text{prop\text{Sylt\text{Lessay}}}} - d_{\text{prop\text{Lessay\text{Sylt}}} \end{bmatrix}
\]

The propagation delays will cancel as \( d_{\text{prop\text{Sylt\text{Lessay}}}} = d_{\text{prop\text{Lessay\text{Sylt}}} \). However, all of the other
delays are related to different hardware setups, and they will not cancel completely. It is
assumed that these errors are constant and can be accurately calibrated.

By using this two-way time transfer method, the timing of the NELS Loran stations is
maintained within 100 ns with respect to UTC_BREST. The timing of the secondary stations is
guaranteed within 30 ns of the master station of the chain.

### 3.3.3 UTC

“UTC” seems the obvious choice for the common clock for a radio-navigation system.
Unfortunately, UTC does not physically exist. It is a “paper-clock” that only provides precise
time of a month ago by averaging the time and frequency of various time-centers around the
world. Some examples of the time standards of these time-centers are: UTC_USNO (United States
Naval Observatory), UTC_GPS, UTC_Paris and UTC_Brest. Most of these timing centers are obliged to
maintain time within 100 ns to UTC, but often do a much better job. Figure 3-7 shows the
relation between UTC_Paris and UTC_GPS and the time measurements derived from the Lessay
Loran station for July 2005.

For radio positioning, the actual time-scale basically does not matter as long as it is common
for all transmitters in the system. However, offsets between time-scales do become an issue
when two or more radionavigation systems are integrated on a pseudorange level, for
example when integrating Loran-C with Chayka. GPS cost and performance have made
UTC\textsubscript{USNO} the de-facto time-standard for most precise timing users. Therefore, for Loran-C to be a backup candidate for sub-100 ns timing, the Loran-C timing users must be supplied with the exact offset between the timing standard used for Loran and UTC\textsubscript{USNO}.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure37.png}
\caption{UTC-Paris compared to UTC-GPS and Loran-time derived from Lessay (6731M) compared to UTC-Paris, July 2005}
\end{figure}

### 3.3.4 Internal timing control

The precise timing of most Loran-C transmitters is adjusted by inserting discrete phase steps, the so-called Local Phase Adjustments or LPAs. The NELS stations use 10 ns LPAs; whereas, the stations in the US apply 20 ns steps. The effect of these steps is clearly visible in Figure 3-4. For the older (hard limiter) Loran-C receivers, the discrete steps were of negligible influence. Modern receivers, however, achieve much higher accuracy than 20 ns, and for these receivers LPAs can be a performance-limiting factor. The use of discrete timing steps in the transmitter is especially disadvantageous for frequency recovery. To mitigate this effect, US Loran stations equipped with new time and frequency equipment will apply time adjustments via a temporary (e.g. 30 minute long) frequency adjustment of the transmitter clock. This causes the timing to change very gradually instead of the traditional LPA step response [2].

The phase and amplitude of the transmitted pulse are continuously monitored and, if needed, the transmitter timing circuitry is optimized automatically. In the older transmitters, the so-called PATCO unit (Pulse Amplitude Timing and Controls) is responsible for this task. The PATCO unit works with discrete time steps (10 ns in Europe, 20 ns in the USA), as is shown in Figure 3-8. The frequency and amplitude of these adjustments can cause the internal transmitter timing control to be the dominant noise-source for Loran-C receivers tracking the signal with high accuracy. The PATCO units of the US Loran-C transmitters will be replaced with more subtle control equipment as part of the Loran recapitalization program [5].
3.4 LF propagation

The propagation of the low-frequency (100 kHz) Loran-C radio waves consists of two parts: the sky wave that is reflected from the ionosphere, and the ground wave. The sky-wave part is not very stable due to the ionospheric changes. Ground-wave propagation is a function of various parameters such as ground conductivity (or surface impedance), terrain variation, and the refractive index profile of the atmosphere. Its stable phase velocity makes the ground wave suitable for navigation and timing purposes. This Section focuses on the separation between ground wave and sky wave, the characteristics of ground-wave propagation, and methods available to model these characteristics.

3.4.1 Ground wave versus sky wave

The ionosphere consists of layers in the earth’s atmosphere containing gasses ionized under the influence of solar radiation. Depending on their frequency, the transmitted signals are refracted or reflected in one of the layers of the ionosphere [9]. The 100 kHz Loran-C signals are reflected in the D-region of the ionosphere as schematically shown in Figure 3-9.

The ionosphere constantly changes due to changing solar activity. Ionospheric absorption is smallest and sky-wave field strengths are largest late at night. The absorption increases rapidly at sunrise and decreases from its day-time value almost as rapidly at sunset. The result is an unpredictable and unstable sky-wave delay, which makes the sky wave unusable for absolute positioning. Furthermore, the sky wave is attenuated less than the ground wave, which makes the sky wave dominant over the ground wave for the more distant stations (e.g. at 1000 km), especially at night.

Figure 3-8: Effect of PATCO-circuits on the Time Of Transmission (TOT) of Malone (7980M) and thereby on its Time Of Arrival (TOA) as measured in Tampa Bay, Florida, April 2004.
3.4.1.1 Ground-wave tracking

Fortunately, the sky-wave signal arrives later than the ground wave, due to its longer path length. The pulsed signal structure of Loran-C then allows the receiver to measure the Time of Arrival (TOA) of the stable ground wave, undisturbed by the sky wave with its unknown delay. This is a significant advantage of Loran-C over Continuous Wave (CW) systems, such as Datatrak and the former DECCA-navigator, where the sky wave severely limits the effective range.

Long delayed sky waves (sky-wave delays larger than 1 ms) are “cancelled” by the Loran-C phase code once a receiver averages over a Phase Code Interval (PCI).

Sky waves with delays shorter than 1 ms are mitigated by tracking only the start of the received pulse. The Loran-MOPS (Minimum Operational Performance Standards) [10] specifies that the receiver needs to track the phase of the pulse at 30 µs, the 3rd negative-to-
positive zero crossing of a pulse with positive phase code. This tracking point is usually sky-wave-free, except under rather extreme circumstances, which are denoted as “early sky-wave” conditions.

Early sky-wave conditions can occur during adverse solar weather in which the ionosphere is disturbed enough to create problems for signals at northern latitudes [11][12]. This causes sky waves with shorter than normal delays, resulting in potential interference with the ground wave at the standard tracking point (30 µs). There are several remedies against this potential integrity hazard. The receiver can reduce the influence of early sky waves by tracking earlier in the pulse. The probability of a sky-wave delay smaller than 20 µs, for example, is significantly lower than that with a delay of 30 µs. By tracking one cycle earlier, for example, the influence of an early sky wave is reduced. However, the amount of energy at an earlier tracking point is also significantly lower, which makes it more difficult to track that zero crossing reliably. A transmitted pulse with a steeper leading edge such as used by the Russian Chayka system mitigates this loss in energy as shown in Figure 3-11.

![Leading part of a Russian Chayka pulse envelope compared to the American Loran-C pulse envelope](image)

Another approach to deal with early sky waves is not by correction, but by detection. Signals contaminated with early sky waves are flagged and discarded. In this way, integrity risks are traded against a lower availability. Reliable early-sky-wave detection by the receiver is a very challenging task; detection by a(n) (existing) monitoring network seems favorable. [35]. Monitor stations at a known position can detect propagation anomalies much easier by comparing measured characteristics, such as ECD and TOA, to reference values. Integrity alarms can be provided to the users by transmission over a Loran data channel such as Eurofix or 9th pulse.

From an SNR point of view, it is beneficial to track the phase of a Loran-pulse as high as possible in the pulse, especially if a narrow bandpass filter is used, which further attenuates the
early portion of the pulse. Special techniques have been developed by Abbas [8] and van Nee [37] to estimate the ground-wave and sky-wave portions of the received signal. Figure 3-12 shows an example of ground-wave-sky-wave separation performed on the pulse received from Ejde, about 1360 km from the receiving site in Delft. This method, implemented by van Nee and Andersen [37] works relatively well up to the top of the estimated ground wave. The residual significantly increases after this point due to the undefined tail of the Loran-C pulse.

![Figure 3-12: Ground-wave/sky-wave separation (Graph reproduced from reference [37])](image)

The measured sky-wave delay can be used to adaptively choose the optimal sky-wave-free tracking point, and hence, optimizing the tracking performance. Note, however, that each part of the pulse consists of slightly different spectral components and is thereby differently influenced by propagation. Therefore, propagation corrections measured at the 30µs tracking point are not entirely correct at a higher tracking point, which will result in a position bias [7]. Knowledge about the sky-wave delay can also be used to detect severe (early) sky-wave conditions and to measure the sky-wave stability and its usability for sky-wave aided ground-wave tracking.

### 3.4.1.2 Sky-wave aided ground-wave tracking

The previous Section concluded that the sky wave can contain a large amount of energy, especially for distant stations, but cannot be used for absolute positioning. The ground wave, on the other hand, can be very weak, but is stable, and should be used to determine the absolute Time of Arrival (TOA). Figure 3-13 shows the tracking results of ten successive negative-to-positive zero crossings from Ejde, measured at Harwich UK. The bottom track depicts the 3rd zero crossing, and is clearly unaffected by sky waves; however, it is noisy. The higher tracks are measured higher in the pulse, and therefore, are less noisy; however, these can be significantly affected by sky waves, especially just after sunset and shortly before sunrise.
**Figure 3-13**: Ten successive negative-to-positive zero crossings from Ejde measured at Harwich UK (1257 km), April 20-21 2006. The bottom track depicts the 3rd zero crossing; the black lines depict the sunset and sunrise. Each blue track has been plotted with an arbitrary offset for presentation purposes; therefore, the distance between individual tracks does not represent the actual time difference between the associated negative-to-positive zero crossings.

**Figure 3-14**: “Total Pulse Tracking” applied on the measurements of Figure 3-13: the absolute stability of the ground wave is combined with the low-noise behavior of the skywave.
Fortunately, the low-noise sky wave is relatively stable over a short time, up to several minutes. This opens the possibility to design an algorithm using both the absolute stability of the ground wave and the low-noise behavior of the sky wave. Figure 3-14 shows the results of such an algorithm, as implemented by the author.

Figure 3-14 clearly shows a significant improvement on the tracking results by merging ground-wave stability with sky-wave strength. The implemented algorithm is not yet perfect, though, as can be seen from the minor sky-wave-induced fluctuations around 20:00 and 4:00 local time.

### 3.4.2 Spatial variations

Traditionally, a Loran-C receiver outputs Time Differences (TDs), which are the differences in the time of arrival of the signal from a master station and that from a secondary station from the same chain. A time-difference measurement results in a hyperbolic Line of Position (LOP). Two TDs provide an intersection between two LOPs, and hence, a position solution. Special maritime charts containing these LOPs have been available for this traditional usage of the Loran-C system.

Modern Loran-C receivers provide the time of arrival of the individual stations. These TOAs can be converted to pseudoranges by multiplication with the propagation speed of a 100 kHz ground wave. At least three pseudoranges are needed to compute a two-dimensional position. Unfortunately, the propagation speed of a 100 kHz ground wave can significantly differ from the speed of light in vacuum. The Loran-C system uses three correction factors to compensate for this difference [6]:

The **Primary Factor (PF)** is the propagation time for the signal to traverse the atmosphere \(V_{pf} = 2.99691162 \times 10^8 \text{ m/s}\).

![Secondary factor as a function of distance](image-url)
The **Secondary Factor (SF)** discounts the extra delay of the signal traveling over salt seawater with a conductivity of 5000 mS/m [6]. The SF varies with distance as can be seen in Figure 3-15.

The **Additional Secondary Factor (ASF)** represents the incremental propagation delay of the Loran signal due to traversing heterogeneous earth versus an all seawater path. The velocity of the ground wave varies according to the type of surface over which it travels; the key parameter is the impedance of the surface. Signals travel most slowly over ice, deserts, or mountains, a little more quickly over good farming land, and most quickly of all over seawater. Further, the velocity of signal propagation varies with distance from the transmitter in a complicated manner. Additionally, the Loran-C signals are forced to travel over mountains and into valleys, which can make the propagation path significantly longer. The ASF embodies all these additional delays and is therefore a function of distance, surface impedance, and topography.

![Figure 3-16: The ASF discounts for the additional propagation delay caused by the conductivity and topography of land masses (Illustration courtesy of Williams and Last [17])](image)

The true propagation time of the Loran-C ground wave is now given by the following:

\[
\text{True propagation time} = PF + SF + ASF
\]  

(3-6)

The Loran-C pulse is not only delayed, but is also distorted as a function of the propagation path. Different frequency components are subjected to different delays and attenuation, which results in a disparity between group-delay and phase-delay, as well as in distortion of the shape of the pulse. A Loran-C receiver is supposed to measure the time of arrival of a Loran pulse on the 3\textsuperscript{rd} zero crossing. This zero crossing is found by matching the received pulse
envelope to a standard envelope (Section 2.4.1). The difference in arrival time between the pulse envelope and pulse phase at this standard tracking point is called the Envelope to Cycle Difference (ECD). This ECD slowly decreases with distance over sea [13], but can undergo more rapid changes over land. Knowledge about the expected ECD is essential to reduce the chance of a faulty cycle identification, which would result in a three-kilometer range error.

### 3.4.3 Modeling ground-wave propagation

The overview presented in this Section is primarily based on a publication by Samaddar [15]. For more information, the reader is referred to this article and the associated references. This Section will briefly discuss the various techniques involved and place them in a historical context.

The phase velocity of 100 kHz ground wave is a complex function of the ground conductivity, terrain variation, and the refractive index profile of the atmosphere. Modeling the exact phase velocity requires solving the electromagnetic wave equation, and thereby incorporating all of the above mentioned parameters. This is an impossible mathematical task. Several approximate methods with various degrees of complexities have been sought by various authors in the past:

1. Sommerfeld and Norton: Homogeneous flat earth model
2. Watson, Van deer Pol and Bremmer: Homogeneous spherical earth model
3. Millington: Empirical model for an earth with mixed surface conductivities for computing signal strength
4. Pressey: Empirical model with mixed surface conductivities for computing phase
5. Wait: Multisection spherical earth model
6. Hufford: Integral equation approach to the propagation over an irregular surface
7. Johler: Integral equation approach
8. Monteath: Method of computationally efficient calculation of ground-wave attenuation
9. Last and Williams: Implementation of Monteath and Wait in their BALOR (BAngor LORan) software package

![Ground wave modes of propagation](image-url)
A ground wave consists of a space wave and a surface wave as shown in Figure 3-17. The space wave is made up of:

- direct wave (direct path between transmitter and receiver)
- ground reflected wave (signal arrives at the receiver after being reflected from the surface of the earth)
- diffracted waves around the earth and the refracted waves in the upper atmosphere

The surface wave is guided along the surface of the earth, which absorbs energy from the wave, causing its attenuation.

The direct and ground reflected terms of the space wave cancel each other if both the transmitting and receiving antenna are very near the surface of the earth. In that case, the Loran-C signal consists entirely of the surface wave, assuming the sky wave is rejected in the receiver. Obviously, this approximation is not allowed for airborne applications where altitude can have a significant impact on the received signal phase.

**Flat Earth Theory**

The “simplest” ground-wave propagation model considers the surface of the earth to be a plane; it neglects the influence of the earth’s curvature on the ground wave. In 1926, Sommerfeld described a solution of flat-earth ground-wave propagation. Norton rewrote Sommerfeld’s solution to make it suitable for numerical computations (1937).

**Spherical Earth Theory**

The spherical earth theory was first presented by MacDonald in 1903, in which he described a solution of the Maxwell’s equation considering the radiation from a vertical electric dipole in
the presence of a perfectly conducting spherical earth. However, his solution showed extremely poor convergence, making it of no practical value. Watson (1918) was able to convert MacDonald’s equations into a new highly convergent series, which was further developed by Van der Pol and Bremmer in 1937.

Figure 3-18 shows the phase of the secondary factor for both the spherical earth theory and the flat-earth theory for various conductivities and distances. The curvature effect causes the secondary phase to be higher when calculated using the spherical earth theory than when calculated with the flat-earth theory for the same conductivity and distance. The all-seawater curve with \( \sigma=5000 \text{ mS} \) is known as the Secondary Factor (SF), which is the standard propagation correction in a Loran-C receiver.

**Mixed path propagation**

The models presented so far deal only with a homogeneous earth surface. Modeling a more realistic mixed path consisting of different conductivities and dielectric constants poses a very difficult problem mathematically. Even the simplified smooth inhomogeneous plane or spherical earth situation has no exact solution.

Millington (1949) developed a semi-empirical method for calculating field strengths over an inhomogeneous path composed of multiple homogeneous segments. Millington’s formula is based on physical reasoning and the principle of reciprocity, and it is applicable when the receiver is far away from the boundary between two media. Pressey (1953) extended Millington’s formula to include phase as well. This method is often called the Millington-Pressey method or the Millington method.

![Figure 3-19: Two-segment mixed path](image)

Figure 3-19 depicts, schematically, a two-segment mixed path consisting of a land path with length \( d_1 \) and a sea path with length \( d_2 \). First, the phase-lag for an all-land and all-sea path is calculated using, for example, Norton’s and Bremmer’s formulas. These results are plotted as the brown “all land” and the blue “all sea” line in Figure 3-20. Note the rapid change in phase close to the transmitter due to the near-field effect. Next, the all-sea line is shifted upward until it intersects the all-land line at the impedance boundary (coastline) depicted by point \( B \).

The total phase lag is \( \theta_t \). Let us now interchange the receiver and transmitter. The reciprocity-principle states that the total phase lag in this new situation should remain the same. First, the signal travels for a distance \( d_2 \) over sea, which was already plotted as the “all sea” path. Next, the all-land line is shifted downward until it intersects the all-sea line at \( B' \). It becomes clear
that the resulting phase $\theta_2$ is different from the previously derived result $\theta_1$. Because the true delay should lie somewhere in between, Millington and Pressey took the true field to be approximately the arithmetic mean, $(\theta_1 + \theta_2)/2$, denoted as the black “composite” line in Figure 3-20.

In 1944, Leontovich introduced the concept that the effect of the earth on an electromagnetic field can conveniently and approximately be described by its surface impedance. The surface impedance $Z$ is defined by

$$n_0 \times E = Z H,$$

where $n_0$ is the unit vector normal to the surface of the earth. $E$ and $H$ are the field tangential to the earth's surface. The surface impedance at a given location on the earth can be derived, in principle, from an equivalent conductivity $\sigma_{eq}$ and dielectric constant $\varepsilon_{eq}$, which represent the properties of the materials at that location if those properties are vertically homogeneous.

Wait (1964) introduced a model for mixed path propagation based on the compensation theorem, which relates the propagation over an inhomogeneous earth to the surface impedance by means of a two-dimensional integral equation. If the following conditions are true:

– the wave reflected from the boundary is neglected,
– the property of the earth's surface does not change in the direction transverse to the propagation path,

then by using the stationary phase principle it is possible to reduce the two-dimensional integral to a one-dimensional integral, which is much more computationally efficient.

![Figure 3-20: Millington-Pressey – Method of determining phase change over an inhomogeneous path](image)
Irregular terrain

The earth surface is not only inhomogeneous; it also consists of irregular terrain. As ground waves are forced to travel over mountains and through valleys, it is desirable to have a propagation model that incorporates this terrain effect. Such a model was first introduced by Hufford in 1952. His solution resulted in a two-dimensional integral over an irregular surface of the earth. This surface integral can be simplified to a line integral using the same argumentation as in Wait’s multi-section model: the terrain properties are supposed to have an azimuthal symmetry.

Johler and Berry used numerical techniques to solve Hufford’s integral and found that the integral equation technique matches more closely the measured data than the smooth Earth theory of Millington-Pressey (1967).

Monteath [16] used Hofford’s method to derive an effective and extremely efficient computation method to model inhomogeneous irregular terrain. An important application at that time (1978) was the coverage prediction of BBC LF broadcast transmitters. The focus, therefore, was on the signal strength rather than on the phase of the received signal. It should be noted that Monteath applied various simplifications to make his model run on 1978’s computers.

Williams and Last [17] implemented Monteath’s method in their software package called “BALOR”. This software accepts topography, conductivity, and coastline-databases, and it calculates the complex attenuation for a user-defined area. This software, combined with the availability of massive computing power, allowed the computation of Loran ASFs over a large area for the first time.

Figure 3-21 shows the BALOR-ASF predictions for the German Loran-C station Sylt up to a range of 1000 km.

Further studies revealed that some of Hufford's simplifications caused unacceptable modeling errors at larger (>1000 km) distances. These discrepancies were removed by using methods developed by Wait [20][21], who starts with a spherical earth, and thereby avoids the simplifying assumption employed by Hufford and Monteath. The resulting improved BALOR model [18] appears to model the ASFs more convincingly, although the results have not yet been thoroughly verified due to the lack of accurate measurement data.

Using essentially the same techniques as described earlier, the phase delay, the group-delay, and pulse distortion can be modeled. Sherman examined the rates of ECD change of different values of (homogeneous) ground conductivity with distance over seawater and over land [13]. BALOR calculates the actual distorted pulse by modeling the complex attenuation factor at 90, 100, and 110 kHz. The intermediate frequencies are obtained via interpolation. A mathematical Loran-C pulse is then “filtered” with this calculated frequency response, and hence, predicts the distorted pulse after propagation [19]. Note that the definition of ECD only describes the phase versus envelope-delay, and it assumes that the shape of the pulse remains unaltered. In practice, this is not the case, which makes the current ECD-definition perhaps less usable.
3.4.3.1 Modeling pitfalls

Modeling low-frequency ground-wave propagation requires various assumptions and simplifications to obtain enough efficiency to be of any practical use. It is generally a good idea to be aware of these simplifications before using the model. This Section discusses some of the author’s concerns regarding the model’s assumptions and simplifications.

Database accuracy

Even the most sophisticated propagation model can only produce results as good as the databases used for the calculations. Precise knowledge of the surface impedance and terrain is essential but not always available. The US conductivity database for example has only 16 levels. Furthermore, the databases used to model the European NELS Loran-C coverage area showed political boundaries: apparently, every nation has its own method of measuring conductivity.

Additionally, the conductivity databases often have very limited resolution. This is somewhat compensated in the BALOR model by using coastline databases to artificially increase the resolution of the conductivity change at the land-sea transitions. Note that this approach introduces a discontinuity in conductivity, where in practice the land-sea interface can show a much more gradual transition. Note also that the fluctuations in surface impedance caused by
ground contours are usually less than those resulting from the presence of something as common as a buried conductor [16]. Therefore, the conductivity database will always have a limited accuracy unless every point is measured. In the latter case it will probably much more convenient to measure the ASFs directly.

**Seasonal and weather-related effects**

The terrain topography will remain constant in time; the impedance, however, is known to change seasonally due to weather changes. The resulting conductivity changes limit the maximum obtainable absolute accuracy of the modeled ASFs.

**Surface integral versus single integral**

The Maxwell's equations have been reduced to a scalar wave equation. This is only valid under the simplifying assumption that the terrain and impedance do not change in the direction perpendicular to the propagation path. In reality, this is not true; therefore, it might introduce severe errors in the predictions.

**Reflections**

One of the simplifying assumptions of Monteath and BALOR is to ignore any reflections due to irregularity of the ground and discontinuities in the surface impedance. It is, for example, assumed that there is negligible reflection back towards the transmitter at a land-sea boundary. This enables the integration to be restricted to the line between transmitter and receiver, which significantly improves the efficiency of the calculations.

**Refraction of the Earth's atmosphere**

The choice of the factor by which the Earth's radius should be multiplied to take account of refraction by the Earth's atmosphere has been studied by Rotterham [22]. His results suggest a factor of 1.18 for the LF band. However, the BALOR software for example uses a value of $4/3$, which is more common for higher frequencies. It would be interesting to perform a sensitivity analysis of this parameter, and based on the results, to further study the refraction index.

**Skin-depth at low frequencies**

According to Leontovich, the effect of the earth on the electromagnetic field can be approximately described by its surface impedance. This surface impedance is a function of the equivalent ground conductivity $\sigma_{eq}$ and dielectric constant $\varepsilon_{eq}$ as long as those properties are vertically homogeneous. Obviously, these properties are never completely vertically homogeneous, which can have an effect on the modeling accuracy. Figure 3-22 shows the skin depth of various soil types for various frequencies.

The skin depth is defined as the distance after which an electromagnetic wave is attenuated to $1/e$ of its original value; therefore, it gives an impression of how deep a wave penetrates into the ground. A couple of meters of poor soil with a skin depth of 100 meters, for example, will have little influence on the effective surface impedance if that layer lays on top of a good soil with a skin depth of 10 meters, for example. As a result, some knowledge of the vertical geological ground structure is required to reliably determine the effective surface impedance, and hence, the propagation properties at 100 kHz.
Discontinuities
In case of discontinuities in terrain and or conductivity, Monteath suggests to either use a small calculation interval or to smooth out the discontinuities [16]. BALOR, however, uses a coastline database to enhance detail of the conductivity database at the coastal interface, and therefore, introduces discontinuities.
It is to be expected that the presented modeling techniques based on a 1D integral will perform relatively poorly close to a coastal interface. However, at this point, little has been done to compare the accuracy of various techniques at some distance after a coastline crossing. The question now arises whether it is better, for example, to smooth the terrain and conductivity in the direction perpendicular to the propagation path, in all directions, or not to apply any smoothing at all.

Calculation interval
The optimal calculation interval of the numerical simulation is a compromise between accuracy and computation time. Rapid changes in surface impedance and topography generally require a shorter calculation interval to maintain modeling accuracy. It has also been found that the relative error increases as the magnitude of the attenuation factor decreases; therefore, the maximum permissible interval decreases as the magnitude of the attenuation factor decreases as its phase retardation increases [16].

Alternate modeling techniques
It would be interesting to compare the results of the modeling techniques described earlier with the more modern Method-Of-Moments techniques such as NEC, for example. The availability of massive computing power should allow for a more elaborate modeling of a small region around a coastline for example.
### 3.4.4 Temporal variations

Spatial variations in propagation are mostly stable in time and, therefore, can be compensated for by using correction maps based on models, measurements, or both. Temporal variations in propagation endanger the effectiveness of these spatial grid corrections.

The USCG has extensively researched, both theoretically and empirically, the temporal variations in ground-wave propagation during the 1970s and 1980s.

According to Johler [23], the causes of ground-wave temporal changes are:
- Change in surface impedance of the ground due to climate and weather
- Influence of weather, climate, and ground elevation on the index of refraction of air at the surface of the ground
- Variations of the gradient of the index of refraction at the surface of the ground with altitude above the surface of ground (vertical lapse of refractive index)
- The more subtle effects of terrain roughness, elevation, soil consistency, geologic underlayment, as it interacts with weather and climate to produce both ground impedance changes, and atmospheric changes such as temperature inversions.

Temporal effects can be diurnal, seasonal, and perhaps even exhibit long-term effects correlated with the sunspot cycle. Additionally, paths over land are much more susceptible to propagation delay fluctuations than paths over water, and these fluctuations are much greater in winter than in summer [27].

---

**Figure 3-23:** Temporal variation regions for the US expressed in ns per 1000 km (Mm) (Illustration from reference [35])
In the early 1980s, the USCG conducted a series of regional Loran-C signal stability studies that developed, and empirically calibrated, a model for explaining the seasonal variation in the signal propagation delays [23]. As part of the US Loran Recapitalization Program, the LORIPP group extended the USCG model with data from the USCG monitoring network, resulting in the assessment of the temporal variations presented in Figure 3-23 [35].

In the US, the temporal variations are largest in the North-East and the Great Lakes, milder in the South-East, even milder along the West Coast and virtually non existent in the high altitude, low atmospheric density of western states.

A local differential reference station can be used to mitigate the effect of the temporal variations. The remaining error is then determined by the difference in experienced variation at the reference site versus that at the user’s location. More information on differential Loran and on the spatial decorrelation of the temporal corrections can be found in Chapter 4.

The temporal variation of ground-wave propagation forms a key element in the error budget of precision positioning (e.g. HEA) and precise time recovery. Accurate knowledge of this phenomenon is crucial to determine the optimal employment of differential reference stations and the resulting expected positioning and timing performance. Although the current knowledge of the temporal variations is mostly based on relatively old measurement data of SAM-controlled Loran chains, new measurements using the modern transmitter timing control and high performance receiver equipment can significantly enhance the quality of the measurement data and potentially yield improved insight in this phenomenon. It also should be noted that most of the research has been conducted in the United States; to the knowledge of the author, little is known about the temporal variations experienced in Europe and other parts of the world, for example. Further research on this subject is recommended to better assess and optimize the applicability of Loran for high precision positioning and timing.

### 3.4.5 Propagation measurements

Actual propagation measurements form a crucial ingredient to validate the various assumptions made in the propagation models. Field measurements of the signal propagation delays have long been hindered by the absence of convenient and affordable position and time reference systems. With the availability of GPS, combined with Rubidium clocks, for example, this problem has been mitigated. This, combined with the introduction of modern receiver technology and processing techniques, now allows for the verification of decades of propagation research. With carefully planned measurements, it should be possible to efficiently validate many aspects of the propagation models, thereby enabling further iterations. Optimized propagation models enhance the capability of such things as coverage prediction, optimal employment of new transmitters, the planning of differential reference stations, and a better assessment of the capabilities of Loran-C.

However, accurate and reliable prediction of ground-wave propagation will always remain challenging. Various assumptions and simplifications still must be made to obtain sufficient computational efficiency. Furthermore, many of the required variables are known with only limited accuracy. As a result, the obtainable accuracy of modeled ground-wave propagation is
limited — most likely too limited for those applications requiring ultimate accuracy, such as the maritime HEA procedures and precise time recovery. It is advisable then to merge actual measurements with the predictions from the model. The measured values form the absolute truth, the modeled values can be used to interpolate and extrapolate beyond the measurement locations, and hence, obtain the desired coverage of the ASF correction map. Section 4.4.1 further details on this merging process.

Chapter 6 contains various examples of highly accurate propagation measurements.

### 3.5 Re-radiation

Section 3.4 assumed far-field propagation, where the E-field and H-field have a fixed orthogonal relationship\textsuperscript{2}:

\[
\frac{|E|}{|H|} = Z_0 = 120\pi \ \Omega
\] (3-8)

This free-space relationship is graphically presented in Figure 3-24:

\[\text{Figure 3-24: In the far field, the electric (E) and magnetic (H) field components have a fixed orthogonal 120}\pi \text{ relation}\]

In the far field, the difference in performance between an H-field and E-field receiver is determined only by the sensor performance and by the potential presence of local interference. Section 6.4, for example, shows measurements in a maritime environment with identical performance for E-field and H-field, as long as no large objects are in close proximity of the receiver.

\textsuperscript{2} Note that Equation 3-8 is only valid in free space, not in atmosphere. However, because of its minor influence, this difference is ignored in this remainder of this dissertation.
The presence of local objects, however, can distort the far-field relation between E-field and H-field. An object with permittivity \( \varepsilon \neq \varepsilon_0 \), permeability \( \mu \neq \mu_0 \), and/or conductivity \( \sigma \neq 0 \) that is irradiated by a (LF) signal will re-radiate this signal. The object needs to be significantly large with respect to the wavelength (Loran-C: \( \lambda = 3 \text{ km} \)) to have “noticeable” effect. However, modern Loran-C receivers redefine “noticeable” because they can now track the radio-navigation signals with nanosecond accuracy, which equals approximately \( 1/10,000 \lambda \) at 100 kHz. Given this level of accuracy, now many more objects are suspected to have noticeable influence on the obtained positioning and timing performance. Examples of such local objects are, for example, buildings, bridges, power lines, coastlines, and mountains.

### 3.5.1 Modeling of re-radiation

A precise prediction of re-radiation requires the accurate knowledge of the characteristics of the objects involved, a sophisticated computer model, and associated massive computing power. The exact knowledge of the composition of the re-radiating object is far from trivial; often it is not possible to retrieve the precise construction and the materials used. Therefore, if the objective of the modeling is scientific; for example, the better understanding of local propagation phenomena and the resulting impact on navigation performance, the more it is advisable to first consider a relative simple, isolated structure. A bridge, for example, is usually a fairly simple and known structure of steel and reinforced concrete. A well-defined ground plane is formed if the bridge is surrounded by saltwater. Such an object forms a good start of further re-radiation research.

Software simulation packages, such as the Numerical Electromagnetics Code (NEC), can be used to predict the effects of the object on the signals as received by the user. Detailed measurements of both amplitude and phase simultaneously in the E-field and H-field can be used as validation and for further model iteration. The measurements of the Sunshine Skyway Bridge (Tampa Bay, Ohio), partly presented in Section 6.4, can form a good starting point for further research.
Instead of the precise and highly complex EM-modeling, it is also possible to use a circuit approximation to understand the observed re-radiation phenomena. Envision, for example, a vertical conducting structure. Over this structure, no potential difference can exist, which forces the equi-potential lines of the vertically polarized E-field over the structure such as depicted in Figure 3-26. At the top edges of the structure larger E-field gradients are noticed, and hence, a slightly higher field-strength. Close to the base of the structure, the opposite effect is experienced.

![Equi-potential lines (red) and electrical field lines (blue) near a grounded conductor (gray)](image)

Measurements close to tall conducting structures confirm this hypothesis (Section 6.3.2). However, note that estimation of the resulting phase changes is less trivial. This simple model, therefore, can help to understand signal attenuation and system availability, but not the resulting positioning performance.

A dual-circuit approximation is possible for the H-field. A conducting surface will not allow a flux to exist on its surface. Therefore, if a magnetic field line has an orthogonal orientation with respect to the conducting surface, it will be severely attenuated. Magnetic field lines parallel to the surface can pass unhindered.

A detailed analysis of individual re-radiation sources is essential for a better understanding of near-field propagation and can, therefore, help in the development of re-radiation detection and mitigation techniques. An assessment of the influence of more complex situations involving a multitude of re-radiators becomes laborious, impractical, or even impossible. Next to a pure empirical approach it is, however, according to past research, still possible to mathematically predict the influence of such complex structures on a macro-level.

Causebrook and others [24][25][26] researched the problem of a built-up terrain on LF and MF ground-wave propagation to predict the effect on the coverage of BBC LF and MF radio broadcasts. The built-up area was modeled as a medium placed between the ground and the “free-space” above that. In this model, many manmade structures (e.g. lamp posts, steel
framing, house-wiring, and plumbing) may be considered as earthed unipoles for LF. Trees are modeled by vertical resistive objects. Moderately built-up areas, such as residential suburbs, are represented by a randomly distributed array of vertical unipoles, whereas more densely built-up areas, such as city centers, are represented by a grooved surface, the grooves corresponding to streets. The resulting signal-attenuation calculations fit the measurements very well although locally rather large differences between model and measurements of the electric fields where observed due to tall buildings. Therefore, this BBC-model can be considered insightful for macro-level coverage prediction, but less for a more detailed micro-level analysis. Furthermore, only the signal attenuation, not signal phase was considered making it again less suitable for the prediction of LF positioning performance.

A crucial parameter in the iterative process of model development is the availability of accurate and reliable measurements. In the past, only signal strength measurements with reasonable accuracy were possible, because phase measurements require accurate knowledge of the position and time, which was at least challenging at the time. Current technical capabilities can help break through these barriers. Therefore, it is the author's approach to not yet continue with further development of re-radiation theory, but to first design and implement an accurate measurement system. This system can then be used in follow-up research for both an empirical re-radiation assessment as also as a crucial ingredient for the iteration of re-radiation modeling techniques.

3.5.2 Sensor response in a re-radiation environment

A re-radiated signal usually originates from a different direction than the direct signal. Therefore, the angular antenna response influences the measured effect of re-radiation. Furthermore, the radiation pattern of the antenna itself can be influenced by local effects, which make the sensor an integral part of the re-radiation measurement, and hence, part of the impact of re-radiation on the positioning and timing performance.

E-field antenna

The LF E-field antenna is usually implemented as an infinitesimal monopole antenna. This antenna, with omni-directional sensitivity, probes the incident field with respect to an arbitrary ground. Imagine, for example, a boat on saltwater: the water forms a conducting ground plane with, as a result, a well-defined antenna phase center. An automotive installation on the other hand has a much more undetermined response. The car chassis can be isolated from the road, but might have parasitic capacitive couplings to objects and structures nearby (Section 5.1.1). This makes the radiation pattern of an infinitesimal monopole E-field antenna mounted on a car also a function of its surroundings, which might explain the large common-mode amplitude and phase fluctuations noticed in among others the Boston measurement campaign presented in Section 6.3. Note that the common mode component of the phase fluctuations cancels in the positioning solution as a receiver clock error, and therefore, does not lead to a positioning error.

1 It is expected that the trees show less attenuation of the electric field in the winter compared to the summer because of the reduction in sap resulting in an increase in resistance. This hypothesis has been confirmed by measurements [24].
**H-field antenna**

The LF H-field loop antenna has a figure-8 radiation pattern. Two loop antennas are combined and the resulting beam can be electronically steered such that an optimal SNR towards a station is obtained. This can be done in software for all the received stations individually (Section 5.7.2).

Figure 3-27 shows a scenario where the receiver is in the presence of a local re-radiator. In the first situation, the re-radiator is located in the opposite direction from the desired signal. The radiation pattern of the H-field antenna causes the re-radiated signal to be inverted and then added to the direct signal. In the second situation, the re-radiated signal is located in the null of the antenna radiation pattern and can therefore be ignored. The third situation will cause a non-inverted, but attenuated re-radiated signal to be added to the direct signal. Obviously, the situation would drastically change if the direct signal would come from another direction, because the electronically formed antenna beams would then have a different orientation. This makes the effect of a single re-radiator entirely different for signals coming from different directions.

The H-field antenna radiation pattern forces the effect of re-radiation to be different towards different stations. Re-radiation in the H-field, therefore, always introduces a positioning error.

![Figure 3-27: The influence of re-radiation on the received H-field signal depends on the heading towards the transmitter and the relative location of the re-radiator with respect to the receiver.](image)

The differential-mode behavior of the H-field antenna forms a plausible explanation of the often-larger positioning errors in the H-field than in the E-field in re-radiation situation as observed in the measurements presented in Sections 6.3 and 6.4.6.

### 3.5.3 Observability of re-radiation

Detection of re-radiation is an important, perhaps even necessary, tool for various applications. For example, it can help to prevent a monitoring station to transmit misleading information to the users. Another use is the added integrity for maritime receivers by preventing misinterpretation of the propagation anomalies caused by large structures such as bridges, for example. Detecting re-radiation increases integrity at the cost of lower availability. As with every integrity tool, misleading information caused by misdetection has to be prevented without increasing the chance of false detection too much.
Re-Radiation can be detected in two ways: by analyzing the relation between the received E-field and H-field and by analyzing the spatial components of the received signals.

### 3.5.3.1 Verification of the $120\pi$ far-field relation

If the far-field condition is satisfied, the E-field and H-field have a fixed relation:

$$\frac{|E|}{|H|} = Z_0 = 120\pi \, \Omega , \tag{3-9}$$

Re-radiation, however, distorts this relation. This makes it trivial to detect re-radiation by comparing the outputs of an E-field and an H-field antenna. Note, however, that a well-designed H-field antenna is a true field probe; whereas, an E-field infinitesimal monopole antenna measures the incoming field with respect to a somewhat arbitrary ground. Therefore, a snapshot measurement of only a single transmitter will not be conclusive. Considering multiple stations, however, should give more confidence in the validity of the far-field relation.

Two applications are likely to benefit from the validation of the E-field vs. H-field relationship:

- **Monitoring.** Monitoring sites are used to send precise time-of-arrival measurements, often with respect to an absolute clock, to the users to enable precise differential positioning or differential timing. These monitoring sites often require an integrity monitor equipped with an autonomous receiver. If the integrity monitor is outfitted with a different antenna type than the primary receiver, then not only the functioning of the receivers, but also the stability of the E-field vs. H-field relation is monitored. In this way, local temporal disturbances of the field can be detected and flagged rather than to send faulty corrections to the users.

- **ASF surveying.** Knowledge of any possible near-field propagation effects is desirable when surveying such places as a harbor to enable precise positioning. Measuring both the E-field and the H-field will enhance the integrity of the required far-field ASF database.

In most cases, it is probably not cost-efficient to outfit a user receiver with both an E-field antenna and an H-field antenna. Fortunately, it is also possible to detect re-radiation in the spatial domain if the receiver is equipped with a dual-loop H-field antenna.

### 3.5.3.2 Spatial domain re-radiation detection

The following analysis assumes a single frequency (CW) radionavigation signal. The wideband Loran-C signal will likely lead to the same results, although it should be noted that pulse distortion caused by such things as dispersion is not taken into account.

For the receiver, a transmitted signal is characterized by its phase, amplitude, and its heading towards the observer (the modulation of the signal is ignored). Figure 3-28 depicts the geometrical situation.

If we assume a perfect H-field antenna with $\cos(\varphi)$ and $\sin(\varphi)$ response for loop 1 respectively loop 2 then the measured signal is depicted by the following:
$loop_1 = A \cdot \cos(\varphi - \delta)$

$loop_2 = A \cdot \sin(\varphi - \delta)$

(3-10)

where $A$ is the complex amplitude (phase + amplitude) and $\delta$ the direction of either the “direct signal” ($\delta_D$) or a re-radiated signal ($\delta_R$), and $\varphi$ is the H-field antenna orientation with respect to North.

We can rewrite 3-10 as:

$\begin{align*}
\cos \sin \\
\sin \cos
\end{align*}$

(3-11)

Or in matrix notation:

$\begin{pmatrix}
\cos \sin \\
\sin \cos
\end{pmatrix} = 
\begin{pmatrix}
A \cdot \cos(\delta) & A \cdot \sin(\delta) \\
-A \cdot \sin(\delta) & A \cdot \cos(\delta)
\end{pmatrix}
\begin{pmatrix}
\cos(\varphi) \\
\sin(\varphi)
\end{pmatrix}$

(3-12)

If we substitute $R_A$ for $A \cdot \cos(\delta)$ and $R_B$ for $A \cdot \sin(\delta)$ then the equation becomes:
The antenna response to the “direct signal” is now represented by two complex numbers $R_A$ and $R_B$, which have the same phase. The same exercise can be repeated for a number of (re-radiated) signals, each resulting in a 2x2 matrix, which will all be added linearly by the antenna.

In summary: if re-radiation can be modeled by a direct signal combined with an infinite number of parasitic re-radiators then the response of an (ideal) H-field antenna will be a 2x2 matrix. This matrix, which we will denote as the “re-radiation matrix” will effectively contain 2 complex numbers. The phase of the different sources will most likely be different, and therefore, also the phase of $R_A$ and $R_B$.

Note that there are an infinite number of re-radiation situations possible, each leading to the same re-radiation matrix. Without additional information, it is not possible to separate the direct signal from the re-radiated signals in the 2D spatial domain.

Detection of re-radiation is now possible in two ways:

$\angle R_A \neq \angle R_B$ or: Antenna loop 1 measures a different phase than Antenna loop 2. This also results in the fact that the signal cannot be completely nulled (electronically steer the antenna beam such that the response is minimized).

Another spatial domain detection method relies on the fact that a re-radiated signal appears to be coming from the “wrong” direction. This method obviously needs some heading reference, which can be provided by using the signals from multiple transmitters. The correct heading differences between the signals from the transmitters can be calculated given an approximate user location and the location of the transmitters. A re-radiation situation is detected if the measured heading differences disagree with the calculated differences. Note that with this method the re-radiation observability is a function of transmitter geometry.

### 3.5.3.3 Influence of antenna imperfections on the spatial domain re-radiation detection

The assumption of a perfect H-field antenna-response is seldom valid. Section 5.8 shows that H-field antenna tuning, gain and cross-talk errors can be modeled as:

\[
\begin{align*}
\text{loop1} &= G_1 (\cos \varphi + A_{21} \cdot \text{loop2}) \\
\text{loop2} &= G_2 (\sin \varphi + A_{12} \cdot \text{loop1})
\end{align*}
\]  

(3-14)

where the complex parameters $G_1$ and $G_2$ depict the antenna tuning and gain and $A_{21}$ and $A_{12}$ the effect of cross-talk.

Formula 3-14 can be rewritten as:

\[
\begin{pmatrix}
\text{Loop1} \\
\text{Loop2}
\end{pmatrix} =
\begin{pmatrix}
R_A & R_B \\
-R_B & R_A
\end{pmatrix}
\begin{pmatrix}
\cos(\varphi) \\
\sin(\varphi)
\end{pmatrix}
\]  

(3-13)
\[
\begin{bmatrix}
\text{Loop1} \\
\text{Loop2}
\end{bmatrix} = \frac{1}{1 - G_1 G_2 A_{21} A_{12}} \begin{bmatrix} G_1 & G_1 G_2 A_{21} \\ G_1 G_2 A_{12} & G_2 \end{bmatrix} \begin{bmatrix} \cos(\varphi) \\ \sin(\varphi) \end{bmatrix}
\] 
(3-15)

After the following substitutions:

\[X_1 = \frac{G_1}{1 - G_1 G_2 A_{21} A_{12}}, X_2 = \frac{G_1 G_2 A_{21}}{1 - G_1 G_2 A_{21} A_{12}},\]
\[X_3 = \frac{G_1 G_2 A_{12}}{1 - G_1 G_2 A_{21} A_{12}}, X_4 = \frac{G_2}{1 - G_1 G_2 A_{21} A_{12}}\]

Formula 3-14 is rewritten as:

\[
\begin{bmatrix}
\text{Loop1} \\
\text{Loop2}
\end{bmatrix} = \begin{bmatrix} X_1 & X_2 \\ X_3 & X_4 \end{bmatrix} \begin{bmatrix} \cos(\varphi) \\ \sin(\varphi) \end{bmatrix}
\] 
(3-17)

where the 2x2 matrix from now on is called the cross-talk or “X-talk” matrix.

The response of a non-ideal antenna to a re-radiated signal can now be given by:

\[
\begin{bmatrix}
\text{Loop1} \\
\text{Loop2}
\end{bmatrix} = \begin{bmatrix} R_A & R_B \\ -R_B & R_A \end{bmatrix} \begin{bmatrix} X_1 & X_2 \\ X_3 & X_4 \end{bmatrix} \begin{bmatrix} \cos(\varphi) \\ \sin(\varphi) \end{bmatrix}
\] 
(3-18)

From this it can be concluded that antenna cross talk becomes indistinguishable from re-radiation when looking at a single station. The antenna first needs to be (auto) calibrated (which is effectively the determination of the inverse of the cross-talk matrix) before spatial domain re-radiation detection is possible.

### 3.5.4 Re-radiation correction

LF re-radiation is physically the same as GPS-multipath, but the consequences depend on the wavelength considered. Current GPS receivers and antennas show that for most applications the GPS-multipath problem has been reduced to manageable proportions. Is the same possible for LF re-radiation? Can special hardware and processing techniques help us to remove the re-radiation from the composite signal?

#### 3.5.4.1 Separation of the re-radiated signal from the direct signal

It might be possible to separate the re-radiated signal from the direct signal if they differ in one or more signal domains.

**Coding and time domain**

It is possible, in principle, to separate the direct signal from the reflected or re-radiated signal in the time domain, but only if the distance to the re-radiator or reflector is significantly large compared to the wavelength of the radio waves and if suitable signal modulation such as CDMA or pulse modulation is used.

The GPS systems meet the described criteria in most situations. At the GPS L1 frequency (1.575 GHz), most objects are far away compared to the wavelength (19 cm). The CA-code...
modulation combined with a suitable correlator (e.g. a 0.1 chip “narrow correlator”) can drastically reduce the impact of multipath.

Loran-C ground-wave-sky-wave separation shows a comparable situation. At distances up to approximately two thousand kilometers, the sky-wave path is significantly longer than the ground-wave path. Usually the path length difference ensures a time-of-arrival difference of at least 30 µs between the ground wave and the sky wave. Sky-wave rejection is ensured if a receiver tracks the Loran-C signal somewhere in the first 30 µs of the pulse. Note that the influence of some rare solar phenomenon on the ionosphere can cause the ground-wave-sky-wave delay to reduce to less than 20 µs. During these circumstances, also referred to as “early sky waves,” reliable ground-wave-only tracking is questionable.

Re-radiation rejection in the time/coding domain analogue to sky-wave rejection is not possible, because the phase difference between the direct signal and the re-radiated signal is not more than a fraction of the carrier period. Unfortunately, due to the long wavelength (3 km), even a small phase error can lead to a significant range measurement error.

**Spatial and polarization domain**

Section 3.5.3.2 shows that the radiation pattern of an H-field antenna can be used to detect re-radiation in the spatial domain. The analysis can be extended with a 3rd (Z) direction to include polarization. The LF signal is vertically polarized, so any horizontal polarized signal is likely to be caused by re-radiation.

The H-field antenna allows for a sharp null in only one direction accompanied with a wide beam (-3 dB beam width of 90°) in the orthogonal direction. Hence, re-radiation rejection is only possible if there is only a single re-radiator and when its heading is known exactly. Unfortunately, the assumption of only a single re-radiator will seldom be valid and furthermore, finding the exact heading toward the re-radiator will be a difficult, perhaps impossible, challenge. Instead of nulling the re-radiation, it would be more desirable to focus a narrow beam on the direct signal. Unfortunately, the implementation of a classical phased-array type antenna at LF will require a structure of immense proportions and is, therefore, not achievable.

In comparison, the spatial domain is much easier to exploit at GPS frequencies. Relative compact antenna configurations such as a choke-ring antenna or NovAtel’s “Pinwheel®” antenna are possible. Phased-array antenna techniques are currently primarily used by military systems, but are likely to come into the commercial sector in the future.

**Frequency domain**

Analogous to the time-domain, the phase and Doppler difference between the direct and the re-radiated signal is too small to enable separation of the signals.

**E-field vs. H-field**

Comparing the received E-field signal with the H-field signal is an easy way to detect re-radiation. Currently, however, there seems to be no method of using the E-field signal, for example, to correct the H-field signal for the influence of re-radiation.
3.5.4.2 Exclusion of measurements polluted with re-radiation

In an over-determined positioning solution (more than three transmitters for a 2D position solution), those ranges polluted with re-radiation can be excluded from the final solution. Methods described in Section 3.5.3 can be used to select those range measurements that suffer most from re-radiation.

The problem with this method is that given the current transmitter constellation there are seldom more than 4, or sometimes even only 3, transmitters close enough to the user to provide a relatively high quality position fix. In addition, if an object causes re-radiation on the signal from one transmitter it is reasonable to assume that the same object will also have some effect on the signals from one or more other transmitters.

The exclusion scenario becomes much more feasible if the (absolute) heading is the primary parameter of interest to the user. Only a single station and an approximate position are needed to derive an absolute orientation. This usually allows multiple stations to be discarded based on suspected re-radiation contamination.

3.5.4.3 Survey-and-correct re-radiation

To meet the stringent accuracy requirements of some applications, such as Harbor Entrance and Approach, it is necessary to survey the propagation effects in the region of interest. It is, of course, also possible to survey the near-field re-radiation propagation effects and to include them in the correction database. When this survey-and-correct method is used, several issues must be taken into account:

Temporal stability

The survey-and-correct method assumes that the influence of re-radiation is constant in time. This will be mostly true for a bridge or a building, for example. Although, it is possible that rain and snow, for example, change the re-radiation effect of these objects. During a survey campaign, it should be decided whether a certain object is likely to show a temporally stable re-radiation effect.

Spatial resolution of correction map

The influence of re-radiation is extremely dependent on the user position. Ten-meter displacement can already show a significant change, especially when the receiver is very close to a large bridge, for example. This results in a significant chance of under-sampling the effects, even at a grid-distance of 10 meters, for example.

Different corrections for E-field and H-field receivers

Re-radiation is a near-field effect, and therefore, requires different corrections for an E-field and H-field receiver.

Risk of ambiguous position solution

If the re-radiation correction is significant, then the risk of ambiguous position solutions exists: two or more physical positions with the same TOA measurements. Based on these
TOAs, it becomes dangerous or even impossible for the receiver to decide what corrections it should use. Section 6.4.6 shows examples of such situations.

**Influence of receiver implementation**

With an H-field receiver, the actual effect of re-radiation is a function of its antenna beam-steering algorithm. In situations of severe re-radiation, it is not trivial to determine the exact antenna orientation, because the individual heading measurements will mutually disagree. The final reported antenna orientation then depends on the weighting-scheme of the receiver. Therefore, the beam-steering algorithm implementation of the receiver is an important factor in the total surveying and correction process.

It can be concluded that, next to several practical issues, the correction of re-radiation by using surveyed values is potentially risky, and it is best avoided.

### 3.6 Noise and Interference

Noise and interference are responsible for the degradation of the repeatable accuracy of a radionavigation system. The system performance and possible processing gain are highly dependent on the detailed statistical characteristics of both the signal and the noise. For example, the effect of zero-mean Gaussian noise can be reduced only by longer integration by the receiver at the cost of reduced dynamic responsiveness; whereas, pulsed noise caused by local lightning strikes can be mitigated by blanking these pulses in the time domain.

Noise and interference can be divided into two categories, noise internal and noise external to the receiving system. The internal noise is produced by the receiver itself: by the antenna and associated amplifier, which is described in detail in Section 5.4. This Section describes various external noise sources and possible mitigation techniques.

#### 3.6.1 Atmospheric noise

External noise can be either man-made or caused by natural phenomena. The latter can be divided into atmospheric and galactic noise. Atmospheric noise will be dominant over galactic noise for frequencies below 30 MHz [28].

Atmospheric noise is generated by electrical discharges between clouds or between the clouds and the ground (in the form of lightning). The energy from these discharges is wide band and peaks at 10 kHz. Sky-wave propagation allows these low-frequency waves to be detected thousands of kilometers from the source. The lightning discharges from large distances, several thousands kilometers (for example), will be received randomly spaced in time and at a rather high rate, although the amplitudes will not vary over extreme values. Addition of all these lightning strikes results in a more or less Gaussian distribution [51]. This part of the atmospheric noise can only be mitigated by averaging, and it forms the lower limit of the external noise.

Lightning discharges at closer proximity to the receiver can be very energetic. There is a significant time between strikes [34] where the background noise is only modest and signal reception is possible [29]. A Loran-C station transmits a series of 8 pulses (9 for the master
station) spaced one millisecond apart. This pattern is repeated every GRI, which lies between 40.0 and 99.99 ms (Section 2.4.2). A single lightning strike will most likely render a maximum of one received Loran pulse useless. However, a Loran receiver integrates many pulses before taking a measurement. Even if only a single pulse is severely distorted by lightning, this may render the whole (linear) integration of perhaps seconds useless. Therefore, it is crucial to detect and drop the distorted pulses before the averaging process. This technique is also known as hole punching, and some research claims that at least 15 dB of impulse-noise reduction is obtainable in this way [29]. At least a considerable noise reduction is needed to meet the availability and continuity requirements of various applications [35][29]. To get more insight in the obtainable processing gain by non-linear noise reduction techniques, it is imperative to subject the statistics of the noise amplitude distribution to further analysis.

3.6.1.1 CCIR

An extensive analysis of atmospheric noise levels has been published in the CCIR 322-2 document. [28]. This report provides estimates for atmospheric noise as the average background noise level to lightning in the absence of other signals, whether intentionally or unintentionally radiated. The CCIR report is based on 4 years of data collected from 1957-1961, at 16 stations around the world, and provides statistics on the noise-factor, the impulsiveness of the noise, and the amplitude probability distribution.

Figure 3-29 depicts the CCIR noise levels at 100 kHz for Europe and Florida for the four seasons, and for different times of the day. A first glance at these plots shows a stunning difference between the two locations: the atmospheric noise during summer evenings in Florida can be as much as 20 dB higher than during the same time in Europe. Note that these plots do not consider the actual noise amplitude distribution. The high noise levels in Florida are primarily caused by high-energy, low-repetition-rate pulses resulting from local thunderstorms. These pulses can probably largely be rejected by the receiver resulting in a more moderate difference between the two locations. A more global view of the atmospheric noise levels is shown in Figure 3-30 [43], which again does not reflect the ability to mitigate...
the impulsive part of the noise. Note that the differences between the absolute noise levels between Figure 3-29 and Figure 3-30 are caused by the different bandwidths considered.

![Noise (dB) Summer @ 1800 Local](image)

**Figure 3-30: Comparison of CCIR noise predictions for North America and Western Europe (Illustration from reference [43])**

Although insightful, there are several limitations to the collected data shown in the CCIR report. First, the measurements were taken using very narrow-band receivers (200 Hz). Such a small bandwidth limits the measurement accuracy of the steep and short pulses associated with local lightning strikes and the effect of these pulses on the relatively wide-band Loran-C system. Another drawback is that only background noise was collected, local thunderstorms were omitted from the results while they caused corona-effects in the receiving equipment. Although the effect of local storms was omitted from the measurement data, it is doubtful whether the CCIR-report provides a sufficiently accurate assessment of the impact of atmospheric noise on the LF radionavigation performance and on the possible processing gain by special non-linear processing.

### 3.6.1.2 Amplitude distributions

Feldman [52] repeated some of the CCIR noise measurements in the 1970s under three different types of weather conditions: “quiet,” “tropical,” and “frontal.” These three types of weather conditions are intended to be representative of real-life situations. He provided the probability densities (pd) of the noise amplitudes for the three types of weather, and showed that the lower amplitudes have a Gaussian character, whereas the probability density of the higher amplitudes depends highly on the type of weather. Van der Wal and van Willigen made a mathematical approximation of the pds provided by Feldman [53]. Figure 3-31 represents these equations graphically.
More recently, the Ohio University Avionics Engineering Center (AEC) has recorded many hours of ground data and airborne data in the period of 2003-2005. Most of the data is recorded in Florida, because that area is infamous for its high thunderstorm activity. The two antennas used for the measurements were an omni-directional quadrature H-field loop antenna and an E-field whip antenna \[31]\[32]. After removal of all Loran-C signals, CW, and high-energy noise spikes (local lightning flashes), the remainder of the noise has, as expected, a close resemblance to Gaussian noise \[30]. At the time of this writing, efforts are being undertaken to derive both the Cumulative Distribution Functions (CDFs) and the statistics of the pulsed part of the measured noise. This should lead to realistic estimates of the possible noise reduction by non-linear processing and of the impact of atmospheric noise on the (airborne) navigation performance after such techniques have been applied.

For determining accurate in-band noise statistics, it is crucial to first remove all the Loran-C signals. Table 3-1 and Figure 3-32 show the effect of various Loran “removal” techniques on the noise statistics and energy. The source data was measured at Delft during “quiet” weather: with minimal atmospheric activity.

**Table 3-1: Effect of different methods of Loran removal**

<table>
<thead>
<tr>
<th>Loran removal method</th>
<th>Median noise level</th>
<th>Pct of data used</th>
</tr>
</thead>
<tbody>
<tr>
<td>No Loran removal</td>
<td>61.6 dB</td>
<td>100%</td>
</tr>
<tr>
<td>Loran cancelled</td>
<td>31.6 dB</td>
<td>100%</td>
</tr>
<tr>
<td>Loran blanked</td>
<td>28.0 dB</td>
<td>8.3%</td>
</tr>
<tr>
<td>Strongest 9 Loran stations blanked</td>
<td>32.6 dB</td>
<td>30.6%</td>
</tr>
</tbody>
</table>
The Loran energy during this measurement was about 33 dB above the background atmospheric noise level in the 90-110 kHz band. Simply throwing away (blanking) all tracked Loran-C signals is the simplest and most effective way to reliably measure the noise amplitude distribution. Loran blanking, however, limits the ability to measure the statistics of the individual noise spikes as most noise spikes will, at least partly, overlap with a Loran pulse that eventually will be blanked. Therefore, the complete noise-spike is not available after blanking. With Loran-C estimate-and-subtract or “canceling,” none of the samples are discarded. Unfortunately, perfect canceling is not possible, as will be explained in the Section 3.6.3. Therefore, the residual Loran-C can still affect the noise measurements, especially if the atmospheric noise levels are relatively low.

From Figure 3-32 it becomes evident that Loran-C cross-rate removal is at least as important as atmospheric noise mitigation. Fortunately, the deterministic character of Loran-C cross-rate allows for effective mitigation, as will be discussed in detail later in this Section.

![Figure 3-32: Effect of different methods of Loran removal on CDF (left) and noise level (right). Note that the noise statistics have been compensated for the actual number of samples used.](image)

**3.6.1.3 Atmospheric Noise Mitigation**

Even a single high-energy atmospheric noise spike can severely deteriorate the performance of a LF radionavigation receiver if no counter measures are taken. Therefore, mitigation is crucial for the receiver to remain functional during moderate to severe atmospheric activity. The non-Gaussian or “impulsive” part of atmospheric noise can be mitigated in several ways. Figure 3-33 depicts several of these methods as described by Feldman [52].

A major advantage of the hard-limiter design is its hardware simplicity. The limiter circuit converts the analog input into two levels, after which only limited computing power is needed to compute the navigation solution. With respect to atmospheric noise, the hard-limiting receiver provides near optimum performance in most cases, especially those with a longer observation interval [52]. This type of receiver is, however, less robust against other types of
interference (e.g. CW), and it generally provides less tracking accuracy. Therefore, the hard-limiter implementation has been abandoned in favor of the linear receiver.

The “clipper” and “hole-punch” techniques are relatively simple to implement in a linear receiver. The “clipper” method clips all signals above a certain threshold value towards that threshold; whereas, the “hole punch” method replaces those samples above the threshold by zero. The implementation of an adaptive-threshold clipper seems to provide near optimal results for most SNRs [52].

Compared to some decades ago, modern receivers are equipped with massive computer power, which allows more advanced types of noise mitigation. Instead of clipping the signal or replacing the signal by zeros, it is also possible to exclude those by noise infected samples from the averaging and tracking process. The calculation of the achievable noise-reduction for this scenario is straightforward. The reduction of noise energy is given by the sum of all the noise samples above the threshold divided by the total noise energy:

\[
\Delta \text{Noise} = 10 \cdot \log \left( \frac{\sum_{A=\text{threshold}} PDF(A) \cdot A^2}{\sum_{A=0} PDF(A) \cdot A^2} \right)
\]

where PDF is the noise amplitude probability density function and A the noise amplitude (given by \( A = \sqrt{I^2 + Q^2} \)). The loss in signal power, due to disregarding those samples that are too noisy, is given by the probability that the noise energy exceeds the threshold.

\[
\Delta \text{Signal} = 10 \cdot \log \left( 1 - \text{CDF} \left( \text{threshold} \right) \right)
\]

Figure 3-34 shows the theoretical obtainable gain when the described method is applied on the “clean” data file measured in Delft.

Two issues surface when further analyzing this mitigation technique: if the threshold is lowered, the theoretical hole punching gain seems always to increase. However, Figure 3-32-left shows that for this dataset, the noise with amplitudes below approximately 100 has predominantly a Gaussian character for which hole punching should not provide any performance increase. Rejection of noise samples within this “Gaussian” region will be practically impossible, because the unknown information carried by the desired signal (the
actual phase, amplitude, and shape of the navigation signal) makes it impossible to actually determine the noise contribution of these individual samples within the “Gaussian” region. Therefore, those hole-punching gains presented in Figure 3-34 belonging to an amplitude threshold below approximately 100 should be discarded as physically unrealistic.

Secondly, selectively disregarding only noisy samples imposes the threat of introducing discontinuities within the averaged Loran-C pulse. This is because user dynamics, propagation, and transmitter fluctuation causes the phase and amplitude of the received Loran-C pulse to fluctuate on a pulse-by-pulse basis. Therefore, updating only a portion of the pulse will likely introduce discontinuities. Instead of selectively rejecting only noisy samples, it is better to reject the (partly) contaminated Loran-C pulse completely. In the remainder of this text, this method will be denoted as “Loran pulse punching.”

Figure 3-35 shows actual data, where some Loran-C pulses are hit by atmospheric noise spikes. The blue plot denotes the input signal and the overlaid red plot shows the same input signal, but with the expected Loran-C signal subtracted.

For tracking the Loran-C signal, the part from just before the start up to the peak of the pulse is of interest. The leading edge is needed for cycle-identification and contains the only valid time-of-arrival of the signal. The peak of the pulse, likely contaminated by sky waves, contains much more energy but only be used only for relative timing, for example for as data demodulation and short-term dynamics tracking. The tail of the pulse is not controlled in the transmitter accurately, and it should not be used. Therefore, if a noise spike hits the tail of a Loran-C pulse it should not be considered as a loss; whereas, if the leading edge of the Loran-C pulse is hit we should discard the whole pulse for tracking. This results in the first and second pulse of Figure 3-35 to be used for tracking while the 3rd pulse is discarded.
Next to the PDF or CDF noise statistics, both the “effective” length of the Loran pulse as also the average length of the noise spikes are needed to calculate the “Loran-punching” gain. Figure 3-36-left shows the effect of bandpass filtering ($10^{th}$ order 20 kHz Butterworth) on a short pulse. Given this filter bandwidth, a filtered noise spike cannot be much shorter than a Loran-C pulse. The duration of the physical phenomenon that causes the atmospheric noise spike is generally also in the order of tenths of microseconds [34]. This combined makes a duration of 100 µs or longer plausible, which is confirmed by some actual measurements shown in Figure 3-36-right.

**Figure 3-35:** The first and second Loran pulse are not significantly affected by noise, the third pulse should be completely discarded for tracking.

**Figure 3-36:** Left: theoretical effect of bandpass filtering on the length of a very short noise spike. Right: measured length of noise spikes in the “clean” dataset presented earlier.
The combined knowledge of both the effective Loran-C pulse length and the noise spike length allows us to calculate the additional signal loss when we punch out a complete Loran-C pulse (Loran Pulse Punching) instead of only the part actually contaminated by the noise spike (Hole Punching). Figure 3-37 shows this relation.

For example: if we are interested in 200 µs of the Loran pulse and if the average noise spike length is approximately 100 µs, than we loose approximately 5 dB with respect to the theoretical hole punching gain if we use “Loran pulse punching.” The theoretical hole-punching gain can be derived directly from the CDF or PDF noise statistics.

This Section has shown the CDF and noise spike statistics of a “quiet” dataset. Ohio University’s Avionics Engineering Center is currently analyzing both ground-based and airborne data of moderate and severe weather conditions [31]. Hopefully, with the statistics derived from that data, it will be possible to assess the actual impact of (local) thunderstorms on the effective noise level after non-linear processing has been applied. The next step will be to verify these analyses with the tracking results of an actual receiver implementation.

### 3.6.2 Continuous wave interference

The frequency band from 90 to 110 kHz is strictly reserved for radionavigation [40], and is therefore, usually free from any intentionally radiated Continuous Wave Interference (CWI).

Figure 3-38 shows part of the LF spectra of both Europe and the United States. It can be clearly seen that the spectrum between 50 and 150 kHz in the USA is significantly cleaner than that of Europe, where some 1000 transmitters are identified, which broadcast close to the Loran-C frequency band [41]. The signal strength of all these interferers is a function of both distance and propagation, and it can become quite severe at nighttime due to the increased sky-wave propagation.
The effect of CW interference on the performance of the Loran-C receiver depends on the signal strength of that interferer, relative to the desired Loran-C signal, and on its frequency as well. The tracking loop in the receiver acts as a comb filter, where each pass band is centered on a spectral line of the transmitted Loran-C signal.

The receiver comb filter can be modeled as depicted in Figure 3-40, where the receiver tracking loop bandwidth is represented by $f_B$. Older receivers often used integration times of a minute or longer, resulting in a tracking loop bandwidth of less than 0.02 Hz. Modern receivers usually integrate much shorter for their phase tracking, for example, between 1 and 10 seconds resulting in tracking loop bandwidths between 1 and 0.1 Hz. Note that the pulse
envelope measurement (used for the cycle identification process) is averaged significantly longer, up to several minutes, resulting in an $f_0$ for the envelope-tracking of usually less than 0.01 Hz.

Three types of CW interference can be identified [33]:

### 3.6.2.1 Synchronous interference

Synchronous interference signals are those signals that coincide with the Loran-C spectral lines exactly, and therefore, fold back to $f=0$ after averaging by the receiver:

$$f_{CWI} = N \cdot \frac{1}{2GRI}, \quad N = 1, 2, 3, \ldots$$

The worst-case phase tracking error occurs if the CWI interferer has a phase of 90° or 270° relative to the carrier Loran-C signal. This will introduce a time-invariant time-measurement error of [50]

$$t_{error} = \frac{10 \mu s}{2\pi} \sin^{-1} \left( \frac{1}{SIR} \right)$$

where $SIR$ is the Signal-to-Interference Ratio. Synchronous interference can cause serious tracking errors even if the interference level is well below the Loran-C signal strengths. Synchronous interference can also add a DC-component to the pulse envelope, thereby deteriorating the cycle identification process. The worst-case envelope distortion occurs if the CW interferer has a phase of 0° or 180° relative to the Loran-C carrier. This envelope distortion, combined with the phase tracking error, leads to a change in measured ECD, which subsequently endangers correct cycle identification.

### 3.6.2.2 Near-synchronous interference

Near-synchronous interference signals are those signals that fold back to frequencies smaller than the tracking bandwidth $f_B$.

$$f_{CWI} = N \cdot \frac{1}{2GRI} + \Delta f, \quad N = 1, 2, 3, \ldots, \quad |\Delta f| < f_B$$

Near synchronous interference will introduce an oscillating phase error. Envelope tracking loops, and hence the Loran-C cycle identification process, are less likely to experience near-synchronous interference as their bandwidth $f_B$ is much smaller.
3.6.2.3 Asynchronous interference

Asynchronous interference signals are those signals that fold back to frequencies larger than the tracking bandwidth \( f_B \).

\[
f_{\text{CWI}} = N \cdot \frac{1}{2 \text{GRI}} + \Delta f, \quad N = 1, 2, 3, \ldots, \quad |\Delta f| > f_B \tag{3-24}
\]

In the long term, asynchronous interference will be attenuated completely due to the low-pass characteristics of the tracking loop. In the short term, however, asynchronous interference will increase the noise level and degrade the data-demodulation performance.

When choosing optimal GRI for a certain area, next to mutual Cross Rate Interference (CRI), CWI is a crucial parameter. All US Loran-C transmitters use a GRI that are multiples of 100 µs, and therefore, are all sensitive to synchronous interference on multiples of 5 kHz. Because Europe has many CWI at those frequencies, the European NELS GRI are multiples of 10 µs, which introduces a line spectrum with a period of 50 kHz. [42]

3.6.2.4 Bandpass filtering

An essential first step of CWI-mitigation is the appliance of a bandpass filter. The situation in Europe favors a relative narrow, steep filter, because many interference sources are present fairly close to the Loran band. However, a narrow filter causes a slower rising pulse, and hence, more severe attenuation of the start of the pulse. This effect is shown in Figure 3-41, which depicts the result of a 20 kHz wide 8th order Butterworth filter. The extra attenuation is undesired considering that the receiver is required to track the phase very early in the pulse, at 30 µs, to prevent sky-wave contamination.

The older receivers, which used analog filtering, were relatively limited in their choice of filtering. First, the filters had a limited order and secondly they needed to be causal. Modern software radio receivers allow engineers to apply any kind of filtering if processing power allows. Batch processing even allows for (non-causal) zero phase distortion filters. This considering, it would be a logical choice to apply a zero-phase 85-115 kHz brick-wall filtering, for example, effectively rejecting all signals below 85 and above 115 kHz [33].

Because almost all the energy of the Loran-C pulse is within the pass-band, the application of such an extreme filter will have limited impact on the Loran-C waveform, and hence, conserving the tracking energy early in the pulse. Unfortunately, this filtering method has its own shortcomings. The time domain equivalent of frequency domain multiplication with a brick-wall-function is the convolution with a sinc function. If the time-domain vector has infinite length and the signal is perfectly periodically then these sinc functions will cancel out exactly.
However, practical receiver implementations can only process finite batch-sizes and Loran-C does not have the required periodicity. Due to these shortcomings, non-causal filtering will cause ringing of the pulse, which can clearly be seen in Figure 3-42. So, although some modern filtering techniques are able to get rid of all off-band interference while conserving the tracking energy at the 3rd zero crossing, they will also cause the sky wave to ring back towards the ground wave, and thereby, potentially even degrading the sky-wave rejection compared to more conservative (causal) filtering approaches.

![Figure 3-41: A steep, narrow bandpass filter causes severe attenuation at the beginning of the pulse](image)

![Figure 3-42: Non-causal “filt-filt” filtering does preserve the phase and most of the tracking energy. However, a steep filter will now produce significant “pre-ringing,” thereby potentially causing sky-wave contamination at the standard tracking point.](image)
3.6.2.5 Notch filtering

Interference too close to the Loran-C band to allow for sufficient attenuation by the bandpass filter, as well as in-band interference is best mitigated by notch filtering. If only a limited number of notches is available, priority should be given to those interferers that are (near-) synchronous with GRIs used for positioning. Note that modern receivers calculate their position based on the measurements of many stations, often belonging to multiple chains. The use of multiple chains increases the possible number of (near-) synchronous interferers. Notch filters can selectively attenuate frequency components. Unfortunately, notch filters will also distort the Loran-C waveform and thereby decrease tracking performance and cycle-identification capability. Batch-processing receivers allow for non-causal notch filters, which will have minimal to no influence outside their stop band. Although the stop-band is usually very narrow, the time-domain sinc function introduced by the non-causal notch will have a very low frequency and will have limited impact.

A discrete Fast Fourier Transform (FFT) analysis can be used to find the CWI interference. Two methods are available to distinguish synchronous interference from non-synchronous interference. The first method is to use a very high resolution (and therefore computationally and memory demanding) FFT analysis. A resolution of 0.1 Hz requires at least 10 seconds of data, for example. The required memory and computation power for this spectral analysis can be reduced by analyzing only a selected part of the spectrum, by using the chirp Z-transform as proposed by Beckman [33], for example. Another method to locate synchronous interference is to synchronize the FFT-sampling process with the tracking loop of a station from the desired GRI [54] [56]. Hereby, the spectrum is effectively averaged over multiple PCI periods, which significantly attenuates the non-synchronous interferers.

However, modern “all-in-view” receivers make the necessity of identifying the synchronicity of interference with GRIs questionable. First, the receiver tracks multiple stations from many chains (GRIs), and synchronicity must be checked with respect to all these GRIs. Secondly, modern receivers integrate much shorter than old hard-limiter receivers; therefore, the modern receivers have a much higher tracking-loop-bandwidth: typically between 0.1 and 1 Hz. This increases the chance for the interference to be (near) synchronous with one of the GRIs used for tracking. Finally, the relative massive amount of computing power available allows for many notches implemented in software. Therefore, it might be more reasonable to simply notch all in-band interference.

The presence of strong local loran-C stations imposes another challenge in the interference-detection process. The spectrum of a local station can effectively mask those interfering lines that are of significant influence on the tracking performance of a more distant station. A simple, but effective, mitigation is found by only analyzing the “quiet” time period(s) between the pulse groups of the local station(s) [55].

3.6.2.6 Estimate and subtract

Maximum likelihood algorithms can be used to estimate and subtract CW interference from the received composite signal [37]. Ideally, amplitude, frequency, phase, and modulation should be estimated to remove the interference effectively and completely. In practical applications, only amplitude, frequency, and phase are estimated. Therefore, the effectiveness
CWI canceling is highly dependent on the complexity of the signal structure and the stability of the CWI.

### 3.6.3 Cross-rate interference

Loran-C is a TDMA-CDMA system, where transmitters are grouped together in chains. Within a chain, each transmitter has its own timeslot to prevent overlap with other transmitters from the same chain. Each chain has a different repetition rate, known as the Group Repetition Interval (GRI). Because each Loran-C station transmits the same pulse shape signal at the same frequency, the signals of a Loran-C chain are often disturbed by those of one or more other chains.

Cross-rate Interference or CRI is largely deterministic and mainly influenced by the position of the receiver and the timing and shape of the composite ground-wave and sky-wave signal from the cross-rating station with respect to the timing and shape of the signal from the desired station. Cross-rate patterns are periodical in nature, with a crossover time given by the following:

\[
T_{cross-over} = \frac{GRI_A \cdot GRI_B}{GCD(GRI_A, GRI_B)} \times 10^{-5} \text{s} \tag{3-25}
\]

where \( GCD \) is the greatest common divisor and \( GRI_A \) and \( GRI_B \) the GRIs in their 4-digit representation (e.g. “7499”). Most GRIs of Loran-C chains in Europe and the USA have been chosen to have large crossover periods with chains in their vicinity. However, cross-rate patterns can still contain large sub-periodicities or “short interleaving” where the pulse groups partly overlap.

As shown in Figure 3-44, the effect of cross rate is a function of the momentary phase and amplitude the cross-rating pulse including its sky wave (red) and the tracked pulse (blue).

At larger distances, the signal strength of the sky wave can easily exceed that of the ground wave, especially at night. Therefore, even if ground-wave attenuation renders a station useless for tracking, its sky wave can still cause severe cross rate, and it might very well be capable of degrading the receiver’s positioning performance. According to Figure 3-45 and [43], all the stations within a range of several thousand kilometers can contribute to cross-rate and must be taken into account.
Cross rate can have significant impact on both the phase and envelope of the tracked signal. Thereby, cross rate influences both the TOA and ECD measurement, as well as the ability to correctly decode data messages (e.g. Eurofix). It is difficult to give an exact mathematical analysis on the effect of cross rate on receiver performance, because it is a function of many propagation and timing variables.

Several CRI mitigation techniques will be described along with an assessment of their effectiveness. Measurements from Delft, the Netherlands, are used for clarification. The measurements were taken at a static location relative to an absolute clock. Nineteen stations at
distances between 400 km and 2500 km were tracked. The effects of various forms of CRI-mitigation on the tracking performance of Værlandet (7499Y, 1000 km from Delft) are compared by processing the same raw data file using different settings. It is assumed that no synchronous CWI is present, which has been verified by comparing the results of both rates of Værlandet (7499Y and 9007Y), which showed similar results.

### 3.6.3.1 Averaging

The Loran-C system is designed such that the cross rate significantly reduces after PCI-synchronous averaging of the measurements. Careful selection of the GRIs and the Phase-Code are crucial for the success of this method.

Figure 3-46 shows the envelope of the averaged signal of Værlandet after various integration times. The cross-over time (3-25) between 7499 and the adjacent chains 6731 and 9007 is approximately 8.4 respectively 11.2 minutes. After this period, the pattern repeats itself and additional CRI-reduction by longer integration becomes impossible.

However, even after approximately two minutes of averaging, the cross rate is not reduced further, which clearly can be seen at the right plot, which zooms in on the beginning of the pulse. A significant amount of cross-rate energy remains present, which will most likely cause the Cycle Identification process to fail.

This system imperfection became apparent soon after the two Loran-C chains began operating in proximity to one another. The performance impact had already been predicted in 1960 [44] and later studies where successful in identifying the root causes of the problem and produced several recommendations [45]. Unfortunately, only one of the recommendations was carried out, namely: choosing chain repetition rates, which were relatively non-interfering. An example of optimal GRI selection was done by Delft University for the European NELS-chains [42]. As Figure 3-46 shows, only this GRI selection does not solve the cross-rate problem. In 1975, Feldman insistently suggested that in addition to choosing optimal GRIs, it would be well to optimize the phase-code of Loran-C [46], thereby dramatically reducing the cross-rate problem. Unfortunately, Feldman’s proposed changes appeared to be too radical for the established Loran-C system and the policy makers involved.

Whereas the classic hard-limiting receivers were relatively capable to handle cross-rate by long integration, the modern linear receivers are much more sensitive to this system imperfection. Even the smallest discrepancy in the cross-rate mitigation can lead to unacceptable errors, especially if the cross-rating signal is significantly stronger than the tracked signal. This can be explained by observing that with linear averaging, a single energetic uncompensated cross-rate pulse can destroy the averaging for a long period. A hard-limiting receiver simply limits the severe cross rate and is, therefore, less susceptible.

Other drawbacks of long integration include the reduction of the dynamic response of the receiver and the inability to reduce cross rate for data demodulation. For the latter, as information is stored in each transmitted pulse separately, averaging of pulses will obviously not help the process of data recovery.
3.6.3.2 Blanking

A simple, and possibly the most obvious, cross-rate mitigation technique is to simply discard all pulses that are hit by cross-rating pulses. This can be either done by detecting abnormal pulses (pulses that do not match the average pulse received from that station), or by tracking the cross-rating station and calculating when the two stations will overlap in time.

The cost of blanking is the loss of tracking energy. If the target signal-to-cross-rate is -30 dB, for example, then all the cross rate coming from stations stronger than -30 dB relative to the tracked station must be blanked. This easily adds up to a loss of 90% of the pulses that can be used for tracking, which translates into a 10 dB reduction in tracking energy. This signal loss can be acceptable for some applications, such as for static monitoring applications and timing receivers. In that instance, blanking is probably the most optimal cross-rate mitigation technique.

According to reference [35], the integrity requirements of Non-Precision Approach (NPA) and the accuracy requirements of the Harbor Entrance and Approach (HEA) do not allow for such a significant drop in signal energy associated with blanking; therefore they require alternate mitigation techniques.

Next to signal loss, another drawback of blanking is its inherent inability to mitigate cross-rate for data demodulation.

3.6.3.3 Beam steering

Provided a dual loop H-field antenna, it is possible to electronically null the strongest cross-rating signal. This can done on a per-pulse basis: for each pulse, it is determined which...
transmitter causes the most severe cross rate for the tracked station, and the H-field antenna radiation pattern is electronically steered such that exactly that interfering signal is in the null. The influence of beam steering on the tracked signal needs to be corrected to achieve constant amplitude. Of course, nulling only removes a single cross-rating signal at a time, and it is only able to do that when the heading towards that cross-rating station does not coincide too closely with the heading (modulo 180°) of the tracked station. Although often more than one cross-rating signal is present at the same instant, there still can be a significant cross-rate residual present. When we apply beam steering, we also assume that the phase response will be heading independent. Antenna errors and local propagation phenomena may invalidate this assumption.

Beam steering is probably the simplest (and most processing-power efficient) cross-rate mitigation technique that also improves the data-demodulation performance of a dual-loop H-field receiver. More information on the possible improvement of data demodulation by using beam steering can be found in reference [49].

3.6.3.4 Frequency domain canceling

Each GRI has its own distinct frequency pattern with spectral lines spaced at a ½ GRI distance. Figure 3-47 shows an example of these spectral lines for both the 7499 and the 6731 GRI. The different spectral lines make cross-rate filtering in the frequency domain possible.

![Figure 3-47: The difference in spectral lines between (for example) the 7499 and 6731 GRI allow for frequency domain cross-rate canceling](image)

Modern digital signal processing allows real-time digital filters with notches at all of the cross-rate harmonics. Figure 3-48 depicts an implementation by Peterson [36] to cancel the 9960 and 7980 GRI to track the 8970 GRI.

![Figure 3-48: Frequency domain cross-rate filtering as implemented by Peterson (Illustration reproduced from reference [36])](image)
The notches must be very narrow, for example, with -3 dB bandwidths of 20 mHz to be sufficiently effective. To accomplish a notch depth of 20 dB, a very stable oscillator with an absolute frequency accuracy better than 1e-8 is necessary. This accuracy can be achieved relatively easily by locking the oscillator to the carrier of one or more received Loran-C transmitters. However, in a dynamic application, the differential Dopplers towards different stations from the same GRI can easily exceed 1e-8 (approximately 10 km/h). This effectively limits the application to fixed monitoring or timing receivers. Additionally, note that any modulation of the Loran-C signals, either by data-transmissions such as Eurofix or by blanking of a dual-rated transmitter, for example, will also modulate the frequency behavior of that station making the notches less effectively.

Although the frequency domain cross-rate filtering is an elegant proof-of-concept, the shortcomings described make it less suitable for most practical applications.

3.6.3.5 Time-domain canceling

Cross-rate interference is largely deterministic. When the cross-rating station is being tracked, a footprint of the cross-rate signal shape can be made available. This cross-rate signal can then be subtracted from the incoming samples, theoretically canceling the cross-rate effects completely [38] [39]. Figure 3-49 shows the effect of cross-rate canceling on the envelope of Værlandet. Even after short integration the cross-rate energy levels are well below that of averaging alone, which was shown earlier in Figure 3-46.

![Figure 3-49: Time domain cross-rate cancellation by estimate-and-subtract. The envelope of Værlandet (7499Y) is plotted for various integration times, ranging from 0.2 minutes to 12.1 minutes. The right plot zooms in to just prior to the start of the pulse.](image)

Although the plots look very promising, the time-domain cross-rate cancellation shown here is not perfect, possibly not even good enough for a reliable cycle identification of the weaker stations. The major drawback of time-domain cross-rate canceling is the assumption that Loran-C is deterministic when in fact it is not, at least not completely. Everything that is not deterministic needs to be measured, and hence, is susceptible to errors.
The first unknown is obviously the unknown time of arrival of the signal. Any discrepancies between the expected and the actual TOA of a pulse will lead to a misalignment of the reference figure with respect to the incoming signal, and hence, decrease the cross-rate cancellation efficiency when subtracting one from the other. Figure 3-50 shows that the reference figure needs to be aligned with the incoming pulse within 150 ns for a cross-rate cancellation of only 20 dB. If the pulses are Eurofix modulated (tri-state Pulse-Position Modulation with a 1 µs modulation index) then an erroneous decoding, and hence, re-modulation will lead to a maximum obtainable cross-rate cancellation of only 4 dB in case (for example) a cross-rating prompt pulse is mistaken for a (1 µs) delayed pulse.

Other challenges for cross-rate cancellation are, for example, the influence of the rotation of the H-field antenna, as well as blinking and blanking of the signal. Changing propagation effects, for example, when driving through a city, also influence the effectiveness of time-domain cross-rate cancellation. Finally, some integration of the reference figure is needed to reduce the noise. During this integration, vehicle dynamics such as displacement and rotation must be compensated for, which might also introduce some errors.

The transmitter itself is also not completely beyond suspicion. Figure 3-51 shows a received GRI of Sylt, 7499M. After Eurofix re-modulation, the same reference figure is subtracted from all 9 pulses. From the right plot, it can clearly be seen that the first 4 pulses cancel completely; whereas, the cancellation of the last 5 gets increasingly worse.

This effect can be accounted for by the limited transmitter stability. On a pulse-by-pulse basis, the transmitted pulses can easily jitter 50 ns in time, especially the pulses later in the GRI. Because this effect is somewhat deterministic, it can be compensated for in the transmitter by firing the pulse circuit a bit earlier or later. Compensating the added timing errors caused by
dual-rating a station, however, is more difficult due to the more complex timing relations, and is not implemented at the time of this writing. This leaves some residual phase-jitter, which is a function of the momentary timing relation between the two GRIs of the dual-rated transmitter.

![Diagram showing the effect of transmitter stability on Loran canceling performance.](image)

**Figure 3-51: Effect of transmitter stability on Loran canceling performance**

The amplitude stability of the transmitted pulses has never been part of any signal specification, but according to Megapulse, the manufacturer of most Loran-C transmitters, can be in the order of 2%.

For example, a 50 ns phase- or a 2% amplitude-fluctuation will both limit the theoretical maximum cross-rate cancellation to about 30 dB.

Due to the described limitations: the limited transmitter stability, unknown TOA, antenna orientation, data modulation, and noise, the cross-rate cancellation is unlikely to exceed 30 dB. In practice, cross-rate reduction by cancellation of 15 to 30 dB should be achievable, depending on local circumstances and the receiver implementation.

### 3.6.3.6 Combination of cross-rate mitigation techniques

The easiest way of removing cross-rate is by blanking all the cross-rated pulses. However, the loss in signal energy might be too severe for some applications. Time-domain cross-rate canceling has the potential of cross-rate cancellation without costing any tracking energy, but it seems very challenging to get sufficient cross-rate reduction of the stronger stations by canceling alone. A likely desired approach would be the combination of canceling of the weaker and blanking of the stronger cross-rate. The distinction between weak and strong cross-rate, and hence, between canceling and blanking, is a trade-off between the desired cross-rate reduction versus the required minimum tracking energy.
3.6.4 Local interference

The Loran-C link-budget is designed with the relatively high atmospheric noise levels present in the LF band in mind. Therefore, for an interference source to be harmful, it needs to either radiate a significant amount of energy or it needs to be very close to the receiving antenna. If the latter is the case, it is denoted as local interference. As the interference source is very nearby, usually well within hundred meters, the receiving antenna is in the near field of the interferer. Therefore, there can be a significant difference between the H-field and the E-field component of the interference.

3.6.4.1 Examples of local interference

Several examples of local interference are as follows:

Switching power regulators

Most modern electronic equipment uses switching power supplies to convert main power to low voltage DC or to convert (for example) 12V DC car power to another voltage. These switching converters usually use a poorly shielded magnetic coil or transformer. Better shielding is often not attractive from a cost perspective and is only done if needed to meet regulations. Because the radiated field from these devices originates from a fluctuating current through a (poorly shielded) coil, the interference is almost exclusively harmful for receivers equipped with an H-field antenna. The radiated interference spectrum of a switching regulator depends on its implementation and can vary from a fixed frequency without modulation (which can, therefore, easily be notched out) to pulse-width or frequency modulated depending on the (fluctuating) current drawn by the sourced equipment. Depending on the frequency used, switching regulators can be extremely harmful for Loran-C reception, but usually only at very short distances (up to several meters at most).

Computer equipment

Modern computers can draw significant power during very short time intervals. For example, an interrupt routine is called every 10 µs, which causes some data from DRAM to be copied. The memory operation causes a very short but significant power drain that radiates some EM-field. Although it sounds unlikely, this kind of process can cause serious trouble in the LF band. Although legislation should prohibit this harmful radiation, experience with several car-navigation computers has proven the opposite. The careful choice of components, PCB design, and shielding are effective means to mitigate this problem by the source.

CRT-monitors and televisions

The electron-beam in a Cathode Ray Tube (CRT) monitor is deflected using powerful magnetic fields. If the horizontal sync of a computer monitor is set to 96 kHz, for example, this monitor becomes a powerful short-range Loran-C jammer. Fortunately, if the front-end of the receiver is not forced outside its linear operation, notching of the interference is often possible.

PAL and NTSC televisions use a line-frequency of 15.734 kHz. The 6th harmonic of this signal, at 94.404 kHz, falls inside the Loran-C band. Some, mostly older, TVs can produce significant
energy at this 6th harmonic, which becomes apparent when performing H-field measurements in close proximity of an apartment complex, for example.

**Power lines**

Many power lines carry data modulation for various applications. This can range from remote meter readout to power grid status. Usually frequencies in the LF-band (30-300 kHz) are used for these low-data, rate-signaling applications. Although fiber-optic cable is now usually the primary means of grid-status communication, still many high-power connections are modulated with some sort of signaling data.

At the time of this writing, the possibility of using the power grid for wide-band Internet services is being extensively researched. The so-called “Power Line Communication” (PLC), “Broadband over Power Lines” (BPL), or “Power Line Telecommunications” (PLT) uses the last stretch of wire from the final transformer to the outlet inside the home to provide an alternative to phone or cable-TV wires. Primarily frequencies in the HF-band are used, above our frequency band of interest [48]. However, the developments in this area must be monitored closely to ensure that the LF radionavigation band remains spared both by the communication signals directly and by their (sub)harmonics.

Other noise sources related to power lines are the distribution transformers that transform from high-voltage to the 110V or 240V required for households. Often, these transformers are attached to the poles carrying the power lines. Newer transformer types often are built for an estimated lifetime of only 10 years. This involves cheaper components, worse shielding, and hence, a significant increase in LF interference. The interference is often so severe that it will significantly distort or even inhibit all LF radio-navigation, both E-field and H-field, in proximity of the transformer.

**Power wires**

Most power wires are very poorly symmetrical. They lack shielding, are not twisted, and sometimes even, (for example) in cars, an alternate path is used for the return power. This combined with a fluctuating power usage makes some power wiring effective H-field radiators. For example, the rear-window heating wires have proven to be excellent in radiating all the modulation present on the car’s power system. This makes, for example, the hood of the trunk of a car a sub-optimal location for the H-field antenna at times that the rear-window heating is being used.

**Overhead lines**

The power transfer from an overhead line to a train or trolley via a wiper frequently generates some sparking and corona effects. During wintertime, this phenomenon often coincides with a clearly visible blue flare. The emanating power fluctuations can be transported via the overhead lines and radiated at some distance. Therefore, it is likely to already receive the wideband noise well before and after the train actually passes.

Next to corona effects, also the switching power regulators of modern trains and trolleys can cause current fluctuations in the overhead lines resulting in a significant increase in interference.
Traffic detection loops
Traffic detection loops are used very frequently for traffic management. In the more metropolitan areas of the Netherlands, there is probably not a single kilometer of road without a traffic detection loop. The loops radiate an LF H-field and detect cars by the change in impedance of the loop.

There does not seem to be a single standard for traffic detection loops. In Delft alone, several types have been detected; some transmit continuously, some are pulsed, some use frequencies around 40 kHz, and others use frequencies around 100 kHz. Figure 3-52 shows two specific types of pulsed detection loops. This type of interferers can be mitigated fairly simply by throwing away all the Loran-C pulses during the interference bursts. In this particular case, this noise censoring would result in a 1.7 respectively 0.4 dB signal loss. Applying notches might be an option for continuously transmitting detection loops, but only if the front-end of the receiver remains linear (is not clipping), which is far from trivial under these circumstances. Interference from traffic detection loops can be especially bothersome while waiting for a traffic light. Reception of the LF radionavigation signals may very well be impossible during these circumstances. The long coasting time then required, up to a couple of minutes, lays a heavy burden on the receiver’s, often cheap, crystal oscillator.

![Figure 3-52: Two types of pulsed traffic detection loop interference: 35 respectively 2.2 ms long pulses with a period of 107 respectively 24 ms](image)

Engines, generators, and alternators
Systems such as engine ignition, fuel pumps, generators, alternators, and motor management computers can potentially generate LF radio noise. The severity of the noise is very dependent on the specific implementation of the vehicle, as well as upon the engine’s revolutions per minute. It is the author’s experience that some cars are completely quiet; whereas, others impose insuperable difficulties for LF radio-navigation. Noise related to sparking tends to
influence both E-field and H-field reception; whereas, noise related to electrical systems such as the generator is primarily present as an H-field interference source.

![Diagram of engine noise](image)

**Figure 3-53: Example of engine noise: every 4.8 ms a 160 µs noise spike**

### Precipitation static

Precipitation static or P-static is the term used to describe electrical noise, which can be generated either by electrically charged rain falling on an antenna or, in an aviation context, by the transport of electrical charge from the airframe of a plane to the surrounding atmosphere. The airframe can become electrically charged during flight; the equalization of this charge can cause arcs between airframe segments, streamer discharges across dielectric surfaces, and corona discharges from airframe elements with relative small diameters. The installation of discharge devices and maintenance measures can drastically reduce the P-static noise. However, even after these countermeasures the P-static effects remain of great concern for the aviation community. [47] Fortunately, the H-field antenna is almost non-susceptible to P-static effects and is thereby the antenna of choice for aviation usage.

P-static during land and sea modal use of LF radionavigation is usually caused by charged raindrops falling on the E-field antenna. Increase of the size of the antenna dome can reduce the P-static noise in these situations somewhat. Total disruption of the E-field reception of LF radio waves is still possible during severe rainstorms. Here again, use of an H-field antenna forms the solution.

### 3.6.4.2 Mitigation methods

#### Mitigation at the source

Often, a series inductor, a parallel capacitor, or some (extra) shielding reduces the emitted harmful interference significantly, possibly even to a negligible level. The cost of these alterations might be acceptable if the changes are carried out during initial assembly. For example, most airplanes are designed such that no harmful interference is produced in the LF and HF band that might endanger the correct functioning of an Automatic Direction Finder (ADF) system. Every component must comply with strict regulations resulting in an environment suitable for the LF/HF ADF system, and therefore, most likely for interference free LF Loran-C reception as well. However, in the automotive industry, every penny that can be saved in production has become crucial in this highly competitive market. Since the declining use of long-wave broadcast radio in cars, there is hardly any need to strive for low EM radiation in the LF-band. This probably makes mitigation of LF interference one of the last points on the priority list of most car manufacturers. Post-assembly altering of the vehicle is
often very expensive; therefore, only an option for a limited range of applications, for example, when safety-of-live is involved. Large-scale use of LF radionavigation implies the need for the receiver itself to be able to deal with the local interference.

**E-field versus H-field**

Some local interference is more observable in the H-field, especially when the source involves a fluctuating current through a wire or coil. Other interference sources, such as local discharges, cause more trouble in the E-field. These differences may motivate the choice for an E-field or H-field antenna. However, other differences between an E-field and H-field receiver also require consideration:

- An E-field antenna needs grounding
- The low profile H-field antenna can be co-located with a GPS antenna.
- The E-field antenna can be combined with such devices as a car-radio antenna
- The P-static sensitivity of an E-field antenna makes it an undesirable option for aviation and questionable for maritime applications.
- H-field receivers are more complex because they must process two channels
- There is a difference between H-field and E-field propagation when in close proximity to large objects. This can result in a significant difference in positioning performance between a receiver equipped with an H-field antenna and a receiver using an E-field antenna.

It is interesting to note that the maritime H-field DGPS radio-beacon receivers are recognized as more robust than their E-field counterparts; whereas, the installation of an H-field Loran-C receiver on a boat often requires more attention than that of an E-field Loran-C receiver. This contradiction can probably be explained by the fact that Loran-C uses a larger bandwidth than DGPS radio-beacon receiver; therefore, it is susceptible to interference in a broader frequency band.

**Time domain**

Interference such as engine ignition noise, (local) lightning flashes, and some types of traffic detection loops have a pulsed character with a relatively low duty cycle. This allows them to be punched out by the receiver without losing too much tracking energy. The challenge for the receiver is to distinguish the harmful pulsed noise from the desired pulsed Loran-C signals.

**Coding domain**

Most interference sources are not coherent with the used Loran-C Group Repetition Intervals (GRIs). Integration over the GRI interval, therefore, reduces the influence of the interference source. However, long integration times limit the responsiveness of the receiver.

**Frequency domain**

If the interference is either located outside the Loran-C frequency band (90-110 kHz) or if it is narrow band, then frequency domain filtering in the form of bandpass filtering and notches can mitigate the effect of the interference. Care should be taken with in-band notch filtering, because this affects the Loran-C waveform, and hence, its phase (timing) and envelope (cycle-identification) characteristics.
**Spatial domain**

The LF wavelength is too long to apply phased-array antenna techniques successfully. LF spatial domain interference cancellation, however, is possible by either using the figure-eight radiation pattern of the H-field loop antenna or by using the near-field distance versus field strength relations.

**Directional nulling with H-field antenna**

A single H-field antenna loop has a figure-eight radiation pattern of which the null can be electronically pointed towards a dominant interference source. Only a single direction can be nulled per station, and obviously all the radionavigation signals from that direction are nulled as well.

Another approach would be to use two single loop H-field antennas at some distance from each other, as shown in Figure 3-54. The distance between the antennas, for example, one meter, is such that we can assume that the phase of the radio-navigation signal as received by both antennas can be considered equal.

![Figure 3-54: Two single loop H-field antennas are used to directionally null an interference source while combined maintaining the ability of an omni-directional radiation pattern](image)

If the angle $\alpha \neq 0$ it is possible to electronically combine both antennas such that a virtual orthogonal dual-loop pair is obtained, after which normal processing can take place, but with the dominant interference source removed.

Let the radiation pattern of a single loop be defined by $\cos(\varphi)$ and both single loops summed after multiplication with $A$ and $B$, then the resulting radiation pattern is given by the following:

$$A \cdot \cos(\varphi + \alpha/2) + B \cdot \cos(\varphi - \alpha/2)$$

$$A \cdot (\cos(\varphi)\cos(\alpha/2) - \sin(\varphi)\sin(\alpha/2)) + B \cdot (\cos(\varphi)\cos(\alpha/2) + \sin(\varphi)\sin(\alpha/2)) \quad (3-26)$$
It is now a trivial step to find the required $A$ and $B$ to form two virtual orthogonal loops

$$
(A + B) \cdot \cos(\phi) \cos(\alpha / 2) + (B - A) \cdot \sin(\phi) \sin(\alpha / 2) = \cos(\phi)
$$

$$
B - A = 0, A + B = 1/\cos(\alpha / 2)
$$

$$
A = B = \frac{1}{2 \cdot \cos(\alpha / 2)}
$$

(3-27)

And for the other virtual loop:

$$
(A + B) \cdot \cos(\phi) \cos(\alpha / 2) + (B - A) \cdot \sin(\phi) \sin(\alpha / 2) = \sin(\phi)
$$

$$
A + B = 0, B - A = 1/\sin(\alpha / 2)
$$

$$
A = -B = -\frac{1}{2 \cdot \sin(\alpha / 2)}
$$

(3-28)

If, for example, two single loop antennas are placed in the rear-bumper of a car, 1.5 meters apart and 4 meters from the engine, then $\alpha = 21^\circ$ requiring $A=B=0.51$ for one virtual loop and $A=-B=-2.71$ for the other loop. Note that in this example the effective noise contribution of the virtual orthogonal antenna will be significantly larger (+11.7 dB!) in the direction of the interference source. The ideal setup would be if $\alpha = 90^\circ$, which results in the removal of the interference source without effectively increasing the antenna noise.

In this example, two fixed, single loop antennas are used. A more versatile solution would be to use two dual loop antennas (and hence a four-channel receiver front-end) and to adaptively combine the two to achieve optimal interference rejection. Because the desired navigation signals remain unaffected, it is also possible to do the actual combining of the two antennas in the frequency domain, thereby allowing multiple directional nulls for multiple interference sources at different frequencies.

**Distance nulling in near field**

Only a small change in distance (e.g. 50 cm) towards a local interference source (e.g. at 2 meter distance) gives a large change in received interference energy. Experiments show that the difference in received interference energy is often more extreme than the expected cubic relation, which is expected in the near field. This extreme relation between distance and field strengths leads to the following local interference mitigation method:

If:
- Two equal antennas $A_1$ and $A_2$ are placed in close proximity of an interference source $S$.
- The antennas are equally aligned, and they both receive virtually the same Loran-C signal $L$.
- Each of the antennas has an other distance towards the interference source $S$ (e.g. 1 meter and 1.5 meter) and, therefore, receive a different amount of interference energy.
- If the two antennas are subtracted, the difference only consists of the local interference as the Loran-C portion completely cancels.
Then the local interference $S$ received on antenna $A_1$ can now be cancelled by:

\[
A_1 = L + \alpha_1 S \\
A_2 = L + \alpha_2 S \\
A_1 - A_2 = (\alpha_1 - \alpha_2) S \\
L = A_1 - \beta (A_1 - A_2), \quad \beta = \frac{\alpha_1}{\alpha_1 - \alpha_2}
\]

The challenge is to find the correction parameter $\beta$. The author has written a software package called SDLIC, Spatial Domain Local Interference Cancellation, to test the stated hypothesis. Advanced Loran-C receiver software is used to track all Loran-C signals and subtract them from the input signal. The remaining residue is compared with the interference footprint, derived from the difference between the two antennas.

**Figure 3-55:** Interference nulling based on difference in distance towards two antennas

**Figure 3-56:** Experimental results of Spatial Domain Local Interference Cancellation (SDLIC) using distance nulling, left the time domain, right the frequency domain. The blue line represents the input signal, the red line the result after cancellation.
This comparison results in an adaptive estimate of the correction parameter, which is then applied to mitigate the interference. All the corrections are done in the frequency domain, thereby allowing to simultaneously canceling multiple interference sources if they have a different frequency.

The software is tested with an experiment using a fluorescent flashlight. The power converter in the flashlight emitted strong interference in the 85-115 kHz frequency band as shown by the blue plot in Figure 3-56-right. Conventional notching of this interference would lead to tracking unacceptable distortion of the Loran-C pulse. The SDLIC software was able to cancel the interference while leaving the Loran-C signal largely in tact.

Spatial domain interference cancellation might be a solution to mitigate local interference: one of the most serious weaknesses of Loran-C. However, a much more complex receiver is required, which consists of multiple antennas and front-ends and a challenging software architecture.

### 3.7 LF radionavigation antenna

The LF radionavigation antenna is discussed in detail in Chapter 5; therefore it is only briefly summarized in this Section.

Three types of sensor errors can be distinguished: noise, heading-independent errors, and heading-dependent errors.

#### 3.7.1 Noise

In a well-designed LF radionavigation system, the noise contribution of the sensor, also denoted as internal noise, should be similar to that of the effective external noise. If the sensor noise is dominant, then a better sensor should have been used or processing power could have been spared. If the external noise is dominant, the sensor could have been noisier, which usually means smaller in size. For example, various interference sources make a car a relative noisy environment. This makes the use of an expensive, high-end antenna excessive. In this case, the usage of a smaller, low profile antenna will usually not result in a degradation of the overall performance. If no significant interference is present, the average atmospheric noise level is usually used as a target for the noise contribution for the sensor.

The sensor noise is directly related to the dimensions of the sensor and to the quality of the associated amplifier. An H-field antenna usually requires a more careful design to achieve similar performance compared to an E-field antenna.

#### 3.7.2 Heading-dependent errors

A single H-field loop has a figure-8 radiation pattern. Two of these loops are needed to receive all transmitters with equal sensitivity. A heading dependent error can occur if:

- there is a difference in resonance frequency of the loops,
the phase response of the filters and amplifiers associated with both loops are not equal,
there is a parasitic coupling between both loops, or
the loops have parasitic E-field susceptibility.

The resulting pseudorange error depends on the heading towards the transmitter and can be as large as 200 ns, which will result in a significant positioning and timing error. Fortunately, it is possible to calibrate for all of the heading dependent errors except the E-field susceptibility. If the E-field susceptibility is sufficiently low, for example, by careful shielding and/or balancing, then calibration of the H-field antenna can reduce the heading dependent errors to a negligible level. The omni-directional E-field antenna does not have any heading-dependent errors.

3.7.3 Heading-independent errors
All the delays of an E-field antenna or those delays that are common for both H-field antenna loops give a heading independent time-of-arrival error. Examples of these errors are:
- analog and digital filter delays (as long as they are common),
- H-field common tuning error,
- cable delays, and
- phase of the E-field ground plane.

These errors cancel out as part of the receiver clock error in the position calculation. Timing users, however, are affected by these errors. Compensation is possible by either calibration or by measuring the feedback response of an injected calibration signal.

3.8 Summary and discussion
In this Chapter, various error sources have been reviewed together with possible mitigation methods and their net impact on the navigation performance:

Coverage and geometry
A 2-dimensional radionavigation system needs at least three transmitters to derive a position and time. The distance between the user and the transmitter has a direct relationship with the received signal strength, and hence, the quality of the individual range measurements. Additionally, the geometry of the transmitters with respect to the user is of importance, because this determines the relation between the quality of the individual pseudoranges and that of the final position solution. The coverage of the Loran-C system, for example, has a crucial impact on its performance although it is not a fundamental issue. More transmitters can be added, although it is often a difficult political and economical struggle.

Transmitter timing errors
Several methods of transmitter timing are currently employed. Europe uses Time-Of-Transmission (TOT) control where two-way Loran-C measurements are used for the time transfers. The USA is migrating from SAM control to TOT, where eventually GPS and two-way satellite time transfer will be used to distribute absolute time. Transmitter timing errors are
not a fundamental problem and can be solved relatively easily by using modern technology. The possible use of the Loran-C system in the future must make the business case for the required investments.

Re-radiation

Objects of significant size close to the receiver can cause re-radiation. Re-radiation can be detected, but not easily corrected.

Noise and interference

Various interference sources and mitigation methods have been identified: Atmospheric noise is caused by thunderstorms all over the world. Significant mitigation can be achieved by non-linear processing.

The deterministic character of Cross-Rate Interference makes this problem largely solvable, but at the cost of a significant increase in receiver complexity.

Continuous Wave Interference is to be battled by bandpass filtering and notch filtering.

Several special techniques have been presented to suppress local interference. This type of interference, however, can be quite bothersome in some situations.

LF radionavigation antenna

The design of a good LF antenna is possible. (Auto) calibration can make a good H-field antenna virtually perfect. The noise contribution of the antenna can be made lower than the external noise, and hence, of little concern.

LF propagation

Loran-C shows a stunning difference between a meter-level repeatable accuracy and an absolute accuracy that can easily have an error of hundreds of meters. The origin lies in the low-frequency ground-wave propagation. The ASF factor is defined as the difference between the actual propagation time and the propagation time if the path were all seawater. This ASF factor can be modeled from the topography and conductivity information of the propagation path. Limited accuracy and detail of the topography and the conductivity information, as well as the simplified model used causes discrepancies between the modeled and the measured values. The model error can easily lead to a pseudorange error of hundred meters or more, depending on the complexity of the propagation path, the quality of the databases, and the level of sophistication of the propagation model used.

Most error sources can be effectively mitigated. However, the errors introduced by unknown propagation effects are bothersome. The next Chapter describes methods to mitigate these errors by:

– Supplying the user with measured propagation delays, which are the result of a surveying campaign in those areas where high accuracy positioning is required
– Real-time differential corrections to resolve the temporal variations in propagation and transmitter timing

Applying these techniques makes it possible to achieve an absolute positioning performance close to the short-term repeatability.
3.9 References

[6] “Specifications of the Transmitted Loran-C Signal”, United States Coast Guard, Department of Transportation, COMDTINST M16562.4, 1994
[17] “Extending the range of Loran-C ASF modelling”, Paul Williams and David Last, International Loran Association, Tokyo, Japan, 26 October 2004
NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY

[26] “Medium frequency propagation: a survey”, P. Knight, Research Department, Engineering Division, British Broadcasting Corporation, May 1983
[33] “Carrier wave signals interfering with Loran-C”, Martin Beckmann, Ph.D. thesis Delft University of Technology, the Netherlands, 1992
LORAN-C ERROR MODEL – REFERENCES

[40] “Technical characteristics of methods of data transmission and interference protection for radionavigation services in the frequency bands between 70 and 130 kHz”, Recommendation ITU-R 589.3, International Telecommunication Union, August 2001


[54] “Loran-C interference Study”, B. Peterson and J. Hartnett, WGA Technical Symposium, October 1987


[57] “Vulnerability Assessment of the Transportation Infrastructure Relying on the Global Positioning System”, John A. Volpe National Transportation System Center, August 20, 2001

NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY
4 dLoran

GPS, especially in combination with augmentation systems such as SBAS or marine radio beacons, has made precise positioning ubiquitous and affordable. Consequently, this has raised the bar for alternative and/or backup navigation systems. This Chapter uses the Harbor Entrance and Approach (HEA) procedure as an example, and it discusses the feasibility of Loran-C to be used as a (secondary) navigation aid for this application.
4.1 Introduction

Table 4-1 depicts the requirements for HEA [1]

<table>
<thead>
<tr>
<th>Performance requirements</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Accuracy (back-up)</td>
<td>20 m, 95%</td>
</tr>
<tr>
<td>Monitor /Alert Limit (back-up)</td>
<td>50 m, 95%</td>
</tr>
<tr>
<td>Integrity (target)</td>
<td>$3 \cdot 10^{-5}$ / hour</td>
</tr>
<tr>
<td>Time-to-alert</td>
<td>10 sec</td>
</tr>
<tr>
<td>Availability (minimum)</td>
<td>99.7%</td>
</tr>
<tr>
<td>Continuity (minimum)</td>
<td>99.85% over 3 hours</td>
</tr>
</tbody>
</table>

The most challenging of the HEA specifications is the required absolute positioning accuracy of 20 meters, 95%. In Section 3.8, it is concluded that the unknown characteristics of the ground-wave propagation, represented by the ASF parameter, are the most dominant component in the total error budget. Both the spatial and temporal variations of the ASFs need to be mitigated to achieve the challenging accuracy demand. Figure 4-1 shows the suggested approach to achieve this goal:

A local reference station measures the (mostly unpredictable) temporal variations of the ground-wave propagation and transmits these measurements to the user in real-time. The spatial ASF variations are mitigated by an ASF correction map, which will be created by survey campaigns and made available to maritime users. In practice, the user equipment will first correct its measurements by using the real-time differential corrections and then apply a position dependent spatial ASF correction resulting in an accurate Loran-C positioning solution.

Figure 4-1: dLoran operation: the temporal fluctuations in transmitter timing and propagation are measured by a local reference station; the spatial component of the ASF is compensated by an ASF correction map.
The described differential Loran-C approach is not new for maritime applications. Extensive tests of the USCG in the 1980s have shown promising results. However, those experiments were troubled with limited position and time references and with computer technology still in its infancy. Current technology such as differential GPS positioning, affordable atomic clocks, and seemingly unlimited computing power allow us to repeat these efforts with much more precision, resulting in an assessment founded upon the capabilities of an upgraded Loran-C system in a maritime environment.

This Chapter describes the various error sources associated with the dLoran approach, as well as the various fundamental, practical, and procedural issues involved with the employment of dLoran for Harbor Entrance and Approach.

Chapter 6 describes various actual performance measurements of the dLoran system and provides an analysis of its suitability to meet the HEA requirements.

### 4.2 dLoran error model

The employment of a local reference station changes the error model as is presented in Section 3.2. Furthermore, in this Chapter, the following is assumed regarding the reference station and transmitter timing control:

**Transmitter**

The transmitter uses TOT (Time Of Transmission) control. TOT control is not a fundamental requirement for dLoran, but it makes more sense, because the historical role of the SAM site (System Area Monitor), namely stabilizing the Loran-C Time Differences in a particular harbor area, is now assumed by the reference stations. Many Loran-C transmitters still use LPAs (Local Phase Adjustments) and other timing circuitry, which results in discrete timing steps of 10 ns or 20 ns. These steps force a relatively short integration time and high update rate of the reference station, which is undesirable and impractical. A better transmitter timing control is possible with upgraded Time and Frequency Equipment (TFE) such as operational in all the US Loran-C stations since 2005. This equipment guarantees smooth timing control by applying temporary frequency adjustments instead of timing steps, which results in the removal of the high-frequency components of the transmitter timing error. The remaining low-frequency timing error can be effectively mitigated by a reference station using a long integration time. In this Chapter, it is assumed that the transmitters use TOT with this smooth timing control.

**Reference Station**

First, the reference station is assumed to be virtually free of any re-radiation effects. Secondly, the antenna errors need to be mitigated by means of calibration. Finally, good interference mitigation techniques — against CRI (Cross Rate Interference), CWI (Continuous Wave Interference), and the impulse part of atmospheric noise — combined with long integration should diminish the noise to a negligible level.

Given these assumptions, the dLoran error model is depicted by Figure 4-2:
transmitter timing
(assumed TOT control)

reference station ASF

re-radiation
(mitigated by site-
selection)

noise / interference
(mitigated by filtering + long integration)

antenna errors
(mitigated by design + calibration)

reference station TOA

UTC

transmitter timing
(assumed TOT control)

ASF
(measured by reference station)

differential ASF
(corrected using ASF-map)

re-radiation
(can be detected using H-field antennas)

external noise
(mitigated using multi-domain filtering)

antenna noise

H-field antenna heading-dependent error
(mitigated by design + calibration)

user TOA

Figure 4-2: dLoran error model
The likely dominant error source is now the difference between the ASF measured at the reference station and the ASF measured at the user-receiver, which can be described as (after [2] but with addition of the spatial decorrelation of the temporal correction):

\[
\text{ASF}_{\text{tot}}^{\text{TXi}} (x_{\text{user}}, x_0, t) = \text{ASF}_{\text{average}}^{\text{TXi}} (x_0) + \text{ASF}_{\text{temporal}}^{\text{TXi}} (x_0, t) + \text{ASF}_{\text{spatial}}^{\text{TXi}} (x_{\text{user}}, x_0) + \text{ASF}_{\text{temporal}}^{\text{dtemporal}} (x_{\text{user}}, x_0, t)
\]  

(4-1)

Where \( x_0 \) is the position of the reference station, \( x_{\text{user}} \) the position of the user-receiver and \( t \) the time.

The first two terms, \( \text{ASF}_{\text{average}}^{\text{TXi}} (x_0) + \text{ASF}_{\text{temporal}}^{\text{TXi}} (x_0, t) \) depict the average ASF and the temporal fluctuations of the ASF at the reference station where they are measured and transmitted in real-time to the user via a data-link.

The third term, \( \text{ASF}_{\text{spatial}}^{\text{TXi}} (x_{\text{user}}, x_0) \), is the spatial variation of the ASF between the reference station and the user. This term is temporal stable by definition, which allows it to be measured during a survey campaign and provided to the user as a spatial ASF correction map.

The last term, \( \text{ASF}_{\text{temporal}}^{dtemporal} (x_{\text{user}}, x_0, t) \), consists of the spatial decorrelation of the temporal correction. It determines the effective range limit of the temporal correction from the reference station. The analysis and implications of this term form a considerable challenge, which will be discussed later in more detail.

In summary, after successful appliance of differential and spatial ASF corrections, the (range) accuracy of the dLoran system is limited by:

- Transmitter timing noise with a relative high frequency
- Reference station errors such as noise, re-radiation and the limited update rate
- Measurement errors and the limited spatial resolution of the ASF map
- Spatial decorrelation of the temporal ASF correction
- Re-radiation effects at the user
- Noise and interference at the user
- (Uncorrected) antenna errors of the user receiver
- Tracking errors of the user receiver

Note that a pseudorange\(^1\) error cannot be directly translated into a positioning error without knowledge of the correlation between the various errors. The common-mode (correlated) part of the TOA-errors cancels as a clock error; whereas, the differential mode (uncorrelated) errors give rise to a position error. This Chapter, however, focuses solely on minimizing of the pseudorange error, not on the correlation between those errors.

\(^1\) The pseudorange is calculated from the TOA by applying the PF, SF and ASF and correcting for the error in Time Of Transmission of the transmitter.
4.3 Reference station

The reference station consists of one or more high-end Loran-C receivers, controlled by a highly stable timing source, at a known (surveyed) location. The purpose of the reference station is to measure the average ASF at the reference site, \( ASF_{\text{average}}(x_0) \), and the temporal fluctuations of this ASF: \( ASF_{\text{temporal}}(x_0,t) \). The measured ASFs, or at least the temporal components, need to be transmitted to the user in real-time.

4.3.1 Absolute time

The reference station needs to have the same time-reference as the transmitter to measure absolute ASFs. However, the absolute ASF measured at the reference station is of no importance for the differential Loran-C positioning application\(^2\). The reference station effectively sets the combined transmitter timing error plus the ASF of all the transmitters individually to zero with respect to an arbitrary timing reference. This arbitrary timing reference is common for all the transmitters; therefore, it cancels as a common clock error in the positioning solution of the dLoran user. However, using an absolute time, for example UTC-USNO, for the reference station is still useful for the following reasons:

Absolute corrections from a dLoran reference station can be used by timing users to calibrate for most of the (unknown) propagation delay. Because most timing users use UTC-GPS (UTC-USNO) as a reference, it is most convenient to use UTC-USNO for the dLoran reference stations as well. It is tempting to use a GPS timing receiver as the timing source, but that will affect the credibility of Loran as a GPS-independent timing backup. A more robust, although significantly more expensive, solution would be the use of two-way time transfers.

For those harbors where there is a severe spatial decorrelation of the temporal corrections, it might be necessary to use multiple reference stations. In some cases, it might even be necessary to use different reference stations to correct for the propagation variations of the signals from different transmitters. Only the common part of the time references of the reference stations will then cancel in the positioning solution. It is, therefore, crucial to use the same time reference for all reference stations in a particular area. This can be achieved by synchronizing the reference stations to UTC-GPS by using GPS timing receivers. The implicit dependency upon GPS, however, may make this solution less favorable. Another solution would be to implement two-way time transfers between the reference stations or between the reference stations and another timing source.

\(^2\) Pseudorange level integration (deep integration) of differential Loran-C and for example GPS imposes additional requirements on the timing of the reference station. If the GPS system and the dLoran reference station are timed using the exact same time-standard (UTC-USNO), if the receiver uses the same oscillator for both systems, and if the receiver delays for both systems are accurately known, then only one receiver clock errors needs to be solved. This effectively provides an additional range measurement. The receiver has to solve time for both systems independently if these stringent timing demands are not fulfilled.
The spatial ASF correction maps used for HEA will be measured relative to one or more dLoran reference stations. Absolute ASF differences can be only measured if the mobile ASF measurement equipment uses the same timing reference as the dLoran reference station. Because (d)GPS will be used as a positioning reference, it makes sense to use UTC-GPS as a timing reference. Note that for the actual dLoran positioning it is not required to know the absolute ASF difference between the user and the reference station. A common offset in the ASF-map will eventually cancel in the positioning solution. However, if the ASF-map contains absolute ASFs, it is possible to observe the repeatability of multiple ASF surveys. This is useful to signal ASF measurements errors and to investigate the effect of seasonal variations on the ASF-map, for example.

In summary, absolute timing is not required for most dLoran positioning situations unless multiple reference stations are used in a single position solution. Timing users benefit most from a reference station with absolute time, preferably UTC-USNO.

4.3.2 Re-radiation detection

Re-radiation at a differential Loran reference station is comparable to multi-path at a differential GPS reference station: in both cases, the effectiveness of the differential setup is severely degraded.

Re-radiation at a Loran-C reference-station does not necessarily lead to a positioning error in the user receiver. The effect of re-radiation will cancel if the ASF map is measured with respect to the same reference station with the same re-radiation and if this re-radiation is constant in time. The latter requirement is doubtful to be guaranteed valid, which makes re-radiation at the reference site an undesired phenomenon that needs to be prevented.

It is recommended to survey a potential reference site first relative to re-radiation before permanently installing the equipment. However, even if the site is free from re-radiation at the time of installation, there is still a possibility of occasional occurrence of re-radiation (Section 6.2.2). Therefore, from an integrity point of view it is desirable to continuously verify the absence of re-radiation. According to Section 3.5.3, this can be done in two ways:

Re-radiation will most likely alter the measured direction of a signal. This change in (absolute) heading can be easily detected with a stationary setup. A more universal approach is the comparison between the measured and calculated heading differences between stations: any discrepancies between the two differences indicate re-radiation. Section 6.4.6 contains various examples of the latter re-radiation detection.

Another re-radiation technique relies on the fixed $90^\circ - 120\pi$ far-field relation between the E-field and H-field in the far field. Any deviation between this relation, which can be measured with a pair of synchronized H-field and E-field Loran-C receivers, is most likely caused by a local phenomenon (Section 6.2.2). Critical infrastructure, such as a dLoran reference setup,

---

1 It is assumed that the H-field antenna is calibrated. See Section 3.5.3.3 for the impact of antenna errors on spatial domain re-radiation detection.
usually requires an independent integrity monitoring system. This requirement can be combined nicely with re-radiation detection if, for example, the primary receiver is equipped with an H-field antenna and the secondary receiver (integrity monitor) with an E-field antenna.

4.3.3 Integration time

Both the transmitter fluctuations and the variations in ground-wave propagation will be very slow, allowing the reference station to use very long integration times of 10 minutes or more, for example. A user receiver on the other hand is limited to seconds of integration time. Therefore, the zero-mean part of the noise experienced by the reference receiver will diminish by integration and have a negligible impact of the total user error compared to the noise contribution of the user receiver. Interference, however, can (after integration) also cause a bias, which then directly degrades the user’s dLoran positioning performance. Synchronous interference, for example, will not disappear after long integration (Section 3.6.2.1).

4.3.4 Data channel

A Loran-C data channel such as Eurofix [3] or 9th pulse [4] can be used to transmit the differential data in real time from the reference station to the user.

Eurofix uses a three-state pulse position modulation (PPM) on the last six of the eight pulses of a GRI. A 7-bit word per GRI is transmitted by 128 balanced modulation patterns. A Eurofix message is 30 GRIIs long, which consists of forward error coding (140 bits), CRC (14 bits) and 56 data bits. Depending on the GRI used, this results in a data-rate of minimal 18.7 bits/sec.

Ninth-pulse modulation uses an extra data-only pulse, which is modulated using a 32-state PPM resulting in one five-bit word per GRI. Nine data and fifteen Reed Solomon parity words compromise a 24-word, 24-GRII message resulting in an effective data rate of 45 bits per 24 GRIIs or minimal 18.8 bits/sec.

Peterson [4] proposes to transmit the differential data from up to 25 reference stations via a single Loran-C transmitter with corrections for about six dual-rated stations per reference station. A single correction needs only few bits: if we bound the total ASF fluctuation to ±1 µs and the resolution to 2 ns, then it requires only 10 bits to transmit the correction for a single transmitter. After taking into account some identifier and control bits, a correction can be sent approximately every two minutes. Due to the slow fluctuation of the correction, this update rate is considered more than enough. More important is the time to first fix, which can be up to three minutes with the suggested scheme [4]. However, for maritime applications, this is probably not a problem.

---

It is assumed that the transmitter uses frequency control rather than time-steps to control its Time-Of-Transmission (see Section 4.2)
4.4 ASF-Map

The ASF-map is a database provided to the user and contains the spatial decorrelation of the ASF between the reference station and the user, or: $ASF_{spatial}^{TXI} \left(x_{user}, x_0\right)$. The use of an accurate ASF map, combined with the real-time differential corrections from a nearby reference station, will dramatically improve the absolute positioning performance of Loran-C.

4.4.1 Coverage

The ASF-map must cover at least all the official routes used for the Harbor Entrance and Approach (HEA). This will consist primarily of the channels and the actual harbor areas. Therefore, the required coverage is more 1D (line) than 2D (surface), which explains the name “Track Line Survey” that is often used for the ASF measurement efforts in harbor areas. Figure 4-3 shows such a “Track-Line” ASF map, created using the Tampa Bay measurements (Section 6.4).

![Figure 4-3: ASF measurement results of Carolina Beach placed in a 0.005 degrees grid. The ASFs are expressed in µs.](image)

4.4.1.1 Radial interpolation and extrapolation

An ASF-map such as the one presented in Figure 4-3 can be suitable for specific procedures such as a harbor entrance through a narrow channel. Any deviation from the projected path,
however, quickly results in a blank spot on the correction map. Of course, the receiver can search for the closest cell with a valid correction, but often this is not the best correction. It is the author’s opinion that is better to provide the user with a complete correction map, without blank spots. It then becomes the responsibility of the provider to make a best estimate of the ASFs between the measurements. These estimated values will most likely have a degraded accuracy, which can be provided to the user by setting a specific flag in those “estimated” cells. At the locations where stringent HEA accuracy requirements must be met, within a channel (for example), measured ASF values are strongly recommended.

There are various methods to “fill the blank spots.” Often, a significant increase in the number of measurements will be not economically viable, and even if the budget is not an issue, it will be a very time consuming effort to cover the complete area. Another method is to increase the cell size. Then, the same number of measurements will provide a larger coverage area at the cost of higher quantization errors. Interpolation seems the obvious answer to the problem, but this does not always make sense physically. Imagine that the ASF of a specific cell consists of a significant land path, while the neighboring cell has an all-sea water path: the propagation characteristics between two adjacent cells can be dramatically different. The difference between two cells on the same transmitter radial, however, is influenced primarily by the surface characteristics between those two cells. Using this knowledge, a new method of radial interpolation and radial extrapolation is introduced, which is further clarified in Figure 4-4.

![Radial interpolation and extrapolation of gridded measurement data](image)

All the green cells in Figure 4-4 contain a number of valid measurements. The red radials, all originating at the transmitter, intersect at least one measured (green) cell. The blue cells are the result of interpolation between two or more measured cells on the same radial. The yellow cells contain the extrapolated value of a measured cell. Note that extrapolation quickly causes the result to spiral out of control; therefore, the last measured value is maintained for the remainder of the radial. Finally, the gray cells have no clear relation with the measured cells;
therefore, they are filled by the value of the closest neighbor with either a measured value or an interpolated value.

This method of radial interpolation and extrapolation has been implemented on the Tampa Bay measurement results (Section 6.4) and displayed for Carolina Beach in Figure 4-5.

The left Figure contains the measured values, surrounded by a white box, as well as the interpolated values. A closer inspection shows that the method of radial interpolation provides a good approximation of the ASFs between the measured cells. Extrapolation, as shown in the right Figure, does not always lead to satisfying results, especially if, for example, the last measurement on a radial lies very close to a coastline. Therefore, radial extrapolation should be avoided, and replaced, for example, with clever planning of the measurement routes, resulting in a maximal interpolated coverage.

**4.4.1.2 Propagation modeling**

Potentially better results can be achieved when the measurements are combined with modeled values. Propagation models of Monteath and Wait (Section 3.4.3), for example, use line integrals originating at the transmitter. Therefore, the modeled values lie on the same radials as those used for the radial interpolation and radial extrapolation. A logical step is the warping of these modeled values such that they optimally fit the measurement results. In this way, the knowledge of the coastlines, topography, and conductivity is optimally merged with the facts provided by the measurement results.
4.4.2 Accuracy

The accuracy of the ASF map is limited by the quality of the measurement setup and by the spatial resolution of the correction map. The resolution of the map must be chosen such that the difference between adjacent cells is within acceptable limits. The user can use an interpolation (linear) between the grid points to further limit the error of the spatial quantization. The actual survey is normally done with a much higher resolution than what is finally used in the database, which allows calculating the influence of the ASF-map cell size on the TOA error and position error in post processing. Section 6.4.4 describes and applies a method of finding a suitable cell size for a given data set.

The following issues are crucial regarding the survey measurement setup:

4.4.2.1 Position reference

An accurate and reliable position reference was an enormous challenge for the measurement campaigns conducted in the 1980s and 90s. Specialized microwave links combined with visual aids were used to obtain a reference position with adequate accuracy. Currently, GPS provides a ubiquitous and reliable positioning reference. The slim overall error margin forces the need for meter-level GPS positioning that can be obtained by using differential corrections from a local dGPS radio beacon, from SBAS, or by using carrier-phase GPS, for example. Note that it is very important to clearly specify the coordinate frame that is used for the measurements (e.g. WGS84), because the differences between various reference frames are usually much larger than the allowable Loran-C positioning inaccuracies.

4.4.2.2 Vessel dynamics

The ASF survey measurements are usually done under dynamic conditions with velocities frequently exceeding 40 km/h or more. Therefore, any time skew between the Loran-C and the GPS measurements will directly lead to an ASF-measurement error. The GPS measurements are always very precisely time stamped as in principle also the absolute Loran-C ASF measurements, which allows for tight synchronization. The effect of acceleration is mitigated by first integrating the Loran-C signals for a short time, during which the velocity can be considered relatively constant. Next, the Loran-C ASF is calculated based on this relatively noisy measurement. These noisy ASFs, which do not contain any vessel dynamics, can be integrated much longer to reduce potential noise.

Note that the Loran-C measurements are not continuous but pulsed according to a GRI-pattern. At a 9990 GRI, new measurements arrive only every 100 ms. At a speed of 50 km/h, 100 ms corresponds to maximal 5 ns TOA change which is probably insignificant in the total ASF measurement error budget. During the same 100 ms but at a speed of 500 km/h, however (e.g. airborne ASF measurements), the TOA can change up to 50 ns (≈15 m), which likely becomes noticeable. The added “noise” due to the GRI structure of Loran-C can be prevented

---

5 SBAS: Space Based Augmentation System, such as the American WAAS, the European EGNOS or the Asian MSAS

6 The Loran-C ASF receiver has exact knowledge of UTC in order to measure absolute ASFs. From this knowledge it is possible to determine the exact timestamp of the ASF measurement.
by taking the sample moment of the individual pulses into account when calculating the effective sample moment of the integrated TOA.

4.4.2.3 Time standard

In principle, absolute time is not required for the ASF-map measurements, because the time information will eventually cancel as a clock error in the user's receiver (see also 4.3.1). Absolute time, however, allows for the measurement of absolute ASFs, which is very valuable for the analysis of the consistency between multiple measurement runs and of the seasonal variations between runs taken at different times during the year, for example. Absolute ASFs also ease the interpretation of the ASF fluctuation and the correlation of these effects with physical properties of the specific area. The actual source of absolute time is not important, as long as it is the same as that of the reference station.

4.4.2.4 Antenna errors

Minimizing of antenna errors, such as cross talk and E-field susceptibility (Chapter 5), is even more important for the ASF measurement setup than for the user receiver. Smaller errors in the ASF-map leave more margin of error for the (low cost) user receiver. Calibration adjustments should be made to compensate for the antenna delay, combined with the filter and cable delay, enabling accurate measurement of the absolute ASFs.

4.4.2.5 Receiver errors

It is essential that the receiver tracks the third zero crossing and that the influence of sky waves is negligible (Section 3.4.1). Finding the correct zero crossing should not be an issue with the help of the correct GPS position combined with absolute time. Sky-wave influences can be prevented by choosing an optimal bandpass filter. Frequency domain filtering can have significant influence on the phase of the measured signal. The effect of bandpass filtering should be common on all stations, and it can easily be calibrated. The influence of notches, however, is slightly different for specific GRIs, because it is a function of synchronicity between the notch frequency and the GRI spectral needles. Therefore, it is important to be cautious when applying notches in a monitoring or ASF-receiver.

Many dLoran proof-of-concept measurement campaigns use the same receiver technology for both mapping the ASF fluctuations as for the actual user equipment. This causes receiver errors related to receiver (algorithm) implementation to largely cancel, which yields a result that might be unrealistically positive.

4.4.2.6 Reference station errors

The ASF map is created relative to the local dLoran reference station. Any error of the reference station results in errors in the correction map. Section 4.3 gives a detailed description of the possible reference-station errors and their countermeasures. However, any reference station error cancels when it has not changed between the time the ASF-map was measured and the time it is used.
4.4.3 E-field versus H-field

In the far field, the E-field and H-field have a fixed relation resulting in a similar TOA measured by an E-field or by an H-field receiver. Therefore, both receiver types can use the same ASF correction map in the far-field. Correction for near-field effects might be possible, but is ill advised (Section 3.5.4), and consequently, it is most logical to exclude all measurements with potential near field from the ASF-map.

Near-field phenomena can arise near a rapid change in surface impedance, such as the type caused by a coastline or a structure of significant size. Fortunately, the Tampa Bay measurements have not revealed significant impact of coastlines on the relation between E-field and H-field (Section 6.4.5). Conversely, bridges and other large structures have shown noticeable influence leading to large positioning errors, but only in very close proximity of the object (Section 6.4.6).

4.4.4 Integrity

There are several ways to assess the integrity of the ASF-map measurements. A primary approach is to validate the repeatability of the measurements, which can be done by comparing the results of both rates of a dual-rated Loran transmitter, for example. This comparison is an excellent way to monitor the potential presence of noise and interference in the measurements. Note that any errors in the reference time or reference position are equal for both rates and will cancel in the difference; therefore, they will remain hidden in the integrity analysis. Another way to verify the integrity of the measurements is by comparing the results of the measurements taken at different dates. For this comparison, it is desirable when the measurements are taken relative to an excellent time base. Ideally, the difference between the various measurement runs should be noise, not a bias.

It is also important to detect any local phenomena caused by re-radiation (Section 3.5.3). These effects can be extremely localized and severe (Section 6.4.6). A more detailed survey at such a “contaminated” location is recommended to assess the extent of the problem.

The best integrity tool is probably a smart operator equipped with suitable real-time and post-processing analysis tools.

4.5 Spatial decorrelation of temporal corrections

The success of dLoran positioning relies on the level of independence between the temporal and spatial fluctuations of propagation: the temporal fluctuations are assumed equal throughout the whole area; therefore, they can be measured at a single point. Next, the spatial fluctuations can be surveyed after successful removal of the temporal component. Another outlook would be to view the ASF correction map as perfectly stable, because all potential fluctuations are assumed homogeneous over the whole coverage area and thus compensated.
for by the reference station. If this simplification of reality would be true, then the dLoran TOA error would be determined by:

- Reference station errors
- ASF-map accuracy
- Re-radiation (not incorporated in the ASF-map)
- Noise and interference
- Receiver errors

Most of these errors, except the re-radiation effects, can be reduced significantly given enough engineering effort. For example, the ASF-map accuracy can be improved by more extensive surveys; noise and interference can be overcome by a smarter receiver design, integration with inertial sensors, or by increasing transmitter power, for example.

Unfortunately, differences in propagation paths, seasonal influences, and (local) weather cause the propagation fluctuations to be less homogeneous. This results in a limited spatial validity of the temporal corrections measured by a single reference station. The (often unknown) discrepancy between the temporal variations measured at the reference station and at the user is denoted by the following:

\[
\Delta_{\text{temporal}} \equiv T_{\text{TX}_i}^\text{user} - T_{\text{TX}_i}^\text{ref} \quad (4-2)
\]

Knowledge of this spatial decorrelation of the temporal corrections is crucial for determining the following:

- How many reference stations are needed for a given area?
- What is the optimal location for the reference stations?
- Given one or more reference station, what then are the expected pseudorange and positioning errors caused by the imperfect temporal correction?

From a positioning perspective, it is crucial to separate the common mode from the differential mode temporal variations of the TOAs. The seasonal influence, for example, will cancel if it has predominantly the same effect on all the TOAs used for the positioning solution. The differential effect, however, impacts positioning accuracy. Obviously, timing users have to deal with both the common and differential variations.

There are various methods with various levels of complexity to approximate the spatial decorrelation of the temporal corrections.

The simplest model assumes uniform propagation between the user, reference station, and the transmitters. If the speed of propagation of the Loran-C signals is uniform and equals \( v \), then the propagation delay is given by the path length divided by that (constant) propagation velocity:

\[
TOA_{\text{user}} = \frac{d_{\text{user}}}{v}, \quad TOA_{\text{ref}} = \frac{d_{\text{ref}}}{v} \quad (4-3)
\]

where \( d_{\text{user}} \) is the distance between the user and the transmitter, \( d_{\text{ref}} \) the distance between the reference station and the transmitter, and \( v \) the uniform propagation velocity.
If the propagation velocity changes from time to time, but in a uniform manner, then the difference between the TOA fluctuation at the user and the fluctuations at the reference station are proportional to the difference in distance between the user and the transmitter \((d_{\text{user}})\) and distance between the reference station and the transmitter \((d_{\text{ref}})\). This simple approach is known as the double range difference model [5]:

\[
\frac{\delta \left( \text{TOA}_{\text{user}} - \text{TOA}_{\text{ref}} \right)}{\delta t} = \frac{d_{\text{user}} - d_{\text{ref}}}{\delta v/\delta t}
\]

where \(d_{\text{user}}\) is the distance between the user and the transmitter, \(d_{\text{ref}}\) the distance between the reference station and the transmitter, and \(\delta v\) the uniform change of the uniform propagation velocity.

The assumption of uniform (change in) propagation velocity results in the hypothesis that the reference station correction would remain equally valid if a user would move away from the reference station while maintaining the same distance towards the transmitter, for example. Obviously, there are numerous situations where the uniform propagation assumptions of the double range difference model are proven wrong. However, this very simple model was quite useful in explaining the majority of the variations found in the St Mary's River Loran-C Mini-Chain Stability Studies [7], for example.

Figure 4-6 shows a more challenging situation. The propagation path from the transmitter to the user quickly changes from almost all sea to all land while moving closer to the reference station. The propagation over the (salt) sea path will be only very mildly influenced by seasonal changes (assuming that the sea path does not freeze); whereas, the land path potentially suffers significantly from seasonal changes. This observation suggests a simple modification of the double range difference model: exclude saltwater paths when calculating ranges. The resulting model is called the Modified Double Range Difference Model.
The Modified Double Range Difference Model has been used satisfactory in the USCG 1980s stability studies \[5\][6][7]. Its simplicity is appealing, but it also limits the applicability.

If both the user and the reference station are on a transmitter radial, then only the local weather between the user and reference station has influence on the validity of the temporal correction. In all other cases, the whole propagation path between both the user and transmitter, as well as between the reference station and transmitter needs to be taken into account. The challenging task of mathematically modeling propagation variations has been described by Cambell, Doherty, and Johler \[8\] (Section 3.4.4), for example. Those discussions are vital to understanding the phenomena, but might have their shortcomings in accurate quantitative predictions. Actual measurements of day-to-day, month-to-month, and year-to-year variations at strategic locations are most likely necessary for assessments that are more reliable.

Most currently available measurements are from the 1980s and are time-difference measurements of (SAM-controlled) Loran signals. The resulting measurement accuracy was, therefore, limited. However, the results, such as those shown in Section 3.4.4, are still insightful and give a first impression of what level of variation can be expected. A more detailed analysis of the spatial decorrelation of the temporal corrections requires new absolute TOA measurements with high accuracy. For those measurements to be useful, they have to be taken simultaneously at multiple strategic locations for an extensive period.

### 4.6 Discussion

To meet the stringent HEA accuracy requirements of 20 meters, 95%, all components of the error model must be taken very seriously. Biases must be as good as removed, which results effectively in upgrading Loran’s absolute positioning accuracy to the level of its (short-term) repeatable accuracy. To achieve this, the spatial ASF variations must be charted, and the temporal fluctuations must be measured and transmitted to the user.

A high quality ASF measurement setup is a vital part of the dLoran- and HEA-related research and implementation. The research, design, and implementation of this measurement equipment constitute a major role in this Ph.D. research, which is reflected by most of the content of this dissertation (Chapter 3, 5 and 6). Chapter 6 contains various exploratory measurements with the developed equipment in various stages of its design. Furthermore, extensive tests have been performed in a maritime environment, thereby providing a proof-of-concept of the HEA application (Section 6.4).

Preferably, the reference station is re-radiation free, which requires careful site-selection. Re-radiation by bridges, for example, forms an interesting challenge, which is further outlined in Section 6.4.6.

The spatial decorrelation of the temporal corrections poses a fundamental problem that requires special attention. The effect has been researched in the 1980s, but those results are limited by the performance of the equipment available at that time. Renewed year-round
measurements by a large number of strategically placed monitor stations will help to understand this phenomenon better. With that knowledge, it should be possible to deduce the best location of the reference stations and to bound the error induced by the spatial decorrelation of the temporal corrections from that reference station.

Finally, the performance of the user equipment is obviously of a major influence on the total error budget. To meet the requirements, the user equipment has only a slim error margin with respect to noise mitigation, antenna calibration, and tracking performance, for example. On the other side, the user equipment needs to be affordable enough to allow for widespread use. An extensive user base is subsequently needed to justify the necessary investments politically and economically.

4.7 References

[2] “Analysis of the Effects of ASF Variations for Loran RNP 0.3”, Sherman Lo, Per Enge, Stanford University, ILA 2003
5 Low-Frequency H-field Antenna

The antenna is perhaps the most important part of a radionavigation receiver. Electromagnetic waves need to be transformed into electrical signals before they can be processed into time-of-arrival, pseudorange, and finally a position. This transformation is always cursed with noise and parasitic effects, such as distortion and undesired couplings, for example. Smart algorithms can compensate for some of these parasitics, other deficiencies cause permanent degradation in positioning and/or timing performance. The antenna plays an important role in the total error budget; therefore, it deserves careful analysis and design.


5.1 Introduction

The LF Electro-Magnetic (EM) wave can be probed using either an E-field or an H-field antenna, both have their particular advantages and challenges.

5.1.1 E-field antenna

The vertically polarized EM-wave can be probed with an E-field antenna consisting of a vertical conductor and associated amplifier. In contrast to the H-field antenna, the vertically oriented E-field antenna has an omni-directional radiation pattern (with respect to the vertically polarized EM-wave), which allows the use of a single-channel receiver.

The relative simplicity of the E-field antenna made it the sensor of choice for decades of LF radionavigation and associated research and development. This Section only briefly describes the E-field antenna; the reader is referred to the extensive literature available, for example [11] and [22], for a more detailed description of this sensor and its design methodology.

The 3 km Loran-C wavelength makes any practical E-field receiving antenna infinitesimal resulting in capacitive source impedance as depicted in Figure 5-1.

![Diagram](image)

**Figure 5-1:** E-field antenna followed by a charge-amplifier. $C_a$ depicts the capacitive source impedance, $C_{fb}$ the feedback capacitance and $C_{iss}$ the total input capacitance of the FET including the diode and wiring capacitances. [11]

The capacitive source impedance of the infinitesimal E-field antenna makes a charge amplifier topology, such as shown in Figure 5-1, a suitable implementation. Literature ([11], [22]) has shown that with this type of active antenna it is possible to achieve an antenna noise contribution lower than the external atmospheric noise while using a short whip (decimeters length or less). Furthermore, the virtual ground of the charge-amplifier configuration makes it robust against large electric discharges and relative insensitive to shortening by splashes of seawater, for example. Unfortunately, the configuration shown in Figure 5-1 has a rather wide-band frequency transfer. The lack of selectivity allows out-of-band interference to endanger the linear operation of the active part of the antenna system. Even cell-phone
transmissions in the GHz band can potentially cause intermodulation, distorting the transfer of the desired 100 kHz signals. Passive frequency-selectivity, prior to the electronic amplification, would significantly ease the active amplifier requirements. However, the resonance circuit required for the desired frequency selectivity requires a rather impractical or even impossibly large inductance (e.g. between 0.1 and 1 H [11]).

Another method of amplification is the use of a high-input impedance voltage amplifier. This type of amplifier is generally better known among analog circuit designers than the charge amplifier; therefore, it is frequently used in this application. However, the voltage amplifier configuration is highly sensitive to salt water splashing against the antenna (see Section 6.4.1.5), is more likely to break down as a result of nearby high-energy discharges, and finally has a poorer noise performance. Therefore, in most situations the charge amplifier is preferred over the voltage amplifier configuration.

A drawback of the E-field monopole antenna is its grounding requirement. The electric field is probed with respect to a rather arbitrary ground potential. In stationary applications, the grounding is often achieved by placing a conducting pin in the ground. In mobile setups, the ground connection usually has to be made to the vehicle chassis. On a metal boat this is relative straightforward; whereas, on a car this can be more problematic. Because a car has rubber tires, there is no clearly defined ground path. It is possible that capacitive coupling between the car and constructions in or under the road cause the ground pattern to change with a different antenna response as a result. This change in antenna response has a common-mode effect on all the received signals. In short: an E-field antenna mounted on a car does not probe the field at a well-defined location, but probes a potential with respect to an arbitrary reference. This makes the radiation pattern of the E-field antenna a function of the antenna itself and its surroundings. Section 6.3 shows examples of common-mode phase fluctuations that might be caused by this grounding problem. In principle, it should be possible to construct a LF dipole E-field antenna that differentially probes the electric field. Using fiber optics for the signal connection between antenna and receiver further enhances the definition of the phase center of this differential probe. The author suggest further research in such a research antenna, as it would help to identify the common mode fluctuations seen in Section 6.3, for example, thereby enabling a better understanding of the propagation phenomena.

A serious drawback of the E-field antenna is its inherent susceptibility to precipitation static or p-static (Section 3.6.4.1). This effect is created by charged raindrops falling on the (capacitive) antenna, and in airborne situations by discharges on the aircraft body. Using dischargers on the wings, combined with careful aircraft maintenance, can reduce the latter [23]. In spite of these mitigation efforts, p-static is, and perhaps always will be, the Achilles heel of LF E-field antennas. Fortunately, the H-field antenna hardly shows any susceptibility to p-static [23][14], which in recent years has motivated extensive research, among others by the author, in the magnetic antenna.

### 5.1.2 H-field antenna

The LF H-field antenna consists of a so-called “loop” antenna, which has some distinct advantages over the E-field “whip” antenna:
– The H-field antenna is virtually insensitive to p-static effects. This is a major advantage in airborne situations where p-static forms a major weakness of Loran-C [14].
– H-field signals tend to penetrate much better in dense urban environments and mountainous areas, for example. This makes the H-field antenna equipped Loran-C receiver much more suitable for applications such as land-mobile track-and-trace [15][16][24].
– H-field antennas do not require grounding and are low profile. This makes them ideal for integration with GPS (patch) antennas or placement in covert locations.

In contrast to the E-field “whip” antenna, the loop antenna does not have an omni-directional radiation pattern:

A dual-loop configuration is required to obtain an omni-directional radiation pattern (Section 5.7). A pleasant advantage of the dual-loop configuration is its ability to be used as an accurate radio heading (compass) device. This compass functionality has proven to be an important commercial benefit for maritime applications [21], for example. Next to a figure-8 radiation pattern, the sensor suffers also from a 180° phase ambiguity [20] that needs to be solved.

The required extra receiver complexity to obtain an unambiguous omni-directional radiation pattern has made receiver manufacturers reticent to use H-field technology for a long time. Advances in technical capabilities and the distinct advantages of H-field antennas have turned the tide resulting in a growing interest in this technology since the 1990s [14][17][18] and an additional push caused by the eLoran efforts initiated by the US-FAA in 2000.

The low noise design of an H-field antenna is even more challenging than that of the E-field antenna. Section 5.4 outlines a structured, orthogonal method of low-noise H-field antenna design. The structured design methodology presented (following [13] and [19]) first investigates the influence of all parameters separately. Next, the design process is split in orthogonal parts, resulting in a straightforward process towards an optimal implementation, while limiting the required iterations.
Unfortunately, the H-field antenna is not only sensitive to H-field, but parasitically to E-field as well. The parasitic E-field pickup will lead to heading-dependent phase and amplitude errors, and therefore, needs to be mitigated. Theory, measurements, and mitigation methods are discussed in Section 5.5.

Usually, an LF H-field antenna intended for Loran-C reception is operated in resonance at 100 kHz. The cause, effect, and compensation of potential tuning mismatches and fluctuations are detailed in Section 5.6.

Finally, parasitic coupling between the two loops and between the vehicle and the antenna causes heading dependent errors. Section 5.8 describes these cross-talk errors and possible mitigation techniques.

### 5.1.3 E-field versus H-field

In a total system perspective, the usage of E-field versus H-field antennas can be evaluated in two ways: regarding propagation and regarding the sensor deficiencies. In far-field situations, the E-field and H-field have a fixed relation, and therefore, render equal performance. Due to the 3 km Loran-C wavelength, however, the receiving antenna is relatively often located in the near field of structures. In these situations, the E-field and H-field propagation is far from similar resulting in significant differences in positioning and timing performance.

The reader is referred to Chapter 6 for various detailed comparative measurements between H-field and E-field in near-field situations. Local interference sources can be more severe in either E-field or H-field radiation as well. Section 3.6.4 discusses this in more detail.

From a sensor perspective, both E-field and H-field antennas have their advantages and disadvantages, among others:

<table>
<thead>
<tr>
<th>Advantage</th>
<th>Disadvantage</th>
</tr>
</thead>
<tbody>
<tr>
<td>+ omni directional radiation pattern</td>
<td>– susceptible to p-static</td>
</tr>
<tr>
<td>+ relative simple low noise design</td>
<td>– needs grounding</td>
</tr>
<tr>
<td>+ can be combined with e.g. car radio antenna</td>
<td>– wide-band sensitivity can cause challenges with respect to e.g. intermodulation</td>
</tr>
</tbody>
</table>

The LF E-field antenna has been discussed extensively in literature; the references [11] and [22] form a good start for more information on the subject. This dissertation focuses further on the LF H-field antenna, the associated challenges, potential solutions, and measurements from laboratory and field where theory has been brought into practice as well.
5.2 Description of an H-field sensor

The loop antenna is the preferable sensor to convert the H-field component of the LF Loran-C EM-waves into an electrical signal. A fairly large coil diameter is required to achieve sufficient low-noise performance. Fortunately, the coil-diameter can be reduced significantly by inserting ferrite material into the loop. This Section will evaluate the design and performance of the ferrite-loaded loop in detail.

5.2.1 Equivalent circuit

Figure 5-3 shows the ferrite-loaded coil and its equivalent electric circuit.

The induced voltage \( E_s \) and the inductance \( L \) can be represented \([1][6]\) as:

\[
E_s = F_A \mu_0 \mu_{rod} \omega ANH
\]

\[
L = K\mu_0 \mu_{rod} \frac{N^2 A}{I_r}
\]
where $N$ is the number of turns, $A$ the rod cross-section area ($\pi d^2/4$), $l$, the length of the ferrite rod, and $H$ is the magnetic field strength in A/m. These two equations play a key-role in the following Sections.

In situations where the far-field relation is valid, $H$ equals:

$$|H| = \frac{|E|}{Z_0}$$  \hspace{1cm} (5-3)

where $E$ is the electrical field strength in [V/m] and $Z_0$ the free space impedance of $120\pi$ Ω. Note that due to re-radiation, 5-3 loses validity in close proximity of large objects such as bridges and building.

5.2.2 $\mu_{\text{rod}}$

The effective permeability $\mu_{\text{nd}}$ depends on the relative permeability $\mu$, of the ferrite material and on $m$, the ratio of the rod-length, and the rod-diameter ($l/d_r$). Figure 5-4 shows $\mu_{\text{nd}}$ as function of $l/d$, with $\mu$, as variable [2].

**Figure 5-4:** Rod permeability $\mu_{\text{nd}}$ as a function of length-to-diameter ratio $l_{\text{rod}}/d_{\text{rod}}$ with material permeability $\mu$ as a parameter (graph based on data from Ferroxcube Data Handbook [2])
In practical situations, where $2 \leq m \leq 20$, $\mu_{rod}$ for a cylindrical rod can be approximated by [5]:

$$\mu_{rod} = \frac{\mu_r}{1 + D(\mu_r - 1)} \tag{5-4}$$

where $\mu_r$ is the relative permeability and $D$ the demagnetization factor of the rod. According to [3][4] $D$ can be modeled as:

$$D \approx 0.37 m^{-1.44} \quad \text{for} \quad 2 \leq m \leq 20 \tag{5-5}$$

### 5.2.3 $F_A$ and $K$

Both $F_A$ and $K$ are not very clearly described in the literature. All non-ideal behavior of the ferrite-loaded rod seems to be united in these two factors. However, for a predictable behavior of the sensor, it is crucial to know the influence of geometry, material, and implementation of the ferrite-loaded rod on the factors $F_A$ and $K$.

As shown in Figure 5-5-right, in a ferrite rod, not all magnetic field-lines enter and leave the rod at the frontal area as would be the case with an ideal ferrite rod, shown in Figure 5-5-left. This effect will lead to a difference in “efficiency” between a turn at the center and a turn at the edge of the rod. The distribution of turns on the rod has a dominant influence on the factors $F_A$ and $K$. $F_A$ and $K$ have been determined experimentally as function of the ratio of the length of the coil and the length of the rod ($l_c/l_r$). For a 10 cm long and 1 cm diameter Philips 3F3 rod, the measurement results are shown in Figure 5-6. Other ferrite material or geometry may lead to slightly different results.
5.2.4 $R_{\text{loss}}$

The resistive part $R_{\text{loss}}$ of the equivalent circuit given in Figure 5-3-right consists of:

$$R_{\text{loss}} = R_{\text{ferrite}} + R_{\text{wire}} + R_{\text{radiation}}$$  \hspace{1cm} (5-6)

$R_{\text{ferrite}}$ is caused by the ferrite losses and can be represented by [8]:

$$R_{\text{ferrite}} = \frac{\omega L \mu''}{\mu'}$$  \hspace{1cm} (5-7)

where $\mu''$ is the imaginary (loss) part and $\mu'$ the real part of the ferrite's permeability.

$R_{\text{wire}}$ is represented by:

$$R_{\text{wire}} = R_{\text{DC}} + R_{\text{skineffect}}$$  \hspace{1cm} (5-8)

The skin depth $\delta$ is given by the equation:

$$\delta = \sqrt{\frac{2}{\omega \sigma \mu}}$$  \hspace{1cm} (5-9)

For copper wire, the permeability $\mu \approx \mu_0$ and the conductivity $\sigma = 5.81 \cdot 10^7$ S. This results in a skin depth $\delta$ of 0.21 mm at 100 kHz. For wire diameters of $d<3\delta$, the skin effect can be neglected in comparison with the DC-resistance $R_{\text{DC}}$ [7], which makes the skin effect negligible for most Loran-C H-field antennas.

The ohmic or DC-part $R_{\text{DC}}$ of $R_{\text{wire}}$ is given by:

$$R_{\text{DC}} = \frac{1}{\sigma S}$$  \hspace{1cm} (5-10)
new potential of low-frequency radionavigation in the 21st century

where $S$ is the cross-section and $\sigma$ the conductivity of the wire. For a coil consisting of a copper wire of 0.3 mm diameter, for example, $R_{dc}$ is approximately 0.25 $\Omega$ per meter wire length at 100 kHz.
The radiation resistance $R_{\text{radiation}}$ is negligibly small and is not taken into account.

## 5.3 Topology

Figure 5-7 and Figure 5-8 show two possible topologies to match and amplify the signal from the loop antenna. The antenna can be operated in resonance according to Figure 5-7, or as a wide-band sensor such as depicted in Figure 5-8.

![Resonance configuration](image1)

**Figure 5-7: Resonance configuration**

![Wideband configuration](image2)

**Figure 5-8: Wideband configuration**

### 5.3.1 Resonance configuration

Figure 5-7 shows a ferrite-loaded rod in resonance followed by an amplifier. The resonance frequency $\omega_0$ and the required tuning capacitance $C$ can be found from:

$$\omega_0 \approx \sqrt{\frac{1}{LC}} \Rightarrow C \approx \frac{1}{L\omega_0^2}$$

(5-11)
An undamped resonance circuit has a relatively small bandwidth. This selectivity can be useful for a small-band AM-broadcast receiver, but it is not suitable for the wideband Loran-C reception. The 3 dB bandwidth of a resonance circuit is specified by the “quality-factor” $Q$, which is defined as the ratio between the center frequency and the 3 dB bandwidth. For Loran-C the required bandwidth is approximately 25 kHz [9], which limits the maximum $Q$ of the resonance-circuit to $100/25=4$. In practice, a $Q$ of 3 is frequently used. To achieve this, a damping resistor $R_L$ can be added as shown in Figure 5-9.

The influence of the damping resistor $R_L$ on the quality factor $Q$ is given by:

$$Q = \frac{R_L}{\omega_0 L} \Rightarrow R_L = Q \omega_0 L \quad \text{for} \quad R_L \gg R_{\text{loss}} \quad (5-12)$$

The implementation of $R_L$ as a physical resistor may lead to an unacceptable decrease of noise performance of the receiver. Fortunately, $R_L$ can also be implemented as the input impedance of a power-to-voltage or a power-to-current amplifier. The improvement of noise performance by this complex type of amplifier will be discussed further in Section 5.4.2. The output voltage $V_{\text{out}}$ at the resonance frequency $\omega_0$ is given by (5-13), where $G$ is the voltage amplification factor of the amplifier.

$$V_{\text{out}(\omega=\omega_0)} \approx E_s \cdot Q \cdot G \quad (5-13)$$

### 5.3.2 Wideband configuration

The topology shown in Figure 5-8 can be used to obtain a wideband active antenna. This can be useful for such applications as CityFix [10], where miscellaneous broadcast transmitters at various frequencies are tracked to obtain a relative position determination. A wideband antenna makes it also possible to receive, for example, Loran-C (90-110 kHz) and DGPS radio beacon (283.5 – 325.0 kHz) signals simultaneously.

The output voltage $V_{\text{out}}$ is given by (5-14) where $Z_{FB}$ is the feedback impedance.

$$V_{\text{out}} \approx -\frac{Z_{FB}}{j\omega L} \cdot E_s \quad \text{for} \quad R_{\text{loss}} \ll j\omega L \quad (5-14)$$

When a resistor $R_{FB}$ is used for $Z_{FB}$ the following transfer is obtained:
Referring to (5-1), $E_s/\omega$ results in a frequency independent transfer from H-field to output voltage. Note that in practice the ferrite permeability is frequency-dependent and so will be the transfer. Unfortunately, the use of a resistor as feedback impedance produces significant noise. Another option is the use of a feedback-inductance $L_{FB}$:

$$V_{out} \approx -\frac{R_{FB}}{j\omega L} \cdot E_s = -\frac{R_{FB}}{jL} \cdot \frac{E_s}{\omega}$$  (5-15)

$$V_{out} \approx -\frac{j\omega L_{FB}}{j\omega L} \cdot E_s = -\frac{L_{FB}}{L} \cdot E_s$$  (5-16)

$V_{out}$ now shows the same linear frequency dependency as $E_s$.

### 5.3.3 Wideband versus resonance

The pulsed Loran-C system requires a relative wide-band antenna and front end, resulting in a much lower Q than generally used for most LF systems such as AM radio receivers. The choice between a damped resonant antenna and a truly wide-band antenna is, therefore, not obvious:

An advantage of the resonance configuration is its frequency selectivity. Off-band interference is largely rejected before passing the first amplifier stages. This eases the amplifier design with respect to distortion and intermodulation. The natural frequency selectivity of the resonant antenna can even be sufficient to meet the Nyquist criteria of modern sigma-delta AD converters making additional analog filtering superfluous. The resonant circuit needs to be tuned accurately to 100 kHz during production. The resonant frequency, and therefore, also the phase response, is a function of temperature and can potentially cause positioning and timing errors.

The wideband configuration is most versatile. The same antenna can be used to receive, for example, Loran-C, DGPS radio beacons, DataTrak signals, DCF timing signals, and “Signals Of Opportunity” such as AM radio broadcasts. No tuning is needed and the antenna is, therefore, potentially less susceptible to temperature fluctuations. Off-band interference is not attenuated by the sensor itself and demands a more careful amplifier design with respect to distortion and intermodulation.

Analysis by the author revealed that the noise performance of a 100 kHz resonant H-field antenna with a Q of 3 is better than that of a wide-band H-field antenna\(^1\). Based on this superior noise performance, together with the favorable theoretical and operational knowledge, the author has chosen the resonant configuration for further analysis, design, and implementation. One should, however, note the appealing versatility of the wide-band configuration and further research in that area is, therefore, highly recommended as well.

---

\(^1\) For this comparison, the wide-band configuration was cascaded with a filter section to obtain a similar frequency transfer function as the resonant antenna. At a Q of 3, the resonant configuration showed superior noise performance, lower Qs were increasingly in favor of the wide-band configuration.
5.4 Noise

5.4.1 Desired antenna noise performance
The application in which the Loran-C sensor is used imposes the minimum receiver performance regarding minimal tracking performance and maximum integration time or tracking loop bandwidth.
From these specifications, the required minimal signal-to-noise ratio (SNR) can be calculated. The SNR is determined by:

\[
SNR = \frac{\text{received Loran-C signal strength}}{\text{external noise + internal noise}}
\]  

(5-17)

5.4.1.1 Received Loran-C signal strength
The received Loran-C signal strength is a function of transmitted field strength (Effective Radiated Power or ERP, the Transmitter power multiplied by the transmitter antenna efficiency), distance to the transmitter, and the propagation path to the transmitter. By using higher power transmitters, the system provider needs fewer transmitters to guarantee a minimal signal level at the user. However, the actual receiver performance is not only a function of received signal level, but also of the receiver position with regard to the transmitter positions, unknown propagation delays, and potential sky-wave contamination of the signal. Therefore, it is usually beneficial to have multiple small transmitters nearby over a minimal configuration of fewer, but stronger, transmitters further away.

5.4.1.2 External versus internal noise
The total noise contribution can be divided into noise that is internal and to noise that is external to the receiving system. The antenna and receiver cause the internal noise, and this noise has the characteristics of thermal noise (white Gaussian noise). The external noise consists of atmospheric, galactic, and man-made noise, and it is generally non-Gaussian. Some of these non-Gaussian components can be removed using special filtering techniques, such as bandpass filters, notch filters, noise sensing, or “hole-punching” techniques (Section 3.6). It is obvious that there is no reason to strive for a noise contribution of the receiving system that is below external noise (after noise-reduction) [12].

Figure 5-10 shows the atmospheric noise levels at 100 kHz as measured by the CCIR [12] for Europe and Florida. Note that the extreme noise levels, in Florida during the summer evenings, for example, are caused primarily by impulsive noise produced by relative near-by thunderstorms. Time-domain filtering can significantly reduce these noise-spikes as discussed in detail in Section 3.6.1.3. However, even after time-domain noise reduction, there is a significant difference in atmospheric noise level between Europe and Florida, between the seasons and between day and night.
The large differences in noise levels make the desired antenna noise specification a non-trivial task. When optimal performance is required, for monitoring purposes, for example, the receiver noise should be less than the external noise under virtually all circumstances. This receiver noise performance is represented in Figure 5-10 by line “A.” However, this low-noise behavior is only beneficial for the user in Europe during the winter at noon. At all other times, a receiver with lower noise specifications (and therefore smaller and/or cheaper) will be sufficient. When the receiver/sensor size is a crucial factor, for pedestrian applications, for example, a smaller but noisier antenna can be most suitable, represented by line “B.” To obtain sufficient signal-to-noise ratio for such a small antenna, it might be necessary to increase Loran-C transmitter power or to have more transmitters in close proximity of the receiver. For most applications, a trade-off will be made like receiver “C.”

The influence of sensor and amplifier configuration and implementation on noise behavior will be discussed in detail hereafter.

### 5.4.2 Receiver noise analysis for resonance topology

As shown in Section 5.3.1, the resonance circuit has to be damped to obtain the necessary Loran-C bandwidth. The most obvious implementation is shown in Figure 5-11.

The noise current $I_{NRL}$ of the damping resistor $R_L$ is given by the equation:

$$I_{NRL}^2 = \frac{4kT}{R_L}$$  \hspace{1cm} (5-18)

where $k$ is the Boltzmann-constant and $T$ the absolute temperature. In this configuration, the contribution of $I_{NRL}$ to the total receiver noise is probably dominant and can be unacceptably high for a low-noise receiver design.
A dual-loop negative-feedback amplifier with the same input impedance \( R_L \) can lower this noise contribution significantly. Depending on the required output impedance, three types of amplifiers can be used for the dual-loop implementation [13]: a power-to-power, a power-to-voltage, or a power-to-current amplifier. The remainder of this analysis is based on the power-to-voltage implementation shown in Figure 5-12. Note that in a practical low-cost antenna, a transformer is undesired. Fortunately, alternate dual-loop configurations are also possible, for example, with an active part in the feedback loop. The principle of the noise analysis presented in this Section is generally also valid for those alternate configurations.

Two feedback-loops, the resistor \( R \), and the transformer with turns-ratio 1:n determine the input impedance and voltage gain:

\[
Z_{\text{in}} = R / n \\
\text{VoltageGain} = n
\]

(5-19)

When the input impedance \( Z_{\text{in}} \) is made equal to \( R_L \), then \( I_{NR} \) can be given by:
\[
I_{NR}^2 = \frac{4kTR}{R} = \frac{4kT}{nR_L} = \frac{I_{NRL}^2}{n}
\] (5-20)

Equation 5-20 shows that the noise-power of damping resistor \( R_L \) has been reduced by the transformer ratio \( n \). The gain-bandwidth product of the active-part of the amplifier and the practical implementation of the transformer determine the maximum value of \( n \). A transformer ratio \( n \) of 20, for example, is feasible. Unfortunately, \( I_{NR} \) is not the only noise source as is shown in Figure 5-13. The noise of the active part is represented by the noise voltage source \( U_{NT} \) and the noise current source \( I_{INT} \). \( UNRloss \) represents the noise caused by \( R_{loss} \). The transformer, designated by the windings “1” and “n,” is considered ideal, which is only allowed in case of perfect coupling and when the inductance of the “1” side of the transformer (drawn left in Figure 5-13) is small in comparison with the antenna-inductance \( L \).

**Figure 5-13: Amplified resonance circuit with noise-sources**

It is now possible to calculate \( U_{Ntot} \), the total equivalent receiver noise voltage source:

\[
\overline{U_{Ntot}^2} \approx \omega_0 \int_{\omega_1}^{\omega_2} \left( U_{NRloss}^2 + \omega^2 L^2 I_{NR}^2 \right) + \left( 1 - \left( \frac{\omega}{\omega_0} \right)^2 \right)^2 U_{NT}^2 + \omega^2 L^2 I_{INT}^2 \right) \cdot d\omega
\] (5-21)

From the combination of equation (5-21), (5-7) and (5-12) it can be deduced that:

\[
U_{NRloss}^2 = 4kT R_{loss} = 4kT \omega L \frac{\mu_0''}{\mu_s'}
\] (5-22)

\[
I_{NR}^2 = \frac{4kT}{R} = \frac{4kT}{nR_L} = \frac{4kT}{nQ \omega_0 L}
\] (5-23)

The combination of (5-21), (5-22) and (5-23) gives:
The noise voltage \( U_{NT} \) and noise current \( I_{NT} \) are assumed constant over the bandwidth \( \omega_1 \leq \omega \leq \omega_2 \). The contribution of \( I_{NT} \) to the total noise \( U_{Not} \) depends on the inductance \( L \) that can be set to any value (within practical limits) by choosing the number of turns \( N \) (5-2). \( N \) also influences the induced voltage \( E_s \) (5-1). To obtain the optimal value for \( L \) it is necessary to calculate the signal-to-noise ratio or the equivalent receiver-noise field-strength.

### 5.4.3 Equivalent receiver-noise field strength

The noise-performance of the receiver is evaluated by comparing the total receiver noise with the atmospheric noise. The obtained equivalent total receiver noise source \( U_{Not}^2 \) (5-24) is expressed in \([V^2]\); whereas, the atmospheric noise is given in \([A/m] / Hz\). A conversion is needed to make comparison possible. The most appropriate conversion is to describe the noise-performance of the receiver as an equivalent receiver-noise field-strength. Figure 5-14 shows this concept.

The noise power is normalized to the unit bandwidth of 1 Hz. The receiver noise performance can now be expressed as an equivalent receiver-noise field strength, which is a single unambiguous parameter that fully specifies the noise behavior of the receiver front end regardless of the amplifier-gain, sensor-gain, and the receiver bandwidth.

The equivalent circuit for the ferrite loaded loop Figure 5-3-right and the equation for the induced voltage \( E_s \) (5-1), are used to calculate the equivalent receiver-noise field strength \( H_{\text{Receiver}} \):
where $B$ is the receiver bandwidth, $E_s/H$ the H-field transfer function, and $U_{Ntot}$ the equivalent total receiver noise source. The influence of the different design parameters on the equivalent noise field-strength can be made comprehensible with the following substitutions:

\[ L = K \mu_0 \mu_{rod} \frac{N^2 A}{I_r} \Rightarrow N = \sqrt{\frac{L}{K \mu_0 \mu_{rod} A I_r}} \tag{5-26} \]

\[ E_s = \frac{F_A \mu_0 \mu_{rod} \omega \alpha N H}{H} = \frac{F_A}{\sqrt{K}} \cdot \sqrt{\mu_0 \mu_{rod} A I_r \cdot \omega \cdot \sqrt{L}} \cdot \left[ \frac{V}{A/m} \right] \tag{5-27} \]

The combination of (5-25), (5-26) and (5-27) then gives:

\[ H_{receiver} \approx \left( \frac{F_A}{\sqrt{K}} \cdot \sqrt{H_0 \cdot \mu_{rod} \cdot A \cdot I_r} \right)^{-1} \cdot \sqrt{\frac{U_{Ntot}^2}{B \cdot \omega_0^2 \cdot L \sqrt{Hz}}} \cdot \left[ \frac{A/m}{A/m} \right] \tag{5-28} \]

where $\omega_0$ is taken as the average of $\omega$.

### 5.4.4 Orthogonal design

With the expression for $H_{receiver}$ (5-28), it is now possible to design the front-end according to the required noise specifications as specified in Figure 5-10, for example. A closer look at equation 5-28 reveals that the first term represents the sensor geometry and the ferrite material; whereas, the second term reflects the influence of the active part on the noise performance. When optimizing this second term for low noise, one should strive for a high inductance $L$ and a low $U_{Ntot}^2$ at the same time. This demand is contradictory, as can be concluded from (5-24), where a higher $L$ implies a larger influence of the noise current $I_{NS}$ of the active part. The optimal inductance is found by solving the following:

\[ \frac{\delta (H_{receiver})}{\delta L} = 0 \tag{5-29} \]
\[
\delta \left( \frac{U_{\text{Not}}^2}{L} \right) = - \frac{U_{\text{NT}}^2}{L^2} \int_{\sigma_1}^{\sigma_2} \left(1 - \left( \frac{\omega}{\omega_0} \right)^2 \right)^2 \cdot d\omega + I_{\text{NT}}^2 \int_{\sigma_1}^{\sigma_2} \omega^2 \cdot d\omega = 0 \tag{5-30}
\]

From 5-30 it follows that:

\[
U_{\text{NT}}^2 \int_{\sigma_1}^{\sigma_2} \left(1 - \left( \frac{\omega}{\omega_0} \right)^2 \right)^2 d\omega = I_{\text{NT}}^2 L_{\text{opt}}^2 \int_{\sigma_1}^{\sigma_2} \omega^2 d\omega \tag{5-31}
\]

In literature, this is known as a noise-matching situation: the contribution of the noise current \(I_{\text{NT}}\) to the total noise equals the contribution of the noise voltage \(U_{\text{NT}}\) to the total noise. The optimal inductance \(L_{\text{opt}}\) is now given by:

\[
L_{\text{opt}} \approx \frac{U_{\text{NT}}}{I_{\text{NT}}} \sqrt{\int_{\sigma_1}^{\sigma_2} \left(1 - \left( \frac{\omega}{\omega_0} \right)^2 \right)^2 d\omega \int_{\sigma_1}^{\sigma_2} \omega^2 d\omega} \tag{5-32}
\]

Concluding, the optimal inductance, \(L_{\text{opt}}\) is imposed by the active part implementation and the bandwidth of the front-end; whereas, \(L_{\text{opt}}\) can be realized independently of the sensor implementation by choosing the correct number of turns \(N\). This approach offers an orthogonal relation between the design of the sensor and that of the amplifier, which implies that both parts can be designed individually.

### 5.4.5 H-field sensor implementation

Based on 5-28 it is now possible to compare the equivalent receiver noise field strength with the atmospheric noise:

\[
\frac{H_{\text{receiver}}}{H_{\text{atmospheric}}} = 10 \cdot \log \left( \frac{U_{\text{Not}}^2}{B \cdot \omega_0^2 \cdot L} \right) + \left( -20 \cdot \log \left( \frac{F_4}{\sqrt{K}} \right) + \right. \\
+ \left. -10 \cdot \log (\mu_0 \cdot \mu_{\text{rod}} \cdot A \cdot l_r) + \right. \\
\left. -20 \cdot \log \left( H_{\text{atmospheric}} \right) \right) \text{ [dB]} \tag{5-33}
\]

These four terms show a large orthogonality. The second and third terms are related to the H-field sensor implementation, and they will be discussed in more detail in the remainder of this Section.
5.4.5.1 Distribution of turns on the rod

Both the parameters \( F_A \) and \( K \) mainly depend on the distribution of turns on the rod (e.g. a fully wound rod versus a narrow coil). The relation between \( F_A \) and the induced voltage \( E_s \) is given by (5-1); whereas, (5-2) gives the relation between \( K \) and the inductance \( L \). The influence of \( F_A \) and \( K \) on the noise performance is given by the second term of equation 5-33:

\[
20 \cdot \log \left( \frac{F_A}{\sqrt{K}} \right) \text{ [dB]} \tag{5-34}
\]

Figure 5-15 shows the relation between equation 5-34 and the measured data shown in Figure 5-6.

This Figure shows that the best noise performance is achieved for a fully wound rod: an improvement in SNR of approximately 2.2 dB is obtained when a fully wound rod is used instead of a narrow coil at the center of the rod.

\[\text{Figure 5-15: Influence of the coil length / rod length on noise performance}\]

5.4.5.2 Rod dimensions and material

The third term of equation 5-33,

\[
10 \cdot \log(\mu_0 \cdot \mu_{rod} \cdot A \cdot l_r) \tag{5-35}
\]

gives the relation between rod geometry, material, and the noise performance.

Figure 5-16 was derived using the data provided by Figure 5-4, and shows the influence of the rod-length and rod-diameter on the noise performance. The rod permeability \( \mu \), has been fixed at 2000.
The rod-length has a larger influence on the noise-performance than the rod diameter. Therefore, in weight-limited applications, rod length is preferred over rod diameter. However, this may conflict with the requirements of volume-limited applications, such as mobile/handheld.

![Figure 5-16: Influence of rod length and rod diameter on noise performance (for $\mu_r=2000$). The lines represent the antenna noise performance in dB given by $10 \cdot \log\left(\mu_0 \cdot \mu_{rod} \cdot A \cdot I_r\right)$](image)

### 5.4.5.3 Influence of sensor optimization on SNR

Until now, only a single cylindrical ferrite rod has been discussed. Other rod geometries or, for example, multiple rods can potentially also increase noise performance. Figure 5-17 shows the process of a convenient substitution to calculate the influence of a change of a sensor-parameter on the noise performance.

![Figure 5-17: Equivalent circuit for optimized sensor](image)

The first step of the substitution process is modifying the H-field sensor while maintaining the same number of turns $N$. In the now obtained equivalent circuit, both $R_{loss}$ and $L$ are corrected...
by $\alpha$, and the induced voltage $E_i$ is corrected by $\beta$. In the second step, the number of turns on the rod $N$ has been changed to regain the original optimal noise-matched inductance $L_{opt}$.

The SNR of the optimized H-field sensor can now be given by:

$$
SNR = 10 \cdot \log \left( \frac{\left( \frac{\beta}{\sqrt{\alpha}} \right)^2 \cdot \frac{E_s^2}{U_{Ntot}^2}}{1} \right)
= 10 \cdot \log \left( \frac{E_s^2}{U_{Ntot}^2} \right) + 20 \cdot \log \left( \frac{\beta}{\sqrt{\alpha}} \right)
$$

(5-36)

Using 5-36 the change in SNR is given by:

$$
\Delta SNR = 20 \cdot \log \left( \frac{\beta}{\sqrt{\alpha}} \right)
$$

(5-37)

5.4.5.4 Influence of multiple rods on SNR

In applications where the antenna size is limited, it can be beneficial to use multiple small ferrite rods instead of a single larger rod. This Section discusses the influence of distance between two rods on the antenna noise performance. The analysis is done from a circuit perspective.

Figure 5-18 depicts the coupling that exists when two rods are brought in close proximity of one another.

![Figure 5-18: Overlapping field-lines and inductive coupling between two rods](image)
The coupling between the coils (red lines) causes a lowering of the total inductance \( L \); whereas, the overlap in “captive area” lowers the effective H-field transfer \( E_s \). Maintaining noise matching requires increasing the inductance, which can be accomplished by more turns on the rods. Figure 5-19 shows this process schematically.

\[
\Delta \text{SNR} = 3 + 20 \cdot \text{LOG} \left( \frac{1 - k_{Es}}{\sqrt{1 - k_L}} \right) \ [dB] \quad (5-38)
\]

As expected, equation 5-38 shows a 3dB SNR improvement in the absence of any mutual coupling between the rods.

A special test setup has been built to measure the influence of rod spacing on \( L \) and \( E_s \).

\[L_{opt} = \frac{\sqrt{2 \cdot \frac{1 - k_{Es}}{\sqrt{1 - k_L}} \cdot E_s}}{2(1 - k_L) \cdot R_{loss}}\]

\[\text{adjust } \# \text{ turns N} \]

**Figure 5-19:** Modelling of overlapping captive areas and inductive coupling by a coupling factor \( k_{Es} \) respectively \( k_L \), the resulting equivalent circuit and the required noise-matching.

**Figure 5-20:** Test setup to measure the relation between the distance between the rods and the inductance \( L \) and the H-field transfer \( E_s \).
The resulting measurements are depicted in Figure 5-21.

![Graph showing mutual coupling and SNR](image)

**Figure 5-21: Influence of rod-distance on mutual coupling and SNR**

The measurements show an asymptotic decay of coupling with distance. Surprisingly, the inductive coupling does not converge to zero, resulting in a SNR improvement of more than 3 dB at 10 cm rod separation. Most likely, this is a measurement error. Based on these empirical results, it is plausible to assume that \( k_E \) and \( k_L \) are identical.

From Figure 5-21, it can be concluded that two 10 cm rods experience virtually no mutual coupling when placed 10 cm apart, and therefore, result in a 3 dB increase in SNR compared to a single 10 cm rod. Note that given a square surface of 10x10 cm, for example, using 2x2 10 cm rods in a square configuration is not necessarily beneficial over a two crossed rods. The crossed rods can be placed in the diagonals, allowing them to be \( \sqrt{2} \) times longer. According to Figure 5-16, this also yields a 3 dB SNR improvement over the single 10 cm rod. Other aspects such as parasitic cross coupling, mechanical challenges, and of production cost will, therefore, be most likely the deciding factor in the choice of configuration.

### 5.5 E-field Susceptibility

Unfortunately, an H-field antenna is not solely sensitive to the incident H-field. Every practical H-field antenna suffers from some parasitic E-field susceptibility as modeled in 5-39.

\[
V_{\text{out}} = f_H (\mathbf{H}) + f_E \left( \frac{\mathbf{E}}{Z} \right) \tag{5-39}
\]

where \( f_H \) is the antenna sensitivity to H-field and \( f_E \) the relative sensitivity to E-field.

If we assume a perfect figure-8 H-field response and correct polarization of the H-field antenna then:

\[
f_H (\mathbf{H}) = G \cdot \cos(\varphi) \cdot |\mathbf{H}| \tag{5-40}
\]
where \( G \) is the antenna gain [\( V m / A \)] and \( \phi \) the orientation of the loop with respect to the incident signal. If we further assume that the far-field condition \( Z = |E| / |H| = Z_0 = 120 \pi \Omega \) is satisfied, and that the antenna E-field susceptibility is omni-directional, then the antenna response is given by 5-41:

\[
V_{\text{out}} = G \cdot \cos(\phi) \cdot |H| + G \cdot 10^{ES/20} \cdot j \frac{|E|}{120\pi}
\]  

(5-41)

where \( ES \) is the parasitic E-field susceptibility of the H-field. Because the far-field relation is assumed valid, 5-41 can be simplified further by replacing \( |E| \) by \( 120\pi \cdot |H| \):

\[
V_{\text{out}} = G \cdot (\cos(\phi) - 10^{ES/20} \cdot j) \cdot |H|
\]  

(5-42)

According to 5-42, an \( ES \) of 0 dB combined with a heading \( \phi \) of 0° will introduce a phase error of 45°, which results in a 338 meter pseudorange error at 100 kHz. Figure 5-22 shows the relation between various E-field susceptibilities and the resulting error in Time-Of-Arrival (TOA) measurement.

Figure 5-22: Angular dependent TOA measurement error due to E-field susceptibility \( ES \)

In a practical crossed loop H-field antenna, an individual H-field loop is effectively used from \( \phi = -45^\circ \) to \( +45^\circ \). An E-field susceptibility \( ES \) of -50 dB, therefore, gives a worst-case TOA error of approximately 10 ns, which at 100 kHz corresponds to a 3-meter pseudorange error.

### 5.5.1 E-field susceptibility measurements

Given 5-42, the phase error \( \gamma \) introduced by E-field susceptibility is:
NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY

\[ \gamma = \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha - \delta_T)} \right) \]  

(5-43)

Where \( \alpha \) is the orientation of the H-field antenna regarding north, and \( \delta_T \) is the heading towards the transmitter regarding north. The influence of the E-field susceptibility can be measured by changing \( \alpha \) (physically rotating the antenna), while measuring the time-of-arrival of a Loran signal relative to a clock with a negligible frequency error (e.g. a Cesium or GPS disciplined Rubidium clock).

\[ TD_{\text{error}} = \left( TOA_A + \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha_A - \delta_T)} \right) \right) - \left( TOA_B + \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha_B - \delta_T)} \right) \right) \]

(5-44)

\( TD_{\text{error}} \) is the time-difference error caused by E-field susceptibility. \( TOA_A \) and \( TOA_B \) are the time of arrival assuming a perfect H-field response without any E-field susceptibility. \( TOA_A = TOA_B \) assuming a perfect clock and the absence of re-radiation. In practice, the H-field transfer of the H-field antenna might give some deviation from the ideal \( \cos(\alpha - \delta_T) \) and re-radiation might be present. However, all the H-field transfer related errors, as also potential re-radiation effects, will show 180° symmetry. Therefore, if the antenna is flipped 180°, the H-field component (including all H-field related errors such as cross-talk, tuning, etc.) will invert exactly; whereas, the E-field component remains unchanged. If we assume:

\[ \alpha_B = \alpha_A + 180^\circ \Rightarrow TOA_A = TOA_B \]

(5-45)

then it is straight-forward to calculate the E-field susceptibility \( ES \):

\[ TD_{\text{error}} = 2 \cdot \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha_A - \delta_T)} \right) \]

(5-46)

\[ ES = 20 \cdot \log_{10} \left( \cos(\alpha_A - \delta_T) \cdot \tan \left( \frac{TD_{\text{error}}}{2} \right) \right) \]

(5-47)

If a receiver controlled by a “perfect” clock is not available, it is also possible to accurately measure the E-field susceptibility by measuring the time-differences (TDs) between two Loran-C stations at two different physical orientations of the antenna (\( \varphi_A \neq \varphi_B \)).

\[ TD_A = \left( TOA_{1,A} + \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha_A - \delta_{T1})} \right) \right) - \left( TOA_{2,A} + \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha_A - \delta_{T2})} \right) \right) \]

\[ TD_B = \left( TOA_{1,B} + \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha_B - \delta_{T1})} \right) \right) - \left( TOA_{2,B} + \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha_B - \delta_{T2})} \right) \right) \]

(5-48)
If the antenna is flipped, $\varphi_B = \varphi_A + 180^\circ$, then the time-of-arrival differences caused solely by H-field transfer remains constant: $\text{TOA}_{1A} - \text{TOA}_{2A} = \text{TOA}_{1B} - \text{TOA}_{2B}$. These assumptions combined with 5-48 gives:

$$TD_A - TD_B = 2 \cdot \left( \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha - \delta_{T1})} \right) - \arctan \left( \frac{-10^{ES/20}}{\cos(\alpha - \delta_{T2})} \right) \right)$$

(5-49)

If the measurements are taken near the baseline between the two transmitters, $\delta_{T1} \approx -\delta_{T2}$, and if the antenna is optimally aligned towards the transmitters ($\alpha - \delta_{T1} = 0$) then 5-49 further reduces to:

$$TD_A - TD_B = 4 \cdot \arctan \left( -10^{ES/20} \right)$$

(5-50)

$$ES = 20 \cdot \log_{10} \left( \tan \left( \frac{TD_A - TD_B}{4} \right) \right)$$

(5-51)

Figure 5-23 depicts an example of measured and simulated TD-errors caused by E-field susceptibility.

![Figure 5-23: Measured and simulated TD-error caused by E-field susceptibility. The red and blue markers depict the actual Sylt-Lessay TD measurements of an unshielded, single loop antenna. The black line is the result of a simulation assuming an E-field susceptibility $ES$ of $-28.7$ dB. Minimizing E-field susceptibility is a crucial part of the H-field antenna design process. Various mitigation methods, such as balancing and shielding, are outlined in the following Sections.](image)
5.5.2 E-field susceptibility compensation

Theoretically, it is possible for the receiver to calculate and compensate for the error introduced by the E-field susceptibility if:

– the E-field susceptibility $ES$ is known from factory calibration (for example),
– the antenna orientation with respect to the received signal is known, and
– far-field is assumed.

However, it is not trivial that $ES$, the parasitic E-field susceptibility of the H-field antenna, is a constant factor that can be measured once during factory calibration and thereafter used indefinitely. It is also doubtful whether the far-field relation can be assumed in, for example, urban environments or when the antenna is mounted on a steel boot or close to various metal objects, for example.

Some of these issues can be overcome by measuring the E-field separately using a dedicated probe. As long as the E-field susceptibility of the H-field antenna is constant, it is then possible to mix the received signal from the probe with the signal received from the H-field antenna in such a way that the E-field component cancels out. However, both the sensitivity of an E-field antenna and the E-field susceptibility of an H-field antenna are sensitive to the grounding of the measurement setup. A change in this antenna ground might have a different influence on the E-field sensitivity of the probe versus the E-field susceptibility of the H-field antenna, and hence, deteriorate the E-field cancellation.

In static situations, it is possible to measure and cancel the E-field influence by physically rotating the antenna as explained in the previous Section. However potentially very accurate, practical usage is doubtful for the majority of applications.

5.5.3 Balancing

As schematically outlined in Figure 5-24-left the E-field coupling is primarily common-mode; whereas, the H-field coupling is differential mode. An amplifier with common mode-rejection, as outlined in Figure 5-24-right, will cancel the common-mode E-field component.
The amplifier can never cancel the differential mode E-field component, because the differential mode E-field component is indistinguishable from the desired (differential) H-field component.

The most straightforward method of building a symmetrical amplifier is by using a transformer. Figure 5-24-right shows this approach.

In Figure 5-25, the transformer is replaced by its equivalent circuit, where $\alpha L$ is the primary induction, $k$ the coupling factor and $n$ the turn-ratio. When the transformer coupling is assumed ideal ($k=1$), and the transformer losses are neglected, the circuit simplifies to Figure 5-26:

$$\Delta \text{SNR} = 20 \log \frac{\alpha}{\alpha + 1} \quad [\text{dB}]$$

Noise matching between amplifier and sensor is assumed. From 5-51, it is concluded that the transformer inductance should be chosen significantly larger than the antenna inductance to prevent degradation of the antenna noise performance.
In modern miniature antenna designs, a relative bulky and expensive transformer is undesired. As schematically shown in Figure 5-27, it is also possible to build a symmetrical amplifier without using a transformer. Because such an amplifier has two separate input stages, the total noise power produced by the input stages is double the amount of that of an asymmetrical amplifier. This leads to degradation of the signal to noise ratio of 3 dB compared to the use of an asymmetrical amplifier. To achieve sufficient common-mode rejection, special care should be taken to match both amplifier input stages as well as possible. This requirement is essentially automatically fulfilled in an IC design, but can be challenging for a design consisting of discrete components.

**Figure 5-27: Symmetrical amplifier**

5.5.4 Shielding

Figure 5-28 shows schematically the concept of shielding.

**Figure 5-28: E-field susceptibility rejection by shielding**

The external E-field couples directly with the shielding around the antenna and is shorted to ground. No potential difference is possible across an infinitely conducting material. For an incident E-field, the only way to comply with this boundary condition is total reflection. Unfortunately, the incident H-field will induce currents in the shield causing reflection of the
H-field component as well. Therefore, if the antenna is hermetically shielded with an infinite conducting material, all the incident EM-waves are reflected and the actual antenna is effectively useless in its Faraday cage.

Fortunately, it is possible to build a shielding structure that attenuates E-field, but leaves the H-field unharmed. Induced currents in the conductive material cause H-field attenuation. Although currents can only flow in a closed loop, the H-field attenuation is minimized if the shield is constructed in such a way that closed loops are prevented. The author has successfully built a prototype antenna based on this principle, which is shown in Figure 5-29. The comb-structure of the shielding plates shorts the E-field while preventing closed loops, thereby leaving the H-field component unharmed. The prototype antenna has an E-field susceptibility better than -55 dB, which results in a maximum TOA error of approximately 4 ns (1.2 m pseudorange) at the worst-case 45° incidence.

![Figure 5-29: Reelektronika / TU-Delft prototype Loran-C H-field antenna with GPS patch](image)

### 5.5.5 Rod construction

Figure 5-30 shows two methods of reducing E-field pickup by the ferrite-loaded loop: First, choosing a conducting ferrite material for the rod and connecting this rod to ground can reduce the potential difference over the rod. Secondly, as outlined in Figure 5-30, any remaining potential difference can be significantly reduced by using two sets of turns for the coil. One set consists of left-hand, the other of right-hand turns. The output voltage due to H-field \( V_{HB} \) is inverted with respect to \( V_{HA} \) due to the different orientation of the coil. In contrast, the output voltages caused by E-field pickup, \( V_{EA} \) and \( V_{EB} \), have the same polarity. By
connecting A and B anti-parallel as outlined in Figure 5-30, the E-field component is effectively shorted; whereas, the H-field component remains unharmed.

![Diagram of cancelling of E-field pickup in the ferrite loaded loop](image)

**Figure 5-30: Cancelling of E-field pickup in the ferrite loaded loop**

### 5.6 Tuning errors in a resonant H-field antenna

The phase response of a resonant H-field antenna is dominantly determined by the bandwidth and tuning of the resonance circuit and the phase response of the associated amplifier. The tuning of a resonant H-field antenna is function of the impedance (inductance) of the ferrite-loaded coil, the tuning-capacitor, and the (complex) input impedance of the amplifier. All these components are susceptible to temperature and, to some extent, aging.

In the following analysis, an inductance of approximately 2.5 mH and a capacitance of approximately 1 nF is assumed, which will cause resonance at approximately 100 kHz. The example resonance circuit has a Q of 3:

\[
\begin{align*}
L_0 &\approx 2.5 \text{ mH} \\
C_0 &\approx 1 \text{ nF} \\
f_0 &\approx 100 \text{ kHz} \\
Q &\approx 3 
\end{align*}
\]  

(5-53)

A one Hz change in resonance frequency will lead to a change in time of arrival (at 100 kHz) of:

\[
\frac{\delta \text{TOA}}{\delta f} \approx 0.1 \text{ ns/Hz}
\]  

(5-54)

#### 5.6.1 Temperature coefficient of tuning capacitance

The relation between the resonant frequency, inductance, and capacitance is given by equation 5-11 and is repeated here:
Given the parameters mentioned in 5-53, it can be calculated that the change in resonance frequency as a function of a change in capacitance, $\delta f / \delta C$, is approximately 50 Hz/pF. Using 5-54 this leads to:

$$\frac{\delta \text{TOA}}{\delta C} \approx 5 \text{ ns/pF}$$ (5-56)

Modern high quality capacitors with a low temperature coefficient (e.g. NP0-types) can relatively easy provide temperature stability better than 100 ppm (parts per million). 100 ppm results in a change in TOA as a function of a change in temperature, $\delta \text{TOA} / \delta T$, of 0.5 ns/°C. An operational temperature range of -20°C - +60°C, for example, leads to a peak-to-peak TOA fluctuation of 40 ns as a result of the 100-ppm capacitor.

### 5.6.2 Temperature coefficient of antenna inductance

The relation between antenna inductance and resonance frequency $\delta f / \delta L$ for the parameters outlined in 5-53 is approximately 20 Hz/μH. Using 5-54 this leads to:

$$\frac{\delta \text{TOA}}{\delta L} \approx 2 \text{ ns/μH}$$ (5-57)

However, the relation between temperature and the actual antenna inductance is not as straightforward as that for temperature versus capacitance.
The ferrite datasheet usually provides the relationship between $\mu_i$ and temperature. Figure 5-31 shows this relationship for two types of Ferroxcube ferrites [2].

As shown in equation 5-2, the inductance $L$ does not directly depend on $\mu_i$ but on $\mu_{rod}$. The relationship between these two parameters is a function of the length/diameter ratio of the rod as shown in Figure 5-4. Figure 5-32 is derived from Figure 5-4 for a rod length/diameter ratio of 10. This Figure clearly shows that for a relative large $\mu_i$, e.g. $\mu_i > 1000$, $\mu_{rod}$ hardly changes with a changing $\mu_i$. Therefore, the temperature stability of the ferrite has only moderate influence on the resonance frequency, and hence, on the TOA.

Table 5-3 shows the temperature stability of an antenna based on a “standard” Ferroxcube 3F3 ferrite and that of an antenna based on the Ferroxcube 3B7 material, the latter specially designed for (temperature-stable) tuning circuits.

**Table 5-3: Influence of temperature on inductance, tuning frequency and TOA for the Ferroxcube 3F3 and 3B7 10x100 ferrite rod. An initial inductance of 2.5 mH and capacitance of 1 nF is assumed.**

<table>
<thead>
<tr>
<th>temp</th>
<th>$\mu_i$</th>
<th>$\mu_{rod}$</th>
<th>$\delta f_0$</th>
<th>$\delta$TOA</th>
<th>$\mu_i$</th>
<th>$\mu_{rod}$</th>
<th>$\delta f_0$</th>
<th>$\delta$TOA</th>
</tr>
</thead>
<tbody>
<tr>
<td>-25°C</td>
<td>1400</td>
<td>66.80</td>
<td>+280 Hz</td>
<td>+28 ns</td>
<td>1950</td>
<td>67.15</td>
<td>+60 Hz</td>
<td>+6 ns</td>
</tr>
<tr>
<td>25°C</td>
<td>2000</td>
<td>67.17</td>
<td>0 Hz</td>
<td>0 ns</td>
<td>2350</td>
<td>67.23</td>
<td>0 Hz</td>
<td>0 ns</td>
</tr>
<tr>
<td>75°C</td>
<td>3000</td>
<td>67.23</td>
<td>-45 Hz</td>
<td>-4.5 ns</td>
<td>2300</td>
<td>67.22</td>
<td>+7.4 Hz</td>
<td>+0.7 ns</td>
</tr>
</tbody>
</table>
From the calculated temperature dependencies, it becomes clear that the temperature stability of the material is not of primary concern when choosing a ferrite material for a LF antenna. Therefore, it makes more sense to shift the focus to the ferrite losses, and hence, noise performance of the material.

### 5.6.3 Temperature coefficient of the active part

The FT of transistors and Gain-Bandwidth (GB) product of Operational Amplifiers (opamps) are temperature dependent. When active components with limited bandwidth (e.g., 10 MHz) are used to achieve significant amplification (e.g., 50x), then it is plausible that these active components cause significant temperature-related phase shifts. Simulation programs such as SPICE can provide quick and fairly accurate assessments on these errors.

### 5.6.4 Measurement and calibration of tuning errors

A convenient way of measuring the tuning errors is by injecting a reference signal into the loops, which can be done in two ways: by injecting an H-field signal into the antenna or by injecting a current into the resonant circuit. The latter method will be explored first.

Figure 5-33 shows the Norton-Parvin equivalent of Figure 5-7. Usually, $R_{\text{loss}}$ is very small compared to $j\omega L$ and is, therefore, ignored in the remainder of this analysis. The combination of 5-1 and 5-2 shows that the H-field induced current is frequency-independent:

$$\frac{E_s}{j\omega L + R_{\text{loss}}} = F_4\mu_b\mu_{ro}A\omega H$$

Therefore, it makes sense to inject a reference-current directly into the resonance loop for calibration purposes, as shown in Figure 5-34.

The easiest way of implementing the reference current-source is by using a voltage source combined with a large series resistor ($R_{\text{calibrate}} \gg R_L$) as shown in Figure 5-35.

Unfortunately, the hardware configuration presented is too simple to be perfect. Parasitic capacitance over $R_{\text{calibrate}}$, for example, can cause a significant phase shift of the reference
current. Also, the exclusion of $R_{\text{loss}}$ from the equation does not completely reflect the physical reality.

Various experiments have proven that the described method is efficient and useful to measure the tuning fluctuations of the H-field loops, but less suitable to measure the absolute tuning difference between two receiving loops. Another method of implementing the injection of a reference signal is by using an H-field transmitting loop inside the antenna. A high degree of channel equality can be achieved when this transmitting loop is oriented at 45° with respect to both receiving loops. However, it is not trivial to prevent parasitic coupling between the injection loop and (for example) the antenna electronics.

**Figure 5-34:** Injection of a reference current into the resonance circuit

**Figure 5-35:** Implementation of current source by a single resistor

**Figure 5-36:** Influence of temperature on the phase of an H-field antenna
Figure 5-36 shows the effect of feedback compensation on the differential tuning errors between two loops. For this experiment, two single loop antennas were placed approximately 10 meters apart. Both loops were simultaneously processed using the same receiver. The blue line depicts the difference in TOA of Sylt (7499M, at approximately 500 km distance) measured by the two loops. At the start of the measurement, both antennas were in the shade and had roughly the same temperature. Soon after the experiment was started, one of the antennas caught some direct sunlight and started heating up.

After 40 minutes, the temperature difference between both antennas was at its peak resulting in a phase difference of 20 ns between both antennas. After approximately 30 minutes, the second antenna also started to heat up resulting in a declining temperature difference, and hence, a declining phase difference between both antennas. During the experiment, a simulated Loran-C signal was injected in the resonance circuitry of both antennas. These signals were processed in the same manner as the signals received from Sylt. Calibration was performed by subtracting the measured simulator TOA from the “real” Loran-C signals resulting in the red line plotted in Figure 5-36. This experiment clearly shows the potential of temperature correction by signal injection.

5.6.5 Common-mode versus differential-mode TOA errors

Usually, two or more ferrite-loaded loops are used to obtain an omni-directional H-field antenna. If the tuning errors, and hence, phase errors or TOA errors are common on both loops, then the error is common for all received transmitters and cancels out in the position solution. Therefore, when a receiver is only used for positioning, tuning errors are not an issue as long as they are completely common on both loops. Naturally, timing receivers require absolute time of arrival and are, therefore, susceptible to all (common and differential) tuning errors.

![Figure 5-37: Common-mode versus differential-mode antenna tuning errors (Tampa Bay, April 2004)](image-url)
Figure 5-37 shows the measured simulator TOA of a dual-loop H-field antenna during a sea trial in Tampa Bay, Florida (Section 6.4). The absolute TOA error, due to changes in antenna-phase response, is relatively large (55 ns, which equals 16.5 m pseudorange). Without calibration, these errors are too large for most ASF survey or monitoring purposes. The differential TOA error, however, is almost insignificant: 5 ns at most, resulting in a maximum pseudorange error of 1.5 meters for the 100 kHz Loran-C signals.

It can be concluded that the most efficient way of preventing positioning errors due to antenna tuning fluctuations is by forcing these errors to be common mode. This can be achieved by thermally coupling the two or more ferrite loaded loops and associated amplifiers. If the absolute TOA of an individual station is required, for an ASF measurement campaign, for example, then it is beneficial to inject a simulator signal into the antenna for compensation. This feedback-compensation method also provides the possibility to calibrate all the other unknown (analog) delays caused by filtering, for example.

5.7 Obtaining an omni-directional radiation pattern

A single H-field loop antenna does not have an omni-directional radiation pattern. The radiation pattern of an air loop is ideally a perfect figure-8; whereas, the ferrite-loaded loop shows slight deviation from this pattern:

\[ A = \cos(\varphi) \]
\[ B = \sin(\varphi) \]  \hspace{1cm} (5-59)

Due to the directional radiation pattern, a single H-field loop is usually not sufficient for a practical LF radio navigation receiver. This problem can be overcome by using two orthogonally placed loops as shown in Figure 5-39. The loop transfer is given by:

\[ A = \cos(\varphi) \]
\[ B = \sin(\varphi) \]
The two loops can be combined either by quadrature or linear addition. Both methods will be discussed in the next Sections.

### 5.7.1 Quadrature addition

In quadrature addition, loops A and B are added as \( A + jB \), where \( j = \sqrt{-1} \). This addition can be implemented in either hardware or software. The advantage of addition in hardware is that only a single-channel receiver is required to process the antenna signals. Hardware addition can be achieved, for example, by cascading a phase shifter of 90° after one of the loops followed by a linear adder that combines both loops.
Given a perfect figure-8 radiation pattern of a single loop, the quadrature addition provides a perfect omni-directional gain and a phase response that is equal to the angle of incidence $\varphi$:

$$R = \cos(\varphi) + j \sin(\varphi)$$  \hspace{1cm} (5-60)

A quadrature H-field antenna is ideal for LF data reception of, for example, Eurofix or DGPS radio beacons. For these applications, the absolute phase of the signal is irrelevant. As long as the quadrature H-field antenna is not rotated too fast, the data modulation is virtually immune to the heading-dependent phase response.

Some extra processing is needed to make the quadrature antenna suitable for radio-navigation purposes. For example, when a Time-Difference or TD between a Loran-C master station at $\varphi=0^\circ$ and a secondary station at $\varphi=50^\circ$ is measured, an error of $10 \mu s \times 50/360 = 1.389 \mu s$ occurs (417 m). In an ideal situation, this TD error is completely predictable and can be easily compensated for when roughly the position of the receiver, and hence, the relative directions towards all transmitters are known. Unfortunately, in practice, situations can seldom be characterized as ideal:

### 5.7.1.1 Imperfection of the figure-8 radiation pattern

As qualitatively shown in Figure 5-38, a ferrite-loaded loop does not have a perfect figure-8 radiation pattern. Fortunately, this imperfection is small enough to cause only minor range errors when using quadrature addition.

### 5.7.1.2 Misalignment of and cross-coupling between the antennas

In a low cost production process, the two supposedly orthogonally placed ferrite rods are, in practice, seldom exactly orthogonal. This directly translates into an error in the amplitude and phase response of the quadrature antenna. Most crossed-loop H-field antennas also suffer from some capacitive and inductive cross coupling between the two loops (see also Section 5.8). When the two loops are treated independently by the receiver, it is possible to measure this cross-talk or “Xtalk” and also the misalignment of the ferrites and compensate for the effects. A quadrature H-field antenna and associated single-channel receiver does not have this luxury, and therefore, both misalignment and capacitive and inductive crosstalk will lead directly to a positioning error.

### 5.7.1.3 Tilting of the antenna

Often, the rods are not entirely oriented in a horizontal plane. When $\alpha$ is the tilt for rod A and $\beta$ the tilt for B, the radiation pattern $R$ will be:

$$R = \cos(\alpha) \cdot \cos(\varphi) + j \cdot \cos(\beta) \cdot \sin(\varphi)$$  \hspace{1cm} (5-61)

When $\alpha$ and $\beta$ are unknown, this will lead to a phase and amplitude error, dependent on the direction of signal arrival. An example for $\alpha=0^\circ$ and $\beta=30^\circ$ tilt is given in Figure 5-41.
The effects of tilting can be compensated for if the amount of tilting is known from additional sensors. Also, when the tilting is dominantly in the same direction, for example, in a car where you can climb or descend a hill, but not often roll over the diagonal axis, the effects of tilting are cancelled when the ferrites are oriented $\pm 45^\circ$ with respect to the dominant tilt-direction. Depending on the application, the effect of tilting might be negligible. The worst-case TD error given a $30^\circ$ tilt is $0.23 \mu s$, which equals a 69 m pseudorange error. For a $10^\circ$ tilt, however, the worst case TD error is only $0.025 \mu s$ (or 7-meter pseudorange).

### 5.7.1.4 Re-radiation

A quadrature H-field antenna can only be used for positioning when the relative directions of the signals with respect to each other are known. This is the case when the antenna position is roughly known and local objects do not hinder the radio-signals. When local objects are present, in urban environments (for example), re-radiation will influence the received amplitude, phase and direction of the radio-signals. Usually, only the effect of re-radiation on the signal phase affects the position solution. However, when quadrature is used, an “error” in signal direction directly translates into an error in signal phase, and hence, position. Therefore, it is likely that a quadrature H-field antenna is extra affected by re-radiation from local objects.

### 5.7.1.5 ECD measurements

A Loran-C receiver is supposed to track the 3$^{rd}$ zero crossing of a Loran-C pulse to determine the time of arrival (Section 3.4.1). The specific shape of the envelope of the pulse is used to find this 3$^{rd}$ zero crossing. The relation between the shape of the envelope and the location of the
$3^\text{rd}$ zero crossing is denoted as the “Envelope-to-Cycle Discrepancy” or ECD. With a quadrature H-field antenna of unknown orientation, the relation between phase and envelope is lost. This severely affects the ability of a Loran-C receiver to find the right cycle and might pose significant challenges for positioning applications.

The quadrature H-field antenna is a very appealing concept; whereas, only a single channel receiver is needed to process the signals. However, the quadrature H-field antenna is also hindered by various error sources that are troublesome or even for which it is impossible to compensate. Whether the introduced errors by the quadrature H-field antenna are acceptable or not depends on the application.

### 5.7.2 Linear addition

With linear addition, the rods are combined as:

$$\cos(\alpha) \cdot \text{loop}1 + \sin(\alpha) \cdot \text{loop}2$$  \hspace{1cm} (5-62)

The signals of the two antennas are combined in software and form a single virtual H-field antenna with a beam direction $\vec{\alpha}$. It is the task of the receiver algorithms to determine the optimal beam direction for each station and create the electronic beams accordingly. In practice, multiple beam directions are needed for multiple stations, and therefore, the signals of the two physical antennas need to be handled separately up to the point that the beam-steering algorithm adds them. For this approach, a dual-channel receiver is needed.

The downside of a dual channel receiver is its increased complexity:

- Two H-field antennas. Both antennas should be (preferably but not necessarily) orthogonally placed (as long as they are not oriented in parallel)
- Two analog front-ends. Both antennas need separate amplification and analog filtering (e.g. anti-aliasing filtering).
- Two Analog to Digital converters
- Double amount of processing power needed for interference cancellation (e.g. bandpass filtering, notches, time-domain filtering, etc.)
- More processing needed for cross-rate canceling, tracking, etc.

However, a dual-channel H-field receiver has also some unique advantages:

- Compass functionality by comparing the received signal strengths of a transmitter on the two loops
- Spatial nulling of one interference source per tracked signal
- Possibility of compensation for miscellaneous H-field antenna related errors

#### 5.7.2.1 Solving the antenna orientation unambiguously

The direction of a LF (Loran-C) radio signal can be calculated by the $\arctan$ of the received signal levels on two orthogonal loops. This method requires the signal direction with an ambiguity of 180°. It is possible to solve the ambiguity by carefully analyzing the Loran-C waveform: the carrier around the $3^\text{rd}$ zero crossing is supposed to go from negative to positive (given a pulse with positive phase-code). A positive-to-negative zero crossing implies a 180°
heading offset, and hence, the antenna ambiguity is solved. In practice, the zero-crossing sign is determined by solving the cycle identification within +-2.5 µs instead of the +-5 µs normally used for E-field antenna operations. The reduction in ECD margin from 5 to 2.5 µs for H-field antenna operation is likely acceptable for the stronger signals coming from nearby stations. However, weak signals from more distant transmitters, propagated over a mostly unknown path, are less forgiving. The reduction in ECD margin for these transmitters might become unacceptable, especially for high-integrity aviation applications where it is crucial to track the correct cycle. Fortunately, it is possible to use the stronger signals to aid the ambiguity resolution of the weaker stations using the following approach:

First determine a coarse position. The accuracy of this position solution is not important as this position is used only to calculate approximate headings towards the transmitters. A rather large position error introduces only a small error in the calculated heading unless the receiver is very close to a transmitter.

Given the coarse position, calculate the directions, with respect to the geographic north, towards all stations.

Next, measure the heading towards all stations with respect to the antenna. This can be done by taking the $\arctan$ of the signal levels received by the two loops.

The antenna orientation (modulo 180°) can now be calculated by taking the difference between the calculated and measured heading of one or more stations. In principle, only one station is needed. The information from multiple stations can be used to smooth the measurements or for RAIM-like integrity purposes.

At this point, the antenna orientation is known with an ambiguity of 180°. The cycle-identification information of the strongest stations can be used to solve this ambiguity. Once the antenna orientation is uniquely solved, it can be tracked by comparing successive heading measurements at short enough time intervals.

The beam direction towards a station is calculated by adding the antenna orientation to the calculated heading towards that station.

From this point on, the remainder of the signal processing has to deal only with a single channel, without antenna ambiguity. The only difference with an E-field receiver (in the far field) is the 90° phase difference between E-field and H-field response: an Envelope-to-Cycle-Discrepancy or ECD of +2.5 µs measured by an E-field receiver corresponds to an ECD of 0 µs for an H-field receiver.

## 5.8 Cross talk

The perfect antenna, without E-field susceptibility, without cross talk and with perfectly orthogonal placed ferrites with perfect figure-8 radiation patterns, amplified by perfect amplifiers, the radiation pattern of a dual-loop H-field antenna would be:
\[
\text{loop1} = G_1 \cos \varphi \\
\text{loop2} = G_2 \sin \varphi
\]

where \( \varphi \) is the direction of the antenna regarding the transmitter, and \( G_1 \) and \( G_2 \) are two complex numbers that discount the absolute gain-response and phase-response of the antenna and associated amplifier.

Unfortunately, enforcing this ideal situation will be either rather costly or perhaps even impossible. Parasitic effects cause the two loops, amplifiers, and wires to be mutually coupled. This undesired effect is called Cross Coupling, Cross Talk or in short: Xtalk. Figure 5-42 shows various causes of cross talk:

- Both antenna coils have some mutual inductive and capacitive coupling.
- The Power Supply Rejection Ratio (PSRR) of the amplifiers is not infinite. If, for example, antenna 1 draws current when amplifying a large input signal, the power supply of Antenna 2 will drop slightly. If the power supply of Antenna 2 is insignificant decoupled, this drop in voltage will cause a change in the output of Antenna 2.
In most applications, the output stages need to deliver some output power to drive the load impedance of the receiver. In case of a high-impedance receiver input, the load impedance will be dominantly capacitive due to the cable capacitance, in case of a low receiver input impedance or characteristic/matched impedance, the load impedance will be predominant resistive. The current through the driver stages, PCB traces, and cable will cause an H-field, which will be picked up by the receiving loops. A well-balanced symmetrical output stage reduces this effect significantly, but it is very challenging to rule out this effect completely.

The wires of the two loops inside the antenna cable will have some amount of mutual capacitive and inductive coupling. Twisted wires are effective in reducing this coupling.

5.8.1 Characterization of cross talk

Most cross-talk related effects are qualitatively relative well understood. However, as far as a quantitative description of these parasitic effects is concerned it is crucial to have an effective characterization and measurement method.

In the following analysis we assume that:

- The antenna itself, without cross-coupling, e-field susceptibility and other errors, has a perfect figure-8 radiation pattern.
- The E-field susceptibility is negligible.
- Only the single-frequency (100 kHz) response is considered, the frequency dependent response within (for example) the wideband Loran-C signal band is ignored.

Given these assumptions, the author concludes that the effect of cross talk on the antenna response can be adequately modeled by equation 5-64. Antenna tuning and gain is modeled by the complex parameters $G_1$ and $G_2$; whereas, the effect of cross talk is represented by the complex parameters $A_{21}$ and $A_{12}$. The real part of the cross-talk-parameters depicts inductive coupling, the imaginary part represents capacitive coupling.

$$
\begin{align*}
\text{loop}1 &= G_1 \left( \cos \phi + A_{21} \cdot \text{loop}2 \right) \\
\text{loop}2 &= G_2 \left( \sin \phi + A_{12} \cdot \text{loop}1 \right)
\end{align*}
$$

Equation 5-64 assumes that the dominant part of the cross coupling happens between the receiving ferrite-loaded loop antennas themselves, and that the cross-coupled signal from, for example, loop 2 to loop 1 is filtered by the resonance circuit of loop 1. This “filtering” is represented by $G_1$ and $G_2$, respectively, and the placement of the brackets in equation 5-64. Initially, it seems plausible that cross talk is reciprocal, in other words, that $A_{21} = A_{12}$. Although this might be the case for cross talk caused by the coupling between the ferrite-loaded loops, it is not the case for (inductive) cross coupling from (for example) an amplifier output to the receiving loops at the input.

5.8.2 Laboratory measurements of cross-talk parameters

Late 2003, the author built a test setup at Delft University of Technologies to measure the heading-dependent response of H-field antennas. The setup, as shown in Figure 5-43, consists of a large transmitting loop with a diameter of 1.6 meters. This size of transmitting loop was
chosen as a compromise between having an area as large as possible in the center of the loop with a relative homogeneous field and having a transmitting loop that is still “portable” and practical in usage.

The transmitting loop itself is a balanced, shielded loop, as shown in Figure 5-44. A balanced, shielded loop effectively reduces the amount of radiation and pickup of electric field.
In the center of the loop is a computer-controlled stepping motor with attached the H-field antenna under test. A Loran-C signal simulator is used to generate a test signal. Both the simulator and the receiver run from the same clock, and are therefore, tightly frequency coupled. Figure 5-45 shows the measurement results of one of the first Reelektronika/TU-Delft prototype H-field antennas.

The blue markers represent the actual phase and amplitude measurements of loop 1, the red markers the measurements of loop 2. The solid blue and red lines show the result of a Matlab© curve fitting of equation 5-64 on the measurements. This curve fitting provided the following cross-talk parameters:

\[
\begin{align*}
G_1 &= 1.00083 - 0.02140i \\
A_{21} &= 0.08672 + 0.00050i \\
G_2 &= 0.99917 + 0.02140i \\
A_{12} &= 0.05553 - 0.01649i
\end{align*}
\]

Figure 5-46 shows the difference between model and simulation. The difference between measured and modeled amplitude is only 1% at maximum, and might be caused by the less-than-perfect figure-8 radiation pattern (due to the influence of ferrite material in the loops). It should also be noted that the stepping motor used in the measurements has a bit of slack, which might have negatively influenced the measurement accuracy. The modeled phase response resembles the measurements very closely. Nearby the null of a loop (within 10°), the

![Balanced shielded loop diagram](image)

**Figure 5-44:** Balanced shielded loop: the transmitting loop is surrounded by a grounded metal shield which is interrupted at the top end to prevent shortening the H-field.

![Amplitude and phase response graph](image)

**Figure 5-45:** Amplitude and phase response of Reelektronika/TU-Delft prototype H-field antenna (blue represents loop 1, red loop 2)
measurements are slightly noisy. Further away from the nulls, the modeled response is well within 5 ns of the measured response.

Figure 5-46: Differences between the measured and modeled amplitude and phase of the H-field antenna (blue represents loop 1, red loop 2)

Over a hundred different H-field antennas have been successfully measured by the measurement method presented here and modeled according to equation 5-64. Therefore, for the remainder of this Chapter, we will assume that using equation 5-64 is an adequate method of describing the behavior of the H-field antenna regarding cross talk, tuning, and gain errors. Note that the E-field susceptibility is not modeled by 5-64. However, because the balanced shielded loop transmits almost exclusively H-field, E-field susceptibility does not pose problem for these factory calibrations.

Figure 5-47 shows four close ups of Figure 5-45 as illustration of the various H-field antenna related errors. The plot in the upper-left corner shows that the mean phase response of loop 2 is significantly lower than that of loop 1. This is probably caused by a difference in resonance frequency between both loops or a difference in the phase-response of the associated amplifier. The difference in tuning can also be seen directly from the imaginary part of the gain-parameters $G_1$ and $G_2$, which are clearly not zero.

In the upper-right plot of Figure 5-48, the phase response of loop 2 is clearly not constant. This can either be caused by E-field susceptibility or by capacitive cross talk. E-field susceptibility can be ruled out by either looking at the 180° symmetry of the plot or because there hardly any E-field present in the measurement setup, so also no E-field to be parasitically picked-up by the H-field antenna. From this, it can be concluded that the decreasing phase of loop 2 in the upper-right plot is caused by capacitive cross talk from loop 1 to loop 2. Loop 1 shows an almost heading-independent phase response, and hence, hardly any capacitive cross talk from loop 2 to loop 1. Parameters $A_{21}$ and $A_{12}$ agree with this assumption: $A_{12}$, the cross talk from
loop 1 to loop 2 shows an imaginary (capacitive) part of approximately 1.6%; whereas, the imaginary part of $A_{21}$ is, with 0.5%, more than 3 times smaller.

The lower-left plot shows that the maximum amplitudes of both loops are very similar. The maximum amplitude value can hardly be influenced by cross talk from the other loop, because that other loop, with orthogonal orientation, is in its null at that point. Therefore, the of the loops’ maximal amplitude responses are directly representative for the absolute value of the gain parameters $G_1$ and $G_2$. The close match of the peaks of both loops is consistent with the similarity of the absolute values of $G_1$ and $G_2$, which are only some tenths of a percent apart.

Finally, the lower-right plot shows that the null of the loops are not exactly at 90 and 180°. This can be caused either due to misalignment of the antenna in the measurement setup, misalignment of the ferrites in the antenna or due to inductive cross-talk. Because these effects are not fully orthogonal, they cannot be distinguished from one another.

Figure 5-47: Effect of tuning, gain differences, capacitive and inductive cross talk on the antenna response
5.8.3 Compensation for cross talk

Assuming equation 5-64 is a valid model for an H-field antenna, then it is straightforward to find the following formula to correct for the H-field antenna errors:

\[
\begin{align*}
\cos \varphi &= \frac{\text{loop}_1}{G_1} - A_{21} \cdot \text{loop}_2 \\
\sin \varphi &= \frac{\text{loop}_2}{G_2} - A_{12} \cdot \text{loop}_1
\end{align*}
\]  

(5-65)

![Corrected antenna response (blue represents loop 1, red loop 2)](image)

**Figure 5-48:** Corrected antenna response (blue represents loop 1, red loop 2)

![TOA errors (left) and compass errors (right) before and after calibration](image)

**Figure 5-49:** TOA errors (left) and compass errors (right) before and after calibration
With formula 5-65, and the correct antenna parameters, it is possible to feed-forward compensate for all the H-field antenna-related errors, except for parasitic E-field susceptibility.

The result of this compensation is shown in Figure 5-48 and in Figure 5-49. The latter shows the result of the antenna corrections, followed by beam steering, exactly what would happen in most receivers. The TOA error reduces from 75 ns before to 6 ns after calibration and the compass error shrinks from approximately 9° peak-to-peak to less than one degree. Clearly, even after calibration, the H-field antenna is still not perfect. The remaining errors are a mix of measurement noise, measurement errors (e.g. slack of the antenna-rotor), and model errors (e.g. the influence of the ferrite material on the radiation pattern is ignored in the model).

### 5.8.4 Field measurements and compensation of cross-talk parameters

Factory calibration, as described in the previous Section, is an effective way to deal with many cross-talk related errors. However, factory calibration also has some drawbacks:

- **Cost:** the production company needs a special setup to perform the calibration, the ambient interference levels need to be within limits, and the calibration takes some time (up to a couple of minutes).
- **Calibration parameters need to be stored somewhere, preferably in some flash-ROM in the antenna itself. If the parameters are stored in the receiver, the receiver-antenna combination needs to remain fixed.**
- **Slight modifications to the antenna, such as a change in cable length, will reduce the effectiveness of the factory calibration.**
- **And perhaps most important of all: the vessel on which the antenna is mounted might also have a significant impact on the antenna behavior. Naturally, it is not possible to compensate for these vessel-related effects by means of factory calibration.**

Ideally, the antenna-receiver combination is calibrated after installation by the end-user. To investigate the possibility of antenna-calibrations in the field, various tests have been performed.

In April 2004, the author, in cooperation with Reelektronika B.V., Delft University of Technologies, and Megapulse Inc., conducted extensive measurements in the Tampa Bay area (Section 6.4). The goal of this measurement campaign was to measure the influence of far-field and near-field effects on the Loran-C H-field and E-field signals.

The measurement setup consisted next to an H-field and an E-field Loran-C receiver of a very good reference system:

- A NovAtel OEM4 GPS/WAAS receiver to provide for a reference position with meter-level accuracy,
- A Koden/CSI Vector-Pro GPS compass with an absolute heading accuracy of 0.5° RMS, and
- An HP 5071A Cesium clock.

At first, the Loran-C H-field antenna was located on the rear end of the ship. However, the high-power diesel engines and the alternator caused interference levels at this location too
high to allow for the desired high-accuracy propagation measurements. After a survey, it was
decided that the best location of the H-field antenna, with respect to noise, was on the bow of
the ship. This location was virtually noise-free but, as can be seen in Figure 5-50, the metal
cabin of the boat is now fully within the radiation pattern of the H-field antenna.

![Figure 5-50: Tampa-Bay measurement setup. The H-field antenna is mounted on the bow of the ship.](image)

To measure the influence of the boat on the antenna, 4 successive circles were sailed, as shown
by the measured GPS track in Figure 5-51.

![Figure 5-51: Measurement track: 4 successive circles with a diameter of approximately 170 meters](image)

Next, the Loran-C measurements were compensated for the movement of the vessel as
measured by the GPS receiver. The Loran-C phase and heading was measured against the
virtually drift-free Cesium clock, the Vector Pro GPS compass continuously provided the
absolute antenna heading.

The closest Loran-C station Yupiter, at 265 km distance, was used as Loran-C reference signal.
Figure 5-52 shows the heading dependent response of the amplitude and phase of Yupiter's
Loran-C signal measured by the H-field antenna mounted on the metal boat.
The influence of the metal structure is clearly visible in the amplitude response of the antenna. Loop 1 is attenuated and also shows some small phase deviation close to the antenna null (at 90 and 270°). The heading-dependent antenna response has been analyzed by using the same curve-fitting algorithm as described in 5.8.2. This provided the following cross-talk parameters:

\[
\begin{align*}
G_1 &= 0.91460 + 0.00555i; \\
A_{21} &= 0.04433 + 0.00792i; \\
G_2 &= 1.08540 - 0.00555i; \\
A_{12} &= 0.05250 + 0.00094i;
\end{align*}
\]

After a feed-forwarding correction using these parameters and equation 5-65, the antenna response shows significant improvement:

\[
\begin{align*}
G_1 &= 0.91460 + 0.00555i; \\
A_{21} &= 0.04433 + 0.00792i; \\
G_2 &= 1.08540 - 0.00555i; \\
A_{12} &= 0.05250 + 0.00094i;
\end{align*}
\]

The uncorrected antenna response (blue=loop 1, red=loop 2) is shown in Figure 5-52. After correction, the response shows significant improvement, as shown in Figure 5-53.
Figure 5-54 shows the antenna response after one-second integrate-and-dump integration and beam steering of the received signal. The difference in gain response shown in Figure 5-53 caused a compass error of more than 12° peak-to-peak. After calibration, this error was reduced to 2.3° peak-to-peak. It should be noted that the remaining error resembles noise more than a deterministic error. The boat did not influence the received phase excessively, because the peak-to-peak phase error before calibration was only 39 ns. Calibration reduces this error to 24 ns peak-to-peak. The phase plot does not show any 180° mirror symmetry and from this it can be concluded that E-field susceptibility was insignificant during these measurements.

The high-frequency oscillations in Figure 5-54 can be explained because not one, but four successive circles were sailed, after which the measurements were sorted regarding the reference heading. Therefore, two successive points in Figure 5-54 might, therefore, relate to two different measurement positions at two different moments. The circumstances during these successive measurements were probably a bit noisy, and not exactly reproducible, resulting in a repeatability of only 5-10 ns.

Next, in post-processing a comparison is made between the measurements with the Tampa-Bay setup using factory-calibration parameters versus field-calibration parameters. Figure 5-55 shows the TOA of various received Loran-C stations while sailing the circles. Two rates coming from the same dual-rated Loran-C “stick” (e.g. Malone, 7980M and 8970W) are grouped together. The TOAs are compensated for the change in position, so ideally, they should remain constant if there is no change in propagation or in the timing of the transmitted signal.
The factory-calibrated antenna response (top plot) clearly shows TOA fluctuations consistent with the sailed circles. These fluctuations are completely removed after applying the correction parameters obtained with the field-calibration (bottom plot). However, other fluctuations are still visible, such as the 20-ns step response of 8970W (top track in the bottom plot). These steps are not caused by the receiving system, however, but by the timing control of the transmitter. The remainder of the TOA-anomalies is assumed to be noise.

**Figure 5-55: TOA measurements after factory calibration (top) and field calibration (bottom)**
Except for the timing community and some scientists, most users are seldom interested in the individual TOA-errors; it is the final positioning error that matters most. The actual positioning error is a function of the individual TOA-errors and the constellation.

Figure 5-57 shows the Loran-C positioning error after factory calibration (left) versus field calibration (right). GPS/WAAS has been used as the position reference. Field-calibration reduces the 95% positioning error slightly with respect to the factory calibration. However, the largest portion of the positioning error is caused by noise and geometry, as can be seen from Figure 5-56. Jupiter and Malone provide an excellent positioning performance in the NW-SE direction; Carolina Beach is too far away to provide adequate accuracy in the NE-SW axis.

**Figure 5-57:** 95% position error (inner solid red circle) and 99% position error (outer dashed red circle) after factory calibration (left) and after field calibration (right)

**Figure 5-56:** Overlay of positioning scatter on geometry map (positioning scatter not on scale)
Whereas the field-calibration only mildly increased positioning accuracy, it reduces the compass error significantly:

With only factory calibration, the Loran-derived compass heading had an offset of 2.8°, a standard deviation of 3.9°, and a 95% error of 8.3° with respect to the Vector Pro GPS compass. After field-calibration, these errors reduced to an offset of -0.1°, a standard deviation of 0.5°, and a 95% heading error of 1.1°, which are acceptable errors for applications such as a wide range of maritime applications, for example.

5.9 Discussion

LF EM fields can be probed with both an E-field and H-field antenna. The simplicity of the E-field antenna made it the sensor of choice for radionavigation applications for decades. The vast amount of available literature combined with more than half a century of operational experience form a comprehensive knowledge base for this antenna type. The major drawback of the E-field antenna is its susceptibility to precipitation static or p-static. This phenomenon causes serious problems, especially for airborne applications during p-static events. Although mitigation is possible to some extent, the solution is most likely the use of an H-field antenna.

The author has spent a considerable part of his Ph.D. research on the H-field antenna, its many error source and potential mitigation techniques. This Chapter has reflected these efforts.

The efficiency of the infinitesimal LF loop antenna is extremely low, even considerably lower than that of an infinitesimal E-field antenna. This low sensitivity requires a low noise antenna design. Section 5.4.4 has revealed a situation rather unique to analog design: the sensor source-impedance can be chosen freely. This is done by changing the number of turns of the
receiving loop allowing an optimal adaptation of the sensor to the associated amplifier. This adaptation process is called noise matching and for the H-field loop antenna allows for an orthogonal design of the passive loop-sensor on one side and the low-noise amplifier on the other side.

The low sensitivity of the H-field antenna poses, next to noise, as well as the threat of parasitic E-field pickup. Fortunately, balancing and/or shielding can overcome this E-field susceptibility. Measurements have proven these methods to be highly effective.

The author has chosen a resonance configuration for the H-field antenna topology. This resonance circuitry is tuned (in contrast to the wide-band setup), which makes it extra susceptible to temperature dependent phase changes. Feedback compensation by measuring the response of an injected simulator signal has shown success in mitigation of this temperature dependency.

Another nuisance of the H-field antenna is its lack of an omni-directional radiation pattern. The figure-8 radiation pattern forces a dual loop configuration to obtain the required omni-directional sensitivity. The two loops can be combined by either quadrature or linear addition. Quadrature addition is ill advised for positioning and timing applications, but is an effective solution for data-only receivers. Conversely, linear addition requires a dual-channel receiver and additional algorithms to electronically beam steer the antennas and solve the 180° phase ambiguity. Fortunately, the methodology explained in Section 5.7 has proven to be a successful implementation to overcome this challenge. The two loops of the dual-loop configuration are likely to suffer from parasitic capacitive and inductive coupling. This coupling causes a heading-dependent phase and gain error, potentially resulting in significant positioning, timing, and compass errors. Section 5.8 introduces a model for these errors and also a successful feed-forward compensation technique. After application of this feed-forward correction, both the cross-talk errors and the tuning-related errors are reduced to a diminishable level. Furthermore, the compensation technique allows mitigation of the influence of the vessel on the antenna radiation pattern as well.

In conclusion, this Chapter has shown the complex challenges of the H-field antenna. Careful analysis has provided novel mitigation methods, for which effectiveness has been proven by the numerous measurements discussed in this Chapter and in Chapter 6. Although perhaps more challenging, the carefully designed H-field antenna has no fundamental performance limitations compared to the E-field sensor. The choice between both antenna types can, therefore, be made based on near-field propagation phenomena, the characteristics of near-field interference and finally on the application-specific demands such as compass functionality and production cost.

5.10 References

[21] "GettingAhead(ing) with eLoran", Erik Johannessen, Megapulse, presented on the 34th Annual Convention and Technical Symposium of the International Loran Association, October 17-19 2005

LF Propagation Measurements

The renewed interest in Loran-C or “eLoran” requires a new assessment of the current and potential capabilities of the system. What level of performance can be expected if we upgrade the system, transmitters, and receivers, using the most modern technology? This dissertation has presented an elaborate analysis of the many error sources involved in LF radionavigation, as well as various methods available to counteract the consequences of these errors. The theory that has been presented is the result of an iterative process of theoretical analysis, hardware and software developments, and measurements. This Chapter presents the measurement.
6.1 Introduction

Low-frequency radionavigation systems have been on-air since the Second World War. Since then, numerous measurements have been performed and reports have been written detailing the propagation of the low-frequency radio waves under various circumstances. However, these measurements had several drawbacks. First, it is only since the availability of GPS that it is possible to easily and very accurately determine the reference position. Secondly, a highly accurate mobile (absolute) timing source has long been unaffordable. Finally, until recently, computing power was very limited, resulting in only modest tracking performance of radionavigation receivers. These past challenges made it virtually impossible to accurately determine the absolute propagation time of the LF radio waves. Furthermore, practically no measurements were done with H-field antennas; almost all measurements were E-field time-difference measurements.

The renewed interest in Loran-C or “eLoran” requires a new assessment of the current and potential capabilities of the system. What level of performance can be expected if we upgrade the system, transmitters, and receivers, using the most modern technology? This dissertation has presented an elaborate analysis of the many error sources involved in LF radionavigation, as well as various methods available to counteract the consequences of these errors. The theory that has been presented is the result of an iterative process of theoretical analysis, hardware and software developments, and measurements. This Chapter presents the measurement.

First, a measurement campaign in Reeuwijk (the Netherlands) is discussed. Differential techniques are used to very accurately measure highly localized propagation effects. Additionally, a method is demonstrated to successfully monitor the stability of near-field propagation by comparing E-field and H-field.

Secondly, the Boston (USA) measurement campaign is presented. Here both E-field and H-field were measured simultaneously, allowing unparalleled comparison of the influence of local objects, such as a cloverleaf, on E-field and H-field propagation. Measurements along a highway provide a sense of the achievable accuracy and the operational characteristics of the navigation system.

Finally, the Tampa Bay (USA) measurements outline a campaign in which all theory and development converges. Among others, the successful antenna calibration and use of a reference station provides excellent measurement quality. The overall performance lies well within the required Harbor Entrance and Approach requirements of 20 m, 95%.

6.2 Reeuwijk measurements

The objective of the “2003 Reeuwijk-measurements” was two-fold: first to prove the feasibility of conducting very accurate, absolute, and relative Loran-C phase measurements. The second objective was to obtain an initial assessment of the presence of re-radiation and of the
temporal repeatability of that potential re-radiation. The required measurement accuracy imposed an interesting challenge.

6.2.1 Spatial repeatability re-radiation measurements

A commonly used figure of merit of a positioning system is the scatter plot. The radionavigation receiver is located at a fixed location, and the position measurements are subsequently plotted against a reference position or against the mean position, for example. This method shows the temporal stability of the system and reveals, therefore, the influence of signal-to-noise ratio, transmitter geometry (DOP), transmitter stability, and receiver quality, for example. With GNSS, the transmitters move, which causes a continuous changing situation. The influence of any potential multipath, therefore, is quickly noticed in a scatter plot. With Loran, the transmitters are stationary; any potential reflections or re-radiation effects at the measurement site are, therefore, likely to remain constant and they will be invisible in the scatter plot. This different behavior is clearly visible in Figure 6-1, which shows GPS and Loran-C scatters measured by Tele Consult Austria [1]. The measurements were taken close to a stack of metal containers. This metal structure caused severe multipath on GPS, noticeable by the wandering tracks in Figure 6-1-left. Most likely, the metal structure had a significant effect on Loran-C as well. However, this effect is not directly noticeable in Figure 6-1-right. The elliptic form of the Loran-C scatter is caused by a poor Loran-C geometry, and it is not much different from any other Loran-C scatter measured in the Netherlands at that time. Potential re-radiation might have caused a Loran-C position offset, but this is difficult to deduct from this picture, because uncompensated ASFs are also likely to cause significant position offsets (Section 3.4.2).

![Figure 6-1: Scatter plot of GPS (left) and Loran-C (right) at the Rotterdam container terminal. Scale is 25 m/div. (Graphs from reference [1])](image)

A Loran-C scatter plot is a good indication of the noise performance of the receiver, the stability of the transmitter, and the fluctuations of the propagation characteristics. However, local propagation disturbances remain hidden, because they are usually temporally stable. An
interesting experiment would be to move the setup 10 meters and see whether the scatter also changes accordingly. Any discrepancies are most likely caused by local effects. This leads to the introduction of spatial repeatability: the agreement of the change in reported position or time-difference with a (known) change in measurement location. It is also possible to analyze spatial repeatability at a pseudorange level; the theoretical change in Time Of Arrival (TOA) of a station is then compared to the measured change in TOA. The direction of movement is hereby crucial:

First, consider if the user stays perfectly on a transmitter radial: the direction towards the transmitter remains constant, only the distance changes. In this case, the change in propagation path is expected to be influenced only by the ground properties between the initial and final position.

Next, consider the situation when the user moves perfectly perpendicular to the transmitter radial: the distance towards the transmitter remains constant, because the direction constantly changes. The whole propagation path alters in this situation. It is theoretically possible that the propagation path changes from an all-sea path in the initial position to an all-land path in the final position, while moving only over a small distance. Such a radical change in propagation path is a rather unlikely scenario, but we should be aware of potential (smaller) changes when analyzing the measurement results.

6.2.1.1 Measurement setup

The antenna will be moved only over small distances to sample the localized effects. Based on previous experience, the changes in TOA are expected to be small: in the order of tens of nanoseconds. Therefore, the total measurement error was targeted to the sub-10 ns level. To achieve this, it is necessary to cancel any far-field, sky-wave, and transmitter timing effects. Furthermore, we must remove the measurement errors caused by potential cross-rate, receiver timing, and temperature effects. Differential measurement techniques show potential in fulfilling these challenges: instead of measuring absolute TOAs at various locations, differential time differences are measured between a mobile “rover” antenna and a fixed reference antenna. The antenna signals are processed by a Reelektronika dual-channel Loran receiver, which handles both channels identically. This differential measurement technique, as shown in Figure 6-2, cancels most of the errors:

- Far-field propagation effects are assumed similar on both the rover and the reference antenna.
- The sky-wave influence received at the reference antenna is assumed to be mostly equal to that received by the rover antenna. Therefore, we are allowed to track the received signals at the (sky-wave polluted) peak of the pulse, which increases the effective SNR.
- The effect of cross-rate is mostly common for both locations and, therefore, cancels significantly

It would be ideal to have two dual loop H-field antennas: one as reference, and the other as the rover antenna. With such a setup, it is possible to measure both the phase and the direction of signals. However, next to optimally calibrated dual-loop antennas, it would also require a four-channel receiver or two tightly synchronized receivers. This was not feasible at the time of this experiment. Therefore, the approach with two single loops was chosen. Both loops were continuously aligned towards the signal under investigation, thereby providing phase,
but no heading information. The two single-loop H-field antennas were specially designed and built for this experiment. The rover antenna is shown in Figure 6-3. Special care has been taken to minimize phase and gain differences between both antennas. A simulated Loran-C signal, injected in both antenna-loops, gives an accurate compensation for temperature influences (Section 5.6.4).

The whole setup is clocked by a GPS disciplined Rubidium oscillator. The presented double-difference measurement technique does not necessarily need such an excellent clock, but using it allows the analysis of individual TOAs, for example, which in turn enables further integrity control.

**Figure 6-2: Spatial repeatability measurement setup**

The whole setup is clocked by a GPS disciplined Rubidium oscillator. The presented double-difference measurement technique does not necessarily need such an excellent clock, but using it allows the analysis of individual TOAs, for example, which in turn enables further integrity control.

**Figure 6-3: Rover antenna, optimally aligned towards Sylt (35.9°)**
The spatial repeatability measurements took place in March and April 2003 at the Reelektronika headquarters in Reeuwijk, the Netherlands. Figure 6-4 shows the surroundings of the measurement site. Reeuwijk, a rural area surrounded by lakes, is situated almost on the baseline between Sylt (388 km, 35.9°) and Lessay (547 km, 236.6°).

At first sight, there are no significant objects close to the measurement site. Most dominant is the fresh water around, but also the groundwater under the islands. The land portion can be so thin that it is hard to estimate where the land begins for a 100 kHz radio wave. An unknown number of communication lines are under the roads. Power lines on poles line the sides of the roads.

6.2.1.3 Measurement results

On a line exactly towards Sylt (35.9°), six measurement points have been defined: spaced 5 meters apart. Figure 6-5-right shows the situation. A single-loop H-field antenna, at a static location acts as the reference antenna; whereas, another single-loop H-field measurement antenna is moveable along the measurement track. At all times, the measurement antenna remains optimally aligned towards Sylt.

Figure 6-5-left depicts the measurement results. The trajectory, presented by the black dashed line, shows that the rover antenna is moved back and forth during the measurements to monitor the repeatability of the measurements. The blue (upper) line represents the measured differential pseudorange of Sylt, and the (lower) red line depicts the measured differential pseudorange\(^1\) of Lessay. The differential pseudorange of Lessay is corrected by \(\cos(35.9-\)

\(^1\) The TOAs are converted into pseudoranges by assuming free space propagation: 1 ns equals approximately 0.3 m
236.6° = -0.94 to compensate for the fact that the measurement trajectory (35.9°) is not optimally aligned towards Lessay (236.6°).

The symmetry of the graph indicates that the measurement setup is temporally stable. With a 60-second integration time, the pseudorange measurement noise is in the meter range.

A close look at the Sylt-track (blue) reveals that:
- A change of approximately 6 meters in differential pseudorange is measured after a 5-meter physical displacement of the antenna; an error of +20%.
- The measured differential pseudorange remains relatively constant when moving the antenna from the 15-meter to the 20-meter position. This missing step is compensated by a 12-meter step when the rover antenna moves from 20 to 25 meters.

In addition, the Lessay track (red) reveals:
- The total change in differential pseudorange of 20 meters corresponds with a total displacement of 25 meters. Therefore, the average step size seems to be 4 meters instead of 5 meters; which is an error of -20%.
- The measured differential pseudorange remains relatively constant when moving the rover antenna from the 5-meter position (via 10 and 15) to the 20-meter position. When moving from the 20-meter position to the 25-meter position, this difference apparently is compensated with a 10-meter step.

Obviously, the measurements show discrepancies between the antenna displacements and the propagation time of the signals from Lessay and Sylt. There is at least some local propagation effect with different influence at the 15-, 20-, and 25-meter mark. There can be numerous reasons for the observed phenomena, but most likely, it is caused by an item such as a buried power-line or another (metal) object.
To observe the local phenomena around the 15-meter mark more closely, a new measurement is conducted. The measurement now starts at the 15-meter mark of the previous test and extends with a heading perpendicular towards Lessay. Again, six measurement points have been defined, spaced 5 meters apart. The differential pseudorange of Lessay “should” be constant for all measurement points, because the distance towards Lessay does not change. Sylt has a 201° different heading with respect to Lessay, so a change in pseudorange of \( \sin(236.5\,\text{°}-35.9)\times5 = -1.8 \) meter per 5-meter step is expected and is compensated for. Every 5 minutes, the rover antenna is moved 5 meters towards the next measurement point, and after 25 minutes, the antenna is moved back towards the starting point. Figure 6-6-right shows the measurement track and Figure 6-6-left the measurement results.

The plot is fairly symmetrical around the 25-meter point, which suggests good repeatability. The fluctuations within each 5-minute interval are explainable by either measurement noise (noise, interference, cross-rate) or temporal instabilities of the re-radiation effects.

The measurements show relative large changes in TOAs between the measurement points. Sylt shows a top-top differential error of 10 meters, whereas the Lessay-track, with a top-top differential error of 20 meters, is even less stable. Note that the propagation path is different for every measurement point; the total distance towards Lessay remains constant, but the angle changes slightly. However, it is probably safe to assume that the change in far-field propagation is negligible and that the observed phenomena, therefore, are caused by local (re-radiation) effects.

### 6.2.1.4 Discussion

Measured spatial repeatability around various objects, such as large buildings, bridges, tunnels, and power lines may give a more stable assessment on the severity of re-radiation under...
various circumstances. The two measurements shown here indicate that something is going on; however, not what and why exactly. The difference in re-radiation at various locations has been plotted, but the absolute influence of re-radiation remains hidden. For the latter to be revealed, one could start the experiment on the lake, for example, far away from any potential disturbances, followed by moving towards the island while continuously measuring the phase of the various signals. Such an experiment, however, requires an entirely different setup, which was not available at the time.

What caused the observed phase jumps remains speculative. The highly localized phenomena suggest objects very close to the measurement trajectory. There are, for example, some power cables buried on the island; however, it is not known where exactly. The antenna cable itself cannot be ignored completely either, although care has been taken to minimize that influence by aligning the cable orthogonal to the measurement trajectory.

The use of single-loop H-field antennas only allows for measuring phase and amplitude. The direction of the signals can be revealed using a double-loop H-field antenna, but that would require a three-channel receiver or a totally different measurement setup.

It is the author’s opinion that the initial objectives of the measurement run, to demonstrate the feasibility of a highly accurate measurement setup and to give an initial assessment of the re-radiation effects, have been met.

### 6.2.2 Temporal stability re-radiation measurements

The previous Section presented a measurement method to detect re-radiation in the spatial domain. If the existent re-radiation is stable in time, one could theoretically map the location-dependent phase offsets during a measurement campaign and apply these corrections in a user application. In this way, it might be possible to mitigate the near-field effects of a bridge in a harbor area\(^1\), for example. This correction becomes useless or might even be counterproductive if the re-radiation effect changes over time. The same is true for re-radiation at a reference-station, where it is possible to calibrate for re-radiation but only if it is stable. Changing re-radiation conditions at a harbor-reference site, for example, might have severe effects on the integrity of the use of Loran-C for Harbor Entrance and Approach (HEA). Therefore, it is desirable to have the ability to monitor the stability of the near-field conditions.

#### 6.2.2.1 Measurement setup

The temporal stability of re-radiation can be measured by either looking at the direction of the signals or by analyzing the difference between H-field and E-field (Section 4.3.2). The latter has been chosen for this experiment. The measurement setup is largely the same as the one used for the Reeuwijk spatial-repeatability measurements (Figure 6-2), but with the H-field rover antenna replaced by a stationary E-field antenna. The single loop H-field antenna is oriented to receive both Sylt and Lessay optimally, and has been compensated for temperature influences.

---

\(^1\) This method is sometimes referred to as ‘micro-ASF’ correction. The term ASFs here is misleading as ASFs are defined as far-field phenomena. Note also that re-radiation correction potentially causes an ambiguous position solution which is demonstrated in Section 6.4.6.1.
by measuring the antenna response of the injected simulator signal. The E-field antenna did not have a suitable connection for signal-injection; however, fortunately it showed only limited temperature influence.

6.2.2.2 Measurement results

Figure 6-7 shows a 75-hour measurement of the phase difference\(^1\) between the E-field and the H-field signals for both Sylt (7499M) and Lessay (6731M). Measurements of the secondary-rates of both stations (6731Z and 7499X) show the same results; therefore, they are not displayed.

![Figure 6-7: Measurement of the temporal stability of near-field propagation by comparison of E-field and H-field](image)

Both tracks show a great deal of commonality, with exception of the outliers of the Lessay track. The common mode effect is probably caused by a changing response of the (not temperature compensated) E-field antenna. The Sylt track behaves nicely: only nanoseconds variation in the difference between E-field and H-field. The Lessay track, however, shows extreme excursions in a daily repeating pattern.

The excursions on the Lessay track can be caused by neither transmitter nor far-field fluctuations, because those would have only caused a common effect on E-field and H-field, which would have been cancelled in the plot. Fluctuations in the sensor response are also very unlikely, because that would also impact the response of Sylt, which remained clearly unaffected. It is possible that Continuous Wave Interference (CWI) synchronous with the 6731 GRI causes an offset on only Lessay (6731M), but not on (7499M) (Section 3.6.2). However, that hypothesis is disproved by the fact that the dual rates (7499X: Lessay and 6731Z: Sylt) show identical results as those plotted in Figure 6-7.

\(^1\) The phase difference is expressed in ns for the 100 kHz Loran-C signal, 1° phase equals 27.8 ns
A plausible explanation is a changing near field with probably a heading-dependent character that affects the Lessay signal, while leaving the Sylt signal unharmed.

Figure 6-8 shows the E-field and H-field TOAs of Lessay separately, zoomed in at the time span of 11 to 25 hours. The TOAs are measured with respect to a GPS-disciplined Rubidium (Rb) oscillator. The local propagation phenomenon influences the E-field (100 ns steps) much more than the H-field (50 ns steps).

It is not yet clear what causes the changing re-radiation. The daily repeating character of the measured excursions on Lessay suggests a man-made source. Perhaps there is a daily repeating change in the power-grid of the nearby overhead-lines that causes changing re-radiation. However, without further information and measurements that are more elaborate, this reasoning remains highly speculative.

6.2.2.3 Discussion

Figure 6-7 and Figure 6-8 show an interesting, but also very disturbing, changing re-radiation effect. Imagine a reference station equipped with an E-field antenna at this location in Reeuwijk. Users using this reference station would see TOA-errors of 100 ns (30 m pseudorange) on one of the most important stations in the position fix. Such large excursions would clearly endanger the capability to meet the HEA positioning requirement of 20 meters, for example. Naturally, timing users would also be significantly affected by the observed phenomenon if Lessay would be used in the timing solution. Signaling such a hazardous re-radiation condition would allow a user to deselect the affected station from the position or timing solution, thereby maintaining integrity at the potential cost of availability.

The measurement technique described proves to be a relatively simple, but effective tool to monitor the stability of the near-field conditions. Re-radiation can still be hidden under a track...
as stable as that from Sylt. With a calibrated antenna setup, it would be theoretically possible to monitor the 120π far-field relation, and hence, measure the absolute presence of re-radiation. An easier approach would be to measure the direction of the signals using a (calibrated) dual loop H-field antenna as well. Then, any deviation from the calculated headings indicates a disturbed near field. Examples of such re-radiation detection can be found in Section 6.4.6.

The results presented prove the feasibility of the near-field stability detection such as described in Section 4.3.2.

### 6.3 Boston highway measurements

In December 2003, a series of land-mobile tests were conducted in the Boston area (Massachusetts, USA) with the following objectives:

- To assess the severity of re-radiation in land mobile environments
- To build and test a mobile setup capable of measuring propagation phenomena simultaneously in the E-field and H-field domain
- To verify the possibility to calibrate an H-field antenna

![Figure 6-9: Measurement location north-west of Boston, Massachusetts, USA (Map: Google Maps [4])](image-url)
6.3.1 Measurement setup

The challenging objectives required a new measurement setup that was specially designed and built for these tests. Figure 6-10 shows this setup.

For mobile ASF measurements, it is essential to have a position and time reference of undisputed quality. A NovAtel OEM4 GPS-WAAS receiver was used as a positioning reference. This Wide Area Augmentation System (WAAS) enabled GPS receiver is capable of positioning accuracies in the order of one to two meters under normal conditions. Note that under bridges and close to steel constructions, the GPS position becomes unreliable or unavailable. The ASF processing software fills these GPS outages by interpolating the position solutions before and after the outage. The resulting reference position error is, thereby, minimal as long as the vehicle moves in a constant direction and at a constant speed.

An HP5071A Cesium clock provided an ultra-stable frequency source to the measurement setup. According to the specifications, this clock provides a frequency accuracy of 5e-13,
which results in a maximum time-drift of only 43 ns per day without any external steering. Calibration can even improve these results.

The setup was not equipped with absolute time, only with absolute frequency. Therefore, the calculation of absolute ASFs was not possible, only ASF fluctuations during a measurement run could be measured. The ASFs will be plotted as rASFs, or relative ASFs while they are plotted against a relative time reference. Relative ASFs are introduced to prevent confusion with dASFs or differential ASFs, which are calculated relative to the ASFs of one station, usually that with the best SNR. Therefore, with differential ASFs it is not possible to single out the ASF-fluctuations of a single station; whereas, the relative ASFs used in this Chapter do allow for this option.

The H-field setup used a prototype H-field antenna that was built by the author according to the noise analysis and E-field shielding theory presented in Chapter 5.

**Figure 6-11:** Prototype H-field antenna on the test-bench at Delft University of Technology. Left the antenna interior, right the antenna with its E-field shielding in place.

This antenna, shown in Figure 6-11, was tested extensively with respect to cross talk at Delft University. Unfortunately, live tests in Boston did not agree with the results of the prior laboratory tests. Analysis revealed that the discrepancy was caused by the antenna cable, which was close to one of the ferrite rods in the Boston-setup. Therefore, the H-field measurement results could not be compensated using the parameters measured in Delft resulting in heading-dependent errors hindering the cloverleaf analysis (Section 6.3.2.2) and an increased overall H-field positioning error. The maximum phase error due to (uncompensated) cross talk is estimated at 60 ns peak-to-peak. Rotation of the antenna showed an excellent 180° repeatability, which indicates an extremely low E-field susceptibility.

A Loran-C simulator was used to inject a Loran-C signal into both H-field loops for temperature compensation, similar as explained in Section 6.2.1.1.
The analog antenna signals were sampled using the same Reelektronika DataSampler as used in Section 6.2.1.1. The Loran-C signal processing was performed in real-time on a powerful laptop.

Unfortunately, all the measurement equipment in the car (three laptops, a PC, power converter, etc.) produced significant H-field interference, which caused the measurement to be somewhat noisy. The quality of the E-field signals was only modestly influenced.

The E-field setup consisted of a Locus E-field antenna connected to a Reelektronika DataSampler similar to the one used for the H-field setup.

Both the E-field and H-field ASF measurements were displayed in real-time, which allowed for measurements that were more exhaustive once an interesting phenomenon was spotted.

There was no stationary reference station setup available during the Boston measurements. This obstructs us from analyzing and correcting for station timing stability in post-processing.

An unpleasant discovery was the occasional presence of extreme interference caused by the high-voltage to 110 VAC transformers attached to the poles carrying the overhead power lines (Section 3.6.4.1). Often these transformers made Loran-C measurements impossible on local roads in the area around the Megapulse facility (Bedford, MA, see Figure 6-20). Therefore, measurement results from these locations are omitted from this analysis. Fortunately, the highways were free from this interference.

### 6.3.2 Cloverleaf measurements

An interesting example of local propagation phenomena was found on the cloverleaf crossing between the 495 and 93, just south of Lawrence (MA) (see also Figure 6-20). Figure 6-12 shows this cloverleaf and the positioning results of five successive measurement runs at low speeds.

The black line represents the GPS-track, which has been carefully repeated five times. The yellow line shows the Loran-C H-field position solution after correction with one averaged ASF-value per station for the whole cloverleaf. The red line shows the same, but then for E-field positioning.

Ten markers have been placed on the plot: the markers B,C,G, and H are positioned exactly at the under-passes, E and J halfway on the overpasses, and the markers A,D,F, and I indicate interesting measurement results of primarily E-field. The plots on the next pages contain the same markers to facilitate easy cross-reference.

On this cloverleaf, the E-field propagation behaved entirely different from H-field. Therefore, the analysis of both domains will be done separately, after which the results will be compared.
6.3.2.1 E-field

Figure 6-13 shows the signal-strength, phase, and positioning results of the E-field measurements on the cloverleaf. Note that the signal-strength is plotted against an arbitrary reference, the actual “effective length” of the antenna is unknown, and therefore, the absolute field strength is unknown as well. The relative ASF plots are obtained by subtracting the median of the measurements and adding an offset to make the plot more readable.
Figure 6-13: Local effects on the Cloverleaf – E-field, five successive runs. The top plot shows the signal strength, the middle plot the relative phase and the bottom plot the resulting positioning error.
Positioning errors are obtained by using the median ASFs as calibration values. The momentary ASFs minus these median values result in pseudorange errors from which a positioning error can be calculated. This position error is 21.9 m 95% and 51.8 m 99%. The 99% position error is dominated by the influence of the bridge at the B, C, G, and H. The 95% position error is primarily determined by the effects at A and F.

An important measurement quality assessment is obtained by looking at repeatability. Figure 6-14 shows the dual-rate repeatability and the repeatability of the five successive runs for both Nantucket and Caribou.

From Figure 6-14 it is clear that the repeatability is better than 50 ns (approximately 15 meters) for Caribou and a fraction of that for Nantucket, the lack of reference-station corrections is, therefore, not a problem for this analysis. The significant phase and amplitude variations shown in Figure 6-14 are under sampled. At a velocity of approximately 10-17 m/s, and an integration time of 5 sec, an independent measurement is taken only at every 50-85 meters. This is insufficient for an accurate (quantitative) observation around the underpasses of the cloverleaf, for example, which are separated by only 50 meters (the distance between B and C and between G and H). However, the quality of the measurements does allow for a qualitative analysis.

The underpasses (B, C and G, H) show large attenuation of 8-10 dB. Note that the actual attenuation under the underpass, in fact, could have been significantly larger while remaining hidden, due to the averaging over approximately 50 meters. At and around the underpasses large differential mode phase changes are noticeable, which result in large positioning errors of up to 50 meters. The local attenuation and phase change of the E-field can be explained by envisioning the bridge as a vertical conductor (metal construction, metal reinforced concrete pillars) over which no potential difference can exist. The near the earth’s surface, equipotential lines of the vertically polarized E-field are elevated by the conducting structure of the...
bridge, causing local larger E-field gradients, which in turn result in a locally slightly higher field strength (E and J). The opposite effects are experienced under the bridge (B,C, G, and H) (Section 3.5.1).

Other fluctuations in amplitude and phase are seen at the points A,D,F, and I. In contrast to the other markers (B,C,G,H,E, and J), these fluctuations do not coincide with obvious structural elements of the cloverleaf. The first observation is the extreme commonality between the various stations at these points. This becomes especially clear if we overlay the measurement results of all the stations on top of each other as can be seen in Figure 6-15. At E and F for example, dips of up to 700 ns in the phase of all stations have been measured. Because the phase changes at these locations are almost entirely common mode, they result in hardly any positioning error, which can be seen in Figure 6-13, bottom plot. From an engineering point of view, we can stop the analysis at this point, because the user will hardly notice these effects in the positioning solution. From a scientific perspective, however, these mysterious dips are quite intriguing.

The large fluctuations in both amplitude and phase are almost completely common mode for signals coming from three entirely different directions. The question now rises whether this can be caused by propagation phenomena, the measurement method, or both.

The E-field antenna is an infinitesimal monopole that senses the E-field relative to a ground potential. During these measurements, the ground connection of the E-field antenna was connected to the chassis of the measurement vehicle. Because the measurement vehicle has rubber tires, there is not always a clearly defined ground path. It is possible that capacitive coupling between the car and constructions in or under the road cause the ground pattern to change with a different antenna response as a result. This change in antenna response has a common-mode effect on all the received signals. In short: the E-field antenna does not probe the field at a well-defined location, but probes a potential with respect to an arbitrary reference. This makes the radiation pattern of the E-field antenna a function of the antenna itself and its surroundings (Section 3.5.2).

A propagation effect with such a large common mode component is hard to envision. Only when a re-radiator is located directly above or below the receiver, then the direct signal and re-radiated signal will form the same interference pattern for the three stations coming from three different directions. If the re-radiator is situated at a different location, it might be possible to have a similar effect at one location, but this similarity is not likely to be the case over a range of hundreds of meters.

On the aerial photograph of Figure 6-12 there is no evidence of local objects that can explain the effects at A,D,F, and I. The locations are largely symmetrical regarding the cloverleaf itself; therefore, it seems plausible that the cause is related to the cloverleaf construction. Figure 6-16 shows the height profile of the cloverleaf track. The height profile has been estimated from the GPS height measurements. These measurements are unreliable close to and under the bridges. Therefore, a constant height has been assumed at these locations. It is interesting to see that two of the four plotted GPS height-tracks show anomalies exactly at point F.
Although the cloverleaf seems quite symmetrical in form, it is not symmetrical in height: point F is approximately four meters lower than point A. The slopes of the off ramps and on ramps are also different. Unfortunately, at this point there is not enough information for a further deterministic breakdown of the measurement results. Nevertheless, various hypotheses are possible, for example:

– Is there a special construction such as extra steel reinforcement in or under the road to support the on ramps and off ramps? Do these constructions start on the on-ramps at D and E and end on the off-ramps at A and F? If such constructions are present, it is quite possible that the metal conductors interact with the radiation pattern of the E-field antenna.

– Is there a relation between the difference in height at A and F and the difference in the depth of the “ASF-dip” at these points?

**Figure 6-15:** The E-field amplitude (upper plot) and phase responses (lower plot) on the cloverleaf are primarily common-mode for the three transmitters. Therefore, the large excursions result foremost in a receiver clock error, not a position error.
The on ramps and off ramps have a different turn-ratio and slope. Is the construction, therefore, also different resulting in a difference in ASF measurements between A,F and D,I?

At a first glance, the cloverleaf E-field measurements show worrisome results: extreme fluctuations in ASFs without a clear cause. It is the author’s opinion that a momentary change in the radiation pattern of the E-field antenna, which is a function of the physical antenna and its surroundings, is probably to blame for the majority of the fluctuations. A true differential E-field probe must be used to confirm this hypothesis.

Further analysis shows that, around the cloverleaf, the local propagation effects in the E-field have only moderate influence on the positioning accuracy. Only in a very close proximity of the underpasses, within 25 meters, a significant positioning error is noticeable. At all other locations, the positioning error is only about 20 meters.

**6.3.2.2 H-field**

Figure 6-17 shows similar information as Figure 6-13 only then for H-field. The rASF plots of Figure 6-13 and Figure 6-17 have the same vertical scale for comparison. H-field interference from the measurement equipment causes the H-field measurements to be slightly noisier (see Section 6.3.1), and therefore less repeatable, then the E-field results.
NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY

Figure 6.17: Local effects on the cloverleaf – H-field, five successive runs. Top plot shows the signal-strength, the middle plot the relative phase and the bottom plot the resulting positioning error.
As stated in Section 6.3.1, the H-field antenna suffered from a heading-dependent phase error of approximately 60 ns peak-to-peak, which could not be corrected for by using the factory calibration parameters. Because the cross-talk effect is perfectly repeatable, the introduced error remains hidden when looking at the repeatability between the five cloverleaf runs. The heading-dependent errors directly degrade the positioning accuracy, but without exact knowledge of the cross talk, it is not possible to quantify this influence.

Figure 6-17 shows, in contrast to the E-field plot, no common mode phase and amplitude variations. This can be explained by the “figure-8” radiation pattern of the H-field antenna and the required beam steering as explained in Section 3.5.2. Therefore, the resulting H-field re-radiation effects are, per definition, differential mode, and hence, always leading to a positioning error. Furthermore, the H-field antenna is a true differential sensor, probing the field at the location of the antenna. This makes common mode phase-fluctuations as seen with the E-field antenna unlikely.

The largest position errors were measured under the bridge at the markers B, C, G, and H. These locations show significant differential mode phase fluctuations and are primary responsible for the 99% position error of 44.4 meters. The 95% H-field position error is, at 37.4 meters, significantly larger than that of the E-field measurements, which was only 21.9 meters.

The signal-strength from Nantucket is hardly influenced under the bridge; whereas, the other stations, Caribou and Seneca, suffer significantly more from signal attenuation. From these observations, we can derive a hypothesis about the orientation of a conductor modeling the dominant effects of the bridge.

Figure 6-18 shows a 3D representation of the bridge and the H-field lines as they cross under marker J. Assume that the effect of the bridge can be modeled by a conducting ribbon such as shown in the Figure. Field lines parallel to this ribbon remain unaffected; they will not be
attenuated even if they are close to the conductor. Field lines that must cross the conducting ribbon orthogonally will be completely attenuated while no flux can exist on the surface of the conductor (Section 3.5.1). The field lines from Nantucket (5930X, 9960X) are almost parallel to the ribbon, and show little attenuation in the signal-strength plot of Figure 6-17 as well. Caribou (5930M, 9960W) is more orthogonal and suffers from significant attenuation.

But why is the effect different at markers B,C vs. G,H? Especially the phase change of Nantucket is almost non-existent at the G,H underpass; whereas, a change of 200 ns is measured at B,C. Notice that the phase of the E-field signals is more affected at B,C than at G,H as well. There are clearly a number of unknown variables, which will not reveal themselves from the shape of the cloverleaf, which is fairly symmetrical, or from photo material such as shown in Figure 6-19.

6.3.2.3 Discussion

The E-field measurements show excellent repeatability, the H-field measurements are noisier but still adequate for analysis. The lack of a reference station did not significantly impact the quality of the analysis. At various locations, large fluctuations in amplitude and phase of the E-field have been measured. Because these fluctuations are primarily common mode, they hardly affect the positioning accuracy. In spite of the lack of precise construction details of the cloverleaf, various hypotheses have been formed to explain the large E-field fluctuations. Testing these hypotheses requires better knowledge of the construction details of the cloverleaf and an E-field measurement setup with a true differential E-field probe.

Significant (differential mode) phase fluctuations in both E-field and H-field have been measured under and close to the bridge. These fluctuations result in positioning errors of up to 70 meters, but only in very close proximity of an underpass. Beyond 25 meters from an underpass, these effects are not noticeable anymore.
The analysis presented focused primarily on the influence of the cloverleaf on positioning accuracy. Analysis of the potential timing accuracy will lead to entirely different results, because the common mode E-field errors would then severely influence the results.

### 6.3.3 Measurements 495 south

The propagation phenomena along a part of the 495-highway have been measured as an example of what can be expected in land-based situations. As can be seen on Figure 6-20, the measurement track started just south of Lowell, MA.

![Figure 6-20: 495-South track (Map from Google Maps [4])](image)

The 495-south was followed for approximately 35 km to exit 27, east of Bolton, where the car stopped, turned around, and began the return trip along the same route. The driving speed during most parts of the measurement run was approximately 95 km/h. With an averaging time of 5 seconds, this resulted in effectively one uncorrelated measurement per 130 m.
6.3.3.1 Raw ASF measurements

Figure 6-21 depicts the ASFs measured simultaneously with the E-field setup (top plot) and the H-field setup (bottom plot). The ASFs are plotted against the distance on the measurement track. Zero is the starting point, 35 km the turning-point at exit 27, and 70 km back at the starting point. At first glance, the plots show large symmetry around the 35 km point, which indicates that probably the same ASF can be used for both sides of the highway. Next, many rapid fluctuations are seen on both plots. Because the same fluctuations are present on both rates of a single transmitter, it is most likely that they are caused by local propagation effects rather than by noise.

**Figure 6-21:** Measured ASFs on the 495-South. The upper plot shows the E-field, the bottom plot the H-field measurements.
6.3.3.2 Dual-rate difference

A more thorough assessment of the measurement noise is achieved by analyzing the dual-rate difference. The dual-rate difference is obtained by subtracting the measurement results of one rate of a transmitter from the results of the other rate. Receiver-timing and propagation effects and antenna errors are common on both rates, and they will cancel. The resulting difference is determined, therefore, by the transmitter timing (e.g. LPAs and PATCO, see Section 3.3) (local) interference and receiver noise. Note that the resulting difference contains 3 dB more noise than the individual measurements, assuming that the fluctuations on both rates are uncorrelated.

**Figure 6-22:** dual-rate differences for E-field (top plot) and H-field (bottom plot)
new potential of low-frequency radionavigation in the 21st century

Figure 6-22 and Table 6-1 show that the dual-rate difference, and hence, measurement noise of the E-field setup is superior to that of the H-field setup. The dual-rate difference of Nantucket is almost equal for both E-field an H-field, which suggests that transmitter-timing fluctuations will play a dominant role for this nearby station. For the more distant and, therefore, weaker stations (Seneca and Caribou) the receiver noise and (local) interference play a more dominant role. As stated earlier, the measurement setup introduced a-typical interference conditions, especially in the H-field domain. Therefore, we should be careful with translating the H-field measurement noise of this measurement run to a more general case.

### Far-field propagation

In the far field, the E and H-field have a fixed relation of $120\pi \Omega^4$ resulting in the same ASF correction for E-field and H-field receivers. Near-field propagation phenomena distort this relation causing discrepancies between both fields. The 495-track has numerous of these local propagation phenomena as can be seen from the measured ASFs shown in Figure 6-21. In this Section, an attempt is made to filter out the local effects to derive true ASFs, which are far-field phenomena by definition.

First, a 35 km long track has been defined, which is used for both driving directions. Next, all the measurements are positioned on this track and interpolated to derive an ASF value every meter. This interpolated ASF-track is filtered with a moving median filter over 5 km (5000 samples). This method ensures spatial domain filtering over a fixed distance rather than time-domain filtering. Furthermore, the moving median filter removes narrow spikes, but preserves potential step-responses. The filtered ASF values of both rates of a single station have been combined and plotted in Figure 6-23 for both E-field (dashed lines) and H-field (solid lines).

Although E-field and H-field follow largely the same trend there is still significant dissimilarity, which points to a remaining presence of local propagation phenomena.

---

4 Strictly speaking, the $120\pi$ relation between E-field and H-field is only valid in vacuum. The earth’s atmosphere slightly changes this relation.

5 The choice for 5 km filter-length has been based among others on the results depicted in Table 6-3.
Figure 6-24 zooms in on a part of the 495 track, where significant steps in the ASFs were measured with the E-field setup. The markers A-B and C-D on Figure 6-23 and Figure 6-25 depict transition regions and B-C a relative constant ASF-offset. Next to a change in heading of the road, there is no obvious visual clue that coincides with the measured ASF anomalies. Perhaps there is a change in road (support) structure between B and C that interacts with the radiation pattern of the E-field antenna.

Based on Figure 6-23 it is doubtful whether corrections measured with an H-field setup, for example, can be used for E-field-positioning. However, a closer look reveals that the dissimilarities between E-field and H-field seem to have a common-mode behavior. Common-
mode ASF offsets cancel as a receiver clock error; therefore they will not influence positioning accuracy. From a positioning point of view; therefore, we can just as well plot the ASFs relative to the ASFs of the strongest station, for example, which is Nantucket in this case. This method is usually referred to as “differential ASFs” or dASFs and is shown in Figure 6-25.

Figure 6-25 reveals that the discrepancies between E-field and H-field ASFs were, in fact, almost entirely common-mode and will cancel as a clock offset. This, for example, allows the use of the derived H-field ASF corrections for E-field positioning and vice versa, as is shown in Table 6-2.

Table 6-2: H-field vs. E-field ASFs used for E-field positioning:

<table>
<thead>
<tr>
<th>Source of ASF correction</th>
<th>95%</th>
<th>99%</th>
</tr>
</thead>
<tbody>
<tr>
<td>H-field ASF table – 5 km grid</td>
<td>31.6 m</td>
<td>57.2 m</td>
</tr>
<tr>
<td>E-field ASF table – 5 km grid</td>
<td>30.8 m</td>
<td>59.1 m</td>
</tr>
</tbody>
</table>

From Figure 6-25 and Table 6-2, it can be concluded that for this measurement run a moving median filter over 5 km removes virtually all local effects leaving only “far-field ASFs,” which are equally applicable for an H-field as for an E-field receiver. The common mode error existing on the ASFs measured with the E-field setup does not result in a positioning error, but can be removed for analysis and display purposes by calculating the differential ASF (dASF).

6.3.3.4 Effect of ASF corrections on positioning accuracy

A detailed analysis of ASF fluctuations is interesting from a scientific perspective, but the end user is only interested in the resulting positioning degradation and possible correction...
methods. Obviously, it is possible to measure the ASFs and store them in the receiver as a correction table. The positioning accuracy is limited, then, by the repeatability of the system (transmitter timing and propagation) and the spatial frequencies of the ASFs in combination with the number of correction points\(^6\). This approach is simulated by using smoothed measurement data as the ASF correction for the same (unsmoothed) data. For practical reasons, the measurement area is first divided into a grid. In this example, the grid is actually 1D instead of 2D, a line survey instead of a surface survey. For every grid cell, an ASF correction value is calculated by taking the median of all the ASFs of that cell. Then, the user receiver derives its correction by an interpolation (linear) of the values of the closest grid cells.

---

\(^6\) This statement is constraint by the fact that it is not always possible to unambiguously correct local propagation phenomena. See also Section 6.4.6.1.

---

**Figure 6-26:** E-field (top) and H-field (bottom) radial position error after correction with a single ASF value (red line, 17 km ASF grid) and by multiple values with a grid-distance of 5 km (blue line).
Figure 6-26 shows the radial positioning error for E-field and H-field after correction with a single ASF correction point, corresponding to a 17 km correction grid (red line), and that for a correction point every 5 km (blue line). Figure 6-27 depicts the same information displayed as a scatter plot. The circles in the scatter plots represent the 95% positioning error.

Correction with a 5 km ASF-grid removes the low-frequency trend of the radial positioning error, which significantly improves the positioning accuracy, especially for the E-field setup. The H-field positioning clearly suffers from more noisy behavior, which is also clearly visible in Figure 6-27.

![E-field and H-field position scatter plots](image)

**Figure 6-27**: E-field (left) and H-field (right) position scatter plot of the Loran positioning error after correction with a single ASF value (red dots, 17 km ASF grid) and multiple values at a 5 km interval (blue dots). The solid lines depict the 95% positioning errors.

The relatively poor performance of the H-field setup can only be explained partly by the poor dual-rate repeatability, and hence, higher measurement noise. Figure 6-28 shows the excursions of the H-field positioning solutions (yellow) on both sides of the road to be fairly similar. This suggests that the H-field positioning suffers more from local propagation phenomena then E-field positioning.

The 5-km grid-distance used in this analysis has been chosen rather arbitrarily. The influence of grid distance on positioning accuracy has been outlined in Table 6-3. Reduction of the cell-size to 2.5 km does not offer significant performance improvement. Further reduction of grid-size to 250m shows improvement, because now part of the local propagation disturbances is compensated as well. Note that these micro-corrections are still spatially undersampled: the cloverleaf analysis shows that near-field phenomena can be much more localized than 250 meters. However, further reduction of the grid-size has limited validity. The raw ASF measurements provide, on average, only one uncorrelated correction every 130 m due to the high velocity of the measurement vehicle and the 5 s integration time.
6.3.4 Discussion

Measurement setup

The cloverleaf and 495 runs have proven the capability to measure Loran ASFs simultaneously with an E-field and H-field setup. Real-time feedback of the measurement results allowed anticipation on interesting phenomena such as seen at the cloverleaf. The combination of E-field and H-field has proven powerful in the analysis phase. Comparing both fields allows identification of near-field phenomena. Unfortunately, the H-field antenna calibration was not successful due to an interfering cable close to the loops. This instrumentation problem was not fundamental, and it was solved in later measurement campaigns (see Section 6.4.1.4). The
H-field setup was also more susceptible to local interference, especially interference radiated by the numerous power converters and the computer equipment present in the measurement vehicle. The unavailability of reference station corrections did not severely affect the analysis as indicated by the good repeatability of the measurement results.

**ASF corrections**

A grid-size of 5 km has been suggested to compensate for the ASF fluctuations around the 495-south track. With this grid-size, the correction values consist solely of far-field effects, which make them equally applicable for E-field and for H-field receivers (see Figure 6-25). A significant further reduction of grid size provides only limited performance enhancement. Given a grid-cell of 5 km by 5 km, less then 2000 points are needed for the entire Netherlands or 500,000 points for the USA or Europe. If each grid cell contains corrections for 5 transmitters and every correction is 2 bytes, then 5 MB of storage space is sufficient for the whole of Europe. Given these assumptions, the storage-requirements of the ASF corrections become, with today’s available memories, an insignificant issue. Measuring all these points still poses a challenge, however. The combination of propagation modeling with actual measurements is a potential solution for this scale-issue (see also Section 4.4.1).

**E-field or H-field for land-mobile**

In the 495-south and cloverleaf measurement run, the E-field ASFs show rather large fluctuations. Due to the common-mode behavior, few of these fluctuations actually result in a positioning error. The ASF fluctuations measured with the H-field setup are often smaller than that of the E-field setup, but due to its differential nature, H-field propagation anomalies always result in positioning errors (Section 3.5.2). The net result is that the E-field setup actually outperforms the H-field setup on both the cloverleaf and 495-south (Figure 6-27). These measurements, on the other hand, were “out in the open.” In urban canyons, for example, the E-field signal is likely to be more attenuated than the H-field signal. This can result in a situation where a position solution remains possible with an H-field setup; whereas, the E-field setup loses track. Superior availability can be a strong advocate for H-field Loran, especially in the context of its role as a backup-system. Another asset of H-field Loran is the unique compass functionality. The combination of this heading functionality with a rate gyro and an odometer, for example, can create an interesting performance enhancement, regarding both accuracy and availability.

Modern electronics, such as switching power supplies, tend to radiate more H-field interference than E-field interference. This was emphasized by the difference in noise between the E-field and H-field setup, as can be seen in Figure 6-22. From a practical perspective, it should also be noted that an E-field antenna could probably be combined with a car-radio/GSM/GPS antenna making it a cost-effective solution. Conversely, E-field antennas are susceptible to P-static noise (Section 3.6.4.1), which endangers the availability of the system. However, this is less of an issue for land-mobile applications than for aviation, for example.

The choice between E-field and H-field contains many dimensions of which only few have been addressed in this Section. The measurement results of the 495-south and cloverleaf favor the E-field setup, but it is quite probable that other measurements and/or practical considerations swing the pendulum back to the H-field setup.
The Boston measurements focused primarily on the positioning domain. Clearly, for a (mobile) timing user, the decision between E-field and H-field has a different argumentation. Common-mode errors do not cancel in the timing solutions, which severely degrades the E-field timing capabilities.

### 6.4 Tampa Bay measurements

In April 2004, extensive measurements were conducted in Tampa Bay, Florida, USA. To meet the scientific challenges of this campaign, most of the equipment has been specially developed based on the analysis described in Chapters 3, 4, and 5. The accurate measurement results derived allowed for an iterative process of detailed analysis, model, and algorithm development.

Figure 6-29 shows the measurement location and the Loran-C stations used for the measurements.

![Figure 6-29: Loran-C stations used for the 2004 Tampa Bay measurement campaign (Map from Google Maps [4])](image)

The objectives of the Tampa Bay trials were as follows:

- to design, develop, and implement equipment and software that is capable of measuring “absolute” ASFs, simultaneously for H-field and E-field
- to implement a reference station controlled by a very accurate clock. This reference station should allow for correction of propagation and transmitter timing variations
- to compensate for temperature and cross talk in the H-field setup
- to assess the <20m 95% Harbor Entrance and Approach (HEA) requirement
- to analyze re-radiation around structures such as bridges
6.4.1 Measurement setup

The Tampa Bay measurement setup was based on the Boston setup (6.3.1), but with some additions and improvements, which will be further detailed in this Section.

6.4.1.1 Reference station

The reference station consisted of a Locus LRS-IIID Loran monitoring receiver controlled by an undisciplined HP 5071A Cesium (Cs) clock. The antenna was placed on the roof of the headquarters of Si-Tex Marine Electronics, approximately at 27.873 latitude, -82.650 longitude (see Figure 6-38). Because the Cs clock only provided frequency, not time, absolute ASFs could not be measured with this reference setup. Temporal variations in transmitter timing and propagation phenomena were monitored, however, with high quality as is further explained in Section 6.4.2.

6.4.1.2 Research Vessel “Subchaser”

The research vessel “Subchaser,” shown in Figure 6-30, was rented from the University of South Florida’s Department of Marine Science, St. Petersburg.

This 10.7m long vessel provided an ideal platform for the measurements. Its shallow draught allowed measurements close to the shoreline and the research character of the vessel provided easy installation and operation of the measurement equipment.
6.4.1.3 Ground truth

A NovAtel OEM4 GPS-WAAS receiver was used as the positioning truth. Under normal conditions, this GPS receiver provides meter-level positioning accuracy. Under bridges, however, the positioning accuracy has been reported to be significantly off, up to several tens of meters. Examples of such outliers can be found in Section 6.4.6.1, for example in Figure 6-45, where GPS shows a 35 m outlier under the Friendship Trail Bridge.

A Koden (CSI) Vector Pro GPS compass, based on a two-antenna array, provided absolute heading with an accuracy of 0.5° RMS. The Vector pro appeared to have functioned rather well, although it showed some sparse convergence issues and outliers under various bridges.

An HP 5071A Cesium clock provided the frequency ground truth in a similar fashion as described in Section 6.3.1.

6.4.1.4 H-field setup

The Reelektronika H-field setup was mostly similar to that used for the Boston setup (Section 6.3.1): A Reelektronika H-field antenna such as shown in Section 5.5.4 was connected to a Reelektronika DataSampler, clocked by the HP cesium. The H-field antenna was positioned on the bow of the ship, next to the GPS antenna. This location was chosen for its minimal interference levels.

Temperature compensation

A Loran-C signal-simulator, controlled by the HP-cesium clock, provided a calibration signal, which was used to compensate for temperature influences. Figure 6-31, which is also shown in Section 5.6.5, shows the response of the antenna to this signal during one of the measurement runs.

![Figure 6-31: H-field antenna response to the injected simulator signal. The common-mode phase variation yields 55 ns peak-to-peak whereas the difference between the loops only measures a maximum of 5 ns.](image)
The temperature change caused primarily a common mode phase change. This would not have caused a degradation of the positioning accuracy, but correction did provide significant improvement in the repeatability between the multiple ASF runs; therefore, it eased the interpretation and processing of the results.

The simulated signal did not have a fixed relation with UTC-time, only with the frequency standard provided by the Cs clock. Therefore, the simulated signal response could only be used to correct for temperature fluctuations, not to compensate for the absolute impact of the total (temperature dependent) antenna and receiver delays.

**Antenna calibration**

As described in Section 5.8.4, the H-field antenna was calibrated using an on-air Loran-C signal and the references provided by the Cs clock, GPS positioning, and GPS compass. Figure 6-32 shows the heading dependent TOA and compass error before and after calibration.

![Figure 6-32: TOA (left) and compass error (right) before and after calibration](image)

The TOA error before calibration was already rather good with a maximum phase error of less than 20 ns (6-m pseudorange error). The compass error, however, significantly improved from more than 6° to less than 1° error. The large metal pilothouse behind the antenna probably caused the large compass error before calibration. Next to the applied corrections, mitigation of the compass error might also have been possible by rotating the antenna 45° such that both antenna loops were more equally influenced by the metal structure.

**6.4.1.5 E-field setup**

For the E-field setup, a Locus E-field antenna connected to a Reelektronika DataSampler was used. The DataSampler was controlled by the same HP Cs clock.

First, the antenna was mounted on the port side of the boat, approximately 8.5 meters behind the H-field and GPS antenna. In most situations this provided adequate results, but with certain courses and wind directions salt water repeatedly splashed against the antenna. This
caused the antenna to short resulting in extreme attenuation and a largely common-mode phase change. These events were removed from the datasets resulting in a data loss of up to 30%. Later, during the measurement campaign, the antenna was moved to a position central on the pilothouse where salt-water splashes could not reach it.

Compensation of the reference (GPS) position was needed, because the Loran-E field and GPS antenna were not at the same location. The position transformation was performed by rotating the position offset using the heading measurements provided by the Vector Pro GPS compass.

### 6.4.1.6 Time synchronization

The HP cesium provides an ultra stable frequency to the Analog to Digital converters and the Loran simulator. Knowledge of the exact UTC moment of the first sample fixes the time relation of the whole setup. This start moment of the H-field ADC is measured by the NovAtel GPS receiver with a resolution of 49 ns. The E-field ADC was synchronous with the H-field ADC within 78 ns (one period of the 12.8 MHz sample-clock).

For every measurement run, first the corrections from the reference station were applied in post-processing. Next, the H-field setup was compensated for temperature drift. The time offset between the various measurement run was estimated by comparing ASFs measured at the same location. This resulted in the following time offset corrections:

**Table 6-4: Applied time offsets**

<table>
<thead>
<tr>
<th>Applied time offset</th>
<th>H-field</th>
<th>E-field</th>
</tr>
</thead>
<tbody>
<tr>
<td>Tuesday April 13, run 1</td>
<td>0 ns</td>
<td>+10 ns</td>
</tr>
<tr>
<td>Tuesday April 13, run 2</td>
<td>+20 ns</td>
<td>+10 ns</td>
</tr>
<tr>
<td>Wednesday April 14, run 1</td>
<td>+60 ns</td>
<td>+50 ns</td>
</tr>
<tr>
<td>Wednesday April 14, run 2</td>
<td>+15 ns</td>
<td>+20 ns</td>
</tr>
<tr>
<td>Thursday April 15, run 1</td>
<td>+10 ns</td>
<td>0 ns</td>
</tr>
<tr>
<td>Thursday April 15, run 2</td>
<td>0 ns</td>
<td>0 ns</td>
</tr>
</tbody>
</table>

The remaining timing uncertainty is formed by the transmitter timing and the unknown delays through the (analog) antenna, front-end, ADCs, and digital filtering. No attempt has been made to conclusively determine this offset due to the lack of sufficient information regarding the Time-Of-Transmission of the transmitters. Instead, an almost all-seawater path towards Loran-C transmitter Raymondville (west of Tampa) has been used as a reference for a correction common for all stations. After applying all the corrections, there was still a significant difference between the ASFs of two rates of a single station. For Malone, for example, this difference amounted to 480 ns, which can only be caused by faulty transmitter timing. This stresses the inability to measure absolute ASFs without proper transmitter timing control. For the dLoran HEA application, however, not the absolute ASFs, but the repeatability
of the ASFs is the figure of merit that will be proven to be excellent during the Tampa Bay measurements.

6.4.1.7 Dual-rate difference

An assessment of the measurement noise is done by analysis of the dual-rate difference, shown in Figure 6-33. Propagation, receiver clock, and receiver timing have identical effects on both rates of a single transmitter; therefore, they cancel in the dual-rate difference. The effects of transmitter timing variations, such as LPAs and PATCO-noise, have been mitigated by applying corrections from the reference station. The resulting dual-rate difference gives direct insight in the receiver’s interference mitigation capabilities and its tracking quality.

Table 6-5 shows the standard deviation of the dual-rate difference for the three stations.
The dual-rate differences for both Malone and Carolina Beach are practically equal for E-field and H-field. Grangeville, however, shows a much worse performance on the H-field setup then on the E-field setup. This is most likely a receiver processing issue related to cross-rate mitigation. Analysis showed that the 9610Z rate of Grangeville caused the high dual-rate difference.

### 6.4.1.8 Real-time analysis

The ability to assess the quality and analyze the measurements in real time greatly enhances the effectiveness of the measurement campaign. This was already acknowledged during the Boston measurement campaign. For the Tampa Bay measurement campaign, new software was written by the author, which allowed for more detailed real-time analysis.

<table>
<thead>
<tr>
<th>Dual-rate difference</th>
<th>distance</th>
<th>E-field</th>
<th>H-field</th>
</tr>
</thead>
<tbody>
<tr>
<td>Malone (7980M-8970W)</td>
<td>435 km</td>
<td>0.011 µs</td>
<td>0.012 µs</td>
</tr>
<tr>
<td>Grangeville (7980W-9610Z)</td>
<td>862 km</td>
<td>0.019 µs</td>
<td>0.033 µs</td>
</tr>
<tr>
<td>Carolina Beach (7980Z-9960Y)</td>
<td>832 km</td>
<td>0.020 µs</td>
<td>0.019 µs</td>
</tr>
</tbody>
</table>

![Real-time analysis of the measurement results](image)
Figure 6-34 shows the data analysis setup of the Tampa Bay measurement trials. The top monitor shows the current position on an accurate water chart, the left monitor the control of the real-time processing software, the center monitor the real-time analysis control, loran tracking information and H-field heading performance, and the right monitor the E-field (top) and H-field (bottom) ASFs and position scatters. Time-synchronized voice recordings allowed researchers to give comments on the measurement result, for example to mark interesting events and relate them to the particular circumstances of that moment, for example, the influences of a large boat passing nearby.

6.4.1.9 Post processing analysis

Real-time analysis software has proven to be an invaluable guide during measurement campaigns. It provides crucial information about the status of the equipment and the quality of the measurements at the only time it matters: in real-time, when changes can still be made. Immediate qualitative feedback of the measurements allows anticipation to interesting phenomena, making the best out of the measurement campaign. The real analysis, however, is done back home, in a more convenient environment and supported by massive computing power.

Usually, many small software programs are written specifically for the current task and only for the measurement campaign under analysis. After numerous measurement campaigns, it became more efficient to develop general-purpose post-processing analysis software that was able to tackle numerous challenges and was re-usable for multiple measurement campaigns. Figure 6-35 shows a screen dump of such a package, which has been written by the author after his first airborne measurement campaign (December 2005) where altitude formed a new information dimension, which could not be ignored, but troubled the software available at that time. The developed package is capable of addressing many domains simultaneously such as the geographical situation (relative position of the user with respect to such information as coastlines and transmitters, transmitter radials etc), position, height, velocity, heading, GPS quality, reference clock quality, ASFs, ECDs, signal-strengths, re-radiation levels, position domain errors, ASF-correction maps, and so forth. All the visualization is highly configurable and visible across multiple monitors allowing optimal correlation of the information at hand. This software enabled a much more exhausting analysis of the Tampa Bay measurement data, two years after the campaign, resulting in a number of new findings.

Other examples of post-processing software that have been written are as follows:
- the fully automatic ASF-correction map generation software (used to generate the ASF-maps discussed in Section 6.4.4)
- the software to superimpose measurement results in mapping programs such as Google Earth© [5] (used to represent the re-radiation in Section 6.4.6.1)
- the software to read, process, and archive the massive amount of data accompanying a measurement campaign

It is the author’s opinion that these powerful post-processing analysis software packages greatly enhance the efficiency and creativity of the researcher, allowing researchers to gain insights that otherwise might have remained unrevealed.
6.4.2 Reference station analysis

As outlined in Section 6.4.1.1, the reference station was equipped with an E-field antenna on the roof of the Si-Tex building. The top plot of Figure 6-36 shows some of the TOAs measured by the reference station. The frequency offset of the Cesium clock was estimated at -1.4e-13 Hz, which was compensated for in the bottom plot.

The reference station setup does not provide absolute ASFs; it provides only ASF fluctuations. During the whole measurement campaign, the reference station ran continuously, which allows the propagation effects and transmitter timing during that period of time to be “frozen,” and hence, to significantly improve the repeatability of the measurements.

The 24-hours representation shown in the bottom plot of Figure 6-36 reveals a daily repeating character that can be explained by diurnal propagation effects and transmitter timing control. At the time of these measurements the US Loran-C transmitters were still under SAM control (Section 3.3.1) and the timing was steered with discrete 20 ns steps. With SAM control, the Time-Difference of a transmitter pair is held constant at a monitoring (SAM) site, which is

![Figure 6-35: Screen-dump of a post-processing analysis software package written by the author. Left-top shows the measured ASFs, left-middle the re-radiation levels, left-bottom the position domain errors, right the map with the current position, the sailed route and the paths towards the transmitters. The information screen contains all the data available for the selected epoch.](image-url)
usually located in a harbor. The SAM sites controlling the 7980 chain are at several hundred kilometers distance from Tampa Bay. Therefore, the reduction of the diurnal effect at these SAM sites actually introduced extra “diurnal” fluctuations in the Tampa Bay area. This is clearly seen by subtracting the 20 ns steps from Figure 6-36. Malone, the master of the 7980 chain, is not controlled by a SAM site, but is directly synchronized to UTC. The numerous timing steps indicate that the timing control of this station is not implemented ideally, which results in a significant increase in noise of the measurements of this station.

Ideally, one should be able to integrate the reference data for a long period (e.g. 10 minutes) to maximally reduce the amount of noise in the reference measurements. The fluctuations in propagation are assumed to be sufficiently slow to allow for such a long integration time. Unfortunately, the presence of the 20 ns transmitter timing steps limits our ability to smooth the reference measurements. Figure 6-37 shows the effectiveness of the removal of the transmitter timing steps from the measured ASF data of Malone. The reference data has been

**Figure 6-36:** Top plot: TOA fluctuations measured by the reference station. Bottom plot: Estimated Cs frequency offset of -1.4e-13 Hz removed and results plotted against time-of-day.
integrated for 5 minutes, which is rather long considering the obvious transients in the corrections from the reference station (black line). The author estimates that, after smoothing, there is still approximately 5-10 ns noise or error in the reference data, which directly impacts the ASF and eventually positioning accuracy. This will be significantly less with the new Time and Frequency control installed on the US transmitters, which will make SAM control and discrete timing steps something of the past.

**Figure 6-37:** Correction of the ASF measurements of Malone with reference station data. The blue line depicts the uncorrected ASFs as measured while sailing in the Tampa Bay area. The TOA fluctuations measured at the reference site (black line) are used to correct these ASFs for temporal propagation and transmitter-timing fluctuations, which results in the red line. Most of the rapid fluctuations, caused by for example LPAs, have been removed. However, the relative long integration time of the reference data does not allow complete mitigation of the more steep transients.

### 6.4.3 Measurements Tampa Bay area

Figure 6-38 shows the chart with the measurement routes that were sailed between Tuesday 13 April and Thursday 15 April 2004.

A total of 346 km was covered in approximately 13 hours resulting in an average speed of about 26.9 km/h. Every 5 seconds an independent integrate-and-dump measurement was taken resulting in a total of almost 10,000 measurements. The channel starting at the Sunshine Skyway bridge and ending in the Tampa Bay harbor was measured three times allowing assessment of the repeatability of this most important track.
Figure 6-38: Measurement tracks in the Tampa Bay area. The red dot depicts the location of the reference station.

The ASFs of Malone, as measured during the three-day trip, are represented in Figure 6-39.
The red line depicts the track as measured by GPS. The ASFs are displayed both in height and in color. As a reference, the horizontal light-red line is plotted at 50 ns. This plot allows assessment of both qualitative and quantitative aspects of the measurement results. The repeatability between multiple measurement runs is in the 10-20 ns range.

In the following Sections, the Tampa Bay measurement results will be transformed into ASF-correction maps, local effects around bridges will be assessed, and the measurements will be used to simulate the ASF-corrected dLoran performance for a typical harbor approach.

### 6.4.4 Creation of ASF correction maps

The Loran HEA concept relies on a real-time differential correction signal and on the availability of ASF correction maps. The latter can be constructed with modeled data, measurements or a combination of both. The ASF models alone do not deliver the required accuracy for the high accuracy demands of the Harbor Entrance and Approach procedures (see also Section 3.4). Therefore, the author suggests basing the HEA ASF correction maps primarily on measurements that will be further detailed in this Section.

This Section outlines the process of “gridding” of the measurement data, the procedure to find the optimal grid size, and the resulting ASF-maps of the Tampa Bay area.
6.4.4.1 Gridding of ASF measurement data

The author obtained the measurement value of a grid cell of the ASF correction map by the following steps:

- Collect all the measurements within a grid-cell of both rates of a transmitter. Note that any difference between the measurement values of both rates of a single transmitter, caused by transmitter timing control, should be constant after applying dLoran reference-station corrections. This offset is stored as a constant for the entire map.

- Throw away all the measurements with significant re-radiation (the process of re-radiation detection used is further outlined in Section 6.4.6). As stated in Section 3.5.4, it is complex and dangerous to attempt to correct for re-radiation by using a highly detailed ASF correction map. Not only is there the risk of spatial under-sampling, it is also quite possible that the resulting corrected position is not clear (see Section 6.4.6.1).

- Throw away 25% of the extreme values

- Take the weighted average of the remaining measurements within the cell. The measurements close to the center of the cell weigh more than those close to the border of the cell.

- The grid-cell is only valid if the measurements have a reasonable quality, such as enough measurements of good repeatability

This results, for example, in the following ASF-map of Malone:

![ASF measurement results of Malone placed in a 0.005 degrees grid](image)

**Figure 6-40: ASF measurement results of Malone placed in a 0.005 degrees grid**
The receiver gets the ASF correction for its current position by linear interpolation of the ASFs of the three closest grid-cells.

### 6.4.4.2 Effect of ASF-map cell-size on positioning accuracy

With the method described in Section 6.4.4.1, the ASF correction map only contains far-field phenomena; whereas, re-radiation has been excluded. Far-field ASF changes will be relatively slow, thereby making extremely small grid-cells superfluous. The optimal grid size will be a compromise. Smaller grid cells will consume more receiver memory and can be noisier, as well, because fewer measurements are used to determine the value of a single cell. Larger cells create the risk of spatial quantization errors.

Theoretically, the ideal grid size is determined by the Nyquist criterion in the spatial domain. Spatial quantization errors will occur if the ASF variations have spatial frequency components higher than \( 1/(2\times\text{grid-size}) \). Note that for perfect reconstruction assumed by Nyquist, an interpolation filter with infinite length is needed. The receiver, however, applies a linear interpolation between the three closest cells, for example, which makes the parallel with the Nyquist criterion not entirely valid.

The Nyquist approach described may be theoretically sound, but translation into practical steps is not trivial. The author suggests a different, more practical procedure to determine the desired grid size:

- Assemble all the measurements (about 10,000 for this Tampa Bay measurement campaign).
- Exclude those locations with known re-radiation (bridges, other constructions, etc.).
- Create maps with different grid-sizes.
- Correct the measurements with the correction maps based on the same measurements.
- Determine positioning errors based on the ASF-corrected measurements.

Then the resulting positioning error is determined by spatial quantization, measurement errors, and noise.

Figure 6-41 shows the result of the described approach for grid-sizes of 0.0025, 0.005, 0.01, 0.02, 0.03, 0.04, and 0.05 degrees. In the Tampa Bay area, a 0.01° grid corresponds to about 1.11 km (latitude) by 0.986 km (longitude).

Note that these graphs do not reflect the actual Loran positioning performance: it is just the simulated influence of grid-size on positioning performance. A more realistic performance scenario is presented in 6.4.7.

Figure 6-41-right clearly shows that re-radiation essentially masks the effects of the change in grid-size. This is not a surprise, because the ASF-correction map does not contain corrections for re-radiation leaving these extreme effects uncompensated. Re-radiation does not happen too often, but in this campaign it happened regularly enough to dominate the 95 and especially the 99 percentile. The dominating presence of re-radiation is primarily caused because the locations with expected re-radiation were visited frequently and repeatedly at very
low speed to study the effects more carefully. This research-approach is not representative for normal operations in which re-radiation will play a less-dominant role.

The left plot of Figure 6-41 provides us with an interesting correlation between grid-size and positioning accuracy. Because the corrections are based on the same data as the positioning, the situation might occur that local measurement errors impact the correction map. This is especially the case with very small grid-cells of 0.0025° (about 278x247 m), for example, resulting in an unrealistically small positioning error. It is interesting to focus on the E-field positioning error after correction with an H-field correction map and vise versa. In this situation, the positioning performance is only degraded for grid-sizes larger than 0.02° (2.22 km x 1.97 km). Based on this data, a grid size of 0.01° of 0.02 degrees is a logical choice for the Tampa Bay ASF correction map. A grid size of 0.01° will result in about 2500 cells for this area.

Some anomalies are observed when using a grid-size of 0.04 degrees. These effects can be explained by some issues with the E-field measurements (caused by salt-water splashes, see Section 6.4.1.5) that become emphasized by an unfortunate placement of the 0.04° grid-cells.

Note that Figure 6-41 shows that an E-field ASF correction-map can be used for H-field positioning and vise versa. The importance of this observation will be further outlined in Section 6.4.5.

### 6.4.4.3 Generated ASF-maps of Tampa Bay

Figure 6-42 shows the ASF-maps of the Tampa Bay. The maps are based on the gridded measurement data (Section 6.4.4.1) and the radial interpolation/extrapolation technique as outlined in Section 4.4.1. Only the measured and interpolated values are shown, because the
extrapolated values often contain too many anomalies, thereby blurring the view. The yellow arrows depict the direction of propagation. Notice the coastal recovery effects: an increase in ASF is measured close to a land-sea interface. On the same transmitter radial, but slightly further away from the coastline, the ASFs recover to their “normal” value.

Figure 6-42: Generated ASF-maps for Tampa Bay. The grid-size is 0.005°.
6.4.5  E-field versus H-field in the far-field

The far field, E-field, and H-field should have per definition a fixed $120\pi$ relation, and consequently, the same ASF. Therefore, an ASF (far-field) correction map created using E-field data should be just as valid for an E-field as for an H-field receiver.

After exclusion of the locations with known re-radiation (bridges), the Tampa Bay measurement data indeed shows a very high similarity between E-field and H-field.

**Table 6-6: Comparison of E-field and H-field ASF data, re-radiation excluded**

<table>
<thead>
<tr>
<th>Station</th>
<th>Identifier</th>
<th>Mean ASF (µs)</th>
<th>Std ASF (µs)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>H-field</td>
<td>E-field</td>
</tr>
<tr>
<td>Malone</td>
<td>7980M</td>
<td>0.552</td>
<td>0.551</td>
</tr>
<tr>
<td>Malone</td>
<td>8970W</td>
<td>1.033</td>
<td>1.031</td>
</tr>
<tr>
<td>Grangeville</td>
<td>7980W</td>
<td>0.231</td>
<td>0.233</td>
</tr>
<tr>
<td>Grangeville</td>
<td>9610Z</td>
<td>0.480</td>
<td>0.481</td>
</tr>
<tr>
<td>Jupiter</td>
<td>7980Y</td>
<td>0.594</td>
<td>0.585</td>
</tr>
<tr>
<td>Carolina Beach</td>
<td>7980Z</td>
<td>0.176</td>
<td>0.173</td>
</tr>
<tr>
<td>Carolina Beach</td>
<td>9960Y</td>
<td>0.036</td>
<td>0.032</td>
</tr>
</tbody>
</table>

The mean ASFs, as presented in Table 6-6, are practically identical for E-field and H-field. Note that the unknown delays of the E-field and H-field setup have been calibrated such that they produce, on average, the same ASF. This has been done by applying one single offset value for all the measurements of all the stations of a measurement run (Section 6.4.1.6). The agreement between E-field and H-field of the individual stations, therefore, still proves the validity of the far-field relation. Unfortunately, the exact time of transmission of the stations was not known and could not be compensated for. This unknown deviation between the supposed TOT synchronized to UTC and the actual TOT results among others in a difference in “ASFs” between the two rates of a single transmitter.

Another experiment was conducted to validate the agreement between E-field and H-field: the E-field ASF measurements were corrected using H-field measurements on a per epoch basis. So, each independent E-field measurement is corrected by subtracting an independent (time-synchronized) H-field measurement. Any errors in GPS positioning, transmitter timing control, and reference station corrections are common on both receivers; therefore, they will cancel. Correct functioning of the GPS compass heading is crucial, because it is used for the E-field antenna position transformation (see Section 6.4.1.5).

The resulting positioning accuracy after correction is:

---

7 ASFs contain per definition only propagation effects. Two rates of the same transmitter therefore have the same ASF. Any discrepancies between the two rates are either a result of erroneous transmitter timing or a measurement error.
Re-radiation is, per definition, not common for E-field and H-field; therefore, it results in a significant increase in the 95% and 99% errors when the re-radiation is included in the statistics.

Finally, Figure 6-41 is further proof that an ASF correction map based on E-field measurements can be successfully used to correct far-field H-field measurements and vice versa.

### 6.4.6 Re-radiation

The only significant re-radiation apparent in the Tampa Bay measurements was caused by bridges. Besides those locations, the E-field and H-field measurements agreed very well and the measured H-field signal directions matched the calculated values indicating far-field. Figure 6-43 shows the locations of the seven bridges that have been analyzed and Table 6-8 depicts their names.

This Section primarily discusses the impact of the bridges on the Loran-C E-field and H-field positioning performance: the magnitude of the position errors and the minimal distance from the bridge, after which the positioning error falls below the 25-meter alert limit specified by the IMO resolution A.915(22) and A.953(22) (See Appendix B).

In addition, a successful attempt is made to autonomously detect the presence of re-radiation by comparing the measured and calculated signal heading differences between stations, as described in Section 3.5.3. The absolute value of the discrepancies between these two values is averaged for the strongest stations resulting in a re-radiation level. A threshold of 2.5° has been chosen: a re-radiation level below 2.5° is designated as re-radiation free, any measurements with a re-radiation level above 2.5° are most likely contaminated with re-radiation, and they should be discarded or at least treated with suspicion by the user.

Naturally, this re-radiation detection technique only works effectively when the impact of antenna errors and the influence of the vessel on the measured heading are very small. This has been achieved by successfully calibrating the antenna (Section 6.4.1.4). Additionally, tilting and rolling of the ship influences the measured heading, resulting in “noise” in the re-radiation estimation. Note that this re-radiation detection method is based on the assumption that re-radiation influences both the direction and phase of the signal simultaneously. However likely, this is not always the case.

<table>
<thead>
<tr>
<th></th>
<th>Re-radiation excluded</th>
<th>Re-radiation included</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>95%</td>
<td>99%</td>
</tr>
<tr>
<td>E-field vs. H-field</td>
<td>10.4m</td>
<td>14.3m</td>
</tr>
</tbody>
</table>

Table 6-7: E-field positioning performance after correction with H-field on a per-epoch basis
After analyzing the impact of re-radiation, three possible scenarios will be discussed:

- Do nothing, take the full hit of re-radiation, and take that into account in the statistics
- Detect the effects of re-radiation, and alarm the user for potentially faulty position information
- Survey the re-radiation, and build a correction map, which will be applied by the user.

**TABLE 6-8 Bridges shown in Figure 6-43**

<table>
<thead>
<tr>
<th>Id</th>
<th>Name</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>E. Howard Frankland Bridge</td>
</tr>
<tr>
<td>B</td>
<td>Friendship Trail Bridge (Gandy Bridge)</td>
</tr>
<tr>
<td>C</td>
<td>Corey Causeway Bridge</td>
</tr>
<tr>
<td>D</td>
<td>Pinellas Bayway Bridge - Tierra Verda</td>
</tr>
<tr>
<td>E</td>
<td>Pinellas Bayway Bridge - St. Petersburg Beach</td>
</tr>
<tr>
<td>F</td>
<td>Sunshine Skyway</td>
</tr>
<tr>
<td>G</td>
<td>Sunshine Skyway Bridge</td>
</tr>
</tbody>
</table>

**FIGURE 6-43: Bridges analyzed in the Tampa Bay area**
6.4.6.1 Measurement results re-radiation Tampa Bay bridges

The impact of re-radiation of each bridge is presented by superimposing the measurement results in two ways on aerial photographs. The first presentation shows a top view of the bridge with superimposed the GPS positioning (blue), Loran-C E-field positioning (red) and Loran-C H-field positioning (yellow). The thin yellow and red lines depict the Loran positioning error vectors, starting at the GPS position and ending at the Loran-C position. A black track is also plotted along with the distance towards the center of the bridge.

The second presentation depicts a 3D side view of the same data, only now the magnitude of the errors is plotted as height regarding the closest location on the track. These plots also show the result of re-radiation detection, plotted in green. This parameter has been scaled such that a re-radiation level of 2.5° corresponds to 25 m. The thicker black gridline at 25m depicts the 25 m alert limit and the 2.5° re-radiation threshold.

Bridge A: E. Howard Frankland Bridge

Figure 6-44: Bridge A: E. Howard Frankland Bridge (Aerial photographs from Google Earth [5])
The E. Howard Frankland Bridge consists of two fixed bridges with a horizontal clearance of 22.6 meters, and a vertical clearance of 13.4 meters. The width of the bridge, including the space between the two bridges, is 72.0 meters. Note that the left bridge is slightly wider than the right bridge. The GPS ground truth is relatively good close to the bridge; only one outlier of 15 m is noticed at the -35 m marker.

The Loran E-field (red) position error is significantly smaller than the H-field error and has a maximum of 42 m. The user needs to be at least 17 meters away from the bridge for a positioning error of less than 25 meters. The Loran H-field position error (yellow) is far more severe close to the left bridge than to the right bridge and yields a maximum error of 111 m. At 17-meter distance, the H-field positioning error has also decreased to less than 25 m. The asymmetrical behavior is also visible in the output of the re-radiation algorithm (green line), which seems to “miss” the effect of the right bridges thereby leaving two erroneous H-field positions unflagged.

**Bridge B: Friendship Trail Bridge (Gandy Bridge)**

---

Figure 6-45: Bridge B: Friendship Trail Bridge (Gandy Bridge) (Aerial photographs Google Earth [5])
The Friendship Trail Bridge consists of an old bridge, left on the photographs, and two bridges of more modern construction on the right. All three bridges are fixed, and their combined width, including the space between the bridges, is 61.0 m.

GPS shows one significant outlier of about 35-meter magnitude. The four successive passes under the bridges show good repeatability, and again, better results for E-field then for H-field. The maximum E-field position error is 54 m, where H-field suffers from positioning errors up to 104 m. The H-field re-radiation detection covers the entire contaminated area nicely, detecting all H-field position outliers.

**Bridge C: Corey Causeway Bridge**

The Corey Causeway Bridge is a bascule bridge; therefore, it has a relatively low vertical clearance of 7 meters when it is closed. The width of the total structure, 28 meters, combined with the low height, causes significant GPS outliers under the bridge. When under the bridge,
GPS reports positions just outside the bridge, which results in two outliers of around 16 meters. The Loran-E field positioning looks quite good at first sight, on the top-view plot, but a closer inspection shows that it contains significant long-track errors of up to 73 meters. Obviously, long-track errors are operationally of less importance than cross-track errors, but this does not change the fact that this bridge has significant influence on E-field positioning. H-field positioning, in contrast, experiences rather extreme cross-track errors with a maximum of 152 meters. The H-field re-radiation detection works properly successfully identifying all erroneous H-field positions.

**Bridge D: Pinellas Bayway Bridge – St. Petersburg Beach**

Bridge D and E, the Pinellas Bayway Bridges, have a similar structure: a vertical clearance of 6.7 m, a horizontal clearance of 18.3 m, and a width of 12.0 m. The GPS reception of the bridge towards St. Petersburg Beach performs well. Loran-E field positioning is hardly affected with a maximum error of only 19 meters. H-field suffers more from the bridge, resulting in a maximum error of 72 meters. Again, the re-radiation detection catches all faulty H-field positions.
Note that when very close to the bridge, either on the right side or on the left side, the Loran H-field position sensor reports a position to the North (left) of the bridge, right into the shipping route. This phenomenon makes correction by means of an accurate correction map impossible, because there are now at least three physical positions resulting in the same measured position. If a receiver measures this position, what correction should it use then? This dilemma of ambiguity, which can be found in many re-radiation situations, is a strong advocate against re-radiation correction.

**Bridge E: Pinellas Bayway Bridge – Terra Verda**

The Pinellas Bayway Bridge – Terra Verda has the same structure as Bridge D. GPS suffers from a significant error of about 25 m under the bridge, which causes the anomaly in the plot around +12 meters from the bridge. Loran-C E-field positioning seems unaffected with a maximum positioning error of only 10 m. H-field positioning has to deal with outliers of up to 46 m. All the H-field positioning outliers are covered by the re-radiation algorithm.
Bridge F: Sunshine Skyway

Bridge F, under the Sunshine Skyway, not to be confused with the Sunshine Skyway Bridge (Bridge G), has a horizontal clearance of 27.4 meters, a vertical clearance of 6.4 meters and a total width of 39 meters. GPS positioning remains good with a maximum error of about 10 meters.

The Loran E-field positioning shows a maximum error of 75 m, thereby outperforming H-field positioning, which has a maximum error of 204 m. The large H-field error is not directly expected when looking at the dimensions of the bridge itself. However, if this bridge is seen as part of the total Sunshine Skyway Bridge, then its length grows to an impressive 16 km. The H-field re-radiation detection misses some outliers, but is still able to warn for the most severe H-field positioning errors.
Bridge G: Sunshine Skyway Bridge

Bridge G, the Sunshine Skyway Bridge is, without doubt, the most impressive bridge of Tampa Bay and possibly in Florida. Its dimensions are shown in Figure 6-51.

The official horizontal clearance is 296.9 m, but the distance between the central pylons measures 366 m. The vertical clearance of 53.6 meters allows huge cruise ships to safely sail under the bridge while two massive pylons of 131 meter support the road using 42 stay cables. The width of the road deck is 39 meters. This enormous, tall-but-slim structure does not cause
too much trouble for GPS. Loran-C E-field positioning, however, is significantly affected by this bridge, causing a maximum error of 110 meters, much more than what is seen by the other bridges. Most likely, the two pylons and the 42 stay cables are primarily responsible for the large E-field errors, because they form a gigantic vertical structure with good conductivity.

In contrast, the H-field positioning shows only a maximum error of 64 meters, but this error decays quite slowly, especially in the northeast direction of the bridge. The decay of the H-field re-radiation detection level is more rapid than that of H-field positioning, which causes a five position errors to slip through the detection, causing misleading information.

6.4.6.2 Numerical results of the influence of bridges

Table 6-9 shows some numerical information about the bridges discussed in the previous Section and the resulting Loran E-field and H-field position errors.

<table>
<thead>
<tr>
<th>Id</th>
<th>Name of bridge</th>
<th>vertical clearance</th>
<th>width</th>
<th>minimal distance</th>
<th>max error</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>E. Howard Frankland Bridge</td>
<td>13.4</td>
<td>72.0</td>
<td>17</td>
<td>42</td>
</tr>
<tr>
<td>B</td>
<td>Friendship Trail Bridge (Gandy Bridge)</td>
<td>?</td>
<td>61.0</td>
<td>6</td>
<td>54</td>
</tr>
<tr>
<td>C</td>
<td>Corey Causeway Bridge</td>
<td>7.0</td>
<td>28.0</td>
<td>5</td>
<td>73</td>
</tr>
<tr>
<td>D</td>
<td>Pinellas Bayway Bridge - Tierra Verda</td>
<td>6.7</td>
<td>12.0</td>
<td>0</td>
<td>19</td>
</tr>
<tr>
<td>E</td>
<td>Pinellas Bayway Bridge - St. Petersb.B.</td>
<td>6.7</td>
<td>11.5</td>
<td>0</td>
<td>10</td>
</tr>
<tr>
<td>F</td>
<td>Sunshine Skyway</td>
<td>6.4</td>
<td>39.0</td>
<td>14</td>
<td>75</td>
</tr>
<tr>
<td>G</td>
<td>Sunshine Skyway Bridge</td>
<td>53.6*</td>
<td>29.0</td>
<td>225</td>
<td>110</td>
</tr>
</tbody>
</table>

* Note: The pylons of the Sunshine Skyway Bridge are 131m high

The “minimal distance” is the distance, measured from the nearest edge of the bridge, after which the position error is less than 25 meters. For example, a user needs to be at least 298 meters away from the Sunshine Skyway Bridge to have an H-field positioning error less than 25 meters. It is not trivial to correlate the dimensions of the bridge with the resulting positioning error. Some general observations are possible though:

For all but the Sunshine Skyway bridge the Loran E-field positioning error is significantly smaller then the H-field positioning error.

The Sunshine Skyway Bridge forms an exception because no other tall vertical structure has been analyzed. All of the other bridges have a more horizontal geometry. This may explain
why close to the Sunshine Skyway Bridge the E-field positioning error is significantly larger than the H-field positioning error; whereas, this is reversed for the other bridges.

Re-radiation only causes a problem very close to a bridge. At such a small distance from a bridge, it is better not to rely on absolute radio positioning in combination with charts, but to use other relative positioning techniques, such as radar and “eye-ball” navigation, combined with common sense. This makes re-radiation caused by bridges, most likely, not a significant operational risk for Harbor Entrance and Approach procedures.

### 6.4.6.3 Success-rate re-radiation detection

Table 6-10 depicts statistical information about the success of re-radiation detection around the Tampa Bay bridges.

<table>
<thead>
<tr>
<th>Id</th>
<th>Name of bridge</th>
<th>total</th>
<th>nominal</th>
<th>justified</th>
<th>false alert</th>
<th>misleading</th>
<th>information</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>#</td>
<td>%</td>
<td>#</td>
<td>%</td>
<td>#</td>
<td>%</td>
</tr>
<tr>
<td>A</td>
<td>E. Howard Frankland Bridge</td>
<td>34</td>
<td>44.1</td>
<td>9</td>
<td>26.5</td>
<td>8</td>
<td>23.5</td>
</tr>
<tr>
<td>B</td>
<td>Friendship Trail Br. (Gandy Br)</td>
<td>59</td>
<td>44.1</td>
<td>16</td>
<td>27.1</td>
<td>17</td>
<td>28.8</td>
</tr>
<tr>
<td>C</td>
<td>Corey Causeway Bridge</td>
<td>25</td>
<td>60.0</td>
<td>3</td>
<td>12.0</td>
<td>7</td>
<td>28.0</td>
</tr>
<tr>
<td>D</td>
<td>Pinellas Bayway Br. - Tierra V.</td>
<td>25</td>
<td>68.0</td>
<td>3</td>
<td>12.0</td>
<td>5</td>
<td>20.0</td>
</tr>
<tr>
<td>E</td>
<td>Pinellas Bayway Br. - St. Pt. Br.</td>
<td>10</td>
<td>80.0</td>
<td>2</td>
<td>20.0</td>
<td>0</td>
<td>0.0</td>
</tr>
<tr>
<td>F</td>
<td>Sunshine Skyway</td>
<td>19</td>
<td>47.4</td>
<td>9</td>
<td>47.4</td>
<td>0</td>
<td>0.0</td>
</tr>
<tr>
<td>G</td>
<td>Sunshine Skyway Bridge</td>
<td>409</td>
<td>58.4</td>
<td>71</td>
<td>17.4</td>
<td>94</td>
<td>23.0</td>
</tr>
</tbody>
</table>

The most important figure is the “misleading information”: position outliers that are not flagged by the re-radiation detection algorithm. Fortunately, most of these “misleading-information”-events of the Tampa Bay dataset have only just exceeded the 25-meter alert limit.

Table 6-10 gives a somewhat distorted view, because it only covers those areas very close to bridges. For example, all the false alert samples of Table 6-10 were contaminated by re-radiation, but coincidentally so that this did not result in significant position errors.

A better statistical view is derived by analyzing the entire dataset. Figure 6-52 shows all the 8565-position errors with respect to the re-radiation detection levels. Table 6-11 contains the numerical information of the same analysis.
Most of the measurements, 95.8%, have been correctly detected as valid, which means that they had both a re-radiation level below 2.5° and a position error smaller than 25 meters, and therefore, are plotted in the lower-left quadrant. Because they had a re-radiation level and position error above the thresholds, 1.9% of the samples have been correctly identified as erroneous, and they ended up in the upper-right quadrant. False alerts happened for 2.1%, or 181 of the measurements, plotted in the upper-left quadrant. The lower-right quadrant shows the misleading information events, which happened for 0.1% of the measurements.

**Table 6-11: H-field re-radiation detection results for the entire Tampa Bay dataset**

<table>
<thead>
<tr>
<th></th>
<th>total</th>
<th>nominal operation</th>
<th>justified alert</th>
<th>false alert</th>
<th>misleading information</th>
</tr>
</thead>
<tbody>
<tr>
<td>#</td>
<td>#</td>
<td>%</td>
<td>#</td>
<td>%</td>
<td>#</td>
</tr>
<tr>
<td>Entire dataset</td>
<td>8565</td>
<td>8208</td>
<td>95.8</td>
<td>162</td>
<td>95.8</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>1.9</td>
<td>183</td>
<td>2.1</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>12</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>0.1</td>
</tr>
</tbody>
</table>

From this data, it can be concluded that the implemented re-radiation detection shows promising results in the capability to identify erroneous and potentially hazardous
measurements. Re-radiation detection can be used as a positioning accuracy/integrity feedback to the user and as a weight factor in an integrated navigation solution.

6.4.7 HEA dLoran showcase

The “HEA showcase” is introduced as a scenario with close resemblance to a realistic HEA procedure. dLoran is used in combination with an ASF correction map to derive a Loran position. In the Tampa Bay measurement campaign, this was achieved by creating ASF-maps using the measurement data of Tuesday 13 April 2003 and applying these correction maps on the measurements of Thursday 15 April 2003. For the measurement track, a representative path has been chosen, which starts in the Tampa Bay harbor and ends just past the Sunshine Skyway Bridge.

![HEA showcase track](image_url)

**Figure 6-53: HEA showcase track, from the Tampa Bay harbor to the Sunshine Skyway Bridge**

The total distance of the selected Thursday track is 42.3 km, which was covered in 1 hour and 11 minutes. This results in an average speed of 36 km/h. Close to the bridge, the boat slowed down significantly to 14 km/h, allowing better analysis of the influence of the bridge on the measurements. Note that slowing down close to the bridge increases the influence of the effects of the bridge on the statistics. Figure 6-54 shows the resulting positioning accuracy as a function of the grid-size of the ASF correction map.
Unfortunately, the Loran positioning results as a function of grid size are less straightforward than those presented in Figure 6-41. Several issues can be identified as likely causes for the somewhat erratic behavior:

- A large part of the E-field measurements of Tuesday could not be used, because they were too severely degraded by salt-water splashing against the E-field antenna. Exclusion of these measurements caused blank spots in the E-field correction map, which were eventually filled by extrapolation. This causes the E-field correction map to be of unrepresentative, degraded quality.

- The blank spots mentioned above will be fewer in number when a larger cell-size is chosen. This explains the significant decrease in position error between a cell-size of 0.0025° and 0.01° when an E-field correction map is used.

- The performance plot with re-radiation (the Sunshine Skyway Bridge) included, Figure 6-54-right, shows a relative small 99% positioning error for a 0.0025° grid size. Apparently, the correction map with this grid-size partly corrects the effects caused by re-radiation; this correction map is, therefore, not entirely re-radiation free.

- The Thursday run contains only 852 independent 5-second measurements. This implies that, for example, the 99% error is determined by only 9 samples. For a more reliable statistic, it is recommended to use more samples, and hence, a longer measurement run or repeating the same run a couple of times.

Because of the degraded E-field ASF correction map, the remainder of this analysis will focus only on the H-field correction map.

---

8 This active E-field antenna utilized a high input-impedance amplifier. A charge coupled amplifier would have probably largely mitigated the effect of the salt water splashes. See also Section 5.1.
Table 6-12: HEA showcase positioning performance after correction with an H-field ASF-map with a 0.01° cell-size

<table>
<thead>
<tr>
<th></th>
<th>re-radiation excluded</th>
<th>re-radiation included</th>
</tr>
</thead>
<tbody>
<tr>
<td>H-field positioning</td>
<td>95%: 9.1m, 99%: 11.4m</td>
<td>95%: 11.7m, 99%: 30.4m</td>
</tr>
<tr>
<td>E-field positioning</td>
<td>95%: 11.1m, 99%: 14.6m</td>
<td>95%: 14.2m, 99%: 38.5m</td>
</tr>
</tbody>
</table>

Figure 6-55: Position accuracy of the Tampa Bay HEA showcase with re-radiation excluded (top plots) and included (bottom plots). The solid line depicts the 95% and the dashed line the 99% positioning accuracy.
The error distributions of the H-field Loran positioning, after correction with an H-field ASF-map with a 0.01° cell-size, are plotted in Figure 6-55.

The difference between the performance with and without re-radiation is determined only by the Sunshine Skyway Bridge, which is further outlined in Section 6.4.6. With the Sunshine Skyway Bridge excluded from the dataset, the error scatter has a cigar-shaped form. This shape, which was also seen in the antenna calibration plots in Section 5.8.4, is caused by the transmitter constellation, combined with measurement noise. Malone and Jupiter provide excellent accuracy in the NW-SE direction, but Carolina Beach is too weak to match this performance in the NE-SW direction. Unfortunately, there is no significant long-term static data available that would allow comparison with the HEA-showcase. The success of the implementation of ASF-correction map can be directly derived from the comparison between a static performance measurement and the dynamic, ASF-corrected performance. Therefore, it is highly recommended to include static measurements in a future campaign.

### 6.4.8 Discussion

#### Setup

The mobile setup consisted of custom-built, tightly synchronized, E-field and H-field ASF measurement equipment. This allowed unprecedented comparison between E-field and H-field in near-field and far-field situations. The H-field antenna was successfully calibrated, thereby, minimizing any cross-talk influences and influences of the boat. An injected Loran simulator signal allowed for compensation of temperature influences of the H-field setup. The E-field setup suffered from salt-water splashing against the antenna, but relocation of the antenna towards the center of the pilothouse proved to be an adequate solution for this problem. Compensation for the propagation and transmitter timing fluctuations was enabled by the installation of a local reference station. Finally, custom-built, real-time, and post-processing analysis software facilitated the gathering of new scientific and operational insights.

#### ASF measurements

Approximately 13 hours of data was recorded during the Tampa Bay measurement campaign. The day-to-day and dual-rate repeatability of the measurements was very good, which proves proper functioning of the measurement setup. As expected from theory, the E-field and H-field measurements show excellent agreement in the far field. These far-field measurements have been successfully converted into ASF correction maps, which are directly suitable for usage in modern eLoran receivers. A method has been presented to determine the optimal grid-size for these ASF correction maps. Blank spots between the measurements are effectively filled by a novel radial interpolation and extrapolation technique, resulting in an ASF-map with full coverage of the desired area. The bridges in the Tampa Bay area caused significant re-radiation. Generally, close to bridges the E-field positioning error was smaller than the H-field error, except in close proximity of the Sunshine Skyway Bridge. This Sunshine Skyway Bridge is the only (extremely) tall vertical structure, which can be an explanation for the relative large E-field positioning errors. Correction of re-radiation effects by using a very detailed ASF-map is discouraged, because of the significant risk of ambiguous position solutions. Detection of
re-radiation, on the other hand, has proven to be very effective: virtually all position outliers were successfully identified by the implemented re-radiation detection technique. Fortunately, re-radiation is principally only a problem in very close proximity of a bridge. In those situations, it is ill advised to primarily use radio-navigation anyhow; it is then preferable to use radar and visual guidance, for example. This makes re-radiation, most likely, not a significant operational issue for Harbor Entrance and Approach procedures.

**HEA showcase performance**

The HEA showcase has been introduced as a realistic assessment of the potential Loran-C positioning performance during a Harbor Entrance and Approach procedure. The measurements taken on Tuesday were used to create an ASF correction map, which was applied on the measurements taken on Thursday. This resulted in a positioning accuracy of 9.1 m 95% and 11.4 m 99%, which is well within the HEA performance target of 20 m 95%. The grid size of the ASF correction map has some influence on the final positioning accuracy; a grid distance of 0.01 or 0.02° is ideal for the Tampa Bay area. The transmitter geometry forces the scatter-plot in a cigar-shape; an additional transmitter in the northeast will significantly improve the performance. Jupiter forms a single point of failure for Loran-C HEA procedures in south of Florida: without this transmitter, the positioning accuracy will degrade dramatically. The chance of losing Jupiter during the hurricane season is quite possible.

**E-field vs. H-field for HEA**

The unprecedented accuracy of the Tampa Bay measurements introduces the unique opportunity of a detailed comparison between Loran-C E-field and H-field performance. In the far field, both show similar results. However, E-field outperforms H-field positioning close to re-radiators such as bridges. On the other hand, H-field allows for detection of potential hazardous re-radiation situations, which is a great benefit for integrity allowing tight system integration. Additionally, H-field provides absolute heading, which can be used for such things as radar chart overlay. This extra compass functionality forms a strong commercial advantage for the H-field sensor. Some other practical issues that need consideration are the fact that an H-field antenna needs to be either calibrated or extremely well placed and carefully placed. An E-field antenna, on the other hand, requires a good ground connection and is susceptible for p-static and possibly for water splashing against the antenna. In short, both E-field and H-field have their advantages. It is the author’s opinion that the robustness against p-static and the commercial benefit of the excellent compass functionality will be likely the decisive advantages of the H-field setup. Obviously, it is also possible to equip a receiver with both an E-field and H-field antenna and to choose the antenna with momentary best performance. Such a combined receiver can also facilitate in additional integrity. The disadvantage of a combined receiver is its inherently higher cost and more difficult installation.

**Recommendations**

The Tampa Bay measurement campaign did not address the spatial decorrelation of the temporal corrections (see Section 4.5): how stable is the ASF correction map? This needs to be addressed by repeated measurements of the area, for example, combined with long-term data recording of one or more reference stations. Harbors in areas with large temporal fluctuations, such as in New England or the Great Lakes, should also be considered for such campaigns.
An attempt should be made to match ASF-modeling with the ASF-measurements. This will lead to a better understanding of the propagation phenomena, such as coastal effects, and as a result to better planning of the survey efforts and finally to better ASF correction maps. A more extensive survey is needed, especially along the coastlines, to make this iteration most successful.

The HEA-showcase presented in this Chapter shows very promising results, but is too limited to have significant statistical weight. For that, many more measurements along the important shipping routes are needed along with data detailing the stability of the ASF corrections as discussed earlier.

The E-field antenna used for the Tampa Bay measurements was susceptible to contact with salt water. Further research is needed to determine whether this phenomenon is related to the particular implementation or if it is a more fundamental problem.

It is desirable to have a reference station capable of measuring absolute ASFs. Analysis of the propagation variations also requires exact knowledge of the time of transmission of the signals, which should be made available by the system provider. For integrity purposes, it will be an advantage to equip the reference station, with both an E-field and an H-field Loran-C receiver (see Section 4.3.2).

Re-radiation needs more study: better understanding of the re-radiation effects leads to better detection algorithms, and hence, more robust receivers with higher integrity. Bridges form excellent study objects, because they are well defined, isolated structures on a conducting ground plane (salt water). This in contrast to land-based structures, where there are many unknown variables. Therefore, an extensive campaign around the Sunshine Skyway Bridge, for example, as well as EM-modeling of the same object, is highly recommended.

The Tampa Bay measurement campaign, with measurements of unprecedented quality, demonstrates the high potential of Loran-C for HEA procedures. It has also given significant insight in the scientific and operational sides of high accuracy LF radio-navigation.

### 6.5 General conclusions and recommendations

This Section contains general conclusions and recommendations regarding the measurements discussed in this Chapter. The reader is referred to the discussions of individual measurement campaigns (Section 6.2.1.4, 6.2.2.3, 6.3.2.3, 6.3.4, and 6.4.8) for more detailed analysis.
**Integrity monitoring**

The integrity of ASF measurements with respect to re-radiation has been successfully monitored by comparing the relation between E-field and H-field (Section 6.2.2, 6.3.3.3, and 6.4.5), and by analyzing the directions of the H-field signals (Section 6.4.6). These techniques are usable for dLoran reference stations (Section 4.3.2), ASF survey equipment, and for user receivers.

**Land-mobile**

Land-mobile ASF measurements in the Boston area have revealed significant LF-interference in many urban areas. This interference, dominantly produced by distribution transformers (Section 6.3.1), can severely limit the availability, and hence, the applicability of LF radionavigation in urban environments.

Buildings, tunnels, bridges, and buried power-lines, for example, cause an almost ubiquitous presence of re-radiation in land-mobile environments. The analyzed measurements show a superior performance of E-field positioning compared to H-field positioning (Section 6.3.4). It should be noted, however, that no measurements have been done inside deep urban canyons where the E-field can be severely attenuated, which thereby significantly compromise E-field availability. The measured E-field response also suffers from severe common-mode phase fluctuations, which are harmless for positioning, but do significantly affect timing capabilities (Section 6.3.2.1). Detection of severe re-radiation seems feasible (Section 6.4.6) and can be of great assistance when integrating Loran-C with other sensors, such as rate-gyro’s and odometers. Correction for re-radiation, however, is very challenging, because of the extremely localized character, and because it can be hazardous when the re-radiation correction is so significant that ambiguous position solutions can occur.

Most of the constructions causing re-radiation contain many unknown variables (e.g. a cloverleaf, Section 6.3.2) or are simple hidden (e.g. buried power-lines, Section 6.3.3). This makes a deterministic analysis extremely challenging. Therefore, a more statistical approach is recommended. For this to be successful, many more land-mobile measurements in numerous “typical” situations are required. With the resulting information, a database can be constructed enabling a performance assessment under various conditions. It is eventually the application that desires a certain accuracy, availability, and integrity and specifies the framework in which the equipment should operate. The Boston land-mobile measurements show 30-60 meters, 95% accuracy while using one ASF correction point per 5 km. For vehicle navigation applications, these results can potentially be enhanced by integration of Loran-C with, for example, inertial navigation sensors such as odometers and gyros. Track-and-trace applications, on the other hand, often do not allow for special installation requirements, the equipment should work out-of-the-box or, perhaps more appropriate, inside the (cargo-) box. The requirements then usually focus more on availability under extremely challenging conditions then on accuracy.

**Maritime**

The Tampa Bay measurement campaign has proven Loran-C positioning performance to be excellent in maritime environments. The achieved 9.1 m, 95% during a realistic scenario
(Section 6.4.7) is well within the HEA performance requirements of 20 meters, 95%. This indicates the high potential of Loran-C as a backup for GNSS during Harbor Entrance and Approach procedures.

The measurements show excellent agreement between E-field and H-field in the far field (Section 6.4.5). These far-field measurements have been converted into ASF correction maps with full coverage of the entire Tampa Bay area (Section 6.4.4). The analysis of the relation between grid-distance and positioning accuracy has shown an optimum cell-size between one and two kilometers (Section 6.4.4.2). It is recommended to analyze the possibility of merging ASF measurements with ASF modeling. Such a hybrid technique is likely to produce ASF-maps with better performance outside the measured areas, and it will lead to better understanding of the propagation effects as well. The spatial decorrelation of temporal corrections has not been addressed. This important parameter, which reflects the stability of the propagation effects, needs further research. This can be done by year-through data collection of multiple reference stations at strategic locations combined with multiple ASF survey campaigns performed in different seasons, for example.

Near-field propagation effects in maritime environments are primarily caused by bridges (Section 6.4.6.1). Fortunately, the bridges degrade the positioning accuracy only very locally. Re-radiation detection has been demonstrated successfully (Section 6.4.6.3), re-radiation correction is ill advised. It is recommended to continue re-radiation analysis, because the knowledge gained will most likely lead to improvement of the re-radiation detection techniques, and hence, higher integrity and thereby safety during HEA-procedures as well. Bridges are ideal re-radiation study objects, because they are well defined and isolated structures.

**Various**

Airborne measurements have not been discussed in this Chapter. This modality will introduce altitude as an extra parameter, which most likely imposes a significant scientific challenge. Next to avionics applications, airborne Loran-C allows for large-scale ASF and ECD measurements, which will be an advantage for the creation of nation-wide ASF correction maps. For such measurement campaigns to be useful, it is essential to first understand the effect(s) of altitude on ASFs and to develop the necessary compensation algorithms.

### 6.6 References

This Ph.D. research consists of an elaborate analysis of the characteristics of Loran-C, of the potential of a future eLoran, the technical advancements needed to obtain eLoran, and successful demonstrations of some of these future capabilities. This Chapter briefly summarizes these results and places them in an application context.
Various reports [1][3] have stated that a sole-means dependency upon GPS is undesirable and that a backup is needed for safety, environmental, and economically critical applications. It is also stated that a Low-Frequency (LF) radionavigation system is potentially suitable to fulfill this backup role due to its large dissimilarity with the Ultra High Frequency (UHF) satellite radionavigation systems [2][3]. Loran-C is currently the only operational LF radionavigation system with regional coverage; however, unfortunately it does not meet the performance standards of most contemporary applications. Various governmental agencies have fueled a renewed research effort into the true potential of Loran-C and have invested into a system upgrade from Loran-C to enhanced Loran or eLoran. This dissertation is the author’s contribution to these efforts.

This Ph.D. research consists of an elaborate analysis of the characteristics of Loran-C, of the potential of a future eLoran, the technical advancements needed to obtain eLoran, and successful demonstrations of some of these future capabilities. This Chapter briefly summarizes these results and places them in an application context. Indepth discussions can be found at the end of each individual Chapter.

In the introduction (Chapter 1), the following research question was formulated:

What are the fundamental limits of LF radionavigation and how do they affect potential applications?

The answer to the first part of the research question, regarding the fundamental limits, is discussed in Section 7.1; the answer to the second part is discussed in Section 7.2. Finally, Section 7.3 contains recommendations for further work.

7.1 Conclusions

This Section will assess the fundamental limitations of LF radionavigation based on the findings of the author’s Ph.D. research.

7.1.1 Loran-C System

This dissertation focuses on Loran-C because currently this it is the only operational low-frequency radionavigation with regional coverage. The system is the result of decades of experience, but present-day technical capabilities and requirements might lead to a more beneficial system design. A hypothesis regarding a potentially better low-frequency radionavigation system is presented in Section 7.2.2.

The most straightforward increase in system performance is obtained by increasing the number of Loran-C transmitters. The resulting shorter distances towards the transmitters and improved transmitter geometry, directly results in better repeatability (accuracy) and lowers the risk of a faulty cycle selection (integrity). Obviously, increasing the number of transmitters is related to cost and is not a fundamental issue.
7.1.2 Transmitter

Technically, modern solid-state transmitter technology is capable of very accurate control of the signal waveform, and is capable of extremely high effective radiated peak power output, which can exceed 1.5 MW if desired. **Whether the best possible transmitter technology is used is a matter of cost.**

Traditionally, Site Area Monitors (SAM) control Loran transmitter timing. This method guarantees optimal repeatability at the SAM site, but allows a decrease of positioning performance at other locations. The Loran-C transmitters in Europe and the USA have been converted to Time Of Transmission (TOT) control, where the transmitter timing is directly related to UTC. The precision with which this is done is related to cost, not to fundamental limitations. Furthermore, the internal timing control of many Loran-C transmitters uses discrete steps or “Local Phase Adjustments” (LPA). These 10 ns (Europe) or 20 ns (USA) timing steps can reduce the performance of high precision positioning (e.g. HEA) and timing applications significantly. The solution to LPAs is frequency control instead of timing control, which is being implemented in the USA, not yet in Europe, and is again a cost-related issue. Another approach to transmitter timing, which is used for GNSS transmitters (for example), would be to notify the user via a data message of the most recent transmitter timing error rather than attempting to perfectly control this error to zero.

In conclusion, the performance of the Loran-C transmitters is primarily cost-related. **The current shortcomings can be — and mostly have been already — technically solved.**

7.1.3 Receiver

The most significant difference between legacy Loran-C and the new eLoran is receiver performance. This performance is highly dependent upon the quality of the antenna, which can be made sensitive to either the E-field or the H-field component of the incident EM field. Traditionally, E-field antennas have been used for Loran-C receivers because of their relative simplicity. Unfortunately, E-field antennas are susceptible to precipitation static, which seriously endangers airborne usage (for example). **The H-field antenna has negligible susceptibility to P-static; therefore, it has become an important element in the eLoran system.**

Unfortunately, the figure-8 radiation pattern makes processing H-field antenna signals more complex than signals from the omni-directional E-field antenna. A dual-loop configuration and special algorithms are required to obtain omni-directional H-field sensitivity. Additionally, the figure-8 radiation pattern makes the H-field antenna susceptible to heading-dependent errors that, if left unmitigated, can cause significant positioning errors. E-field susceptibility has been mitigated effectively by shielding. Parasitic cross-talk, gain, and tuning differences between the two H-field loops have been modeled successfully and compensated by feed-forward correction. **The presented measurements prove successful mitigation of all heading-dependent errors to a negligible level.**

The noise behavior of the H-field antenna and associated amplifier has been studied extensively, which has lead to an orthogonal design methodology. This methodology
facilitates in the design of an optimal sensor and allows detailed analysis of the impact of antenna parameters such as rod-size. As expected, reducing the size of an H-field antenna has a significant effect on its efficiency. **Loss in noise performance can only be prevented partially by using better amplifiers.** Therefore, the noise performance of an H-field antenna has fundamental limitations; for example, it will not be possible to shrink the antenna to be fitted into a cell phone without seriously affecting the link-budget, hence reducing the receiver's performance.

Modern technology provides the receiver with a significant amount of processing power, which is used eagerly by novel signal processing algorithms. For example, these algorithms enable fast acquisition (time-to-first-fix within seconds), effective interference mitigation, and high-quality tracking resulting in meter-level positioning-repeatability.

### 7.1.4 Interference

A Loran-C receiver suffers from various forms of interference. The presence of high levels of atmospheric noise in the LF-band is indisputable. This interference, caused by thunderstorms all over the world, has forced the Loran-C system designers to employ extremely high transmitter output powers. Non-linear noise mitigation techniques can further reduce the impact of atmospheric noise. However, in areas such as Florida, USA, atmospheric noise can still hinder the performance of Loran-C and other Low Frequency radionavigation significantly. Research is ongoing to assess this difficulty.

Cross-Rate Interference (CRI), interference caused by in time overlapping Loran-C chains, is not fundamental to low-frequency radionavigation, but an unfortunate by-product of the Loran-C system implementation. The deterministic character of this interference allows modern receiver algorithms to mitigate the impact at the cost of a significant increase in receiver complexity or at the cost of a significant reduction of signal energy.

The 70-130 kHz frequency band is reserved strictly for radionavigation purposes. This international regulation has freed the Loran-C frequency band (90-110 kHz) from most interfering radio broadcasts. Remaining Continuous Wave Interference can be eliminated by notch filtering.

Local interference, originating from sources very close to the receiver, can significantly hinder reception. Regulations do not prohibit sufficiently the radiation of worrisome levels of LF interference of for example switching power electronics. Multi-domain filtering is required to suppress the local interference, but the often-erratic behavior limits the effectiveness of the mitigation attempts. Interference originating from the user vehicle itself hinders “plug-and-play” antenna installation. Interference originating from sources outside the vehicle, such as from power transformers for households, limits the potential of applications primarily in the land-mobile sector.

**In conclusion, atmospheric noise and local interference negatively impact the applicability of low-frequency radionavigation.**
7.1.5 Far-field propagation

Low-frequency radio waves propagate in the far field via both a ground wave over the earth’s surface and a sky wave, reflected against the ionosphere. The ground wave is highly stable, but often significantly attenuated, whereas the sky wave is potentially strong, but has an unknown and heavily fluctuating propagation delay. The combination of the pulse modulation of Loran-C and special receiver-tracking algorithms allows the use of the stability of the ground wave as well as the potential strength of the sky wave.

Unfortunately, the ground wave delay suffers from spatial and temporal variations. The diurnal and seasonal variations are difficult to model, but can be measured effectively by local reference stations and transmitted to the user as differential corrections. The spatial decorrelation of the temporal corrections determines its effectiveness. While only limited research has been done on this phenomenon, it is difficult to assess this fundamental limitation of LF radionavigation.

The spatial variations in propagation delay are a function of conductivity and topography and, if left uncompensated, can cause significant positioning offsets of up to hundreds of meters. The spatial variations, depicted by the Additional Secondary Factor (ASF), can be determined by measurement, modeling, or both. The lack of precise measurement equipment has disabled researchers in the past from verifying the existing propagation models. One of the objectives of this Ph.D. program was the research, design, and implementation of the necessary measurement equipment to enable an iteration process towards more insight in propagation phenomena and better propagation models.

The measurement campaigns performed in Tampa Bay and Boston suggest that an ASF-value measured every two kilometers (approximately) is sufficient to compensate for most of the spatial variation in far-field propagation. Radial interpolation between the measurement points has been successful in obtaining coverage between the measurement points. Further enhancement of the ASF correction map is expected when propagation models are calibrated with measurements.

7.1.6 Near-field propagation

Objects of significant size can very locally influence the low-frequency propagation severely by re-radiation. The E-field and H-field are decoupled in the resulting near-field propagation. Therefore, re-radiation can have an entirely different effect on a receiver equipped with an E-field antenna than on a receiver equipped with an H-field antenna.

The measurement results presented in this dissertation generally show superior E-field over H-field positioning performance in re-radiation situations and mostly similar performance in far-field conditions. There is, however, insufficient data to make a conclusive distinction in performance between the two systems.

Furthermore, the H-field tends to penetrate better in areas such as urban environments, which can be a decisive advantage for land-mobile applications.
propagation phenomena has been demonstrated by a novel technique based on the directional sensitivity of the H-field antenna. This re-radiation detection increases integrity at the cost of availability and allows for a better integration with systems such as dead-reckoning systems. The latter systems can potentially coast the user through re-radiation polluted areas.

7.1.7 In summary

Some fundamental limitations
- **Re-radiation.** Local objects influence the LF radio waves, potentially causing significant positioning errors. Detection of these errors is possible using novel H-field antenna technology; correction for the influence of re-radiation is troublesome.
- **Spatial decorrelation of temporal corrections.** The applicability of differential corrections, and hence the absolute positioning and timing accuracy, decreases with the distance towards the dLoran reference station.
- **Local interference.** The often-erratic behavior of local interference combined with limited regulations can potentially limit the applicability of LF radionavigation in primarily land-mobile applications. Furthermore, local interference from the user’s vehicle itself hinders plug-and-play antenna installation.
- **Antenna noise.** A smaller antenna will be noisier. Successful integration of an LF H-field antenna in, for example, a cell-phone will require more transmitters and/or an increase in transmitter power.

Some practical considerations
- **Transmitter geometry.** Adding more transmitters has to be politically and/or economically justifiable.
- **Receiver complexity.** Modern technology enables performance enhancements by a wide variety of hardware and software techniques. Power, cost, and practical constraints enforced by the application limits the applicability of these techniques.
- **ASF-correction map.** Absolute positioning and timing accuracy is primarily determined by the availability and accuracy of an ASF correction map.
- **dLoran.** Ultimate positioning and timing performance relies on the availability of a relatively nearby dLoran reference station.

7.2 Discussion

This Section analyzes the impact of the fundamental and practical limitations discussed in Section 7.1 on various applications. Section 7.2.2 discusses the potential of alternate low-frequency radionavigation systems.

7.2.1 Applications

GPS with its extraordinary performance has changed the radionavigation industry and its applications.
Required performance

The required positioning performance has increased for many applications. The excellent positioning accuracy of GPS can, next to increased safety and user friendliness, also enable cost savings. For example, boats equipped with modern GPS can now follow shipping routes more accurately. As a result, the channels can be dredged narrower. However, as a result of these potential narrower channels, shipping now requires GPS-quality primary and backup positioning; traditional means of navigation are not sufficient anymore. Furthermore, new regulations often use GPS performance as a starting point instead of assessing the actual required performance. The ever-increasing and sometimes unnecessarily stringent performance requirements have the risk of excessive cost, especially once a non-GNSS backup system becomes necessary.

Expected performance

Additionally, the expected positioning performance has changed significantly over the past decade. A user is familiar with the performance of his low-cost car-navigation system, conveniently forgetting outliers and failures, and expects nothing less than meter-level accuracy from any kind of navigation system under any circumstance. Hollywood further fuels the conception of unlimited possibilities with, for example, its hypothetical miniature GPS locator chips operating deep under ground in a shielded subway. Although these expectations might be unrealistic, they have to be dealt with when marketing an alternative radionavigation system.

Plug-and-Play

GPS has enabled “plug-and-play” navigation. The relatively easy installation, acceptable cost, and fancy moving-map displays have conquered not only the general public, but many professional users as well. Although reverting to traditional navigation aids, such as paper charts, lighthouses, and buoys, is in principle possible, the necessary skills might not always be sufficiently available.

Primary versus backup

Low-frequency radionavigation or Loran-C can fulfill two roles in this new era of radionavigation: as a primary system and as a backup system. Function as a primary system is only possible if Loran-C outperforms GPS in price and/or performance. While we envision the upgraded Loran — eLoran — for this role, it is not more than fair to also assume an upgraded GPS system and a fully operational GALILEO. It will be extremely hard to compete with these satellite systems and only few possibilities come to mind: the usage of LF radionavigation for heading determination (compass), indoor frequency and timing, and covert track-and-trace applications. These applications will be discussed in more detail hereafter.

Mandatory backup

The business case for Loran-C as a backup system can be split into two scenarios: those applications for which a stand-alone backup system is mandatory and those applications where it is the user’s decision to employ a backup system.
Loran-C has relatively good chances for success in the first scenario, when a stand-alone, non-GNSS backup is mandatory, most likely for safety and environmentally related uses. For the land-mobile, maritime, and timing modalities, Loran-C can form a feasible, cost-effective backup-scenario. For aviation, many systems are already in place to guarantee full redundancy. For aviation, it is mainly a governmentally determined, cost-related trade-off that determines which systems will be part of the future aeronautical system-mix and which systems will be abandoned.

In summary, considering mandatory requirements and assuming sufficient enforcement, the business case for Loran-C is determined by the legislators rather than the users.

Non-mandatory backup

If a backup system is not mandatory — for example, for most land-mobile applications — it is the user that determines the applicability of Loran-C, usually with an economical motivation. The cost-benefit analysis of the user is based on the risk of a failure multiplied by the cost of this potential failure.

Assume, for example, that a taxi company has switched its entire operation to GPS/GSM/GPRS-based planning, employment, billing, and navigation. A GPS outage could cripple this system causing serious loss in revenue. A backup system such as Loran-C could potentially mitigate this vulnerability. However, the backup system would have to be very affordable: the contemporary competitive markets allow only slim margins. This puts high pressure on the operational costs: although the risk can be known, a backup system simply might not be affordable enough.

The same reasoning can be applied for such things as backup timing for cell-phone operators. Loran-C is an ideal candidate to deliver backup timing with GPS performance, but the system needs to be sufficiently affordable.

In summary: the non-mandatory backup market has potential, but it requires good performance for sharp prices and also the user’s awareness of the necessity of a backup.

The role of Loran-C for various applications is discussed in the remainder of this Section.

7.2.1.1 Land-mobile

Success in land-mobile applications opens the door to mass-market employment, which is an appealing scenario for investors. Land-mobile use of low-frequency radionavigation is most challenging, primarily caused by the almost ubiquitous presence of re-radiation.

Measurements have shown that especially the position derived by the H-field setup is sensitive to local disturbances. A buried cable can be sufficient to cause significant position outliers. Detection of these anomalies appears feasible; on the other hand, correction is difficult or even impossible. The land-mobile measurements described in this dissertation were generally in favor of the E-field setup, but the dataset is too limited for a conclusive
assessment. From an availability perspective, H-field is probably superior over E-field due to the better penetration of the H-field in urban environments, for example.

The available measurements results are too limited for a more detailed assessment of the availability and accuracy, and hence the applicability of Loran-C in land-mobile environments.

**Automotive**

The author has some experience with the automotive usage of Loran-C based on measurements in, among others, the Boston area, as well as in London, Rotterdam, and Delft. The positioning accuracy on highways is usually very good, except under bridges. County roads in the USA are cursed with many overhead power lines and noisy power transformers. The resulting increase in noise and re-radiation causes limited availability and decreased accuracy. In the Netherlands, most power lines are underground and the power transformers are more centralized, potentially resulting in overall better automotive positioning performance. Urban environments are hindered more by re-radiation causing a more “noisy” positioning. The information currently available is unfortunately insufficient for a more quantitative performance assessment.

**Automotive usage of Loran-C allows the integration with odometers and rate-gyros, which can significantly increase the overall positioning performance.** For most applications, the accuracy of the positioning system must be sufficient to uniquely identify the road on which the user is located. Map-matching techniques can further increase the quality of the navigation feedback to the user.

**Indoor**

**Successful indoor positioning technology has the potential to enable “killer-applications”** and therefore receives much attention from scientists, engineers, and investors. An example of an indoor positioning application is the mandatory E911 (USA) and E112 (Europe) requirement that demands an accurate position fix of the mobile phone used to place an emergency call. Another example is the use of electronic navigation aids by firefighters. Indoor navigation combines a most difficult environment with very high 3D accuracy demands.

Low-frequency H-field penetrates relatively well into buildings, but is significantly affected by re-radiation from the building structure. Furthermore, the electromagnetic interference originating from the electrical systems severely decreases the obtainable signal-to-noise ratio. It is the author’s opinion, for example, that Loran-C might be capable to determine which building the user is in, but not where the user is inside the building.

It would be an interesting experiment to measure the indoor propagation characteristics for relatively low-frequency signals, from (for example) 100 kHz to 100 MHz, while the current focus is more on the 1.5 GHz GPS signals and on Ultra Wide Band (3 – 10 GHz).
Track and trace

The international transport industry is a multi-billion-dollar business. A well-organized infrastructure is essential in this highly competitive market. Real-time feedback of the position of (important) packages enables process optimizing and helps to prevent late deliveries along with the associated penalties. These penalties can be extreme: for example, €10,000 per minute for automotive parts if the delay results in a halt of production. The positioning of the packages does not necessarily require meter-level positioning [4]. In many situations, it can be sufficient to know whether the package is stuck in traffic and if so, its approximate location in the traffic.

The radionavigation receiver/communication unit is preferably placed inside the package, which subsequently can be inside a container, truck, or airplane. Recent tests have shown promising results of the usage of Loran-C for track and trace applications: Loran positioning showed a significantly higher availability than a next-generation, highly sensitive GPS receiver showed [4]. It is advisable to explore the applicability of low-frequency radionavigation for track-and-trace applications further; the market potential is considerable.

7.2.1.2 Maritime

Originally, Loran-C was primarily intended for maritime use on open waters and has had a considerable user base. Many fishermen still have their favorite fishing spots written down in Loran TDs. Not surprisingly, GPS became more popular in this user group when Garmin added GPS-Lat/Lon to Loran-TD conversion in their marine product line. The conversion algorithm, however, assumes only all-sea water propagation, potentially resulting in a significant conversion error. Therefore, Loran-C can still rely upon a “legacy” user group that uses Loran-C in a traditional fashion to retrace locations recorded decades ago. However, the potential future availability of accurate ASF-maps will enable this legacy user group to make a more precise conversion from their fishing-spot TDs to Lat/Lon coordinates, thereby making Loran-C receiving equipment superfluous. It is an interesting phenomenon that the availability of Loran ASF-maps can actually reduce the size of certain existing Loran user groups.

High-performance backup

Loran’s maritime potential lies in its stand-alone backup functionality. Shipping routes are used more extensively, resulting in increasingly stringent navigation demands. The increased importance of high-quality navigation is also stressed in, for example, the eNavigation\(^1\) concept[5]. The dependency upon accurate positioning combined with safety requirements results in the need of a high-performance backup positioning system as well. The performance demonstrated in the Tampa Bay measurement campaign (Section 6.4.7) and in other campaigns [5] suggests potential for Loran-C to fulfill this maritime backup role. It should be noted, however, that these measurement campaigns were conducted under good

\(^1\) eNavigation is the cost-effective collection, integration, and display of maritime information onboard and ashore by electronic means to enhance berth-to-berth navigation and related services for safety and security at sea, and to protect the marine environment.
conditions. More measurements, under more challenging conditions, are needed for a more definite performance assessment.

**Compass**

An H-field Loran-C receiver can, along side positioning and timing, provide a very accurate compass functionality. Increasingly, accurate heading devices are used in a maritime environment; for example, these devices are used to enable chart-overlay on radar imagery. An LF radio compass can provide the heading information with the necessary accuracy while using only a small antenna. A GPS-compass, on the other hand, requires at least two antennas with a separation of usually 50 cm or more. The **Loran-C compass functionality has potential as a primary heading sensor**, which makes the purchase of a Loran-C receiver more appealing for the maritime user.

### 7.2.1.3 Aviation

The safety requirements for aviation require extreme **integrity** demands of the navigation system. This dissertation, however, primarily focuses on **accuracy** and is therefore less suitable for an assessment of the applicability of Loran-C for aviation applications.

Loran-C has been approved as an enroute, supplemental air navigation system for both Instrument Flight Rules (IFR) and Visual Flight Rules (VFR) operations. The FAA published a report in 2004 [7] stating that eLoran will be technically capable to meet the requirements of the Non-Precision Approach procedure.

**Technically, eLoran can fulfill a backup-role for aviation.** It is a political decision, to be made by governmental bodies such as the American FAA and the European Euro Control, whether eLoran will be used as such².

### 7.2.1.4 Time and frequency

The availability of accurate time and frequency has become increasingly important in modern society. For example, many communication infrastructures, such as cell-phone base stations, require an accurate frequency source and often accurate time as well. GPS very affordably fulfills these time and frequency requirements; therefore, it is used extensively. As a result, there is a growing dependency upon GPS-time for many critical systems. Short GPS outages can be bridged by Rubidium clocks, for example; longer outages require either a highly expensive Cesium clock or an alternate (non-GNSS) radio-system.

Alternatives to GPS (GNSS) time are the use of geo-stationary satellites or low-frequency radio systems. A high-gain dish antenna can be used to receive the signals from the geostationary satellites (e.g. SBAS) making the system robust against jamming, but also cumbersome in installation. **Low-frequency radio systems can provide potentially a more cost-effective solution for backup time and frequency.**

---

² For commercial aviation in Europe, Euro Control has expressed its preference for the aviation specific DME-DME system over the multi-modal eLoran system.
Low-accuracy timing is available in large parts of the world via dedicated LF timing transmitters such as the DCF77 transmitter in Mainflingen, Germany. The receivers for these signals are extremely small and cheap, making them suitable for integration in such things as wall clocks, coffee machines, and even in wristwatches. Sky wave influences and the unknown user position limit the obtainable timing accuracy of these LF systems to more than 1 ms [8]; in practice, the accuracy provided by most units is much worse.

**Stationary**

Currently, most applications requiring accurate time and/or frequency are stationary. This significantly eases the technical challenges. The unknown part of the propagation delay can be calibrated once by GPS during installation, an integrated Loran-C / GPS timing unit can perform this task fully automated. Highly accurate timing performance can be achieved by using a nearby dLoran reference station. The fundamental limitation of the achievable timing performance is then determined by the spatial decorrelation of the temporal corrections, hence: a larger distance towards the dLoran reference stations leads to a potentially larger timing error.

Given a calibrated receiver and the availability of dLoran corrections, a timing accuracy better than 100 ns is likely [8]. The frequency error is more related to the transmitter timing control. Timing steps or LPAs inserted at the transmitter significantly degrade the receiver’s frequency stability. Fortunately, the new time and frequency equipment installed at the US Loran-C transmitters employ frequency control rather than timing control. The resulting obtainable frequency stability by eLoran is better than $2 \cdot 10^{-13}$ [8].

**Indoor**

eLoran also opens the door to indoor timing. The LF H-field signals are attenuated much less than the L-band GPS signals. Furthermore, a stationary application allows extremely long integration of the weakened LF signals, thereby mitigating most of the potential interference and noise. The economical advantage of indoor timing can be significant, considering the sometimes-excessive cost of installing an antenna on a roof. Therefore, it is worthwhile to research the potential of eLoran indoor timing further.

The potential role of eLoran for the time-and-frequency infrastructure is significant. The timing performance delivered by modern eLoran receivers using the current Loran-C system infrastructure is already significant for most applications. Upgrading the transmitter timing equipment and the availability of dLoran reference stations will further enhance the performance, meeting the requirements of even the most demanding applications. This makes eLoran unique in its ability to backup GNSS timing, a role that is often undervalued.

**7.2.2 Is Loran-C the optimal LF radionavigation system?**

This dissertation has used Loran-C as an example for low-frequency radionavigation. This choice has been practical: Loran-C is the currently the only LF radionavigation system with
regional coverage. The question now arises: If we would start from scratch, is Loran-C then the LF system we would build?

**Frequency**

The 1943 Loran-A radionavigation system operated at 1950 KHz, whereas transmitters for lower frequencies were technically not possible at that time. Ground-wave stability studies revealed that a lower frequency provided better performance over a longer range. This motivated the choice for 100 kHz for the 1958 Loran-C system. Although Loran-C functions over longer distances, it is still a regional system. In 1968, the Omega system was put into service using continuous wave signals around 10 kHz. Omega did provide global coverage, but with reduced accuracy, whereas the limited frequency band available at those extremely low frequencies did not allow for a pulsed signal modulation which is needed to separate the ground wave from the sky wave. In retrospect, the choice of the 100 kHz band appears to be optimal with respect to the highly stable ground wave over a long distance, combined with the possibility of ground wave/sky wave separation.

However, the argument for 100 kHz is based on stable far-field propagation. At this low frequency, the tiniest phase-change caused by re-radiation results in a significant positioning error. It would be an interesting experiment to compare the near-field effects at 100 kHz, 1 MHz and 10 MHz, for example. A higher frequency might prove beneficial under these local conditions. On the other hand, a higher frequency will also significantly reduce the transmitter's effective range forcing an increase in the number transmitters. Fortunately, at those higher frequencies less transmitter power is needed: the antenna efficiency is higher, and the atmospheric background noise lower. Such an investment in infrastructure is only opportune, however, if the novel system provides a significant performance increase compared to both eLoran and GPS, such as for track-and-trace and indoor applications. The author is aware of the speculative nature of this argument, but thinks that it might be refreshing to also consider novel low-frequency technology next to the present focus on technology such as Ultra Wide Band (UWB) research.

**Modulation**

LF radionavigation systems can be divided into the pulsed Loran-C system and the Continuous Wave (CW) systems such as DataTrak and the former Omega and Decca Navigator. Both technologies have their advantages.

The advantage of CW systems is their simplicity. Straightforward frequency domain filtering is sufficient to discard most noise and interference; a simple tracking algorithm is sufficient to accurately determine the phase of the CW. The phases of multiple CW signals from a transmitter are compared to solve the ambiguity and hence determine the pseudo-range; the pseudo-ranges from three or more transmitters are used to calculate the user's position. The main drawback of LF CW systems is the systems' inability to separate the ground wave from the sky wave, which severely limits the operational range of the systems and the obtainable absolute accuracy, primarily at night.
Loran-C uses pulse modulation to enable ground wave/sky wave separation. Tracking only the ground wave provides more stability over a significantly larger range. The pulse modulation requires significant bandwidth: Loran-C, with its center frequency at 100 kHz, uses 20 kHz for the pulse modulation. Processing these wide-band signals by the receiver is not a trivial task. Only moderate bandpass filtering is allowed to avoid disturbing the cycle-identification process. Notch filtering is required to suppress in-band CW interference. Finally, spatial-domain and time-domain filtering is employed to mitigate cross-rate and remaining interference. The tracking algorithm is complex: only the low-energy start of the pulse is sky wave free and therefore stable while the remaining part of the pulse contains the majority of the energy, but is potentially polluted with sky waves. **Combining the stability of the ground wave and the energy of the sky wave is non-trivial, but is required for the high demands of contemporary applications.**

Fortunately, modern signal-processing techniques, combined with the currently available computing power in modern receivers, allow employment of the described complex filtering and tracking algorithms. Under normal conditions, the obtainable performance is impressive as demonstrated, for example, by the Tampa Bay measurement campaign (Section 6.4). Accurate and robust positioning becomes challenging, however, under less favorable conditions: for example, when forced to using distant stations, in deep urban canyons, during abnormal sky wave conditions or under the presence of significant interference. **A change in modulation can increase the performance, especially under difficult conditions.**

A steeper rising pulse, with a rise-time of 40 µs instead of 65 µs, for example, provides more energy at the ground-wave tracking point, resulting in a more robust cycle-identification process. Unfortunately, regulations do not allow the increased spectrum usage required for such a change. Perhaps other modulation techniques such as CDMA or FM are possible, which potentially more efficiently use the 90-110 kHz band while still providing the necessary ground wave/sky wave separation.

Finally, **a mixed form of both pulse modulation and CW might be worth considering.** The CW is then used for the short-term tracking loop, providing a robust, low-noise track, but with an arbitrary and slowly changing offset. The pulsed signal is used to provide a stable, but noisy ground-wave track.

The Loran-C pulse modulation currently used has been established based on the allowable spectral usage and available transmitter technology. Although the transmitter technology has significantly evolved over the past five decades, the spectral regulations have been significantly less progressive.

**Timing structure**

Loran-C transmitters are organized in chains, which have a unique Group Repetition Interval (GRI), after which the timing is repeated. Station identification is done by the GRI and the timing with respect to the master station of the chain (Section 2.4.2). This concept has been developed for single-chain hard-limiter receivers based on 1960s technology. Modern receivers are all-in-view; they use stations from multiple chains. Furthermore, Eurofix or 9th pulse modulation allows station identification and UTC synchronization by a data message.
As a result, stations can be used in pseudo-rho-rho positioning in a similar fashion to GNSS, independently of the GRI-timing. This allows restructuring the system’s timing structure to fit the capabilities of current and future eLoran receivers optimally.

A major drawback of the chain configuration of Loran-C is cross rate: stations from multiple chains overlap in time. Legacy (hard-limiter) receivers were relatively robust against this type of impulsive interference. Modern linear receivers, on the other hand, require cross-rate mitigation techniques for proper functioning. Averaging does not provide sufficient suppression, blanking cross-rated pulses costs significant tracking energy, and the estimate-and-subtract method is highly complex and also has its limitations (Section 3.6.3). It would be beneficial for the overall system performance to redesign the system’s timing structure with the focus on minimal cross-rate and efficient acquisition.

**Signals of opportunity**

Although implementing a new LF radionavigation system is an interesting thought, it will most likely remain a hypothetical scenario. The ongoing upgrade process from Loran-C to eLoran, while maintaining compatibility, probably has more chance of success, especially within the current political climate. However, the thought of alternate LF radionavigation systems should not be abandoned. In fact, these systems are already operational, but not yet identified as such. The spectrum is filled with radio transmissions, many of which can, in principle, be used in for ranging. This “Signals of Opportunity” concept is not new; it has, for example, been demonstrated in the “CityFix” project [9]. A small upgrade of strategically chosen broadcast transmitters can strengthen the potential of the “Signals of Opportunity,” for example, by equipping those transmitters with a highly stable clock and by piggybacking an additional navigation signal on the radio broadcast, if possible.

### 7.3 Recommendations

**Position domain**

This dissertation has focused on pseudorange errors rather than on position domain errors. Studying the latter involves a complicated analysis of the correlation between the various errors and of the influence of geometry on the position error. Comprehending these relations is an essential extension to the research presented.

**Temporal stability of propagation**

The temporal stability of LF propagation is largely unknown. The assessments available are based on old TD measurements; more-accurate information is needed to meet the contemporary requirements. It is suggested to use multiple, strategically placed monitor receivers to collect year-round measurements. The agreement between the measurements of the multiple monitors can be used to assess the spatial decorrelation of the temporal variations, a crucial parameter for precise timing and positioning. A better understanding of temporal propagation variation is crucial for the determination of the optimal locations of
dLoran reference stations and for a better assessment of the potential positioning and timing performance.

**Spatial fluctuations of propagation**
Accurate ASF measurements are now possible; the next step is the validation of the available propagation models. Airborne ASF measurements are probably the quickest method to cover a large area. However, the use of airborne measurements for model verifications requires knowledge of the influence of receiver-altitude on the measured ASF. Once this relation has been established, it is possible to iterate the propagation models by specific, well-planned measurements. Once the model shows reasonable agreement with the measured ASFs, it will be worthwhile to research the optimal method of merging the modeled and measured ASFs to obtain an ASF correction map with a high accuracy and large coverage, while requiring only a minimal number of measurements.

**Re-radiation**
Bridges surrounded by saltwater form interesting re-radiation study objects. The isolated structures on a conducting ground plane form well-defined objects enabling detailed electromagnetic modeling. Comparing the modeled propagation effects with the measurements can lead to better understanding of re-radiation, and hence better re-radiation detection and potentially correction techniques.

**Receiver**
New receiver technology is the most important part of the upgrade process from Loran-C to eLoran. Many improvements are still in a research stage; some have already found their way into production units. For receiver manufacturers to move forward, it is important to define the receiver Minimum Operational Performance Standards (MOPS) for eLoran. Obviously, receiver manufacturers also require a healthy business case, which is highly dependent upon the political commitment to maintain and upgrade the system infrastructure.

**Land-mobile applications**
The technology is now available to assess the capabilities of Loran-C for land-mobile applications. It is suggested to survey typical environments extensively, and to build a database with the derived statistics. This database could be used to predict the performance of potential track-and-trace applications, for example. Furthermore, automotive usage of Loran-C allows integration with sensors such as rate-gyros and odometers. Integration of Loran-C with these sensors might provide a leap in performance, especially in the case of deep integration combined with feedback from re-radiation detection algorithms.

**Timing applications**
Stationary timing is already possible with the current infrastructure. The addition of dLoran reference stations will further enhance the obtainable accuracy. Perhaps most important is the stimulation of the general awareness of the critical need for backup timing. The business case for backup timing using eLoran will probably build itself once this awareness becomes reality.
It is further suggested to investigate eLoran’s potential to be used for indoor timing; success in this area will create an interesting advantage over the current GPS timing equipment.

**Interference**

The LF radionavigation band from 70-130 kHz increasingly suffers from local interference sources. For example, classic transformers make place for poorly shielded switching power supplies; the transformers used to convert high-voltage to household power are being built significantly cheaper, but they also create significantly more LF EM pollution, and developments with Power Line Telecommunications (PLT), for example, further endanger the low end of the radio spectrum. Without more stringent regulations, increasing interference levels might make land-mobile usage of LF radionavigation practically impossible in the near future.

**Political**

The future of eLoran will be determined mainly by politics. An acknowledgement of the need for Loran combined with the accompanying investments in the infrastructure and legislative actions will motivate receiver manufacturers to invest and motivate users to purchase new equipment. The result is an acknowledgement of the need for and success of eLoran. Refraining from initiative will have the opposite effect: industry will be reluctant to invest because the system is not guaranteed; the eLoran user base will diminish to a negligible level, and apparently there was no need for eLoran after all. Both arguments appear to be valid; the difference is the willingness to invest in insurance.

### 7.4 References


### List of Abbreviations and Acronyms

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>2D</td>
<td>2-dimensional</td>
</tr>
<tr>
<td>3D</td>
<td>3-dimensional</td>
</tr>
<tr>
<td>ADC</td>
<td>Analog-to-Digital Converter</td>
</tr>
<tr>
<td>ASF</td>
<td>Additional Secondary Factor</td>
</tr>
<tr>
<td>bps</td>
<td>bits per second</td>
</tr>
<tr>
<td>CA</td>
<td>Coarse / Acquisition</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>CEP</td>
<td>Circular Error Probability</td>
</tr>
<tr>
<td>CONUS</td>
<td>Conterminous United States</td>
</tr>
<tr>
<td>CRI</td>
<td>Cross Rate Interference</td>
</tr>
<tr>
<td>Cs</td>
<td>Cesium</td>
</tr>
<tr>
<td>CWI</td>
<td>Continuous Wave Interference</td>
</tr>
<tr>
<td>dASF</td>
<td>differential ASF (ASF measured relative to the ASF of one of the stations)</td>
</tr>
<tr>
<td>DCF77</td>
<td>German 77.5 kHz timing service</td>
</tr>
<tr>
<td>DGPS</td>
<td>Differential GPS</td>
</tr>
<tr>
<td>DOD</td>
<td>Department of Defense</td>
</tr>
<tr>
<td>DOP</td>
<td>Dilution of Precision</td>
</tr>
<tr>
<td>DOT</td>
<td>Department of Transportation</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital Signal Processor</td>
</tr>
<tr>
<td>ECD</td>
<td>Envelope-to-Cycle Difference</td>
</tr>
<tr>
<td>ED</td>
<td>Emission Delay</td>
</tr>
<tr>
<td>E-field</td>
<td>Electric field</td>
</tr>
<tr>
<td>EGNOS</td>
<td>European Geostationary Navigation Overlay Service</td>
</tr>
<tr>
<td>eLoran</td>
<td>Enhanced LORAN</td>
</tr>
<tr>
<td>EM</td>
<td>Electro Magnetic</td>
</tr>
<tr>
<td>ERNP</td>
<td>European Radio Navigation Plan</td>
</tr>
<tr>
<td>FAA</td>
<td>Federal Aviation Administration</td>
</tr>
<tr>
<td>FERNS</td>
<td>Far East Radionavigation System</td>
</tr>
<tr>
<td>FIR</td>
<td>Finite Impulse Response</td>
</tr>
<tr>
<td>FM</td>
<td>Frequency Modulation</td>
</tr>
<tr>
<td>FRP</td>
<td>Federal Radionavigation Plan</td>
</tr>
<tr>
<td>GCD</td>
<td>Greatest Common Devisor</td>
</tr>
<tr>
<td>GDOP</td>
<td>Geodetic Dilution Of Precision</td>
</tr>
<tr>
<td>GEO</td>
<td>Geostationary Earth Orbiting</td>
</tr>
<tr>
<td>GLONASS</td>
<td>Global Orbiting Navigation System</td>
</tr>
<tr>
<td>GNSS</td>
<td>Global Navigation Satellite System</td>
</tr>
<tr>
<td>GPS</td>
<td>Global Positioning System</td>
</tr>
<tr>
<td>GRI</td>
<td>Group Repetition Interval</td>
</tr>
<tr>
<td>HCPR</td>
<td>Halve Cycle Peak Ratio</td>
</tr>
<tr>
<td>HDOP</td>
<td>Horizontal Dilution Of Precision</td>
</tr>
<tr>
<td>HEA</td>
<td>Harbor Entrance and Approach</td>
</tr>
<tr>
<td>HF</td>
<td>High Frequency</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
</tr>
<tr>
<td>--------------</td>
<td>-----------</td>
</tr>
<tr>
<td>H-field</td>
<td>Magnetic field</td>
</tr>
<tr>
<td>HMI</td>
<td>Hazardous Misleading Information</td>
</tr>
<tr>
<td>HPL</td>
<td>Horizontal Protection Limit</td>
</tr>
<tr>
<td>IALA</td>
<td>International Association of Lighthouse Authorities</td>
</tr>
<tr>
<td>ICAO</td>
<td>International Civil Aviation Organization</td>
</tr>
<tr>
<td>IFR</td>
<td>Instrument Flight Rules</td>
</tr>
<tr>
<td>ILA</td>
<td>International Loran-C association</td>
</tr>
<tr>
<td>ILS</td>
<td>Instrument Landing System</td>
</tr>
<tr>
<td>IMO</td>
<td>International Maritime Organization</td>
</tr>
<tr>
<td>INS</td>
<td>Inertial Navigation System</td>
</tr>
<tr>
<td>IRS</td>
<td>Inertial Reference System</td>
</tr>
<tr>
<td>ITU</td>
<td>International Telecommunication Union</td>
</tr>
<tr>
<td>LF</td>
<td>Low Frequency</td>
</tr>
<tr>
<td>LOP</td>
<td>Line Of Positioning</td>
</tr>
<tr>
<td>LORAN</td>
<td>LOng RAnge Navigation</td>
</tr>
<tr>
<td>LORAPP</td>
<td>LORan-C Accuracy Performance Panel</td>
</tr>
<tr>
<td>LORIPP</td>
<td>LORan-C Integrity Performance Panel</td>
</tr>
<tr>
<td>LPA</td>
<td>Local Phase Adjustment</td>
</tr>
<tr>
<td>MOPS</td>
<td>Minimal Operational Performance Standard</td>
</tr>
<tr>
<td>N/A</td>
<td>Not Applicable</td>
</tr>
<tr>
<td>NAS</td>
<td>National Airspace System</td>
</tr>
<tr>
<td>NELS</td>
<td>Northwest European Loran System</td>
</tr>
<tr>
<td>NEC</td>
<td>Numerical Electromagnetic Code</td>
</tr>
<tr>
<td>NM</td>
<td>Nautical Mile</td>
</tr>
<tr>
<td>NPA</td>
<td>Aviation Non Precision Approach</td>
</tr>
<tr>
<td>ns</td>
<td>nanosecond</td>
</tr>
<tr>
<td>OU</td>
<td>Ohio University</td>
</tr>
<tr>
<td>PC</td>
<td>Phase Code</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed Circuit Board</td>
</tr>
<tr>
<td>PCCIP</td>
<td>President’s Commission on Critical Infrastructure Protection</td>
</tr>
<tr>
<td>PCI</td>
<td>Phase Code Interval</td>
</tr>
<tr>
<td>PDF</td>
<td>Probability Density Function</td>
</tr>
<tr>
<td>PF</td>
<td>Primary Factor</td>
</tr>
<tr>
<td>PVT</td>
<td>Position, Velocity and Time</td>
</tr>
<tr>
<td>PNT</td>
<td>Position, Navigation and Time</td>
</tr>
<tr>
<td>ppm</td>
<td>parts per million</td>
</tr>
<tr>
<td>PPM</td>
<td>Pulse Position Modulation</td>
</tr>
<tr>
<td>p-static</td>
<td>Precipitation Static</td>
</tr>
<tr>
<td>PVT</td>
<td>Position, Velocity and Time</td>
</tr>
<tr>
<td>RAIM</td>
<td>Receiver Autonomous Integrity Monitoring</td>
</tr>
<tr>
<td>rASF</td>
<td>relative ASF (ASF measured with respect to an arbitrary time but absolute frequency)</td>
</tr>
<tr>
<td>RASIM</td>
<td>Receiver Autonomous Signal Integrity Monitoring</td>
</tr>
<tr>
<td>Rb</td>
<td>Rubidium</td>
</tr>
<tr>
<td>RMS</td>
<td>Root Mean Square</td>
</tr>
<tr>
<td>RNP</td>
<td>Required Navigation Performance</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Definition</td>
</tr>
<tr>
<td>--------------</td>
<td>------------</td>
</tr>
<tr>
<td>RTCA</td>
<td>Radio Technical Commission for Aeronautical Services</td>
</tr>
<tr>
<td>RTCM</td>
<td>Radio Technical Commission for Maritime Services</td>
</tr>
<tr>
<td>SA</td>
<td>Selective Availability</td>
</tr>
<tr>
<td>SAM</td>
<td>System Area Monitor</td>
</tr>
<tr>
<td>SBAS</td>
<td>Space-Based Augmentation Systems</td>
</tr>
<tr>
<td>SDLIC</td>
<td>Spatial Domain Local Interference Cancellation</td>
</tr>
<tr>
<td>SF</td>
<td>Secondary Factor</td>
</tr>
<tr>
<td>SIR</td>
<td>Signal to Interference Ratio</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
</tr>
<tr>
<td>TD</td>
<td>Time Difference</td>
</tr>
<tr>
<td>TDMA</td>
<td>Time Division Multiple Access</td>
</tr>
<tr>
<td>TFE</td>
<td>Time and Frequency Equipment</td>
</tr>
<tr>
<td>TOA</td>
<td>Time Of Arrival</td>
</tr>
<tr>
<td>TOT</td>
<td>Time Of Transmission</td>
</tr>
<tr>
<td>TRAIM</td>
<td>Time Receiver Autonomous Integrity Monitoring</td>
</tr>
<tr>
<td>TU</td>
<td>Technical University</td>
</tr>
<tr>
<td>UHF</td>
<td>Ultra High Frequency</td>
</tr>
<tr>
<td>US</td>
<td>United States</td>
</tr>
<tr>
<td>USA</td>
<td>United States of America</td>
</tr>
<tr>
<td>USCG</td>
<td>United States Coast Guard</td>
</tr>
<tr>
<td>USNO</td>
<td>United States Naval Observatory</td>
</tr>
<tr>
<td>UTC</td>
<td>Coordinated Universal Time</td>
</tr>
<tr>
<td>UWB</td>
<td>Ultra Wide Band</td>
</tr>
<tr>
<td>VFR</td>
<td>Visual Flight Rules</td>
</tr>
<tr>
<td>VHF</td>
<td>Very High Frequency</td>
</tr>
<tr>
<td>Volpe</td>
<td>Volpe National Transportation Systems Center</td>
</tr>
<tr>
<td>WAAS</td>
<td>Wide Area Augmentation System</td>
</tr>
<tr>
<td>WGS</td>
<td>World Geodetic System</td>
</tr>
<tr>
<td>Xtalk</td>
<td>cross-talk</td>
</tr>
</tbody>
</table>
NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY
Appendix A  RNP Definitions

A.1 RNP Definitions

In the 1990s, the International Civil Aviation Organization (ICAO) and Radio Technical Commission for Aeronautical Services (RTCA) developed the Required Navigation Performance parameters or RNP. Since then, these parameters have also been largely adopted by the other modalities: maritime and land-mobile.

The RNP parameters define the minimum requirements in terms of accuracy, integrity, availability, and continuity of service. This Section summarizes these definitions as specified in [1].

Accuracy

“The degree of conformance between the estimated or measured position and/or the velocity of a platform at a given time and its true position or velocity.”

Radio navigation performance accuracy is usually presented as a statistical measure of system error and is specified as:

a) Predictable. The accuracy of a position in relation to the geographic or geodetic coordinates of the earth. Predictable accuracy is commonly also denoted as absolute accuracy.

b) Repeatable. The accuracy with which a user can return to a position whose coordinates have been measured at a previous time with the same navigation system.

c) Relative. The accuracy with which a user can determine one position relative to another position regardless of any error in their true positions.

The accuracy requirement is a statistical measure of performance and is usually expressed in a 95-percentile confidence level. For example, the RNP 0.3 requirements for Non Precision Approach states a target absolute accuracy of 307 m 95%; 95 out of 100 reported positions need to be within 307 meters of the true position. Note that the error distribution of a navigation system is seldom Gaussian, which makes determination of the 95-percent accuracy limit a non-trivial task.

Integrity

“The ability to provide timely and valid warnings to users when the system should not be used for navigation”

Integrity is the key safety parameter within RNP; it describes the risk associated with latent system failure. The integrity is usually specified in terms of the protection limit and time to alarm, being the absolute accuracy limit on which the system should raise an alarm, and the
maximum delay allowed between the occurrence of exceeding this limit, and the user’s notification, respectively.

**Continuity of service**

“The ability of the system to perform its function without interruption during the intended operation, if it did so at the beginning of the operation”

Reversion to a backup navigation system or procedure is inconvenient and potentially not without risks. This is especially true for most demanding operations, such as landing an aircraft, where reversion is usually not possible and an alert leads to such situations as a missed-approach procedure. The continuity parameter specifies the risk that such an undesirable situation occurs. Note that landing an aircraft takes only a very limited amount of time resulting in continuity requirements usually specified in minutes at most. In a maritime context, however, entering a harbor, for example, will take considerably longer resulting in a continuity requirement specified in hours.

**Availability**

“Availability is an indication of the ability of the system to provide usable service within the specified coverage area and is defined as the portion of time during which the system is to be used for navigation during which reliable navigation information is presented to the flight crew, autopilot or other system managing the aircraft.”

Availability is defined as the percentage of time that the services of a system are usable. In the context of the earlier discussed parameters, a system can be said to be available when it meets the accuracy, integrity, and continuity requirements. In principle, equipment failure and maintenance reduce system availability. Further, the user might experience a reduced availability under local disturbances such as interference, multipath, shadowing, or re-radiation.

**A.2 Additional definitions**

Although not strictly defined within the RNP concept, the following parameters are also very important in defining the performance of a radionavigation system:

**Coverage**

*The coverage provided by a radio navigation system is the surface area or space volume in which the signals are adequate to permit the user to determine position to a specified level of accuracy.* Coverage is influenced by system geometry, signal power levels, receiver sensitivity, atmospheric noise conditions, and other factors that affect signal availability.

**Ambiguity**

*System ambiguities exist when the navigation system identifies two or more possible positions of the vehicle, with the same set of measurements, with no indication of which is the most nearly correct position.*
potential for system ambiguities should be identified together with a provision for users to identify and resolve them.

**Alternate systems**
Alternate means of navigation may be provided at various levels, fully redundant backup and contingency [2]:

- A *truly redundant system* provides the same functionality as the primary system, allowing a seamless transition with no change in procedures.
- A *backup system* ensures continued operation, but not necessarily with the full functionality of the primary system, and it may necessitate some changes in procedures by the user.
- A *contingency system* allows safe completion of a maneuver, but it may not be adequate for long-term use.

**A.3 References**


NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY
Appendix B  Applications and Their Requirements

This Appendix briefly summarizes various applications and their requirements, and it is sorted by modality. The focus is on those applications of which the navigation requirements can be fulfilled by either the current Loran-C system, or potentially, by a future eLoran and when the usage of Loran-C is beneficial from an economical, environmental, or safety perspective. The list should not be considered comprehensive, merely indicative. Also, many of the requirements listed are subject to continuous change as function of what is technically achievable and what is desirable from a safety and cost perspective.

B.1 Maritime

Maritime requirements are defined by professional international bodies such as the International Maritime Organization (IMO) and by national organizations such as the United States Coast Guard (USCG).

Loran-C was originally developed to provide radionavigation service for coastal waters and provides better than 0.25 nautical mile absolute accuracy for suitably equipped users within the published areas. A modernized Loran-C can significantly improve those figures, allowing the system to be used for more challenging applications such as harbor entrance.

The USCG “Harbor Entrance and Approach” (HEA) procedure has been used throughout this dissertation as an example of an application with highly demanding requirements:

<table>
<thead>
<tr>
<th>Performance requirement</th>
<th>value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Accuracy (back-up)</td>
<td>20 m 95%</td>
</tr>
<tr>
<td>Monitor / alert limit (backup)</td>
<td>50 m</td>
</tr>
<tr>
<td>Integrity (target)</td>
<td>$3 \times 10^{-5}$/hour</td>
</tr>
<tr>
<td>Time-to-alert</td>
<td>10 sec</td>
</tr>
<tr>
<td>Availability (minimum)</td>
<td>99.7%</td>
</tr>
<tr>
<td>Continuity (minimum)</td>
<td>99.85% over 3 hours</td>
</tr>
</tbody>
</table>
The most challenging figure is the accuracy requirement of 20 m, with 95% confidence. The accuracy requirement is usually defined as a Circular Error Probability or CEP. Such a figure of merit is very useful in considering large-scale, general-accuracy questions such as those involved in general Loran chain planning. However, sometimes, the requirements are too demanding for general metrics. In those situations, it can be more suitable to concentrate on specific requirements, which in a maritime context often relates to the cross-track error. An example of such a requirement is that the cross-track error of the navigation system should be 5 meters less than 20% of the channel width – 99% of the time, which was suggested in the St. Lawrance Seaway study [2]. Such a more specific requirement is especially useful for Loran-C, which most often shows an error ellipse rather than a circle.

Table B-2: IMO A.953(23)

<table>
<thead>
<tr>
<th></th>
<th>Absolute accuracy</th>
<th>Time to alarm</th>
<th>Availability</th>
<th>Continuity</th>
<th>Coverage</th>
<th>Update rate*</th>
</tr>
</thead>
<tbody>
<tr>
<td>Harbor entrances, harbor approaches, coastal waters with a high volume of traffic and/or significant degree of risk</td>
<td>10 m, 95%</td>
<td>10 sec</td>
<td>99.8%</td>
<td>99.97%</td>
<td>Adequate to provide position-fixing throughout this phase of navigation</td>
<td>&gt; 0.1 Hz</td>
</tr>
<tr>
<td>Harbor entrances, harbor approaches, coastal waters with a low volume of traffic and/or a less significant degree of risk</td>
<td>10 m, 95%</td>
<td>10 sec</td>
<td>99.5%</td>
<td>99.85%</td>
<td>Adequate to provide position-fixing throughout this phase of navigation</td>
<td>&gt; 0.1 Hz</td>
</tr>
<tr>
<td>Ocean waters</td>
<td>100 m, 95%</td>
<td>As soon as practicable by MSI** systems</td>
<td>99.8%</td>
<td>per 30 days</td>
<td>Adequate to provide position-fixing throughout this phase of navigation</td>
<td>&gt; 0.1 Hz</td>
</tr>
</tbody>
</table>

* If the computed data is used for AIS, graphical display or for direct control of the ship, then the update rate should be greater than once every 2 sec.
** Maritime Safety Information
The IMO has adopted two resolutions that specify the requirements for future radionavigation systems in a maritime environment.

Resolution A.953(23) gives the formal requirements and procedures for accepting new systems as components of the World-Wide Radionavigation System (WWRNS) and is depicted in Table B-2. Resolution A.915(22) sets the requirements for future developments of GNSS to be considered within the framework of A.953(23).

Other use of Loran-C in a maritime context can include its compass functionality. Accurate heading information is, for example, increasingly used to enable chart overlay on radar imagery. The requirements of the heading sensor in that application are related more to the user's convenience than to safety. Modern high-performance heading sensors usually provide heading accuracies better than 1° at an update rate of 10 Hz.

**B.2 Aviation**

The FAA has approved Loran as a supplement system in the NAS (National Airspace System) for en-route and terminal phases of flight.

The applicability of a future eLoran for Non Precision Approach (NPA) has been researched by the LORIPP-team [1]. In a Non-Precision Approach, the navigation system provides guidance in the horizontal plane only. The pilot uses this to maneuver the aircraft onto the final approach track. Descent, in accordance with the approach procedure, is controlled using the measure of distance from the touch-down point provided by the navigation system, by
reference to a pressure altimeter. The term “non-precision” indicates that no descent guidance is available from the navigation system itself [5].

This LORIPP team concluded that the requirements for NPA are RNP 0.3, which equals to the following:

### Table B-4 Aviation RNP 0.3 Requirements [1]

<table>
<thead>
<tr>
<th>Performance requirement</th>
<th>value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Accuracy (target)</td>
<td>307 m</td>
</tr>
<tr>
<td>Monitor limit (HPL) (target)</td>
<td>556 m</td>
</tr>
<tr>
<td>Integrity</td>
<td>$10^7$/hour</td>
</tr>
<tr>
<td>Time-to-alert</td>
<td>10 seconds</td>
</tr>
<tr>
<td>Availability (minimum)</td>
<td>99.9%</td>
</tr>
<tr>
<td>Availability (target)</td>
<td>99.99%</td>
</tr>
<tr>
<td>Continuity (minimum)</td>
<td>99.9%</td>
</tr>
<tr>
<td>Continuity (target)</td>
<td>99.99%</td>
</tr>
</tbody>
</table>

* HPL is horizontal protection limit

Furthermore, the integrity demand of “seven nines” has been identified as the most challenging requirement, and that fulfillment of this requirement is possible assuming various upgrades to the Loran-C system.

### B.3 Land mobile

Loran-C can be useful for various land-mobile applications, especially if there is a safety, environmental, or economic benefit in having a system available when a GNSS outage occurs.

Examples of potential land-mobile usage of Loran-C include:
- Car navigation, for example backup navigation for rescue vehicles
- Track-and-trace
- Tracking hazardous cargo

In general, the applicability of eLoran increases with its performance. Although meter-level absolute eLoran positioning will be most interesting, a significantly less performing system can be still viable in this modality.
B.4 Time and frequency

The timing-and-frequency users have no known published government requirements that equipment must meet. However, timing-and-frequency applications, including those used by government agencies, employ applications with specific timing-and-frequency requirements.

Frequency performance is usually expressed in Stratum levels:

**Table B-5: Definition of the Stratum Levels, frequency expressed in Hz [6]**

<table>
<thead>
<tr>
<th>Stratum</th>
<th>Frequency accuracy</th>
<th>Frequency stability</th>
<th>Timing accuracy</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>±1.0 x 10^{-11}</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>2</td>
<td>±1.6 x 10^{-8}</td>
<td>1.0 x 10^{-10}/day</td>
<td>N/A</td>
</tr>
<tr>
<td>3</td>
<td>±4.6 x 10^{-6}</td>
<td>3.7 x 10^{-7}/day</td>
<td>N/A</td>
</tr>
<tr>
<td>4</td>
<td>±32 x 10^{-6}</td>
<td>Same as accuracy</td>
<td>N/A</td>
</tr>
</tbody>
</table>

*Frequency accuracy* is defined as the maximum long-term deviation from the definition of the second without external calibration. This is measured as the frequency difference from a recognized and maintained source, for example, the U.S. Naval Observatory (USNO) or le Bureau International des Poids et Mesures (BIPM).

*Frequency stability* is defined as the change in frequency over a given time interval.

*Timing accuracy* is defined as the absolute offset in time from a recognized and maintained time source (USNO, BIPM, etc.).

Figure B-1 and Figure B-2 depict various applications as a function of the performance requirements [1]. Note that these user surveys are not intended to be complete representations of all users. The requirements have been generalized and averaged over the user groups.
Based on the DOT Task Force Report [7], the following performance metrics are desired for eLoran to have significant potential as a primary or backup time and frequency source [1]:
### B.5 References


[3] “Revised maritime policy and requirements for a future global navigation satellite system (GNSS)”, IMO Resolution A.915(22), adopted on 29 November 2001


---

**TABLE B-6: Time and Frequency requirements [1]**

<table>
<thead>
<tr>
<th>Performance requirement</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Accuracy (target)</td>
<td>$1 \times 10^{-13}$ averaged over 24 hours</td>
</tr>
<tr>
<td>Frequency Accuracy (desired)</td>
<td>$1 \times 10^{-12}$ averaged over 6 hours</td>
</tr>
<tr>
<td>Frequency Accuracy (minimum)</td>
<td>$1 \times 10^{-11}$ averaged over 1 hours</td>
</tr>
<tr>
<td>Antenna</td>
<td>No external antenna (desired)</td>
</tr>
<tr>
<td>Legacy use</td>
<td>Backward compatible (desired)</td>
</tr>
<tr>
<td>Integrity data</td>
<td>Minimum “use/no use” flag</td>
</tr>
<tr>
<td>Timing data</td>
<td>Time tag, leap second info</td>
</tr>
<tr>
<td>Timing accuracy at the user’s receiver</td>
<td>&lt; 100 ns (RMS)</td>
</tr>
<tr>
<td>Differential data update rate</td>
<td>&lt; once /hour</td>
</tr>
</tbody>
</table>
### Appendix C  Timeline

<table>
<thead>
<tr>
<th>Date</th>
<th>Event Description</th>
</tr>
</thead>
</table>
| 14 May 2001        | Master thesis: H-field antenna design  
master thesis: H-field antenna design  
**Orthogonal design process of LF H-field antenna and associated amplifier** |
| 2001               | H-field antenna research  
- E-field susceptibility  
- prototype antenna  
**Development of highly effective E-field shielding** |
| February 2002      | Start PhD research at Delft University                                              |
| April 2002         | LORADD “DataGrabber” receiver + antenna  
**A Loran-C research receiver is born**                                             |
| September 2002     | Paris measurement campaign  
**Loran-C positioning inside an underground parking garage**                         |
| October 2002       | Spatial Domain Local Interference Cancellation                                      |
| February 2003      | **Automatic local interference cancellation based on distance towards interferer**  |
| April 2003         | Reeuwijk measurements  
**Analysis of the spatial and temporal repeatability of re-radiation**             |
| April 2003         | Best presentation award for “Loran-C Challenges GNSS: From a Quarter Nautical Mile Down to Meter-Level Accuracy” at the ENC-GNSS 2003  
**Loran beats GPS with best presentation at a ‘GNSS’ conference**                      |
| June 2003          | Boston measurements – I                                                            |
| June 2003          | Loran-C receiver development, e.g.:  
- Cycle identification  
- Total Pulse tracking                                                             |
| September 2003     | **H-field antenna processing**  
- beam-steering  
- Cross-talk measurement & compensation  
**Successful mitigation of H-field heading dependent errors**                         |
| October 2003       |                                                                                   |
December 2003

Boston measurements – II
Accurate simultaneous H-field and E-field land-mobile Loran-C measurements

April 2004
Tampa Bay measurements
- Unprecedented Loran-C measurement performance
- Highly accurately synchronized H-field and E-field measurements
- Successful mitigation of all H-field antenna related errors, including the influence of the vessel

May 2004
Invited speech “The Integration Game” during the opening session of the ENC-GNSS 2004

July 2004
H-field antenna cross-talk analysis

April 2005
Reelektronika LORADD
- firmware 1.0 finished, first LORADDs shipped to customers
High performance integrated eLoran-GPS receiver

July 2005
Research at Ohio University

December 2005

December 2005
Ohio University flight trials

March 2006
Reelektronika LORADD firmware 2.0

April 2006
Harwich measurements
Demonstration of the potential of eLoran on European soil

May 2006
Further enhancements H-field calibration and re-radiation detection

November 2006
PhD defense
“New Potential for Low Frequency Radionavigation in the 21st Century”
Appendix D  Related Publications and Presentations by the Author

D.1 Primary author


W.J. Pelgrum, "The Integration Game", Invited presentation at the ENC-GNSS, Rotterdam, the Netherlands, May 2004.

W. J. Pelgrum, A.W.S. Helwig, D. van Willigen, E. Johannessen, A. Grebnev, “eLoran For Harbor Entrance and Approach – the Tampa Bay Trials”, Presented at the ENC-GNSS, Rotterdam, the Netherlands, May 2004
NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY


W.J. Pelgrum, “Noise – From a Receiver Perspective”, Proceedings of the 34th Annual Convention and Technical Symposium, Santa Barbara, California, USA, October 2005

D.2 Co-author


A.W.S. Helwig, W.J. Pelgrum, G.W.A. Offermans, “Improving Availability through Integration of GNSS and Loran-C”, Presented at the ENC-GNSS, Rotterdam, the Netherlands, May 2004


Wouter Johan Pelgrum was born in Bussum, The Netherlands, on October 18 1976. He received his M.Sc. degree in Electrical Engineering cum laude from Delft University of Technology in 2001, for his research on an H-field antenna for low-frequency radionavigation systems. Wouter Pelgrum started his Ph.D. research at Delft University in 2002 on low-frequency radionavigation. From 2001 until 2006 he worked for Reelektronika, a company specialized in the research and design of integrated navigation, for which he contributed to the development of an integrated GPS-eLoran receiver.

De steeds verder verbeterende kwaliteit van GPS heeft lange tijd voeding gegeven aan de gedachte dat GPS – en GPS alleen – de toekomst zou zijn van radioplaats- en tijdsbepaling. De in 2001 uitgebrachte Volpe-studie, en later het in 2004 voorgestelde ERNP (European Radio Navigation Plan), bevatten een ander toekomstbeeld. Hoewel GPS en andere plaatsbepalingsystemen op basis van satellieten heel nauwkeurig zijn, worden ze onvoldoende betrouwbaar geacht om gebruikt te kunnen worden als enige systeem voor die applicaties waarbij veiligheid, economie en/of het milieu in het geding zijn. Voor die applicaties is een backupsysteem nodig waarvan de foutkarakteristieken verschillen van die van GNSS.

De door het Volpe-rapport en ERNP-voorstel gesuggereerde oplossing is – wellicht verrassend – een oud en vrijwel vergeten radioplaatsbepalingssysteem: Loran-C. Dit systeem verschilt in grote mate van GPS door het gebruik van laagfrequente pulsen met hoge energie. De combinatie van Loran-C en GPS heeft daarom de potentie om veel robuuster te zijn dan ieder van de systemen afzonderlijk. Echter, de “officiële” specificaties van het uit 1958 stammende Loran-C systeem zijn niet toereikend om te voldoen aan de zware eisen van de meeste moderne applicaties.

Gelukkig zijn de officiële specificaties van het systeem gebaseerd op de beperkte mogelijkheden van inmiddels achterhaalde technologie en niet op wat fundamenteel mogelijk is met laagfrequente radioplaatsbepaling. Dit leidt tot de volgende vraag:

**Wat zijn de fundamentele limieten van laagfrequente radioplaatsbepaling en hoe hebben die invloed op mogelijke toepassingen?**

Loran-C is momenteel het enige beschikbare, operationele en publiek toegankelijke laagfrequente radioplaatsbepaling systeem. Daarom richt deze dissertatie zich primair op Loran-C. De meeste resultaten zijn echter ook toepasbaar op andere laagfrequente plaatsbepalingsystemen. Hoofdstuk 2 introduceert de systeemeigenschappen van Loran-C.

De zoektocht naar de fundamentele limieten begint bij het identificeren van de mogelijke foutbronnen. Hoofdstuk 3 bevat een grondige systeemanalyse die achtereenvolgens de zender, de signaalpropagatie, de antenne, de ontvangeralgoritmen en de uiteindelijk berekende plaats en tijd behandelt.
Laagfrequente grondgolven ondergaan een vertraging als functie van de grondgeleidbaarheid, de topografie, de seizoenen en het weer. Indien er niet voor deze vertragingen gecompenseerd wordt, kunnen significante positiefouten ontstaan. Hoofdstuk 4 bespreekt het gebruik van een differentieel referentiestation om de tijdafhankelijke vertragingen te compenseren. Door het gebruik van een plaatsafhankelijke correctietabel worden de propagatiegerelateerde positiefouten nog verder gereduceerd. De resulterende positienauwkeurigheid is potentiëel voldoende voor bijvoorbeeld de stringente nauwkeurigheidseis van 20 meter met een betrouwbaarheid van 95% van de maritieme Harbor Entrance and Approach (havennadering) procedure.


Deze dissertatie sluit af met een beoordeling van het potentiële van laagfrequente plaatsbepaling, gebaseerd op de resultaten van het promotieonderzoek gecombineerd met de persoonlijke visie van de promovendus.

Wouter J. Pelgrum
Delft, november 2006
Acknowledgements

I would like to acknowledge the following persons and organizations:

Henk, Bea and Michiel, for their continuous support, for their sound advice, and for being there when I needed it most.

Prof. Ligthart, for providing a place at IRCTR to do my research, for getting and keeping me on track to actually finish the Ph.D., and for his interest in the person behind the scientist.

Durk, for introducing me into the field of radionavigation, and for his endless enthusiasm for science.

Frank, for his hospitality at Ohio University, for providing the ambiance needed to seriously start writing this dissertation, and for facilitating moments of reflection with regard to my personal and professional live.

GAUSS Research Foundation, for funding my Ph.D. research.

Reelektronika, that is, Arthur, Gerard, Rene and Durk, with whom I formed a team that has been able to make a difference in the world of radionavigation. Arthur, for sharing the numerous moments of frustration during the measurement campaigns as well as the successive moments of joy when everything still worked out in the end.

Megapulse, for their support of the measurement campaigns in Boston and at Tampa Bay.

Sanne, for designing and applying the layout of the dissertation.

My friends, for cycling, rowing, running, talking and just relaxing together.
NEW POTENTIAL OF LOW-FREQUENCY RADIONAVIGATION IN THE 21ST CENTURY