

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

BEFORE THE PATENT TRIAL AND APPEAL BOARD

ERICSSON INC.
NOKIA OF AMERICA CORPORATION F/K/A ALCATEL-LUCENT USA INC.
Petitioners,

v.

XR COMMUNICATIONS LLC
Patent Owner.

U.S. PATENT NO. 7,177,369
Title: MULTIPATH COMMUNICATION METHODS AND APPARATUSES

Inter Partes Review No.: IPR2024-00314

DECLARATION OF INGRID HSIEH-YEE, PH.D.

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I. Introduction

I, Ingrid Hsieh-Yee, Ph.D., do hereby declare as follows:

1. I have personal knowledge of all of the matters about which I testify in this declaration.

2. I have been retained as an independent expert witness on behalf of Nokia and Ericsson (“NOKIA & ERICSSON”) for proceedings before the United States Patent and Trademark Office (“PTO”) regarding the authenticity and public availability of certain prior art references in relation to US Patent No. 7,177,369 (“the ’369 patent”).

3. I am being compensated for my work in this matter at my customary hourly rate. I am also being reimbursed for any reasonable expenses associated with my work and testimony in this investigation. My compensation is not contingent on the results of my study, the substance of my opinions, or the outcome of this matter.

A. Qualifications and Professional Experience

4. My complete qualifications and professional experience are described in my academic curriculum vitae (**Appendix A**). The following is a brief summary of my relevant qualifications and professional experience.

5. I was a professor of the Department of Library and Information Science at the Catholic University of American for 32 years. I retired in August 2022 and was conferred the title of Professor Emerita in December 2022. I have experience

working in an academic library, a medical library, and a legislative library. I hold a Ph.D. and a Masters in Library and Information Studies from the University of Wisconsin-Madison.

6. I am an expert on library cataloging and classification and have published two editions on this subject, *Organizing Audiovisual and Electronic Resources for Access: A Cataloging Guide* (2000 and 2006). I taught a variety of courses, including Cataloging and Classification, Advanced Cataloging and Classification, Organization of Internet Resources, Organization of Information, Digital Content Creation and Management, Metadata, Internet Searches and Web Design, Information Literacy Instruction, Advanced Information Retrieval and Analysis Strategies, and The Information Professions in Society. I am familiar with metadata schema design and implementation. In my teaching I covered the design and implementation of metadata in libraries, databases, search engines, digital repositories, digital libraries, and digital archives. I also covered how information organization affects the discovery and access to digital resources on the Internet. My research interests cover cataloging and classification, information organization, metadata, information retrieval, information architecture, digital collections, scholarly communication, social media, user interaction with information systems, and others.

7. I am very familiar with a library cataloging encoding standard known as the “Machine-Readable Cataloging” standard, also known as “MARC,” which became the national standard for sharing bibliographic data in the United States by 1971 and the international standard by 1973. MARC is the primary communications protocol for the transfer and storage of bibliographic metadata in libraries. It is widely accepted in my field to rely upon MARC records to determine whether and when a publication would have been publicly accessible.

8. A MARC record consists of several fields, each of which contains specific data about the work. Each field is identified by a standardized, unique, three-digit code corresponding to the type of data that follows. **Appendix B** is a true and correct copy of Parts VII to X of “Understanding MARC Bibliographic: Machine-Readable Cataloging” (<http://www.loc.gov/marc/umb/um07to10.html>) from the Library of Congress that explains commonly used MARC fields. For example, the personal author of a work is recorded in Field 100, the title is recorded in Field 245, publisher information is recorded in Field 260, the physical volume and characteristics of a publication are recorded in Field 300, and topical subjects are recorded in Fields 650.

9. I am very familiar with library cataloging and processing practices and the relationships between OCLC (Online Computer Library Center) and library

catalogs, and how libraries create and share their records to help the public discover and access information.

10. The Online Computer Library Center (OCLC) is the largest bibliographic network of the world, with more than 548 million records and thousands of member institutions (many of which are libraries of some type) in more than 100 countries. OCLC was founded in 1967 to promote and support library cooperation. According to the “Third Article, Amended Articles of Incorporation of OCLC Online Computer Library Center, Inc.,” OCLC was created “to establish, maintain and operate a computerized library network and to promote the evolution of library use, of libraries themselves, and of librarianship, and to provide processes and products for the benefit of library users and libraries, including such objectives as increasing availability of library resources to individual library patrons and reducing the rate of rise of library per-unit costs, all for the fundamental public purpose of furthering ease of access to and use of the ever-expanding body of worldwide scientific, literary and educational knowledge and information” (source: <https://www.oclc.org/content/dam/oclc/membership/articles-of-incorporation.pdf>).

11. OCLC members can contribute original cataloging records in MARC to the system or derive cataloging records from existing records, an activity referred to as “copy cataloging.” When an OCLC participating institution acquires a work, it can create an original MARC record for the work in OCLC’s Connexion system (a

system for catalogers to create and share MARC records), and the system will automatically generate a code for the date of record creation in the yymmdd format, and the creating library's OCLC symbol is recorded in subfield "a" of the 040 field. Once the MARC record is in Connexion, it becomes available to other OCLC members for adoption to their local online catalogs (*i.e.*, copy cataloging).

12. After a MARC record is created in Connexion, it also becomes searchable and viewable on WorldCat, which is a free web portal for users to explore more than 10,000 libraries worldwide. Records in WorldCat, however, are not presented in MARC fields. Instead, the data elements are labeled to help users interpret the records. Thus, the information stored in MARC records in Connexion is available to the interested public through the user-friendly WorldCat web portal.

13. WorldCat (<http://www.worldcat.org>) is "the world's largest network of library content and services" and its features are summarized in "What is WorldCat" (<https://www.worldcat.org/about>). WorldCat began operation in January 1998 as a free web portal to more than 10,000 libraries worldwide. Through WorldCat, users can search for information resources in their local libraries and libraries around the world. WorldCat allows users to search for books, CDs, videos, and many new types of digital content, such as audiobooks, in many languages. Users can also retrieve research materials and article citations with links to their full text. After an item is retrieved, WorldCat helps users identify a library nearby that holds the item or all

the libraries that hold the item. WorldCat is an efficient way to explore the content held by more than 10,000 libraries around the world. WorldCat is publicly accessible, and there is no fee to perform a search on WorldCat.

14. Library online catalogs gained acceptance in the early 1980s and many libraries migrated their systems to the World Wide Web in the mid-1990s. Library online catalogs are based on MARC records that represent their collections in order to help the public understand what materials are publicly accessible in those libraries. Most libraries with online catalogs have made their catalogs freely available on the Web. These online catalogs offer user-friendly search interfaces. Strong user interest in keyword searches and the popularity of Google have led to the “googlization” of library search systems. As a result, many library catalogs now provide a single search box for users to conduct keyword searches, with additional support for searches by author, title, subject terms, and other data elements such as ISBN (International Standard Book Number). Library catalogs these days also offer features for users to narrow their search results by language, year, format, and other elements. Many libraries display MARC records on their online catalogs with labels for the data elements to help the public interpret MARC records. Many libraries also offer the option to display MARC records in MARC fields.

15. Libraries create MARC records for works they acquire, including books, serials, motion pictures, and publications in other formats. MARC records

are representations of materials held by libraries and are critical for the discovery and access to those materials. Monograph cataloging is fairly common in libraries, and most libraries make a newly cataloged monograph available to the public soon after the cataloging work is completed, usually within a week. Libraries can create original cataloging records or use an existing record in OCLC to create a copy cataloging record. As soon as the cataloging record is completed, it is added to the library's online catalog for users. If the record is an original record, it is also entered into OCLC WorldCat. If it is a copy cataloging record, the library's holding symbol is attached to the existing original record in OCLC WorldCat to facilitate searching and interlibrary loan.

16. The cataloging of serials and the serial check-in process are discussed here to show how libraries usually provide access to newly received serial issues. According to the glossary of the *RDA: Resource Description and Access* cataloging standard, a serial is “a mode of issuance of a manifestation issued in successive parts, usually bearing numbering, that has no predetermined conclusion. A serial includes a periodical, monographic series, newspaper, etc.” Because the publisher of a serial makes new issues of the serial available successively, a customary cataloging practice is to create one bibliographic record for the serial, and the serial record (encoded in MARC) typically provides information on the beginning date and frequency of the serial, not the dates of individual issues. In other words, libraries

typically do not create MARC records for individual issues of a serial. Instead, they rely on a serial check-in system to track the receipt of new issues. A common check-in practice is to date stamp a new issue when it arrives. This practice has become automated since the late 1990s, and libraries now vary in how they share the receipt date of a new serial issue with the public. Some libraries use a date stamp, some affix a label to indicate the receipt date, some pencil in the receipt date, and some do not provide the information to the public.

17. The serial check-in process usually takes less than an hour, and one of the steps involves placing a date stamp on the new issue to document the date the issue is checked in. After that, the holdings information of the serial is updated in the library's catalog so that users know which issues are available for request or access. After serial check-in is completed, the new issue is placed on the shelf with the previous issues of the serial. Libraries with a public periodical room typically place new issues in the periodical room for easy user access. Because information presented in serials often reflects latest discovery, a general practice of libraries is to make new issues of serials available for user access soon after they are checked in, usually within a week.

18. I am personally familiar with many online catalogs, databases, and search engines. In preparing for this declaration, I used the following authoritative information systems to search for records:

- Google Scholar (<https://scholar.google.com>)
- Internet Archive (<https://archive.org/>)
- Online catalog of the Linda Hall Library (https://catalog.lindahall.org/discovery/search?vid=01LINDAHALL_INST:LHL)
- Online catalog of the University of Iowa Libraries (<https://search.lib.uiowa.edu/primo-explore/search?vid=01IOWA>)
- Online catalog of the Wisconsin Library System (<https://search.library.wisc.edu/search/system>)
- Semantic Scholar (<https://www.semanticscholar.org/>)
- WorldCat, the free web portal (<https://www.worldcat.org>)

B. Scope of This Declaration

19. I have been asked to offer opinion on the authenticity and public availability date of the following documents:

- (1) Wong, C. Y., Cheng, R. S., Lataief, K. B., & Murch, R. D. (1999, October). Multiuser OFDM with adaptive subcarrier, bit, and power allocation. *IEEE Journal on Selected Areas in Communications*, vol. 17, no. 10, pp. 1747-1758, **Ex. 0001** (“Wong”);
- (2) Minn, H., & Bhargava, V. K. (2000, December). An investigation into time-domain approach for OFDM channel estimation. *IEEE*

Transactions on Broadcasting, vol. 46, no. 4, pp. 240-248, **Ex. 0002**

(“Minn”);

(3) Yeh, C. S., & Lin, Y. (1999, December). Channel estimation using pilot tones in OFDM systems. *IEEE Transactions on Broadcasting*, vol. 45, no. 4, pp. 400-409, **Ex. 0003** (“Yeh”);

(4) Lehne, H., & Pettersen, M. An overview of smart antenna technology for mobile communications systems, *IEEE Communications Surveys*, Fourth Quarter 1999, vol. 2, no. 4, pp. 2-13, **Ex. 0004** (“Lehne”).

20. I am rendering my expert opinion on the authenticity of the document referenced herein and on whether it was published. I will provide my opinion on when and how this document was publicly disseminated or otherwise made publicly available to the extent that persons interested and ordinarily skilled in the subject matter or art, exercising reasonable diligence, could have located and retrieved this document.

21. I am informed by counsel that an item is considered authentic if there is sufficient evidence to support a finding that the item is what it is claimed to be. I am also informed that authenticity can be established based on the contents of the documents themselves, such as the appearance, content, substance, internal patterns, protocols followed or other distinctive characteristics of the item.

22. I am informed by counsel that a given reference is “published” if it is publicly accessible upon a satisfactory showing that such a document has been disseminated or otherwise made publicly available to the extent that persons interested and ordinarily skilled in the subject matter or art could locate it exercising reasonable diligence and obtain the document. I have also been informed by counsel that materials available in a library constitute “printed publications” if they are cataloged and indexed according to general library practices and protocols that make the references available and accessible to members of the interested public.

23. I have been informed by counsel that the ’369 patent was filed on April 25, 2002, in the United States, and the earliest priority date of the ’369 patent is April 25, 2001. I have assessed public accessibility of the documents referenced below as of this claimed priority date.

C. Person of Ordinary Skills in the Arts

24. I am told by counsel that the subject matter of this proceeding relates to wireless communication such as transmissions between a wireless base station and a subscriber’s wireless device.

25. I have been informed by counsel that a “person of ordinary skill in the art at the time of the inventions” (POSA) is a hypothetical person who is presumed to be familiar with the relevant field and its literature at the time of the inventions.

This hypothetical person is also a person of ordinary creativity, capable of understanding the scientific principles and literature applicable to the pertinent field.

26. I am told by counsel that persons of ordinary skill in this subject matter or art would have had at least a Bachelor's degree in Electrical Engineering or an equivalent field, and at least two years of work experience in developing OFDM-based wireless communications. Alternatively, a POSITA would have had a more advanced degree, such as a Master's degree in Electrical Engineering or an equivalent field, combined with at least one year of work experience in developing OFDM-based wireless communications. More education can supplement practical experience and vice versa.

27. It is my opinion that such a person would have been actively engaged in learning about the field, possibly through formal instruction through the bibliographic resources. By at least April 25, 2001, such a person would have had access to a vast array of print and electronic resources, including at least the documents referenced below.

D. Evidence Considered in Forming My Opinions

28. In the preparation of this declaration, I have reviewed the documents referenced below and any other documents I reference herein, and each of these is a type of material that experts in my field would reasonably rely upon when forming their opinions:

- (1) The document referenced above in Section I.B;
- (2) Library copy of *Wong*, obtained from Linda Hall Library, **Appendix 0001A**;
- (3) Bibliographic and MARC records for *IEEE Journal on Selected Areas in Communications* that contains *Wong*, available at https://catalog.lindahall.org/permalink/01LINDAHALL_INST/1jnv6m8/alma992566013405961 from the online catalog of the Linda Hall Library, accessed and obtained on October 31, 2023, **Appendix 0001B**;
- (4) Library copy of *Wong*, obtained from the University of Iowa Library, **Appendix 0001C**;
- (5) Bibliographic and MARC records for *IEEE Journal on Selected Areas in Communications* that contains *Wong*, available at https://search.lib.uiowa.edu/permalink/f/9i2ftm/01IOWA_ALMA21352900050002771 from the online catalog of the University of Iowa Library, accessed and obtained on October 31, 2023, **Appendix 0001D**;
- (6) Citations to *Wong*, obtained from Google Scholar, **Appendix 0001E**;
- (7) Library copy of *Minn*, obtained from Linda Hall Library, **Appendix 0002A**;
- (8) Bibliographic and MARC records for *IEEE Transactions on Broadcasting* that contains *Minn*, available at https://catalog.lindahall.org/permalink/01LINDAHALL_INST/191da7s/alma992718793405961 from the online catalog of the Linda

Hall Library, accessed and obtained on October 27, 2023, **Appendix 0002B**;

- (9) Library copy of *Minn*, obtained from the University of Wisconsin-Madison Library, **Appendix 0002C**;
- (10) Bibliographic and MARC records for *IEEE Transactions on Broadcasting* that contains *Minn*, available at <https://search.library.wisc.edu/catalog/999482629302121> from the online catalog of the University of Wisconsin-Madison Library, accessed and obtained on October 27, 2023, **Appendix 0002D**;
- (11) Library copy of *Yeh*, obtained from Linda Hall Library, **Appendix 0003A**;
- (12) Bibliographic and MARC records for *IEEE Transactions on Broadcasting* that contains *Yeh*, available at https://catalog.lindahall.org/permalink/01LINDAHALL_INST/191da7s/alma992718793405961 from the online catalog of the Linda Hall Library, accessed and obtained on October 27, 2023, **Appendix 0003B**;
- (13) Library copy of *Yeh*, obtained from the University of Wisconsin-Madison Library, **Appendix 0003C**;
- (14) Bibliographic and MARC records for *IEEE Transactions on Broadcasting* that contains *Yeh*, available at <https://search.library.wisc.edu/catalog/999482629302121> from the online catalog of the University of Wisconsin-Madison Library, accessed and obtained on October 27, 2023, **Appendix 0003D**;
- (15) Copy of *Lehne* archived on August 18, 2000, by the Internet

Archive, the archived HTML version is available at https://web.archive.org/web/20000818222434fw_/http://www.comsoc.org/pubs/surveys/4q99issue/lehne.html, the file is also available on the website of the journal *IEEE Communications Surveys & Tutorials* that was archived on August 15, 2000, by the Internet Archive (follow the “Quarterly Issues” link, then select the “Fourth Quarter 1999” link to reach “Fourth Quarter 1999, Vol. 2, No. 4” and select the *Lehne* article to view the article on the journal website), **Appendix 0004A**,

- (16) Number of citations to *Lehne*, obtained from Google Scholar and Semantic Scholar, and a citation to *Lehne* published in 2000, **Appendix 0004B**;
- (17) Affidavit of Mr. Nathaniel E. Frank-White, **Ex. 0005**.

These records are identified and discussed in this declaration. Experts in the field would reasonably rely on the data described herein to form their opinions.

II. Authenticity and Public Availability of *Wong* (Ex. 0001)

A. Authentication

29. **Ex. 0001** is a true and correct copy of “Multiuser OFDM with adaptive subcarrier, bit, and power allocation,” (“*Wong*”), by Wong et al., *IEEE Journal on Selected Areas in Communications*, vol. 17, no. 10, pp. 1747-1758, that I obtained from the IEEE Xplore Digital Library. When I began preparing this declaration, I searched Google Scholar for *Wong* for records, and a record link led me to an IEEE

metadata record for this article at <https://ieeexplore.ieee.org/abstract/document/793310>. I then followed the PDF link in the record to purchase this article from the Institute of Electrical and Electronics Engineers (“IEEE”). This IEEE copy is presented as **Ex. 0001** in this declaration.

30. **Ex. 0001** is a true and correct copy of *Wong* obtained from the publisher IEEE. Page 1 shows the title, authors, abstract, and index terms (“Adaptive modulation, frequency selective fading channel, multiaccess communication, multiuser channel, orthogonal frequency division multiplexing (OFDM), resource management”). It also shows the manuscript was “received October 15, 1998; revised March 27, 1999,” the Publisher Item Identifier is “S 0733-8716(99)07861-0,” *Wong* appears in a journal whose ISSN (International Standard Serial Number) is “0733-8716,” the article has a “1999” copyright date and IEEE is the copyright holder. Page 1 (internal page 1747) and even-numbered pages show a header that identifies the journal issue as “IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, VOL. 17, NO. 10, OCTOBER 1999.” **Ex. 0001** shows *Wong* has a total of 12 pages, including six sections, eight figures, 23 equations, and 18 references.

31. **Ex. 0001** is in a condition that creates no suspicion about its authenticity. Specifically, this copy is not missing any intermediate pages, the text on each page appears to flow seamlessly from one page to the next, and there are no

visible alterations to the document. Moreover, **Ex. 0001** was found in the custody of the publisher, a place where, if authentic, it would likely be found. I therefore see no reason to question the authenticity of this IEEE copy of *Wong* (**Ex. 0001**).

B. Linda Hall Library Copy

32. To determine if *Wong* was available in libraries, I searched WorldCat by the ISSN of the journal containing *Wong*, and found a record at <https://search.worldcat.org/title/8649040> that shows Linda Hall Library is one of the 830 holding libraries of this journal. After verifying the holdings information in their online catalog, I requested a copy of *Wong* from the library's Document Delivery Service. This library copy is presented as **Appendix 0001A** in this declaration.

33. **Appendix 0001A** is a true and correct copy of *Wong* obtained from Linda Hall Library. Pages 1 to 12 (internal pages 1747 to 1758) are the full text of *Wong*. I have closely compared this library copy with the IEEE copy (**Ex. 0001**) and found them to have the same content. Page 13 is the cover that identifies this issue as volume 17, number 10 of *IEEE Journal on Selected Areas in Communications*, which is a publication of the IEEE Communications Society. This page also shows the issue has an "October 1999" issue date, the journal's CODEN (a six-character alphanumeric code that provides unique and concise identifications of serials and non-serial publications) is ISACEM, and the journal's ISSN is "0733-8716." The

cover shows this issue is about “Broadband Wireless Techniques” and “Multiuser OFDM with Adaptive Subcarrier, Bit, and Power Allocation” by Wong et al. begins on page 1747. The cover carries a library stamp of “Linda Hall Library OCT 07 1999” that indicates this issue was received by the library on October 7, 1999. Page 14 is the copyright page that shows this journal is published monthly by IEEE, and this issue has a “1999” copyright date with IEEE as the copyright holder.

C. Linda Hall Library Records

34. **Appendix 0001B** is a true and correct copy of the bibliographic and MARC records for the journal *IEEE Journal on Selected Areas in Communications* that contains *Wong*. I personally located, identified and obtained these records from the online catalog of the Linda Hall Library. These are the type of records experts in my field would reasonably rely upon when forming their opinions. I used screen captures to produce this appendix.

35. Pages 1 to 3 of **Appendix 0001B** are the bibliographic record that shows this publication is a journal that is available at “Linda Hall Library Closed Stacks - Serials (I.E.E.E. journal on selected areas in communications) and other locations.” Page 2 shows details of this journal, including title, other titles, creators/contributors, subjects, identifiers, publisher, creation date, format, frequency, general notes, and citations/references.

36. Pages 4 and 5 are the MARC record. The first six digits of Field 008

show the record was created on “820802” (*i.e.*, August 2, 1982), and the “c19839999” code following the record creation date indicates this publication is a continuing resource that began publication in 1983 and has no publication end date yet. Subfield “a” of Field 040 shows “NSD” (the National Serial Database Program of the Library of Congress) created the original MARC serial record, and Field 049 shows this journal is in Collection A of “LHL.” According to the *Directory of OCLC Members* (<https://www.oclc.org/en/contacts/libraries.html>), LHL is the OCLC library symbol for Linda Hall Library. Data in Fields 008 and 040 inform my opinion that Linda Hall Library used the original MARC record created by NSD to create their copy cataloging record for this journal.

37. Field 022 of the MARC record shows the ISSN of this journal, Field 030 shows the journal’s CODEN, Field 035 shows the record’s OCLC number is “8649040.” Field 245 shows the title, Fields 246 show three variant titles are included to help users discover this journal, Field 260 shows IEEE of New York has published this journal since 1983, Field 362 shows the journal began in January 1983, and Field 310 shows the journal has published nine issues a year since January 1989. Fields 510 show this journal is fully indexed by four abstracting and indexing services, and selectively indexed by four other services. Fields 710 show IEEE and the parent organization of this journal are included to help users discover this journal.

38. The original MARC record from NSD includes a classification number

in Field 050, but Linda Hall Library organizes their journals by title, so this classification number is not used at this library. At this library, subject of this journal is represented in Field 650 by a Library of Congress subject heading, using “Telecommunication” as the main topical heading, followed by “Periodicals” as a form subheading. This subject representations enables users to search for this journal by subject.

39. This MARC record (**Appendix 0001B**) has made this journal discoverable at Linda Hall Library. Interested users would have been able to discover this work by the journal’s ISSN, CODEN, title, variant titles, corporate contributors, and subject.

40. As discussed earlier, a customary library practice is to process a newly received journal issue through the serial check-in system, and process the new issue promptly to make it available to users, usually within a week after serial check-in. The date stamp of “October 7 1999” on the Linda Hall Library copy of *Wong* (**Appendix 0001A**) informs my opinion that the issue containing *Wong* was received by the library on October 7, 1999, and the physical copy, after processing, would have become publicly accessible no later than October 14, 1999.

D. University of Iowa Library Copy

41. To further determine the public availability date of the journal issue containing *Wong*, I obtained a copy of *Wong* from the University of Iowa Library

through a document delivery service. **Appendix 0001C** is a true and correct copy of *Wong*. Page 1 is the issue cover that carries a date stamp of “The University of Iowa Libraries OCT 07 1999,” page 2 is the copyright page, and pages 3 to 14 are the full text of *Wong*. I have closely compared this copy with the publisher’s copy of *Wong* (**Ex. 0001**) and the Linda Hall Library copy (**Appendix 0001A**), and found these three copies to have the same content for *Wong*. In addition, the cover and copyright page of the University of Iowa Library copy (**Appendix 0001C**) are the same as those in the Linda Hall Library copy (**Appendix 0001A**), and the date stamps show these two libraries received this issue containing *Wong* on October 7, 1999.

E. University of Iowa Library Records

42. **Appendix 0001D** is a true and correct copy of the bibliographic and MARC records for *IEEE Journal on Selected Areas in Communications* that I personally located, identified, and obtained from the online catalog of the University of Iowa Libraries. This is the type of record experts in my field would reasonably rely upon when forming their opinions.

43. **Appendix 0001D** identifies this publication as a journal and shows the same bibliographic details as the Linda Hall Library serial records (**Appendix 0001B**), including the journal’s ISSN, journal title, variant titles, publisher, beginning year of publication, subject representation and corporate contributors. Field 040 of the MARC record shows the University of Iowa Libraries also used the

record originally created by the National Serial Database Program to create their copy cataloging serial record. The MARC record has made this journal discoverable in the online catalog of the University of Iowa Libraries. Interested users would have been able to discover this journal by its ISSN, title, variant titles, corporate contributors and subject.

44. The date stamp of “OCT 07 1999” on **Appendix 0001C** informs my opinion that this issue containing *Wong* was received by the University of Iowa Libraries on October 7, 1999. After processing, the physical copy would have been publicly accessible no later than October 14, 1999.

F. Usage Records

45. Actual usage of a publication is reflected by other publications that make reference to it. Citation history on Google Scholar shows *Wong* has been cited 3,804 times, and 23 times between 1999 and 2000. I used Google Scholar to compile a list of early citations. **Appendix 0001E** shows six IEEE citations published in 2000, and the earliest citation appeared in June 2000.

G. Summary of My Opinion on *Wong*

46. Based on my review of **Ex. 0001** and the library copies (**Appendices 0001A** and **0001C**), it is my opinion that *Wong* is authentic and has been publicly accessible at libraries since October 1999. Library serial records (**Appendices 0001B** and **0001D**) show the journal containing *Wong* can be discovered by the

journal's ISSN, title, variant titles, corporate contributors, and subject, and the date stamps on the library copies show both libraries received their copies on October 7, 1999. After processing, these library copies would have been publicly accessible no later than October 14, 1999. The citation list I compiled shows at least six IEEE citations to *Wong* were published in 2000, and the earliest one appeared in June 2000.

III. Authenticity and Public Availability of *Minn* (Ex. 0002)

A. Authentication

47. **Ex. 0002** is a true and correct copy of “An investigation into time-domain approach for OFDM channel estimation,” (“*Minn*”), by H. Minn and V. K. Bhargava, *IEEE Transactions on Broadcasting*, vol. 46, no. 4, pp. 240-248, that I obtained from the IEEE Xplore Digital Library. When I began preparing this declaration, I searched Google Scholar for *Minn* for records, and a record link led me to an IEEE metadata record for this article at <https://ieeexplore.ieee.org/abstract/document/898744>. I then followed the PDF link in the record to purchase this article from the Institute of Electrical and Electronics Engineers (“IEEE”). This IEEE copy is presented as **Ex. 0002** in this declaration.

48. **Ex. 0002** is a true and correct copy of *Minn* obtained from the publisher IEEE. Page 1 shows the title, authors, abstract, and index terms (“Channel estimation, intra-symbol time-averaging, linear minimum mean square error

estimator, most significant taps selection, orthogonal frequency division multiplexing (OFDM).” It also shows the manuscript was “received October 24, 2000,” the Publisher Item Identifier is “S 0018-9316(00)11686-5,” *Minn* appears in a journal whose ISSN (International Standard Serial Number) is “0018-9316,” the article has a “2000” copyright date and IEEE is the copyright holder. Page 1 (internal page 240) and even-numbered pages show a header that identifies the journal issue as “IEEE TRANSACTIONS ON BROADCASTING, VOL. 46, NO. 4, DECEMBER 2000.” **Ex. 0002** shows *Minn* has a total of nine pages, including eight sections, eight figures, 23 equations, three tables, and 20 references.

49. **Ex. 0002** is in a condition that creates no suspicion about its authenticity. Specifically, this copy is not missing any intermediate pages, the text on each page appears to flow seamlessly from one page to the next, and there are no visible alterations to the document. Moreover, **Ex. 0002** was found in the custody of the publisher, a place where, if authentic, it would likely be found. I therefore see no reason to question the authenticity of this IEEE copy of *Minn* (**Ex. 0002**).

B. Linda Hall Library Copy

50. To determine if *Minn* was available in libraries, I searched WorldCat by the ISSN of the journal containing *Minn*, and found a record at <https://search.worldcat.org/title/1642875> that shows Linda Hall Library is one of the 694 holding libraries of this journal. After verifying the holdings information in their

online catalog, I requested a copy of *Minn* from the library's Document Delivery Service. This library copy is presented as **Appendix 0002A** in this declaration.

51. **Appendix 0002A** is a true and correct copy of *Minn* obtained from Linda Hall Library. Pages 1 to 9 (internal pages 240 to 248) are the full text of *Minn*. I have closely compared this library copy with the IEEE copy (**Ex. 0002**) and found them to have the same content for *Minn*. Page 10 is the cover that identifies this issue as volume 46, number 4 of *IEEE Transactions on Broadcasting*, which is a publication of the IEEE Broadcast Technology Society. This page also shows the issue has an "December 2000" issue date, the journal's CODEN is IETBAC, and the journal ISSN is "0018-9316." The cover includes the table of contents of this issue and shows that "An Investigation into Time-Domain Approach for OFDM Channel Estimation" begins on page 240. The cover carries a library stamp of "Linda Hall Library FEB 07 2001" that indicates this issue was received by the library on February 7, 2001. Page 12 is the copyright page that shows this journal is published quarterly by IEEE, and this issue has a "1999" copyright date with IEEE as the copyright holder.

C. Linda Hall Library Records

52. **Appendix 0002B** is a true and correct copy of the bibliographic and MARC records for the journal *IEEE Transactions on Broadcasting* that contains *Minn*. I personally located, identified and obtained these records from the online

catalog of the Linda Hall Library. These are the type of records experts in my field would reasonably rely upon when forming their opinions. I used screen captures to produce this appendix.

53. Pages 1 to 3 of **Appendix 0002B** are the bibliographic record that shows this publication is a journal that is available at “Linda Hall Library Closed Stacks - Serials (I.E.E.E. transactions on broadcasting.) and other locations.” Page 2 shows details of this journal, including title, other titles, creators/contributors, subjects, identifiers, related title, publisher, creation date, format, frequency, and citations/references.

54. Pages 4 and 5 are the MARC record. The first six digits of Field 008 show the record was created on “750921” (*i.e.*, September 21, 1975), and the “c19639999” code following the record creation date indicates this publication is a continuing resource that began publication in 1963 and has no publication end date yet. Subfield “a” of Field 040 shows “MUL” (Minnesota Union List of Serials) created the original MARC serial record, the record was revised by NSD (National Serial Data Program of the Library of Congress) and other cataloging agencies. Field 049 shows this journal is in Collection A of “LHL.” According to the *Directory of OCLC Members* (<https://www.oclc.org/en/contacts/libraries.html>), LHL is the OCLC library symbol for the Linda Hall Library. Data in Fields 008 and 040 inform my opinion that Linda Hall Library used the original MARC record created by MUL

to create their copy cataloging record for this journal.

55. Field 022 of the MARC record shows the ISSN of this journal, Field 030 shows the journal's CODEN, Field 035 shows the record's OCLC number is "1642875." Field 245 shows the title, Fields 246 show two variant titles are included to help users discover this journal, Field 260 shows IEEE of New York is the publisher, Field 362 shows the journal began in "Feb. 1963," and Field 310 shows the current publication frequency of this journal is "Quarterly." Fields 510 show five indexing and abstracting services have fully indexed this journal, while five other services have selectively indexed this journal. Fields 710 show IEEE Broadcasting Group and IEEE Broadcast, Cable, and Consumer Electronics Society are included to help users discover this journal.

56. The original MARC record from MUL includes a classification number in Field 090, but Linda Hall Library organizes their journals by title, so this classification number is not used at this library. At this library, subject of this journal is represented in Fields 650 by two Library of Congress subject headings, with "Radio \$x Transmitters and transmission" ("x" indicates a topical subdivision) and "Television \$x Transmitters and transmission" as the main headings, each followed by "Periodicals" as a form subdivision. These subject representations enables users to conduct subject searches to find this journal.

57. This MARC record (**Appendix 0002B**) has made this journal

discoverable at Linda Hall Library. Interested users would have been able to discover this journal by the its ISSN, CODEN, title, variant titles, corporate contributors, and subjects.

58. As discussed earlier, a customary library practice is to process a newly received journal issue through the serial check-in system, and process the new issue promptly to make it available to users, usually within a week after serial check-in. The date stamp of “FEB 7 2001” on the Linda Hall Library copy of *Minn* (**Appendix 0002A**) informs my opinion that the issue containing *Minn* was received by the library on February 7, 2001, and the physical copy, after processing, would have become publicly accessible no later than February 14, 2001.

D. University of Wisconsin Library Copy

59. To further determine the public availability date of the journal issue containing *Minn*, I obtained a copy of *Minn* from the University of Wisconsin Library through a document delivery service. **Appendix 0002C** is a true and correct copy of *Minn*. Page 1 is the issue cover of volume 46, number 4 that has a “December 2000” issue date and page 2 is the copyright page of this issue that shows a “1999” copyright date with IEEE as the copyright holder. Page 3 is internal page 233 that carries a library date stamp of “Kurt F. Wendt Library College of Engineering, FEB 08 2001, UW-Madison” that indicates this issue was received by the Engineering Library on February 8, 2001. Pages 3 to 12 (internal pages 240 to 248) are the full

text of *Minn*. I have closely compared this library copy with the publisher's copy of *Wong* (**Ex. 0002**) and the Linda Hall Library copy (**Appendix 0002A**), and found the three copies to have the same content for *Minn*. In addition, the cover and copyright page of the University of Wisconsin Library copy (**Appendix 0002C**) are the same as those in the Linda Hall Library copy (**Appendix 0002A**), and the date stamps on the library copies show they received this issue containing *Wong* on February 7 and February 8 of 2001, respectively.

E. University of Wisconsin Library Records

60. **Appendix 0002D** is a true and correct copy of the bibliographic and MARC records for *IEEE Transactions on Broadcasting* that I personally located, identified, and obtained from the online catalog of the University of Wisconsin Library system. This is the type of record experts in my field would reasonably rely upon when forming their opinions. I used screen captures to produce this appendix.

61. **Appendix 0002D** shows the format of this publication is "Journals, Magazines, Newspapers" and the records have the same bibliographic details as the Linda Hall Library serial records (**Appendix 0002B**), including the journal's ISSN, journal title, variant titles, publisher, beginning year of publication, subject representation and corporate contributors. The University of Wisconsin record (**Appendix 0002D**) contains other details, but the essential bibliographic data elements in both records are the same. The record creation date in Field 008 and the

creator of the original MARC record in Field 040 are the same in both records, indicating both libraries used the record originally created by MUL (Minnesota Union List of Serials) to create their copy cataloging serial record. The MARC record has made this journal discoverable in the online catalog of the University of Wisconsin Library system. Interested users would have been able to discover this journal by its ISSN, title, variant titles, corporate contributors and subjects.

62. The date stamp of “FEB 08 2001” on **Appendix 0002C** informs my opinion that this issue containing *Minn* was received by the University of Wisconsin Library system on February 8, 2001. After processing, the physical copy would have been publicly accessible no later than February 15, 2001.

F. Summary of My Opinion on *Minn*

63. Based on my review of **Ex. 0002** and the library copies (**Appendices 0002A** and **0002C**), it is my opinion that *Minn* is authentic and has been publicly accessible at libraries since February 2001. Library serial records (**Appendices 0002B** and **0002D**) show the journal containing *Minn* can be discovered by its ISSN, title, variant titles, corporate contributors, and subjects, and the date stamps on the library copies show these two libraries received their copies on February 7 and February 8 of 2001, respectively. After processing, the library copies would have become publicly accessible no later than February 15, 2001.

IV. Authentication and Public Availability of *Yeh* (Ex. 0003)

A. Authentication

64. **Ex. 0003** is a true and correct copy of “Channel estimation using pilot tones in OFDM systems,” (“*Yeh*”), by C. S. Yeh and Y. Lin, *IEEE Transactions on Broadcasting*, vol. 45, no. 4, pp. 400-409, that I obtained from the IEEE Xplore Digital Library. When I began preparing this declaration, I searched Google Scholar for *Yeh* for records, and a record link led me to an IEEE metadata record for this article at <https://ieeexplore.ieee.org/document/825535>. I then followed the PDF link in the record to purchase this article from the Institute of Electrical and Electronics Engineers (“IEEE”). This IEEE copy is presented as **Ex. 0003** in this declaration.

65. **Ex. 0003** is a true and correct copy of *Yeh* obtained from the publisher IEEE. Page 1 shows the title, authors, and abstract. It also shows the Publisher Item Identifier is “S 0018-9316(99)10444-X” and *Yeh* appears in a journal whose ISSN (International Standard Serial Number) is “0018-9316,” the article has a “1999” copyright date and IEEE is the copyright holder. The header on page 1 (internal page 400) shows the journal containing *Yeh* is “IEEE Transactions on Broadcasting, Vol. 45, No. 4, December 1999.” **Ex. 0003** shows *Yeh* has a total of 10 pages, including six sections, 15 equations, two tables, eight figures, and 11 references.

66. **Ex. 0003** is in a condition that creates no suspicion about its authenticity. Specifically, this copy is not missing any intermediate pages, the text

on each page appears to flow seamlessly from one page to the next, and there are no visible alterations to the document. Moreover, **Ex. 0003** was found in the custody of the publisher, a place where, if authentic, it would likely be found. I therefore see no reason to question the authenticity of this IEEE copy of *Yeh* (**Ex. 0003**).

B. Linda Hall Library Copy

67. To determine if *Yeh* was available in libraries, I searched WorldCat by the ISSN of the journal containing *Yeh*, and found a record at <https://search.worldcat.org/title/1642875> that shows Linda Hall Library is one of the 694 holding libraries of this journal. After verifying the holdings information in their online catalog, I requested a copy of *Yeh* from the library's Document Delivery Service. This library copy is presented as **Appendix 0003A** in this declaration.

68. **Appendix 0003A** is a true and correct copy of *Yeh* obtained from Linda Hall Library. Pages 1 to 10 (internal pages 400 to 409) are the full text of *Yeh*. I have closely compared this library copy with the IEEE copy (**Ex. 0003**) and found them to have the same content for *Yeh*. Page 11 is the cover that identifies this issue as volume 45, number 4 of *IEEE Transactions on Broadcasting*, which is a publication of the IEEE Broadcast Technology Society. This page also shows the issue has a "December 1999" issue date, the journal's CODEN is IETBAC, and the journal ISSN is "0018-9316." The cover includes the table of contents of this issue and shows that "Channel Estimation Using Pilot Tones in OFDM Systems" begins on

page 400. The cover carries a library stamp of “Linda Hall Library FEB 07 2000” that indicates this issue was received by the library on February 7, 2000. Page 12 is the copyright page that shows this journal is published quarterly by IEEE, and this issue has a “1999” copyright date with IEEE as the copyright holder.

C. Linda Hall Library Records

69. **Appendix 0003B** is a true and correct copy of the bibliographic and MARC records for the journal *IEEE Transactions on Broadcasting* that contains *Yeh*. I personally located, identified and obtained these records from the online catalog of the Linda Hall Library. These are the type of records experts in my field would reasonably rely upon when forming their opinions. I used screen captures to produce this appendix.

70. Pages 1 to 3 of **Appendix 0003B** are the bibliographic record that shows this publication is a journal that is available at “Linda Hall Library Closed Stacks - Serials (I.E.E.E. transactions on broadcasting.) and other locations.” Page 2 shows details of this journal, including the title, other titles, creators/contributors, subjects, identifiers, related title, publisher, creation date, format, frequency, and citations/references.

71. Pages 4 and 5 are the MARC record. The first six digits of Field 008 show the record was created on “750921” (*i.e.*, September 21, 1975), and the “c19639999” code following the record creation date indicates this publication is a

continuing resource that began publication in 1963 and has no publication end date yet. Subfield “a” of Field 040 shows “MUL” (Minnesota Union List of Serials) created the original MARC serial record, the record was revised by NSD (National Serial Data Program of the Library of Congress) and other cataloging agencies. Field 049 shows this journal is in Collection A of “LHL.” According to the *Directory of OCLC Members* (<https://www.oclc.org/en/contacts/libraries.html>), LHL is the OCLC library symbol for the Linda Hall Library. Data in Fields 008 and 040 inform my opinion that Linda Hall Library used the original MARC record created by MUL to create their copy cataloging record for this journal.

72. Field 022 of the MARC record shows the ISSN of this journal, Field 030 shows the journal’s CODEN, Field 035 shows the record’s OCLC number is “1642875.” Field 245 shows the title, Fields 246 show two variant titles are included to help users discover this journal, Field 260 shows IEEE of New York is the publisher, Field 362 shows the journal began in “Feb. 1963,” and Field 310 shows the current publication frequency of this journal is “Quarterly.” Fields 510 show five indexing and abstracting services have fully indexed this journal, while five other services have selectively indexed this journal. Fields 710 show IEEE Broadcasting Group and IEEE Broadcast, Cable, and Consumer Electronics Society are included to help users discover this journal.

73. The original MARC record from MUL includes a classification number

in Field 090, but Linda Hall Library organizes their journals by title, so this classification number is not used at this library. At this library, subject of this journal is represented in Fields 650 by two Library of Congress subject headings, with “Radio \$x Transmitters and transmission” (“\$x” indicates a topical subdivision) and “Television \$x Transmitters and transmission” as the main headings, each followed by “Periodicals” as a form subdivision. These subject representations enables users to conduct subject searches to find this journal.

74. This MARC record (**Appendix 0002B**) has made this journal discoverable at Linda Hall Library. Interested users would have been able to discover this journal by the its ISSN, CODEN, title, variant titles, corporate contributors, and subjects.

75. As discussed earlier, a customary library practice is to process a newly received journal issue through the serial check-in system, and process the new issue promptly to make it available to users, usually within a week after serial check-in. The date stamp of “FEB 7 2000” on the Linda Hall Library copy of *Yeh* (**Appendix 0003A**) informs my opinion that the issue containing *Yeh* was received by the library on February 7, 2000, and the physical copy, after processing, would have become publicly accessible no later than February 14, 2000.

D. University of Wisconsin Library Copy

76. To further determine the public availability date of the journal issue

containing *Yeh*, I obtained a copy of this article from the University of Wisconsin Library through a document delivery service. **Appendix 0003C** is a true and correct copy of *Yeh*. Page 1 is the issue cover of volume 45, number 4 that has a “December 1999” issue date and page 2 is the copyright page of this issue that shows a “1999” copyright date with IEEE as the copyright holder. Page 3 is internal page 353 that carries a library date stamp of “Kurt F. Wendt Library College of Engineering, FEB 07 2000, UW-Madison, WI 53706” that indicates this issue was received by the Engineering Library on February 7, 2000. Pages 4 to 13 (internal pages 400 to 409) are the full text of *Yeh*. I have closely compared this library copy with the publisher’s copy of *Yeh* (**Ex. 0003**) and the Linda Hall Library copy (**Appendix 0003A**), and found the three copies to have the same content for *Yeh*. In addition, the cover and copyright page of the University of Wisconsin Library copy (**Appendix 0003C**) are the same as those in the Linda Hall Library copy (**Appendix 0003A**), and the date stamps from these two library copies show both libraries received this issue containing *Yeh* on February 7, 2000.

E. University of Wisconsin Library Records

77. **Appendix 0003D** is a true and correct copy of the bibliographic and MARC records for *IEEE Transactions on Broadcasting* that I personally located, identified, and obtained from the online catalog of the University of Wisconsin Library system. This is the type of record experts in my field would reasonably rely

upon when forming their opinions. I used screen captures to produce this appendix.

78. **Appendix 0003D** shows the format of this publication is “Journals, Magazines, Newspapers” and the records have the same bibliographic details as the Linda Hall Library serial records (**Appendix 0003B**), including the journal’s ISSN, journal title, variant titles, publisher, beginning year of publication, subject representation and corporate contributors. The University of Wisconsin record (**Appendix 0003D**) contains other details, but the essential bibliographic data elements in both records are the same. The record creation date in Field 008 and the creator of the original MARC record in Field 040 are the same in both records, indicating both libraries used the record originally created by MUL (Minnesota Union List of Serials) to create their copy cataloging serial record. The MARC record has made this journal discoverable in the online catalog of the University of Wisconsin Library system. Interested users would have been able to discover this journal by its ISSN, title, variant titles, corporate contributors and subjects.

79. The date stamp of “FEB 07 2000” on **Appendix 0003C** informs my opinion that this issue containing *Yeh* was received by the University of Wisconsin Library system on February 7, 2000. After processing, the physical copy would have been publicly accessible no later than February 14, 2000.

F. Usage Record

80. Actual usage of a publication is reflected by other publications that

make reference to it. Citation history on Google Scholar shows *Yeh* has been cited for 143 times, and one IEEE article referenced *Yeh* in December 2000: Minn, H., & Bhargava, V. K. (2000). An investigation into time-domain approach for OFDM channel estimation. *IEEE Transactions on Broadcasting*, 46(4), 240-248. Minn is included as **Ex. 0002** in this declaration, and its citation to *Yeh* (Reference 17) is as follows: C. S. Yeh and Y. Lin, “Channel estimation using pilot tones in OFDM systems,” *IEEE Trans. On Broadcasting*, vol. 45, no. 4, pp. 400-409, Dec. 1999.

G. Summary of My Opinion on *Yeh*

81. Based on my review of **Ex. 0003** and the library copies (**Appendices 0003A** and **0003C**), it is my opinion that *Yeh* is authentic and has been publicly accessible at libraries since February 2000. Library serial records (**Appendices 0003B** and **0003D**) show the journal containing *Yeh* can be discovered by its ISSN, title, variant titles, corporate contributors, and subjects, and the date stamps on the library copies show these two libraries received their copies on February 7, 2000. After processing, the library copies would have become publicly accessible no later than February 14, 2000. The citation history of *Yeh* on Google Scholar shows that the earliest citation is a December 2000 IEEE article by Minn.

V. Authenticity and Public Availability of *Lehne* (Ex. 0004)

A. Authentication

82. **Ex. 0004** is a true and correct copy of “An overview of smart antenna

technology for mobile communications systems,” (“*Lehne*”), by Per H. Lehne and Magne Pettersen, *IEEE Communications Surveys*, Fourth Quarter 1990, vol. 2, no. 4, that I received from counsel. Page 1 (internal page 2) shows the title, authors, and abstract, and indicates *Lehne* appears in *IEEE Communications Surveys*, which is “The Electronic Magazine of Original Peer-Reviewed Survey Articles.” The footer on all the pages of *Lehne* identifies the journal issue as vol. 2, no. 4 that has a “Fourth Quarter 1999” issue date. **Ex. 0004** shows *Lehne* has a total of 12 pages, including four equations, 17 figures in color, and 29 references.

83. Because the journal containing *Lehne* is an electronic magazine, I searched Internet Archive for records. Internet Archive is a digital library that archives cultural heritage materials, scholarly research, and other web resources from the Internet. It provides the public with free access to its collections. I have used Internet Archive in the past and am familiar with their practice of assigning an extended URL to an archived file in the format of [http://www.archive.org/web/\[Year in yyyy\]\[Month in mm\]\[Day in dd\]\[Time code in hh:mm:ss\]/\[Archived URL\]](http://www.archive.org/web/[Year in yyyy][Month in mm][Day in dd][Time code in hh:mm:ss]/[Archived URL]) to show the point in time when the archived file was captured. Because Internet Archive usually does not capture web resources right after they appear on the Internet, the archive date does not represent the first time a web resource appeared on the Internet. Rather, it means the archive resource was available prior to the archive date and was publicly available no later than the archive date.

84. **Appendix 0004A** is a true and correct copy of the webpage of “Magazines at a Glance” archived by the Internet Archive on March 3, 2000, and a copy of *Lehne* archived by the Internet Archive on August 15, 2000. I personally located, identified, and obtained these records. They are the type of material experts in my field would reasonably rely upon when forming their opinions.

85. Pages 1 and 2 of **Appendix 0004A** are screen captures of the IEEE Communications Society’s webpage of “Magazines at a Glance” that shows the webpage was archived 26 times and this particular copy was archived on “MAR 03 2000” and is available at <https://web.archive.org/web/20000303101516/http://www.comsoc.org/pubs/magazines.html>. This archived webpage shows a link to “EEE Communications Surveys” which is “the first electronically published journal of the IEEE Communications Society” that provides “a continuously-available source of peer-reviewed, comprehensive, leading-edge surveys covering all active areas of communications” and “[t]he electronic medium allows rapid publication of up-to date survey articles and the capability of electronic searching.” The archive date of “20000303101516” (*i.e.*, March 3, 2000, at 10 a.m., 15 minutes, 16 seconds) of this archived webpage informs my opinion that the journal containing *Lehne* would have been publicly accessible no later than March 3, 2000, on the Internet.

86. Pages 3 to 6 are screenshots of contents presented at the journal website.

Because the website was designed with frame, pages 3 to 6 are webpages displayed with an “AUG 15 2000” Wayback Machine calendar, but these webpages themselves were archived on different dates. The URLs and archive dates of these webpages are confirmed by an Internet Archive affidavit (**Ex. 0005**) that is discussed below.

87. Page 3 is a screen capture of the journal website of *IEEE Communications Surveys & Tutorials*. The Wayback Machine calendar shows this website was archived 343 times between “14 Jun 1997” and “16 Mar 2023” and this particular copy was archived on “AUG 15 2000” and is available at <https://web.archive.org/web/20000815093908/http://www.comsoc.org/pubs/surveys/>. The “Quarterly Issues” link on the navigation panel on the left side led me to the frame for “quarterly issues” (page 4). This frame shows that “[t]he quarterly issues are currently FREE OF CHARGE. The first quarterly issue was published in October 1998.” I selected “Fourth Quarter 1999” to view that issue. Page 5 is a screen capture of vol. 2, no. 4, the Fourth Quarter 1999 issue of the journal, and “An Overview of Smart Antenna Technology for Mobile Communications Systems” by Lehne and Pettersen is the first feature article. I followed the article link to reach the article webpage. Page 6 is a screen capture that shows *Lehne* has a “1999” copyright date and IEEE is the copyright holder. To obtain the full article, I saved the content from the main display frame. Pages 8 to 21 show the PDF of *Lehne* that I obtained from

the display frame of this journal website. The URL of this frame is https://web.archive.org/web/20000818222434fw_/http://www.comsoc.org/pubs/surveys/4q99issue/lehne.html. The “20000818222434fw” shows the archive date of this frame was August 18, 2000, at 10 p.m., 24 minutes, 34 seconds. This archive date informs my opinion that the webpage of *Lehne* would have been publicly accessible by August 18, 2000. I also obtained the source code for this frame from https://web.archive.org/web/20000818222434fw_/http://www.comsoc.org/pubs/surveys/4q99issue/lehne.html, and it includes the note below that confirms *Lehne* was archived by the Internet Archive on August 18, 2000:

```
<!--  
    FILE ARCHIVED ON 22:24:34 Aug 18, 2000 AND RETRIEVED FROM THE  
    INTERNET ARCHIVE ON 19:18:49 Oct 26, 2023.  
    JAVASCRIPT APPENDED BY WAYBACK MACHINE, COPYRIGHT INTERNET ARCHIVE  
  
    ALL OTHER CONTENT MAY ALSO BE PROTECTED BY COPYRIGHT (17 U.S.C.  
    SECTION 108(a)(3)).  
-->
```

88. I have closely compared this archived copy of *Lehne* (**Appendix 0004A**) with the counsel copy (**Ex. 0004**), and found the text, equations, and references to be the same in both copies. The differences between these two copies are that some of the Greek characters in **Appendix 0004A** are not displayed as clearly as **Ex. 0004**, and that while **Appendix 0004A** includes links to the figures, Internet Archive did not archive those figures. For the purposes of my analysis, I

have found the copy of *Lehne* archived by the Internet Archive (**Appendix 0004A**) to have the same content as the counsel copy (**Ex. 0004**), with the understanding that the figures are not archived. I therefore see no reason to question the authenticity of **Ex. 0004**.

B. Internet Archive Affidavit (Ex. 0005)

89. My analysis of the archived copy of *Lehne* and contents on the journal website (**Appendix 0004A**) above is supported by the affidavit of Mr. Nathaniel E. Frank-White of the Internet Archive (**Ex. 0005**) that I received from counsel. Mr. Frank-White states that “[t]he Internet Archive is a website that provides access to a digital library of Internet sites and other cultural artifacts in digital form. Like a paper library, we provide free access to researchers, historians, scholars, and the general public” (Page 1, ¶ 2). He further explains that “[t]he Internet Archive assigns a URL on its site to the archived files in the format `http://www.archive.org/web/[Year in yyyy][Month in mm][Day in dd][Time code in hh:mm:ss]/[Archived URL]` aka an ‘extended URL’ ... The date indicated by an extended URL applies to a preserved instance of a file for a given URL” (Page 1, ¶ 5). He then states that Exhibit A to his declaration “are true and accurate copies of browser screenshots of the Internet Archive’s records of the archived files for the URLs and the dates specified in the attached coversheet of each printout” (Page 1, ¶ 6).

90. Mr. Frank-White’s Exhibit A (Pages 9 to 25) confirms the authenticity

and accuracy of the screenshots presented in **Appendix 0004A** that I discussed above. Because the journal website was designed with frame, screenshots of the journal (pages 3 to 6 of **Appendix 0004A**) show the editor page, quarterly issues, table of contents, and the archived copy of *Lehne* are displayed with an “AUG 15 2000” Wayback Machine calendar. To authenticate these webpages, I copied their frame addresses and asked, through counsel, the Internet Archive to verify the URLs, contents and archive dates of these webpages. The affidavit of Mr. Frank-White confirms the screenshots in **Appendix 0004A** are true and accurate copies archived by the Internet Archive. Specifically, the following webpages and archive dates are confirmed by Exhibit A:

- Pages 9 to 15 of Exhibit A. A copy of *Lehne* archived on August 18, 2000, available at https://web.archive.org/web/20000818222434fw_/http://www.comsoc.org/pubs/surveys/4q99issue/lehne.html.
- Pages 16 and 17 of Exhibit A. The Table of Contents of vol. 2, no. 4 (Fourth Quarter 1999) of *IEEE Communications Surveys*, archived on September 29, 2000, available at https://web.archive.org/web/20000929054741fw_/http://www.comsoc.org/pubs/surveys/4q99issue/4q99-lead.html.
- Pages 18 and 19 of Exhibit A. A list of “Quarterly Issues” archived on August 24, 2000, available at https://web.archive.org/web/20000824011905fw_/http://www.comsoc.org/pubs/surveys/4q99issue/4q99-lead.html.

org/pubs/surveys/surv-quart-iss.html.

- Pages 20 and 21 of Exhibit A. The journal's main frame archived on August 16, 2000, available at https://web.archive.org/web/20000816203109fw_/http://www.comsoc.org/pubs/surveys/frame277698.html.
- Pages 22 and 23 of Exhibit A. The journal website archived on August 15, 2000, available at <https://web.archive.org/web/20000815093908/http://www.comsoc.org/pubs/surveys/>.
- Pages 24 and 25 of Exhibit A. IEEE List of Magazines including *IEEE Communications Surveys*, archived on March 3, 2000, available at <https://web.archive.org/web/20000303101516/http://www.comsoc.org/pubs/magazines.html>.

91. The affidavit of Mr. Frank-White (**Ex. 0005**) states that “Exhibit A are true and accurate copies of browser screenshots of the Internet Archive’s records of the archived files for the URLs and the dates specified in the attached coversheet of each printout” (**Ex. 0005**, Page 1, ¶ 6). This affidavit supports my analysis that *Lehne* was indeed archived on August 18, 2000, by the Internet Archive.

C. Citation Records

92. Actual usage of a publication is reflected by other publications that make reference to it. I used Google Scholar and Semantic Scholar to review citations and produced **Appendix 0004B**. Google Scholar shows *Lehne* has been cited 372 times (page 1), Semantic Scholar shows *Lehne* has been cited 244 times, and one

citation appeared in 2000 (pages 2 and 3). Pages 4 to 9 are the full text of the 2000 citation to *Lehne*: Krishnamurthy, P., & Karimi, H. A. (2000, September). Telegeoinformatics for efficient resource allocation and protocol development in wireless networks. In *Vehicular Technology Conference Fall 2000. IEEE VTS Fall VTC2000. 52nd Vehicular Technology Conference* (Cat. No. 00CH37152) (Vol. 5, pp. 2126-2131). IEEE. *Lehne* is listed as Reference 20: P.H. Lehne and M. Pettersen, “An overview of smart antenna technology for mobile communications systems,” IEEE Communication Surveys, Vol. 2, No. 4, Fourth Quarter, 1999.

D. Summary of My Opinion on *Lehne*

93. The copy of *Lehne* archived by the Internet Archive on August 18, 2000 (**Appendix 0004A**) shows the same full text, equations, and references as the counsel copy (**Ex. 0004**), it is therefore my opinion that, for the purposes of my analysis, **Ex. 0004** is authentic. The archive date of the IEEE Communications Society’s “Magazines at a Glance” webpage shows that the journal containing *Lehne* would have been publicly accessible by March 3, 2000, the “quarterly issues” frame informed users in August 2000 that the journal was freely accessible, and the archive date for the archived *Lehne* copy shows this article would have been publicly accessible on the Internet no later than August 18, 2000 (**Appendix 0004A**). The earliest citation to *Lehne* is an IEEE article that appeared in September 2000 (**Appendix 0004B**), further supporting my opinion that *Lehne* would have been

publicly accessible by August 18, 2000, on the Internet.

VI. Conclusion

94. In signing this declaration, I recognize that the declaration will be filed as evidence in a Petition for *Inter Partes* Review before the Patent Trial and Appeal Board of the United States Patent and Trademark Office. I also recognize that I may be subject to deposition in the case. If deposition is required of me, I will appear at a reasonable time and place to be agreed upon.

95. I hereby declare that all statements made herein on my own knowledge are true and that all statements made on information and belief are believed to be true, and further, that these statements were made with the knowledge that willful false statements and the like so made are punishable by fine or imprisonment, or both, under Section 1001 of Title 18 of the United States Code.

Date: 12-20-2023

Executed: Ingrid Hsieh-Yee
Ingrid Hsieh-Yee, Ph.D.

Hsieh-Yee Declaration

Exhibit 0001

Multiuser OFDM with Adaptive Subcarrier, Bit, and Power Allocation

Cheong Yui Wong, Roger S. Cheng, *Member, IEEE*,
Khaled Ben Letaief, *Senior Member, IEEE*, and Ross D. Murch, *Senior Member, IEEE*

Abstract—Multiuser orthogonal frequency division multiplexing (OFDM) with adaptive multiuser subcarrier allocation and adaptive modulation is considered. Assuming knowledge of the instantaneous channel gains for all users, we propose a multiuser OFDM subcarrier, bit, and power allocation algorithm to minimize the total transmit power. This is done by assigning each user a set of subcarriers and by determining the number of bits and the transmit power level for each subcarrier. We obtain the performance of our proposed algorithm in a multiuser frequency selective fading environment for various time delay spread values and various numbers of users. The results show that our proposed algorithm outperforms multiuser OFDM systems with static time-division multiple access (TDMA) or frequency-division multiple access (FDMA) techniques which employ fixed and predetermined time-slot or subcarrier allocation schemes. We have also quantified the improvement in terms of the overall required transmit power, the bit-error rate (BER), or the area of coverage for a given outage probability.

Index Terms—Adaptive modulation, frequency selective fading channel, multiaccess communication, multiuser channel, orthogonal frequency division multiplexing (OFDM), resource management.

I. INTRODUCTION

RECENTLY, intense interest has focused on modulation techniques which can provide broadband transmission over wireless channels for applications including wireless multimedia, wireless Internet access, and future-generation mobile communication systems. One of the main requirements on the modulation technique is the ability to combat intersymbol interference (ISI), a major problem in wideband transmission over multipath fading channels. There are many methods proposed to combat the ISI, e.g., [1]–[3]. Multicarrier modulation techniques, including orthogonal frequency division multiplex (OFDM), (e.g., [4]) are among the more promising solutions to this problem.

Assuming that the transmitter knows the instantaneous channel transfer functions of all users, many papers [5]–[7] have demonstrated that significant performance improvement can be achieved if adaptive modulation is used with OFDM. In

particular, subcarriers with large channel gains employ higher order modulation to carry more bits/OFDM symbol, while subcarriers in deep fade carry one or even zero bits/symbol. Integrated design of forward error correcting code and adaptive modulation has also been studied using BCH code and trellis coded modulation (TCM) in [8] and [9], respectively. Although both references considered only time-varying flat fading channels, the same coded adaptive modulation design can be easily applied to OFDM systems. As different subcarriers experience different fades and transmit different numbers of bits, the transmit power levels must be changed accordingly. The problem of optimal power allocation has also been studied in [10].

In this paper, we consider extending OFDM with adaptive modulation to multiuser frequency selective fading environments. When OFDM with adaptive modulation is applied in a frequency selective fading channel, a significant portion of the subcarriers may not be used. These are typically subcarriers which experience deep fade and are not power efficient to carry any information bit. In multiuser systems using static time-division multiple access (TDMA) or frequency-division multiple access (FDMA) as multiaccess schemes, each user is allocated a predetermined time slot or frequency band to apply OFDM with adaptive modulation. Consequently, these unused subcarriers (as a result of adaptive modulation) within the allocated time slot or frequency band of a user are wasted and are not used by other users. However, the subcarriers which appear in deep fade to one user may not be in deep fade for other users. In fact, it is quite unlikely that a subcarrier will be in deep fade for all users, as the fading parameters for different users are mutually independent. This motivates us to consider an adaptive multiuser subcarrier allocation scheme where the subcarriers are assigned to the users based on instantaneous channel information. This approach will allow all the subcarriers to be used more effectively because a subcarrier will be left unused only if it appears to be in deep fade to all users.

We consider a multiuser subcarrier, bit, and power allocation scheme where all users transmit in all the time slots. Our objective is to minimize the overall transmit power by allocating the subcarriers to the users and by determining the number of bits and the power level transmitted on each subcarrier based on the instantaneous fading characteristics of all users. In this paper, we formulate the multiuser subcarrier, bit, and power allocation problem and propose an iterative algorithm to perform the multiuser subcarrier allocation. Once

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The authors are with the Department of Electrical and Electronic Engineering, Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong (e-mail: ceyui@ust.hk; eecheng@ust.hk; eekhaled@ust.hk; and deermurch@ust.hk).

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the subcarrier allocation is determined, the bit and power allocation algorithm can be applied to each user on its allocated subcarriers. We also compare the performance of our proposed solution to various other static subcarrier allocation schemes.

The results of the work can be applied, for instance, to the downlink transmission in a time division duplex (TDD) wireless communication system to improve the downlink capacity. In such a system, the base station (BS) can estimate the instantaneous channel characteristics of all the BS-to-mobile links based on the received uplink transmissions. The multiuser subcarrier, bit, and power allocation can then be used. It is clear that there is a certain amount of transmission overhead as the BS has to inform the mobiles about their allocated subcarriers and the number of bits assigned to each subcarrier.¹ However, this overhead can be relatively small, especially if the channels vary slowly (e.g., in an indoor low mobility environment), and the assignment is done once every many OFDM symbols. To further reduce the overhead, we can assign a contiguous band of subcarriers with similar fading characteristics as a group, instead of assigning each individual subcarrier. In this paper, we will not focus on how the subcarrier allocation information is transmitted. Instead, we will focus on how—and by how much—this new strategy can reduce the required transmit power; or how and by how much this new scheme can improve the bit-error rate (BER) for a fixed transmit power. Alternately, we also consider how and by how much this new scheme can increase the area of coverage for a given transmit power and target BER.

While the bit allocation algorithm can be viewed as a practical implementation of the water-pouring interpretation for achieving the Shannon capacity of an ISI channel [13], the multiuser subcarrier and bit allocation algorithm presented in this paper is the counterpart of the multiuser water-pouring solution given in [14]. In information theoretic studies, the usual approach is to maximize the capacity (or information rate) under the power constraint. In this study, we focus on deriving practical algorithms that can support real-time multimedia data whose bit rates are generally fixed by the compression algorithms. Hence, we assume a given set of user data rates and attempt to minimize the total transmit power under a fixed performance requirement.

The organization of this paper is as follows. In Section II, we will first give the system model and formulate the minimum overall transmit power problem. The optimization problem seeks to minimize the overall transmit power using combined subcarrier, bit, and power allocation schemes for multiuser OFDM systems. The bit and power allocation algorithm for a single-user system is studied in Section III. In Section IV, we derive a lower bound to the minimum overall transmit

power by relaxing some of the constraints in the original problem. We also derive a suboptimal subcarrier allocation algorithm. In Section V, we compare the performance between our proposed method and other static approaches via Monte Carlo simulations. Finally, we conclude in Section VI.

II. SYSTEM MODEL

The configuration of our multiuser adaptive OFDM system is shown in Fig. 1. We assume that the system has K users and the k th user has a data rate equal to R_k bit per OFDM symbol. In the transmitter, the serial data from the K users are fed into the subcarrier and bit allocation block which allocates bits from different users to different subcarriers. We assume that each subcarrier has a bandwidth that is much smaller than the coherence bandwidth of the channel and that the instantaneous channel gains on all the subcarriers of all the users are known to the transmitter. Using the channel information, the transmitter applies the combined subcarrier, bit, and power allocation algorithm to assign different subcarriers to different users and the number of bits/OFDM symbol to be transmitted on each subcarrier. Depending on the number of bits assigned to a subcarrier, the adaptive modulator will use a corresponding modulation scheme, and the transmit power level will be adjusted according to the combined subcarrier, bit, and power allocation algorithm. We define $c_{k,n}$ to be the number of bits of the k th user that are assigned to the n th subcarrier. As we do not allow more than one user to share a subcarrier, it follows that for each n , if $c_{k',n} \neq 0$, $c_{k,n} = 0$ for all $k \neq k'$. We also assume that the adaptive modulator allows $c_{k,n}$ to take values in the set $\mathbf{D} = \{0, 1, 2, \dots, M\}$ where M is the maximum number of information bits/OFDM symbol that can be transmitted by each subcarrier.

The complex symbols at the output of the modulators are transformed into the time domain samples by inverse fast Fourier transform (IFFT). Cyclic extension of the time domain samples, known as the guard interval, is then added to ensure orthogonality between the subcarriers, provided that the maximum time dispersion is less than the guard interval. The transmit signal is then passed through different frequency selective fading channels to different users.

We assume that the subcarrier and bit allocation information is sent to the receivers via a separate control channel. At the receiver, the guard interval is removed to eliminate the ISI, and the time samples of the k th user are transformed by the FFT block into modulated symbols. The bit allocation information is used to configure the demodulators while the subcarrier allocation information is used to extract the demodulated bits from the subcarriers assigned to the k th user.

In the frequency selective fading channel, different subcarriers will experience different channel gains. We denote by $\alpha_{k,n}$ the magnitude of the channel gain (assuming coherent reception) of the n th subcarrier as seen by the k th user. We assume that the single-sided noise power spectral density (PSD) level N_0 is equal to unity (i.e., $N_0 = 1$), for all

¹Note that the power level used does not need to be transmitted to the receiver in such a TDD system. As the subcarrier gain is known to the transmitter, it can adjust the transmit power level to achieve a predetermined receiver power level based on the number of bits allocated to that subcarrier. However, in FDD systems, the transmit power levels determined by the receiver have to be sent back to the transmitter. In such systems, the additional performance gain achieved by power allocation may not justify the cost of sending the transmit power level information to the transmitter.

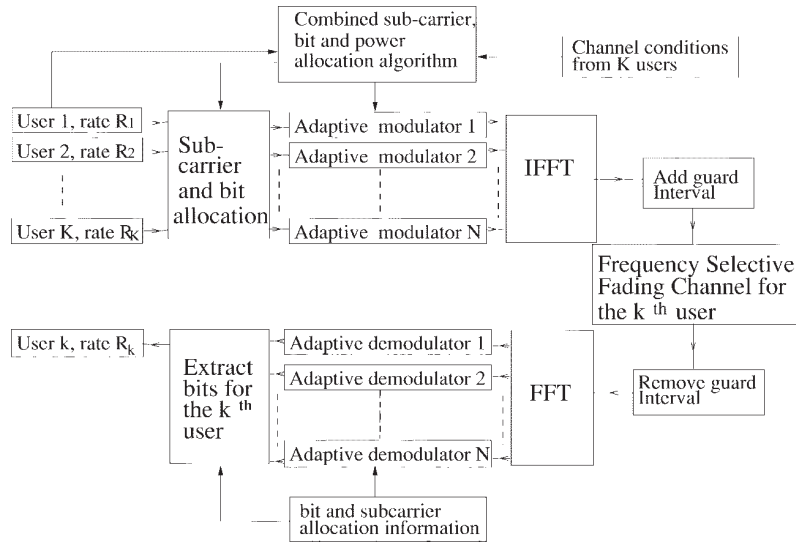


Fig. 1. Block diagram of a multiuser OFDM system with subcarrier, bit, and power allocation.

subcarriers and is the same for all users. Furthermore, we denote by $f_k(c)$ the required received power (in energy per symbol) in a subcarrier for reliable reception of c information bits/symbol when the channel gain is equal to unity. Note that the function $f_k(c)$ depends on k , and this allows different users to have different quality-of-service (QoS) requirements and/or different coding and modulation schemes. In order to maintain the required QoS at the receiver, the transmit power, allocated to the n th subcarrier by the k th user must equal

$$P_{k,n} = \frac{f_k(c_{k,n})}{\alpha_{k,n}^2}. \quad (1)$$

Using these transmit power levels, the receiver can demodulate the modulated symbols at the output of the FFT processor and achieve the desired QoS's of all users.

The goal of the combined subcarrier, bit, and power allocation algorithm is then to find the best assignment of $c_{k,n}$ so that the overall transmit power, the sum of $P_{k,n}$ over all subcarriers and all users, is minimized for given transmission rates of the users and given QoS requirements specified through $f_k(\cdot)$, $k = 1, \dots, K$. In order to make the problem tractable, we further require that $f_k(c)$ is a convex and increasing function with $f_k(0) = 0$. This condition essentially means that no power is needed when no bit is transmitted and that the required additional power to transmit an additional bit increases with c [i.e., $f_k(c+1) - f_k(c)$ is increasing in c]. Almost all popular coding and modulation schemes satisfy this condition.

It is important to note that even though the problem is formulated to minimize the overall transmit power for given QoS requirements, the same solution can be applied to improve the QoS's of the users for a given overall transmit power. The latter can simply be achieved by increasing the power proportionally for all the subcarriers, while using the same set of $c_{k,n}$.

Mathematically, we can formulate the problem as

$$P_T^* = \min_{c_{k,n} \in \mathbf{D}} \sum_{n=1}^N \sum_{k=1}^K \frac{1}{\alpha_{k,n}^2} f_k(c_{k,n}) \quad (2)$$

and the minimization is subjected to the constraints

$$\mathbf{C1:} \text{ For all } k \in \{1, \dots, K\}, R_k = \sum_{n=1}^N c_{k,n} \quad (3)$$

and

$$\begin{aligned} \mathbf{C2:} \text{ For all } n \in \{1, \dots, N\}, \\ \text{if there exists } k' \text{ with } c_{k',n} \neq 0, \text{ then } c_{k,n} = 0, \\ \forall k \neq k'. \end{aligned} \quad (4)$$

Note that constraint (3) is the data rate requirement and constraint (4) ensures that each subcarrier can only be used by one user. Moreover, $\mathbf{D} = \{0, 1, 2, \dots, M\}$ is the set of all possible values for $c_{k,n}$, and $c_{k,n} = 0$ means that the k th user does not use the n th subcarrier to transmit any information.

III. BIT ALLOCATION ALGORITHM FOR SINGLE USER CHANNEL

Before we try to solve the multiuser allocation problem, we will first derive the bit allocation algorithm for the single-user environment. The single-user problem not only gives better understanding of the issues involved, but also provides a bit allocation algorithm that we will use in our multiuser solution.

We can rewrite the optimization problem in (2) for the single-user case as

$$P_T^* = \min_{c_n \in \mathbf{D}} \sum_{n=1}^N \frac{1}{\alpha_n^2} f(c_n) \quad (5)$$

and the minimization is under the constraint

$$R = \sum_{n=1}^N c_n. \quad (6)$$

Note that we have dropped the subscript k , which denotes the user in all notations.

As the power needed to transmit a certain number of bits in a subcarrier is independent of the numbers of bits allocated to other subcarriers, it turns out that a greedy approach is optimal. A greedy algorithm assigns bits to the subcarriers one bit at a time, and in each assignment, the subcarrier that requires the least additional power is selected. The bit allocation process will be completed when all R bits are assigned. Several papers (e.g., [15] and [16]) have provided various algorithms for this problem, and the basic structure of most algorithms are similar and can be described as follows:

Initialization:

For all n , let $c_n = 0$ and $\Delta P_n = [f(1) - f(0)]/\alpha_n^2$;

Bit Assignment Iterations:

Repeat the following R times:

$\hat{n} = \arg \min_n \Delta P_n$;

$c_{\hat{n}} = c_{\hat{n}} + 1$;

$\Delta P_{\hat{n}} = [f(c_{\hat{n}} + 1) - f(c_{\hat{n}})]/\alpha_{\hat{n}}^2$;

End;

Finish:

$\{c_n\}_{n=1}^N$ is the final bit allocation solution.

The initialization stage computes, for each subcarrier, the additional power needed to transmit an additional bit. For each bit assignment iteration, the subcarrier that needs the minimum additional power is assigned one more bit, and the new additional power for that subcarrier is updated. After R iterations, the final bit assignment gives the optimal bit allocation for each subcarrier. It is important to note that the bit allocation is optimal only for the given function $f(c)$, which depends on the selected modulation scheme. Different modulation schemes will lead to different $f(c)$, different bit allocation, and possibly lower transmit power P_T^* .

The concept of this algorithm is fairly simple, and many similar algorithms based on the same principle have been obtained before. In particular, there exist faster and less complex algorithms which can speed up the bit allocation process significantly (e.g., [15] and [16]). In our simulations, we use the algorithm given in [16].

IV. MULTIUSER SUBCARRIER AND BIT ALLOCATION

We have observed that, in the single-user case, a greedy approach which assigns one bit at a time to the subcarrier that requires the least additional power gives the optimal allocation in the sense of minimizing the overall transmit power. Unfortunately, the problem becomes more difficult in the multiuser environment. As users cannot share the same subcarrier, allocating bits to a subcarrier essentially prevents other users from using that subcarrier. This dependency makes any greedy algorithm a nonoptimal solution. It turns out that the optimal solution may not assign any of a user's bits to the best subcarrier seen by that user. This may happen when the best subcarrier of a user is also the best subcarrier of another user who happens to have no other good subcarriers. Hence, the multiuser subcarrier and bit allocation problem is much more complicated to solve than that of the single-user case.

It turns out that the optimization problem in (2) is a combinatorial optimization problem. To make the problem tractable, we consider a different but similar optimization problem. We relax the requirement $c_{k,n} \in \mathbf{D}$ to allow $c_{k,n}$ to be a real number within the interval $[0, M]$. Moreover, in order to deal with constraint (4), K variables, $\rho_{k,n}$, $k = 1, \dots, K$, with values within the interval $[0, 1]$, are introduced to the cost function as sharing factors of the n th subcarrier. The new optimization problem becomes

$$\underline{P}_T = \min_{\substack{c_{k,n} \in [0, M] \\ \rho_{k,n} \in [0, 1]}} \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k(c_{k,n}) \quad (7)$$

where $c_{k,n}$ and $\rho_{k,n}$ have to satisfy

$$R_k = \sum_{n=1}^N \rho_{k,n} c_{k,n}, \quad \text{for all } k \in \{1, \dots, K\} \quad (8)$$

and

$$1 = \sum_{k=1}^K \rho_{k,n}, \quad \text{for all } n \in \{1, \dots, N\}. \quad (9)$$

For any valid set of $c_{k,n} \in \mathbf{D}$ satisfying the constraints (3) and (4) in the original optimization problem, we can let

$$\rho_{k,n} = \begin{cases} 1, & \text{if } c_{k,n} \neq 0, \\ 0, & \text{if } c_{k,n} = 0. \end{cases} \quad (10)$$

Then, it is easy to show that the same set of $c_{k,n}$ and the corresponding $\rho_{k,n}$ defined in (10) satisfy the constraints (8) and (9) in the new optimization problem. Moreover, with $\rho_{k,n}$ defined in (10), the new cost function in (7) is equal to the cost function in (2). Hence, the minimization problem in (7) is the same as the original optimization problem, except that the minimization is done over a larger set. Consequently, the minimum power obtained in (7) $\underline{P}_T$ is a lower bound to the minimum power obtained in (2), P_T^* .

Another way to interpret the optimization in (7) is to consider $\rho_{k,n}$ as the time-sharing factor for the k th user of the n th subcarrier. For example, in every L OFDM symbol (L being a very large number), user k uses the n th subcarrier in $L\rho_{k,n}$ symbols. Clearly, the average (over L symbols) information data rate and the average transmit power has to be scaled by the same factor $\rho_{k,n}$. Hence, we can consider (7) as the optimization problem when the users are allowed to time-share each subcarrier over a large number of OFDM symbols. However, most wireless communication channels are time varying, and the channels may not stay unchanged long enough for timesharing to be feasible. Hence, in this paper, we will continue to consider the original problem in (2) and use the optimization problem in (7) as a lower bound, even though it has its own physical interpretation.

The modified optimization problem in (7) is more tractable. However, even though the function $f_k(c)$ is convex in c , the terms in the cost function have the form $\rho f_k(c)$, and as a function of (ρ, c) , $\rho f_k(c)$ is not convex in (ρ, c) . To proceed further, we let $r_{k,n} = c_{k,n} \rho_{k,n}$ and rewrite the cost function in terms of $r_{k,n}$ and $\rho_{k,n}$. The constraint on $r_{k,n}$

becomes $r_{k,n} \in [0, M\rho_{k,n}]$, and it can be easily shown that $\rho f_k(c) = \rho f_k(r/\rho)$ is convex in (ρ, r) within the triangular region specified by $\rho \in [0, 1]$ and $r \in [0, M\rho]$. In particular, the Hessian evaluated at any point within this region is a positive semidefinite matrix. Hence, we can reformulate the optimization problem in (7) as a convex minimization problem over a convex set. That is

$$P_T = \min_{\substack{r_{k,n} \in [0, M\rho_{k,n}] \\ \rho_{k,n} \in [0, 1]}} \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k\left(\frac{r_{k,n}}{\rho_{k,n}}\right) \quad (11)$$

where $r_{k,n}$ and $\rho_{k,n}$ have to satisfy

$$R_k = \sum_{n=1}^N r_{k,n}, \quad \text{for all } k \in \{1, \dots, K\} \quad (12)$$

and

$$1 = \sum_{k=1}^K \rho_{k,n}, \quad \text{for all } n \in \{1, \dots, N\}. \quad (13)$$

Using standard optimization techniques in [17], we obtain the Lagrangian

$$L = \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k\left(\frac{r_{k,n}}{\rho_{k,n}}\right) - \sum_{k=1}^K \lambda_k \left(\sum_{n=1}^N r_{k,n} - R_k \right) - \sum_{n=1}^N \beta_n \left(\sum_{k=1}^K \rho_{k,n} - 1 \right) \quad (14)$$

where λ_k and β_n are the Lagrangian multipliers for the constraints (12) and (13), respectively.

After differentiating L with respect to $r_{k,n}$ and $\rho_{k,n}$, respectively, we obtain the necessary conditions for the optimal solution, $r_{k,n}^*$ and $\rho_{k,n}^*$. Specifically, if $\rho_{k,n}^* \neq 0$, we have

$$\begin{aligned} \frac{\partial L}{\partial r_{k,n}} \Big|_{(r_{k,n}, \rho_{k,n})=(r_{k,n}^*, \rho_{k,n}^*)} &= \frac{1}{\alpha_{k,n}^2} f'_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) - \lambda_k \begin{cases} > 0, & \text{if } r_{k,n}^* = 0 \\ = 0, & \text{if } r_{k,n}^* \in (0, M\rho_{k,n}^*) \\ < 0, & \text{if } r_{k,n}^* = M\rho_{k,n}^* \end{cases} \end{aligned} \quad (15)$$

and

$$\begin{aligned} \frac{\partial L}{\partial \rho_{k,n}} \Big|_{(r_{k,n}, \rho_{k,n})=(r_{k,n}^*, \rho_{k,n}^*)} &= \frac{1}{\alpha_{k,n}^2} \left[f_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) - f'_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) \frac{r_{k,n}^*}{\rho_{k,n}^*} \right] \\ &\quad - \beta_n \begin{cases} = 0, & \text{if } \rho_{k,n}^* \in (0, 1) \\ < 0, & \text{if } \rho_{k,n}^* = 1. \end{cases} \end{aligned} \quad (16)$$

On the other hand, if $\rho_{k,n}^* = 0$, then $r_{k,n}^* = 0$, and we have

$$\begin{aligned} r_{k,n} \frac{\partial L}{\partial r_{k,n}} + \rho_{k,n} \frac{\partial L}{\partial \rho_{k,n}} &\geq 0, \\ \text{for all } \rho_{k,n} \in (0, 1] \text{ and } r_{k,n} \in (0, M\rho_{k,n}]. \end{aligned} \quad (17)$$

These necessary conditions can be interpreted by the fact that if the minimum occurs within the constrained region $[(0, 1)$ for $\rho_{k,n}$ and $(0, M\rho_{k,n})$ for $r_{k,n}]$, then the derivative evaluated at the minimum point must be zero. On the other hand, if the optimal solution occurs at a boundary point, then the derivative must be positive along all directions pointing toward the interior of the constraint set. Then, (17) follows from considering the boundary point at $(r_{k,n}^*, \rho_{k,n}^*) = (0, 0)$.

From (15) and (17), we can conclude that

$$r_{k,n}^* = \rho_{k,n}^* f_k'^{-1}(\lambda_k \alpha_{k,n}^2) \quad (18)$$

where

$$\lambda_{q,k} = \begin{cases} f'_k(0)/\alpha_{k,n}^2, & \text{if } f_k'^{-1}(\lambda_k \alpha_{k,n}^2) < 0; \\ \lambda_k, & \text{if } 0 \leq f_k'^{-1}(\lambda_k \alpha_{k,n}^2) \leq M; \\ f'_k(M)/\alpha_{k,n}^2, & \text{if } f_k'^{-1}(\lambda_k \alpha_{k,n}^2) > M. \end{cases}$$

Moreover, from (16) and (17), it follows that

$$\rho_{k,n}^* = \begin{cases} 0, & \text{if } \beta_n < H_{k,n}(\lambda_{q,k}) \\ 1, & \text{if } \beta_n > H_{k,n}(\lambda_{q,k}) \end{cases} \quad (19)$$

where

$$H_{k,n}(\lambda) = \frac{1}{\alpha_{k,n}^2} [f_k(f_k'^{-1}(\lambda \alpha_{k,n}^2)) - \lambda \alpha_{k,n}^2 f_k'^{-1}(\lambda \alpha_{k,n}^2)]. \quad (20)$$

Since constraint (13) must be satisfied, we find from (19) that for each n , if $H_{k,n}(\lambda_{q,k})$ for $k = 1, \dots, K$ are all different, then only the user with the smallest $H_{k,n}(\lambda_{q,k})$ can use that subcarrier. In other words, for the n th subcarrier, if $H_{k,n}(\lambda_{q,k})$ are different for all k , then

$$\rho_{k',n}^* = 1, \quad \rho_{k,n}^* = 0, \quad \text{for all } k \neq k' \quad (21)$$

where

$$k' = \arg \min_k H_{k,n}(\lambda_{q,k}). \quad (22)$$

Hence, it follows that for a fixed set of Lagrange multipliers λ_k , $k = 1, \dots, K$, we can use them to determine k' for each n using (22). The $r_{k,n}^*$ and $\rho_{k,n}^*$ obtained will then form an optimal solution for the optimization problem; however, the individual rate constraint (12) may not be satisfied.

In order to find the set of λ_k such that the individual rate constraints are satisfied, we have obtained an iterative searching algorithm. Starting with some small values for all λ_k , this iterative procedure increases one of the λ_k until the data rate constraint (12) for user k is satisfied. Then, we switch to another user and go through the users one at a time. This process repeats for all users until the data rate constraint for all users are satisfied. This algorithm converges because for a given k , as λ_k increases, $H_{k,n}(\lambda_{q,k})$ for all n decreases, and more $\rho_{k,n}^*$ in (19) become one while $r_{k,n}^*$ in (18) increases for those n where $\rho_{k,n}^* > 0$. Hence, $\sum_{n=1}^N r_{k,n}^*$ increases. During this process, some of the other $\rho_{k',n}^*$ may change from one to zero and consequently decrease the total data rate for other users. However, as all the λ_k increase, $r_{k,n}^*$ increases accordingly. As long as the total data rate is less than MN bits/symbol, which is the total number of bits possibly transmitted within an OFDM symbol, the algorithm

will converge to a solution that satisfies all the constraints. Since the optimization problem is a convex optimization problem over a convex set, the set of necessary conditions is also sufficient, and the solution that satisfies all the necessary conditions is the unique optimal solution.

In the process of adjusting λ_k for $k = 1, \dots, K$, the situation where, for a fixed n , more than one $H_{k,n}(\lambda_{q,k})$ has the same values cannot be ignored. In that case, $\rho_{k,n}^*$ has to take values within the interval $(0, 1)$. This solution suggests that the subcarrier should be shared by multiple users. In practice, this can be done by having these users with $\rho_{k,n}^* > 0$ time share the n th subcarrier, and the ratio of the symbols used by different users are set proportionally to $\rho_{k,n}^*$. The detailed flow chart of the algorithm is given in the Appendix.

Now, we have an algorithm to obtain the optimal values of $\rho_{k,n}^*$ and

$$c_{k,n}^* = \begin{cases} r_{k,n}^* / \rho_{k,n}^*, & \text{if } \rho_{k,n}^* \neq 0 \\ 0, & \text{otherwise.} \end{cases} \quad (23)$$

This solution, when substituted in (7), gives a lower bound to the minimum overall transmit power. However, we cannot use these results immediately in (2). One problem is that $c_{k,n}^*$ may not be in $\mathbf{D}$, and the other is that some $\rho_{k,n}^*$ may be within $(0, 1)$, indicating a time-sharing solution. Furthermore, simply quantizing $c_{k,n}^*$ and $\rho_{k,n}^*$ will not satisfy the individual rate constraints in (3).

To solve this problem, we propose a multiuser adaptive OFDM (MAO) scheme where the subcarrier allocation follows essentially the solution to the lower bound in (7), and then the single-user bit allocation algorithm given in Section III is applied to each user on the allocated subcarriers. Specifically, we modify $\rho_{k,n}^*$ for the optimization problem in (7) by letting for each n $\rho_{k',n}^* = 1$ where $k' = \arg \max_k \rho_{k,n}^*$, and $\rho_{k,n}^* = 0$ for $k \neq k'$. Then, we apply the single-user bit allocation algorithm on each user using the assigned subcarriers. We denote the total transmit power (in energy/symbol) obtained using this MAO scheme by P_T . It is easy to see that $\underline{P}_T \leq P_T^* \leq P_T$, where P_T^* is the minimum power in the original problem, and $\underline{P}_T$ is the minimum power for the modified problem with the relaxed constraints. More specifically, the difference between P_T and the minimum $\underline{P}_T$ gives an upper bound to how far away our MAO scheme is from the solution of our original optimization problem.

V. PERFORMANCE COMPARISON

In this section, we obtain and compare the performance of the MAO scheme with other static subcarrier allocation schemes. We consider a system that employs M-ary quadrature amplitude modulation (MQAM) with $\mathbf{D} = \{0, 2, 4, 6\}$. Square signal constellations (4-QAM, 16-QAM, and 64-QAM) are used to carry two, four, or six bits/symbol. The bit-error probability is upper bounded by the symbol error probability, which is tightly approximated by $4Q[\sqrt{d^2/(2N_0)}]$ [12, p. 281], where d is the minimum distance between the points in the signal constellation. Since the average energy of a M-QAM symbol is equal to $(M-1)d^2/6$, it follows that the required power for supporting c bits/symbol at a given BER

P_e is

$$f(c) = \frac{N_0}{3} \left[Q^{-1} \left(\frac{P_e}{4} \right) \right]^2 (2^c - 1)$$

where we recall that

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt.$$

It is easy to see that $f(c)$ is convex and increasing in c and that $f(0) = 0$.

To evaluate the performance of our scheme, we have simulated 1000 sets of five-path frequency selective Rayleigh fading channels with an exponential power delay profile. Each set of channels consists of K independent channels, one for each user. We use an OFDM system with 128 subcarriers over a 5 MHz band along with a total (over all users) transmission rate equal to 512 bits/symbol (or equivalently, an average of four bits/subcarrier). Recall that the single-sided power spectral density level N_0 is equal to unity, and we assume that the average subcarrier channel gain $\mathbf{E}[\alpha_{k,n}]^2$ is equal to unity for all k and n .

For comparison purposes, we have also considered three other static multiuser subcarrier allocation methods. Two of them are based on the multiple access methods described in [7]. The methods are presented as follows.

- OFDM-TDMA: each user is assigned a predetermined TDMA time slot and can use all the subcarriers within that time slot exclusively.
- OFDM-FDMA: each user is assigned a predetermined band of subcarriers and can only use those subcarriers exclusively in every OFDM symbol.
- In a frequency selective fading channel, there is a high correlation between the channel gains of adjacent subcarriers. In order to avoid the situation where all subcarriers of a user are in deep fade, we propose an enhanced version of OFDM-FDMA, which we shall refer to as OFDM Interleaved-FDMA.
- OFDM Interleaved-FDMA: this is the same as OFDM-FDMA except that subcarriers assigned to a user are interleaved with other users' subcarriers in the frequency domain.

The time and subcarrier assignment of these three multiuser OFDM schemes are illustrated in Fig. 2. Note that these static schemes have predetermined subcarrier allocations which are independent of the channel gains of the users. The main difference between the proposed MAO scheme and these static schemes is that MAO assigns subcarriers adaptively based on the instantaneous channel gains. To ensure a fair comparison, we use the optimal single-user bit allocation (OBA) for each user on the assigned subcarriers. For comparison purposes, we also show the results when equal bit allocation (EBA) is employed on the assigned subcarriers for these three OFDM schemes. Notice that when using EBA, all three schemes will have the same performance in an uncoded system. This is because the average bit signal-to-noise ration (SNR) needed is a function of only the marginal probability density function of each subcarrier gain.

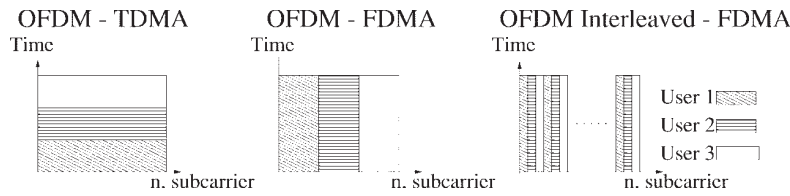


Fig. 2. Subcarrier and time-slot allocations of OFDM-TDMA, OFDM-FDMA, and OFDM interleaved-FDMA schemes.

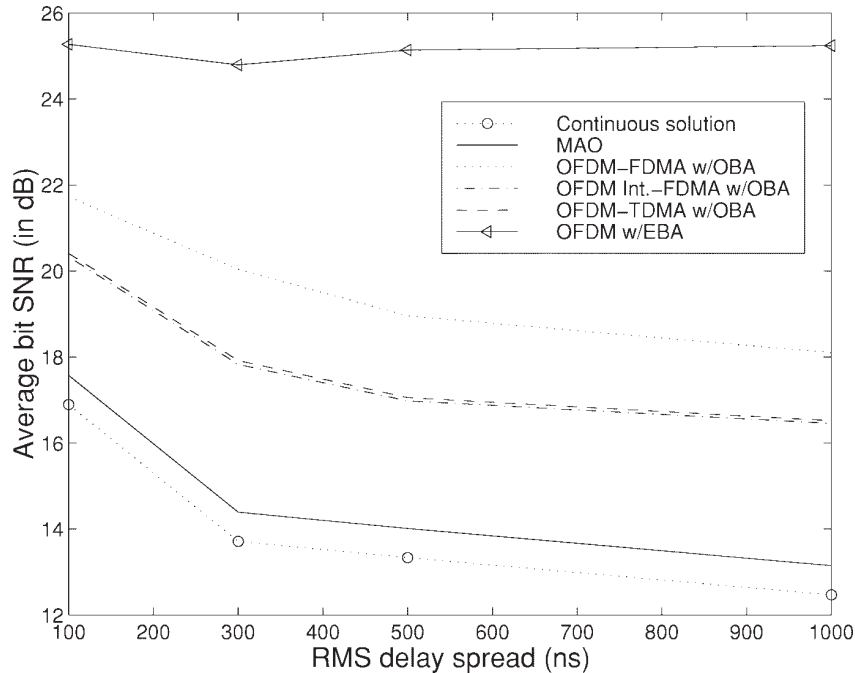


Fig. 3. Average bit signal-to-noise ratio (SNR) required by different schemes in various root mean square (RMS) delay spreads in a five-user system with $P_e = 10^{-4}$.

Fig. 3 shows the average bit SNR needed to achieve a BER at $P_e = 10^{-4}$ for a five-user system versus the root mean square (RMS) delay spread (for definition, see for example [18, p. 160]) for different multiuser OFDM schemes. The average required transmit power (in energy per bit) is defined as the ratio of the overall transmit energy per OFDM symbol (including all subcarriers and all users) to the total number of bits transmitted per OFDM symbol. Moreover, we define the average bit SNR as the ratio of the average transmit power to the noise PSD level N_0 . As we assume that the data rate is fixed and that N_0 is just a constant, the overall transmit power is proportional to the average bit SNR. For ease of comparison, we have used the average bit SNR for comparison. We find in Fig. 3 that the MAO scheme is never more than 0.6 dB from the lower bound. Since the bit SNR of the optimal combined subcarrier, bit, and power allocation algorithm must lie between the bit SNR's achieved by the lower bound and the MAO scheme, we find that the MAO scheme is never more than 0.6 dB away from the optimal solution. On the other hand, we observe that our proposed MAO scheme is 3–5 dB better than the static subcarrier allocation schemes with OBA, which are in turn 5–10 dB better than that with EBA. We also find that when OBA is used, the OFDM interleaved-FDMA

scheme and the OFDM-TDMA scheme have very similar performance, and both of them outperform the OFDM-FDMA scheme.² A closer observation of Fig. 3 also indicates that the gains achieved by optimal bit allocation and optimal multiuser subcarrier allocation increase with the RMS delay spread. This is mainly because the larger the RMS delay spread, the more the fading variation and hence higher gains can be obtained when the allocation is performed adaptively.

Fig. 4 shows the average bit SNR (in dB) needed to achieve the same BER versus the number of users when the RMS delay spread is 100 ns. We find that the savings in the required bit SNR achieved by MAO when compared to other schemes are roughly the same, independent of the number of users in the system.

While these two figures show the improvement in the required bit SNR, the results can perhaps be more easily understood using the more familiar BER versus bit SNR curves. For each BER requirement, we compute $f(c)$ for all $c \in \mathcal{D}$ and then use our algorithm to calculate the subcarrier

²OFDM-FDMA refers only to the specific FDMA scheme that assigns to each user a contiguous band of subcarriers as shown in Fig. 2, but not the general FDMA schemes. In fact, both OFDM interleaved-FDMA and MAO can be considered as different forms of FDMA and they are not outperformed by the OFDM-TDMA scheme.

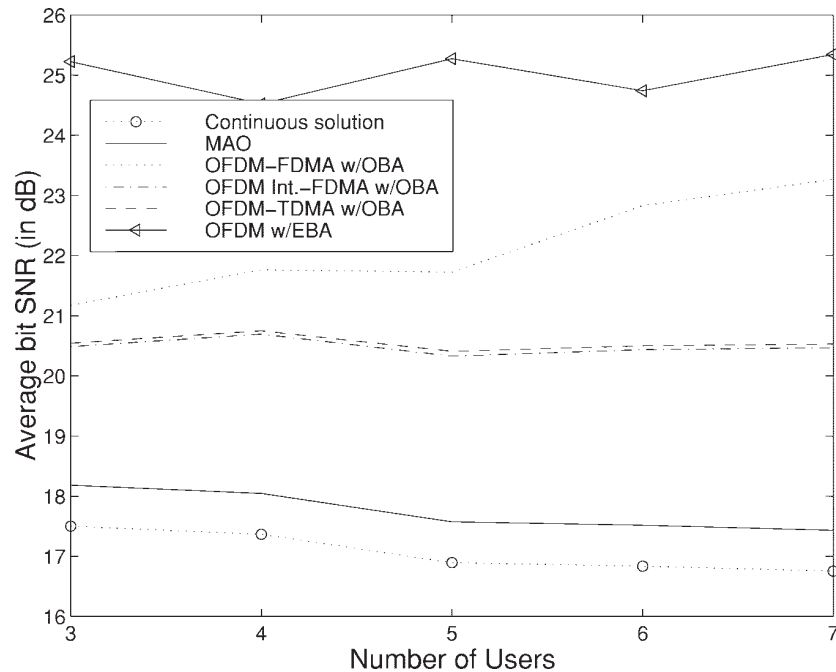


Fig. 4. Average bit SNR required by different schemes versus the number of users in a multiuser OFDM system with 100 ns RMS delay spread, and $P_e = 10^{-4}$.

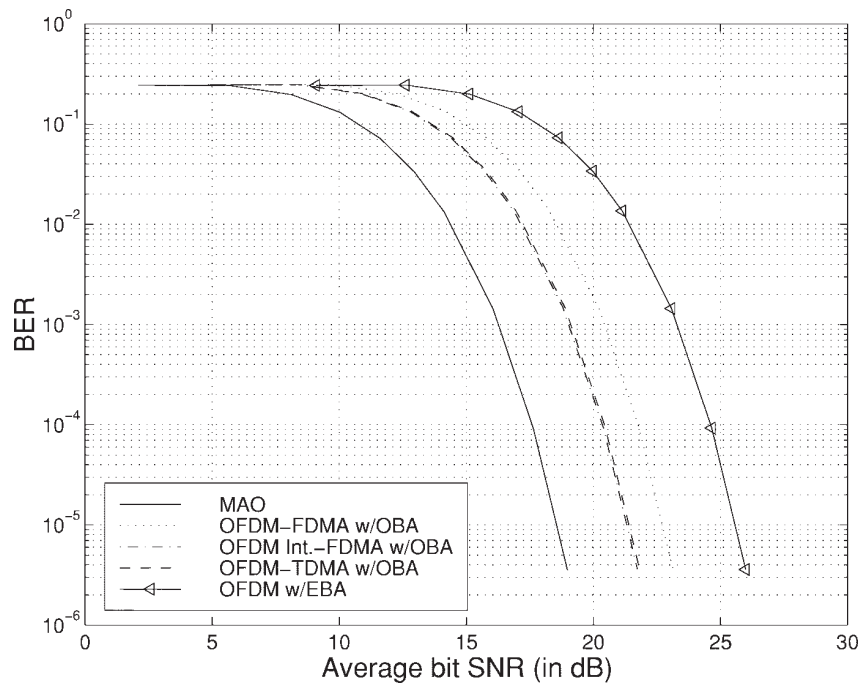


Fig. 5. BER versus average bit SNR for various subcarrier allocation schemes.

allocation for the MAO case. For all other static subcarrier allocation schemes, the allocations are independent of the BER. Once the subcarrier allocation is fixed, we apply the optimal bit and power allocation algorithm to every user. The final average power per bit divided by the noise power spectral density level gives the average bit SNR. We repeat this procedure for different BER values, and the results are

plotted in Fig. 5 for a five-user system with an RMS delay spread equal to 100 ns. We find that our proposed MAO has at least 3–4 dB advantage over all other schemes.

Another way to illustrate the impact of the bit and subcarrier allocation is to consider the area of coverage for a given outage probability, assuming that the BS has a maximum transmit power. We consider a circular cell with five users, indepen-

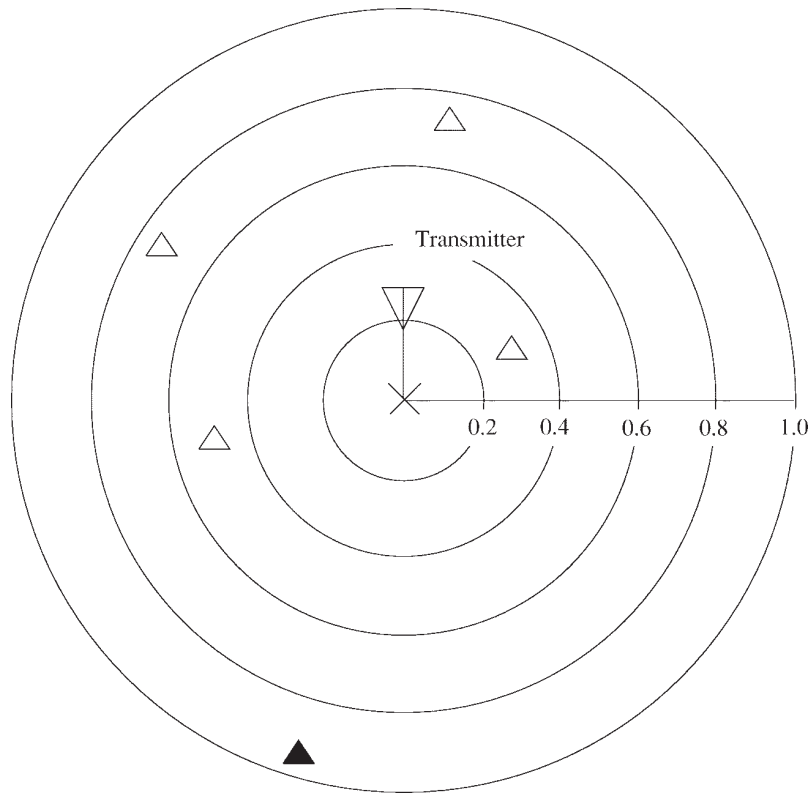


Fig. 6. Cell for analyzing the outage probability.

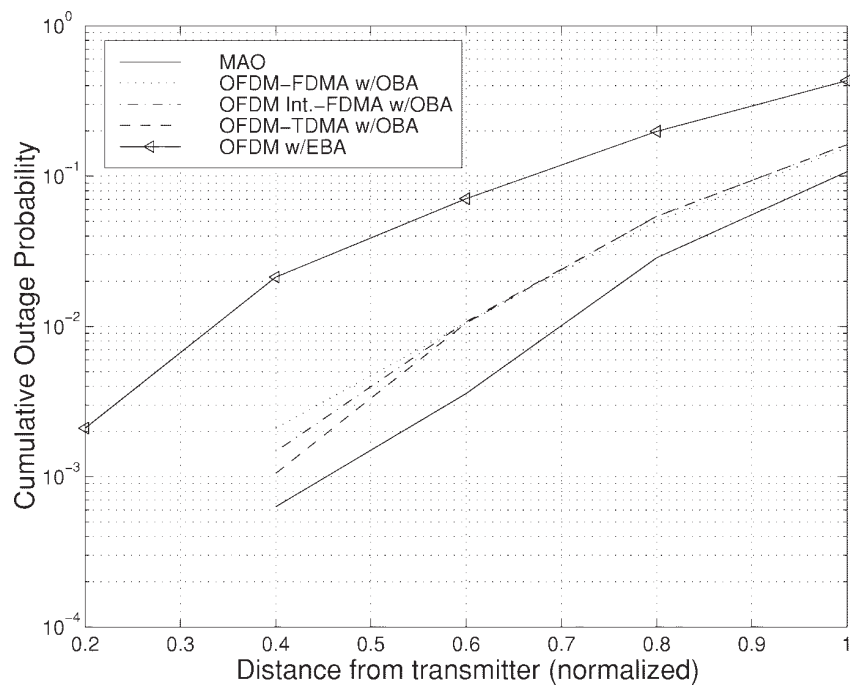


Fig. 7. Outage probability at 17 dB fading margin.

dently and uniformly distributed within the cell. A typical scenario is shown in Fig. 6, where the triangles represent the five users. In addition to frequency selective fading, path loss and log-normal shadowing are also included in simulating the

actual channel gains seen by the users. Using these channel gains, subcarriers and bits assigned to each user are determined by the various multiple access schemes and the total required transmit power is calculated. If the total power for all five users

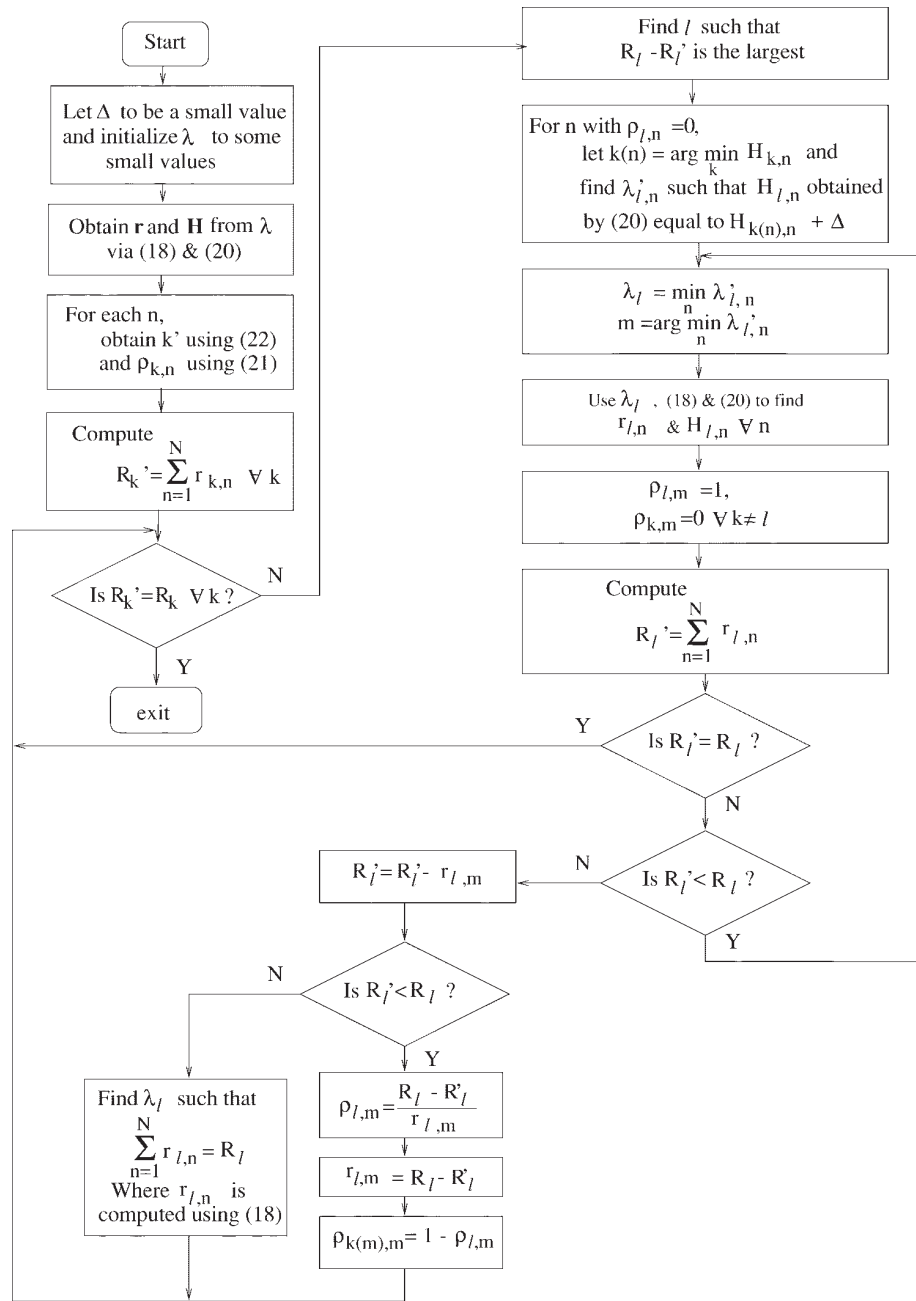


Fig. 8. Flow chart of the multiuser subcarrier allocation algorithm.

exceeds the maximum power of the BS, the user requiring the largest transmit power (in this case, the black one) is dropped and counted as one outage event occurring at a distance equal to the distance between the BS and the dropped user. This process continues until the transmit power is smaller than the maximum power of the BS. In this example, the maximum transmit power is set to the transmit power required for all five users assuming that they are all located at the boundary of the cell, taking into account the path loss effect and a 17 dB fading margin for shadowing.

The cumulative outage probabilities at various normalized distances, normalized to the cell radius, are plotted in Fig. 7. A

cumulative outage probability of 5% at a normalized distance of 0.8 means that there is a 5% chance of outage for a mobile located more than $0.8R$ away from the BS where R is the radius of the cell. We observe that MAO outperforms others with a large reduction in the outage probability at all distances. Alternatively, if the same outage probability is maintained, say at 1%, the coverage are provided by MAO is 36% larger than the best of all other schemes.

VI. CONCLUSION

In this paper, we considered OFDM transmission in a multiuser environment and formulated the problem of min-

imizing the overall transmit power by adaptively assigning subcarriers to the users along with the number of bits and power level to each subcarrier. In particular, we derived a multiuser adaptive subcarrier and bit allocation algorithm. Given the instantaneous channel information, the algorithm obtains a suboptimal subcarrier allocation, and then single-user bit allocation is applied on the allocated subcarriers. Using this scheme, the overall required transmit power can be reduced by about 5–10 dB from the conventional OFDM without adaptive modulation. Likewise, the transmit power can be reduced by about 3–5 dB from the conventional OFDM with adaptive modulation and adaptive bit allocation, but without adaptive subcarrier allocation. The reduction in transmit power can also be translated to a significant reduction in the required bit SNR for a given BER. Moreover, the same improvement can also be translated to a reduction in the outage probability or to an increase in the area of coverage.

The results in this paper assume perfect channel estimation, and we have not considered issues related to imperfect implementation, such as imperfect synchronization. As channel estimation in wireless fading channels is in general not very accurate, the effect of nonideal channel information on the performance of our proposed MAO scheme is a very important issue. We have started looking at this issue, and our preliminary results have indicated that the MAO scheme is not very sensitive to channel estimation errors. Nevertheless, detailed sensitivity studies will be needed before the algorithm can be applied to practical systems.

APPENDIX

A flow chart providing the detailed description of the multiuser subcarrier allocation algorithm is shown in Fig. 8.

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Cheong Yui Wong was born in Hong Kong, in 1975. He received the B.Eng. degree in electronic engineering from the Hong Kong University of Science and Technology (HKUST), Clear Water Bay, Hong Kong in 1997. He is currently pursuing the M.Phil. degree at HKUST and expects to graduate in 1999.

During his master studies, he joined a project that aims at achieving wireless multimedia communication using OFDM. His research interests include adaptive modulation, OFDM, and optimization.



Roger S. Cheng (S'85–M'91) received the B.S. degree from Drexel University, Philadelphia, PA, in 1987, and both the M.A. and Ph.D. degrees from Princeton University, Princeton, NJ, in 1988 and 1991, respectively, all in electrical engineering.

From 1987 to 1991, he was a Research Assistant in the Department of Electrical Engineering, Princeton University. From 1991 to 1995, he was an Assistant Professor in the Electrical and Computer Engineering Department of University of Colorado, Boulder. In June 1995, he joined the Faculty of Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong, where he is currently an Assistant Professor in the Department of Electrical and Electronic Engineering. He has also held visiting positions with Qualcomm, San Diego, in the summer of 1995 and with Institute for Telecommunication Sciences, NTIA, Boulder, CO, in the summers of 1993 and 1994. His current research interests include wireless communications, multiuser communications, spread spectrum CDMA, digital implementation of communication systems, wireless multimedia communications, information theory, and coding.

Dr. Cheng was the recipient of the Meitec Junior Fellowship Award from the Meitec Corporation in Japan, the George Van Ness Lothrop Fellowship from the School of Engineering and Applied Science in Princeton University, and the Research Initiation Award from the National Science Foundation. He was a member of the Technical Program Committees of GLOBECOM'98 and IEEE Wireless Communications and Networking Conference (WCNC'99). He has served as Guest Editor of the special issue on Multimedia Network Radios in the IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, as Associate Editor of the IEEE TRANSACTIONS ON SIGNAL PROCESSING, and as Membership Chair for of the IEEE Information Theory Society.



Khaled Ben Letaief (S'85–M'86–SM'97) received the B.S. degree (with distinction), the M.S. degree, and the Ph.D. degrees in electrical engineering from Purdue University, West Lafayette, IN, in 1984, 1986, and 1990, respectively.

In 1985, he was a Graduate Instructor in the School of Electrical Engineering at Purdue University and he taught courses in communications and electronics. From 1990 to 1993, he was on the Academic Staff at the Department of Electrical and Electronic Engineering, University of Melbourne,

Australia, where he was also a Member of the Center for Sensor Signal and Information Systems. Since 1993, he has been with the Department of Electrical and Electronic Engineering, Hong Kong University of Science and Technology (HKUST), Clear Water Bay, Hong Kong, where is currently an Associate Professor. During several semesters from 1995 to 1998, he served as the Departmental Director of the Undergraduate Studies program. His research interests include wireless personal and mobile communications, spread spectrum systems, optical fiber networks, multiuser detection, wireless multimedia communications, and CDMA systems. He is an Editor of the *Wireless Personal Communications Journal* and is also a Guest Editor of the 1999 *Wireless Personal Communications Journal's* Special Issue on Intelligent Multimedia Systems, Terminals, and Components.

Dr. Letaief has been an active member of various professional societies and has published papers in several journals and conference proceedings. In 1990, he was awarded the Mangoon Teaching Award at Purdue University. From 1995 to 1997, he was awarded the Teaching Excellence Appreciation Award by the School of Engineering, HKUST. He was also the 1998 University Recipient of the Michael G. Gale Medal for Distinguished Teaching (the highest teaching award at the university). He is the Editor for Wireless Systems of the IEEE TRANSACTIONS ON COMMUNICATIONS and a Technical Editor of the *IEEE Communications Magazine*. He has been a Guest Editor of the 1997 IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS' Special Issue on Computer-Aided Modeling, Analysis, and Design of Communications Links. In addition, he served as the Technical Program Chair of the 1998 IEEE Mini-Conference on Communications Theory (CTMC'98) which was held in Sydney, Australia. He is currently serving as an Officer in the IEEE Communications Society Technical Committee on Personal Communications.



Ross D. Murch (S'85–M'87–SM'98) received the Ph.D. degree in electrical and electronic engineering from the University of Canterbury in 1990 for research involving electromagnetic inverse scattering.

From 1990 to 1992, he was a Post Doctoral Fellow at the Department of Mathematics and Computer Science, Dundee University, Scotland, where he worked on approximate approaches to electromagnetic scattering from three-dimensional objects. Currently he is an Associate Professor

in the Department of Electrical and Electronic Engineering at the Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong. He is also the founding Director of the Center for Wireless Information Technology, which began in August 1997. From August to December 1998, he was on sabbatical leave at Allgon Mobile Communications, Sweden and AT&T Research Labs, NJ. He also acts as a Consultant for various industrial projects on wireless communications and has two U.S. patents related to wireless communication. His most recent research activities center around the application of electromagnetics and signal processing to wireless communications. Research interests include: antenna design for mobile telephone handsets, smart antenna algorithms and propagation prediction for wireless communications.

Dr. Murch is an Editor of IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, Wireless Series, and he has been involved in the organization of several international conferences. He is a Chartered Engineer, a Member of the Institute of Electrical Engineers, and also an URSI correspondent. He won an URSI Young Scientist Award in 1993 and an Engineering Teaching Excellence Appreciation award in 1996.

Hsieh-Yee Declaration

Exhibit 0002

An Investigation into Time-Domain Approach for OFDM Channel Estimation

Hlaing Minn, *Student Member* and Vijay K. Bhargava, *Fellow, IEEE*

Abstract—A time-domain based channel estimation for OFDM system with pilot-data multiplexed scheme is investigated. As an approximation to linear minimum mean square estimator (LMMSE), a time-domain based channel estimation is proposed where intra-symbol time-averaging and most significant channel taps selection are applied. The relation and differences of the proposed method to DFT-based LMMSE methods are discussed. The performances of the proposed method, DFT-based LMMSE method [15] and the methods of [16] and [17] are evaluated in multipath fading channels. The simulation results show that proposed method achieves almost the same performance as DFT-based LMMSE method and better BER performance than the other methods while keeping less complexity.

Index Terms—Channel estimation, intra-symbol time-averaging, linear minimum mean square error estimator, most significant taps selection, orthogonal frequency division multiplexing (OFDM).

I. INTRODUCTION

ORTHOGONAL frequency division multiplexing (OFDM) [1]–[3] has recently achieved much popularity due to its desirable properties such as its robustness to multipath delay spread and impulse noise, its high data rate transmission capability with high bandwidth efficiency, and its feasibility in application of adaptive modulation and power allocation across the subcarriers according to the channel conditions. It has been adopted in wireline applications such as ADSL (Asymmetric Digital Subscriber Line) [4], broadcasting services such as European DAB (Digital Audio Broadcasting) [5], DVB-T (Terrestrial Digital Video Broadcasting) [6] and Japanese ISDB-T (Terrestrial Integrated Services Digital Broadcasting) [7], high rate wireless LAN standards such as ETSI HiperLAN 2 and IEEE 802.11(a), and multimedia wireless services such as Japanese MMAC (Multimedia Mobile Access Communications) [8], [9].

If noncoherent OFDM system is used, the system complexity will be reduced at the cost of 3–4 dB performance loss [10]. If coherent OFDM system is adopted, channel estimation becomes a requirement and usually pilot tones are used for channel (frequency response) estimation. Pilot tones can be inserted in all subcarriers of a particular OFDM symbol forming an OFDM training symbol, in which case training symbols are transmitted at an appropriate regular rate determined by the time varying nature of the wireless channel. Another approach is that, instead of using all subcarriers, the pilot tones are multiplexed

with data to form OFDM symbols. The subcarrier spacing between pilot tones is usually determined by the frequency selectivity of the wireless channel. The pilot multiplexing can be allowed for all OFDM symbols (i.e., all the time of transmission) or at an appropriate rate depending on the time selectivity of the wireless channel. In channel estimation using training symbols, decision directed approach has to be used. If complexity is affordable, time interpolation (e.g., [11]) can be used to improve the performance. If pilot tone multiplexing is used, the frequency interpolation has to be performed [12]–[14]. Similarly, time domain interpolation can be performed at the cost of complexity. Most of the channel estimation approaches may be viewed as DFT-based approaches [15], [16], [10], [11], where LS (Least square) channel (frequency response) estimates are fed to IFFT block to get time domain channel impulse response estimate, and then appropriately processed and transformed back to frequency domain by FFT. A DFT-based approach for OFDM system with transmit and receive antenna diversity has been discussed in [10].

In this paper, we investigate a time-domain channel estimation approach, namely FPTA (Frequency Pilot Time Average) [17] which applies intra-symbol time-domain averaging of identical parts of the pilot signal. We also propose a time-domain approach for OFDM channel estimation which achieves performance gain over LS or FPTA approaches.

The rest of the paper is organized as follows. In Section II, the OFDM system and LS channel estimation are presented for notational description. The method of time-averaging the identical parts of a pilot signal is briefly presented in Section III, and our analysis on this intra-symbol time-averaging is given in Section IV. Section V presents the proposed time-domain channel estimation and its relation to DFT-based approaches are discussed in Section VI. Simulation results are discussed in Section VII and conclusions are given in Section VIII.

II. SYSTEM DESCRIPTION

Suppose the pilot tones $P[k]$ are multiplexed with data $D[k]$ in all OFDM symbols at a pilot ratio $1/K$ (ratio of the number of pilot tones to the total number of subcarriers) where k is subcarrier index $(0, 1, \dots, N-1)$ with N being the total number of subcarriers, and $P[k]$ and $D[k]$ are zeros except at their corresponding subcarriers. Then the transmitted OFDM signal in discrete-time domain, excluding guard-interval, can be expressed as

$$s[n] = \text{IFFT}_N\{P[k]\} + \text{IFFT}_N\{D[k]\} \\ = p[n] + d[n] \quad (1)$$

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The authors are with the Department of Electrical and Computer Engineering, University of Victoria, Victoria, B.C., Canada V8W 3P6.

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where $IFFT_N\{\}$ is N -point inverse Fast Fourier transform and n is the time-domain index $(0, 1, \dots, N-1)$ of an OFDM symbol. Suppose the wireless channel has a discrete-time impulse response given by

$$h[n] = \sum_{l=0}^{L-1} \alpha_l \delta[n - \tau_l] \quad (2)$$

where

- α_l is complex path gain of l th path,
- τ_l is the delay of l th path, and
- L is the total number of channel paths.

For simplicity, time dependence nature of the channel impulse response is suppressed in the notation.

After passing through a multipath wireless channel, the time-domain received samples of an OFDM symbol, if appropriate cyclic prefix guard samples are used, is given by

$$r[n] = s[n] \otimes h[n] + w[n] \quad (3)$$

where $\otimes$ represents N -point circular convolution, $\{w[n]\}$ are independent and identically distributed (*iid*) AWGN samples with zero mean and variance of σ_w^2 . Assuming perfect synchronization, the FFT output frequency-domain subcarrier symbols can be expressed as

$$R[k] = FFT_N\{r[n]\} = H[k] P[k] + H[k] D[k] + W[k] \quad (4)$$

where $W[k] = FFT_N\{w[n]\}$ is frequency-domain AWGN noise samples with zero mean and variance $\sigma_f^2 = N\sigma_w^2$. Then the channel frequency response at the pilot tones can be estimated by

$$\hat{H}[m] = \frac{R[m]}{P[m]} = H[m] + \frac{W[m]}{P[m]} \quad (5)$$

where m is the subcarrier index for pilot tones. This channel estimate is called LS (least square) estimate. The channel responses at other subcarriers can be obtained by interpolation.

III. FREQUENCY PILOT TIME AVERAGE (FPTA) METHOD

In FPTA [17] approach, positive and negative alternatively polarized pilot tones are multiplexed with data at a pilot ratio of $1/K$. The frequency-domain pilot tones can be expressed as

$$P[k] = \begin{cases} (-1)^m A, & k = Km \\ 0, & k = Km + i \end{cases} \quad (6)$$

where

- K, m and i are integers, $1 \leq i \leq K-1, 0 \leq m \leq M-1$,
- A is pilot amplitude and
- $M = N/K$ is an integer.

The corresponding time-domain pilot samples can be expressed as

$$\begin{aligned} p[n] &= IFFT_N\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] W_N^{-nk} \\ &= \frac{A}{K} \sum_{r=0}^{K-1} \delta[n - Mr - M/2] \end{aligned} \quad (7)$$

where $W_N = \exp(-j2\pi/N)$.

In FPTA approach, since there are K identical parts of time-domain pilot samples, the corresponding parts of received samples are averaged over K parts. This intra-symbol time-domain averaging reduces the variance of noise samples by K times, i.e., $\sigma_{t,avg}^2 = \sigma_t^2/K$. Based on this observation, [17] defined time-domain pilot to noise ratio as $p[n]/\sigma_{t,avg}$ and compared it with frequency domain pilot to noise ratio $P[k]/\sigma_f$ resulting in the following expression

$$\frac{p[n]/\sigma_{t,avg}}{P[k]/\sigma_f} = \sqrt{\frac{N}{K}} \quad (8)$$

which was mentioned as the gain of FPTA approach over frequency domain approach such as LS estimation discussed in Section II. However, this result is optimistic and more details on it will be discussed in the following section.

IV. ANALYSIS OF INTRA-SYMBOL TIME AVERAGING IN FPTA

In order to investigate the pilot to noise ratio insightfully, we neglect the data part and consider only the pilot affected by multipath channel and Gaussian noise. Then the time-domain received samples vector of an OFDM signal can be given by

$$\mathbf{r} = \mathbf{p}' + \mathbf{w} \quad (9)$$

where $\mathbf{r} = [\mathbf{r}_0, \mathbf{r}_1, \dots, \mathbf{r}_{K-1}]$ with $\mathbf{r}_i = [r_i[0], r_i[1], \dots, r_i[M-1]]$, $\mathbf{p}'$ is N -point circular convolution of pilot signal $\mathbf{p}$ and channel impulse response $\mathbf{h}$ and can be expressed as $\mathbf{p}' = [\mathbf{p}'_0, \mathbf{p}'_1, \dots, \mathbf{p}'_{K-1}]$ with $\mathbf{p}'_i = [p'_i[0], p'_i[1], \dots, p'_i[M-1]]$ and $\mathbf{w} = [w[0], w[1], \dots, w[N-1]]$ is *iid* Gaussian noise samples.

Then averaging the received samples over K parts, assuming the channel impulse response is constant over the OFDM symbol interval, gives

$$\mathbf{r}_{avg} = \frac{1}{K} \sum_{i=0}^{K-1} \mathbf{r}_i = \mathbf{p}'_0 + \mathbf{w}_{avg} \quad (10)$$

where $\mathbf{w}_{avg} = [w_{avg}[0], w_{avg}[1], \dots, w_{avg}[M-1]]$ with $w_{avg}[i] = (1/K) \sum_{j=0}^{K-1} w[i + jM]$. Since $\{w[n]\}$ are *iid* zero mean complex Gaussian random variables with variance σ_t^2 , $\{w_{avg}[i]\}$ are *iid* zero mean complex Gaussian random variables with variance $\sigma_{t,avg}^2 = \sigma_t^2/K$.

Now, we investigate two possible approaches for channel estimation using intra-symbol time-averaged received samples. The first approach, which will be denoted by FPTA-1 in the rest, is to reconstruct the received samples of length N by repeating the $\mathbf{r}_{avg}, K$ times. Then the reconstructed samples $\mathbf{y}$ can be expressed as

$$\mathbf{y} = [\mathbf{r}_{avg}, \mathbf{r}_{avg}, \dots, \mathbf{r}_{avg}] = \mathbf{p}' + \mathbf{v} \quad (11)$$

where $\mathbf{v} = [\mathbf{w}_{avg}, \mathbf{w}_{avg}, \dots, \mathbf{w}_{avg}]$ is K times repeated version of $\mathbf{w}_{avg}$. The corresponding frequency domain samples are

$$Y[k] = FFT_N\{y[n]\} = H[k] P[k] + V[k] \quad (12)$$

where $\{P[k]\}$ are original pilot tones, $H[k]$ is channel frequency response and $V[k] = FFT_N\{\mathbf{v}\}$ is frequency-domain zero

mean Gaussian noise term. The channel frequency responses at pilot tones can be estimated by

$$\hat{H}[m] = Y[m]/P[m] = H[m] + V[m]/P[m] \quad (13)$$

where $m = 0, K, 2K, \dots, (M-1)K$.

The variance of $V[m]$ can be given as follows:

$$\begin{aligned} \text{var}\{V[m]\} &= E\{V[m]V^*[m]\}, \\ &\quad m = 0, K, 2K, \dots, (M-1)K \\ &= E\left\{\sum_{i=0}^{N-1} v[i] W_N^{mi} \sum_{j=0}^{N-1} v^*[j] W_N^{-mj}\right\} \\ &= E\left\{\sum_{i=0}^{M-1} \sum_{j=0}^{K-1} v[i] W_N^{m(i+lM)} \sum_{j=0}^{M-1} v^*[j] \right. \\ &\quad \left. \cdot \sum_{n=0}^{K-1} W_N^{-m(j+nM)}\right\} \\ &= \sum_{i=0}^{M-1} \sum_{j=0}^{M-1} E\{v[i]v^*[j]\} \sum_{l=0}^{K-1} W_N^{m(i+lM)} \\ &\quad \cdot \sum_{n=0}^{K-1} W_N^{-m(j+nM)} \\ &= \sum_{i=0}^{M-1} \sigma_{t,avg}^2 \left| \sum_{l=0}^{K-1} W_N^{m(i+lM)} \right|^2 \\ &= \sum_{i=0}^{M-1} \sigma_{t,avg}^2 K^2 = N\sigma_t^2 \end{aligned} \quad (14)$$

which is the same as the variance of frequency-domain noise term without intra-symbol time-averaging (i.e., $W[k]$ in (4)). The pilot to noise ratio is also the same, given by $A/\sqrt{N}\sigma_t^2$.

The second possible approach, which will be denoted by FPTA-2 in the rest, is to use $FFT_N\{r_{avg}\}$ together with $FFT_N\{p_0\}$ where p_0 is one of the identical parts of time-domain pilot samples. The channel frequency response is estimated as follows:

$$P_0[k] = FFT_N\{p_0[n]\} = A/K, \quad \text{for } k = 0, 1, \dots, N-1 \quad (15)$$

$$R_{avg}[k] = FFT_N\{r_{avg}[n]\} = H[k]P_0[k] + W_{avg}[k] \quad (16)$$

$$\hat{H}[k] = R_{avg}[k]/P_0[k] = H[k] + W_{avg}[k]/P_0[k] \quad (17)$$

where $\{W_{avg}[k]\}$ are zero mean complex Gaussian noise terms and $\{P_0[k]\}$ are the equivalent pilot tones for this approach. It can be shown that the variance of $W_{avg}[k]$ is $M\sigma_{t,avg}^2 (= (M/K)\sigma_t^2)$. The pilot to noise ratio is the same as the previous case.

The channel estimation error is given by $e_H = \hat{H}[k] - H[k]$ which is the second term in $\hat{H}[k]$ equation. Since its mean is zero, the mean square error (*mse*) of channel estimate, $mse\{\hat{H}[k]\} = E\{e_H e_H^*\}$, is equal to the variance of the second term. Hence the channel estimation *mse*'s of LS method and intra-symbol time-averaging methods are the same and equal to $N\sigma_t^2/A^2$.

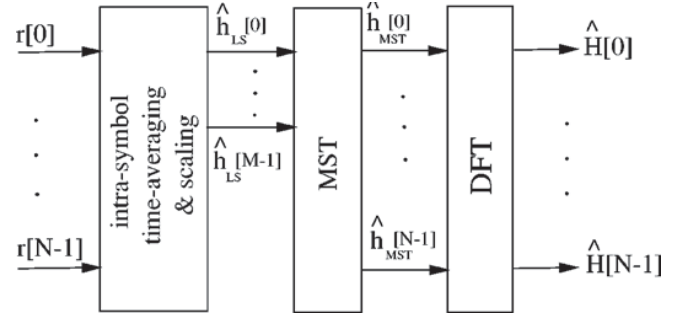


Fig. 1. Most Significant Taps (MST) Method.

The above discussion also indicates that intra-symbol time-averaging alone does not achieve channel estimation performance gain and the result in (8) is optimistic. Another way of explanation for the result in (8) is that it considers only one time-domain sample (after averaging) while other samples also affect the channel estimate. The result would be appropriate if $p[n]/\sigma_{t,avg}$ in (8) is obtained by averaging over all M samples (after averaging). In the rest, the gain of a channel estimation method over another (if not mentioned, LS method is assumed) will be expressed by the *mse* gain (i.e., the ratio of *mse*'s) or the (frequency-domain) pilot-to-noise power ratio instead of the pilot amplitude to noise standard deviation ratio.

V. MOST SIGNIFICANT TAPS APPROACH

For practical multipath wireless channels, there are not so many channel paths with significant energy (if compared to the FFT size N). Hence, among N samples (taps) of the channel impulse response estimate, many samples (taps) will have little or no energy at all except noise perturbation. Neglecting those nonsignificant channel taps in channel estimation may introduce some performance degradation if some of the channel energy is missed, but at the same time it will eliminate the noise perturbation from those taps. Usually total noise perturbation from those neglected channel estimate taps is much higher than the multipath energy contained in them, especially for low SNR values. Hence, neglecting those nonsignificant channel estimate taps can improve the channel estimation performance significantly and this fact is applied in the proposed method as shown in Fig. 1.

We consider the same scenario as in [17] where pilot tones are multiplexed with data in each OFDM symbol at a pilot ratio of $1/K$. The pilot tone used in the proposed *MST* approach is

$$P[k] = \sum_{m=0}^{M-1} A\delta[k - mK], \quad k = 0, 1, \dots, N-1 \quad (18)$$

and the corresponding time-domain samples contain K identical parts and are given by

$$p[n] = \sum_{m=0}^{K-1} \frac{A}{K} \delta[n - mM], \quad n = 0, 1, \dots, N-1 \quad (19)$$

If the maximum channel delay spread is less than the length of an identical part, which can be designed to satisfy this, then the time-domain received samples corresponding to the time-domain pilot samples contain K parts, each representing a

scaled channel impulse response for the respective part corrupted by AWGN. If the channel path gains remain essentially the same over an OFDM symbol interval, which is usually the case since OFDM systems are usually designed to satisfy this in order to maintain orthogonality among subcarriers, then the received samples corresponding to time-domain pilot samples contain K repeated version of scaled channel impulse response which are independently corrupted by AWGN. In order to choose most significant channel taps, those K parts can be averaged so that the noise variance is reduced by K times and more reliable most significant channel taps can be obtained. In mathematical expression, the time-domain received samples corresponding to time-domain pilot samples, (For simplicity, data part is neglected in the received samples expression. The mean of data part after averaging is zero), can be given by

$$\begin{aligned} r[n] &= h[n] \otimes p[n] + w[n] \\ &= \frac{A}{K} \sum_{m=0}^{K-1} h[n - mM] + w[n], \\ n &= 0, 1, \dots, N-1 \end{aligned} \quad (20)$$

After averaging, we have the noise-corrupted scaled channel impulse response

$$r_{avg}[n] = \frac{A}{K} h[n] + w_{avg}[n], \quad n = 0, 1, \dots, M-1 \quad (21)$$

Then the raw channel impulse response estimate is given by

$$\hat{h}_{LS}[n] = \frac{K}{A} r_{avg}[n] = h[n] + \frac{K}{A} w_{avg}[n], \quad n = 0, 1, \dots, M-1 \quad (22)$$

Now, the most significant J channel taps are chosen as the J largest amplitude channel taps. Let the channel tap indexes for those most significant J taps be denoted by $n_0, n_1, \dots, n_{J-1}$. Then the time-domain channel impulse response estimate of proposed MST method is obtained by setting the other channel tap gains to zero as shown below:

$$\hat{h}_{MST}[n] = \sum_{i=0}^{J-1} \hat{h}_{LS}[n_i] \delta[n - n_i], \quad n = 0, 1, \dots, N-1 \quad (23)$$

The channel frequency response estimate is directly obtained by applying FFT to $\{\hat{h}_{MST}[n]\}$ as

$$\begin{aligned} \hat{H}_{MST}[k] &= FFT_N\{\hat{h}_{MST}[n]\}, \quad k = 0, 1, \dots, N-1 \\ &= H[k] - H_{res}[k] + W_{avg}[k] \end{aligned} \quad (24)$$

where

$$\begin{aligned} H[k] &= \sum_{n=0}^{N-1} h[n] W_N^{kn} \\ H_{res}[k] &= \sum_{t=0}^{N-J-1} h[n'_t] W_N^{kn'_t} \\ W_{avg}[k] &= \frac{K}{A} \sum_{i=0}^{J-1} w_{avg}[n_i] W_N^{n_i k} \end{aligned}$$

with $\{n_i\}$ being the most significant channel tap indexes and $\{n'_t\}$ being the indexes of the other channel taps.

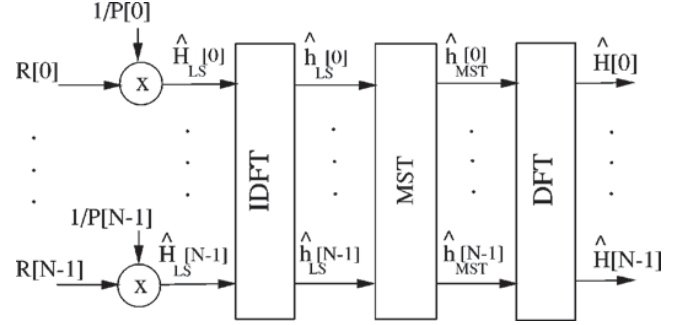


Fig. 2. MST's dual form DFT-based method.

By using suitable number of most significant taps, $H_{res}[k]$ can be kept very small (or zero) and hence the channel estimation error is mostly dominant (or totally caused) by the noise term $W_{avg}[k]$. For this case, the mse of the channel estimate can be approximated (or expressed) as

$$\begin{aligned} mse\{\hat{H}[k]\} &= E\{W_{avg}[k] W_{avg}^*[k]\} = \frac{K^2}{A^2} J \sigma_{t,avg}^2 \\ &= \frac{K J \sigma_t^2}{A^2} = \frac{K J}{N} mse\{\hat{H}_{LS}[k]\} \end{aligned} \quad (25)$$

Hence, the mse performance gain of proposed MST method over LS method (similarly over FPTA method) is ideally N/JK . The actual mse performance gain would be less than this amount due to the interference from data part and some (if any) excluded channel taps of nonsignificant energy.

The choice of the number of most significant taps J in the channel estimation depends on the application scenario. Broadcasting environment such as single frequency network can have larger number of multipaths with significant energy than non-broadcasting cases such as wireless LAN's environment. In any case, J should be chosen larger than the (designed) number of multipaths in order to prevent channel estimation error caused by missed channel taps. The channel estimation error caused by the noise from an additional tap in channel estimation is much less than that caused by missing one of the multipaths. A suitable choice for J may be two times or more of the (designed) number of multipaths (as will be seen in the simulation results) in order to ensure no channel energy missing.

Another MST tap selection approach can be implemented by selecting the channel taps whose energy is above a threshold. The threshold may be set as η times the maximum channel tap's energy in the raw channel estimate. The suitable choice of η depends on the operating SNR and more details will be discussed in the simulation results section.

VI. SIMILAR APPROACHES IN DFT-BASED METHODS

In this section, we relate the proposed MST method to the DFT-based approaches [15], [16], [10]. Essentially, due to the one-to-one relationship of DFT and IDFT, the MST method can be related to DFT-based approach as shown in Fig. 2. Consider the system with training symbol (i.e., pilot tones on all subcarriers). First, LS estimates are obtained and then input to IDFT block resulting in N samples LS estimate of channel impulse response. The largest amplitude J channel taps among the N samples (taps) are chosen as J most significant channel taps and

the other taps are set to zero. The resulting MST channel impulse response estimate is input to FFT block to get the MST channel frequency response estimate. This MST's dual form DFT-based approach has been applied in [10]. A similar concept by using singular value decomposition can be found in [18].

The difference between proposed MST approach and its dual form DFT-based approach is that MST uses pilot-data multiplexed approach while its dual form DFT based approaches use training symbol approach. It can be shown that the potential gain of the latter approach is N/J . Using total pilot power of NA^2 , the latter approach achieves potential gain of N/J whereas MST achieves potential gain of $N/(JK)$ with total pilot power of NA^2/K ; hence, on the basis of the same total pilot power, both methods have the same potential gain for channel estimation. Another difference is the complexity. In MST approach, operations involved are time-averaging, most significant channel taps selection and one FFT operation whereas its dual form DFT-based approach requires LS estimation, one IFFT operation, most significant channel taps selection and one FFT operation. Hence, the proposed MST approach saves some complexity.

One approach which is similar to MST's dual form DFT-based approach is the method of [16]. The difference is that [16] uses the first N_{cp} channel taps where N_{cp} is the number of cyclic prefix samples whereas MST's dual form DFT-based approach uses only J most significant channel taps, ($J < N_{cp}$). Since multipath channels usually have much less channel taps than N_{cp} , [16] has more noise perturbation than MST.

Other DFT-based approaches use linear minimum mean square error (LMMSE) estimators or approximate LMMSE estimators with reduced complexity [15]. If complexity can be afforded, these LMMSE or approximate LMMSE estimators can be implemented in MST approach. Let us consider a wide-sense stationary uncorrelated scattering (WSSUS) multipath channel with power delay profile given by $\sigma_{h_i}^2$ at delays of i OFDM sample intervals. Due to the uncorrelated multipaths, the correlation matrix of the channel impulse response becomes a diagonal matrix with the diagonal elements given by the power delay profile. Then the LMMSE's matrix multiplication (or scalar multiplication on corresponding channel taps) is performed as follows [19], [15]:

$$\hat{h}_{mmse} = \text{diag} \left\{ \frac{\sigma_{h_0}^2}{\sigma_{h_0}^2 + \beta/SNR}, \frac{\sigma_{h_1}^2}{\sigma_{h_1}^2 + \beta/SNR}, \dots, \frac{\sigma_{h_{N-1}}^2}{\sigma_{h_{N-1}}^2 + \beta/SNR} \right\} \hat{h}_{LS} \quad (26)$$

where $\text{diag}\{a_0, a_1, \dots, a_{N-1}\}$ is a diagonal matrix with diagonal elements $[a_0, a_1, \dots, a_{N-1}]$. In DFT-based LMMSE approach, the first symbol is composed of all pilot tones of equal amplitude and the following data symbols are composed of M-ary QAM symbols on all subcarriers. Then for the initial channel estimation based on the training symbol, SNR/β in (26) is replaced by the pilot-to-noise power ratio. For the following channel estimation operated in the decision directed mode, β in (26) is given by $\beta = E[|X|^2]E[|X|^{-2}]$ where X is M-ary QAM symbol. For 64-QAM, $\beta = 2.6854$. For MST-LMMSE, SNR/β is replaced by its pilot-to-noise power

ratio (which is $1/K$ times that of DFT-based LMMSE due to its pilot ratio of $1/K$).

There is a delicate difference between DFT-based LMMSE and MST-LMMSE. In DFT-based LMMSE, all N samples of $\hat{h}_{LS}$ are corrupted by *iid* AWGN noise samples of zero mean and variance σ_t^2/A^2 . In MST-LMMSE, the first M samples of $\hat{h}_{LS}$ are corrupted by *iid* AWGN noise samples of zero mean and variance $K\sigma_t^2/A^2$ and the rest are zeros. Hence, if compared for the training symbol alone, DFT-based LMMSE will have a better channel estimation performance due to the larger pilot power used in channel estimation. But for the following data symbols, DFT-based LMMSE has to use decision directed approach which may degrade the channel estimation performance and BER performance due to the decision directed errors. For MST-LMMSE, the channel estimation and BER performances remain the same for all symbols.

In LMMSE implementation, since multipath channel correlation and SNR are usually unknown at the receiver, some fixed values have to be used for them. In [18], it is suggested to use high dummy SNR value and uniform multipath channel correlation which are robust to the channel correlation mismatch. If high dummy SNR value is used and uniform multipath channel correlation is assumed over guard interval, then LMMSE approach would be similar to [16]'s method. From (26) with high dummy SNR value, the LMMSE approach can be viewed as setting the channel taps with no energy to zeros and bypassing the other taps. Since most practical multipath channels have only a few significant paths (if compared to N), a suitable way of implementing an approximate LMMSE is choosing a predefined number of most significant taps and setting the others to zero which is the underlying idea of MST method. In complexity aspect, MST-LMMSE estimator still saves one N -point IFFT operation if compared to DFT-based LMMSE estimators.

VII. CHANNEL ESTIMATION PERFORMANCE EVALUATION BY SIMULATION

The proposed MST channel estimation is evaluated by computer simulation for two multipath fading channel models, namely Channel-A and Channel-B. The Channel-A is the ATTC (Advanced Television Technology Center) and the Grand Alliance DTV laboratory's ensemble E model whose channel impulse response for the static case is given by

$$\begin{aligned} h[n] = & \delta[n] + 0.3162 \delta[n-2] + 0.1995 \delta[n-17] \\ & + 0.1296 \delta[n-36] + 0.1 \delta[n-75] \\ & + 0.1 \delta[n-137] \end{aligned} \quad (27)$$

where unit delay is assumed to be the same as OFDM sample period. The Channel-B is a simplified version of DVB-T channel model P_1 [20] and its channel impulse response for the static case is given in Table I. In the simulation, the given channel tap gains of both channels represent the standard deviations of the gains of the complex Gaussian random variables.

The OFDM system parameters are as follows: the number of subcarriers $N = 8192$, pilot ratio $1/K = 1/8$, guard interval ratio $= 1/8$ and carrier modulation 64-QAM. Perfect synchronization is assumed in order to observe the channel estimation performance alone. The considered methods are

TABLE I
CHANNEL IMPULSE RESPONSE FOR CHANNEL-B

Delay (OFDM samples)	Gain	Phase (radians)
0	0.2478	-2.5694
1	0.1287	-2.1208
3	0.3088	0.3548
4	0.4252	0.4187
5	0.4900	2.7201
7	0.0365	-1.4375
8	0.1197	1.1302
12	0.1948	-0.8092
17	0.4187	-0.1545
24	0.3170	-2.2159
29	0.2055	2.8372
49	0.1846	2.8641

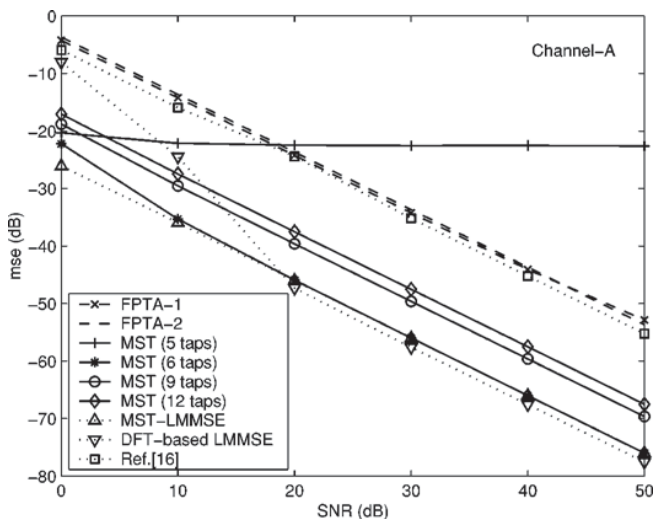


Fig. 3. Channel estimation mean square error (*mse*) in Channel-A.

MST, FPTA-1, FPTA-2, MST-LMMSE, DFT-based LMMSE and [16]. In the last two methods, one OFDM symbol is used for training symbol (pilot tones on all subcarriers) in every K OFDM symbols (i.e., the same pilot ratio $1/K$). In both LMMSE methods, ideal channel correlation and SNR values are used in order to evaluate the relative performance of MST method. In all methods, the same pilot tone symbol of $7 + j7$ signal point in 64-QAM constellation is used. For SNR of 40 and 50 dB, 10^5 independent simulation trials are used and for the other SNR values, 10^4 independent trials are used. In our simulation, floating point computation is used. However, it should be noted that due to the limited quantization levels in real hardware implementation, the computation error in real situation may degrade the performance obtained in our simulation.

Fig. 3 shows the *mse* performances of different channel estimation methods in Channel-A. The required interpolation in FPTA-1 is performed using Matlab's default *interp* function. Due to some noise reduction of this interpolation, the *mse* of FPTA-1 is slightly less than FPTA-2. The method of [16] has

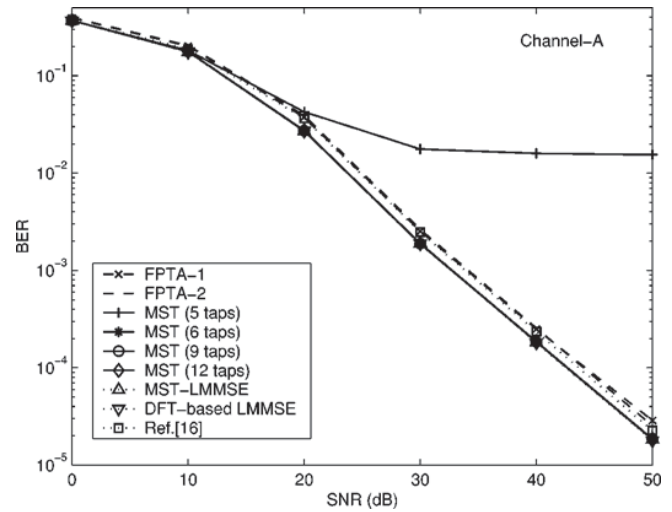


Fig. 4. BER performances of OFDM system with different channel estimation methods in Channel-A.

some improvement over FPTA approaches even though it is associated with decision directed errors. MST method with 6 taps, which is the same as the number of taps in Channel-A, has approximately 18 to 22 dB *mse* gain over FPTA approaches. MST with 5 taps shows an irreducible channel estimation error floor caused by missing some of the channel energy. It has a better performance than FPTA approaches for SNR less than 20 dB since for this SNR region the gain in noise suppression is greater than the loss of channel energy missing. However, for higher SNR region where noise has smaller impact than the channel energy missing, the channel estimation error floor results in a worse performance for MST with 5 taps. For the cases of MST with larger number of taps than the actual channel taps, MST achieves performance gain over FPTA and [16] approaches. The channel estimation performance gains are not as high as MST with 6 taps case due to the additional noise perturbation from the extra taps.

Also shown in Fig. 3 are the performances of MST-LMMSE and DFT-based LMMSE methods. LMMSE's approaches are of similar performance to MST with 6 taps case. For SNR less than 20 dB, DFT-based LMMSE has a worse performance than MST with 6 taps due to the decision directed errors. For higher SNR region, it has a slightly better performance due partly to the fact that its decision directed errors become insignificant and partly to the fact that MST suffers some interference from data part. MST-LMMSE has a slight performance gain over MST with 6 taps case for low SNR region and almost the same performance for high SNR region. This slight performance gain of MST-LMMSE in low SNR region is due to the utilization of (ideal) channel and noise statistics in MST-LMMSE approach by which the selection of channel taps are always correct, whereas in MST approach, the selection of channel taps may not be always correct due to the large noise perturbation.

The actual *mse* gain of MST method is slightly less than the potential gain of MST method described in previous section since data signal interference to training signal is neglected in the derivation of the potential gain. The actual *mse* gain is less in low SNR region than in high SNR region. The reason is

TABLE II
BER FOR DIFFERENT CHANNEL ESTIMATION METHODS IN CHANNEL-A

SNR (dB)	0	10	20	30	40	50
FPTA-1	0.3926	0.2015	3.7501×10^{-2}	2.5364×10^{-3}	2.4960×10^{-4}	2.8721×10^{-5}
FPTA-2	0.3957	0.2049	3.9122×10^{-2}	2.6389×10^{-3}	2.5513×10^{-4}	2.5377×10^{-5}
MST (5 taps)	0.3693	0.1814	4.262×10^{-2}	1.7651×10^{-2}	1.5964×10^{-2}	1.5441×10^{-2}
MST (6 taps)	0.3689	0.1770	2.7049×10^{-2}	1.8693×10^{-3}	1.8358×10^{-4}	1.8309×10^{-5}
MST (9 taps)	0.3695	0.1776	2.7291×10^{-2}	1.8841×10^{-3}	1.8505×10^{-4}	1.8502×10^{-5}
MST (12 taps)	0.3700	0.1781	2.7487×10^{-2}	1.8958×10^{-3}	1.8598×10^{-4}	1.8574×10^{-5}
MST-LMMSE	0.3689	0.1753	2.7023×10^{-2}	1.9081×10^{-3}	1.8309×10^{-4}	1.8299×10^{-5}
DFT-based LMMSE	0.3670	0.1762	2.7051×10^{-2}	1.8911×10^{-3}	1.8052×10^{-4}	1.7851×10^{-5}
Ref. [16]	0.3711	0.1887	3.6529×10^{-2}	2.4789×10^{-3}	2.3445×10^{-4}	2.2593×10^{-5}

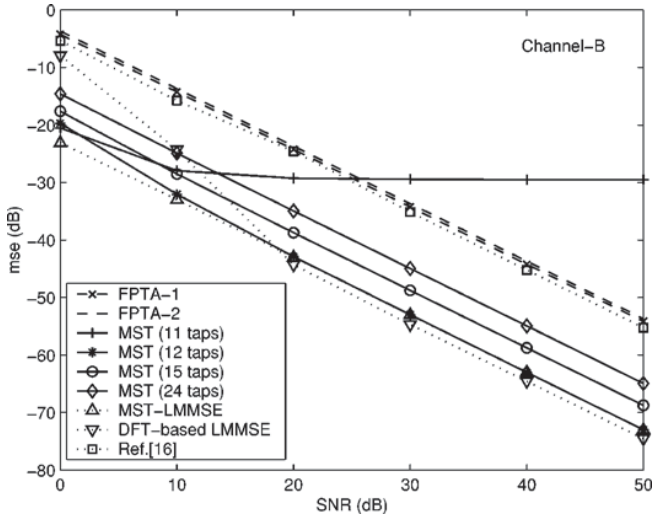


Fig. 5. Channel estimation mean square error (*mse*) in Channel-B.

that most significant channel taps selection would more likely choose some taps with no channel energy at low SNR region than at high SNR region, resulting in less *mse* gain at low SNR region.

In Fig. 4, the BER performances in Channel-A are presented for different channel estimation methods. Due to the channel estimation error floor and the BER sensitivity of 64-QAM to channel estimation error, MST with 5 taps case shows a BER floor while the others do not. The method of [16] is slightly better than FPTA approaches as is the case in *mse* performance. The other MST and LMMSE approaches have better BER performances than FPTA and [16] approaches, but among themselves, the BER performances are just slightly different and hence their BER curves are almost the same. For a better presentation of their slight differences, their BER values are tabulated in Table II. These slight differences can be ascribed to the corresponding different channel estimation performances. It is also noted that when the channel estimation *mse* is much smaller than $1/\text{SNR}$, then some deviation in channel estimation *mse* does not significantly affect the BER performance since noise is the dominant contributor to BER in this case.

The channel estimation performance and BER performance of the considered methods in Channel-B are plotted in Figs. 5 and 6 respectively and the BER values are given in Table III for the sake of clarity. The performances are of the same trend as in

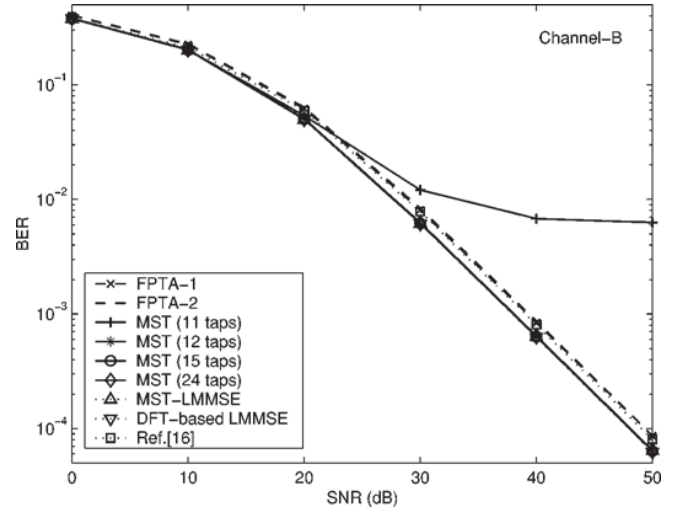


Fig. 6. BER performances of OFDM system with different channel estimation methods in Channel-B.

Channel-A. It is also noted that the channel estimation and BER performances in Channel-B is worse than Channel-A except for the case of missing some of the channel energy (i.e., MST with 5 taps in Channel-A and MST with 11 taps in Channel-B). This exception is due to the larger percentage of the channel energy missed in Channel-A than in Channel-B.

From the simulation results, it is clear that the number of MST taps J should not be smaller than the number of actual channel taps. On the other hand, using more taps may slightly degrade the BER performance due to the more noise perturbation. However, even up to double of the number of actual channel taps, MST has almost the same BER performance as using the number of actual channel taps. Hence, a suitable choice for J might be double of the (designed) number of channel taps for the considered channel environment.

Another way of selecting MST taps by threshold decision is also evaluated in Channel-B and the channel estimation *mse* and BER results are shown in Figs. 7 and 8 respectively. The performance of MST with 12 taps is also included for reference. It is observed that the suitable choice of threshold also depends on the operating SNR, as expected. The results also indicate that almost the same performance as MST with 12 taps case can be obtained if the threshold for an operating SNR is set within approximately 20 to 23 dB below the operating $1/\text{SNR}$. This low

TABLE III
BER FOR DIFFERENT CHANNEL ESTIMATION METHODS IN CHANNEL-B

SNR (dB)	0	10	20	30	40	50
FPTA-1	0.4014	0.2250	6.1986×10^{-2}	8.0725×10^{-3}	8.3867×10^{-4}	8.6296×10^{-5}
FPTA-2	0.4041	0.2283	6.3793×10^{-2}	8.3684×10^{-3}	8.6701×10^{-4}	8.6831×10^{-5}
MST (11 taps)	0.37828	0.2019	5.3833×10^{-2}	1.2092×10^{-2}	6.7836×10^{-3}	6.2976×10^{-3}
MST (12 taps)	0.3785	0.2011	4.9783×10^{-2}	6.1269×10^{-3}	6.2924×10^{-4}	6.3026×10^{-5}
MST (15 taps)	0.3790	0.2017	5.0082×10^{-2}	6.1707×10^{-3}	6.3410×10^{-4}	6.3421×10^{-5}
MST (24 taps)	0.3805	0.2032	5.0749×10^{-2}	6.2714×10^{-3}	6.4442×10^{-4}	6.4503×10^{-5}
MST-LMMSE	0.3781	0.2011	4.9847×10^{-2}	6.1509×10^{-3}	6.3000×10^{-4}	6.3806×10^{-5}
DFT-based LMMSE	0.3775	0.2018	4.969×10^{-2}	6.1261×10^{-3}	6.2774×10^{-4}	6.2568×10^{-5}
Ref. [16]	0.3809	0.2123	5.9644×10^{-2}	7.7969×10^{-3}	8.0554×10^{-4}	8.0195×10^{-5}

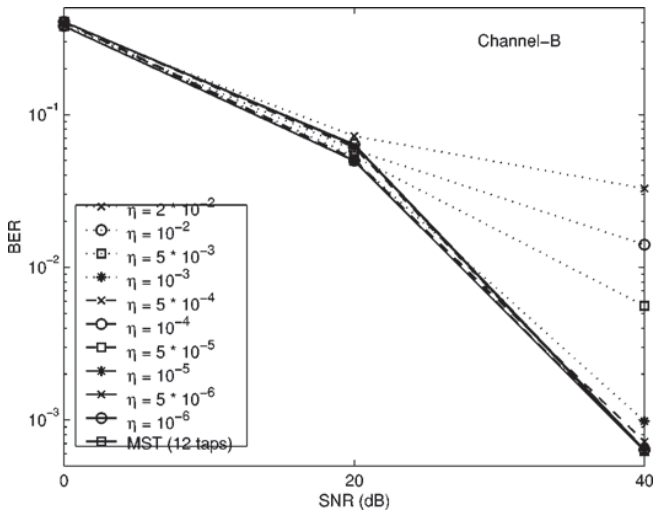


Fig. 7. Channel estimation mean square error (*mse*) for MST with threshold setting η .

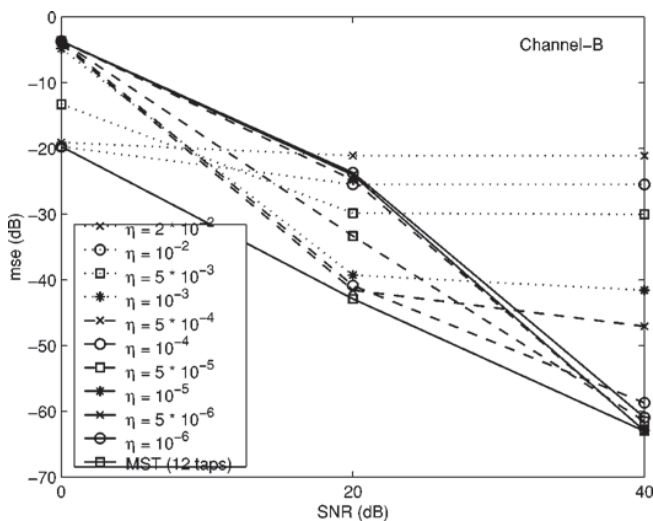


Fig. 8. BER performance of MST with threshold setting η .

value of the threshold in ensuring no channel energy miss is also due to the gain variations in the channel taps.

It is also remarked that for relatively slow fading environments, the decision feedback approaches such as [16] and

LMMSE can increase spectrum efficiency by using smaller pilot ratio. Similarly, MST approach can increase the spectrum efficiency by not inserting pilot tones in every OFDM symbol. For this case, the decision feedback approach as used in MST's dual form DFT-based approach can easily be applied for pure data symbols.

VIII. CONCLUSION

In this paper, intra-symbol time-domain averaging approaches for time-domain channel estimation of OFDM system are investigated. As an approximation to LMMSE type estimator, a time-domain approach, called MST, is proposed where intra-symbol time-domain averaging and most significant taps selection are applied. The similarities and differences of the proposed MST method versus DFT-based LMMSE methods are discussed. Two approaches for MST taps selection are described where the first uses a fixed number of taps while the second chooses all taps above a threshold. The simulation results suggest that a suitable choice for the fixed number of MST taps is using double of the (designed) number of channel taps for the considered channel environment. And a suitable choice of the threshold is within the range of approximately 20 to 23 dB below the operating $1/\text{SNR}$. In terms of BER performance in multipath fading channels, MST without channel energy missing and LMMSE approaches have almost the same performance which is better than the other considered approaches. In terms of complexity, MST approach keeps minimum complexity among the considered methods.

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- H. Minn** received his B.E.(Electronics) degree with honors from Yangon Institute of Technology, Myanmar, in 1995, and M.Eng. (Telecommunications) degree from Asian Institute of Technology (AIT), Thailand, in December 1997. He worked as a laboratory supervisor at Telecommunications Program, AIT, for one year. Since 1999, he has been a research assistant in the Department of Electrical and Computer Engineering at the University of Victoria, Canada. Currently, he is a Ph.D. student at University of Victoria. His research interests include wireless communications, OFDM, synchronization, channel estimation, equalization, modulation and error control coding.
- Vijay K. Bhargava** received the B.Sc., M.Sc., and Ph.D. degrees from Queen's University, Kingston, Canada in 1970, 1972 and 1974 respectively. Currently, he is a Professor of Electrical and Computer Engineering at the University of Victoria. He is a co-author of the book *Digital Communications by Satellite* (New York: Wiley, 1981) and co-editor of the IEEE Press Book *Reed-Solomon Codes and Their Applications*. He is an Editor-in-Chief of WIRELESS PERSONAL COMMUNICATION, a Kluwer Periodical. His research interests are in multi-media wireless communications.
- Dr. Bhargava is very active in the IEEE and is currently the President of the IEEE Information Theory Society. He was Co-chair for IEEE ISIT'95, Technical Program Chair for IEEE ICC'99 and is the Chair of IEEE VTC'2002 Fall.
- Dr. Bhargava is a Fellow of the B.C. Advanced Systems Institute, Engineering Institute of Canada (EIC), IEEE and the Royal Society of Canada. He is a recipient of the IEEE Centennial Medal (1984), IEEE Canada's McNaughton Gold Medal (1995), the IEEE Haraden Pratt Award (1999) and the IEEE Third Millennium Medal (2000).

Hsieh-Yee Declaration

Exhibit 0003

Channel Estimation Using Pilot Tones in OFDM Systems

Che-Shen Yeh and Yinyi Lin
Department of Electrical Engineering
National Central University, Chung Li
Taiwan 32054, R. O. C.

Abstract –

Channel estimation techniques using pilot tones in OFDM systems are investigated. Two time-domain channel estimation methods are proposed. Based on theoretical analysis and computer simulation, for a channel with multipath distortion and AWGN, the proposed time-domain channel estimation schemes perform better than the conventional frequency domain channel estimation techniques.

I. Introduction

Since terrestrial broadcast channels are often affected by multipath distortion, in a single carrier modulation system, an adaptive equalizer is usually employed to combat channel variations. However, in the Orthogonal Frequency Division Multiplexing (OFDM) modulation system, the channel estimation is accomplished using pilot tones inserted in OFDM symbols. In an OFDM system, the incoming signal is transmitted using N equally spaced carriers with a carrier spacing of $1/(NT)$ Hz, where T is the input data bit period. Multipath causes the channel transfer function to have variation in both spectral magnitude and spectral phase linearity. The number of carriers is chosen such that a typical channel transfer function is essentially constant over the bandwidth of each individual carrier. The effect of multipath distortion is then reduced to amplitude attenuation and phase rotation of data in each carrier, and the channel equalization is reduced to a complex multiplication bank in front of the decision circuit at the receiver [1-2].

An OFDM system can eliminate Intersymbol Interference (ISI) between OFDM symbols by inserting a cyclically extended guard interval, where each OFDM symbol is preceded by a periodic extension of data signal itself [3, 10]. The total symbol duration becomes $T_{total} = T_g + T_s$, where T_g is the guard interval and T_s is the useful symbol duration. When the guard interval T_g is longer than the channel impulse response $h[n]$, or the multipath delay spread, the ISI between OFDM symbols can be entirely eliminated. The ratio of the guard interval to useful symbol duration is dependent on application. A longer guard interval will reduce the system data throughput, and a commonly adopted compromise length is

$$T_g \leq \frac{1}{4} T_s. \text{ A cyclic prefix in the guard interval can be used to}$$

maintain the receiver carrier synchronization during a long silence. In addition, cyclic convolution can be used to model the transmission system.

However, the in-band fading will still exist. Therefore, let $H_k = \lambda_k e^{j\theta_k}$ be the response within the bandwidth of the k^{th} carrier. Equalization of the channel in the receiver requires that the FFT output of the k^{th} carrier be multiplied by a complex coefficient $c_k = \frac{1}{H_k}$. This operation will scale both

signal and noise/interference. In the OFDM system, the estimation of the channel response uses in-band pilot tones inserted into each symbol with a pilot ratio $PR = 1/R$. In this scheme, the channel response can be estimated rapidly and the tuner phase noise can be significantly reduced [10-11]. The optimum choice for the numbers of pilots ($=N/R$) for the AWGN channel is described in [4]. This scheme is referred to as scheme A, as depicted in Fig. 1a. In the AWGN channel every optimally spaced set of pilots can provide good channel estimation, so a fixed set of optimally spaced pilot tones is used in scheme A.

An alternative is to insert pilot tones into all of the carriers in an OFDM symbol with a period of R , as depicted in Fig. 1b. The channel response is then estimated using the pilot symbol, and the estimated response is used for subsequent OFDM symbols until the next pilot symbol is received. The channel response can also be estimated by averaging some consecutive pilot symbols. This scheme is referred to as scheme B. As will be shown, it is not suitable for fading channels.

II. Proposed Channel Estimation Techniques

A typical OFDM system is depicted in Fig. 2. On the transmitter site, a multiplexer (MUX) combines the data symbols and pilot signals as a data stream in the frequency domain, and the data stream is fed into an Inverse Fast Fourier Transform (IFFT) block. The IFFT block converts the data stream from a serial format into a parallel format, and transforms the signal from frequency domain to time domain. The guard interval is then inserted after each OFDM symbol to eliminate ISI between OFDM symbols.

Pilot tones are modulated with a PN sequence in the frequency domain and are inserted into each OFDM symbol with a ratio $PR = 1/R$ using scheme A. The channel response is estimated in the frequency domain using FFT processing and comparing received pilots with the locally stored reference pilots. This is referred to as the Frequency Pilot Frequency Interpolation (FPFI) technique since the channel response is obtained via frequency interpolation. The pilot tones can also be inserted into the OFDM symbols using scheme B. In this scheme, all pilot tones are inserted in the first OFDM symbol (pilot symbol) of a data frame. The channel response is obtained by averaging several consecutive pilot symbols and comparing the average with a locally stored reference symbol [5]. This technique is referred to as the Frequency Pilot Frequency Average (FPFA) technique.

In this paper, channel estimation conducted in the time domain is proposed. In one of the two proposed techniques, instead of the PN sequence the positive and negative polarization pilot signals are alternatively inserted into OFDM symbols in the frequency domain using FPFI. An OFDM symbol can be regarded as several time slots, with each time slot having a pilot. Assuming that the impulse response of each time slot is identical, one can obtain the channel response in the time domain by averaging these time slots. This technique is referred to as the Frequency Pilot Time Average (FPTA) technique, as depicted in Fig. 3a.

In the other technique a PN sequence is periodically inserted into the OFDM symbol in the time domain, and the channel response can be obtained via correlation in the time domain. This technique is referred to as the Time Pilot Time Correlation (TPTC) technique, as depicted in Fig. 3b. In this technique the channel response is estimated using the correlation of received pilot signal and the locally stored PN sequence [8]. If the transmitted pilot signal $p[n]$ is a PN sequence, the received pilot signal $\tilde{p}[n]$ can be expressed as $\tilde{p}[n] = p[n] \otimes h[n]$ in the time domain (where $\otimes$ represent the convolution operator) or $\tilde{P}[k] = P[k] \cdot H[k]$ in the frequency domain. Thus the channel response is given by $H[k] = \tilde{P}[k] / P[k] = \tilde{P}[k] \cdot P[k] / |P[k]|^2$. As the spectrum of a PN sequence is nearly constant at the frequency occupied by the received pilot signal, $|P[k]|^2$ can be approximated as a constant, K , which is equal to the length of the PN sequence. Thus the channel response can be estimated using

$$H[k] = \frac{\tilde{P}[k] \cdot P[k]}{K} \text{ in the frequency domain}$$

or

$$h[n] = \frac{1}{K} \sum_i \tilde{p}[n] \cdot p[n+i] \text{ in the time domain.}$$

III. Pilot to Noise (P/N) Ratio

This section is an analysis of the pilot to noise ratio for both time domain channel estimation and frequency domain channel estimation. Referencing to Fig 2 and Fig 3a, the pilot signal is $P[k]$. Suppose pilot ratio $PR = 1/R$, $P[k]$ can be shown as:

$$P[k] = \begin{cases} A, & k = Rm \\ 0, & k = Rm + \ell \end{cases} \quad (1)$$

Where R , m and ℓ are integers, $1 \leq \ell \leq R-1$, $0 \leq m \leq (N/R)-1$, A is the pilot amplitude and N/R is an integer.

$$p[n] = \text{IFFT}\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] e^{j \frac{2\pi nk}{N}} \quad (2)$$

Where N is the IFFT transformation length

The multipath channel impulse response $h[n]$ is:

$$h[n] = \delta[n] + \sum_{i=1}^M \alpha_i \cdot \delta[n - \tau_i] = \sum_{i=0}^M \alpha_i \cdot \delta[n - \tau_i] \quad (3)$$

Where M is the total number of echoes

α_i represents the amplitude of i^{th} path echo

τ_i represents the delay of i^{th} path echo

$\alpha_0 = 1$, $\tau_0 = 0$, (representing the principle pilot signal)

$p[n]$ after multipath channel $h(n)$ are:

$$\tilde{p}[n] = p[n] * h[n] = p[n] + \sum_{i=1}^M \alpha_i p[n - \tau_i] \quad (4)$$

On the Rx side, the received frequency domain pilot signal $\tilde{P}[k]$ and multipath channel response $H[k]$ will be:

$$\tilde{P}[k] = P[k] \cdot H[k] = \begin{cases} \sum_{i=0}^M \alpha_i P[k] e^{j \frac{2\pi \tau_i k}{N}}, & k = mR \\ 0, & k \neq mR \end{cases} \quad (5)$$

$$H[k] = 1 + \sum_{i=1}^M \alpha_i e^{-j \frac{2\pi \tau_i k}{N}} \quad (6)$$

Assuming the time domain channel noise $n[n]$ is wide sense stationary AWGN with zero mean and variance:

$$\sigma_{tn}^2 = \text{Var}\{n[n]\} = E\{n[n] - E\{n[n]\}\} \quad (7)$$

Where $n[n]$ represents the time domain white noise at the n^{th} sample, $E\{n[n]\}$ represents the expected value of $n[n]$.

In radio channel, suppose the noise is zero mean AWGN,

$E\{n[n]\} = 0$. On the Rx side, after the FFT, the $n[n]$ signal will become the frequency domain noise

$$N[k] = \sum_{n=0}^{N-1} n[n] W_N^{kn}, W_N = e^{-j\frac{2\pi}{N}} \quad (8)$$

Assuming the variance of frequency domain noise $N[k]$ is σ_{FN}^2 , according to variance definition:

$$E\{N[k]\} = \sum_{n=0}^{N-1} E\{n[n]\} W_N^{kn} = 0$$

$$\begin{aligned} \sigma_{FN}^2 &= \text{Var}\{N[k]\} \\ &= E\{(N[k] - E\{N[k]\}) \cdot (N[k] - E\{N[k]\})^*\} \\ &= E\{N[k] \cdot N^*[k]\} \\ &= E\left\{\left(\sum_{n=0}^{N-1} n[n] \cdot W_N^{kn}\right) \cdot \left(\sum_{m=0}^{N-1} n[m] \cdot W_N^{km}\right)^*\right\} \\ &= \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} E\{n[n] \cdot n^*[m]\} W_N^{k(n-m)} \end{aligned} \quad (9)$$

Because $n[n]$ and $n[m]$ are independent and zero mean, Eq.(9) can be re-written as:

$$\begin{aligned} \sigma_{FN}^2 &= \sum_{n=0}^{N-1} \left\{ E\{n[n] \cdot n[n]\} W_N^{0} + \sum_{\substack{m=0 \\ m \neq n}}^{N-1} E\{n[n]\} \cdot E\{n^*[m]\} W_N^{k(n-m)} \right\} \\ &= \sum_{n=0}^{N-1} (\sigma_{tn}^2 + 0) = N \cdot \sigma_{tn}^2 \end{aligned} \quad (10)$$

From the previous mathematical deductions, in the radio channel of frequency domain, the noise variance is being enhanced N times while the noise amplitude is amplified $\sqrt{N}$ times.

In the FPTA, pilot ratio $PR = 1/R$ and the pilot signal $P[k]$ is shown as:

$$\begin{aligned} P[k] &= \begin{cases} (-1)^m \cdot A, & k = Rm \\ 0, & k = Rm + \ell \end{cases} \\ &= \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \delta[k - Rm] \end{aligned} \quad (11)$$

Where $m = 0, 1, 2, \dots, (N/R)-1$, ℓ is an integer and $1 \leq \ell \leq R-1$. The time domain TX pilot signal $p[n]$ is shown as:

$$\begin{aligned} p[n] &= \text{IFFT}\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] \cdot e^{+j\frac{2\pi kn}{N}} \\ &= \frac{1}{N} \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot e^{+j\frac{2\pi Rmn}{N}} \\ &= \frac{A}{R} \cdot \frac{1}{N/R} \sum_{m=0}^{(N/R)-1} e^{+j2\pi m \left(\frac{n - (N/2R)}{N/R}\right)} \\ &= \frac{A}{R} \sum_{r=0}^{R-1} \delta\left[n - \frac{Nr}{R} - \frac{N}{2R}\right] \\ &= \begin{cases} A/R, & n = Nr/R + N/2R \\ 0, & \text{otherwise} \end{cases} \end{aligned} \quad (12)$$

From the result of Eq.(12), when n is equal to $N/2R$ plus the integer time of N/R , the time domain pilot will be added in-phase. The Rx pilot signal $\tilde{p}[n]$ in the time domain and $\tilde{P}[k]$ in the frequency domain will be:

$$\begin{aligned} \tilde{p}[n] &= p[n] \otimes h[n] \\ &= \frac{A}{R} \sum_{u=0}^{N-1} \left\{ \sum_{i=0}^{R-1} \delta\left[u - \frac{Nr}{R} - \frac{N}{2R}\right] \cdot \left(\sum_{i=0}^M \alpha_i \delta[n - u - \tau_i] \right) \right\} \\ &= \frac{A}{R} \sum_{r=0}^{R-1} \sum_{i=0}^M \alpha_i \cdot \delta\left[n - \left(\tau_i + \frac{Nr}{R} + \frac{N}{2R}\right)\right] \end{aligned} \quad (13)$$

$$\begin{aligned} \tilde{P}[k] &= P[k] \cdot H[k] \\ &= \begin{cases} \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \delta[k - Rm] \cdot \left(1 + \sum_{i=1}^M \alpha_i \cdot e^{-j\frac{2\pi \tau_i k}{N}} \right) \\ \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \left[\sum_{i=1}^M \alpha_i \cdot e^{-j\frac{2\pi \tau_i m}{N/R}} \right], & k = Rm \\ 0, & k \neq Rm \end{cases} \end{aligned} \quad (14)$$

where $\alpha_0=1$, $\tau_0=0$, (it is the subscript 0 that represents the principle pilot component).

In the FPFI method, $P[k]$ = PN-Sequence, $p[n]$ and $P[k]$ are both noise-like and randomly distributed in the time domain OFDM symbol. Hence, after FFT on the Rx end, the result is similar to AWGN case described above, and $\tilde{P}[k]$ and $\tilde{p}[n]$ both have $\sqrt{N}$ gain. Therefore, in the second (FPFI) method the P/N in the time domain and in the frequency domain are about the same, so using pilot signal for channel model estimation in the time domain is as good as in the frequency domain.

If the FPFI method is used for channel estimation with the pilot signal in the frequency domain, the pilot to noise ratio is $\tilde{P}[k]/\sigma_{FN}$. If the FPTA method is used for channel estimation with the pilot signal in the time domain, the pilot to noise ratio

is $\tilde{p}[n]/\sigma_{in}$, which implies that $\tilde{P}[k]$ and $\tilde{p}[n]$ are affected by the echo. The amplitude of $\tilde{p}[n]$ has different variation, which depends upon both the amplitude of echo and delay spread. Different multipath models result in different variation of $\tilde{P}[k]$. Hence, it is difficult to characterize generally the variation. This variation may differ with different channel models, hence, it is difficult to describe or derive with such unpredictable value compared with the principle pilot signal to noise variance ratio. However, the third method repeats an average of R times, after which it reduces σ_{in} to $\sigma'_{in} = \sigma_{in}/\sqrt{R}$. Assuming that the pilot interference of data to pilot tone can be ignored, we may compare the principle pilot to noise of these two methods:

$$\frac{p[n]/\sigma'_{in}}{P[k]/\delta_{FN}} = \frac{A/R \cdot (\sigma_{in}/\sqrt{R})}{A \cdot (\sqrt{N} \cdot \sigma_{in})} = \sqrt{\frac{N}{R}} \quad (15)$$

The FPTA estimation performance is better than the FPFA and FPFI. This improvement will be verified in the next section. In the derivation of Eq. (15), the FPTA does not take the "data ghosts" into account, thus, the gain of the time domain $\tilde{p}[n]/\delta_{in}$ over the frequency domain $\tilde{P}[k]/\delta_{FN}$ is actually lower than the value of $\sqrt{N/R}$ indicated in Eq.(15).

The interference of training sequence coming from data is referenced to as "data ghosts". A specific comparison of channel estimation will be given in the next section.

IV. Performance Comparisons and Discussions

The performance comparison of previously mentioned estimation techniques is made on several channel models recommended by the Advanced Television Technology Center (ATTC) in the U. S. A. and by the Digital Audio Broadcasting (DAB) project in Europe [6-7]. During simulations, assume that both transmitter and receiver are perfectly synchronized and the guard interval is long enough so that ISI between OFDM symbols can be ignored. The channel is assumed to be the additive white Gaussian noise (AWGN) channel.

Static Channel

Two channel models: the multipath Ensembles A and E of the ATTC, as recommended by the US ATTC and the Grand Alliance DTV laboratory [6][7], are used in simulations: Ensemble A as channel model 1:

$$h_1[n] = 0.1\delta[n+16] + \delta[n] + 0.1\delta[n-2] + 0.3162\delta[n-16] + 0.1995\delta[n-52] + 0.1295\delta[n-168]$$

Ensemble E as channel model 2:

$$h_2[n] = \delta[n] + 0.3162\delta[n-2] + 0.1995\delta[n-17] + 0.1296\delta[n-36] + 0.1\delta[n-75] + 0.1\delta[n-137]$$

In the above two formulas, the unit delay length $T = 1.1 \times 10^{-7}$ seconds. In addition, the simplified non-hilly urban channel model recommended by DAB test (pr EN50248) is also used in simulations as channel model 3:

$$h_3[n] = 0.707\delta[n+2] + \delta[n] + 0.707\delta[n-3] + 0.5\delta[n-13] + 0.4\delta[n-19] + 0.31625\delta[n-44]$$

Parameters	Specifications
Number of carriers (N)	8192
Pilot ratio (PR=1/R)	1/8, 1/32
Guard interval	1/8, 1/32
Signal constellation	64QAM
Channel model	h1(n), h2(n), h3(n)
S/N	0 ~ 30 dB

Table 1. OFDM system performance simulation parameters

The OFDM system parameters used in simulations are summarized in Table.1. All data carriers in the OFDM are 64-QAM modulated signals (signal constellation (x,y) with $x, y \in \{\pm 1, \pm 3, \pm 5, \pm 7\}$, and the average power of a data symbol is $P_s = 42$ [9]). The pilot signals with amplitude ± 7 are used.

The error performance is displayed in Figs 4 and 5 for various estimation techniques on the channel model 1 with pilot ratio 1/8 and 1/32 respectively. A similar result was found on the other two channel models and is not shown here. The results demonstrate that the proposed estimation techniques (i.e., FPTA and TPTC) outperform the conventional estimation techniques (i.e., FPFA and FPFI). A SNR improvement gain of 2-3 dB can be achieved using the time-domain estimation technique. This is because a higher P/N ratio can be obtained in the time domain. For pilot ratio $PR = 1/32$ on the television broadcasting channels (channel models 1 and 2), the error floor occurs using FPFI technique, as shown in Fig. 5. This is due to the aliasing effect introduced by the frequency interpolation. However, no error floor was found on the audio broadcasting channel (channel model 3). Since this channel has a shorter delay spread and the aliasing effect is not severe.

Extensive simulations using FPFI technique were conducted using different pilot ratios (i.e., 1/4, 1/8, 1/16, and 1/32, respectively). It was found that with pilot ratios $PR = 1/4$ and 1/8, no error floor occurs for all the three channel models. However with pilot ratio $PR = 1/16$, the error floor occurs at

8.2×10^{-4} and 1.35×10^{-4} for channels 1 and 2 respectively, although no error floor was found for channel 3. As the pilot ratio PR was reduced to 1/32, a more severe error floor occurs (0.335 and 0.0665 for the two DTV channel models), although there was still no error floor found on the DAB channel.

Channel Estimation Accuracy

In addition to symbol error rate performance, the mean square error between estimated response and real channel response was also compared to demonstrate the superiority of the time-domain estimation techniques. The system parameters used for comparison are displayed in Table 2 and the results are illustrated in Fig. 6. As shown, a smaller mean square error was found in the time-domain channel estimation techniques. The FPFA technique demonstrates a larger estimation error than the FPGI technique since in the FPFA the estimation error extends over several OFDM symbols, while in the FPGI the estimation error impacts only adjacent carriers within one OFDM symbol. The estimation error in the FPTA is larger than that in the TPTC because the FPTA is affected by the components of "data ghosts" which are non-zero, even after averaging several OFDM symbols. According to equation (15), the FPTA method outperforms FPGI method with a P/N gain about 30 dB. Figure (6) shows a P/N gain about 25.2 dB. The difference between 4.8 dB is due to "data ghosts".

Parameters	Specifications
Number of carriers (N)	8192
Pilot ratio (PR = 1/R)	1/8
Guard interval	1/8
Signal constellation	64QAM
Channel model	$h_1[n], h_2[n], h_3[n]$
S/N	10 dB, 20dB
Mean Average Number	16

Table 2 System parameters

Slow Fading Channel

The error performance of the estimation techniques is also compared on a mobile time-varying channel using the autoregressive (AR) model given as

$$h[k+1] = ah[k] + w[k]$$

Where a and $w[k]$ represent slow time variations and the noise vector sequence of the AWGN, respectively. It is assumed that the channel fading is slow enough so that the channel impulse response doesn't change within the duration of one OFDM symbol [4].

During simulations, parameters of the AR model are

assumed as $a = 0.99\pi$ and $w[k] = 0.0001\pi$, where π is the unit matrix. The OFDM parameters displayed in Table 1 and the three channel models previously mentioned are used. The error rate is shown in Fig. 7 with pilot ratio PR = 1/8. The results also reveal that both FPTA and TPTC provide better performance than those estimated in the frequency domain (FPFI and FPFA). Although both FPFA and FPFI techniques achieve almost the same error performance in static channels, the performance of the FPFA is slightly worse than that of the FPFI. This is because in the FPFA the channel response is estimated by averaging R pilot symbols (in this case, 8 pilot symbols were averaged), which cannot accurately estimate the response of each OFDM symbol. The effect of the pilot ratio on the error performance is also simulated, and the result with pilot ratio PR = 1/32 is demonstrated in Fig. 8 for channel 2. Again the error floor was found for the FPFI due to the aliasing effect. The error floor was also found in the FPFA, in which the channel was estimated by averaging 32 OFDM symbols. The error floor occurs when the accumulated estimation error due to channel variations becomes not negligible as compared to the estimation error caused by channel noise.

V. Summary

In this paper the channel estimation techniques (FPTA and TPTC) in the time domain using pilot tones for the OFDM system are presented. It is shown that a higher P/N ratio can be obtained in the time domain and that the proposed estimation techniques outperform the conventional estimation techniques (FPFI and FPFA) in which the channel response is estimated in the frequency domain. In addition, both FPFI and FPFA suffer from the aliasing effect and channel mis-estimation resulting in error floors, especially for low pilot ratios.

The proposed FPTA and TPTC techniques have advantages of low computational complexity and high P/N ratio. FPTA is suitable for fading channels with shorter transmission range, i.e., Local Multipoint Distribution Systems (LMDS), which has shorter channel response. FPFI and FPFA can be used to reduce overhead for systems with lower pilot ratios (i.e., 1/32, 1/64). However, in the FPTA the pilot signal consumes more RF power and interferes with data signal when the pilot ratio is high. The TPTC does not have this disadvantage.

VI. Acknowledgement

We are grateful to Dr. Yiyan Wu of CRC (Communication Research Centre of Industry Canada, Canada), who gave us advice and assistance in this paper over the past few months, as well as his long term counsel.

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Che-Shen Yeh received B.S. and M.S.E.E. in Electrical Engineering from Chung-Yuan Christian University, Jung-Li, Taiwan in 1980 and 1986 respectively, and is currently a Ph.D. candidate in Electrical Engineering at the National Central University, Jung-Li, Taiwan. His major research interests are digital transmission for television and audio. He leads the DTV/DAB transmission project at Telecommunications Labs of Chunghwa Telecom Co., Ltd. Taiwan, and is also responsible for DTV/DAB field trials and the development of DTV/DAB transmission system standards and regulations. He also is a member of the Task Force for SDTV/HDTV transmission standards and regulation at the Ministry of Transportation and Communications, Taiwan.

Yinyi Lin was born in Chuang Hua, Taiwan, R.O.C., in 1959. He received the B.S. degree in communication engineering from National Chiao Tung University, Hsin Chu, Taiwan in 1981, and M.S. and Ph.D. degrees in electrical engineering from the University of California, San Diego in 1986 and 1989, respectively. From 1983 to 1985, he has employed by the Electronics Research and Service Organizations (ERSO), I.T.R.I., Taiwan. During 1988, he spent six months at the IBM Almaden Research Center, San Jose, CA working in the area of signal processing for digital magnetic recording. From 1990 to 1991, he was with Archive Corporation, Costa Mesa, CA, where he engaged in the research of quarter inch cartridge (QIC) tape system. Since 1991, he has been with the Electrical Engineering Department of National Central University, Jung-Li, Taiwan, where he is now a Professor. His current research interests include coding and signal processing for digital magnetic recording systems, digital compression, and error correction for documents.

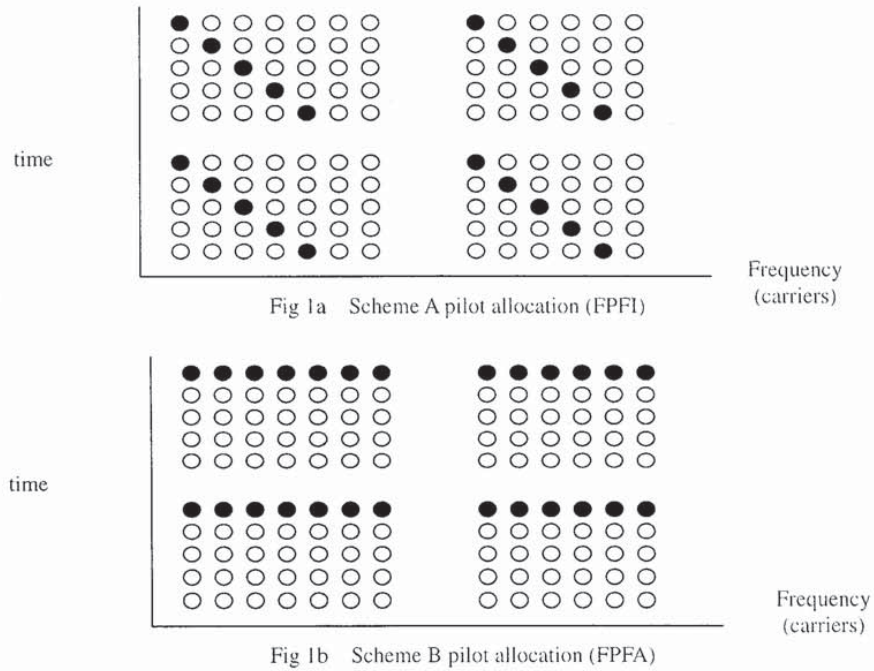
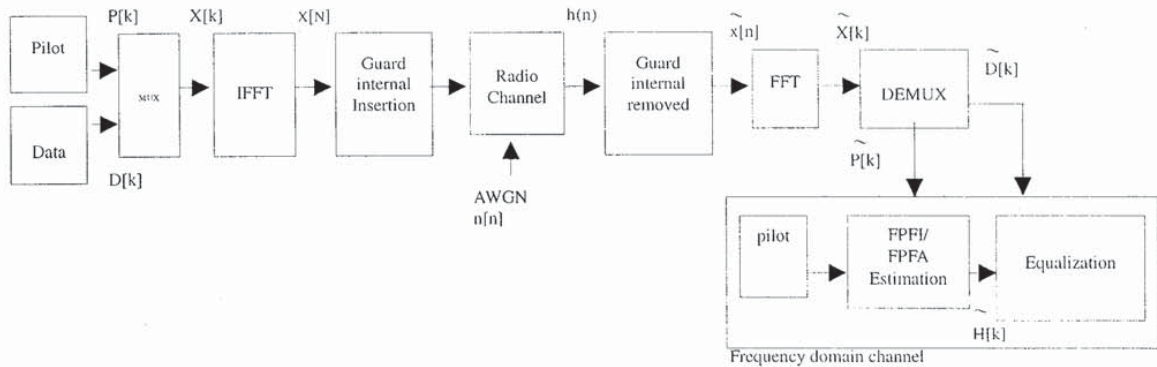


Figure 1 OFDM system frequency domain pilot signal allocation



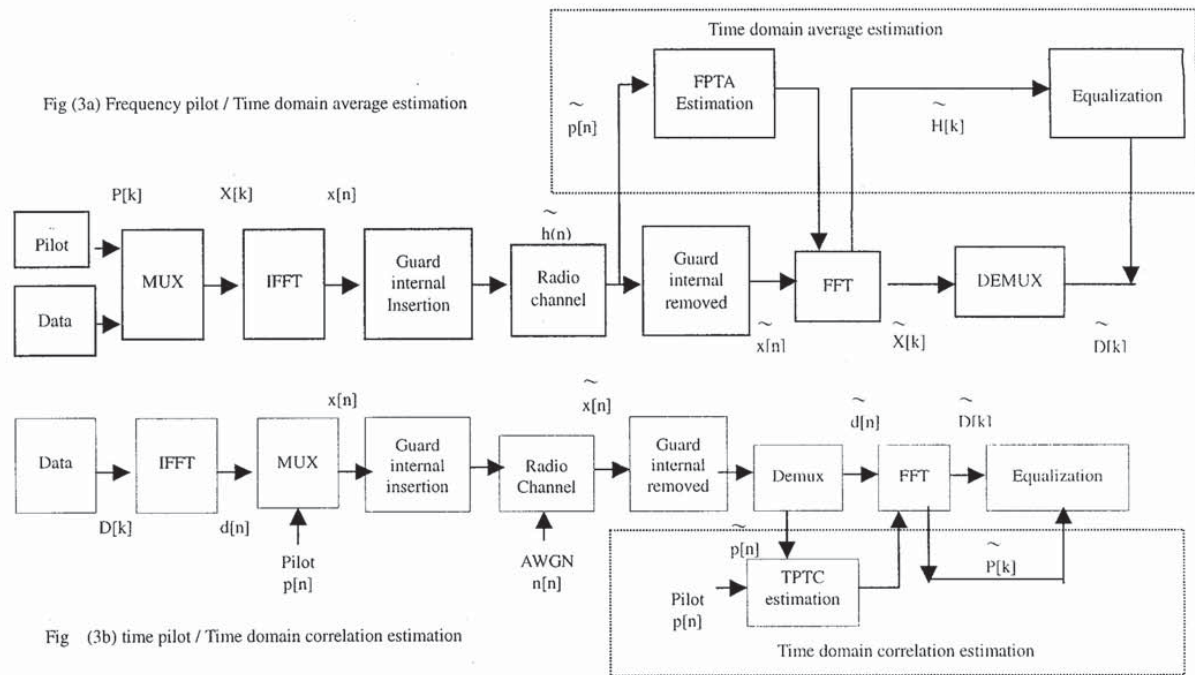


Figure. 3 Proposed Estimation Techniques (a) FPTA (b) TPTC

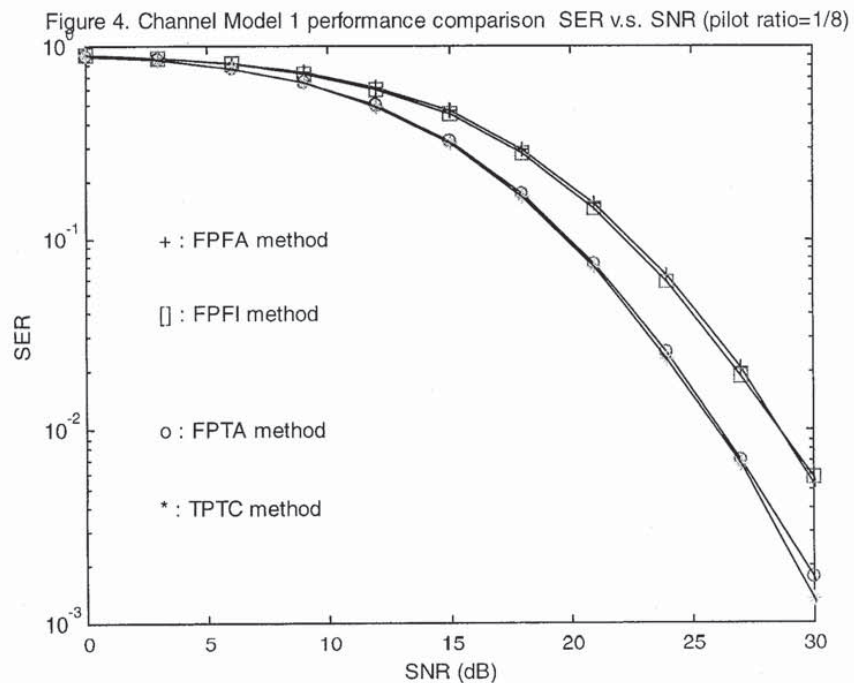


Figure 5. Channel Model 1 performance comparison SER v.s. SNR (pilot ratio=1/32)

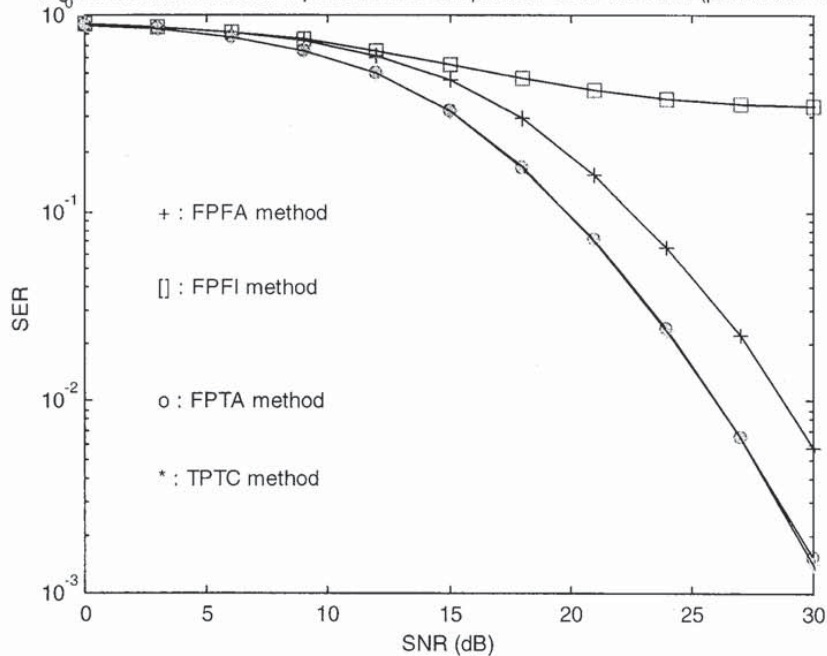


Figure 6. Channel Model 5 estimation MSE (CNR=10dB)

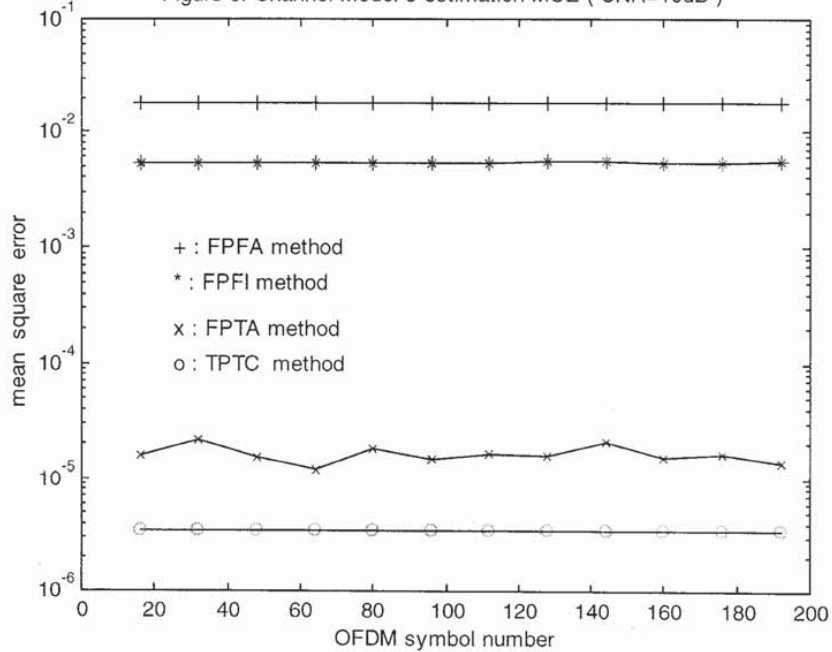


Figure 7. Channel Model 1 performance comparison SER v.s. SNR (pilot ratio=1/8)

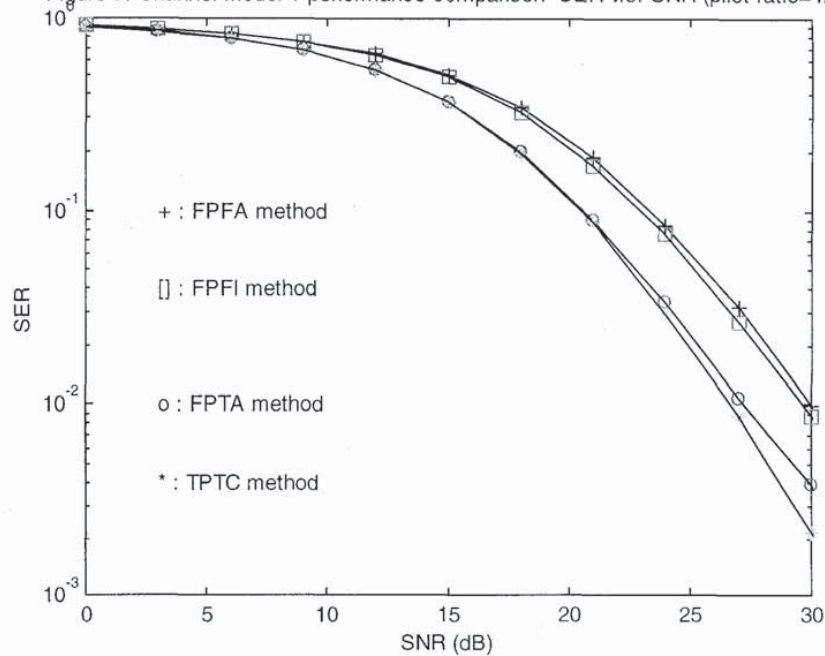
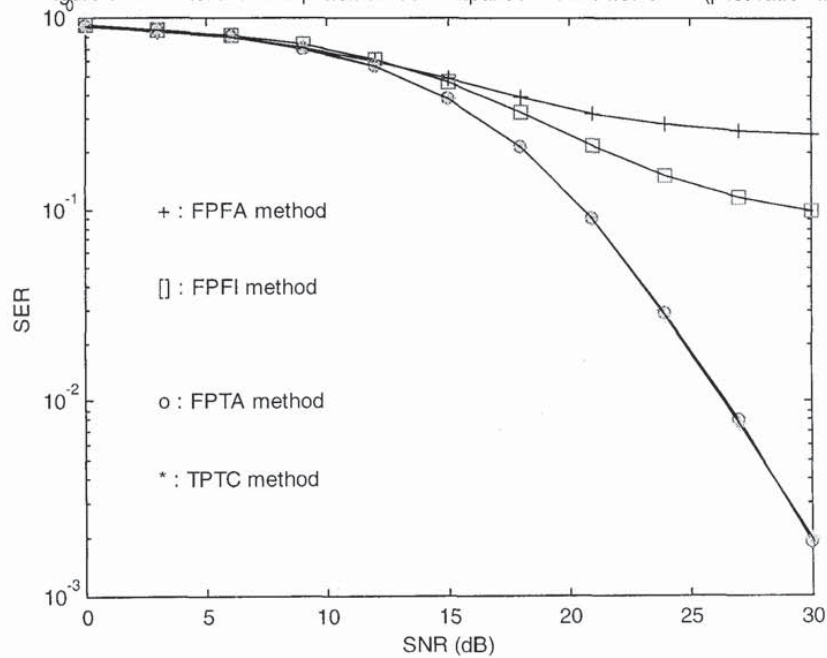


Figure 8. Channel Model 2 performance comparison SER v.s. SNR (pilot ratio=1/32)



Hsieh-Yee Declaration

Exhibit 0004

AN OVERVIEW OF SMART ANTENNA TECHNOLOGY FOR MOBILE COMMUNICATIONS SYSTEMS

PER H. LEHNE AND MAGNE PETTERSEN, TELENOR RESEARCH AND DEVELOPMENT

ABSTRACT

The technology of smart or adaptive antennas for mobile communications has received enormous interest worldwide in recent years. The purpose of this article is to give an overview of the technology and of the different ways in which it is foreseen to influence mobile communications systems. A broad definition of smart antennas is used in this article, namely base station antennas with a pattern that is not fixed, but adapts to the current radio conditions. The article starts with an explanation of the basic principles of smart antennas. Different levels of intelligence are introduced, ranging from simple switching between predefined beams to optimum beamforming. Also, an evolutionary path for the development of smart antennas is suggested. A discussion of the consequences for mobile systems is given. The principle reason for introducing smart antennas is the possibility for a large increase in capacity: an increase of three times for TDMA systems and five times for CDMA systems has been reported. Other advantages include increased range and the potential to introduce new services. Major drawbacks and cost factors include increased transceiver complexity and more complex radio resource management. The basic principles for implementation of smart antennas at the base stations are explained. First, the general principles for beamforming using array antennas are explored, then the special cases of different smart antenna implementations are discussed. Also, the fundamental difference between performing beamforming on uplink (base station receiving) and downlink (base station transmitting) is dealt with. Special attention is given to the critical factors and technological challenges, including achieving equal performance on uplink and downlink as well as real-time calibration of the receiver and transmitter chains. The article concludes with a short description of some of the trials and testbeds that have been implemented. In the autumn of 1998 Ericsson, in cooperation with Mannesmann Mobilfunk in Germany, became the first to allow commercial traffic through a base station employing smart antennas.

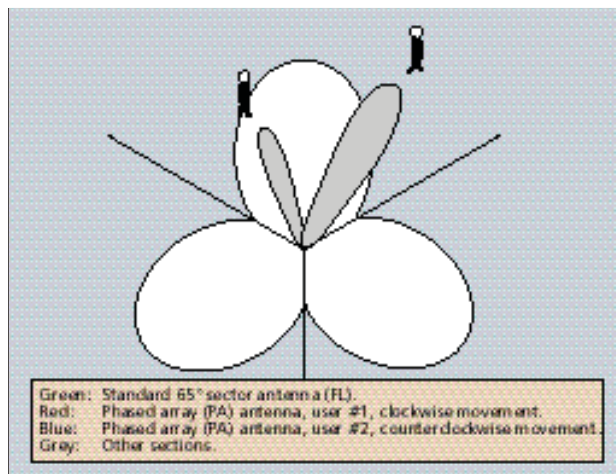
It is foreseen that in the future an enormous increase in traffic will be experienced for mobile and personal communications systems. This is due both to an increased number of users as well as new high bit rate data services being introduced. This trend is observed for second-generation systems and will most certainly continue for third-generation systems being introduced worldwide within a few years.¹ The increase in traffic will put a demand on both manufacturers and operators to provide enough capacity in the networks. Presently, one of the most promising techniques for increasing the capacity in cellular systems by the amount necessary is smart or adaptive antennas.

¹ While the data calls in the GSM network are in the order of 4 percent of the total traffic, data calls contribute to more than 50 percent of the total traffic in the fixed network (1998). It is expected that in the future a similar trend will be visible for cellular networks.

This article will describe the basic principles behind the technology as well as provide a discussion about the advantages and technical challenges. Although it is feasible to also use these techniques in, for example, satellite systems and high frequency wireless local area networks, the discussion here is limited to cellular systems in the frequency range 1-2 GHz. At these frequencies it is difficult to introduce smart antennas into the mobile stations due to the limited size of the terminal in terms of wavelengths.

CAPACITY ASPECTS

In a cellular system the radio communication is between the user and a *base station*, which provides radio coverage within a certain area, which is called a *cell*. Capacity in such a system can be defined as the total bit rate per unit bandwidth per unit area, or bit/s/Hz/m². Because the available frequency band is limited, the capacity is given by the *cell density*, the *fre-*



■ FIGURE 1. Illustration of the difference between a traditional base station radiation pattern and a smart antenna base station.

quency reuse distance and the number of users that can be served simultaneously by each base station. Techniques for increasing capacity in cellular systems have included using smaller cells, so-called *microcells*, and *frequency hopping*, a technique that disperses interference and averages the fading rate.

WHAT ARE SMART ANTENNAS?

Base station antennas have up till now been omnidirectional or sectored. This can be regarded as a “waste” of power as most of it will be radiated in other directions than toward the user. In addition, the power radiated in other directions will be experienced as interference by other users. The idea of smart antennas is to use base station antenna patterns that are not fixed, but adapt to the current radio conditions. This can be visualized as the antenna directing a beam toward the communication partner only. The difference between the fixed and the smart antenna concept is illustrated in Fig. 1. Smart antennas will lead to a much more efficient use of the power and spectrum, increasing the useful received power as well as reducing interference.

SHARING THE RADIO SPECTRUM

Traditionally, users communicating via the same base station have been separated by frequency, as in FDMA (frequency division multiple access); by time, as in TDMA (time division multiple access); or by code, as in CDMA (code division multiple access). Smart antennas add a new way of separating users, namely by space, through SDMA (*space division multiple access*). As will be explained later, SDMA, which means that users in the same cell can use the same *physical communication channel*, is the final step in an evolutionary path toward increasingly more advanced utilization. The *physical communication channel* is defined as a combination of carrier frequency, time slot and spreading code.

WHY NOW?

In the last couple of years smart antennas for mobile communications have received enormous interest worldwide. So why is the interest appearing now, and not five or ten years ago? The answer probably lies with the fact that for most operators there has not been much reason to worry about capacity and spec-

trum efficiency. Also, as will be apparent later on, if the base station is to track a large number of users simultaneously, the computational cost will be large. Only recently are processors with sufficient power becoming available. In addition to increased capacity, smart antennas also introduce a number of other advantages to cellular networks, including *increased range*, *a higher level of security*, and *the possibility for new services*. This will be discussed in more detail later.

OUTLINE

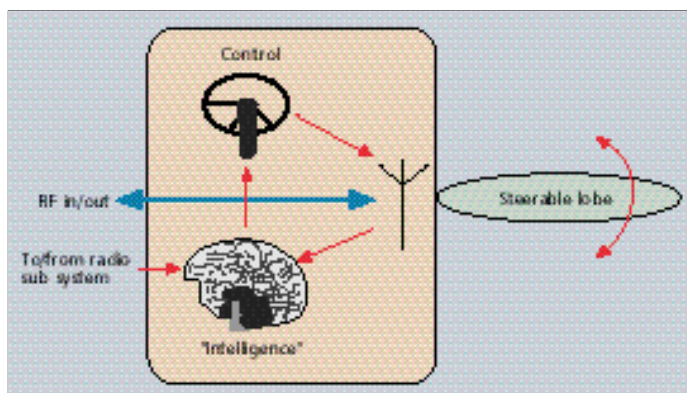
This article consists mainly of four parts. In the next section definitions of terms and descriptions of the basic principles are given. There is a discussion of the consequences of introducing the new technology into mobile communications networks. Later, implementation issues are discussed in more detail. We also describe some of the field trials and testbeds for smart antennas that have been carried out.

The general principles are valid for both TDMA systems (e.g., GSM) and CDMA systems (e.g., IS-95 and UMTS). In this article, most reported techniques and test results stem from TDMA-based testbeds, so the discussion is somewhat limited to this case. A comprehensive treatment of smart antennas applied to CDMA is, however, given by Liberti and Rappaport in [1].

BASIC PRINCIPLES

What do we mean by the term “smart antenna?” The theory behind smart antennas is not new. The technique has for many years been used in electronic warfare (EW) as a countermeasure to electronic jamming. In military radar systems similar techniques were already used during World War II. There are in principle a number of ways in which an adaptively adjustable antenna beam can be generated, for instance by mechanically steered antennas. However, the technology almost exclusively suggested for land-based mobile and personal communications systems is *array antennas*. This technology will be explained in more detail later.

The main philosophy is that interferers rarely have the same geographical location as the user. By maximizing the antenna gain in the desired direction and simultaneously placing minimal radiation pattern in the directions of the interferers, the quality of the communication link can be significantly improved. In personal and mobile communications, the interferers are other users than the user being addressed.



■ FIGURE 2. Principle of a smart antenna.

CONCEPTS

Several different definitions for smart antennas are used in the literature. One useful and consistent definition can be that the difference between a smart/adaptive antenna and a “dumb”/fixed antenna is the property of having an adaptive and fixed lobe-pattern, respectively. This is the definition to be used in this article. Figure 2 illustrates the concept of a smart antenna.

Normally, the term “antenna” comprises only the mechanical construction transforming free electromagnetic (EM) waves into radio frequency (RF) signals traveling on a shielded cable and vice versa. We may call it *the radiating element*. In the context of smart antennas, the term “antenna” has an extended meaning. It consists of a number of radiating elements, a combining/dividing network and a control unit. The control unit can be called the smart antenna’s intelligence, normally realized using a digital signal processor (DSP). The processor controls feeder parameters of the antenna, based on several inputs, in order to optimize the communications link. Different optimization criteria can be used, as will be explained in the next section. This shows that smart antennas are more than just the “antenna,” but rather a complete transceiver concept.

LEVELS OF INTELLIGENCE

Based on the definition above, one can define “levels of intelligence.” One set of definitions often used are described below and illustrated in Fig. 3.

- **Switched lobe (SL):** This is also called switched beam. It is the simplest technique, and comprises only a basic switching function between separate directive antennas or predefined beams of an array. The setting that gives the best performance, usually in terms of received power, is chosen. Because of the higher directivity compared to a conventional antenna, some gain is achieved. Such an antenna will be easier to implement in existing cell structures than the more sophisticated adaptive arrays, but it gives a limited improvement.
- **Dynamically phased array (PA):** By including a *direction of arrival* (DoA) algorithm for the signal received from the user, continuous tracking can be achieved and it can be viewed as a generalization of the switched lobe concept. In this case also, the received power is maximized.

- **Adaptive array (AA):** In this case, a DoA algorithm for determining the direction toward interference sources (e.g., other users) is added. The radiation pattern can then be adjusted to null out the interferers. In addition, by using special algorithms and space diversity techniques, the radiation pattern can be adapted to receive multipath signals which can be combined. These techniques will maximize the *signal to interference ratio* (SIR) (or signal to interference and noise ratio (SINR)).

Conventional mobile systems usually employ some sort of antenna diversity (e.g., space or polarization diversity). Adaptive antennas can be regarded as an extended diversity scheme, having more than two diversity branches. In this context, phased arrays will have a greater gain potential than switched lobe antennas because all elements can be used for diversity combining [2, 3].

All these concepts are analyzed in more detail later.

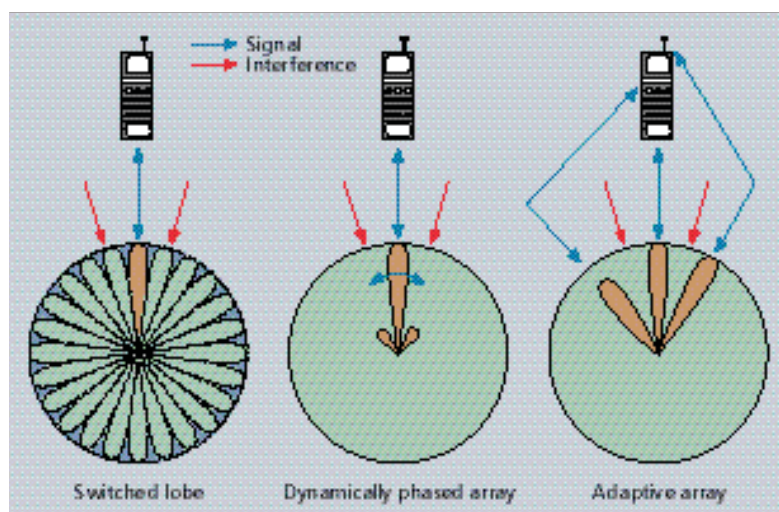
EVOLUTIONARY PATH

All the levels of intelligence described in the previous section are technologically realizable today. However, in the domain of personal and mobile communications, an evolution can be foreseen in the utilization of smart antennas toward gradually more advanced solutions. The evolution can be divided into three phases [4]:

- Smart antennas are used on *uplink* only (uplink means that the user is transmitting and the base station is receiving). By using a smart antenna to increase the gain at the base station, both the sensitivity and range are increased. This concept is called *high sensitivity receiver* (HSR) and is in principle not different from the diversity techniques implemented in today’s mobile communications systems.
- In the second phase, directed antenna beams are used on the *downlink direction* (base station transmitting and user receiving) in addition to HSR. In this way, the antenna gain is increased both on uplink and downlink, which implies a spatial filtering in both directions. Frequencies can be more closely reused, thus the system capacity increases. The method is called *spatial filtering for interference reduction* (SFIR). It is possible to introduce this in second-generation systems.
- The last stage in the development will be full *space division multiple access* (SDMA). This implies that more than one user can be allocated to the same physical communications channel simultaneously in the same cell, only separated by angle. In a TDMA system, two users will be allocated to the same time slot and carrier frequency at the same time and in the same cell.

In phase 2, the capacity is increased due to closer frequency reuse allowing more carriers per base station. In phase 3, an additional increase in capacity is achieved by allowing more users per carrier. Introducing SDMA in second-generation TDMA systems will be difficult and maybe undesirable [5], but it may become a natural component in third-generation systems.

The “levels of intelligence” in the previous section describe the level of technological development, while the steps described here can be regarded as part of a system evolution.



■ FIGURE 3. Different smart antenna concepts.

CONSEQUENCES

The introduction of smart antennas will have a large impact on the performance of cellular networks. It will also affect many aspects of both the planning and deployment of mobile systems. This chapter will discuss the potential benefits and cost factors, and will also briefly describe the implications on radio planning.

IMPROVEMENTS AND BENEFITS

Capacity Increase — The principle reason for the growing interest in smart antennas is the capacity increase. In densely populated areas mobile systems are normally interference-limited, meaning that interference from other users is the main source of noise in the system. This means that the signal to interference ratio, SIR, is much larger than the signal to thermal noise-ratio, SNR. Smart antennas will on average, by simultaneously increasing the useful received signal level and lowering the interference level, increase the SIR. Especially, the adaptive array will give a significant improvement. Experimental results report up to 10 dB increases in average SIR in urban areas [6].

In TDMA systems, the implication of the increased SIR is the possibility for reduced frequency reuse distance. An example is shown in Fig. 4, where the traditional seven-cell cluster has been reduced to a three-cell cluster. This will lead to a capacity increase of 7/3 as all the cells can be allocated this amount more carriers.

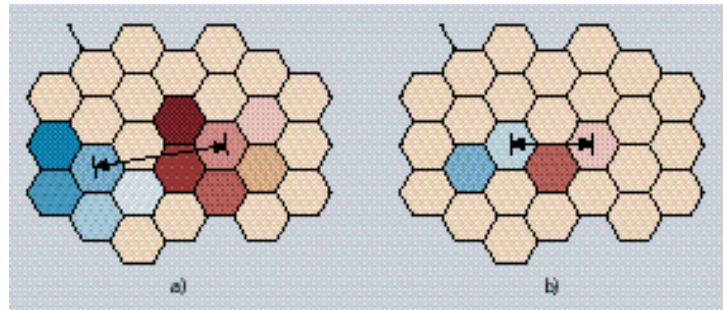
Simulations performed on a FH-GSM network with 1/3 reuse distance utilizing SFIR reports that a capacity increase of 300 percent can be expected [5].

CDMA systems, such as IS-95 or UMTS, are more inherently interference-limited than TDMA systems. The main source of noise in the system is the interference from other users due to the spreading codes being non-ideally orthogonal. This means that the expected capacity gain is even larger for CDMA than for TDMA. A fivefold capacity gain has been reported for CDMA in [7].

Range Increase — In rural and sparsely populated areas radio coverage rather than capacity will give the premises for base station deployment. Because smart antennas will be more directive than traditional sector or omnidirectional antennas, a range increase potential is available. This means that base stations can be placed further apart, potentially leading to a more cost-efficient deployment. The antenna gain compared to a single element antenna can be increased by an amount equal to the number of array elements, e.g., an eight-element array can provide a gain of eight (9 dB).

New Services — When using smart antennas the network will have access to spatial information about the users. This information can be used to estimate the positions of the users much more accurately than in existing networks. Positioning can be used in services such as emergency calls and location-specific billing. The FCC (Federal Communications Commission) in the United States has decided that by October 2001 user location information with an accuracy of 125 meters RMS error must be provided [8].

Security — It is more difficult to tap a connection when smart antennas are used. To successfully tap a connection the intruder must be positioned in the same direction as the user as seen from the base station.



■ FIGURE 4. Illustration of reduced frequency reuse distance: a) Traditional 7-cell cluster; b) cluster enabled by interference reduction like e.g. smart antennas.

Reduced Multipath Propagation — By using a narrow antenna beam at the base station the multipath propagation can be somewhat reduced. The actual reduction depends on the scenario, and is not always significant. Although channel equalizers and RAKE receivers most often will handle and even exploit the multipath components, on very high-speed connections this may not be the case. Potentially, the reduction of multipath propagation can be used to ease the requirement on future modem design.

COST FACTORS

Although the benefits of using smart antennas are many, there are also drawbacks and cost factors. The gain should always be evaluated against the cost.

Transceiver Complexity — It is obvious that a smart antenna transceiver is much more complex than a traditional base station transceiver. The antenna will need separate transceiver chains for each of the array antenna elements and accurate real-time calibration of each of them. In addition, the antenna beamforming is a computationally intensive process, especially if *adaptive arrays* are to be used (see earlier section). This means that the smart antenna base station must include very powerful numeric processors and control systems. There will be a growing need for the development of efficient algorithms for real-time optimizing and signal tracking. Smart antenna base stations will no doubt be much more expensive than conventional base stations.

Resource Management — Smart antennas are mainly a radio technology, but they will also put new demands on network functions such as resource and mobility management. When a new connection is to be set up or the existing connection is to be handed over to a new base station, no angular information is available to the new base station and some means to “find” the mobile station is necessary. This can be handled by letting the base station continuously sweep through the cell with a “search” beam looking for candidates for a new connection or a handover. Another possibility is to use an external system for positioning, e.g., GPS. As far as handover is concerned, a third possibility is available: directional information from the existing cell can be used by the network to provide an “educated guess” about which cell to hand the connection over to.

As was explained earlier, SDMA involves different users using the same physical communication channel in the same cell, separated only by angle. When angular collisions between these users occur, one of them must quickly switch to another channel so that the connection is not broken. This means that in systems providing full SDMA, there will be much more *intracell handovers* than in conventional TDMA or CDMA



■ FIGURE 5. Picture of an 8-element array antenna at 1.8 GHz. (Antenna property of Telia Research AB, Sweden).

systems, and more monitoring by the network is necessary. The animation in Fig. 1 shows what is meant by an angular collision.

Physical Size — For the smart antenna to obtain a reasonable gain, an array antenna with several elements is necessary. Typically arrays consisting of six to 10 horizontally separated elements have been suggested for outdoor mobile environments [9–11]. The necessary element spacing is 0.4–0.5 wavelengths. This means that an eight-element antenna would be approximately 1.2 meters wide at 900 MHz and 60 cm at 2 GHz. With a growing public demand for less visible base stations, this size, although not excessive, could provide a problem. Figure 5 shows a picture of an eight-element antenna array at 1.8 GHz.

RADIO PLANNING

For the smart antennas to have the desired gain the base stations must be able to separate the users in the angular domain. This means that some of the current strategies for radio planning need to be modified. In existing networks base stations are often positioned along highways and railway lines. From a smart antenna point of view it is much more efficient to position the base station away from the road or railway. In this way the spatial dimension is better exploited. This strategy is illustrated in Fig. 6.

Another difficult point is small urban cells with base stations positioned below rooftop level. In these cases the signal from both the desired users and from the interferers will tend to arrive along the street canyons, making it difficult to separate them in the angular domain. Figure 7 shows an example of an average *power azimuth spectrum* (PAS)² measured in urban area microcells in Oslo [12] and shows that the signal tends to arrive from two perpendicular directions given by the street grid.

Radio channel models and radio planning tools need renewed attention upon the introduction of smart antennas. Statistical radio channel models are used in performance testing and design of radio systems. However, existing models do not describe the channel in the angular domain and can therefore

not be used in the evaluation of systems employing smart antennas. Work is currently ongoing to provide statistical radio channel models describing the directional domain.

Similarly, existing commercial radio planning tools are normally only concerned with estimating the signal level and possibly the time dispersion on the link. However, in the future it will be beneficial for the tools to predict the angular dispersion as well. Some analytical radio prediction methods, such as *ray tracing*, have the possibility for estimation of angular information inherent.

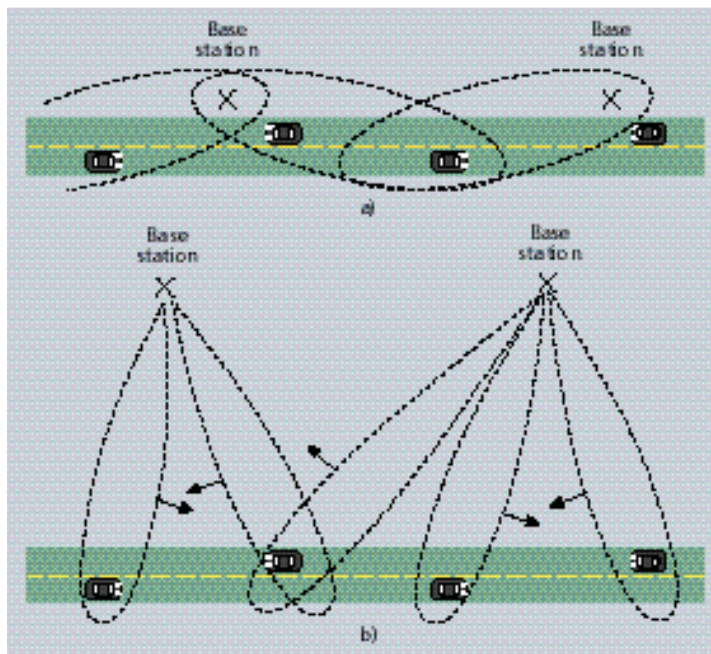
IMPLEMENTATION

This section explains some of the underlying technology employed in order to implement smart antennas. The treatment given here is not comprehensive, but should give an impression of the basic ideas. The technology is based on array antennas where the radiation pattern is altered by adjusting the amplitude and relative phase on the different array elements. Next we explain the fundamental principles of array beamforming. Then we explain how this is exploited in smart antennas. Even though the reception and transmission parts most often are integrated, and much of the same hardware normally will be used, they will be explained separately. This is because conceptually the uplink and downlink are quite different in terms of smart antennas.

ARRAY ANTENNAS

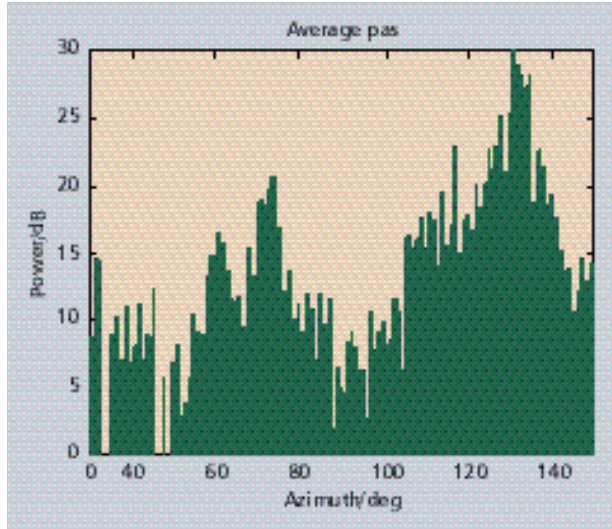
Electronically steerable antenna patterns are most often generated using array antennas. These are antennas consisting of a number of antenna elements on which the signal is divided or combined in both phase and amplitude. Generally, any combination of elements can form an array. However, usually equal elements in a regular geometry are used.

Arrays can be one-, two- and three-dimensional, depending



■ FIGURE 6. Illustration of base station positioning along roads and railways: a) Traditional deployment; b) Possible smart antenna deployment.

² An average power azimuth spectrum shows an area average of the power received from the azimuth antenna view angles.



■ FIGURE 7. Example of average power azimuth spectrum (PAS), recorded in Oslo.

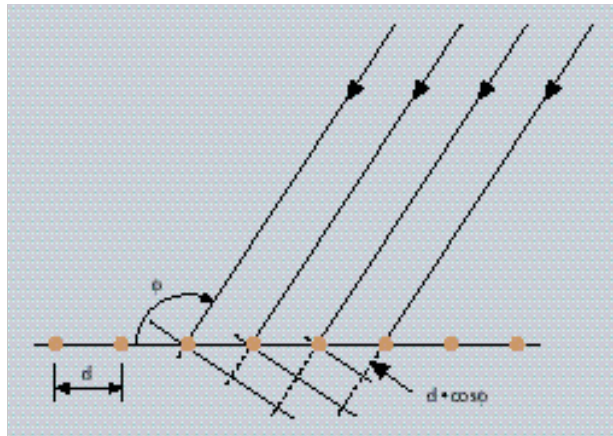
on the dimension of the space one wants to access. Figure 13 shows some examples of array geometries. Using regular geometries simplifies both analysis and synthesis of the array. In this section, the linear array principle will be worked through as an example. The mathematics can be extended to more dimensions and other geometries, but the principle is the same.

The total radiation pattern is given by the *element type*, the *relative positions* and the excitation (amplitude and phase). Using an array antenna, it is possible to obtain a very good control of the radiation pattern, e.g., the shape of the main lobe and the side lobe level (SLL). The radiation pattern is uniquely given by the aforementioned parameters.

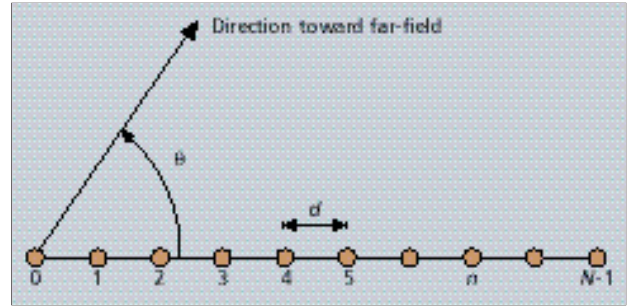
Linear Equidistant Arrays — The “simplest” geometry is the *one-dimensional linear equidistant array*. All the array elements are placed along a line with equal distance between them, as shown in Fig. 8. The figure also shows which parameters are available.

The far-field expression of the electrical field from a single array element is given by Eq. 1:

$$\vec{E}_n(r, \theta, \varphi) = I_n \cdot \frac{e^{-jkr}}{r} \cdot \vec{e}(\theta, \varphi) \quad (1)$$



■ FIGURE 9. Using the phase angle of each element to steer the lobe.



■ FIGURE 8. Geometry of the equidistant linear array.

where I_n is the complex element current and $\vec{e}(\theta, \varphi)$ is the element antenna pattern, usually called the *element factor*, and r , θ and φ are the spherical coordinates of the observation point.

The total field from an array consisting of N elements is the superposition of the element contributions as expressed in Eq. 2:

$$\vec{E}(r, \theta, \varphi) = \sum_{n=0}^{N-1} I_n \cdot \frac{e^{-jkr}}{r} \cdot e^{jkr'_n \cos \psi_n} \cdot \vec{e}(\theta, \varphi) \quad (2)$$

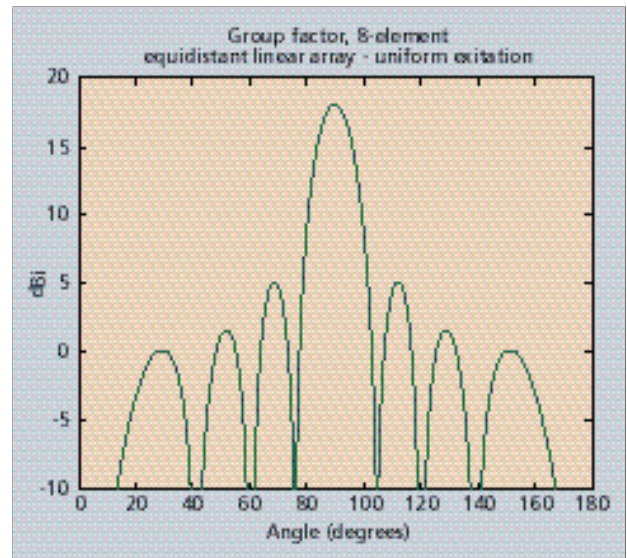
where

$$r'_n \cos \psi_n = z'_n \cos \theta = nd \cos \theta \quad (3)$$

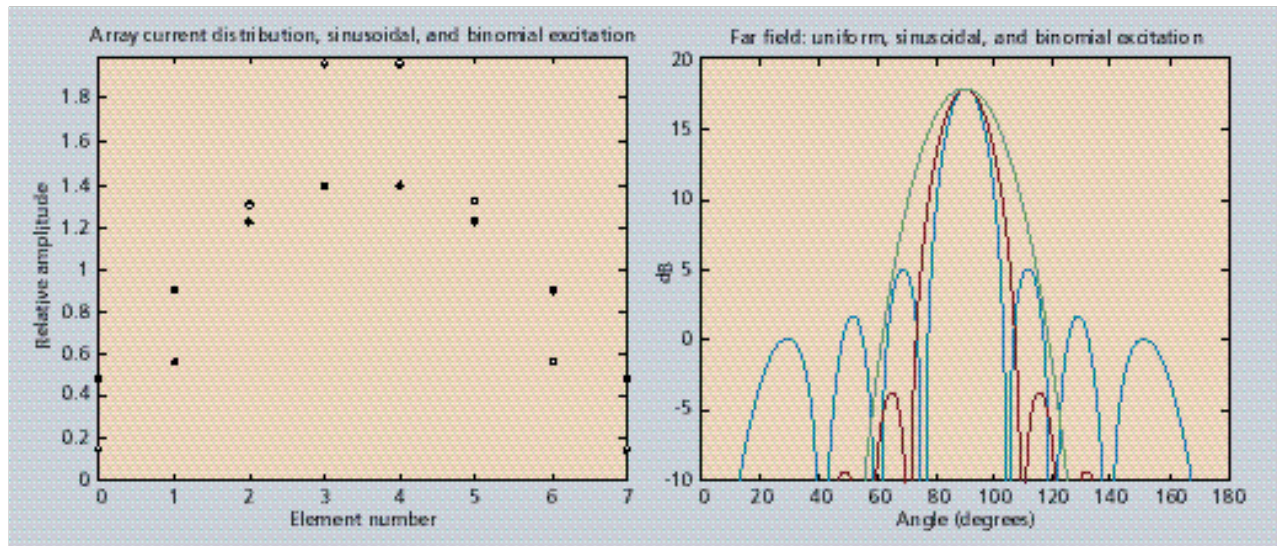
is the difference in distance for element n compared to element 0 toward the observation point in the direction θ due to the spatial separation in the array. The element spacing is d . (r'_n is the distance between the coordinate systems origin and element n , ψ_n is the corresponding angle, and z'_n , the distance along the array axis.) In the case of an array with equal elements, the far-field expression becomes

$$\vec{E}(r, \theta, \varphi) = \frac{e^{-jkr}}{r} \cdot \vec{e}(\theta, \varphi) \sum_{n=0}^{N-1} I_n \cdot e^{jkrnd \cos \theta} \quad (4)$$

The sum expression in Eq. 4 is the array factor which is independent of both the element type and the distance from the antenna. It is only dependent on the *array geometry* and the *element excitation*. Thus, the total far field is given by the product of the element factor and the array factor. The array



■ FIGURE 10. The resulting array factor with uniform excitation ($\alpha = 0$).



■ FIGURE 11. Examples of non-uniform excitation and resulting far-field.

factor can be understood as the radiation pattern from an array antenna consisting of isotropic elements.

As the radiation pattern is uniquely given by the excitation parameters, a careful choice can give us the required effect. Consider, for example, that it is desired to steer the main lobe of the antenna in a certain direction, φ . Thus, we want the far-field signals from all the elements of the array to be added *in phase* in the wanted direction. Figure 9 shows what this means. If element number 0 is given the reference phase angle 0, then the phase angle of element number 1 must be shifted (in this case negatively) with a value of $k d \cos \varphi$, where $k = 2\pi/\lambda$ is the wave constant. We can call this angle α . Thus element number 2 must be shifted 2α , element 3 must be shifted 3α , and so on. In order to obtain maximum gain in the

direction of φ , the amplitudes of all the elements must be equal. This example is called *uniform excitation* ($|I_0| = |I_1| = \dots = |I_n| = \dots = |I_{N-1}|$). Figure 10 shows the resulting array factor in this case.

A drawback of the linear array is the front-back ambiguity. A mirror of the antenna pattern will appear around the antenna axis, i.e., in the direction $-\varphi$ ($\cos \varphi = \cos(-\varphi)$). Most often we want only to have one main lobe, and this is usually obtained by using directive elements. It is also worth noting that for skew angles, the gain decreases and the main lobe width increases, because the element separation appears smaller in the maximum direction of the main lobe (apparent distance: $d \sin \varphi$). In practice, the linear array is only useful in a sector of at most 120° .

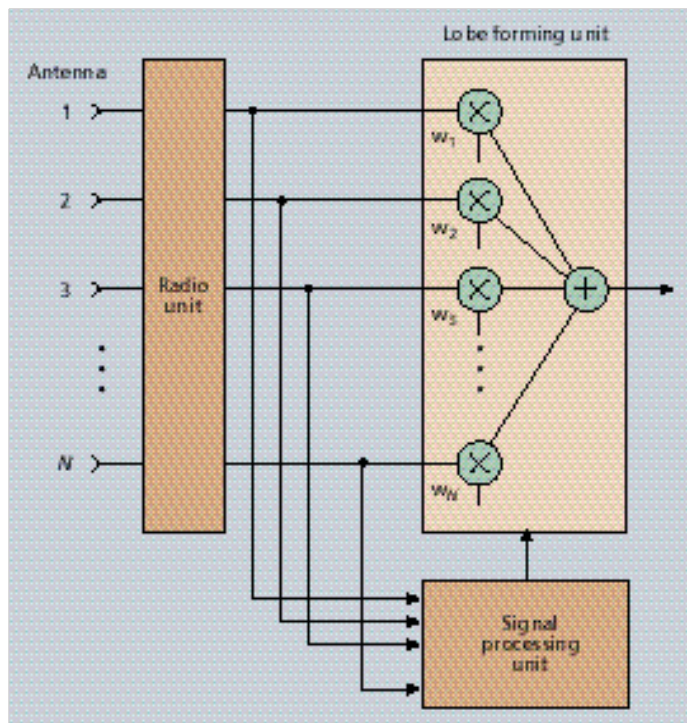
Using uniform excitation gives the highest gain but also the highest sidelobe level (SLL) of -3.2 dB. From the expression for the array factor one sees that the far-field is given as the Fourier-transform of the current distribution along the array. Thus, main lobe width and width of the current distribution are inverse. Using a non-uniform excitation with an amplitude roll-off toward the end of the array results in both a wider main lobe and a lower SLL. Figure 11 shows some examples.

Practical Considerations — Above we have anticipated that the element patterns are all identical and “perfect.” In implementing an array antenna, this is not the case. The different elements of the antenna will have slightly different patterns due to production tolerances and so on. Additionally, when mounting several elements close together (typically less than $\lambda/2$) they influence each other. This mutual coupling gives changes both to the element patterns and the impedance match. This means that the array performance will be somewhat degraded.

RECEIVER

Figure 12 shows schematically the elements of the reception part of a smart antenna. The antenna array contains N elements. The N signals are being combined into one signal, which is the input to the rest of the receiver (channel decoding, etc.).

As the figure shows, the smart antenna reception

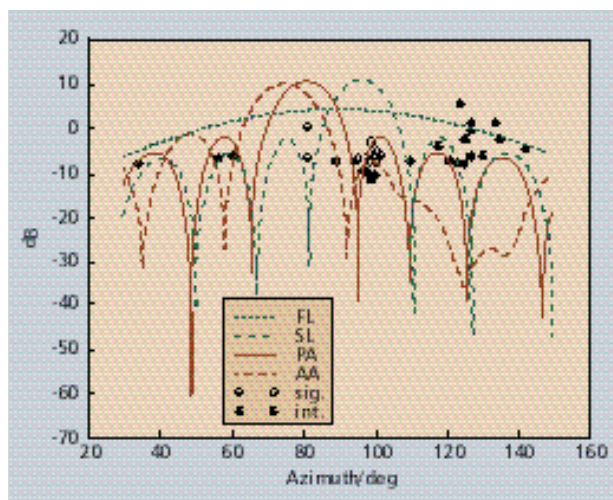


■ FIGURE 12. Reception part of a smart antenna.

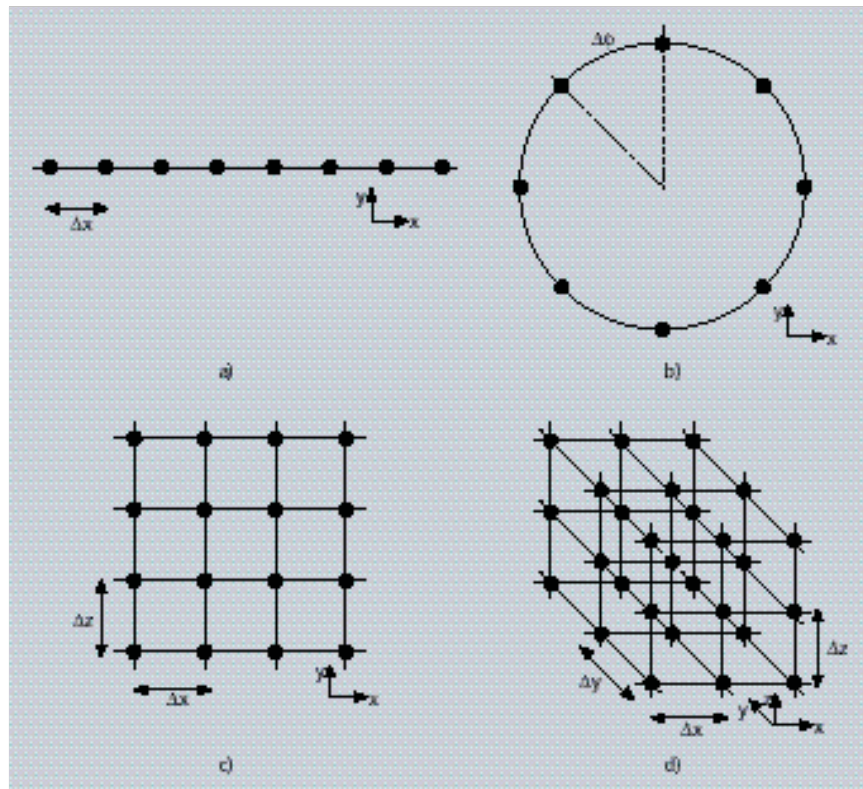
part consists of four units. In addition to the antenna itself it contains a radio unit, a lobe forming unit, and a signal processing unit [13].

The array will often have a relatively low number of elements in order to avoid unnecessarily high complexity in the signal processing. Figure 13 shows four examples of different array geometries. The first two structures are used for beamforming in the horizontal plane (azimuth) only. This will normally be sufficient for outdoor environments, at least in large cells. The first example (a) shows a one-dimensional linear array with uniform element spacing of Δx . This structure can perform beamforming in azimuth angle within an angular sector. This is the most common structure due to its low complexity. The second example (b) shows a birds eye view of a circular array with angular element spacing of $\Delta\Phi = 2\pi/N$. This structure can perform beamforming in all azimuth angles. The last two structures are used for performing two-dimensional beamforming, in both azimuth and elevation angles. This may be desirable for indoor or dense urban environments. Figure 13c is the front view of a two-dimensional linear array with horizontal element spacing of Δx and vertical element spacing of Δz . This structure can perform beamforming within a solid angle. Beamforming in the entire space, within all solid angles, requires some sort of cubic or spherical structure. Figure 13d shows an example of a cubic structure with element separations of Δx , Δy and Δz .

The radio unit consists of down-conversion chains and (complex) analog-to-digital conversion (A/D). There must be N down-conversion chains, one for each of the array elements.



■ FIGURE 14. Resulting antenna patterns for SL, PA, and AA for measured signal and interference channels. The figure also shows a so-called “fixed lobe” (FL) conventional antenna pattern for comparison.



■ FIGURE 13. Different array geometries for smart antennas: a) uniform linear array; b) circular array; c) 2 dimensional grid array; d) 3 dimensional grid array.

The signal processing unit will, based on the received signal, calculate the complex weights $w_1 - w_N$ with which the received signal from each of the array elements is multiplied. These weights will decide the antenna pattern in the uplink direction as described in the previous section. The weights can be optimized from two main types of criteria: maximization of received signal from the desired user (e.g., switched lobe or phased array) or maximization of the SIR by suppressing the signal from interference sources (adaptive array). In theory, with M antenna elements one can “null out” $M-1$ interference sources, but due to multipath propagation this number will normally be lower.

The method for calculating the weights will differ depending on the type of optimization criterion. When *switched lobe* (SL) is used the receiver will test all the pre-defined weight vectors (corresponding to the lobe set) and choose the one giving the strongest received signal level. If the *phased array* approach (PA) is used, which consists of directing a maximum gain beam toward the strongest signal component, the *direction-of-arrival* (DoA) is first estimated and then the weights are calculated as described in the previous section with uniform amplitude and phase in accordance with the desired steering angle. A number of well documented methods exist for estimating the DoA, for instance, MUSIC, ESPRIT, or SAGE [4, 14].

If maximization of SIR is to be done (AA), the optimum weight vector (of dimension N) W_{opt} is given by the dominant generalized eigenvector of the matrix pair $[R_k, Q_k]$ where R_k is the $N \times N$ correlation matrix of the total received signal and Q_k is the $N \times N$ correlation matrix of the total interference. This is called *optimum combining* [15]. Figure 14 shows an example of the resulting antenna patterns for PA, SL, and AA in a case where both the signal and interference are measured in a real channel. Note how the AA directs nulls toward the

strongest interference sources. Other techniques are also possible, for example, well documented diversity algorithms such maximum-ratio in multiple dimensions [16].

For the base station to be able to estimate the radio channel, a reference or training sequence is normally necessary, i.e., a known bit sequence must be transmitted periodically. The methods that maximize the SIR do in principle require knowledge of the instantaneous channel response from both the desired user and all the interference sources, which means the training sequence must be unique for each user.

In the lobe forming unit the actual weighting of the received signal from each of the array elements is done. In the most advanced case this unit will be an integration of the channel equalizer/RAKE receiver and the smart antenna. In this case $N \cdot D$ weights will be required, where D is the number of symbol periods (depth) in the channel equalizer or number of fingers in the RAKE receiver. This is called a *spatio-temporal filter* [17, 18], because it removes the undesired signal components and keeps the desired ones both in time and space domains. Such a lobe forming unit is shown schematically in Fig. 15 [19]. In this example the number of weights is $N \cdot 3$, showing that the depth of the equalizer is 3. The time between the taps on each antenna element is T . In a RAKE receiver, the time delay between the taps is not uniform.

When the lobe forming is done digitally (after A/D) the lobe forming and signal processing units will normally be integrated in the same unit, namely the DSP. The separation in Fig. 12 is done to clarify the functionality. It is also possible to perform the lobe forming in hardware at *radio frequency* (RF) or *intermediate frequency* (IF) [20].

The three aforementioned methods (SL, PA, and AA) dif-

fer in complexity and computational burden. However, a detailed analysis of this is out of the scope of this article because many parameters and assumptions must be defined in order to quantify this difference. Generally, SL is less complex than PA, which in turn is less complex than AA. Conditions that influence this are, for example, the required accuracy and performance in terms of power or C/I-gain. It is also expected that the radio access method makes a difference, i.e., TDMA vs. CDMA.

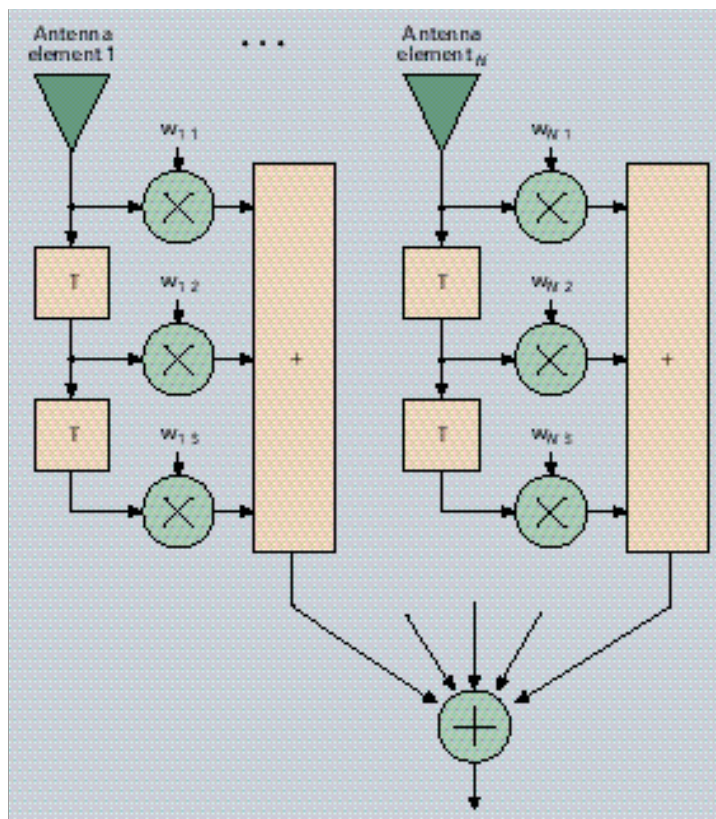
TRANSMITTER

The transmission part of the smart antenna will be schematically very similar to the reception part. An illustration is shown in Fig. 16. The signal is split into N branches, which are weighted by the complex weights $z_1 - z_N$ in the lobe forming unit. The weights, which decide the radiation pattern in the downlink direction, are calculated by the signal processing unit. The radio unit consists of D/A converters and the up converter chains. In practice, some components, such as the antennas themselves and the DSP, will of course be the same as on reception.

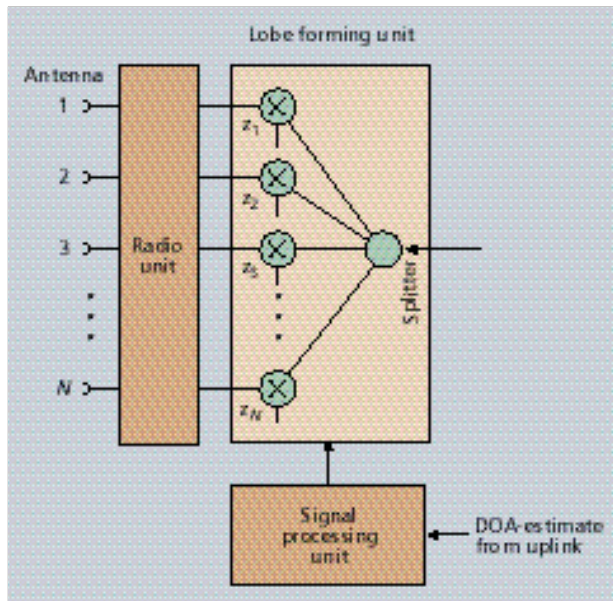
The principal difference between uplink and downlink is that since there are no smart antennas applied to the user terminals (mobile stations) no knowledge of the spatial channel response is available on downlink. In a time division duplex (TDD) system the mobile station and base station use the same carrier frequency only separated in time. In this case the weights calculated on uplink will be optimal on downlink if the channel does not change during the period from uplink to downlink transmission. However, this can not be assumed to be the case in general, at least not in systems where the users are expected to move at high speed. If frequency division duplex (FDD) is used, the uplink and downlink are separated in frequency. In this case the optimal weights will generally not be the same because of the channel response dependency on frequency.

Thus optimum beamforming (i.e., AA) on downlink is difficult, and the technique most frequently suggested is the geometrical approach of estimating the direction-of-arrival (DoA). The assumption is directional reciprocity, i.e., the direction from which the signal arrived on the uplink is the direction in which the signal should be transmitted to reach the user on downlink. This assumption has been strengthened by recent experimental results [21]. The strategy used by the base station is to estimate the DoA of the direction (or directions) from which the main part of the user signal is received. This direction is used on downlink by choosing the weights $z_1 - z_N$ so that the radiation pattern is a lobe (or lobes) directed toward the desired user. This is similar to PA. In addition, it is possible to position zeros in the direction toward other users so that the interference suffered by these users is minimized. Due to fading on the different signal paths, it has been suggested to choose the downlink direction based on averaging the uplink channel over a period of time [11]. This will, however, be sub-optimum compared to the uplink situation where knowledge about the instantaneous radio channel is available.

It should be stressed that in the discussion above it is assumed that the interferers observed by the base stations are mobile stations and that the interferers observed by the mobile stations are base stations. This means that when the base station on



■ FIGURE 15. Illustration of the lobe forming unit in the case of a spatio-temporal filter with three time bins and N antenna elements.



■ FIGURE 16. Transmission part of a smart antenna.

transmission positions zeros in the direction toward other mobile stations than the desired one, it will reduce the interference suffered by these mobiles. If, however, the interferers observed by mobiles are other mobiles, as may be the case, for instance, in UTRA TDD [22], there will be a much more fundamental limitation in the possibility for interference reduction at the mobile.

CRITICAL FACTORS AND CHALLENGES

A number of technical challenges remain to be solved for smart antennas to reach their full potential. These are aspects on both modem implementation as well as the algorithms for beamforming.

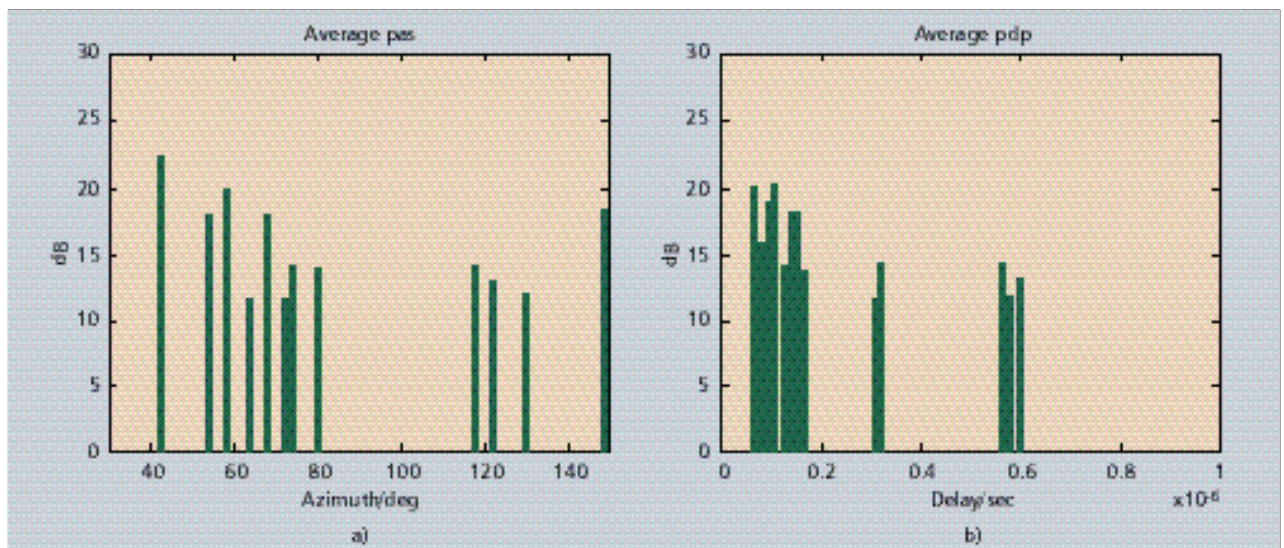
As mentioned in the previous section, beamforming on the downlink is made more difficult by the lack of knowledge about the instantaneous channel response. The strategy of directing the lobe (or lobes) in the direction(s) of the user sig-

nal DoA is often used, but this makes the downlink performance very dependent on the radio channel and its rate of change. If the channel suffers from much angular spread, i.e., the signal arrives from many directions, directing a single lobe in the direction of maximum reception is far from optimum in terms of maximizing the power transfer to the user. Directing zeros toward other users to minimize the interference suffered by these is difficult for the same reason. When the channel is changing rapidly, which will be the case for fast moving mobiles and for very dense, complicated environments, the difficulties on downlink will be worsened. Figure 17 shows a recording of an angular radio channel from a suburban microcellular environment. The distance between each channel measurement was 2 cm corresponding to 10 ms between measurements at a speed of 2 m/s (walking speed).

Another critical factor is the linearity in reception and transmission chains. The transfer functions of all the up- and down-converter chains need to be perfectly known for the beamforming to be accurate. However, since there are active components involved the responses will change with time and temperature. This means that for the smart antenna to function properly frequent on-line calibration of the radio units needs to be performed [20].

The requirements on linearity in transceiver chains can possibly be relaxed by performing the A/D (and D/A) conversion as close to the array element as possible, that is sampling on RF or IF. This requires very fast and specialized hardware.

The complexity of the smart antenna is a great challenge because the most advanced smart antenna implementations involve simultaneously maximizing the useful signal and nulling out the interference sources. In addition, this beamforming needs to be done for each user communicating via the base station in question. Even with the powerful signal processing units available today it is a very challenging task to perform this in real time. Much work will need to be done on optimizing code and developing efficient algorithms for this to work properly. However, developments in computer technology over the past few years have undoubtedly made it possible to perform even more complex processing in real time. In the last three years processor speed has increased more than five times, and if this trend continues, there will be processors with 2-3 GHz clock frequency available when UMTS is to be introduced.



■ FIGURE 17. Recording of an angular radio channel in a suburban microcellular environment. The distance between measurements is 2 cm.: a) Power azimuth spectrum (PAS); b) Power delay profile (PDP).

STATUS: TRIALS AND TESTBEDS

Smart antennas for mobile communications are not yet available commercially. But it is a well known fact that many manufacturers are working on solutions for both second-generation systems and third-generation systems. This section will give a short overview of some of the documented trials and testbeds, but should not be considered as a complete list.

ERICSSON/MANNESMANN TRIAL IN GIESSEN, GERMANY

Ericsson has performed a trial in cooperation with the German GSM1800 operator Mannesmann in the city of Giessen. Detailed results from the trial have not yet been published, but preliminary results are promising [9]. Three base stations were equipped with smart antennas. The base stations were placed at high positions, giving a large coverage. In the autumn of 1998 the Ericsson/Mannesmann trial became the first trial to demonstrate commercial traffic through base stations equipped with smart antennas.

Physically, the smart antennas were six-element stacked dipoles. Each element was dual polarized, with $\pm 45^\circ$ polarized elements. The combining on uplink was based on maximal ratio combining, while the downlink beam forming was based on switching between eight lobes.

TSUNAMI II TRIAL IN BRISTOL, UK

TSUNAMI II is a project in the European 4th framework program ACTS³ dealing with the development of a smart antenna testbed and the implementation of a smart antenna base station. Some of the participants in this project have been the University of Aalborg (DK), Bosch Telekom (D), Orange (UK), and the University of Bristol (UK).

The project has finished and performed trials near Bristol (UK) in both urban and rural environments. Even though not all data analyses have been published yet, the trials confirm the capacity increase of a smart antenna. In the rural environment (macrocell) an extension of the range of up to 54 percent was reported, using the HSR principle. Interference reduction was better than 30 dB, which is much higher than expected. However, only one interferer was used, which in addition was placed more than 10 degrees away from the transmitter system [10, 23].

The antenna was eight-element dual polarized. Different strategies for combining were tested. It was shown that the adaptive array-based combining techniques, which included some sort of interference rejection, showed the best results [10].

OTHER

The U.S. company Arraycomm has developed smart antenna solutions for the GSM standard and for the Japanese Personal Handyphone System (PHS). In June 1997 a smart antenna wireless local loop (WLL) solution based on using PHS was commercially available [24]. Field trials have been performed in several countries, including the U.S., Brazil, and Malaysia. The Canadian company Nortel also claims to have been the first to implement a smart antenna for GSM [25].

Other manufacturers and research institutions are also developing testbeds and trials. Alcatel, in cooperation with the University of Vienna, is planning a testbed [11]. Much work

has also been performed in the U.S., including at Telcordia [26] and Virginia Tech [27]. Other test results have been reported from Korea [28] and Japan [29].

CONCLUSIONS

The main goal of this article has been to provide an overview of smart antenna technologies with respect to third-generation land mobile systems, i.e., IMT2000 and UMTS. It is obvious that smart antennas at the base stations will be an important technology to provide the necessary capacity and coverage. It also helps to realize new services, e.g., based on user location.

From a technology point of view, smart antennas can be seen as an extension of the "conventional" resource allocation schemes used in radio communications. In addition to dividing the space into cells, it will now also be possible to employ space division inside each cell. Different degrees of utilization of the spatial dimension are possible, and different steps have been described here.

Smart antenna technology is a broad concept and implementations range from simple techniques that involve switching between lobes to advanced algorithms maximizing the received signal-to-interference ratio. Implementation of smart antennas is done using array antennas. The techniques for beamforming with array antennas are well known, and must be employed in both duplex directions for the improvements to be substantial. However, with rapid channel variations it is not a trivial task to provide optimum beamforming, especially for the downlink direction.

The use of smart antennas is not purely a radio transmission issue. It also influences network services such as handover and connection setup. Introducing the spatial domain in the resource management system makes this more complex.

Several smart antenna testbeds and field trials have been set up and run by manufacturers and research institutions. The first tests allowing commercial traffic over a base station employing smart antennas were performed by Ericsson and Mannesmann Mobilfunk in Germany in the autumn of 1998.

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BIOGRAPHIES

PER H. LEHNE (per-hjalmar.lehne@telenor.com) graduated from the Norwegian Institute of Technology (NTH) in 1988 from the Department of Telecommunications, Radio Technology. Since then, he has been employed at Telenor Research and Development as a research scientist working on several aspects of mobile and personal communications. From 1988 to 1991 he was involved in the standardization of the pan-European radio paging system ERMES within ETSI. Later, the emphasis was on radio propagation aspects for cellular mobile communications. He was involved in the RACE MBS project, on millimeter-wave propagation. He has been a member of COST 259 — "Wireless Flexible Personalized Communications" since 1997. Currently, the main focus of his work is on smart antennas and access network technology for UMTS.

MAGNE PETTERSEN (magne.pettersen@telenor.com) received his M.Sc. from the Norwegian University of Science and Technology (NTNU) in 1994. From 1994 to 1995 he was a research assistant at the Department of Telecommunications, NTNU. In 1995 he joined Telenor Research and Development, Kjeller, where he is currently employed as a research scientist. He is working in the field of personal communications, in particular in the area of antennas and radio wave propagation. He has a special interest in modeling and characterization of the mobile radio channel, an area in which he is currently pursuing a Ph.D. at NTNU.

Hsieh-Yee Declaration

Appendix 0001A

Multiuser OFDM with Adaptive Subcarrier, Bit, and Power Allocation

Cheong Yui Wong, Roger S. Cheng, *Member, IEEE*,
Khaled Ben Letaief, *Senior Member, IEEE*, and Ross D. Murch, *Senior Member, IEEE*

Abstract—Multiuser orthogonal frequency division multiplexing (OFDM) with adaptive multiuser subcarrier allocation and adaptive modulation is considered. Assuming knowledge of the instantaneous channel gains for all users, we propose a multiuser OFDM subcarrier, bit, and power allocation algorithm to minimize the total transmit power. This is done by assigning each user a set of subcarriers and by determining the number of bits and the transmit power level for each subcarrier. We obtain the performance of our proposed algorithm in a multiuser frequency selective fading environment for various time delay spread values and various numbers of users. The results show that our proposed algorithm outperforms multiuser OFDM systems with static time-division multiple access (TDMA) or frequency-division multiple access (FDMA) techniques which employ fixed and predetermined time-slot or subcarrier allocation schemes. We have also quantified the improvement in terms of the overall required transmit power, the bit-error rate (BER), or the area of coverage for a given outage probability.

Index Terms—Adaptive modulation, frequency selective fading channel, multiaccess communication, multiuser channel, orthogonal frequency division multiplexing (OFDM), resource management.

I. INTRODUCTION

RECENTLY, intense interest has focused on modulation techniques which can provide broadband transmission over wireless channels for applications including wireless multimedia, wireless Internet access, and future-generation mobile communication systems. One of the main requirements on the modulation technique is the ability to combat intersymbol interference (ISI), a major problem in wideband transmission over multipath fading channels. There are many methods proposed to combat the ISI, e.g., [1]–[3]. Multicarrier modulation techniques, including orthogonal frequency division multiplexing (OFDM), (e.g., [4]) are among the more promising solutions to this problem.

Assuming that the transmitter knows the instantaneous channel transfer functions of all users, many papers [5]–[7] have demonstrated that significant performance improvement can be achieved if adaptive modulation is used with OFDM. In

particular, subcarriers with large channel gains employ higher order modulation to carry more bits/OFDM symbol, while subcarriers in deep fade carry one or even zero bits/symbol. Integrated design of forward error correcting code and adaptive modulation has also been studied using BCH code and trellis coded modulation (TCM) in [8] and [9], respectively. Although both references considered only time-varying flat fading channels, the same coded adaptive modulation design can be easily applied to OFDM systems. As different subcarriers experience different fades and transmit different numbers of bits, the transmit power levels must be changed accordingly. The problem of optimal power allocation has also been studied in [10].

In this paper, we consider extending OFDM with adaptive modulation to multiuser frequency selective fading environments. When OFDM with adaptive modulation is applied in a frequency selective fading channel, a significant portion of the subcarriers may not be used. These are typically subcarriers which experience deep fade and are not power efficient to carry any information bit. In multiuser systems using static time-division multiple access (TDMA) or frequency-division multiple access (FDMA) as multiaccess schemes, each user is allocated a predetermined time slot or frequency band to apply OFDM with adaptive modulation. Consequently, these unused subcarriers (as a result of adaptive modulation) within the allocated time slot or frequency band of a user are wasted and are not used by other users. However, the subcarriers which appear in deep fade to one user may not be in deep fade for other users. In fact, it is quite unlikely that a subcarrier will be in deep fade for all users, as the fading parameters for different users are mutually independent. This motivates us to consider an adaptive multiuser subcarrier allocation scheme where the subcarriers are assigned to the users based on instantaneous channel information. This approach will allow all the subcarriers to be used more effectively because a subcarrier will be left unused only if it appears to be in deep fade to all users.

We consider a multiuser subcarrier, bit, and power allocation scheme where all users transmit in all the time slots. Our objective is to minimize the overall transmit power by allocating the subcarriers to the users and by determining the number of bits and the power level transmitted on each subcarrier based on the instantaneous fading characteristics of all users. In this paper, we formulate the multiuser subcarrier, bit, and power allocation problem and propose an iterative algorithm to perform the multiuser subcarrier allocation. Once

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The authors are with the Department of Electrical and Electronic Engineering, Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong (e-mail: eeyui@ust.hk; eecheng@ust.hk; eekhaled@ust.hk; andeermurch@ust.hk).

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the subcarrier allocation is determined, the bit and power allocation algorithm can be applied to each user on its allocated subcarriers. We also compare the performance of our proposed solution to various other static subcarrier allocation schemes.

The results of the work can be applied, for instance, to the downlink transmission in a time division duplex (TDD) wireless communication system to improve the downlink capacity. In such a system, the base station (BS) can estimate the instantaneous channel characteristics of all the BS-to-mobile links based on the received uplink transmissions. The multiuser subcarrier, bit, and power allocation can then be used. It is clear that there is a certain amount of transmission overhead as the BS has to inform the mobiles about their allocated subcarriers and the number of bits assigned to each subcarrier.¹ However, this overhead can be relatively small, especially if the channels vary slowly (e.g., in an indoor low mobility environment), and the assignment is done once every many OFDM symbols. To further reduce the overhead, we can assign a contiguous band of subcarriers with similar fading characteristics as a group, instead of assigning each individual subcarrier. In this paper, we will not focus on how the subcarrier allocation information is transmitted. Instead, we will focus on how—and by how much—this new strategy can reduce the required transmit power; or how and by how much this new scheme can improve the bit-error rate (BER) for a fixed transmit power. Alternately, we also consider how and by how much this new scheme can increase the area of coverage for a given transmit power and target BER.

While the bit allocation algorithm can be viewed as a practical implementation of the water-pouring interpretation for achieving the Shannon capacity of an ISI channel [13], the multiuser subcarrier and bit allocation algorithm presented in this paper is the counterpart of the multiuser water-pouring solution given in [14]. In information theoretic studies, the usual approach is to maximize the capacity (or information rate) under the power constraint. In this study, we focus on deriving practical algorithms that can support real-time multimedia data whose bit rates are generally fixed by the compression algorithms. Hence, we assume a given set of user data rates and attempt to minimize the total transmit power under a fixed performance requirement.

The organization of this paper is as follows. In Section II, we will first give the system model and formulate the minimum overall transmit power problem. The optimization problem seeks to minimize the overall transmit power using combined subcarrier, bit, and power allocation schemes for multiuser OFDM systems. The bit and power allocation algorithm for a single-user system is studied in Section III. In Section IV, we derive a lower bound to the minimum overall transmit

power by relaxing some of the constraints in the original problem. We also derive a suboptimal subcarrier allocation algorithm. In Section V, we compare the performance between our proposed method and other static approaches via Monte Carlo simulations. Finally, we conclude in Section VI.

II. SYSTEM MODEL

The configuration of our multiuser adaptive OFDM system is shown in Fig. 1. We assume that the system has K users and the k th user has a data rate equal to R_k bit per OFDM symbol. In the transmitter, the serial data from the K users are fed into the subcarrier and bit allocation block which allocates bits from different users to different subcarriers. We assume that each subcarrier has a bandwidth that is much smaller than the coherence bandwidth of the channel and that the instantaneous channel gains on all the subcarriers of all the users are known to the transmitter. Using the channel information, the transmitter applies the combined subcarrier, bit, and power allocation algorithm to assign different subcarriers to different users and the number of bits/OFDM symbol to be transmitted on each subcarrier. Depending on the number of bits assigned to a subcarrier, the adaptive modulator will use a corresponding modulation scheme, and the transmit power level will be adjusted according to the combined subcarrier, bit, and power allocation algorithm. We define $c_{k,n}$ to be the number of bits of the k th user that are assigned to the n th subcarrier. As we do not allow more than one user to share a subcarrier, it follows that for each n , if $c_{k',n} \neq 0$, $c_{k,n} = 0$ for all $k \neq k'$. We also assume that the adaptive modulator allows $c_{k,n}$ to take values in the set $D = \{0, 1, 2, \dots, M\}$ where M is the maximum number of information bits/OFDM symbol that can be transmitted by each subcarrier.

The complex symbols at the output of the modulators are transformed into the time domain samples by inverse fast Fourier transform (IFFT). Cyclic extension of the time domain samples, known as the guard interval, is then added to ensure orthogonality between the subcarriers, provided that the maximum time dispersion is less than the guard interval. The transmit signal is then passed through different frequency selective fading channels to different users.

We assume that the subcarrier and bit allocation information is sent to the receivers via a separate control channel. At the receiver, the guard interval is removed to eliminate the ISI, and the time samples of the k th user are transformed by the FFT block into modulated symbols. The bit allocation information is used to configure the demodulators while the subcarrier allocation information is used to extract the demodulated bits from the subcarriers assigned to the k th user.

In the frequency selective fading channel, different subcarriers will experience different channel gains. We denote by $\alpha_{k,n}$ the magnitude of the channel gain (assuming coherent reception) of the n th subcarrier as seen by the k th user. We assume that the single-sided noise power spectral density (PSD) level N_0 is equal to unity (i.e., $N_0 = 1$), for all

¹Note that the power level used does not need to be transmitted to the receiver in such a TDD system. As the subcarrier gain is known to the transmitter, it can adjust the transmit power level to achieve a predetermined receiver power level based on the number of bits allocated to that subcarrier. However, in FDD systems, the transmit power levels determined by the receiver have to be sent back to the transmitter. In such systems, the additional performance gain achieved by power allocation may not justify the cost of sending the transmit power level information to the transmitter.

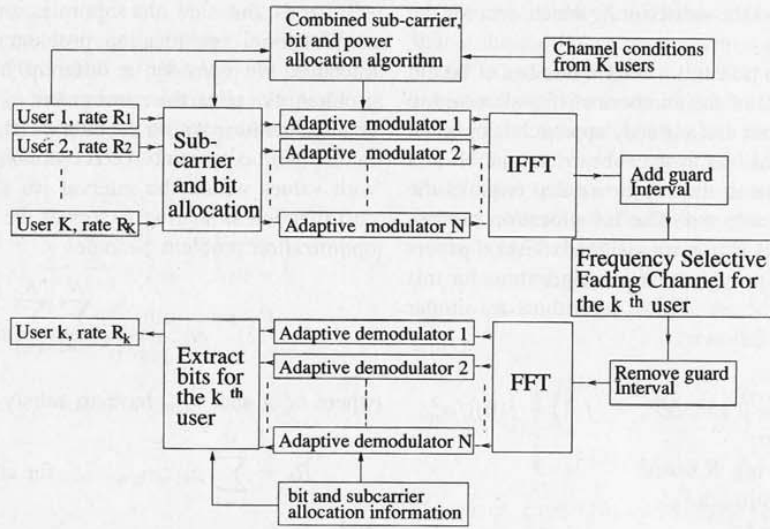


Fig. 1. Block diagram of a multiuser OFDM system with subcarrier, bit, and power allocation.

subcarriers and is the same for all users. Furthermore, we denote by $f_k(c)$ the required received power (in energy per symbol) in a subcarrier for reliable reception of c information bits/symbol when the channel gain is equal to unity. Note that the function $f_k(c)$ depends on k , and this allows different users to have different quality-of-service (QoS) requirements and/or different coding and modulation schemes. In order to maintain the required QoS at the receiver, the transmit power, allocated to the n th subcarrier by the k th user must equal

$$P_{k,n} = \frac{f_k(c_{k,n})}{\alpha_{k,n}^2}. \quad (1)$$

Using these transmit power levels, the receiver can demodulate the modulated symbols at the output of the FFT processor and achieve the desired QoS's of all users.

The goal of the combined subcarrier, bit, and power allocation algorithm is then to find the best assignment of $c_{k,n}$ so that the overall transmit power, the sum of $P_{k,n}$ over all subcarriers and all users, is minimized for given transmission rates of the users and given QoS requirements specified through $f_k()$, $k = 1, \dots, K$. In order to make the problem tractable, we further require that $f_k(c)$ is a convex and increasing function with $f_k(0) = 0$. This condition essentially means that no power is needed when no bit is transmitted and that the required additional power to transmit an additional bit increases with c [i.e., $f_k(c+1) - f_k(c)$ is increasing in c]. Almost all popular coding and modulation schemes satisfy this condition.

It is important to note that even though the problem is formulated to minimize the overall transmit power for given QoS requirements, the same solution can be applied to improve the QoS's of the users for a given overall transmit power. The latter can simply be achieved by increasing the power proportionally for all the subcarriers, while using the same set of $c_{k,n}$.

Mathematically, we can formulate the problem as

$$P_T^* = \min_{c_{k,n} \in \mathbf{D}} \sum_{n=1}^N \sum_{k=1}^K \frac{1}{\alpha_{k,n}^2} f_k(c_{k,n}) \quad (2)$$

and the minimization is subjected to the constraints

$$\text{C1: For all } k \in \{1, \dots, K\}, R_k = \sum_{n=1}^N c_{k,n} \quad (3)$$

and

$$\begin{aligned} \text{C2: For all } n \in \{1, \dots, N\}, \\ \text{if there exists } k' \text{ with } c_{k',n} \neq 0, \text{ then } c_{k,n} = 0, \\ \forall k \neq k'. \end{aligned} \quad (4)$$

Note that constraint (3) is the data rate requirement and constraint (4) ensures that each subcarrier can only be used by one user. Moreover, $\mathbf{D} = \{0, 1, 2, \dots, M\}$ is the set of all possible values for $c_{k,n}$, and $c_{k,n} = 0$ means that the k th user does not use the n th subcarrier to transmit any information.

III. BIT ALLOCATION ALGORITHM FOR SINGLE USER CHANNEL

Before we try to solve the multiuser allocation problem, we will first derive the bit allocation algorithm for the single-user environment. The single-user problem not only gives better understanding of the issues involved, but also provides a bit allocation algorithm that we will use in our multiuser solution.

We can rewrite the optimization problem in (2) for the single-user case as

$$P_T^* = \min_{c_n \in \mathbf{D}} \sum_{n=1}^N \frac{1}{\alpha_n^2} f(c_n) \quad (5)$$

and the minimization is under the constraint

$$R = \sum_{n=1}^N c_n. \quad (6)$$

Note that we have dropped the subscript k , which denotes the user in all notations.

As the power needed to transmit a certain number of bits in a subcarrier is independent of the numbers of bits allocated to other subcarriers, it turns out that a greedy approach is optimal. A greedy algorithm assigns bits to the subcarriers one bit at a time, and in each assignment, the subcarrier that requires the least additional power is selected. The bit allocation process will be completed when all R bits are assigned. Several papers (e.g., [15] and [16]) have provided various algorithms for this problem, and the basic structure of most algorithms are similar and can be described as follows:

Initialization:

For all n , let $c_n = 0$ and $\Delta P_n = [f(1) - f(0)]/\alpha_n^2$;

Bit Assignment Iterations:

Repeat the following R times:

$\hat{n} = \arg \min_n \Delta P_n$;

$c_{\hat{n}} = c_{\hat{n}} + 1$;

$\Delta P_{\hat{n}} = [f(c_{\hat{n}} + 1) - f(c_{\hat{n}})]/\alpha_{\hat{n}}^2$;

End;

Finish:

$\{c_n\}_{n=1}^N$ is the final bit allocation solution.

The initialization stage computes, for each subcarrier, the additional power needed to transmit an additional bit. For each bit assignment iteration, the subcarrier that needs the minimum additional power is assigned one more bit, and the new additional power for that subcarrier is updated. After R iterations, the final bit assignment gives the optimal bit allocation for each subcarrier. It is important to note that the bit allocation is optimal only for the given function $f(c)$, which depends on the selected modulation scheme. Different modulation schemes will lead to different $f(c)$, different bit allocation, and possibly lower transmit power P_T^* .

The concept of this algorithm is fairly simple, and many similar algorithms based on the same principle have been obtained before. In particular, there exist faster and less complex algorithms which can speed up the bit allocation process significantly (e.g., [15] and [16]). In our simulations, we use the algorithm given in [16].

IV. MULTIUSER SUBCARRIER AND BIT ALLOCATION

We have observed that, in the single-user case, a greedy approach which assigns one bit at a time to the subcarrier that requires the least additional power gives the optimal allocation in the sense of minimizing the overall transmit power. Unfortunately, the problem becomes more difficult in the multiuser environment. As users cannot share the same subcarrier, allocating bits to a subcarrier essentially prevents other users from using that subcarrier. This dependency makes any greedy algorithm a nonoptimal solution. It turns out that the optimal solution may not assign any of a user's bits to the best subcarrier seen by that user. This may happen when the best subcarrier of a user is also the best subcarrier of another user who happens to have no other good subcarriers. Hence, the multiuser subcarrier and bit allocation problem is much more complicated to solve than that of the single-user case.

It turns out that the optimization problem in (2) is a combinatorial optimization problem. To make the problem tractable, we consider a different but similar optimization problem. We relax the requirement $c_{k,n} \in \mathbf{D}$ to allow $c_{k,n}$ to be a real number within the interval $[0, M]$. Moreover, in order to deal with constraint (4), K variables, $\rho_{k,n}$, $k = 1, \dots, K$, with values within the interval $[0, 1]$, are introduced to the cost function as sharing factors of the n th subcarrier. The new optimization problem becomes

$$P_T = \min_{\substack{c_{k,n} \in [0, M] \\ \rho_{k,n} \in [0, 1]}} \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k(c_{k,n}) \quad (7)$$

where $c_{k,n}$ and $\rho_{k,n}$ have to satisfy

$$R_k = \sum_{n=1}^N \rho_{k,n} c_{k,n}, \quad \text{for all } k \in \{1, \dots, K\} \quad (8)$$

and

$$1 = \sum_{k=1}^K \rho_{k,n}, \quad \text{for all } n \in \{1, \dots, N\}. \quad (9)$$

For any valid set of $c_{k,n} \in \mathbf{D}$ satisfying the constraints (3) and (4) in the original optimization problem, we can let

$$\rho_{k,n} = \begin{cases} 1, & \text{if } c_{k,n} \neq 0, \\ 0, & \text{if } c_{k,n} = 0. \end{cases} \quad (10)$$

Then, it is easy to show that the same set of $c_{k,n}$ and the corresponding $\rho_{k,n}$ defined in (10) satisfy the constraints (8) and (9) in the new optimization problem. Moreover, with $\rho_{k,n}$ defined in (10), the new cost function in (7) is equal to the cost function in (2). Hence, the minimization problem in (7) is the same as the original optimization problem, except that the minimization is done over a larger set. Consequently, the minimum power obtained in (7) P_T is a lower bound to the minimum power obtained in (2), P_T^* .

Another way to interpret the optimization in (7) is to consider $\rho_{k,n}$ as the time-sharing factor for the k th user of the n th subcarrier. For example, in every L OFDM symbol (L being a very large number), user k uses the n th subcarrier in $L\rho_{k,n}$ symbols. Clearly, the average (over L symbols) information data rate and the average transmit power has to be scaled by the same factor $\rho_{k,n}$. Hence, we can consider (7) as the optimization problem when the users are allowed to time-share each subcarrier over a large number of OFDM symbols. However, most wireless communication channels are time varying, and the channels may not stay unchanged long enough for timesharing to be feasible. Hence, in this paper, we will continue to consider the original problem in (2) and use the optimization problem in (7) as a lower bound, even though it has its own physical interpretation.

The modified optimization problem in (7) is more tractable. However, even though the function $f_k(c)$ is convex in c , the terms in the cost function have the form $\rho f_k(c)$, and as a function of (ρ, c) , $\rho f_k(c)$ is not convex in (ρ, c) . To proceed further, we let $r_{k,n} = c_{k,n} \rho_{k,n}$ and rewrite the cost function in terms of $r_{k,n}$ and $\rho_{k,n}$. The constraint on $r_{k,n}$

becomes $r_{k,n} \in [0, M\rho_{k,n}]$, and it can be easily shown that $\rho f_k(c) = \rho f_k(r/\rho)$ is convex in (ρ, r) within the triangular region specified by $\rho \in [0, 1]$ and $r \in [0, M\rho]$. In particular, the Hessian evaluated at any point within this region is a positive semidefinite matrix. Hence, we can reformulate the optimization problem in (7) as a convex minimization problem over a convex set. That is

$$P_T = \min_{\substack{r_{k,n} \in [0, M\rho_{k,n}] \\ \rho_{k,n} \in [0, 1]}} \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k\left(\frac{r_{k,n}}{\rho_{k,n}}\right) \quad (11)$$

where $r_{k,n}$ and $\rho_{k,n}$ have to satisfy

$$R_k = \sum_{n=1}^N r_{k,n}, \quad \text{for all } k \in \{1, \dots, K\} \quad (12)$$

and

$$1 = \sum_{k=1}^K \rho_{k,n}, \quad \text{for all } n \in \{1, \dots, N\}. \quad (13)$$

Using standard optimization techniques in [17], we obtain the Lagrangian

$$L = \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k\left(\frac{r_{k,n}}{\rho_{k,n}}\right) - \sum_{k=1}^K \lambda_k \left(\sum_{n=1}^N r_{k,n} - R_k \right) - \sum_{n=1}^N \beta_n \left(\sum_{k=1}^K \rho_{k,n} - 1 \right) \quad (14)$$

where λ_k and β_n are the Lagrangian multipliers for the constraints (12) and (13), respectively.

After differentiating L with respect to $r_{k,n}$ and $\rho_{k,n}$, respectively, we obtain the necessary conditions for the optimal solution, $r_{k,n}^*$ and $\rho_{k,n}^*$. Specifically, if $\rho_{k,n}^* \neq 0$, we have

$$\begin{aligned} \frac{\partial L}{\partial r_{k,n}} \bigg|_{(r_{k,n}, \rho_{k,n})=(r_{k,n}^*, \rho_{k,n}^*)} &= \frac{1}{\alpha_{k,n}^2} f'_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) - \lambda_k \begin{cases} > 0, & \text{if } r_{k,n}^* = 0 \\ = 0, & \text{if } r_{k,n}^* \in (0, M\rho_{k,n}^*) \\ < 0, & \text{if } r_{k,n}^* = M\rho_{k,n}^* \end{cases} \end{aligned} \quad (15)$$

and

$$\begin{aligned} \frac{\partial L}{\partial \rho_{k,n}} \bigg|_{(r_{k,n}, \rho_{k,n})=(r_{k,n}^*, \rho_{k,n}^*)} &= \frac{1}{\alpha_{k,n}^2} \left[f_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) - f'_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) \frac{r_{k,n}^*}{\rho_{k,n}^*} \right] \\ &\quad - \beta_n \begin{cases} = 0, & \text{if } \rho_{k,n}^* \in (0, 1) \\ < 0, & \text{if } \rho_{k,n}^* = 1. \end{cases} \end{aligned} \quad (16)$$

On the other hand, if $\rho_{k,n}^* = 0$, then $r_{k,n}^* = 0$, and we have

$$\begin{aligned} r_{k,n} \frac{\partial L}{\partial r_{k,n}} + \rho_{k,n} \frac{\partial L}{\partial \rho_{k,n}} &\geq 0, \\ \text{for all } \rho_{k,n} \in (0, 1] \text{ and } r_{k,n} \in (0, M\rho_{k,n}]. \end{aligned} \quad (17)$$

These necessary conditions can be interpreted by the fact that if the minimum occurs within the constrained region $[(0, 1)$ for $\rho_{k,n}$ and $(0, M\rho_{k,n})$ for $r_{k,n}$], then the derivative evaluated at the minimum point must be zero. On the other hand, if the optimal solution occurs at a boundary point, then the derivative must be positive along all directions pointing toward the interior of the constraint set. Then, (17) follows from considering the boundary point at $(r_{k,n}^*, \rho_{k,n}^*) = (0, 0)$.

From (15) and (17), we can conclude that

$$r_{k,n}^* = \rho_{k,n}^* f_k'^{-1}(\lambda_k \alpha_{k,n}^2) \quad (18)$$

where

$$\lambda_{q,k} = \begin{cases} f'_k(0)/\alpha_{k,n}^2, & \text{if } f_k'^{-1}(\lambda_k \alpha_{k,n}^2) < 0; \\ \lambda_k, & \text{if } 0 \leq f_k'^{-1}(\lambda_k \alpha_{k,n}^2) \leq M; \\ f'_k(M)/\alpha_{k,n}^2, & \text{if } f_k'^{-1}(\lambda_k \alpha_{k,n}^2) > M. \end{cases}$$

Moreover, from (16) and (17), it follows that

$$\rho_{k,n}^* = \begin{cases} 0, & \text{if } \beta_n < H_{k,n}(\lambda_{q,k}) \\ 1, & \text{if } \beta_n > H_{k,n}(\lambda_{q,k}) \end{cases} \quad (19)$$

where

$$H_{k,n}(\lambda) = \frac{1}{\alpha_{k,n}^2} [f_k(f_k'^{-1}(\lambda \alpha_{k,n}^2)) - \lambda \alpha_{k,n}^2 f_k'^{-1}(\lambda \alpha_{k,n}^2)]. \quad (20)$$

Since constraint (13) must be satisfied, we find from (19) that for each n , if $H_{k,n}(\lambda_{q,k})$ for $k = 1, \dots, K$ are all different, then only the user with the smallest $H_{k,n}(\lambda_{q,k})$ can use that subcarrier. In other words, for the n th subcarrier, if $H_{k,n}(\lambda_{q,k})$ are different for all k , then

$$\rho_{k',n}^* = 1, \quad \rho_{k,n}^* = 0, \quad \text{for all } k \neq k' \quad (21)$$

where

$$k' = \arg \min_k H_{k,n}(\lambda_{q,k}). \quad (22)$$

Hence, it follows that for a fixed set of Lagrange multipliers λ_k , $k = 1, \dots, K$, we can use them to determine k' for each n using (22). The $r_{k,n}^*$ and $\rho_{k,n}^*$ obtained will then form an optimal solution for the optimization problem; however, the individual rate constraint (12) may not be satisfied.

In order to find the set of λ_k such that the individual rate constraints are satisfied, we have obtained an iterative searching algorithm. Starting with some small values for all λ_k , this iterative procedure increases one of the λ_k until the data rate constraint (12) for user k is satisfied. Then, we switch to another user and go through the users one at a time. This process repeats for all users until the data rate constraint for all users are satisfied. This algorithm converges because for a given k , as λ_k increases, $H_{k,n}(\lambda_{q,k})$ for all n decreases, and more $\rho_{k,n}^*$ in (19) become one while $r_{k,n}^*$ in (18) increases for those n where $\rho_{k,n}^* > 0$. Hence, $\sum_{n=1}^N r_{k,n}^*$ increases. During this process, some of the other $\rho_{k',n}^*$ may change from one to zero and consequently decrease the total data rate for other users. However, as all the λ_k increase, $r_{k,n}^*$ increases accordingly. As long as the total data rate is less than MN bits/symbol, which is the total number of bits possibly transmitted within an OFDM symbol, the algorithm

will converge to a solution that satisfies all the constraints. Since the optimization problem is a convex optimization problem over a convex set, the set of necessary conditions is also sufficient, and the solution that satisfies all the necessary conditions is the unique optimal solution.

In the process of adjusting λ_k for $k = 1, \dots, K$, the situation where, for a fixed n , more than one $H_{k,n}(\lambda_{q,k})$ has the same values cannot be ignored. In that case, $\rho_{k,n}^*$ has to take values within the interval $(0, 1)$. This solution suggests that the subcarrier should be shared by multiple users. In practice, this can be done by having these users with $\rho_{k,n}^* > 0$ time share the n th subcarrier, and the ratio of the symbols used by different users are set proportionally to $\rho_{k,n}^*$. The detailed flow chart of the algorithm is given in the Appendix.

Now, we have an algorithm to obtain the optimal values of $\rho_{k,n}^*$ and

$$c_{k,n}^* = \begin{cases} r_{k,n}^*/\rho_{k,n}^*, & \text{if } \rho_{k,n}^* \neq 0 \\ 0, & \text{otherwise.} \end{cases} \quad (23)$$

This solution, when substituted in (7), gives a lower bound to the minimum overall transmit power. However, we cannot use these results immediately in (2). One problem is that $c_{k,n}^*$ may not be in $\mathbf{D}$, and the other is that some $\rho_{k,n}^*$ may be within $(0, 1)$, indicating a time-sharing solution. Furthermore, simply quantizing $c_{k,n}^*$ and $\rho_{k,n}^*$ will not satisfy the individual rate constraints in (3).

To solve this problem, we propose a multiuser adaptive OFDM (MAO) scheme where the subcarrier allocation follows essentially the solution to the lower bound in (7), and then the single-user bit allocation algorithm given in Section III is applied to each user on the allocated subcarriers. Specifically, we modify $\rho_{k,n}^*$ for the optimization problem in (7) by letting for each n $\rho_{k',n}^* = 1$ where $k' = \arg \max_k \rho_{k,n}^*$, and $\rho_{k,n}^* = 0$ for $k \neq k'$. Then, we apply the single-user bit allocation algorithm on each user using the assigned subcarriers. We denote the total transmit power (in energy/symbol) obtained using this MAO scheme by P_T . It is easy to see that $\underline{P}_T \leq P_T^* \leq P_T$, where P_T^* is the minimum power in the original problem, and $\underline{P}_T$ is the minimum power for the modified problem with the relaxed constraints. More specifically, the difference between P_T and the minimum $\underline{P}_T$ gives an upper bound to how far away our MAO scheme is from the solution of our original optimization problem.

V. PERFORMANCE COMPARISON

In this section, we obtain and compare the performance of the MAO scheme with other static subcarrier allocation schemes. We consider a system that employs M-ary quadrature amplitude modulation (MQAM) with $\mathbf{D} = \{0, 2, 4, 6\}$. Square signal constellations (4-QAM, 16-QAM, and 64-QAM) are used to carry two, four, or six bits/symbol. The bit-error probability is upper bounded by the symbol error probability, which is tightly approximated by $4Q[\sqrt{d^2}/(2N_0)]$ [12, p. 281], where d is the minimum distance between the points in the signal constellation. Since the average energy of a M-QAM symbol is equal to $(M-1)d^2/6$, it follows that the required power for supporting c bits/symbol at a given BER

P_e is

$$f(c) = \frac{N_0}{3} \left[Q^{-1} \left(\frac{P_e}{4} \right) \right]^2 (2^c - 1)$$

where we recall that

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt.$$

It is easy to see that $f(c)$ is convex and increasing in c and that $f(0) = 0$.

To evaluate the performance of our scheme, we have simulated 1000 sets of five-path frequency selective Rayleigh fading channels with an exponential power delay profile. Each set of channels consists of K independent channels, one for each user. We use an OFDM system with 128 subcarriers over a 5 MHz band along with a total (over all users) transmission rate equal to 512 bits/symbol (or equivalently, an average of four bits/subcarrier). Recall that the single-sided power spectral density level N_0 is equal to unity, and we assume that the average subcarrier channel gain $\mathbb{E}|\alpha_{k,n}|^2$ is equal to unity for all k and n .

For comparison purposes, we have also considered three other static multiuser subcarrier allocation methods. Two of them are based on the multiple access methods described in [7]. The methods are presented as follows.

- OFDM-TDMA: each user is assigned a predetermined TDMA time slot and can use all the subcarriers within that time slot exclusively.
- OFDM-FDMA: each user is assigned a predetermined band of subcarriers and can only use those subcarriers exclusively in every OFDM symbol.
- OFDM Interleaved-FDMA: this is the same as OFDM-FDMA except that subcarriers assigned to a user are interlaced with other users' subcarriers in the frequency domain.

The time and subcarrier assignment of these three multiuser OFDM schemes are illustrated in Fig. 2. Note that these static schemes have predetermined subcarrier allocations which are independent of the channel gains of the users. The main difference between the proposed MAO scheme and these static schemes is that MAO assigns subcarriers adaptively based on the instantaneous channel gains. To ensure a fair comparison, we use the optimal single-user bit allocation (OBA) for each user on the assigned subcarriers. For comparison purposes, we also show the results when equal bit allocation (EBA) is employed on the assigned subcarriers for these three OFDM schemes. Notice that when using EBA, all three schemes will have the same performance in an uncoded system. This is because the average bit signal-to-noise ratio (SNR) needed is a function of only the marginal probability density function of each subcarrier gain.

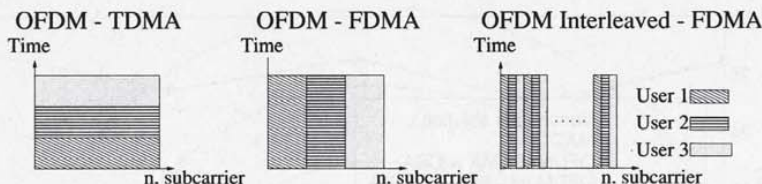


Fig. 2. Subcarrier and time-slot allocations of OFDM-TDMA, OFDM-FDMA, and OFDM interleaved-FDMA schemes.

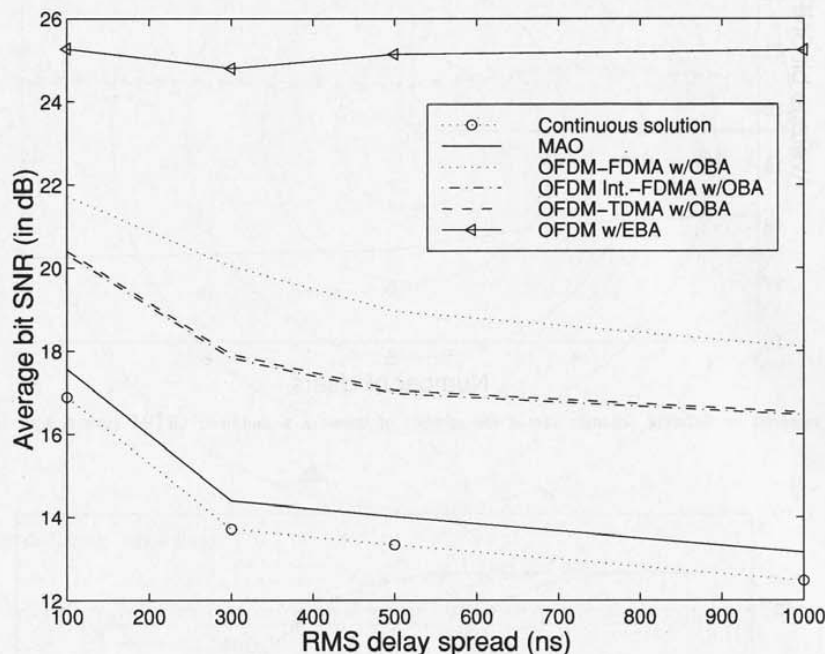


Fig. 3. Average bit signal-to-noise ratio (SNR) required by different schemes in various root mean square (RMS) delay spreads in a five-user system with $P_e = 10^{-4}$.

Fig. 3 shows the average bit SNR needed to achieve a BER at $P_e = 10^{-4}$ for a five-user system versus the root mean square (RMS) delay spread (for definition, see for example [18, p. 160]) for different multiuser OFDM schemes. The average required transmit power (in energy per bit) is defined as the ratio of the overall transmit energy per OFDM symbol (including all subcarriers and all users) to the total number of bits transmitted per OFDM symbol. Moreover, we define the average bit SNR as the ratio of the average transmit power to the noise PSD level N_0 . As we assume that the data rate is fixed and that N_0 is just a constant, the overall transmit power is proportional to the average bit SNR. For ease of comparison, we have used the average bit SNR for comparison. We find in Fig. 3 that the MAO scheme is never more than 0.6 dB from the lower bound. Since the bit SNR of the optimal combined subcarrier, bit, and power allocation algorithm must lie between the bit SNR's achieved by the lower bound and the MAO scheme, we find that the MAO scheme is never more than 0.6 dB away from the optimal solution. On the other hand, we observe that our proposed MAO scheme is 3–5 dB better than the static subcarrier allocation schemes with OBA, which are in turn 5–10 dB better than that with EBA. We also find that when OBA is used, the OFDM interleaved-FDMA

scheme and the OFDM-TDMA scheme have very similar performance, and both of them outperform the OFDM-FDMA scheme.² A closer observation of Fig. 3 also indicates that the gains achieved by optimal bit allocation and optimal multiuser subcarrier allocation increase with the RMS delay spread. This is mainly because the larger the RMS delay spread, the more the fading variation and hence higher gains can be obtained when the allocation is performed adaptively.

Fig. 4 shows the average bit SNR (in dB) needed to achieve the same BER versus the number of users when the RMS delay spread is 100 ns. We find that the savings in the required bit SNR achieved by MAO when compared to other schemes are roughly the same, independent of the number of users in the system.

While these two figures show the improvement in the required bit SNR, the results can perhaps be more easily understood using the more familiar BER versus bit SNR curves. For each BER requirement, we compute $f(c)$ for all $c \in \mathcal{D}$ and then use our algorithm to calculate the subcarrier

²OFDM-FDMA refers only to the specific FDMA scheme that assigns to each user a contiguous band of subcarriers as shown in Fig. 2, but not the general FDMA schemes. In fact, both OFDM interleaved-FDMA and MAO can be considered as different forms of FDMA and they are not outperformed by the OFDM-TDMA scheme.

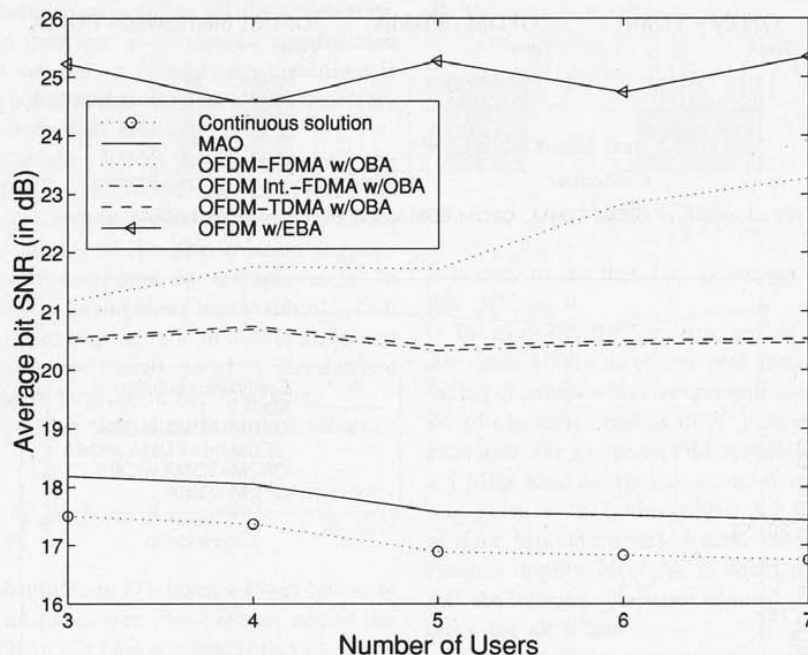


Fig. 4. Average bit SNR required by different schemes versus the number of users in a multiuser OFDM system with 100 ns RMS delay spread, and $P_e = 10^{-4}$.

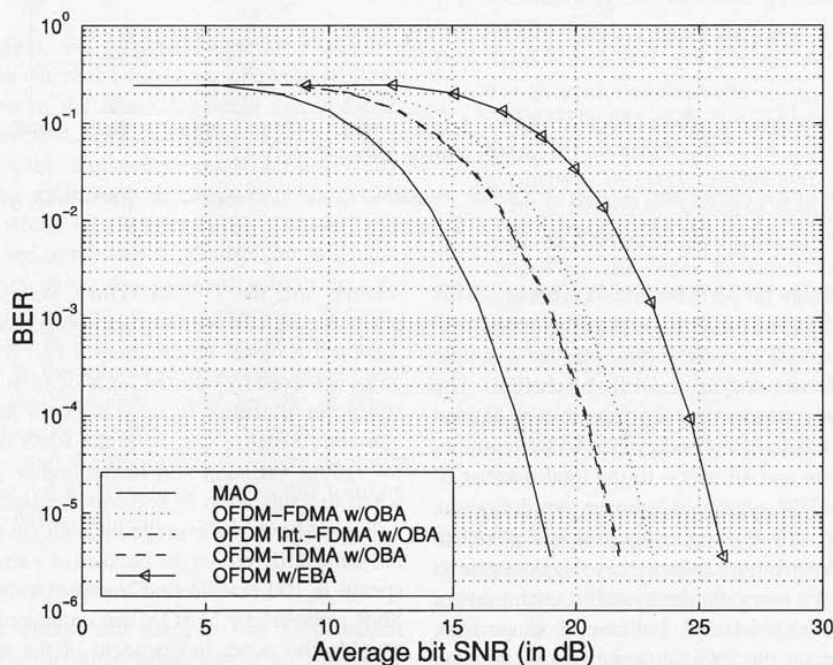


Fig. 5. BER versus average bit SNR for various subcarrier allocation schemes.

allocation for the MAO case. For all other static subcarrier allocation schemes, the allocations are independent of the BER. Once the subcarrier allocation is fixed, we apply the optimal bit and power allocation algorithm to every user. The final average power per bit divided by the noise power spectral density level gives the average bit SNR. We repeat this procedure for different BER values, and the results are

plotted in Fig. 5 for a five-user system with an RMS delay spread equal to 100 ns. We find that our proposed MAO has at least 3–4 dB advantage over all other schemes.

Another way to illustrate the impact of the bit and subcarrier allocation is to consider the area of coverage for a given outage probability, assuming that the BS has a maximum transmit power. We consider a circular cell with five users, indepen-

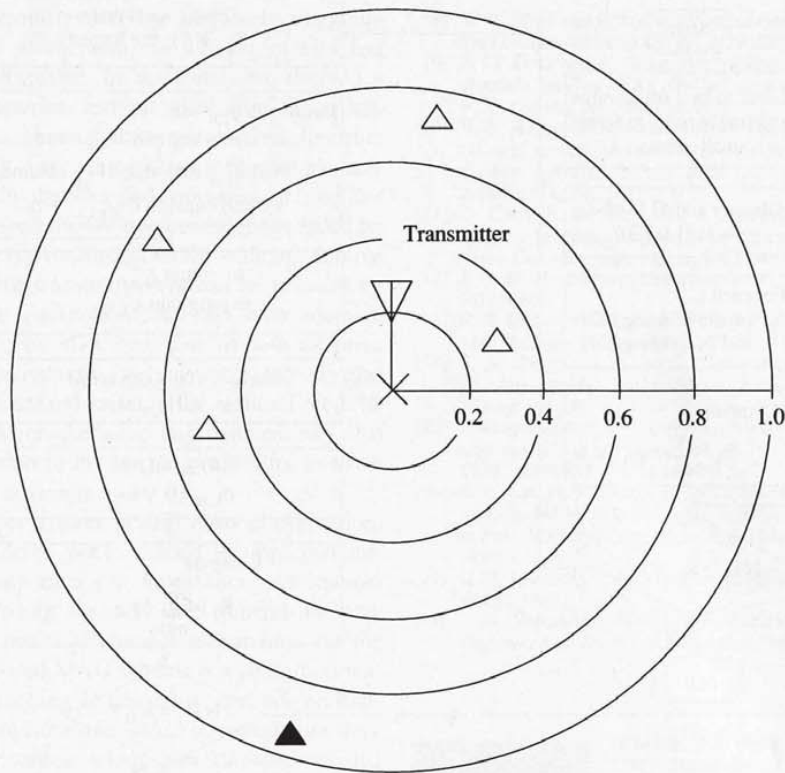


Fig. 6. Cell for analyzing the outage probability.

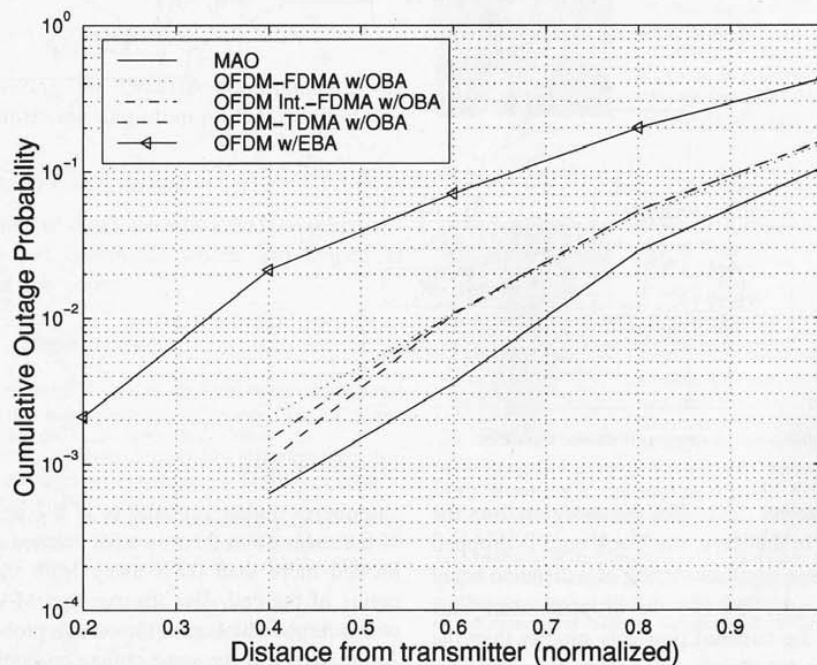


Fig. 7. Outage probability at 17 dB fading margin.

dently and uniformly distributed within the cell. A typical scenario is shown in Fig. 6, where the triangles represent the five users. In addition to frequency selective fading, path loss and log-normal shadowing are also included in simulating the

actual channel gains seen by the users. Using these channel gains, subcarriers and bits assigned to each user are determined by the various multiple access schemes and the total required transmit power is calculated. If the total power for all five users

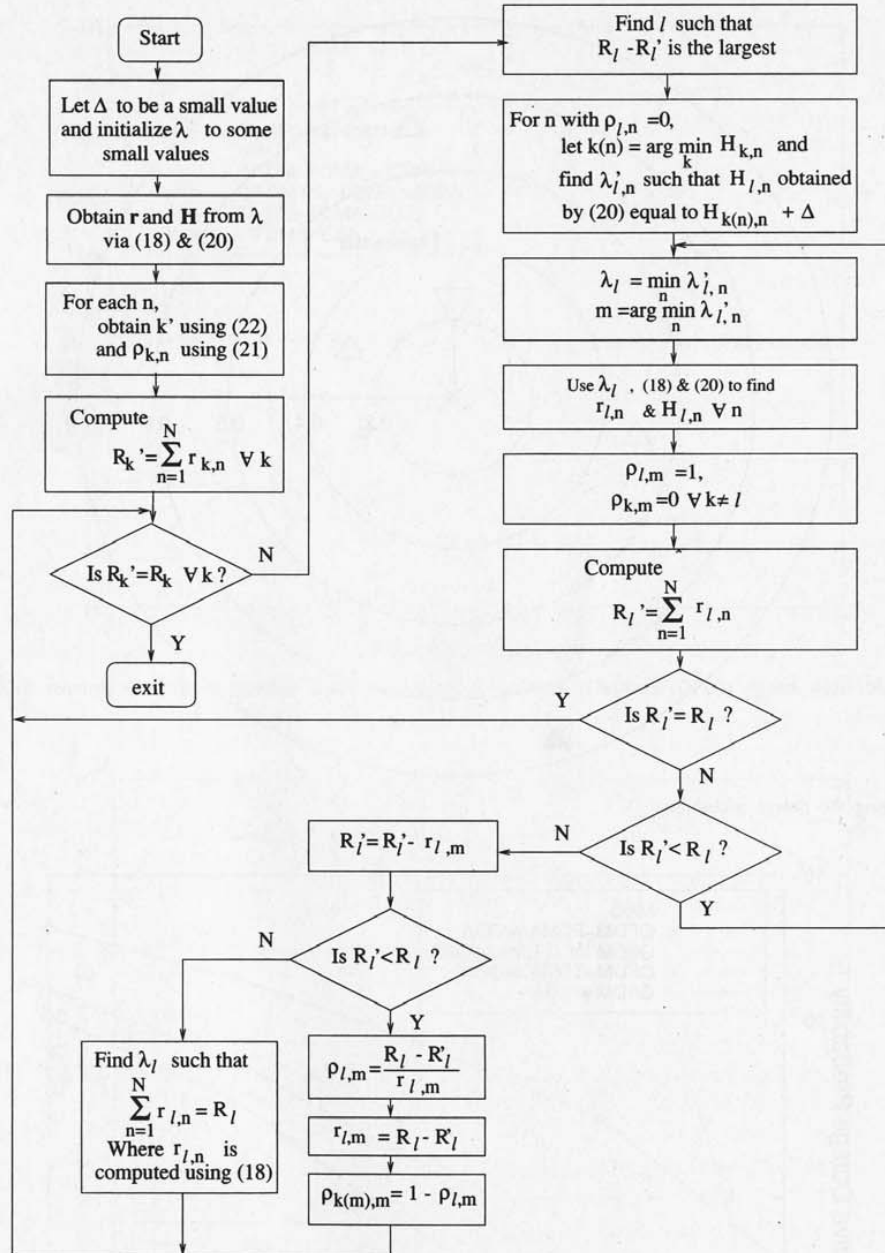


Fig. 8. Flow chart of the multiuser subcarrier allocation algorithm.

exceeds the maximum power of the BS, the user requiring the largest transmit power (in this case, the black one) is dropped and counted as one outage event occurring at a distance equal to the distance between the BS and the dropped user. This process continues until the transmit power is smaller than the maximum power of the BS. In this example, the maximum transmit power is set to the transmit power required for all five users assuming that they are all located at the boundary of the cell, taking into account the path loss effect and a 17 dB fading margin for shadowing.

The cumulative outage probabilities at various normalized distances, normalized to the cell radius, are plotted in Fig. 7. A

cumulative outage probability of 5% at a normalized distance of 0.8 means that there is a 5% chance of outage for a mobile located more than $0.8R$ away from the BS where R is the radius of the cell. We observe that MAO outperforms others with a large reduction in the outage probability at all distances. Alternatively, if the same outage probability is maintained, say at 1%, the coverage are provided by MAO is 36% larger than the best of all other schemes.

VI. CONCLUSION

In this paper, we considered OFDM transmission in a multiuser environment and formulated the problem of min-

imizing the overall transmit power by adaptively assigning subcarriers to the users along with the number of bits and power level to each subcarrier. In particular, we derived a multiuser adaptive subcarrier and bit allocation algorithm. Given the instantaneous channel information, the algorithm obtains a suboptimal subcarrier allocation, and then single-user bit allocation is applied on the allocated subcarriers. Using this scheme, the overall required transmit power can be reduced by about 5–10 dB from the conventional OFDM without adaptive modulation. Likewise, the transmit power can be reduced by about 3–5 dB from the conventional OFDM with adaptive modulation and adaptive bit allocation, but without adaptive subcarrier allocation. The reduction in transmit power can also be translated to a significant reduction in the required bit SNR for a given BER. Moreover, the same improvement can also be translated to a reduction in the outage probability or to an increase in the area of coverage.

The results in this paper assume perfect channel estimation, and we have not considered issues related to imperfect implementation, such as imperfect synchronization. As channel estimation in wireless fading channels is in general not very accurate, the effect of nonideal channel information on the performance of our proposed MAO scheme is a very important issue. We have started looking at this issue, and our preliminary results have indicated that the MAO scheme is not very sensitive to channel estimation errors. Nevertheless, detailed sensitivity studies will be needed before the algorithm can be applied to practical systems.

APPENDIX

A flow chart providing the detailed description of the multiuser subcarrier allocation algorithm is shown in Fig. 8.

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Cheong Yui Wong was born in Hong Kong, in 1975. He received the B.Eng. degree in electronic engineering from the Hong Kong University of Science and Technology (HKUST), Clear Water Bay, Hong Kong in 1997. He is currently pursuing the M.Phil. degree at HKUST and expects to graduate in 1999.

During his master studies, he joined a project that aims at achieving wireless multimedia communication using OFDM. His research interests include adaptive modulation, OFDM, and optimization.



Roger S. Cheng (S'85–M'91) received the B.S. degree from Drexel University, Philadelphia, PA, in 1987, and both the M.A. and Ph.D. degrees from Princeton University, Princeton, NJ, in 1988 and 1991, respectively, all in electrical engineering.

From 1987 to 1991, he was a Research Assistant in the Department of Electrical Engineering, Princeton University. From 1991 to 1995, he was an Assistant Professor in the Electrical and Computer Engineering Department of University of Colorado, Boulder. In June 1995, he joined the Faculty of Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong, where he is currently an Assistant Professor in the Department of Electrical and Electronic Engineering. He has also held visiting positions with Qualcomm, San Diego, in the summer of 1995 and with Institute for Telecommunication Sciences, NTIA, Boulder, CO, in the summers of 1993 and 1994. His current research interests include wireless communications, multiuser communications, spread spectrum CDMA, digital implementation of communication systems, wireless multimedia communications, information theory, and coding.

Dr. Cheng was the recipient of the Meitac Junior Fellowship Award from the Meitac Corporation in Japan, the George Van Ness Lothrop Fellowship from the School of Engineering and Applied Science in Princeton University, and the Research Initiation Award from the National Science Foundation. He was a member of the Technical Program Committees of GLOBECOM'98 and IEEE Wireless Communications and Networking Conference (WCNC'99). He has served as Guest Editor of the special issue on Multimedia Network Radios in the IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, as Associate Editor of the IEEE TRANSACTIONS ON SIGNAL PROCESSING, and as Membership Chair for of the IEEE Information Theory Society.



Khaled Ben Letaief (S'85-M'86-SM'97) received the B.S. degree (with distinction), the M.S. degree, and the Ph.D. degrees in electrical engineering from Purdue University, West Lafayette, IN, in 1984, 1986, and 1990, respectively.

In 1985, he was a Graduate Instructor in the School of Electrical Engineering at Purdue University and he taught courses in communications and electronics. From 1990 to 1993, he was on the Academic Staff at the Department of Electrical and Electronic Engineering, University of Melbourne,

Australia, where he was also a Member of the Center for Sensor Signal and Information Systems. Since 1993, he has been with the Department of Electrical and Electronic Engineering, Hong Kong University of Science and Technology (HKUST), Clear Water Bay, Hong Kong, where is currently an Associate Professor. During several semesters from 1995 to 1998, he served as the Departmental Director of the Undergraduate Studies program. His research interests include wireless personal and mobile communications, spread spectrum systems, optical fiber networks, multiuser detection, wireless multimedia communications, and CDMA systems. He is an Editor of the *Wireless Personal Communications Journal* and is also a Guest Editor of the 1999 *Wireless Personal Communications Journal's* Special Issue on Intelligent Multimedia Systems, Terminals, and Components.

Dr. Letaief has been an active member of various professional societies and has published papers in several journals and conference proceedings. In 1990, he was awarded the Mangoon Teaching Award at Purdue University. From 1995 to 1997, he was awarded the Teaching Excellence Appreciation Award by the School of Engineering, HKUST. He was also the 1998 University Recipient of the Michael G. Gale Medal for Distinguished Teaching (the highest teaching award at the university). He is the Editor for Wireless Systems of the IEEE TRANSACTIONS ON COMMUNICATIONS and a Technical Editor of the *IEEE Communications Magazine*. He has been a Guest Editor of the 1997 IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS' Special Issue on Computer-Aided Modeling, Analysis, and Design of Communications Links. In addition, he served as the Technical Program Chair of the 1998 IEEE Mini-Conference on Communications Theory (CTMC'98) which was held in Sydney, Australia. He is currently serving as an Officer in the IEEE Communications Society Technical Committee on Personal Communications.



Ross D. Murch (S'85-M'87-SM'98) received the Ph.D. degree in electrical and electronic engineering from the University of Canterbury in 1990 for research involving electromagnetic inverse scattering.

From 1990 to 1992, he was a Post Doctoral Fellow at the Department of Mathematics and Computer Science, Dundee University, Scotland, where he worked on approximate approaches to electromagnetic scattering from three-dimensional objects. Currently he is an Associate Professor

in the Department of Electrical and Electronic Engineering at the Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong. He is also the founding Director of the Center for Wireless Information Technology, which began in August 1997. From August to December 1998, he was on sabbatical leave at Allgon Mobile Communications, Sweden and AT&T Research Labs, NJ. He also acts as a Consultant for various industrial projects on wireless communications and has two U.S. patents related to wireless communication. His most recent research activities center around the application of electromagnetics and signal processing to wireless communications. Research interests include: antenna design for mobile telephone handsets, smart antenna algorithms and propagation prediction for wireless communications.

Dr. Murch is an Editor of IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, Wireless Series, and he has been involved in the organization of several international conferences. He is a Chartered Engineer, a Member of the Institute of Electrical Engineers, and also an URSI correspondent. He won an URSI Young Scientist Award in 1993 and an Engineering Teaching Excellence Appreciation award in 1996.



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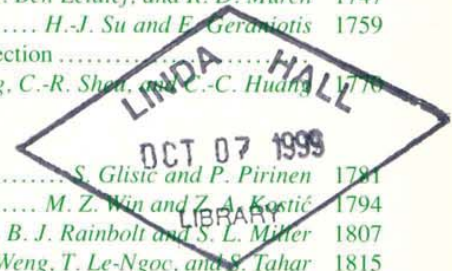
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Multiuser OFDM with Adaptive Subcarrier, Bit, and Power Allocation

Cheong Yui Wong, Roger S. Cheng, *Member, IEEE*,
Khaled Ben Letaief, *Senior Member, IEEE*, and Ross D. Murch, *Senior Member, IEEE*

Abstract—Multiuser orthogonal frequency division multiplexing (OFDM) with adaptive multiuser subcarrier allocation and adaptive modulation is considered. Assuming knowledge of the instantaneous channel gains for all users, we propose a multiuser OFDM subcarrier, bit, and power allocation algorithm to minimize the total transmit power. This is done by assigning each user a set of subcarriers and by determining the number of bits and the transmit power level for each subcarrier. We obtain the performance of our proposed algorithm in a multiuser frequency selective fading environment for various time delay spread values and various numbers of users. The results show that our proposed algorithm outperforms multiuser OFDM systems with static time-division multiple access (TDMA) or frequency-division multiple access (FDMA) techniques which employ fixed and predetermined time-slot or subcarrier allocation schemes. We have also quantified the improvement in terms of the overall required transmit power, the bit-error rate (BER), or the area of coverage for a given outage probability.

Index Terms—Adaptive modulation, frequency selective fading channel, multiaccess communication, multiuser channel, orthogonal frequency division multiplexing (OFDM), resource management.

I. INTRODUCTION

RECENTLY, intense interest has focused on modulation techniques which can provide broadband transmission over wireless channels for applications including wireless multimedia, wireless Internet access, and future-generation mobile communication systems. One of the main requirements on the modulation technique is the ability to combat intersymbol interference (ISI), a major problem in wideband transmission over multipath fading channels. There are many methods proposed to combat the ISI, e.g., [1]–[3]. Multicarrier modulation techniques, including orthogonal frequency division multiplex (OFDM), (e.g., [4]) are among the more promising solutions to this problem.

Assuming that the transmitter knows the instantaneous channel transfer functions of all users, many papers [5]–[7] have demonstrated that significant performance improvement can be achieved if adaptive modulation is used with OFDM. In

particular, subcarriers with large channel gains employ higher order modulation to carry more bits/OFDM symbol, while subcarriers in deep fade carry one or even zero bits/symbol. Integrated design of forward error correcting code and adaptive modulation has also been studied using BCH code and trellis coded modulation (TCM) in [8] and [9], respectively. Although both references considered only time-varying flat fading channels, the same coded adaptive modulation design can be easily applied to OFDM systems. As different subcarriers experience different fades and transmit different numbers of bits, the transmit power levels must be changed accordingly. The problem of optimal power allocation has also been studied in [10].

In this paper, we consider extending OFDM with adaptive modulation to multiuser frequency selective fading environments. When OFDM with adaptive modulation is applied in a frequency selective fading channel, a significant portion of the subcarriers may not be used. These are typically subcarriers which experience deep fade and are not power efficient to carry any information bit. In multiuser systems using static time-division multiple access (TDMA) or frequency-division multiple access (FDMA) as multiaccess schemes, each user is allocated a predetermined time slot or frequency band to apply OFDM with adaptive modulation. Consequently, these unused subcarriers (as a result of adaptive modulation) within the allocated time slot or frequency band of a user are wasted and are not used by other users. However, the subcarriers which appear in deep fade to one user may not be in deep fade for other users. In fact, it is quite unlikely that a subcarrier will be in deep fade for all users, as the fading parameters for different users are mutually independent. This motivates us to consider an adaptive multiuser subcarrier allocation scheme where the subcarriers are assigned to the users based on instantaneous channel information. This approach will allow all the subcarriers to be used more effectively because a subcarrier will be left unused only if it appears to be in deep fade to all users.

We consider a multiuser subcarrier, bit, and power allocation scheme where all users transmit in all the time slots. Our objective is to minimize the overall transmit power by allocating the subcarriers to the users and by determining the number of bits and the power level transmitted on each subcarrier based on the instantaneous fading characteristics of all users. In this paper, we formulate the multiuser subcarrier, bit, and power allocation problem and propose an iterative algorithm to perform the multiuser subcarrier allocation. Once

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The authors are with the Department of Electrical and Electronic Engineering, Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong (e-mail: eeyui@ust.hk; eecheng@ust.hk; eekhaled@ust.hk; andeermurch@ust.hk).

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the subcarrier allocation is determined, the bit and power allocation algorithm can be applied to each user on its allocated subcarriers. We also compare the performance of our proposed solution to various other static subcarrier allocation schemes.

The results of the work can be applied, for instance, to the downlink transmission in a time division duplex (TDD) wireless communication system to improve the downlink capacity. In such a system, the base station (BS) can estimate the instantaneous channel characteristics of all the BS-to-mobile links based on the received uplink transmissions. The multiuser subcarrier, bit, and power allocation can then be used. It is clear that there is a certain amount of transmission overhead as the BS has to inform the mobiles about their allocated subcarriers and the number of bits assigned to each subcarrier.¹ However, this overhead can be relatively small, especially if the channels vary slowly (e.g., in an indoor low mobility environment), and the assignment is done once every many OFDM symbols. To further reduce the overhead, we can assign a contiguous band of subcarriers with similar fading characteristics as a group, instead of assigning each individual subcarrier. In this paper, we will not focus on how the subcarrier allocation information is transmitted. Instead, we will focus on how—and by how much—this new strategy can reduce the required transmit power; or how and by how much this new scheme can improve the bit-error rate (BER) for a fixed transmit power. Alternately, we also consider how and by how much this new scheme can increase the area of coverage for a given transmit power and target BER.

While the bit allocation algorithm can be viewed as a practical implementation of the water-pouring interpretation for achieving the Shannon capacity of an ISI channel [13], the multiuser subcarrier and bit allocation algorithm presented in this paper is the counterpart of the multiuser water-pouring solution given in [14]. In information theoretic studies, the usual approach is to maximize the capacity (or information rate) under the power constraint. In this study, we focus on deriving practical algorithms that can support real-time multimedia data whose bit rates are generally fixed by the compression algorithms. Hence, we assume a given set of user data rates and attempt to minimize the total transmit power under a fixed performance requirement.

The organization of this paper is as follows. In Section II, we will first give the system model and formulate the minimum overall transmit power problem. The optimization problem seeks to minimize the overall transmit power using combined subcarrier, bit, and power allocation schemes for multiuser OFDM systems. The bit and power allocation algorithm for a single-user system is studied in Section III. In Section IV, we derive a lower bound to the minimum overall transmit

power by relaxing some of the constraints in the original problem. We also derive a suboptimal subcarrier allocation algorithm. In Section V, we compare the performance between our proposed method and other static approaches via Monte Carlo simulations. Finally, we conclude in Section VI.

II. SYSTEM MODEL

The configuration of our multiuser adaptive OFDM system is shown in Fig. 1. We assume that the system has K users and the k th user has a data rate equal to R_k bit per OFDM symbol. In the transmitter, the serial data from the K users are fed into the subcarrier and bit allocation block which allocates bits from different users to different subcarriers. We assume that each subcarrier has a bandwidth that is much smaller than the coherence bandwidth of the channel and that the instantaneous channel gains on all the subcarriers of all the users are known to the transmitter. Using the channel information, the transmitter applies the combined subcarrier, bit, and power allocation algorithm to assign different subcarriers to different users and the number of bits/OFDM symbol to be transmitted on each subcarrier. Depending on the number of bits assigned to a subcarrier, the adaptive modulator will use a corresponding modulation scheme, and the transmit power level will be adjusted according to the combined subcarrier, bit, and power allocation algorithm. We define $c_{k,n}$ to be the number of bits of the k th user that are assigned to the n th subcarrier. As we do not allow more than one user to share a subcarrier, it follows that for each n , if $c_{k',n} \neq 0$, $c_{k,n} = 0$ for all $k \neq k'$. We also assume that the adaptive modulator allows $c_{k,n}$ to take values in the set $D = \{0, 1, 2, \dots, M\}$ where M is the maximum number of information bits/OFDM symbol that can be transmitted by each subcarrier.

The complex symbols at the output of the modulators are transformed into the time domain samples by inverse fast Fourier transform (IFFT). Cyclic extension of the time domain samples, known as the guard interval, is then added to ensure orthogonality between the subcarriers, provided that the maximum time dispersion is less than the guard interval. The transmit signal is then passed through different frequency selective fading channels to different users.

We assume that the subcarrier and bit allocation information is sent to the receivers via a separate control channel. At the receiver, the guard interval is removed to eliminate the ISI, and the time samples of the k th user are transformed by the FFT block into modulated symbols. The bit allocation information is used to configure the demodulators while the subcarrier allocation information is used to extract the demodulated bits from the subcarriers assigned to the k th user.

In the frequency selective fading channel, different subcarriers will experience different channel gains. We denote by $\alpha_{k,n}$ the magnitude of the channel gain (assuming coherent reception) of the n th subcarrier as seen by the k th user. We assume that the single-sided noise power spectral density (PSD) level N_0 is equal to unity (i.e., $N_0 = 1$), for all

¹ Note that the power level used does not need to be transmitted to the receiver in such a TDD system. As the subcarrier gain is known to the transmitter, it can adjust the transmit power level to achieve a predetermined receiver power level based on the number of bits allocated to that subcarrier. However, in FDD systems, the transmit power levels determined by the receiver have to be sent back to the transmitter. In such systems, the additional performance gain achieved by power allocation may not justify the cost of sending the transmit power level information to the transmitter.

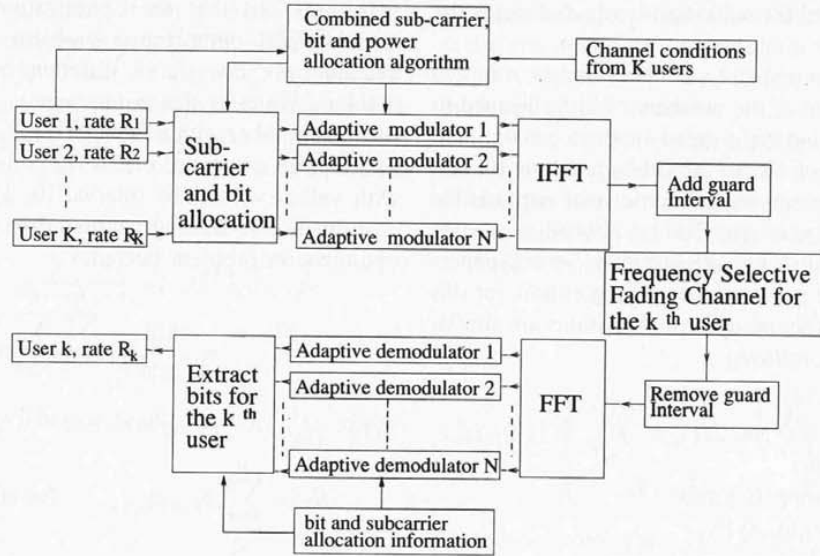


Fig. 1. Block diagram of a multiuser OFDM system with subcarrier, bit, and power allocation.

subcarriers and is the same for all users. Furthermore, we denote by $f_k(c)$ the required received power (in energy per symbol) in a subcarrier for reliable reception of c information bits/symbol when the channel gain is equal to unity. Note that the function $f_k(c)$ depends on k , and this allows different users to have different quality-of-service (QoS) requirements and/or different coding and modulation schemes. In order to maintain the required QoS at the receiver, the transmit power, allocated to the n th subcarrier by the k th user must equal

$$P_{k,n} = \frac{f_k(c_{k,n})}{\alpha_{k,n}^2}. \quad (1)$$

Using these transmit power levels, the receiver can demodulate the modulated symbols at the output of the FFT processor and achieve the desired QoS's of all users.

The goal of the combined subcarrier, bit, and power allocation algorithm is then to find the best assignment of $c_{k,n}$ so that the overall transmit power, the sum of $P_{k,n}$ over all subcarriers and all users, is minimized for given transmission rates of the users and given QoS requirements specified through $f_k(c)$, $k = 1, \dots, K$. In order to make the problem tractable, we further require that $f_k(c)$ is a convex and increasing function with $f_k(0) = 0$. This condition essentially means that no power is needed when no bit is transmitted and that the required additional power to transmit an additional bit increases with c [i.e., $f_k(c+1) - f_k(c)$ is increasing in c]. Almost all popular coding and modulation schemes satisfy this condition.

It is important to note that even though the problem is formulated to minimize the overall transmit power for given QoS requirements, the same solution can be applied to improve the QoS's of the users for a given overall transmit power. The latter can simply be achieved by increasing the power proportionally for all the subcarriers, while using the same set of $c_{k,n}$.

Mathematically, we can formulate the problem as

$$P_T^* = \min_{c_{k,n} \in \mathbf{D}} \sum_{n=1}^N \sum_{k=1}^K \frac{1}{\alpha_{k,n}^2} f_k(c_{k,n}) \quad (2)$$

and the minimization is subjected to the constraints

$$\text{C1: For all } k \in \{1, \dots, K\}, R_k = \sum_{n=1}^N c_{k,n} \quad (3)$$

and

$$\begin{aligned} \text{C2: For all } n \in \{1, \dots, N\}, \\ \text{if there exists } k' \text{ with } c_{k',n} \neq 0, \text{ then } c_{k,n} = 0, \\ \forall k \neq k'. \end{aligned} \quad (4)$$

Note that constraint (3) is the data rate requirement and constraint (4) ensures that each subcarrier can only be used by one user. Moreover, $\mathbf{D} = \{0, 1, 2, \dots, M\}$ is the set of all possible values for $c_{k,n}$, and $c_{k,n} = 0$ means that the k th user does not use the n th subcarrier to transmit any information.

III. BIT ALLOCATION ALGORITHM FOR SINGLE USER CHANNEL

Before we try to solve the multiuser allocation problem, we will first derive the bit allocation algorithm for the single-user environment. The single-user problem not only gives better understanding of the issues involved, but also provides a bit allocation algorithm that we will use in our multiuser solution.

We can rewrite the optimization problem in (2) for the single-user case as

$$P_T^* = \min_{c_n \in \mathbf{D}} \sum_{n=1}^N \frac{1}{\alpha_n^2} f(c_n) \quad (5)$$

and the minimization is under the constraint

$$R = \sum_{n=1}^N c_n. \quad (6)$$

Note that we have dropped the subscript k , which denotes the user in all notations.

As the power needed to transmit a certain number of bits in a subcarrier is independent of the numbers of bits allocated to other subcarriers, it turns out that a greedy approach is optimal. A greedy algorithm assigns bits to the subcarriers one bit at a time, and in each assignment, the subcarrier that requires the least additional power is selected. The bit allocation process will be completed when all R bits are assigned. Several papers (e.g., [15] and [16]) have provided various algorithms for this problem, and the basic structure of most algorithms are similar and can be described as follows:

Initialization:

For all n , let $c_n = 0$ and $\Delta P_n = [f(1) - f(0)]/\alpha_n^2$;

Bit Assignment Iterations:

Repeat the following R times:

$\hat{n} = \arg \min_n \Delta P_n$;

$c_{\hat{n}} = c_{\hat{n}} + 1$;

$\Delta P_{\hat{n}} = [f(c_{\hat{n}} + 1) - f(c_{\hat{n}})]/\alpha_{\hat{n}}^2$;

End;

Finish:

$\{c_n\}_{n=1}^N$ is the final bit allocation solution.

The initialization stage computes, for each subcarrier, the additional power needed to transmit an additional bit. For each bit assignment iteration, the subcarrier that needs the minimum additional power is assigned one more bit, and the new additional power for that subcarrier is updated. After R iterations, the final bit assignment gives the optimal bit allocation for each subcarrier. It is important to note that the bit allocation is optimal only for the given function $f(c)$, which depends on the selected modulation scheme. Different modulation schemes will lead to different $f(c)$, different bit allocation, and possibly lower transmit power P_T^* .

The concept of this algorithm is fairly simple, and many similar algorithms based on the same principle have been obtained before. In particular, there exist faster and less complex algorithms which can speed up the bit allocation process significantly (e.g., [15] and [16]). In our simulations, we use the algorithm given in [16].

IV. MULTIUSER SUBCARRIER AND BIT ALLOCATION

We have observed that, in the single-user case, a greedy approach which assigns one bit at a time to the subcarrier that requires the least additional power gives the optimal allocation in the sense of minimizing the overall transmit power. Unfortunately, the problem becomes more difficult in the multiuser environment. As users cannot share the same subcarrier, allocating bits to a subcarrier essentially prevents other users from using that subcarrier. This dependency makes any greedy algorithm a nonoptimal solution. It turns out that the optimal solution may not assign any of a user's bits to the best subcarrier seen by that user. This may happen when the best subcarrier of a user is also the best subcarrier of another user who happens to have no other good subcarriers. Hence, the multiuser subcarrier and bit allocation problem is much more complicated to solve than that of the single-user case.

It turns out that the optimization problem in (2) is a combinatorial optimization problem. To make the problem tractable, we consider a different but similar optimization problem. We relax the requirement $c_{k,n} \in \mathbf{D}$ to allow $c_{k,n}$ to be a real number within the interval $[0, M]$. Moreover, in order to deal with constraint (4), K variables, $\rho_{k,n}$, $k = 1, \dots, K$, with values within the interval $[0, 1]$, are introduced to the cost function as sharing factors of the n th subcarrier. The new optimization problem becomes

$$\underline{P}_T = \min_{\substack{c_{k,n} \in [0, M] \\ \rho_{k,n} \in [0, 1]}} \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k(c_{k,n}) \quad (7)$$

where $c_{k,n}$ and $\rho_{k,n}$ have to satisfy

$$R_k = \sum_{n=1}^N \rho_{k,n} c_{k,n}, \quad \text{for all } k \in \{1, \dots, K\} \quad (8)$$

and

$$1 = \sum_{k=1}^K \rho_{k,n}, \quad \text{for all } n \in \{1, \dots, N\}. \quad (9)$$

For any valid set of $c_{k,n} \in \mathbf{D}$ satisfying the constraints (3) and (4) in the original optimization problem, we can let

$$\rho_{k,n} = \begin{cases} 1, & \text{if } c_{k,n} \neq 0, \\ 0, & \text{if } c_{k,n} = 0. \end{cases} \quad (10)$$

Then, it is easy to show that the same set of $c_{k,n}$ and the corresponding $\rho_{k,n}$ defined in (10) satisfy the constraints (8) and (9) in the new optimization problem. Moreover, with $\rho_{k,n}$ defined in (10), the new cost function in (7) is equal to the cost function in (2). Hence, the minimization problem in (7) is the same as the original optimization problem, except that the minimization is done over a larger set. Consequently, the minimum power obtained in (7) $\underline{P}_T$ is a lower bound to the minimum power obtained in (2), P_T^* .

Another way to interpret the optimization in (7) is to consider $\rho_{k,n}$ as the time-sharing factor for the k th user of the n th subcarrier. For example, in every L OFDM symbol (L being a very large number), user k uses the n th subcarrier in $L\rho_{k,n}$ symbols. Clearly, the average (over L symbols) information data rate and the average transmit power has to be scaled by the same factor $\rho_{k,n}$. Hence, we can consider (7) as the optimization problem when the users are allowed to time-share each subcarrier over a large number of OFDM symbols. However, most wireless communication channels are time varying, and the channels may not stay unchanged long enough for timesharing to be feasible. Hence, in this paper, we will continue to consider the original problem in (2) and use the optimization problem in (7) as a lower bound, even though it has its own physical interpretation.

The modified optimization problem in (7) is more tractable. However, even though the function $f_k(c)$ is convex in c , the terms in the cost function have the form $\rho f_k(c)$, and as a function of (ρ, c) , $\rho f_k(c)$ is not convex in (ρ, c) . To proceed further, we let $r_{k,n} = c_{k,n} \rho_{k,n}$ and rewrite the cost function in terms of $r_{k,n}$ and $\rho_{k,n}$. The constraint on $r_{k,n}$

becomes $r_{k,n} \in [0, M\rho_{k,n}]$, and it can be easily shown that $\rho f_k(c) = \rho f_k(r/\rho)$ is convex in (ρ, r) within the triangular region specified by $\rho \in [0, 1]$ and $r \in [0, M\rho]$. In particular, the Hessian evaluated at any point within this region is a positive semidefinite matrix. Hence, we can reformulate the optimization problem in (7) as a convex minimization problem over a convex set. That is

$$P_T = \min_{\substack{r_{k,n} \in [0, M\rho_{k,n}] \\ \rho_{k,n} \in [0, 1]}} \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k\left(\frac{r_{k,n}}{\rho_{k,n}}\right) \quad (11)$$

where $r_{k,n}$ and $\rho_{k,n}$ have to satisfy

$$R_k = \sum_{n=1}^N r_{k,n}, \quad \text{for all } k \in \{1, \dots, K\} \quad (12)$$

and

$$1 = \sum_{k=1}^K \rho_{k,n}, \quad \text{for all } n \in \{1, \dots, N\}. \quad (13)$$

Using standard optimization techniques in [17], we obtain the Lagrangian

$$L = \sum_{n=1}^N \sum_{k=1}^K \frac{\rho_{k,n}}{\alpha_{k,n}^2} f_k\left(\frac{r_{k,n}}{\rho_{k,n}}\right) - \sum_{k=1}^K \lambda_k \left(\sum_{n=1}^N r_{k,n} - R_k \right) - \sum_{n=1}^N \beta_n \left(\sum_{k=1}^K \rho_{k,n} - 1 \right) \quad (14)$$

where λ_k and β_n are the Lagrangian multipliers for the constraints (12) and (13), respectively.

After differentiating L with respect to $r_{k,n}$ and $\rho_{k,n}$, respectively, we obtain the necessary conditions for the optimal solution, $r_{k,n}^*$ and $\rho_{k,n}^*$. Specifically, if $\rho_{k,n}^* \neq 0$, we have

$$\begin{aligned} \frac{\partial L}{\partial r_{k,n}} \bigg|_{(r_{k,n}, \rho_{k,n})=(r_{k,n}^*, \rho_{k,n}^*)} &= \frac{1}{\alpha_{k,n}^2} f'_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) - \lambda_k \begin{cases} > 0, & \text{if } r_{k,n}^* = 0 \\ = 0, & \text{if } r_{k,n}^* \in (0, M\rho_{k,n}^*) \\ < 0, & \text{if } r_{k,n}^* = M\rho_{k,n}^* \end{cases} \end{aligned} \quad (15)$$

and

$$\begin{aligned} \frac{\partial L}{\partial \rho_{k,n}} \bigg|_{(r_{k,n}, \rho_{k,n})=(r_{k,n}^*, \rho_{k,n}^*)} &= \frac{1}{\alpha_{k,n}^2} \left[f_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) - f'_k\left(\frac{r_{k,n}^*}{\rho_{k,n}^*}\right) \frac{r_{k,n}^*}{\rho_{k,n}^*} \right] \\ &\quad - \beta_n \begin{cases} = 0, & \text{if } \rho_{k,n}^* \in (0, 1) \\ < 0, & \text{if } \rho_{k,n}^* = 1. \end{cases} \end{aligned} \quad (16)$$

On the other hand, if $\rho_{k,n}^* = 0$, then $r_{k,n}^* = 0$, and we have

$$\begin{aligned} r_{k,n} \frac{\partial L}{\partial r_{k,n}} + \rho_{k,n} \frac{\partial L}{\partial \rho_{k,n}} &\geq 0, \\ \text{for all } \rho_{k,n} \in (0, 1] \text{ and } r_{k,n} \in (0, M\rho_{k,n}]. \end{aligned} \quad (17)$$

These necessary conditions can be interpreted by the fact that if the minimum occurs within the constrained region $[(0, 1)$ for $\rho_{k,n}$ and $(0, M\rho_{k,n})$ for $r_{k,n}]$, then the derivative evaluated at the minimum point must be zero. On the other hand, if the optimal solution occurs at a boundary point, then the derivative must be positive along all directions pointing toward the interior of the constraint set. Then, (17) follows from considering the boundary point at $(r_{k,n}^*, \rho_{k,n}^*) = (0, 0)$.

From (15) and (17), we can conclude that

$$r_{k,n}^* = \rho_{k,n}^* f_k'^{-1}(\lambda_k \alpha_{k,n}^2) \quad (18)$$

where

$$\lambda_{k,n} = \begin{cases} f'_k(0)/\alpha_{k,n}^2, & \text{if } f_k'^{-1}(\lambda_k \alpha_{k,n}^2) < 0; \\ \lambda_k, & \text{if } 0 \leq f_k'^{-1}(\lambda_k \alpha_{k,n}^2) \leq M; \\ f'_k(M)/\alpha_{k,n}^2, & \text{if } f_k'^{-1}(\lambda_k \alpha_{k,n}^2) > M. \end{cases}$$

Moreover, from (16) and (17), it follows that

$$\rho_{k,n}^* = \begin{cases} 0, & \text{if } \beta_n < H_{k,n}(\lambda_{k,n}) \\ 1, & \text{if } \beta_n > H_{k,n}(\lambda_{k,n}) \end{cases} \quad (19)$$

where

$$H_{k,n}(\lambda) = \frac{1}{\alpha_{k,n}^2} [f_k(f_k'^{-1}(\lambda \alpha_{k,n}^2)) - \lambda \alpha_{k,n}^2 f_k'^{-1}(\lambda \alpha_{k,n}^2)]. \quad (20)$$

Since constraint (13) must be satisfied, we find from (19) that for each n , if $H_{k,n}(\lambda_{k,n})$ for $k = 1, \dots, K$ are all different, then only the user with the smallest $H_{k,n}(\lambda_{k,n})$ can use that subcarrier. In other words, for the n th subcarrier, if $H_{k,n}(\lambda_{k,n})$ are different for all k , then

$$\rho_{k',n}^* = 1, \quad \rho_{k,n}^* = 0, \quad \text{for all } k \neq k' \quad (21)$$

where

$$k' = \arg \min_k H_{k,n}(\lambda_{k,n}). \quad (22)$$

Hence, it follows that for a fixed set of Lagrange multipliers λ_k , $k = 1, \dots, K$, we can use them to determine k' for each n using (22). The $r_{k,n}^*$ and $\rho_{k,n}^*$ obtained will then form an optimal solution for the optimization problem; however, the individual rate constraint (12) may not be satisfied.

In order to find the set of λ_k such that the individual rate constraints are satisfied, we have obtained an iterative searching algorithm. Starting with some small values for all λ_k , this iterative procedure increases one of the λ_k until the data rate constraint (12) for user k is satisfied. Then, we switch to another user and go through the users one at a time. This process repeats for all users until the data rate constraint for all users are satisfied. This algorithm converges because for a given k , as λ_k increases, $H_{k,n}(\lambda_{k,n})$ for all n decreases, and more $\rho_{k,n}^*$ in (19) become one while $r_{k,n}^*$ in (18) increases for those n where $\rho_{k,n}^* > 0$. Hence, $\sum_{n=1}^N r_{k,n}^*$ increases. During this process, some of the other $\rho_{k',n}^*$ may change from one to zero and consequently decrease the total data rate for other users. However, as all the λ_k increase, $r_{k,n}^*$ increases accordingly. As long as the total data rate is less than MN bits/symbol, which is the total number of bits possibly transmitted within an OFDM symbol, the algorithm

will converge to a solution that satisfies all the constraints. Since the optimization problem is a convex optimization problem over a convex set, the set of necessary conditions is also sufficient, and the solution that satisfies all the necessary conditions is the unique optimal solution.

In the process of adjusting λ_k for $k = 1, \dots, K$, the situation where, for a fixed n , more than one $H_{k,n}(\lambda_{q,k})$ has the same values cannot be ignored. In that case, $\rho_{k,n}^*$ has to take values within the interval $(0, 1)$. This solution suggests that the subcarrier should be shared by multiple users. In practice, this can be done by having these users with $\rho_{k,n}^* > 0$ time share the n th subcarrier, and the ratio of the symbols used by different users are set proportionally to $\rho_{k,n}^*$. The detailed flow chart of the algorithm is given in the Appendix.

Now, we have an algorithm to obtain the optimal values of $\rho_{k,n}^*$ and

$$c_{k,n}^* = \begin{cases} r_{k,n}^* / \rho_{k,n}^*, & \text{if } \rho_{k,n}^* \neq 0 \\ 0, & \text{otherwise.} \end{cases} \quad (23)$$

This solution, when substituted in (7), gives a lower bound to the minimum overall transmit power. However, we cannot use these results immediately in (2). One problem is that $c_{k,n}^*$ may not be in $\mathbf{D}$, and the other is that some $\rho_{k,n}^*$ may be within $(0, 1)$, indicating a time-sharing solution. Furthermore, simply quantizing $c_{k,n}^*$ and $\rho_{k,n}^*$ will not satisfy the individual rate constraints in (3).

To solve this problem, we propose a multiuser adaptive OFDM (MAO) scheme where the subcarrier allocation follows essentially the solution to the lower bound in (7), and then the single-user bit allocation algorithm given in Section III is applied to each user on the allocated subcarriers. Specifically, we modify $\rho_{k,n}^*$ for the optimization problem in (7) by letting for each n $\rho_{k',n}^* = 1$ where $k' = \arg \max_k \rho_{k,n}^*$, and $\rho_{k,n}^* = 0$ for $k \neq k'$. Then, we apply the single-user bit allocation algorithm on each user using the assigned subcarriers. We denote the total transmit power (in energy/symbol) obtained using this MAO scheme by P_T . It is easy to see that $\underline{P}_T \leq P_T^* \leq P_T$, where P_T^* is the minimum power in the original problem, and $\underline{P}_T$ is the minimum power for the modified problem with the relaxed constraints. More specifically, the difference between P_T and the minimum $\underline{P}_T$ gives an upper bound to how far away our MAO scheme is from the solution of our original optimization problem.

V. PERFORMANCE COMPARISON

In this section, we obtain and compare the performance of the MAO scheme with other static subcarrier allocation schemes. We consider a system that employs M-ary quadrature amplitude modulation (MQAM) with $\mathbf{D} = \{0, 2, 4, 6\}$. Square signal constellations (4-QAM, 16-QAM, and 64-QAM) are used to carry two, four, or six bits/symbol. The bit-error probability is upper bounded by the symbol error probability, which is tightly approximated by $4Q[\sqrt{d^2/(2N_0)}]$ [12, p. 281], where d is the minimum distance between the points in the signal constellation. Since the average energy of a M-QAM symbol is equal to $(M-1)d^2/6$, it follows that the required power for supporting c bits/symbol at a given BER

P_e is

$$f(c) = \frac{N_0}{3} \left[Q^{-1} \left(\frac{P_e}{4} \right) \right]^2 (2^c - 1)$$

where we recall that

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-t^2/2} dt.$$

It is easy to see that $f(c)$ is convex and increasing in c and that $f(0) = 0$.

To evaluate the performance of our scheme, we have simulated 1000 sets of five-path frequency selective Rayleigh fading channels with an exponential power delay profile. Each set of channels consists of K independent channels, one for each user. We use an OFDM system with 128 subcarriers over a 5 MHz band along with a total (over all users) transmission rate equal to 512 bits/symbol (or equivalently, an average of four bits/subcarrier). Recall that the single-sided power spectral density level N_0 is equal to unity, and we assume that the average subcarrier channel gain $E|\alpha_{k,n}|^2$ is equal to unity for all k and n .

For comparison purposes, we have also considered three other static multiuser subcarrier allocation methods. Two of them are based on the multiple access methods described in [7]. The methods are presented as follows.

- OFDM-TDMA: each user is assigned a predetermined TDMA time slot and can use all the subcarriers within that time slot exclusively.
 - OFDM-FDMA: each user is assigned a predetermined band of subcarriers and can only use those subcarriers exclusively in every OFDM symbol.
- In a frequency selective fading channel, there is a high correlation between the channel gains of adjacent subcarriers. In order to avoid the situation where all subcarriers of a user are in deep fade, we propose an enhanced version of OFDM-FDMA, which we shall refer to as OFDM Interleaved-FDMA.
- OFDM Interleaved-FDMA: this is the same as OFDM-FDMA except that subcarriers assigned to a user are interlaced with other users' subcarriers in the frequency domain.

The time and subcarrier assignment of these three multiuser OFDM schemes are illustrated in Fig. 2. Note that these static schemes have predetermined subcarrier allocations which are independent of the channel gains of the users. The main difference between the proposed MAO scheme and these static schemes is that MAO assigns subcarriers adaptively based on the instantaneous channel gains. To ensure a fair comparison, we use the optimal single-user bit allocation (OBA) for each user on the assigned subcarriers. For comparison purposes, we also show the results when equal bit allocation (EBA) is employed on the assigned subcarriers for these three OFDM schemes. Notice that when using EBA, all three schemes will have the same performance in an uncoded system. This is because the average bit signal-to-noise ratio (SNR) needed is a function of only the marginal probability density function of each subcarrier gain.

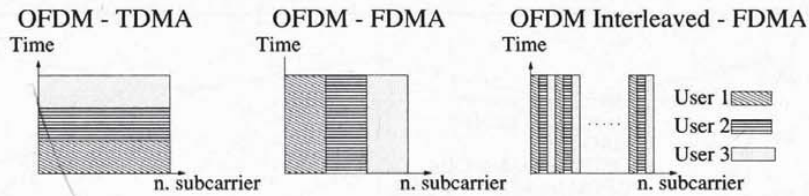


Fig. 2. Subcarrier and time-slot allocations of OFDM-TDMA, OFDM-FDMA, and OFDM interleaved-FDMA schemes.

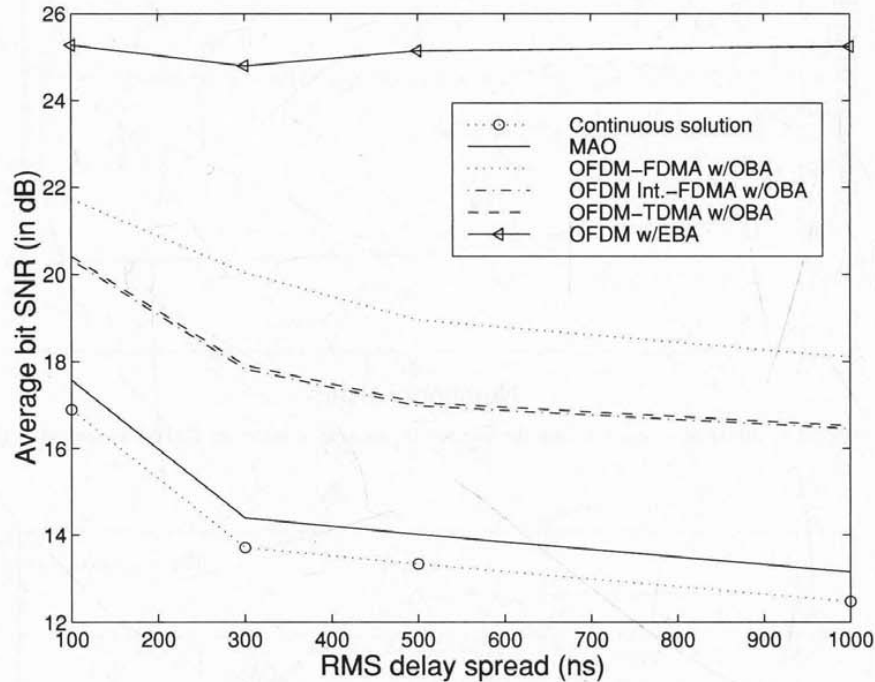


Fig. 3. Average bit signal-to-noise ratio (SNR) required by different schemes in various root mean square (RMS) delay spreads in a five-user system with $P_e = 10^{-4}$.

Fig. 3 shows the average bit SNR needed to achieve a BER at $P_e = 10^{-4}$ for a five-user system versus the root mean square (RMS) delay spread (for definition, see for example [18, p. 160]) for different multiuser OFDM schemes. The average required transmit power (in energy per bit) is defined as the ratio of the overall transmit energy per OFDM symbol (including all subcarriers and all users) to the total number of bits transmitted per OFDM symbol. Moreover, we define the average bit SNR as the ratio of the average transmit power to the noise PSD level N_0 . As we assume that the data rate is fixed and that N_0 is just a constant, the overall transmit power is proportional to the average bit SNR. For ease of comparison, we have used the average bit SNR for comparison. We find in Fig. 3 that the MAO scheme is never more than 0.6 dB from the lower bound. Since the bit SNR of the optimal combined subcarrier, bit, and power allocation algorithm must lie between the bit SNR's achieved by the lower bound and the MAO scheme, we find that the MAO scheme is never more than 0.6 dB away from the optimal solution. On the other hand, we observe that our proposed MAO scheme is 3–5 dB better than the static subcarrier allocation schemes with OBA, which are in turn 5–10 dB better than that with EBA. We also find that when OBA is used, the OFDM interleaved-FDMA

scheme and the OFDM-TDMA scheme have very similar performance, and both of them outperform the OFDM-FDMA scheme.² A closer observation of Fig. 3 also indicates that the gains achieved by optimal bit allocation and optimal multiuser subcarrier allocation increase with the RMS delay spread. This is mainly because the larger the RMS delay spread, the more the fading variation and hence higher gains can be obtained when the allocation is performed adaptively.

Fig. 4 shows the average bit SNR (in dB) needed to achieve the same BER versus the number of users when the RMS delay spread is 100 ns. We find that the savings in the required bit SNR achieved by MAO when compared to other schemes are roughly the same, independent of the number of users in the system.

While these two figures show the improvement in the required bit SNR, the results can perhaps be more easily understood using the more familiar BER versus bit SNR curves. For each BER requirement, we compute $f(c)$ for all $c \in \mathcal{D}$ and then use our algorithm to calculate the subcarrier

²OFDM-FDMA refers only to the specific FDMA scheme that assigns to each user a contiguous band of subcarriers as shown in Fig. 2, but not the general FDMA schemes. In fact, both OFDM interleaved-FDMA and MAO can be considered as different forms of FDMA and they are not outperformed by the OFDM-TDMA scheme.

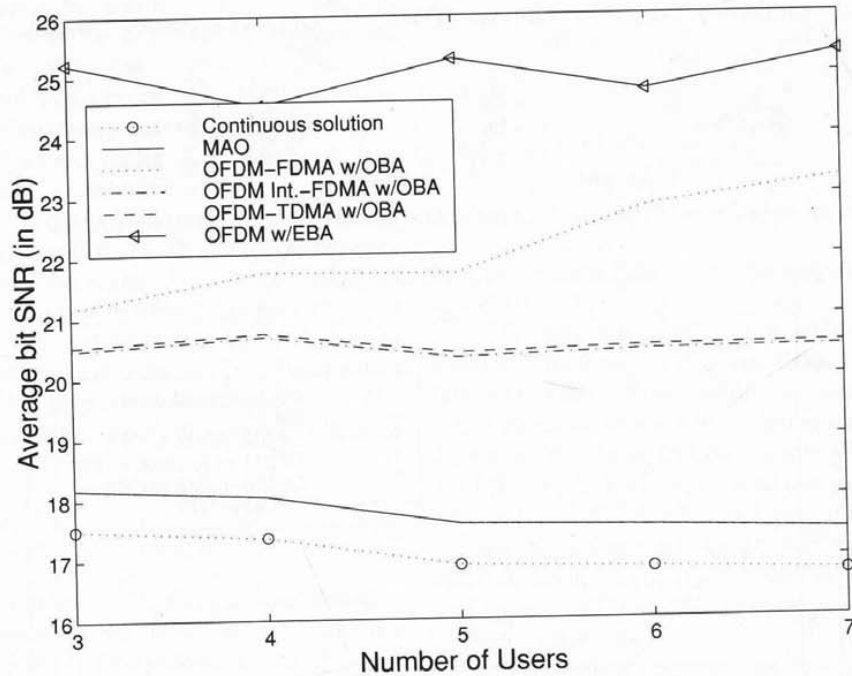


Fig. 4. Average bit SNR required by different schemes versus the number of users in a multiuser OFDM system with 100 ns RMS delay spread, and $P_e = 10^{-4}$.

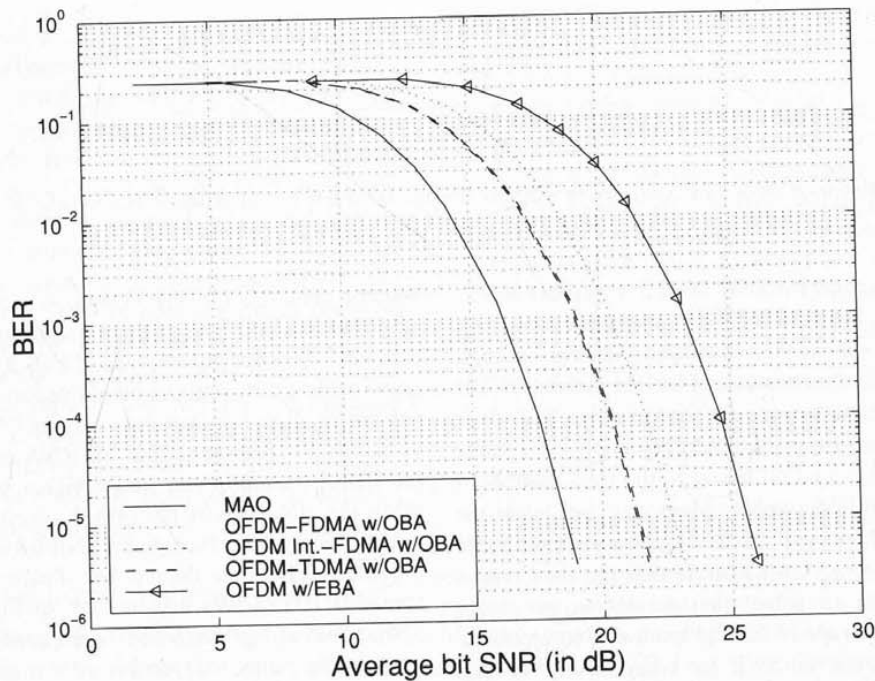


Fig. 5. BER versus average bit SNR for various subcarrier allocation schemes.

allocation for the MAO case. For all other static subcarrier allocation schemes, the allocations are independent of the BER. Once the subcarrier allocation is fixed, we apply the optimal bit and power allocation algorithm to every user. The final average power per bit divided by the noise power spectral density level gives the average bit SNR. We repeat this procedure for different BER values, and the results are

plotted in Fig. 5 for a five-user system with an RMS delay spread equal to 100 ns. We find that our proposed MAO has at least 3–4 dB advantage over all other schemes.

Another way to illustrate the impact of the bit and subcarrier allocation is to consider the area of coverage for a given outage probability, assuming that the BS has a maximum transmit power. We consider a circular cell with five users, indepen-

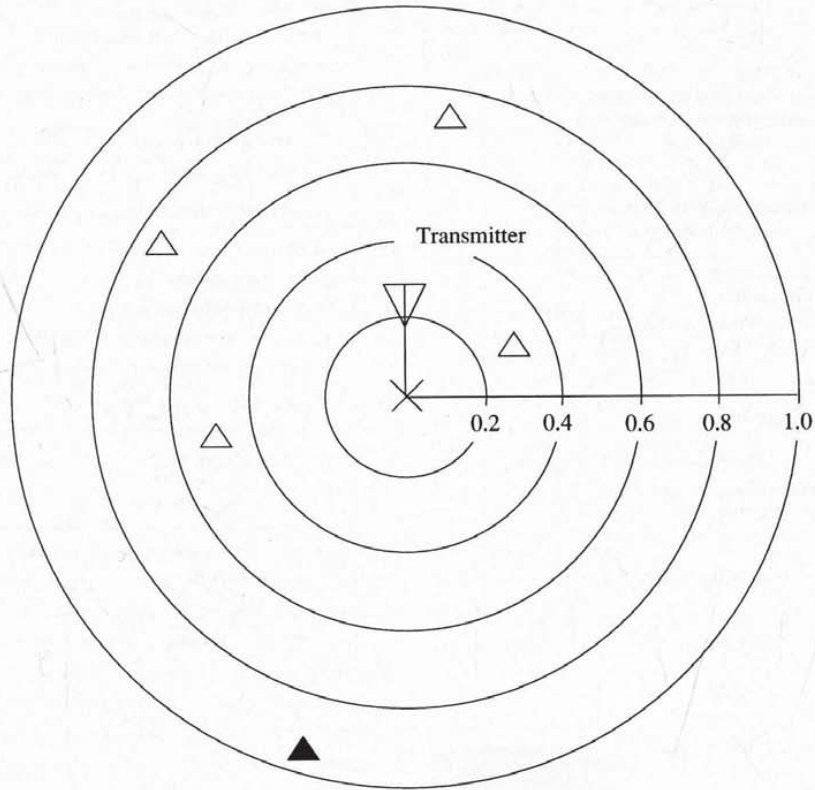


Fig. 6. Cell for analyzing the outage probability.

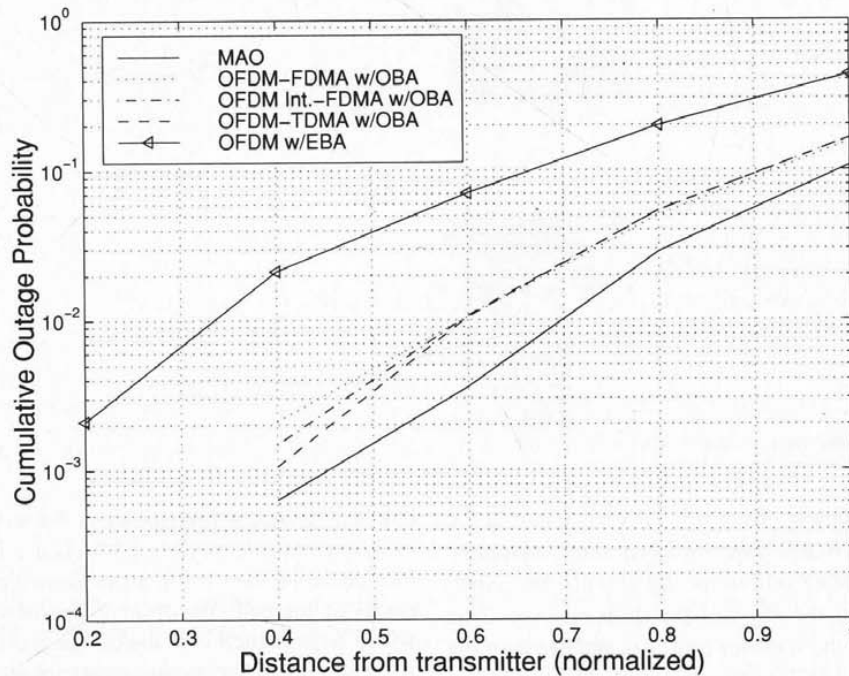


Fig. 7. Outage probability at 17 dB fading margin.

dently and uniformly distributed within the cell. A typical scenario is shown in Fig. 6, where the triangles represent the five users. In addition to frequency selective fading, path loss and log-normal shadowing are also included in simulating the

actual channel gains seen by the users. Using these channel gains, subcarriers and bits assigned to each user are determined by the various multiple access schemes and the total required transmit power is calculated. If the total power for all five users

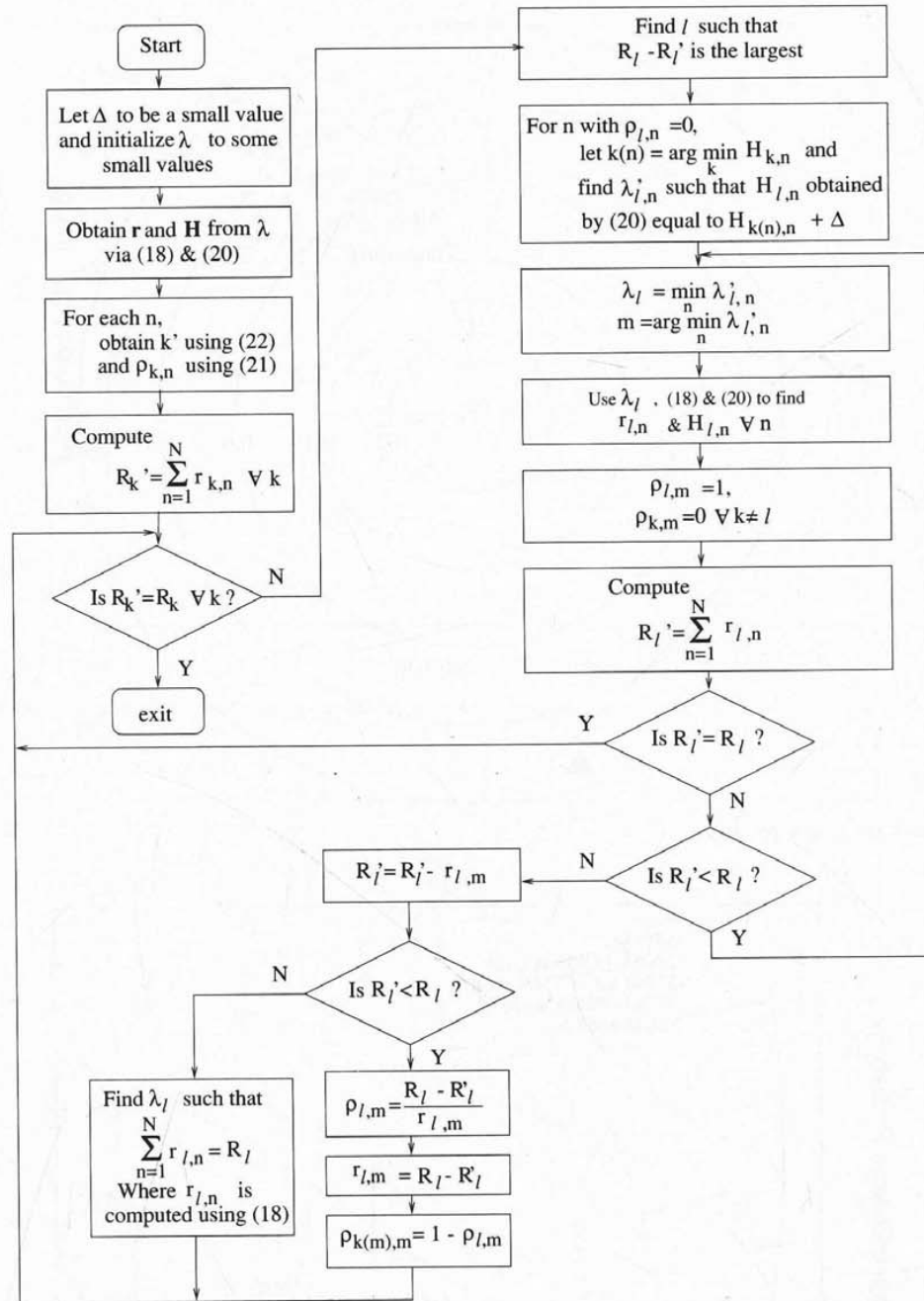


Fig. 8. Flow chart of the multiuser subcarrier allocation algorithm.

exceeds the maximum power of the BS, the user requiring the largest transmit power (in this case, the black one) is dropped and counted as one outage event occurring at a distance equal to the distance between the BS and the dropped user. This process continues until the transmit power is smaller than the maximum power of the BS. In this example, the maximum transmit power is set to the transmit power required for all five users assuming that they are all located at the boundary of the cell, taking into account the path loss effect and a 17 dB fading margin for shadowing.

The cumulative outage probabilities at various normalized distances, normalized to the cell radius, are plotted in Fig. 7. A

cumulative outage probability of 5% at a normalized distance of 0.8 means that there is a 5% chance of outage for a mobile located more than $0.8R$ away from the BS where R is the radius of the cell. We observe that MAO outperforms others with a large reduction in the outage probability at all distances. Alternatively, if the same outage probability is maintained, say at 1%, the coverage are provided by MAO is 36% larger than the best of all other schemes.

VI. CONCLUSION

In this paper, we considered OFDM transmission in a multiuser environment and formulated the problem of min-

imizing the overall transmit power by adaptively assigning subcarriers to the users along with the number of bits and power level to each subcarrier. In particular, we derived a multiuser adaptive subcarrier and bit allocation algorithm. Given the instantaneous channel information, the algorithm obtains a suboptimal subcarrier allocation, and then single-user bit allocation is applied on the allocated subcarriers. Using this scheme, the overall required transmit power can be reduced by about 5–10 dB from the conventional OFDM without adaptive modulation. Likewise, the transmit power can be reduced by about 3–5 dB from the conventional OFDM with adaptive modulation and adaptive bit allocation, but without adaptive subcarrier allocation. The reduction in transmit power can also be translated to a significant reduction in the required bit SNR for a given BER. Moreover, the same improvement can also be translated to a reduction in the outage probability or to an increase in the area of coverage.

The results in this paper assume perfect channel estimation, and we have not considered issues related to imperfect implementation, such as imperfect synchronization. As channel estimation in wireless fading channels is in general not very accurate, the effect of nonideal channel information on the performance of our proposed MAO scheme is a very important issue. We have started looking at this issue, and our preliminary results have indicated that the MAO scheme is not very sensitive to channel estimation errors. Nevertheless, detailed sensitivity studies will be needed before the algorithm can be applied to practical systems.

APPENDIX

A flow chart providing the detailed description of the multiuser subcarrier allocation algorithm is shown in Fig. 8.

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Cheong Yui Wong was born in Hong Kong, in 1975. He received the B.Eng. degree in electronic engineering from the Hong Kong University of Science and Technology (HKUST), Clear Water Bay, Hong Kong in 1997. He is currently pursuing the M.Phil. degree at HKUST and expects to graduate in 1999.

During his master studies, he joined a project that aims at achieving wireless multimedia communication using OFDM. His research interests include adaptive modulation, OFDM, and optimization.



Roger S. Cheng (S'85–M'91) received the B.S. degree from Drexel University, Philadelphia, PA, in 1987, and both the M.A. and Ph.D. degrees from Princeton University, Princeton, NJ, in 1988 and 1991, respectively, all in electrical engineering.

From 1987 to 1991, he was a Research Assistant in the Department of Electrical Engineering, Princeton University. From 1991 to 1995, he was an Assistant Professor in the Electrical and Computer Engineering Department of University of Colorado, Boulder. In June 1995, he joined the Faculty of Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong, where he is currently an Assistant Professor in the Department of Electrical and Electronic Engineering. He has also held visiting positions with Qualcomm, San Diego, in the summer of 1995 and with Institute for Telecommunication Sciences, NTIA, Boulder, CO, in the summers of 1993 and 1994. His current research interests include wireless communications, multiuser communications, spread spectrum CDMA, digital implementation of communication systems, wireless multimedia communications, information theory, and coding.

Dr. Cheng was the recipient of the Meitec Junior Fellowship Award from the Meitec Corporation in Japan, the George Van Ness Lothrop Fellowship from the School of Engineering and Applied Science in Princeton University, and the Research Initiation Award from the National Science Foundation. He was a member of the Technical Program Committees of GLOBECOM'98 and IEEE Wireless Communications and Networking Conference (WCNC'99). He has served as Guest Editor of the special issue on Multimedia Network Radios in the IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS, as Associate Editor of the IEEE TRANSACTIONS ON SIGNAL PROCESSING, and as Membership Chair for of the IEEE Information Theory Society.



Khaled Ben Letaief (S'85-M'86-SM'97) received the B.S. degree (with distinction), the M.S. degree, and the Ph.D. degrees in electrical engineering from Purdue University, West Lafayette, IN, in 1984, 1986, and 1990, respectively.

In 1985, he was a Graduate Instructor in the School of Electrical Engineering at Purdue University and he taught courses in communications and electronics. From 1990 to 1993, he was on the Academic Staff at the Department of Electrical and Electronic Engineering, University of Melbourne, Australia, where he was also a Member of the Center for Sensor Signal and Information Systems. Since 1993, he has been with the Department of Electrical and Electronic Engineering, Hong Kong University of Science and Technology (HKUST), Clear Water Bay, Hong Kong, where is currently an Associate Professor. During several semesters from 1995 to 1998, he served as the Departmental Director of the Undergraduate Studies program. His research interests include wireless personal and mobile communications, spread spectrum systems, optical fiber networks, multiuser detection, wireless multimedia communications, and CDMA systems. He is an Editor of the *Wireless Personal Communications Journal* and is also a Guest Editor of the 1999 *Wireless Personal Communications Journal's* Special Issue on Intelligent Multimedia Systems, Terminals, and Components.

Dr. Letaief has been an active member of various professional societies and has published papers in several journals and conference proceedings. In 1990, he was awarded the Mangoon Teaching Award at Purdue University. From 1995 to 1997, he was awarded the Teaching Excellence Appreciation Award by the School of Engineering, HKUST. He was also the 1998 University Recipient of the Michael G. Gale Medal for Distinguished Teaching (the highest teaching award at the university). He is the Editor for Wireless Systems of the *IEEE TRANSACTIONS ON COMMUNICATIONS* and a Technical Editor of the *IEEE Communications Magazine*. He has been a Guest Editor of the 1997 *IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS* Special Issue on Computer-Aided Modeling, Analysis, and Design of Communications Links. In addition, he served as the Technical Program Chair of the 1998 IEEE Mini-Conference on Communications Theory (CTMC'98) which was held in Sydney, Australia. He is currently serving as an Officer in the IEEE Communications Society Technical Committee on Personal Communications.



Ross D. Murch (S'85-M'87-SM'98) received the Ph.D. degree in electrical and electronic engineering from the University of Canterbury in 1990 for research involving electromagnetic inverse scattering.

From 1990 to 1992, he was a Post Doctoral Fellow at the Department of Mathematics and Computer Science, Dundee University, Scotland, where he worked on approximate approaches to electromagnetic scattering from three-dimensional objects. Currently he is an Associate Professor in the Department of Electrical and Electronic Engineering at the Hong Kong University of Science and Technology, Clear Water Bay, Hong Kong. He is also the founding Director of the Center for Wireless Information Technology, which began in August 1997. From August to December 1998, he was on sabbatical leave at Allgon Mobile Communications, Sweden and AT&T Research Labs, NJ. He also acts as a Consultant for various industrial projects on wireless communications and has two U.S. patents related to wireless communication. His most recent research activities center around the application of electromagnetics and signal processing to wireless communications. Research interests include: antenna design for mobile telephone handsets, smart antenna algorithms and propagation prediction for wireless communications.

Dr. Murch is an Editor of *IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS*, *Wireless Series*, and he has been involved in the organization of several international conferences. He is a Chartered Engineer, a Member of the Institute of Electrical Engineers, and also an URSI correspondent. He won an URSI Young Scientist Award in 1993 and an Engineering Teaching Excellence Appreciation award in 1996.

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Appendix 0002A

An Investigation into Time-Domain Approach for OFDM Channel Estimation

Hlaing Minn, *Student Member* and Vijay K. Bhargava, *Fellow, IEEE*

Abstract—A time-domain based channel estimation for OFDM system with pilot-data multiplexed scheme is investigated. As an approximation to linear minimum mean square estimator (LMMSE), a time-domain based channel estimation is proposed where intra-symbol time-averaging and most significant channel taps selection are applied. The relation and differences of the proposed method to DFT-based LMMSE methods are discussed. The performances of the proposed method, DFT-based LMMSE method [15] and the methods of [16] and [17] are evaluated in multipath fading channels. The simulation results show that proposed method achieves almost the same performance as DFT-based LMMSE method and better BER performance than the other methods while keeping less complexity.

Index Terms—Channel estimation, intra-symbol time-averaging, linear minimum mean square error estimator, most significant taps selection, orthogonal frequency division multiplexing (OFDM).

I. INTRODUCTION

ORTHOGONAL frequency division multiplexing (OFDM) [1]–[3] has recently achieved much popularity due to its desirable properties such as its robustness to multipath delay spread and impulse noise, its high data rate transmission capability with high bandwidth efficiency, and its feasibility in application of adaptive modulation and power allocation across the subcarriers according to the channel conditions. It has been adopted in wireline applications such as ADSL (Asymmetric Digital Subscriber Line) [4], broadcasting services such as European DAB (Digital Audio Broadcasting) [5], DVB-T (Terrestrial Digital Video Broadcasting) [6] and Japanese ISDB-T (Terrestrial Integrated Services Digital Broadcasting) [7], high rate wireless LAN standards such as ETSI HiperLAN 2 and IEEE 802.11(a), and multimedia wireless services such as Japanese MMAC (Multimedia Mobile Access Communications) [8], [9].

If noncoherent OFDM system is used, the system complexity will be reduced at the cost of 3–4 dB performance loss [10]. If coherent OFDM system is adopted, channel estimation becomes a requirement and usually pilot tones are used for channel (frequency response) estimation. Pilot tones can be inserted in all subcarriers of a particular OFDM symbol forming an OFDM training symbol, in which case training symbols are transmitted at an appropriate regular rate determined by the time varying nature of the wireless channel. Another approach is that, instead of using all subcarriers, the pilot tones are multiplexed

with data to form OFDM symbols. The subcarrier spacing between pilot tones is usually determined by the frequency selectivity of the wireless channel. The pilot multiplexing can be allowed for all OFDM symbols (i.e., all the time of transmission) or at an appropriate rate depending on the time selectivity of the wireless channel. In channel estimation using training symbols, decision directed approach has to be used. If complexity is affordable, time interpolation (e.g., [11]) can be used to improve the performance. If pilot tone multiplexing is used, the frequency interpolation has to be performed [12]–[14]. Similarly, time domain interpolation can be performed at the cost of complexity. Most of the channel estimation approaches may be viewed as DFT-based approaches [15], [16], [10], [11], where LS (Least square) channel (frequency response) estimates are fed to IFFT block to get time domain channel impulse response estimate, and then appropriately processed and transformed back to frequency domain by FFT. A DFT-based approach for OFDM system with transmit and receive antenna diversity has been discussed in [10].

In this paper, we investigate a time-domain channel estimation approach, namely FPTA (Frequency Pilot Time Average) [17] which applies intra-symbol time-domain averaging of identical parts of the pilot signal. We also propose a time-domain approach for OFDM channel estimation which achieves performance gain over LS or FPTA approaches.

The rest of the paper is organized as follows. In Section II, the OFDM system and LS channel estimation are presented for notational description. The method of time-averaging the identical parts of a pilot signal is briefly presented in Section III, and our analysis on this intra-symbol time-averaging is given in Section IV. Section V presents the proposed time-domain channel estimation and its relation to DFT-based approaches are discussed in Section VI. Simulation results are discussed in Section VII and conclusions are given in Section VIII.

II. SYSTEM DESCRIPTION

Suppose the pilot tones $P[k]$ are multiplexed with data $D[k]$ in all OFDM symbols at a pilot ratio $1/K$ (ratio of the number of pilot tones to the total number of subcarriers) where k is subcarrier index $(0, 1, \dots, N-1)$ with N being the total number of subcarriers, and $P[k]$ and $D[k]$ are zeros except at their corresponding subcarriers. Then the transmitted OFDM signal in discrete-time domain, excluding guard-interval, can be expressed as

$$\begin{aligned} s[n] &= \text{IFFT}_N\{P[k]\} + \text{IFFT}_N\{D[k]\} \\ &= p[n] + d[n] \end{aligned} \quad (1)$$

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The authors are with the Department of Electrical and Computer Engineering, University of Victoria, Victoria, B.C., Canada V8W 3P6.

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where $IFFT_N\{\}$ is N -point inverse Fast Fourier transform and n is the time-domain index ($0, 1, \dots, N-1$) of an OFDM symbol. Suppose the wireless channel has a discrete-time impulse response given by

$$h[n] = \sum_{l=0}^{L-1} \alpha_l \delta[n - \tau_l] \quad (2)$$

where

- α_l is complex path gain of l th path,
- τ_l is the delay of l th path, and
- L is the total number of channel paths.

For simplicity, time dependence nature of the channel impulse response is suppressed in the notation.

After passing through a multipath wireless channel, the time-domain received samples of an OFDM symbol, if appropriate cyclic prefix guard samples are used, is given by

$$r[n] = s[n] \otimes h[n] + w[n] \quad (3)$$

where $\otimes$ represents N -point circular convolution, $\{w[n]\}$ are independent and identically distributed (*iid*) AWGN samples with zero mean and variance of σ_w^2 . Assuming perfect synchronization, the FFT output frequency-domain subcarrier symbols can be expressed as

$$R[k] = FFT_N\{r[n]\} = H[k] P[k] + H[k] D[k] + W[k] \quad (4)$$

where $W[k] = FFT_N\{w[n]\}$ is frequency-domain AWGN noise samples with zero mean and variance $\sigma_f^2 = N\sigma_w^2$. Then the channel frequency response at the pilot tones can be estimated by

$$\hat{H}[m] = \frac{R[m]}{P[m]} = H[m] + \frac{W[m]}{P[m]} \quad (5)$$

where m is the subcarrier index for pilot tones. This channel estimate is called LS (least square) estimate. The channel responses at other subcarriers can be obtained by interpolation.

III. FREQUENCY PILOT TIME AVERAGE (FPTA) METHOD

In FPTA [17] approach, positive and negative alternatively polarized pilot tones are multiplexed with data at a pilot ratio of $1/K$. The frequency-domain pilot tones can be expressed as

$$P[k] = \begin{cases} (-1)^m A, & k = Km \\ 0, & k = Km + i \end{cases} \quad (6)$$

where

- K, m and i are integers, $1 \leq i \leq K-1, 0 \leq m \leq M-1$,
- A is pilot amplitude and
- $M = N/K$ is an integer.

The corresponding time-domain pilot samples can be expressed as

$$\begin{aligned} p[n] &= IFFT_N\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] W_N^{-nk} \\ &= \frac{A}{K} \sum_{r=0}^{K-1} \delta[n - Mr - M/2] \end{aligned} \quad (7)$$

where $W_N = \exp(-j2\pi/N)$.

In FPTA approach, since there are K identical parts of time-domain pilot samples, the corresponding parts of received samples are averaged over K parts. This intra-symbol time-domain averaging reduces the variance of noise samples by K times, i.e., $\sigma_{t,avg}^2 = \sigma_t^2/K$. Based on this observation, [17] defined time-domain pilot to noise ratio as $p[n]/\sigma_{t,avg}$ and compared it with frequency domain pilot to noise ratio $P[k]/\sigma_f$ resulting in the following expression

$$\frac{p[n]/\sigma_{t,avg}}{P[k]/\sigma_f} = \sqrt{\frac{N}{K}} \quad (8)$$

which was mentioned as the gain of FPTA approach over frequency domain approach such as LS estimation discussed in Section II. However, this result is optimistic and more details on it will be discussed in the following section.

IV. ANALYSIS OF INTRA-SYMBOL TIME AVERAGING IN FPTA

In order to investigate the pilot to noise ratio insightfully, we neglect the data part and consider only the pilot affected by multipath channel and Gaussian noise. Then the time-domain received samples vector of an OFDM signal can be given by

$$\mathbf{r} = \mathbf{p}' + \mathbf{w} \quad (9)$$

where $\mathbf{r} = [\mathbf{r}_0, \mathbf{r}_1, \dots, \mathbf{r}_{K-1}]$ with $\mathbf{r}_i = [r_i[0], r_i[1], \dots, r_i[M-1]]$, $\mathbf{p}'$ is N -point circular convolution of pilot signal $\mathbf{p}$ and channel impulse response $\mathbf{h}$ and can be expressed as $\mathbf{p}' = [\mathbf{p}'_0, \mathbf{p}'_1, \dots, \mathbf{p}'_{K-1}]$ with $\mathbf{p}'_i = [p'_i[0], p'_i[1], \dots, p'_i[M-1]]$ and $\mathbf{w} = [w[0], w[1], \dots, w[N-1]]$ is *iid* Gaussian noise samples.

Then averaging the received samples over K parts, assuming the channel impulse response is constant over the OFDM symbol interval, gives

$$\mathbf{r}_{avg} = \frac{1}{K} \sum_{l=0}^{K-1} \mathbf{r}_l = \mathbf{p}'_0 + \mathbf{w}_{avg} \quad (10)$$

where $\mathbf{w}_{avg} = [w_{avg}[0], w_{avg}[1], \dots, w_{avg}[M-1]]$ with $w_{avg}[i] = (1/K) \sum_{j=0}^{K-1} w[i + jM]$. Since $\{w[n]\}$ are *iid* zero mean complex Gaussian random variables with variance σ_t^2 , $\{w_{avg}[i]\}$ are *iid* zero mean complex Gaussian random variables with variance $\sigma_{t,avg}^2 = \sigma_t^2/K$.

Now, we investigate two possible approaches for channel estimation using intra-symbol time-averaged received samples. The first approach, which will be denoted by FPTA-1 in the rest, is to reconstruct the received samples of length N by repeating the $\mathbf{r}_{avg}$, K times. Then the reconstructed samples $\mathbf{y}$ can be expressed as

$$\mathbf{y} = [\mathbf{r}_{avg}, \mathbf{r}_{avg}, \dots, \mathbf{r}_{avg}] = \mathbf{p}' + \mathbf{v} \quad (11)$$

where $\mathbf{v} = [\mathbf{w}_{avg}, \mathbf{w}_{avg}, \dots, \mathbf{w}_{avg}]$ is K times repeated version of $\mathbf{w}_{avg}$. The corresponding frequency domain samples are

$$Y[k] = FFT_N\{y[n]\} = H[k] P[k] + V[k] \quad (12)$$

where $\{P[k]\}$ are original pilot tones, $H[k]$ is channel frequency response and $V[k] = FFT_N\{\mathbf{v}\}$ is frequency-domain zero

mean Gaussian noise term. The channel frequency responses at pilot tones can be estimated by

$$\hat{H}[m] = Y[m]/P[m] = H[m] + V[m]/P[m] \quad (13)$$

where $m = 0, K, 2K, \dots, (M-1)K$.

The variance of $V[m]$ can be given as follows:

$$\begin{aligned} \text{var}\{V[m]\} &= E\{V[m]V^*[m]\}, \\ &\quad m = 0, K, 2K, \dots, (M-1)K \\ &= E\left\{\sum_{i=0}^{N-1} v[i] W_N^{mi} \sum_{j=0}^{N-1} v^*[j] W_N^{-mj}\right\} \\ &= E\left\{\sum_{i=0}^{M-1} v[i] \sum_{l=0}^{K-1} W_N^{m(i+lM)} \sum_{j=0}^{M-1} v^*[j] \right. \\ &\quad \left. \cdot \sum_{n=0}^{K-1} W_N^{-m(j+nM)}\right\} \\ &= \sum_{i=0}^{M-1} \sum_{j=0}^{M-1} E\{v[i]v^*[j]\} \sum_{l=0}^{K-1} W_N^{m(i+lM)} \\ &\quad \cdot \sum_{n=0}^{K-1} W_N^{-m(j+nM)} \\ &= \sum_{i=0}^{M-1} \sigma_{t,avg}^2 \left| \sum_{l=0}^{K-1} W_N^{m(i+lM)} \right|^2 \\ &= \sum_{i=0}^{M-1} \sigma_{t,avg}^2 K^2 = N\sigma_t^2 \end{aligned} \quad (14)$$

which is the same as the variance of frequency-domain noise term without intra-symbol time-averaging (i.e., $W[k]$ in (4)). The pilot to noise ratio is also the same, given by $A/\sqrt{N\sigma_t^2}$.

The second possible approach, which will be denoted by FPTA-2 in the rest, is to use $FFT_N\{r_{avg}\}$ together with $FFT_N\{p_0\}$ where p_0 is one of the identical parts of time-domain pilot samples. The channel frequency response is estimated as follows:

$$P_0[k] = FFT_N\{p_0[n]\} = A/K, \quad \text{for } k = 0, 1, \dots, N-1 \quad (15)$$

$$R_{avg}[k] = FFT_N\{r_{avg}[n]\} = H[k]P_0[k] + W_{avg}[k] \quad (16)$$

$$\hat{H}[k] = R_{avg}[k]/P_0[k] = H[k] + W_{avg}[k]/P_0[k] \quad (17)$$

where $\{W_{avg}[k]\}$ are zero mean complex Gaussian noise terms and $\{P_0[k]\}$ are the equivalent pilot tones for this approach. It can be shown that the variance of $W_{avg}[k]$ is $M\sigma_{t,avg}^2 (= (M/K)\sigma_t^2)$. The pilot to noise ratio is the same as the previous case.

The channel estimation error is given by $e_H = \hat{H}[k] - H[k]$ which is the second term in $\hat{H}[k]$ equation. Since its mean is zero, the mean square error (*mse*) of channel estimate, $mse\{\hat{H}[k]\} = E[e_H e_H^*]$, is equal to the variance of the second term. Hence the channel estimation *mse*'s of LS method and intra-symbol time-averaging methods are the same and equal to $N\sigma_t^2/A^2$.

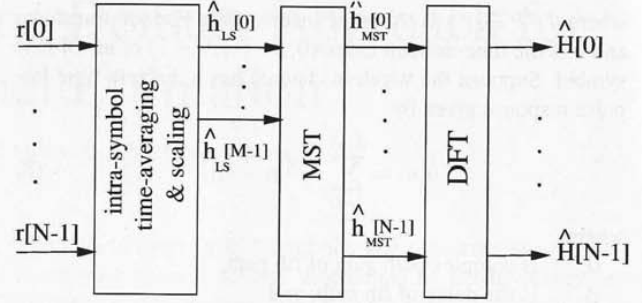


Fig. 1. Most Significant Taps (MST) Method.

The above discussion also indicates that intra-symbol time-averaging alone does not achieve channel estimation performance gain and the result in (8) is optimistic. Another way of explanation for the result in (8) is that it considers only one time-domain sample (after averaging) while other samples also affect the channel estimate. The result would be appropriate if $p[n]/\sigma_{t,avg}$ in (8) is obtained by averaging over all M samples (after averaging). In the rest, the gain of a channel estimation method over another (if not mentioned, LS method is assumed) will be expressed by the *mse* gain (i.e., the ratio of *mse*'s) or the (frequency-domain) pilot-to-noise power ratio instead of the pilot amplitude to noise standard deviation ratio.

V. MOST SIGNIFICANT TAPS APPROACH

For practical multipath wireless channels, there are not so many channel paths with significant energy (if compared to the FFT size N). Hence, among N samples (taps) of the channel impulse response estimate, many samples (taps) will have little or no energy at all except noise perturbation. Neglecting those nonsignificant channel taps in channel estimation may introduce some performance degradation if some of the channel energy is missed, but at the same time it will eliminate the noise perturbation from those taps. Usually total noise perturbation from those neglected channel estimate taps is much higher than the multipath energy contained in them, especially for low SNR values. Hence, neglecting those nonsignificant channel estimate taps can improve the channel estimation performance significantly and this fact is applied in the proposed method as shown in Fig. 1.

We consider the same scenario as in [17] where pilot tones are multiplexed with data in each OFDM symbol at a pilot ratio of $1/K$. The pilot tone used in the proposed *MST* approach is

$$P[k] = \sum_{m=0}^{M-1} A \delta[k - mK], \quad k = 0, 1, \dots, N-1 \quad (18)$$

and the corresponding time-domain samples contain K identical parts and are given by

$$p[n] = \sum_{m=0}^{K-1} \frac{A}{K} \delta[n - mM], \quad n = 0, 1, \dots, N-1 \quad (19)$$

If the maximum channel delay spread is less than the length of an identical part, which can be designed to satisfy this, then the time-domain received samples corresponding to the time-domain pilot samples contain K parts, each representing a

scaled channel impulse response for the respective part corrupted by AWGN. If the channel path gains remain essentially the same over an OFDM symbol interval, which is usually the case since OFDM systems are usually designed to satisfy this in order to maintain orthogonality among subcarriers, then the received samples corresponding to time-domain pilot samples contain K repeated version of scaled channel impulse response which are independently corrupted by AWGN. In order to choose most significant channel taps, those K parts can be averaged so that the noise variance is reduced by K times and more reliable most significant channel taps can be obtained. In mathematical expression, the time-domain received samples corresponding to time-domain pilot samples, (For simplicity, data part is neglected in the received samples expression. The mean of data part after averaging is zero), can be given by

$$\begin{aligned} r[n] &= h[n] \otimes p[n] + w[n] \\ &= \frac{A}{K} \sum_{m=0}^{K-1} h[n - mM] + w[n], \\ n &= 0, 1, \dots, N-1 \end{aligned} \quad (20)$$

After averaging, we have the noise-corrupted scaled channel impulse response

$$r_{avg}[n] = \frac{A}{K} h[n] + w_{avg}[n], \quad n = 0, 1, \dots, M-1 \quad (21)$$

Then the raw channel impulse response estimate is given by

$$\begin{aligned} \hat{h}_{LS}[n] &= \frac{K}{A} r_{avg}[n] = h[n] + \frac{K}{A} w_{avg}[n], \\ n &= 0, 1, \dots, M-1 \end{aligned} \quad (22)$$

Now, the most significant J channel taps are chosen as the J largest amplitude channel taps. Let the channel tap indexes for those most significant J taps be denoted by $n_0, n_1, \dots, n_{J-1}$. Then the time-domain channel impulse response estimate of proposed MST method is obtained by setting the other channel tap gains to zero as shown below:

$$\begin{aligned} \hat{h}_{MST}[n] &= \sum_{i=0}^{J-1} \hat{h}_{LS}[n_i] \delta[n - n_i], \\ n &= 0, 1, \dots, N-1 \end{aligned} \quad (23)$$

The channel frequency response estimate is directly obtained by applying FFT to $\{\hat{h}_{MST}[n]\}$ as

$$\begin{aligned} \hat{H}_{MST}[k] &= FFT_N\{\hat{h}_{MST}[n]\}, \quad k = 0, 1, \dots, N-1 \\ &= H[k] - H_{res}[k] + W_{avg}[k] \end{aligned} \quad (24)$$

where

$$\begin{aligned} H[k] &= \sum_{n=0}^{N-1} h[n] W_N^{kn} \\ H_{res}[k] &= \sum_{l=0}^{N-J-1} h[n'_l] W_N^{kn'_l} \\ W_{avg}[k] &= \frac{K}{A} \sum_{i=0}^{J-1} w_{avg}[n_i] W_N^{n_i k} \end{aligned}$$

with $\{n_i\}$ being the most significant channel tap indexes and $\{n'_l\}$ being the indexes of the other channel taps.

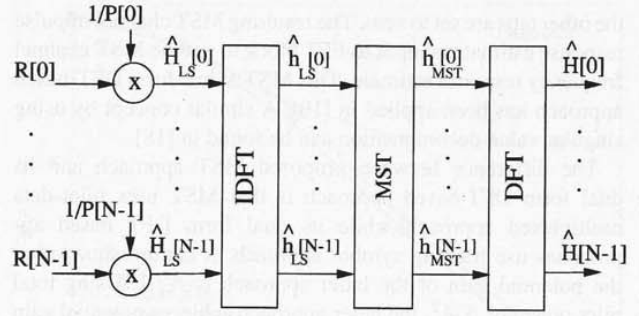


Fig. 2. MST's dual form DFT-based method.

By using suitable number of most significant taps, $H_{res}[k]$ can be kept very small (or zero) and hence the channel estimation error is mostly dominant (or totally caused) by the noise term $W_{avg}[k]$. For this case, the mse of the channel estimate can be approximated (or expressed) as

$$\begin{aligned} mse\{\hat{H}[k]\} &= E\{W_{avg}[k] W_{avg}^*[k]\} = \frac{K^2}{A^2} J \sigma_{t,avg}^2 \\ &= \frac{K J \sigma_t^2}{A^2} = \frac{K J}{N} mse\{\hat{H}_{LS}[k]\} \end{aligned} \quad (25)$$

Hence, the mse performance gain of proposed MST method over LS method (similarly over FPTA method) is ideally N/JK . The actual mse performance gain would be less than this amount due to the interference from data part and some (if any) excluded channel taps of nonsignificant energy.

The choice of the number of most significant taps J in the channel estimation depends on the application scenario. Broadcasting environment such as single frequency network can have larger number of multipaths with significant energy than non-broadcasting cases such as wireless LAN's environment. In any case, J should be chosen larger than the (designed) number of multipaths in order to prevent channel estimation error caused by missed channel taps. The channel estimation error caused by the noise from an additional tap in channel estimation is much less than that caused by missing one of the multipaths. A suitable choice for J may be two times or more of the (designed) number of multipaths (as will be seen in the simulation results) in order to ensure no channel energy missing.

Another MST tap selection approach can be implemented by selecting the channel taps whose energy is above a threshold. The threshold may be set as η times the maximum channel tap's energy in the raw channel estimate. The suitable choice of η depends on the operating SNR and more details will be discussed in the simulation results section.

VI. SIMILAR APPROACHES IN DFT-BASED METHODS

In this section, we relate the proposed MST method to the DFT-based approaches [15], [16], [10]. Essentially, due to the one-to-one relationship of DFT and IDFT, the MST method can be related to DFT-based approach as shown in Fig. 2. Consider the system with training symbol (i.e., pilot tones on all subcarriers). First, LS estimates are obtained and then input to IDFT block resulting in N samples LS estimate of channel impulse response. The largest amplitude J channel taps among the N samples (taps) are chosen as J most significant channel taps and

the other taps are set to zero. The resulting MST channel impulse response estimate is input to FFT block to get the MST channel frequency response estimate. This MST's dual form DFT-based approach has been applied in [10]. A similar concept by using singular value decomposition can be found in [18].

The difference between proposed MST approach and its dual form DFT-based approach is that MST uses pilot-data multiplexed approach while its dual form DFT based approaches use training symbol approach. It can be shown that the potential gain of the latter approach is N/J . Using total pilot power of NA^2 , the latter approach achieves potential gain of N/J whereas MST achieves potential gain of $N/(JK)$ with total pilot power of NA^2/K ; hence, on the basis of the same total pilot power, both methods have the same potential gain for channel estimation. Another difference is the complexity. In MST approach, operations involved are time-averaging, most significant channel taps selection and one FFT operation whereas its dual form DFT-based approach requires LS estimation, one IFFT operation, most significant channel taps selection and one FFT operation. Hence, the proposed MST approach saves some complexity.

One approach which is similar to MST's dual form DFT-based approach is the method of [16]. The difference is that [16] uses the first N_{cp} channel taps where N_{cp} is the number of cyclic prefix samples whereas MST's dual form DFT-based approach uses only J most significant channel taps, ($J < N_{cp}$). Since multipath channels usually have much less channel taps than N_{cp} , [16] has more noise perturbation than MST.

Other DFT-based approaches use linear minimum mean square error (LMMSE) estimators or approximate LMMSE estimators with reduced complexity [15]. If complexity can be afforded, these LMMSE or approximate LMMSE estimators can be implemented in MST approach. Let us consider a wide-sense stationary uncorrelated scattering (WSSUS) multipath channel with power delay profile given by $\sigma_{h_i}^2$ at delays of i OFDM sample intervals. Due to the uncorrelated multipaths, the correlation matrix of the channel impulse response becomes a diagonal matrix with the diagonal elements given by the power delay profile. Then the LMMSE's matrix multiplication (or scalar multiplication on corresponding channel taps) is performed as follows [19], [15]:

$$\hat{h}_{lmmse} = \text{diag} \left\{ \frac{\sigma_{h_0}^2}{\sigma_{h_0}^2 + \beta/SNR}, \frac{\sigma_{h_1}^2}{\sigma_{h_1}^2 + \beta/SNR}, \dots, \frac{\sigma_{h_{N-1}}^2}{\sigma_{h_{N-1}}^2 + \beta/SNR} \right\} \hat{h}_{LS} \quad (26)$$

where $\text{diag}\{a_0, a_1, \dots, a_{N-1}\}$ is a diagonal matrix with diagonal elements $[a_0, a_1, \dots, a_{N-1}]$. In DFT-based LMMSE approach, the first symbol is composed of all pilot tones of equal amplitude and the following data symbols are composed of M-ary QAM symbols on all subcarriers. Then for the initial channel estimation based on the training symbol, SNR/β in (26) is replaced by the pilot-to-noise power ratio. For the following channel estimation operated in the decision directed mode, β in (26) is given by $\beta = E[|X|^2] E[|X|^{-2}]$ where X is M-ary QAM symbol. For 64-QAM, $\beta = 2.6854$. For MST-LMMSE, SNR/β is replaced by its pilot-to-noise power

ratio (which is $1/K$ times that of DFT-based LMMSE due to its pilot ratio of $1/K$).

There is a delicate difference between DFT-based LMMSE and MST-LMMSE. In DFT-based LMMSE, all N samples of $\hat{h}_{LS}$ are corrupted by iid AWGN noise samples of zero mean and variance σ_t^2/A^2 . In MST-LMMSE, the first M samples of $\hat{h}_{LS}$ are corrupted by iid AWGN noise samples of zero mean and variance $K\sigma_t^2/A^2$ and the rest are zeros. Hence, if compared for the training symbol alone, DFT-based LMMSE will have a better channel estimation performance due to the larger pilot power used in channel estimation. But for the following data symbols, DFT-based LMMSE has to use decision directed approach which may degrade the channel estimation performance and BER performance due to the decision directed errors. For MST-LMMSE, the channel estimation and BER performances remain the same for all symbols.

In LMMSE implementation, since multipath channel correlation and SNR are usually unknown at the receiver, some fixed values have to be used for them. In [18], it is suggested to use high dummy SNR value and uniform multipath channel correlation which are robust to the channel correlation mismatch. If high dummy SNR value is used and uniform multipath channel correlation is assumed over guard interval, then LMMSE approach would be similar to [16]'s method. From (26) with high dummy SNR value, the LMMSE approach can be viewed as setting the channel taps with no energy to zeros and bypassing the other taps. Since most practical multipath channels have only a few significant paths (if compared to N), a suitable way of implementing an approximate LMMSE is choosing a predefined number of most significant taps and setting the others to zero which is the underlying idea of MST method. In complexity aspect, MST-LMMSE estimator still saves one N -point IFFT operation if compared to DFT-based LMMSE estimators.

VII. CHANNEL ESTIMATION PERFORMANCE EVALUATION BY SIMULATION

The proposed MST channel estimation is evaluated by computer simulation for two multipath fading channel models, namely Channel-A and Channel-B. The Channel-A is the ATTC (Advanced Television Technology Center) and the Grand Alliance DTV laboratory's ensemble E model whose channel impulse response for the static case is given by

$$h[n] = \delta[n] + 0.3162 \delta[n-2] + 0.1995 \delta[n-17] + 0.1296 \delta[n-36] + 0.1 \delta[n-75] + 0.1 \delta[n-137] \quad (27)$$

where unit delay is assumed to be the same as OFDM sample period. The Channel-B is a simplified version of DVB-T channel model P_1 [20] and its channel impulse response for the static case is given in Table I. In the simulation, the given channel tap gains of both channels represent the standard deviations of the gains of the complex Gaussian random variables.

The OFDM system parameters are as follows: the number of subcarriers $N = 8192$, pilot ratio $1/K = 1/8$, guard interval ratio = $1/8$ and carrier modulation 64-QAM. Perfect synchronization is assumed in order to observe the channel estimation performance alone. The considered methods are

TABLE I
CHANNEL IMPULSE RESPONSE FOR CHANNEL-B

Delay (OFDM samples)	Gain	Phase (radians)
0	0.2478	-2.5694
1	0.1287	-2.1208
3	0.3088	0.3548
4	0.4252	0.4187
5	0.4900	2.7201
7	0.0365	-1.4375
8	0.1197	1.1302
12	0.1948	-0.8092
17	0.4187	-0.1545
24	0.3170	-2.2159
29	0.2055	2.8372
49	0.1846	2.8641

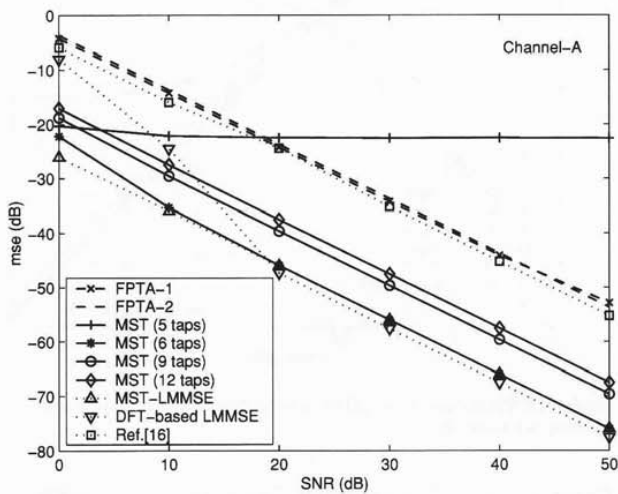


Fig. 3. Channel estimation mean square error (*mse*) in Channel-A.

MST, FPTA-1, FPTA-2, MST-LMMSE, DFT-based LMMSE and [16]. In the last two methods, one OFDM symbol is used for training symbol (pilot tones on all subcarriers) in every K OFDM symbols (i.e., the same pilot ratio $1/K$). In both LMMSE methods, ideal channel correlation and SNR values are used in order to evaluate the relative performance of MST method. In all methods, the same pilot tone symbol of $7 + j7$ signal point in 64-QAM constellation is used. For SNR of 40 and 50 dB, 10^5 independent simulation trials are used and for the other SNR values, 10^4 independent trials are used. In our simulation, floating point computation is used. However, it should be noted that due to the limited quantization levels in real hardware implementation, the computation error in real situation may degrade the performance obtained in our simulation.

Fig. 3 shows the *mse* performances of different channel estimation methods in Channel-A. The required interpolation in FPTA-1 is performed using Matlab's default *interp* function. Due to some noise reduction of this interpolation, the *mse* of FPTA-1 is slightly less than FPTA-2. The method of [16] has

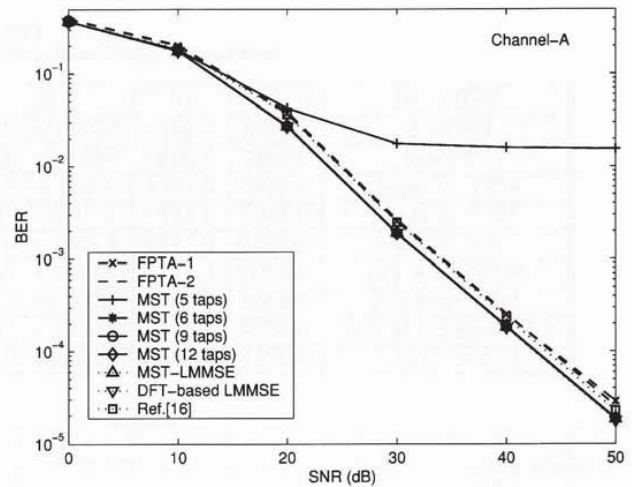


Fig. 4. BER performances of OFDM system with different channel estimation methods in Channel-A.

some improvement over FPTA approaches even though it is associated with decision directed errors. MST method with 6 taps, which is the same as the number of taps in Channel-A, has approximately 18 to 22 dB *mse* gain over FPTA approaches. MST with 5 taps shows an irreducible channel estimation error floor caused by missing some of the channel energy. It has a better performance than FPTA approaches for SNR less than 20 dB since for this SNR region the gain in noise suppression is greater than the loss of channel energy missing. However, for higher SNR region where noise has smaller impact than the channel energy missing, the channel estimation error floor results in a worse performance for MST with 5 taps. For the cases of MST with larger number of taps than the actual channel taps, MST achieves performance gain over FPTA and [16] approaches. The channel estimation performance gains are not as high as MST with 6 taps case due to the additional noise perturbation from the extra taps.

Also shown in Fig. 3 are the performances of MST-LMMSE and DFT-based LMMSE methods. LMMSE's approaches are of similar performance to MST with 6 taps case. For SNR less than 20 dB, DFT-based LMMSE has a worse performance than MST with 6 taps due to the decision directed errors. For higher SNR region, it has a slightly better performance due partly to the fact that its decision directed errors become insignificant and partly to the fact that MST suffers some interference from data part. MST-LMMSE has a slight performance gain over MST with 6 taps case for low SNR region and almost the same performance for high SNR region. This slight performance gain of MST-LMMSE in low SNR region is due to the utilization of (ideal) channel and noise statistics in MST-LMMSE approach by which the selection of channel taps are always correct, whereas in MST approach, the selection of channel taps may not be always correct due to the large noise perturbation.

The actual *mse* gain of MST method is slightly less than the potential gain of MST method described in previous section since data signal interference to training signal is neglected in the derivation of the potential gain. The actual *mse* gain is less in low SNR region than in high SNR region. The reason is

TABLE II
BER FOR DIFFERENT CHANNEL ESTIMATION METHODS IN CHANNEL-A

SNR (dB)	0	10	20	30	40	50
FPTA-1	0.3926	0.2015	3.7501×10^{-2}	2.5364×10^{-3}	2.4960×10^{-4}	2.8721×10^{-5}
FPTA-2	0.3957	0.2049	3.9122×10^{-2}	2.6389×10^{-3}	2.5513×10^{-4}	2.5377×10^{-5}
MST (5 taps)	0.3693	0.1814	4.262×10^{-2}	1.7651×10^{-2}	1.5964×10^{-2}	1.5441×10^{-2}
MST (6 taps)	0.3689	0.1770	2.7049×10^{-2}	1.8693×10^{-3}	1.8358×10^{-4}	1.8309×10^{-5}
MST (9 taps)	0.3695	0.1776	2.7291×10^{-2}	1.8841×10^{-3}	1.8505×10^{-4}	1.8502×10^{-5}
MST (12 taps)	0.3700	0.1781	2.7487×10^{-2}	1.8958×10^{-3}	1.8598×10^{-4}	1.8574×10^{-5}
MST-LMMSE	0.3689	0.1753	2.7023×10^{-2}	1.9081×10^{-3}	1.8309×10^{-4}	1.8299×10^{-5}
DFT-based LMMSE	0.3670	0.1762	2.7051×10^{-2}	1.8911×10^{-3}	1.8052×10^{-4}	1.7851×10^{-5}
Ref. [16]	0.3711	0.1887	3.6529×10^{-2}	2.4789×10^{-3}	2.3445×10^{-4}	2.2593×10^{-5}

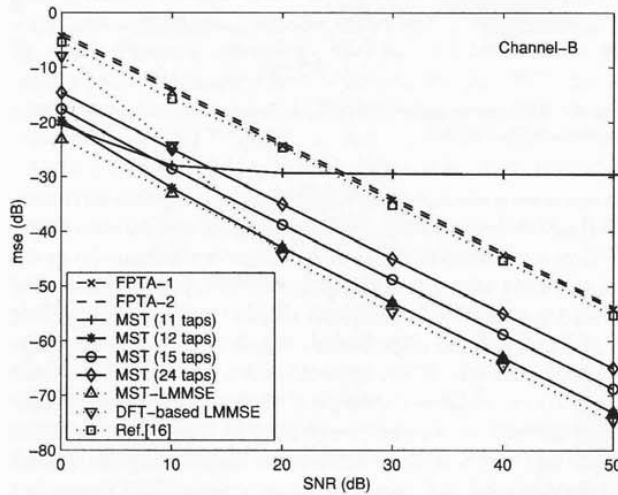


Fig. 5. Channel estimation mean square error (*mse*) in Channel-B.

that most significant channel taps selection would more likely choose some taps with no channel energy at low SNR region than at high SNR region, resulting in less *mse* gain at low SNR region.

In Fig. 4, the BER performances in Channel-A are presented for different channel estimation methods. Due to the channel estimation error floor and the BER sensitivity of 64-QAM to channel estimation error, MST with 5 taps case shows a BER floor while the others do not. The method of [16] is slightly better than FPTA approaches as is the case in *mse* performance. The other MST and LMMSE approaches have better BER performances than FPTA and [16] approaches, but among themselves, the BER performances are just slightly different and hence their BER curves are almost the same. For a better presentation of their slight differences, their BER values are tabulated in Table II. These slight differences can be ascribed to the corresponding different channel estimation performances. It is also noted that when the channel estimation *mse* is much smaller than $1/\text{SNR}$, then some deviation in channel estimation *mse* does not significantly affect the BER performance since noise is the dominant contributor to BER in this case.

The channel estimation performance and BER performance of the considered methods in Channel-B are plotted in Figs. 5 and 6 respectively and the BER values are given in Table III for the sake of clarity. The performances are of the same trend as in

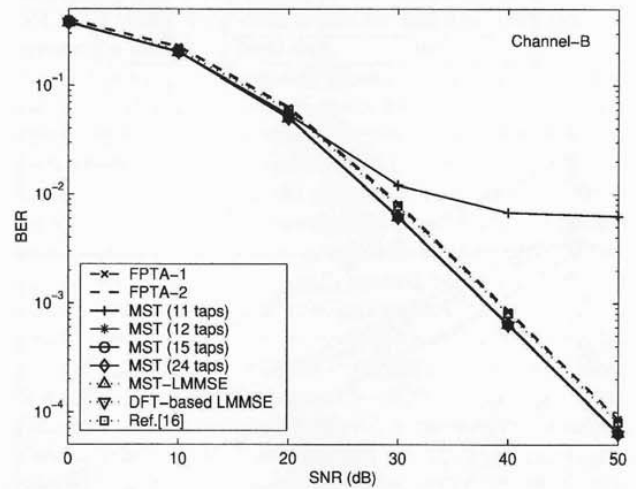


Fig. 6. BER performances of OFDM system with different channel estimation methods in Channel-B.

Channel-A. It is also noted that the channel estimation and BER performances in Channel-B is worse than Channel-A except for the case of missing some of the channel energy (i.e., MST with 5 taps in Channel-A and MST with 11 taps in Channel-B). This exception is due to the larger percentage of the channel energy missed in Channel-A than in Channel-B.

From the simulation results, it is clear that the number of MST taps J should not be smaller than the number of actual channel taps. On the other hand, using more taps may slightly degrade the BER performance due to the more noise perturbation. However, even up to double of the number of actual channel taps, MST has almost the same BER performance as using the number of actual channel taps. Hence, a suitable choice for J might be double of the (designed) number of channel taps for the considered channel environment.

Another way of selecting MST taps by threshold decision is also evaluated in Channel-B and the channel estimation *mse* and BER results are shown in Figs. 7 and 8 respectively. The performance of MST with 12 taps is also included for reference. It is observed that the suitable choice of threshold also depends on the operating SNR, as expected. The results also indicate that almost the same performance as MST with 12 taps case can be obtained if the threshold for an operating SNR is set within approximately 20 to 23 dB below the operating $1/\text{SNR}$. This low

TABLE III
BER FOR DIFFERENT CHANNEL ESTIMATION METHODS IN CHANNEL-B

SNR (dB)	0	10	20	30	40	50
FPTA-1	0.4014	0.2250	6.1986×10^{-2}	8.0725×10^{-3}	8.3867×10^{-4}	8.6296×10^{-5}
FPTA-2	0.4041	0.2283	6.3793×10^{-2}	8.3684×10^{-3}	8.6701×10^{-4}	8.6831×10^{-5}
MST (11 taps)	0.37828	0.2019	5.3833×10^{-2}	1.2092×10^{-2}	6.7836×10^{-3}	6.2976×10^{-3}
MST (12 taps)	0.3785	0.2011	4.9783×10^{-2}	6.1269×10^{-3}	6.2924×10^{-4}	6.3026×10^{-5}
MST (15 taps)	0.3790	0.2017	5.0082×10^{-2}	6.1707×10^{-3}	6.3410×10^{-4}	6.3421×10^{-5}
MST (24 taps)	0.3805	0.2032	5.0749×10^{-2}	6.2714×10^{-3}	6.4442×10^{-4}	6.4503×10^{-5}
MST-LMMSE	0.3781	0.2011	4.9847×10^{-2}	6.1509×10^{-3}	6.3000×10^{-4}	6.3806×10^{-5}
DFT-based LMMSE	0.3775	0.2018	4.969×10^{-2}	6.1261×10^{-3}	6.2774×10^{-4}	6.2568×10^{-5}
Ref. [16]	0.3809	0.2123	5.9644×10^{-2}	7.7969×10^{-3}	8.0554×10^{-4}	8.0195×10^{-5}

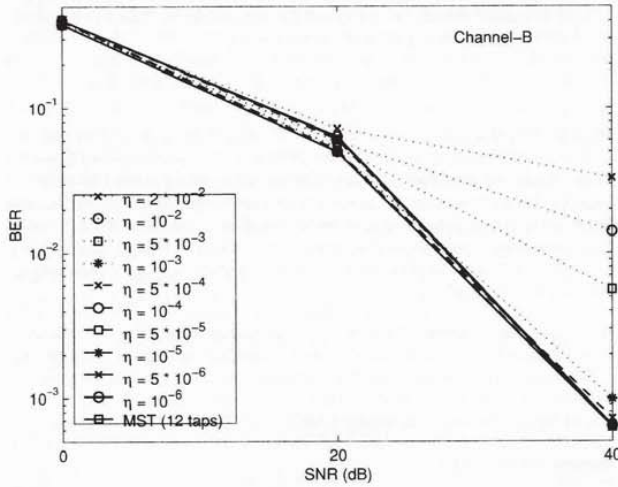


Fig. 7. Channel estimation mean square error (mse) for MST with threshold setting η .

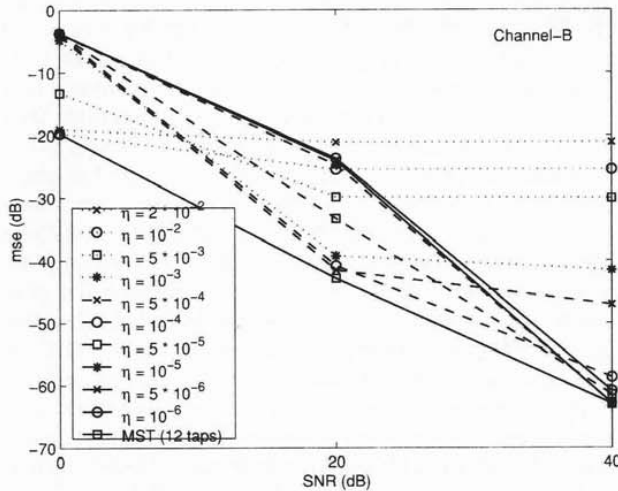


Fig. 8. BER performance of MST with threshold setting η .

value of the threshold in ensuring no channel energy miss is also due to the gain variations in the channel taps.

It is also remarked that for relatively slow fading environments, the decision feedback approaches such as [16] and

LMMSE can increase spectrum efficiency by using smaller pilot ratio. Similarly, MST approach can increase the spectrum efficiency by not inserting pilot tones in every OFDM symbol. For this case, the decision feedback approach as used in MST's dual form DFT-based approach can easily be applied for pure data symbols.

VIII. CONCLUSION

In this paper, intra-symbol time-domain averaging approaches for time-domain channel estimation of OFDM system are investigated. As an approximation to LMMSE type estimator, a time-domain approach, called MST, is proposed where intra-symbol time-domain averaging and most significant taps selection are applied. The similarities and differences of the proposed MST method versus DFT-based LMMSE methods are discussed. Two approaches for MST taps selection are described where the first uses a fixed number of taps while the second chooses all taps above a threshold. The simulation results suggest that a suitable choice for the fixed number of MST taps is using double of the (designed) number of channel taps for the considered channel environment. And a suitable choice of the threshold is within the range of approximately 20 to 23 dB below the operating $1/SNR$. In terms of BER performance in multipath fading channels, MST without channel energy missing and LMMSE approaches have almost the same performance which is better than the other considered approaches. In terms of complexity, MST approach keeps minimum complexity among the considered methods.

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H. Minn received his B.E.(Electronics) degree with honors from Yangon Institute of Technology, Myanmar, in 1995, and M.Eng. (Telecommunications) degree from Asian Institute of Technology (AIT), Thailand, in December 1997. He worked as a laboratory supervisor at Telecommunications Program, AIT, for one year. Since 1999, he has been a research assistant in the Department of Electrical and Computer Engineering at the University of Victoria, Canada. Currently, he is a Ph.D. student at University of Victoria. His research interests include wireless communications, OFDM, synchronization, channel estimation, equalization, modulation and error control coding.

Vijay K. Bhargava received the B.Sc., M.Sc., and Ph.D. degrees from Queen's University, Kingston, Canada in 1970, 1972 and 1974 respectively. Currently, he is a Professor of Electrical and Computer Engineering at the University of Victoria. He is a co-author of the book *Digital Communications by Satellite* (New York: Wiley, 1981) and co-editor of the IEEE Press Book *Reed-Solomon Codes and Their Applications*. He is an Editor-in-Chief of *WIRELESS PERSONAL COMMUNICATION*, a Kluwer Periodical. His research interests are in multi-media wireless communications.

Dr. Bhargava is very active in the IEEE and is currently the President of the IEEE Information Theory Society. He was Co-chair for IEEE ISIT'95, Technical Program Chair for IEEE ICC'99 and is the Chair of IEEE VTC'2002 Fall.

Dr. Bhargava is a Fellow of the B.C. Advanced Systems Institute, Engineering Institute of Canada (EIC), IEEE and the Royal Society of Canada. He is a recipient of the IEEE Centennial Medal (1984), IEEE Canada's McNaughton Gold Medal (1995), the IEEE Haraden Pratt Award (1999) and the IEEE Third Millennium Medal (2000).



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WILLIAM ZOU
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Cavell, Mertz & Davis, Inc.

10300 Eaton Place, Suite 200

Fairfax, VA 22030, USA

Tel: +1 703 591 0110

Fax: +1 703 591 0115

t.kuligowski@ieee.org

ASSOCIATE EDITORS

JAMES KUTZNER

Advanced Television Technology Center

(ATTC)

1330 Braddock Place, Suite 200

Alexandria, VA 22314 USA

Tel: +1 703 739 3866

Fax: +1 703 739 3230

jkutzner@attc.org

DR. YIYAN WU

Communications Research Centre

Canada

3701 Carling Ave., Box 11490,

Station H

Ottawa, Ontario, Canada K2H 8S2

Tel: +1 613 998 2870

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yiy.wu@crc.ca

DAVID H. LAYER

National Association of Broadcasters

1771 N. Street NW

Washington, DC 20036, USA

Tel: +1 202 429 5339

Fax: +1 202 775 4981

dlayer@nab.org

WILLIAM ZOU

Director, Video Systems Integration

Next Level Communications

6540 Lusk Blvd. Suite C-240

San Diego, CA 92121

Tel: +1 619 384 3028

Fax: +1 435 302 7377

email: wzou@nlc.com

email: wzou@inetworld.net

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Analysis of Polarization Diversity at Digital TV Indoor Receivers

Mercedes Sánchez and Manuel García Sánchez

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Abstract—The reduction of signal fading by using polarization diversity at a digital terrestrial TV receiver is evaluated. Two different diversity techniques, one operating at carrier (TV channel) level and the other at subcarrier level, are compared. A measurement campaign has been carried out in three different indoor environments using two antennas with orthogonal linear polarizations at the receiver. The polarization diversity gain has been calculated when maximum selection technique is used. As expected, the level crossing rate and the average fade duration are considerably reduced by the diversity system.

Index Terms—AFD, digital terrestrial television, DVB-T, LCR, polarization diversity, short term fading.

I. INTRODUCTION

DIGITAL television broadcasting is the adventure of our time. Ahead is the transition period. The major advantages of digital technology, baseband efficiency, flexibility, and RF performance, mean that all broadcasts will eventually fall to the digital wave. The only issue is how long it will take. A major catalyst of the digital era is the DVB project. It is a collaborative project that began in Europe, but now involves a large number of organizations from around the world. Not only has DVB scaled technical heights in digital television systems development, it has also taught us much about working together to achieve standardization [1]. In Europe there are different terrestrial broadcasting environments largely linked to the extent to which single frequency networks (SFN's) are needed, and the type of SFN's needed.

In the DVB-T system, different reception situations can be considered. With fixed reception the receiver is always at the same position. In a portable system the receiver can be moved around, but when it is in operation, it is in one fixed position. In a mobile system the receiver can be in movement during the reception. Although the DVB-T was not initially designed for the mobile reception, recently the Commercial module (CM) of DVB identified a growing interest in mobile applications of DVB-T. Therefore, the Steering Board asked the Technical Module (TM) to investigate the potential of the DVB-T transmission standard for mobile reception. The performance of a mobile terrestrial digital TV receiver, as other mobile or portable radio receivers that use omnidirectional antennas, is affected by signal fading. As a result of multipath propagation several radio signals interfere at the receiver, and the received signal level

varies with the position of the receiver and the frequency. Multipath propagation effects can be reduced by the use of directional antennas, but this is not the case for mobile terrestrial digital TV receivers, where omnidirectional antennas have to be used. In this case, signal fading can be mitigated with either "time-interleaving" or diversity techniques. The OFDM modulation multiplexes many narrow band signals into one wide-band signal using a fast Fourier transform (FFT) [1]. This type of modulation reduces the impact of the frequency selective fading caused by multipath propagation in the terrestrial VHF/UHF channel. However, a spatial fading related to the location of the receiving antenna is still present in the DVB-T channel.

Polarization diversity is expected to feature prominently in future communications systems because it provides a means of realizing two independently fading signals without the need of physical separation of antennas. The mechanism of decorrelation for the signals in each polarization is the multiple reflections undertaken between the transmitter and the receiver. The reflection coefficients for both orthogonal polarizations are in general different, which results in the phases of orthogonal polarizations undergoing different changes for each or at least some of the reflections. Some studies [2] have shown that the transmission of dual polarizations with comparable powers provides received signals with comparable levels and small cross-correlation, hence the full benefit of diversity can be realized. An easier implementation of polarization diversity is based on the transmission of only one polarization and the reception of both, the transmitted and the orthogonal one. This latter case is the one considered in this paper, where the performance of two different diversity techniques at the receiver is evaluated. One of them operates at the carrier level (TV channel) and the other one operates at the subcarrier level. A measurement campaign has been carried out in three different indoor environments, in the frequency band assigned for terrestrial digital video broadcasting (DVB-T) [1]. Two orthogonal linear polarized antennas, vertical and horizontal, have been used at the receiver. To combine the signal from the two antennas at the receiver a maximum level selection (MAP) technique has been considered [3]. The diversity gain, the average fade duration (AFD) and the level crossing rate (LCR), have been obtained and are analyzed.

The measurements were carried out with both transmit and receiver antennas indoors, in a situation similar to that of a domestic gap-filler. Further work is needed to extend the conclusions presented in this paper to other propagation environments.

II. MEASUREMENT SETUP AND ENVIRONMENT DESCRIPTION

In order to characterize the radio channel in the 470–862 MHz frequency band a HP 8510C Vector Network

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The authors are with the Departamento de Tecnologías de Comunicaciones, E.T.S.E. Telecomunicación, Universidade de Vigo, Campus Universitario S/N, 36200 Vigo, Spain.

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An Investigation into Time-Domain Approach for OFDM Channel Estimation

Hlaing Minn, *Student Member* and Vijay K. Bhargava, *Fellow, IEEE*

Abstract—A time-domain based channel estimation for OFDM system with pilot-data multiplexed scheme is investigated. As an approximation to linear minimum mean square estimator (LMMSE), a time-domain based channel estimation is proposed where intra-symbol time-averaging and most significant channel taps selection are applied. The relation and differences of the proposed method to DFT-based LMMSE methods are discussed. The performances of the proposed method, DFT-based LMMSE method [15] and the methods of [16] and [17] are evaluated in multipath fading channels. The simulation results show that proposed method achieves almost the same performance as DFT-based LMMSE method and better BER performance than the other methods while keeping less complexity.

Index Terms—Channel estimation, intra-symbol time-averaging, linear minimum mean square error estimator, most significant taps selection, orthogonal frequency division multiplexing (OFDM).

I. INTRODUCTION

ORTHOGONAL frequency division multiplexing (OFDM) [1]–[3] has recently achieved much popularity due to its desirable properties such as its robustness to multipath delay spread and impulse noise, its high data rate transmission capability with high bandwidth efficiency, and its feasibility in application of adaptive modulation and power allocation across the subcarriers according to the channel conditions. It has been adopted in wireline applications such as ADSL (Asymmetric Digital Subscriber Line) [4], broadcasting services such as European DAB (Digital Audio Broadcasting) [5], DVB-T (Terrestrial Digital Video Broadcasting) [6] and Japanese ISDB-T (Terrestrial Integrated Services Digital Broadcasting) [7], high rate wireless LAN standards such as ETSI HiperLAN 2 and IEEE 802.11(a), and multimedia wireless services such as Japanese MMAC (Multimedia Mobile Access Communications) [8], [9].

If noncoherent OFDM system is used, the system complexity will be reduced at the cost of 3–4 dB performance loss [10]. If coherent OFDM system is adopted, channel estimation becomes a requirement and usually pilot tones are used for channel (frequency response) estimation. Pilot tones can be inserted in all subcarriers of a particular OFDM symbol forming an OFDM training symbol, in which case training symbols are transmitted at an appropriate regular rate determined by the time varying nature of the wireless channel. Another approach is that, instead of using all subcarriers, the pilot tones are multiplexed

with data to form OFDM symbols. The subcarrier spacing between pilot tones is usually determined by the frequency selectivity of the wireless channel. The pilot multiplexing can be allowed for all OFDM symbols (i.e., all the time of transmission) or at an appropriate rate depending on the time selectivity of the wireless channel. In channel estimation using training symbols, decision directed approach has to be used. If complexity is affordable, time interpolation (e.g., [11]) can be used to improve the performance. If pilot tone multiplexing is used, the frequency interpolation has to be performed [12]–[14]. Similarly, time domain interpolation can be performed at the cost of complexity. Most of the channel estimation approaches may be viewed as DFT-based approaches [15], [16], [10], [11], where LS (Least square) channel (frequency response) estimates are fed to IFFT block to get time domain channel impulse response estimate, and then appropriately processed and transformed back to frequency domain by FFT. A DFT-based approach for OFDM system with transmit and receive antenna diversity has been discussed in [10].

In this paper, we investigate a time-domain channel estimation approach, namely FPTA (Frequency Pilot Time Average) [17] which applies intra-symbol time-domain averaging of identical parts of the pilot signal. We also propose a time-domain approach for OFDM channel estimation which achieves performance gain over LS or FPTA approaches.

The rest of the paper is organized as follows. In Section II, the OFDM system and LS channel estimation are presented for notational description. The method of time-averaging the identical parts of a pilot signal is briefly presented in Section III, and our analysis on this intra-symbol time-averaging is given in Section IV. Section V presents the proposed time-domain channel estimation and its relation to DFT-based approaches are discussed in Section VI. Simulation results are discussed in Section VII and conclusions are given in Section VIII.

II. SYSTEM DESCRIPTION

Suppose the pilot tones $P[k]$ are multiplexed with data $D[k]$ in all OFDM symbols at a pilot ratio $1/K$ (ratio of the number of pilot tones to the total number of subcarriers) where k is subcarrier index $(0, 1, \dots, N-1)$ with N being the total number of subcarriers, and $P[k]$ and $D[k]$ are zeros except at their corresponding subcarriers. Then the transmitted OFDM signal in discrete-time domain, excluding guard-interval, can be expressed as

$$\begin{aligned} s[n] &= \text{IFFT}_N\{P[k]\} + \text{IFFT}_N\{D[k]\} \\ &= p[n] + d[n] \end{aligned} \quad (1)$$

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The authors are with the Department of Electrical and Computer Engineering, University of Victoria, Victoria, B.C., Canada V8W 3P6.

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where $IFFT_N\{\}$ is N -point inverse Fast Fourier transform and n is the time-domain index $(0, 1, \dots, N-1)$ of an OFDM symbol. Suppose the wireless channel has a discrete-time impulse response given by

$$h[n] = \sum_{l=0}^{L-1} \alpha_l \delta[n - \tau_l] \quad (2)$$

where

- α_l is complex path gain of l th path,
- τ_l is the delay of l th path, and
- L is the total number of channel paths.

For simplicity, time dependence nature of the channel impulse response is suppressed in the notation.

After passing through a multipath wireless channel, the time-domain received samples of an OFDM symbol, if appropriate cyclic prefix guard samples are used, is given by

$$r[n] = s[n] \otimes h[n] + w[n] \quad (3)$$

where $\otimes$ represents N -point circular convolution, $\{w[n]\}$ are independent and identically distributed (*iid*) AWGN samples with zero mean and variance of σ_t^2 . Assuming perfect synchronization, the FFT output frequency-domain subcarrier symbols can be expressed as

$$R[k] = FFT_N\{r[n]\} = H[k] P[k] + H[k] D[k] + W[k] \quad (4)$$

where $W[k] = FFT_N\{w[n]\}$ is frequency-domain AWGN noise samples with zero mean and variance $\sigma_f^2 = N\sigma_t^2$. Then the channel frequency response at the pilot tones can be estimated by

$$\hat{H}[m] = \frac{R[m]}{P[m]} = H[m] + \frac{W[m]}{P[m]} \quad (5)$$

where m is the subcarrier index for pilot tones. This channel estimate is called LS (least square) estimate. The channel responses at other subcarriers can be obtained by interpolation.

III. FREQUENCY PILOT TIME AVERAGE (FPTA) METHOD

In FPTA [17] approach, positive and negative alternatively polarized pilot tones are multiplexed with data at a pilot ratio of $1/K$. The frequency-domain pilot tones can be expressed as

$$P[k] = \begin{cases} (-1)^m A, & k = Km \\ 0, & k = Km + i \end{cases} \quad (6)$$

where

- K, m and i are integers, $1 \leq i \leq K-1, 0 \leq m \leq M-1$,
- A is pilot amplitude and
- $M = N/K$ is an integer.

The corresponding time-domain pilot samples can be expressed as

$$\begin{aligned} p[n] &= IFFT_N\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] W_N^{-nk} \\ &= \frac{A}{K} \sum_{r=0}^{K-1} \delta[n - Mr - M/2] \end{aligned} \quad (7)$$

where $W_N = \exp(-j2\pi/N)$.

In FPTA approach, since there are K identical parts of time-domain pilot samples, the corresponding parts of received samples are averaged over K parts. This intra-symbol time-domain averaging reduces the variance of noise samples by K times, i.e., $\sigma_{t,avg}^2 = \sigma_t^2/K$. Based on this observation, [17] defined time-domain pilot to noise ratio as $p[n]/\sigma_{t,avg}$ and compared it with frequency domain pilot to noise ratio $P[k]/\sigma_f$ resulting in the following expression

$$\frac{p[n]/\sigma_{t,avg}}{P[k]/\sigma_f} = \sqrt{\frac{N}{K}} \quad (8)$$

which was mentioned as the gain of FPTA approach over frequency domain approach such as LS estimation discussed in Section II. However, this result is optimistic and more details on it will be discussed in the following section.

IV. ANALYSIS OF INTRA-SYMBOL TIME AVERAGING IN FPTA

In order to investigate the pilot to noise ratio insightfully, we neglect the data part and consider only the pilot affected by multipath channel and Gaussian noise. Then the time-domain received samples vector of an OFDM signal can be given by

$$\mathbf{r} = \mathbf{p}' + \mathbf{w} \quad (9)$$

where $\mathbf{r} = [\mathbf{r}_0, \mathbf{r}_1, \dots, \mathbf{r}_{K-1}]$ with $\mathbf{r}_i = [r_i[0], r_i[1], \dots, r_i[M-1]]$, $\mathbf{p}'$ is N -point circular convolution of pilot signal $\mathbf{p}$ and channel impulse response $\mathbf{h}$ and can be expressed as $\mathbf{p}' = [\mathbf{p}'_0, \mathbf{p}'_1, \dots, \mathbf{p}'_{K-1}]$ with $\mathbf{p}'_i = [p'_i[0], p'_i[1], \dots, p'_i[M-1]]$ and $\mathbf{w} = [w[0], w[1], \dots, w[N-1]]$ is *iid* Gaussian noise samples.

Then averaging the received samples over K parts, assuming the channel impulse response is constant over the OFDM symbol interval, gives

$$\mathbf{r}_{avg} = \frac{1}{K} \sum_{l=0}^{K-1} \mathbf{r}_l = \mathbf{p}'_0 + \mathbf{w}_{avg} \quad (10)$$

where $\mathbf{w}_{avg} = [w_{avg}[0], w_{avg}[1], \dots, w_{avg}[M-1]]$ with $w_{avg}[i] = (1/K) \sum_{j=0}^{K-1} w[i + jM]$. Since $\{w[n]\}$ are *iid* zero mean complex Gaussian random variables with variance σ_t^2 , $\{w_{avg}[i]\}$ are *iid* zero mean complex Gaussian random variables with variance $\sigma_{t,avg}^2 = \sigma_t^2/K$.

Now, we investigate two possible approaches for channel estimation using intra-symbol time-averaged received samples. The first approach, which will be denoted by FPTA-1 in the rest, is to reconstruct the received samples of length N by repeating the $\mathbf{r}_{avg}$, K times. Then the reconstructed samples $\mathbf{y}$ can be expressed as

$$\mathbf{y} = [\mathbf{r}_{avg}, \mathbf{r}_{avg}, \dots, \mathbf{r}_{avg}] = \mathbf{p}' + \mathbf{v} \quad (11)$$

where $\mathbf{v} = [\mathbf{w}_{avg}, \mathbf{w}_{avg}, \dots, \mathbf{w}_{avg}]$ is K times repeated version of $\mathbf{w}_{avg}$. The corresponding frequency domain samples are

$$Y[k] = FFT_N\{y[n]\} = H[k] P[k] + V[k] \quad (12)$$

where $\{P[k]\}$ are original pilot tones, $H[k]$ is channel frequency response and $V[k] = FFT_N\{\mathbf{v}\}$ is frequency-domain zero

mean Gaussian noise term. The channel frequency responses at pilot tones can be estimated by

$$\hat{H}[m] = Y[m]/P[m] = H[m] + V[m]/P[m] \quad (13)$$

where $m = 0, K, 2K, \dots, (M-1)K$.

The variance of $V[m]$ can be given as follows:

$$\begin{aligned} \text{var}\{V[m]\} &= E\{V[m] V^*[m]\}, \\ &\quad m = 0, K, 2K, \dots, (M-1)K \\ &= E \left\{ \sum_{i=0}^{N-1} v[i] W_N^{mi} \sum_{j=0}^{N-1} v^*[j] W_N^{-mj} \right\} \\ &= E \left\{ \sum_{i=0}^{M-1} v[i] \sum_{l=0}^{K-1} W_N^{m(i+lM)} \sum_{j=0}^{M-1} v^*[j] \right. \\ &\quad \left. \cdot \sum_{n=0}^{K-1} W_N^{-m(j+nM)} \right\} \\ &= \sum_{i=0}^{M-1} \sum_{j=0}^{M-1} E\{v[i] v^*[j]\} \sum_{l=0}^{K-1} W_N^{m(i+lM)} \\ &\quad \cdot \sum_{n=0}^{K-1} W_N^{-m(j+nM)} \\ &= \sum_{i=0}^{M-1} \sigma_{t,avg}^2 \left| \sum_{l=0}^{K-1} W_N^{m(i+lM)} \right|^2 \\ &= \sum_{i=0}^{M-1} \sigma_{t,avg}^2 K^2 = N \sigma_t^2 \end{aligned} \quad (14)$$

which is the same as the variance of frequency-domain noise term without intra-symbol time-averaging (i.e., $W[k]$ in (4)). The pilot to noise ratio is also the same, given by $A/\sqrt{N\sigma_t^2}$.

The second possible approach, which will be denoted by FPTA-2 in the rest, is to use $FFT_N\{r_{avg}\}$ together with $FFT_N\{p_0\}$ where p_0 is one of the identical parts of time-domain pilot samples. The channel frequency response is estimated as follows:

$$P_0[k] = FFT_N\{p_0[n]\} = A/K, \quad \text{for } k = 0, 1, \dots, N-1 \quad (15)$$

$$R_{avg}[k] = FFT_N\{r_{avg}[n]\} = H[k] P_0[k] + W_{avg}[k] \quad (16)$$

$$\hat{H}[k] = R_{avg}[k]/P_0[k] = H[k] + W_{avg}[k]/P_0[k] \quad (17)$$

where $\{W_{avg}[k]\}$ are zero mean complex Gaussian noise terms and $\{P_0[k]\}$ are the equivalent pilot tones for this approach. It can be shown that the variance of $W_{avg}[k]$ is $M\sigma_{t,avg}^2 (= (M/K)\sigma_t^2)$. The pilot to noise ratio is the same as the previous case.

The channel estimation error is given by $e_H = \hat{H}[k] - H[k]$ which is the second term in $\hat{H}[k]$ equation. Since its mean is zero, the mean square error (mse) of channel estimate, $mse\{\hat{H}[k]\} = E\{e_H e_H^*\}$, is equal to the variance of the second term. Hence the channel estimation mse 's of LS method and intra-symbol time-averaging methods are the same and equal to $N\sigma_t^2/A^2$.

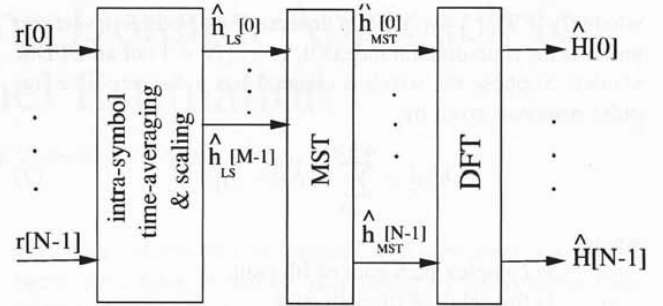


Fig. 1. Most Significant Taps (MST) Method.

The above discussion also indicates that intra-symbol time-averaging alone does not achieve channel estimation performance gain and the result in (8) is optimistic. Another way of explanation for the result in (8) is that it considers only one time-domain sample (after averaging) while other samples also affect the channel estimate. The result would be appropriate if $p[n]/\sigma_{t,avg}$ in (8) is obtained by averaging over all M samples (after averaging). In the rest, the gain of a channel estimation method over another (if not mentioned, LS method is assumed) will be expressed by the mse gain (i.e., the ratio of mse 's) or the (frequency-domain) pilot-to-noise power ratio instead of the pilot amplitude to noise standard deviation ratio.

V. MOST SIGNIFICANT TAPS APPROACH

For practical multipath wireless channels, there are not so many channel paths with significant energy (if compared to the FFT size N). Hence, among N samples (taps) of the channel impulse response estimate, many samples (taps) will have little or no energy at all except noise perturbation. Neglecting those nonsignificant channel taps in channel estimation may introduce some performance degradation if some of the channel energy is missed, but at the same time it will eliminate the noise perturbation from those taps. Usually total noise perturbation from those neglected channel estimate taps is much higher than the multipath energy contained in them, especially for low SNR values. Hence, neglecting those nonsignificant channel estimate taps can improve the channel estimation performance significantly and this fact is applied in the proposed method as shown in Fig. 1.

We consider the same scenario as in [17] where pilot tones are multiplexed with data in each OFDM symbol at a pilot ratio of $1/K$. The pilot tone used in the proposed MST approach is

$$P[k] = \sum_{m=0}^{M-1} A \delta[k - mK], \quad k = 0, 1, \dots, N-1 \quad (18)$$

and the corresponding time-domain samples contain K identical parts and are given by

$$p[n] = \sum_{m=0}^{K-1} \frac{A}{K} \delta[n - mM], \quad n = 0, 1, \dots, N-1 \quad (19)$$

If the maximum channel delay spread is less than the length of an identical part, which can be designed to satisfy this, then the time-domain received samples corresponding to the time-domain pilot samples contain K parts, each representing a

scaled channel impulse response for the respective part corrupted by AWGN. If the channel path gains remain essentially the same over an OFDM symbol interval, which is usually the case since OFDM systems are usually designed to satisfy this in order to maintain orthogonality among subcarriers, then the received samples corresponding to time-domain pilot samples contain K repeated version of scaled channel impulse response which are independently corrupted by AWGN. In order to choose most significant channel taps, those K parts can be averaged so that the noise variance is reduced by K times and more reliable most significant channel taps can be obtained. In mathematical expression, the time-domain received samples corresponding to time-domain pilot samples, (For simplicity, data part is neglected in the received samples expression. The mean of data part after averaging is zero), can be given by

$$\begin{aligned} r[n] &= h[n] \otimes p[n] + w[n] \\ &= \frac{A}{K} \sum_{m=0}^{K-1} h[n - mM] + w[n], \\ n &= 0, 1, \dots, N-1 \end{aligned} \quad (20)$$

After averaging, we have the noise-corrupted scaled channel impulse response

$$r_{avg}[n] = \frac{A}{K} h[n] + w_{avg}[n], \quad n = 0, 1, \dots, M-1 \quad (21)$$

Then the raw channel impulse response estimate is given by

$$\hat{h}_{LS}[n] = \frac{K}{A} r_{avg}[n] = h[n] + \frac{K}{A} w_{avg}[n], \quad n = 0, 1, \dots, M-1 \quad (22)$$

Now, the most significant J channel taps are chosen as the J largest amplitude channel taps. Let the channel tap indexes for those most significant J taps be denoted by $n_0, n_1, \dots, n_{J-1}$. Then the time-domain channel impulse response estimate of proposed MST method is obtained by setting the other channel tap gains to zero as shown below:

$$\begin{aligned} \hat{h}_{MST}[n] &= \sum_{i=0}^{J-1} \hat{h}_{LS}[n_i] \delta[n - n_i], \\ n &= 0, 1, \dots, N-1 \end{aligned} \quad (23)$$

The channel frequency response estimate is directly obtained by applying FFT to $\{\hat{h}_{MST}[n]\}$ as

$$\begin{aligned} \hat{H}_{MST}[k] &= FFT_N\{\hat{h}_{MST}[n]\}, \quad k = 0, 1, \dots, N-1 \\ &= H[k] - H_{res}[k] + W_{avg}[k] \end{aligned} \quad (24)$$

where

$$\begin{aligned} H[k] &= \sum_{n=0}^{N-1} h[n] W_N^{kn} \\ H_{res}[k] &= \sum_{l=0}^{N-J-1} h[n'_l] W_N^{kn'_l} \\ W_{avg}[k] &= \frac{K}{A} \sum_{i=0}^{J-1} w_{avg}[n_i] W_N^{n_i k} \end{aligned}$$

with $\{n_i\}$ being the most significant channel tap indexes and $\{n'_i\}$ being the indexes of the other channel taps.

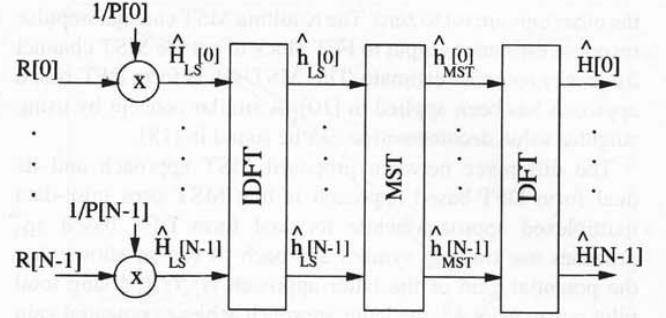


Fig. 2. MST's dual form DFT-based method.

By using suitable number of most significant taps, $H_{res}[k]$ can be kept very small (or zero) and hence the channel estimation error is mostly dominant (or totally caused) by the noise term $W_{avg}[k]$. For this case, the mse of the channel estimate can be approximated (or expressed) as

$$\begin{aligned} mse\{\hat{H}[k]\} &= E\{W_{avg}[k] W_{avg}^*[k]\} = \frac{K^2}{A^2} J \sigma_{t,avg}^2 \\ &= \frac{K J \sigma_t^2}{A^2} = \frac{K J}{N} mse\{\hat{H}_{LS}[k]\} \end{aligned} \quad (25)$$

Hence, the mse performance gain of proposed MST method over LS method (similarly over FPTA method) is ideally N/JK . The actual mse performance gain would be less than this amount due to the interference from data part and some (if any) excluded channel taps of nonsignificant energy.

The choice of the number of most significant taps J in the channel estimation depends on the application scenario. Broadcasting environment such as single frequency network can have larger number of multipaths with significant energy than non-broadcasting cases such as wireless LAN's environment. In any case, J should be chosen larger than the (designed) number of multipaths in order to prevent channel estimation error caused by missed channel taps. The channel estimation error caused by the noise from an additional tap in channel estimation is much less than that caused by missing one of the multipaths. A suitable choice for J may be two times or more of the (designed) number of multipaths (as will be seen in the simulation results) in order to ensure no channel energy missing.

Another MST tap selection approach can be implemented by selecting the channel taps whose energy is above a threshold. The threshold may be set as η times the maximum channel tap's energy in the raw channel estimate. The suitable choice of η depends on the operating SNR and more details will be discussed in the simulation results section.

VI. SIMILAR APPROACHES IN DFT-BASED METHODS

In this section, we relate the proposed MST method to the DFT-based approaches [15], [16], [10]. Essentially, due to the one-to-one relationship of DFT and IDFT, the MST method can be related to DFT-based approach as shown in Fig. 2. Consider the system with training symbol (i.e., pilot tones on all subcarriers). First, LS estimates are obtained and then input to IDFT block resulting in N samples LS estimate of channel impulse response. The largest amplitude J channel taps among the N samples (taps) are chosen as J most significant channel taps and

the other taps are set to zero. The resulting MST channel impulse response estimate is input to FFT block to get the MST channel frequency response estimate. This MST's dual form DFT-based approach has been applied in [10]. A similar concept by using singular value decomposition can be found in [18].

The difference between proposed MST approach and its dual form DFT-based approach is that MST uses pilot-data multiplexed approach while its dual form DFT based approaches use training symbol approach. It can be shown that the potential gain of the latter approach is N/J . Using total pilot power of NA^2 , the latter approach achieves potential gain of N/J whereas MST achieves potential gain of $N/(JK)$ with total pilot power of NA^2/K ; hence, on the basis of the same total pilot power, both methods have the same potential gain for channel estimation. Another difference is the complexity. In MST approach, operations involved are time-averaging, most significant channel taps selection and one FFT operation whereas its dual form DFT-based approach requires LS estimation, one IFFT operation, most significant channel taps selection and one FFT operation. Hence, the proposed MST approach saves some complexity.

One approach which is similar to MST's dual form DFT-based approach is the method of [16]. The difference is that [16] uses the first N_{cp} channel taps where N_{cp} is the number of cyclic prefix samples whereas MST's dual form DFT-based approach uses only J most significant channel taps, ($J < N_{cp}$). Since multipath channels usually have much less channel taps than N_{cp} , [16] has more noise perturbation than MST.

Other DFT-based approaches use linear minimum mean square error (LMMSE) estimators or approximate LMMSE estimators with reduced complexity [15]. If complexity can be afforded, these LMMSE or approximate LMMSE estimators can be implemented in MST approach. Let us consider a wide-sense stationary uncorrelated scattering (WSSUS) multipath channel with power delay profile given by $\sigma_{h_i}^2$ at delays of i OFDM sample intervals. Due to the uncorrelated multipaths, the correlation matrix of the channel impulse response becomes a diagonal matrix with the diagonal elements given by the power delay profile. Then the LMMSE's matrix multiplication (or scalar multiplication on corresponding channel taps) is performed as follows [19], [15]:

$$\hat{h}_{lmmse} = \text{diag} \left\{ \frac{\sigma_{h_0}^2}{\sigma_{h_0}^2 + \beta/SNR}, \frac{\sigma_{h_1}^2}{\sigma_{h_1}^2 + \beta/SNR}, \dots, \frac{\sigma_{h_{N-1}}^2}{\sigma_{h_{N-1}}^2 + \beta/SNR} \right\} \hat{h}_{LS} \quad (26)$$

where $\text{diag}\{a_0, a_1, \dots, a_{N-1}\}$ is a diagonal matrix with diagonal elements $[a_0, a_1, \dots, a_{N-1}]$. In DFT-based LMMSE approach, the first symbol is composed of all pilot tones of equal amplitude and the following data symbols are composed of M-ary QAM symbols on all subcarriers. Then for the initial channel estimation based on the training symbol, SNR/β in (26) is replaced by the pilot-to-noise power ratio. For the following channel estimation operated in the decision directed mode, β in (26) is given by $\beta = E[|X|^2] E[|X|^{-2}]$ where X is M-ary QAM symbol. For 64-QAM, $\beta = 2.6854$. For MST-LMMSE, SNR/β is replaced by its pilot-to-noise power

ratio (which is $1/K$ times that of DFT-based LMMSE due to its pilot ratio of $1/K$).

There is a delicate difference between DFT-based LMMSE and MST-LMMSE. In DFT-based LMMSE, all N samples of $\hat{h}_{LS}$ are corrupted by iid AWGN noise samples of zero mean and variance σ_t^2/A^2 . In MST-LMMSE, the first M samples of $\hat{h}_{LS}$ are corrupted by iid AWGN noise samples of zero mean and variance $K\sigma_t^2/A^2$ and the rest are zeros. Hence, if compared for the training symbol alone, DFT-based LMMSE will have a better channel estimation performance due to the larger pilot power used in channel estimation. But for the following data symbols, DFT-based LMMSE has to use decision directed approach which may degrade the channel estimation performance and BER performance due to the decision directed errors. For MST-LMMSE, the channel estimation and BER performances remain the same for all symbols.

In LMMSE implementation, since multipath channel correlation and SNR are usually unknown at the receiver, some fixed values have to be used for them. In [18], it is suggested to use high dummy SNR value and uniform multipath channel correlation which are robust to the channel correlation mismatch. If high dummy SNR value is used and uniform multipath channel correlation is assumed over guard interval, then LMMSE approach would be similar to [16]'s method. From (26) with high dummy SNR value, the LMMSE approach can be viewed as setting the channel taps with no energy to zeros and bypassing the other taps. Since most practical multipath channels have only a few significant paths (if compared to N), a suitable way of implementing an approximate LMMSE is choosing a predefined number of most significant taps and setting the others to zero which is the underlying idea of MST method. In complexity aspect, MST-LMMSE estimator still saves one N -point IFFT operation if compared to DFT-based LMMSE estimators.

VII. CHANNEL ESTIMATION PERFORMANCE EVALUATION BY SIMULATION

The proposed MST channel estimation is evaluated by computer simulation for two multipath fading channel models, namely Channel-A and Channel-B. The Channel-A is the ATTC (Advanced Television Technology Center) and the Grand Alliance DTV laboratory's ensemble E model whose channel impulse response for the static case is given by

$$h[n] = \delta[n] + 0.3162 \delta[n-2] + 0.1995 \delta[n-17] + 0.1296 \delta[n-36] + 0.1 \delta[n-75] + 0.1 \delta[n-137] \quad (27)$$

where unit delay is assumed to be the same as OFDM sample period. The Channel-B is a simplified version of DVB-T channel model P_1 [20] and its channel impulse response for the static case is given in Table I. In the simulation, the given channel tap gains of both channels represent the standard deviations of the gains of the complex Gaussian random variables.

The OFDM system parameters are as follows: the number of subcarriers $N = 8192$, pilot ratio $1/K = 1/8$, guard interval ratio = $1/8$ and carrier modulation 64-QAM. Perfect synchronization is assumed in order to observe the channel estimation performance alone. The considered methods are

TABLE I
CHANNEL IMPULSE RESPONSE FOR CHANNEL-B

Delay (OFDM samples)	Gain	Phase (radians)
0	0.2478	-2.5694
1	0.1287	-2.1208
3	0.3088	0.3548
4	0.4252	0.4187
5	0.4900	2.7201
7	0.0365	-1.4375
8	0.1197	1.1302
12	0.1948	-0.8092
17	0.4187	-0.1545
24	0.3170	-2.2159
29	0.2055	2.8372
49	0.1846	2.8641

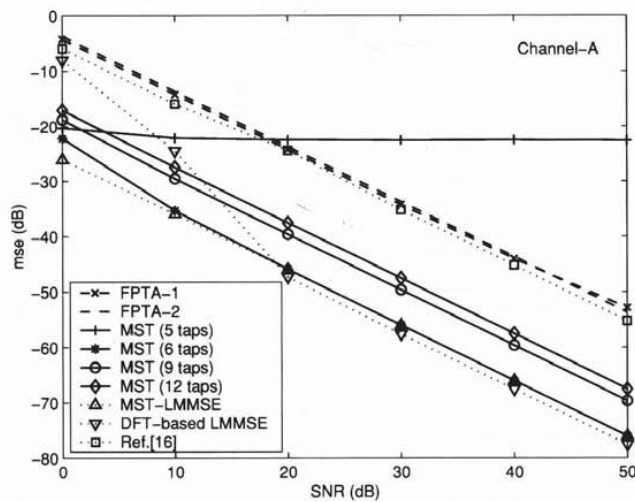


Fig. 3. Channel estimation mean square error (*mse*) in Channel-A.

MST, FPTA-1, FPTA-2, MST-LMMSE, DFT-based LMMSE and [16]. In the last two methods, one OFDM symbol is used for training symbol (pilot tones on all subcarriers) in every K OFDM symbols (i.e., the same pilot ratio $1/K$). In both LMMSE methods, ideal channel correlation and SNR values are used in order to evaluate the relative performance of MST method. In all methods, the same pilot tone symbol of $7 + j7$ signal point in 64-QAM constellation is used. For SNR of 40 and 50 dB, 10^5 independent simulation trials are used and for the other SNR values, 10^4 independent trials are used. In our simulation, floating point computation is used. However, it should be noted that due to the limited quantization levels in real hardware implementation, the computation error in real situation may degrade the performance obtained in our simulation.

Fig. 3 shows the *mse* performances of different channel estimation methods in Channel-A. The required interpolation in FPTA-1 is performed using Matlab's default *interp* function. Due to some noise reduction of this interpolation, the *mse* of FPTA-1 is slightly less than FPTA-2. The method of [16] has

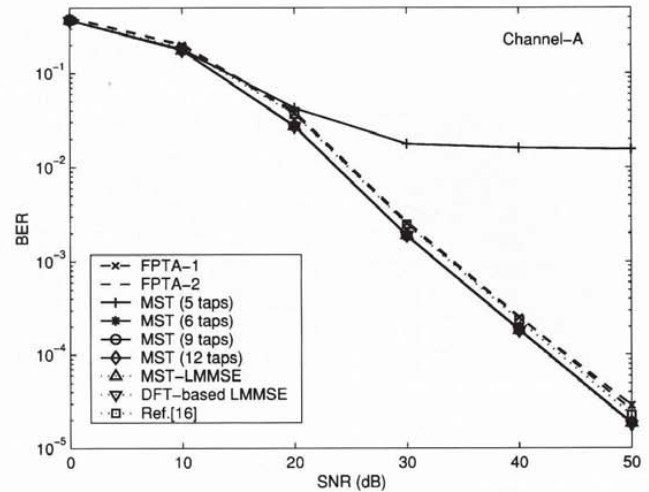


Fig. 4. BER performances of OFDM system with different channel estimation methods in Channel-A.

some improvement over FPTA approaches even though it is associated with decision directed errors. MST method with 6 taps, which is the same as the number of taps in Channel-A, has approximately 18 to 22 dB *mse* gain over FPTA approaches. MST with 5 taps shows an irreducible channel estimation error floor caused by missing some of the channel energy. It has a better performance than FPTA approaches for SNR less than 20 dB since for this SNR region the gain in noise suppression is greater than the loss of channel energy missing. However, for higher SNR region where noise has smaller impact than the channel energy missing, the channel estimation error floor results in a worse performance for MST with 5 taps. For the cases of MST with larger number of taps than the actual channel taps, MST achieves performance gain over FPTA and [16] approaches. The channel estimation performance gains are not as high as MST with 6 taps case due to the additional noise perturbation from the extra taps.

Also shown in Fig. 3 are the performances of MST-LMMSE and DFT-based LMMSE methods. LMMSE's approaches are of similar performance to MST with 6 taps case. For SNR less than 20 dB, DFT-based LMMSE has a worse performance than MST with 6 taps due to the decision directed errors. For higher SNR region, it has a slightly better performance due partly to the fact that its decision directed errors become insignificant and partly to the fact that MST suffers some interference from data part. MST-LMMSE has a slight performance gain over MST with 6 taps case for low SNR region and almost the same performance for high SNR region. This slight performance gain of MST-LMMSE in low SNR region is due to the utilization of (ideal) channel and noise statistics in MST-LMMSE approach by which the selection of channel taps are always correct, whereas in MST approach, the selection of channel taps may not be always correct due to the large noise perturbation.

The actual *mse* gain of MST method is slightly less than the potential gain of MST method described in previous section since data signal interference to training signal is neglected in the derivation of the potential gain. The actual *mse* gain is less in low SNR region than in high SNR region. The reason is

TABLE II
BER FOR DIFFERENT CHANNEL ESTIMATION METHODS IN CHANNEL-A

SNR (dB)	0	10	20	30	40	50
FPTA-1	0.3926	0.2015	3.7501×10^{-2}	2.5364×10^{-3}	2.4960×10^{-4}	2.8721×10^{-5}
FPTA-2	0.3957	0.2049	3.9122×10^{-2}	2.6389×10^{-3}	2.5513×10^{-4}	2.5377×10^{-5}
MST (5 taps)	0.3693	0.1814	4.262×10^{-2}	1.7651×10^{-2}	1.5964×10^{-2}	1.5441×10^{-2}
MST (6 taps)	0.3689	0.1770	2.7049×10^{-2}	1.8693×10^{-3}	1.8358×10^{-4}	1.8309×10^{-5}
MST (9 taps)	0.3695	0.1776	2.7291×10^{-2}	1.8841×10^{-3}	1.8505×10^{-4}	1.8502×10^{-5}
MST (12 taps)	0.3700	0.1781	2.7487×10^{-2}	1.8958×10^{-3}	1.8598×10^{-4}	1.8574×10^{-5}
MST-LMMSE	0.3689	0.1753	2.7023×10^{-2}	1.9081×10^{-3}	1.8309×10^{-4}	1.8299×10^{-5}
DFT-based LMMSE	0.3670	0.1762	2.7051×10^{-2}	1.8911×10^{-3}	1.8052×10^{-4}	1.7851×10^{-5}
Ref. [16]	0.3711	0.1887	3.6529×10^{-2}	2.4789×10^{-3}	2.3445×10^{-4}	2.2593×10^{-5}

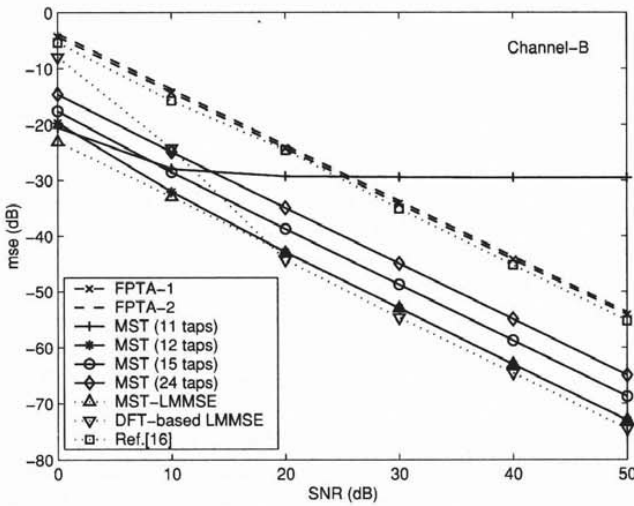


Fig. 5. Channel estimation mean square error (*mse*) in Channel-B.

that most significant channel taps selection would more likely choose some taps with no channel energy at low SNR region than at high SNR region, resulting in less *mse* gain at low SNR region.

In Fig. 4, the BER performances in Channel-A are presented for different channel estimation methods. Due to the channel estimation error floor and the BER sensitivity of 64-QAM to channel estimation error, MST with 5 taps case shows a BER floor while the others do not. The method of [16] is slightly better than FPTA approaches as is the case in *mse* performance. The other MST and LMMSE approaches have better BER performances than FPTA and [16] approaches, but among themselves, the BER performances are just slightly different and hence their BER curves are almost the same. For a better presentation of their slight differences, their BER values are tabulated in Table II. These slight differences can be ascribed to the corresponding different channel estimation performances. It is also noted that when the channel estimation *mse* is much smaller than $1/\text{SNR}$, then some deviation in channel estimation *mse* does not significantly affect the BER performance since noise is the dominant contributor to BER in this case.

The channel estimation performance and BER performance of the considered methods in Channel-B are plotted in Figs. 5 and 6 respectively and the BER values are given in Table III for the sake of clarity. The performances are of the same trend as in

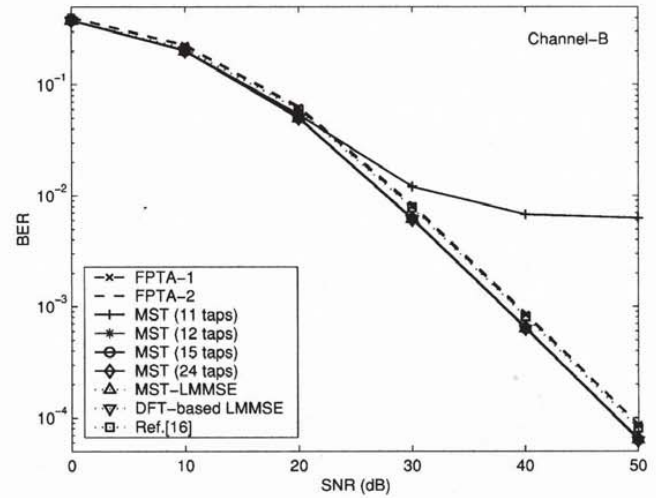


Fig. 6. BER performances of OFDM system with different channel estimation methods in Channel-B.

Channel-A. It is also noted that the channel estimation and BER performances in Channel-B is worse than Channel-A except for the case of missing some of the channel energy (i.e., MST with 5 taps in Channel-A and MST with 11 taps in Channel-B). This exception is due to the larger percentage of the channel energy missed in Channel-A than in Channel-B.

From the simulation results, it is clear that the number of MST taps J should not be smaller than the number of actual channel taps. On the other hand, using more taps may slightly degrade the BER performance due to the more noise perturbation. However, even up to double of the number of actual channel taps, MST has almost the same BER performance as using the number of actual channel taps. Hence, a suitable choice for J might be double of the (designed) number of channel taps for the considered channel environment.

Another way of selecting MST taps by threshold decision is also evaluated in Channel-B and the channel estimation *mse* and BER results are shown in Figs. 7 and 8 respectively. The performance of MST with 12 taps is also included for reference. It is observed that the suitable choice of threshold also depends on the operating SNR, as expected. The results also indicate that almost the same performance as MST with 12 taps case can be obtained if the threshold for an operating SNR is set within approximately 20 to 23 dB below the operating $1/\text{SNR}$. This low

TABLE III
BER FOR DIFFERENT CHANNEL ESTIMATION METHODS IN CHANNEL-B

SNR (dB)	0	10	20	30	40	50
FPTA-1	0.4014	0.2250	6.1986×10^{-2}	8.0725×10^{-3}	8.3867×10^{-4}	8.6296×10^{-5}
FPTA-2	0.4041	0.2283	6.3793×10^{-2}	8.3684×10^{-3}	8.6701×10^{-4}	8.6831×10^{-5}
MST (11 taps)	0.37828	0.2019	5.3833×10^{-2}	1.2092×10^{-2}	6.7836×10^{-3}	6.2976×10^{-3}
MST (12 taps)	0.3785	0.2011	4.9783×10^{-2}	6.1269×10^{-3}	6.2924×10^{-4}	6.3026×10^{-5}
MST (15 taps)	0.3790	0.2017	5.0082×10^{-2}	6.1707×10^{-3}	6.3410×10^{-4}	6.3421×10^{-5}
MST (24 taps)	0.3805	0.2032	5.0749×10^{-2}	6.2714×10^{-3}	6.4442×10^{-4}	6.4503×10^{-5}
MST-LMMSE	0.3781	0.2011	4.9847×10^{-2}	6.1509×10^{-3}	6.3000×10^{-4}	6.3806×10^{-5}
DFT-based LMMSE	0.3775	0.2018	4.969×10^{-2}	6.1261×10^{-3}	6.2774×10^{-4}	6.2568×10^{-5}
Ref. [16]	0.3809	0.2123	5.9644×10^{-2}	7.7969×10^{-3}	8.0554×10^{-4}	8.0195×10^{-5}

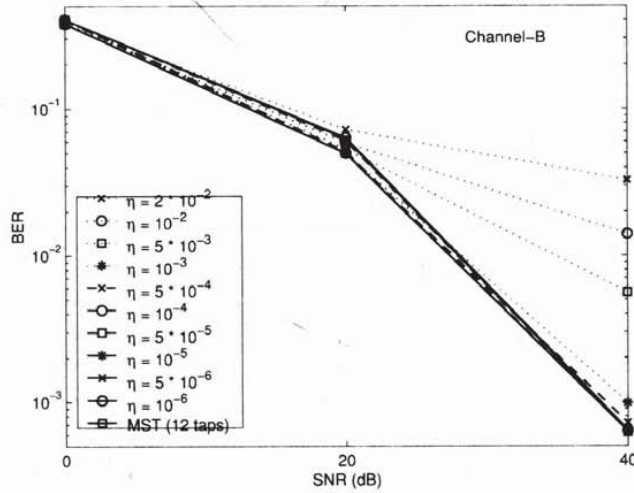


Fig. 7. Channel estimation mean square error (*mse*) for MST with threshold setting η .

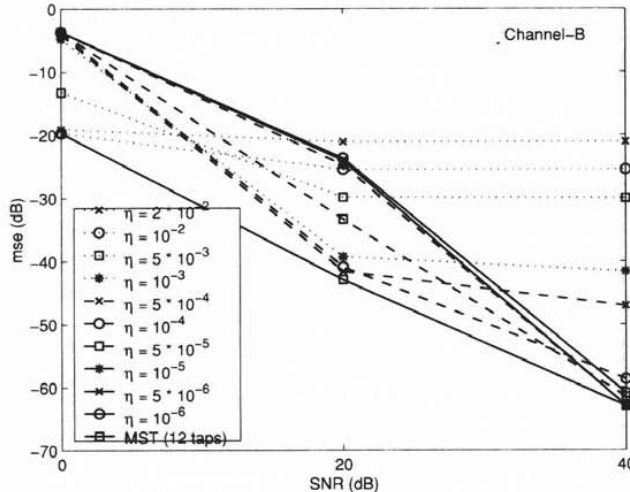


Fig. 8. BER performance of MST with threshold setting η .

value of the threshold in ensuring no channel energy miss is also due to the gain variations in the channel taps.

It is also remarked that for relatively slow fading environments, the decision feedback approaches such as [16] and

LMMSE can increase spectrum efficiency by using smaller pilot ratio. Similarly, MST approach can increase the spectrum efficiency by not inserting pilot tones in every OFDM symbol. For this case, the decision feedback approach as used in MST's dual form DFT-based approach can easily be applied for pure data symbols.

VIII. CONCLUSION

In this paper, intra-symbol time-domain averaging approaches for time-domain channel estimation of OFDM system are investigated. As an approximation to LMMSE type estimator, a time-domain approach, called MST, is proposed where intra-symbol time-domain averaging and most significant taps selection are applied. The similarities and differences of the proposed MST method versus DFT-based LMMSE methods are discussed. Two approaches for MST taps selection are described where the first uses a fixed number of taps while the second chooses all taps above a threshold. The simulation results suggest that a suitable choice for the fixed number of MST taps is using double of the (designed) number of channel taps for the considered channel environment. And a suitable choice of the threshold is within the range of approximately 20 to 23 dB below the operating $1/SNR$. In terms of BER performance in multipath fading channels, MST without channel energy missing and LMMSE approaches have almost the same performance which is better than the other considered approaches. In terms of complexity, MST approach keeps minimum complexity among the considered methods.

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H. Minn received his B.E.(Electronics) degree with honors from Yangon Institute of Technology, Myanmar, in 1995, and M.Eng. (Telecommunications) degree from Asian Institute of Technology (AIT), Thailand, in December 1997. He worked as a laboratory supervisor at Telecommunications Program, AIT, for one year. Since 1999, he has been a research assistant in the Department of Electrical and Computer Engineering at the University of Victoria, Canada. Currently, he is a Ph.D. student at University of Victoria. His research interests include wireless communications, OFDM, synchronization, channel estimation, equalization, modulation and error control coding.

Vijay K. Bhargava received the B.Sc., M.Sc., and Ph.D. degrees from Queen's University, Kingston, Canada in 1970, 1972 and 1974 respectively. Currently, he is a Professor of Electrical and Computer Engineering at the University of Victoria. He is a co-author of the book *Digital Communications by Satellite* (New York: Wiley, 1981) and co-editor of the IEEE Press Book *Reed-Solomon Codes and Their Applications*. He is an Editor-in-Chief of *WIRELESS PERSONAL COMMUNICATION*, a Kluwer Periodical. His research interests are in multi-media wireless communications.

Dr. Bhargava is very active in the IEEE and is currently the President of the IEEE Information Theory Society. He was Co-chair for IEEE ISIT'95, Technical Program Chair for IEEE ICC'99 and is the Chair of IEEE VTC'2002 Fall.

Dr. Bhargava is a Fellow of the B.C. Advanced Systems Institute, Engineering Institute of Canada (EIC), IEEE and the Royal Society of Canada. He is a recipient of the IEEE Centennial Medal (1984), IEEE Canada's McNaughton Gold Medal (1995), the IEEE Haraden Pratt Award (1999) and the IEEE Third Millennium Medal (2000).

Hsieh-Yee Declaration

Appendix 0002D

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Appendix 0003A

Channel Estimation Using Pilot Tones in OFDM Systems

Che-Shen Yeh and Yinyi Lin
Department of Electrical Engineering
National Central University, Chung Li
Taiwan 32054, R. O. C.

Abstract –

Channel estimation techniques using pilot tones in OFDM systems are investigated. Two time-domain channel estimation methods are proposed. Based on theoretical analysis and computer simulation, for a channel with multipath distortion and AWGN, the proposed time-domain channel estimation schemes perform better than the conventional frequency domain channel estimation techniques.

I. Introduction

Since terrestrial broadcast channels are often affected by multipath distortion, in a single carrier modulation system, an adaptive equalizer is usually employed to combat channel variations. However, in the Orthogonal Frequency Division Multiplexing (OFDM) modulation system, the channel estimation is accomplished using pilot tones inserted in OFDM symbols. In an OFDM system, the incoming signal is transmitted using N equally spaced carriers with a carrier spacing of $1/(NT)$ Hz, where T is the input data bit period. Multipath causes the channel transfer function to have variation in both spectral magnitude and spectral phase linearity. The number of carriers is chosen such that a typical channel transfer function is essentially constant over the bandwidth of each individual carrier. The effect of multipath distortion is then reduced to amplitude attenuation and phase rotation of data in each carrier, and the channel equalization is reduced to a complex multiplication bank in front of the decision circuit at the receiver [1-2].

An OFDM system can eliminate Intersymbol Interference (ISI) between OFDM symbols by inserting a cyclically extended guard interval, where each OFDM symbol is preceded by a periodic extension of data signal itself [3, 10]. The total symbol duration becomes $T_{total} = T_g + T_s$, where T_g is the guard interval and T_s is the useful symbol duration. When the guard interval T_g is longer than the channel impulse response $h[n]$, or the multipath delay spread, the ISI between OFDM symbols can be entirely eliminated. The ratio of the guard interval to useful symbol duration is dependent on application. A longer guard interval will reduce the system data throughput, and a commonly adopted compromise length is

$T_g \leq \frac{1}{4}T_s$. A cyclic prefix in the guard interval can be used to maintain the receiver carrier synchronization during a long silence. In addition, cyclic convolution can be used to model the transmission system.

However, the in-band fading will still exist. Therefore, let $H_k = \lambda_k e^{j\theta_k}$ be the response within the bandwidth of the k^{th} carrier. Equalization of the channel in the receiver requires that the FFT output of the k^{th} carrier be multiplied by a complex coefficient $c_k = \frac{1}{H_k}$. This operation will scale both

signal and noise/interference. In the OFDM system, the estimation of the channel response uses in-band pilot tones inserted into each symbol with a pilot ratio $PR = 1/R$. In this scheme, the channel response can be estimated rapidly and the tuner phase noise can be significantly reduced [10-11]. The optimum choice for the numbers of pilots ($=N/R$) for the AWGN channel is described in [4]. This scheme is referred to as scheme A, as depicted in Fig. 1a. In the AWGN channel every optimally spaced set of pilots can provide good channel estimation, so a fixed set of optimally spaced pilot tones is used in scheme A.

An alternative is to insert pilot tones into all of the carriers in an OFDM symbol with a period of R , as depicted in Fig. 1b. The channel response is then estimated using the pilot symbol, and the estimated response is used for subsequent OFDM symbols until the next pilot symbol is received. The channel response can also be estimated by averaging some consecutive pilot symbols. This scheme is referred to as scheme B. As will be shown, it is not suitable for fading channels.

II. Proposed Channel Estimation Techniques

A typical OFDM system is depicted in Fig. 2. On the transmitter site, a multiplexer (MUX) combines the data symbols and pilot signals as a data stream in the frequency domain, and the data stream is fed into an Inverse Fast Fourier Transform (IFFT) block. The IFFT block converts the data stream from a serial format into a parallel format, and transforms the signal from frequency domain to time domain. The guard interval is then inserted after each OFDM symbol to eliminate ISI between OFDM symbols.

Pilot tones are modulated with a PN sequence in the frequency domain and are inserted into each OFDM symbol with a ratio $PR = 1/R$ using scheme A. The channel response is estimated in the frequency domain using FFT processing and comparing received pilots with the locally stored reference pilots. This is referred to as the Frequency Pilot Frequency Interpolation (FPFI) technique since the channel response is obtained via frequency interpolation. The pilot tones can also be inserted into the OFDM symbols using scheme B. In this scheme, all pilot tones are inserted in the first OFDM symbol (pilot symbol) of a data frame. The channel response is obtained by averaging several consecutive pilot symbols and comparing the average with a locally stored reference symbol [5]. This technique is referred to as the Frequency Pilot Frequency Average (FPFA) technique.

In this paper, channel estimation is conducted in the time domain is proposed. In one of the two proposed techniques, instead of the PN sequence the positive and negative polarization pilot signals are alternatively inserted into OFDM symbols in the frequency domain using FPFI. An OFDM symbol can be regarded as several time slots, with each time slot having a pilot. Assuming that the impulse response of each time slot is identical, one can obtain the channel response in the time domain by averaging these time slots. This technique is referred to as the Frequency Pilot Time Average (FPTA) technique, as depicted in Fig. 3a.

In the other technique a PN sequence is periodically inserted into the OFDM symbol in the time domain, and the channel response can be obtained via correlation in the time domain. This technique is referred to as the Time Pilot Time Correlation (TPTC) technique, as depicted in Fig. 3b. In this technique the channel response is estimated using the correlation of received pilot signal and the locally stored PN sequence [8]. If the transmitted pilot signal $p[n]$ is a PN sequence, the received pilot signal $\tilde{p}[n]$ can be expressed as $\tilde{p}[n] = p[n] \otimes h[n]$ in the time domain (where $\otimes$ represent the convolution operator) or $\tilde{P}[k] = P[k] \cdot H[k]$ in the frequency domain. Thus the channel response is given by $H[k] = \tilde{P}[k] / P[k] = \tilde{P}[k] \cdot P[k]^*$. As the spectrum of a PN sequence is nearly constant at the frequency occupied by the received pilot signal, $|P[k]|^2$ can be approximated as a constant, K , which is equal to the length of the PN sequence. Thus the channel response can be estimated using

$$H[k] = \frac{\tilde{P}[k] \cdot P[k]^*}{K} \text{ in the frequency domain}$$

or

$$h[n] = \frac{1}{K} \sum_i \tilde{p}[n] \cdot p[n+i]^* \text{ in the time domain.}$$

III. Pilot to Noise (P/N) Ratio

This section is an analysis of the pilot to noise ratio for both time domain channel estimation and frequency domain channel estimation. Referencing to Fig 2 and Fig 3a, the pilot signal is $P[k]$. Suppose pilot ratio $PR = 1/R$, $P[k]$ can be shown as:

$$P[k] = \begin{cases} A, & k = Rm \\ 0, & k = Rm + \ell \end{cases} \quad (1)$$

Where R , m and ℓ are integers, $1 \leq \ell \leq R-1$, $0 \leq m \leq (N/R)-1$, A is the pilot amplitude and N/R is an integer.

$$p[n] = \text{IFFT}\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] e^{j \frac{2\pi nk}{N}} \quad (2)$$

Where N is the IFFT transformation length

The multipath channel impulse response $h[n]$ is:

$$h[n] = \delta[n] + \sum_{i=1}^M \alpha_i \cdot \delta[n - \tau_i] = \sum_{i=0}^M \alpha_i \cdot \delta[n - \tau_i] \quad (3)$$

Where M is the total number of echoes

α_i represents the amplitude of i^{th} path echo

τ_i represents the delay of i^{th} path echo

$\alpha_0 = 1$, $\tau_0 = 0$, (representing the principle pilot signal)

$p[n]$ after multipath channel $h(n)$ are:

$$\tilde{p}[n] = p[n] * h[n] = p[n] + \sum_{i=1}^M \alpha_i p[n - \tau_i] \quad (4)$$

On the Rx side, the received frequency domain pilot signal $\tilde{P}[k]$ and multipath channel response $H[k]$ will be :

$$\tilde{P}[k] = P[k] \cdot H[k] = \begin{cases} \sum_{i=0}^M \alpha_i P[k] e^{j \frac{2\pi \tau_i k}{N}}, & k = mR \\ 0, & k \neq mR \end{cases} \quad (5)$$

$$H[k] = 1 + \sum_{i=1}^M \alpha_i e^{-j \frac{2\pi \tau_i k}{N}} \quad (6)$$

Assuming the time domain channel noise $n[n]$ is wide sense stationary AWGN with zero mean and variance:

$$\sigma_{n[n]}^2 = \text{Var}\{n[n]\} = E\{n[n] - E\{n[n]\}\} \quad (7)$$

Where $n[n]$ represents the time domain white noise at the n^{th} sample, $E\{n[n]\}$ represents the expected value of $n[n]$.

In radio channel, suppose the noise is zero mean AWGN,

$E\{n[n]\} = 0$. On the Rx side, after the FFT, the $n[n]$ signal will become the frequency domain noise

$$N[k] = \sum_{n=0}^{N-1} n[n] W_N^{kn}, W_N = e^{-j\frac{2\pi}{N}} \quad (8)$$

Assuming the variance of frequency domain noise $N[k]$ is σ_{FN}^2 , according to variance definition:

$$E\{N[k]\} = \sum_{n=0}^{N-1} E\{n[n]\} W_N^{kn} = 0$$

$$\begin{aligned} \sigma_{FN}^2 &= \text{Var}\{N[k]\} \\ &= E\{(N[k] - E\{N[k]\}) \cdot (N[k] - E\{N[k]\})^*\} \\ &= E\{N[k] \cdot N^*[k]\} \\ &= E\left\{\left(\sum_{n=0}^{N-1} n[n] \cdot W_N^{kn}\right) \cdot \left(\sum_{m=0}^{N-1} n[m] \cdot W_N^{km}\right)^*\right\} \\ &= \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} E\{n[n] \cdot n^*[m]\} W_N^{k(n-m)} \end{aligned} \quad (9)$$

Because $n[n]$ and $n[m]$ are independent and zero mean, Eq.(9) can be re-written as:

$$\begin{aligned} \sigma_{FN}^2 &= \sum_{n=0}^{N-1} \left\{ E\{n[n] \cdot n[n]\} W_N^0 + \sum_{\substack{m=0 \\ m \neq n}}^{N-1} E\{n[n]\} \cdot E\{n^*[m]\} W_N^{k(n-m)} \right\} \\ &= \sum_{n=0}^{N-1} (\sigma_{in}^2 + 0) = N \cdot \sigma_{in}^2 \end{aligned} \quad (10)$$

From the previous mathematical deductions, in the radio channel of frequency domain, the noise variance is being enhanced N times while the noise amplitude is amplified $\sqrt{N}$ times.

In the FPTA, pilot ratio $PR = 1/R$ and the pilot signal $P[k]$ is shown as:

$$\begin{aligned} P[k] &= \begin{cases} (-1)^m \cdot A & , \quad k = Rm \\ 0 & , \quad k = Rm + \ell \end{cases} \\ &= \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \delta[k - Rm] \end{aligned} \quad (11)$$

Where $m = 0, 1, 2, \dots, (N/R)-1$, ℓ is an integer and $1 \leq \ell \leq R-1$. The time domain TX pilot signal $p[n]$ is shown as:

$$\begin{aligned} p[n] &= \text{IFFT}\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] \cdot e^{+j\frac{2\pi kn}{N}} \\ &= \frac{1}{N} \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot e^{+j\frac{2\pi Rmn}{N}} \\ &= \frac{A}{R} \cdot \frac{1}{N/R} \sum_{m=0}^{(N/R)-1} e^{+j2\pi m \left(\frac{n - (N/2R)}{N/R}\right)} \\ &= \frac{A}{R} \sum_{r=0}^{R-1} \delta\left[n - \frac{Nr}{R} - \frac{N}{2R}\right] \\ &= \begin{cases} A/R & , \quad n = Nr/R + N/2R \\ 0 & , \quad \text{otherwise} \end{cases} \end{aligned} \quad (12)$$

From the result of Eq.(12), when n is equal to $N/2R$ plus the integer time of N/R , the time domain pilot will be added in-phase. The Rx pilot signal $\tilde{p}[n]$ in the time domain and $\tilde{P}[k]$ in the frequency domain will be:

$$\begin{aligned} \tilde{p}[n] &= p[n] \otimes h[n] \\ &= \frac{A}{R} \sum_{u=0}^{N-1} \left\{ \sum_{r=0}^{R-1} \delta\left[u - \frac{Nr}{R} - \frac{N}{2R}\right] \cdot \left(\sum_{i=0}^M \alpha_i \delta[n - u - \tau_i] \right) \right\} \\ &= \frac{A}{R} \sum_{r=0}^{R-1} \sum_{i=0}^M \alpha_i \cdot \delta\left[n - \left(\tau_i + \frac{Nr}{R} + \frac{N}{2R}\right)\right] \end{aligned} \quad (13)$$

$$\begin{aligned} \tilde{P}[k] &= P[k] \cdot H[k] \\ &= \begin{cases} \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \delta[k - Rm] \cdot \left\{ 1 + \sum_{i=1}^M \alpha_i \cdot e^{-j\frac{2\pi \tau_i k}{N}} \right\} \\ \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \left[\sum_{i=0}^M \alpha_i \cdot e^{-j\frac{2\pi \tau_i m}{N/R}} \right] & , \quad k = Rm \\ 0 & , \quad k \neq Rm \end{cases} \end{aligned} \quad (14)$$

where $\alpha_0=1$, $\tau_0=0$, (it is the subscript 0 that represents the principle pilot component).

In the FPFI method, $P[k] = PN$ -Sequence, $p[n]$ and $P[k]$ are both noise-like and randomly distributed in the time domain OFDM symbol. Hence, after FFT on the Rx end, the result is similar to AWGN case described above, and $\tilde{P}[k]$ and $\tilde{p}[n]$ both have $\sqrt{N}$ gain. Therefore, in the second (FPFI) method the P/N in the time domain and in the frequency domain are about the same, so using pilot signal for channel model estimation in the time domain is as good as in the frequency domain.

If the FPFI method is used for channel estimation with the pilot signal in the frequency domain, the pilot to noise ratio is $\tilde{P}[k]/\sigma_{FN}$. If the FPTA method is used for channel estimation with the pilot signal in the time domain, the pilot to noise ratio

is $\tilde{p}[n]/\sigma_{tn}$, which implies that $\tilde{P}[k]$ and $\tilde{p}[n]$ are affected by the echo. The amplitude of $\tilde{p}[n]$ has different variation, which depends upon both the amplitude of echo and delay spread. Different multipath models result in different variation of $\tilde{P}[k]$. Hence, it is difficult to characterize generally the variation. This variation may differ with different channel models, hence, it is difficult to describe or derive with such unpredictable value compared with the principle pilot signal to noise variance ratio. However, the third method repeats an average of R times, after which it reduces σ_{tn} to $\sigma'_{tn} = \sigma_{tn}/\sqrt{R}$. Assuming that the pilot interference of data to pilot tone can be ignored, we may compare the principle pilot to noise of these two methods:

$$\frac{p[n]/\sigma'_{tn}}{P[k]/\delta_{FN}} = \frac{A/R \cdot (\sigma_{tn}/\sqrt{R})}{A \cdot (\sqrt{N} \cdot \sigma_{tn})} = \sqrt{\frac{N}{R}} \quad (15)$$

The FPTA estimation performance is better than the FPFA and FPFI. This improvement will be verified in the next section. In the derivation of Eq. (15), the FPTA does not take the "data ghosts" into account, thus, the gain of the time domain $\tilde{p}[n]/\delta_{tn}$ over the frequency domain $\tilde{P}[k]/\delta_{FN}$ is actually lower than the value of $\sqrt{N/R}$ indicated in Eq.(15).

The interference of training sequence coming from data is referenced to as "data ghosts". A specific comparison of channel estimation will be given in the next section.

IV. Performance Comparisons and Discussions

The performance comparison of previously mentioned estimation techniques is made on several channel models recommended by the Advanced Television Technology Center (ATTC) in the U. S. A. and by the Digital Audio Broadcasting (DAB) project in Europe [6-7]. During simulations, assume that both transmitter and receiver are perfectly synchronized and the guard interval is long enough so that ISI between OFDM symbols can be ignored. The channel is assumed to be the additive white Gaussian noise (AWGN) channel.

Static Channel

Two channel models: the multipath Ensembles A and E of the ATTC, as recommended by the US ATTC and the Grand Alliance DTV laboratory [6][7], are used in simulations: Ensemble A as channel model 1:

$$h_1[n] = 0.1\delta[n+16] + \delta[n] + 0.1\delta[n-2] + 0.3162\delta[n-16] + 0.1995\delta[n-52] + 0.1295\delta[n-168]$$

Ensemble E as channel model 2:

$$h_2[n] = \delta[n] + 0.3162\delta[n-2] + 0.1995\delta[n-17] + 0.1296\delta[n-36] + 0.1\delta[n-75] + 0.1\delta[n-137]$$

In the above two formulas, the unit delay length $T = 1.1 \times 10^{-7}$ seconds. In addition, the simplified non-hilly urban channel model recommended by DAB test (pr EN50248) is also used in simulations as channel model 3:

$$h_3[n] = 0.707\delta[n+2] + \delta[n] + 0.707\delta[n-3] + 0.5\delta[n-13] + 0.4\delta[n-19] + 0.31625\delta[n-44]$$

Parameters	Specifications
Number of carriers (N)	8192
Pilot ratio (PR = 1/R)	1/8, 1/32
Guard interval	1/8, 1/32
Signal constellation	64QAM
Channel model	h1(n), h2(n), h3(n)
S/N	0 ~ 30 dB

Table 1. OFDM system performance simulation parameters

The OFDM system parameters used in simulations are summarized in Table.1. All data carriers in the OFDM are 64-QAM modulated signals (signal constellation (x,y) with $x, y \in \{\pm 1, \pm 3, \pm 5, \pm 7\}$, and the average power of a data symbol is $P_s = 42$ [9]). The pilot signals with amplitude ± 7 are used.

The error performance is displayed in Figs 4 and 5 for various estimation techniques on the channel model 1 with pilot ratio 1/8 and 1/32 respectively. A similar result was found on the other two channel models and is not shown here. The results demonstrate that the proposed estimation techniques (i.e., FPTA and TPTC) outperform the conventional estimation techniques (i.e., FPFA and FPFI). A SNR improvement gain of 2-3 dB can be achieved using the time-domain estimation technique. This is because a higher P/N ratio can be obtained in the time domain. For pilot ratio $PR = 1/32$ on the television broadcasting channels (channel models 1 and 2), the error floor occurs using FPFI technique, as shown in Fig. 5. This is due to the aliasing effect introduced by the frequency interpolation. However, no error floor was found on the audio broadcasting channel (channel model 3). Since this channel has a shorter delay spread and the aliasing effect is not severe.

Extensive simulations using FPFI technique were conducted using different pilot ratios (i.e., 1/4, 1/8, 1/16, and 1/32, respectively). It was found that with pilot ratios $PR = 1/4$ and 1/8, no error floor occurs for all the three channel models. However with pilot ratio $PR = 1/16$, the error floor occurs at

8.2×10^{-4} and 1.35×10^{-4} for channels 1 and 2 respectively, although no error floor was found for channel 3. As the pilot ratio PR was reduced to 1/32, a more severe error floor occurs (0.335 and 0.0665 for the two DTV channel models), although there was still no error floor found on the DAB channel.

Channel Estimation Accuracy

In addition to symbol error rate performance, the mean square error between estimated response and real channel response was also compared to demonstrate the superiority of the time-domain estimation techniques. The system parameters used for comparison are displayed in Table 2 and the results are illustrated in Fig. 6. As shown, a smaller mean square error was found in the time-domain channel estimation techniques. The FPFA technique demonstrates a larger estimation error than the FPFI technique since in the FPFA the estimation error extends over several OFDM symbols, while in the FPFI the estimation error impacts only adjacent carriers within one OFDM symbol. The estimation error in the FPTA is larger than that in the TPTC because the FPTA is affected by the components of "data ghosts" which are non-zero, even after averaging several OFDM symbols. According to equation (15), the FPTA method outperforms FPFI method with a P/N gain about 30 dB. Figure (6) shows a P/N gain about 25.2 dB. The difference between 4.8 dB is due to "data ghosts".

Parameters	Specifications
Number of carriers (N)	8192
Pilot ratio (PR = 1/R)	1/8
Guard interval	1/8
Signal constellation	64QAM
Channel model	$h_1[n], h_2[n], h_3[n]$
S/N	10 dB, 20dB
Mean Average Number	16

Table 2 System parameters

Slow Fading Channel

The error performance of the estimation techniques is also compared on a mobile time-varying channel using the autoregressive (AR) model given as

$$h[k+1] = ah[k] + w[k]$$

Where a and $w[k]$ represent slow time variations and the noise vector sequence of the AWGN, respectively. It is assumed that the channel fading is slow enough so that the channel impulse response doesn't change within the duration of one OFDM symbol [4].

During simulations, parameters of the AR model are

assumed as $a = 0.99\Pi$ and $w[k] = 0.0001\Pi$, where Π is the unit matrix. The OFDM parameters displayed in Table 1 and the three channel models previously mentioned are used. The error rate is shown in Fig. 7 with pilot ratio PR = 1/8. The results also reveal that both FPTA and TPTC provide better performance than those estimated in the frequency domain (FPFI and FPFA). Although both FPFA and FPFI techniques achieve almost the same error performance in static channels, the performance of the FPFA is slightly worse than that of the FPFI. This is because in the FPFA the channel response is estimated by averaging R pilot symbols (in this case, 8 pilot symbols were averaged), which cannot accurately estimate the response of each OFDM symbol. The effect of the pilot ratio on the error performance is also simulated, and the result with pilot ratio PR = 1/32 is demonstrated in Fig. 8 for channel 2. Again the error floor was found for the FPFI due to the aliasing effect. The error floor was also found in the FPFA, in which the channel was estimated by averaging 32 OFDM symbols. The error floor occurs when the accumulated estimation error due to channel variations becomes not negligible as compared to the estimation error caused by channel noise.

V. Summary

In this paper the channel estimation techniques (FPTA and TPTC) in the time domain using pilot tones for the OFDM system are presented. It is shown that a higher P/N ratio can be obtained in the time domain and that the proposed estimation techniques outperform the conventional estimation techniques (FPFI and FPFA) in which the channel response is estimated in the frequency domain. In addition, both FPFI and FPFA suffer from the aliasing effect and channel mis-estimation resulting in error floors, especially for low pilot ratios.

The proposed FPTA and TPTC techniques have advantages of low computational complexity and high P/N ratio. FPTA is suitable for fading channels with shorter transmission range, i.e., Local Multipoint Distribution Systems (LMDS), which has shorter channel response. FPFI and FPFA can be used to reduce overhead for systems with lower pilot ratios (i.e., 1/32, 1/64). However, in the FPTA the pilot signal consumes more RF power and interferes with data signal when the pilot ratio is high. The TPTC does not have this disadvantage.

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Che-Shen Yeh received B.S. and M.S.E.E. in Electrical Engineering from Chung-Yuan Christian University, Jung-Li, Taiwan in 1980 and 1986 respectively, and is currently a Ph.D. candidate in Electrical Engineering at the National Central University, Jung-Li, Taiwan. His major research interests are digital transmission for television and audio. He leads the DTV/DAB transmission project at Telecommunications Labs of Chunghwa Telecom Co., Ltd. Taiwan, and is also responsible for DTV/DAB field trials and the development of DTV/DAB transmission system standards and regulations. He also is a member of the Task Force for SDTV/HDTV transmission standards and regulation at the Ministry of Transportation and Communications, Taiwan.

Yinyi Lin was born in Chuang Hua, Taiwan, R.O.C., in 1959. He received the B.S. degree in communication engineering from National Chiao Tung University, Hsin Chu, Taiwan in 1981, and M.S. and Ph.D. degrees in electrical engineering from the University of California, San Diego in 1986 and 1989, respectively.

From 1983 to 1985, he has employed by the Electronics Research and Service Organizations (ERSO), I.T.R.I., Taiwan. During 1988, he spent six months at the IBM Almaden Research Center, San Jose, CA working in the area of signal processing for digital magnetic recording. From 1990 to 1991, he was with Archive Corporation, Costa Mesa, CA, where he engaged in the research of quarter inch cartridge (QIC) tape system. Since 1991, he has been with the Electrical Engineering Department of National Central University, Jung-Li, Taiwan, where he is now a Professor. His current research interests include coding and signal processing for digital magnetic recording systems, digital compression, and error correction for documents.

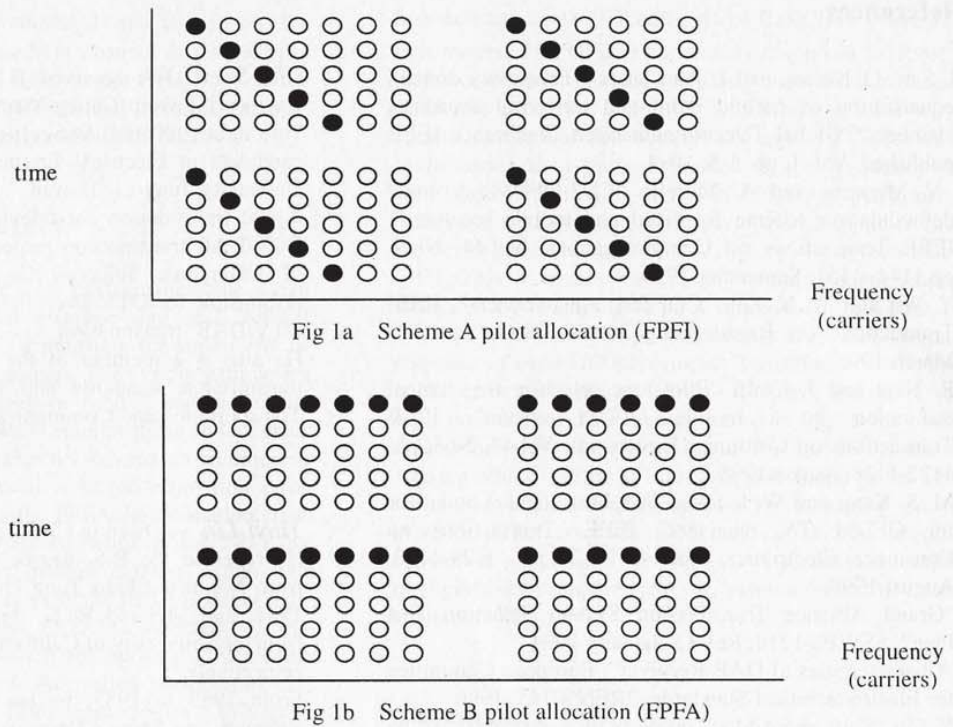


Figure 1 OFDM system frequency domain pilot signal allocation

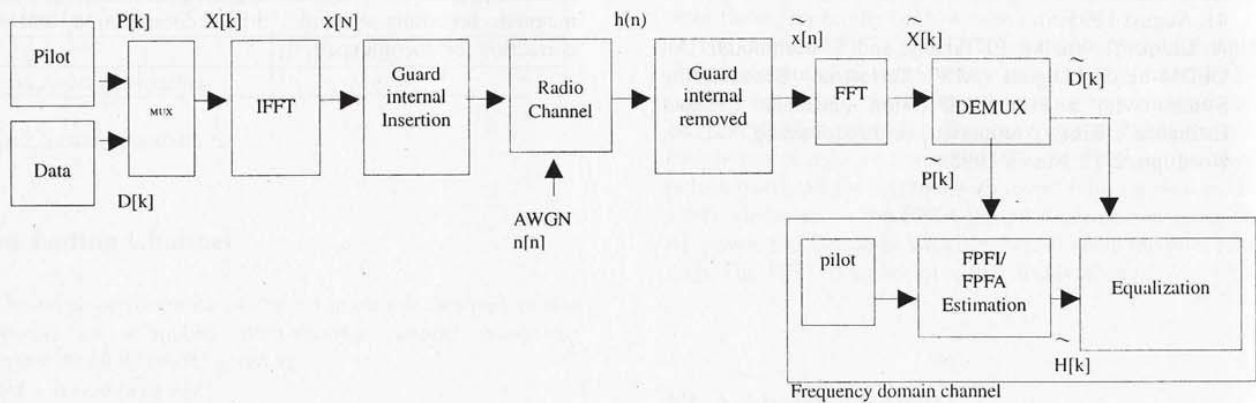


Figure 2 Channel model estimation of frequency pilot in the frequency domain

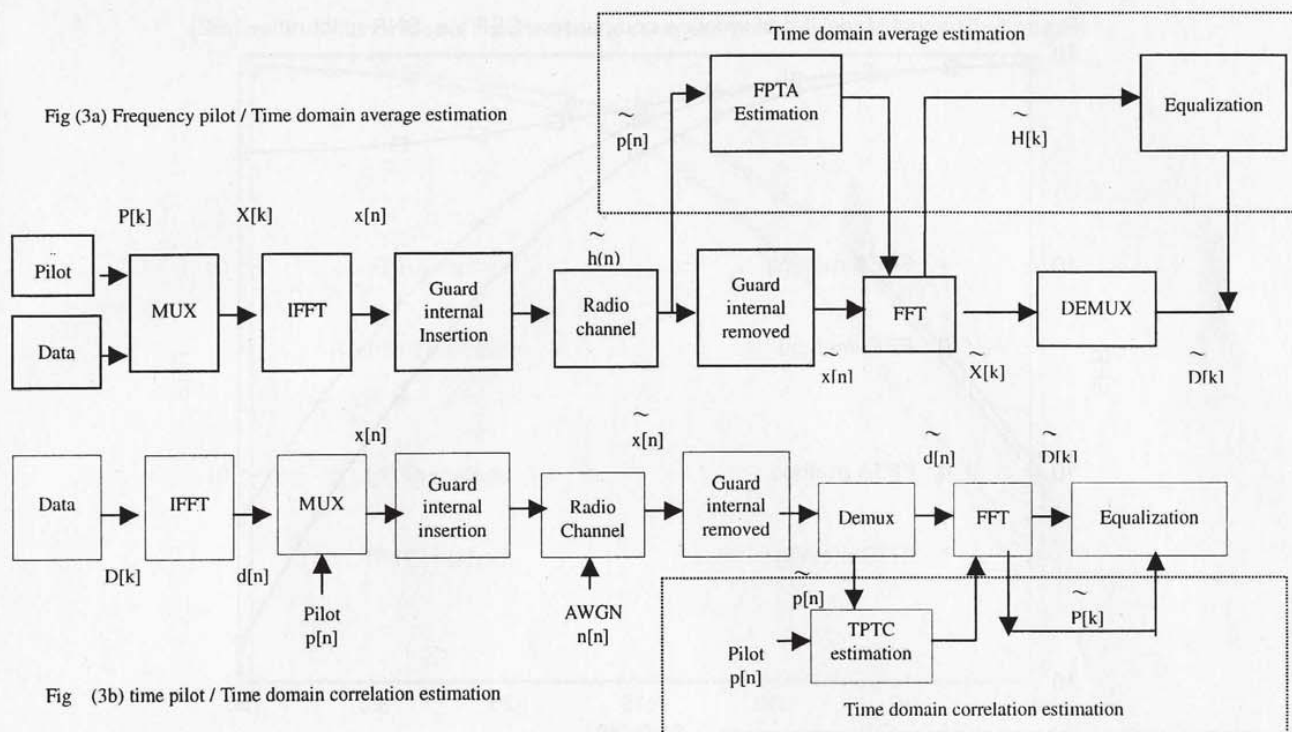


Figure. 3 Proposed Estimation Techniques (a) FPTA (b) TPTC

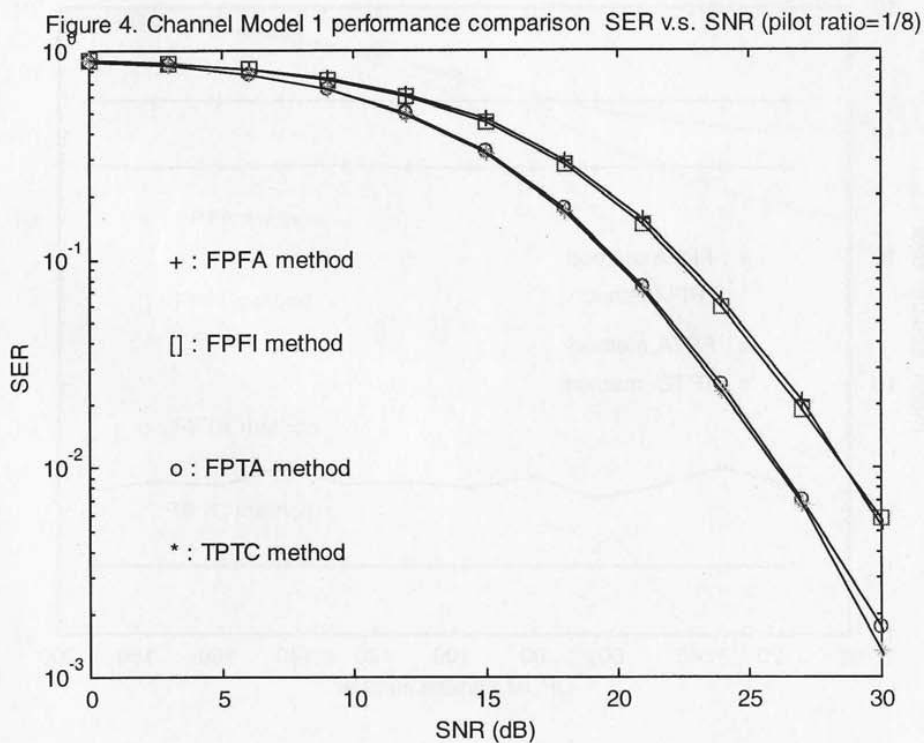


Figure 5. Channel Model 1 performance comparison SER v.s. SNR (pilot ratio=1/32)

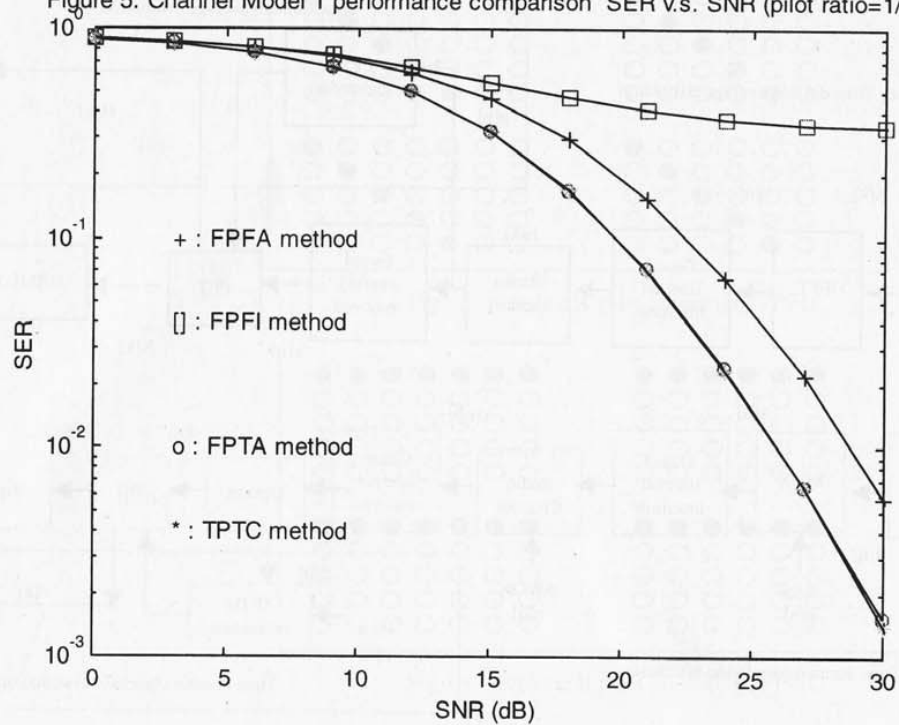


Figure 6. Channel Model 5 estimation MSE (CNR=10dB)

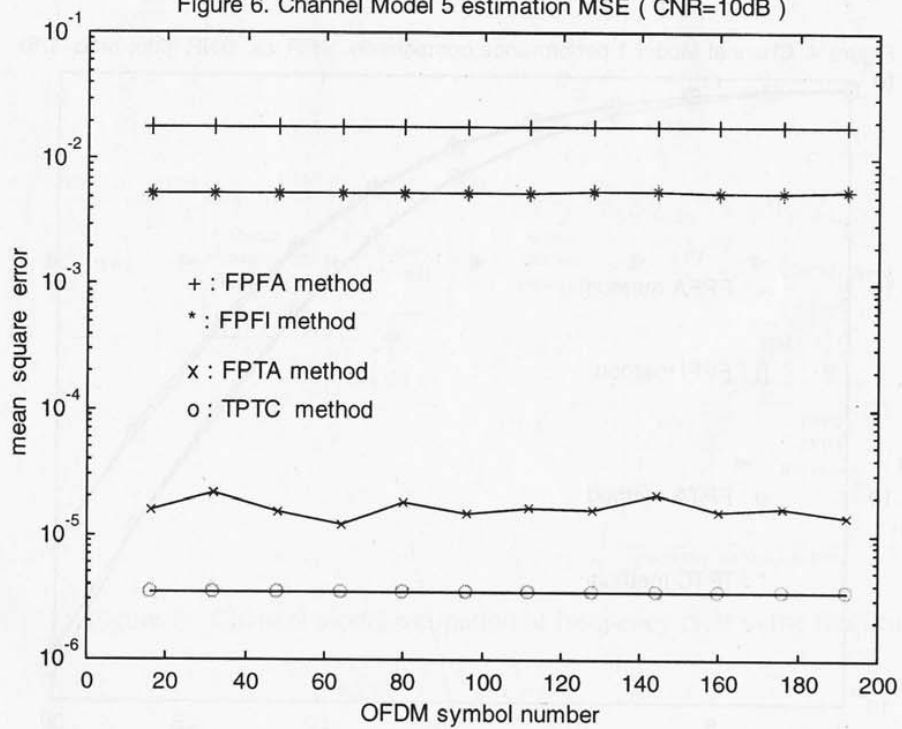


Figure 7. Channel Model 1 performance comparison SER v.s. SNR (pilot ratio=1/8)

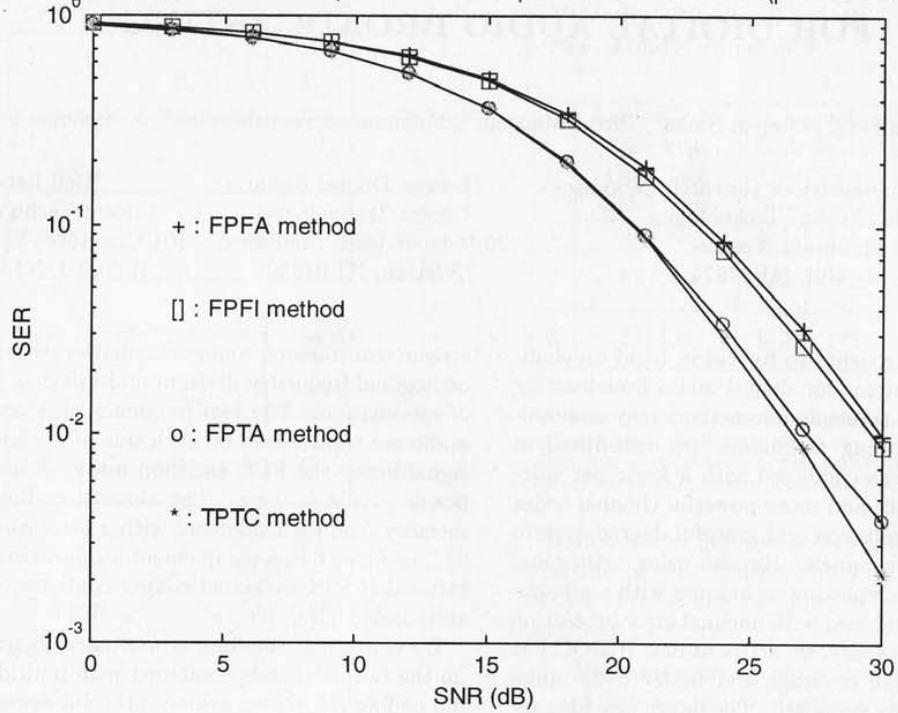
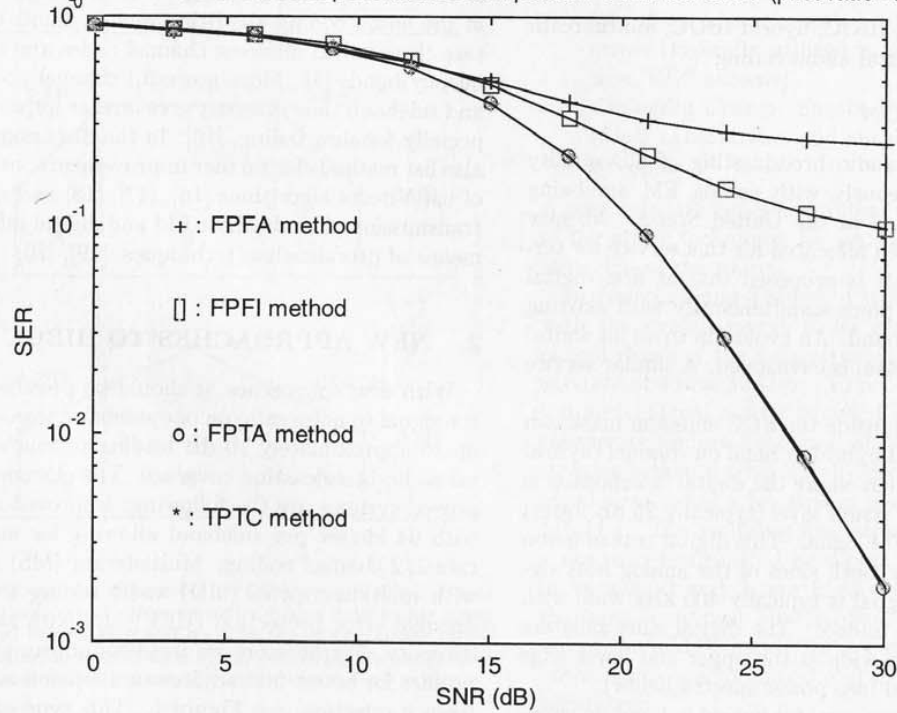


Figure 8. Channel Model 2 performance comparison SER v.s. SNR (pilot ratio=1/32)





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ASSOCIATE EDITORS

JAMES KUTZNER
Comark Digital Services
Thomcast Communications
601 N. Fairfax St.
Alexandria, VA 22314, USA
Tel: +1 703 838 5659
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jkutzner@thomcastcom.com

DR. YIYAN WU
Communications Research Centre Canada
3701 Carling Ave., Box 11490, Station H
Ottawa, Ontario, Canada K2H 8S2
Tel: +1 613 998 2870
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DAVID H. LAYER
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 yiy.wu@crc.ca

DAVID H. LAYER
 National Association of Broadcasters
 1771 N. Street NW
 Washington, DC 20036, USA
 Tel: +1 202 429 5339
 Fax: +1 202 775 4981
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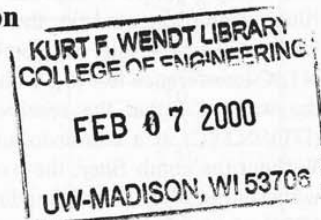
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Canadian Digital Terrestrial Television System Technical Parameters

Yiyan Wu, Pierre Bouchard, and Bernard Caron
Communications Research Centre Canada
Ottawa, Canada

Donald Tyrie and Royce Trenholm
Spectrum Engineering, Industry Canada
Ottawa, Canada



1. Abstract

On November 8, 1997, Canada formally adopted the ATSC Digital Television Standard for terrestrial transmission [1] as defined in document A/53 of the Advanced Television Systems Committee of the USA [2] and as modified by the FCC document MM Docket No. 87-268 [3]. This paper will discuss the Canadian digital terrestrial television system technical parameters for implementation and frequency planning: interference protection ratios, reference system parameters, and radio frequency emission masks, as well as their design methodology. First, the Canadian terrestrial DTV design guidelines are presented, followed by the introduction of DTV reference receiving system set up and the parameter selection of the interference protection ratios. Last, the design methodology of the emission masks is discussed.

2. Canadian DTV Service Objectives and Protection Ratios

The basic objective of the DTV service is to provide a coverage area replicating as closely as possible the existing NTSC service areas out to the protected or Grade B

contour. This implies that reliable service be provided to all locations presently served and that the quality of service be maintained throughout the coverage area.

The protected or Grade B contour of NTSC service is defined as the outer geographic limits within which the field strength at 10 metres above ground is sufficient to provide an acceptable picture at least 90 percent of the time at the best 50 percent of receiving locations, i.e., F(50, 90) availability. During the NTSC to DTV transition period, the desired service availability for DTV is F(50, 90). The interference criteria is F(50,10). Considerations for the future DTV-only plan is incorporated by providing DTV allotments using separation distances based on DTV service availability of F(90, 90) and DTV-DTV interference determination at F(10, 10) [4,5]. Table 1 lists the Canadian [4,5] and the USA [6, 14] DTV system protection ratios. The values in bracket are the measurement results from the Advanced Television Test Center (ATTC) [6]. In Table 1, it can be seen by comparing the first row with the fourth, that the co-channel DTV to DTV interference behaves very much like random noise [17].

Parameters	Canadian Parameters	USA Parameters
Carrier-to-Noise Ratio	+19.5 dB	+15.19 dB
Co-Channel DTV into NTSC	+33.8 dB	+34.44 dB (+33.81 dB)
Co-Channel NTSC into DTV	+7.2 dB	+1.81 dB
Co-Channel DTV into DTV	+19.5 dB	+15.27 dB
Lower Adj. Ch. DTV into NTSC	-16 dB	-17.43 dB (-15.96 dB)
Upper Adj. Ch. DTV into NTSC	-12 dB	-11.95 dB
Lower Adj. Ch. NTSC into DTV	-48 dB	-47.33 dB
Upper Adj. Ch. NTSC into DTV	-49 dB	-48.71 dB
Lower Adj. Ch. DTV into DTV	-27 dB	-28 dB
Upper Adj. Ch. DTV into DTV	-27 dB	-26 dB

Table 1: DTV and NTSC system protection ratios

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Channel Estimation Using Pilot Tones in OFDM Systems

Che-Shen Yeh and Yinyi Lin
Department of Electrical Engineering
National Central University, Chung Li
Taiwan 32054, R. O. C.

Abstract –

Channel estimation techniques using pilot tones in OFDM systems are investigated. Two time-domain channel estimation methods are proposed. Based on theoretical analysis and computer simulation, for a channel with multipath distortion and AWGN, the proposed time-domain channel estimation schemes perform better than the conventional frequency domain channel estimation techniques.

I. Introduction

Since terrestrial broadcast channels are often affected by multipath distortion, in a single carrier modulation system, an adaptive equalizer is usually employed to combat channel variations. However, in the Orthogonal Frequency Division Multiplexing (OFDM) modulation system, the channel estimation is accomplished using pilot tones inserted in OFDM symbols. In an OFDM system, the incoming signal is transmitted using N equally spaced carriers with a carrier spacing of $1/(NT)$ Hz, where T is the input data bit period. Multipath causes the channel transfer function to have variation in both spectral magnitude and spectral phase linearity. The number of carriers is chosen such that a typical channel transfer function is essentially constant over the bandwidth of each individual carrier. The effect of multipath distortion is then reduced to amplitude attenuation and phase rotation of data in each carrier, and the channel equalization is reduced to a complex multiplication bank in front of the decision circuit at the receiver [1-2].

An OFDM system can eliminate Intersymbol Interference (ISI) between OFDM symbols by inserting a cyclically extended guard interval, where each OFDM symbol is preceded by a periodic extension of data signal itself [3, 10]. The total symbol duration becomes $T_{total} = T_g + T_s$, where T_g is the guard interval and T_s is the useful symbol duration. When the guard interval T_g is longer than the channel impulse response $h[n]$, or the multipath delay spread, the ISI between OFDM symbols can be entirely eliminated. The ratio of the guard interval to useful symbol duration is dependent on application. A longer guard interval will reduce the system data throughput, and a commonly adopted compromise length is

$T_g \leq \frac{1}{4} T_s$. A cyclic prefix in the guard interval can be used to maintain the receiver carrier synchronization during a long silence. In addition, cyclic convolution can be used to model the transmission system.

However, the in-band fading will still exist. Therefore, let $H_k = \lambda_k e^{j\theta_k}$ be the response within the bandwidth of the k^{th} carrier. Equalization of the channel in the receiver requires that the FFT output of the k^{th} carrier be multiplied by a complex coefficient $c_k = \frac{1}{H_k}$. This operation will scale both

signal and noise/interference. In the OFDM system, the estimation of the channel response uses in-band pilot tones inserted into each symbol with a pilot ratio $PR = 1/R$. In this scheme, the channel response can be estimated rapidly and the tuner phase noise can be significantly reduced [10-11]. The optimum choice for the numbers of pilots ($=N/R$) for the AWGN channel is described in [4]. This scheme is referred to as scheme A, as depicted in Fig. 1a. In the AWGN channel every optimally spaced set of pilots can provide good channel estimation, so a fixed set of optimally spaced pilot tones is used in scheme A.

An alternative is to insert pilot tones into all of the carriers in an OFDM symbol with a period of R , as depicted in Fig. 1b. The channel response is then estimated using the pilot symbol, and the estimated response is used for subsequent OFDM symbols until the next pilot symbol is received. The channel response can also be estimated by averaging some consecutive pilot symbols. This scheme is referred to as scheme B. As will be shown, it is not suitable for fading channels.

II. Proposed Channel Estimation Techniques

A typical OFDM system is depicted in Fig. 2. On the transmitter site, a multiplexer (MUX) combines the data symbols and pilot signals as a data stream in the frequency domain, and the data stream is fed into an Inverse Fast Fourier Transform (IFFT) block. The IFFT block converts the data stream from a serial format into a parallel format, and transforms the signal from frequency domain to time domain. The guard interval is then inserted after each OFDM symbol to eliminate ISI between OFDM symbols.

Pilot tones are modulated with a PN sequence in the frequency domain and are inserted into each OFDM symbol with a ratio $PR = 1/R$ using scheme A. The channel response is estimated in the frequency domain using FFT processing and comparing received pilots with the locally stored reference pilots. This is referred to as the Frequency Pilot Frequency Interpolation (FPFI) technique since the channel response is obtained via frequency interpolation. The pilot tones can also be inserted into the OFDM symbols using scheme B. In this scheme, all pilot tones are inserted in the first OFDM symbol (pilot symbol) of a data frame. The channel response is obtained by averaging several consecutive pilot symbols and comparing the average with a locally stored reference symbol [5]. This technique is referred to as the Frequency Pilot Frequency Average (FPFA) technique.

In this paper, channel estimation conducted in the time domain is proposed. In one of the two proposed techniques, instead of the PN sequence the positive and negative polarization pilot signals are alternatively inserted into OFDM symbols in the frequency domain using FPFI. An OFDM symbol can be regarded as several time slots, with each time slot having a pilot. Assuming that the impulse response of each time slot is identical, one can obtain the channel response in the time domain by averaging these time slots. This technique is referred to as the Frequency Pilot Time Average (FPTA) technique, as depicted in Fig. 3a.

In the other technique a PN sequence is periodically inserted into the OFDM symbol in the time domain, and the channel response can be obtained via correlation in the time domain. This technique is referred to as the Time Pilot Time Correlation (TPTC) technique, as depicted in Fig. 3b. In this technique the channel response is estimated using the correlation of received pilot signal and the locally stored PN sequence [8]. If the transmitted pilot signal $p[n]$ is a PN sequence, the received pilot signal $\tilde{p}[n]$ can be expressed as $\tilde{p}[n] = p[n] \otimes h[n]$ in the time domain (where $\otimes$ represent the convolution operator) or $\tilde{P}[k] = P[k] \cdot H[k]$ in the frequency domain. Thus the channel response is given by $H[k] = \tilde{P}[k] / P[k] = \tilde{P}[k] \cdot P[k]^* / |P[k]|^2$. As the spectrum of a PN sequence is nearly constant at the frequency occupied by the received pilot signal, $|P[k]|^2$ can be approximated as a constant, K , which is equal to the length of the PN sequence. Thus the channel response can be estimated using

$$H[k] = \frac{\tilde{P}[k] \cdot P[k]^*}{K} \text{ in the frequency domain}$$

or

$$h[n] = \frac{1}{K} \sum_i \tilde{p}[n] \cdot p[n+i]^* \text{ in the time domain.}$$

III. Pilot to Noise (P/N) Ratio

This section is an analysis of the pilot to noise ratio for both time domain channel estimation and frequency domain channel estimation. Referencing to Fig 2 and Fig 3a, the pilot signal is $P[k]$. Suppose pilot ratio $PR = 1/R$, $P[k]$ can be shown as:

$$P[k] = \begin{cases} A, & k = Rm \\ 0, & k = Rm + \ell \end{cases} \quad (1)$$

Where R , m and ℓ are integers, $1 \leq \ell \leq R-1$, $0 \leq m \leq (N/R)-1$, A is the pilot amplitude and N/R is an integer.

$$p[n] = \text{IFFT}\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] e^{+j \frac{2\pi nk}{N}} \quad (2)$$

Where N is the IFFT transformation length

The multipath channel impulse response $h[n]$ is:

$$h[n] = \delta[n] + \sum_{i=1}^M \alpha_i \cdot \delta[n - \tau_i] = \sum_{i=0}^M \alpha_i \cdot \delta[n - \tau_i] \quad (3)$$

Where M is the total number of echoes

α_i represents the amplitude of i^{th} path echo

τ_i represents the delay of i^{th} path echo

$\alpha_0 = 1$, $\tau_0 = 0$, (representing the principle pilot signal)

$p[n]$ after multipath channel $h[n]$ are:

$$\tilde{p}[n] = p[n] * h[n] = p[n] + \sum_{i=1}^M \alpha_i p[n - \tau_i] \quad (4)$$

On the Rx side, the received frequency domain pilot signal $\tilde{P}[k]$ and multipath channel response $H[k]$ will be :

$$\tilde{P}[k] = P[k] \cdot H[k] = \begin{cases} \sum_{i=0}^M \alpha_i P[k] e^{j \frac{2\pi \tau_i k}{N}}, & k = mR \\ 0, & k \neq mR \end{cases} \quad (5)$$

$$H[k] = 1 + \sum_{i=1}^M \alpha_i e^{-j \frac{2\pi \tau_i k}{N}} \quad (6)$$

Assuming the time domain channel noise $n[n]$ is wide-sense stationary AWGN with zero mean and variance:

$$\sigma_{tn}^2 = \text{Var}\{n[n]\} = E\{n[n] - E\{n[n]\}\}^2 \quad (7)$$

Where $n[n]$ represents the time domain white noise at the n^{th} sample, $E\{n[n]\}$ represents the expected value of $n[n]$.

In radio channel, suppose the noise is zero mean AWGN,

$E\{n[n]\} = 0$. On the Rx side, after the FFT, the $n[n]$ signal will become the frequency domain noise

$$N[k] = \sum_{n=0}^{N-1} n[n] W_N^{kn}, W_N = e^{-j\frac{2\pi}{N}} \quad (8)$$

Assuming the variance of frequency domain noise $N[k]$ is σ_{FN}^2 , according to variance definition:

$$E\{N[k]\} = \sum_{n=0}^{N-1} E\{n[n]\} W_N^{kn} = 0$$

$$\begin{aligned} \sigma_{FN}^2 &= \text{Var}\{N[k]\} \\ &= E\{(N[k] - E\{N[k]\}) \cdot (N[k] - E\{N[k]\})^*\} \\ &= E\{N[k] \cdot N^*[k]\} \\ &= E\left\{\left(\sum_{n=0}^{N-1} n[n] \cdot W_N^{kn}\right) \cdot \left(\sum_{m=0}^{N-1} n[m] \cdot W_N^{km}\right)^*\right\} \\ &= \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} E\{n[n] \cdot n^*[m]\} W_N^{k(n-m)} \end{aligned} \quad (9)$$

Because $n[n]$ and $n[m]$ are independent and zero mean, Eq.(9) can be re-written as:

$$\begin{aligned} \sigma_{FN}^2 &= \sum_{n=0}^{N-1} \left\{ E\{n[n] \cdot n[n]\} W_N^0 + \sum_{\substack{m=0 \\ m \neq n}}^{N-1} E\{n[n]\} \cdot E\{n^*[m]\} W_N^{k(n-m)} \right\} \\ &= \sum_{n=0}^{N-1} (\sigma_{in}^2 + 0) = N \cdot \sigma_{in}^2 \end{aligned} \quad (10)$$

From the previous mathematical deductions, in the radio channel of frequency domain, the noise variance is being enhanced N times while the noise amplitude is amplified $\sqrt{N}$ times.

In the FPTA, pilot ratio $PR = 1/R$ and the pilot signal $P[k]$ is shown as :

$$\begin{aligned} P[k] &= \begin{cases} (-1)^m \cdot A, & k = Rm \\ 0, & k = Rm + \ell \end{cases} \\ &= \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \delta[k - Rm] \end{aligned} \quad (11)$$

Where $m = 0, 1, 2, \dots, (N/R)-1$, ℓ is an integer and $1 \leq \ell \leq R-1$. The time domain TX pilot signal $p[n]$ is shown as :

$$\begin{aligned} p[n] &= \text{IFFT}\{P[k]\} = \frac{1}{N} \sum_{k=0}^{N-1} P[k] \cdot e^{+j\left(\frac{2\pi kn}{N}\right)} \\ &= \frac{1}{N} \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot e^{+j\left(\frac{2\pi Rmn}{N}\right)} \\ &= \frac{A}{R} \cdot \frac{1}{N/R} \sum_{m=0}^{(N/R)-1} e^{+j2\pi m \left(\frac{n - (N/2R)}{N/R}\right)} \\ &= \frac{A}{R} \sum_{r=0}^{R-1} \delta\left[n - \frac{Nr}{R} - \frac{N}{2R}\right] \\ &= \begin{cases} A/R, & n = Nr/R + N/2R \\ 0, & \text{otherwise} \end{cases} \end{aligned} \quad (12)$$

From the result of Eq.(12), when n is equal to $N/2R$ plus the integer time of N/R , the time domain pilot will be added in-phase. The Rx pilot signal $\tilde{p}[n]$ in the time domain and $\tilde{P}[k]$ in the frequency domain will be :

$$\begin{aligned} \tilde{p}[n] &= p[n] \otimes h[n] \\ &= \frac{A}{R} \sum_{u=0}^{N-1} \left\{ \left(\sum_{r=0}^{R-1} \delta\left[u - \frac{Nr}{R} - \frac{N}{2R}\right] \right) \cdot \left(\sum_{i=0}^M \alpha_i \delta[n - u - \tau_i] \right) \right\} \\ &= \frac{A}{R} \sum_{r=0}^{R-1} \sum_{i=0}^M \alpha_i \cdot \delta\left[n - \left(\tau_i + \frac{Nr}{R} + \frac{N}{2R}\right)\right] \end{aligned} \quad (13)$$

$$\begin{aligned} \tilde{P}[k] &= P[k] \cdot H[k] \\ &= \begin{cases} \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \delta[k - Rm] \cdot \left(1 + \sum_{i=1}^M \alpha_i \cdot e^{-j\frac{2\pi \tau_i k}{N}} \right) \\ \sum_{m=0}^{(N/R)-1} (-1)^m \cdot A \cdot \left[\sum_{i=1}^M \alpha_i \cdot e^{-j\frac{2\pi \tau_i m}{N/R}} \right], & k = Rm \\ 0, & k \neq Rm \end{cases} \end{aligned} \quad (14)$$

where $\alpha_0=1$, $\tau_0=0$, (it is the subscript 0 that represents the principle pilot component).

In the FPFI method, $P[k]$ = PN-Sequence, $p[n]$ and $P[k]$ are both noise-like and randomly distributed in the time domain OFDM symbol. Hence, after FFT on the Rx end, the result is similar to AWGN case described above, and $\tilde{P}[k]$ and $\tilde{p}[n]$ both have $\sqrt{N}$ gain. Therefore, in the second (FPFI) method the P/N in the time domain and in the frequency domain are about the same, so using pilot signal for channel model estimation in the time domain is as good as in the frequency domain.

If the FPFI method is used for channel estimation with the pilot signal in the frequency domain, the pilot to noise ratio is $\tilde{P}[k]/\sigma_{FN}$. If the FPTA method is used for channel estimation with the pilot signal in the time domain, the pilot to noise ratio

is $\tilde{p}[n]/\sigma_{in}$, which implies that $\tilde{P}[k]$ and $\tilde{p}[n]$ are affected by the echo. The amplitude of $\tilde{p}[n]$ has different variation, which depends upon both the amplitude of echo and delay spread. Different multipath models result in different variation of $\tilde{P}[k]$. Hence, it is difficult to characterize generally the variation. This variation may differ with different channel models, hence, it is difficult to describe or derive with such unpredictable value compared with the principle pilot signal to noise variance ratio. However, the third method repeats an average of R times, after which it reduces σ_{in} to $\sigma'_{in} = \sigma_{in}/\sqrt{R}$. Assuming that the pilot interference of data to pilot tone can be ignored, we may compare the principle pilot to noise of these two methods:

$$\frac{p[n]/\sigma'_{in}}{P[k]/\delta_{FN}} = \frac{A/R \cdot (\sigma_{in}/\sqrt{R})}{A \cdot (\sqrt{N} \cdot \sigma_{in})} = \sqrt{\frac{N}{R}} \quad (15)$$

The FPTA estimation performance is better than the FPFA and FPFi. This improvement will be verified in the next section. In the derivation of Eq. (15), the FPTA does not take the "data ghosts" into account, thus, the gain of the time domain $\tilde{p}[n]/\delta_{in}$ over the frequency domain $\tilde{P}[k]/\delta_{FN}$ is actually lower than the value of $\sqrt{N/R}$ indicated in Eq.(15).

The interference of training sequence coming from data is referenced to as "data ghosts". A specific comparison of channel estimation will be given in the next section.

IV. Performance Comparisons and Discussions

The performance comparison of previously mentioned estimation techniques is made on several channel models recommended by the Advanced Television Technology Center (ATTC) in the U. S. A. and by the Digital Audio Broadcasting (DAB) project in Europe [6-7]. During simulations, assume that both transmitter and receiver are perfectly synchronized and the guard interval is long enough so that ISI between OFDM symbols can be ignored. The channel is assumed to be the additive white Gaussian noise (AWGN) channel.

Static Channel

Two channel models: the multipath Ensembles A and E of the ATTC, as recommended by the US ATTC and the Grand Alliance DTV laboratory [6][7], are used in simulations: Ensemble A as channel model 1:

$$h_1[n] = 0.1\delta[n+16] + \delta[n] + 0.1\delta[n-2] + 0.3162\delta[n-16] + 0.1995\delta[n-52] + 0.1295\delta[n-168]$$

Ensemble E as channel model 2:

$$h_2[n] = \delta[n] + 0.3162\delta[n-2] + 0.1995\delta[n-17] + 0.1296\delta[n-36] + 0.1\delta[n-75] + 0.1\delta[n-137]$$

In the above two formulas, the unit delay length $T = 1.1 \times 10^{-7}$ seconds. In addition, the simplified non-hilly urban channel model recommended by DAB test (pr EN50248) is also used in simulations as channel model 3:

$$h_3[n] = 0.707\delta[n+2] + \delta[n] + 0.707\delta[n-3] + 0.5\delta[n-13] + 0.4\delta[n-19] + 0.31625\delta[n-44]$$

Parameters	Specifications
Number of carriers (N)	8192
Pilot ratio (PR=1/R)	1/8, 1/32
Guard interval	1/8, 1/32
Signal constellation	64QAM
Channel model	h1(n), h2(n), h3(n)
S/N	0 ~ 30 dB

Table 1. OFDM system performance simulation parameters

The OFDM system parameters used in simulations are summarized in Table.1. All data carriers in the OFDM are 64-QAM modulated signals (signal constellation (x,y) with $x, y \in \{\pm 1, \pm 3, \pm 5, \pm 7\}$), and the average power of a data symbol is $P_s = 42$ [9]. The pilot signals with amplitude ± 7 are used.

The error performance is displayed in Figs 4 and 5 for various estimation techniques on the channel model 1 with pilot ratio 1/8 and 1/32 respectively. A similar result was found on the other two channel models and is not shown here. The results demonstrate that the proposed estimation techniques (i.e., FPTA and TPTC) outperform the conventional estimation techniques (i.e., FPFA and FPFi). A SNR improvement gain of 2-3 dB can be achieved using the time-domain estimation technique. This is because a higher P/N ratio can be obtained in the time domain. For pilot ratio PR = 1/32 on the television broadcasting channels (channel models 1 and 2), the error floor occurs using FPFi technique, as shown in Fig. 5. This is due to the aliasing effect introduced by the frequency interpolation. However, no error floor was found on the audio broadcasting channel (channel model 3). Since this channel has a shorter delay spread and the aliasing effect is not severe.

Extensive simulations using FPFi technique were conducted using different pilot ratios (i.e., 1/4, 1/8, 1/16, and 1/32, respectively). It was found that with pilot ratios PR = 1/4 and 1/8, no error floor occurs for all the three channel models. However with pilot ratio PR = 1/16, the error floor occurs at

8.2×10^{-4} and 1.35×10^{-4} for channels 1 and 2 respectively, although no error floor was found for channel 3. As the pilot ratio PR was reduced to 1/32, a more severe error floor occurs (0.335 and 0.0665 for the two DTV channel models), although there was still no error floor found on the DAB channel.

Channel Estimation Accuracy

In addition to symbol error rate performance, the mean square error between estimated response and real channel response was also compared to demonstrate the superiority of the time-domain estimation techniques. The system parameters used for comparison are displayed in Table 2 and the results are illustrated in Fig. 6. As shown, a smaller mean square error was found in the time-domain channel estimation techniques. The FPFA technique demonstrates a larger estimation error than the FPFI technique since in the FPFA the estimation error extends over several OFDM symbols, while in the FPFI the estimation error impacts only adjacent carriers within one OFDM symbol. The estimation error in the FPTA is larger than that in the TPTC because the FPTA is affected by the components of "data ghosts" which are non-zero, even after averaging several OFDM symbols. According to equation (15), the FPTA method outperforms FPFI method with a P/N gain about 30 dB. Figure (6) shows a P/N gain about 25.2 dB. The difference between 4.8 dB is due to "data ghosts".

Parameters	Specifications
Number of carriers (N)	8192
Pilot ratio (PR = 1/R)	1/8
Guard interval	1/8
Signal constellation	64QAM
Channel model	$h_1[n], h_2[n], h_3[n]$
S/N	10 dB, 20dB
Mean Average Number	16

Table 2 System parameters

Slow Fading Channel

The error performance of the estimation techniques is also compared on a mobile time-varying channel using the autoregressive (AR) model given as

$$h[k+1] = ah[k] + w[k]$$

Where a and $w[k]$ represent slow time variations and the noise vector sequence of the AWGN, respectively. It is assumed that the channel fading is slow enough so that the channel impulse response doesn't change within the duration of one OFDM symbol [4].

During simulations, parameters of the AR model are

assumed as $a = 0.99\pi$ and $w[k] = 0.0001\pi$, where π is the unit matrix. The OFDM parameters displayed in Table 1 and the three channel models previously mentioned are used. The error rate is shown in Fig. 7 with pilot ratio PR = 1/8. The results also reveal that both FPTA and TPTC provide better performance than those estimated in the frequency domain (FPFI and FPFA). Although both FPFA and FPFI techniques achieve almost the same error performance in static channels, the performance of the FPFA is slightly worse than that of the FPFI. This is because in the FPFA the channel response is estimated by averaging R pilot symbols (in this case, 8 pilot symbols were averaged), which cannot accurately estimate the response of each OFDM symbol. The effect of the pilot ratio on the error performance is also simulated, and the result with pilot ratio PR = 1/32 is demonstrated in Fig. 8 for channel 2. Again the error floor was found for the FPFI due to the aliasing effect. The error floor was also found in the FPFA, in which the channel was estimated by averaging 32 OFDM symbols. The error floor occurs when the accumulated estimation error due to channel variations becomes not negligible as compared to the estimation error caused by channel noise.

V. Summary

In this paper the channel estimation techniques (FPTA and TPTC) in the time domain using pilot tones for the OFDM system are presented. It is shown that a higher P/N ratio can be obtained in the time domain and that the proposed estimation techniques outperform the conventional estimation techniques (FPFI and FPFA) in which the channel response is estimated in the frequency domain. In addition, both FPFI and FPFA suffer from the aliasing effect and channel mis-estimation resulting in error floors, especially for low pilot ratios.

The proposed FPTA and TPTC techniques have advantages of low computational complexity and high P/N ratio. FPTA is suitable for fading channels with shorter transmission range, i.e., Local Multipoint Distribution Systems (LMDS), which has shorter channel response. FPFI and FPFA can be used to reduce overhead for systems with lower pilot ratios (i.e., 1/32, 1/64). However, in the FPTA the pilot signal consumes more RF power and interferes with data signal when the pilot ratio is high. The TPTC does not have this disadvantage.

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Che-Shen Yeh received B.S. and M.S.E.E. in Electrical Engineering from Chung-Yuan Christian University, Jung-Li, Taiwan in 1980 and 1986 respectively, and is currently a Ph.D. candidate in Electrical Engineering at the National Central University, Jung-Li, Taiwan. His major research interests are digital transmission for television and audio. He leads the DTV/DAB transmission project at Telecommunications Labs of Chunghwa Telecom Co., Ltd. Taiwan, and is also responsible for DTV/DAB field trials and the development of DTV/DAB transmission system standards and regulations. He also is a member of the Task Force for SDTV/HDTV transmission standards and regulation at the Ministry of Transportation and Communications, Taiwan.

Yinyi Lin was born in Chuang Hua, Taiwan, R.O.C., in 1959. He received the B.S. degree in communication engineering from National Chiao Tung University, Hsin Chu, Taiwan in 1981, and M.S. and Ph.D. degrees in electrical engineering from the University of California, San Diego in 1986 and 1989, respectively.

From 1983 to 1985, he has employed by the Electronics Research and Service Organizations (ERSO), I.T.R.I., Taiwan. During 1988, he spent six months at the IBM Almaden Research Center, San Jose, CA working in the area of signal processing for digital magnetic recording. From 1990 to 1991, he was with Archive Corporation, Costa Mesa, CA, where he engaged in the research of quarter inch cartridge (QIC) tape system. Since 1991, he has been with the Electrical Engineering Department of National Central University, Jung-Li, Taiwan, where he is now a Professor. His current research interests include coding and signal processing for digital magnetic recording systems, digital compression, and error correction for documents.

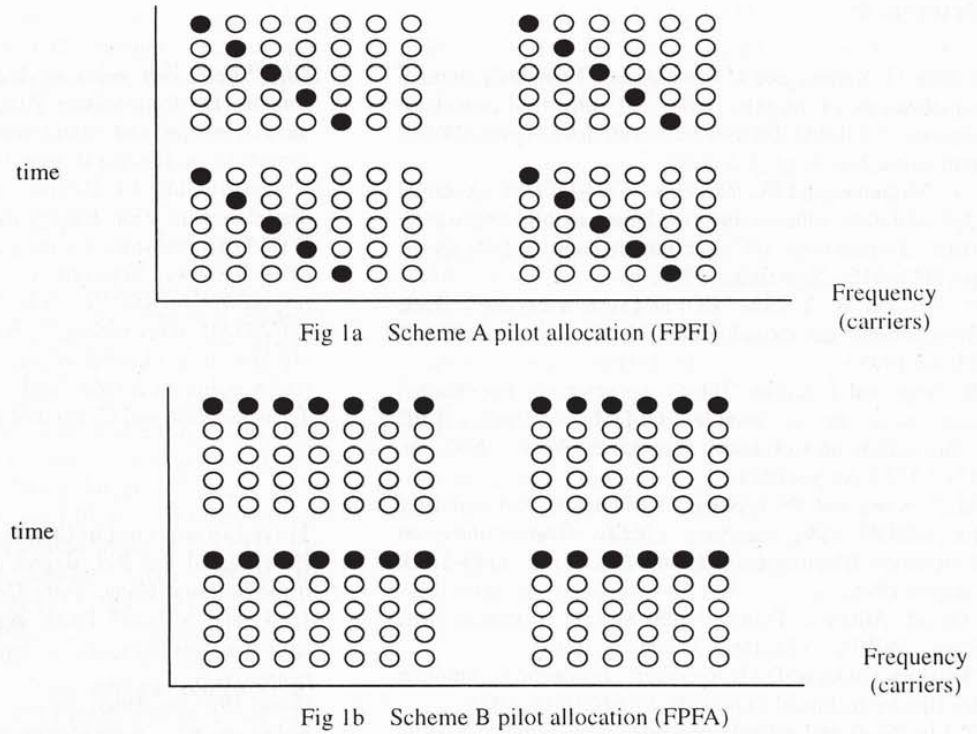


Figure 1 OFDM system frequency domain pilot signal allocation

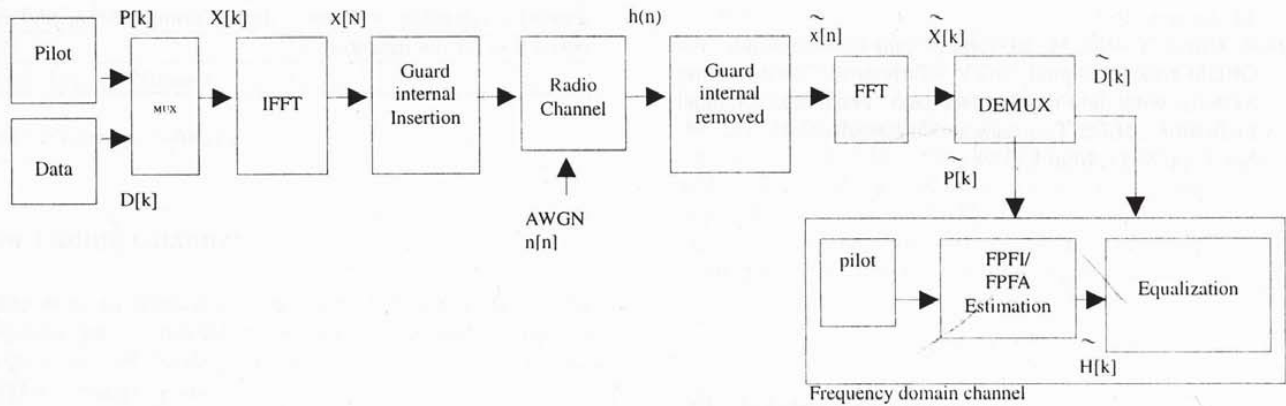


Figure 2 Channel model estimation of frequency pilot in the frequency domain

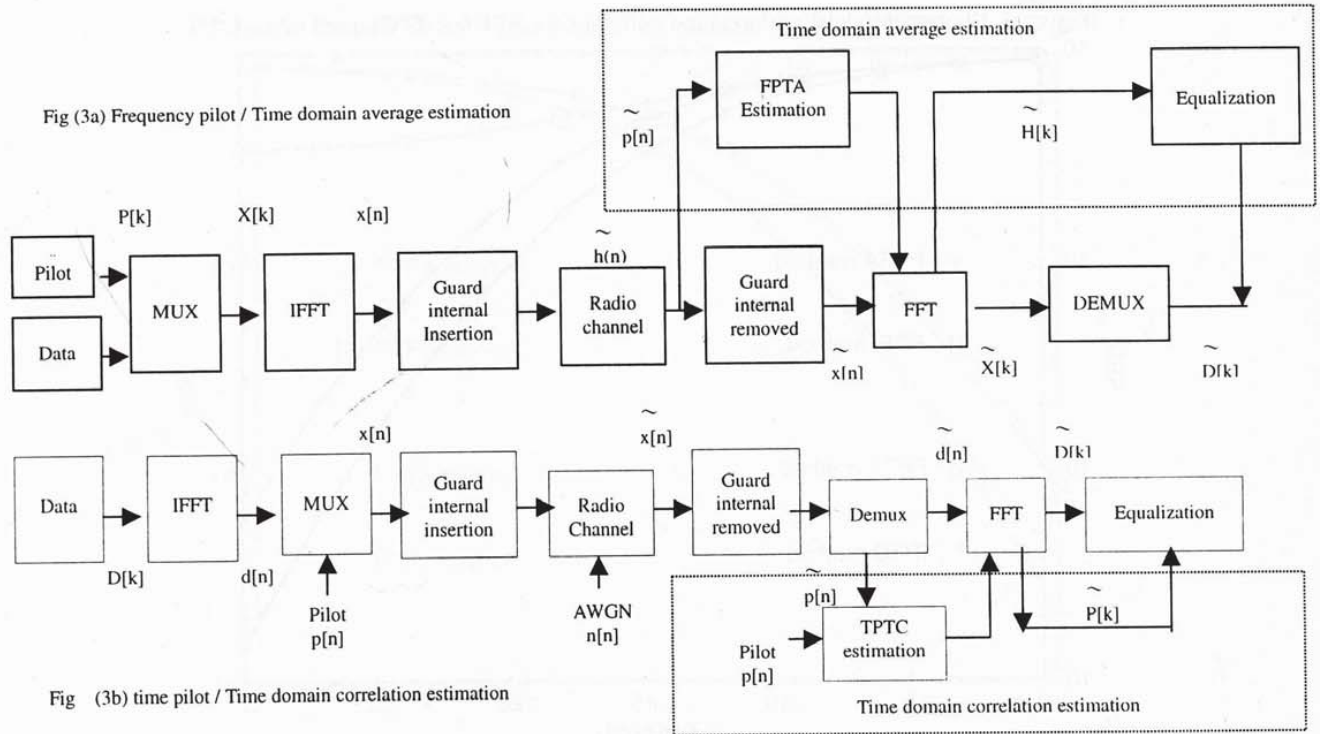


Figure. 3 Proposed Estimation Techniques (a) FPTA (b) TPTC

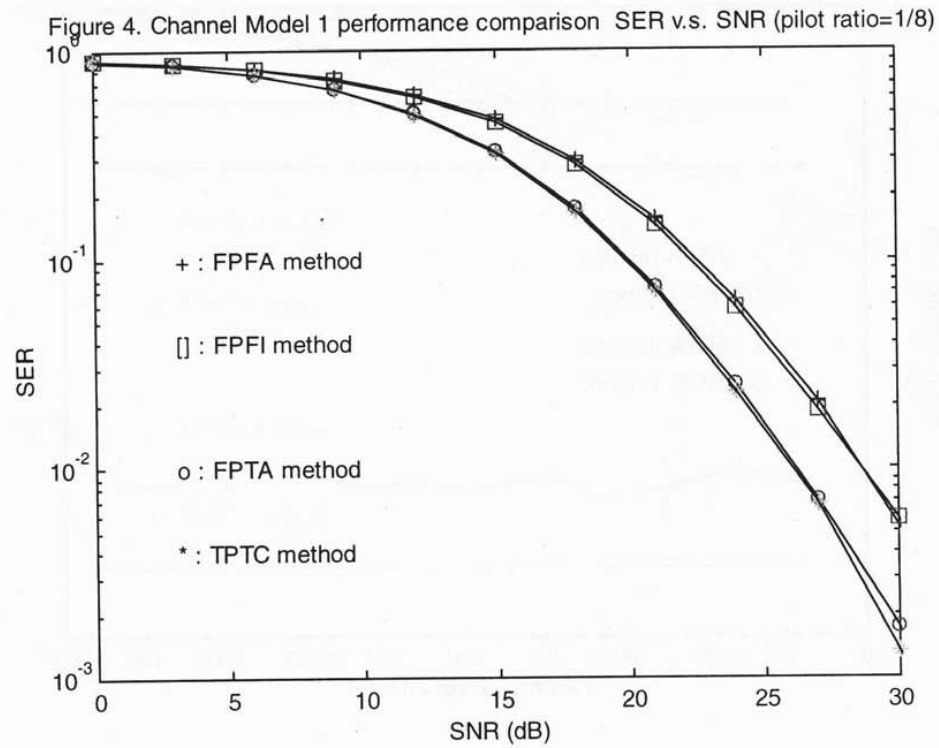


Figure 5. Channel Model 1 performance comparison SER v.s. SNR (pilot ratio=1/32)

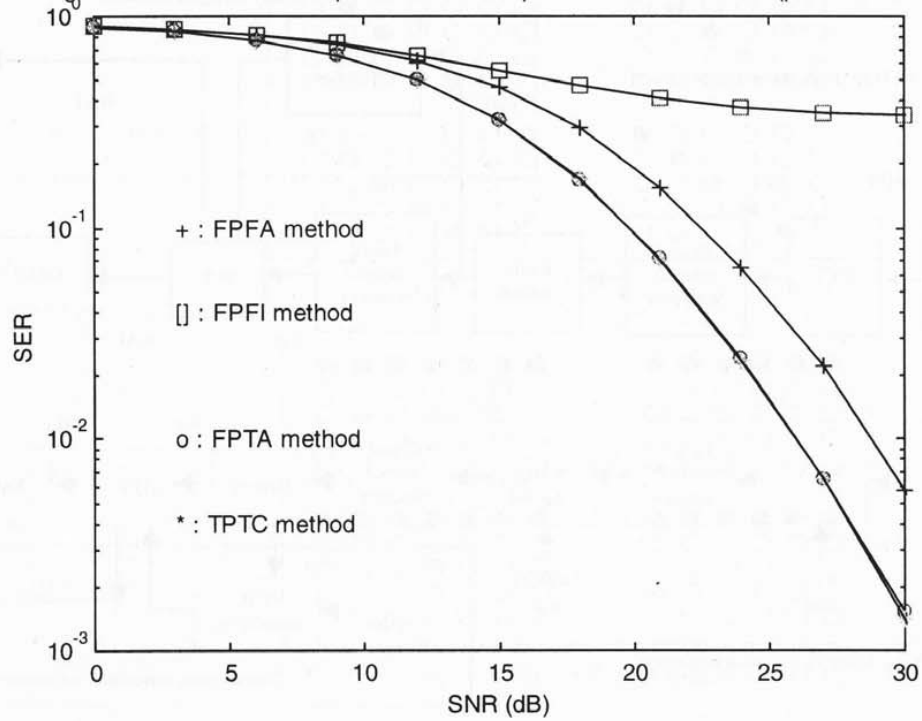


Figure 6. Channel Model 5 estimation MSE (CNR=10dB)

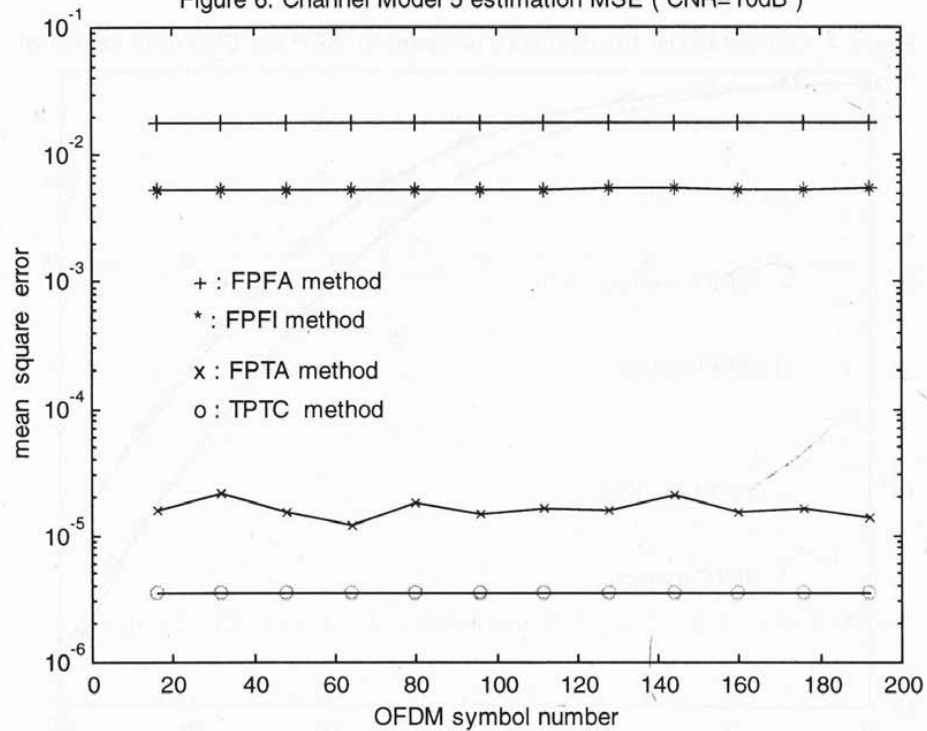


Figure 7. Channel Model 1 performance comparison SER v.s. SNR (pilot ratio=1/8)

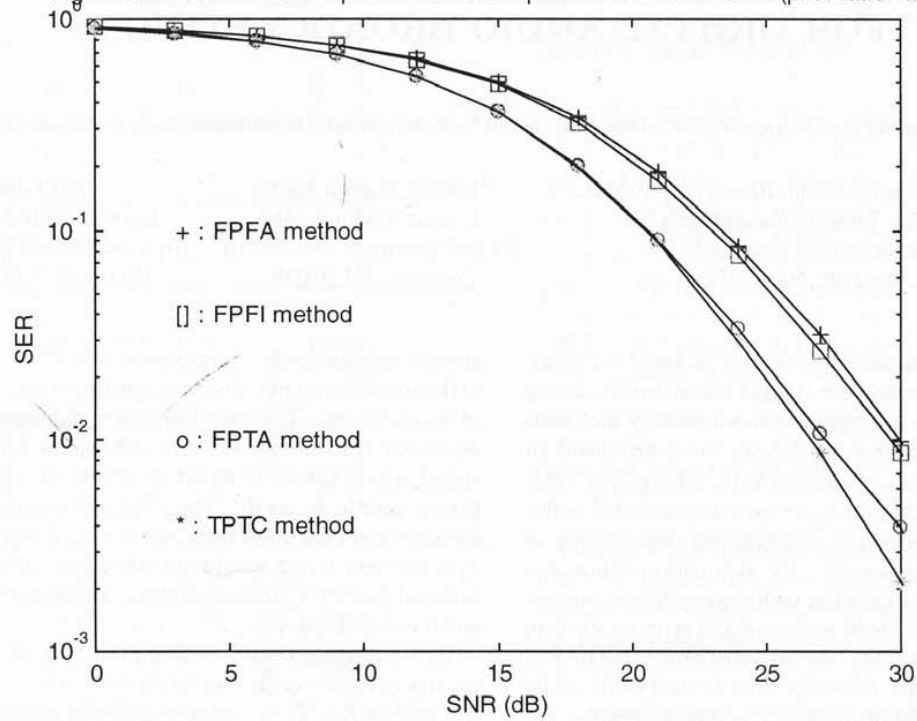
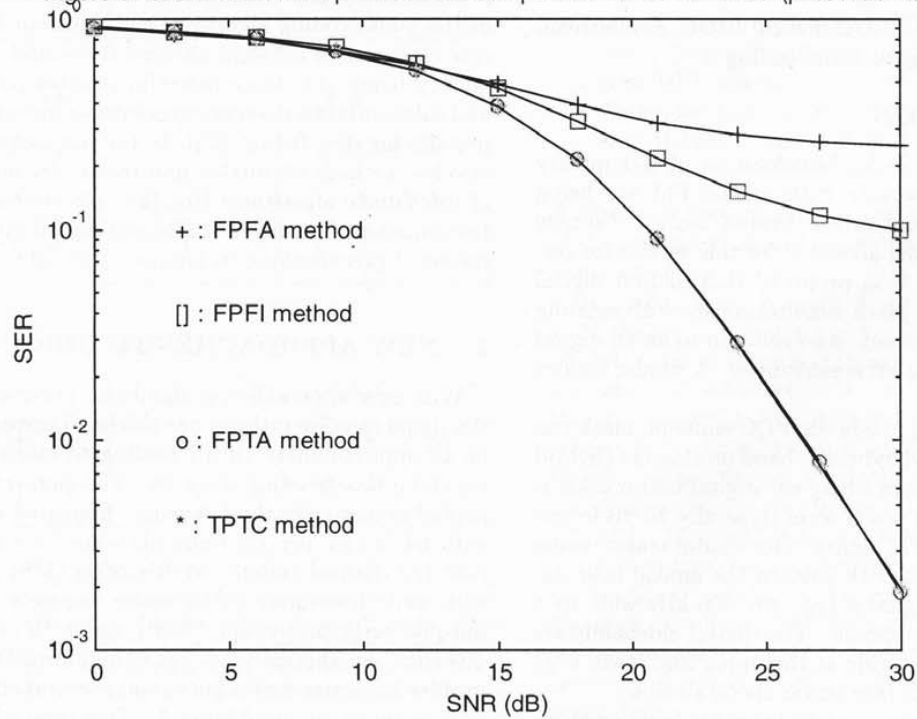


Figure 8. Channel Model 2 performance comparison SER v.s. SNR (pilot ratio=1/32)



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Appendix 0003D

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An Overview of Smart Antenna Technology for

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An Overview of Smart Antenna Technology for

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An Overview of Smart Antenna Technology for Mobile Communications Systems

Per H. Lehne and Magne Pettersen, Telenor Research and Development

Abstract

The technology of smart or adaptive antennas for mobile communications has received enormous interest worldwide in recent years. The purpose of this article is to give an overview of the technology and of the different ways in which it is foreseen to influence mobile communications systems. A broad definition of smart antennas is used in this article, namely base station antennas with a pattern that is not fixed, but adapts to the current radio conditions. The article starts with an explanation of the basic principles of smart antennas. Different levels of intelligence are introduced, ranging from simple switching between predefined beams to optimum beamforming. Also, an evolutionary path for the development of smart antennas is suggested. A discussion of the consequences for mobile systems is given. The

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An Overview of Smart Antenna Technology for Mobile Communications Systems

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The technology of smart or adaptive antennas for mobile communications has received enormous interest worldwide in recent years. The purpose of this article is to give an overview of the technology and of the different ways in which it is foreseen to influence mobile communications systems. A broad definition of smart antennas is used in this article, namely base station antennas with a pattern that is not fixed, but adapts to the current radio conditions. The article starts with an explanation of the basic principles of smart antennas. Different levels of intelligence are introduced, ranging from simple switching between predefined beams to optimum beamforming. Also, an evolutionary path for the development of smart antennas is suggested. A discussion of the consequences for mobile systems is given. The principle reason for introducing smart antennas is the possibility for a large increase in capacity: an increase of three times for TDMA systems and five times for CDMA systems has been reported. Other advantages include increased range and the potential to introduce new services. Major drawbacks and cost factors include increased transceiver complexity and more complex radio resource management. The basic principles for implementation of smart antennas at the base stations are explained. First, the general principles for beamforming using array antennas are explored, then the special cases of different smart antenna implementations are discussed. Also, the fundamental difference between performing beamforming on uplink (base station receiving) and downlink (base station transmitting) is dealt with. Special attention is given to the critical factors and technological challenges, including achieving equal performance on uplink and downlink as well as real-time calibration of the receiver and transmitter chains. The article concludes with a short description of some of the trials and testbeds that have been implemented. In the autumn of 1998 Ericsson, in cooperation with Mannesmann Mobilfunk in Germany, became the first to allow commercial traffic through a base station employing smart antennas.

It is foreseen that in the future an enormous increase in traffic will be experienced for mobile and personal communications systems. This is due both to an increased number of users as well as new high bit rate data services being introduced. This trend is observed for second-generation systems and will most certainly continue for third-generation systems being introduced worldwide within a few years.¹ The increase in traffic will put a demand on both manufacturers and operators to provide enough *capacity* in the networks. Presently, one of the most promising techniques for increasing the capacity in cellular systems by the amount necessary is *smart* or *adaptive antennas*.

This article will describe the basic principles behind the technology as well as provide a discussion about the advantages and technical challenges. Although it is feasible to also use these techniques in, for example, satellite systems and high frequency wireless local area networks, the discussion here is limited to cellular systems in the frequency range 1-2 GHz. At these frequencies it is difficult to introduce smart antennas into

the mobile stations due to the limited size of the terminal in terms of wavelengths.

Capacity Aspects

In a cellular system the radio communication is between the user and a *base station*, which provides radio coverage within a certain area, which is called a *cell*. Capacity in such a system can be defined as the total bit rate per unit bandwidth per unit area, or bit/s/Hz/m^2 . Because the available frequency band is limited, the capacity is given by the *cell density*, the *frequency reuse distance* and the number of users that can be served simultaneously by each base station. Techniques for increasing capacity in cellular systems have included using smaller cells, so-called *microcells*, and *frequency hopping*, a technique that disperses interference and averages the fading rate.

What are Smart Antennas?

Base station antennas have up till now been omnidirectional or sectored. This can be regarded as a "waste" of power as most of it will be radiated in other directions than toward the user. In addition, the power radiated in other directions will be experienced as interference by other users. The idea of smart antennas is to use base station antenna patterns that are not fixed, but adapt to the current radio conditions. This can be visualized as the antenna directing a beam toward the communication partner only. The difference between the fixed and the smart antenna concept is illustrated in [Fig. 1](#). Smart antennas will lead to a much more efficient use of the power and spectrum, increasing the useful received power as well as reducing interference.

Sharing the Radio Spectrum

Traditionally, users communicating via the same base station have been separated by frequency, as in FDMA (frequency division multiple access); by time, as in TDMA (time division multiple access); or by code, as in CDMA (code division multiple access). Smart antennas add a new way of separating users, namely by space, through SDMA (*space division multiple access*). As will be explained later, SDMA, which means that users in the same cell can use the same *physical communication channel*, is the final step in an evolutionary path toward increasingly more advanced utilization. The *physical communication channel* is defined as a combination of carrier frequency, time slot and spreading code.

Why Now?

In the last couple of years smart antennas for mobile communications have received enormous interest worldwide. So why is the interest appearing now, and not five or ten years ago? The answer probably lies with the fact that for most operators there has not been much reason to worry about capacity and spectrum efficiency. Also, as will be apparent later on, if the base station is to track a large number of users simultaneously, the computational cost will be large. Only recently are processors with sufficient power becoming available. In addition to increased capacity, smart antennas also introduce a number of other advantages to cellular networks, including *increased range*, *a higher level of security*, and the *possibility for new services*. This will be discussed in more detail later.

Outline

This article consists mainly of four parts. In the next section definitions of terms and descriptions of the basic principles are given. There is a discussion of the consequences of introducing the new technology into mobile communications networks. Later, implementation issues are discussed in more detail. We also describe some of the field trials and testbeds for smart antennas that have been carried out.

The general principles are valid for both TDMA systems (e.g., GSM) and CDMA systems (e.g., IS-95 and UMTS). In this article, most reported techniques and test results stem from TDMA-based testbeds, so the discussion is somewhat limited to this case. A comprehensive treatment of smart antennas applied to CDMA is, however, given by Liberti and Rappaport in [1].

Basic Principles

What do we mean by the term "smart antenna?" The theory behind smart antennas is not new. The technique has for many years been used in electronic warfare (EW) as a countermeasure to electronic jamming. In military radar systems similar techniques were already used during World War II. There are in principle a number of ways in which an adaptively adjustable antenna beam can be generated, for instance by mechanically steered antennas. However, the technology almost exclusively suggested for land-based mobile and personal communications systems is *array antennas*. This technology will be explained in more detail later.

The main philosophy is that interferers rarely have the same geographical location as the user. By maximizing the antenna gain in the desired direction and simultaneously placing minimal radiation pattern in the directions of the interferers, the quality of the communication link can be significantly improved. In personal and mobile communications, the interferers are other users than the user being addressed.

Concepts

Several different definitions for smart antennas are used in the literature. One useful and consistent definition can be that the difference between a smart/adaptive antenna and a "dumb"/fixed antenna is the property of having an adaptive and fixed lobe-pattern, respectively. This is the definition to be used in this article. [Figure 2](#) illustrates the concept of a smart antenna.

Normally, the term "antenna" comprises only the mechanical construction transforming free electromagnetic (EM) waves into radio frequency (RF) signals traveling on a shielded cable and vice versa. We may call it *the radiating element*. In the context of smart antennas, the term "antenna" has an extended meaning. It consists of a number of radiating elements, a combining/dividing network and a control unit. The control unit can be called the smart antenna's intelligence, normally realized using a digital signal processor (DSP). The processor controls feeder parameters of the antenna, based on several inputs, in order to optimize the communications link. Different optimization criteria can be used, as will be explained in the next section. This shows that smart antennas are more than just the "antenna," but rather a complete transceiver concept.

Levels of Intelligence

Based on the definition above, one can define "levels of intelligence." One set of definitions often used are described below and illustrated in [Fig. 3](#).

- *Switched lobe (SL)*: This is also called switched beam. It is the simplest technique, and comprises only a basic switching function between separate directive antennas or predefined beams of an array. The setting that gives the best performance, usually in terms of received power, is chosen. Because of the higher directivity compared to a conventional antenna, some gain is achieved. Such an antenna will be easier to implement in existing cell structures than the more sophisticated adaptive arrays, but it gives a limited improvement.
- *Dynamically phased array (PA)*: By including a *direction of arrival* (DoA) algorithm for the signal received from the user, continuous tracking can be achieved and it can be viewed as a generalization of the switched lobe concept. In this case also, the received power is maximized.
- *Adaptive array (AA)*: In this case, a DoA algorithm for determining the direction toward interference

sources (e.g., other users) is added. The radiation pattern can then be adjusted to null out the interferers. In addition, by using special algorithms and space diversity techniques, the radiation pattern can be adapted to receive multipath signals which can be combined. These techniques will maximize the *signal to interference ratio* (SIR) (or signal to interference and noise ratio (SINR)).

Conventional mobile systems usually employ some sort of antenna diversity (e.g., space or polarization diversity). Adaptive antennas can be regarded as an extended diversity scheme, having more than two diversity branches. In this context, phased arrays will have a greater gain potential than switched lobe antennas because all elements can be used for diversity combining [2, 3]. All these concepts are analyzed in more detail later.

Evolutionary Path

All the levels of intelligence described in the previous section are technologically realizable today. However, in the domain of personal and mobile communications, an evolution can be foreseen in the utilization of smart antennas toward gradually more advanced solutions. The evolution can be divided into three phases [4]:

- Smart antennas are used on *uplink* only (uplink means that the user is transmitting and the base station is receiving). By using a smart antenna to increase the gain at the base station, both the sensitivity and range are increased. This concept is called *high sensitivity receiver* (HSR) and is in principle not different from the diversity techniques implemented in today's mobile communications systems.
- In the second phase, directed antenna beams are used on the *downlink direction* (base station transmitting and user receiving) in addition to HSR. In this way, the antenna gain is increased both on uplink and downlink, which implies a spatial filtering in both directions. Frequencies can be more closely reused, thus the system capacity increases. The method is called *spatial filtering for interference reduction* (SFIR). It is possible to introduce this in second-generation systems.
- The last stage in the development will be full *space division multiple access* (SDMA). This implies that more than one user can be allocated to the same physical communications channel simultaneously in the same cell, only separated by angle. In a TDMA system, two users will be allocated to the same time slot and carrier frequency at the same time and in the same cell.

In phase 2, the capacity is increased due to closer frequency reuse allowing more carriers per base station. In phase 3, an additional increase in capacity is achieved by allowing more users per carrier. Introducing SDMA in second-generation TDMA systems will be difficult and maybe undesirable [5], but it may become a natural component in third-generation systems.

The "levels of intelligence" in the previous section describe the level of technological development, while the steps described here can be regarded as part of a system evolution.

Consequences

The introduction of smart antennas will have a large impact on the performance of cellular networks. It will also affect many aspects of both the planning and deployment of mobile systems. This chapter will discuss the potential benefits and cost factors, and will also briefly describe the implications on radio planning.

Improvements and Benefits

Capacity Increase -- The principle reason for the growing interest in smart antennas is the capacity increase. In densely populated areas mobile systems are normally interference-limited, meaning that interference from

other users is the main source of noise in the system. This means that the signal to interference ratio, SIR, is much larger than the signal to thermal noise-ratio, SNR. Smart antennas will on average, by simultaneously increasing the useful received signal level and lowering the interference level, increase the SIR. Especially, the adaptive array will give a significant improvement. Experimental results report up to 10 dB increases in average SIR in urban areas [6].

In TDMA systems, the implication of the increased SIR is the possibility for reduced frequency reuse distance. An example is shown in [Fig. 4](#), where the traditional seven-cell cluster has been reduced to a three-cell cluster. This will lead to a capacity increase of 7/3 as all the cells can be allocated this amount more carriers.

Simulations performed on a FH-GSM network with 1/3 reuse distance utilizing SFIR reports that a capacity increase of 300 percent can be expected [5].

CDMA systems, such as IS-95 or UMTS, are more inherently interference-limited than TDMA systems. The main source of noise in the system is the interference from other users due to the spreading codes being non-ideally orthogonal. This means that the expected capacity gain is even larger for CDMA than for TDMA. A fivefold capacity gain has been reported for CDMA in [7].

Range Increase -- In rural and sparsely populated areas radio coverage rather than capacity will give the premises for base station deployment. Because smart antennas will be more directive than traditional sector or omnidirectional antennas, a range increase potential is available. This means that base stations can be placed further apart, potentially leading to a more cost-efficient deployment. The antenna gain compared to a single element antenna can be increased by an amount equal to the number of array elements, e.g., an eight-element array can provide a gain of eight (9 dB).

New Services -- When using smart antennas the network will have access to spatial information about the users. This information can be used to estimate the positions of the users much more accurately than in existing networks. Positioning can be used in services such as emergency calls and location-specific billing. The FCC (Federal Communications Commission) in the United States has decided that by October 2001 user location information with an accuracy of 125 meters RMS error must be provided [8].

Security -- It is more difficult to tap a connection when smart antennas are used. To successfully tap a connection the intruder must be positioned in the same direction as the user as seen from the base station.

Reduced Multipath Propagation -- By using a narrow antenna beam at the base station the multipath propagation can be somewhat reduced. The actual reduction depends on the scenario, and is not always significant. Although channel equalizers and RAKE receivers most often will handle and even exploit the multipath components, on very high-speed connections this may not be the case. Potentially, the reduction of multipath propagation can be used to ease the requirement on future modem design.

Cost Factors

Although the benefits of using smart antennas are many, there are also drawbacks and cost factors. The gain should always be evaluated against the cost.

Transceiver Complexity -- It is obvious that a smart antenna transceiver is much more complex than a traditional base station transceiver. The antenna will need separate transceiver chains for each of the array antenna elements and accurate real-time calibration of each of them. In addition, the antenna beamforming is a computationally intensive process, especially if *adaptive arrays* are to be used (see earlier section). This means that the smart antenna base station must include very powerful numeric processors and control systems. There will be a growing need for the development of efficient algorithms for real-time optimizing and signal tracking. Smart antenna base stations will no doubt be much more expensive than conventional base stations.

Resource Management -- Smart antennas are mainly a radio technology, but they will also put new demands on network functions such as resource and mobility management. When a new connection is to be set up or the existing connection is to be handed over to a new base station, no angular information is available to the new base station and some means to "find" the mobile station is necessary. This can be handled by letting the

base station continuously sweep through the cell with a "search" beam looking for candidates for a new connection or a handover. Another possibility is to use an external system for positioning, e.g., GPS. As far as handover is concerned, a third possibility is available: directional information from the existing cell can be used by the network to provide an "educated guess" about which cell to hand the connection over to. As was explained earlier, SDMA involves different users using the same physical communication channel in the same cell, separated only by angle. When angular collisions between these users occur, one of them must quickly switch to another channel so that the connection is not broken. This means that in systems providing full SDMA, there will be much more *intracell handovers* than in conventional TDMA or CDMA systems, and more monitoring by the network is necessary. The animation in [Fig. 1](#) shows what is meant by an angular collision.

Physical Size -- For the smart antenna to obtain a reasonable gain, an array antenna with several elements is necessary. Typically arrays consisting of six to 10 horizontally separated elements have been suggested for outdoor mobile environments [911]. The necessary element spacing is 0.40.5 wavelengths. This means that an eight-element antenna would be approximately 1.2 meters wide at 900 MHz and 60 cm at 2 GHz. With a growing public demand for less visible base stations, this size, although not excessive, could provide a problem. [Figure 5](#) shows a picture of an eight-element antenna array at 1.8 GHz.

Radio Planning

For the smart antennas to have the desired gain the base stations must be able to separate the users in the angular domain. This means that some of the current strategies for radio planning need to be modified. In existing networks base stations are often positioned along highways and railway lines. From a smart antenna point of view it is much more efficient to position the base station away from the road or railway. In this way the spatial dimension is better exploited. This strategy is illustrated in [Fig. 6](#).

Another difficult point is small urban cells with base stations positioned below rooftop level. In these cases the signal from both the desired users and from the interferers will tend to arrive along the street canyons, making it difficult to separate them in the angular domain. [Figure 7](#) shows an example of an average *power azimuth spectrum* (PAS)² measured in urban area microcells in Oslo [12] and shows that the signal tends to arrive from two perpendicular directions given by the street grid.

Radio channel models and radio planning tools need renewed attention upon the introduction of smart antennas. Statistical radio channel models are used in performance testing and design of radio systems. However, existing models do not describe the channel in the angular domain and can therefore not be used in the evaluation of systems employing smart antennas. Work is currently ongoing to provide statistical radio channel models describing the directional domain.

Similarly, existing commercial radio planning tools are normally only concerned with estimating the signal level and possibly the time dispersion on the link. However, in the future it will be beneficial for the tools to predict the angular dispersion as well. Some analytical radio prediction methods, such as *ray tracing*, have the possibility for estimation of angular information inherent.

Implementation

This section explains some of the underlying technology employed in order to implement smart antennas. The treatment given here is not comprehensive, but should give an impression of the basic ideas. The technology is based on array antennas where the radiation pattern is altered by adjusting the amplitude and relative phase on the different array elements. Next we explain the fundamental principles of array beamforming. Then we explain how this is exploited in smart antennas. Even though the reception and transmission parts most often are integrated, and much of the same hardware normally will be used, they will be explained separately. This is because conceptually the uplink and downlink are quite different in terms of smart antennas.

Array Antennas

Electronically steerable antenna patterns are most often generated using array antennas. These are antennas consisting of a number of antenna elements on which the signal is divided or combined in both phase and amplitude. Generally, any combination of elements can form an array. However, usually equal elements in a regular geometry are used.

Arrays can be one-, two- and three-dimensional, depending on the dimension of the space one wants to access. [Figure 13](#) shows some examples of array geometries. Using regular geometries simplifies both analysis and synthesis of the array. In this section, the linear array principle will be worked through as an example. The mathematics can be extended to more dimensions and other geometries, but the principle is the same.

The total radiation pattern is given by the *element type*, the *relative positions* and the excitation (amplitude and phase). Using an array antenna, it is possible to obtain a very good control of the radiation pattern, e.g., the shape of the main lobe and the side lobe level (SLL). The radiation pattern is uniquely given by the aforementioned parameters.

Linear Equidistant Arrays -- The "simplest" geometry is the *one-dimensional linear equidistant array*. All the array elements are placed along a line with equal distance between them, as shown in [Fig. 8](#). The figure also shows which parameters are available.

The far-field expression of the electrical field from a single array element is given by Eq. 1:

$$\vec{E}_n(r, \theta, \varphi) = I_n \cdot \frac{e^{-jk r}}{r} \cdot \vec{e}(\theta, \varphi) \quad (1)$$

where I_n is the complex element current and $\vec{e}(\theta, \varphi)$ is the element antenna pattern, usually called the

element factor, and r , θ and  are the spherical coordinates of the observation point.

The total field from an array consisting of N elements is the superposition of the element contributions as expressed in Eq. 2:

$$\vec{E}(r, \theta, \varphi) = \sum_{n=0}^{N-1} I_n \cdot \frac{e^{-jk r}}{r} \cdot e^{jk r'_n \cos \Psi_n} \cdot \vec{e}(\theta, \varphi) \quad (2)$$

where



is the difference in distance for element n compared to element 0 toward the observation point in the direction θ due to the spatial separation in the array. The element spacing is d . (r'_n is the distance between the coordinate systems origin and element n , Ψ_n is the corresponding angle, and z'_n , the distance along the array axis.) In the case of an array with equal elements, the far-field expression becomes

$$\vec{E}(r, \theta, \varphi) = \frac{e^{-jk r}}{r} \cdot \vec{e}(\theta, \varphi) \sum_{n=0}^{N-1} I_n \cdot e^{jk n d \cos \theta} \quad (4)$$

The sum expression in Eq. 4 is the array factor which is independent of both the element type and the distance from the antenna. It is only dependent on the *array geometry* and the *element excitation*. Thus, the total far field is given by the product of the element factor and the array factor. The array factor can be understood as the radiation pattern from an array antenna consisting of isotropic elements.

As the radiation pattern is uniquely given by the excitation parameters, a careful choice can give us the required effect. Consider, for example, that it is desired to steer the main lobe of the antenna in a certain

direction, φ . Thus, we want the far-field signals from all the elements of the array to be added *in phase* in the wanted direction. [Figure 9](#) shows what this means. If element number 0 is given the reference phase angle 0, then the phase angle of element number 1 must be shifted (in this case negatively) with a value of $k d \cos \varphi$, where $k = 2\pi/\lambda$ is the wave constant. We can call this angle φ_1 . Thus element number 2 must be shifted $2\varphi_1$, element 3 must be shifted $3\varphi_1$, and so on. In order to obtain maximum gain in the direction of φ , the amplitudes of all the elements must be equal. This example is called *uniform excitation* ($|I_0| = |I_1| = \dots = |I_n| = \dots = |I_{N-1}|$). [Figure 10](#) shows the resulting array factor in this case.

A drawback of the linear array is the front-back ambiguity. A mirror of the antenna pattern will appear around the antenna axis, i.e., in the direction φ ($\cos \varphi = \cos(\varphi)$). Most often we want only to have one main lobe, and this is usually obtained by using directive elements. It is also worth noting that for skew angles, the gain decreases and the main lobe width increases, because the element separation appears smaller in the maximum direction of the main lobe (apparent distance: $d \sin \varphi$). In practice, the linear array is only useful in a sector of at most 120° .

Using uniform excitation gives the highest gain but also the highest sidelobe level (SLL) of 3.2 dB. From the expression for the array factor one sees that the far-field is given as the Fourier-transform of the current distribution along the array. Thus, main lobe width and width of the current distribution are inverse. Using a non-uniform excitation with an amplitude roll-off toward the end of the array results in both a wider main lobe and a lower SLL. [Figure 11](#) shows some examples.

Practical Considerations -- Above we have anticipated that the element patterns are all identical and "perfect." In implementing an array antenna, this is not the case. The different elements of the antenna will have slightly different patterns due to production tolerances and so on. Additionally, when mounting several elements close together (typically less than $\lambda/2$) they influence each other. This mutual coupling gives changes both to the element patterns and the impedance match. This means that the array performance will be somewhat degraded.

Receiver

[Figure 12](#) shows schematically the elements of the reception part of a smart antenna. The antenna array contains N elements. The N signals are being combined into one signal, which is the input to the rest of the receiver (channel decoding, etc.).

As the figure shows, the smart antenna reception part consists of four units. In addition to the antenna itself it contains a radio unit, a lobe forming unit, and a signal processing unit [13].

The array will often have a relatively low number of elements in order to avoid unnecessarily high complexity in the signal processing. [Figure 13](#) shows four examples of different array geometries. The first two structures are used for beamforming in the horizontal plane (azimuth) only. This will normally be sufficient for outdoor environments, at least in large cells. The first example (a) shows a one-dimensional linear array with uniform element spacing of $\lambda/2$. This structure can perform beamforming in azimuth angle within an angular sector. This is the most common structure due to its low complexity. The second example (b) shows a birds eye view of a circular array with angular element spacing of $2\pi/N$. This structure can perform beamforming in all azimuth angles. The last two structures are used for performing two-dimensional beamforming, in both azimuth and elevation angles. This may be desirable for indoor or dense urban environments. [Figure 13c](#) is the front view of a two-dimensional linear array with horizontal element spacing of $\lambda/2$ and vertical element spacing of $\lambda/2$. This structure can perform beamforming within a solid angle. Beamforming in the entire space, within all solid angles, requires some sort of cubic or spherical structure. [Figure 13d](#) shows an example of a cubic structure with element separations of $\lambda/2$ in x , y and z .

The radio unit consists of down-conversion chains and (complex) analog-to-digital conversion (A/D). There must be N down-conversion chains, one for each of the array elements.

The signal processing unit will, based on the received signal, calculate the complex weights $w_1 w_N$ with which the received signal from each of the array elements is multiplied. These weights will decide the antenna pattern in the uplink direction as described in the previous section. The weights can be optimized from two main types of criteria: maximization of received signal from the desired user (e.g., switched lobe or phased array) or maximization of the SIR by suppressing the signal from interference sources (adaptive array). In theory, with M antenna elements one can "null out" $M-1$ interference sources, but due to multipath propagation this number will normally be lower.

The method for calculating the weights will differ depending on the type of optimization criterion. When *switched lobe* (SL) is used the receiver will test all the pre-defined weight vectors (corresponding to the lobe set) and choose the one giving the strongest received signal level. If the *phased array* approach (PA) is used, which consists of directing a maximum gain beam toward the strongest signal component, the *direction-of-arrival* (DoA) is first estimated and then the weights are calculated as described in the previous section with uniform amplitude and phase in accordance with the desired steering angle. A number of well documented methods exist for estimating the DoA, for instance, MUSIC, ESPRIT, or SAGE [4, 14].

If maximization of SIR is to be done (AA), the optimum weight vector (of dimension N) $\mathbf{W}_{\text{opt}}$ is given by the

dominant generalized eigenvector of the matrix pair $[\mathbf{R}_k, \mathbf{Q}_k]$ where $\mathbf{R}_k$ is the $M \times N$ correlation matrix of the total received signal and $\mathbf{Q}_k$ is the $M \times N$ correlation matrix of the total interference. This is called *optimum combining* [15]. Figure 14 shows an example of the resulting antenna patterns for PA, SL, and AA in a case where both the signal and interference are measured in a real channel. Note how the AA directs nulls toward the strongest interference sources. Other techniques are also possible, for example, well documented diversity algorithms such maximum-ratio in multiple dimensions [16].

For the base station to be able to estimate the radio channel, a reference or training sequence is normally necessary, i.e., a known bit sequence must be transmitted periodically. The methods that maximize the SIR do in principle require knowledge of the instantaneous channel response from both the desired user and all the interference sources, which means the training sequence must be unique for each user.

In the lobe forming unit the actual weighting of the received signal from each of the array elements is done. In the most advanced case this unit will be an integration of the channel equalizer/RAKE receiver and the

smart antenna. In this case $M \times D$ weights will be required, where D is the number of symbol periods (depth) in the channel equalizer or number of fingers in the RAKE receiver. This is called a *spatio-temporal filter* [17, 18], because it removes the undesired signal components and keeps the desired ones both in time and space domains. Such a lobe forming unit is shown schematically in Fig. 15 [19]. In this example the number of

weights is $M \times 3$, showing that the depth of the equalizer is 3. The time between the taps on each antenna element is T . In a RAKE receiver, the time delay between the taps is not uniform.

When the lobe forming is done digitally (after A/D) the lobe forming and signal processing units will normally be integrated in the same unit, namely the DSP. The separation in Fig. 12 is done to clarify the functionality. It is also possible to perform the lobe forming in hardware at *radio frequency* (RF) or *intermediate frequency* (IF) [20].

The three aforementioned methods (SL, PA, and AA) differ in complexity and computational burden. However, a detailed analysis of this is out of the scope of this article because many parameters and assumptions must be defined in order to quantify this difference. Generally, SL is less complex than PA, which in turn is less complex than AA. Conditions that influence this are, for example, the required accuracy and performance in terms of power or C/I-gain. It is also expected that the radio access method makes a difference, i.e., TDMA vs. CDMA.

Transmitter

The transmission part of the smart antenna will be schematically very similar to the reception part. An illustration is shown in [Fig. 16](#). The signal is split into N branches, which are weighted by the complex weights $z_1 z_N$ in the lobe forming unit. The weights, which decide the radiation pattern in the downlink direction, are calculated by the signal processing unit. The radio unit consists of D/A converters and the up converter chains. In practice, some components, such as the antennas themselves and the DSP, will of course be the same as on reception.

The principal difference between uplink and downlink is that since there are no smart antennas applied to the user terminals (mobile stations) no knowledge of the spatial channel response is available on downlink. In a time division duplex (TDD) system the mobile station and base station use the same carrier frequency only separated in time. In this case the weights calculated on uplink will be optimal on downlink if the channel does not change during the period from uplink to downlink transmission. However, this can not be assumed to be the case in general, at least not in systems where the users are expected to move at high speed. If frequency division duplex (FDD) is used, the uplink and downlink are separated in frequency. In this case the optimal weights will generally not be the same because of the channel response dependency on frequency. Thus optimum beamforming (i.e., AA) on downlink is difficult, and the technique most frequently suggested is the geometrical approach of estimating the direction-of-arrival (DoA). The assumption is directional reciprocity, i.e., the direction from which the signal arrived on the uplink is the direction in which the signal should be transmitted to reach the user on downlink. This assumption has been strengthened by recent experimental results [21]. The strategy used by the base station is to estimate the DoA of the direction (or directions) from which the main part of the user signal is received. This direction is used on downlink by choosing the weights $z_1 z_N$ so that the radiation pattern is a lobe (or lobes) directed toward the desired user. This is similar to PA. In addition, it is possible to position zeros in the direction toward other users so that the interference suffered by these users is minimized. Due to fading on the different signal paths, it has been suggested to choose the downlink direction based on averaging the uplink channel over a period of time [11]. This will, however, be sub-optimum compared to the uplink situation where knowledge about the instantaneous radio channel is available.

It should be stressed that in the discussion above it is assumed that the interferers observed by the base stations are mobile stations and that the interferers observed by the mobile stations are base stations. This means that when the base station on transmission positions zeros in the direction toward other mobile stations than the desired one, it will reduce the interference suffered by these mobiles. If, however, the interferers observed by mobiles are other mobiles, as may be the case, for instance, in UTRA TDD [22], there will be a much more fundamental limitation in the possibility for interference reduction at the mobile.

Critical Factors and Challenges

A number of technical challenges remain to be solved for smart antennas to reach their full potential. These are aspects on both modem implementation as well as the algorithms for beamforming.

As mentioned in the previous section, beamforming on the downlink is made more difficult by the lack of knowledge about the instantaneous channel response. The strategy of directing the lobe (or lobes) in the direction(s) of the user signal DoA is often used, but this makes the downlink performance very dependent on the radio channel and its rate of change. If the channel suffers from much angular spread, i.e., the signal arrives from many directions, directing a single lobe in the direction of maximum reception is far from optimum in terms of maximizing the power transfer to the user. Directing zeros toward other users to minimize the interference suffered by these is difficult for the same reason. When the channel is changing rapidly, which will be the case for fast moving mobiles and for very dense, complicated environments, the difficulties on downlink will be worsened. [Figure 17](#) shows a recording of an angular radio channel from a suburban microcellular environment. The distance between each channel measurement was 2 cm corresponding to 10 ms between measurements at a speed of 2 m/s (walking speed).

Another critical factor is the linearity in reception and transmission chains. The transfer functions of all the

up- and down-converter chains need to be perfectly known for the beamforming to be accurate. However, since there are active components involved the responses will change with time and temperature. This means that for the smart antenna to function properly frequent on-line calibration of the radio units needs to be performed [20].

The requirements on linearity in transceiver chains can possibly be relaxed by performing the A/D (and D/A) conversion as close to the array element as possible, that is sampling on RF or IF. This requires very fast and specialized hardware.

The complexity of the smart antenna is a great challenge because the most advanced smart antenna implementations involve simultaneously maximizing the useful signal and nulling out the interference sources. In addition, this beamforming needs to be done for each user communicating via the base station in question. Even with the powerful signal processing units available today it is a very challenging task to perform this in real time. Much work will need to be done on optimizing code and developing efficient algorithms for this to work properly. However, developments in computer technology over the past few years have undoubtedly made it possible to perform even more complex processing in real time. In the last three years processor speed has increased more than five times, and if this trend continues, there will be processors with 2-3 GHz clock frequency available when UMTS is to be introduced.

Status: Trials and Testbeds

Smart antennas for mobile communications are not yet available commercially. But it is a well known fact that many manufacturers are working on solutions for both second- generation systems and third-generation systems. This section will give a short overview of some of the documented trials and testbeds, but should not be considered as a complete list.

Ericsson/Mannesmann Trial in Giessen, Germany

Ericsson has performed a trial in cooperation with the German GSM1800 operator Mannesmann in the city of Giessen. Detailed results from the trial have not yet been published, but preliminary results are promising [9]. Three base stations were equipped with smart antennas. The base stations were placed at high positions, giving a large coverage. In the autumn of 1998 the Ericsson/Mannesmann trial became the first trial to demonstrate commercial traffic through base stations equipped with smart antennas.

Physically, the smart antennas were six-element stacked dipoles. Each element was dual polarized, with ± 455 polarized elements. The combining on uplink was based on maximal ratio combining, while the downlink beam forming was based on switching between eight lobes.

Tsunami II Trial in Bristol, UK

Tsunami II is a project in the European 4th framework program ACTS³ dealing with the development of a smart antenna testbed and the implementation of a smart antenna base station. Some of the participants in this project have been the University of Aalborg (DK), Bosch Telekom (D), Orange (UK), and the University of Bristol (UK).

The project has finished and performed trials near Bristol (UK) in both urban and rural environments. Even though not all data analyses have been published yet, the trials confirm the capacity increase of a smart antenna. In the rural environment (macrocell) an extension of the range of up to 54 percent was reported, using the HSR principle. Interference reduction was better than 30 dB, which is much higher than expected. However, only one interferer was used, which in addition was placed more than 10 degrees away from the transmitter system [10, 23].

The antenna was eight-element dual polarized. Different strategies for combining were tested. It was shown that the adaptive array-based combining techniques, which included some sort of interference rejection,

showed the best results [10].

Other

The U.S. company Arraycomm has developed smart antenna solutions for the GSM standard and for the Japanese Personal Handyphone System (PHS). In June 1997 a smart antenna wireless local loop (WLL) solution based on using PHS was commercially available [24]. Field trials have been performed in several countries, including the U.S., Brazil, and Malaysia. The Canadian company Nortel also claims to have been the first to implement a smart antenna for GSM [25].

Other manufacturers and research institutions are also developing testbeds and trials. Alcatel, in cooperation with the University of Vienna, is planning a testbed [11]. Much work has also been performed in the U.S., including at Telcordia [26] and Virginia Tech [27]. Other test results have been reported from Korea [28] and Japan [29].

Conclusions

The main goal of this article has been to provide an overview of smart antenna technologies with respect to third-generation land mobile systems, i.e., IMT2000 and UMTS. It is obvious that smart antennas at the base stations will be an important technology to provide the necessary capacity and coverage. It also helps to realize new services, e.g., based on user location.

From a technology point of view, smart antennas can be seen as an extension of the "conventional" resource allocation schemes used in radio communications. In addition to dividing the space into cells, it will now also be possible to employ space division inside each cell. Different degrees of utilization of the spatial dimension are possible, and different steps have been described here.

Smart antenna technology is a broad concept and implementations range from simple techniques that involve switching between lobes to advanced algorithms maximizing the received signal-to-interference ratio. Implementation of smart antennas is done using array antennas. The techniques for beamforming with array antennas are well known, and must be employed in both duplex directions for the improvements to be substantial. However, with rapid channel variations it is not a trivial task to provide optimum beamforming, especially for the downlink direction.

The use of smart antennas is not purely a radio transmission issue. It also influences network services such as handover and connection setup. Introducing the spatial domain in the resource management system makes this more complex.

Several smart antenna testbeds and field trials have been set up and run by manufacturers and research institutions. The first tests allowing commercial traffic over a base station employing smart antennas were performed by Ericsson and Mannesmann Mobilfunk in Germany in the autumn of 1998.

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Biographies

Per H. Lehne graduated from the Norwegian Institute of Technology (NTH) in 1988 from the Department of Telecommunications, Radio Technology. Since then, he has been employed at Telenor Research and Development as a research scientist working on several aspects of mobile and personal communications. From 1988 to 1991 he was involved in the standardization of the pan-European radio paging system ERMES within ETSI. Later, the emphasis was on radio propagation aspects for cellular mobile communications. He was involved in the RACE MBS project, on millimeter-wave propagation. He has been a member of COST 259 -- "Wireless Flexible Personalized Communications" since 1997. Currently, the main focus of his work is on smart antennas and access network technology for UMTS.

Magne Pettersen received his M.Sc. from the Norwegian University of Science and Technology (NTNU) in 1994. From 1994 to 1995 he was a research assistant at the Department of Telecommunications, NTNU. In 1995 he joined Telenor Research and Development, Kjeller, where he is currently employed as a research scientist. He is working in the field of personal communications, in particular in the area of antennas and radio wave propagation. He has a special interest in modeling and characterization of the mobile radio channel, an area in which he is currently pursuing a Ph.D. at NTNU.

Hsieh-Yee Declaration

Appendix 0004B

Hsieh-Yee Declaration

Appendix A

APPENDIX A

Ingrid Hsieh-Yee

Professor Emerita
Dept. of Library and Information Science
Catholic University of America
Washington, D.C. 20064
E-mail: hsiehyee@cua.edu

Education

Ph.D. 1990. Library and Information Studies, University of Wisconsin-Madison
Minors: Sociology and Psychology

M.A. 1984. Library and Information Studies, University of Wisconsin-Madison.

M.A. 1983. Comparative Literature, University of Wisconsin-Madison.

B.A. 1980. Foreign Languages and Literature, National Taiwan University.

Work Experience

Professor, School/Dept. of Library and Information Science, Catholic University of America,
1990-2022 (Assistant Professor, 1990-1996; Associate Professor, 1997-2004; Professor
2004-2022; Professor Emerita 2022)

Co-Chair, Dept. of Library and Information Science, Catholic University of America, June 2015-
August 2016.

Acting Dean, School of Library and Information Science, Catholic University of America,
January 2010-June 2012.

Cataloger, Dept. of Legislative Reference Library, Annapolis, Maryland, 1989-1990.

Lecturer, School of Library and Information Studies, University of Wisconsin-Madison, 1988.

Teaching Assistant, School of Library and Information Studies, University of Wisconsin-
Madison, 1986-1988.

Cataloger, Health Sciences Library, University of Wisconsin-Madison, 1984-1986.

Areas of Teaching and Research Interests

Information Organization and Access; Metadata; Cataloging & Classification; Information
Architecture; Information Retrieval; Digital Collections; Social Media; Scholarly

Communication; Information Behavior; Health Informatics; Human Computer Interaction; Usability Studies

Grants & Honors

Cultural Heritage Information Management Project. IMLS grant. Amount: \$498,741. Period: Aug. 2012 to July 2015. Co-PI with Dr. Youngok Choi.

D.C. Health Information Technology (HIT4): Building Capacity & Providing Access in Our Nation's Capital. Dept. of Labor H2B Training Grant. Grant amount: \$4,175,500. Grant period: Nov. 2011 to Dec. 2015. Partner with the Metropolitan School of Professional Studies of the Catholic University of America, Children's National Medical Center, D.C. Department of Employment Services, Holy Cross Hospital, Howard University, Center for Urban Progress, Providence Hospital, and Sibley Memorial Hospital.

Capital Health Careers Project. Department of Labor Healthcare Sector and Other High Growth and Emerging Industries Grant. Grant amount: \$4,953,999. Grant period: March 2010 – February 2013. Awarded to a group of healthcare organizations and educational institutions in Washington, D.C. Providence Health Foundation of Providence Hospital (Lead institution). Part of the grant supported the development of a Master's degree program in Information Technology with a concentration in Health Information Technology offered by the School of Library and Information Science.

The Washington D.C. School Librarians Project. IMLS grant. Grant amount: \$412,660. Grant period: Aug. 2007 – June 2011. The School partnered with the District of Columbia Public Schools (DCPS) and the District of Columbia Library Association to educate and mentor school media specialists for the DCPS system. PI, Jan. 2010 to June 2011.

SIG Member of the Year, American Society for Information Science and Technology (2009).

Most Outstanding Paper of *OCLC Systems & Services* (2001).

ALISE Research Grant (2001).

Most Outstanding Paper of *OCLC Systems & Services* (2000).

Research Grant from ERIC (1999-2000).

Best Research Paper Award; Association for Library and Information Science Education (1998).

Research Grants, Catholic University of America. 1991, 1992, 1993, 1996, 1998, 1999, 2004, 2005, 2006, 2007, 2013-14.

Cooperative Faculty Research Grant, Consortium of Universities in the Washington Metropolitan Area (1993-1994).

Cooperative Research Grant, Council on Library Resources (1993-1994).

Journal of the American Society for Information Science Best Paper Award (1993).

ASIS/ISI Information Science Doctoral Dissertation Scholarship (1989).

HEA Title IIB Fellowship (Dept. of Education) (1989)

Chinese-American Librarians Association Scholarship (1987).

Beta Phi Mu (1985).

Vilas Fellowship, University of Wisconsin-Madison. 1984

Publications

Hsieh-Yee, I. (2021). Can We Trust Social Media? *Internet Reference Services Quarterly*, 1-15.
Published online: 28 July, 2021. 10.1080/10875301.2021.1947433.

Bailey, T., & Hsieh-Yee, I. (2020). Combating the Sharing of False Information: History, Framework & Literacy Strategies, *Internet Reference Services Quarterly*, 24 (1-2), 9-30.

Choi, Y., and Hsieh-Yee, I. (2010). Finding Images in an OPAC: Analysis of User Queries, Subject Headings, and Description Notes. *Canadian Journal of Information and Library Science*, 34(3): 271 – 295.

Hsieh-Yee, I. (2008). Educating Cataloging Professionals in a Changing Information Environment. *Journal of Education for Library and Information Science*, 46(2): 93-106.

Vellucci, S. L., Hsieh-Yee, I., and Moen, W.E. (2007). The Metadata Education and Research Information Commons (MERIC): A Collaborative Teaching and Research Initiative. *Education for Information*, 25(3&4): 169-178.

NISO Framework for Guidelines for Building Good Digital Collections. 3rd ed. Baltimore, MD: National Information Standards Organization, 2007. Also available online: <http://www.niso.org/framework/framework3.pdf> (NISO Working Group members: Priscilla Caplan (chair), Grace Agnew, Murtha Baca, Tony Gill, Carl Fleischhauer, Ingrid Hsieh-Yee, Jill Koelling, and Christie Stephenson.)

Choi, Y., Hsieh-Yee, I., and Kules, B. (2007). Retrieval Effectiveness of TOC and LCSH. *Proceedings of the Joint Conference on Digital Libraries*, pp. 233-234.

Vellucci, S., and Hsieh-Yee, I. (2007). They Didn't Teach Me That in Library School! Building

a Digital Teaching Commons to Enhance Metadata Teaching, Learning and Research. *Proceedings of the National Conference of the Association of College and Research Libraries, Baltimore, MD*, pp. 26-31.

Mitchell, Vanessa, and Ingrid Hsieh-Yee. (2007). Converting Ulrich's Subject Headings to FAST Headings: A Feasibility Study. *Cataloging & Classification Quarterly*, 45(1): 59-85.

Hsieh-Yee, I., Tang, R., and Zhang, S. (2007). User Perceptions of a Federated Search System. *IEEE Technical Committee on Digital Libraries Bulletin*, Summer 3(2) (URL = <http://www.ieee-tcdl.org>).

Tang, R., Hsieh-Yee, I., and Zhang, S. (2007). User Perceptions of MetaLib Combined Search: An Investigation of How Users Make Sense of Federated Searching." *Internet Reference Services Quarterly*, 12(12): 211-236.

Hsieh-Yee, I., Tang, R., and Zhang, S. (2006). User Perceptions of a Federated Search System. *Proceedings of the Joint Conference on Digital Libraries, June 11-15, 2006, Chapel Hill*, p. 338.

Hsieh-Yee, I. (2006). *Organizing Audiovisual and Electronic Resources for Access: A Cataloging Guide*. 2nd ed. Westport, Conn.: Libraries Unlimited.

NISO A Framework of Guidance for Building Good Digital Collections. 2nd ed. Bethesda, MD: National Information Standards Organization, 2004. Framework Advisory Group: Grace Agnew, Liz Bishoff, Priscilla Caplan (Chair), Rebecca Gunther and Ingrid Hsieh-Yee.

Hsieh-Yee, I. (2004). Cataloging and Metadata Education in North American LIS Programs. *Library Resources & Technical Services*, 48(1): 59-68.

Hsieh-Yee, I. (2004). Cataloging and Metadata Education. In Gary E. Gorman (Ed.), *International Yearbook of Library and Information Management 2003: Metadata Applications and Management*, (pp.204-234). London: Facet Publishing.

Yee, P. L., Hsieh-Yee, I., Pierce, G.R., Grome, R., and Schantz, L. (2004). Self-Evaluative Intrusive Thoughts Impede Successful Searching on the Internet. *Computers in Human Behavior*, 20(1): 85-101.

Hsieh-Yee, I. (2003). Cataloging and Metadata Education: A Proposal for Preparing Cataloging Professionals of the 21st Century. A report submitted to the ALCTS-Education Task Force in response to Action Item 5.1 of the *Bibliographic control of Web Resources: A Library of Congress Action Plan*. Approved by the Association for Library Collections and Technical Services. Web version available since April 2003 at <http://lcweb.loc.gov/catdir/bibcontrol/CatalogingandMetadataEducation.pdf>.

- Hsieh-Yee, I. (2002). Cataloging and Metadata Education: Asserting a Central Role in Information Organization. *Cataloging & Classification Quarterly* 34(½): 203-222.
- Hsieh-Yee, I., and Smith, M. (2001). The CORC Experience: Survey of Founding Libraries, Part I. *OCLC Systems & Services*, 17: 133-140. (Received "The Most Outstanding Paper of OCLC Systems & Services in 2001" award.)
- Hsieh-Yee, I., and Smith, M. (2001). The CORC Experience: Survey of Founding Libraries, Part II, Automated Tools and Usage. *OCLC Systems & Services*, 17: 166-177. (Received "The Most Outstanding Paper of OCLC Systems & Services in 2001" award.)
- Hsieh-Yee, I. (2001). ERIC User Services: Changes and Evaluation for the Future. *Government Information Quarterly*, 18: 31-42.
- Hsieh-Yee, Ingrid. (2001). Research on Web Search Behavior. *Library and Information Science Research*, 23: 167-185.
- Logan, E., and Hsieh-Yee, I. (2001). Library and Information Science Education in the Nineties. *Annual Review of Information Science and Technology*, 35: 425-477.
- Hsieh-Yee, I. (Ed.) (2001). *Library and Information Science Research*, 23 (2). A special issue in honor of the retirement of Douglas L. Zweizig.
- Hsieh-Yee, I. (2000). *ERIC User Services: Evaluation in a Decentralized Environment*. Washington, D.C.: Dept. of Education.
- Hsieh-Yee, Ingrid. (2000). *Organizing Audiovisual and Electronic Resources for Access: A Cataloging Guide*. Littleton, CO: Libraries Unlimited.
- Hsieh-Yee, I. (2000). Organizing Internet Resources: Teaching Cataloging Standards and Beyond. *OCLC Systems & Services*, 16: 130-143. (Received "The Most Outstanding Paper of OCLC Systems & Services in 2000" award.)
- Hsieh-Yee, I. (1998). The Retrieval Power of Selected Search Engines: How Well Do They Address General Reference Questions and Subject Questions? *Reference Librarian*, 60: 27-47.
- Hsieh-Yee, I. (1998). Search Tactics of Web Users in Searching for Texts, Graphics, Known Items and Subjects: A Search Simulation Study. *Reference Librarian*, 60: 61-85. (Received the 1997 Best ALISE Research Paper Award.)
- Hsieh-Yee, I. (1997). Access to OCLC and Internet Resources: LIS Educators' Views and Teaching Practices. *RQ*, 36: 569-86.

- Hsieh-Yee, I. (1997). Teaching Online and CD-ROM Resources: LIS Educators' Views and Practices. *Journal of Education for Library and Information Science*, 38: 14-34.
- Hsieh-Yee, I. (1996). The Cataloging Practices of Special Libraries and Their Relationship with OCLC. *Special Libraries*, 87: 10-20.
- Hsieh-Yee, I. (1996). Modifying Cataloging Practice and OCLC Infrastructure for Effective Organization of Internet Resources. In *Proceedings of the OCLC Internet Cataloging Colloquium*. [Online]. Available: <http://www.oclc.org/oclc/man/colloq/hsieh.htm>
- Hsieh-Yee, I. (1996). Student Use of Online Catalogs and Other Information Channels. *College & Research Libraries*, 57: 161-175.
- Hsieh-Yee, I. (1995). Ten entries in James S. C. Hu (Ed.), *Encyclopedia of Library & Information Science*, 913, 1028-29, 1036, 1037, 1145-46, 1514, 1575, 1763-64, 2216-27, 2378-79. Taipei, Taiwan: Sino-American Publishing. (Topics include "Advanced Technology/Libraries," "Information Ethics," "Instruction on Cataloging and Classification," "Instruction on Reference Services.")
- Hsieh-Yee, I. (1993). Effects of Search Experience and Subject Knowledge on Online Search Behavior: Measuring the Search Tactics of Novice and Experienced Searchers. *Journal of the American Society for Information Science*, 44: 161-174. (Received the 1993 Best JASIS Paper Award.)

Works in Progress

- Hsieh-Yee, I. Actionable lessons for debunking false information: A comparative analysis of literacy lessons in communication and library and information science
- Hsieh-Yee, I. Representing diverse voices and communities in cataloging and indexing: The evolution of controlled vocabularies
- Hsieh-Yee, I. Information Organization and Retrieval in Practice (a monograph scheduled for completion and publication in 2023)

Presentations

- Bailey, T. & Hsieh-Yee, I. (February 2020). The Phenomena of Sharing Misinformation and the Need for Information Literacy. Presented at the 2020 Bridging the Spectrum Symposium, Washington, D.C., February 14, 2020.

- Hsieh-Yee, I. and Fragan-Fly, J. (May 2018) Trends, Design & Strategies for Digital Scholarship Services. Presented at the 2018 Maryland/Delaware Library Association Conference, Cambridge, MD.
- Hsieh-Yee, I. (February, 2018) Research Data Management: What It Takes to Succeed. Presented at the 10th Bridging the Spectrum Symposium, Washington, D.C.
- Hsieh-Yee, I. (February, 2017) *Research Data Management: New Competencies and Opportunities for Information Professionals*. Presented at the 9th Bridging the Spectrum Symposium, Washington, D.C.
- Hsieh-Yee, I. and Lawton, P. (February, 2017) *Enhancing Catholic Portal Searches with User Terms and LCSH*. Presented at the 9th Bridging the Spectrum Symposium, Washington, D.C.
- Hsieh-Yee, I. (2016, October) *Visualizing Data for Information*. Presented at the 2016 Virginia Library Association Conference, Hot Springs, VA.
- Hsieh-Yee, I. (2016, August) *Religious Materials Toolbox for Archivists: Solutions to Problems Facing the Profession*. Presented at Archives * Records 2016, Atlanta, GA.
- Hsieh-Yee, I. and Lawton, P. (2016, March) *Enhancing Retrieval of Catholic Materials with LCSH Knowledge Structure*. Presented at the 2016 Catholic Library Association Conference, San Diego, CA.
- Fagan-Fry, J. and Hsieh-Yee, I. (2016, February) *Approaches to Digital Scholarship at Top Universities around the World: Scholarly Publishing in the Digital Age*. Presented at the 8th Bridging the Spectrum Symposium, Washington, D.C.
- Hsieh-Yee and Fagan-Fry, J. (2016, January) *Innovative Services for Digital Scholarship at Top 100 Research Libraries of the World*. Poster presented at the 2016 Annual Conference of the Association for Library and Information Science Education, Boston, Mass.
- Hsieh-Yee, I. and Lawton, P. (2015, June). *Crowdsourcing terms for CRRA portal themes*. Poster presented at the third CRRA symposium and annual meeting, Bringing the created toward the Creator: Liturgical art and design since Vatican II. Catholic Theological Union, Chicago, Illinois.
- Hsieh-Yee, I. and Lawton, P. (2015, February). *Crowdsourcing terms for thematic exploration in the Catholic Portal*. Poster presented at the 7th Annual Bridging the Spectrum Symposium, Washington, D.C.
- Hsieh-Yee, I., James, R., and Fagan-Fry, J. (2015, February). *Support for digital scholarship at top university libraries of the world*. Poster presented at the 7th Annual Bridging the Spectrum Symposium, Washington, D.C.

- Hsieh-Yee, I., Zhang, S., Lin, K., and Cherry, S. (2015, February). *Thus said the end users: Summon experience and support for research workflows*. Poster presented at the 7th Annual Bridging the Spectrum Symposium, Washington, D.C.
- Yontz, E., Hsieh-Yee, I., & Houston, S. (2015, February). *Healthy Heroes Summer Reading Club: Developing healthy youth at public libraries*. 11th Annual Jean Mills Health Symposium, Greenville, North Carolina.
- Yontz, E., Hsieh-Yee, I., and Houston, S. (2015, January). *Healthy youth and libraries: A pilot study*. Association for Library & Information Science Education (ALISE) Annual Conference, Chicago, Illinois.
- Hsieh-Yee, I. (2014, May). *Linking CRRA resources to portal themes via authority files*. Presented at the Catholic Research Resources Alliance 2014 Membership Meeting, Marquette, WI.
- Hsieh-Yee, I. (2014, April). *Enhancing subject access to CRRA resources*. Presented at the 2014 Catholic Library Association Conference, Pittsburgh, PA.
- Hsieh-Yee, I. (2014, January). *Health Information Technology Program: Educational entrepreneurship in action*. Presented at the 2014 annual Conference of the Association for Library and Information Science Education, Philadelphia, PA.
- Hsieh-Yee, I., Zhang, S., Lin, K., and Cherry, S. (2014, January). *Discovering information through Summon: An analysis of user search strategies and search success*. Paper presented at the 6th Bridging the Spectrum Symposium, Washington, D.C.
- Hsieh-Yee, I. (2012, December). *National Digital Stewardship Alliance and SLIS at CUA: An Educational Partnership*. Paper presented at Best Practices Exchange: Acquiring, Preserving, and Providing Access to Government Information in the Digital Era, Annapolis, MD.
- Choi, Y. and Hsieh-Yee, I. (2010, January). *Finding Images in an OPAC: Analysis of User Queries, Subject Headings, and Description Notes*. Paper presented at 2nd Annual Bridging the Spectrum Symposium, Catholic University of America, Washington, D.C.
- Hsieh-Yee, I. and Coogan, J. (2010, January). *Google Scholar vs. Academic Search Premier: What Libraries and Searchers Need to Know*. Paper presented at 2nd Annual Bridging the Spectrum Symposium, Catholic University of America, Washington, D.C.
- Hsieh-Yee, I. (2009, November). *Information Science Education: An LIS School's Perspective*. Paper presented at Annual Meeting of the American Society for Information Science and Technology, Vancouver, British Columbia, Canada.

- Hsieh-Yee, I., Menard, E., Ya-Ning Chen, A., Shu-Jiun Chen, S., Kalfatovic, M. R., Wisser, K. M. (2009, November). *Information Organization in Libraries, Archives and Museums: Converging Practices and Collaboration Opportunities*. Presented at Annual Meeting of the American Society for Information Science and Technology, Vancouver, British Columbia, Canada. (Organizer and moderator of this panel.)
- Hsieh-Yee, I. and Coogan, J. (2009, July). *Catching up to Google Scholar: The Retrieval Power of Academic Search Premier and Google Scholar*. Poster presented at American Library Association Conference, Chicago, Illinois.
- Hsieh-Yee, I., with the CUA Scholarly Communications Project Team. (2009, January). *Digital Scholarship@CUA: Developing an Institutional Repository for CUA*. Poster presented at 1st Annual Bridging the Spectrum Symposium, Catholic University of America, Washington, D.C.
- Wise, M., Cylke, K., and Hsieh-Yee, I. (2009, January). *Digital Talking Books: Meeting the Needs of the Blind and the Handicapped*. Paper presented at the Bridging the Spectrum Symposium, Catholic University of America, Washington, D.C.
- Hsieh-Yee, I. (2009, January). *User Expectations of MERIC*. Presented at the Information Organization Competencies for the 21st Century Discussion Session of the 2009 Conference of the Association for Library and Information Science Education, Denver, Colorado.
- Choi, Y., and Hsieh-Yee, I. (2008, November). *Subject Access for Images in an OPAC*. Annual Meeting of the American Society for Information Science and Technology, Columbus, Ohio. (Also co-organized a panel on Retrieving and Using Visual Resources: Challenges and Opportunities for Research and Education.)
- Hsieh-Yee, I. (2008, June). *Educating Cataloging Professionals in a Changing Information Environment*. National Taiwan University, Taipei, Taiwan.
- Vellucci, S. L., Moen, W.E., Hsieh-Yee, I., Marson, B., and Wisser, K. (2008, January) *Building a Metadata Education and Research Community through MERIC (Metadata Education and Research Information Commons): Demo and Stakeholder Input*. A panel presented at the 2008 Conference of the Association for Library and Information Science Education, Philadelphia, Pennsylvania.
- Hsieh-Yee, I., Choi, Y. and Kules, B. (2007, October). *Searching for Books and Images in OPAC: Effects of LCSH, TOC and Subject Domains*. A poster presented at the American Society for Information Science and Technology Annual Meeting, Milwaukee, Wisconsin.
- Hsieh-Yee, I. and Coogan, J. (2007, August) *Google Scholar vs. Academic Search Premier: A Comparative Analysis*. Presented to the Faculty and Staff of the University of the District

of Columbia.

- Hsieh-Yee, I. and Coogan, J. (2007, June). *Google Scholar vs. Academic Search Premier: A Comparative Analysis*. Presented to the Washington Research Library Consortium Community, Catholic University of America, Washington, D.C.
- Hsieh-Yee, I., Choi, Y., and Kules, B.. (2007, June). *What Users Need for Subject Access: Table of Contents or Subject Headings?* A poster presented at the 2007 American Library Association Annual Conference, Washington, D.C., June 2007.
- Choi, Y., Hsieh-Yee, I., and Kules, B. (2007, June). *Retrieval Effectiveness of TOC and LCSH*. A paper presented at the Joint Conference on Digital Libraries 2007, Vancouver, Canada.
- Vellucci, S. L., Hsieh-Yee, I., and Moen, W.E. (2007, May). *If We Build It, Will They Come? Building a Community of Practice for Metadata Stakeholders*. A poster presented at the Rutgers University Research Day, Bridgeton, New Jersey.
- Hsieh-Yee, I. (2007, May). *Federated Searching: User Experience & Perceptions*. International Conference on Information Organization & Retrieval, National Taiwan University, Taipei, Taiwan.
- Hsieh-Yee, I. (2007, May). *Search Performance of Google Scholar and Academic Search Premier*. International Conference on Information Organization & Retrieval, National Taiwan University, Taipei, Taiwan.
- Hsieh-Yee, I. (2007, May) *MERIC: Building a Digital Commons for Metadata Education & Research*. International Conference on Information Organization & Retrieval, National Taiwan University, Taipei, Taiwan.
- Hsieh-Yee, I., and Coogan, J. (2007, March/April). *A Comparative Analysis of Google Scholar and Academic Search Premier*. Poster presented at the Association of College & Research Libraries 13th National Conference, Baltimore, Maryland.
- Vellucci, S. L. and Hsieh-Yee, I. (2007, March/April) *They Didn't Teach Me That in Library School! Building a Digital Teaching Commons to Enhance Metadata Teaching, Learning and Research*. On-site presentation and Webcast by Elluminate. A contributed paper presented at the Association of College & Research Libraries 13th National Conference, Baltimore, Maryland. The acceptance rate for contributed paper was 20%. This paper was one of 10 conference papers chosen for live webcast during the conference.
- Moen, W., Hsieh-Yee, I. and Vellucci, S.L. (2007, January) *A DSpace Foundation for a Teaching & Research Commons: The Metadata Education and Research Information Commons*. A poster session presented at the Open Repositories Conference 2007, San Antonio, Texas.

- Tang, R., Hsieh-Yee, I., and Zhang, S. (2006, November) *User Perception of MetaLib Combined Search*. Paper presented at the Annual Meeting of the American Society for Information Science and Technology, Austin, Texas, Nov. 2006.
- Hsieh-Yee, I. (2006, November). *Federated Searching: User Perceptions, System Design, and Library Instructions*. Paper presented at the Annual Meeting of the American Society for Information Science and Technology, Austin, Texas. (Panel organizer, moderator, presenter).
- Hsieh-Yee, I. (2006, November). *Building a Digital Teaching Commons to Enhance Teaching and Learning: The MERIC Experience and Challenges*. Paper presented at the Annual Meeting of the American Society for Information Science and Technology, Austin, Texas. (Panel organizer, moderator, presenter)
- Hsieh-Yee, I. (2006, September). *Search Performance of Google Scholar and Academic Search Premier*. Paper presented at the ERIC Publishers Meeting, Washington, D.C.
- Hsieh-Yee, I., Zhang, S., and Rong Tang, R. (2006, June). *User Perceptions of a Federated Search System*. Poster presented at Joint Conference on Digital Libraries, Chapel Hill, North Carolina.
- Hsieh-Yee, I. and Zhang, S. (2006, June). *Preparing Users for Federated Search: Implications of a MetaLib User Perceptions Study*. Paper presented at the 2006 Ex Libris User Groups of North America Conference, Knoxville, Tennessee.
- Hsieh-Yee, I. (2006, January). *MERIC Organizations and Navigation*. Paper presented at the 2006 ALISE Annual Conference, San Antonio, Texas.
- Hsieh-Yee, I. (2006, January). *Metadata and Cataloging Education: Recommended Competencies*. Paper presented at the 2006 ALISE Annual Conference, San Antonio, Texas.
- Hsieh-Yee, I. (2005, November). *Digital Library Evaluation: Progress & Next Steps*. Presentation at the Annual Meeting of the American Society for Information Science & Technology, Charlotte, North Carolina.
- Hsieh-Yee, I. (2005, August). *Providing Access to Digital Content: Issues for DL Managers*. Presentation at MDK12 Digital Library Steering Committee Meeting, Columbia, Maryland.
- Hsieh-Yee, I. (2005, April). *Enhancing Teaching and Learning: The Role of School Library Media Specialists*. Presentation at Meeting of the Baltimore County Public School System School Media Specialists, Baltimore, Maryland.

- Hsieh-Yee, I. (2005, January). *Subject Access and Users: Insights & Inspirations from Marcia J. Bates*. Paper presented at the Historical Perspectives SIG, 2005 Conference of the Association for Library and Information Science Education, Boston, Massachusetts.
- Hsieh-Yee, I. (2005, January). *Electronic Resource Management: Practice, Employer Expectations, & CE Interests*. Paper presented at Technical Services Education SIG, 2005 Conference of the Association for Library and Information Science Education, Boston, Massachusetts.
- Hsieh-Yee, I. (2004, October). *Library Professionals for the Digital Age: Competencies & Preparation*. Paper presented at Bibliographic Access Management Team meeting, Library of Congress, Washington, D.C.
- Hsieh-Yee, I. (2004, January). *Cataloging and metadata expertise for the digital era*. Presented at Preparing 21st Century Cataloging and Metadata Professionals: A Workshop for Educators and Trainers, San Diego and sponsored by ALCTS, ALISE, LC, and OCLC.
- Hsieh-Yee, I. (2004, January). *Educating catalogers for the digital era*. Paper presented at the Technical Services SIG, 2004 Conference of the Association for Library and Information Science Education, San Diego.
- Hsieh-Yee, I. (2003, July). *Cataloging Education for the 21st Century*. A presentation at the Library of Congress, Washington, D.C.
- Hsieh-Yee, I. (2002, January) *Metadata Education and Research Priorities: A Delphi Study of Metadata Experts*. Presentation at the 2002 Conference of the Association for Library and Information Science Education, New Orleans.
- Hsieh-Yee, I. (2001, November). *A Delphi Study of Metadata: Preliminary Findings*. Poster session at the 2001 Annual Meeting of the American Society for Information Science & Technology, Washington, D.C.
- Hsieh-Yee, I. (2001, June). *Resources on Asian American Children: Analysis of Retrieval by Search Engines and WorldCat*. Presentation at the National Conference on Asian Pacific American Librarians, San Francisco.
- Hsieh-Yee, I. (2001, January). *Delphi Study on Metadata: Project Design*. Presentation at Research Awards Session, Association for Library & Information Science Education, Washington, D.C.
- Hsieh-Yee, I. (2000, May). *Web Search Behavior Research: Progress and Implications*. Presentation at the Symposium on Evaluating Library and Information Science Research, University of Wisconsin-Madison, Madison, Wisconsin.

- Hsieh-Yee, I. (2000, March). *ERIC User Services: Evaluation in a Decentralized Environment*. Presentation at the National ERIC Joint Directors/Technical Meeting, Arlington, Virginia.
- Hsieh-Yee, I. (2000, January). *Enhancing Learning with Web Technology*. Presentation at Faculty Conversations, Catholic University of America, Washington, D.C.
- Hsieh-Yee, I. (2000, January). *From Surrogates to Objects: CUA's Approaches to Organizing Electronic Resources*. Paper presentation at the Annual Conference of the Association for Library and Information Science Education, San Antonio, Texas.
- Yee, P., and Hsieh-Yee, I. (1997, November). *Individual Differences in Search Behavior on the WWW*. A poster session presented at the 38th Annual Meeting of the Psychonomic Society, Philadelphia, Pennsylvania.
- Hsieh-Yee, I. (1997, April). *Research + Marketing + Preparation = Job!* Presented at the "Workshop on Resume and Interview Techniques," Special Libraries Association, Student Chapter, Catholic University of America, Washington, D.C.
- Hsieh-Yee, I. (1997, February). *Creating CyberCatalogers: Education and Training*. Presentation at ALA's Midwinter Meeting, Washington, D.C.
- Hsieh-Yee, I. (1997, February). *Search Tactics of Web Users in Searching for Texts, Graphics, Known Items and Subjects: A Search Simulation Study*. Presented at the Conference of the Association for Library and Information Science Education, Washington, D.C.
- Hsieh-Yee, Ingrid. "Beginning Your Special Library/Information Center Career." Presented at SLA's "Career Day," Jan. 11, 1997, Catholic University of America.
- Hsieh-Yee, I. (1996, September). *The Roles of Library and Information Scientists in Managing Electronic Information*. Presentation at Hamilton College, Clinton, New York.
- Hsieh-Yee, I. (1996, May). *The Future of Cataloging as a Profession*. Presented at "The Cataloging Forum, Library of Congress, Washington, D.C.
- Hsieh-Yee, I. (1994, October). *The Impact of the Internet on OPACs*. Presented at the Third Workshop on User Interfaces for OPACs, Library of Congress, Washington, D.C.

Reports

- Hsieh-Yee, I., with Knowledge Management Competencies and Performance Action Group of the Federal Knowledge Management Initiative. "From Knowledge Management Competencies to Improved Organizational Performance." April 9, 2009.

Hsieh-Yee, I., with Knowledge Practices Action Group of the Federal Knowledge Management Initiative. "KM Practice in Government Agencies: Findings and Recommendations." April 9, 2009.

Hsieh-Yee, I. "Delphi Study on Metadata." 2001. Three quarterly reports submitted to the Association for Library and Information Science Education.

Hsieh-Yee, I. "College Students' Information Channels: Patterns of Use and Possible Factors in Channel Selection." 1995. Submitted to the Catholic University of America.

Hsieh-Yee, I. "The Information-Seeking Patterns of Scholars and Their Use of an Online Information System." 1994. Submitted to the Council on Library Resources.

Book Reviews

Review of *The Measurement and Evaluation of Library Services*, by Sharon L. Baker and F. Wilfrid Lancaster. *Information Processing and Management* 30 (1994): 450-52.

Review of *Subject Access to Films and Videos*, by Sheila S. Intner and William E. Studwell; and *Cataloging Unpublished Nonprint Materials*, by Verna Urbanski with Bao Chu Chang and Bernard L. Karon. *Information Processing and Management* 30 (1994): 449-50.

Review of *Automated Information Retrieval in Libraries: A Management Handbook*, by Vicki Anders. *Journal of Library and Information Science* 19 (1993): 98-100.

Review of *Full Text Databases*, by Carol Tenopir and Jung Soon Ro. *Information Processing and Management* 28 (1992): 667-68.

Review of *Descriptive Cataloging for the AACR2R And USMARC: A How-to-Do It Workbook*, by Larry Millsap and Terry Ellen Ferl. *Information Processing and Management* 28 (1992): 809-11.

Review of *MARC Manual: Understanding and Using MARC Records*, by Deborah J. Byrne. *Information Processing and Management* 28 (1992): 537-38.

Service

Professional Associations and Societies

- Library of Congress. RDA Training Program for the Profession. Co-authored with Tim Carlton. 2013-2014.
- 2014 Digital Preservation Outreach & Education Survey. Contributed to the design of the survey, 2014.
- National Digital Stewardship Alliance. Outreach Committee. 2011-2014.
- National Digital Stewardship Residency Program. Advisory Group, 2012-2013.
- FEDLINK Health Information Technology Advisory Council, 2011-2015.

- 2012 Joint Conference on Digital Libraries. Program Planning Committee, Pre-Conference Proposals Review Committee, 2012
- Catholic Research Resources Alliance. Five-Year Strategic Plan Task Force, 2011-2012
- Institute of Museum and Library Services. Grant reviewer. 2004, 2005, 2010.
- Association for Library and Information Science Education.
 - * ALISE Bodan Wynar Research Paper Award Committee, 2015, 2016, 2017
 - * ALISE Eugene Garfield Dissertation Award Competition, Jury, 2013, 2014
 - * ALISE Research Grant Competition Committee. Chair, 2012
 - * Pratt-Severn Faculty Innovation Award. Chair, 2009, 2010
 - * ALISE Doctoral Poster Jury, 2012
 - * “Information Organization Competencies for the 21st Century” Discussion session leader. 2009 Conference of the Association for Library and Information Science Education.
 - * Assisted Technical Services SIG Convener in organizing a program, ““Building a Metadata Education and Research Community through MERIC (Metadata Education and Research Information Commons): Demo and Stakeholder Input” for the 2008 ALISE conference.
 - * Association for Library Collections and Technical Services/Association for Library and Information Science Education (ALCTS/ALISE) Metadata Education and Research Information Center (MERIC) Advisory Board, Co-Chair (with Sherry Vellucci), 2005-2007. Chair, 2008-2009 (leading the effort to build MERIC, a repository and collaborative space for metadata educators, practitioners, and researchers)
 - * Technical Services SIG, Convener, 2004-2005. Organized a program on “Electronic Resources Management: Current Practices, Employer Expectations, and Teaching Strategies” for the 2005 conference in Boston, Massachusetts.
 - * Technical Services SIG, Convener, 2003-2004. Organized a program on “Organizing Information with Metadata: Desired Competencies and Teaching Innovations” for the 2004 conference.
 - * Technical Services SIG, Convener, 1999-2000. Organized a program on "Teaching the Organization of Electronic Resources" for the 2000 conference.
 - * Curriculum SIG, Co-convener (with Sibyl Moses), 1996-97. Organized a program on “Government Information Policy” for the 1997 conference.
- American Society for Information Science & Technology.
 - * Reviewer, Conference program panel submissions and poster submissions, 2005, 2006, 2007, 2009, 2011, 2012, 2013, 2014, 2015, 2016, 2017
 - * Nomination Committee, 2009-2011
 - * Information Science Education Special Interest Group. American Society for Information Science and Technology. Chair-Elect, 2007-2008. Chair 2008-2009.
 - * Committee on Information Science Education. 1999-2006.
 - * Committee on Information Science Education. Organizing Committee for an orientation program for students at ASIS annual meetings, 1999-2001
 - * Committee on Information Science Education. Sub-committee on Student Welfare (focusing on issues related to master's education), 1998-2001

- * SIG ED. Organizing Committee for the "Seminar on Research and Career Development" for junior researchers. 1995-96 (chair), 1997-2001
- * ISI Doctoral Dissertation Proposal Scholarship Jury, 1997; 2001, 2002
- * Pratt-Severn Best Student Research Paper Award Jury. Chair. 1997
- * 1998 Midyear Meeting (referee of contributed papers), 1997
- * Organizer and moderator of the ASIS Doctoral Forum and the Doctoral Research Seminar 1994-1995
- * SIG Human Computer Interaction. Chair-Elect, Chair, 1993-1995
- * Doctoral Forum Award Jury, 1995
- * Best Student Paper Award Jury, 1995
- American Library Association.
 - * Committee on Accreditation, External Review Panelist, 2009- (site visiting team 2013-2014; site visiting team 2016-2017)
 - * Association for Library Collections and Technical Services Task Force on Competencies and Education for a Career in Cataloging, member, 2008-2009
 - * Facilitator for "What They Don't Teach in Library School: Competencies, Education and Employer Expectations for a Career in Cataloging," an Association for Library Collections and Technical Services Preconference, June 22, 2007 in Washington, D.C. Also a local liaison for bringing this program to the Catholic University of America.
 - * Facilitator for a discussion on "Effect of Electronic Resources on Technical Services" at ALA's Midwinter Meeting held in Feb. 1997 in Washington, D.C.
 - * International Relations Committee, Subcommittee Task Force for IFLA and China, 1994-1997
- Virginia Association of School Librarians. Scholarships and Awards Committee. 2010-2012
- Federal Knowledge Management Initiative, Knowledge Management Practices Action Group. Member. 2009 (leading the effort to build a knowledge management repository)
- Federal Knowledge Management Initiative, Knowledge Management Competencies & Learning Action Group. Member. 2009 (developing an action plan for helping government knowledge workers and government agencies to develop knowledge management competencies)
- National Center for Education Statistics. Technical Review Panel. 2008.
- External evaluator for a case of promotion to full professorship. University of Tennessee. 2008.
- National Information Standards Organization (NISO). Advisory Board, Revision of "IMLS Framework of Guidance for Building Good Digital Collections," 2004, 2007.
- Library of Congress, Bibliographic Control of Web Resources: A Library of Congress Action Plan. Principal Investigator of Action Item 5.1, focusing on cataloging and metadata education for students and new librarians, 2002-2003. (worked with the Association for Library Collections and Technical Services, Education Task Force)

- Chinese American Librarians Association
 - * Chinese American Librarians Association Outstanding Library Leadership Award in Memory of Dr. Margaret Chang Fung, Award Committee, 2016-2017
 - * Achievement Award Jury, 2000-2001
 - * CALA Goal 2000 Task Force, 1997
 - * Scholarship Committee, 1995, 1996-1997 (chair)
 - * Board of Directors, 1994-1997
 - * Publication Committee, 1993-1995
 - * International Relations Committee, 1993-1996
- SailorSM Assessment Advisory Group (An impact study of Sailor, Maryland's Public Information Network), 1995
- Editorial boards
 - Journal of Library & Information Science. Editorial Board, 2012-
 - Chinese American Librarians Association, *Occasional Papers Series*. Editorial Board, 2009-2016.
 - Library Quarterly*. Editorial Board, 2003-2008
 - Bulletin of the Medical Library Association*, 1994-97
 - Newsletter editor for the Chinese American Librarians Association, 1989-92
- Referee for the following journals
 - Information Processing and Management*
 - Journal of Digital Information*
 - Journal of Education for Library and Information Science*
 - Journal of Information Science*
 - Journal of Library & Information Science*
 - Journal of Library Metadata*
 - Journal of the American Society for Information Science & Technology*
 - Library and Information Science Research*
 - Library Quarterly*
- Expert reviewer, “Digital Library” course, Evaluation module, University of North Carolina, Chapel Hill, 2007-2008.
- Expert reviewer, “Information Organization” course, University of Michigan, Ann Arbor. 2007.

Catholic University of America

- Academic Senate Committee on Appointments and Promotions, 2005-2008
- Academic Senate Committee on Committees and Rules, 2009-2012, 2019-2022
- Academic Senate representative (for School of Arts & Sciences), 2017-2020

- Academic Senate representative (for School of Library & Information Science) 2003-2012
- Academic Senate Library Committee, Interim Chair (2007), Member, 2008-2012
- Academic Senate Committee on Computing, 1995-2003
- President's Administrative Council, 2010-2012
- Deans' Council, 2010-2012
- Academic Leadership Group, 2010-2012
- Graduate Board, 2010-2012
- School of Arts & Sciences, Committee on Appointments and Promotions, 2015-2019
- School of Arts & Sciences, Academic Council, 2015-2016.
- School of Arts & Sciences, Ordinary Professor Group, 2013-
- Doctoral Dissertation Defense Committee, Chair, Dept. of Psychology, 2016, 2017, 2018, 2019, 2020, 2021, 2022
- Doctoral Dissertation Defense Committee, Chair, Dept. of Education, 2014, 2015, 2017, 2018, 2019
- Doctoral Dissertation Defense Committee, Chair, National Catholic School of Social Services, 2019, 2020, 2021
- Doctoral Dissertation Defense Committee, Chair, School of Nursing, 2006, 2008
- CUA Scholarly Communication Project Team, Member (2007), Chair, 2008-2009
- Dean Search Committee, 1992-1994, 1998-1999, 2002-2003, 2006-2007
- Fulbright Review Panel, 2006
- CUA Service Learning Advisory Board, 2001-2002
- CUA Faculty Conversations on Enhancing Teaching and Learning through Technology, Planning Group, 1999-2001
- CUA Initiative on Technology and Teaching, 1998-2001

Dept. of Library and Information Science

- Admissions Committee, 2007-2009, Chair 2010-2012, Member 2013-2015, Member 2018-2022
- Appointments and Promotions Committee, 1991-2022
- Student advisement, 1990-2022
- 4+1 Bachelor to Master program, Program Advisor, 2014-2022
- LIS Minor, Advisor, 2017-2022
- Curriculum Committee, 2020-21 (Member), 2021-2022 (Chair)
- Scholarship & Awards Committee, 2016-2018, 2020-21
- Comprehensive examination editor, 2020-21
- Symposium, Colloquium, Lecture Series Committee, May 2018-May 2020 (Chair)
- Symposium and Colloquium Committee, fall 2016-May 2018
- Bridging the Spectrum Symposium session moderator, 2018, 2019, 2020, 2021

- Organization of Information Core Course Outcomes Assessment, 2019, 2020
- The Information Professions in Society Core Course Outcomes Assessment, 2019, 2020
- Community Services Librarianship, Course of Study Review and Revision, 2020
- Organization of Information, Course of Study Review and Revision, 2019
- Organization of Information, Competency and Course Offerings Review, 2018-2019
- Accreditation presentation, Chair, June 2015-August 2016
- Interim Co-Chair, June 2015-August 2016.
- Blended/OWL Learning Committee, spring 2016-2018
- Technology Committee, fall 2016-2017
- Comprehensive examination editor, 2016-2017, reader (every year since 1990)
- LIS Advisory Board, 2015-2016 (chair); fall 2016- May 2018 (member)
- Committee on Planning and Assessment, 2015-2016 (chair)
- Senior Faculty Committee, 2014-2016.
- Accreditation Steering Committee, 2014-2016 (Chair, 2015-2016)
- Accreditation Students Standard Committee, co-chair, 2014-2016
- Accreditation Mission, Goals, and Objectives Standard Committee, co-chair, 2014-2016
- Accreditation Curriculum Standard, member 2014-2-16
- Accreditation Administration and Finance Standard, member 2014-2016
- Cultural Heritage Information Management Project (IMLS-funded), Co-PI, 2012-2015
- Cultural Heritage Information Management Forum (scheduled for June 2015), Co-Organizer, 2013-2015
- Health Information Technology Interim Review Committee, 2015 (chair)
- Health Sciences Librarianship Advisory Group, 2015- (chair)
- Comprehensive examination editors, 2013-2014, 2016-2017
- National Digital Stewardship Alliance liaison, 2011-2014
- Advisory Board, Chair 2010-2012
- Academic Honesty Committee, Chair, 2008-2012
- Blended Learning Committee, 2010-2012
- Colloquium Committee, 2010-2012
- Comprehensive Examination Administration, 2010-2012
- Cultural Heritage Information Management Advisory Committee, 2010-2012 (chair), 2013-
- Curriculum Committee, 1991-2003, 2007-2009, Chair 2010-2012, member 2013-
- Curriculum Subcommittee on Comprehensive Examination, Chair 2009-2012
- Health Information Technology Advisory Board, Chair 2010-2012. Member 2013-
- Health Sciences Advisory Committee, 2009, Chair 2010-2012. Member 2013-
- HIT Expert Forum, Chair 2012. Member 2013-2015
- Health Information Technology Student Group Advisor, 2011-2012
- State Council for Higher Education of Virginia, SLIS Representative, 2010-2012
- Symposium Planning Committee, 2010-2012
- Website Management Team, Chair, 2010-2012
- Urban School Librarianship Project (IMLS-Funded), PI, 2007-2011 (chair, 2010-11)

- Failing Grades Committee, 1995-1997 (chair), 2000-2001 (chair), 2004-2005 (chair), 2007 (chair)-2011
- Faculty Search Committee, 1994-1998, 2002-2004, 2006 (chair), Fall 2007-2009, Chair fall 2009-2012
- Recruitment Committee, Chair 2010-2012
- Strategic Planning Committee, Chair 2010-2012
- Technology Committee, 2010-2012
- Accreditation Advisory Committee, 2007-2009
- Accreditation Coordinating Committee, 2007-2009
- Accreditation Steering Committee, 2007-2009
- SLIS Advisory Group, 2007-2009
- Accreditation Curriculum Standard Committee, Co-chair, 2007-2009
- Accreditation Faculty Standard Committee, Co-chair, 2007-2009
- LSC 551 Information Organization Review Team, Co-chair, 2008-2009, 2015-2016.
- Curriculum Subcommittee on Portfolios, 2009
- LSC 555 Information Systems in Libraries and Information Centers Review Team, contributor, 2008-2009
- Redesign of LSC 730 Use and Users of Libraries and Information. 2009-
- Development of a metadata institute that was taught as LSC 715 Organization of Internet Resources in 2008. The institute is being revised and will be offered in 2010 under a new course title.
- Development of lesson plans, assignments, and evaluation rubrics for LSC 606, Cataloging and Classification, for the School's NCATE accreditation. 2008
- Howard and Mathilde Rovelstad Scholarship Committee, Chair, 2004-2007
- Assistant Dean Search Committee, Chair, Fall 2007
- Liaison to the Association for Library Collections and Technical Services to bring its preconference program, Cataloging Education and Employer Expectations, to CUA during the 2007 American Library Association Annual Meeting in Washington, D.C.
- Organizer of the colloquium presentation and reception for Tamar Sadeh of Ex Libris on PRIMO June 2007
- Practicum review and design (work with potential supervisors, such as the American Indian Museum internship description revision) 2006-
- Comprehensive examinations (edits, proctoring, and grading), 1990-
- SLIS Web site redesign: Comments and suggestions. Fall 2007
- Conducted surveys of current students and alumni in preparation for the 2005 re-accreditation, 2004-2005
- Technology Committee, 1992-1999 (chair, 1996-1998), 2002-2003 (member)
- Colloquia Committee 1997-1999, 2002-2003.
- Advisor of the CUA Student Chapter of the American Society for Information Science and Technology, 2002-2003
- Visiting Professor Search Committee, 1999, 2000, 2001
- Leader, Participation in the CORC experiment, 1999-2000
- Advisor of the Special Libraries Association Student Chapter, 1993-1999; the group was recognized for outstanding leadership by SLA in 1999.

- COA planning Committee, Task Force on Electronic Presentation of SLIS Reports (team leader) 1997-1998
- COA Planning Committee, Subcommittee on Technology 1996-1998
- NLM practicum coordinator, 1997-1998
- Computer Literacy Workshops: Assisted with the development and evaluation of the workshops, 1996-1998
- Leader, Participation in the InterCat project, 1995-1997

Hsieh-Yee Declaration

Appendix B

APPENDIX B

[Library of Congress](#) >> [MARC](#) >> [Understanding MARC](#)

MARC 21 Reference Materials

[Part VII: A Summary of Commonly Used MARC 21 Fields](#)

[Part VIII: A List of Other Fields Often Seen in MARC Records](#)

[Part IX: The Leader](#)

[Part X: Field 008 for Books](#)

Part VII:

A Summary of Commonly Used MARC 21 Fields

This is a summary of the MARC 21 tags used most frequently by libraries in entering their own bibliographic records. For full listings of all MARC 21 tags, indicators, and subfield codes, see *MARC 21 Format for Bibliographic Data*.

In the explanations on these pages:

Tags -- The tags (3-digit numbers) are followed by the names of the fields they represent. In this summary, and in the *MARC 21 Format for Bibliographic Data*, if a tag can appear more than once in one bibliographic record, it is labeled repeatable (R). If it can only be used once, it is labeled non-repeatable (NR). For example, a catalog record can have several subjects, so the tags for subject added entries (6XX) are labeled repeatable (R).

Indicators -- The use of indicators is explained in fields where they are used. Indicators are one-digit numbers. Beginning with the 010 field, in every field -- following the tag -- are two character positions, one for Indicator 1 and one for Indicator 2. The indicators are not actually defined in all fields, however. And it is possible that a 2nd indicator will be used, while the 1st indicator remains undefined (or vice versa). When an indicator is undefined, the character position will be represented by the character # (for blank space).

Subfield codes -- All the data in each field (beginning with the 010 field) is divided into subfields, each of which is preceded by a delimiter-subfield code combination. The most common subfield codes used with each tag are shown. Each subfield code is preceded by the character \$, signifying a delimiter. The name of the subfield follows the code.

In general, every field **MUST** have a subfield 'a' (**\$a**). One exception that is often seen is in Field 020 (ISBN), when the ISBN information (subfield **\$a**) is unavailable but the price (subfield **\$c**) is known. Some subfields are repeatable. In this summary, repeatability is noted for only the more common repeatable subfields.

Examples: Examples follow the explanation for each field. For clarity, one space has been placed between the tag and the first indicator, one space has been placed between the second indicator and the first delimiter-subfield code, and one space has been inserted between the delimiter-subfield code and the subfield data.

010 Library of Congress Control Number -- (LCCN)

(NR, or Not Repeatable

Indicators undefined.

*Subfield used most often:***\$a** -- Library of Congress control number

Example: 010 ## \$a ###86000988#

020 International Standard Book Number -- (ISBN)

(R, or Repeatable)

*Indicators undefined.**Subfields used most often:***\$a** -- International Standard Book Number**\$c** -- Terms of availability (often a price)**\$z** -- Cancelled/invalid ISBN (R)

Example: 020 ## \$a 0877547637

040 Cataloging source -- (NR)*Indicators undefined.**Subfields used most often:***\$a** -- Original cataloging agency**\$c** -- Transcribing agency**\$d** -- Modifying agency (R)

Example: 040 ## \$a DLC
 \$c DLC
 \$d gwhs

100 Main entry -- Personal name -- (primary author)

(NR; there can be only one main entry)

Indicator 1: Type of personal name entry element

0 -- Forename

1 -- Surname (this is the most common form)

3 -- Family name

Indicator 2 undefined.

Indicator 2 became obsolete in 1990. Older records may display 0 or 1

*Subfields used most often:***\$a** -- Personal name**\$b** -- Numeration**\$c** -- Titles and other words associated with a name (R)**\$q** -- Fuller form of name**\$d** -- Dates associated with a name (generally, year of birth)

Example: 100 1# \$a Gregory, Ruth W.

\$q (Ruth Wilhelme), \$d 1910-

130 Main entry -- Uniform title -- (NR)*Indicator 1:* Nonfiling characters

0-9 -- Number of nonfiling characters present (for initial articles, including spaces)

Indicator 2 undefined.

Indicator 2 became obsolete in 1990. (See 100 above.)

*Subfields used most often:***\$a** -- Uniform title**\$p** -- Name of part/section of a work (R)**\$l** -- Language of a work**\$s** -- Version**\$f** -- Date of a work

<i>Example:</i> 130 0# \$a Bible. \$p O.T. \$p Psalms.
--

240 Uniform title (NR)*Indicator 1:* Uniform title printed or displayed

0 -- Not printed or displayed

1 -- Printed or displayed (most common)

Indicator 2: Nonfiling characters

0-9 -- Number of nonfiling characters present (for initial articles, including spaces)

*Subfields used most often:***\$a** -- Uniform title**\$l** -- Language of a work**\$f** -- Date of a work

<i>Example:</i> 240 10 \$a Ile mystérieuse. \$l English. \$f 1978

245 Title Statement (NR)*Indicator 1:* Title added entry

(Should the title be indexed as a title added entry?)

0 -- No title added entry

(indicates a title main entry; i.e. no author is given)

1 -- Title added entry

(the proper indicator when an author given in 1XX; the most common situation)

Indicator 2: Nonfiling characters

0-9 -- Number of nonfiling characters present, including spaces; usually set at zero, except when the title begins with an article; e.g., for *The robe*, the second indicator would be set to 4. The letters *T*, *h*, *e*, and the space following them are then ignored in alphabetizing titles. The record will be automatically filed under "*r*" -- for *Robe*.

Subfields used most often:

\$a -- Title proper
\$h -- Medium (often used for non-book media)
\$p -- Name of part/section of a work (R)
\$b -- Reminder of title (subtitles, etc.)
\$c -- Remainder of title page transcription/Statement of responsibility

Example: 245 14 \$a The DNA story :
 \$b a documentary history of gene
 cloning /
 \$c James D. Watson, John Tooze.

246 Varying form of title (R)

Indicator 1: Note/title added entry controller

1 -- Note, title added entry
 3 -- No note, title added entry

Indicator 2: Type of title

-- No information provided
 0 -- Portion of title
 1 -- Parallel title
 4 -- Cover title
 8 -- Spine title

Subfield used most often:

\$a -- Title proper

Example: 246 3# \$a Four corners power review

250 Edition statement (NR)

Indicators undefined.

Subfield used most often:

\$a -- Edition statement

Example: 250 ## \$a 6th ed.

260 Publication, distribution, etc. (Imprint) (R)

Indicator 1: Sequence of publishing statements

-- No information provided

Indicator 2: Undefined

Subfields used most often:

\$a -- Place of publication, distribution, etc. (R)

\$b -- Name of publisher, distributor, etc. (R)

\$c -- Date of publication, distribution, etc. (R)

Example: 260 ## \$a New York :
 \$b Chelsea House,
 \$c 1986.

300 Physical description (R)

Indicators undefined.

Subfields used most often:

\$a -- Extent (number of pages) (R)

\$b -- Other physical details (usually illustration information)

\$c -- Dimensions (cm.) (R)

\$e -- Accompanying material (for example, "teacher's guide" or "manual")

Example: 300 ## \$a 139 p. :
 \$b ill. ;
 \$c 24 cm.

440 Series statement / Added entry--Title

This field was made obsolete in 2008 to simplify the series statement. See 490 and 830.

490 Series statement (No added entry is traced from field) (R)

Indicator 1: Specifies whether series is traced (whether an 8XX tag is also present)

0 -- Series not traced

1 -- Series traced (8XX is in record)

Indicator 2 undefined.

Subfield used most often:

\$a -- Series statement (R)

\$v -- Volume number (R)

Example: 490 1# \$a Colonial American craftsmen

500 General note (R)

Indicators undefined.

Subfield used most often:

\$a -- General note (Used when no specialized note field has been defined for the information. Examples: Notes regarding the index; the source of the title; variations in title; descriptions of the nature, form, or scope of the item.)

Example: 500 ## \$a Includes index.

504 Bibliography, etc. note (R)

Indicators undefined.

Subfield used most often:

\$a -- Bibliography, etc. note

Example: 504 ## \$a Includes bibliographical references.

505 Formatted contents note (R)

Indicator 1: Type of contents note

0 -- Complete contents

1 -- Incomplete contents (used with multivolume set when some volumes are not yet published)

2 -- Partial contents

Indicator 2: Level of content designation

-- Basic

Subfield used most often:

\$a -- Formatted contents note

Example: 505 0# \$a Pride and prejudice -- Emma -- Northanger Abbey.

520 Summary, etc. note (R)

Indicator 1: Display constant controller

-- Summary

1 -- Review

2 -- Scope and content

3 -- Abstract

Indicator 2 undefined

Subfields used most often

\$a -- Summary, abstract, or annotation

\$b -- Expansion of summary note

Example: 520 ## \$a This basic guide to parliamentary procedure tells how to conduct and participate in a meeting properly.

600 Subject added entry -- Personal name (R)

Indicator 1: Type of personal name entry element

0 -- Forename

- 1 -- Surname (this is the most common form)
 - 3 -- Family name
- Indicator 2:* Subject heading system/thesaurus (identifies the specific list or file which was used)
- 0 -- Library of Congress Subject Headings
 - 1 -- LC subject headings for children's literature
 - 2 -- Medical Subject Headings
 - 3 -- National Agricultural Library subject authority file
 - 4 -- Source not specified
 - 5 -- Canadian Subject Headings
 - 6 -- Répertoire de vedettes-matière
 - 7 -- Source specified in subfield \$2

(Note regarding Sears subject headings: The MARC 21 format does not provide an assigned indicator for Sears subject headings. Therefore, an indicator of 7 is used, and the MARC defined code "sears" is placed in subfield \$2.)

Subfields used most often:

- \$a** -- Personal name (surname and forename)
- \$b** -- Numeration
- \$c** -- Titles and other words associated with a name (R)
- \$q** -- Fuller form of name
- \$d** -- Dates associated with a name (generally, year of birth)
- \$t** -- Title of a work
- \$v** -- Form subdivision (R)
- \$x** -- General subdivision (R)
- \$y** -- Chronological subdivision (R)
- \$z** -- Geographic subdivision (R)
- \$2** -- Source of heading or term (used with 2nd indicator of 7)

Example: 600 10 \$a Shakespeare, William,
 \$d 1564-1616
 \$x Comedies
 \$x Stage history.

Example: 600 10 \$a Shakespeare, William,
 \$d 1564-1616
 \$x Knowledge
 \$z Rome
 \$v Congresses.

Notice that subfields \$v, \$x, and \$z in the 600 field are repeatable. Subfields \$v, \$x, \$y, and \$z do not have to be in alphabetical order. They will be in the order prescribed by the instructions given by the subject heading system.

610 Subject added entry -- Corporate name (R)

Indicator 1: Type of corporate name entry element

- 0 -- Inverted name (not used with AACR2)
- 1 -- Jurisdiction name
- 2 -- Name in direct order

Indicator 2: Subject heading system/thesaurus.

See indicator 2 under 600

Subfields used most often:

- \$a** -- Corporate name or jurisdiction name as entry element
- \$b** -- Subordinate unit (R)
- \$v** -- Form subdivision (R)
- \$x** -- General subdivision (R)
- \$y** -- Chronological subdivision (R)
- \$z** -- Geographic subdivision (R)
- \$2** -- Source of heading or term (used with 2nd indicator of 7)

Example: 610 10 \$a United States.
 \$b Army Air Forces
 \$v Biography.

650 Subject added entry -- Topical term (Most subject headings fit here.) (R)

Indicator 1: Level of subject

- # -- No information provided

Indicator 2: Subject heading system/thesaurus

(identifies the specific list or file which was used)

- 0 -- Library of Congress Subject Headings
- 1 -- LC subject headings for children's literature
- 2 -- Medical Subject Headings
- 3 -- National Agricultural Library subject authority file
- 4 -- Source not specified
- 5 -- Canadian Subject Headings
- 6 -- Répertoire de vedettes-matière
- 7 -- Source specified in subfield \$2

Note regarding Sears subject headings: The MARC 21 format does not provide an assigned indicator for Sears subject headings. Therefore, an indicator of 7 is used, and the MARC defined code "sears" is placed in subfield \$2.)

Subfields used most often:

- \$a** -- Topical term
- \$v** -- Form subdivision (R)
- \$x** -- General subdivision (R)
- \$y** -- Chronological subdivision (R)
- \$z** -- Geographic subdivision (R)
- \$2** -- Source of heading or term used with 2nd indicator of 7)

Example: 650 #0 \$a Theater
 \$z United States
 \$v Biography
 \$v Dictionaries.

Notice that subfields \$v, \$x, and \$z in the 650 field are repeatable. Subfields \$v, \$x, \$y, and \$z do not have to be in alphabetical order. They will be in the order prescribed by the instructions given by the subject heading system.

651 Subject added entry -- Geographic name (R)

Indicator 1: undefined.

Indicator 2: Subject heading system/thesaurus.

See indicator 2 under 600

Subfields used most often:

\$a -- Geographic name

\$v -- Form subdivision (R)

\$x -- General subdivision (R)

\$y -- Chronological subdivision (R)

\$z -- Geographic subdivision (R)

\$2 -- Source of heading or term (used with 2nd indicator of 7)

Example: 651 #0 \$a United States
 \$x History
 \$v Chronology.

Notice that subfields \$v, \$x, and \$z in the 651 field are repeatable. Subfields \$v, \$x, \$y, and \$z do not have to be in alphabetical order. They will be in the order prescribed by the instructions given by the subject heading system.

700 Added entry -- Personal name (R)

Indicator 1: Type of personal name entry element

0 -- Forename

1 -- Surname (this is the most common form)

3 -- Family name

Indicator 2: Type of added entry

-- No information provided (most common; co-authors, editors, etc.)

2 -- Analytical entry (The values for Indicator 2 changed in 1994 with Format Integration, and older records may display additional values. An analytical entry involves an author/title of an item contained in a work.)

Subfields used most often:

\$a -- Personal name

\$b -- Numeration

\$c -- Titles and other words associated with a name (R)

\$q -- Fuller form of name

\$d -- Dates associated with a name (generally, year of birth)

\$e -- Relator term (such as ill.) (R)

\$4 -- Relator code (R)

Example: 700 1# \$a Baldrige, Letitia.

710 Added entry -- Corporate name (R)

Indicator 1: Type of corporate name entry element

0 -- Inverted name (not used with AACR2)

1 -- Jurisdiction name

2 -- Name in direct order

Indicator 2: Type of added entry.

See Indicator 2 under 700

-- No information provided

2 -- Analytical entry

Subfields used most often:

\$a -- Corporate name or jurisdiction name as entry element

\$b -- Subordinate unit (R)

Example: 710 2# \$a Sunburst Communications (Firm)

740 Added entry -- Uncontrolled related/analytical title (R)

Indicator 1: Nonfiling characters

0-9 -- Number of nonfiling characters present (for initial articles, including spaces)

Indicator 2: Type of added entry. See Indicator 2 under 700

-- No information provided

2 -- Analytical entry

(This field was redefined in 1994 with Format Integration. Prior to 1994, the field was also used for variant titles, such as a different wording on a spine title. In records created since Format Integration, those variant titles appear in a 246 field.)

Subfield used most often:

\$a -- Title

Example: 740 02 \$a Uncle Vanya.

800 Series added entry -- Personal name (R)

Indicator 1: Type of personal name entry element

0 -- Forename

1 -- Surname

3 -- Family name

Indicator 2 undefined.

Subfields used most often:

\$a -- Personal name

\$b -- Numeration

\$c -- Titles and other words associated with a name (R)

\$q -- Fuller form of name

\$d -- Dates associated with a name (generally, year of birth)

\$t -- Title of a work (the series)

\$v -- Volume number

Example: 800 1# \$a Fisher, Leonard Everett.
 \$t Colonial American craftsmen.

830 Series added entry -- Uniform title (R)

Indicator 1 undefined.

Indicator 2: Nonfiling characters

0-9 -- Number of nonfiling characters present (for initial articles, including spaces)

Subfield used most often:

\$a -- Uniform title

\$v -- Volume number

Example: 830 #0 \$a Railroads of America (Macmillan)

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Part VIII:

A List of Other Fields Often Seen in MARC Records

- 001 Control number
- 003 Control number identifier
- 005 Date and time of latest transaction
- 006 Fixed-length data elements -- additional material characteristics
- 007 Physical description fixed field
- 008 Fixed length data elements (See [Part X](#))
- 022 International Standard Serial Number (ISSN)
- 037 Source of acquisition
- 041 Language code
- 043 Geographic area code
- 050 Library of Congress call number
- 060 National Library of Medicine call number
- 082 Dewey Decimal classification number (the one recommended by the Library of Congress; locally-assigned call numbers may appear elsewhere)
- 110 Main entry -- Corporate name (less frequent under AACR2 rules)
- 256 Computer file characteristics
- 263 Projected publication date
(indicates a CIP -- Cataloging in Publication -- record)

- 306 Playing time
- 508 Creation/production credits note
- 510 Citation/references note (review sources)
- 511 Participant or performer note
- 521 Target audience note (first indicator: 0 = reading grade level, 1 = interest age level, 2 = interest grade level, 3 = special audience characteristics, 4 = motivation interest level)
- 530 Additional physical form available note
- 538 System details note
- 586 Awards note
- 656 Index term -- Occupation
- 730 Added entry -- Uniform title
- 852 Location
- 856 Electronic location and access
- 9XX Reserved for local use. (They are used by vendors, systems, or individual libraries to exchange additional data)

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Part IX:

The Leader

There are 24 positions in the Leader, numbered from 00 to 23. For fuller explanation, see the *MARC 21 Format for Bibliographic Data*.

- 00-04 Record length (calculated by the computer for each record)
- 05 Record status
 - a = increase in encoding level
 - c = corrected or revised
 - d = deleted
 - n = new
 - p = increase in encoding from prepublication (previous CIP)
- 06 Type of record
 - a = language material
 - c = printed music
 - d = manuscript music
 - e = cartographic material
 - f = manuscript cartographic material
 - g = projected medium
 - i = nonmusical sound recording
 - j = musical sound recording
 - k = 2-dimensional nonprojectable graphic
 - m = computer file
 - o = kit
 - p = mixed materials
 - r = 3-dimensional artifact or naturally occurring object
 - t = manuscript language material
- 07 Bibliographic level
 - a = monographic component part
 - b = serial component part

- c = collection
- d = subunit
- i = integrating resource
- m = monograph/item
- s = serial
- 08 **Type of control**
 - # = no specified type
 - a = archival
- 09 **Character coding scheme**
 - # = MARC-8
 - a = UCS/Unicode
- 10 **Indicator count** (always "2")
- 11 **Subfield code count** (always "2")
- 12-16 **Base address of data** (calculated by the computer for each record)
- 17 **Encoding level**
 - # = full level
 - 1 = full level, material not examined
 - 2 = less-than-full level, material not examined
 - 3 = abbreviated level
 - 4 = core level
 - 5 = partial (preliminary) level
 - 7 = minimal level
 - 8 = prepublication level (CIP)
 - u = unknown
 - z = not applicable
- 18 **Descriptive cataloging form**
 - # = non-ISBD
 - a = AACR2
 - i = ISBD
 - u = unknown
- 19 **Multipart resource record level**
 - # = Not specified or not applicable
 - a = Set
 - b = Part with independent title
 - c = Part with dependent title
- 20 **Length of the length-of-field portion** (always "4")
- 21 **Length of the starting-character-position portion** (always "5")
- 22 **Length of the implementation-defined portion** (always "0")
- 23 **Undefined** (always "0")

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Part X:

Field 008 for Books

Field 008 is used for Fixed Length Data Elements ("Fixed Field Codes"). There are 40 character positions in field 008, numbered from 00-39. Undefined positions must contain either a blank (#) or a fill character (|). Positions 00-17 and 35-39 are defined the same way for all media.

The information shown here for positions 18-34 applies only to books. For explanation of all the positions below and for positions 18-34 for other media, see the *MARC 21 Format for Bibliographic Data*.

Note that field 008 has no indicators or subfield codes.

- 00-05 **Date entered on file (YYMMDD)**,
 where Y=year, M=month, and D=day
- 06 **Type of date/publication status:**
 b = no dates given; B.C. date involved
 e = detailed date
 s = single known date/probable date
 m = multiple dates
 r = reprint/reissue date (Date 1) and original date (Date 2)
 n = dates unknown
 q = questionable date
 t = publication date and copyright date
 | = no attempt to code
- 07-10 **Date 1/beginning date of publication**
- 11-14 **Date 2/ending date of publication**

Date fields contain the year(s) of publication. The type of date(s) in these elements are specified in fixed field element 06: Type of date/publication status. (For further details, see the field 008 description in the *MARC 21 Format for Bibliographic Data*.)

- 15-17 **Place of publication, production, or execution**
 For example:
 pk# = Pakistan
 cau = California (US)

(For a full list of codes used in these positions, see the [MARC Code List for Countries](#).)

- 18-21 **Illustrations (up to 4 codes):**
 # = no illustrations
 a = illustrations
 b = maps
 c = portraits
 d = charts
 e = plans
 f = plates
 g = music
 h = facsimiles
 i = coats of arms
 j = genealogical tables
 k = forms
 l = samples
 m = phonodisc, phonowire, etc.
 o = photographs
 p = illuminations
 | = no attempt to code
- 22 **Target audience:**
 # = unknown or not specified
 a = preschool

- b = primary
- c = pre-adolescent
- d = adolescent
- e = adult
- f = specialized
- g = general
- j = juvenile
- | = no attempt to code
- 23 **Form of item:**
 - # = none of the following
 - a = microfilm
 - b = microfiche
 - c = microopaque
 - d = large print
 - f = braille
 - r = regular print reproduction
 - s = electronic
 - | = no attempt to code
- 24-27 **Nature of contents (up to 4):**
 - # = no specified nature of contents
 - a = abstracts/summaries
 - b = bibliographies (is one or contains one)
 - c = catalogs
 - d = dictionaries
 - e = encyclopedias
 - f = handbooks
 - g = legal articles
 - i = indexes
 - j = patent document
 - k = discographies
 - l = legislation
 - m = theses
 - n = surveys of literature
 - o = reviews
 - p = programmed texts
 - q = filmographies
 - r = directories
 - s = statistics
 - t = technical reports
 - u = standards/specifications
 - v = legal cases and notes
 - w = law reports and digests
 - z = treaties
 - | = no attempt to code
- 28 **Government publication:**
 - # = not a government publication
 - i = international intergovernmental
 - f = federal/national
 - a = autonomous or semi-autonomous component
 - s = state, provincial, territorial, dependent, etc.
 - m = multistate
 - c = multilocal

- 1 = local
- z = other type of government publication
- o = government publication -- level undetermined
- u = unknown if item is government publication
- | = no attempt to code
- 29 **Conference publication:**
 - 0 = not a conference publication
 - 1 = conference publication
 - | = no attempt to code
- 30 **Festschrift:**
 - 0 = not a festschrift
 - 1 = festschrift
 - | = no attempt to code
- 31 **Index:**
 - 0 = no index
 - 1 = index present
 - | = no attempt to code
- 32 **Undefined (since 1990)** (Earlier records may contain the values 0 or 1)
 - # = Undefined
 - | = no attempt to code
- 33 **Literary form:**
 - 0 = not fiction (not further specified)
 - 1 = fiction (not further specified)
 - c = comic strips
 - d = dramas
 - e = essays
 - f = novels
 - h = humor, satires, etc.
 - i = letters
 - j = short stories
 - m = mixed forms
 - p = poetry
 - s = speeches
 - u = unknown
 - | = no attempt to code
- 34 **Biography:**
 - # = no biographical material
 - a = autobiography
 - b = individual biography
 - c = collective biography
 - d = contains biographical information
 - | = no attempt to code
- 35-37 **Language:**

A three-letter code. For example: eng fre ger spa rus ita

(For a full list of codes used in these positions, see the [MARC Code List for Languages](#).)

- 38 **Modified record:**
 - # = not modified
 - x = missing characters (because of characters unavailable in MARC character set)
 - s = shortened
 - d = "dashed-on" information omitted

r = completely romanized/printed cards in script
o = completely romanized/printed cards romanized
| = no attempt to code

39

Cataloging source:

= national bibliographic agency
c = cooperative cataloging program
d = other sources
u = unknown
| = no attempt to code

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