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44th Annual Broadcast Engineering Conference Proceedings

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Atlanta, Georgia

National Association of
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March 1990

Dear Industry Engineer,

On behalf of NAB's department of Science & Technology, I am pleased to present the 1990 NAB Engineering Conference *Proceedings.* We are very proud of these state-of-the-art technical papers covering so many important areas of U.S. broadcasting. In radio, we present papers on Digital Audio Broadcasting (DAB); the Radio Data System technology proposed for U.S. FM broadcasters; a progress report on NAB's experimental AM antenna project; and, of course, many fine papers on new radio engineering techniques and technologies. For television, we present papers on NTSC ghost canceling; Advanced Television; and new TV engineering techniques and technologies.

We live in an age where the pace of technology development seems to accelerate every year. Our challenge is to decide, as an industry, which, if any, of these technologies we should embrace. Let NAB's *Proceedings* help answer these questions. It's a very exciting time to be in broadcasting.

Best regards,

Michael Rau

Michael C. Rau

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NEWS AUTOMATION AND MACHINE CONTROL: THE MARRIAGE OF JOURNALISM, PRODUCTION AND ENGINEERING

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Introduction

With an increasingly competitive market, the need for lowering costs while maintaining quality and accuracy of news broadcasting has become of paramount importance. The broadcast industry has gone through many changes over the last few years. There is an ever increasing number of choices offered to the viewing public, such as the diversity of cable programming, pay-per-view and video tape rental. As the "pie" of television revenues is being cut into many more pieces than ever before, the end result is a reduction in each station's proportional share.

Competition for the viewer has never been higher. For the broadcast news operation, quality of programming, accuracy of information and the importance of being first with a story has become even more important.

Machine control systems have recently been developed as a tool to help news organizations meet the goal of doing a better job for less. In conjunction with newsroom computer systems, many are exploring the benefits of expanding the role of the computer in news gathering and production to also include control of on-air production devices.

With computers extending their influence over the entire spectrum of news coverage, many mundane tasks can be handled automatically with greater accuracy, giving more time to focus on overall quality and lowering overall operating costs.

Traditional Staffing Duties

Traditionally, there are clearly defined boundaries surrounding the various departments involved in broadcasting. The News department is responsible for gathering and reporting the events in a timely, accurate and informative manner. They rely upon the Production and Engineering departments to bring their stories to life on the air. Video tape, graphics, titles, special effects, cameras and lighting are usually handled by the Production department. The Engineering department is responsible for the on-air operation of production devices and maintenance of all electronic systems associated with the broadcast organization. The producer and director are there to make sure that all of the elements come together and stay together during the dynamics of a news broadcast.

Machine control systems are re-defining the roles of these departments. For example, these systems allow a reporter to input character generator information into his script, and this text is transferred automatically into the production device. The producers can query the graphics system from their newsroom terminals, assign pictures to scripts and automatically build these into a playlist of pictures for the broadcast. The status of video tapes can be monitored automatically from the newsroom. With the advent of systems and communications to offer this kind of automation, a closer association between departments becomes necessary.

The Newsroom Computer System

Computerization in broadcast news is by no means a novel concept. For over ten years, newsroom computer systems have streamlined the news gathering process for journalists. These systems receive information from wire services, file and distribute it. Word processing features have eliminated the use of typewriters. Database features rapidly put information at the desk of whoever needs it. Seamless connections from remote sites and to other data sources offer writers a world of research material. Aired scripts can be fed to electronic teleprompters and closed caption systems, allowing story revisions, additions or deletions. Producers can re-organize the show on-air, and have all the timing parameters computed automatically.

This all adds up to faster access of vital information, greater accuracy and more efficient use of people's time. Everyone who works with a newsroom computer system shares a common communications, information and production tool.

Machine Control Systems for News

By expanding the scope of the newsroom computer system to include communication with production devices, the same benefits of increased efficiency, accuracy and lower overhead offered to the news department can be put forward to both production and engineering departments.

With machine control systems, the newsroom computer system becomes the central source of all information regarding the news broadcast. Anyone can, for instance get a current listing of correctly sequenced video tapes required for a particular show. An order for Character Generator titling is never re-typed, or lost, and can be immediately transformed into a C.G. super.

All video elements for a particular story are associated with that story in one place. As the story changes, all changes are made in one system, eliminating discrepancies in story versions. As the order of stories is changed throughout the broadcast, the sequence of elements in the production devices is automatically kept in sequence with the producer's rundown of events.

Machine control systems also offer logical, universal playback interfaces for production equipment. These systems allow for playback control of many different devices from a single control position. This allows for greater staff overlap between device types. For example, the same operator could operate the C.G. and electronic still store for the 3:45 cut-in, then run videotape playback for the 6:00 show from the same location in the control room.

Current Machine Control Installations

Notable European broadcasters have taken significant steps toward automating the news broadcasting process. Swiss Radio and Television began controlling a Sony Betacart videotape machine from their newsroom computer system in 1987. They were followed in short order by TV2 in Finland. Independent Television News in the U.K. is approaching news broadcast automation with a total systems approach. They are re-tooling their entire operation to take advantage of integrated newsroom computers, a machine control system and resource management systems.

SKY Television, Europe's 24-hour news operation began automation of tape playback in early 1988 and has expanded its control of production gear to include a network of Chyron Scribe character generators and a Quantel DLS graphics system.

KYW Television in Philadelphia has recently installed a machine control system to interface their newsroom computers to Chyron 4100EXB character generators, Quantel DLS 6000 electronic still store, a Sony Betacart machine and eventually to a robotic camera system. Acceptance of these changes by both management and staff at KYW has been very positive. Their adaptation to machine control required only minor changes to standard operating procedure and has streamlined the news process considerably.

Overcoming the Obstacles

Machine control systems automate considerable amounts of work once handled by people in the Production and Engineering departments.

The newsroom staff becomes directly responsible not only for the accuracy of the scripts but for the visual content as well. This gives them greater responsibility as complete story managers, and gives them the tools to control all elements within the news broadcast.

The production department is no longer responsible for the data entry of C.G. text, still store stacks, cart machine playlists or camera and lighting cues for the news broadcast. This provides more time for these highly trained individuals to concentrate on the overall production, as well as time to create the subtle nuances which make the program one of higher quality. They have more energy to devote to the creative, because less time is devoted to information management.

The engineering staff can afford to have fewer highly skilled operations engineers, as playback control of production equipment becomes easier. This allows for greater staff overlap, reducing operating costs. Also, the newsroom computer and machine control system become an integral part of the broadcast equipment, bringing these into the realm of the maintenance engineer.

Fears and Anticipation of Further Automation

It is inevitable that as computer systems in broadcasting become more advanced, our reliance upon them will increase. Many steps are being taken to standardize computer network protocols to pave the way toward a more automated "Total Systems" approach. Automation will never replace the creative elements of a process, only reduce the mechanical, the repetitious, the mundane. There are obviously financial benefits to automating these tasks, but the real benefit comes from offering more time for fewer people to do a better job with less effort.

There are many considerations to be pondered as computer systems build bridges across departmental gaps. Obviously, these systems must be easy to use. Also, the more associated any computer system gets to the actual on-air product, the higher the degree of reliability it must maintain. How will the unions adapt to these changes? Consider the increased stress factor when machine control systems put more power in the hands of fewer people.

Summary

With computer systems diminishing some of the complexity involved in the broadcaster's daily routine, a better program is inevitable, with much greater information accuracy and a lower cost. Despite the negative implications that come with further computerization , machine control systems offer exciting new opportunities for news broadcasting.

A TECHNICAL DIRECTOR'S WORKSTATION— THE FINAL INTEGRATOR

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INTRODUCTION

The ever-increasing demand placed on broadcasters to maximise the utilisation of studio equipment, whilst ensuring optimum staffing, has accelerated the acceptance and implementation of automating the control of broadcast equipment.

During the past two years, the broadcast industry has experienced a sudden surge of hardware and software products intended to provide the broadcaster with the ability to automate the control of selected items of equipment. In particular has been the increased functionality of newsroom computing systems which now integrate the control of news presentation equipment with the programme rundown schedule. Similarly, major advances in broadcast studio camera robotics have allowed camera positional control to be driven by instructions embedded within the rundown. The overall effect of this high degree of automation has been the response time necessary to incorporate fast-breaking news stories has been improved, with the added benefit of reducing the manpower required to control the equipment.

Up to the present time, significant developments have been targeted towards the control of equipment used for front line programme production. However, little attention has been given to automating the control of second line equipment, not usually directly associated with on-air control. It is believed that this situation will change and due consideration will be given to both areas of operation in order that a fully integrated approach can be implemented. As the current evolution in automation continues, it will be necessary to extend the options available. This paper describes the concept of control that can be achieved from a Technical Directors workstation, providing centralised control and supervision of a wide range of broadcast equipment used for programme preparation and production.

THE CONCEPT

With the proliferation of different products available today, one of the major problems facing the Technical Director of the 90s will be that associated with the issue relating to centralised control. It is a recognised fact that each individual item of equipment currently has its own control panel supported by operator display heiroglyphics, unique mode of operation and, in many cases, its own visual monitor display.

The result is a myriad of control panels taking up considerable and vital real estate on the console and demanding an intimate knowledge of their operation. It is logical to assume that this situation could be considerably reduced by the concentration of input and activity status-output control functions. This final integration process will be achieved by the Technical Directors workstation. The workstation will allow control access to be easily and directly achieved, enabling the Technical Director to easily assimilate status and provide the necessary monitored feedback.

Complex operator commands can be met by the introduction of touch screen technology for control input, coupled with the extensive use of ICON displays for status and activity confirmation. It will also be necessary that system functionality would include the capability to enable the selection of single input instructions to generate multiple commands to any number of related and interconnected devices. Although a phased approach to total integration of control automation would be the conventional path to follow, with the control of routers and VTRs being given highest priority, it is arguable that importance should also be given to extending this control to include a complete range of miscellaneous control room systems, ranging from monitor displays to frame synchronisers, each achieved with similar ease.

TECHNICAL DIRECTOR WORKSTATION

The Technical Director Workstation comprises a dedicated computer which has direct access to a high speed data interchange network, structured to allow real time operation between the Host Computer, Workstation Terminal and broadcast equipment. A number of Workstation Terminals will be capable of communicating with the Host Computer, thus providing regulated access by other key broadcast personnel. The Host Computer will interface with Workstation Terminals via an Ethernet, Local Area Network (LAN). Connected to the LAN will be a number of Machine Control Units (MCUs). Each MCU is identical with regards to hardware configuration and is characterised within the network by its resident software configuration. The software configuration or machine driver software identifies the specific machine under the direct control of the host. Examples of these machines would include:

Cart machines
VTRs
Routers
Intercom systems

A typical system configuration is shown by Figure 1 which illustrates the interconnection by the Ethernet of the three component parts:

(a) Host Computer
(b) Workstation Terminals
(c) Machine Control Units

Host Computer

To meet the control demands of the average studio, the Host Computer would require the power of a DEC VAX 3100. If, as is often the case, a high degree of fault tolerance is required, two VAX 3100s would be configured to operate in a main and standby mode.

Workstation Terminals

A Workstation Terminal would be an IBM or equivalent 286 PC, equipped with a high quality colour touch screen overlay. The use of hard disc storage can be avoided as the run time software is downloaded from the Host Computer.

Machine Control Units

At the heart of a Machine Control Unit are two 68000 microprocessors, configured to provide interfacing control via four serial i/o ports and eight parallel lines. The typical configuration is illustrated by Figure 2.

Application software for the MCU is held on the Host VAX and downline loaded on request or during the initialisation process. This design concept allows numerous machines to be controlled by identical MCUs. Connection to the Ethernet allows a number of MCUs to be distributed throughout the studio. External connections by third parties can be made to the Ethernet in order that status monitoring can be achieved. However, such access would be restricted so as not to affect system response time. Use of Ethernet bridges would assist in access restriction and security.

Having now outlined the concept and necessary hardware and software, the key issue is how such a system can meet the demands imposed by controlling numerous items of equipment. For the purpose of this paper, two high level and two second line devices are discussed in detail.

The Technical Director workstation has been orientated around the use of a touch screen input supported by a keyboard. The keyboard is used to set up text files and to enter instructions which are not required to initiate time critical activities. Each item of broadcast equipment to be controlled is allocated a defined icon profile. Controlled access of each item of equipment is achieved by touching the selected icon displayed within the icon menu. When control is passed to the Technical Director, the screen is refreshed by a new display which is orientated towards displaying the device under control, with a graphical representation of the functionality now available to the Technical Director.

Input Assignment Routers

News and Sports programmes require an ever increasing number of external lines to be fed into the studio, these lines all arriving at the main input matrix. Economic considerations, coupled with practical realisation, often result in the associated vision mixer not being able to accommodate all the incoming lines simultaneously. In order to handle such conflicting situations, it is necessary to pass external lines through a pre-assignment router. This would be accomplished in a way that at any point in time a pre-selected number of lines can be presented to the switcher.

By selection of the pre-assignment switcher icon at the workstation, the Technical Director will be presented with a visual display of all the input lines available, each source being boxed and identified, examples being "LANDLINE 1-WHITE HOUSE" or "O.S.6-LONDON", the text having been previously entered via the keyboard, typically coupled with the ancillary information which may include a correspondent or interviewee's name.

The presentation of crosspoints available enables the Technical Director to directly select or deselect followed by select of the appropriate crosspoint selection. This action being achieved by touching the relevant positions on the screen.

A database can be constructed of sources available to the Technical Director. This would reduce the amount of keyboard work required. Such a shared database reduces the need for each operator to input identical information.

Video Tape Recorders

Similar to the initial stages of control for the router, the first process in selection is by touching the VTR icon. This in turn presents the Technical Director with a graphical display of VTRs available for selection, hence allowing the Technical Director to select the required VTR icon. At this point, displayed on the screen would be a pictorial representation of the VTR controls made available to the Technical Director. Figure 3 illustrates the format displayed to the Technical Director. Control is exercised by touching the relevant box. Typical functions include Play, Fast Forward, Rewind, Record, Stop, Jog, Eject, In/Out Cue points

This method of control is applicable to either single or multiple configurations of VTRs.

Facilities also allow both serial and parallel recording to be achieved, enabling recording of both long items or even multiple recording.

Second level systems can include Monitor Designation Displays. Monitor stacks in both News and Sports studios are becoming unmanageable in both size and number as an ever-increasing number of sources is being made available for programmes. In many cases, control rooms now include multi-character alphanumeric LED displays under each monitor to identify the source. In the case of monitors showing external sources, it is of significant importance that a true identification is shown at the monitor. This requires that when the source is deselected and replaced by another source, the system is sufficiently dynamic to follow the change and update the associated monitor LED.

An important aspect of the Technical Directors workstation is that single input commands result in simultaneous instructions being passed to all affected equipment and displays. When an external source is routed into the studio, it is often required to simultaneously route intercom channels to provide single or two way communication. This presents yet another case where the Technical Directors workstation must know and issue the relevant switching commands to the studio's intercom system such that the required channel can be correctly configured. In addition to passing the necessary switching commands, the system would generate the abbreviated display information that would be used by the intercom system if fitted with LED displays.

FUTURE

This paper has outlined the concept of a Technical Directors Workstation capable of undertaking the role of a centralised control point for a Technical Director. It is recognised that the movement and evolution into fully integrated control will take time and will, in many cases, be as the result of a phased approach. However, the requirements of a Technical Directors Workstation will also evolve and be influenced by implementation. Of prime consideration will be the continuity of display which is in essence irrelevant to the machine under control but vital to the Technical Director. The approach described is based on hardware that is independent of the machine under control, with characterisation of the machine controller being held in software. This approach allows new machines to be readily interfaced and will be of special interest when second level machines such as Frame Synchronisers and Dish Positioning Systems become incorporated with automation control.

FIGURE 1

TECHNICAL DIRECTORS WORKSTATION

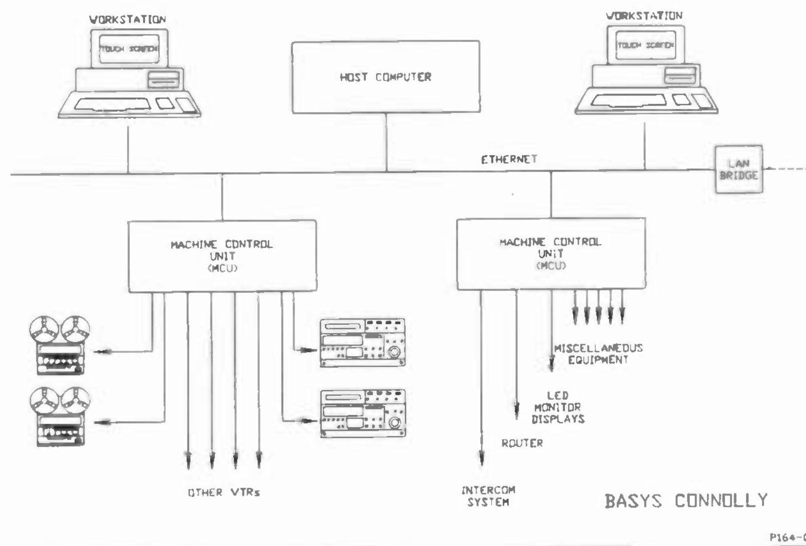
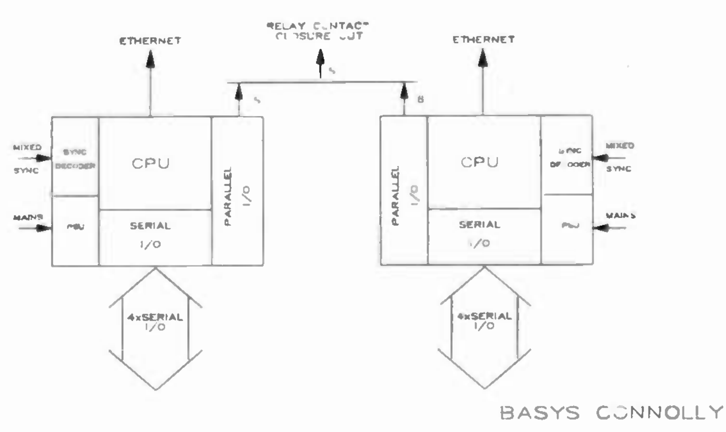


FIGURE 2

MACHINE CONTROL UNIT



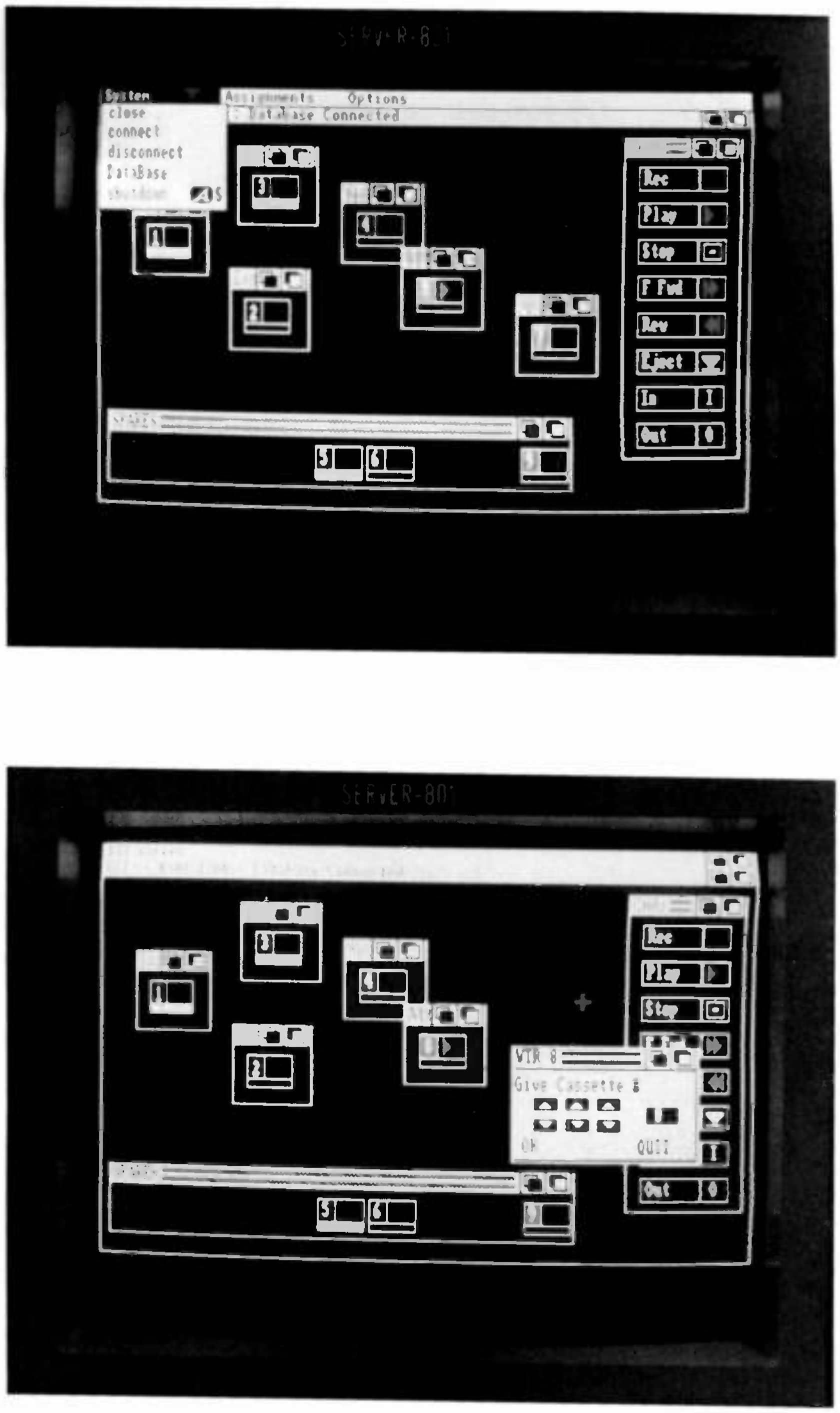

System
close
connect
disconnect
Assignments
Options
Database Connected
Rec
Play
Stop
F Fwd
Rev
Eject
In
Out
SERVER-801
VTR 8
QUIT

FIGURE 3

RECENT ADVANCES CAUSE INCREASE IN CAPTIONING OF LOCAL TV NEWS

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Madison, Wisconsin

ABSTRACT

Because of community pressures and competitive forces, closed captioning of local news is a subject of great current interest to TV station executives. But as recently as August of 1987, fewer than one dozen affiliate TV stations were offering closed captioning of local news for their hearing-impaired audience. This is in direct contrast to the high proportion of closed captioned programs available in national network broadcasts.

The reason for such low numbers was because the only methods available then were either too expensive or operationally unacceptable. But recent advances in newsroom automation technology have caused a dramatic increase in the number of stations now offering local news captioning.

This paper will review all methodologies currently available for the captioning of local news, giving the advantages and disadvantages of each. It will also explain why captioning via electronic newsroom systems has become so widespread. Finally, it will cover the actions that a station should take when it commences closed captioning its local news.

HISTORY OF CLOSED CAPTIONING

For most of us, it is almost impossible to imagine a world without sound, but millions of deaf Americans live in silence. That silence was at least partially broken in 1980 with the launch of the national closed captioning service, making television accessible to deaf and hearing-impaired people. The impact of closed captioning on the lives of deaf people over the last ten years has been dramatic, because for them, it has brought meaning to the most important medium of our time—television. The tremendous growth in the availability of closed captioned programming since 1980 confirms the commitment of the broadcast industry to this segment of the television audience.

The precursor to closed captioning was open captioning—captions visible to all viewers—a service introduced in 1972 by The Caption Center[1] in Boston, Mass. The first open captioned news program, *The Captioned ABC News*, premiered in 1973, giving the hearing-impaired audience access to national television news for the first time. The captioned version was rebroadcast each evening on PBS five hours after the newscast aired on ABC.[2] At the time, this was the best that technology had to offer, and the delayed captioned version of the program provided a valuable service to the audience.

The concept of closed captioning—captions "hidden" in the television signal until decoded by an adapter on a television set—originated in the 1970s at ABC, thanks to the leadership of then-owner and chairman, Leonard Goldenson. A system which had just been developed by the National Bureau of Standards for transmitting time and frequency information in the vertical blanking interval[3] led two ABC executives, Julius Barnathan and Leonard Maskin,[4] to speculate that a similar system could transmit captions for deaf people. They envisioned a system that would allow hearing-impaired viewers to watch captioned programming through a decoder while allowing hearing viewers to watch the same programming without captions.

Shortly thereafter, with support from the U.S. Department of Health, Education, and Welfare (HEW), development of the line-21 closed captioning method began, with significant cooperation from ABC and PBS. In 1976, the FCC reserved line 21 for the transmission of closed captioning data. In 1979, with seed money from HEW, the National Captioning Institute (NCI) was established as a private, non-profit organization to provide a national closed captioning service and ensure its permanence in the television industry. To fulfill its mission, NCI sells the captioning service to the television industry, obtains Federal and private funds to underwrite captioning costs, and manufactures and distributes the TeleCaption decoder.[5] NCI now operates as an independent non-profit corporation which supports itself through the sale of captioning services and decoders.

In 1980, closed captioning premiered with 16 hours of captioned programs airing each week on ABC, NBC, and PBS. At that time, technology allowed the captioning of programs only when the captions could be prepared in advance of air. This "off-line" method of captioning is still used today on prerecorded programming. Caption editors, watching a 3/4" work cassette of a program with timecode, use a computer to enter the program dialogue as captions, editing the dialogue when necessary to present the captions at a readable speed. Each caption is assigned an *appear* and *disappear* time based on the timecode. The captioning data is stored on a floppy disk. Before the program airs, the data is encoded onto line 21 in the vertical blanking interval of the master tape. (The encoding process requires a special piece of equipment, called an encoder, manufactured by EEG Enterprises.[6]) The captioning data is then broadcast as part of the program video, but the captions are visible only when decoded by a TeleCaption decoder.

In 1981, The Caption Center became a second provider, after NCI, of closed captioning services to the television industry. Closed captioning quickly developed a loyal following in the deaf community, and the service grew. The audience most frequently requested that television movies and national news programming be captioned, according to NCI. Recognizing the need to make national television news and other live events accessible to hearing-impaired viewers, NCI then developed "real-time" captioning technology, which allows captions to be produced almost simultaneously as words are spoken. Real-time writers originally trained as court reporters phonetically transcribe program audio into a computer which translates the phonetic symbols into English words. In 1982, ABC's *World News Tonight* was the first daily program ever to be real-time closed captioned.[7]

With real-time captioning technology a reality, it was now possible to closed caption live television programs such as news, sports, and talk shows in addition to the taped programs captioned with the off-line method. Beginning in 1982 and continuing today, presidential speeches and press conferences, special news reports, and the Academy and Emmy Awards are captioned using the real-time method. By the mid-1980s, the mix of captioned programs grew more varied. Important nationally televised events, such as space shuttle launches, the 1984 Winter Olympics, the 1984 political conventions, and primary election coverage were captioned for the first time.[8] CBS began providing line 21 captions on a variety of programming, joining the other two commercial networks. The home video industry became a major supporter, with studios funding the captioning of popular home video titles. Thousands of commercial "spots" had been captioned. HBO and Showtime provided captioned programming. Private funders of closed captioning, primarily the networks, advertisers, producers, and corporations, recognized closed captioning for its ability to enhance corporate public image. As closed captioning came of age, the hearing-impaired audience focused its loyalty on those organizations funding closed captioning,[9] further reinforcing the value of the service to its supporters.

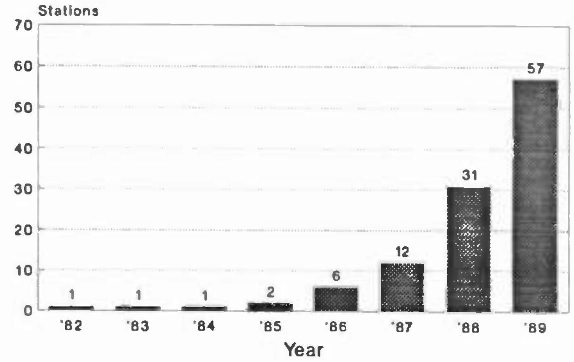
Even more captioning milestones were achieved in the late 1980s. In 1987, *Wheel of Fortune* and *Jeopardy!* became the first game shows captioned. The amount of captioned syndicated programming increased, and in 1988, the *Oprah Winfrey Show* became the first talk show captioned with the real-time method. By early 1988, all three network evening newscasts were captioned, giving hearing-impaired viewers true equal access to national news. Later that same year, a grant to fund the captioning of thousands of hours of sports programming over a three-year period was awarded by the U.S Department of Education. The grant money is combined with private funds to allow the captioning of various sports events including basketball, football, golf, baseball, and bowling. Basic cable programmers and operators joined the list of captioning supporters. At the start of the 1989-90 television season, for the first time ever, the entire prime-time schedule on all three networks was closed captioned.[10]

LOCAL NEWS CAPTIONING—BACKGROUND

1988 brought another captioning trend that continues today: the explosive growth of local news captioning. At the start of 1988, only 12 television stations closed captioned one or more newscasts each day for their hearing-impaired viewers.[11] By January 1, 1990, that number had grown to 57—a 375% increase in just 24 months.[12] A number of factors have contributed to the recent growth of local news captioning, including community pressure, station recognition of the public service value of captioning, competitive considerations, and most importantly, the development of electronic newsroom system captioning.

By early 1988, the hearing-impaired audience could choose between all three network evening newscasts, so the demand for captioned *national* news had been satisfied. However, this milestone in national news made the lack of captioned local news even more obvious. Viewers could finally understand what Dan Rather (or Peter Jennings or Tom Brokaw) was saying about the Soviet Union, but they

U.S. TV Stations
Captioning Local News



couldn't understand their local anchor describing the robbery that occurred down the street. In a survey of decoder owners conducted by NCI in 1988, 100% of the respondents indicated they would watch a captioned local newscast.[13] According to NCI, local news captioning was in demand by the audience, and beginning in 1988, hearing-impaired groups in many cities launched grassroots efforts to demonstrate the need for captioned local news to their television stations. The television stations, known for their community spirit, listened.

Closed captioning holds particular appeal to television stations, since they are interested in projects which demonstrate their commitment to the community. Closed captioning had already proven itself at the national level as a tremendous public service and an image enhancer to those providing it. Television station general managers were interested in taking advantage of this same opportunity to serve a new and loyal audience, and reap the rewards of providing such an important public service. In a few markets, captioned local news was already a reality, so competitive pressure also inspired some stations to consider matching their competitor's offerings.

The first station to provide captioned local news was KCTV in Kansas City, Mo. in 1982. From 1982 until the beginning of 1988, the number of stations offering captioned local news had only grown to 12.[14] Late in 1987, however, tremendous growth in local news captioning began because of the availability of electronic newsroom system captioning. Suppliers of electronic newsroom systems began selling an option to their systems that interfaced the electronic teleprompter to the EEG encoder, resulting in automatic captioning of the scripted portions of the newscast. Due primarily to this new captioning method, the number of stations offering captioned local news soared. Nineteen stations were added to the list in 1988, and 26 were added in 1989, bringing the total to 57 by the end of 1989.[15]

Electronic newsroom system captioning has been responsible for the tremendous growth in local news captioning because it eliminates problems inherent in other captioning methods. This can best be explained by a description of all captioning methods currently used for local news captioning and the relative advantages and disadvantages of each.

LOCAL NEWS CAPTIONING METHODS

There are four primary methods for captioning local news. These are *live-display, real-time, electronic newsroom,* and *hybrid* captioning. All are discussed below.

Live-display Captioning

Several companies manufacture software which interfaces a microcomputer to an EEG encoder. Program scripts are transcribed before air time into the microcomputer. The text is then manually sent by an operator, one line at a time, to the encoder as the program airs. This captioning method is most effectively used on programming which is scripted in advance and does not change once the text had been entered.

Advantages

- Cost: The live-display captioning system is relatively inexpensive, with software costs generally in the $5,000-$10,000 range.
- **Timing:** Captions for the scripted portion of the program will appear in sync with the audio.
- **Accuracy:** Depending on the skill of the operator, the captions should be complete and spelled correctly.

Disadvantages

- **Coverage:** Any unscripted portions of the newscast will not be captioned.
- **Flexibility:** The system offers no flexibility to change the text while the newscast is on the air. Any last minute changes to the script before air time must be re-entered into the captioning system.
- **Efficiency and Personnel:** Depending on the type of live-display system and the type of teleprompter, this captioning method may cause a duplication of effort. That is, the script might be typed once for the paper tray teleprompter, then typed again into the microcomputer. At least one person must be dedicated in the hours preceding the newscast to entering the final script into the live-display system. One person must also be dedicated to "punching" the captions line by line during the newscast.

Industry Acceptance

Of the 57 stations offering captioned local news, today four are using the live-display method.

Vendors

Live-display systems are offered by these companies:

The Caption Center
WGBH Educational Foundation
125 Western Avenue
Boston, MA 02134
(617) 492-9225

CaptionAmerica
312 Avenue of the Allies
Suite 200
Pittsburgh, PA 15222
(412) 261-1458

EEG Enterprises
1 Rome Street
Farmingdale, NY 11735
(516) 293-7472

Real-time Captioning

With real-time captioning, captions are created as words are spoken. This method is sometimes referred to as online captioning. As a program airs, a specially trained real-time writer uses a stenographic keypad (commonly used in courtrooms) to key in words phonetically as they are spoken. A real-time writer can transcribe audio at the rate of 250 words per minute or more—much faster than even the best typist. The steno keypad is connected to a computer which translates the phonetic shorthand into English words and arranges the words into captions. To accomplish this translation, the computer searches its dictionary to find a match between the shorthand and a word. If no match is found, a mistranslate will occur, and a phonetic representation of the word will appear in the caption instead of the intended word. The computer sends the captioning data to the encoder, and the captions appear on the screen two to three seconds after they are spoken.

Real-time writers must first be skilled court reporters. They require additional training and/or practice to acquire the accuracy of transcription required for real-time captioning. To help prevent words from mistranslating, the real-time writers must spend time each day before the newscast entering proper names and unique words that might be included in the newscast into the computer's dictionary.

Advantages

- **Coverage:** All portions of the newscast, live and scripted, can be captioned.
- **Flexibility:** With this method, changes to the newscast do not affect the captioning procedure. However, information regarding program content should be provided to the real-time writer in advance of air.

Disadvantages

- **Cost:** The cost of real-time captioning involves one-time hardware and software costs for the real-time equipment, and personnel costs of the real-time writers. Real-time captioning equipment configurations range from $15,000 to $35,000. Personnel costs for court reporters are high, and they vary from city to city. It is reasonable to expect that the cost to real-time caption one half-hour newscast a day for one year will be $90,000 to $100,000.
- **Personnel:** Real-time captioning requires highly skilled court reporters. Finding court reporters with the potential to develop into good real-time writers can be a major obstacle. It is often difficult to locate court reporters in a particular city who are willing to give up full-time positions in order to fulfill the time requirements of preparing for and captioning an early evening newscast. Once interested court reporters have been identified, there is no guarantee that they will have the skills needed to be good real-time writers.
- **Timing:** The captions appear on the screen two to three seconds behind the audio. This can be confusing to the viewer.
- **Accuracy:** Because of the phonetic nature of the system, words will mistranslate from time to time. When this occurs, the captions are difficult to decipher, and the viewer is distracted.

Industry Acceptance

Of the 57 stations offering captioned local news, eight are using the real-time method. (One additional station offers both real-time captioning and electronic newsroom captioning of different newscasts.)

Vendors

There are two types of vendors: those providing *equipment* and those providing the *service*.

Service Providers—The organizations that supply real-time captioning services to national programming have varying levels of involvement in local news captioning. They are:

The Caption Center
WGBH Educational Foundation
125 Western Avenue
Boston, MA 02134
(617) 492-9225

Caption America
312 Avenue of the Allies
Suite 200
Pittsburgh, PA 15222
(412) 261-1458

Captions, Inc.
2619 Hyperion Avenue
Suite A
Los Angeles, CA 90027
(213) 665-4860

National Captioning Institute, Inc.
5203 Leesburg Pike
Suite 1500
Falls Church, VA 22041
(703) 998-2400

Real-Time Captioning, Inc.
7107 Sepulveda Boulevard
Van Nuys, CA 91405
(818) 376-0406

Equipment Providers—The following companies supply real-time captioning equipment:

Xscribe Corporation
6160 Cornerstone Court East
San Diego, CA 92121
(619) 457-5091

Stenograph Corporation
1500 Bishop Court
Mount Prospect, IL 60056
(312) 803-1400
(800) 323-4247

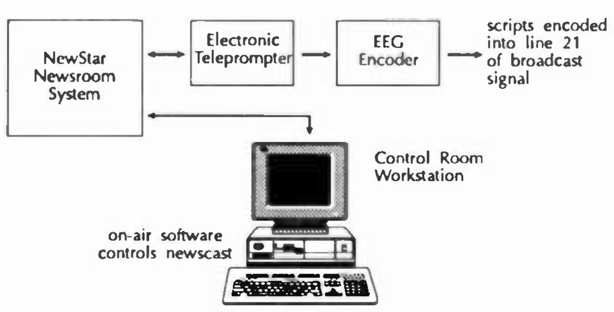
Digitext, Inc.
325 East Hillcrest Drive #250
Thousand Oaks, CA 91360
(805) 495-3456

Electronic Newsroom Captioning

Electronic newsroom captioning is a by-product of the technology that is revolutionizing television newsrooms. Electronic newsroom systems, also referred to as newsroom computer systems or newsroom automation systems, make newsrooms more efficient. Some systems are simple, consisting only of script writing software with the ability to communicate with a teleprompter. The larger systems are more versatile and flexible, helping stations gather, process, produce, and archive the news.

In a typical newsroom system, information is accessed from central data files by workstations available to each person. Wire service copy can be accessed by all workstations, and the system will sort and search the wires as directed. Communication is improved through the electronic mail feature. The rundown, or the "plan" for the newscast, is written on the system, and it can be updated and changed at any workstation authorized to make such changes. Scripts are written, approved, and inserted into the rundown. The rundown and script copy can be changed easily before and even during the newscast. With interfaces to electronic teleprompters, character generators, still store machines and robotic cameras, newsroom systems can provide total automation of the newscast, controlling all these devices needed during the newscast.

Closed captioning is accomplished as a by-product of such on-air newscast automation. When the closed captioning option is added to the system, scripts are sent from the



system to the electronic teleprompter *and* to the encoder, with the timing of the captions controlled by the teleprompter or by a control room workstation.

In order to produce high-quality closed captions, the newsroom system captioning interface must perform the following functions[16]:

- Transmit the scripts to the encoder, but delete talent cues and other internal information that appear within the scripts on the teleprompter.
- Allow for information in the scripts to be coded to prevent transmission if desired.
- Begin the transmission of captions for each new story as the anchor begins to read that story.
- Transmit the captions so that they are timed to the speaker's delivery.
- Allow captions to be created and transmitted for video segments used in the newscast.
- Allow messages to be displayed saying "This segment not captioned."

Often, newscasts contain story segments which are not read from the teleprompter—a package video piece, for example. Many stations that have implemented electronic newsroom system captioning require reporters to enter the scripts for these segments so that they will be captioned.

Advantages

- **Cost:** The incremental cost to add the electronic newsroom captioning interface is low, and it is a one-time charge. (Typical cost for the NewStar option is $15,000.)
- **Timing:** Since the captions are driven by the teleprompter, they are timed to the speaker's delivery and are in sync with the audio.
- **Efficiency and Personnel:** With the exception of script writing for segments normally not included on the teleprompter, closed captioning with this method creates no extra work for newsroom staff. Additional or dedicated staff to accomplish the captioning is not needed, since the captions are a by-product of the electronic newsroom system's newscast production process.
- **Flexibility:** With at least one of the larger electronic newsroom systems (NewStar), closed captioning in no way prevents or affects the producer's ability to change the newscast at any time, even when the newscast is on air. Using simple keystrokes on the control room workstation, stories can be moved or changed instantly. The changes are automatically sent to the teleprompter and to the captioning encoder. Closed captioning does not inhibit the flow of newscast production or of changes made to the newscast.
- **Archiving:** A record of package stories is created when the video segments are scripted for closed captioning. This benefit to the station is a by-product of the captioning process.
- **Accuracy:** The captions will be complete and accurate, and, depending on the care taken by the anchor or reporter, spelled correctly.

Disadvantages

- **Coverage:** Live segments of the news and stories completed without enough time to enter scripts will not be closed captioned.

Industry Acceptance

Of the 57 stations offering captioned local news, 42 are using the electronic newsroom method. (A few stations offer both electronic newsroom captioning and real-time captioning of the same newscasts. This is the Hybrid method.)

Vendors

It is important to note that electronic newsroom capabilities vary with the system and the vendor. The following companies claim to offer closed captioning options to their newsroom systems:

Dynatech Newstar, Inc. 6400 Enterprise Lane Suite 200 Madison, WI 53719 (608) 274-8686	**BASYS, Inc.** 5 Odell Plaza Yonkers, NY 10701 (914) 376-4800
SISCOM 100 Arapahoe Avenue Suite One Boulder, CO 80302 (303) 449-0442	**Columbine Systems** 1707 Cole Boulevard Golden, CO 80401 (303) 237-4000
Data Center Management (DCM) 1017 Kenilworth Avenue Charlotte, NC 28204 (800) 367-4312	**Computer Engineering Assoc.** 3922 Vero Road Baltimore, MD 21227 (301) 247-4000

Hybrid Captioning

Two stations, KARE and KYW, have implemented a hybrid method of captioning which combines electronic newsroom captioning and real-time captioning in one newscast. During prescripted portions of the newscast, the electronic newsroom system sends the captioning data to the encoder. During live or unscripted portions of the newscast, the real-time system sends the captioning data to the encoder, with captions generated by a real-time writer. The changeover can be accomplished with a simple A-B switch in the control room, or by a more sophisticated software interface between the real-time or electronic newsroom systems. Such interfaces are available from some electronic newsroom and real-time equipment vendors.

Advantages

- All parts of the newscast are captioned, with little or no additional work for the news staff.

Disadvantages

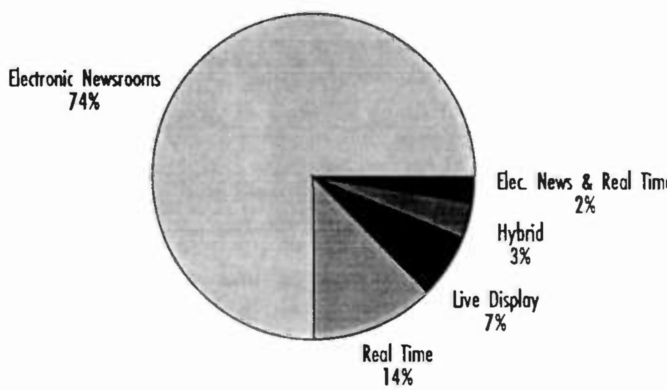
- The cost is generally high, as with real-time captioning. Even though the real-time writer has less work, he or she still must be there for the entire newscast and the preparation period before the newscast.

SUCCESS OF THE ELECTRONIC NEWSROOM SYSTEM METHOD

As of this writing, the four primary methods described above are being used with widely varying frequency, as depicted in the chart that follows.

The reason for the success of electronic newsroom system captioning of local news is that it offers solutions to problems encountered with other captioning methods. Live-display captioning, while relatively inexpensive, usually

News Captioning Methods



requires scripts to be typed twice—once for the live-display system and once for the teleprompter. This is clearly extra work for the newsroom staff. And, once the newscast is on the air, the system offers no flexibility to change the captions, even when the newscast changes. Live-display also requires dedicated personnel prior to and during the newscast. Electronic newsroom captioning overcomes all of these problems. Scripts are *created once* on the system by the reporters and anchors, revised throughout the day, then *fed automatically* to the teleprompter and the encoder during the newscast. Additional staff is not needed, and the captioning creates no extra work for the newsroom. While on the air, some systems permit the newscast to be changed by manipulating the rundown, with the captions automatically and instantly changed as well. Such electronic newsroom systems gives television stations the ability to caption without eliminating their ability to make last-minute or on-air changes to the newscast.

Last-minute changes can also be accommodated by the real-time captioning method, with the captions following the changes. However, the main problem encountered by stations with the real-time method is that it is very expensive. Real-time captioning was initially developed for national programming, with sources such as networks, network advertisers, and the Federal government funding the high cost. Real-time captioning is an important and viable service for national programming, but the expense is clearly an obstacle for the local TV station. Few can afford to pay $90,000 to $100,000 per year for captioning a one-half-hour daily newscast.

Some television stations have been successful in raising funds from the community to pay for real-time captioning, but fundraising can be a lengthy process. Also, some stations are reluctant to fundraise, fearful of replacing ad revenue with the pass-through captioning cost. Because of this financial obstacle, the U.S. Department of Education has offered grants to stations to help implement local news captioning. These grants ($50,000 per year for three years) have been awarded to more than 15 stations over the last four years. (The grants initially funded real-time local news captioning, but recent awards have been made for stations captioning with the electronic newsroom system method as well.)

Electronic newsroom captioning solves the cost problem. The incremental cost to add electronic newsroom captioning is low, and it is a one-time charge, versus a very high annual cost for real-time captioning. Furthermore, electronic newsroom captioning is accomplished automatically for all newscasts produced by the station, every day of the week, every week of the year. The live and unscripted portions of the newscast will not be captioned with the electronic newsroom system method, but roughly 75% of every newscast can be captioned. Electronic newsroom systems provide significantly more captioning for significantly less money.

Stations captioning with electronic newsroom systems can establish a number of procedures to ensure that the deaf audience receives as much information as possible during each newscast. For example, for package stories that are finished too late in the day to allow the entire script to be entered into the system, reporters can enter a short summary of the story or segment. This way, the audience will have a general understanding of the news event. Summaries can also be entered for sports and weather—two segments usually not scripted in advance. For live segments, it is helpful to the audience for the station to transmit a caption explaining that the segment is live and captions are not available. Practices such as these can increase the percentage of the newscast captioned, and help the audience understand why some segments are not available to them.

KEY IMPLEMENTATION POINTS

The closed captioning of local news provides a tremendous service to the hearing-impaired community—access to news and information about local events. By providing closed captioned local news, stations identify themselves as leaders serving the entire community. Captioning is most effectively implemented by stations when combined with a simple promotional campaign. Promotional activities such as those suggested below increase public awareness of the closed captioned newscasts and the station's commitment to the hearing-impaired audience.

Produce PSAs

Stations should produce one or more PSAs announcing the launch of closed captioned local news. The PSAs should begin airing approximately four weeks before the launch, then periodically thereafter. The PSAs should explain closed captioning and include information about the availability of decoders in the area. The National Captioning Institute encourages stations to include its toll-free numbers in all PSAs and printed materials. Viewers can call the toll-free number to learn where decoders can be purchased in their areas: (800) 533-WORD (voice) and (800) 321-TDDS (TDD).

Install TDD

A TDD, or telecommunications device for the deaf, will allow the station to take phone calls from callers using a TDD. The TDD features an acoustic coupler and a keyboard, and the phone message is printed out on an LED display or on paper. Contact telephone retailers in your area for more information. Whenever the station's phone number is printed, the TDD number should also be printed, followed by "(TDD)."

Internal Notification Of Captioning Launch

All station employees should understand the concept of closed captioning. They should be aware of the launch date and the TDD phone number. The receptionist should know where in the area decoders can be purchased.

Press Release

A press release announcing the launch of closed captioned local news should be distributed to the station's usual mailing list, in addition to organizations in the area serving the deaf and hearing-impaired community.

Print Ads

Any print ads produced to announce the launch of captioned local news should contain the "CC" designation, which is commonly used to denote closed captioned programs. The station's TDD number should also be included in the ad.

Special News Stories

During the week closed captioning is launched, the station should run special news stories each day featuring information about closed captioning and stories about the deaf community. The availability of decoders should be mentioned.

Advisory Committee

An advisory committee, comprised of leaders in the deaf and hearing-impaired community, can be formed by the station before or after the launch of its captioned local news. The formation of the committee will prove the station's commitment to the audience, and the committee can provide valuable feedback to the station regarding captioning procedures.

Station Tours

Deaf groups should be invited to the station during the week captioning is implemented for a tour of the newsroom and an explanation of the closed captioning procedure. This will help to include the audience in the closed captioning process and generate goodwill for the station.

Decoder Giveaway

A number of stations have purchased decoders, either on their own or with a corporate sponsor, to give away to needy families, usually at a media event such as a community reception. The decoder giveaway adds another dimension to the service provided by the station, and the station obtains further recognition for its efforts.

ENDNOTES

[1] The Caption Center is a non-profit service of the WGBH Educational Foundation.

[2] Dittman, Doug and Dan Glisson, Larry Goldberg. "Time-of-Air Closed Captioning for News Programs." ©1990, The Caption Center/WGBH Educational Foundation. (This paper is an excellent source of information on local news captioning.)

[3] Ibid.

[4] National Captioning Institute. 1989. Caption. Special Fifth Anniversary Edition. (Summer)

[5] For more information on the TeleCaption decoder, contact: National Captioning Institute, 5203 Leesburg Pike, Falls Church, VA 22041. Phone: 703/998-2400.

[6] For more information on the EEG Encoder, contact: EEG Enterprises, Inc., 1 Rome Street, Farmingdale, NY 11735. Phone: 516/293-7472.

[7] Okrand, Marc. 1985. "Real-Time Captioning: A Major Breakthrough." Falls Church, Va.: The National Captioning Institute.

[8] Ibid.

[9] National Captioning Institute. 1988. "The Awareness of Corporate Support Among Closed-Caption Viewers." Closed-Caption Bulletin. (November)

[10] National Captioning Institute. 1989. Caption. (Fall)

[11] National Captioning Institute. 1989. "Local Stations Providing Captioned Local News Broadcasts." (December)

[12] Ibid.

[13] National Captioning Institute. 1988. "The Closed Caption Viewer's Reaction to Captioned News Broadcasts." Closed Caption Bulletin. (November)

[14] National Captioning Institute, *supra* note 11.

[15] National Captioning Institute, *supra* note 11.

[16] Smith, L. Sanders. 1988. "Newsroom Automation Opportunities." In: Proceedings of the National Association of Broadcasters 1988 Engineering Conference.

APPLICATION OF LIBRARY MANAGEMENT SYSTEMS AT THE NEW CBS BROADCAST ORIGINATION CENTER

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CBS Inc.
New York, New York

CBS is currently installing a new automated master control facility at the Broadcast Center in New York. This paper will present the system design considerations for commercial playback systems capable of maintaining a large commercial inventory, and suggest some approaches for simplifying the operation. In our system, which will originate up to ten channels of programming with commercials, the job of scheduling and handling so many spots is immense. The present system of tape traffic will be described, and several future approaches given, which would significantly reduce the complexity of commercial handling.

Broadcast Origination Center

I will begin with a description of the Broadcast Origination Center (BOC) and the current commercial playback system using Library Management Systems (LMS). BOC will produce ten output channels from self contained control and playback facilities. BOC is designed to integrate a main program channel with different commercial streams for regionalized networks. Figure 1 shows how a main channel re-enters other channel switchers for insertion of different commercials. One typical example of regional commercials is to broadcast in the same break an advertisement for snow tires to the Northeast region network and an advertisement for all season tires to the Southeast region network.

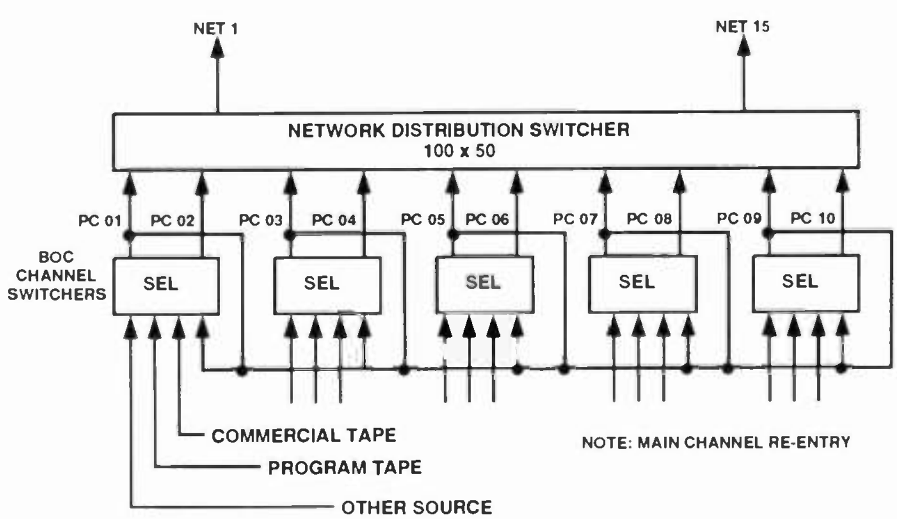


FIGURE 1

BROADCAST ORIGINATION CENTER

Within an individual BOC channel, two program outputs are fed from each switcher. The main output has full mix effects and key capability. Figure 2 shows how this main output is re-entered into the master control switcher for cuts-only insertion of separate commercials over the main program.

Program sources for BOC are 1" Type-C or D2 VTRs for program playback and LMS D2 commercial playback systems. Other input sources to the channel switcher are live studios, plant routing switchers, electronic still stores, and electronic character generators.

Commercial Tape Traffic

With this introduction to the system, next I will describe the path of tape traffic through the system. Both commercial and promotional spots are handled by the system. Spot material enters the system at dubbing stations which consist of a Type-C VTR connected for automatic transfer of spots to D2 cassettes.

Commercials and promos entering the system must be transferred to D2 cassettes. The dubbing process is automated as much as is possible. After the operator loads the one inch Type-C source tape, a computer controller begins a transfer cycle to record black leader on the LMS cassettes, then edit in the source material at a preassigned time code start time, recording one spot per cassette. Two cassette copies are made in each pass. After the dub, the computer recues the cassettes for the operator to view, one at a time, to verify the quality of the dub.

Spots are identified in the system by an identification number (ID) which gives a shelf position in our videotape library. A barcode label with this ID is printed and attached to blank tapes before they are recorded in the dubbing process. An inventory database matches the ID number with the commercial material, and tracks secondary identifying information for the commercials.

After the cassette spots are dubbed and checked, they are stored in the videopate library, until scheduled for broadcast. At that time a play list generated from the program schedule determines which cassettes are needed in each LMS. The tape librarian picks the cassettes scheduled, and a videotape operator loads them into the LMS. Remember that to provide a backup copy of each spot, two cassettes are picked in the video tape library and loaded into two LMSs. After use, provided a spot is not scheduled for fairly immediate reuse, the cassette must be purged from the LMS, and refiled in the tape library.

While the new playback systems with their large cassette inventories can greatly reduce the operators task, the preparation and handling of cassettes still requires considerable manpower and is vulnerable to human error. Figure 3 shows this system using manual tape transfer.

Advantage of Digital Transfer

Now I want to switch from the description of our present system to a new approach which will considerably simplify the commercial playback process.

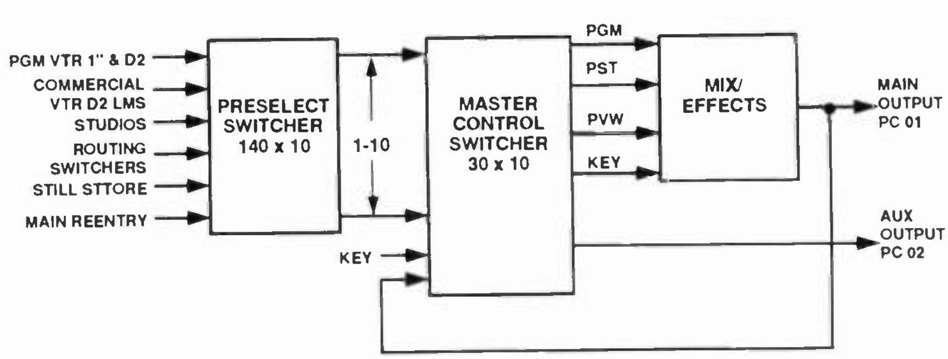


FIGURE 2

BOC CHANNEL

One of the main attractions in choosing the D2 format for BOC LMS machines is to use the raw bit error rate (RBER) diagnostic for automatic verification of digital transfers. By monitoring peak error rates of the source playback material and dub copy confidence playback, unattended verification of video transfer may be achieved.

This technology combined with multiple spots per cassette leads to some very attractive system proposals. The first step improvement is illustrated in Figure 4. Shown here is the elimination of physical use of cassettes within the system. The central commercial library consists of two one-thousand cassette LMS, main and backup, with approximately ten spots in each cassette. This gives an on line capacity of ten thousand spots. The only manually assisted part of the operation would be the original transfer of material into the central library. Thereafter, spots are prepared for air playback by digital transfer to playback LMSs. Sophisticated software would be required to manage the transfer of spots during program segments for playback during upcoming breaks.

Solid State Playback Device

A further improvement in the commercial playback system may be had by replacing the LMS cassette playback devices with semiconductor memory storage devices. Such a "VTR" would be highly reliable since it would be without any moving parts and use no magnetic media. This solid state video recorder would accept D2 digital inputs and provide spot playback under automation control. Figure 5 shows a commercial playback system using solid state players. To replace an 80 cassette LMS with an average fifteen second spot, approximately 20 minutes of memory would be required.

An advantage of this approach of moving data instead of tapes is that any form of high density mass storage may be used for the central library. You are not limited to technologies capable of playing back video in real time. Data may be retrieved and transferred at rates slower or faster than that required to generate pictures. Only the playback unit must convert the data back to audio and video.

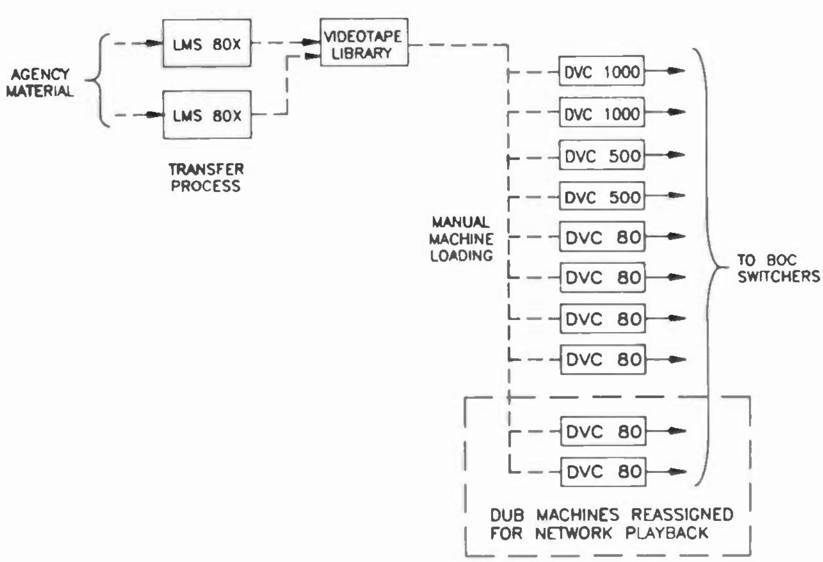


FIGURE 3

MANUAL TAPE MOVEMENT

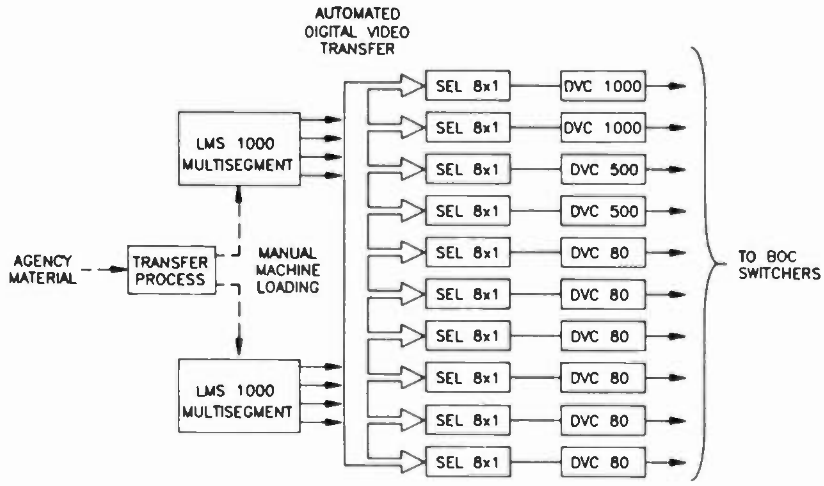


FIGURE 4

DIGITAL DUBBING

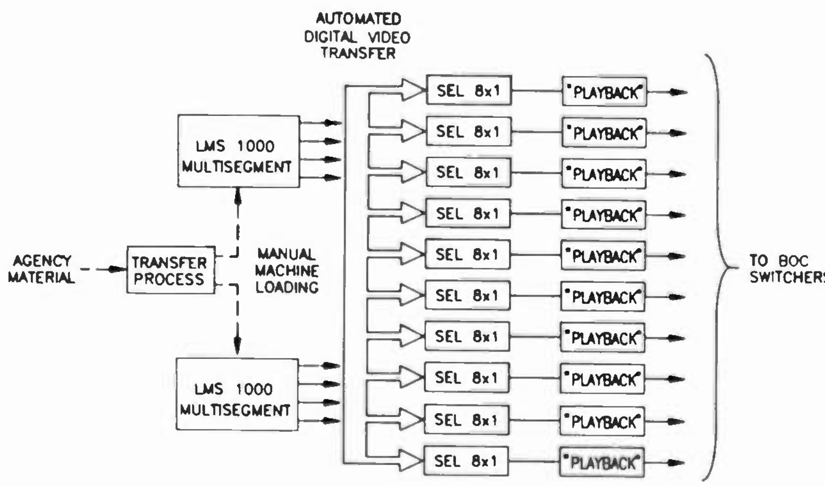


FIGURE 5

SOLID STATE PLAYBACK

Conclusion

The new generation of commercial spot playback equipment offers performance and labor saving improvements. Full development of all the features in this equipment will give significant additional improvements in future system designs.

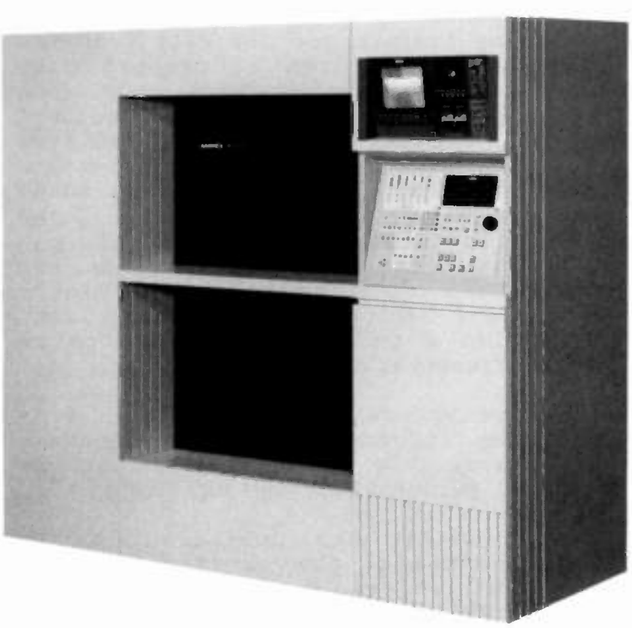
DATABASE MANAGEMENT FOR AN AUTOMATED CASSETTE RECORDER/PLAYER SYSTEM

William F. Carpenter
Ampex Corporation
Redwood City, California

A modern automated cassette recorder/player system, when it is designed around an extensive database management program, can provide many operational and financial benefits as compared to systems which are just a group of studio recorder/players combined with a robot and a computer to provide the controlling functions.

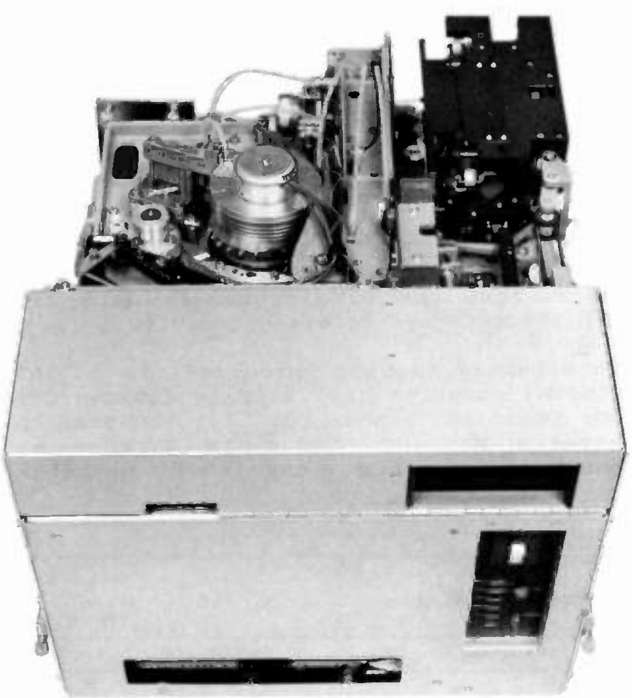
The definitions for the key points:

> **Data**: Factual information used as a basis for reasoning, discussion, or calculation
>
> **Database**: A collection of data organized especially for rapid search and retrieval (as by a computer).



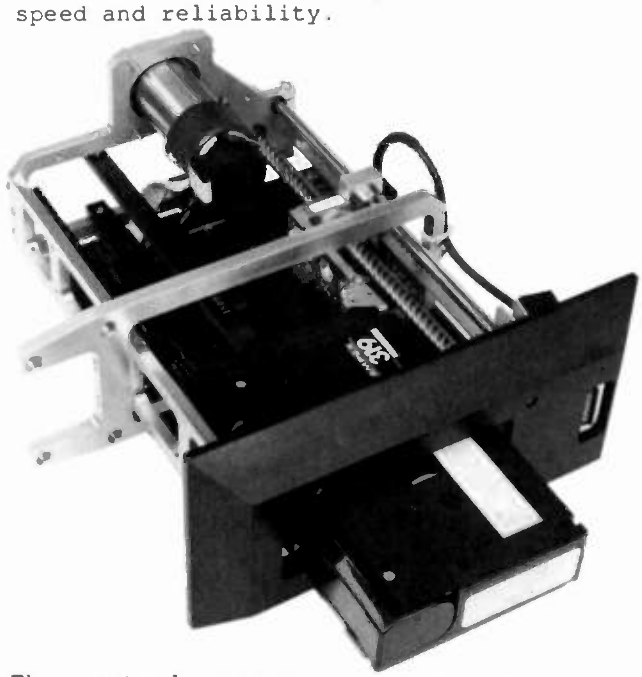
The system components should be designed as smart devices, with their own data processing components and the necessary data outputs to keep the central database current at all times.

The system has the following individual components, which include their own fast and powerful computers.



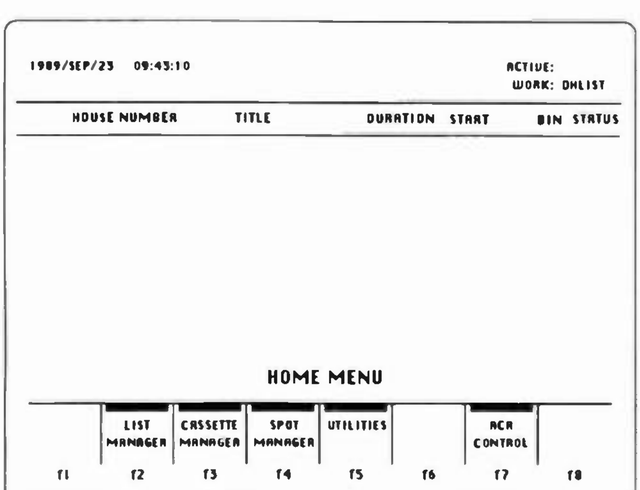
The unique application specific tape transports, three or four, which are designed to handle the tape in a fast and gentle manner with cueing speeds which are twice as fast as typical studio cassette transports. The tape is protected from damage during these cuing functions through the use of air lubricated tape guides and a direct drive, pinch rollerless capstan.

The robotics system, since it is moving a cassette designed for automated handling, also operates in a fast efficient manner. The cassette handling portion also carries the bar code reader, facilitating the reading of the simple six digital code, without moving the cassette. The robot features a simple, but effective method of mechanically coupling the cassette to the handling system to move it from the storage bin to the tape transport with maximum speed and reliability.



The central computer directs the system, and manages the bi-directional data flow with the external database computer.

The database manager which resides in this external computer basically is divided into four basic sections: the List Manager, the Cassette Manager, the Spot Manager and though it is not obvious to the operator, the Resource Manager.



The modern television broadcast facility has many diverse areas whose efforts are all basically aimed in the direction of presenting programming and commercials ,which in turn provide the revenue to support the operation.

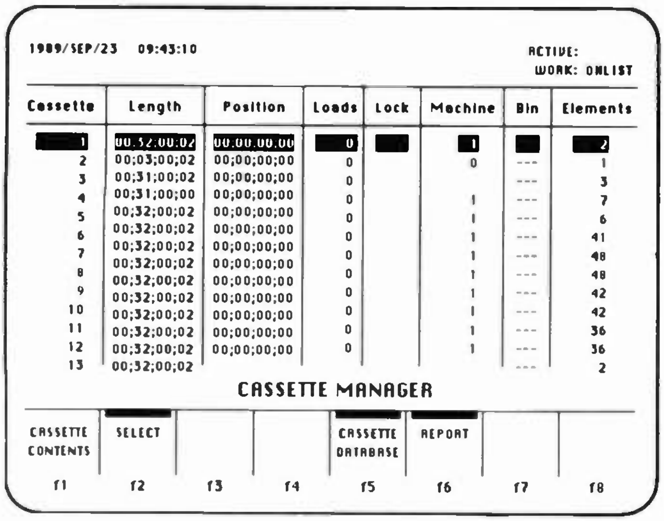
Since many of these functional areas have previously moved to computers to improve their operational efficiency, the database system should also allow easy interface with these systems already in use. The database runs in the commonly available, worldwide accepted, MDOS environment. This was chosen to allow simple, and extremely effective interface with the existing systems. The database is written in method to be directly compatible with Informix, which is a well known, commercially available database which runs on many different computers.

The database computer now can be used as a terminal point to allow the system to be expanded into these other areas. By using a system comprised of proven hardware and software components, namely eithernet or thin net interconnections and Novell network software, the expansion can be extremely cost effective as compared to other typical television system expansion costs.

This external system allows the database computer connected to computers in the traffic department, which can interface to the existing traffic for the easy transfer of information required to prepare play lists. The computer in tape room when interconnected with a D2 cassette recorder, allows the material to be transferred from the incoming format to the cassette format, and the necessary information to be added to the database at the same time. The computer in presentation control can provide the necessary play commands and also be used to make any last minute modifications to the current play list. This is also a point of interface to an Automated Presentation System.

Now that we understand the system, let's examine the information in the database which is available for use by either the operational personnel or the equipment.

When we look at the database via the cassette manager the information is arranged in reference to the cassette number. This is the number which is indicated by both the barcode and the numeric label on the ends of the cassette. These labels are applied to the cassette only at time that it starts to be used in the system and are not changed or modified or removed for the life of the cassette. The simple six digit two of five interleave code allows the machine to quickly identify to the database what cassette is available for use.



1989/SEP/23 09:43:10 ACTIVE: WORK: ONLIST

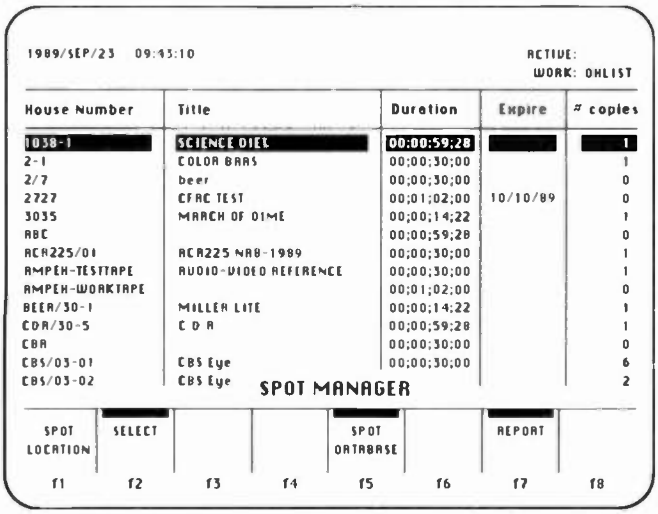
Cassette	Length	Position	Loads	Lock	Machine	Bin	Elements
1	00:32:00:02	00:00:00:00	0		1		2
2	00;03;00;02	00;00;00;00	0		0	---	1
3	00;31;00;02	00;00;00;00	0			---	3
4	00;31;00;00	00;00;00;00	0		1	---	7
5	00;32;00;02	00;00;00;00	0		1	---	6
6	00;32;00;02	00;00;00;00	0		1	---	41
7	00;32;00;02	00;00;00;00	0		1	---	48
8	00;32;00;02	00;00;00;00	0		1	---	48
9	00;32;00;02	00;00;00;00	0		1	---	42
10	00;32;00;02	00;00;00;00	0		1	---	42
11	00;32;00;02	00;00;00;00	0		1	---	36
12	00;32;00;02	00;00;00;00	0		1	---	36
13	00;32;00;02					---	2

CASSETTE MANAGER

CASSETTE CONTENTS	SELECT			CASSETTE DATABASE	REPORT		
f1	f2	f3	f4	f5	f6	f7	f8

Additional information displayed in this area includes, cassette length, number of times loaded into a transport and whether it is locked to prevent removal from the system (ideal for station I.D.s and emergency info, which is either used very often or must be immediately available).

The Spot Manager is referenced to the sixteen alpha-numeric character house identification which has been assigned to a specific spot, or segment. One of the interesting pieces of information available in this area is the "Q" or quality statement which relates to the amount of action of the digital error correction system during the last playback. This information is used two ways, by the operations personnel as a non-destructive early warning system and by the machine to determine which copy of a commercial which is stored in multiple locations to use to provide the best playback quality.



1989/SEP/23 09:43:10 ACTIVE: WORK: ONLIST

House Number	Title	Duration	Expire	# copies
1038-1	SCIENCE DIET	00:00:59;28		1
2-1	COLOR BARS	00;00;30;00		1
2/7	beer	00;00;30;00		0
2727	CFAC TEST	00;01;02;00	10/10/89	0
3035	MARCH OF DIME	00;00;14;22		1
ABC		00;00;59;28		0
ACA225/01	ACA225 NAB-1989	00;00;30;00		1
AMPEX-TESTTAPE	AUDIO-VIDEO REFERENCE	00;00;30;00		1
AMPEX-WORKTAPE		00;01;02;00		0
BEER/30-1	MILLER LITE	00;00;14;22		1
CDA/30-5	C D A	00;00;59;28		1
CBA		00;00;30;00		0
CBS/03-01	CBS Eye	00;00;30;00		6
CBS/03-02	CBS Eye			2

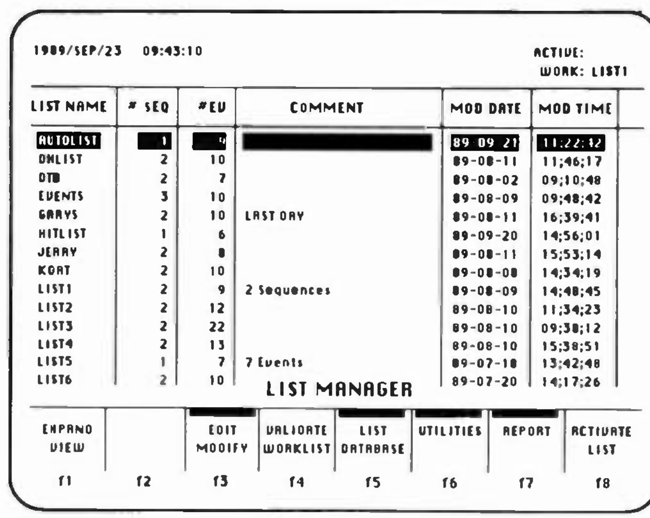
SPOT MANAGER

SPOT LOCATION	SELECT			SPOT DATABASE		REPORT	
f1	f2	f3	f4	f5	f6	f7	f8

In both the Cassette and the Spot Manager areas, facilities have been provided to "select" or "Query by example" the database to find a specific item or group. Also manual control of functions like thread, rewind, and preview, have been made conveniently available in these areas.

The List Manager is arranged in alphabetical list name order, and allows for the creation, storage, editing, and validation of the list which provides to the system the required playback sequence.

The editing portion of the List Manager allows the display of a list event number in ascending order which is also the normal order of playback. Additional fields in the basic display show the House number, title field, the duration of playback, the time of playback, which is valid if the unit is running from the house clock in a basic automation mode. In this mode all list items are timed events.



1989/SEP/23 09:43:10 ACTIVE: WORK: LIST1

LIST NAME	# SEQ	#EV	COMMENT	MOD DATE	MOD TIME
AUTOLIST	1	9		89-09-21	11:22:42
ONLIST	2	10		89-08-11	11;46;17
OTB	2	7		89-08-02	09;10;48
EVENTS	3	10		89-08-09	09;48;42
GARYS	2	10	LAST DAY	89-08-11	16;39;41
HITLIST	1	6		89-09-20	14;56;01
JERRY	2	8		89-08-11	15;53;14
KORT	2	10		89-08-08	14;34;19
LIST1	2	9	2 Sequences	89-08-09	14;48;45
LIST2	2	12		89-08-10	11;34;23
LIST3	2	22		89-08-10	09;38;12
LIST4	2	13		89-08-10	15;38;51
LIST5	1	7	7 Events	89-07-18	13;42;48
LIST6	2	10		89-07-20	14;17;26

LIST MANAGER

EXPAND VIEW		EDIT MODIFY	VALIDATE WORKLIST	LIST DATABASE	UTILITIES	REPORT	ACTIVATE LIST
f1	f2	f3	f4	f5	f6	f7	f8

Also the List Manager allows for the validation of a list at any time prior to use, and the resultant displays indicate the resources required, time to activate, and the availability of the required elements. A printed report is also available to be used to locate any missing elements. This function is also providing the most basic form of "Conflict Recognition", since it recognizes and display any "conflicts" caused by requesting the system to provide functions which are not possible due to element or system time restraints.

This synopsis, is a very basic outline of the Database Management System, which is much easier to understand when it is actively demonstrated to show these major operational values. This paper at presentation will be centered around a fully operational database displayed using a system to allow viewers to see the actual operation.

THE ENGINEER'S ROLE IN IMPLEMENTING A NEW GRAPHIC LOOK—TECHNICAL, PRODUCTION AND MANAGEMENT CONSIDERATIONS

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INTRODUCTION

Like many projects in broadcasting, implementing a new graphic look at a station is a team effort. To be successful, creative, technical, and management skills must mesh to achieve the desired result.

The Engineer has a great deal to contribute to this process. This is especially true in the present era of computerized graphics, a technology which has brought forth tremendous possibilities, but which present operational challenges within the scope of the Engineer. Contributions based on a knowledge of the technical strengths and weaknesses of the hardware and software are often critical to the success of the implementation.

The Engineer needs be familiar with the overall creative objectives as well as the technical considerations in order to be an effective member of the team. This paper will explore the technical, operational, and management considerations involved.

A NEW GRAPHIC LOOK

Overall Goals

The implementation of a new graphic look should achieve a design objective which conveys or supports the desired station position.

Be unique. With the many sources of programming which now compete for viewers, it is of increasing importance that the implementation of a graphic look be uniquely identifiable with the station.

Be consistent. The look should maintain elements of consistency across all aspects of programming associated with the station. The on-air aspect of this generally includes News, Promotion, public service and station continuity elements. It is desirable to have a complete implementation within these categories so that the viewer is aware of what station he or she is watching.

Engineering's Contribution

Implementing such a look is a stationwide concern. Since the majority of the on-air graphic look is achieved through the technology of electronic graphics, input from Engineering in the planning stages can make the difference between a successful and a unsuccessful implementation of new graphics. Engineering management must interface with the "creative" people on the team to get the best results.

Sources of Design Elements

In assembling elements to change the look of a station, one decision which must be weighed is how much of the project can be designed and executed in-house versus out of house. The limiting factors which influence this are availability of resources, time, and money.

Out-of-house. There are several benefits to the use of out-of-house production which make this option desirable to station management. Key design talent or advanced production equipment may be present in the station, but going out-of-house allows the project to proceed without interfering with ongoing operations. Outside contractors are also motivated by the bottom line to work efficiently and to meet specified deadlines.

If out-of-house production is considered, it is important that any technical or operational concerns about the elements are considered in advance. These things may be very easy to specify up front, but difficult to correct after the production is completed. Some of these issues are:

1. What format and brand of tape stock will be used to deliver taped items?

2. Time code on the delivered tapes is often critical for smooth integration into in-house production, logging, and match-frame edits from source material. Will the tapes be timecoded, and if so, will drop-frame, or non-drop frame be used? Is the time code and video color framing aligned to the CF pulse on the tape.

3. If super black is to be used for keyable elements, what is the desired level for best results in-house?

4. Are animated elements designed with run-on time which is more than sufficient for all applications?

5. It is always desirable to take delivery of images in a digital format. How will stills be delivered? On tape, disk packs, or streamer cartridge?

6. Has adequate time been allotted to integrate the elements into the in-house system?

7. What expectations are there for out-of-house elements to be manipulated in-house. For example, to update the talent faces on a news open, will it be necessary to go out-of-house, or is this expected to be done with in-house equipment.

EQUIPMENT CONSIDERATIONS

One of the overall goals mentioned for a new graphic look is to implement a particular design. From the creative standpoint and ideally, the design achieves a particular aesthetic goal without regard to the limitations of equipment. The reality of the situation is that the design passes through all the equipment of the station, the transmission media, and the viewer's receiver before reaching the viewer. Designs which are formulated with total disregard for the technical aspects of the medium are bound to run up against the limitations of the equipment. One of the contributions engineers can make to the design process is to help educate creative members of the team as to the technical constraints of the medium. Good design can exploit the "limitations" of television and turn them into advantages.

All equipment in the video chain can introduce degradation to the video. Electronic graphics tend to manifest these degradations subjectively before other video.

It is important to test graphics by passing them through the various pieces of equipment which will be used to store and retrieve the graphics.

Video Tape. Rendered, animated graphics are often stored and retrieved on videotape. The retrieval process can be live, in the case of a newscast, in preproduction, or in editing. There are several considerations to be aware of when using videotape as the storage medium.

First, image degradation due to tape artifacts are always a factor in analog tape formats. Taped images are subject to dropouts and increasing distortion of color information with repeated use. The use of color-under formats to store and retrieve graphics is undesirable because of the effects of limited luminance bandwidth and chroma delay.

There are also logistical considerations when graphics are rolled-into a live production from videotape, particularly in live broadcasts. First, preroll time must be allowed for the image to stabilize. It is unfortunate that many ACR-25 cart machines and AVR-1 quad machines, which provided instant preroll are coming out of service only to be replaced by media which do require a preroll. In live productions, there is the issue of cuing multiple graphic clips.

Digital media, which include digital video tape, digital disk recorders, and video rams get around these problems. Unfortunately, with the possible exception of digital video tape, these media are not yet cost effective alternatives for television stations.

Super Black. Integrating keyable graphic objects from tape into live productions is another area of logistic concern. The most common, and the most convenient method of accomplishing this is by surrounding the object to be keyed with "negative", or "super" black. In my experience, the best results are obtained when the level of the superblack is about minus five IRE. This provides a margin of safety against tape noise and finite switcher key gain in achieving a clean key. Unfortunately, not all TBCs are capable of passing minus 5 IRE black without either clipping at some higher level, or interfering with sync separation on record or playback. Some TBCs that clip super black can be adjusted or modified to pass it.

There are some disadvantages associated with the use of super black. First, antialiased and semitransparent keys are not possible. The key has a hard edge. Second, there always seems to be some way the raw super black tape finds its way to the transmitter without being keyed out. This results in the transmission of a beautiful graphic surrounded by a richly black, and highly illegal background.

One other problem with taped graphics is that the clips are of finite length. Freezing graphics on tape via Auto Tracking is usually unsatisfactory. For live use, it's important to make sure that there is more than pad time after the graphic resolves for any conceivable duration of use.

In most live situations, the operative means of storing and recalling static graphic elements is the electronic still store. Still stores, while internally digital, introduce artifacts which arise from four basic areas; Analog-to-digital conversion, luminance/chrominance separation, chroma inversion and digital to analog conversion. These factors are more or less important, depending on whether the still store is internally of a component, or a composite design.

Still stores which are internally of a component design introduce artifacts when an encoded input is separated into luminance and chrominance. One strategy to avoid this problem is to transfer graphics into the still store from the source in a component form, or better yet, digitally. If this is not possible, the effect of any artifacts should be tested on the graphics to make sure that some fundamental design element of the look is not degraded.

Composite still stores which store only two fields of video degrade graphics due to artifacts of so-called chroma inversion. Chroma inversion is necessary to reconstruct the 4-field (NTSC) color sequence from the two stored fields. Invariably, imperfect luminance-chrominance separation in the digital comb filter circuit leads to graphics that flicker, particularly on vertical edges. Four-field still stores eliminate the flicker problem by maintaining the complete color sequence. Since twice as much data is required to accomplish this, there is a tradeoff against recall time and disk capacity when using four field images.

Digital Effects. Another category of equipment which is notorious for degrading graphic images is digital effects. Digital effects devices, depending on their design, are prone to introduce all of the same problems still stores can, and in addition manifest artifacts due to image size interpolation and aliasing.

When one considers that the average news shoulder graphic passes through both the still store and the DVE when being used on-air, the combined degradation is of particular concern.

Graphic designs which contain fine detail, lines or vertical edges are especially susceptible to the effects of DVE's and still stores.

Character generators can contribute greatly to the consistency of the look by implementing a particular typographic design. Doing this successfully involves exploiting the strengths and hiding the weaknesses of the particular character generator.

Character generators which are not capable of generating antialiased type are still at this point in very common use. One has to be very careful about selecting, sizing, and cleaning-up fonts which are used to be used without antialiasing so that the end result best reproduces the intended typeface, but also so that fine vertical detail does not cause field rate flicker to occur. Achieving this is often a compromise situation, but good results can be obtained.

Font Acquisition for CGs

Acquiring fonts for use in character generators has advanced from being a matter of little choice, to the present day where many typefaces are available.

Fonts in ROM. When character generators came into common use in television stations in the early seventies, the high end equipment provided the luxury of enough resolution to actually reproduce the curved contours of one or two typefaces. The fonts were stored in read-only memory, and few choices were available. Custom logos were obtainable, but these too had to be burned into ROMs, a process reserved for the manufacturer.

Fonts on disk. The next advance was the introduction of downloadable fonts, which were supplied on a diskette. More typefaces were available, but these fonts were supplied in specific sizes. Only the vendor was capable of generating the data for fonts on the disks. It was often possible to identify the character generator in use by at a particular station by seeing which fonts were used.

Font Compose. The introduction of a font compose option allowed the user to create new custom fonts by digitizing artwork. More often this feature was, and still is, used by stations to capture the station logo and incorporate it into news supers.

Font Scaling. After the introduction of font compose, font scaling utilities became available. Typefaces could now be scaled more easily, and one was therefore not limited to particular sizes to implement a design. Since the actual scaling algorithms were not antialiased, quite a bit of pixel-level cleanup was required on fonts scaled this way.

Foundry Type. Up until this point, typefaces supplied by character generator vendors were derived only indirectly from actual type foundry letterforms. The next leap in character generator font acquisition came when CG vendors were able to offer fonts derived from high-resolution digital representations of foundry typefaces. The data, supplied in the form of vectors or large bitmaps, can be now be used to scale antialiased or standard versions of virtually any typeface.

While the advent of foundry-based type has improved the quality and quantity of type on character generators, there is a downside.

The type foundries charge a royalty for each master font sold, which must be passed-on to the end user. Charges for master fonts can range from 100 to 1000 dollars per typeface depending on the vendor and whether fonts are purchased as part of a package. Multiply this cost over several different pieces of equipment to be made consistent, such as weather graphics machines, and paint systems, and the cost becomes even more significant. Also, agreements between the type foundries and OEMs have given rise to various software protection schemes which allow fonts to be used only on authorized machines. This can be a logistic problem.

MAKING THE TRANSITION

One of the most strenuous periods in implementing a new graphic look is the transition phase. The goal is to achieve a transition that is timely, smooth, and complete.

Often there is a tradeoff between a timely transition and a complete one.

Pitfalls

One frequent pitfall involves rushing the transition on to the air without a sufficient variety of elements in place. There is a tradeoff between working with fewer, but more consistent graphics, as apposed to many inconsistent graphics. It is important to be sure that before the transition, all the new elements to continue operating are in place. If particular elements are missing there should be some agreed-upon means of working around the missing elements.

Transition Strategies

One transition strategy which seems to work is one of replacement and physical removal: All of the new elements, including tapes, stills, and font disks, are lined up on the day of the transition, and the outdated elements physically removed from the operational areas.

It's very difficult to keep those last few old logos from popping up on the air after the transition. It is a good idea to delegate one person the task of screening the on-air cart inventory for occurrences of old logos or graphics which have escaped normal channels. There should also be a policy in place about automatically pulling materials which contain old graphics, should they accidentally air.

Look for old graphics in obscure places, and arrange for replacements. These may include emergency slides, EBS materials, standby filler reels, and Network ID systems, to name a few.

CONCLUSION

The benefits of implementing a consistent graphic look are of stationwide importance. To accomplish this goal, a team effort between creative and technical people is required. The engineer can contribute productively to this team effort, by providing operational, technical, and logistic support.

REAL-TIME WEATHER INFORMATION IN THE 90'S

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INTRODUCTION

The weather information industry is now just ten years old. In the past decade, the amount of raw data available from all sources has grown by at least two orders of magnitude. In the next decade, even greater growth can be expected. New systems such as NEXRAD, ASOS, the Profiler network, and GOES NEXT will make the data assimilation/processing problem one of our greatest challenges.

Will the rapidly changing communications and computing technologies be able to handle the growth? Even as personal computers become smaller and more powerful, will individual users be able to cope with the flood of new information?

This paper will review these trends, and explore the options that are available. In particular, emphasis will be placed on preprocessing these datasets, to provide "value-added raw data", which can then be further processed and analyzed to meet individual users' needs.

BACKGROUND

As recently as the late 1970's, end-users of real-time weather information received their data via low-speed (75 words per minute) teletype circuits. Alphanumeric data arrived via the Service "A" (hourly surface observations), Service "C" (synoptic and upper air data), and RAWARC (Radar and Warning Coordination) circuits. Meteorologists had to constantly tear paper, and sift through yards of printouts to find those pieces of data that were of interest. Furthermore, all of the data were in raw form; the user had to decode the information from the standard meteorological formats. Preparing time series of the data was a time consuming, and tedious process.

Satellite imagery at that time were typically received via GOES-TAP, provided by the National Environmental Satellite, Data, and Information Service (NESDIS). GOES-TAP provided (as it still does today) "channels" of various GOES satellite imagery sectors. As GOES-TAP is an analog land-line circuit, acquisition and maintenance is costly; the imagery are in black and white only, and manipulation of the imagery is difficult.

Radar in the late 1970's was obtained via dial-up from WSR-57 and WSR-74 digitizers located at the NWS sites. Only one site could be obtained at a time, and were subject to severe ground clutter and anomalous propagation contamination.

Facsimile charts were received via the National Facsimile (NAFAX) circuit, which was limited in capacity; the charts lacked resolution.

In the early 1980's the National Weather Service (NWS) instituted what is today called the Family of Services (FoS). The FoS is a group of medium-speed circuits designed to deliver bulk data to the weather information industry, which in turn can provide these raw data - or value-added products and services - to end-users. In 1990, the FoS consists of three 2400 baud circuits (the Public Products Service, or PPS; the Domestic Data Service, or DDS; the International Data Service, or IDS); and two 4800 baud circuits (the Numerical Product Service, or NPS; and the Digital Facsimile Service, or DIFAX). Today, nearly all end-users of meteorological data receive their information from the weather information industry, either via request-reply (dial-up) or satellite distribution. End-users benefit because the data can be obtained in value-added formats, allowing easier assimilation of the information. Valuable manpower can be applied to the tasks at hand, rather than managing the incoming data.

Satellite and radar imagery acquisition have changed over the past decade as well. The weather information industry provides access to high-resolution GOES imagery and radar composites via request-reply and satellite distribution, which can be manipulated by workstations ranging from sophisticated mainframe systems to personal "desk-top" computers.

Government's Role

Over the past decade, a distinction of the roles of the Government and private sector has developed. This development has benefitted all parties: The Government because it can concentrate its efforts on what is its charter; the private sector because it stimulates economic growth; the end-user community because it receives high-value products and services.

The primary charter of the NWS is to provide warnings of severe weather so as to protect lives and property, and to provide general forecasts for the public. Whereas a decade ago the NWS and its sister agency NESDIS provided end products to the user community (media, utilities, agriculture, etc.), now it has turned this task over to the private sector. The Government now provides raw products, basic research and development, and the deployment and operation of the major sensor systems with which the data are collected.

Private Sector's Role

Armed with real-time delivery of the raw data from the Government collection systems, the private sector now utilizes its resources to generate high-value products. The private sector's ability to take advantage of the rapid advances in computing and communications technology, as well as the need to remain competitive in the field, ensures that new and cost-effective products will become available.

NEW MAJOR SENSOR SYSTEMS

In the 1990's, the NWS is embarking on a major modernization program. It is sponsoring three new sensor systems; its sister agency NESDIS a fourth.

NEXRAD

The Next Generation Radar System (NEXRAD) is a tri-service (NWS, Federal Aviation Administration and Department of Defense/Air Weather Service) project which will field more than 175 Doppler radar units by 1995. In addition to the raw data types of reflectivity, turbulence, and radial velocity, NEXRAD can provide up to 70 derived products. A new private sector distribution system, NIDS, is being developed to ensure that external users have easy and cost-effective access to the data.

GOES NEXT

The Next Generation GOES Satellite System (GOES NEXT) will begin with a launch in the early 1990's. It will provide improved resolution imagery (1km visible and 4km infra-red), and have a simultaneous imaging and sounding capability.

Profiler Network

The Profiler Network will by the mid-90's, provide a national network of static, vertical radars capable of obtaining six-minute updates of the vertical profile of the atmosphere. Value-added products will include longer period averages and other time series information.

ASOS

The Automated Surface Observing System (ASOS) will be a network of up to 1300 automated units deployed by 1994. It will be capable of providing one and five minute updates of various meteorological phenomena.

THE DATA EXPLOSION

What does the deployment of the new sensor systems mean to the weather information industry, and the end-user community? It is clear that the amount of data to be processed will continue to grow exponentially.

In 1979, WSI Corporation's mainframe computers processed approximately 8.5 megabytes of information each day from the low-speed alphanumeric circuits. By 1984, the transition to the first stage of the FoS had increased this amount to 24 megabytes per day. In 1989, including the upgraded FoS, as well as real-time GOES imagery and the raw digital data from all 128 NWS radar sites, the number increased to 132 megabytes.

What will it be in 1994? Including all alphanumeric data, DIFAX, ASOS, Profilers, GOES NEXT, NEXRAD, and gridded model output data, it is estimated that WSI will need to ingest, process, and store over 11,000 megabytes of data per day!

VALUE-ADDED RAW PRODUCTS

How will we cope? The answer, in part, relies on the private sector's ability to provide what can be called "value-added raw products". This is pre-processed information which can be transmitted to end-users for reprocessing on their own meteorological workstations. For example, the National Meteorological Center (NMC) of the NWS provides an FoS circuit with gridded output of the various numerical models (Nested Grid Model, Aviation Model, Medium Range Forecast, etc.). This four dimensional dataset can be pre-processed, allowing end-users to obtain specific data points, decoded, and ready for input to their own models or analysis systems. WSI currently provides this type of product. A sample of Medium Range Forecast data is shown in Figure 1.

SUPERfax™

In addition to making the "value-added raw data" available, additional emphasis will be placed on creating new products for end-users. These products should be highly flexible, allowing the end-users to specify what they want to see, and how they want to see it. The user should not be limited by what the computer can do; only by his or her imagination.

WSI's SUPERfax is a first step in this direction. Users can generate a facsimile chart for any data type in the NWS/NPS data set. The chart can be a composite of two different data types, or the same data type at two different time steps. A sample SUPERfax chart is shown in Figure 2.

NOWrad™

For more than 30 years, users of radar have been limited by their ability to acquire composites of radar imagery. The primary reason for this is unprocessed radar imagery are subject to contamination by ground clutter and anomalous propagation - false echoes which can be difficult to distinguish from actual precipitation returns. WSI's NOWrad advanced radar composite product is the first to provide high-resolution national and regional radar composites on which much of the false echoes have been eliminated.

Every 15 minutes, WSI's computers receive the latest raw digital radar data from all reporting NWS radars. A national composite is

```
gdata -50/90 -40/60 500 0z mrf 168hr hght
DATA FROM MRF MODEL RUN: 00Z 12/27/89
SW corner: -50/0090, NE corner: -40/0060, level: 500 mb,Fcst: 168HR Parm: HGHT

534.957275 536.738525 535.144775 531.051025 525.426025 520.738525 518.426025
551.488525 551.676025 549.051025 544.957275 539.144775 533.207275 531.801025
564.519775 563.426025 561.082275 557.957275 552.957275 549.519775 552.019775
```

Fig. 1. Decoded NWS Medium Range Forecast model data: 168-hour forecast of 500mb heights in the area bounded by 40-50 South latitude and 60-90 West longitude.

```
+sfaxi par=fht lev=500 for=144 par=fp lev=sfc lat=40 lon=0 mod=ecm

 analyzing data... ...done

 drawing contours... ...done
[0w
```

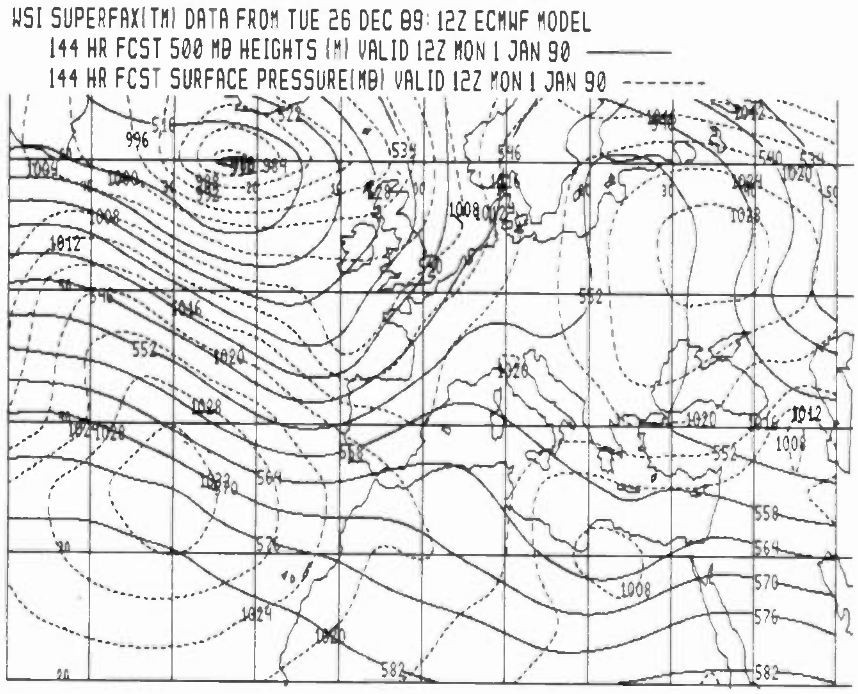

Fig. 2. WSI SUPERfax chart showing a 144-hour forecast of 500mb heights (solid lines) and surface pressure (dashed lines) from the European Centre for Medium-Range Forecasting (ECMWF) model. The raw data arrive via the NWS Family of Services Numerical Product Service.

immediately generated, and special algorithms are applied which substantially reduce the false echoes caused by ground clutter and anomalous propagation. Users can then choose an 8km national composite, or a 2km regional composite for display on a personal computer or other graphics workstation.

CONCLUSIONS

The key to managing the explosion of weather information in the coming years will be the sophisticated data processing and value-added product generation demonstrated here. Once the information is made available to the end-users in this pre-processed, value-added form, they will be better able to assimilate the products into useable information. The challenge for the future is to ensure that Government and the private sector work together to ensure that information flows from the sensor systems to the end user in a timely manner; the products provide useful information; and the users are trained in using the products.

ANIMATION SYSTEMS AT NBC

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Abstract
During the last few years NBC has developed the ability to create sophisticated graphic animation sequences using in-house facilities. There are several methods used to create these animations. Two of those methods will be discussed in this paper. The first uses a network of 3D computer workstations linked together on a single "graphic" Ethernet Lan. These systems are used to create the spectacular graphic effects used for special segment animations, show openings, and on-air promos. The second method uses the frame by frame editing compabilities of a Quantel Harry combined with the Quantel Encore and Paintbox systems to create complex animations, that would be difficult to create on a computer based 3D system. These two methods can be also used in combination with each other. Animations created on a 3D system can be transferred to a Harry and layered with other effects.

This paper will discuss the hardware systems, software and the support issues that arise with the use of these systems.

3D Animation Systems

Since the early 1980's NBC has used a variety of systems to create 3D animation sequences. Although these systems were successfully used at NBC they were not compatible with each other. In 1985, NBC evaluated several 3D animation software packages developed by independent software vendors. As a result of these evaluations it was decided to purchase a 3D software package developed by Wavefront Corp., to run on a Silicon Graphic IRIS 2400T Workstation. One of the major reasons for this decision was the fact that Wavefront software ran under the UNIX operating system. At that time, UNIX was the operating software of choice on most high-end workstations. It was portable across different hardware platforms. This meant that when better, more powerful hardware became available, we could move the UNIX-based software onto those systems and thus protect the software investment.

The IRIS 2400T was installed in the Computer Imaging Lab at NBC in New York. In this lab, artists and software programmers could work together. The programmers could supply the technical knowledge lacking in the artists, while the artists could supply the artistic talents absent in the programmers. This close relationship between technical and artistic people has been a major factor in the success of these systems.

In the lab, various DEC computers running the ULTRIX version of UNIX were linked together via an Ethernet Lan. The 3D workstation was incorporated into this network, and one computer on the network, a DEC 750, became a file server used for storing and retrieval of 3D files.

Gradually, the software bugs, the artists' problems and the programmers' issues were resolved, and in 1986 another Silicon Graphic 2400T, purchased by the Network News Graphic Dept. was added to the network. As the demand for 3D animations increased, the issue of rendering time began to surface. Rendering is the final step in creating a 3D animation sequence. It is the step when all the data about the various objects in a video frame are brought together, pixel by pixel, to create a complete frame of video. This process is computer-intensive. Information regarding the objects' shape, color, shading and the effect of light sources must be processed before each pixel can be rendered. In the latter part of 1986 a Celerity 1260 compute-server was added to the network. It increased the compute power available to the network by a factor of three, and helped relieve the rendering bottleneck.

Animations created on these 3D systems were first used on-air during 1986. They were aired on several special segment News shows and used for show openings on election night. In 1987 and 1988, a team of artists and engineers put together a 3D graphic facility to create many of the 3D animations that were used during the telecasts of the 1988 Olympic Games. The 3D workstation used in that facility was a Silicon Graphics 3130 initially installed in a production house in Toronto, Canada. There, other graphic equipment used for the Olympics were also being assembled. In 1988 it was moved to Seoul, Korea where it could be used to create "last minute" animation sequences during the Games. Although it was a newer, faster system than the ones in New York, it used Wavefront software and therefore, its files were compatible with New York. Work on an animation sequence could be started on any

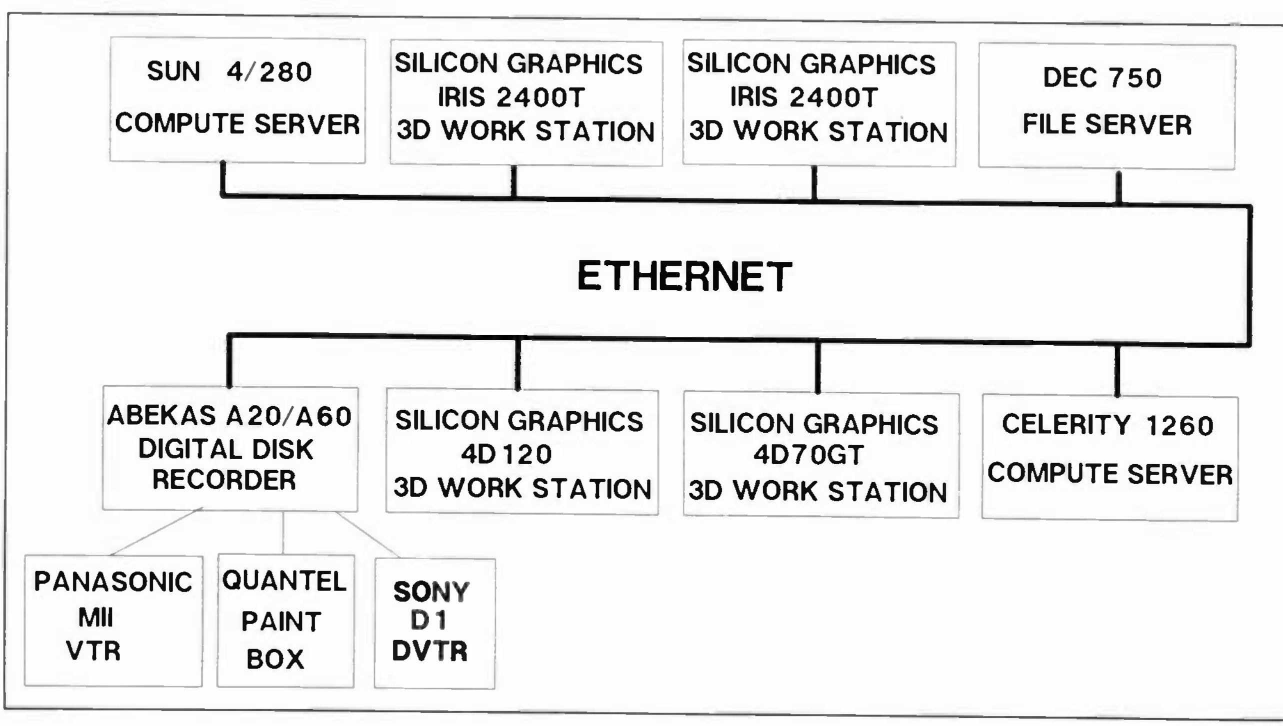


FIGURE 1

NBC NEW YORK
3D GRAPHICS NETWORK

system and finished another. If the backload of work in Korea had become overwhelming, files could have been transferred between Korea and New York via phone.

Because of the successful use of 3D graphic systems during the Olympics, and the growing demand within NBC for this type of animation, the graphics department of NBC's Operations and Technical Services Division purchased two new graphics systems in 1988. These systems, the Silicon Graphics 4D-70 Workstation and the Silicon Graphics 4D-120 Compute-server were also added to the network. In the same period a Sun 4-280 compute-server was acquired by the Systems Development Department as part of a separate software effort. This Sun 4-280 was also added to the Ethernet Lan and the rendering power available to the 3D animation systems was again increased.

The importance of the decision to use UNIX based software had now become apparent. Hardware systems from four different vendors were linked together on a single graphic network servicing the needs of three different departments in NBC. (Figure 1).

Video Systems

Four additional devices play important roles in the 3D facility. A Quantel Paintbox is used to provide textures or patterns for the 3D systems. A Sony D1 (CCIR 601) recorder is used both for production and for archival storage of images in a digital component format. A Panasonic M-II analog component recorder is used for storing production copies of the animation effects, and finally an Abekas A60 is used as a buffer between the 3D animation systems, the Paintbox, the Sony D1 system and the M-II.

The Abekas A60 Digital Disc recorder and Abekas A20 Component Analog to D1 Converter were added to the network in 1988. These systems have become very important in the 3D facility. Images can be transmitted between the Abekas A60 and the 3D systems via an Ethernet port. The Abekas A60 stores video frames in the CCIR 601 format and can transfer these images to and from the Sony D1 system via a parallel data port. With the A20, the A60 can capture images from the Quantel Paint Box and transfer them to the 3D animation systems. With an NTSC encoder, the A60 is used to transfer images to the Panasonic M-II. Finally, because the Abekas A60 can play back images in both real time or frame by frame, it is used for previewing the 3D animations.

Cel Animation

The second method of producing animations at NBC uses the frame by frame editing capabilities of the Quantel Harry System, along with a Paintbox, an Encore digital effects system. and a Sony D1 digital component recorder. (Figure 2).

Cels or frames of animation can be transferred between the Paintbox and the Harry. Using the Paintbox an artist can create an animation sequence frame by frame, or he can modify a frame of video that was transferred from the Harry. Using the Harry's frame editing capabilities, the artist can add or delete frames from an animation sequence as he desires. When the artist wants to move an object in an X, Y or Z dimension, he can use the Encore to create the movement automatically, instead of having to draw the movement on a frame by frame basis. A most important aspect of the system is that it is a digital component system. Images can be composited without loss of quality. For example, an artist can create a moving background in one sequence of animation and then layer moving foregrounds on top of it without effecting the resolution of the background.

This system allows an artist to create animation that would be too complex, if not impossible, to create on a computer based 3D system. Two classic examples of such animations are a series of air bubbles rising through water, where the sizes and shapes of bubbles are constantly changing and finally popping at the surface, or an animation containing flames that vary in size, shape, color and intensity.

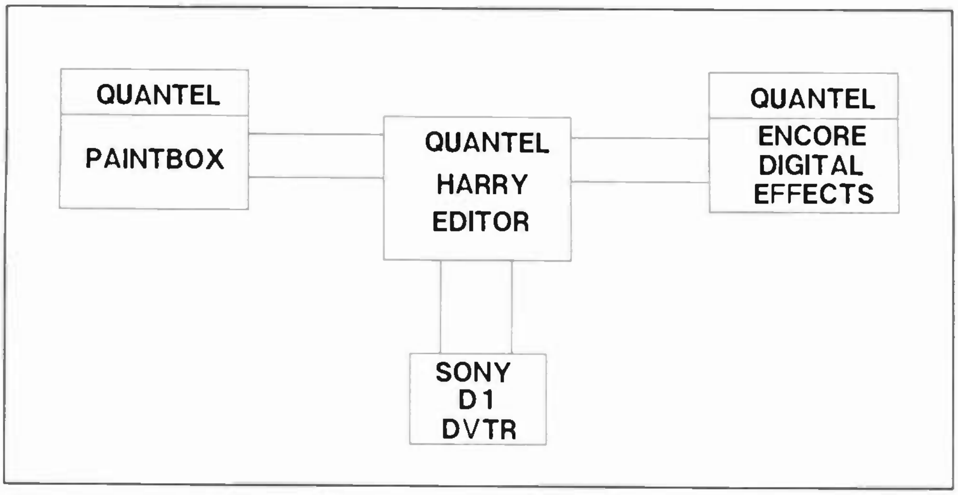


FIGURE 2

NBC NETWORK NEWS ANIMATION GRAPHICS

Summary

To insure that 3D animation systems will be used effectively, two important issues must be considered. The first is portability of software. Every week seems to bring an announcement about a newer, more powerful workstation. If the software packages used to create 3D animation cannot take advantage of these new systems, they will quickly become outdated. The second issue is the need for programmers to support the staff of artists. Programmers are needed to isolate artists from computer issues. They are needed to resolve computer and network problems. They are needed to install new releases of software and train artists in their use. In addition, some of the 3D software packages available today contain software "hooks". These "hooks" enable competent programmers to add new effects to the artists' toolkit. This allows a facility to create unique animation effects and develop "signature" looks. One of the questions that must be asked when deciding to invest in 3D graphic systems is "why bother?" There are certainly many good production houses that are capable of doing this work. Why bother investing in these expensive systems and then have to worry about training artists to use them, and training engineers to maintain them? The answer, of course, is that the investment is not a good one unless you can create 3D animation sequences for significantly less than an outside facility, with at least the same quality and turnaround time.

During the past four years artists at NBC have created over two hundred 3D animation sequences that have been used on air by our Sports, News, and Entertainment divisions. This represents over 4,000 seconds of animations. If these 3D sequences were purchased from outside facilities at estimated rates of $2,000 to $4,000 per second, the cost incurred by these divisions would have greatly exceeded our total investment.

Acknowledgements

The author would like to acknowledge contributions to this paper by the following NBC personnel. W. Colvard, M. Collins, L. Thaler and A. Siegel.

EVERYTHING YOU ALWAYS WANTED TO KNOW ABOUT COMPUTER ANIMATION, BUT WERE AFRAID TO ASK

Rick Lehtinen
Broadcast Engineering Magazine
Overland Park, Kansas

ABSTRACT

Computer animation has become so deeply entrenched in the video production business that it is difficult to imagine a newscast, news feature or sporting event without it. This paper discusses computer animation, classified by type. The paper then touches on how animation segments, or elements, are integrated into a finished production. The hardware used to produce computer animations is examined in a generic fashion, starting with the computer, and moving into how animations systems integrate into the "video backbone" of a broadcast plant.

I- Introduction

This paper briefly overviews the techniques used to generate computer animations. We will look at several types of animation, review the techniques used to create finished animation segments from raw graphic elements, then survey the video and computer hardware with which these graphic elements are created.

II- Types of animation

One can not lump all computer animation into a single category, any more than one can properly lump oils, acrylics, watercolors, tempera, airbrush and Chinese calligraphy as "paint".

One classification scheme might be to break down computer animation as follows:

1) Color-cycle animation

This type of animation works by changing the colors of individual segments of the screen. In some systems this is done by quickly changing the definitions in the color look-up table, or color palette. Changing to the background color makes the object seem to disappear. This can give the illusion of movement.

2) Two-dimensional or Cel animation

The artist first establishes key frames. The computer then performs the "in-betweening", that is adjusting the elements in the picture to complete the movement between keyframes. Several elements of the picture -- position, size, orientation, color, etc. -- may be animated.

3) Two-and-a-half-dimension animation

Not all art has to be "flat". Shapes can be drawn so as to appear three dimensional. These can then be moved as in two dimensional animation.

4) Three-dimensional animation

A set of points and the interconnecting curves is defined to be an "object". The object is then moved about in space, or the camera moved with respect to the object. Textures can be mapped to the objects surface, the object can be painted, and backgrounds can be supplied.

5) Metamorphosis

The shape of an object is changed over time to create a new object. In two dimensions, this is performed by using masks and careful painting. In three dimensions, this is done by varying the location of the points and curves in the underlying object.

6) Behavioral animation

It is difficult to keep track of the path of each object in a group. In behavioral animation, each object determines its own path with respect to the other members of the group, following a set of rules laid down by the artist.

7) Displacement animation

Points are bound to their neighbors according to artist-defined rules. When one point is dragged, the other points follow, resulting in surface distortions that mimic life. This means that one can animate the surface of objects, not necessarily the objects themselves.

8) Live Actors with animation keyed-in

Live actors can be mixed with animation for unique effects. This can be accomplished in post production with a switcher, or by "Rotoscoping" (see below).

9) Rotoscoping

Video images are frame-grabbed by the computer, processed, and then returned to the tape. This is greatly facilitated by the use of digital VTRs, or digital disk recorders, since these allow the use of multiple layers and re-recording without generation loss. Rotoscoping can be used to "retouch" video.

10) Combinations of animation

Just as an artist may chose to mix media for unique effects, computer graphic designers may chose to merge several forms of animation to create many variations of a design.

III- Creating Animation Segments

Only the most recent computer systems are capable of producing finished "pieces" of video. These modern systems automatically produce the "in-between" frames in real time. One system even includes a MIDI (Musical Instrument Digital Interface) output, which could eventually operate a synthesizer, for dropping in sound effects. Conceivably, these systems could find application in news graphics, where they would allow rapid creation of animations.

By far the majority of animations, however, exit the computer as single frames, which are either stored in the computer as picture files, or recorded onto video tape one frame at a time.

These elements are sometimes rendered as complete frames, including backgrounds, but other times are left as discrete images which are merged together in post production. The reason for this is that video techniques are much faster at creating wipes, fades, and digital effects moves than are computers.

In some cases, audio post production, or sweetening, is used after the video elements are combined. The result is finished video production.

IV- An Overview of the Hardware

One of the first big shocks awaiting anyone trying to produce professional quality animation, is that top quality art can only be conveyed with top quality, technically accurate, video signals. As broadcasters know, equipment that produces accurate video costs money. In fact, it is not unlikely that the "video backbone" of a computer graphics facility could cost as much as the computer graphics equipment itself. Fortunately, most broadcasters have access to much of the required equipment already.

The hardware elements of a generic computer graphics system are depicted in figure 1.

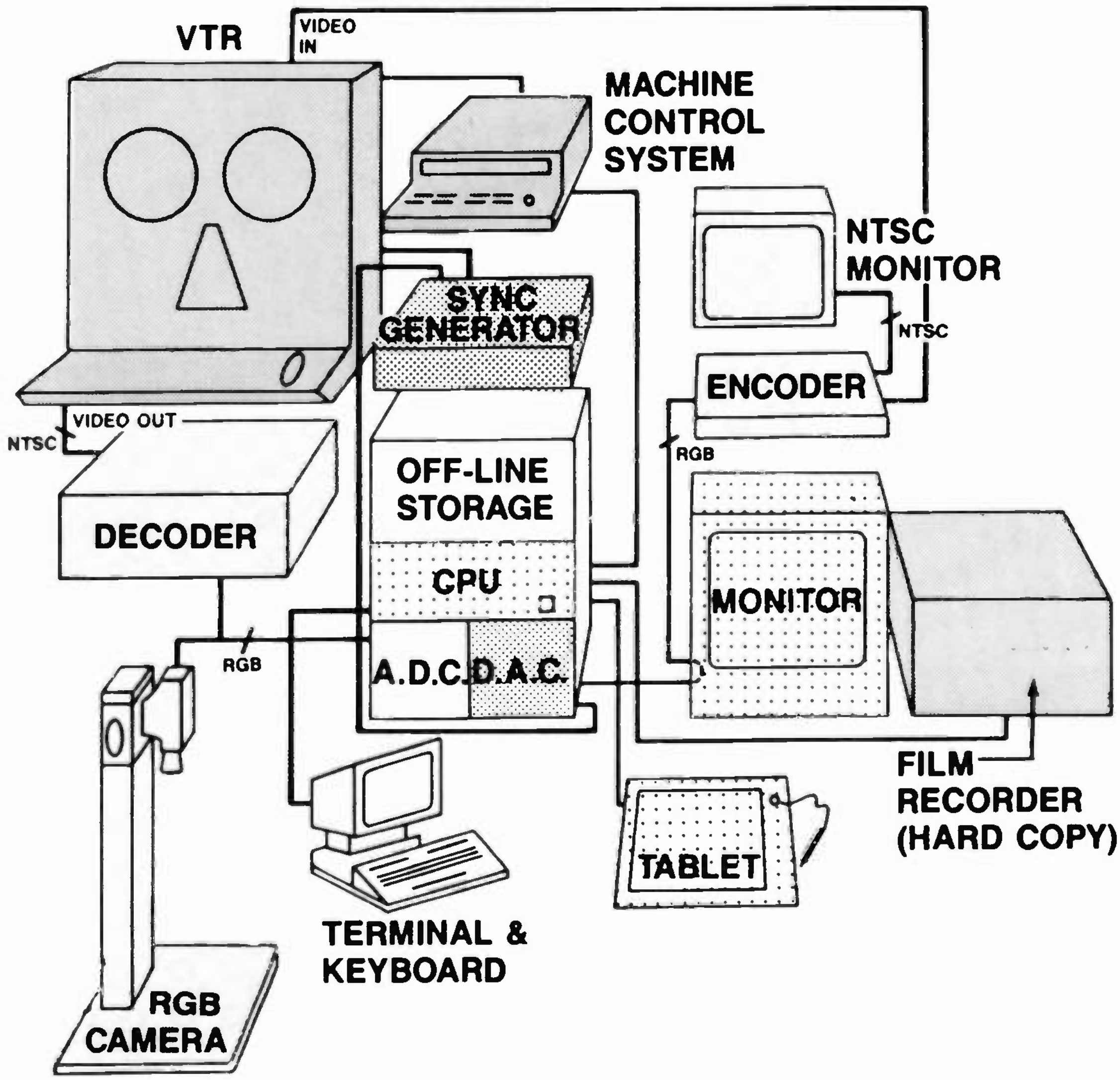
REFERENCES:

Video Systems Magazine, July, 1988, p. 62-76. (A further explanation of computer graphics hardware)

Broadcast Engineering Magazine, April, 1989, p.244-256. (A detailed discussion of NTSC decoders and encoders)

Journal of the Society of Broadcast Engineers--Proceedings of the 1988 SBE & Broadcast Engineering Conference, p. 131-149. (An additional discussion of computer graphics hardware, with a graphical development resembling the slides presented in this paper.)

(-All of the preceding are by the author.)



■**Section 1.** A computer graphics system is basically a computer configured to produce video. A graphics tablet gives the artist a tool for drawing. The monitor displays work in progress.

■**Section 2.** The DAC section receives digital output from the computer and converts it to analog video out. The output is shown here as RGB, but some systems output composite (NTSC) video, 4:2:2 (D-1) or D-2 digital video. A sync generator is needed to make stable, recordable output video. It also enables the system to be gen-locked to house and timed to destination.

□**Section 3.** Off-line storage (removable disk pack, floppy disk or tape drive) allows archival of images, backup (protection copies) of system's memory, transportation of images from one system to another and system security. With the terminal, the artist can perform system-level functions such as moving files and backing up the system. Artists also use the terminal to enter text information and motion macros. Engineers use it to run diagnostic programs and install software updates.

■**Section 4.** The encoder (or in analog component systems, the transcoder) converts RGB from the system into a format suitable for tape machines and switchers. The NTSC monitor is shown here as a separate unit, but may be part of a color paint monitor. Using a camera as a video source greatly increases the effectiveness of the artist. Analog-to-digital conversion hardware allows the outside world to talk to the computer. A decoder allows any composite video source to be fed into the system for digitization. RGB from the decoder follows the same path as RGB from the camera.

□**Section 5.** Because even simple animation segments may require hundreds of single-frame edits, a machine-control system is generally included with the system. A completed computer graphics system is capable of video or camera scan-in and video or hard-copy output.

Figure 1:
Computer graphic system block diagram

ZOOM LENS DESIGN FOR CCD CAMERAS—THE IMPLICATIONS ......AND THE CHALLENGE

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In addition to their well known operational benefits, CCD cameras now reach, and even sometimes surpass, the level of image performance of tube cameras.

However CCDs place new constraints on the lens performance. Unlike tubes, they are in a fixed location, and they do not allow any geometric processing of images.

Furthermore the CCD offers a completely flat response from the center to the edge.

This paper explores the new demands that are thus placed on optics (chromatic aberrations, distortion, MTF, vignetting) and illustrates how well presently available lenses can meet this challenge.

In addition to the inherent advantages of using chips instead of tubes, the level of image performance of CCD cameras continue to improve year after year. In this respect it can be said that CCD cameras now equal, and sometimes surpass, the tube cameras they replace.

Smear, image lag, blooming have been improved. The number of pixels has increased and resolution hits high levels.

Furthermore resolution is uniform from the center to the edges of the picture and stable with time. It is little sensitive to environmental conditions, which is an indisputable operational advantage.

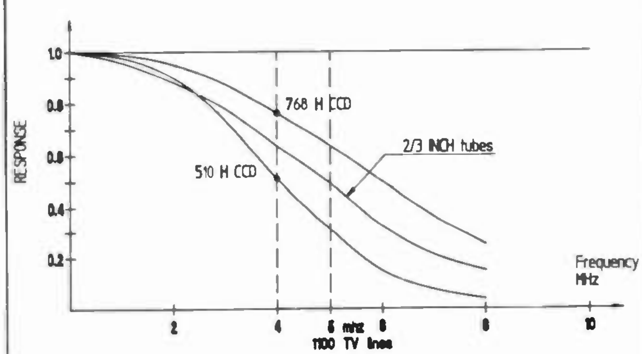
On the other hand some features of tubes have gone, thus placing an additional burden on optics. Lenses must now be fully corrected for chromatism and geometry.

It is of course worth mentioning here that trends of the future such as 16/9 aspect ratio, enhanced TV and HDTV will call for further progress in the very same directions.

This paper will therefore review what optics must do to cope with the cameras of today and how they reach the required level of quality. As a conclusion we will briefly see how these will have to evolve for the television of tomorrow.

1.- CCD MTF RESPONSE

Exhibit 1 shows the MTF response for a 510 horizontal pixels, a 768 horizontal pixels and a 2/3" PBO tube. The gain in MTF between the 510 pixels CCD and the 768 pixels CCD is considerable.



MTF CURVE OF 768 H CCD
COMPARED WITH 2/3 INCH PBO TUBE

EXHIBIT 1

The pixel dimensions being 11 x 11 microns, the optical image spot of a point must always be smaller. Furthermore the differential deviations, i.e. the distances between the three RGB images, must be much less, since there is absolute superposition of the three RGB pixels.

2.- CCD POSITIONING

For 1/2" cameras all the three CCDs are at the same distance from the lens. On the contrary the CCDs of 2/3" cameras are at different distance as can be seen on Exhibit 2.

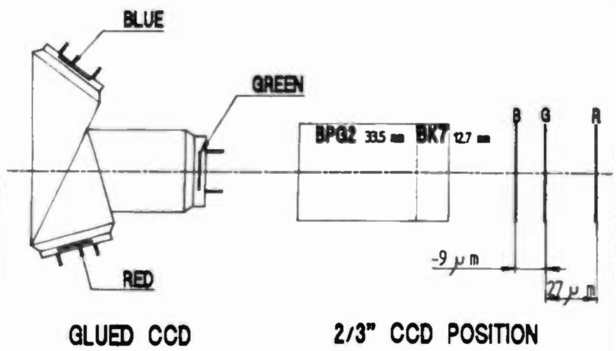
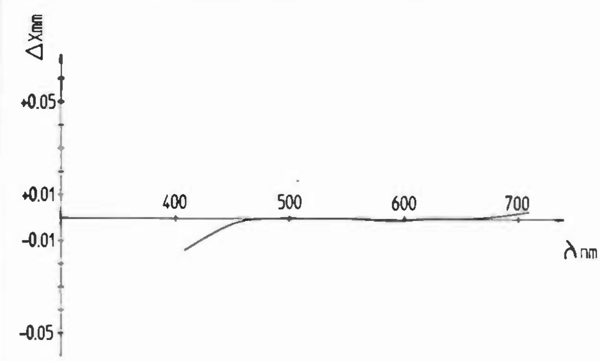


EXHIBIT 2



LONGITUDINAL CHROMATIC ABERRATION (APO)
angenieux 20x8.5 - tube 2/3" - F=8.5mm

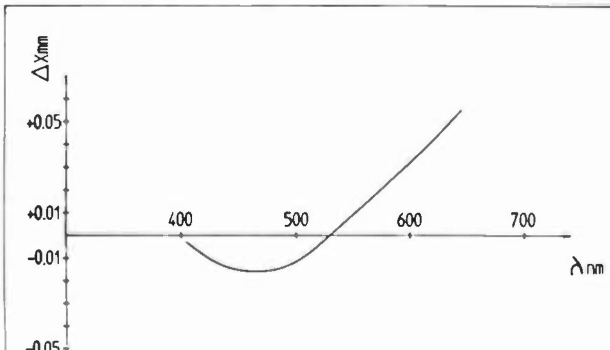
EXHIBIT 4

3.- LONGITUDINAL CHROMATIC ABERRATIONS

Tubes could be moved, CCDs cannot. The lens must therefore deliver a fixed image at the exact location of the CCD of each channel.

With tube cameras there always was some possibility to compensate for longitudinal chromatic aberration by relocating the tubes on each channel. With CCDs all the lenses of all types must have the same longitudinal color correction (although different for 1/2" and 2/3" as the sensors are not in the same plane).

This can be well illustrated with the curves of longitudinal chromatic aberrations for the Angenieux 20 x 8.5, a lens for 2/3" studio, when designed for a tube camera (Exhibit 3), or for a CCD camera (Exhibit 4).



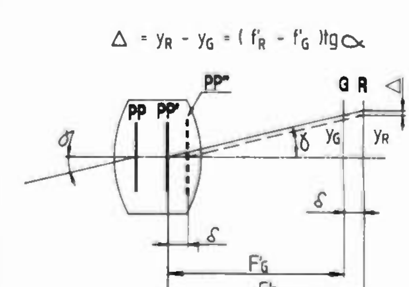
LONGITUDINAL CHROMATIC ABERRATION (standard)
angenieux 20x8.5 - CCD 2/3" - F=8.5mm

EXHIBIT 3

It must be mentionned that the constraint of differential back focus in the 2/3" cameras for the three color channels is not a problem, perhaps even to the contrary, for the short focal lengths, i.e. lenses for ENG and studio.

However this is more and more of a penalty on the lens MTF performance as longer and longer focal lengths are required. For instance, present 2/3" lenses for OB sports reach a focal length of slightly more than 1000 mm. At these focal lengths and for a perfectly corrected lens, at 5 MHz and in the center, the MTF shows a loss in the blue of 18% with F5 glass, or 6% in the red with BPG2 glass as compared to what it would be if the three CCDs were in the same plane.

The other problem (Exhibit 5) with the three chips in different planes is that different back focus, i.e. different focal lengths, result in different image sizes, or lateral chromatic aberrations. To avoid this the optical designer must shift the principal plane. Something much easier to say... than to do!



EFFECT OF THE LONGITUDINAL DIFFERENCE
ON LATERAL CHROMATISM

EXHIBIT 5

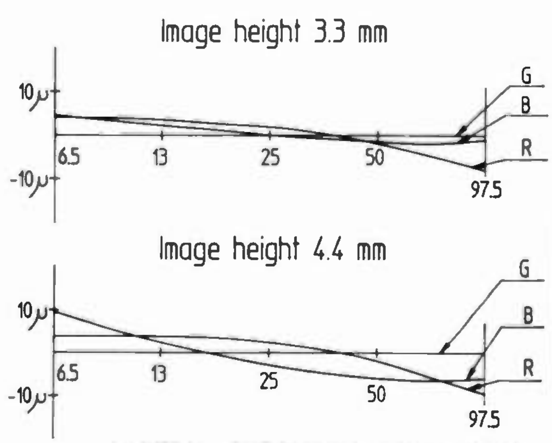
4.- LATERAL CHROMATIC ABERRATIONS

The three images of the tube cameras can be magnified independently. Not with CCD cameras.

A tube camera can thus correct to some extent, the lateral chromatic aberration of lenses (i.e. RGB images of different sizes). With a zoom lens lateral chromatic aberration varies with focal length. Sophisticated tube cameras do provide for real time correction during zooming.

Now CCD cameras, unlike tube cameras, call for lenses which are totally corrected for lateral color. Total correction means that deviations of the blue and the red images from the green image must not exceed the CCD resolution of one pixel (11 microns).

As an illustration, Exhibit 6 shows it is well the case with the Angenieux 15 x 6.5, a new budget high performance studio lens which is introduced in 1990.



LATERAL CHROMATIC ABERRATION
angenieux 15x6.5

EXHIBIT 6

These remarkable performances can only be achieved with exotic low dispersion glasses (i.e. with lowest change in refractive power with light wavelength), as well as with flint glasses of highly specific dispersion laws.

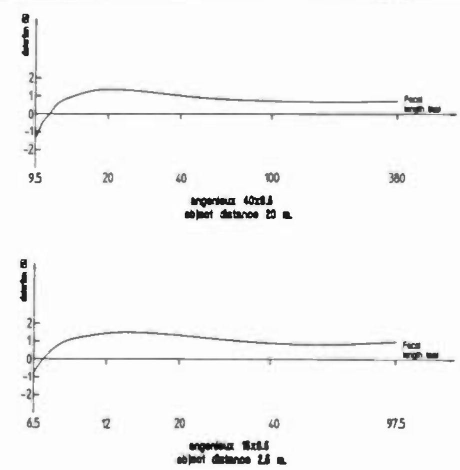
5.- OPTICAL DISTORTION

The geometry of tubes is... what it is, but the image can be processed geometrically. The CCD will show the image of the lens, as is, neither better, neither worst.

The lens must therefore show a straight line as a straight line and a circle as a circle. With a zoom lens the greatest difficulty is to maintain this over the whole zoom range.

This is achieved with more lens elements and with aiming at stabilizing the entrance and exit pupils when the angles of field vary. Aspherics would be ideal, but unfortunately are not available today within the dimensions and the cost that would make them useful.

Exhibit 7 shows the level of distortion of the new Angenieux 15 x 6.5 at the usual studio working distance of 2.5 meters. Although this lens achieves the extremely wide angle of 68.5° it is extremely well corrected for distortion.

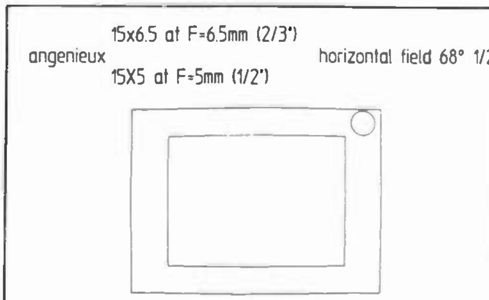


DISTORTION

EXHIBIT 7

As another reference Exhibit 7 shows also the Angenieux 40 x 9.5. a microprocessor controlled OB lens at a distance of 20 meters. It is to be noted how low the distortion is, particularly at the long focal lengths.

Exhibit 8 shows the actual shape of a rectangle, as will be seen with the new Angenieux 15 x 6.5 (for 2/3") of 15 x 5.5 (for 1/2") at the extreme wide angle of 68.5°. The lack of visible distortion is quite remarkable.



EFFECT OF THE DISTORTION ON THE ASPECT OF THE IMAGE

EXHIBIT 8

6.- EVEN PERFORMANCE FROM CENTER TO THE EDGE

With regards to image quality, tubes show a loss from the center of the image to the edge. CCDs provide a picture which is totally uniform.

To take full advantage of this, lenses should give images which are more even in resolution, illumination and colorimetry.

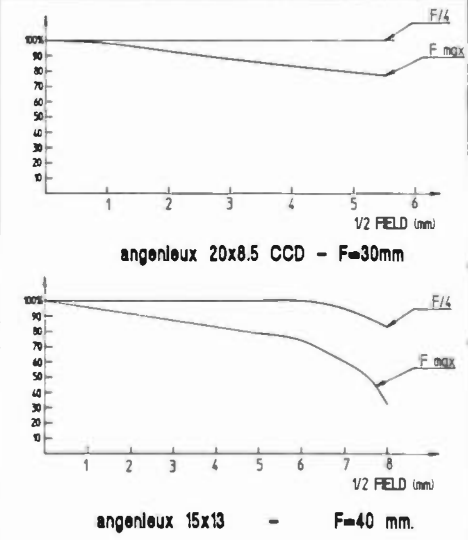
Considerable progress has been achieved in maintaining a high optical quality from center to the edges in the lenses which have been recently developped for CCDs. Coupled with the even performance of the chips the results are spectacular. Exhibit 9 compares the typical numbers of MTF for a 15 x studio lens of the previous generation (but still in production in 1989), with the new Angenieux 20 x 8.5. For a similar sensor MTF normalized response of 100% at the center, the value at the edge improves from 20% to 85%!.

5 MHz MTF	CENTER	CORNER
Tube normalised	100%	50%
15x13 t:1.5 actual	90%	40%
Chain	90%	20%
CCD normalised	100%	100%
20x8.5 t:1.3 actual	90%	85%
Chain	90%	85%

MTF VALUES
WITH TUBE AND CCD CAMERA

EXHIBIT 9

The comparison of these two lenses, both of which were developped to be used with high performance studio cameras is also interesting in another aspect: vignetting. Vignetting is the loss in the uniformity of illumination from the center to the edge of the picture. Exhibit 10 shows the substantial improvement which happened between these two lenses.



VIGNETTING

EXHIBIT 10

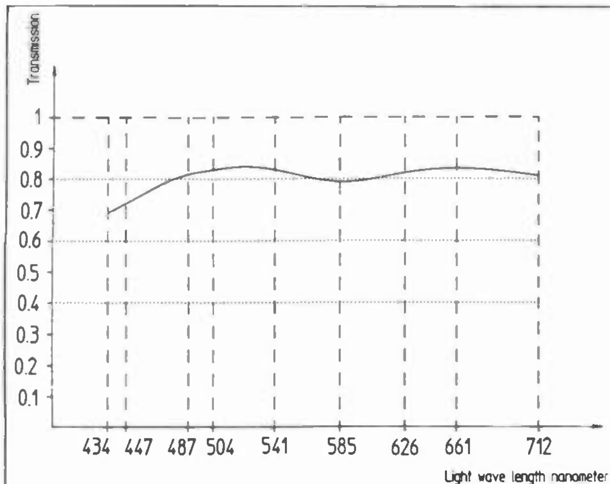
Vignetting is a direct function of the diameter of lens elements. In this case the diameters should be increased by more than half their present values to totally eradicate vignetting. For obvious reasons of weight and cost this is not possible and cameras now provide for a compensation of up to about 50%.

Concerning image uniformity a final point is to be made. In order that image colorimetry is uniform over the whole picture, it is necessary that beams of light reflected by the prism dichroics surfaces impact these surfaces at similar angles, whichever part of the image is concerned. Principal rays must be inclined by no more than 2 degrees on axis. This implies that particular care is taken in the physical location of the iris.

7.- LENS SPECTRAL TRANSMISSION TO FIT THE SENSORS

With CCD cameras the blue channel has intrinsically a low sensitivity. Optical materials and coatings will therefore be carefully selected to avoid penalizing the blue in any way.

The spectral transmission of the Angenieux 14x7, and ENG lens for 1/2" CCD (Exhibit 11) shows how flat a response can be obtained despite the strong blue absorption found in all optical glasses.



SPECTRAL TRANSMISSION angenieux 14x7/14x9

EXHIBIT 11

8.- 16/9 ASPECT RATIO WITH THE CCD OF TODAY

On the way to HDTV, there is an interest in the 16/9 aspect ratio. Chips of this geometry are not presently available. Besides, broadcasters may wish to use the same camera for both 4/3 and 16/9 aspect ratios.

A similar solution as with film is available: anamorphot optics. With cylindrical elements it is possible to increase the horizontal field of view by one third and condense the information on the length of horizontal lines.

This can be done either with placing the cylindrical elements in front or at the rear of the lens. Like with all optical attachments, front anamorphot maintains the lens full aperture but is bulky. Front anamorphot is therefore to be preferred as long as size remains acceptable. Rear anamorphot might be necessary for OB lenses.

9.- HDTV

CCDs clearly appear to be one of the keys of HDTV in the future. All which has been said here will apply, but with much tighter requirements.

At the same time, there will be a need for higher quality level and yet operational demands cannot be anticipated to go down. Directors will insist, and very rightly so, on long zoom ranges, high apertures, very wide and narrow angles of view, very close minimum focusing distances etc.

These demands are, more often than not, inherently technically contradictory. It is thus a totally safe prediction that there will be no shortage of activity for the television optical engineer in the years to come.

NEW ELEMENTS THAT PROVIDE PATTERN VERSATILITY IN THE COAX AND WAVESTAR ANTENNAS

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ABSTRACT: Bent dipoles and loop elements attached to the mast and excited by a slot, have been found to have very useful properties for cylindrical slot array antennas employed in TV broadcasting. By inserting these elements, the inherent slot pattern can be controlled and, in addition, a vertical component can be obtained for CP broadcasting.

A variety of azimuthal H-pol and V-pol patterns have been achieved, showing great versatility, including more directional pattern of the "cosϕ" type for higher gain per element and less mutual coupling between adjacent slots. In the case of full CP, good axial ratio in the overall radiation pattern has also been accomplished.

Finally, negligible changes in the slot array bandwidth were observed by introducing these forementioned elements.

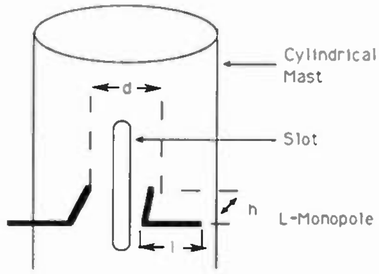
INTRODUCTION

In 1974 Alvin Clavin, D.A. Huebner and F.J. Kilburg (1) suggested the use of a new element composed of a pair of monopoles (straight or L shaped) excited by a slot in array applications due to many desirable properties exhibited. Among these properties, E and H plane element pattern control as well as reduction in inter-element mutual coupling have been achieved by using these elements in rectangular waveguides (X band). Also a complete analysis of two straight monopoles excited by a narrow slot located in an infinite conducting ground plane was carried out by A.B. Papierz, S.M. Sanzgiri, and S.R. Laxpati (4). In both papers, the slot is excited by a TE_{10} mode. In this approach, the same type of elements have been used, but with some change in geometry to account for the cylindrical surface of the pipe, the array factor and the generation and shape of both the H and V pol pattern. In this paper, the slots were excited by a TEM mode (coax) and a TM_{01} (waveguide), respectively. Some of the forementioned properties have been confirmed and, in addition, V-pol pattern control (amplitude percentage, phase and axial ratio) has been achieved. In addition, a slot-loop radiating element with a slightly different shape has been used as compared to the one suggested by M.L. Fee (3).

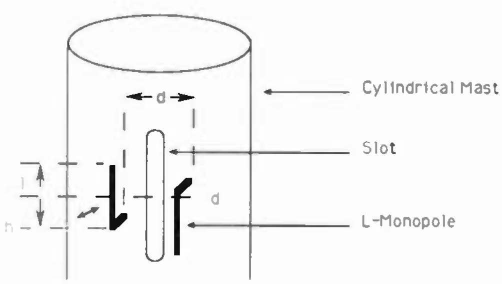
ELEMENT DESCRIPTIONS AND CORRESPONDING PATTERNS

Pair of Monopole Elements:

The basic elements, used for these experiments, consisted of a pair of L inverted monopole rods connected to the mast on both sides of the slot. Two different configurations have been used, one symmetrical and horizontal, as shown in Figure 1 and a second, asymmetrical and vertical (Figure 2a) for the generation of both H and V-pol E-field. By adjusting the distance d between the elements, the height h and length l, the V-pol amplitude and phase, as well as the H-pol pattern can be controlled.

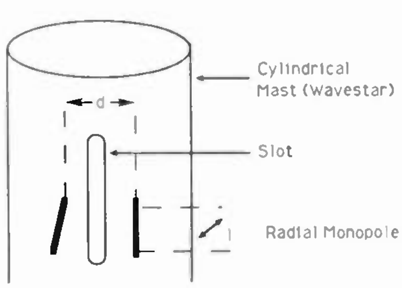


SLOT EXCITED HORIZONTAL L-MONOPOLES
FIGURE 1



SLOT EXCITED VERTICAL L MONOPOLES
FIGURE 2a

In addition, some promising results (not published) have been obtained by using straight radial monopoles on each side of the slot as shown in Figure 2b.



SLOT EXCITED RADIAL MONOPOLES
FIGURE 2b

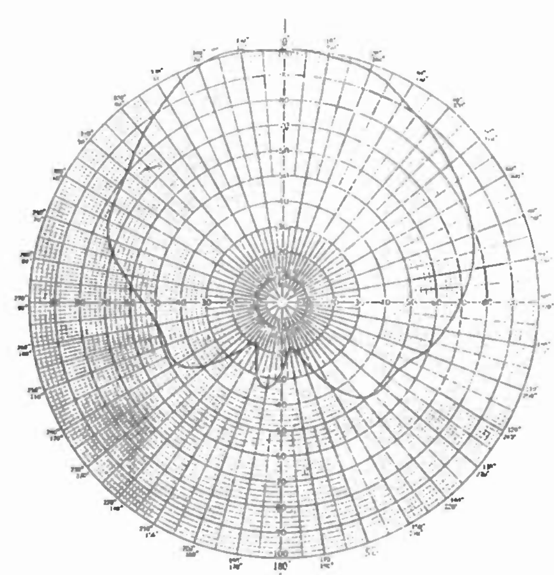
Test Antennas

The patterns on these elements were taken at the Harris Antenna Test Range in Palmyra, Missouri using a short range facility.

Basically, two types of slot antennas were used as test pieces: a Coax CH 53 (OD = 8-5/8"), 4 bay peanut array (for full CP), and a CH 61 (OD = 15", TM_{01} mode), 1 bay trilobe array (H-pol). The forementioned antennas were used in the transmitting mode and a rotating log periodic antenna was used as a receiving element for H or V-pol. The antennas were 90' apart and 20' above the ground (short range). Some of the patterns were measured and the others calculated on a PC computer by using actual measured single element amplitude and phase E-field data.

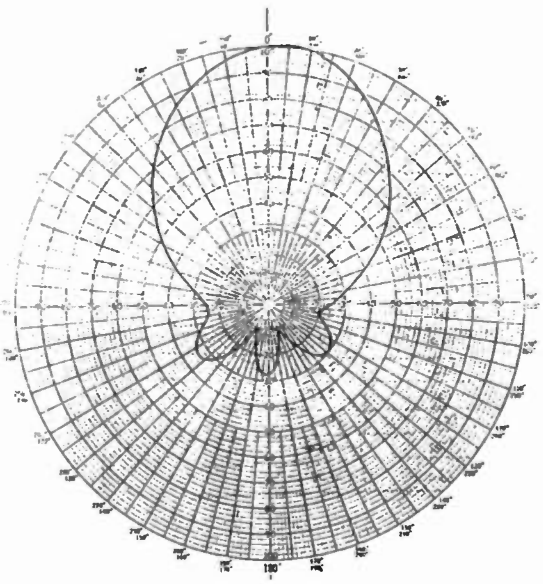
Trilobe Pattern Tests

Figure 3 shows a single slot element pattern corresponding to the CH 61 pipe without monopoles. By inserting two horizontal L monopoles on each side of the slot, and by adjusting the forementioned dimensions, a "cosΦ type" of pattern has been achieved

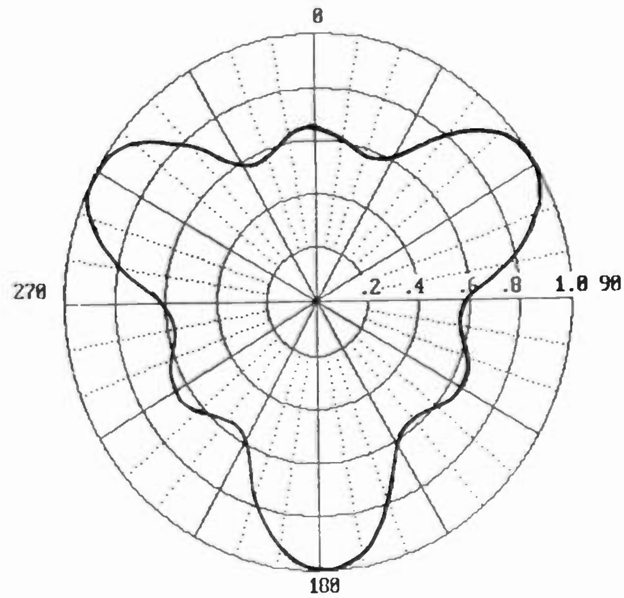


SINGLE SLOT WAVESTAR CH 61 ELEMENT PATTERN
FIGURE 3

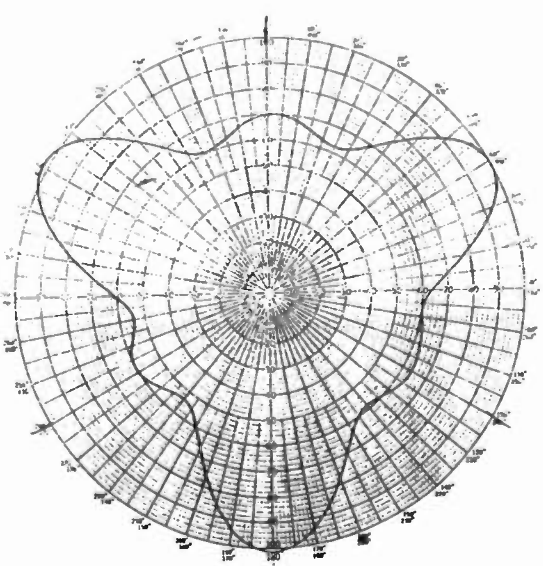
(Figure 4). As shown, less radiation occurs in the tangential direction as well as in the back of the slot plane. Using the same element, in all three slots 1 bay around, a trilobe pattern has been obtained (Figure 5a). Figure 5b shows the corresponding calculated pattern by using single element amplitude and phase data. The nulls in the calculated pattern are slightly deeper than in the measured, probably due to neglected mutual coupling and mast scattering. In contrast, 5c (measured pattern) shows a narrow six lobe pattern with deep nulls when no monopoles are inserted. By comparing the forementioned figures, a drastic improvement in the overall pattern has been achieved without changing the pipe size and by using only three slots.



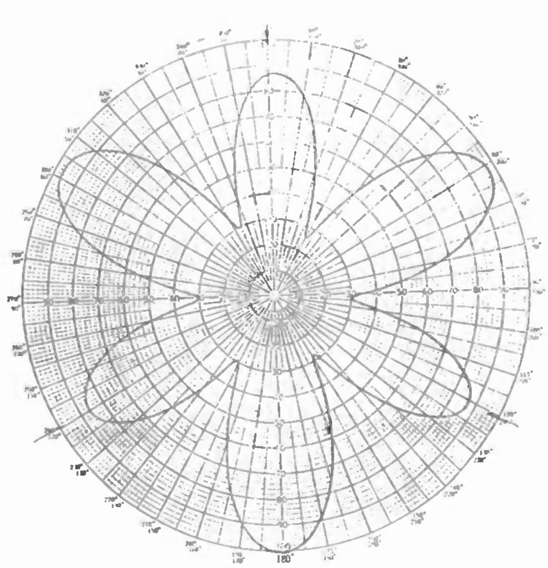
SINGLE SLOT AND DIPOLE CH 61 ELEMENT PATTERN
FIGURE 4



CALCULATED SLOT
AND H-DIPOLE TRILOBE PATTERN
FIGURE 5b



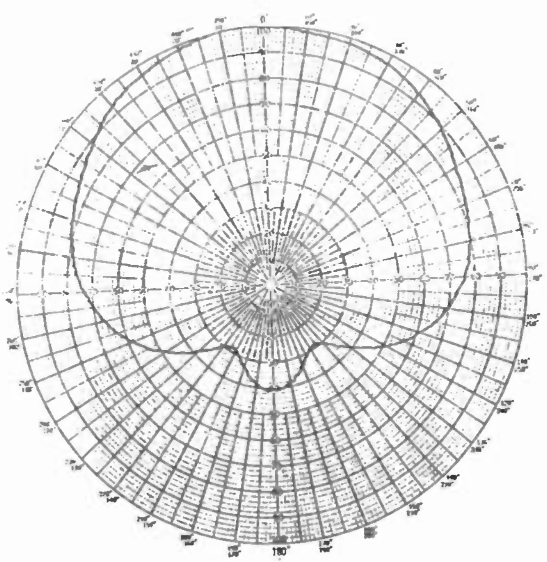
SLOT AND DIPOLE CH 61 TRILOBE PATTERN
FIGURE 5a



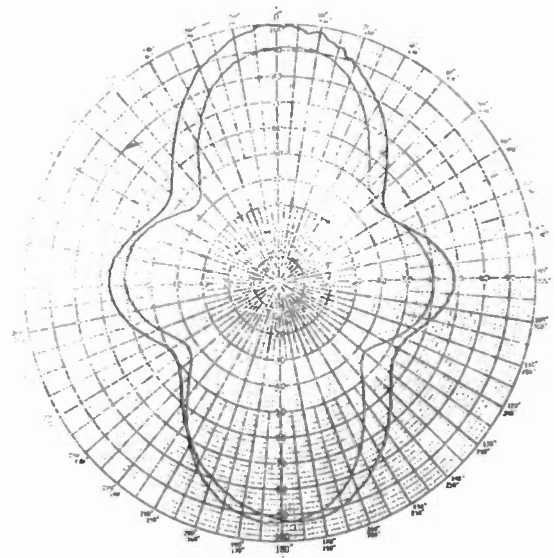
HEXALOBE PATTERN WITH THREE SLOTS ONLY
FIGURE 5c

Cardioid and Peanut Pattern Tests

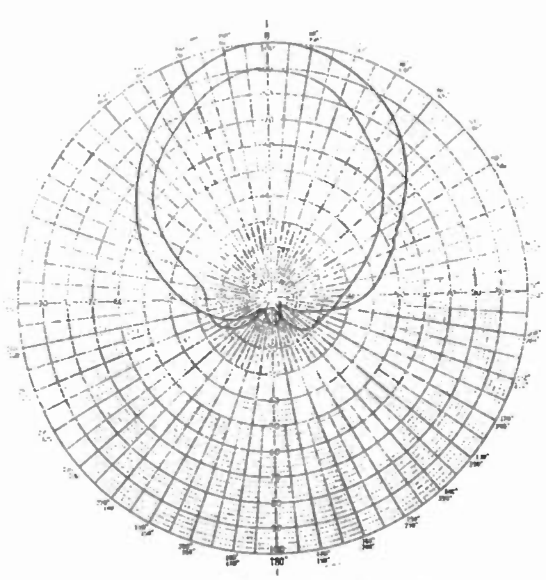
Figure 6 shows a cardioid pattern corresponding to a single element in a CH 53 coax antenna. By inserting the two vertical L monopoles radially and vertically (11" overall length) on both sides of one slot and displaced off center, and by adjusting h,e,d and l, a directional H as well as V-pol pattern, almost full CP was achieved, as shown in Figure 7. In Figure 8a, a full CP peanut pattern was accomplished by using the same forementioned elements on all 4 bays of the CH 53 coax antenna. As can be noticed, a good tracking between H and V pol is achieved due to the similarity between the element H and V pol patterns. Figure 8b shows a measured H-pol peanut pattern without the dipoles. By comparing the patterns of Figure 8a and b respectively, it can be immediately noticed that the mainlobes are rotated 90°. In case of



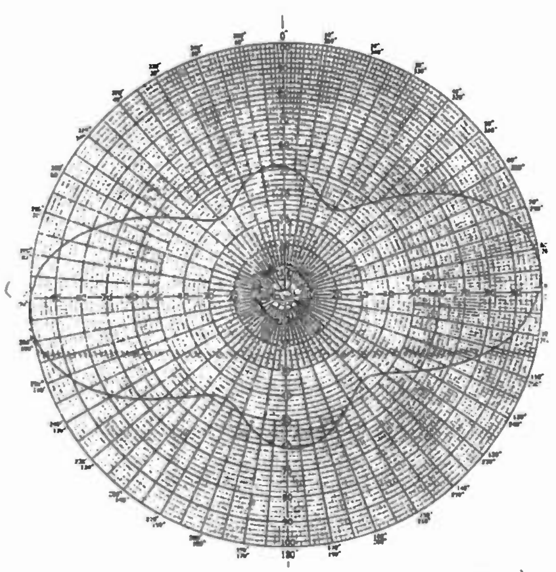
SINGLE SLOT PATTERN OF COAX CH 53
FIGURE 6



H AND V-POL PATTERN OF 4-BAY PEANUT ARRAY
(SLOT AND DIPOLE ELEMENTS)
FIGURE 8a

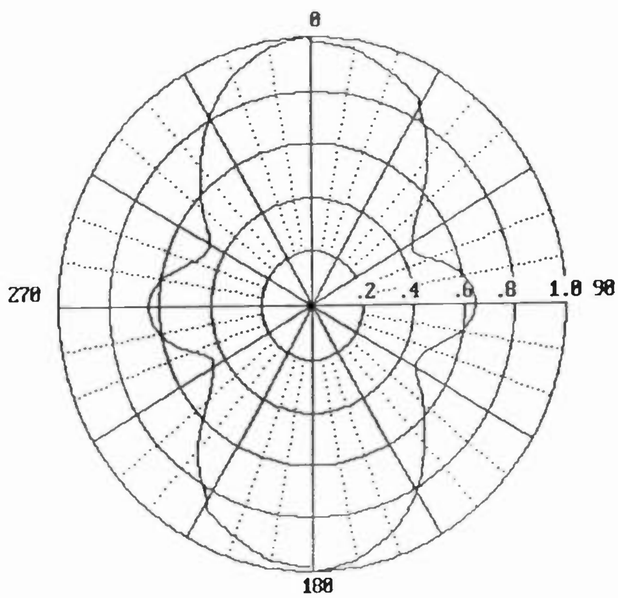


SINGLE H AND V-POL PATTERN
OF SLOT AND V-DIPOLE ELEMENTS
FIGURE 7

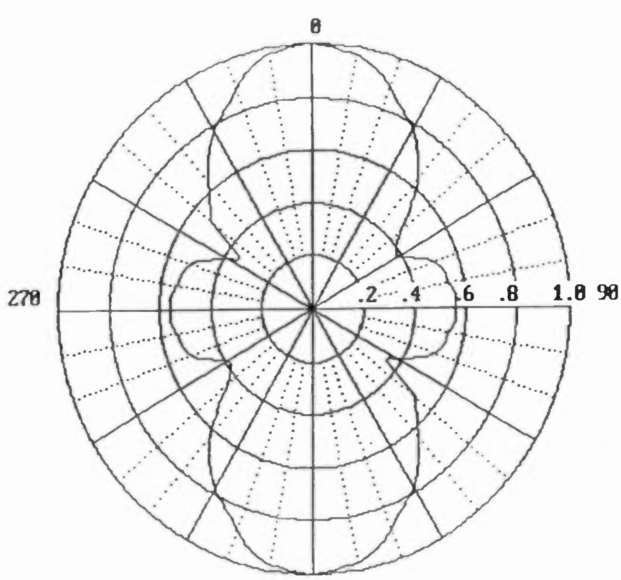


4-BAY PEANUT ARRAY PATTERN (SLOTS ONLY)
FIGURE 8b

Figure 8a, they are located on boresight of each slot which means less mutual pattern interaction (coupling) between adjacent slots per bay occurs due to the insertion of the monopole elements. Figures 8c and 8d show the V and H-pol patterns, calculated from measured data, which are very similar to the Figure 8a patterns, respectively. The main difference again is in the null depth as discussed before in the trilobe pattern. As previously shown, control in the H-pol as well as in the V-pol pattern has been achieved.



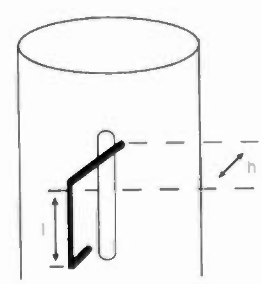
CALCULATED H-POL PEANUT PATTERN
FIGURE 8c



CALCULATED V-POL PEANUT PATTERN
FIGURE 8d

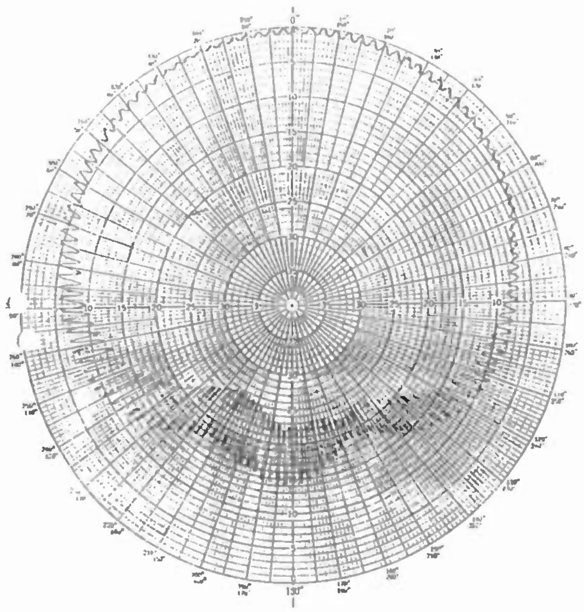
Loop Element

The use of a crossed loop excited by a slot was first suggested by M.L. Fee (3) for generating a V-pol component in the slot E-field. Figure 9 shows the actual loop configuration used in our tests. By adjusting the distance (d), Height (h), and length (l), different percentages of V-pol components can be achieved.

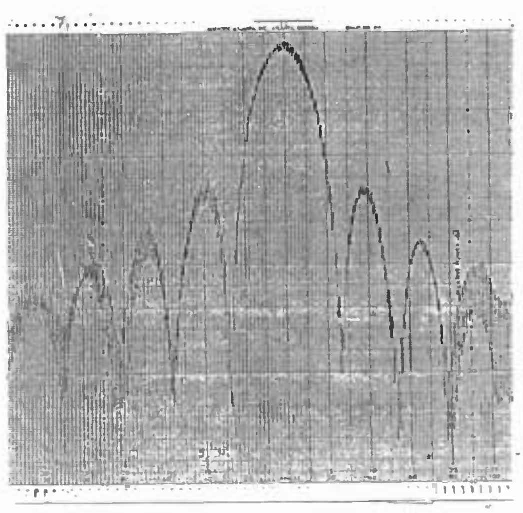


SLOT EXCITED LOOP
FIGURE 9

Tests were conducted on the 4 bay CH 53 Coax antenna and a 10-bay rectangular waveguide (WR 1150) CH 40 antenna. Figures 9a and 9b are axial ratio measurements of the azimuth and elevation pattern respectively, of the 10 bay full CP cardioid array. The azimuth pattern has an axial ratio of 1 dB in ±40 degrees off boresight. On the other hand, the elevation pattern of both H and V-pol are very similar within the first two sidelobes as

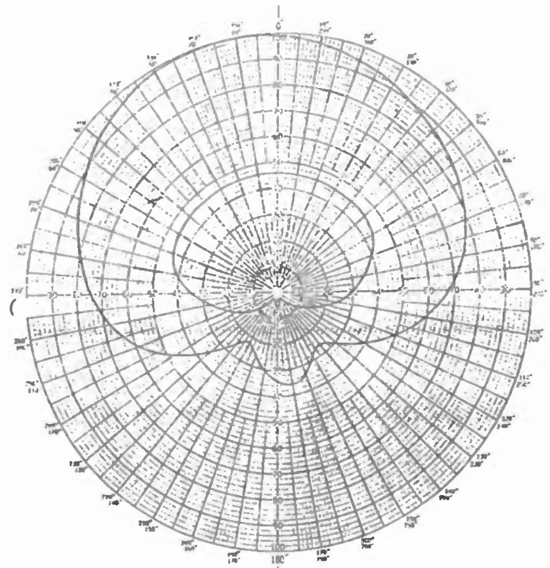


CP CARDIOID AZIMUTH PATTERN
OF SLOT AND LOOP ELEMENTS
FIGURE 9a



CP CARDIOID ELEVATION PATTERN
OF SLOT AND LOOP ELEMENTS
FIGURE 9b

shown in Figure 9b. In addition, Figure 10 shows a cardioid pattern with 20% V-pol obtained on the 4 bay coax antenna. It was found that the loop requires fewer adjustments to obtain a given pattern as compared to the monopole, however, it has less pattern control flexibility.



H AND V-POL (2092) CARDIOD PATTERN
FIGURE 10

IMPEDANCE AND PATTERN BANDWIDTH TEST

The CH 53 coax antenna peanut array (Figure 11) was used for the impedance and power tests. The antenna was laid horizontal on workhorses and the match was optimized with the aid of a network analyzer. The piece was tested with and without radomes as shown in Figure 12a and Figure 12b respectively.

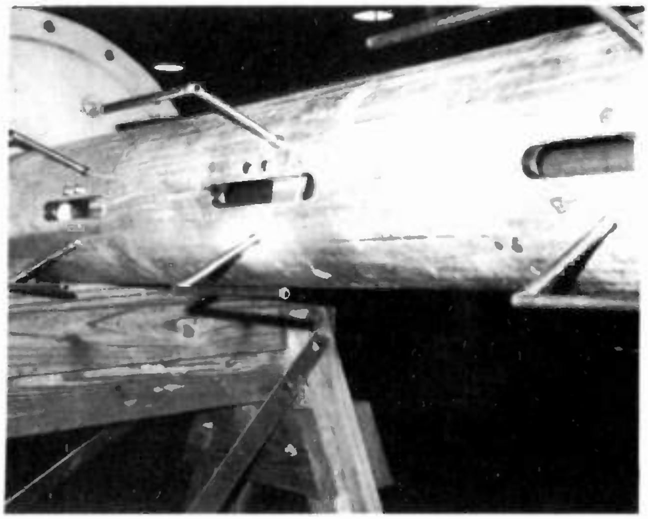


FIGURE 11

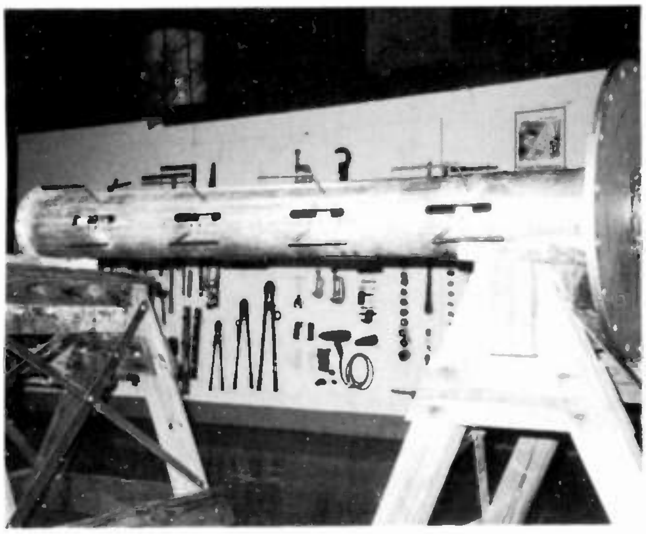


FIGURE 12a

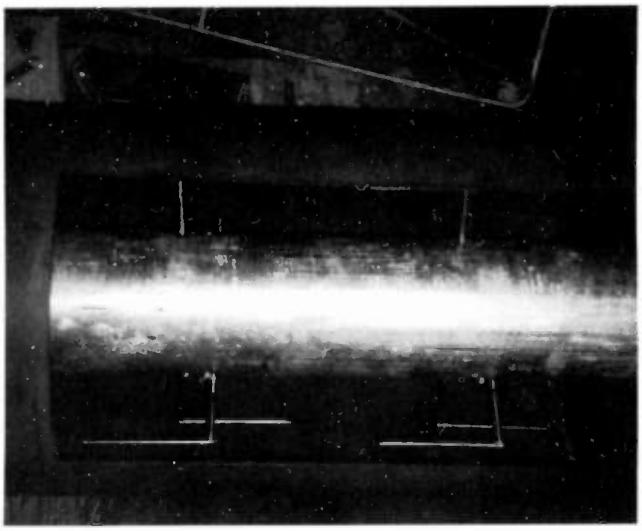


FIGURE 13

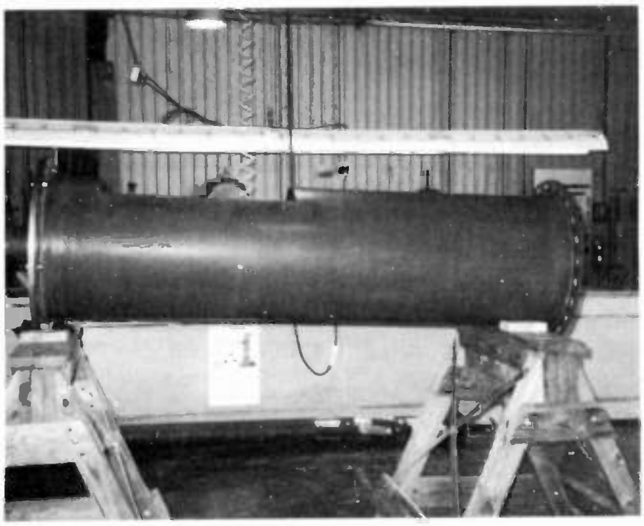
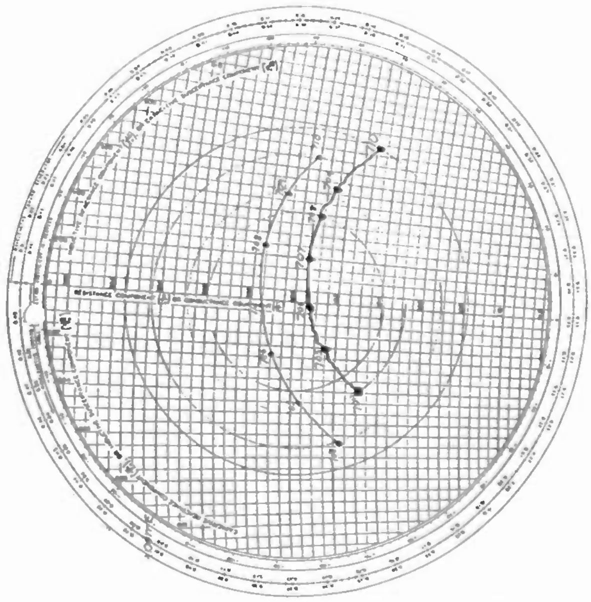


FIGURE 12b

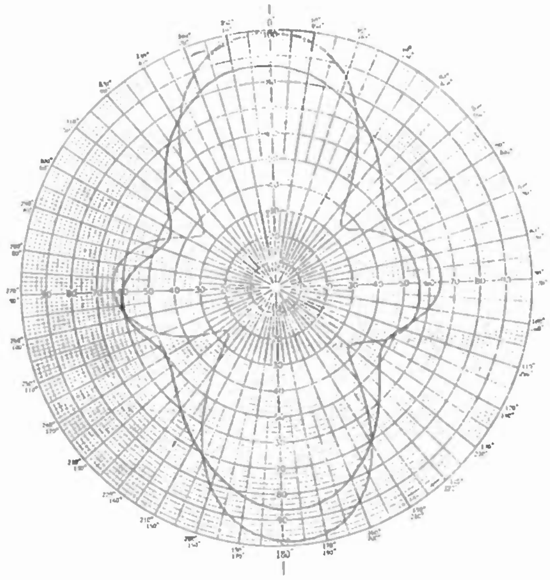


MATCHING PLOTS WITH AND WITHOUT RADOME
FIGURE 14

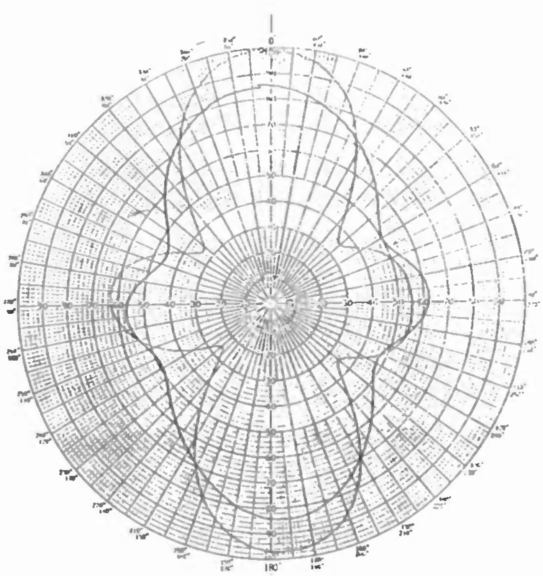
Figure 13 depicts the coax mast and the corresponding dipole elements inside the radome. The match under both conditions is shown in Figure 14. As can be immediately observed, the bandwidth is practically unchanged (1.07 and 1.08 respectively).

The pattern bandwidth was measured on the peanut array with thinner elements, but keeping the same former dimensions. As a consequence, a change in the peanut pattern did occur, as can be noticed by comparing

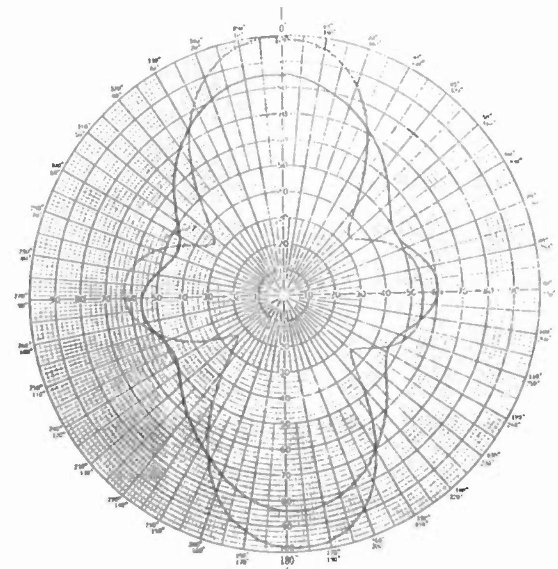
Figures 15 and 8 respectively (2). Figures 15a, 15b, and 15c show the corresponding H and V patterns for the frequencies 704, 707, and 710 MHz respectively. As can be noticed, only negligible changes occurred in both H and V-pol patterns when the frequency was increased from 704 to 710 MHz.



H AND V-POL CH 53 PEANUT PATTERN
AT f = 704 MHz
FIGURE 15a



H AND V-POL CH 53 PEANUT PATTERN
AT f = 707 MHz
FIGURE 15b



H AND V-POL CH 53 PEANUT PATTERN
AT f = 710 MHz
FIGURE 15c

Power Test

The 4 bay CH 53 coax peanut array with slot and monopole elements was also tested for power performances in the Harris transmitter lab. The antenna was located outside the building, laying on workhorses as shown in Figure 16a and Figure 16b respectively.

FIGURE 16a

Figure 16a shows the tested antenna fed to a 60 kW transmitter located inside the building, via a rigid transmission line. Figure 16b is a front view of the coax antenna laying on wood supports to separate itself from the radome. The antenna was tested under a radome and with the transmitter running at full power for three hours. No corona or hot spots were observed.

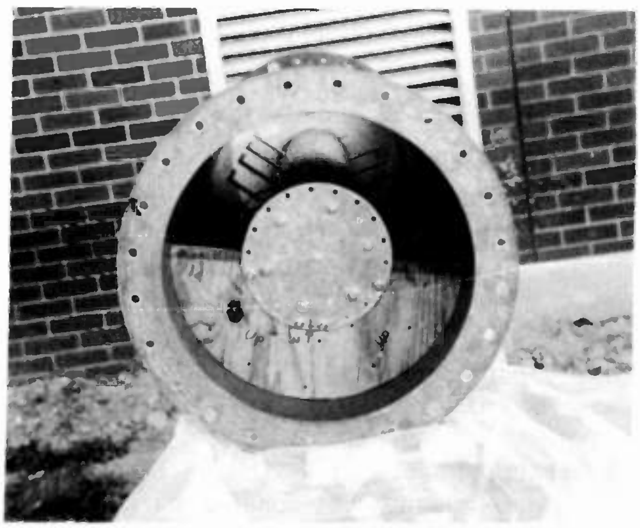


FIGURE 16b

COMMENTS AND CONCLUSIONS

Very interesting and encouraging results have been obtained with the forementioned monopole configurations on the coax and wavestar antennas. Other geometries have been tried to expand the range of different patterns available.

Also, the use of straight radial monopoles looks like an effective and efficient way to achieve, besides some different patterns, better circularity in current omnidirectional patterns (for the same pipe size) especially at frequencies off center.

From the obtained results, the following conclusions can be drawn:

a) The monopole and loop geometries lead to great pattern flexibility in H-pol and V-pol for both the Wavestar and coax antenna while maintaining the same pipe size.

b) As the elements are excited by the slot, both H and V-pol elevation patterns are practically the same.

c) Also due to the forementioned excitation, no additional coupling devices are required, which increases the reliability.

d) Larger diameter radomes have to be used in order to host the excited elements.

e) As shown in the test, the new elements have good power handling capability.

f) Acceptable impedance and pattern bandwidth performance has been shown.

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ACKNOWLEDGEMENT

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NEW ADAPTIVE DIGITAL TECHNOLOGY FOR INTERFACING PRODUCTION INTERCOM SYSTEMS TO DIAL-UP TELCO LINES

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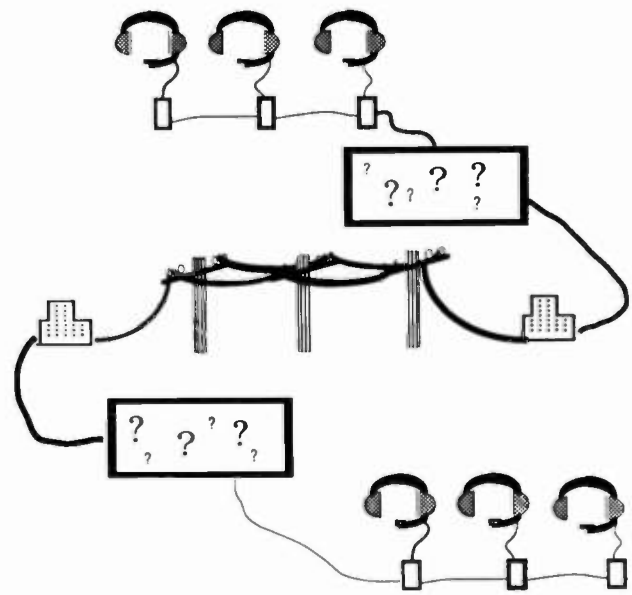
A state-of-the-art solution to the difficult problem of interfacing production intercom systems to dial-up telephone lines is presented. In contrast to ad-hoc solutions which often exhibit problems with achieving desired levels and/or feedback, this solution uses a digital signal processing approach in order to allow appropriate gain to be added to each of the talk paths. A prototype unit which uses the digital approach and integrates all of the functions required for interfacing is described.

INTRODUCTION

Often, in television production, we have the desire to connect a "belt-pack" intercom system to dial-up telephone lines. Smooth integration of live news remote feeds, for instance, requires that production personnel at all locations be able to communicate with each other in a simple, troublefree fashion. This is especially true when multiple remote sites are involved, as for election coverage, major sporting events, and telethons. Ideally, we would like to have our production crews at each location use the familiar intercom systems, such as those made by RTS, Clearcom and others, as if they were simply connected to a common bus - without regard for the distances involved. Most often, access to the dial-up phone network is available by wire or cellular. So, if we could find a way to reliably make the two systems play together, we would have the solution.



We want to connect our production intercom systems to the dial-up phone network. What do we need to put in the boxes?

It would seem, at first thought, to be a simple matter to accomplish this. Unfortunately, as many experi-menters have discovered, some important difficulties intrude. Let's first catalog what needs to be done in an interface appropriate to the task:

- Provide effective hybrid functions for both the intercom and phone line so that gain may be inserted without feedback.

• Provide physical and electrical interface to the intercom system.

• Provide interface to the phone line, with an auto-answer and disconnect function desirable.

• Correct for varying levels with AGC, preferably in both directions.

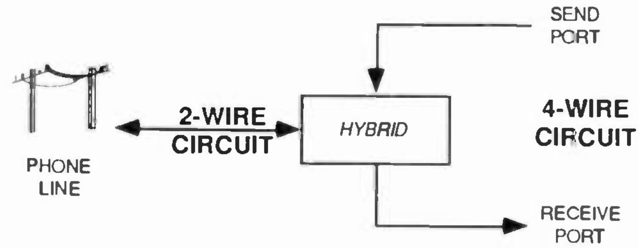
• Provide metering for signal indication and troubleshooting.

It is item number one which causes the most trouble - if the hybrids do not have high performance, level problems and/or feedback are sure to result. Ad hoc solutions are often beset with problems in this area. Since effective hybrid performance is critical to troublefree operation, this discussion will center upon the issue of hybrid use and implementation in an intercom interface system.

HYBRIDS AND INTERCOM INTERFACE

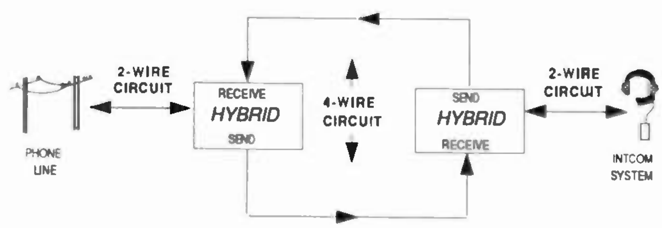
A standard phone line is a *two-wire* circuit. The term comes from early telephone engineering and refers to the fact that a single circuit consisting of two wires is used for both the send and receive speech signals. This is in contrast to a *four-wire* system which would use two circuit pairs with separate, independent paths for each signal direction.

The purpose of a hybrid is to turn a two-wire circuit into a four-wire circuit. That is, to separate the two signal directions into the individual components. We need to do this in order to provide amplification and gain control processing in the two signal paths. Without a hybrid, it would be impossible for our intercom interface to provide any gain, since an amplifier is certainly one-way only!



The purpose of a hybrid is to separate send and receive signals from the phone line.

In fact, our interface is going to require *two* hybrids since the commonly used belt-pack intercom systems are two-wire also. We will want to connect the hybrids "back-to-back" so that the intercom hybrid's receive output is fed to the phone line hybrid's send input and vice-versa with appropriate gain and processing stages inserted in the middle. This system is what telephone engineers call a *two-wire to two-wire repeater*.



Back-to-back hybrids form a 2-wire to 2-wire repeater *which allows gain and signal proc-essing to be inserted into the 4-wire path.*

From the diagram, it is clear to see where problems with feedback can result. If the send-to-receive isolation, or *trans-hybrid loss*, of the hybrids is not at least as great as the gain in the amplifiers, the signals feed around the loop and feedback easily builds-up.

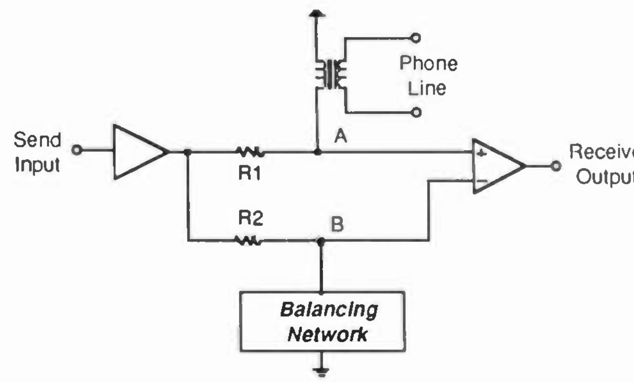
Why gain is necessary

Telephone circuits have widely varying and unpredictable end-to-end transmission characteristics. According to statistical studies of the telephone network, average telephone connection loss has been found to be 16 dB (8 dB per loop at each end), with some connections in the 25 to 35 dB loss range. Indeed, telephone loops are generally designed to have at least 5 dB loss. Why? Because the phone company has the hybrid problems just described and wishes to avoid feedback at all cost. Clearly, connecting intercom systems to phone lines without gain and without AGC is not likely to work very well. Furthermore, the loss of data gives us a good indication of how much trans-hybrid loss we'll need to get the job done reliably.

Hybrids 101

Fortunately, the problem of implementing hybrids with good trans-hybrid loss has been investigated at length for the purpose of providing two-way interface for radio talk shows. In that application, insufficient trans-hybrid loss results in the distorted hybrid leakage signal being combined with the announcer's microphone signal to produce undesired coloration. Perhaps surprisingly, the problem of making a hybrid with consistently high performance is a very difficult one which has only recently been solved. The main difficulty was due to a hybrid's sensitivity to the load connected to it and the necessity that widely varying telephone line impedances needed to be accommodated.

The first hybrids were made from transformers with multiple windings. Nowadays, most hybrids are made with active components - usually op-amps - and are known as *active hybrids* (naturally). Both circuit types use the same principle, and achieve the same effect. Let's look at an op-amp version:



Simplified active op-amp hybrid. When the signals at A and B are matched in amplitude and phase, perfect cancellation results.

The first op-amp is simply a buffer and telephone line driver. The second is used as a differential amplifier - the two inputs are added out of phase (or subtracted, if you prefer). If the phone line and the *balancing network* have identical characteristics, then the signals at the input to the differential amplifier will be identical, and no send audio will appear at the output of the second op-amp.

The balancing network is a circuit consisting of capacitance and resistance, and sometimes inductance, forming an impedance network. Depending on the hybrid's application, this circuit can be very simple, or be made of a large number of components, with a very complex impedance characteristic.

Notice that R1 and the phone line form a voltage divider, as does R2 and the balancing network. If the phone line and the balancing network are pure resistances, and if they have the same value, the signals at A and B, the differential amp inputs, will have the same amplitude, and full cancellation will occur.

In the real world however, the phone line is not purely resistive, but rather is a complex impedance, causing both the amplitude and phase to vary at A as the send signal frequency varies.

Only when the impedance of the balancing network is the same as the phone line, and the signal at B is matched to that at A in both amplitude and phase, will full cancellation be achieved. The amplitude and phase must, of course, be matched across the entire telephone frequency range of 300 - 3,400 Hz. Otherwise, leakage results - the scourge of hybrids.

Thus, we've seen that the problem with hybrids is that they must deal with widely varying phone line impedance vs. frequency characteristics. When the balancing network cannot produce the same transfer function as the phone line, you get poor subtraction of the send audio from the output, which leads to the feedback problem described above.

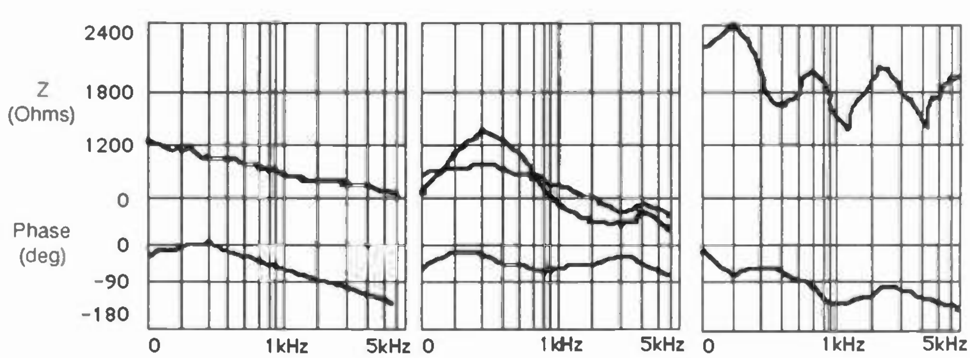
At one time, we made a study of a number of phone lines in order to determine whether there was any kind of similarity that could be exploited in hybrid design. In short, there was none. The lines were all over the place! Loading coils, line extenders, length, and telephone exchange termination vary considerably from line to line and cause wide impedance variation.

While some of the lines had smooth curves which could be simulated with a simple resistor/capacitor balancing network, others would have required a large number of coils and/or caps to get close. The lines with the smooth curves would work reasonably well with a simple hybrid since a practical RC balance network would make the cancellation of send audio at the receive output port high enough to prevent feedback. Of course, if the hybrid is to be switched among a number of lines, they would all have to have nearly the same curve.

Another consideration: the line characteristic would have to be consistent from call to call. While it would theoretically be possible to make a balance network to match the more difficult lines, practical considerations keep this approach from being used - the line impedances are often inconsistent from call to call, or the impedance characteristic required is too difficult to produce.

The hybrid solution

What, then, is the answer? It is to be found in the new technology called *digital signal processing*. Digital signal processing combines digital audio with high-speed computing to create a very powerful tool for the processing and manipulation of audio signals.



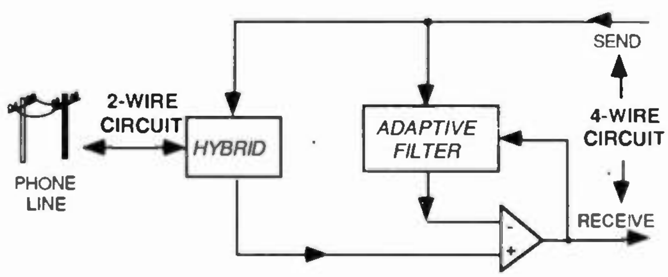
A line with a fairly smooth impedance characteristic.

This line varied from call-to-call.

This line and an analog hybrid would not get along!

Phone line impedances vary considerably - these lines were all measured at one location.

For the hybrid cancellation function, a process called *adaptive filtering* is used. Adaptive filtering is a very advanced signal processing technique which can produce a balancing network with an infinite variety of impedance transfer functions "on the fly." Since an adaptive filter adjusts as required, it can create the frequency vs. phase and amplitude characteristic necessary for maximum cancellation. It does this by comparing the simple hybrid's input and output signals digitally in order to adjust a *transversal filter*. A transversal filter combines signals from multiple time-delayed taps in varying magnitudes to simulate the desired impedance. A feedback scheme is employed to cause the filter to adapt itself to the phone line characteristic as a continuous process.



A digital adaptive filter greatly improves hybrid cancellation by adjusting to the phone line impedance characteristic automatically.

The adaptive filter, then, has two advantages over simpler hybrid approaches: It is able to automatically adjust itself to conform to changing telephone line conditions and it is able to create a wide variety of "bumpy" impedance vs. frequency curves as required for maximum cancellation.

How much better is the digital approach? Simple analog hybrids must be designed with compromise values of resistance and capacitance. Sometimes the resistive portion of the balancing network is manually adjustable. Generally, these produce an average 12-20 dB of rejection, with as little as 6 dB possible on very difficult lines. Our experience with digital hybrids is that 30-40 dB trans-hybrid loss is reliably produced - even on difficult telephone lines.

A REAL SYSTEM

An interface device has been constructed which incorporates the digital hybrid approach described above as well as a number of other features appropriate for the intercom-to-telephone application. The hardware and software characteristics of this device will be described.

The adaptive filter hybrid approach requires that the system be true digital. The telephone and intercom signals are converted to and from the digital domain and are operated on by a special DSP processor. Support circuits provide interface to the intercom and phone line, auto-answer and disconnect, metering, signalling, and power conversion.

Unlike a traditional analog system which is comprised of multiple sequential circuit stages, a digital signal processing-based audio system has the hardware topology of a microcomputer. There is a processor, some memory, and I/O ports as would be present in any computing machine. In a DSP system, there are analog-to-digital and digital-to-analog converters and other circuits which support real-time operation added to the basic microcomputer architecture.

In contrast to traditional analog design, in a DSP system, the hardware exists only to support the software. The system's functional characteristics are determined primarily by the algorithms embodied in the software code which executes in real time. Software can be written to simulate all of the traditional analog processing functions such as AGC and filtering as well as to implement functions which are impractical to produce with analog components such as the digital adaptive hybrid function described above.

Hardware

Processor and support. The processor must be fast enough to be able to perform all of the required operations in the period between signal samples. In the case of a telephone audio digital system, sampling is usually at an 8kHz rate - meaning that there is 125uS between each sample and the next. A special purpose DSP microcomputer such as the Texas Instruments 320C25 is used in this implementation. It has a cycle length of 100ns and an instruction set optimized for signal processing applications such as this one. High speed EPROMS provide program store for the operating software and I/O ports are used for control and metering.

Analog-to-Digital and Digital-to-Analog conversion. As with any digital audio system, there must be a way to go to-and-from the digital domain. Also, low-pass bandlimiting filters must be used to prevent input aliasing and for reconstruction of the output signal. In this implementation, IC's intended for the telephone industry called CODECS (CODer/DECoders) are employed. These IC's are a single-chip solution which provide filtering and conversion in both directions.

Intercom line interface. As described earlier, the usual intercom system uses a two-wire bus with a bilateral current source topology used to allow variable numbers of stations to be easily connected. For compatibility with such systems, the interface includes a circuit similar to that found in the intercom stations to provide the four-wire audio signals to the processor. An XLR connector with the standard two-channel intercom format is provided for system physical connection.

Telco line interface. A circuit is required which provides low impedance drive and two-to-four wire conversion for the telephone line. An op-amp hybrid very much like the simplified circuit presented above is employed.

Software

The software is programmed to provide:

- Auto-nulling hybrids on both the telco and intercom ports.
- Automatic Gain Control on both signal directions: send to phone and send to intercom.
- A 4Hz downward pitch-shift on the send-to-phone signal path.
- Smart gain switching to further enhance feedback margin.
- High and low-pass filtering to reduce phone line noise.
- Call signal generation.
- Metering and control.

The integration of these functions in the system is shown in the overall block diagram.

The large and varied number of processing functions indicated above demonstrates the power of the digital processing approach. In a traditional analog design, each function would represent a complex circuit section; in DSP, each merely exists as a software module.

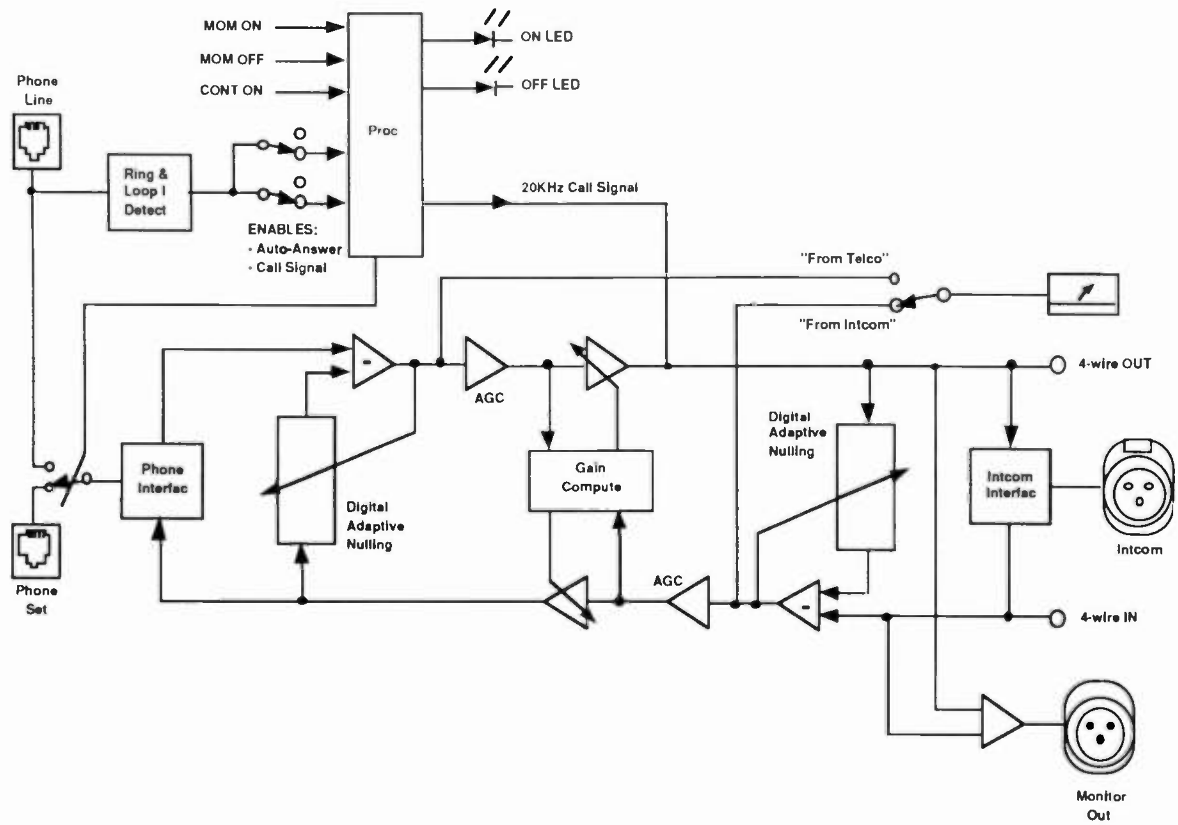
Auto-nulling hybrids. Described above. Provides consistently reliable send/receive isolation so that gain may be inserted into the two speech paths.

AGC. The gain processing insures that the levels to and from the intercom system are maintained consistently without operator intervention. The primary purpose is to adjust for varying telephone line levels. The AGC is fairly sophisticated - incorporating floating platform gating, etc.

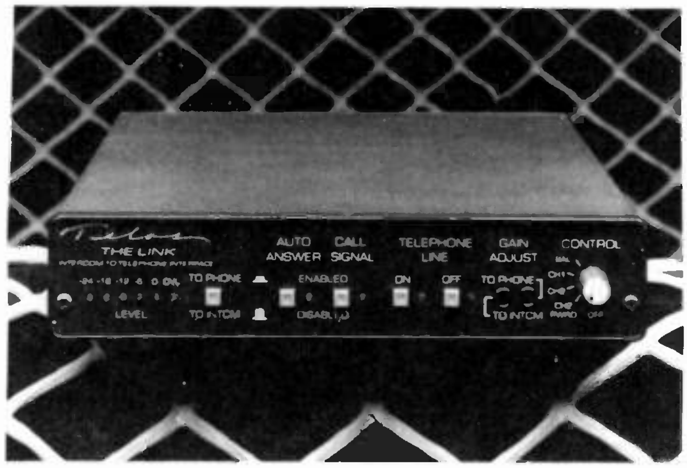
4Hz pitch-shifter. Perhaps you are wondering about the purpose of this function? It is to allow yet more gain margin by suppressing feedback. It is an old trick used in live-sound engineering. Each time the signal makes a trip around the potential feedback loop it gets downshifted in frequency and thus feedback is damped out rather than built-up.

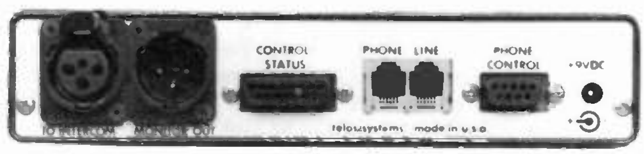
Gain switching. Also used to increase feedback margin. It inserts very shallow (approximately 8dB) ducking in each signal path when the other direction is active. Thus, another 8dB gain is possible. The switching is sufficiently shallow and fast that no effect is generally noticed; the system appears to be entirely full-duplex to the user.

The other software modules serve to support various system operations.



System block diagram of Telos Link *prototype unit showing integration of the various processing functions. Note that most of the functions are provided by software modules running on a DSP processor rather than by the usual analog circuit sections.*





Photos showing front and rear panels of prototype unit. The digital approach allows the entire system to be packaged in a small modem-style box.

SUMMARY

This high-technology approach to intercom-to-phone interfacing which makes use of new digital signal processing technology significantly improves communication capability compared to older techniques which relied upon multiple boxes and primitive manually-adjusted analog hybrids.

In some instances significant cost savings could be realized since inexpensive dial-up lines could be substituted for the special four-wire telco circuits often used for remote production projects.

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A NEW SMALL FORMAT DIGITAL VTR USING 1/2 INCH TAPE

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In order to achieve a single format VTR that can be used in the studio, in post-production, in on-air cart machines as well as in ENG/EFP equipments, a new composite digital VTR using half-inch tape has been developed on the basis of the latest advanced technologies. This paper describes the background to the need for a single format and the advanced technologies for high density digital recording to achieve it.

Introduction

As the nation-wide public broadcaster, NHK has a Broadcasting Center in Tokyo for nation-wide service and 69 broadcasting stations throughout Japan for regional services. In many applications, three different formats have been introduced, because of trade-offs involving quality, cost, portability, easiness of operation and so on. Type C and MII have been used for production and playback facilities, while Betacam has been introduced for ENG purposes.

However, it is not desirable to use various kinds of format in day-to-day operation. On the other hand, the outcome of recent dramatic progress in digital recording has made it possible to produce a single format for use in the studio, post-production, cart machines and ENG/EFP equipments.

Now, with the prospect for replacement of Type C, which was introduced from 1977, we have studied the next generation VTR system. The summarized requirements are as follows.

1. A single format should be adopted in all applications including ENG/EFP, studio recording, post-production and cart machine.

2. Tape width should be 1/2 inch or less so that a single format can be realized.

3. Linear PCM should be adopted, because even state-of-the-art technology of bit rate reduction cannot avoid visible artifacts particularly in post-production applications.

4. Total cost of introduction and operation should decrease remarkably compared with the conventional system.

With the target mentioned above, we have conducted a feasibility study for a small format digital VTR. Consequently, we have developed high performance heads and advanced digital recording technologies such as channel coding and error correction techniques and so on. We accomplished a compact digital VTR using half-inch tape. This format provides two sizes of cassette: the S-size cassette, which is as small as a Betacam cassette, can record a 1 hour program for ENG/EFP and cart machine; and the L-size cassette, which is slightly larger than a VHS or MII cassette, can record a 2 hour program for mainly studio use.

Background

NHK owns 300 Type C VTRs including portable ones, 300 MII VCRs and 2700 Betacam VCRs including camcorders. Type C is used in production, post-production and playback facilities at the Broadcasting Center in Tokyo (Figure 1). MII plays the

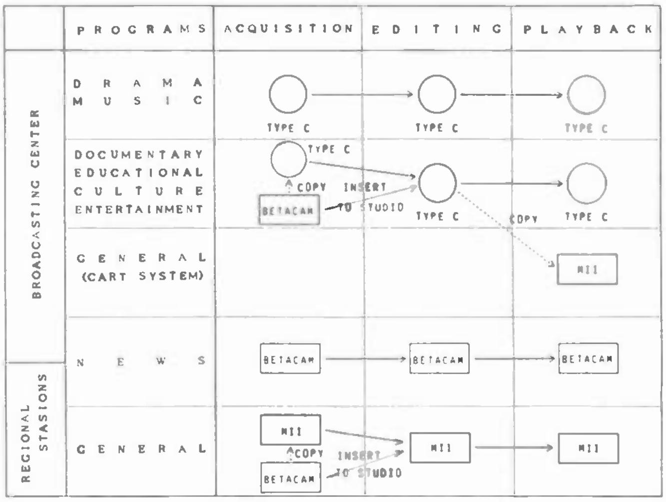

Figure 1. VTR systems at NHK

same role as Type C at the regional stations. Betacam is used in news production.

As half-inch CAV formats are superior to Type C format in operational ability, many CAV VCRs are applied to field recording. When materials are recorded by these CAV VCRs, it is necessary to copy them into Type C format in order to maintain their picture quality.

Considering total cost of investment and labor-saving, a single format is eagerly desired in all applications. Experience has convinced us to adopt half-inch or less tape width for a single format.

The number of tapes which are stored for historical purposes, as program materials or for application in other fields is increasing remarkably. NHK keeps, at present, 110,000 tapes including one-inch and other formats. The number is rising at the rate of approximately 18,000 tapes a year (Figure 2). NHK needs an extra 1,500 to 2,000 square feet of storage space to keep them. Compact format would reduce the space to a fifth or less.

A digital signal-processing camera has been introduced into the market. Digital processing reduces the time of adjustment and maintenance. We have confidence in developing a one-piece digital camcorder which equals an analog camcorder in weight and power consumption.

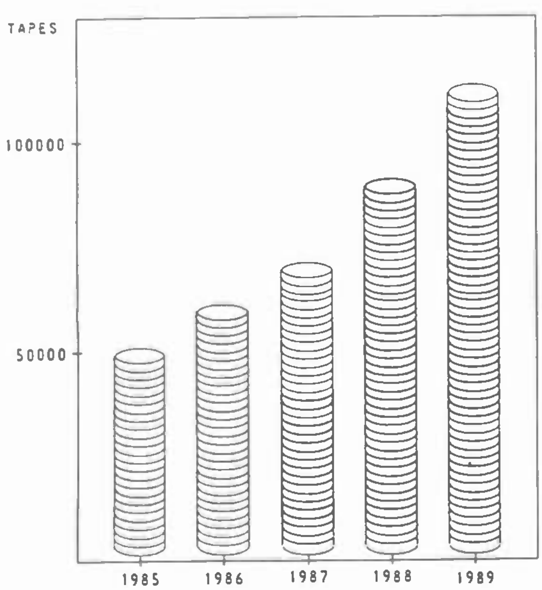

Figure 2. Increase of stored tapes

Advanced Technologies in the New Format

1. A New Channel Coding: 8-14 Block Code. We have developed a new 8-14 block code which improves a drawback of the Scrambled NRZ of the D-1 format in the low frequency region and that of Miller^2 channel coding of the D-2 format in the high frequency region. This 8-14 code limits the transition free run from 2 data intervals to 7 data intervals.

The features of the new 8-14 block code are as follows:

a) DC-free and narrow bandwidth.
These factors are suitable for slant azimuth recording and overwrite, especially, for easy equalization. Figure 3 shows power spectra of the 8-14 code, compared with those of S-NRZ and Miller^2.

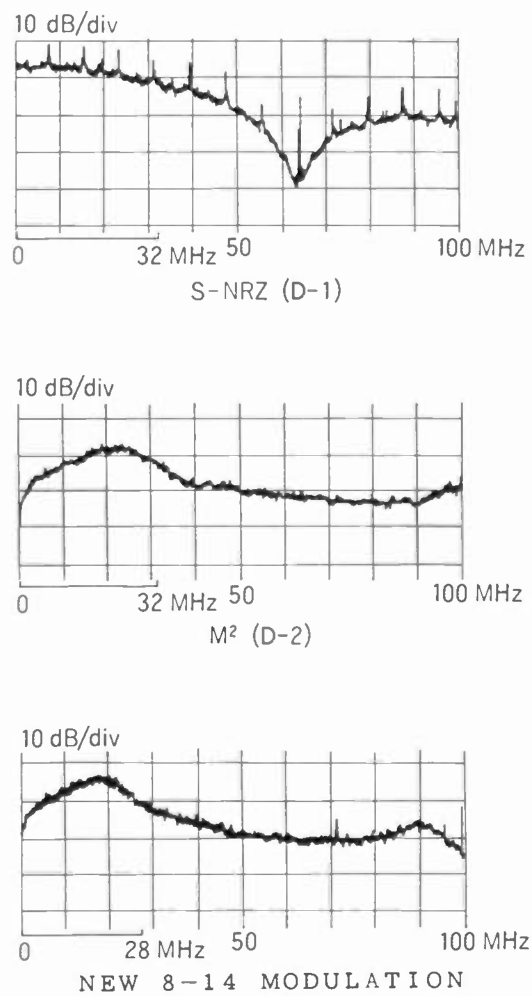

Figure 3. Power spectra of S-NRZ, Miller^2 and New 8-14 coad

b) Longer flux reversal of 14% compared with $Miller^2$ and S-NRZ.
This factor is effective to obtain the high S/N ratio.

c) Large margin of off-track error rate.
Figure 4 shows the characteristics of the error rates versus off-track distance in the 8-14, S-NRZ and $Miller^2$. The 8-14 coding is insensitive to off-track error.

d) Free from error propagation in decoding.
As played-back bits are decoded word by word, bit errors in a coded word have no influence on decoding of other words.

e) Error detection capability.
The number of code words used in this modulation is only 4.7% of the possible combination of 14 bits. So, it is possible to detect erroneous words in decoding. Adding this result to a Reed-Solomon(RS) product code, error correction capability improves remarkably.

2. Error-Correcting Code
High performance error correction technologies are indispensable for digital recording. Three kinds of RS product code with similar coding efficiency are compared in Figure 5. This illustrates that the larger block size of RS product code has a better error correction capability. When the error detection capability of the 8-14 channel code is applied, it remarkably reduces the remaining errors after correction .

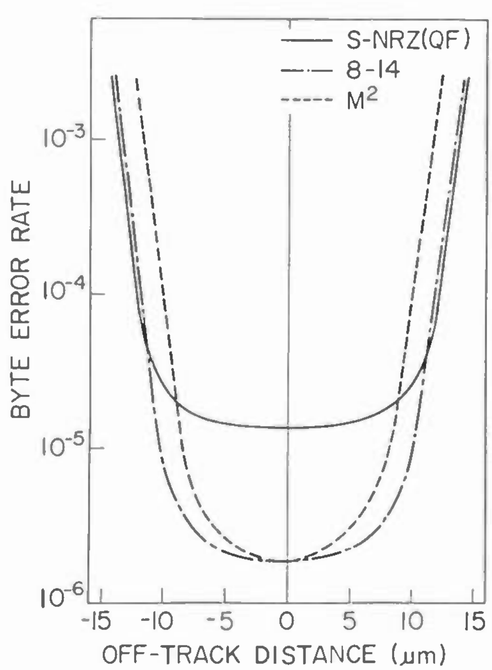

Figure 4. Off-track error rate

3. Improvement of off-track error rate at the editing points.
In order to achieve a high density digital VTR, the tape format having the narrow helical tracks without guard band should be adopted. On the other hand, the track width is restricted by tracking accuracy along the helical tracks, and the tracks at the editing points cause a critical condition.

Because "editing" reduces track width or remains a part of old track with the same azimuth, above all, the latter affects error rate approximately twice as bad (depending on channel coding) as the former.

To solve this problem, a technique introducing a guard band between the old track and the new track has been developed. Figure 6 shows a brief sketch of the recorded tracks at the editing points.

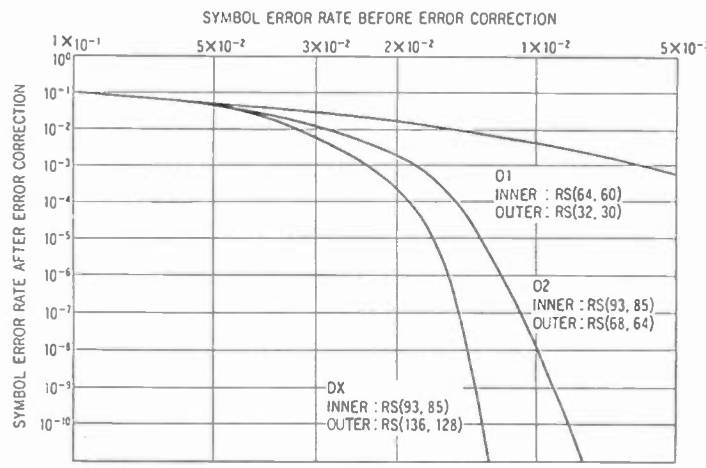

Figure 5. Error correction capability

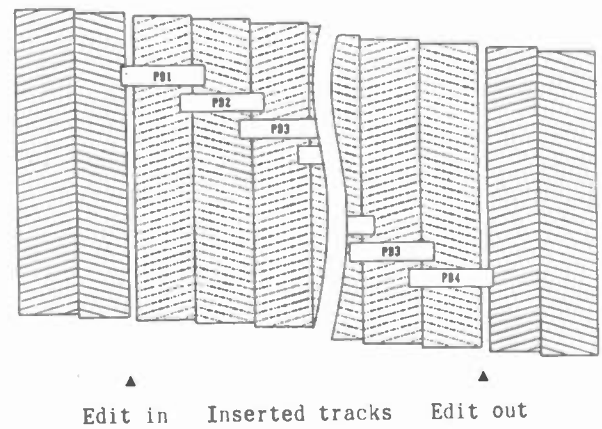

Figure 6. Guard band at edit-in/out points

Outline of Specifications of the New Format

On the basis of results of an experimental machine and developed technologies, and considering operations with high reliability, economic element etc., the format as shown in Table 1 and Figure 7 has been selected.

This format provides technical and economic elements to meet users' requirements such as high quality of video and audio signal, transparent multi-copying, small size, light weight, robustness and low cost.

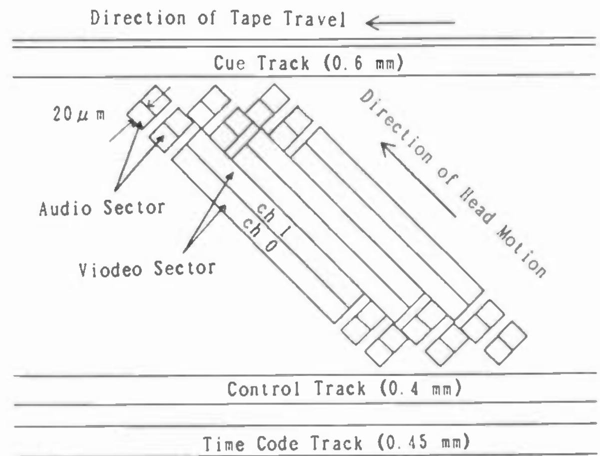

Figure 7. Location of recorded tracks

Table 1. Specifications

ITEM		DESCRIPTION
Play Time		L: 125 min (212x124x25mm) S: 64 min (161x 98x25mm)
Tape	Tape Width	12.65 mm(1/2-in)
	Material	Metal Particle
	Thickness	11μm (Max)
Video Signals		
Input Signal		NTSC composite signal
Sampling Frequency		4Fsc (14.32 MHz)
Sampling Phase		123°, 33°, -57°, -147°
Form of Coding		Linear PCM 8 bits/sample
Recorded Samples/Line		768
Recorded Lines/Field		255
Audio Signals		
Sampling Frequency		48KHz
Form of Coding		Linear PCM 16~20 bits/sample
Number of Channels		4 channels
Channel Coding		NEW 8-14 Code
Error Correction Code		
Type		Reed-Solomon Product Code
Check Bytes		Inner:8 bytes Outer:8 bytes
Track Pitch		20 μm
Minimum Wavelength		0.77 μm
Azimuth Angle		±15°
Tape Speed		83.88 mm/s
Helical Tracks/Field		6
Longitudinal Tracks		Time Code, Control and Cue

Conclusion

An ideal goal of broadcasters is a single format applicable to all uses.
The advanced technologies to achieve a compact and small size digital VTR and a new format have been discussed. This format will meet the requirements of broadcasters, production houses and others. With the cooperation between broadcasters and manufacturers a full scale line-up of a studio recorder, a camcorder and a portable recorder should be completed.

BROADENING THE APPLICATIONS OF ZONE PLATE GENERATORS

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The greater information content of High Definition, advanced, or enhanced television requires sophisticated processing to make recording and transmission practical. This processing is likely to use some form of bandwidth compression, scan rate changes, motion detection and compensation algorithms, or other "new" techniques. Zone plate patterns are well suited to exercising a complex video system in the three dimensions of its signal spectrum: Horizontal, Vertical, and Temporal. Electronic zone plate techniques have been in the technical literature for the past decade; they are not new. They are not, however, yet enjoying widespread application or understanding. The increasing complexity of video signal processing, and the advent of less expensive, easy to use zone plate generators argues for taking another look at these useful patterns.

Introduction

Zone plate signals, unlike most conventional test signals, can be complex and dynamic. Because of this, they are capable of simulating much of the detail and movement of actual picture video, exercising the system under test in three dimensions: horizontal, vertical, and temporal (or motion). These digitally generated and controlled signals also have other important characteristics needed in test signals.

A signal intended for meaningful testing of a television system must be carefully controlled, so that any departure from a known parameter of the signal is attributable to a distortion, or change, in the system under test. The test signal must also be predictable, so it can be accurately reproduced at other times or places. These constraints have usually led to test signals which are electronically generated. In a few special cases, a standardized picture has been televised by a camera or monoscope -- usually for a subjective, but more detailed evaluation of a system's performance.

Actually, a zone plate is a physical optical pattern, which was first used by televising it in this way. Now that electronic generators are capable of generating very similar patterns, the label "zone plate" is being applied to the wide variety of patterns created by this single piece of equipment.

Conventional test signals, for the most part limited by the practical considerations of electronic generation, have represented quite simple images. Each signal is capable of testing a narrow range of possible distortions; several test signals are needed for a more complete evaluation. Even with several signals, this method may not reveal all possible distortions, or allow study of all pertinent characteristics. This is especially true in modern systems employing new forms of sophisticated signal processing.

Efforts to employ a more "wide spectrum" test signal, especially for study of standards converters, led to the development of the electronic zone plate generator several years ago. Details of the pertinent theory are, in fact, well documented in the technical literature of the past decade. Even so, these techniques have not seen widespread use or understanding, especially in the United States.

This situation is likely based on several factors. The earliest implementations of zone plate generators were expensive and difficult to use. Maybe our European counterparts, with their long-standing concern with complex processing, such as standards conversion, had more impetus. Perhaps other factors played a part, as well, but things may be different in the 1990s.

Today, there are several reasons to take a new look at zone plate signal testing. Various High Definition Television (HDTV) systems, around the world, are in the proposal or early implementation stages. More locally, several systems are now being evaluated for advanced or enhanced broadcast television. These new technologies depend much more heavily on complex processing. Motion detection and motion compensation algorithms, conversion of scan parameters between transmission (or recording) and display, and various forms of bandwidth reduction (compression) are no longer just researchers' dreams.

Simple, stationary images are usually handled quite well by these processes. This is another way of saying, "They usually work quite well with conventional test signals." Image distortions, or "artifacts," may appear as image complexity is increased. Motion is often even more a problem if one is concerned with small disturbances in the display. To put this yet another way: The real world is often a lot less forgiving than our familiar test signals.

Demonstration of many of these effects, and help in evaluating and understanding them is enhanced by modern, easier to use, and more affordable zone plate generators.

Simple Zone Plate Patterns

The basic testing of a communication channel historically has involved the application of several single frequencies -- in effect, spot checking the spectrum of interest. A well known, and quite practical adaptation of this idea is the "multiburst" signal (*Figure 1, waveform; and Figure 2, picture*) which has been around since the earliest days of video. The multiburst signal provides several discrete frequencies along a TV line.

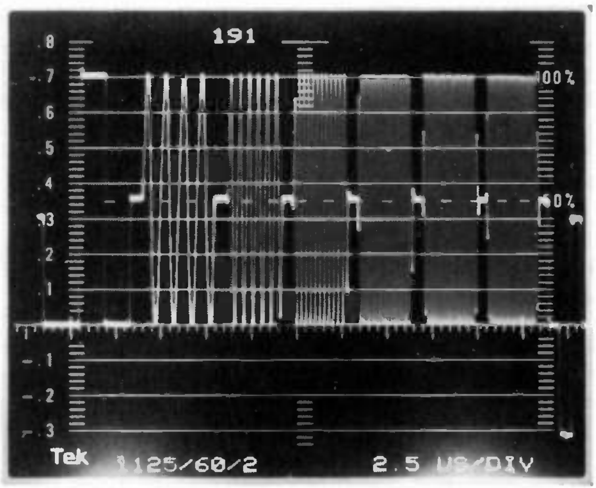

Figure 1. Conventional multiburst waveform (1H).

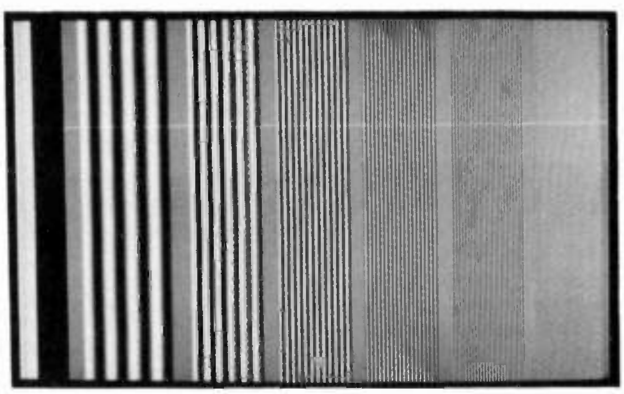

Figure 2. Conventional multiburst picture display.

A signal which was harder to implement in earlier generators, but easier to use, is the frequency sweep (*Figure 3, waveform; and Figure 4, picture*) which varies the signal frequency continuously along the TV line. In some cases, the signal is swept as a function of the vertical position (field time). Even in these cases, the signal being swept is appropriate for testing the spectrum of the horizontal dimension of the picture.

(Figure 4, and several others in this paper, show "beat" effects introduced by the screening process used for the photographs. This is largely unavoidable, since the screening process is quite similar to scanning or sampling a television image -- the patterns are designed to find this sort of problem.)

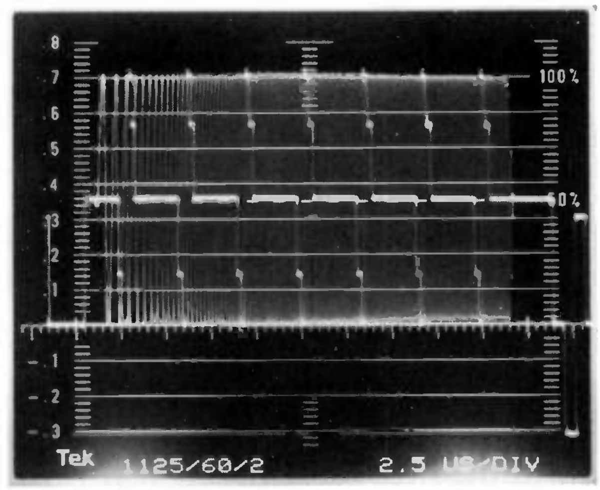

Figure 3. Conventional frequency sweep waveform, with markers (1H).

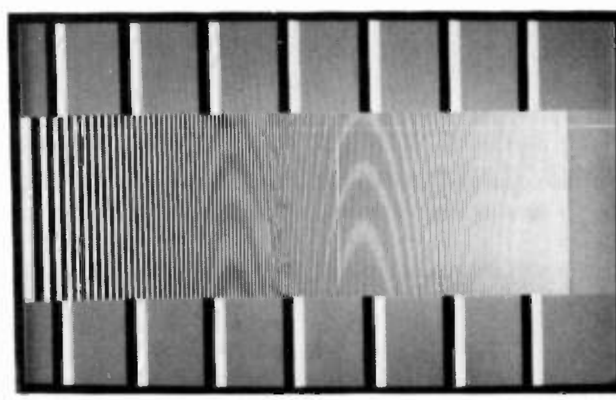

Figure 4. Conventional frequency sweep picture display.

These first four photos were taken from the output of a conventional signal generator. Horizontal single frequency and frequency sweep signals similar to these can be produced with a zone plate generator which has been appropriately configured (*Figure 5 through Figure 8*). On the Tektronix TSG-1000 Family of HD Generators, for example, these patterns are button selectable and a single knob controls frequency or maximum sweep frequency from 0 to well over 30 MHz.

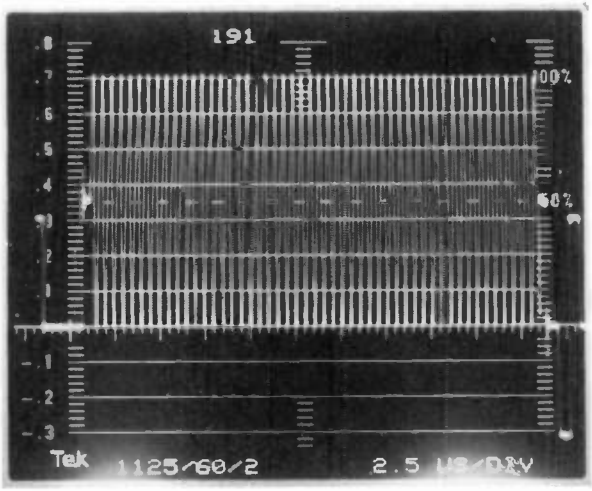

Figure 5. Single horizontal frequency waveform (1H).

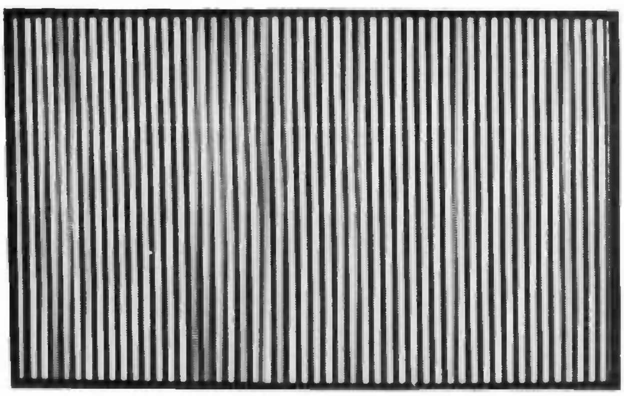

Figure 6. Single horizontal frequency picture display.

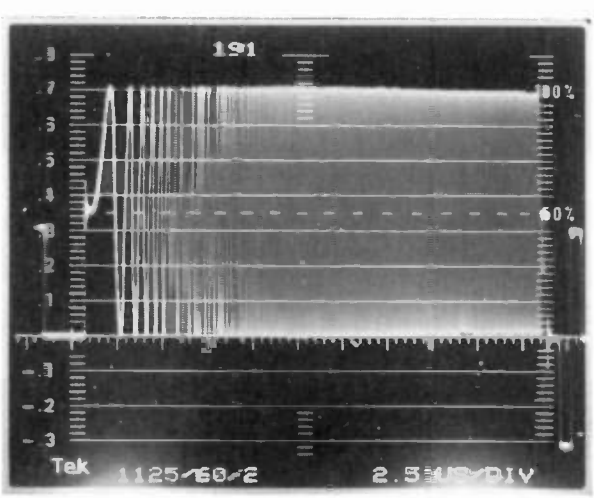

Figure 7. Horizontal frequency sweep waveform (1H).

Horizontal patterns, such as these, may be used to evaluate channel frequency response, horizontal resolution, possible Moire' effects in recorders and displays, or other system characteristics.

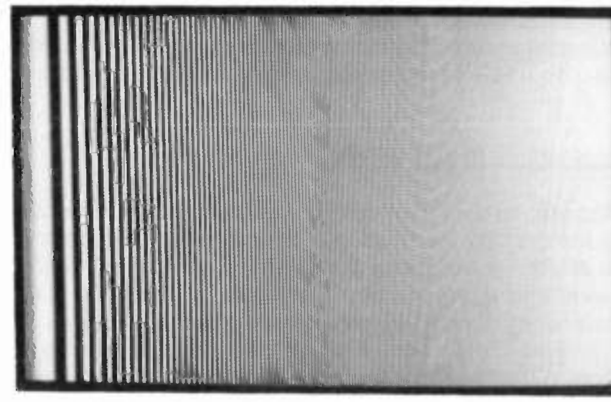

Figure 8. Horizontal frequency sweep picture display.

Patterns which test the vertical (field) related response have been much less frequently used. As new technologies use conversion from progressive to interlaced scan, line doubling display techniques, vertical anti-aliasing filters, scan conversion, motion detection, or other processes which combine information from line to line, these vertical testing patterns should be more in demand.

In the vertical dimension, as well as the horizontal, tests may be done with single frequency or frequency sweep signals (*Figure 9 through Figure 11*). Note that Figure 9 is a magnified vertical rate waveform display; each "dash" in this photo represents one horizontal scan line. Sampling of vertical frequencies is inherent in the scanning process; this photo shows the effects on the signal waveform. Note also that the signal voltage remains constant during each line, and changes only from line to line in accord with the vertical dimension sine function of the signal.

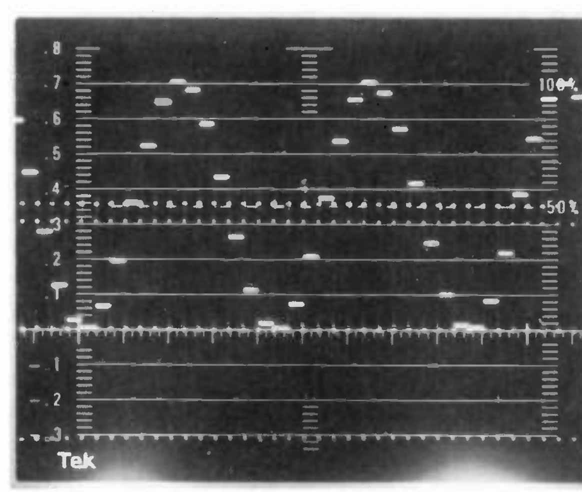

Figure 9. Single vertical frequency, magnified vertical rate waveform, showing the effects of scan sampling.

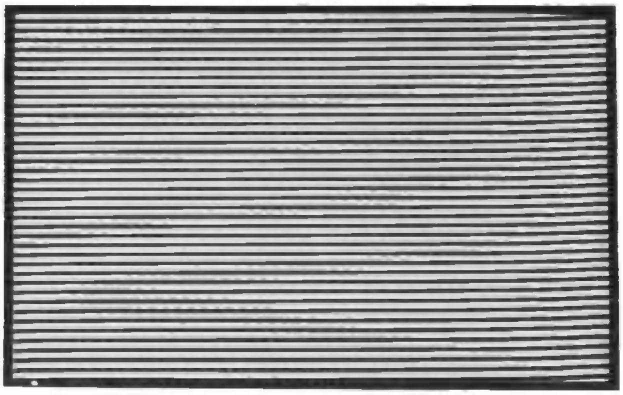

Figure 10. Single vertical frequency picture display.

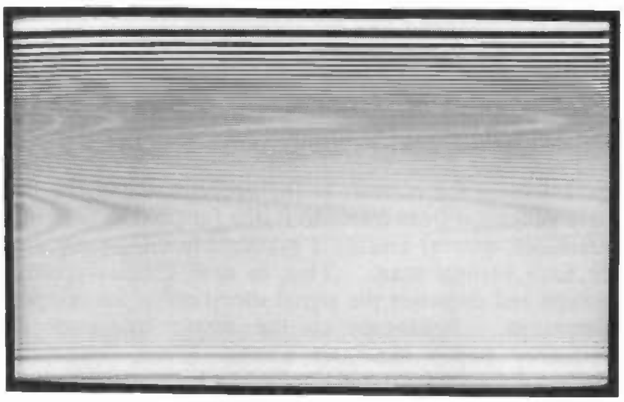

Figure 11. Vertical frequency sweep picture display.

These horizontal and vertical sinewaves and sweeps are quite useful; but they do not use the full potential of a zone plate signal source. Sharing some thoughts on the basic hardware implementation of a zone plate "engine" will aid in understanding the more complex and interesting patterns.

Basic Implementation of a Zone Plate Generator

Originally, test signal generators used analog function generators, switching, and waveform shaping filters. More recent devices use digital signal memory, which is sequentially accessed to create digital, or by means of D to A converters, analog outputs. In each of these cases, the signal has been designed (or programmed) into the device and is simply recreated to derive the outputs. The zone plate generator differs significantly.

A zone plate signal is created in real time; it is computed "on the fly." The value of the signal at any instant in time is represented by a number in the digital hardware. This number is incremented as the scan progresses through the three dimensions which define a point in the television image; Horizontal position, Vertical position, and Time.

The exact way in which these dimensions alter the number is controlled by a set of 10 coefficients.[1] These coefficients determine the initial value of the number, and control the size of the increments as the scan progresses along each horizontal line, from line to line vertically, and from field to field temporally. A set of coefficients uniquely determines a pattern, or a sequence of patterns when the time dimension is active.

The process just described actually produces a sawtooth waveform; overflow in the accumulator holding the signal number effectively resets the value to zero at the end of each cycle of the waveform. Usually it is desirable to minimize the harmonic energy content of the output signal; in this case, the actual output is a sine function of the number generated by the incrementing process. Other functions have been used for special applications.[2]

Many operators, of course, would not consider the setting of 10 coefficients to be a "friendly" control system. Actually, most operators will have no need to understand just how the coefficients control the signal.

In the TSG-1000 generators, single buttons allow selection of a variety of useful patterns, while a knob alters several coefficients simultaneously to control appropriate aspects of the pattern in a sensible way. It is still possible and convenient, for those who desire, to access each coefficient individually to "customize" patterns to a specific task, and save the setup for later retrieval and use.

Complex Patterns

It is quite practical to create a pattern of sinewaves or sweeps in multiple dimensions, using the unique architecture of the zone plate generator. The pattern shown in Figure 12, for example is a single signal sweeping both horizontally and vertically.

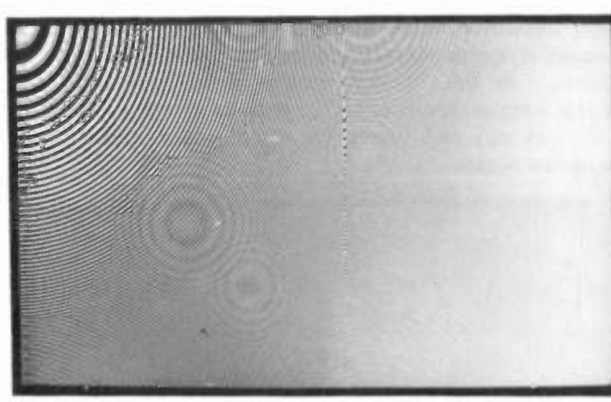

Figure 12. Combined horizontal and vertical frequency sweep picture display.

Figure 13 shows the waveform of a single selected line (line 263 in the 1125/60/2 system being used here for demonstration). Note the horizontal waveform is identical to Figure 7, even though the vertical dimension sweep is now also active.

Actually, different lines will give slightly different waveforms. The horizontal frequency and sweep characteristics will be identical, but the starting phase must be different from line to line to construct the vertical signal.

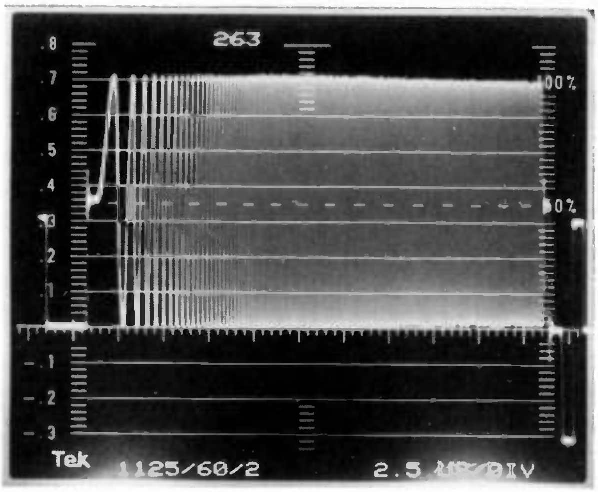

Figure 13. Combined horizontal and vertical frequency sweeps, selected line waveform display (1H). This figure show the maintenance of horizontal structure in the presence of a vertical sweep (Compare to Figure 7).

A waveform taken vertically (at a given horizontal position), although not shown here, would be identical to one taken from the pattern in Figure 11. Vertical waveforms taken at different horizontal positions will similarly show different initial phase because of the simultaneous horizontal signal; the vertical frequency and sweep characteristics would be identical, however.

Figure 14 shows a two-axis sweep pattern which is most often identified with zone plate generators; perhaps because it quite closely resembles the original optical pattern. In this "circle" pattern, both horizontal and vertical frequencies start high, sweep to zero (in the center of the screen) and sweep up again to the end of their respective scans.

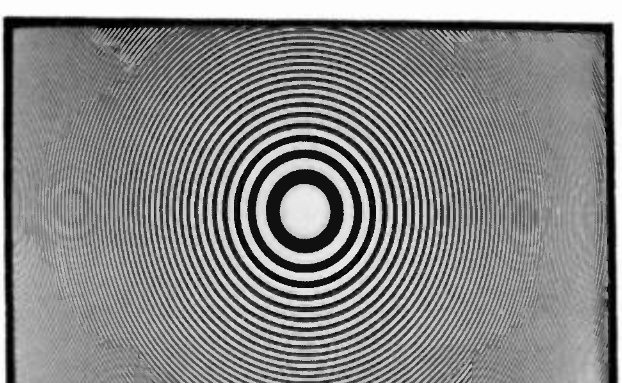

Figure 14. The best known zone plate pattern, combined horizontal and vertical frequency sweeps with zero frequency in center screen.

The concept of two-axis sweeps is actually more powerful than it might, at first, seem. In addition to purely horizontal or vertical effects, there are possible distortions or artifacts which are only apparent with simultaneous excitation in both axes. In other words, a system's response to diagonal detail may not be predictable from information just about its horizontal and vertical responses.

Consider an example from NTSC: A comb filter will suppress crosscolor effects from a horizontal luminance frequency signal near the subcarrier frequency (such as the higher frequency packets in a NTSC multiburst signal). If, however, the right vertical component is added, creating a specific diagonal luminance pattern, even the most complex decoders will interpret the pattern as colored. In this case, the two-axis sweep shows very different effects than the same sweeps shown individually. A multi-dimensional sweep is a powerful tool for showing analogous effects in other complex signal processing systems.

The Time (Motion) Dimension

Incrementing the number in the accumulator of the zone plate generator from frame to frame (or field to field in an interlaced system) creates a predictably different pattern for each vertical scan. This, in turn, creates apparent motion and exercises the signal spectrum in the temporal dimension. Analogous to the single frequency and frequency sweep examples already given, appropriate setting of the time related coefficients will create constant motion or motion sweep (acceleration).

Specific motion detection and interpolation algorithms in a system under test may be exercised by determining the coefficients of a critical sequence of patterns. These patterns may then be saved for subsequent testing during development or adjustment. In an operational environment, appropriate response to a critical sequence could ensure expected operation of the equipment or detect faults.

While motion artifacts are difficult to portray in the still image constraints of this paper, the following example gives some idea of the potential of a versatile generator: In Figure 11 the vertical sweep maximum frequency has been increased to the point where it is zero beating with the scan at the bottom of the screen (cycles per picture height of the pattern matches the lines per picture height per field of the scan). Actually, in direct viewing, there is another noticeable artifact in the vertical center of the screen; it is a harmonic beat related to the gamma of the display CRT. Because of interlace, this beat flickers at the field rate. The photograph integrates the interfield flicker and thereby hides the artifact, which is readily apparent in real time.

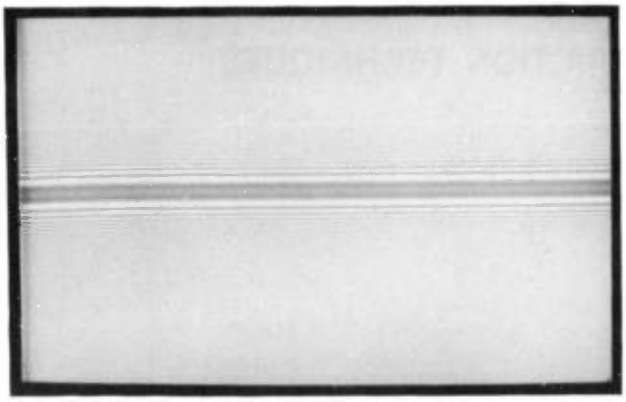

Figure 15. The same vertical sweep as Figure 11, except appropriate pattern motion has been added to "freeze" the beat pattern in center screen for photographing or other analysis.

Figure 15 is identical, but for one important difference: Upward motion of 1/2 cycle per field has been added to the pattern. Now the sweep pattern itself is integrated out, as is the first order beat at the bottom. The harmonic effects in center screen no longer flicker, since the change of scan vertical position from field to field is compensated by a change in position of the image. The resulting beat pattern does not flicker and is easily photographed or, perhaps, scanned to determine depth of modulation.

A change in coefficients produces hyperbolic, rather than circular two-axis patterns (*see Figure 16*). Another interesting pattern, which has been suggested for checking complex codecs, is shown in Figure 17. This is also a moving pattern, which was slightly altered to freeze some aspects of the movement to take this photograph.

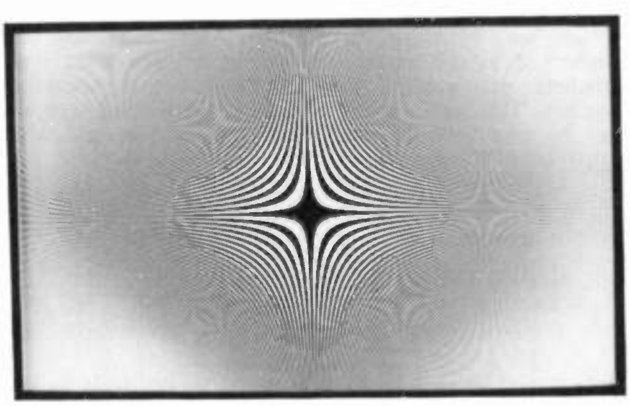

Figure 16. A hyperbolic variation of the two-axis zone plate frequency sweep.

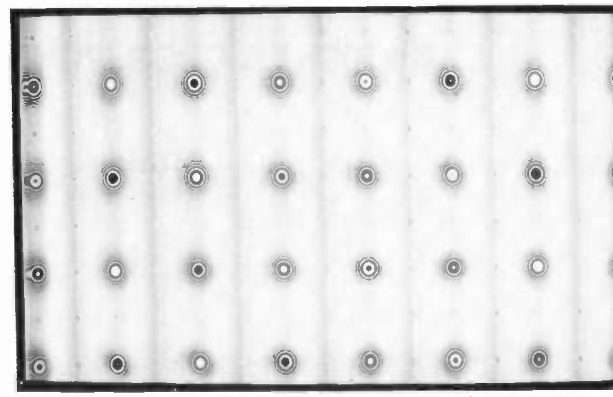

Figure 17. A two-axis frequency sweep in which the range of frequencies is swept several times in each axis. Complex patterns such as this may be user created for specific test requirements.

Summary

The trend is toward more complex signal processing in television technology. With its facility for creating multi-dimensional frequency sweeps, the zone plate generator offers an appropriately complex test signal, one capable of exercising the horizontal, vertical, and temporal spectrum which may be occupied by a moving video image.

The few examples in this introductory paper can only hint at the ultimate utility of zone plate techniques. Other, more interesting and useful, examples will emerge as users pursue real time observation of complex, dynamic effects on a greater range of equipment.

The recent and continuing changes in both test and operational equipment technology create a new environment for this unique tool.

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SWITCHER CROSSPOINT REDUCTION TECHNIQUES

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The signal distribution and routing system is the heart of a modern tele-production facility. New switching architectures combined with powerful control systems create very large signal switching arrays with greater flexibility at greatly reduced costs. This paper includes examples of actual routing switchers using crosspoint reduction techniques.

RECTANGULAR SWITCHING MATRIXES

A rectangular signal switching array has rows and columns corresponding to system inputs and outputs. Where the rows and columns intersect a crosspoint switch is included.

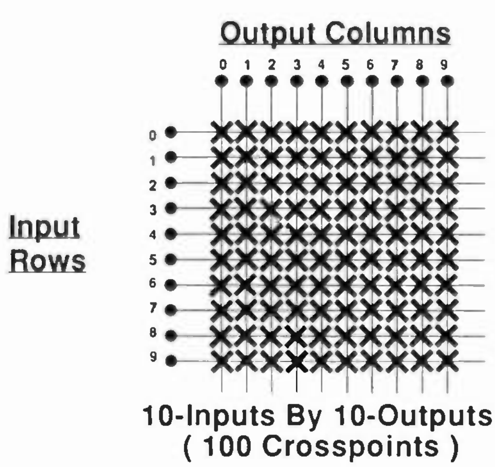

Figure-1. Rectangular Switching Matrix

Any row (input) may be selected to any column (output). The product of the number of inputs and the number of outputs determines the total number of switching crosspoints required. As a matrix size is increased, the number of crosspoints increases dramatically, i.e., by the square of the increase in size.

Increasing The Number Of Inputs

Equipment rack configurations limit the number of signal inputs available. A typical routing switcher rack system supports 130 inputs with internal power supplies and 160 inputs with external power supplies. The number of switcher inputs can not be increased beyond these sizes without using oversized rack frames.

A switcher with more inputs than can be built in a single rack frame may be created by cascading a number of individual switchers. This arrangement may be used with a small number of outputs and when signal timing is not a concern.

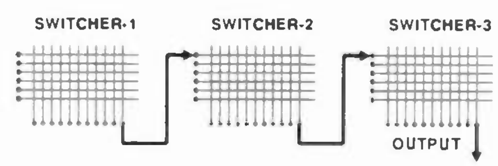

Figure-2. Cascaded Switching System

In another switching arrangement for increasing the number of signal inputs, the output of individual input switchers are entered into a single-output secondary switcher. This arrangement is called a switching tree and may be used with a small number of outputs and when signal timing is important.

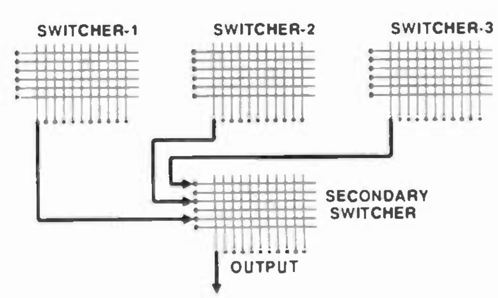

Figure-3. Tree Switching System

Increasing the number of outputs

The number of switcher outputs is limited by the physical space allowed. Typically, the inputs to the switching matrix loop through blocks of system outputs.

The signal quality is degraded as the number of looping connections increase. In a typical routing switcher the maximum number of looping outputs is around 80-90. This means that input distribution amplifiers are required for the additional switching blocks greater than 80-90 outputs.

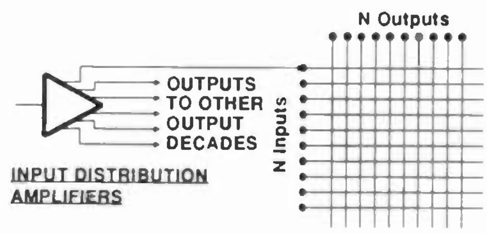

Figure-4. Switcher With Input Distribution

The cost of a signal matrix with greater than 100 outputs is heavily burdened by the large number of input distribution amplifiers and interconnection wiring. A distribution amplifier is required for each system input. In addition, if the number of system inputs requires matrix re-entry, signal timing is lost.

DESIGN GOALS FOR CROSSPOINT REDUCTION

The goal of any crosspoint reduction scheme is to create a very large input and output array, but with significant reduction in the total number of crosspoints required. Reduction of the number of crosspoints reduces hardware costs and may result in reduced space requirements.

A distribution switcher using a crosspoint reduction scheme must appear to the outside user as a simple rectangular matrix. There should be un-restricted access to any system input by any system output.

The output signal quality must be at least comparable to the signal quality of a simple switching array.

Using crosspoint reduction only makes sense if there is a major economical advantage in doing so. The sophisticated control systems and complex interconnection wiring may add considerable hardware and labor cost to a switcher installation.

PATH-FINDING SWITCHERS

The simplest method for creation of a large switching array uses signal tie-lines to interconnect a number of distribution switchers. This method of switching is call **path-finding**. Tie-lines permit signals to be exchanged between routing switchers. Usually the individual switchers are located in operational areas and the tie-lines allow the operator to select sources in other areas of the tele-production facility.

Within path-finding switching system a user has access to all signal inputs on the local matrix. However, if an input signal in another matrix is required, the signal must be routed via one of the system tie-lines. If all tie-lines are active the signal can not be accessed and the system blocks. Careful consideration must be given to the minimum number of tie-lines included. A severe system blockage can occur if the system is under-designed.

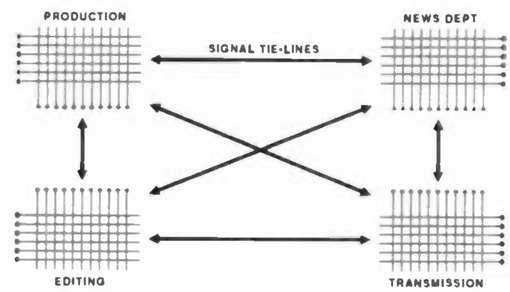

Figure-5. Path-Finding Switcher

MULTI-STAGE SWITCHING ARCHITECTURE

Multi-stage switching arrays have been in use for many years in the telephone switching industry. In a multi-stage switcher a number of small switching arrays are interconnected to create a very large virtual switching matrix. This architecture is designed to be non-blocking. This means that any input can access any un-used output. Telephone switchers using three, five, seven and mores stages have been designed.

In order to maintain signal integrity in a tele-production facility, a maximum of three switching stages are used. Since a signal always passes through three crosspoints, each system element much be as transparent as possible.

A three-stage switching matrix consists of Input, Intermediate and Output matrixes. Each output from an Input matrix is connected to the input of an Intermediate matrix. Each output from an Intermediate matrix is connected to the input of an Output Matrix.

A three-stage switcher is designed as follows:

- The number of inputs per Input matrix is calculated by taking the square root of the total number of system inputs. This calculation also determines the required number of Input matrixes.
- The number of outputs per Output matrix is calculated by taking the square root of the total number of system outputs. This calculation also determines the required number of Output matrixes.
- The number of Intermediate matrixes is found by taking the sum of the number of inputs per Input matrix plus the number of outputs per Output matrix minus one. This relationship insures that there is always at least one path available to any input or output.
- The size of the Intermediate matrixes is determined by the number of Input and Output matrixes.

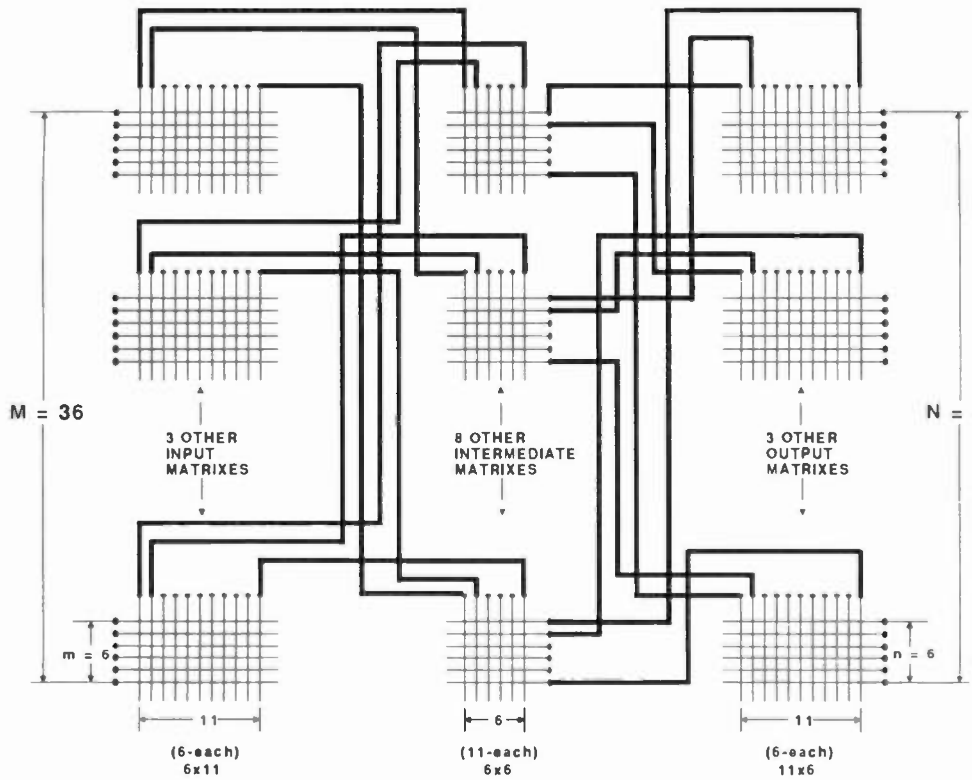

Figure-6. Three Stage Switching Matrix

In Figure-6, each Input and Output matrix has six (6) inputs or outputs. The number of Intermediate matrixes is the number of inputs per Input matrix plus the number of outputs per Output matrix minus one; i.e., 6+6-1=11.

A rectangular 36x36 (square) matrix would require 1296 crosspoints. The 36x36 three-stage matrix requires 1188 crosspoints. The difference in crosspoint count increases as the size of the matrix increases.

Distribution switchers are built using building block sizes; i.e., 8x8, 10x10, 16x16, etc. The actual design of a three-stage switcher must make trade-offs against the input/output block sizes, starting system size and final system size. Consider an example of a switching matrix built for 200 inputs by 200 outputs expandable to 300 inputs by 300 outputs. Using a block size of 10x10, the matrix is organized as follows. Each Input matrix will have ten (10) inputs and each Output matrix will have ten (10) outputs. The matrixes use ten (10) inputs or outputs instead of value of the square root to maximize the usage of the 10x10 switcher block size without wasting crosspoints. There will be twenty (20) Input and twenty (20) Output matrixes. The number of Intermediate matrixes is nineteen (19); i.e., 10+10-1=19. The size of the Intermediate matrixes will determine the final expansion size. The Intermediate matrixes only require 19 inputs and outputs per matrix. But the expansion size will require an Intermediate matrix having 30 inputs and outputs. Therefore, the Intermediate matrixes will be designed as 30x30 but only installed with circuit cards for 20 input and outputs. Since there are 19 Intermediate matrixes, each Input matrix must have 19 outputs and each Output matrix must have 19 inputs. Therefore, the Input matrixes will be 10 inputs by 20 outputs and the Output matrixes will be 20 inputs by 10 outputs.

It is interesting to note that the switcher matrix sizes described maximize the use of current BTS TVS/TAS routing switcher rack hardware. Assuming that this is a video switcher, a pair of Input and Output matrixes may be built in one rack frame with power supplies. The 30x30 Intermediate matrix can be built in a single rack frame with power supply. Audio routing switchers allow greater flexibility than video switchers in the organization of switcher rack frames.

A 200x200 rectangular matrix requires 40,000 crosspoints, the three-stage switcher as described requires only 15,600 crosspoints. On a cost-per-crosspoint basis, the three-stage switcher would be 39% the cost of a rectangular matrix.

If a 200x200 rectangular matrix were expanded to 300x300, requiring 90,000 crosspoints, the number of crosspoints is increased by 50,000 crosspoints or increase of 125%. Many more rack frames would be required to make such an expansion. The three-stage switcher requires ten (10) more Input and Output matrixes, each taking one rack frame, plus the additional switch cards to increase the size of the Intermediate matrixes. This reflects an increase from 15,600 to 29,100 crosspoints, a difference of only 13,500 crosspoints or an increase of 87%. Keep in mind that this is only 32% of the number of crosspoints in a corresponding rectangular matrix.

The following table compares rectangular and three-stage video switchers of various sizes.

Switcher Size	Number Of Crosspoints Square	3-Stage		Number Of Racks Used
150x150	22,500	13,600	<60%>	34 (19 more)
200x200	40,000	15,600	<39%>	39 (1 less)
250x250	62,500	27,100	<43%>	41 (9 less)
300x300	90,000	29,100	<32%>	46 (14 less)

Table-1. Comparison Of Crosspoint Counts In Several Routing Switcher Sizes

Note that some configurations are not as efficient in use of equipment rack hardware. The rack frames and power supplies are a major portion of the cost of an actual switcher.

THREE-STAGE SWITCHING TRADEOFFS

It appears from the preceding examples that there is significant economic reason to use three-stage switching in large matrixes. The cost-per-crosspoint basis is useful for general comparison, but does not reflect the true system costs. The fixed hardware costs must be evaluated based on the system design and growth sizes.

Control of the switching matrix is a major concern and requires sophisticated computer processing power. Recent advances in system control and virtually transparent large-scale switching systems make three-stage switching practical.

Three-stage switching requires a complex system of signal interconnection wiring. Each cable connection is a possible trouble location. However, a large rectangular array with input distribution amplifiers requires as complex a wiring interconnection. The signals always pass through three-switch crosspoints, each switch may degrade the signal quality.

Three-stage switching offers some interesting system advantages other than economic. A system with a large number of outputs does not require input distribution amplifiers and associated cabling. Input loading is greatly reduced. In the example 200x200 array, the maximum input signal loading is twenty (20) outputs. The performance of small routing switchers is superior to that of large switchers. Dense matrix packaging is possible, maximizing the use of available real estate.

A signal may take any possible path through a three-stage switcher. Any differences in path delays could cause serious problems. In the example matrix, using five (5) defined cable lengths, the matrix is zero-timed input to output regardless of the path the signal may take. This is true even from output decade to output decade. Such path timing is important with video as well as stereo audio when the channels are split between switcher levels. Component video switching requires that the channels track both in absolute timing as well as DC levels. This degree of control is not achieved with rectangular switching arrays using looping output decades.

THREE-STAGE SIGNAL BLOCKAGE

A multi-stage switching matrix is designed to support telephone exchange switching. In a telephone system, a subscriber (input) selects another subscriber (output). A telephone exchange is **input-oriented** with an input connected to only one system output. The input can always connect to an output, provided it is not busy.

A signal distribution and routing switcher is **output-oriented**. Users at the output of the switcher select from any of the defined system inputs. It is common for an input to be selected to more than one system output. Consider the use of Tone or Color Bars, typically these signals appear on many system outputs at the same time.

The three-stage architecture does not anticipate an input being used by multiple outputs. Computer simulation demonstrates that a three-stage switcher can be blocked when used in an output-oriented system. Blockage prevents a user from accessing a desired input signal.

This condition does not meet the basic crosspoint reduction requirement of unlimited input signal access. A method was required to prevent a three-stage switching system from blocking.

The approach taken by BTS combines additional hardware and sophisticated matrix controller software.

The number of Intermediate matrixes in a three-stage switcher is normally defined as the **sum** of the number of inputs per input matrix plus the number of outputs per output matrix **minus** one. The use of one more intermediate matrix greatly decreases the probability of signal blockage and reduces the controller software task. This results in what is referred to as a modified three-stage switching matrix.

The matrix controller software uses a complex path-finding algorithm with five (5) distinct search modes. The basic requirement is to re-use existing signal paths whenever possible. In this manner, a connection may be completed with less than three crosspoint switches and redundant signal paths are reduced. However, with random switching such as occurs in a tele-production facility it is possible to create duplicate paths which may prevent access to a required input source. In this case, the controller performs two types of switcher re-organization or signal re-routing. This requires that one or more existing signal paths be re-formed allowing access to the required input signal. Since this change may affect a current user, the signal switches must be transparent to the user. Careful switcher design and a smart controller meet this requirement.

Although the probability of signal blockage in a modified three-stage switcher is very low, a complete control system must be able to deal with all possible switcher conditions.

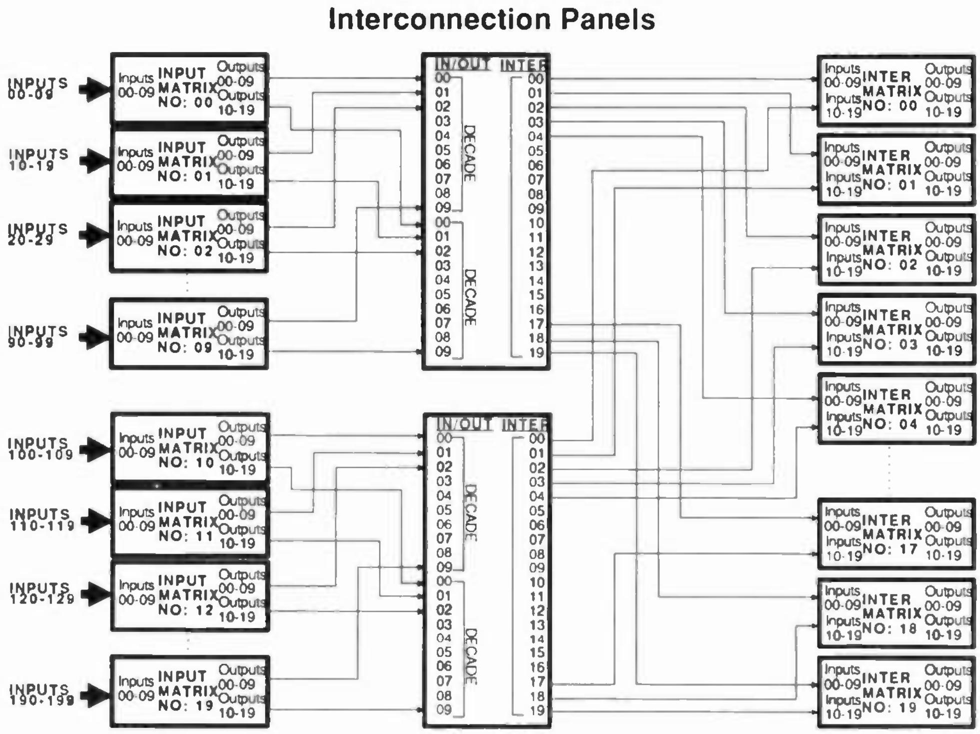

Figure-7. Audio Interconnection System

PRACTICAL THREE-STAGE SWITCHERS

BTS/Broadcast Television Systems is currently building and installing video and audio three-stage switchers. The TVS/TAS-3000 Routing Switcher provides the transparency required for superb signal quality. The BCS-3000 Control System includes the CE-3000 matrix controller which is designed for control of switchers using three-stage and path-finding arrays.

BTS has tested several three-stage routing switchers including video and audio and range in size from 100x100 (expandable to 200x200) to 200x200 (expandable to 400x400). In each of these switchers, the final expansion size was the major factor in choosing multi-stage switching.

A unique digital audio routing switcher using the AES Digital Audio signal standard is part of the ULTIMO facility in Sydney, Australia. This switcher combines high-speed digital switching with three-stage matrix design to create a powerful distribution system which include 18-bit coders/decoders for analog sources. This multi-stage switcher uses flat-wire cables to special interconnection panels to reduce the number of interconnection cables and greatly simplifies matrix installation. The digital matrix is designed for 180 inputs and outputs (expandable to 200 inputs and outputs). Since the AES digital audio signal is inherently two-channel, BTS was able to eliminate an entire switcher level, further reducing system cost and physical rack size.

3-STAGE AUDIO SWITCHER PERFORMANCE

The testing of the 200x200 (expandable to 400x400) audio three-stage switcher showed that the design goals for crosspoint reduction signal quality can be met. The following table contains several typical performance values for the three-stage switcher compared to the BTS published specifications for a standard audio switcher. The three-stage audio switcher compares favorably or exceeds the standard audio routing switcher specifications.

System Parameter	Standard Audio Switcher	Three-Stage Audio Switcher
Frequency Response (20 Hz to 20 kHz, ref. 1kHz)	+/-0.1 dB	+/-0.2 dB
Harmonic Distortion (20 Hz To 20 kHz, 24 dBv)	0.007%	0.010%
SMPTE IMD (0.0 To +24.0 dBv)	0.005%	0.005%
Crosstalk (20 kHz, most hostile conditions)	>-90 dB	>-100 dB

Table-2. Comparison Of Audio Performance For 200x200 (expandable to 400x400) Three-Stage Audio Switcher

CONCLUSION

Powerful control systems, transparent signal switching hardware and compact packaging make very large signal distribution systems practical. Path-finding and Three-Stage architectures offer significant hardware costs saving and space reductions. By proper planning the matrixes can be easily and economically expanded. In a very large facility, rectangular, path-finding and three-stage switching matrixes may be combined, further reducing system costs.

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1. Clos, Charles, "A Study of Non-Blocking Switching Networks, The Bell System Technical Journal, Volume XXXII, 1953, AT&T, pp. 406-424

CHANNEL 69 FILTERING FOR LAND MOBILE COMPATIBILITY—MIAMI, FLORIDA

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John F.X. Browne Associates
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Dielectric Communications
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ABSTRACT

Continued growth of communications systems has put a strain on the available spectrum with new services coming on line as needs grow. Recent allocations of land mobile channels adjacent to the UHF television band have met with varying degrees of success because of the large RF power difference involved. This paper describes the implementation of a filtering system for a Channel 69 transmission system with a 120 KW transmitter and suitable antenna gain to give 5 megawatts ERP, ellipically polarized.

THEORETICAL MODELS AND FIRST TESTS

Out of band radiation products are produced in several ways in a television transmitter. Analysis of spurious products begins with inspection of the baseband video input.

Video is composed of luminance fundamentals interleaved with chrominance and synchronization signals. Each element occupies a different segment of the video spectrum. The sync signal contains odd harmonics of the horizontal and vertical scan rates which decrease linearly in amplitude with increasing frequency. These harmonics are present at very low, but significant levels at frequencies higher than the 4.2 MHz band limitation. (See Figure 1)

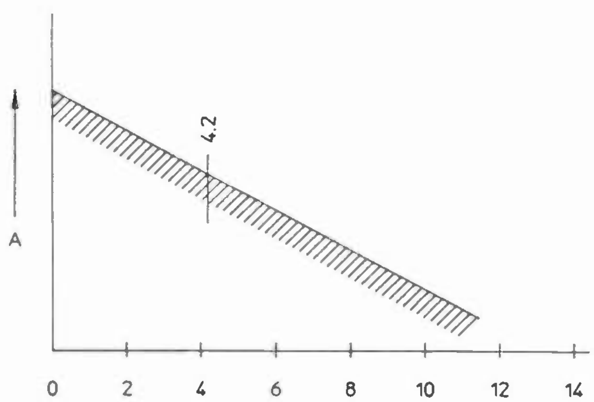


FIGURE 1

Since the sync signal is fixed amplitude, its' harmonic level remains constant. Active video and chrominance signals, on the other hand, are constantly changing with picture and detail content, and can be much larger than the fixed level of the sync, and also extend beyond the 4.2 limit. When all of these are amplified, nonlinearities in the amplifier create harmonic and intermodulation products that increase the out-of-band levels considerably. (See Figure 2)

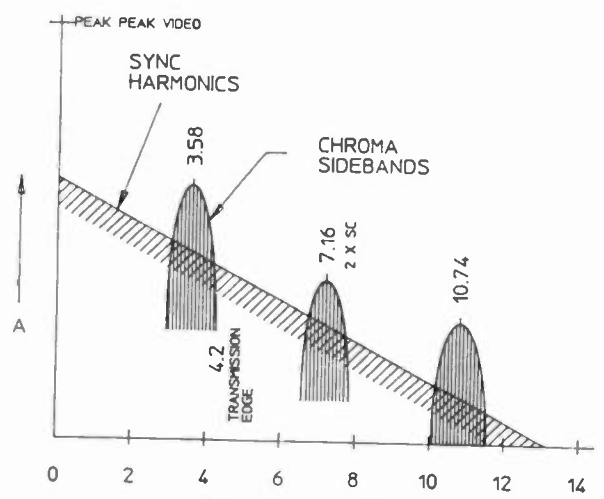
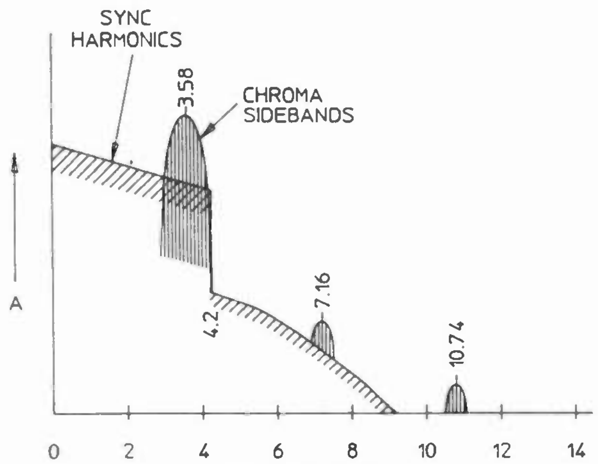


FIGURE 2

Although most of these video harmonics and distortion products are filtered in the exciter by the video low-pass filter and vestigal sideband filter (See Figure 3), they continue to dominate as main producers of transmitter spurious output, but by slightly different mechanisms than what we see in a single amplifier.

Preliminary tests performed on an existing transmitter and specifications provided by the manufacturer showed that out-of-band spurious products range from 65 to 70 dB below peak for some video related frequencies, to around 85 dB below peak for wide band sync harmonic.

The sync spurious level corresponds well with a predicted level based on the input when multiplied by the exciter, klystron, and diplexer response curves. The video related spur, although they can be predicted, are more directly related to the klystron operating parameters, exciter pre-correction, and diplexer cross coupling, and accurate levels are more difficult predict. Measurements made in both the pulsed, and un-pulsed transmitter modes, show an increase of 5-10 dB in wideband sync harmonics with the pulser operating.



USB FILTER AND KLYSTRON SPECTRUM SHAPING.

FIGURE 3

SETTING ATTENUATION GOAL

Arriving at an acceptable attenuation ratio for spurious products is a compromise between cost, physical size, and insertion loss of the filters. To determine the filter ratio, we first calculated the overall range and then subtracted the known losses to end up with a required range for the filter.

As before, the system is to have an E.R.P. of 5 megawatts horizontal, with any axial ration of 22.1% to give an E.R.P. of 1.107 MW vertical. Current communications receivers have sensitivity of approximately -120 dBm or better. For the horizontal signal; 5 megawatts = 97dBm, 97dBm + (-120dBm) = 217dB dynamic range!

For the vertical signal; 1.107 megawatts = 90.5dBm, 90.5dBm + (-120dBm) = 210.5dB dynamic range.

If we assume the closest receiver to be 1/4 mile, the free space loss with be approximately 83dB. Therefore, ignoring the antenna vertical pattern gain change, the worst case additional spur filtering needed would be;

Total Range	217dB
Highest Xmtr Spur	-65dB
Free Space Loss	-83dB
=	69dB

The minimum for additional filtering would be;

Total Range	217dB
Lowest Xmtr	-85dB
Free Space Loss	-83dB
Normal Cross Pol Loss	-20
=	29dB

And for vertical the worst case would be;

Total Range	210.5dB
Highest Xmtr Spur	-65dB
Free Space Loss	-83dB
=	62dB

The minimum additional for vertical would be;

Total Range	210.5dB
Lowest Xmtr Spur	-85dB
Free Space Loss	-83dB
=	42.5dB

Notch filters for narrow frequency bands with depths to near 70dB are possible with as few as 2 High-Q cavities. In addition, mutual coupling between wider spaced cavities can provide an average rejection of around 45dB.

With these concepts, a seven cavity filter could provide a high average rejection with notches placed directly on identified spur frequencies for even greater attenuation.

From the above examples, we see that it is possible to reduce spurious outputs to more than 110dB below the peak sync carrier level. Measurement of such a quantity is made possible by a prefilter to notch the visual carrier by greater than 30dB, the aural carrier by 20dB or more, and the upper sideband by a nominal 15dB, with no more than 2dB insertion loss at any of the spur frequencies. (See Figure 4)

From these rough calculations, we set 100dB below peak visual carrier as the maximum for any spurious products from the transmitter.

DETERMINING SPURIOUS PRODUCT FREQUENCIES

Examination of the exact mechanisms that generate spurious products reveals that there are 3 distinct sources.

1. Direct harmonics of the sync pulse.
2. Harmonic distortion in the visual amplifier. (i.e. 2 x subcarrier)
3. Intermodulation due to cross coupling of the visual and aural carriers.

The first two may be lumped together and dealt with simultaneously since they are both passed by the visual circuits. The third is a special case and must be separated. Applying the standard 2A-B intermod calculations to the visual signal we find that all intermod products fall on the low side of the upper band edge;

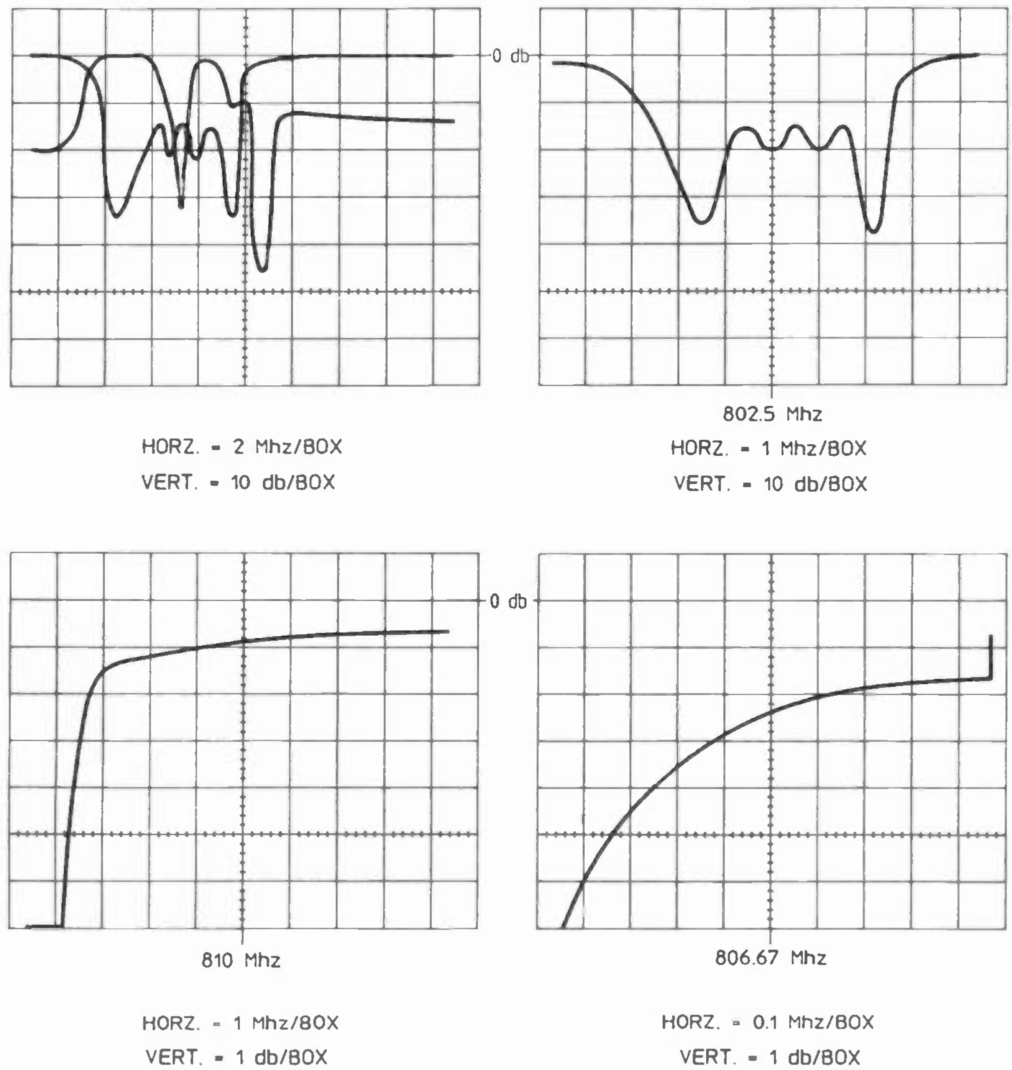


FIGURE 4

V + A, (2 x 801.25) - 805.75 = 796.75 MHz

+SC + A, (2 x 804.83) - 805.75 = 803.91 MHz

With leakage of the visual into aural we get;

A + V, (2 x 805.75) - 801.25 = 810.25 MHz

A + 3.58, (2 x 805.75) - 804.83 = 806.67 MHz

A + (-3.58), (2 x 805.75) - 797.67 = 813.83 MHz

These are the main spurs produced by this particular action, but not the only ones. Vestigal sideband response remains to 1.25 MHz below visual, as well as lower amplitude components below the VSB cut off. With a goal at 100 dB below peak for spurs, these components are important and must be recognized in the filter implementation. (See Figure 5.)

When the actual transmitter to be used became available, measurements were made using the prefilter before the spectrum analyzer to verify frequencies, band widths and relative levels. Several minor unpredicted spurs were found, but they were close to other spurs and small enough in amplitude to be eliminated by the existing filter design.

The exciters were also measured to gauge their contribution to the spurious output. Again several excess products were found. Another set of filters were designed to be added in the visual drive line to the klystrons.

SPECIAL FILTERING FOR PULSER OPERATION

Because a transmitter of the same type was not available at the filter design stage, plans for pulsing to reduce electric costs were put on hold until after the initial installation and testing with filters. When testing could be performed, it was found that additional filtering would be needed for spurs that were not present in the unpulsed mode. Another set of filters were designed and added to the 2 visual output lines.

To aid in further reducing spurious products due to the pulser, a special filter was designed to be added between the pulser output and the beam control element of the klystron. Key factors in this filter were; low input and high output impedance, high peak current capability, fast risetime, low cutoff frequency and less than 1% overshoot (equalized group delay). Computer optimization produced a design which met <u>all</u> of the requirements, and final tests show that it reduces the broad band harmonics of sync by 2 to 13dB in the pulsed mode.

DIPLEXER COUPLING

As before, a simple mathematical analysis shows that all I.M. products above the upper band edge are produced in the aural amplifier, while the visual produced predominantly harmonic distortion in this range. The notch diplexer is responsible because it is here that the two are combined and supposedly isolated. The actual isolation is roughly -30dB from the antenna port to the input port.

With this amount of cross coupling, the visual carrier is injected into the aural klystron 20dB below the aural carrier level. Assuming a high conversion loss of -20dB an intermod product is produced that is now 50dB below visual. It travels back out the aural line, looses another 30dB to cross coupling and is conducted to the antenna at 80dB below the visual.

Most of the spurious energy is dissipated in the notch diplexer reject load, as would be predicted by a perfect model, but some does escape. While this may be fine for normal transmitter use, we require all spurs to be below 100dB and this would obviously need to be filtered.

SUMMARY OF THEORETICAL MODEL

Through careful planning and analysis, it is possible to operate a high power UHF television system directly adjacent to low power land mobile systems without interference.

DESIGN OF THE HIGH POWER FILTERS

In order to reduce the spurious output to an acceptable level Home Shopping Network generated two plots of frequency vs. attenuation. Figure #6 shows the proposed visual filter response. The station wanted 50dB attenuation at 806 MHz, 806.67 MHz, 808.41 MHz and 809.33 MHz as well as 62 dB attenuation at 810.25 MHz. Two of these frequencies are clearly spurs produced by the mixing of the

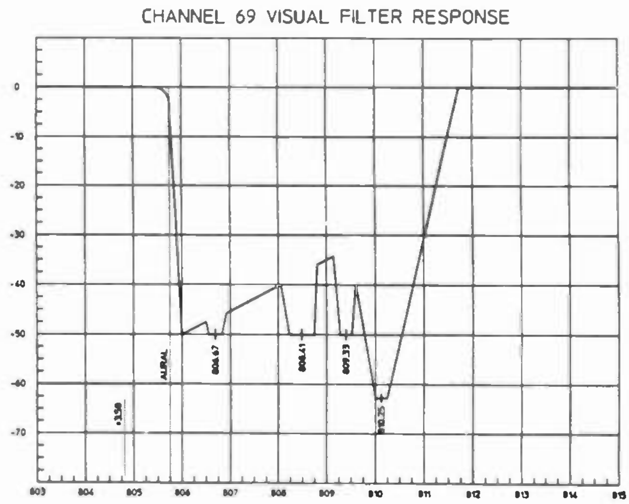
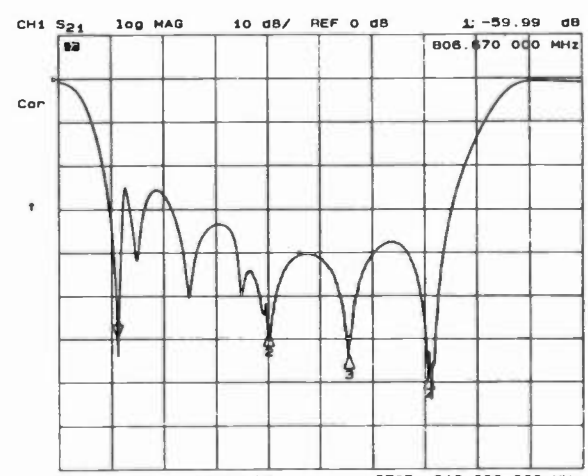


FIGURE 6

chrominance carrier with the aural carrier resulting in a spur at 806.67 MHz and the visual mixing with the aural producing a spur at 810.25 MHz. In addition, they wanted as much bandwidth as possible on either side of the above frequencies.

VISUAL FILTER DESIGN

Figure #7 illustrates the response of the visual filter designed by Dielectric to meet the requirement. It was necessary to tune the notches to a maximum depth in excess of 60dB



MARKER 1	806.670000 MHz -60.035 dB	806.670000 MHz OFF
MARKER 2	808.410000 MHz -59.711 dB	808.410000 MHz OFF
MARKER 3	809.330000 MHz -61.38 dB	809.330000 MHz OFF
MARKER 4	810.250000 MHz -65.831 dB	810.250000 MHz OFF
MKR STIMULUS OFFSET	0.000000 MHz 0 dB	0.000000 MHz 0 dB

FIGURE 7

in order to obtain any bandwidth at the desired frequencies. It was not possible to obtain any significant attenuation at 806 MHz and still obtain any reasonable insertion loss in band. The following table lists the bandwidth obtained at each of the other required frequencies.

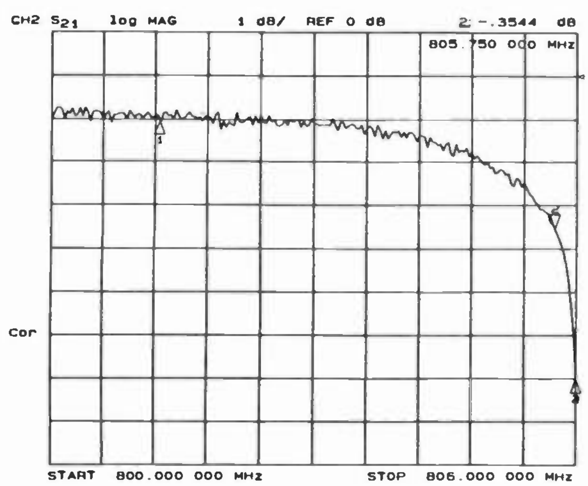
Frequency	Attenuation	Bandwidth
806.67 MHz	> 50 dB	30 KHz
808.41 MHz	> 50 dB	185 KHz
809.33 MHz	> 50 dB	180 KHz
810.25 MHz	> 50 dB	110 KHz

Within the passband from 800 MHz to 806 MHz the insertion loss ranged between .1 dB at the visual carrier to .2 dB at the high end of visual. This is plotted in Figure #8.

The return loss is plotted in Figure #9 and is in excess of 30dB for the whole band and exceeds 40dB at the visual and aural carriers.

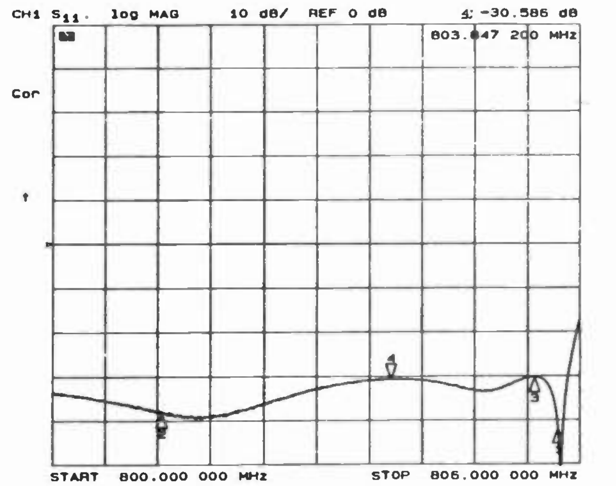
AURAL FILTER DESIGN

The second plot provided to Dielectric by Home Shopping Network is shown in Figure #10. This is the required attenuation for the aural input line of the diplexer. 35dB of attenuation was required at 804.83 MHz, 806.67 MHz and 813.83 MHz and 60dB at 810.25 MHz. These are intended to reduce the spurs created by the mixing of the +3.58 MHz chrominance carrier with aural, the visual with aural and the -3.58 chrominance carrier with aural. Three of these notches attenuate the spurs directly while the leftmost notch attenuates the incoming +3.58 MHz chrominance carrier.



MARKER 1	801.250000 MHz OFF	801.250000 MHz -.0975 dB
MARKER 2	805.750000 MHz OFF	805.750000 MHz -.3553 dB
MARKER 3	806.000000 MHz OFF	806.000000 MHz -.6892 dB
MARKER 4	806.000000 MHz OFF	806.000000 MHz -.6892 dB
MKR STIMULUS OFFSET	0.000000 MHz 0 dB 0 °	0.000000 MHz 0 dB

FIGURE 8



MARKER 1	805.750000 MHz -41.846 dB	805.750000 MHz OFF
MARKER 2	801.250000 MHz -37.965 dB	801.250000 MHz OFF
MARKER 3	805.484000 MHz -30.293 dB	805.484000 MHz OFF
MARKER 4	803.847200 MHz -30.507 dB	803.847200 MHz OFF
MKR STIMULUS OFFSET	0.000000 MHz 0 dB	0.000000 MHz 0 dB

FIGURE 9

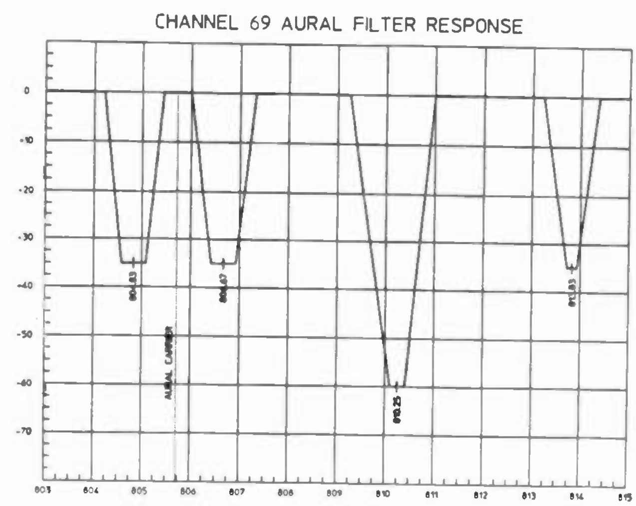
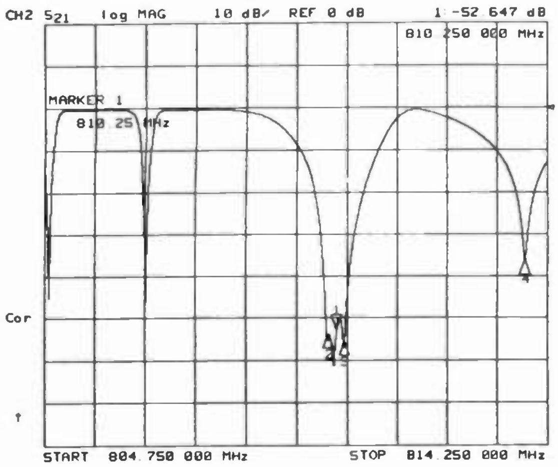


FIGURE 10

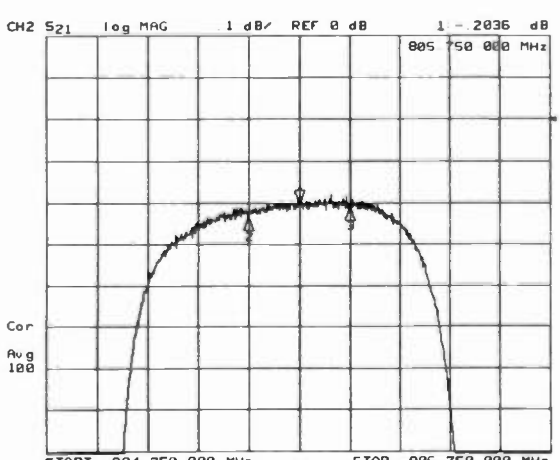
Figure #11 illustrates the attenuation of the aural filter built to meet this requirement. The following table lists the frequencies, bandwidth and attenuation for that bandwidth for each of the points of interest.

Frequency	Attenuation	Bandwidth
804.83 MHz	> 35 dB	20 KHz
806.67 MHz	> 35 dB	20 KHz
810.25 MHz	> 60 dB	400 KHz
813.83 MHz	> 35 dB	13 KHz



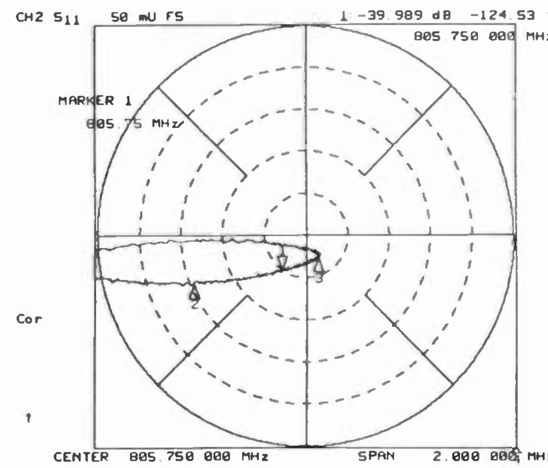
MARKER PARAMETERS	Channel 1	Channel 2
MARKER 1	810.250000 MHz OFF	810.250000 MHz -52.904 dB
MARKER 2	810.100000 MHz OFF	810.100000 MHz -54.841 dB
MARKER 3	810.400000 MHz OFF	810.400000 MHz -56.233 dB
MARKER 4	813.830000 MHz OFF	813.830000 MHz -35.961 dB
MKR STIMULUS OFFSET	0.000000 MHz 0 dB 0 °	0.000000 MHz 0 dB

FIGURE 11



MARKER PARAMETERS	Channel 1	Channel 2
MARKER 1	805.750000 MHz OFF	805.750000 MHz -.2036 dB
MARKER 2	805.550000 MHz OFF	805.550000 MHz -.2322 dB
MARKER 3	805.950000 MHz OFF	805.950000 MHz -.2080 dB
MARKER 4	806.750000 MHz OFF	806.750000 MHz OFF
MKR STIMULUS OFFSET	0.000000 MHz 0 dB 0 °	0.000000 MHz 0 dB

FIGURE 12



MARKER PARAMETERS	Channel 1	Channel 2
MARKER 1	805.750000 MHz OFF	805.750000 MHz -39.879 dB -122.83 °
MARKER 2	805.950000 MHz OFF	805.950000 MHz -30.727 dB -157.14 °
MARKER 3	805.550000 MHz OFF	805.550000 MHz -44.679 dB -63.419 °
MARKER 4	806.750000 MHz OFF	806.750000 MHz -2.8187 dB -52.103 °
MKR STIMULUS OFFSET	0.000000 MHz 0 dB 0 °	0.000000 MHz 0 dB 0 °

FIGURE 13

The insertion loss at aural as shown in Figure #12 is .20dB at carrier and <.23dB at ±200 KHz.

The return loss as plotted in an exaggerated scale in polar form on Figure #13 shows a value of 40dB at carrier, >35dB ±100 KHz and >30dB ±200 KHz. This translates to a VSWR of less than 1:05:1 over the bulk of the band.

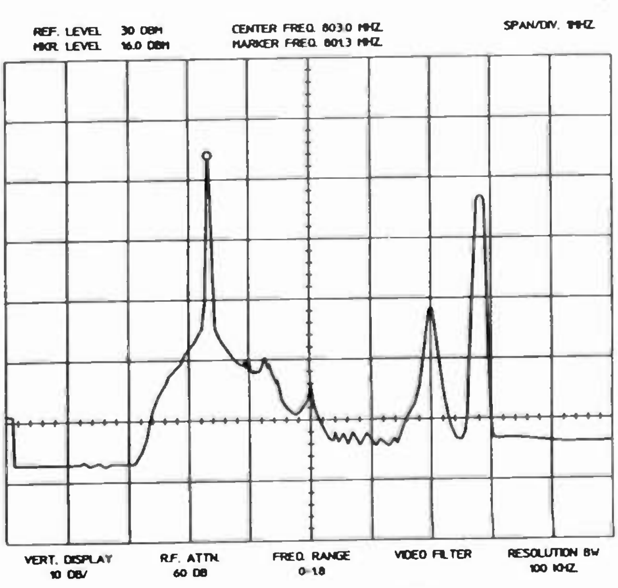


FIGURE 14

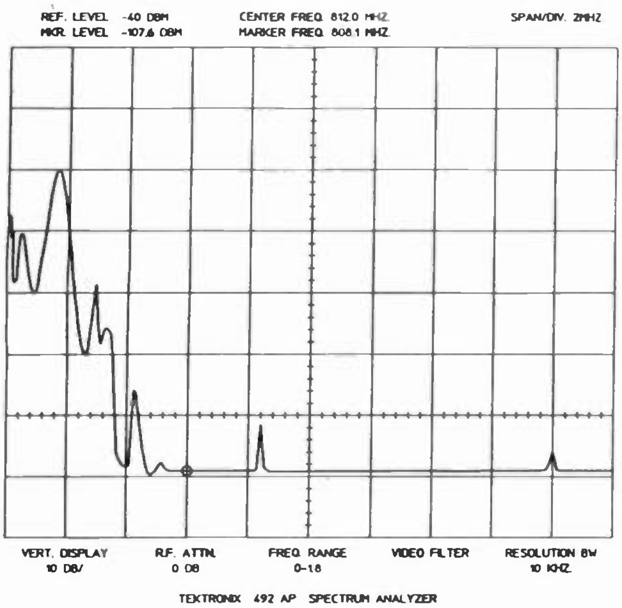


FIGURE 15

The specific high power filters discussed to this point were designed to allow normal transmitter operation that is, with high power multiplexing and without pulsing. The resulting output of the entire system is shown in Figure #14 for the in band response. You can see all frequencies above aural in this figure and attenuated down to at least the noise floor of the measuring instrucment. For the frequencies between 802 MHz and 822 MHz figure #15 illustrates the output. In this figure the aural carrier has been attenuated with a notch to improve the resolution of the measuring instrument. By attenuating the aural the noise floor in this figure is -107 dBm as opposed to -38dBm in the previous plot. The sharp spike in the second column from the left is at 805.75 MHz and is the remnant of the

aural carrier as seen through the notch in the line of the measuring instrument.

ATTENUATION SUMMARY

Since the visual carrier in Figure #14 was referenced at 16dBm and the noise floor in Figure #15 is referenced at -107.6dBm, there is 123.6dB difference between carrier and noise floor in Figure #15. There are only 4 small spikes that rise above the noise floor at frequencies above aural. The one at 806.67 MHz is at -110dB below carrier. The one at 810.25 MHz is at -115dB below carrier. All other spikes are greater than -120dB below carrier which was the original goal for the above listed filters.

PHOTOS OF FILTERS

Figures #16, 17 and 18 show views of the two visual filters and the aural filter used to achieve these results. The visual filters are a matched set of waveguide filters about 12 feet long that feed the input hybrid of the 120 KW diplexer. They were built as a matched set to reduce the risk of excessive heating and possible resulting drift. In retrospect they have run so cool that a single filter could have been used between the hybrid and the diplexer at a reduced cost with no adverse performance inpact.

SUPPLEMENTARY FILTERS FOR PULSING

As mentioned earlier, this series of filters accommodates the normal mode of operation with normal diplexing. In addition, it was determined that two additional modes of operation should be allowed. Low level muliplexing will be necessary in the event of failure of the aural tube and pulsing will be required to improve the efficiency of the transmitter. Operation in each of these modes produces an additional series of spurs just above 860 MHz and also between 810 MHz and 818 MHz. When this was determined, supplementary filter was designed by Dielectric to further attenuate the visual transmitter. Figure #19 shows the additional rejection. It supplied attenuation in accordance with the following table:

Frequency	Attenuation	Bandwidth
812.09 MHz	> 35 dB	1.517 MHz
806.43 MHz	> 20 dB	.032 MHz
816.75 MHz	> 30 dB	.217 MHz
814.75 MHz	> 40 dB	.072 MHz

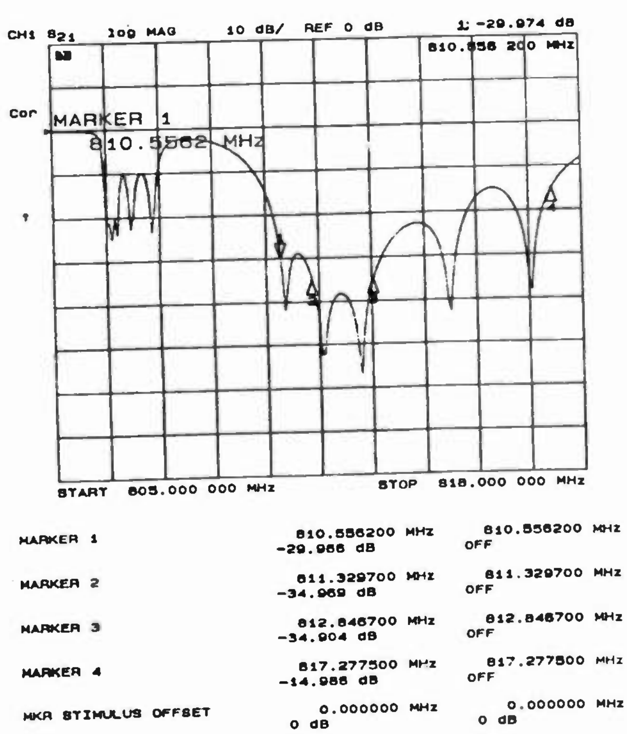


FIGURE 19

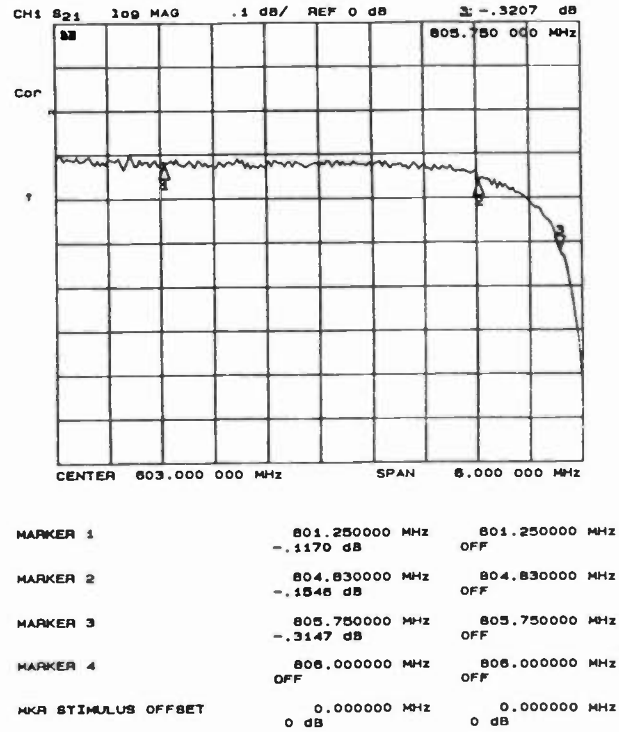


FIGURE 20

The insertion loss for the supplementary visual filter is shown in Figure #20 to be less than .15dB for the entire visual band and < .32 at aural. Aural will pass through this and the original visual filter when low level multiplexing is utilized so the filter is

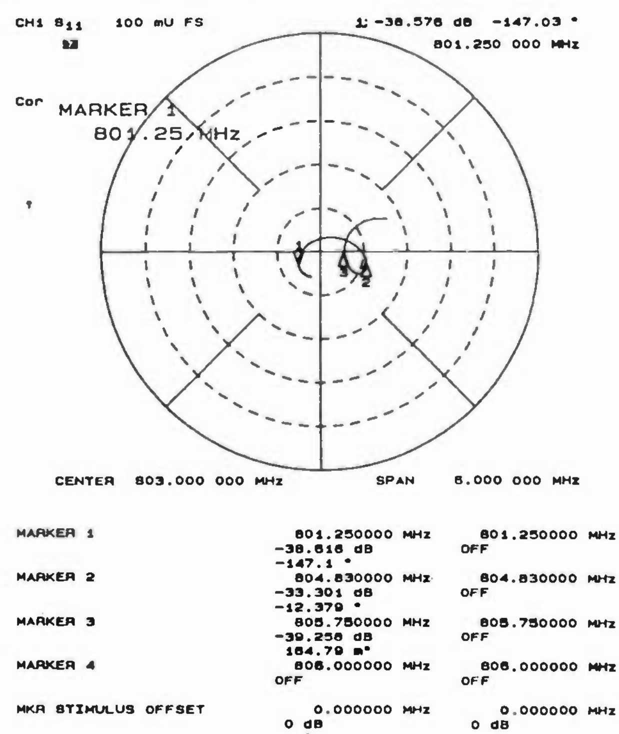


FIGURE 21

designed to accommodate that mode of operation. Return loss is plotted in polar format on Figure #21 and is better than 33dB across the band or better than 1.05:1 VSWR. Figure #22 illustrates these filters again in a matched set mounted near the ceiling in coaxial line.

The resulting RF output spectrum with these modes of operation in use was not available at press time, but will be covered at the presentation of the paper in Atlanta.

LAND MOBILE COORDINATION

The firm of John F.X. Brown & Associates was engaged to coordinate the land mobile aspects of the activation of Channel 69 in Miami. This project had major components which included specifying the out-of-band characteristics of the transmitter, output combiner and diplexer, designing and manufacturing filters to be installed in the transmitter and, external to it, installing, testing and measuring those devices, and, finally, to coordinate construction activities with the land mobile community and the Federal Communications Commission. Our firm was chiefly responsible for these latter activities.

The FCC placed a very stringent condition on the Channel 69 construction permit which required that the permittee prove that no interference was being caused to existing 800 MHz land mobile users in the Miami area. Program test authority was conditional on the FCC receiving of documentation of non-interference.

The first step in the process was to identify all of the 800 MH Land Mobile licenses from the FCC's Private Radio data base. To our great surprise the Commission listed some 2500 licensees. The good news was that most of these licensees turned out to be individual users of trunked systems. The bad news was that several major base stations were located within a mile or so of the TV transmitter site including two which were colocated. One licensee's base station was situated 0.2 mile distant and operated on 806.41325 MHz only 662.5 KHz above the Channel 69 aural carrier. The Miami Police Department had a ase station 8.0 miles distant on 806.125 MHz only 375.0 KHz above the Channel 69 aural carrier. With receiver sensitivities better than -110 dBm, operating in close physical and spectral proximity to the 5 megawatt (+97 dBm) Channel 69 facility it was apparent that there was going to be problems. This, of course, is a 200 dB spread between the ERP of Channel 69 and the receiver threshold of the base station.

After learning of the capabilities of the Dielectric filters and the NEC transmitter and calculating the received undesired signal at these base stations, we had a reasonable assurance that the objectives could be met.

In addition to the spurious products which might emanate from the transmitter, we were also concerned about the potential for intermodulation products resulting from combinations of Channel 69 and the other UHF sttions on the tower which included Channel 33 and 51. Also, a nearby Channel 45 was considered in this process.

All licensees were notified by mail approximately 60 days prior to the anticipated date for first equipment tests and were given a general description of what was being constructed, and assurance that interference was unlikely and a contact telephone number was provided in case there were any questions. Of the approximately 2500 notices mailed less than two dozen generated any kind of an initial response.

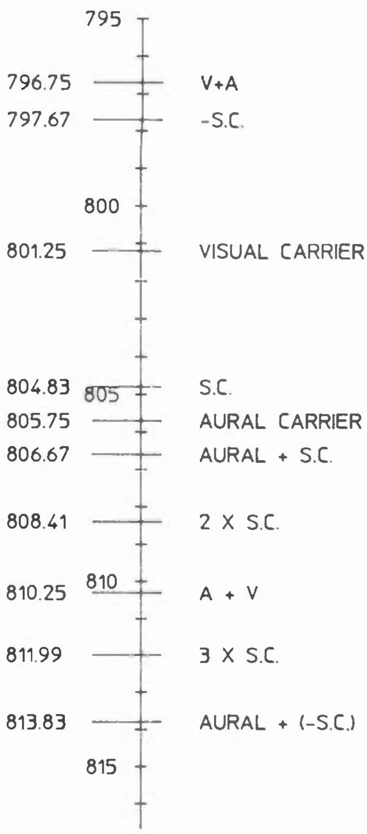
Many months prior to this mailing the real coordination began with major users such as Motorola and the Miami Police Department. In fact, one of the facilities colocated with the TV station was operated by Motorola and employed a base station antenna situated less than 50 feet below the TV antenna.

The major land mobile users were skeptical but very cooperative; they appreciated the long-lead notice of almost one year and worked with us to help assure that the project would have a good chance of succeeding.

When it appeared that the station was ready for operation, the licensees were again notified with more specific estimates of the actual date of the first on-air tests. Prior to actual on-air testing the spectral characteristics of the transmitter output were studied carefully while operating at full power into a dummy load and under varying conditions of visual and aural modulation. The results of the filtering efforts appeared very promising (as out-of-band products appeared to be attenuated at least 100dB below the visual carrier). However, several discrete spurious products were slightly above the objective. Since these spurs did not fall on any known frequency in use, it was decided to attempt initial on-air equipment tests. A specturm analyzer was connected to the colocated land mobile system's antenna and was monitored as the power was gradually increased. It was immediatley noted that the noise floor was increasing with the presence of the Channel 69 signal even though discrete spurs did not seem to be a problem. At full power, the receiver noise floor had degraded over 10dB reducing receiver sensitivity to less than -100dBm. In anticipation of this event, a special bandpass filter had been secured for installation in the receiving system. The filter had approximately 30dB attenuation below 806 MHz and low insertion loss above this frequency. Installation of this filter essentially restored the land mobile system's receiver performance to its pre-Channel 69 level.

It is worth mentioning here that the original transmitting antenna was an omni-directional, horizontally-polarized, slot radiator producing an ERP of 2.5 megawatts that was being used on an interim basis. The ermanent antenna is an elliptically polarized directional antenna producing a maximum ERP of 5 megawatts.

Field measurements were taken to determine the effects of the antenna system on the out-of-band performance. In order to measure spurious signals in the range of -110dBm to



MAJOR SPUR FREQUENCIES.

FIGURE 5

-120dBm "off-air" it was necessary to employ a specially fabricated Low Noise Amplifier to provide an effective reduction in the noise floor of the specturm analyzer. It was also necessary to use a custom filter at the input of the analyzer in order to remove the TV carriers so as to prevent intermodulation in the front end of the spectrum analyzer and LNA. Slightly better results than the dummy load measurements were noted due to the power out-of-band performance of the antenna.

The station operated successfully for nine months on the interim antenna until the permanent antenna was installed. One concern with the permanent antenna was the addition of the vertically polarized antennas. The measurement and observation process was repeated after installation of the new antenna at which time additional filters were installed in the transmitter system to further reduce broadband noise and discrete spurs.

In over fourteen months of operation, there have been no legitimate compliants received from land mobile users. Two or three complaints were received before the station went on the air--in a wish to be the first to anticipate a problem. Two additional complaints were received after the start of interim operations and both complainants advised on follow-up that the problem was in their mobile equipment. Thus, we are happy to report that there have been zero incidents and on no occasion has the TV station had to reduce power or alter its operation to appease land mobile interests.

We believe this demonstrates that, with careful pre-planning and coordination with the land mobile licensees, Channels 14 and 69 can be used in high density land mobile environments.

FIELD PERFORMANCE OF A MULTIPLE STAGE DEPRESSED COLLECTOR KLYSTRON

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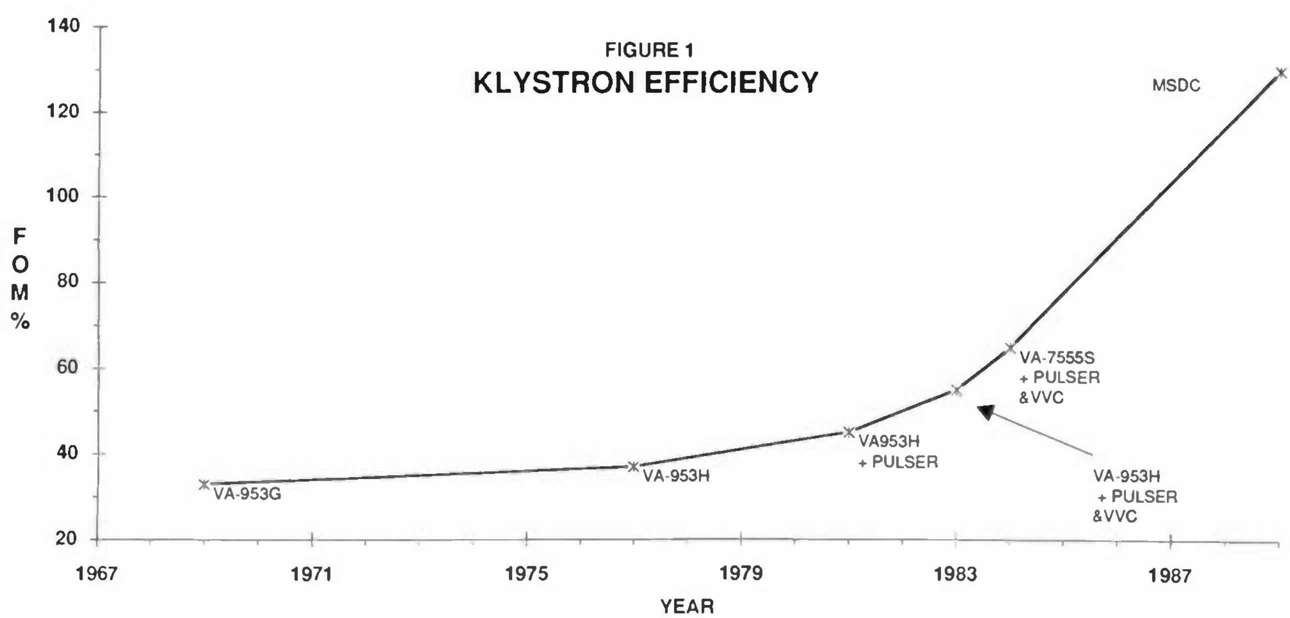
INTRODUCTION

UHF television klystron amplifiers have been with us for many years and improvements in operating efficiency has been the goal by a number of people for as many years. The purpose of this paper is to show you the latest results in this endeavor. I will show the hard facts which are now available on the field performance of the multiple stage depressed collector (MSDC) klystron in operation at WNVT-TV, CH 53 at Falls Church, Virginia. The transmitter built by Harris Broadcast Division went on the air February 4, 1990 using two Varian made MSDC external cavity klystrons. It replaced an old General Electric 55 kW transmitter using an Amperex external four cavity visual tube in the nonpulsed mode. The 60 kW Harris transmitter will be upgraded to 120 kW using three MSDC klystrons operating at 103 kW by the time this paper is presented.

EFFICIENCY

Let's start out by defining the term "figure of merit" as applied to a television klystron amplifier and the assumption that you already have a general working knowledge of how the MSDC klystron operates. Remember "figure of merit" (FOM) used to be called efficiency but when the number exceeded 100, some of us thought we were getting something for nothing and that doesn't happen. The definition of "figure of merit" is the ratio of peak synchronizing power output to power input during average picture



conditions. In mathematical form it looks like this:

$$FOM = \frac{\text{Peak sync power output}}{\text{Power input @ 50\% APL}}$$

Figure 1 shows the chronological improvements made in klystron operating efficiency up to 1990. In this illustration, the "figure of merit" is shown on the vertical axis.

The early internal cavity klystrons started out of a FOM of 33 in 1969. Then in 1977 the H series klystrons were introduced to bring the FOM up to 37. Mod anode pulsing and variable visual couples increased the FOM to 55 in the early 1980's. Further klystron improvements in the mid 1980's enhanced the FOM by 10 points to the 65 area. As this chart shows the 1990's using the MSDC has doubled the FOM to approximately 130.

Most of the power consumed in a UHF television station is the klystron amplifiers, therefore, most of the effort through the years has been directed toward improvements in the klystron FOM In the design of the new Harris MSDC klystron transmitter, other areas of reducing power consumption have been included. We feel a kilowatt saved is a kilowatt that won't have to be paid for year after year. A solid state annular ring (BCD) pulser is used to modulate the klystron beam current for further power reduction. Also, a ring air compressor consuming 2-1/2 horse power was replaced with a squirrel cage blower using only 1/3 of a horse power, and in the process substantially increasing the cooling provided to the klystron cavities and magnets. Cooler operation of the tubes and cavities should improve the life of these components.

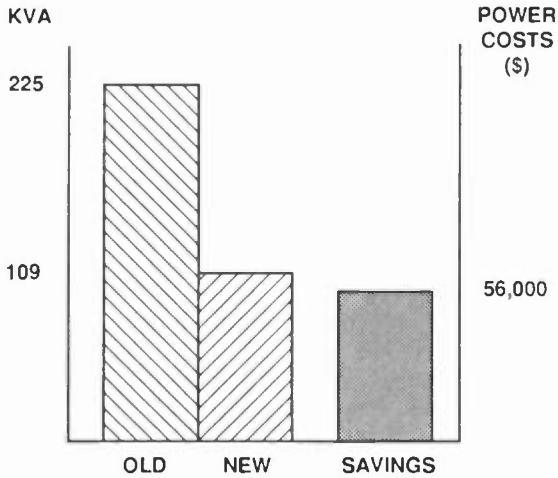
INSTALLATION

Power consumption data was obtained from Garr Johnson, Director of Engineering at WNVT, based on the operation prior to any changes to incorporate the MSDC klystrons. 225 KVA of energy was being used at the station. It should be pointed out at this time that the installation was made in several phases to enhance transmitter availability. There were zero hours of lost air time during the installation phase. The first phase of the installation was to install the new heat exchanger on an outside concrete pad and water/glycol pump module. This system was then used to cool the old General Electric transmitter which allowed the removal of the large old indoor cooling system. This change alone caused the power meter to slow perceptibly.

Three new beam power supplies were also set on the pad adjacent to the building along with the line control cabinets inside to have a completely separate power supply system for each klystron. This redundant system allows the complete and safe shut down of any klystron amplifier for service without interfering with the other klystrons. Much of the internal building AC interconnecting wiring was run during this phase.

The second installation phase brought the new TV60UM transmitter into the building and located the cabinets in the space provided. The new RF system was physically hung below the existing filterplexer and the transmitter connected to water, electrical, and RF system ready to go on the air. On February 4, 1990 WNVT went on the air for the first time with the MSDC transmitter. Power consumption figures now read 109 KVA. The difference between the new and old consumption readings would be 122 kW hours assuming a power factor of .95. Mr. Johnson is paying $0.07 per kilowatt hour which would give an annual savings of $56,000 based on an 18 hour day.

These typical annual savings could be realized in your station if you are presently operating an older unpulsed transmitter (see Figure 2).



WNVT Power / Cost savings
Using Harris TV60UM
Transmitter
FIGURE 2

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1) C. Smiley and G. Best, "Improved Efficiency in UHF Television Transmitters", 1984 NAB Broadcasting.

2) E.W. McCune, "UHF-TV Klystron Multistage Depressed Collector Development Program", Final Report NASA Lewis Research Center, 2100 Brookpark Road, Cleveland, OH.

PROGRESS REPORT ON KLYSTRODE® EQUIPPED TRANSMITTERS AT GEORGIA PUBLIC TELECOMMUNICATIONS NETWORK

Al Korn
Georgia Public Telecommunications Commission
Atlanta, Georgia

I. INTRODUCTION

Comark Communication, Inc., a Thomson-CSF Company, in cooperation with Eimac, a division of Varian Associates, Inc., has developed and delivered a line of Klystrode equipped transmitters to the Georgia Public Telecommunications Commission. The installations now on the air in Georgia are: WCES-TV, Channel 20, Wrens/Augusta, June 1988; WABW-TV, Channel 14, Pelham/Albany, October 1988; and WCLP-TV, Channel 18, Chatsworth, October 1989. A fourth installation will be on the air at WDCO-TV, Channel 29, Cochran/Macon, in June 1990. Each transmitter is 120 kW with an ERP of 5 MW.

This paper details the installation, testing, and performance of these transmitters sites to date.

II. THE EVOLUTION OF THE KLYSTRODE DEVELOPMENT

In 1979, the Eimac Division of Varian completed a survey of UHF broadcasters and OEM's. The message Eimac received was clear--UHF broadcasters, with their relatively poor efficiencies compared to the VHF TV industry, were finding it difficult to compete. A program was started to develop a 50 kW UHF Tetrode. Several designs were proposed, and it became clear to the Eimac engineering staff that it was possible to design such a tube, but it would be very difficult and costly. The result would be some improvement in efficiency due to the fact that tetrodes can be operated Class B, but only a marginal improvement at best on the high end of the UHF band. An even greater concern was the relatively short life compared to klystrons due to the increasing power densities required for 50 kW at the short wave lengths involved at UHF. Concurrent with this program, Eimac was doing R & D on a high voltage switch tube which involved a gridded strip beam gun and klystron-like collector. With some reservations about the feasibility of such a device, an experimental model was built, and the first tests gave an indication of better than expected results. These tests launched serious work on the Klystrode; the tetrode project was abandoned, and the stage was set for a series of developments. First came an aural only Klystrode, then the 30 kW vision, and finally, 9 years later, the 60 kW version that is now on the air at 3 of the 9 GPTV sites. (See figure 1.)

III. THE FIRST KLYSTRODE INSTALLATION

On October 18, 1987, GPTV engineers went to the Comark factory at Southwick, MA. for a test bed session. The configuration for testing was a 60 kW water cooled Klystrode, with both Comark and Eimac representatives on hand. After 2 days of extensive testing for efficiency, vision and aural quality, the conclusion reached by GPTV was to proceed with the manufacture and installation of the first transmitter. WCES-TV, Wrens/Augusta, Ga. was selected for this installation. The new transmitter, a 120 kW vision, 12 kW aural, water cooled unit replaced the 30 kW GE.

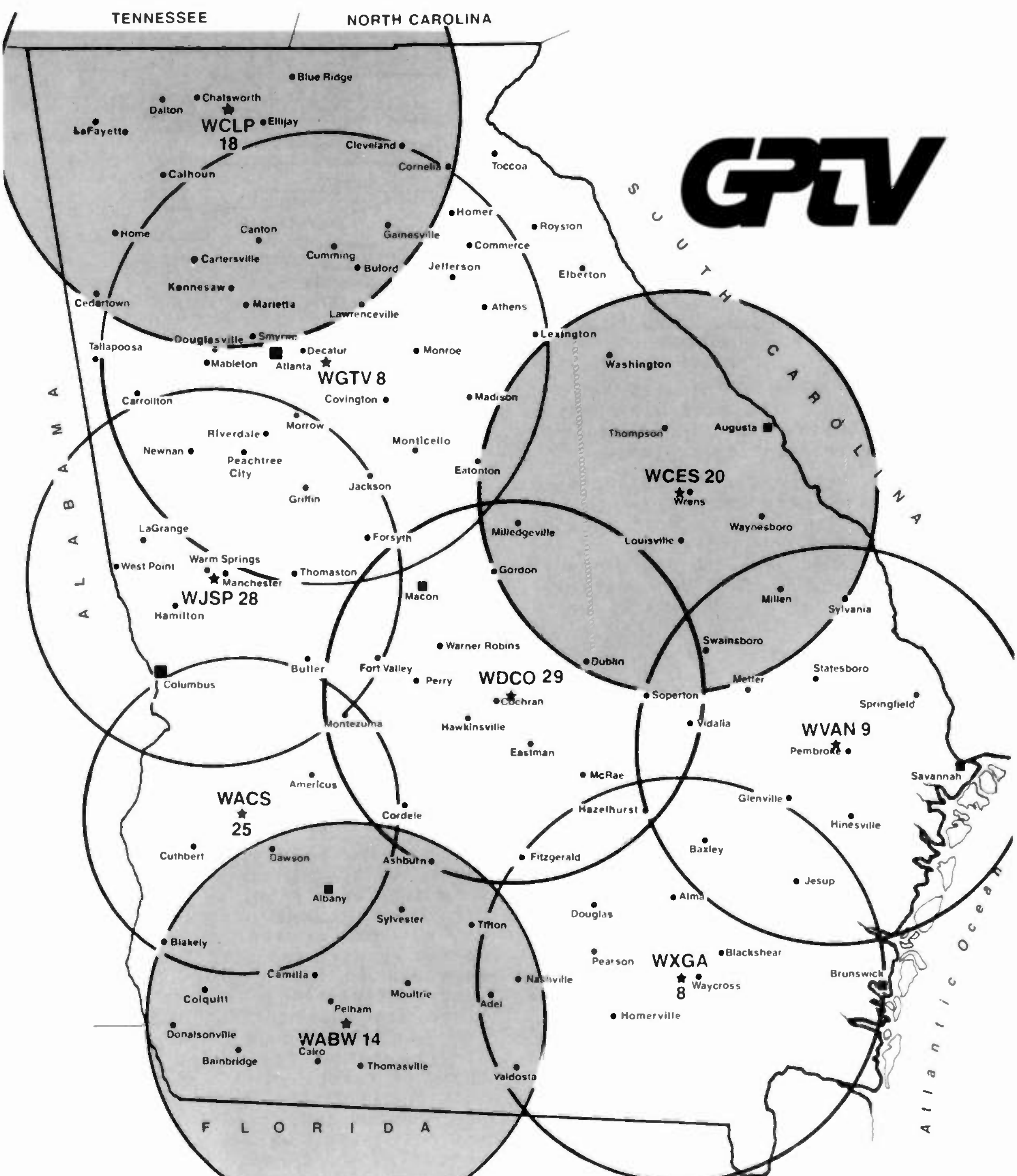
Construction began on May 1, 1988 with the removal of the 30 Kw GE transmitter, 1465 feet of transmission line and antenna. With the increase of transmitter power from 30 kW to 120 kW the total RF system needed to be replaced. The decision was made to go off the air for the changeover which was scheduled for 30 days. Five of the network chief engineers, the five man station staff, and representatives from Comark and Eimac were on hand for the construction phase.

On June 5, 1988, at 2:15 A.M., the proof of performance was completed after 34 days of construction. The transmitter design, (Figure 2) was developed to assure as much redundancy as possible. As with any new technology, problems can and do arise, so great care was taken to have 3 fall back

Klystrode is a registered trademark of Varian Associates, Inc.

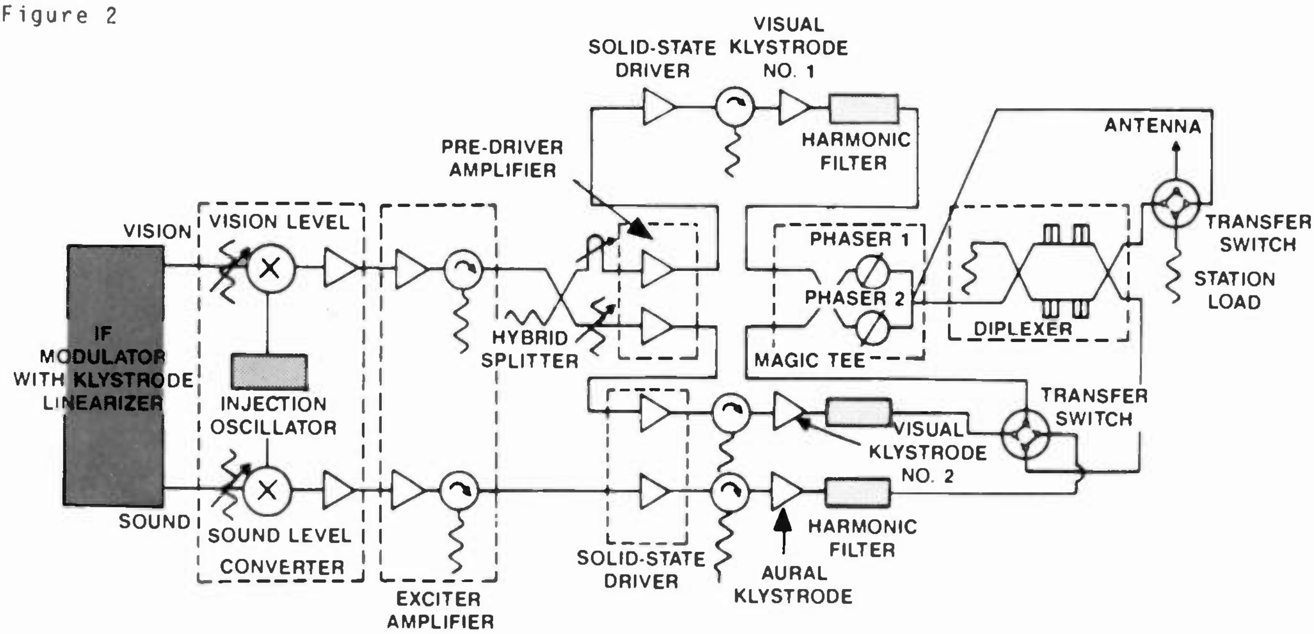
positions should one of the devices fail in operation. First, should either vision device fail, the transmitter can be operated with one vision and aural in a 60 kW configuration. Should the aural device fail, vision 2 can be brought on line as aural and still maintain the 60 kW transmitter output. If only one vision device

Figure 1



GEORGIA PUBLIC TELECOMMUNICATIONS COMMISSION
STATE TELEVISION NETWORK

Figure 2



Block diagram of the 120 kW SK-Series transmitter installed at WCES, Wrens, GA. The same basic design was used at WABW, Pelham, GA., and WCLP, Chatsworth, GA. Redundancy was a primary concern in the design of the transmitter systems.

is operational, either can be multiplexed for continued air service. Even though there is a reduction in ERP, most viewers are able to receive the signal during the repair process. This same configuration is carried through on all of GPTV's installations.

IV. NUMBER TWO

During the process of shaking down the new Wrens installation, we were preparing for the second unit at WABW-TV in Pelham, Ga. Having learned the process at Wrens, the crew was moved to Pelham in September of 1988 and installed the second transmitting plant. We were aware of a potential problem with Channel 14 and the possible interference with 2 two-way installations. When we completed the installation and went on the air our fears were confirmed. With a lower edge of Channel 14 at 470 MHz, one two-way radio 13 miles away at 469.95 MHz, and another 18 miles away at 469.75 MHz, there was a need for extra filtration to skirt the lower edge of the band. Comark built a 5 cavity wave guide filter to sharpen the lower edge of the band, and with its installation we were able to run at 100% power, thus eliminating the interference.

With these two installations on the air it was time to see if all the data collected on the Wrens transmitter could be duplicated. We had learned that at Wrens the average beam current at 50% APL and 104% peak RF output was 1.65 Amperes per tube, and 2.31 Amperes per tube at black. With a beam voltage of 31.5kV the figure of merit was 120%, which was extremely close to the projected numbers. At Pelham the same 50% APL measurement was taken and the results were 1.45 Amperes per tube and a beam voltage of 31.6kV, which increased the figure of merit to 131%. This data confirmed the efficiency of the Klystrode device and led us to the third installation at Chatsworth, Ga.

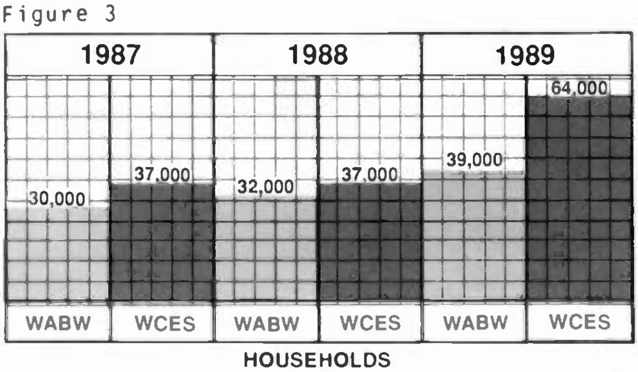
During the year between the Pelham and Chatsworth installation we were able to study the data collected daily at these sites, and we have come to some interesting solutions. First was a problem with multipacting. This process causes the ceramic of the tube to crack and thus go to air. Eimac replaced 4 tubes at Wrens because of this problem. Also, we found that at Pelham, with a lower frequency, the problem was of even greater concern. To date at Pelham we have replaced 5 tubes, 4 of which were due to multipacting. We were aware that multipacting could be a problem, and Eimac has now modified their tube with a ring in the anode area to change the electrostatic field and thus break the chain of events which led to the failures. This modification appears to have solved the problem and, as the hours build up on the new tubes, we are more confident that this fix is correct. In each and every case the redundancy built into the transmitters was brought into play, and we lost no air time to tube failures, which was one of the original goals of the project and its technology.

Another question was how often to scrub the grid of the tubes. Comark recommended that we scrub every 10 days. In one case we let a tube at Wrens go far beyond that, and the scrub process took 20 minutes. With the installation at Chatsworth, we decided to scrub every day at sign on and learned that for the first 1100 hours the process took about 5 minutes. After the 1100 hour mark, the scrub cycle dropped to 70 seconds and it has remained there to date. This leads us to believe that the infancy period of the device is critical. So to maintain the same procedures at all sites we have modified our scrubbing cycle to daily, and the stability of the Klystrode at all sites is now proven. By removing the barium deposits on the grid and the ion pump keeping the vacuum clean, crowbars are kept to a minimum. Should a crowbar occur, a recycle process is quick and easy, and down time is very short.

V. AUDIENCE INCREASES

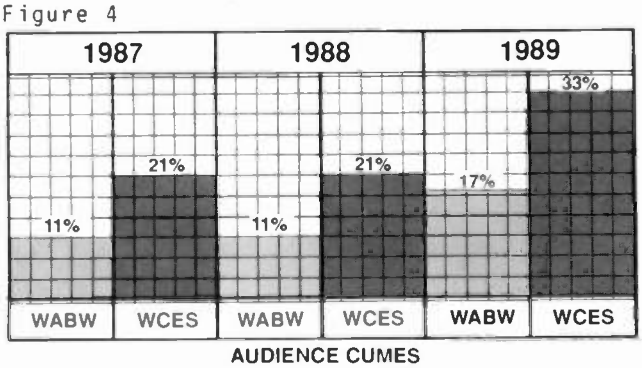
While engineering was studying the data from the transmitters the programming department was collecting the numbers on the coverage of the two new installations. We had hoped for an increase in the audience figures with the increase from 331 kW to 5 MW ERP at each installation. Figure 3 represents the average audience

Figure 3



increase in households for 1989, over the two previous non-Klystrode years. Figure 4 indicates the cume increase for the same time frame. So, along with the energy savings, we accomplished the goal of increased viewers.

Figure 4

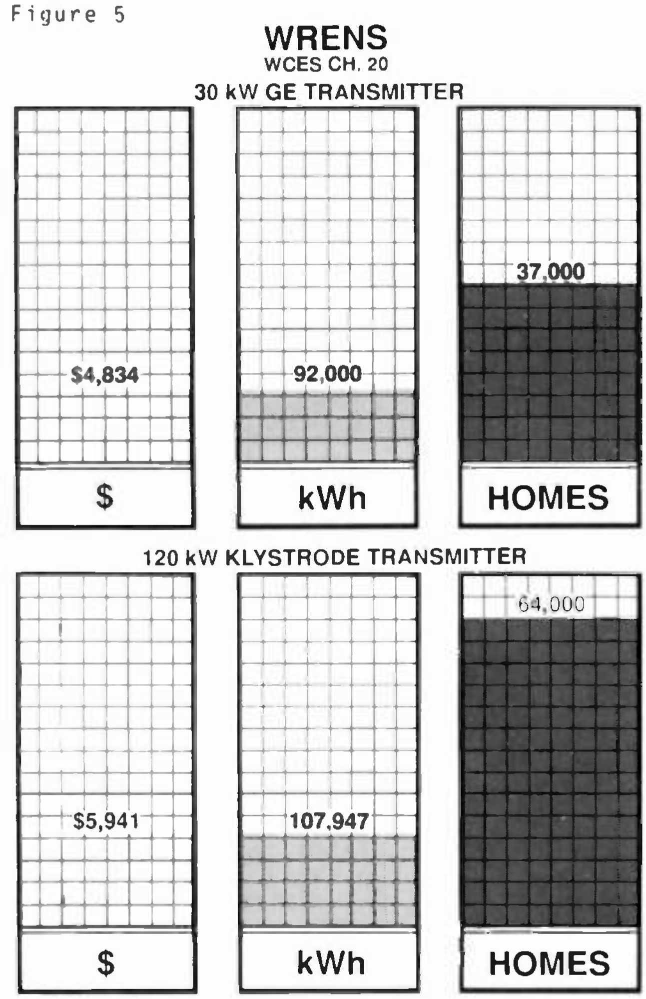


Figures 5 and 6 show the comparison between the 30 kW and 120 kW installations at Wrens and Pelham, for energy cost, energy usage, and homes. In both cases the energy cost average increased by approximately 24% for the transmitting plant and the ERP increased 13 times. The additional homes to each viewing area accomplished the programming goal.

VI. THIRD INSTALLATION

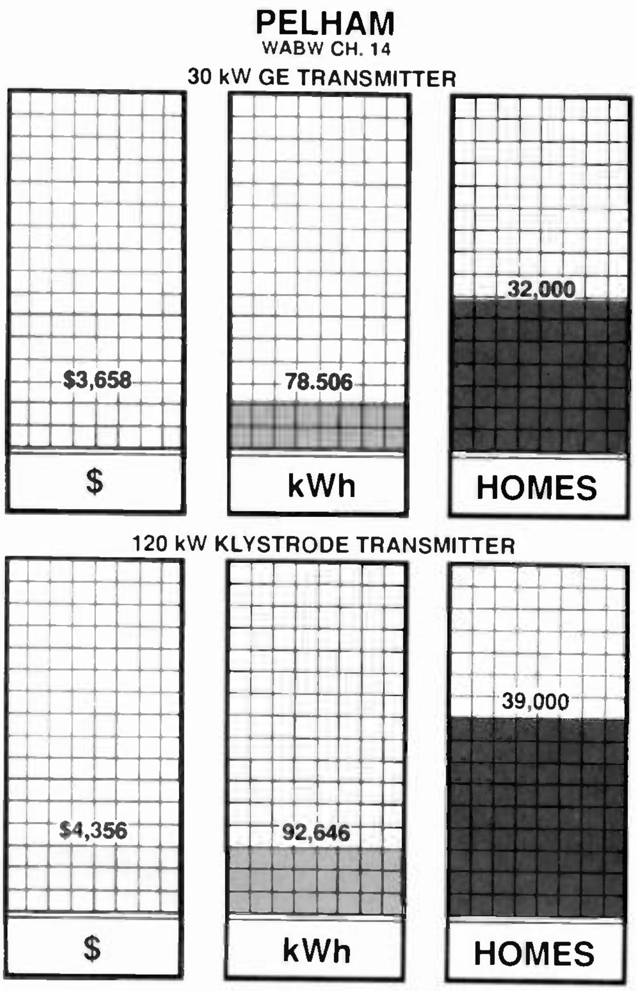
The third installation took place in September of 1989 at WCLP-TV, Chatsworth, Ga. Taking what we learned from the first two installations, this one went as smoothly as one could expect. The station staff had some extra time to learn the signal flow and was more involved with the initial tuning of the Klystrodes. We decided to scrub the grids each day, as mentioned earlier, and the stability of the infancy period dramatically increased. We also noted that the 3 fan heat exchangers would only run with one fan during the broadcast day indicating that the efficiency of the Klystrode was matching the first two installations. The cooling setup uses two

Figure 5



Liebert outdoor heat exchangers. One is used for aural and vision 1, the other for vision 2 and the system load. There is a three pump system setup, so a spare is able to be valved on line in the event of a pump failure. At any time the total transmitter heat load could be dissipated by either heat exchangers.

Figure 6

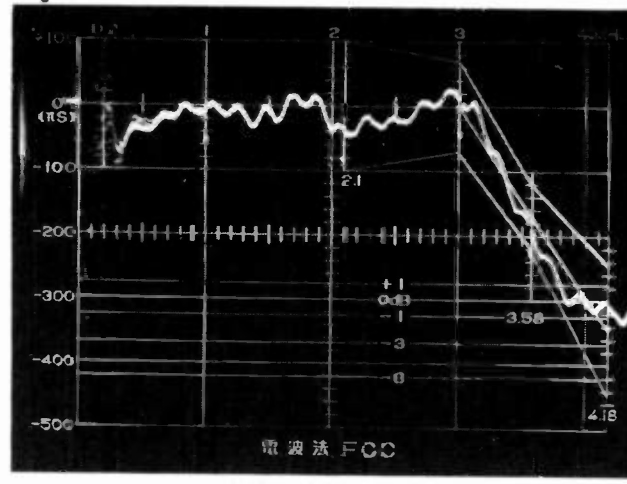


Figures 7-10 are actual proof-of-performance pictures for the Chatsworth transmitter. As with the first two transmitters the parameters were all within the expected range.

VII. FUTURE PLANS

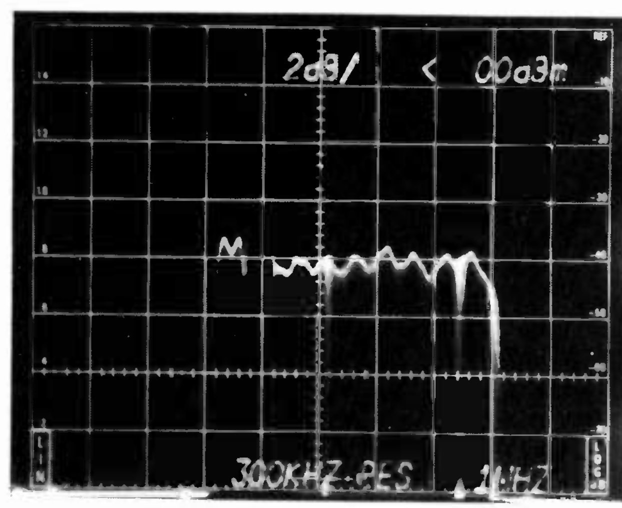
A fourth installation is scheduled for May 1990 at WDCO-TV, Cochran, Ga. This will be the same configuration as the first three, with an air date of June 1, 1990. Beyond that there are two more UHF transmitter plants to be considered for GPTV, and if funding goes well we will have completely replaced our 6 transmitters with the new, more efficient Klystrode installations. Our estimate is that if all 6 UHF systems are on line using the same formula of ef-

Figure 7



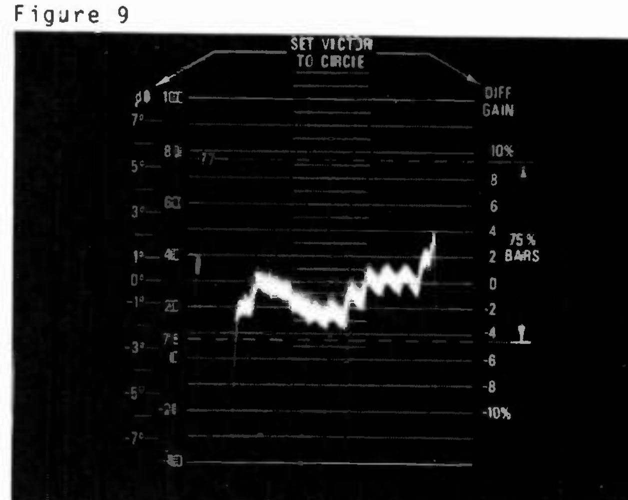
ENVELOPE DELAY

Figure 8



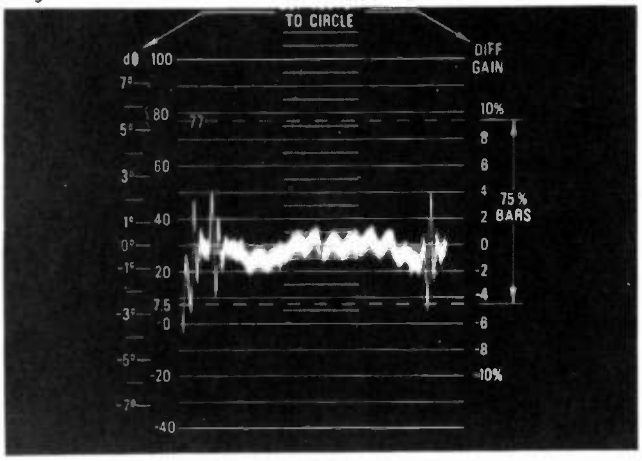
2db/DIVISION BAND-PASS RESPONSE

Figure 9



DIFFERENTIAL GAIN

Figure 10



DIFFERENTIAL PHASE

ficiency as in the past, our state network will have a far more powerful signal and realize a savings in energy expenses of $300,000 per year over conventional Klystron equipment.

VIII. CONCLUSION

In conclusion, the goals of a more powerful signal, greater coverage, energy savings, and new transmitting plants for GPTV were accomplished with the Comark Klystrode equipped transmitter. Each broadcast hour that goes by moves the technology forward, and the only question left to answer is that of the Klystrode's longevity. Only time will answer that question, but all indications are that the technology is here now, and is improving, which is all that any broadcaster can ask.

The author wishes to thank Nat Ostroff and Merrald Shrader for their assistance in providing information for this presentation and LaRoy Gee for the illustrations.

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RECENT ADVANCES IN KLYSTRODE® EQUIPPED TRANSMITTERS

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Comark Communications, Inc.
Colmar, Pennsylvania

ABSTRACT

This paper details the most recent advances in High Power UHF Klystrode Transmitter technology at Comark. Information on the first 240kW installation is provided. The use of dual phase locked modulators with independent linearity correction is also detailed.

Air-cooled multiplexed transmitters to 60kW are described and a detailed analysis of intermodulation distortion and its correction is also provided.

I. INTRODUCTION

In 1985 Comark began its development of a klystrode based transmitter line. At that time we had high hopes for this new device. We saw a highly efficient UHF amplifier with possibilities for great energy savings and unique equipment configurations.

Today our expectations for the new technology are more than fulfilled. Inductive output tubes (Klystrode) are now in service in Comark equipment at over a dozen broadcast installations. Power levels range from 10kW to 240kW in both separate and common amplification configurations, both liquid and air-cooled.

This paper will discuss the most recent developments in this technology including 240kW applications and 60kW air-cooled configurations. In addition, a discussion of multiplex operation and specifications will be covered.

II. 240KW KLYSTRODE APPLICATIONS

The true advantage of high efficiency is best realized at high output power levels. In an earlier paper[1], the concept of the "Super Power" transmitter driving a low gain antenna was developed. The klystrode's efficiency makes it an ideal device for this type of application.

At the time of this writing, early 1990, a 240kW transmitter using klystrode tubes is undergoing installation at WDRB-TV, Channel 41, in Louisville, Kentucky. Completion is scheduled for February 1990.

This installation is the first "Super Power" use of the klystrode device. Four 60kW water-cooled tubes form the visual amplifier and one 60kW klystrode tube provides the aural output.

Figure 1 is a block diagram of the WDRB-TV 240kW transmitter. Much of this system is similar to other 240kW klystron based transmitters manufactured and installed by Comark. The unique aspects of the installation include phase locked modulators, an output AGC system and the overall plant efficiency that is provided by the klystrode configuration.

The 240kW klystrode transmitter uses two independent, but phase locked modulators, one for each pair of 60kW tubes. This configuration provides significant advantages over an active-standby modulator configuration which requires switching in the event of a failure.

Since each modulator is independent and contains its own linearity correction circuits, each individual pair of output tubes can be optimized for linearity. This reduces the inevitable compromise that usually has to be made to match four parallel tubes with one linearity correction system. The independent nature of each modulator in conjunction with the switchless output combiner eliminates loss of air time in the event of a single modulator failure.

Another unique aspect of the 240kW transmitter is its output AGC system. Each pair of 60kW visual klystrode tubes has its own AGC loop back to its independent modulator. This fixes each tube pair at 120kW output power, thus

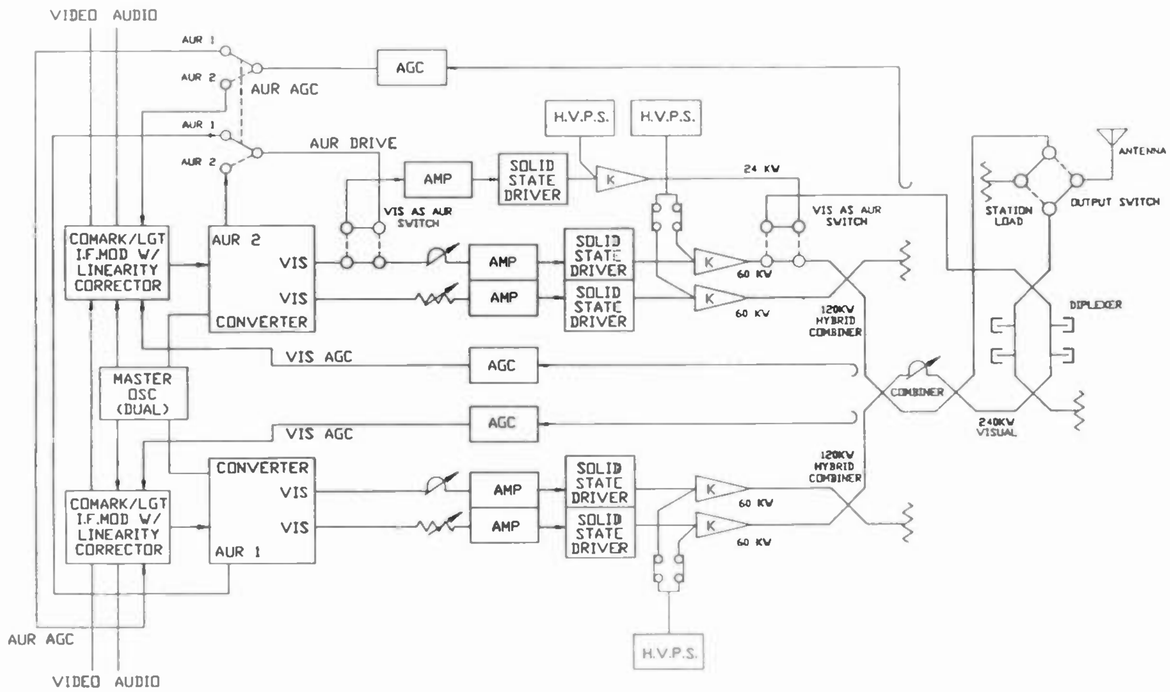

Figure 1 - Block Diagram, 240kW Klystrode equipped transmitter
as installed at WDRB-TV 41, Louisville, KY

stabilizing the overall transmitter output and preventing the RF equivalent of load hogging. The Class B nature of the klystrode means that beam current will increase with increasing drive power. Thus, the elimination of load hogging eliminates beam current unbalance and possible beam current overloads. The output AGC system uses the back porch as a reference level.

The phase locked nature of the dual modulators requires that a common upconverting local oscillator be used to achieve on channel output. Figure 2 shows the array of oscillators used in the system. Note that redundant oscillators are provided to insure continuous on air performance in the event of oscillator failure or frequency drift. An operator selection is also provided to permit the on channel aural carrier to be supplied by either independent modulator. The modulators are fed both video and audio in parallel.

At the time this paper was prepared, the installation at WDRB-TV was still underway. Therefore, final plant efficiency numbers are not available. However, based on several prior 120kW klystrode installations and preliminary tests at half power at WDRB-TV, the expected average plant power consumption for 240kW output should be 296kW. This will yield an overall plant efficiency at 50% APL of 85%. If this transmitter were implemented using the most advanced fully pulsed klystron tubes (5 cavity external), the plant power consumption could be expected to be 412kW*. If a pulsed 4 cavity klystron tube was used, the plant power consumption would be 452**. Thus the advantage of this klystrode system clearly comes down to lower operating costs.

It is interesting to note that the 5-cavity klystron while reducing power consumption by 40kW over its 4-cavity cousin, does not rival the klystrode. The 5-cavity klystron still requires more than 100kW over the power required by the klystrode to achieve the same output power.

* Assumed pulsed klystron efficiency of 82%

** Assumed pulsed klystron efficiency of 74%

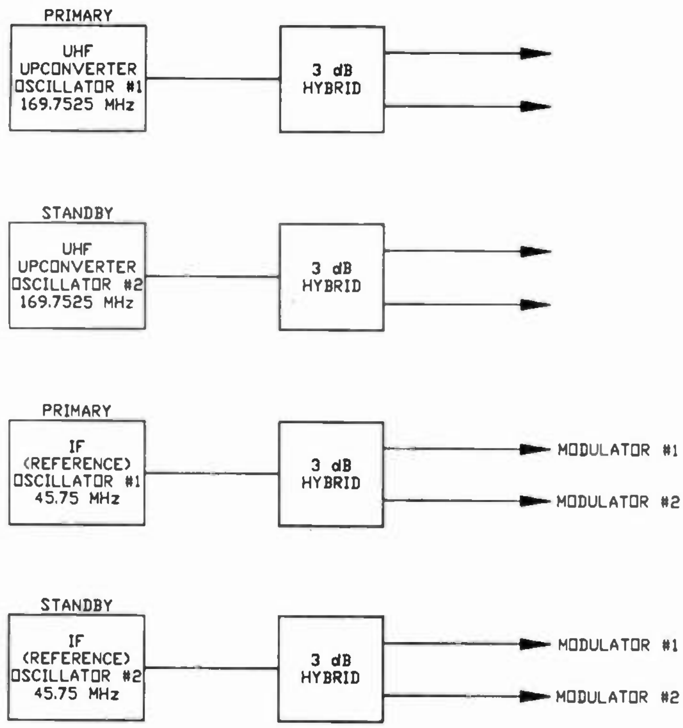

Figure 2 - Master Oscillator System

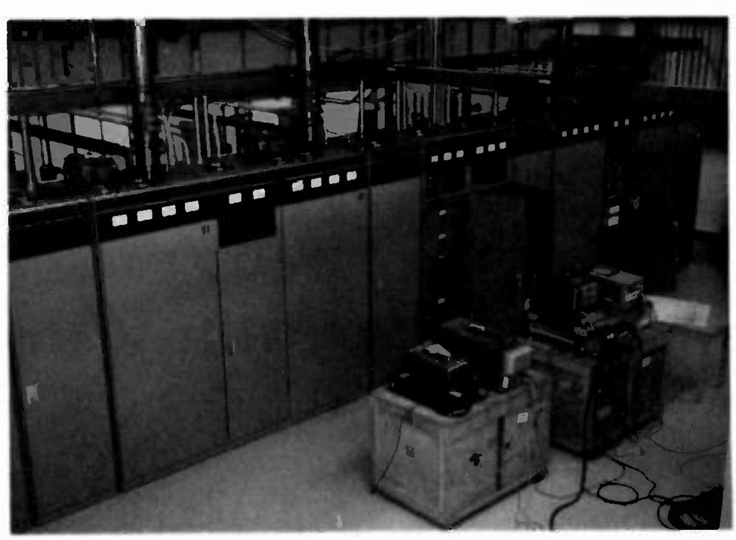

Figure 3 - Comark 240kW Klystrode Transmitter during installation at WDRB-TV, Channel 41, Louisville, KY
(4 visual amplifiers plus modulator/exciter shown)

III. AIR COOLING

Another new development in 1989 was the introduction of a 30/40kW multiplexed/ diplexed klystrode tube that uses air cooling instead of water.

Comark had earlier provided a 10kW air-cooled klystrode transmitter to WHTJ-TV in Charlottesville, VA[2,3]. Thus, the 30/40kW tube easily found its way into the product line. Air cooling is particularly attractive since the collector is at ground potential with no RF voltage.

The 30/40kW tube Eimac #2KDX40LA/LF uses the gun and body of the 60kW water-cooled device with a new air-cooled collector. Thus, much of the tube is already field proven. The ability to use an air-cooled collector arises from the fact that the klystrode tube is Class B in operation and is not pulsed to achieve its efficiency. Thus, no reserve is required in the collector design for non-pulsed operation in the event of pulser failure. The tube is rated at 40kW visual and 30kW multiplexed at a 10db aural-visual ratio.

While 40kW each or 80kW per pair, visual power, is possible, the most interesting transmitter configuration for this device is multiplex application at 30kW and 60kW. At the time of this writing, Comark is manufacturing two 60kW air-cooled transmitters each using a pair of 30kW air-cooled tubes, for shipment to an international customer in early Spring of 1990. Figure 4 is a block diagram of the 60kW transmitter and Figure 5 is a view of the air-cooled tube. Figure 6 is a picture of the air-cooled circuit assembly and Figure 7 lists some of the Eimac specifications for the 2KDX40LA/LF.

These 60kW air-cooled, multiplexed transmitters eliminate the output diplexer and most of the output RF system. The phase locked dual modulators and the Magic Tee combiner permit 50% power operation in the event of a tube or driver failure without loss of air time or hard contact RF switching. The simplicity of the transmitter and RF system should make these transmitters very appealing to the cost conscious UHF broadcaster.

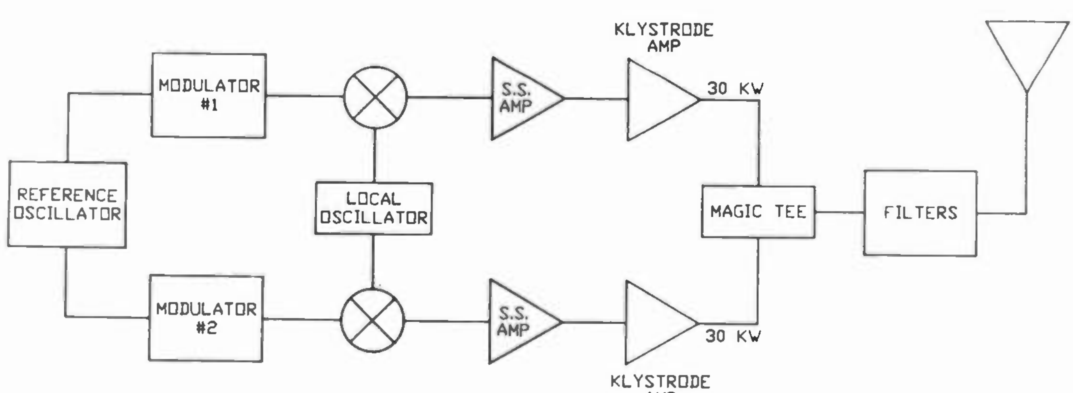

Figure 4 - 60kW Air-Cooled, Multiplexed Klystrode Equipped Transmitter

Figure 5 - 30/40kW Air-Cooled
Klystrode Tube

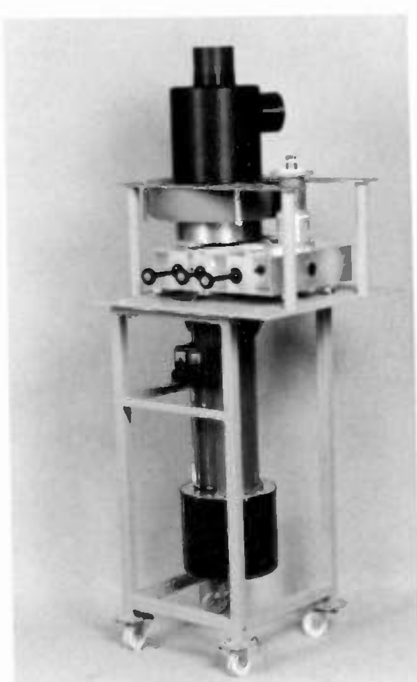

Figure 6 - 30/40kW Air-Cooled Klystrode
Circuit Assembly

GENERAL CHARACTERISTICS

ELECTRICAL

Cathode: Dispenser Type	
Heater Voltage	10 to 12 Vdc
Heater Current (nominal)	10 Adc
Maximum Cold Start Heater Current	20 Adc
Heater Warmup Time (minimum)	300 Sec
Magnet - Part of Circuit Assembly:	
Voltage (maximum)	7 Vdc
Current (maximum)	28 Adc

MECHANICAL

Cooling	Forced Air
Input rf Connector	Type N
Output rf Connector	4-1/16 In. EIA Standard

VISUAL SERVICE, ESTIMATED

Beam Voltage	27 kVdc
Beam Current (peak sync)	2.8 a
Beam Current (black level)	1.7 a
rf Power Output (peak sync)	40 kW
Conversion Efficiency (peak sync)	53 %
rf Drive Power (peak sync)	200 W
Grid Bias Voltage	50 ± 10 Vdc
Bandwidth	8 MHz

Figure of Merit = Peak Sync Power Out / Average Picture Power Input Including sync & blanking pulses = 120%

AURAL SERVICE - 2KDX40LF and 2KDX40LA

	Klystrode Performance		
FM CW rf Power Output	4.0	8.0	kW
Beam Voltage	27	27	kVd
Beam Current	0.5	0.85	Adc
rf Drive Power	20	40	W
rf Power Gain	23	23	dB

Note: Beam current is set by the rf drive level which is adjusted for the required power output. The beam voltage may be the same as that used on the associated visual klystron. Efficiency, power gain, power level and beam voltage are interrelated.

Figure 7 - 30/40kW Air-Cooled Klystrode Tube
Selected Specifications

IV. MULTIPLEX OPERATION
(Common Amplification)

As any broadcast engineer who has run a klystron in emergency multiplex knows, multiplex output signals can be full of intermodulation distortions. Even in a more linear device like the klystrode, or a tetrode, uncorrected intermodulation between the visual, aural and color carriers can create unacceptable levels of distortion. For a multiplex transmitter to be acceptable for broadcast use the intermodulation must be reduced to well below perceptible levels.

Figure 8 details the typical intermodulation signals that are created when a standard NTSC combined vision, sound and color RF signal is amplified in a power amplifier.

Most of these intermodulation products can be eliminated by filtering outside of the bandpass of the channel. A multi-section waveguide trap filter, having low inband insertion loss and high notch rejection usually reduces the out of band signals to well below 60db below the peak visual carrier (FCC spec).

Thus, only two signals need to be addressed that cannot be filtered by brute force. These are the $f_v \pm 0.92$MHZ. The f_v-0.92MHZ is too close to the lower information band edge for practical filtering and f_v+0.92MHZ is in the active information passband. A non-filtering technique for reducing these intermodulation products is well established. The technique makes use of low level linearization circuits that introduce amplitude distortions to the overall system transfer function. These "linearizing distortions" introduce sidebands in the frequency domain that are equal in amplitude but opposite in phase to the distortion created in the power amplifying system. When done properly, an intermodulation product cancellation of 15 to 20db can be achieved with long term stability. The wideband nature of the klystrode amplifier along with the wideband solid state drive system, preserves phase linearity through the system and thus permits straight forward intermodulation cancellation. Narrow band systems are not readily capable of 20db of intermodulation product cancellation. The use of an overall output AGC system also

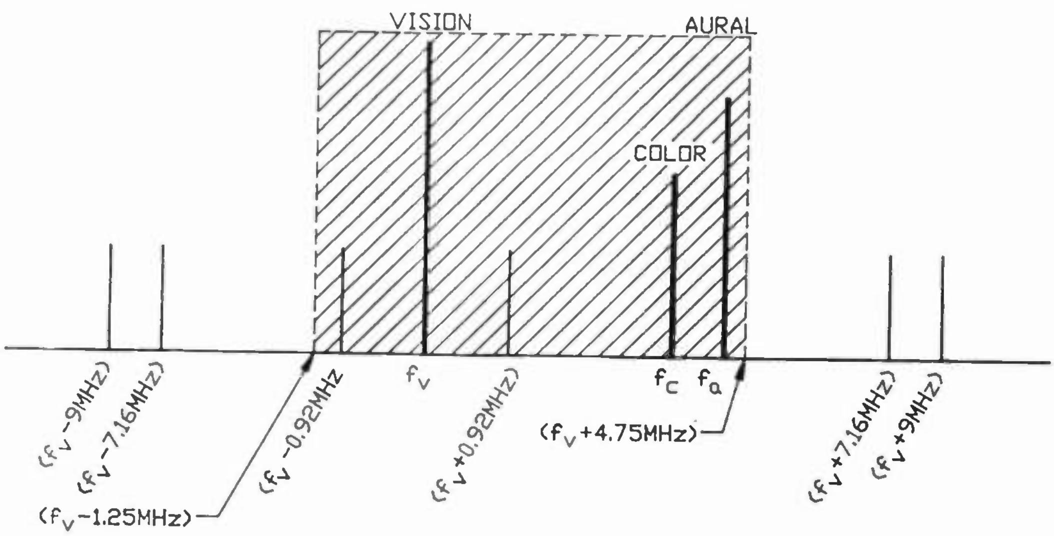

Figure 8 - Typical Intermodulation Products

stabilizes the output levels and helps to insure correction stability. Of course, the "S" shaped transfer characters of a Class B amplifier like the klystrode takes special correction circuits not found in Class A applications like tetrodes or klystrons. Comark has spent a great deal of R&D effort establishing what the special linearization needs of klystrodes are with respect to multiplex operation. The Comark/Thomson-LGT modulator is the product of this research and includes independent correction circuits for ICPM, Differential Gain, Differential Phase, Low Frequency Linearity (5 steps of adjustment) and bandpass amplitude response. Adjustments for all of these circuits are not interactive with each other and are available from the front panel of a plug in module in the modulator.

Figure 9 is a view of the Comark/Thomson-LGT modulator and linearity corrector in its exciter cabinet. This unit is used in both diplexed and multiplex systems with minor modifications.

Figure 9 - Comark/Thomson-LGT Modulator linearity corrector in its exciter cabinet

An analysis of the T.V. signal based on signal to noise considerations, indicates that intermodulation products that are better than 52db below the peak visual carrier will not be perceptible.
The 52dB level has been a standard for the translator industry for many years. However, for full service broadcast use, a level three to four times better should be the standard. Thus, Comark multiplex transmitters will provide 56 to 60db intermodulation performance at an aural visual ratio of 10db.

One final look at the IM problem. The intermodulation product that most effects picture quality is f_v+0.92MHZ. The f_v-0.92MHZ can usually be ignored. The reason for this is illustrated in Figure 10 and Figure 11.

The pre-detection bandpass of a standard T.V. receiver (Figure 10) incorporates a lower sideband rejection response called the Nyquist slope. This is required to suppress the effect of the 0.75MHZ of double sideband spectrum contained in the transmitter vestigial lower sideband RF signal.

The impact of the Nyquist slope to the received T.V. signal is illustrated in Figure 11. The f_v+0.92MHZ falls in the upper sideband portion of the curve and its level is enhanced with respect to the visual carrier by 6db. The f_v-0.92MHZ falls outside of the receiver bandpass and is suppressed by the Nyquiest slope. Thus it is reduced by more than 6db with respect to the visual carrier level. The relative difference between the upper and lower intermodulation products after passing through the receiver IF is thus greater than 12db.

The intermodulation product that must be suppressed in the transmitter RF output signal is therefore f_v+0.92MHZ.
Thus, the picture quality of a klystrode transmitter operating in multiplex at a 10db aural visual ratio can be maintained at levels well above acceptable broadcast quality. The received results from Comark multiplex klystrode transmitters observed from at least three sites has confirmed that both picture and stereo sound quality is indistinguishable from a fully diplexed transmitter.

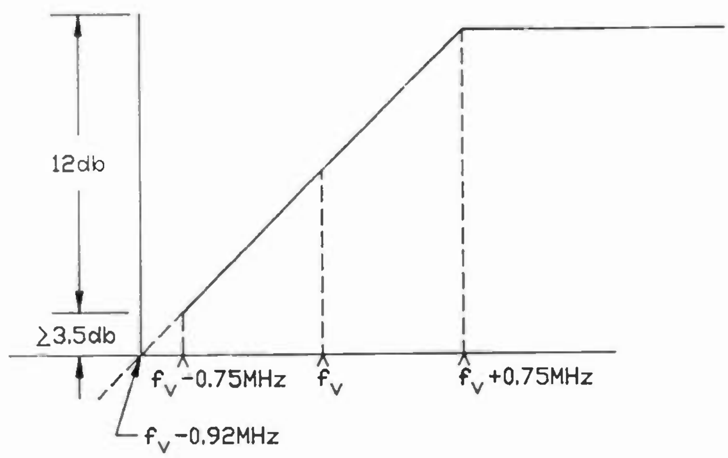

Figure 10 - TV Receiver Predetection Bandpass Response (Nyquist Slope)

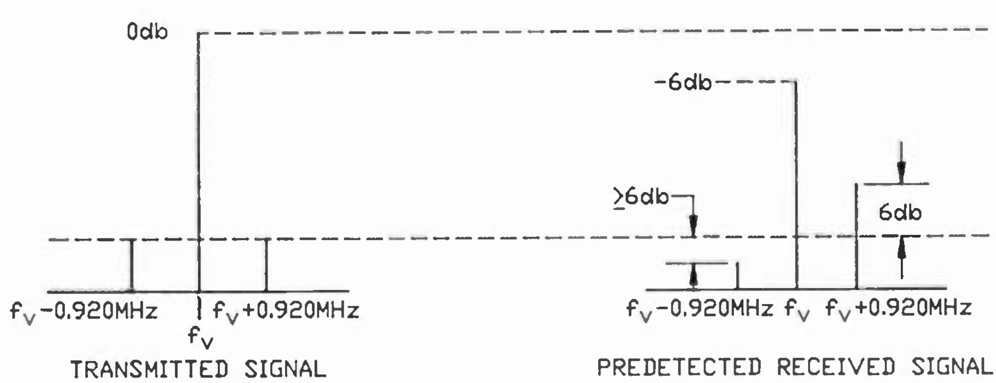

Figure 11 - Nyquist Slope effect on received signal prior to detection

V CONCLUSIONS

During 1989 the klystrode tube has proven its versatility as a high efficiency UHF amplifier. A number of broadcast stations purchased and placed Comark transmitters using klystrodes into service. The variety of equipment configurations now in service includes power levels from 10kW to 240kW water cooled, and air cooled to 60kW.

Clearly, the promise of the klystrode device first made in 1985 has been fulfilled in 1989. Comark is proud to have played a part in bringing this technology to the industry and establishing its viability. The future for this technology is clearly a bright one.

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GOING BEYOND TECHNOLOGICAL FADS: OBJECTIVE CRITERIA FOR SELECTING HIGH-POWER UHF TV TRANSMITTERS

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Abstract:

THOMSON-LGT has the highest number of equipments in operation with transistor devices. More recently we have been involved in a very thorough analysis of the thermal as well as the electrical environment of a very rugged transistor device developed by THOMSON-CSF in US. The results of this study even though theoretical will show how the behaviour of components in the field can be predicted ahead of time, thus preventing expensive and time consuming field problems.

A heavy emphasis on the maintenance design side of this range of new transmitters will be discussed along with design for low operating cost.

INTRODUCTION

Transmitter technology is changing slowly but surely with the introduction of power transistors driving out the tube technologies for increasingly powerful transmitters. The present paper outlines our views on this matter and tries to answer three questions: first, what are the benefits of transistorization in terms of operating dependability, second, up to what UHF power level is transistor technology economically viable, and, third, does such a reliable transmitter exist?

1. Operating dependability

THOMSON-LGT is well aware of the importance of On-Air Availability of its transmitters. Depending on the country and the population density in the broadcasting region, a transmission break may cost the operator between $9,000 and $20,000 per hour. Losses of over $65,000 per minute can even be incurred for transmitters in especially strategic locations from the audience viewpoint. It is harder to quantify all the spin-off effects of a break in transmission with respect to both the operator's image and the image of the transmitter manufacturer.

This factor can greatly affect the cost of ownership of a transmitter. The sectorial graphs enclosed (Figure 1) show that based on a cost of $10,000 for a one-hour break in transmission, an average time of 4 hours for site intervention and standard replacement of the faulty module, the costs involved in a break in transmission can amount to 50% of the cost of ownership over 15 years for a 10 kW unbacked-up tube transmitter, while the cost is less than 0.03% when the transmitter is backed up in passive reserve by a transmitter of the same type. Finally, it should be noted that in this case the cost of ownership of the unbacked-up transmitter is nearly 60% greater than the cost of ownership of the system with passive reserve.

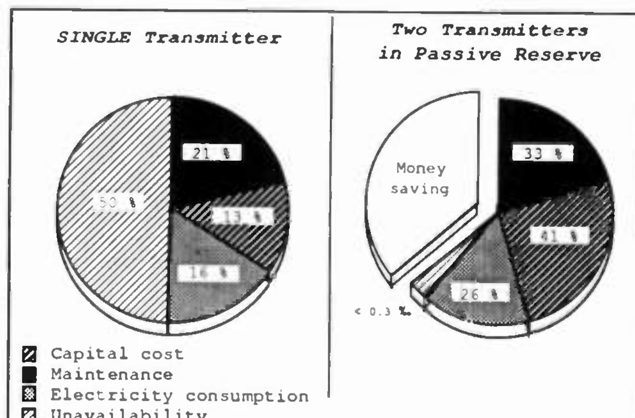

Figure 1
EXAMPLE OF COST OF OWNERSHIP FOR TUBE TRANSMITTER
Influence of transmitter availability

To meet the operating cost requirements of the operators, THOMSON-LGT has developed analysis and computing methods to solve such questions of Operating Dependability. On-air

Availability can be quantified from the operator's viewpoint in terms of the percentage of time during which the programme is actually transmitted or could be transmitted. To improve the On-Air **Availability** of a transmission system, one can improve its **Reliability** characteristics (architecture, thermal qualities, etc.) or its **Maintainability** characteristics (testability, modularity, etc.), as shown in Figure 2.[1].

For a non-redundant system, the mathematical relationship between Reliability, Maintainability and Availability is as follows:

Unavailability = MTTR / (MTTR + MTBF)

where MTTR = Mean Time To Repair
and MTBF = Mean Time Between Failures.

To the denominator in the above formula can be added a term related to preventive maintenance time. This term no longer applies when preventive maintenance inspection requires no break in transmission.

For complex systems, the formula soon becomes unworkable; THOMSON-LGT has therefore adopted an approach by Monte-Carlo type simulations. The transmitter system and its maintenance are formalized in the form of a Petri Net [2]. As shown by the simplified example in Figure 3, maintenance staff and equipment are represented by their status (operating, fault, repair, standby, etc.). Transitions between statuses are enabled when the upstream statuses and the upstream messages are valid, using the statistical law which models the phenomenon: hence, the transition from operating status to failure status takes place in accordance with exponential law exp(-t/MTBF).

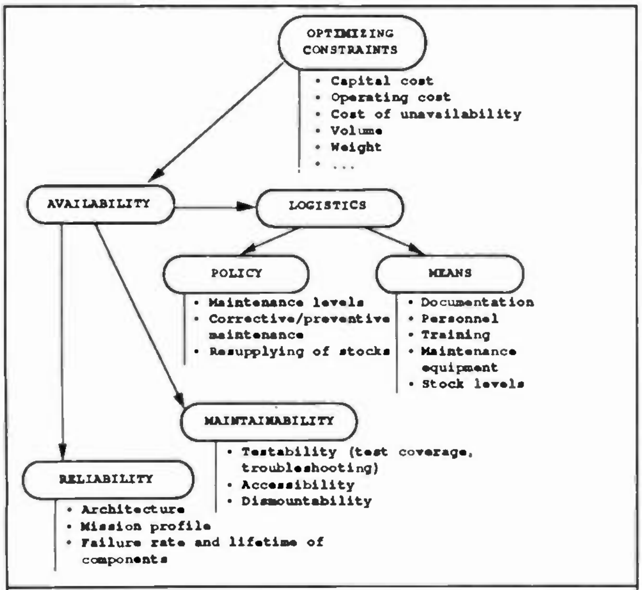

Figure 2
R.A.M. APPROACH

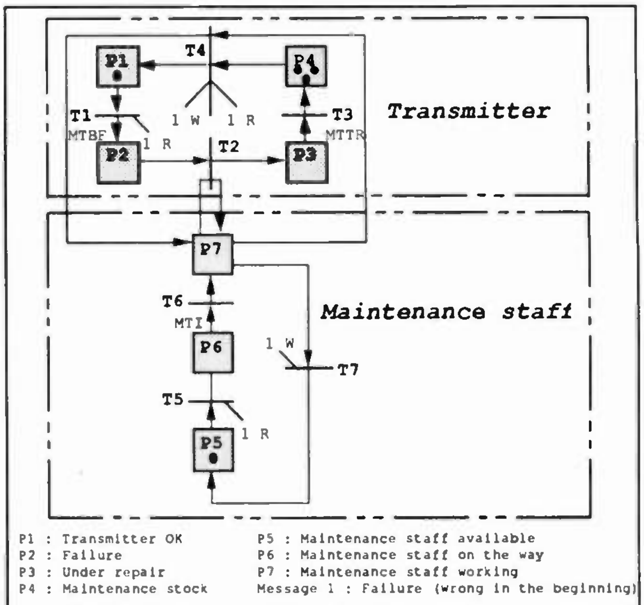

Figure 3
USE OF PETRI NETS IN OPERATING DEPENDABILITY

Such simulations allow us to assess On-Air Availability for a given level of reliability and a given maintenance policy. In this way, we can propose the most efficient redundant structure and provide our customers with information on the optimum maintenance policy required for our transmitters (in terms of required maintenance facilities, frequency of preventive maintenance inspections, etc.).

Given the importance of the on-air availability factor in the cost of ownership, it is especially important, when comparing the various available technologies, that the transmitters compared should have equivalent operating dependability. For example, given the redundance provided by the use of transistors, a solid state transmitter shall be compared with two tube transmitters in passive reserve configuration.

One may well ask whether it is not short-sighted to take into account merely the economic aspect of purchase cost of a TV transmitter when selecting a power component technology. One shall therefore assess, for various output power levels, the operating cost on the basis of the most recent knowledge.

The customers have to be informed about the optimum economic solution best suited to his operating constraints, and making extensive allowance for the concept of Operating Dependability.

2. Solid state solution within the Multi-technology Approach

2.1 TH-LGT Expertise

The term "multitechnology", as applied to terrestrial TV transmitters, refers to the different technologies used in the final amplification stage to obtain the required output power.

Partnership relations between equipment manufacturers and their customers have to be improved. This change is chiefly due to the major progress made in transistors and power tubes, achieved as part of military research projects or space and radio communications projects.

Faced with this profusion of new technologies, companies having the necessary experience to provide their customers with clear information concerning operating costs could be well appreciated.

Considering the huge power range we are capable of handling (from a few Watts to 240 kW in UHF, using the amplification techniques most suited to the desired power level), we can adopt an objective and discerning approach.

To shed light on this question, we have examined on the one hand well established technologies such as the klystron (10 to 70 kW) and tetrode (1 to 20 kW), and on the other hand emerging technologies such as the power transistor (100-300 W), the klystrode and the MSDC klystron.

With COMARK, we have become involved in the Klystrode ® project since 1987. The two companies work together closely: research and development work on modulation, correction and the transistorized amplifier stages has been carried out in France, while the klystron and klystrode stages are mainly developped in the United States [3].

Moreover, the experience acquired by THOMSON-LGT in the design of tetrode amplifiers has consolidated the Group's expertise in this field (see Figure 4).

There is no single answer to a customer's requirements. It is best first to consider the customer's requirements and then look for the appropriate technology, and not vice versa. Technology is a means, not an end in itself.

It is therefore interesting to compare operating costs for different power levels.

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2.2 Operating Costs

As an example, we have taken into account the following three factors [4] [5]:

◊ Estimated annual cost of tubes or silicon components, calculated on the basis of their lifetime or MTBF in their operating conditions in our equipment;

◊ Power consumption, estimated for an average picture, for 18 hours of transmission per day, with an electricity price of 7.3 cents per kW/h;

◊ Cost of maintenance, calculated on the basis of a fixed number of interventions per year for tube transmitters and two interventions per year for solid-state transmitters (assuming that two technicians or engineers are required for each intervention).

In the case of Solid State transmitters, our design based on theoretical reliability projections using the MTBF figures quoted by the transistor manufacturers assume that for such transmitters less than one intervention will be needed per year. Nevertheless it seems wise, given the fact that we have only limited previous experience of site operation of such power components, to adopt the pessimistic figure of two interventions per year for our simulations. (A reliability specialist would say that we are at the start of the reliability growth curve.)

We have assumed that intervention only occurs following a power drop by more than 3 dB in vision power and more than 6 dB in sound power.

For transmitters equipped with a single power component in active or passive reserve operation, intervention occurs as soon as a fault appears. For transmitters equipped with multi-power amplifiers working in parallel, maintenance teams are sent out once the number of non-working amplifiers reaches the -3 dB zone, which gives free scope for action.

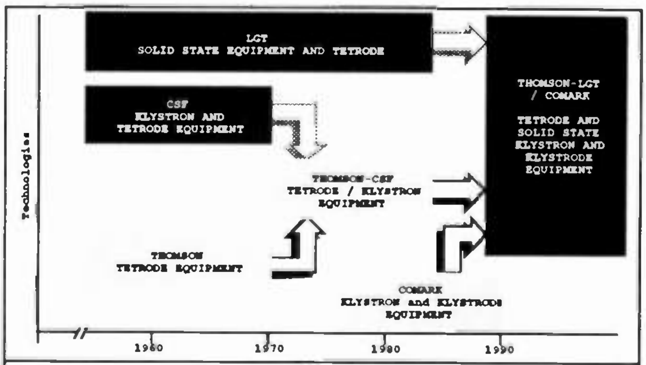


Figure 4
THOMSON-LGT EXPERTISE

For solid-state transmitters, the spare parts holdings can be envisaged as one complete module per transmitter and, at a regional base, a complete set of transistors corresponding to one stage.

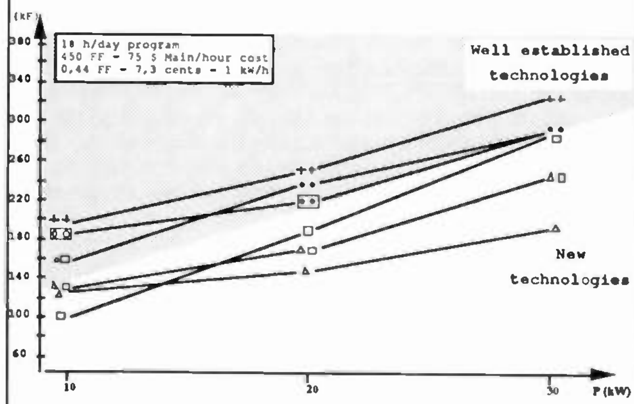
The enclosed curves (figure 5) show the operating cost of one transmitter with an output power of 10, 20 and 30 kW for well-established technologies such as pulsed klystron and tetrode and for the emerging technologies such as klystrode and solid-state.

It can be seen that the emerging technologies have far lower operating costs, which must make them attractive to operators.

Pulsed klystron and tetrode components have parallel curves, with an advantage for the tetrode (this comparison includes the new THOMSON-TTE TH 563 tube).

For the emerging technologies, when a fully solid-state transmitter is compared with a hybrid transmitter designed with a klystrode tube as vision carrier amplifier and solid-state amplifiers for the sound carrier(s), it can be seen that their operating costs are identical at an RF power of 15 kW (video), but that costs soon become economical for mixed technology transmitter at higher power levels, for an average electricity costs of 7,3 cents / kW/h.

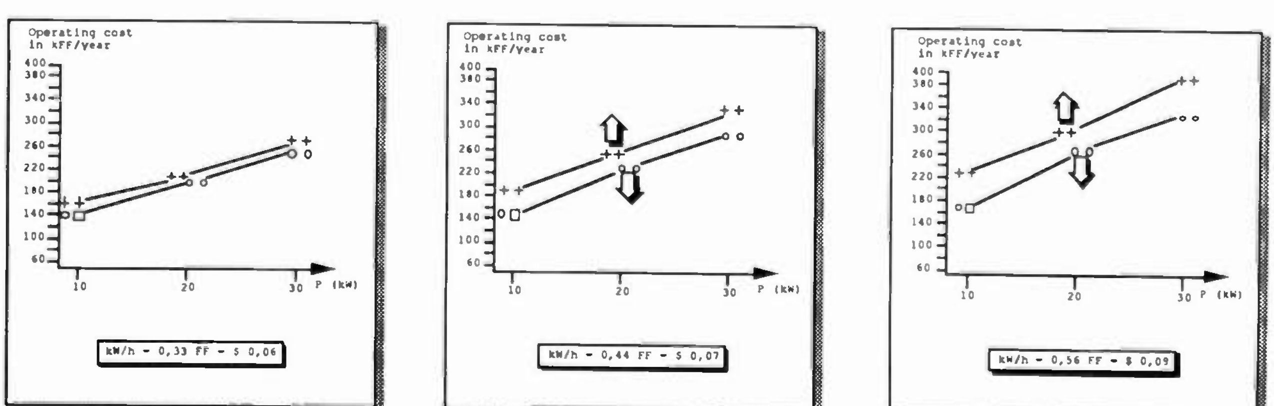
By means of the simulation shown in Figure 6, it is easy to monitor the influence of energy costs on operating costs for the well-established technologies.



Combined amplification	Separate amplification
△ Multiplex klystrode (1 tube)	△□ Klystrode for vision carrier SS PA for sound carrier(s)
⊡⊡ Multiplex tetrode (2 tubes)	++ Klystron (2 tubes)
o□ Tetrode for vision carrier SS PA for sound carrier(s)	□ All solid state
	oo Tetrode (2 tubes)

Figure 5
OPERATING COST
WELL ESTABLISHED AND EMERGING TECHNOLOGIES

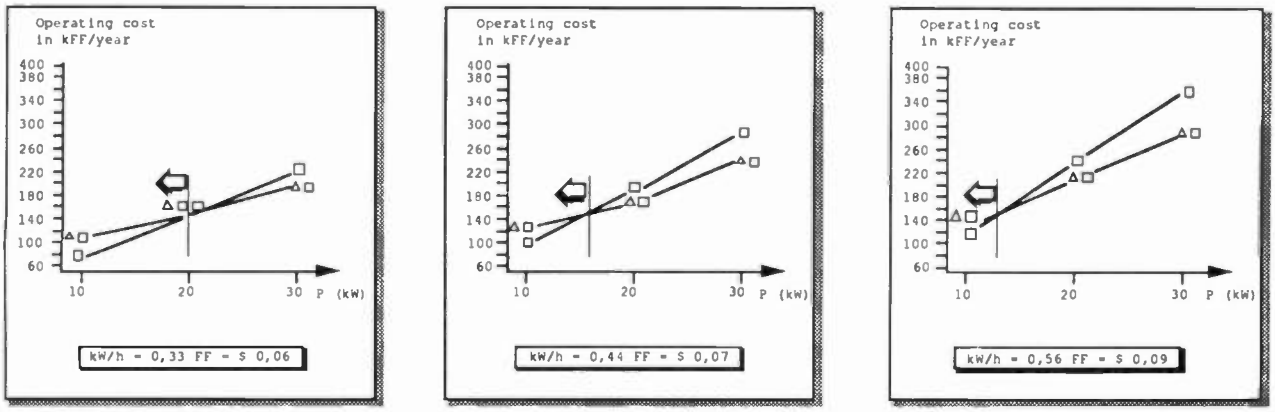
It can be seen that for low energy costs, the operating cost of a tetrode transmitter is not very different from that of a klystron transmitter for RF power levels of 20 kW. As energy costs rise to 9 cents per kW/h, the tetrode technology undoubtedly has the advantage.



++ 2 klystron tubes (pulsed)
o□ 1 tetrode tube for vision carrier
1 SS PA for sound carrier(s)
oo 2 tetrode tubes

1 $ = 6 FF

Figure 6
OPERATING COST FOR WELL ESTABLISHED TECHNOLOGIES. PARAMETER = COST OF ENERGY



△□ Klystrode for vision carrier solid state for sound carrier(s)

□ All solid state

1 $ = 6 FF

Figure 7

OPERATING COST OF EMERGING TECHNOLOGIES. PARAMETER = COST OF ENERGY

If we now analyze the influence of energy costs on the operating cost of emerging technologies (Figure 7), we observe that in those countries where energy costs are high, the klystrode solution is better at power levels above 12 kW, while the fully solid-state transmitter is acceptable when energy costs are lower.

To remain consistent with this data, we worked on the basis of equal redundancy between the various solutions. We rejected any fully solid-state solution for which the transistors would not operate in good conditions of reliability.

3. Solid-state transmitter

For more than ten years, our company has mastered the problems of transistor amplification of the video signal. For all frequency bands, THOMSON-LGT has, since 1972, installed over 1,200 fully solid-state transmitters, or a total of over 9,600 kW years (figure calculated by multiplying the output powers of solid-state transmitters by the corresponding years of operation). This figure is especially significant knowing that in the past only low-power transmitters (less than 1 kW) were manufactured using solid-state amplifiers.

3.1 A new approach: modularity

In designing our new range of high-power UHF transmitters, then, we were working on the basis of our actual experience in the field of solid-state technology. The idea behind all this research is as follows. The starting point must be an impeccable mechanical structure, both simple and modular, so that the amplifier modules can be interchanged for various power components. By this approach, one can guarantee customers long transmitter life and real progress in their operating conditions.

The primary advantage of multiple modular transmitter power stages operating in parallel is on-air availability. But that's not all. This new concept will eventually bring about a revolution in maintenance.

It can be assumed that power transistors will evolve, and new developments can be expected each year. The graph shown in Figure 8 illustrates the speed with which transistor manufacturers are progressing, taking into account the figure of merit. One observes that the design cycle for a transistor is far shorter than that of a tube, research on which has now reached a certain level of maturity (8 different RF transistors in 10 years and only two different klystrons).

This almost annual development in power transistors is disproportionate with the lifetime of a transmitter, which is approximately 15 to 20 years. It is no longer vital to select (or wait for) the latest component, as was the case for tubes, provided that the transmitter manufacturer ensures that the components are interchangeable. Besides, it is unrealistic to try and have our customers keep up with the latest technological fads. If technological progress were to lead to a significant drop in power consumption, users could take advantage of their equipment's modularity to replace old amplifiers (either as the modules fail or systematically), in order to reduce operating costs.

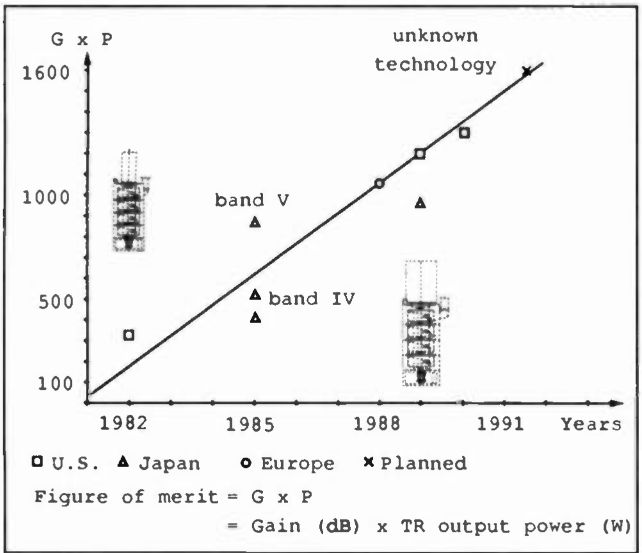


Figure 8
DESIGN OF RF TRANSISTORS

This new outlook is quite different from that which involved making a choice, once and for all, between the klystron or the tetrode, which often led to bitter struggles between rival technologies.

We provide an original solution based on a modular, upgradable amplifier with constant gain and power. Equipment maintainability and long life is now our main concern.

3.2 Reliability

As explained in Paragraph I., the first way of improving the on-air availability of an equipment is to increase its reliability. The parameters on which depends the reliability of a fully solid-state transmitter are mainly the following:

- Transmitter architecture.
- Junction temperature of the semiconductors (see figure 9).

This is why THOMSON-LGT has developed a basic UHF cabinet with eight 800 W amplifiers in parallel, enabling power levels of up to 20 kW to be reached. These amplifiers use a new heat sink (patented design) having a case-ambient thermal resistance of less than 0.02°C/W, when used with its new ventilation system providing an uniform air pressure and flow rate both for amplifiers and power supplies. This system is called "Air Isobarique Repartiteur" System, or A.I.R. system.

a. Architecture

The advantage of a transmitter architecture using parallel transistors for improved reliability can easily be demonstrated [6]. If all amplifier modules are identical from the reliability viewpoint, then the reliability Rs(t) of a "k/n parallel" device (where n is the number of modules and k the minimum number of modules in working condition) can be expressed by the formula:

$$Rs(t) = \sum_{i=k}^{n} C_n^k \, (Re(t))^i \, (1 - Re(t))^{n-i}$$

Where all the components follow an exponential law of parameter λe, the expression becomes:

$$Rs(t) = \sum_{i=k}^{n} C_n^k \exp(-i\,\lambda_e\, t)\,(1-\exp(-\lambda_e\, t))^{n-i}$$

and MTBFs = MTBFc (1/n + ... + 1/k)

where MTBFs = system MTBF

MTBFc = component MTBF

Note: The rule for calculating power loss in dB with our fully isolated coupling system is expressed by the formula:

$$PL(dB) = -10 \log n^2/k^2$$

The following table shows, for three 5 kW systems with 10, 8 or 6 amplifiers in parallel, the power loss in dB when 1, 2 or 3 amplifiers have failed.

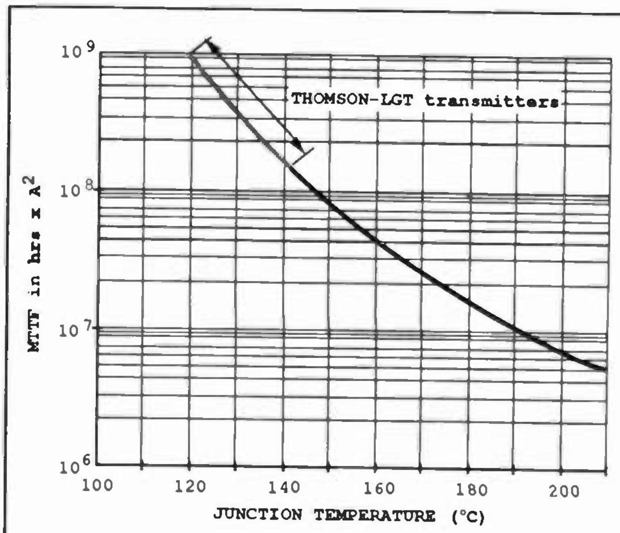


Figure 9
MTTF OF RF TRANSISTOR

Number of failed modules:	1	2	3
10-amplifiers system	0.9	1.9	3.1
8-amplifiers system	1.1	2.5	4.1
6-amplifiers system	1.5	3.5	6.0
	Power loss in dB		

It can be seen that with 6 amplifiers in parallel, the -3 dB fault criterion allows failure for only one amplifier. With 8 amplifiers this criterion is met with 2 faulty amplifiers, or with one faulty power supply (if a power supply feeds two amplifiers). Therefore the optimum cost-reliability compromise can be achieved.

Example: The MTBF of an amplifier is calculated as 100,000 hours for an ambient temperature of 35°C and a black picture. Given a -3 dB fault criterion, the failure of two amplifiers out of eight or one power supply is not regarded as a fault, so that we obtain as MTBF for the 6/8 system:

MTBFs = MTBFc (1/8 + 1/7 + 1/6) = MTBFc x 0.435 = 43452 h

For 5/6 system we obtain:
MTBFs = MTBFc (1/6 + 1/5) = MTBFc 0.367 = 36667 h

Two remarks could be made at this stage of the comparison:

◊ Firstly, two amplifiers with equal MTBF are compared, but this is not actually true since for the same output power the power dissipation and hence the junction temperature is greater in the six-amplifier system. It can be demonstrated that the MTBF of a given amplifier decreases from 100,000 to 70,000 hours for the same total output power when you use 6 amplifiers instead of 8.

◊ Secondly, the active redundancy is mainly interesting during the first years after maintenance. The MTBF is not always significant during this period. The right parameter is the probability of good operating (figure 10), which, in the case of parallel systems, does not follow an exponential law.
The curve shows that there is negligeable risk of failure over one year of operation of a 16 amplifier combination compared to 6 months for a 12 amplifier combination. Of course, after maintenance the reliability curve is reset to the original 100%.

b) Results of the thermal study of the heat sink

To cope with increasing transistor power, a new heat sink has been designed capable

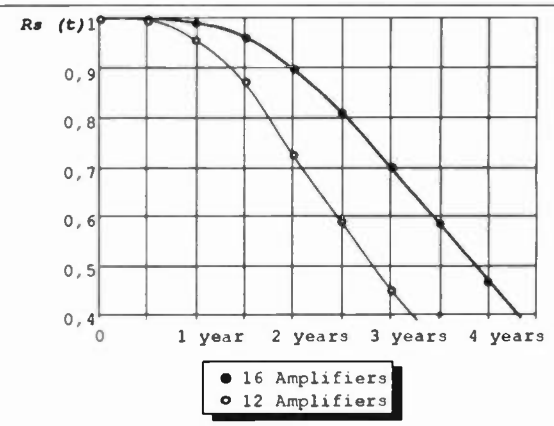


Figure 10
GOOD OPERATING PROBABILITY
Example: 10 kW UHF

of dissipating a total power of 1400 W with a maximum temperature build-up of 25°C at the component case.

Initial studies were performed using a thermal simulator (cf. Figure 11). This simulation shows the thermal unbalance between components located at the top of the heat sink and those at the bottom, if a conventional extruded-fin heat sink is used. In view of such results, research was made to manufacture a new design of heat sink composed of two parts to ensure:

1) uniform top and bottom temperatures,
2) a head loss of less than 200 Pa for a velocity of 6 m/s,
3) weight of less than 14 kg.

Finally we use, for power transistors, a heat sink having an exchange coefficient of approximately 10,000 W/m^2/°C for an air velocity of 6 m/s, which gives a fin outlet air temperature of 12°C for a flow rate of 350 m^3/h.

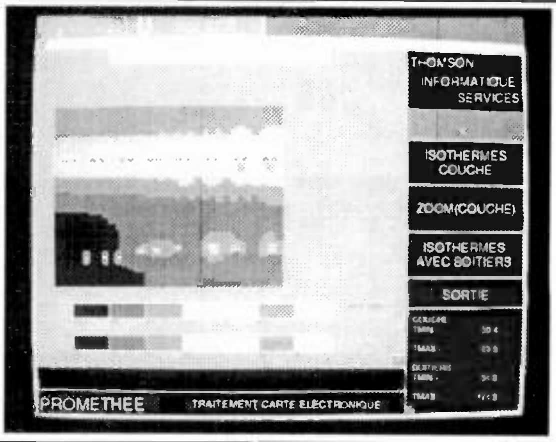


Figure 11
UHF 800 W AMPLIFIER THERMAL SIMULATION

Hence, for transistors dissipating 130 W over a contact surface of 4.57 cm^2, the heat build-up on the package flange is only 24°C.

The RF output power transistors have a thermal resistance of 0.5°C/W at 60°C, which gives a junction temperature of 130°C for a dissipated power of 130 W and an ambient temperature of 35°C. At such a temperature, the failure rate of the transistor is 128 x 10-9 hours, or almost 8 million hours (900 years).

More commonly, for a black picture, a 480 W output power and an ambient temperature of 35°C, an amplifier's MTBF is 100,000 hours.

The same approach was made for power supply and figure 12 shows the infrared thermograph of part of it. This picture shows that the difference between the ambient temperature (25°C) and the temperature at the base of the heat sink is only 22°C when a current of 200 A is supplied to the load under 28 V.

c) Results of the ventilation study

For an average picture corresponding to 30% luminance and a power supply efficiency of 86%, approximately 17 kW must be dissipated for each 6 kW unit cabinet (23% efficiency). Therefore a 12°C temperature build-up requires a flow rate of 4250 m^3/h per cabinet.

By using very reliable centrifugal fans capable of delivering 5000 m3/h at 400 Pa in a compression chamber, we were able to design a centrally controlled ventilation system, which has the following advantages:

◊ Air distribution among the various modules (amplifiers and power supplies), without "loopback" which is encountered when several ventilation systems are contained in the same cabinet;

◊ Minimum cabinet wiring;

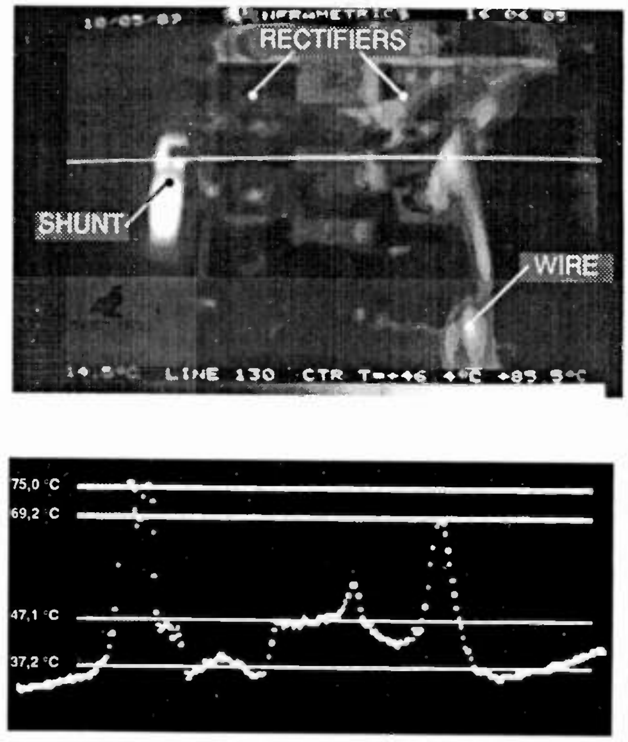
◊ Smaller equipment size;

◊ Minimal and uniform pressure of 250 Pa in the compression chamber;

◊ Low acoustic level (< 63 dBA for 10 kW) by using low pressure fans instead of high pressure fans;

◊ Both internal and external ventilation capability.

Using a fluid simulator (cf. Figure 13), we optimized the shape and size of the compression chamber to achieve the desired pressures for a given fan. The results of these simulations show the perfect uniformity at the amplifier level. At this same amplifier level, measured pressure is 300 Pa and the delivery per amplifier is 400 m3/h. In the power supplies, head losses are 250 Pa, while for such head losses the measured delivery is approximately 300 m3/h. It has been stated that measurements were within ± 10% of the results of simulations.



T ambient = 25 °C
I = 200 A
V = 28 V

Figure 12
INFRARED PICTURE OF POWER SUPPLY RECTIFIERS

3.3 Maintainability

3.3.1 Solid-state amplifier maintainability

We have explained above how very good reliability was achieved. Since the factor that concerns the operator is on air availability, it was essential to improve transmitter maintainability at the same time.

Maintenance of high-power UHF solid-state transmitters is greatly simplified by the fact that all modules (amplifiers and power supplies) can be removed when energized.

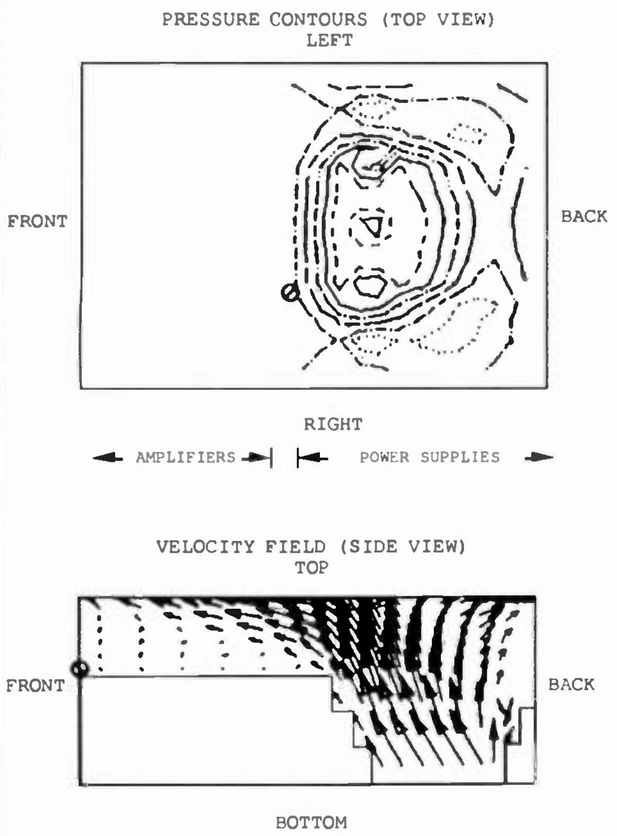


Figure 13
AIR ISOBARIQUE REPARTITEUR
THEBES SIMULATION

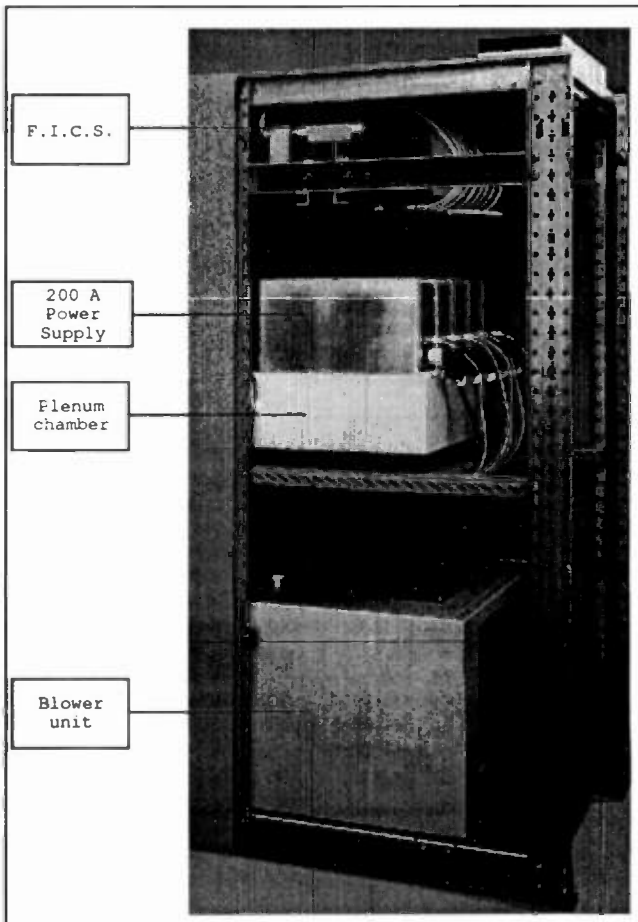


Figure 14
ACCESSIBILITY OF A 6 kW UNIT

This has been made possibly notably by the FICS* (Fully Isolated Coupling System).

Since the transmitters were designed to meet the -3 dB fault criterion for vision and -6 dB for sound, in the event of module failure or when a module is removed for maintenance purposes, downtime is reduced and the operator can perform maintenance while the transmitter continues to broadcast. Maintenance is performed by standard replacement of modules, and therefore requires less skilled technicians than in the past.

Emphasis has been placed on module accessibility and dismountability: the amplifiers can be accessed on the front panel and the power supplies on the rear panel (see figure 14). These modules can be removed without using any tools.

Troubleshooting and testability are made far easier by the firmware amplifier monitoring unit which can monitor for each amplifier the temperature, V.S.W.R., any overdrive, power supplies (+28 V and +5 V for bias), transistor currents and the power at various levels in the amplification system (total output and output of the various intermediate stages). All this information could be processed remotely, but, for a maintenance policy based on exchange of the faulty module, remote pinpointing of the type and quantity of faulty modules is sufficient.

For a 10 kW transmitter, the sound amplifier is of the same type as the vision amplifier. This means easier maintenance parts' management, since only one type of amplifier is needed for each station.

Further tests upon compatibility with MAC signals have been carried out with success, which means that amplifiers and coupling systems can handle digital datas if necessary [7].

3.3.2 Transmitter drive maintainability

Another sub-unit to which we paid careful attention in the design stage is the trans-

* Patented

mitter drive. The quality of this sub-unit's links to the power amplifier is of primary importance; the concept of linearization has been completely revised. Stable overall performance of a transmitter can be achieved through a good understanding of the problem of gain and phase linearity as a function of the modulation level, for the whole vision or sound spectrum.

The new concept proposes to consider IF linearity corrector as a part of the amplifier. To that respect, the IF modulator including the required channel filters delivers a perfect modulated signal to the RF linearized amplifier as shown in figure 15. This helps maintenability. Therefore the amplifier and its complementary amplifier (IF corrector) can be easily tested with a new method for setting up the quality of the amplifier path [8].

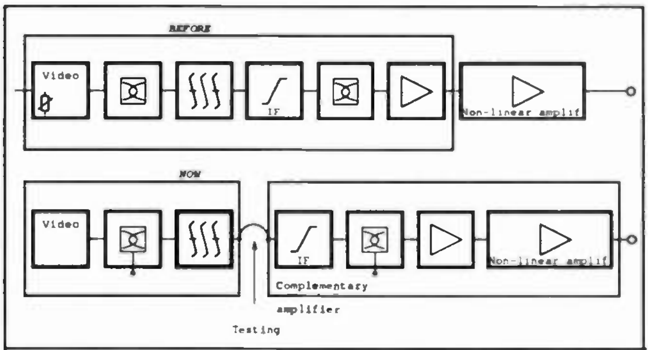


Figure 15
NEW CONCEPT IN LINEARITY CORRECTION

The transmitters drive's circuit design is based on functional chips employing standard SMD components. Such chips are easily dismountable, making for easier maintenance throughout the transmitter's lifetime. All the required single and double band tests can be performed thanks to switchable filters, while a "go home" selector switch system allows the operator to be sure his filters are operational before bringing his transmitter back on the air.

Solid-state transmitters are equipped with two transmitter drives to maintain vision and sound performance and provide total redundancy.

Conclusion

We have emphasized the importance of Availability in selecting a TV transmitting facility. In the specific case of UHF high-power transmitters, this meets the customer's immediate needs. There is no single solution, and certainly no miracle technology which could be used at all power levels. Operating costs are a decisive factor, since they take into account Reliability, Maintainability and power consumption. As was seen in the case of the emerging technologies, it is only when energy costs fall to 4.67 cents per kWh or less that the operating cost of a solid-state transmitter can compete with a combined Klystrode-transistor transmitter at a power of 30 kW. Transistorization at the 30 kW level is today technically feasible, but generally not economically viable.

A solid-state transmitter for intermediate power levels of 5, 10 and 15 kW provides the best compromise for an average energy cost, provided that the facilities used give every guarantee of reliability and maintainability. THOMSON-LGT has endeavoured to solve these two problems with its modern computing facilities.

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PAVING THE WAY FOR ADVANCED TV SERVICES IN CANADA

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In Canada, a unique proactive initiative for development and introduction of advanced television services has been seized by Telesat, the country's pioneer satellite service provider. The company has built an experimental, end-to-end HDTV network and is well embarked upon a two-year applications development program aimed at fostering new markets that would benefit from the technology in closed-circuit applications. Telesat is seeking the participation of interested partners in these endeavours.

INTRODUCTION

In June of 1989, at the Canadian Satellite User Conference, Telesat Canada announced it would conduct a two-year applications development program in advanced television. To do it, the company would invest close to $10 million ($ Cdn) to build the complete satellite-based infrastructure necessary to give Canadian entrepreneurs and video professionals hands-on experience with this powerful new communications medium.

The trial began in November, 1989.

This experimental program is "technology neutral." Telesat is at this time taking no stand on the complex international issues surrounding production and transmission standards. With more than one competing transmission system serving the trial, the company expects to develop an understanding of the advantages of each, particularly as they relate to closed-circuit applications.

Regardless of how standards issues are ultimately resolved, Telesat believes HDTV/ATV will be the future of television and it is best now for the company to concentrate on its ***applications***. Foreseeing early business opportunities, Telesat is seeking to act as a catalyst in educating and supporting potential users and investors through various demonstrations, experiments, testing, trials and market research.

Throughout the trial, market data collection and analysis will be carried out in parallel with these activities. This research is already yielding valuable results.

Why is Telesat Doing This?

As the world's first domestic satellite system, Telesat has from day one remained on the leading edge of satellite/television technology and applications. It plans to stay there.

The company places a high premium on providing excellent customer service and subscribes to action and innovation as guiding values.

A major development such as HDTV will inevitably shape the futures of Telesat's customers, as surely as it will impact the company itself. We therefore believe it essential that we not be led by events beyond our control, but rather that we attempt to understand and capitalize on the developments that will drive change.

Design, procurement and operation of its end-to-end HDTV network will give Telesat world class expertise in advanced television and prepare it to better serve both new and existing customers of many kinds. It will provide a comprehensive education in both technical and marketing issues associated with the medium and better enable Telesat to understand the future.

As the commercial owner-operator of Canada's current fleet of five Anik satellites (operating in both C and Ku Bands), Telesat has traditionally derived close to 60 per cent of its annual revenues from the provision of satellite distribution services to the broadcasting industry. It is the carrier of choice for TV broadcasters, providing network feed and program assembly services to virtually every regional and national network in the country.

Besides its large base of broadcast customers, it has a fast-growing business television clientele and sees great potential for advanced TV systems in such closely related areas as videoconferencing, product introductions, and point-of-sale promotions.

Telesat has played a leading role in the advanced television debate in Canada. Since 1982, the company has been a major force in the staging, by it and other government and industry players, of three international colloquia on HDTV. It will co-sponsor a fourth, HDTV-'90, in Ottawa June 25-29, 1990. The company's new HDTV production mobile will be on display at this year's colloquium, and both entertainment and business television events will be staged.

Telesat is also a key player in the Canadian Advanced Broadcast Systems Committee and was instrumental in North America's first demonstrations of HDTV to the general public of several cities in 1987.

TRIAL OBJECTIVES

Specifically, the objectives of Telesat's HDTV/ATV trial program are these:

1. To stimulate production by increasing the level of technical and applications expertise in the Canadian entertainment, sports, business communications and broadcasting industries, and to enhance their marketing knowledge in the process.

2. To identify and, if possible, give birth to, profitable new business ventures that can best capitalize on the technology.

3. To investigate technical issues and gain experience that will enable Telesat and other trial participants to speak with knowledge and authority about such matters.

4. To develop world class expertise within Telesat in the transmission of advanced television systems; become Canada's HDTV/ATV technical leader and affirm the company as the preferred supplier of such services to existing and future customers.

5. To act as a catalyst in creation of an HDTV production and distribution infrastructure in Canada.

By the end of the trial, in late 1991, Telesat hopes to have facilitated the launch of early-entry business ventures. With its experience gained to date, the company is convinced there exists a myriad of applications that can benefit from the unique attributes of the medium. We see the fields of closed-circuit, special event sports or entertainment broadcasts to large screens in "teletheatres," clubs, pubs, and similar venues as being particularly promising. A number of test productions are in the process of being fitted into the trial schedule.

By 1992, the basic infrastructure for advanced television's introduction to Canada will have been created and exercised, and hopefully new partnerships to exploit it established.

Above all, a core of people will have been given an understanding of the financial, technical and creative potential of the medium.

THE NETWORK

Telesat's ATV network consists of four major components:

- a mobile production unit;
- an encoding and transmission system;
- a transportable uplink;
- transportable receive and display units.

It was apparent to Telesat early in the concept definition of its network that both the production unit and the uplink should be mobile: this would maximize availability to users and expand the number and types of events that could be covered.

Production Mobile

With technical standards for ATV production in rapid ongoing evolution, commercially available equipment is in limited supply and very expensive. It is virtually all Japanese and compatible with the SMPTE 240M standard.

Telesat's tractor-trailer HDTV mobile is equipped with two Sony HDC300 CS1 cameras and two Sony HDD1000 CS1 digital VTRs, but is designed to handle three of each if need be. This equipment scans at 1125 lines in an interlaced format (2:1), at a 60 Hz field rate. With VTR playback at 1125, 2:1, 59.94 field rate, conversion to the current NTSC standard is facilitated.

The mobile has all the hardware needed for both live-to-air and live-to-tape productions, as well as limited post-production capability.

Because Telesat does not have resident expertise in video production, it is contracting with a leading production company to operate the mobile during the trial.

The Uplink
The large, multi-format transportable uplink is equipped with higher power Ku Band amplifiers required for improved carrier to noise ratio (C/N). It has the extra rack space needed to accommodate a variety of transmission systems -- including NTSC, MUSE-E and HDB-MAC. Two separate transmit chains provide complete flexibility and failure protection. The unit also has loop-back monitoring and test capability.

Transmission Systems
The HDTV signal from the production unit contains up to 30 MHZ of information per component. This has to be reduced by advanced signal processing to a bandwidth suitable for satellite transmission.

A baseband bandwidth in the 8 to 10 MHz range is preferred, so that FM modulation in a 27 MHz wide transponder channel can be used.

Telesat has purchased, and was the first commercial user of, the HDB-MAC system manufactured by Digital Video Systems, of Toronto, a subsidiary of Scientific-Atlanta. The uplink is also compatible with the MUSE-E system, developed by Japan's NHK and manufactured by Toshiba.

Receive & Display Units
The transportable receive and display units include both high definition projectors that can produce a bright, sharp image on a cinema-sized screen and Sony HD monitors. These interface with a transmission processor-decoder, satellite receiver and a trailer-mounted dish.
Audio equipment needed to optimize performance in different kinds of environments is also included.
All equipment except the antenna is stowed in transmit cases to facilitate frequent shipping.

EVENTS TO DATE

Production Seminar/Teleconference
Telesat recognized early in the game that the development and introduction of advanced television systems was at such an early stage in Canada that, before productions could be made, many of the people who would have to invest in and use the necessary hardware first had to be shown the medium and given experience in its use.

In this endeavour, Telesat found a natural and ready partner in the National Arts Centre (NAC) in Ottawa.

The NAC is a performing arts centre with a mandate to support and showcase Canadian musical, theatrical, dance and other talent -- both on its own commercial stages and throughout the country. To facilitate this national mandate, the NAC was itself embarking on a program to test and if possible capitalize on HDTV to help it better share with all Canadians this country's performing arts excellence.

Thus, on Nov. 21-23, 1989, the kick-off event in Telesat's two-year ATV/HDTV trial was a unique production seminar, co-sponsored and funded with the NAC. It was held at the NAC, and via a satellite link to the Banff Centre for the arts in Alberta.

Some 80 leading Canadian video and film professionals eagerly soaked up introductory lectures by the world's leading experts on ATV production, standards development, hardware and technology.

Next, they enjoyed three days of hands-on experience with the hardware in areas including lighting, audio and camera composition. Using the equipment destined for its new HDTV/ATV production mobile, Telesat had set up Canada's first operational HDTV control room in a dressing room adjacent to the Centre's studio theatre.

While the Japanese-developed MUSE transmission system was used as well, the event marked both the first commercial use of HDB-MAC --- to transmit the seminar to Banff --- and Canada's first tele-education event in HDTV.

More than half the participants surveyed right after the seminar by Telesat predicted that their industry would use advanced television within the next five years. Participants also rated HDTV as a highly-effective new medium for enhanced teleconferencing.

Sports in HDTV
In order to investigate the potential of advanced television for closed-circuit sports events, Telesat secured Canadian rights for a North American first, staged Dec. 7, 1989.

The much-touted super middleweight boxing match in Las Vegas between Roberto Duran and Sugar Ray Leonard was relayed via Hughes satellite to HDTV viewing sites in

Toronto, New York, Miami,and Los Angeles. It was North America's first live sports broadcast in the new medium.

Telesat chose to participate for many reasons, amongst them to acquire experience in obtaining special event rights, to better understand the roles of promoters and rights holders, to gain visibility and to get practical experience in selecting venues, selling tickets, dealing with theatrical unions and similar requirements of staging such an event.

NHK, the state-owned Japanese broadcasting system, had obtained North American HD rights to the fight. Telesat approached and obtained Canadian rights, through NHK, on condition we make our "peace" with the local closed-circuit NTSC promoter in Toronto. The event was to be co-produced with HDTV Sports Network, a consortium of Zbig Vision and Platinum Productions.

Telesat then had to find and access a venue for the event with desirable theatrical parameters. Provisions for the wider screen to accommodate HD's 16:9 aspect ratio, the G.E. projection system and its ideal distance from the screen, the type of screen and its attitude were all characteristics which had to be reviewed.

Some 300 leading Canadian sports entrepreneurs, investors, broadcasters and others watched the fight on a 25-foot screen at Toronto's Queen Elizabeth Theatre.

Their enthusiastic reaction to the demonstration reinforced Telesat's impression that there may well be a near-term market for such networked sports events to "pubs and clubs," small theatres and similar venues.

Immediately after the event, Telesat surveyed attendees to measure their reactions and impressions of the application of HDTV for entertainment services.

Two thirds of respondents stated they would be likely or very likely to attend a future HDTV broadcast. The clarity, colour quality and brightness of the picture were the most highly rated attributes, and all attributes were ranked as superior to large-screen NTSC.

Respondents were also questioned about their programming preferences for large-screen HDTV events. Live sports and first-run movies were the most highly rated, followed by orchestras and rock concerts.

Discussions with leading Canadian producers of commercial sports television programming are ongoing and it is expected similar events originating in Canada will be trialed during the two-year program.

FUTURE EVENTS

Off-Track Betting

Canada's racetrack owners have been following the Telesat Advanced Television Trial with increasing interest. Federal laws limiting off-track wagering on horse racing have recently been liberalized and discussions with a view to trialing HDTV networking from major tracks to remote entertainment and betting sites have been under way for several months.

Electronic Cinemas

The trend in the movie theatre business in recent years has been to build, under one roof, a complex of several mini-theatres. Both the more intimate size of these smaller theatres and their screen dimensions are ideal for projected HDTV.

With the high costs of producing, duplicating and distributing movies on film, the day may be closer than many think when the major studios will be distributing their products electronically via satellite to thousands of local theatres. But this will just cover the theatres' traditional night-time business.

Today, nearly all movie theatres sit idle throughout most or all of the "9-to-5" working day.

Why not link several of them, in different major cities, in a multi-purpose electronic cinema and teleconferencing network?

They might be used for a major corporation's private sales and product information meeting one day; for a live rock concert the next; and for a special sports event the next, catering to entirely different audiences for each event.

Telesat is in discussions with a major Canadian theatre chain about just such ideas.

CONCLUSIONS

The timing of the introduction of the advanced or HDTV set to the home as a mass market phenomenon is still open to debate. A consumer move to HDTV must await the outcome of standards decisions and significant reductions in the prices charged for sets.

But there seems little doubt that, by forging ahead now with applications tests and demonstrations, such as those being carried out by Telesat, new early-entry opportunities for advanced television systems operating in the closed-circuit mode have been and will be found and new industries and business ventures will be launched.

Satellites are important distribution tools for the broadcasting industry. When the satellite is linked to a large screen, a powerful new medium is created.

It is likely, then, that there will be new opportunities for both broadcasters and satellite service providers resulting from the introduction of advanced television systems.

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THE ANTENNA/TRANSMISSION LINE SYSTEM AND HDTV

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ABSTRACT

To produce the improved picture quality required for ATV transmission, extra information beyond that needed for the NTSC system is required. This paper examines how the transmission line and antenna system might affect HDTV transmission, especially in systems using the presently allocated 6-MHz bandwidth.

Some important antenna system parameters and their influence on signal transmission are examined, such as: 1) Return loss (VSWR) and ghosts; 2) RF pulse testing, and what to look for; 3) Group delay: how much is too much? 4) Group delay distortion as a measure of system quality; 5) Nonlinear components, and the generation of harmonics; 6) Antenna gain and gain stability; 7) Frequency sensitivity; 8) Time varying parameters (wind deflection, amplitude modulation caused by mechanical deficiencies, etc.); and 9) The multi-antenna environment.

The question of improvements likely to be needed in the present systems for the accommodation of future requirements is examined, and steps to achieve these improvements are suggested.

INTRODUCTION

Improved quality ("high-density") TV broadcast systems are now on the horizon. In the U.S., in Europe and in Japan, new designs are nearing readiness for market. Whatever form HDTV finally assumes, one thing is certain: Improved resolution will require transmission of television signals over a different frequency spectrum from that presently in use. Modification of existing hardware is likely to be required as well.

Most if not all of the proposed transmission schemes require a wider frequency spectrum to accommodate increased signal density. Some of the proposed HDTV transmission systems will fit within the existing 6-MHz-wide channel, but will need greater system linearity than the NTSC transmission standard. Some experts think that the transmission system specifications will impose an amplitude linearity of better than 0.25 dB and a maximum phase linearity of 2.0 degrees over the 5- or 6-MHz bandwidth of the video signal.

Transmitter manufacturers have claimed for some time that the transmitters will comply with the above specifications. However, nonlinear distortion caused by the transmitter may well be far less than the echo distortion caused by the antenna and transmission line connected to it.

The purpose of this paper is to examine the transmission line and antenna system of a typical broadcast facility, with a view toward 1) its effect on the transmitted signal, and 2) how this will impact HDTV signal transmission. The nature and degree of various types of signal distortion will be evaluated, and improvements suggested. These will be related to the problem of transmitting signals for the high-density receiving systems currently envisioned.

TRANSMISSION LINE

Transmission line may affect a signal traversing it by:

- Reducing the amplitude of the signal (line attenuation).
- Changing the arrival time at the antenna (phase fluctuations).
- Generating signals of new frequency by mixing existing existing signals (nonlinear mixing).
- Introducing echo distortion (mismatched voltage standing wave ratio [VSWR] of components and termination).

Transmission line signal distortion varies greatly by whether the line is dispersive or nondispersive.

Coaxial transmission line is nondispersive: waves travelling through it have the same velocity of propagation throughout its usable frequency range. On the other hand, waveguide of any kind is dispersive, that is, velocity of wave propagation varies as a function of wave frequency.

Attenuative Distortion

The degree of signal reduction by attenuation varies by frequency, i.e., attenuation and signal reduction become greater as frequency rises. For most transmission lines, the change in attenuation over a 6-MHz band is negligible; for dispersive lines such as waveguide, however, attenuation changes may be significant near the waveguide's cut-off frequency. High-power coaxial lines and well-designed waveguide runs, though, will produce attenuation changes of less than 0.05 dB over a television channel. Figure 1 shows typical values for attenuation as a function of frequency for waveguide and coaxial line.

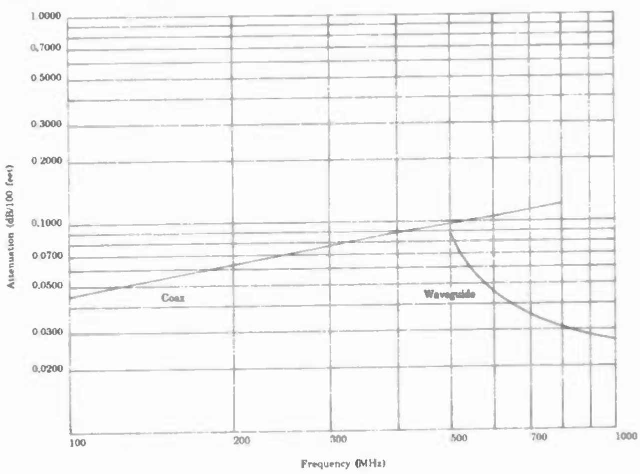


Figure 1. Typical values for attenuation as a function of frequency for waveguide (dispersive) and coaxial (nondispersive) line.

Linear Group Delay Distortion

Since wave propagation velocity does not vary in coaxial lines, there is no problem caused by changes in the arrival time of signals having different frequencies. Waveguide, however, may introduce a 10- to 30-nanosecond time difference between the travel time of the lowest- and the highest-frequency signals in a television channel. Differential time delays of this magnitude are generally either negligible or correctable.

The expression for the velocity of propagation (V_P) in waveguide is:

$$V_P = C \times \sqrt{1 - (F_c/F)^2} \qquad (1)$$

where

C = Speed of light in free space (0.9835 ft./nsec)

Fc = Cutoff frequency of the waveguide

F = Frequency of operation

As an example of the foregoing, assume Channel 30, with 1600-foot-long run of 15-inch-diameter waveguide that has a cutoff frequency of 461.13 MHz. The lowest video frequency is 651.25 MHz; highest video frequency is 655.75 MHz.

With F at 651.25 MHz, the expression is:

$$V_P = 0.9835 \times \sqrt{1 - (461.13/651.25)^2} = 0.69450 \text{ ft/nsec}$$

With F at 655.75 MHz,

$$V_P = 0.9835 \times \sqrt{1 - (461.13/655.75)^2} = 0.69926 \text{ ft/nsec}$$

With F at 651.25 MHz, 1600 feet of waveguide, time of propagation (T_1) is expressed as

$$T_1 = 1600/0.69450 = 2303.8 \text{ nsec}$$

For the same length, at 655.75 MHz,

$$T_2 = 1600/0.69926 = 2288.13 \text{ nsec}$$

Time difference $(T_1 - T_2) = 2303.8 - 2288.13 = 15.67$ nsec.

Figure 2 presents the velocity of propagation as a function of frequency for wave- guide having a cutoff frequency of 461.13 MHz (15-inch-diameter circular waveguide).

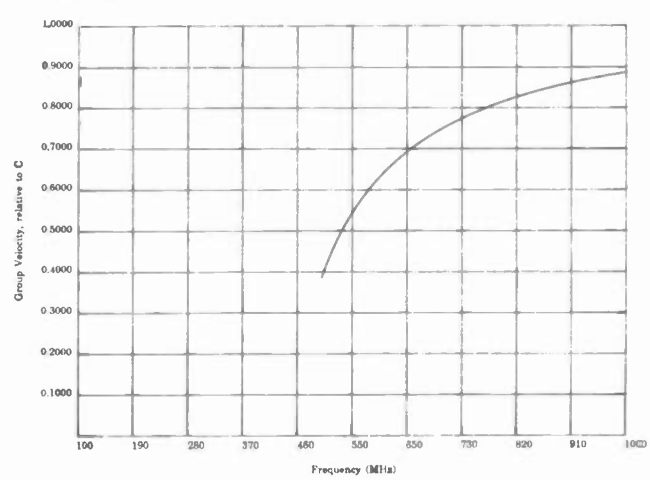
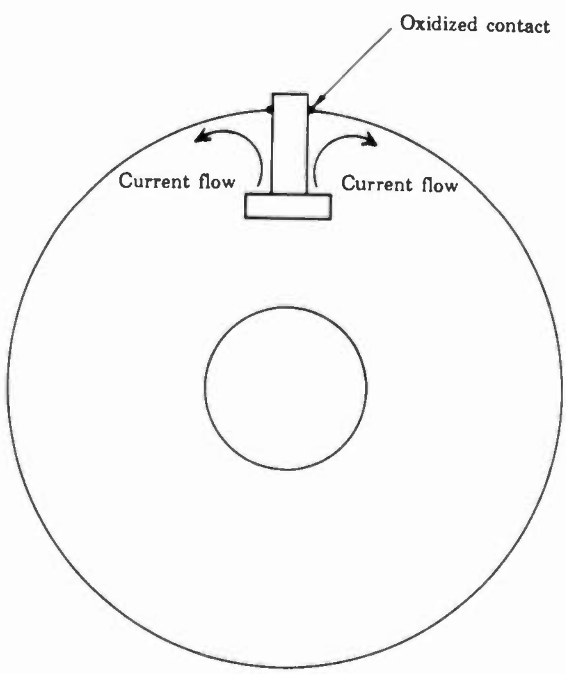


Figure 2. Velocity of propagation as a function of frequency for waveguide having a cutoff frequency of 461.13 MHz. (Normalized to the speed of light.)

Mix-Generated Frequencies

Most transmission line has some contacting surfaces made of copper or copper alloys. Oxidized copper is a rectifier, and can act as a nonlinear device in transmission line. When current, induced by more than one frequency, flows through an oxidized copper contact area, the resultant mixing of frequencies will generate a new frequency. The amplitude of the new frequency depends on the power level of the original signals and on the efficiency of the mixer. This is not a problem with NTSC television signals, but may merit consideration in other systems. Figure 3 depicts current flow in a tuner.



Cross-sectional view, coaxial cable

Figure 3. Current flow in a tuner.

Echo Distortion

As their name implies, echo distortions are caused by echoes in the transmission line. These may be reflections of the signal from some discontinuity in the line or from a mismatch of the load. The extent of distortion depends on the amplitude of the echo and the attenuation and electrical length of the line.

Echo distortions may be increased by the intentional mismatch that exists between a high-power transmitter (which has a low impedance) and the transmission line (which has a relatively high impedance), as follows:

The signal generated by the transmitter enters the transmission line, travels toward the antenna, and encounters a discontinuity in the transmission line. Some portion of the signal will be reflected by the discontinuity, and will travel back toward the transmitter. When the returning signal arrives at the transmitter port, it encounters a large mismatch, and is reflected again toward the antenna. Eventually this signal will be radiated by the antenna as a delayed image, experienced by viewers as the phenomenon of "ghosts" in the received image. (See Figure 4.)

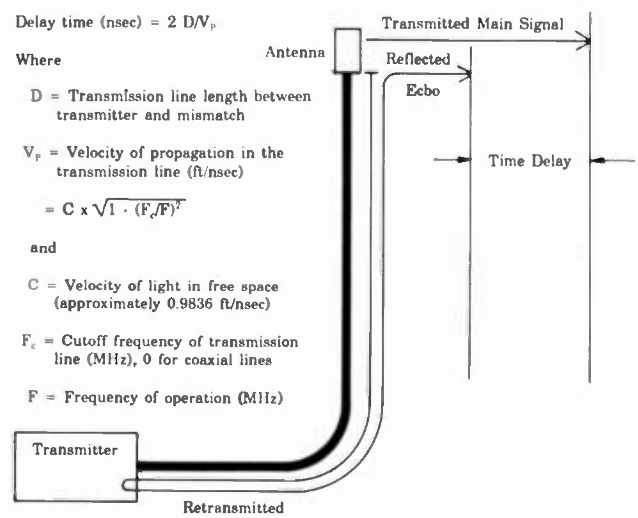


Figure 4. Echo in a transmission line.

Another aspect of distortion caused by echoes is the "monster echo." Consider the signal entering the antenna terminal. This signal will be the sum of the main signal -- the directly radiated signal -- and of all of the echoes that may be created by the line.

These effects may combine in phase to produce a monster echo, here termed E_{max}, producing the greatest disturbance of all. Figure 5 shows vector addition of echoes.

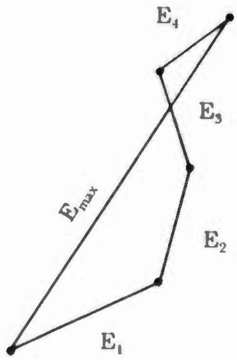
The expression for summing the echoes is:

$$E_{max} = E_1 + E_2 + \ldots E_n \tag{2}$$

E_1 and E_2 are the echoes caused by various discontinuities numbered

$1 \ldots n$

and are expressed as a fraction of the main signal. Naturally, these are complex quantities having both amplitude and phase parameters. The main signal will



In random phase, E_{max} is smaller than when echoes add in phase, as below

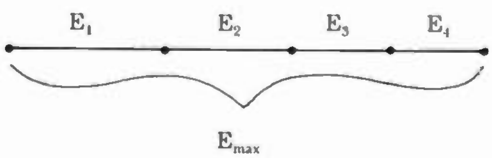


Figure 5. Vector addition of echoes.

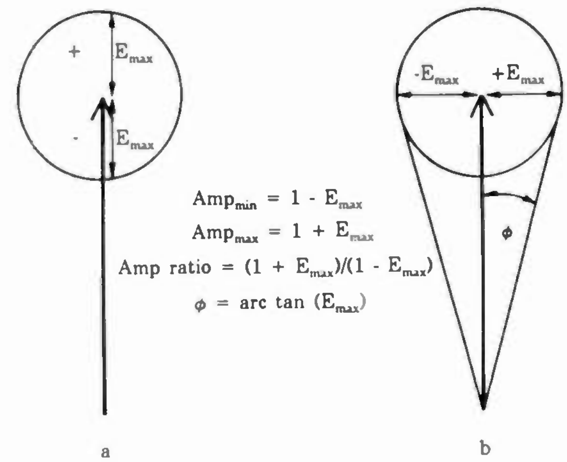


Figure 6. Vector addition of main signal and E_{max}.

serve as the point of reference, and is assumed to have a magnitude of 1.0.

The actual signal amplitude will be the sum of the main signal and E_{max}. From Figure 6a, the resultant signal will have the largest amplitude when E_{max} adds in phase with the main signal, and have the smallest amplitude when E_{max} is out of phase with the main signal.

The amplitude variation is expressed as a ratio of the minimum signal to the maximum signal:

$$\text{Amp ratio} = (1 + E_{max})/(1 - E_{max}) \tag{3}$$

Or, in decibels,

$$\text{Amp ratio, dB} = 20 \log 10 \text{ (Amp ratio)} \tag{4}$$

The phase of the signal entering the antenna will also be affected. The greatest phase change will result when E_{max} is in phase quadrature with the main signal. (See Figure 6b.)

There are two possibilities, one each for the positive (+90°) and negative (-90°) phase quadrature. The magnitude of the resulting phase change may be expressed as:

$$\text{Phase change} = \text{Arc tan } (E_{max}) \tag{5}$$

As there are two cases, one for $+E_{max}$ and one for $-E_{max}$, the total phase change is expressed as:

$$\text{Total phase change} = 2 \text{ phase change} \tag{6}$$

Assuming a near-perfect antenna/transmission line system, we may expect that the antenna and elbow complex VSWR is 1.02, and the rest of the transmission line is perfect.

A VSWR of 1.02 indicates an echo level of

$$(1.02 - 1)/(1.02 + 1) = 0.01$$

The amplitude change therefore is

$$20 \log (1.01/0.99) = 0.172 \text{ dB}$$

The phase change is

$$2.0 \text{ arc tan } (0.1) = 1.146^\circ$$

TESTING

Tests to determine how well a transmission line meets system requirements, aside from attenuation measurement, fall into two kinds. The first is designed to show how well the transmission line will work together with other system components; the second can be used to determine how much signal distortion the line will produce.

VSWR or Return Loss Testing

Voltage standing wave ratio (VSWR) testing is the most common method. Testing may be done at a number of discrete frequencies to cover a TV chnnel, or sweep frequency measurement may be used. The magnitude of the VSWR indicates how closely the line impedance remains the same as the termination, or how well the line matches the load as frequency is varied. This method will provide some insight into how well the impedance of the transmitter output and other components such as the antenna will mate with the line. The magnitude of the VSWR alone is not a reliable indicator of possible ghosting. Skillful interpretation of the frequency vs. VSWR curve is necessary to draw any meaningful conclusions about the timing of the echoes in the line.

RF Pulse Testing

This method is used to find the sources of reflections, and to measure the amplitude of the reflected signal in a transmission line. It is an excellent indicator of possible ghosting, but will not directly show the impedance behavior of the transmission line.

In this method, a short RF pulse is generated and injected into one end of the line. The reflected pulse is then compared in amplitude to the incident one. The time it takes for the reflected pulse to appear indicates the electrical distance to the source of the reflection; the amplitude indicates the severity of the problem. The pulse is shaped, so it represents the frequency content of an NTSC transmission.

The amplitude of the reflected pulse is proportional to the magnitude of the discontinuity that caused the reflection. It is also an excellent indicator of the signal distortion one may expect. The greatest echo level is the sum of all of the reflected pulses. We may safely assume that all of the echoes will add in phase at least once, especially when they originate from a point far from the transmitter. Figure 7 is the result of a typical pulse test.

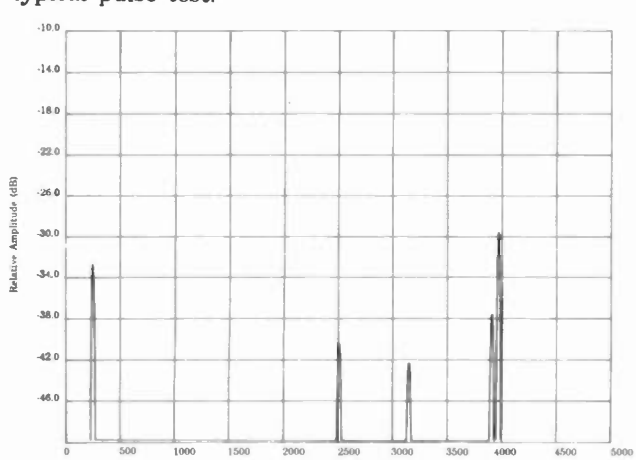


Figure 7. Results of an RF pulse test.

The location of the source of reflection is also very important. A reflected signal generated very close to the transmitter will travel about the same length of line as the main signal. The "ghost" produced by this echo appears barely displaced from the original picture, so it is undiscernible to the naked eye. The phase relationship between the main and reflected signals will also be essentially constant constant with frequency. The sum of the main signal and the echo will no longer exhibit the cyclic amplitude and phase variation damaging to picture quality.

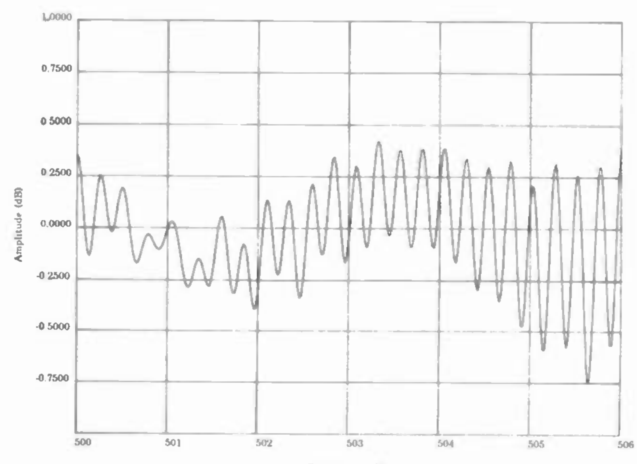


Figure 8. Amplitude ripple vs. frequency caused by echoes.

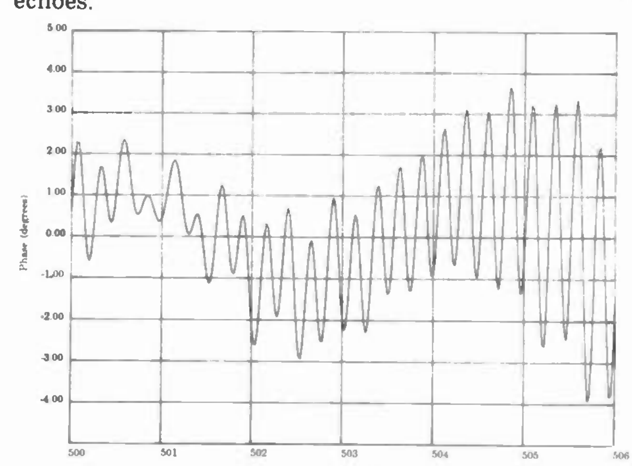


Figure 9. Phase ripple vs. frequency caused by echoes.

A reflection far away from the transmitter causes the reflected signal to travel a much longer distance than the main signal. Even a small change in frequency will cause a large change in the phase between them. The summation of the reflected and main signals will go through as many cycles as the phase between them does. Figures 8 and 9 are the calculated amplitude and phase vs. frequency behavior of a 2000-foot transmission line with an antenna.

THE ANTENNA

For purposes of simplicity, think of the antenna as the device that connects the end of the transmission line to the customer. Of course the connection is not hard-wired, but it is helpful to view the antenna in this manner, omitting any other influencing factors such as propagation, receiving system, terrain, etc.

Input Impedance

The signal passing through the antenna will have to enter it, so the input impedance of the antenna will affect its amplitude and phase.

The amplitude of the signal entering the antenna will be reduced by the amount which is reflected due to the mismatch between the transmission line and the antenna. The phase of the signal will be proportional to real and reactive components of the complex input impedance. Figure 10 shows a typical antenna's impedance on a Smith Chart. Figure 11 shows the corresponding reflection coefficient vs. frequency characteristics for the same antenna.

The amplitude reduction (i.e., the amount of signal rejected) is not very much for the VSWR levels normally encountered. The reduction of signal level in dB is:

$$\text{Loss in dB} = 10 \log (1 - \{[\text{VSWR} - 1]/[\text{VSWR} + 1]\}^2)$$

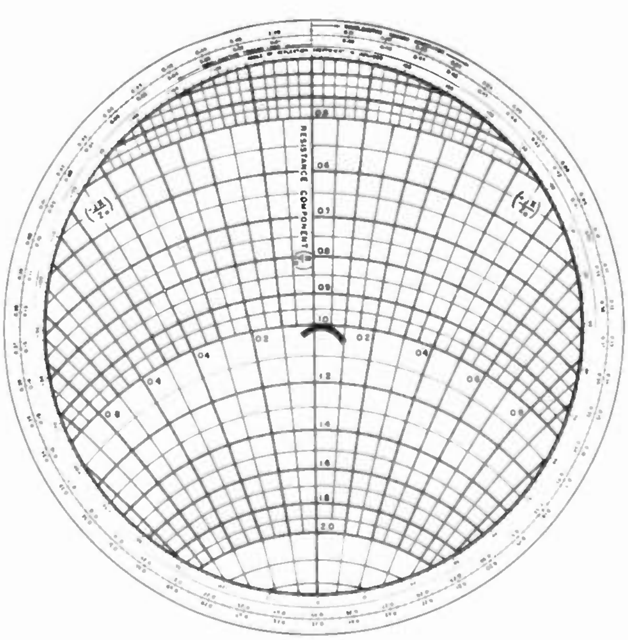


Figure 10. Impedance vs. frequency for a typical antenna.

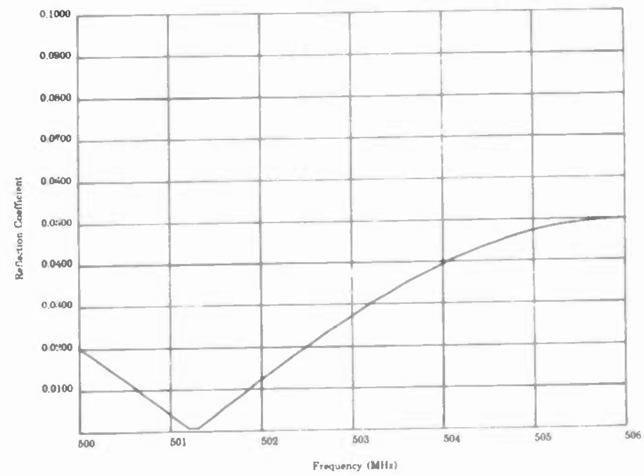


Figure 11. Reflection coefficient vs. frequency for a typical antenna.

When the VSWR is at 1.1, the loss is negligible, about 0.01 dB.

The phase of the signal is determined by the ratio of real to reactive components of the input impedance, or

$$\text{Phase} = \arctan (\text{reactance}/\text{resistance}) \tag{7}$$

Assume a case where the VSWR of the antenna is 1.0 at midband, and 1.1 at the band edge. The phase at midband will be the norm, or zero degrees, since

$$\arctan (0/1) = 0$$

The phase at the low end is

$$\arctan (0.1/1.05) = 5.44^\circ$$

The phase at the high end is

$$\arctan (-0.1/1.05) = -5.44^\circ$$

The total phase variation is 10.88°!

Even for the most perfect system, mentioned before, the total phase change of the signal entering the antenna is

$$\text{total phase} = 2 \arctan (0.02/1) = 2.29^\circ$$

(It is assumed that the reactance is about 0.02 for a VSWR of 1.02.)

Antenna Gain

Antennas are frequency-dependent devices. The aperture of an antenna is set, but the aperture size in terms of wavelength is a function of frequency. The electrical size and therefore the electrical parameters of a television antenna will change with frequency. This also includes the gain of the antenna. Careful design will minimize the gain change to be less than about 0.1 dB over a channel.

Radiation Pattern

The signal amplitude received by the viewer depends on the antenna gain and the relative pattern amplitude radiated toward the receiving site. Change of the pattern amplitude as frequency changes will cause a corresponding amplitude modification; in effect superimposing an amplitude modulation on the received signal. The amplitude amplitude relationship among the frequencies contained within the transmission will be modified. The originally transmitted picture now has been modified to some degree.

There has been some concern over the stability of antenna patterns for some time. The signal from most antennas will be stable and not dependent on frequency when viewed from small depression angles (20 to 100 miles away). Close to the antenna, at large depression angles, the antenna pattern may be very cyclic. This may cause the signal to be greatly dependent on frequency.

Center-fed antennas are vulnerable to angular shifts of the side lobe positions. The side-lobe amplitudes are also fequency-dependent. These changes may introduce a signal variation of as much as 10 to 12 dB in the angular range where the side lobes dominate. End-fed antennas will exhibit some movement in the angular position of the peak of beam. However, an end-fed antenna will produce less amplitude variation at low elevation angles. Although the total signal variation from an end-fed antenna may be 5 to 8 dB, careful design will limit this to about 2 dB.

How much of this is tolerable? That depends on the kind of transmission scheme employed, and on the correction circuits that may be used. The correction may be reasonably simple, provided that the amplitude variation is fairly small and linear or nearly linear with frequency. The NTSC television system uses amplitude modulation for conveying video information. The secondary amplitude modulation, superimposed on the signal by the antenna, may degrade the picture quality. This degradation may affect any system, including HDTV, which relies on amplitude modulation.

Wind Deflection

Antennas are specified to withstand -- that is, survive -- wind velocities in excess of 100 mph, depending on their location. The generally accepted wind velocity specification for operational wind velocity is 50 mph. Top-mounted units are not supported by the tower, and so will bend or deflect in wind. They bend to a parabolic shape. The displacement of the antenna from the vertical varies exponentially with height. The "bending" of the antenna changes the apparent phase relationship among the radiating elements.

The result is that the main beam of the antenna will be displaced; the antenna pattern is modified and the gain is effected. All of these cause the amplitude of the signal to vary as the antenna sways in the wind. Careful design will minimize this problem (stiff antenna, smooth pattern, etc.)

Multi-Antenna Environment

When there are two or more antennas in proximity at the same height, there will be some interaction. The signal radiated by one will be reflected by the others. The reflected signal, travelling a long route, arrives at the receiver delayed in time. This of course is the same as an echo in the transmission system; the same problems result. The magnitude of the reflected signal depends on the distance between the antennas and the electrical cross section of the reflecting antenna. The time delay caused by the extra distance is a function of the physical layout of the installation and the location of the viewer in relation to it.

Calculating Amplitude, Phase Distortion

The worksheet presented in Figure 15 will help in assessing the amplitude and phase distortion one may expect to be caused by the transmission line/antenna system. The prediction of visible ghosting is left solely to the RF pulse test method.

SUMMARY

Attenuation and velocity of propagation are parameters that change slowly with frequency. Even in dispersive transmission lines, the changes caused by these attributes are not considered to be significant.

Transmission lines cause nonlinear distortions mainly due to reflections (echoes). A typical transmission line system having a VSWR of 1.05 will produce a phase variation of about 2.8 degrees, and an amplitude change of about 0.45 dB.

The HDTV systems currently under proposal may need transmission line systems with improved performance. The improvement is not likely to result from improved component specification. Correction circuits may be used, or the mismatch between the transmitter and the transmission line needs to be reduced.

The correction circuits may either reduce the signal distortion caused by the transmission system by some method of pre-emphasis or post-emphasis correction, or interrelate the various signal levels to estimate the correct one.

Echo distortions may be reduced in a number of ways, all aimed at reducing or eliminating the echoes:

■ Increase transmitter output impedance to correspond to the impedance of the transmission line. This would be a highly unpopular solution, because the transmitter efficiency would be drastically reduced. Only about one-half of the transmitter power would be available as input to the transmission line, but the echoes would be absorbed by the transmitter, and eliminated.

Transmission Line

Pulse	Location	Distance (feet)	Reflected Pulse Amplitude (dB)	Reflected Pulse Amplitude (Ratio)
1	bottom elbow	120	-34	0.0200
2	line	1278	-44	0.0063
3	line	1600	-47	0.0045
4	top elbow complex	1990	-40	0.0100
	Antenna			
1	antenna input	2000	---	0.0500
	TOTAL (sum of all reflected signals)			0.0808

Use the maximum measured VSWR within the 6-MHz channel.

RF pulse test measurements produced the amplitude values shown. The reflected pulse amplitude relative to the incident, or main signal, is calculated using the expression:

Relative amplitude = $10^{-dB/20}$

The reflection coefficient for a VSWR of 1.1 = 0.05

Max/min amp ratio = (1 + 0.0808)/(1 - 0.0808) = 1.175

Amplitude ratio, dB = 20 log (1.1418) = 1.41 dB

Phase change, deg = arc tan (0.0662) = 4.62 deg

Total phase change, deg = 2 (3.785) = 9.23 deg

Figure 12. Worksheet: Calculating amplitude and phase distortion.

■ Reduce echo reflections from the transmitter by absorbing the signal travelling toward the transmitter. This may be accomplished with an isolator. The isolator is a directional device capable of allowing signal propagation in one direction, but greatly attenuating it in the other. High-power isolators are now becoming available for this purpose.

■ Improve the antenna and transmission system by reducing or eliminating the mismatch at the point where it occurs. This requires the use of the pulse technique, or perhaps an improved system design, so the location of mismatch may be determined with a high degree of accuracy (within 24 inches).

■ Improve the antenna input impedance so that the match between the antenna and the transmission line is made lower than is provided by the currently accepted VSWR specification of 1.08 to 1.1, bringing it to a range of from 1.02 to 1.03. The antenna tuning should not be optimized to a single frequency (i.e., the visual carrier). The broader spectrum of high-density TV will require a low VSWR over a wider band. Depending on the modulation technique used, this may encompass most of the available 6-MHz channel.

The antenna will cause phase and amplitude changes, due to its frequency-dependent input impedance and radiation pattern. The input impedance is more likely to produce phase changes, the radiation pattern to produce amplitude variations. Careful antenna design reduces the effect of both. The absolute worst-case results are shown in Figure 13. The calculation was performed for the parameters shown on the worksheet, Figure 12.

(In each case, variations are calculated by simple addition)

Maximum amplitude variation due to echoes	1.41 dB
Maximum amplitude variation due to antenna	2.10 dB
Maximum differential gain	3.51 dB
Maximum phase variation due to echoes	9.23 deg
Maximum phase variation due to antenna impedance	2.86 deg
Maximum differential phase	12.06 deg

Figure 13. Worksheet: Calculating worst case distortion.

A CODEC FOR HDTV SIGNAL TRANSMISSION THROUGH TERRESTRIAL AND SATELLITE DIGITAL LINKS

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Summary - *A codec for digital transmission of HDTV signals is described. The bit-rate compression performed by the codec is based on advanced techniques such as: spatial Discrete Cosine Transform (DCT), temporal Differential PCM (DPCM), variable length coding. The codec is designed to operate with both the interlaced studio systems 1125/60 and 1250/50 and, thanks to the inherent flexibility of the packet structure of information, a wide range of line bit- rates can be used as a compromise between video quality and bit- rate constraints of the digital transmission link. The flexibility of the HDTV codec is highlighted through examples of applications over satellite digital links in practical situations.*

INTRODUCTION

High Definition Television (HDTV) has entered the world of electronic film and television production and is expected to become the future visual communication medium for professional, commercial and domestic applications. A cumbersome problem of HDTV is the need to compress the source high bandwidth for transmission over physical channels. In presently proposed production systems with interlaced scanning, the needed analogue bandwidth is around 60 MHz (30 MHz for luminance and 15 MHz for each one of the color differences) if full horizontal definition is transmitted (1920 active pixels), or around 44 MHz if the horizontal definition is nearly twice the 4:2:2 standard (1440 active pixels). If digital techniques for HDTV transmission are considered, the PCM coding of the active portion of the picture, with 8 bits per sample, entails a bit-rate of about 950 Mbit/s or 700 Mbit/s for the two cases above mentioned. For analogue transmission of HDTV (particularly for satellite broadcasting), signal bandwidth reduction systems have been devised, that combine some degree of loss in the horizontal resolution of the source signal with subsampling in space and time for static portions of pictures (offset in time subsampling, together with the use of two frame memories in the reciever, allows a good reproduction of static pictures) and with additional spatial subsampling only for the moving parts of the picture. For some specific applications, like contribution links to connect HDTV studios, or for some high quality demanding non-broadcast applications, the digital transmission of the HDTV signal offers significant advantages, taking into account the evolution of terrestrial links towards the digital technology and (thanks to the advance in digital integrated circuits) the possibility of implementing, in a convenient manner, sophisticated algorithms for significant bit-rate reduction with negligible loss in picture quality. The present paper describes a codec for HDTV applications, in particular for the 1125/60 and 1250/50 interlaced systems, considered by the CCIR[1,2].
The codec is based on the application of Discrete Cosine Transform (DCT), combined with temporal differential PCM (DPCM), on blocks of 8x8 samples for both luminance and color differences. The line bit-rate can range from a minimum value of about 60 Mbit/s up to more than 140 Mbit/s. This range includes the standard digital CCITT hierarchies of 2x34, 2x45, 140 Mbit/s. For higher bit-rates the full studio quality (1920 pixels/line) can be transmitted, while for lower bit-rates, a horizontal source definition of about 1440 pixels/line is a better compromise, in order not to stress the compression algorithm. In any case the codec is very easily adapted to the two source horizontal definitions, and the compression algorithm adapts automatically to any line bit-rate. After a description of the structure of the HDTV codec (the encoder only will be described, as the decoder performs mainly the inverse operations with respect to the encoder) and of the main algorithmic parts, the application of the codec on several transmission links will be considered, with main emphasis on satellite transmission.

CODING ALGORITHM

The HDTV codec hardware is based on a few basic concepts and algorithms, that are tight together in a more or less sophisticated manner[3]. Main building blocks of the encoder are described with reference to Fig. 1 and Fig. 2.

Parallel processing

To match the high sample rate of the HDTV signal with the high integration level of presently available HCMOS technology (having a speed limit of the system clock in the range of 30-40 MHz) a parallel processing of the video picture is performed. The picture is divided through vertical lines in 4 (1440 pixels/line) or 5 (1920 pixels/ line) vertical stripes, which are parallel-processed with a sample rate, for each processor, divided by 4 or 5, respectively. Only the active part of the picture is processed. The digital codec interface of Fig. 1 performs, as its main task, the splitting of the video picture for parallel processing.

DCT Algorithm

Each picture part of the HDTV field is divided into stripes of 8 consecutive lines, and each stripe is rearranged into blocks of 8x8 samples of luminance of color differences that are processed in time along the sequence Y, P_R, Y, P_B. The DCT is applied to blocks of a field (intrafield coding) for moving areas, or to the difference, at block level, between the present field and the copositioned field in the previous frame in the case of blocks which are not affected by motion (interframe coding). As a further possibility the DCT can be applied to time adjacent fields, using line interpolation on the previous field (interfield coding). The selection of which coding mode has to be used (intrafield, interfield, or interframe) is performed a priori on the basis of the energy comparison of the blocks or of block differences. The predicted block can be the block in the previous field, or a zero value, if intrafield DCT has to be performed. DCT processed data are properly quantized, sent to the following encoder part, and processed in an inverse manner (and eventually added to the predicted value if different from zero) to rebuild the block portion of the field, to be stored in a frame memory (this will be the previous field when next field is coded). The picture block reconstruction is the same as performed at the decoder side. The quantizer is a quasi-linear quantizer with a variable step size, that is changed according to the multiplying factor 2 exp(n/8), where n is controlled by the line buffer occupancy. The mean value of n (in the long term and for the same order of picture complexity) depends on the line bit-rate, while in the short term it depends on the picture complexity. Even though not presently implemented, the codec can include all needed circuits for motion compensation so that in the case of blocks affected by translational motion, of limited amount, from field to field, the more efficient interframe coding can be still used. Except for motion compensation, the schematic diagram of the DCT processing part of the codec is shown in Fig. 2.

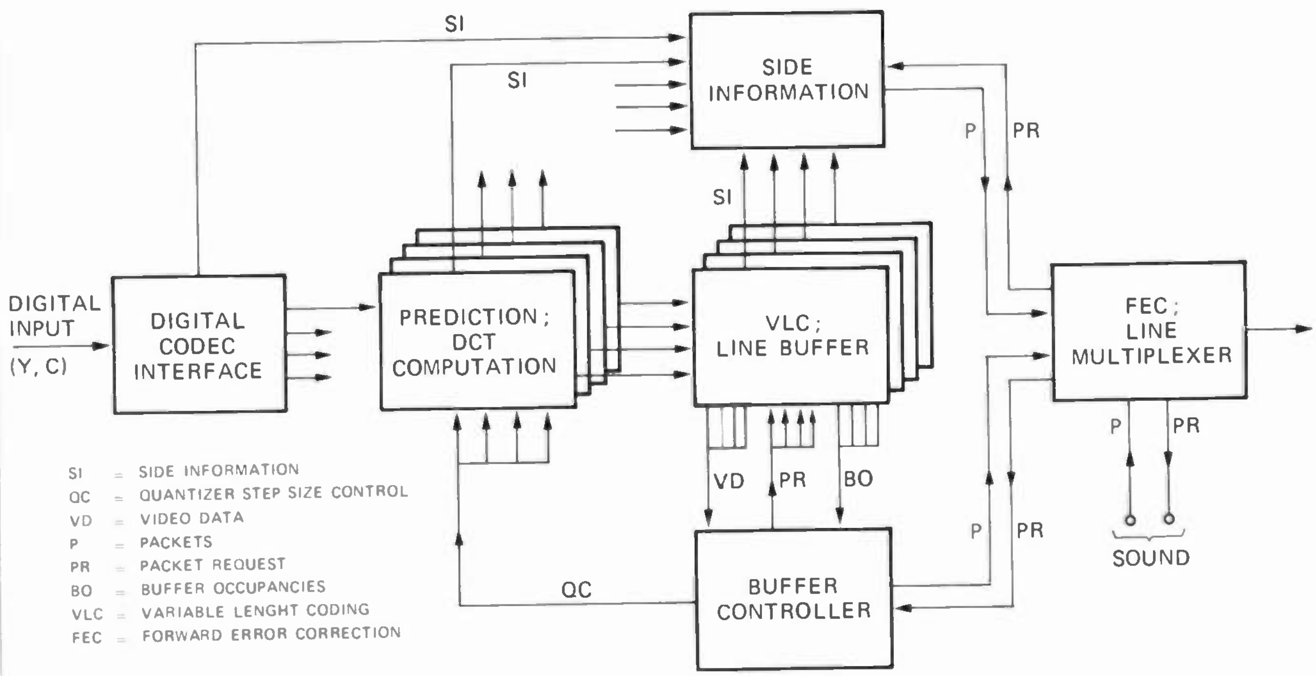

Fig. 1 - Block diagram of the HDTV encoder.

Variable length coding (VLC)

The DCT transformed data have a very steep two-sided probability distribution and a reduction in the line bit- rate is obtained if a variable length coding is used. The length of the coded words ranges from 2 to 18 bits, in steps of 2 bits, and shorter word lengths are assigned to more frequently occurring DCT data. The VLC together with the inverse

VLC operation performed at the receiving side is in itself a transparent operation.

first type is related to the VLC data concerning the video picture. Data belonging to each 8x8 luminance block or to

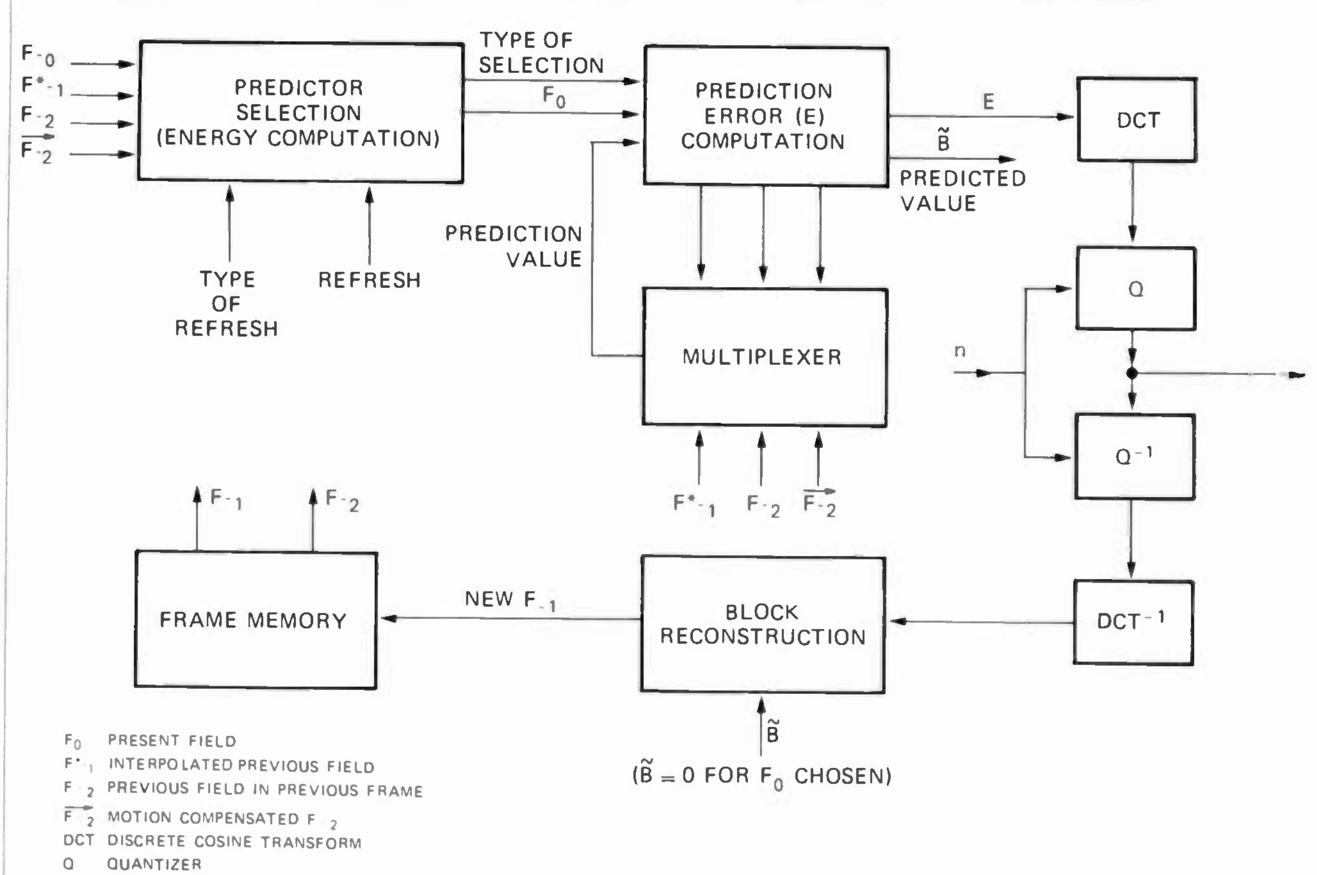

Fig. 2 - Block diagram of spatial DCT and temporal differential PCM computation

Line buffer

For each processor, related to a vertical stripe of the HDTV picture, a line buffer of 4 Mbit/s capacity is provided; the function of the buffer is to smooth the information rate, coming out from the VLC, that changes due to the VLC itself and to the amount of data coming from DCT computation, in its turn related to the picture block complexity. The value of n, in the expression 2 exp(n/8) of the quantizer step size, is a monotonic function of global buffer occupancy; special provisions are undertaken to avoid buffer overflow and underflow. The buffers inserted in each processor are globally controlled as a single buffer (in terms of overflow, underflow and value of n) by the buffer controller.

Handling of information for transmission

The line multiplexer receives three types of information. The first type is related to the VLC data concerning the video picture. Data belonging to each 8x8 luminance block or to each one of the color difference blocks are separated by special words (2 of them are used) belonging to the VLC Table; these special words cannot be used for coding DCT transformed data. The second type of information concerns the description of the codec working mode on the basis of a field, of a stripe of 8 lines or of a block. These data are not VLC coded and are sent to the transmission line as a separate data stream. A third type of information concerns the auxiliary data, the most important of them being sound channels. A separate bit stream at 1.5 or 2 Mbit/s is foreseen for sound transmission. All the information data streams are arranged in packets of a length of about 3800 bits, each packet having a header to identify the source (for picture information one source address for each processor is needed). The packet header of video data packets conveys also the informa- tion of global buffer occupancy and of video source clock. The packets are sent at a continuous rate, and their insertion into the line multiplexer follows a priority assignment (for example the sound packets have the highest priority, as sound transmission cannot be arbitrarily delayed). The packet structuring of the information has several advantages. Two of them are of particular importance: a new source of information can be added without any rearrangement of the codec, and dummy information can be inserted into the multiplexer in the case of a transmission link with non-homogeneous rate capacity; for example, when a terrestrial link is followed by a satellite link, with a lower bit-rate capacity with respect to the terrestrial link, the standard hierarchical capacity of the terrestrial link can be filled up with dummy packets (coming from a dummy source) which can be easily eliminated before entering the satellite link. The video signal source packets are sent from the line buffers to the multiplexer through the buffer controller which provides for the packet header insertion.

Line Multiplexer

The line multiplexer performs the following functions:

1. Generation of the frequency for reading data from several sources (video, sound, side information); depending on this frequency is the packet request frequency, which assures a continuous steady flow of packets into the line. The packet rate for side information and sound is a constant, independent from line bit-rate, while the video data packet rate is automatically changed to fill the line rate capacity, through the feedback from buffer controller to the variable step size quantizer following DCT computation.
2. Insertion of Forward Error Correction (FEC) at packet level by a two error correcting BCH code with 16 bit interleaving. A Reed-Solomon code is also foreseen, working at byte level with two byte interleaving depth. Both error correcting codes increase the data rate by 6.3%.
3. Build-up of line framing structure, in particular the insertion of frame synchronization word, service bits, order wire channel.

The multiplexer can be thought as formed by two different parts. The first one performs the above mentioned functions and works at byte level, the second part depends on the line characteristic and provides for byte partition into two data streams, in the case of line transmission at 2x34 or 2x45 Mbit/s, for parallel serial con- version and for a proper introduction of a line code. In the case of line transmission through two bit streams (at 34 or 45 Mbit/s), at the receiving side (line demultiplexer) a specific circuit provides for the realignment of the two data streams, with a realignment capability up to several hundred bits of relative delay.

APPLICATION OF THE HDTV CODEC ON SATELLITE LINKS

The use of satellite links, associated with the adoption of "digital" techniques and "flexible" codecs, plays a fundamental role in the HDTV signal transport and distribution, allowing high and constant picture quality and harmonised integration with the existing digital terrestrial networks and future Broadband ISDN. The digital solution also offers important advantages over the analogue FM, i.e: higher ruggedness against noise and interferences, better exploitation of the transponder capacity for transmission of two or more TV programmes in alternative to one HDTV, flexible use of the multiplex capacity for picture, sound and data, encryption and scrambling for Pay-TV[4].

Transmission capacity

INTELSAT and EUTELSAT communication satellites carry transponders in Ku-band (14/11 GHz), with 36 MHz and 72 MHz bandwidths, for FM/TV transmissions and TDMA telecommunication services, at line bit-rates of 60 and 120 Mbit/s with QPSK modulation. Some domestic satellites, in North America, are equipped with of 54 MHz transponders for FM/TV and TDMA/QPSK transmissions at 90 Mbit/s. Since the transponders are potentially transparent to the signal format, transmission of digital HDTV at 60, 90 and 120 Mbit/s QPSK is possible without operational constraints. Unfortunately these transmission rates do not allow "transparent" interworking with terrestrial digital networks operating at the CCITT hierarchy levels III and IV, adopted in Europe and North America (34, 45 and 140 Mbit/s). Therefore, in the circumstances where the HDTV signal is conveyed from the studio to the satellite up-link station via a terrestrial CCITT network the availability of "flexible" HDTV codecs capable to be adapted, with minor hardware modifications, to the satel- lite transmission rate is a fundamental requirement. On the other hand, when the up-link station is close to the site where the HDTV signal is produced, and a CCITT digital link is not required, the codec can directly operate at the satellite transmission rate, i.e. 60, 90, 120 Mbit/s, or at any other suitable rate. The HDTV codec, described in this paper, is capable of meeting both the requirements of "mixed" terrestrial/ satellite operation and "independent" satellite operation. Moreover the inherent flexibility of the packet structure allows ease interfacing between CCITT digital links at 140 Mbit/s and satellite transmission at 120 Mbit/s by introduction/extraction of dummy packets in the multiplex.

Channel coding for error protection

For a digital HDTV system via satellite the overall service quality jointly depends on the performance of the picture coding algorithm and on the available margin against noise and interferences, allowed by the RF-channel performance.

The system optimisation then requires a trade-off in the bit-rate allocation between "video coding" and "channel coding" to achieve the highest picture quality with minimum outage times, also at the beam edge and when small transmitting earth stations and receiving installations are used. As regards the modulation techniques no serious alternatives to QPSK, at 1.5 Bit/Hz, are foreseen in a short time-scale although significant efforts are being done to allow transmission at 140 Mbit/s over a 72 MHz transponder. A possible candidate is 180 Mbit/s/8PSK with (7/9) trellis coding, which unfortunately requires significant hardware complexity in

N	Inner code (FEC 2)	Outer code (FEC 1)	E_{bu}/N_o at 10^{-8} (dB)	Spectral efficiency %
1	----	----	15	100
2	----	RS(255,239)	10	94
3	Convol. 3/4	RS(255,239)	6.6	70
4	Convol. 1/2	RS(255,239)	5.4	47

Table 1 - Tipical performance of some channel techniques (QPSK modulation with coherent demodulation).

HDTV SYSTEM		Bit-rate for picture coding (Mbit/s)	C/N (dB) required at 10^{-8}	Receiving antenna diameter (meter)	
				Clear sky	99% of the worst month
A	120 Mb/s	110	14.9	4.6	8.2
B1	90 Mb/s	80	13.5	3.7	6
B2	120 Mb/s + 3/4 conv	80	10.2	2.4	3.3
C1	70 Mb/s	62	12.4	3.1	4.7
C2	90 Mb/s + 3/4 conv	62	9.1	2.1	2.9
D	60 Mb/s	53	11.7	2.9	4.2

Table 2 - Diameter for the receiving antenna to achieve the required C/N (in 36 MHz) for some HDTV systems via 72 MHz Eutelsat II transponder (QPSK modulation with coherent demodulation).

the decoder. For use over terrestrial networks and for a wide range of applications over satellite links, the error protection scheme of the HDTV codec using the interleaved RS(255,239) code, recommended by the CMTT for (4:2:2) codecs, assures adequate service quality. A BER of 10^{-8} after correction (approximately corresponding to grade 4 of the CCIR 5 level impairment scale) is achieved at a channel BER of 4.5 10^{-4} with a coding gain of about 1.3 dB over an interleaved BCH(255,239) code[5]. Moreover, the interleaved RS(255,239) code can correct very long error bursts (up to 128 bits). The concatenation of a convolutional inner code (FEC 2), with rate 3/4 or 1/2 and soft-decision Viterbi decoding, and the RS(255,239) outer code (FEC 1), capable of correcting the long error bursts originated after the convolutional decoder, offers high coding gains (at the expense of reduced bit-rate capacity for picture coding) and allows reliable operation even in dif- ficult reception conditions. Table 1 indicates the expected performances of the three coding techniques with respect to uncoded QPSK, for an equal useful bit-rate, over a typical satellite channel with saturated TWTA, in terms of the Ebu/No ratio required for a BER of 10^{-8} after the outer code at the receiving side. To be noted that technique no. 2 is adopted in the HDTV codec described in this paper. The advanced techniques no 3 and 4 are adopted in some TDMA/ modems for telecommunication services. The implementation of the soft-decision Viterbi decoder associated with the QPSK demodulator, at the high bit-rates required for HDTV, is actually the main problem to be solved. However, the good perspectives for high speed VLSI chips allows to consider these techniques as realistic solutions for the near future.

Examples of digital HDTV systems

Table 2 gives some examples of HDTV systems for satellite applications based on the described HDTV codec. The RS(255,239) error correcting code is assumed as basic FEC for all systems. The various systems are grouped in four families according to the available bit-rate for picture coding. Systems A, B1, C1 and D represent direct applications of the basic HDTV codec and only differ on the transmission bit-rate. The input signals from the codec to the QPSK modem, and vice versa, are data streams at 120, 2x45, 2x34 and 60 Mbit/s, respectively. System C1 and system A, in a version operating at 140 Mbit/s, are at an advanced stage of development. System C1 (70 Mbit/s QPSK) will be used by the RAI for experimental transmission of digital HDTV by satellite[4]. Systems B2 and C2 are advanced versions of systems B1 and B2 respectively and, for the time being, have not yet been developed. Both systems require the implementation of the 3/4 convolutional codec which could be directly associated with the QPSK modem. In Table 2 the expected performances of the various HDTV systems on a 72 MHz transponder of Eutelsat II (European beam, 14.5/11.2 GHz) are compared in terms of required receiving antenna diameter, at the beam edge, for a BER of 10^{-8} ofter the outer code, under the following link-budget assumptions:

- E.i.r.p. earth station (transportable): 74 dBW
- Satellite G/T: 2 dB/K
- Satellite e.i.r.p. at saturation: 44 dBW
- Receiving system N.F.: 1.6 dB
- Rain attenuation (99% of worst month): 1.9 dB up; 1.2dB down
- Noise increase due to precipitation: 1.2 dB
- C/N impairment due to interferences: 2 dB
- Available C/N (36 MHz) in clear sky: 15.5 dB; for 99% of worst month: 12.9 dB

The satellite coverage, with 44 dBW e.i.r.p., includes most of Western and Central Europe countries. Under clear sky reception, all HDTV systems of Table 2 can satisfactorily operate with a transportable up-link station (74 dBW e.i.r.p.) of the kind normally used in Europe for FM/TV, and with simple receiving installations not requiring antenna tracking facilities. In these conditions, system A would allow the operation of very high quality HDTV (110 Mbit/s useful bit-rate for picture coding) in a European context. Under unfavourable receiving conditions, with rain attenuation on up and down links, both systems A and B1 require rather large receiving installations with antenna diameters of 8.2 and 6 meters, respectively, and tracking facilities. In these receiving conditions, system B2, based on an advanced error correction scheme (RS concatenated with 3/4 convolutional code and soft-decision Viterbi decoding) is then particularly attractive since it allows operation of high-quality HDTV (80 Mbit/s useful bit-rate) for most of the applications, using small receiving antennas. No reception problems, even under unfavourable conditions, should exist for HDTV systems C1, C2 and D, which allow useful bit-rates of 62 and 53 Mbit/s.

Conclusions

The general structure of a HDTV codec suitable for a variety of applications has been described. The codec exploits the three-dimensional correlation of the source signal using spatial DCT and temporal DPCM. The adoption of a packet structure to convey the information (video, sound and data) allows a very high flexibility in terms of possible line bit-rates, of advanced error protection strategies, of video quality, and of additional capacity for auxiliary data. The codec automatically adapts to both the interlaced signal sources: 1125/60 and 1250/50. For applications requiring low line bit-rates, a reduced sampling frequency of the source, with respect to the studio systems, can be more convenient in terms of the achievable video quality. The codec is provided with forward error correction (FEC) inside the multiplexer (and in the demultiplexer at the receiving side), which requires only a reasonably low capacity of the line bit-rate. The basic FEC, adopted in the codec, is in a certain way redundant for transmission over optical fiber links, which, even for long distances, present error rates lower than 10^{-8}, whilst for terrestrial radio links it assures a very good quality with channel error rates up to 10^{-4}. The requirements for satellite transmission are more complex, when compared with terrestrial transmission, due to the wider set of parameters which characterise this type of links. Since professional applications of digital HDTV by satellite are expected in the near future, a more detailed description of this topic, accompained by some examples, has been given in the paper. Specific applications of the described HDTV codec over satellite links can be envisaged in the following areas:

- exchange of high quality HDTV material among Production Centres
- broadcasting from outside locations to a Production Centre (e.g. satellite new gathering)
- electronic movie distribution
- live transmission of special events (e.g. sport events) to selected locations
- transmission from Production Centres to optical fiber distribution headends, to up-link earth stations (for conversion to analogue HDTV distribution systems) or to terrestrial broadcast stations (for conversion to standard TV systems, like ATV, NTSC, PAL, etc.).

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THE COST OF CONVERTING A BROADCAST FACILITY TO HDTV: AN UPDATE

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Abstract

One year ago here at the NAB convention I authored a paper on the problems and costs of converting a broadcast facility to HDTV. It was a snapshot in time of the equipment costs. This year the snapshot was taken in late January just as two of the proponents have joined forces. This paper will review some of the unsolved problems and examine the relative costs of the proposed systems. You will not find a complete answer to the question 'how much?' because the systems' varied complexity still has a direct relationship to the total system cost.

Changes: What is New This Year

Last year's paper had a bottom line figure of 38 million dollars. It was a "worst case" figure, a first pass at projecting the cost until the FCC sets a transmission standard or further information becomes available from the ATV system proponents.

In the past year, we have almost no new information on the hardware for the transmission side of a future ATV system. We also have very little information on the black boxes needed to convert between formats.

Several companies have introduced new equipment for ATV production in the 1125/60 SMPTE 240M format but none qualify as a price breakthrough.

In January of 1989 there were 17 systems from 14 proponents; as of September 1989 there were 9 systems from 7 proponents.

What has not changed:

BANDWIDTH = CO$T.

Transmission Standard vs. Production Standard

The only current format for HDTV production is 1125/60. It is also the only format for which you can buy hardware and build a complete studio system today. If the FCC picks a transmission system that uses other than 1125/60 they will create a double conversion system. It will add equipment and it might look like Figure 1.

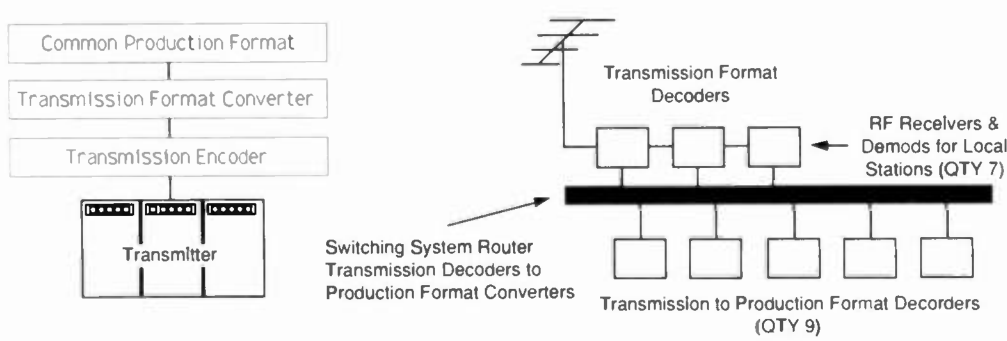


Figure 1

The double conversion process would need only one format converter on the input to the transmitter. On the receiving end, most stations have many receivers/decoders in TV's and VCR's that convert the RF NTSC to baseband video. This signal can be fed back into our system as needed. If the production standard is not the same as the transmission standard, either we will need a switching system to switch the transmission format into transmission/production format converters or many stand alone

bandwidth required, three signals, optics, image devices, and limited quantities. This adds up to similar problems in handling any of them in a broadcast plant.

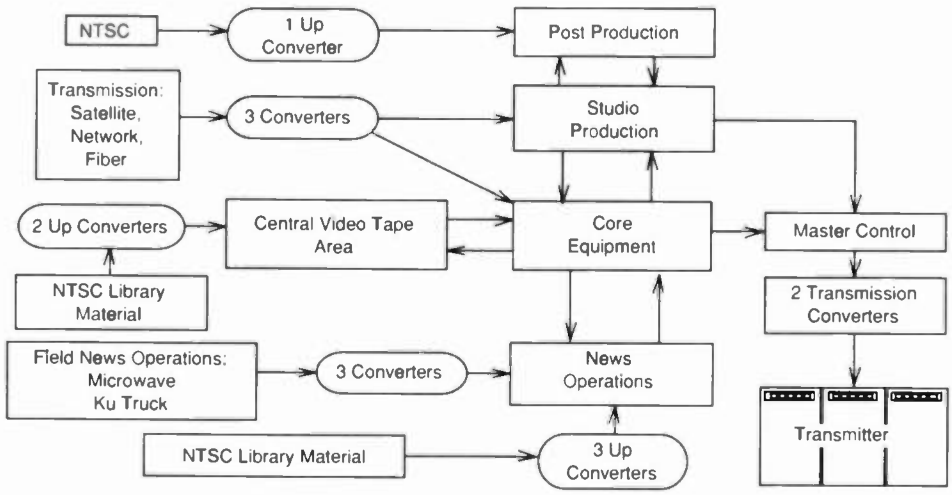


Figure 2

decoder/converters. Those converters are not "simple" black boxes and, for some proposed systems, would require conversion from 59.94 Hz vertical rate to a production standard at a 60.00 Hz frame rate.

No station wants to buy and maintain converters to convert to and from a production standard to the format needed for transmission. When the FCC sets the transmission standard they will, by default, create a production standard and my current costing assumes the production standard is the same as the transmission format. The cost and complexity of these standard converters is still a major unknown factor at this time.

Production Formats

The current ATV systems scheduled to be tested range from modified NTSC to full bandwidth HDTV. To simplify the costing we can break them down into two basic groups based on how easy they integrate into the current plant.

> HDTV comes in many formats but they all share two parameters: wide bandwidth (30 MHz) and three signal cables. Some of those systems include 1125/60, 2:1 - 1050/59.94 2:1, and 787/59.94, 1:1. Although only one format has equipment available they all share the same high costs because of the wide

> EDTV or modified NTSC are single cable systems with 10 MHz of bandwidth or less. This signal will pass through a typical broadcast plant with little or no modification.

Format Conversion

There will be several types of format converters in the ATV broadcast plant: library material up-converted to the production format, conversion from the satellite and microwave system to the production format, and material generated in other than 4:3.

Figure 2 shows some of the locations and expected quantity of converters that will be needed.

The NTSC up-converter would convert the existing library of NTSC material to the production/transmission format with a fixed aspect ratio.

Existing satellite and microwave systems will require converters.

At some point in the conversion process the aspect ratio difference between the many 4:3 NTSC receivers and the 16:9 receivers must be dealt with. That conversion process is more complex because it will need manpower. There is no computer system that can automate the Pan and Scan process. HDTV programming material could be shot for the 4:3 format and ignore the wide aspect ratio, or material can be prepared in post production for broadcast in the 4:3 format. Live programming may require a new job description, "Pan and Scan Specialist," to get the most from both formats.

General Issues in Converting to ATV

Signal Continuity Monitoring

If the cost of an ATV television set is $3,000, how many can we afford for signal continuity monitoring? There are over 75 color tv's in offices and various other locations around my station. When we converted from B&W to color you could at least use your existing B&W monitor to view for signal continuity and buy a new color TV for the General Manager's office and the lobby. The conversion to an ATV system, however, will need a new monitor or television at every location that needs to view the wide aspect ratio.

Equipment Wiring

Today, the only in-plant cabling system for implementing full HDTV is separate cables for the Red, Blue and Green signals. HDTV RBG calls for triple the number of wideband distribution amplifiers, three router levels, and three times the number of connections for each input and output termination.

Off-Air Monitoring

In most stations the in-house RF antenna system will not pass the new ATV signal. If your system modulates the cable from baseband you will need new modulators for the approved standard.

Core Equipment Area

The core equipment area contains the central signal router and its support equipment. If a local station originates in ATV and simulcasts NTSC, it may require most of the core equipment to be duplicated. Even a modest entry into ATV will call for quite a few new racks to house all the RGB distribution. Many plants will need major reworking of their central equipment areas. The electrical requirements for this new equipment must be provided for and that electricity will generate more heat and thus the need for more air conditioning.

Monitor Size

A control room monitor wall is either made up of standard EIA 19-inch wide racks or wood construction. If a new HDTV monitor slides into the existing hole, the horizontal dimension of the monitor will be about the same as the NTSC monitor it replaces. If the production staff sits at same distance from the monitor wall, the viewed image on the new HDTV monitor will appear smaller because of the wider 16:9 aspect ratio. Producer/Directors want to see larger images, not smaller ones. To install 30 or 40-inch wide racks to fit the larger monitors, you will need a larger control room or rebuild the interior to move the production staff closer to the central monitors and forego the increased viewing angle to the outer monitors.

Towers

FCC Working Party Two and the NAB have conducted separate surveys to determine the available tower space if a 2 channel ATV systen were to be approved. All the data has not been compiled as of this time but it appears that less than half of the towers can accommodate another full power antenna. Somewhere between 30% and 50% of those questioned could locate a second antenna or tower nearby. Knowing the windloading capacity of your tower will be important information for your station management.

Conversion Costs

Just as the proponents systems vary in complexity so do their conversion costs. To simplify the task I broke down the proponents into 5 groups with each group representing a major step in cost.

Group 1 - NTSC

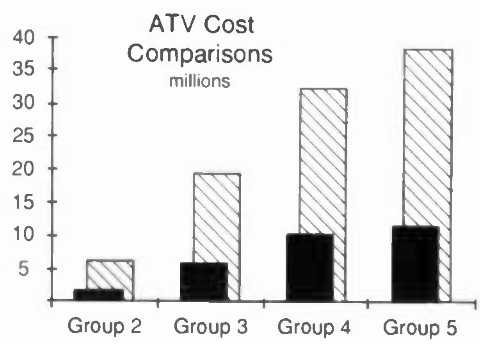
Group 2 - The production signal will pass through the core of today's broadcast plant requiring some modifications to tape equipment, cameras, and the transmitter exciter.

Group 3 - The production signal will pass through today's broadcast core equipment, but origination is in HDTV. It requires no modifications to the core plant or transmitter.

Group 4 - The production signal is HDTV but converted to fit into a modified transmitter.

Group 5 - The production signal is HDTV. The plant will require replacement to carry the production format. It needs a second transmitter for either augmentation or simulcast.

Equipment costs for NTSC and HDTV products used in this paper have been developed for use by the FCC Advisory Committee on Advanced Television Service, Systems Subcommittee, Working Party Three on Economic Assessment. The master spread sheet I have used for the estimation of conversion costs has changed greatly over the year. It now has over 130 types of equipment with NTSC prices and, if currently being built, the HDTV product cost. A multiplier is used to estimate costs for HDTV equipment not currently built. The sheet now has equipment quantities for stations in markets from 8 to



156. Many hours have been spent in collecting and estimating the numbers. But, as shown they are only a first pass at costing until the FCC sets a transmission standard or further information becomes available from the ATV system proponents. It is a snapshot of where we are today.

Because I have equipment quantities for several stations the chart shows by group the range of costs for full conversion. The solid bar is for a small station or a very small commitment and the hashed bar is the estimated cost for a full conversion.

Several assumptions have been made to create this chart:

It does not include costs for proponent transmission encoders, there is just not enough information today.

No installation has been provided for. Those costs will vary wildly because of location, unions, degree of system documentation, etc.

All prices are list price for quantity of one.

If a second RF system for augmentation or simulcast is necessary the needed spectrum will be available at whatever frequency or channel spacing is required. It is a full power UHF and will require a new tower, building and land to locate it (worst case).

The antenna will not need special bandwidth, linearity, etc.

The proponent system will be encoded at the studio and the signal will pass through a standard microwave STL system and not need a second microwave.

No operational or maintenance costs are included.

Pass the Network

From the chart above we can get an idea on what a full conversion might cost. But, conversion will happen over several years in a phase in process. The first step would be to pass along a network signal. The minimum equipment necessary to pass a network signal includes:

1. A second signal delivered to the studio from the network.

   (The costs of signal delivery to the studio may or might not be included as part of your current network agreement.)
2. Signal processing at the studio, minimum videotape and switching between ATV and NTSC.
3. Delivery to transmitter site.
4. Transmission over the air.
5. Minimum monitoring.

To pass the network for groups 2 or 3 requires on the order of $250,000 to $500,000 depending on equipment needed. Augmentation or simulcast - which add a second transmitter, antenna, and tower - causes the cost to go to as much as 8 million dollars.

Conclusion

These numbers do not make any assessment of the picture quality of the proposed systems. Some proponents systems are enhanced NTSC and are much simpler to implement than full HDTV based ones. There is still no inexpensive path to any ATV system: converting a major market station still costs almost 40 million dollars. The first in the market to convert will pay a premium. The necessity to triple-wire for RGB is very costly. Fiber or digital transmission must be developed to lower the installation manpower costs. The possibility of needing a complete new transmitter plant and very costly format converters all add to the expense of conversion.

THE PROPOSED SC-HDTV PROGRAM PRODUCTION STANDARD

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The Spectrum-Compatible High-Definition Television (SC-HDTV) System is designed for a progressively scanned source signal of 787.5 lines per frame, 59.94 frames per second and a horizontal line rate of exactly three times NTSC horizontal line rate.

Superficial comparison of the SC-HDTV line rate with the line rate of other proposed systems can easily lead to the conclusion that "787.5 lines is not as good as 1050 or 1125". This overlooks, however, the mode of scanning. An interlaced scanned picture loses vertical resolution when compared to a picture of the same number of lines per frame when progressively scanned. The amount of loss is expressed by the Kell factor which has a value that ranges between 0.6 and 0.7 for interlaced scan systems and 0.9 for progressive scan systems. A 1125-line 2:1 interlaced picture has no better vertical resolution than a 787.5 progressively scanned picture!

Scanning with a vertical rate equal to NTSC and a horizontal rate of three times NTSC is a part of the SC-HDTV System [1]. These rates still are considered to have the most desirable relation to the NTSC scanning rates for conversion from HDTV to NTSC for simulcast as well as for broadcasting a HDTV signal which minimizes interference into and from existing NTSC broadcasts.

The proposed standard includes a few changes with respect to those mentioned in [1] which are minor with the exception of the aspect ratio. The aspect ratio now has been set at 16:9. Another change is in the size of the pixels which now are "square." This expresses the fact that horizontal and vertical resolution in the display are equal. The previous pixel size was based on the CCIR Recommendation 601 which deviates slightly from square. The new square size as well as the 16:9 aspect ratio are based on a recently proposed image format, the Common Image Format (CIF). The CIF includes the 16:9 aspect ratio and a picture of 1080 lines and 1920 pixels per line.

Other changes include a longer retrace time for cameras, monitors and receivers and additional edge guard pixels at the studio to accommodate video filter edge effects.

The total active sample rate per frame is still slightly less than that of existing HDTV digital video recorders. This allows the use of existing technology to obtain the benefits of progressive scan.

The following Table lists the relevant parameters for the new SC-HDTV Program Production Standard under the column marked ZIF. The CIF column lists the parameters for the Common Image Format which are only a few . For comparison, a NTSC column is included. This column also lists (in parenthesis) a few parameters derived from CCIR Recommendation 601-1 "Encoding Parameters of Digital Television for Studios" (1986). CCIR 601-1 pertains to 525 and 625 line systems, thus provides a NTSC-digital bridge between NTSC-analog, and HDTV-digital.

The last line in the Table lists the number of video samples per field. It is interesting to compare this number to the one cited in SMPTE Standard 240M for digital tape recording: 993,600. The lower number in the ZIF column indicates the current technical feasibility of the proposed standard.

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Also:

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TABLE

Comparison of the Proposed SC-HDTV Program Production Standard (ZIF) to NTSC and to CIF

Parameter		NTSC (CCIR 601)	ZIF	CIF
Aspect Ratio, Camera and Emitted		4:3	16:9	16:9
Field Ratio (= 60 x 1000/1001)		59.94	59.94	
Line Rate (NTSC: f_H - 4.5/286 MHz				
ZIF: 3 x f_H)	kHz	15.734266	47.202797	
Total Lines Per Field		262.5	787.5	
Interlace		2:1	1:1	
Active Lines Per Field, Emitted		241.5	720	1080
Camera (ZIF: 720 + 11 Guard Space)		241.5	731	
Total Pixels Per Line		(858)	1596	
Active Pixels Per Line, Emitted		(720)	1280	1920
Camera (1280 + 20 Guard Space)		(720)	1300	
Useful Resolution (Kell Factor = .9)				
Vertical (ZIF: 0.9 x 720)	(Lines/Picture Height)	338	648	
Horizontal	(Lines/Picture Width)		1152	
	(Lines/Picture Height)	338 (443)	648	
Pixel Rate (ZIF: 2596 x 3 x f_H)	MHz	(13.5)	75.335664	
Nyquist Frequency	MHz	(6.75)	37.67	
Video Bandwidth (ZIF: 0.9 x Nyquist)	MHz	4.2 (5.5)	33.9	
Pixel Aspect (Width: Height)		8:9	1:1	1:1
Pixel Width Ratio to ZIF		(4:3)	1:1	2:3
Height Ratio to ZIF		(3:2)	1:1	2:3
Active Line Time, Receiver	Microsec	52.66	16.991	
Camera (Includes Guard Space)	Microsec	52.66	17.256	
Retrace Time Receiver	Microsec	10.9	4.195	
Camera	Microsec	10.9	3.929	
Video Samples Per Field (ZIF: 1300 x 731) (NTSC: 483 x 720)		(347,800)	950,300	

HDTV IMAGE COMPRESSION FOR REDUCING BANDWIDTH AND IMPROVING RECEIVED IMAGE QUALITY

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ABSTRACT

A HDTV image can be subdivided into a lower-resolution component and fidelity components. The lower resolution components can be scan converted and broadcast over a conventional television channel. A method has been developed to transmit most of the visible high-resolution information in real time over a 6 MHz augmentation channel. The remaining detail is transferred and displayed before the observer can detect it's omission. The high definition image can therefore, be reconstructed in full detail without visible temporal degradation. Image transmission redundancies are utilized to enhance the image quality by improving the signal to noise ratio. These techniques may also be applied to PAL or SECAM transmissions, and many of the improved definition transmission schemes.

INTRODUCTION

The New York Institute of Technology (NYIT) has been developing methods to reduce the bandwidth needed to broadcast HDTV by augmenting the image information contained in a conventional terrestrial broadcast channel. Supplemental information is intended to be placed on an independent channel of comparable bandwidth but with sufficiently different characteristics so as to minimize co-channel and taboo channel interference. During the past year we have concentrated on developing an approach which has attributes of our previous ideas,[1,2,3] with further concepts based upon psycophysical studies. Processing techinques which heretofore have been impractical can now be implemented using state-of-the-art digital signal processing elements.

This paper describes an improved transmission technique which allows most of the perceptual HDTV image fidelity to be broadcast, received and displayed in real time over a conventional NTSC broadcast channel and an augmentation channel. The NTSC channel will provide improved diagonal resolution (though this may be of little value to the viewer of a conventional television) and include encoded information which will allow significant improvements in the signal-to-noise ratio of stationary portions of the picture. The augmentation channel is used to pass the luminance fidelity and the R-Y color detail components. Data compression is used to reduce the augmentation channel bandwidth by removing unneeded spatial frequencies and adaptively adjusting the transmission time of high spatial frequencies to the needs of the human vision system. Additional bandwidth reduction can be achieved by statistically multiplexing two or more stations augmentation signals.

SAMPLE RATE VS. LINE RESOLUTION

Television engineers commonly refer to the quality of an image in terms of the number of unique video lines displayed on a screen. Vertical information on a video screen must be described in terms of discrete samples because raster scanning separates video into unique lines. The highest vertical frequency which can be unambiguously resolved by raster scanning systems is half the sample rate (242 cycles per picture height for NTSC). The apparent contradiction in resolution (484 lines of viewable video vs. 242 cycles) can be explained by noting that two lines can be displayed per cycle, one at it's peak and the other at its trough. Resolution described in terms of lines per picture height actually refers to spatial frequencies at half the rate.

Precautions must be taken to insure frequencies above the NTSC resolution of 484 lines per picture height (l/ph) are removed before they are sampled by the camera to avoid aliasing. If a sharp filter were used to abruptly remove frequencies at and above 484 l/ph, unacceptable ringing would occur on the image. High frequency vertical signals present an even more serious problem for interlaced systems. A single horizontal line is displayed in only one field and results in an obvious and annoying flicker. It is easy to verify that some of the new CCD cameras have resolutions exceeding that of NTSC by simply observing the flicker on horizontal edges of an image. Limiting vertical resolution to about 330 l/ph reduces flicker to levels which are not perceptible and allows practical filters to be designed which minimize ringing.

The limiting vertical resolution for SMPTE 240M, having 1036 active video lines (2*(517 lines/field+ 1-half line/field)), should be approximately 1036*330/484=706 l/ph. Interlacing further reduces the vertical sample rate for areas where motion is present to that of one field, i.e. 518 l/ph. Although it is possible to adaptively process interlaced video in moving areas on the screen, the best processing results can be obtained using a progressively scanned HDTV camera with a limiting resolution of 706 l/ph and then providing scan conversion in the processing circuitry.

The sampling operation resulting from raster scanning defines the vertical image characteristics but is not applicable to horizontal components since they are passed in an analog format. It has been found that the human visual system exhibits the same degree of visual acuity along the horizontal axis as it does along the vertical (note that we specify resolution in terms of lines per picture height, in order to provide the same line density for any angle of image rotation, regardless of the display's aspect ratio). To provide a natural looking picture having the same degree of clarity horizontally as it does vertically, the horizontal video response should also be limited to 706 l/ph.

EXAMINATION OF THE 2D FREQUENCY SPECTRUM

The two dimensional Fourier transform of a test pattern composed of 700 horizontal bars shaded to provide a sinusoidal waveshape results in two impulses on the vertical frequency axis at +/- 700 l/ph. Rotating a spatial image about an axis perpendicular to the image causes the 2-dimensional Fourier transform of the image to be rotated about the origin by the same amount.[5] If the test pattern is rotated by 180 degrees, the 700 l/ph impulse traces a circle in the 2D frequency domain as shown in FIGURE 1.

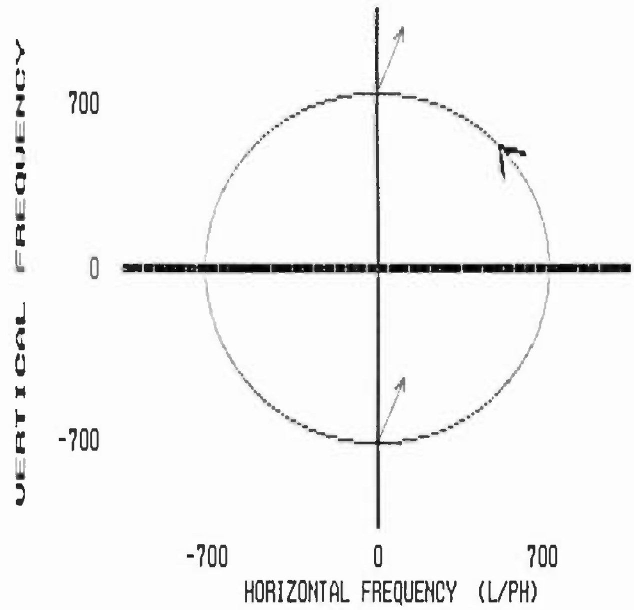

FIGURE 1. Path traced in the 2-dimensional spectrum when a 700 l/ph image is rotated about its axis.

The spectrum of the camera video can be obtained by performing a discrete Fourier transform (DFT) on its raster scanned output. To obtain the same spatial resolution horizontally, it is necessary to use the same sampling frequency of 1036 l/ph. A square image composed of an array of 1036 by 1036 cardinally sampled pixels is first considered. The concepts are then extended to the entire HDTV image area.

FIGURE 2 shows the discrete spectrum space of the detail portion of a high definition image. A circle has been inscribed at 700 l/ph to show the limiting resolution to which the camera and optical system is designed. A second circle drawn at 1036 l/ph is used to indicate the horizontal and vertical Nyquist limit. All the frequency information necessary to describe any image with 700 lines or less resolution positioned at any angle about the axis of the image plane will be contained within the inner circle. We should consider all frequency components up to 1036 l/ph due to practical filter limitations. The area of the discrete spectrum space outside the 1036 l/ph circle encompases (4-Π)/4 =22% of the total. It can be shown that the transformation on an N by N spatial array results in N by N unique frequency components. If the frequency components broadcast are restricted to be within the Nyquist limit circle, a compression of 22% is obtained.

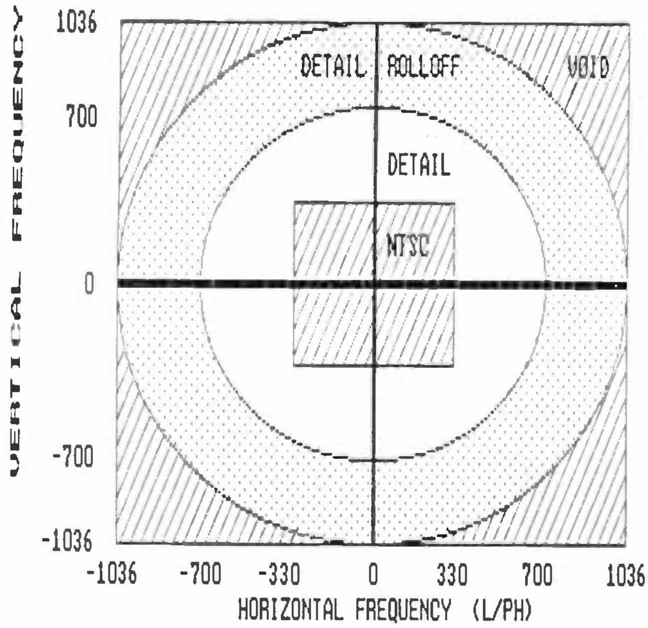

FIGURE 2. HDTV spectrum space with NTSC components removed.

The HDTV image can be passed through a 2D complementary spatial filter to separate frequency components below 330 l/ph both horizontally and vertically. The low passed output is then formatted to obtain NTSC video. Notice that the diagonal detail in the image will be 40% greater than from cameras having a resolution of 330 l/ph and free of spherical aberations as shown in FIGURE 3. An additional 10% of the HDTV frequency components are not needed by the augmentation channel since they are already present in the NTSC channel.

The amount of augmentation information necessary to send HDTV signals can be reduced by one third prior to using compression techniques by sending image frequency components instead of actual images. The reduction causes no degradation whatsoever to the reconstructed image since we merely ignored frequency components which never existed in the detail spectrum to start with.

Psycophysical studies performed at NYIT and elsewhere[4,10,11] indicate an oblique effect exists in vision which results from the eye being less sensitive to detail along diagonal lines or diagonal portions of the image. We are continuing to investigate this possibility which could significantly reduce the necessary spectral components without any perceptual degradation in image quality.

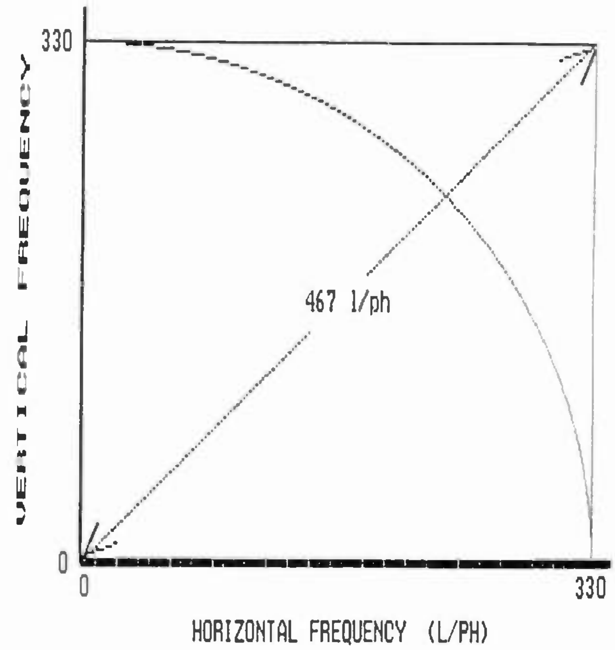

FIGURE 3. The low frequency components of HDTV video can provide 40% greater diagonal without any perceptual degredation in image quality.

Transform compression operates on images, which are highly correlated in the spatial domain and transforms them into a different domain having de-correlated components. Natural images tend to have most of their spectral energy concentrated into the lower frequency components. The higher frequency components for most images are of such a low value that some of them may be omitted or compressed without significantly altering the image when an inverse transform is performed.[8]

One measure of quality of an image transmission scheme is the mean squared error of the final image relative to the original.[5]

$$E^2 = \sum_{i=0}^{N-1} \sum_{j=0}^{N-1} [R(i,j)-P(i,j)]^2 \qquad (1)$$

where:

$R(i,j)$ = pixels of the reference image
$P(i,j)$ = pixels of the processed image

The Fourier transform/inverse transform pair is among a set of functions which produce the least error by this definition. It is not the only transform exhibiting the least squared error result. The Walsh-Haddamard, slant, sine, cosine, Haar, and

Karhunen-Loeve are other examples of transforms exhibiting the same property.

Processing techinques which recover the original picture information are called restoring algorithms. If we are willing to accept some small amount of error in the received signal, the transmission bandwidth may be further reduced by using non- restoring algorithms. Such algorithms can provide excellent subjective results by removing information which can not be detected by the eye. Removing frequency components containing the least energy will cause the least degradation to the image by the criteria established in equation 1. The most desirable choice of transform technique will be the one which best de-correlates the spatial image, thereby causing the power spectral density to fall off the fastest with frequency. It has been shown that the Karhunen-Loeve is the optimum in this respect and surpasses the Fourier transform. Unfortunately there is no fast algorithm to perform this transform. Ahmed, Natarajan and Rao[6] showed that the cosine transform closely matches the performance obtanined by the Karhunen-Loeve transform. The discrete cosine transform (DCT) for N spatial samples can be obtained from a modified N point fast Fourier transform (FFT)[7] and exhibits the same properties as FFT. Subsequent discussions will be based upon discrete Fourier transforms (DFTs) but apply equally well to DCTs.

FREQUENCY COMPONENT SPREADING

The sampling frequency f_S is rarely a multiple of the frequencies in the image being sampled. Non-integral sampling results in the energy in the spatial domain signal being spread out over all the components in the frequency domain and is referred to as leakage. The spreading characteristics can be determined by recognizing any sine wave placed in a gating window of N units length will have a Fourier spectrum composed of an impulse at the sinusoidal frequency convolved with the sin(Nx)/x spectrum of the gating window. If this signal is sampled at unit intervals, the convolution of the original time signal with the sampler produces frequency components only at integral multiples of the $2*\Pi/N$. The gate window response has nulls placed at intervals of $2*\Pi/N$ except for the main lobe. When an input sinusoid which is a multiple of the base frequency $2*\Pi/N$ is applied, the nulls of its composite response will coincide with all the discrete frequency components except for the main lobe component and no spreading occurs. The worst case spreading occurs when the input frequency is midway between two discrete frequency components, (i.e. $f_{worstSpread} = (2*n+1)*\Pi/N$), because the peaks of the sidelobes will coincide with the discrete frequency components.

FIGURES 4A and 4B show the worst-case sprectrum spreading which could occur from sampling windows of 16 and 64 points. It is apparent that the normalized width of the smearing is inversely proportional to the number of samples. When a reasonable number of samples are used, the leakage energy falls off rapidly with frequency, as shown in Table 1. Weighting functions are commonly used to reduce spreading, in sampled data systems by applying a symmetrical taper to data within the input window, however, this is clearly not acceptable for video processing since it would result in very obvious changes in image intensity near the edges of the spatial window.

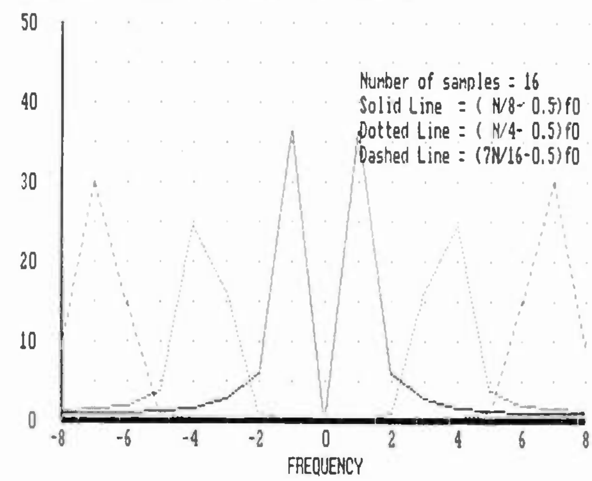

FIGURE 4A. Worst-case spectral spreading for a small sample space.

A one dimensional spatial image, and it's Fourier transform, will be used to examine the effect of removing selected spectral components. Suppose an image is composed of (700+1/2) horizontal lines in the picture height (a worst case spectral spreading frequency). If we then truncate the frequency spectrum above 701 l/ph and want to determine how this affect the spatial image. The frequency domain truncation could also be obtained by adding another set of spectral components between 702 and 1036 l/ph of equal amplitude and everwhere 180 degrees out of phase with the original components. Fourier transformation is a linear and distributive process, therefore, addition of the spectral cancellation elements in the frequency domain results in the superposition of the original image with an error image containing only high frequency components. It is also apparent that an error image created by removing low frequency terms will be composed exclusively of low frequencies. The effects of truncating a spread frequency spectrum will be negligible if sufficiently large video windows are used and the input video has a reasonable roll-off.

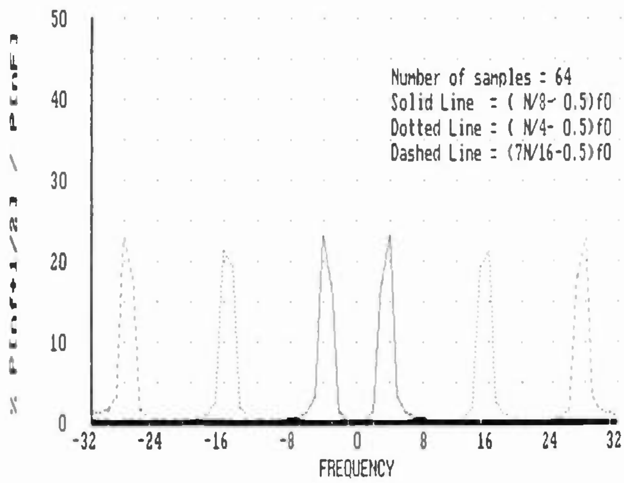

FIGURE 4B. Worst-case spectral spreading for a larger sample space.

Table 1 indicates that most of the energy in a non-integral frequency component is found in the adjacent integral frequencies. The spectral response of a Fourier transform can therefore be considered as a series of bandpass filters, each approximately 1/(NumberOfSamples) 1/ph wide. A transform performed on a large number of points can be subdivided into K smaller transformed each operating on 1/K th the data. The results obtained from the smaller transforms are similar to the original except the frequency resolution is degraded by a factor of K. When smaller windows are used, it is possible to reduce the effects of spatial errors generated when frequency spread spectral components are truncated by passing the processed image through a spatial filter designed to attinuate out-of-band frequency components.

TEMPORAL, SPATIAL AND REDUNDANCY TRADE-OFFS

The response of the human vision system has been the subject of much research. It is an established fact that the eye is not as sensitive to detail in moving portions of a scene (see Glenn[1,2] and Seyler[14]). In other words, the visual time constant for various levels of image fidelity increases with increasing fidelity. This characteristic can be used to an advantage in further reducing the bandwidth of the augmentation channel. All the relevant detail frequency components needed to reconstruct a high definition image are broadcast, however they don't all have to be sent at the 30 frames per second update rate used to convey low resolution motion. The conventional broadcast channel is used to send the lower resolution spatial video signal in real time, freeing the augmentation channel to send detail frequency components at rates which closely match that of the vision system. In areas of the screen where motion is detected, frequency bands containing the highest energy content are selected to be output first in order to minimize the mean square error. When motion has ceased, the remaining frequency detail is added in subsequent frames until the entire image is reconstructed. This technique therefore allows the major components of the detail spectrum to be sent in real time while the complete picture is reconstructed before any image degradation can be observed.

TABLE I.

Percent of spectral energy as a function of bandwidth for worst-case spread frequency of N/4-0.5 1/ph.

Number of Samples	8	16	64	256
Bandwidth (centered at N/4-0.5)				
1/N	0.83	0.82	0.81	0.81
3/N	0.95	0.92	0.90	0.91
5/N	-	0.96	0.94	0.93
7/N	-	0.99	0.95	0.95

The spatial detail can either be processed as a whole or subdivided into arrays of smaller image windows called tiles. There are several advantages of working with tiles instead of operating on an entire image. Performing real-time transforms (or their inverse) on large spatial arrays involves a large number of computations which cannot be carried out in real time, however, the computations necessary to process smaller tile sizes can be performed in

VLSI integrated circuits. Processing smaller areas of the screen also allows for greater freedom in adaptive processing. Areas containing motion can be separated from stationary areas and treated differently.

Additional reductions in the augmentation channel bandwidth can be realized by sharing an augmentation channel among two or more broadcasters. Since programing content is statistically independent, tile update information can be sent for the portion of each program containing motion while sacrificing redundant transmissions on the stationary areas of other programs which would have otherwise occurred. At times motion in two or more of the programs will cause reduction in the full quality high definition augmentation rate, however augmentation information is used to improve the image fidelity but is non-essential. It is very difficult for an observer to determine if detail on certain image tiles is occasionally degraded, since the worst possible degradation merely reduces the image quality to NTSC video-half that of HDTV horizontally and vertically.

Although this paper addresses the needs of terrestrial broadcasters, it should be mentioned that the statistical program characteristic/frame tile redundancy tradeoff can be applied to provide even greater compression ratios for cable and satellite transmissions where multiple stations could share the same augmentation channel(s).

THE EFFECTS OF NOISE

Although the area covered by the NTSC components in FIGURE 2 is only 10% of the total spectral space, the energy contained therein is at least an order of magnitude greater than the remaining spectrum of natural images because of the decorrolation properties of the transform. A gain must be applied to the augmentation components prior to broadcasting if the average power in the conventional and augmentation channel are to be the same. Assuming that each of the channels contain the same noise power, the signal to noise ratio in the augmentation channel will be improved by the gain used (typically 10 to 20 db). Since little is achieved by sending a detail component which exceeds the quality of the base signal, the power in the augmentation channel can be reduced thereby substantially reducing cross and co-channel interference. Additional reduction in co-channel and cross channel interference occurs because of the difference in the characteristics of the data transmitted.

When the received frequency domain signal is transformed back into the spatial domain, noise de-correlates and tends to average out over the spatial domain. A post processing spatial filter operating on the spatial detail signal will further attenuate the effects of noise. Image degradation due to ghosting will also be reduced in this manner. Detail components passed in an adaptive format do not have to be broadcast with a constant time delay relative to the beginning of the frame. The ghost response will be different for every frame and will take on the characteristics of random noise in the temporal image display.

Our transform coding scheme operates on images subdivided into smaller window tiles. Tile status information, which is included in the information broadcast can be utilized to determine if motion has occurred on each tile. When no motion is present, the NTSC portion of the image can be improved by integrating the signal received for each pixel within the tile area. The improvement can be shown to be on the order of $M_{1/2}$ where M is the number of frames over which the integration occured. The frame to frame spatial redundancy is therefore used to advantage in improving the signal to noise ratio by typically 5 to 10 db. Most detail components will also be sent redundantly and therefore can benefit from the same type of processing. If tile motion information is included in the NTSC signal, an improved television receiver can be produced which will provide higher quality pictures in fringe areas by using tile integration filtering.

PRACTICAL IMPLEMENTATION OF TRANSFORMS IN HARDWARE

The most time consuming portion of performing discrete Fourier transforms (DFTs) is involved with the complex multiplications necessary. A normal DFT requires N^2 multiplications for a N length sample set. The fast Fourier transforms (FFT) can reduce the number to approximately $N \log_2 N$. Two dimensional transforms require the compilation of $2N^2 \log_2 N$ multiplies to process a N by N spatial sample. 2048 by 1024 high resolution video operating at 30 frames per second would require approximately 14 billion complex multiplies each second. It is not now economically feasible to include this level of processing power in consumer television receivers, however if such an image were sub-divided into arrays of tiles, each tile containing an 8 by 8 sample window, the multiply rate is reduced to a mere 90 million per second. Two manufacturers have introduced cosine transform/inverse transform integrated circuit processors which approach these rates. It is rumored that 16 by 16 processors are being developed. These recent technical advances thus now make it feasible to consider bandwidth reduction using frequency domain processing.

Frequency spreading effect becomes significant on very small sample windows and some of the advantages gained by eliminating out of band components can no longer be realized. Post processing spatial filters can be used to improve the resulting image to acceptable levels. It is not as easy to detect errors on smaller tiles however. Researchers involved with generalized forms of image compression[5] indicate that minimal improvement is realized by going to tile sizes larger than about 16 by 16 and subjectively little is gained for tiles larger than 4 by 4.

CONCLUSION

The FCC will require any new terrestrial HDTV broadcast to offer a NTSC compatible version as well. The transmission bandwidth needed HDTV can be substantially reduced by utilizing this channel in the HDTV broadcast. HDTV video can be subdivided into the NTSC quality video and the higher fidelity components. The augmentation channel can then send the spectral fidelity elements at rates which satisfy the needs of the human vision system. Detail components are obtained for small areas of the image called tiles. The sequence chosen for broadcasting tile components is determined as a function of motion within the tile and individual component contribution to picture quality. Frequency bands containing the highest energy components are chosen to be sent first (in real time) for tiles in which motion is occuring thus minimizing the mean square error in the received signal. When the motion ceases, the remaining components are sent before the eye can detect any degradation in image fidelity. Even more compression is possible by combining the augmentation signals from several sources and statistically multiplexing their motion components. Satellite broadcasting could especially benefit from this technique since only a few channels would be sacrificed to provide high definition for all the remaining signals.

Some control information required for specifying the augmentation frequency bands can be used by NTSC (only) receivers to improve the signal to noise ratio in stationary regions of the picture. Diagonal resolution of NTSC picture is also increased by up to 40%.

Although some of the techniques described above were not practical in the past, state of the art components are now available to provide appropriate processing in real time. We will be able to provide full high definition transmission over broadcast channels of reasonable bandwidths and yet, still provide NTSC compatibility.

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THE COMMON IMAGE FORMAT AND THE COMMON DATA RATE APPROACHES TO HDTV STUDIO STANDARDS— A EUROPEAN VIEW

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ABSTRACT

Although everyone agrees that a single world-wide standard for HDTV studio production is highly desirable, discussions in the CCIR on what parameters to adopt are currently stalemated owing to a difference of opinion between PAL/SECAM countries and NTSC countries on the HDTV field-rate to be adopted.

Two ways forward have been identified, known as Common Image Format (CIF) and Common Data Rate (CDR) respectively. Both of these approaches envisage that 50 Hz and 60/59.9 Hz HDTV standartds will co-exist in the first instance, but that the standards adopted for these would be such as to allow a progression to a single HDTV standard at a later stage. This paper outlines the CIF and CDR concepts and philosophies, gives an account of some recent studies from around the world, and draws some provisional conclusions. Particular emphasis is given to the impact on the economics of HDTV studio production and equipment (including DVTRs, CCD cameras, displays, special effects, and standards conversion).

Following discussion of the broadcasting issues, the possible relationship between HDTV broadcast and non-broadcast applications is reviewed with a view to setting the scene for future studies in these areas. The topics reviewed include applications in printing, non-broadcast computer graphics applications, cinematography, and the relationship to possible hierarchies of service in the ISDN and BISDN and to the transport mechanisms (for example, Asynchronous Transport Mode).

While differing views on the above topics are noted, the main studies drawn upon are European in origin, and the overall perspective portrayed is broadly that of the European Eureka 95 HDTV project. The paper is intended as a platform for further debate on these important HDTV issues.

1. INTRODUCTION

At the Extraordinary Meeting in Geneva in May 1989, CCIR Study Group 11 reaffirmed that "a unique studio standard is a clearly identified need and must continue to be the main target". The administrations of the European Community fully support that goal, and consider that the 1250/50 1:1 system best meets this target.

In report XE/11, agreed at the same meeting[1], Study Group 11 defined approaches to a single worldwide HDTV standard, and among them the concept of a two-step approach. Two main paths have been identified, namely the Common Data Rate and the Common Image Format. This paper analyses the relative merits of these two approaches.

The Common Data Rate (CDR) assumes that HDTV standards would initially be adopted based on 50 Hz and 59.94 Hz field rates related to current emission standards, having a maximum of commonality in other parameters such as line frequency, based on the principles of SMPTE RP 125, leading to a common data rate. The concept of the HDTV delivery environment is illustrated in Fig. 1. Central to any approach which commences with more than one frame rate

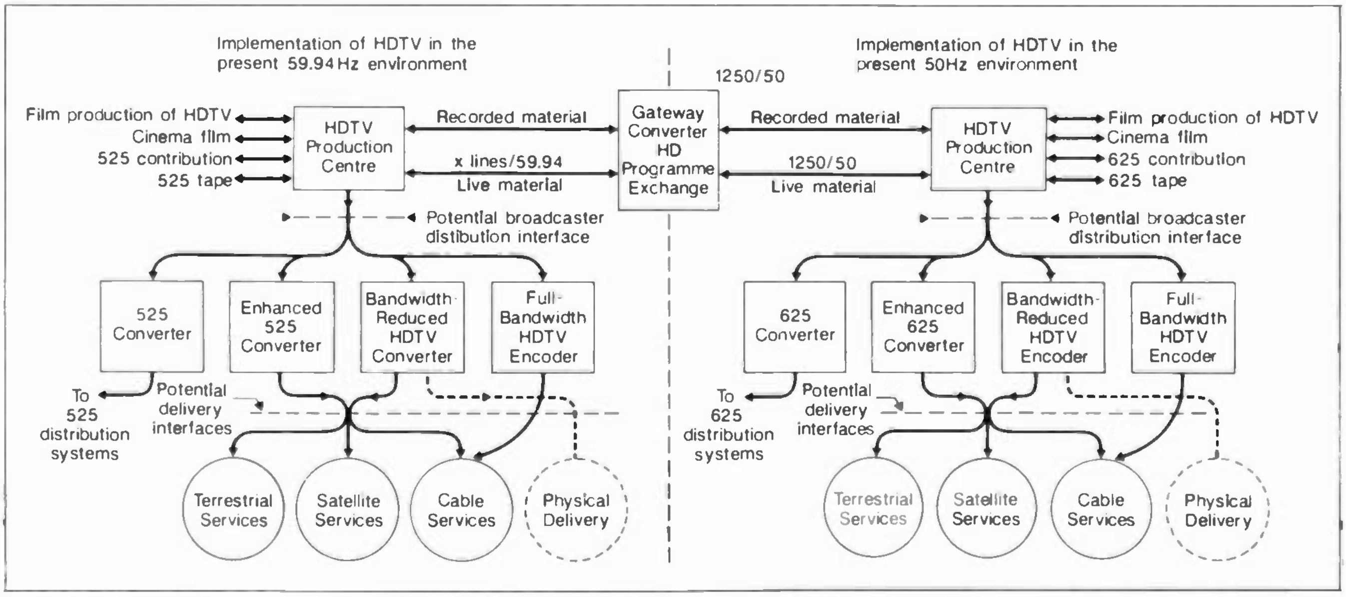

Fig. 1 Implementations of common image format at 30 and 25 frames per second using a possible set of parameters.

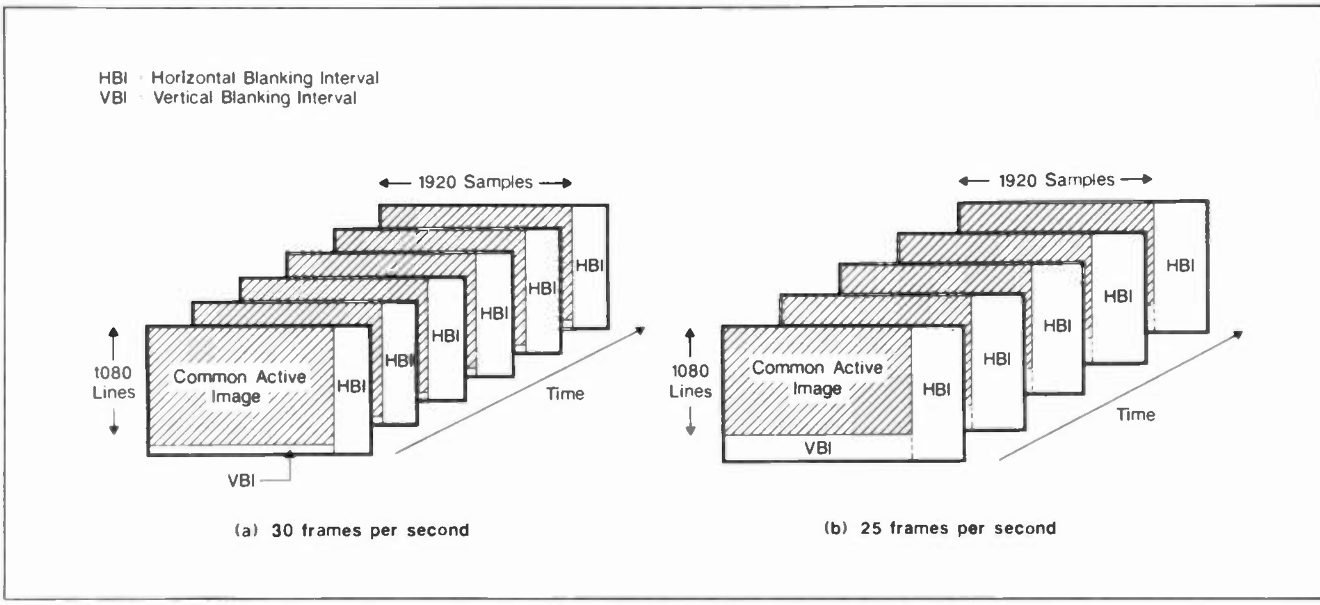

Fig. 2 An example of the common data rate approach.

is the need for high quality conversion as a gateway between the different HDTV standards.

The concept of the Common Image Format (CIF) is based on the premise that the development of a single standard is presently constrained by the desire for compatibility with the frame rates of existing television systems, but that this constraint is likely eventually to be removed by developments in technology leading to the setting of a standard based on a higher frame rate for better motion rendition, and easier conversion to current frame rates. This involves defining common parameters for the image area; for example, aspect ratio of the image and of individual samples, and the number of samples in both the horizontal and vertical directions. There would be complete freedom to choose the scanning method, frame rate and size of blanking intervals for individual HDTV distribution formats (see Fig. 2).

Since the Extraordinary Meeting, Administrations have continued their studies on these topics. Taking account of relevant material in agreed CCIR reports[1,2], and of various recent submissions to CCIR study groups, this paper reviews the relative advantages of the two approaches from several points of view:

(i) A simple relationship to proposed emission and display standards

(ii) Relationship to SMPTE RP 125 and Existing Studio Facilities

(iii) Field Rate Standards Conversion

(iv) Digital Recording

(v) CCD Cameras

(vi) CRT Displays

(vii) LCD Displays

(viii) Relationship to Computer Graphics for Non-Broadcast Applications

(ix) Electronic Graphics and Digital Effects

(x) Electronic Cinematography

(xi) Relationship to the ISDN

(xii) Conversion to-and-from Other Media

(xiii) Combined CDR and CIF Proposals

Following this review some conclusions are drawn as to the balance of the argument between the two approaches on the evidence accumulated to date.

2. THE REVIEW

2.1 The Proposed Emission and Display Standards

a) *The Terrestrial Broadcasting* issues are closely related to the need for compatability with existing terrestrial infrastructures.

In the USA, for example, of the 18 proposals for ATV systems which comply with the FCC Tentative Decision of 1st September 1988, three are based upon the simulcast concept (one NTSC channel + one incompatible ATV channel) and 15 are based upon the concepts of direct NTSC compatibility or NTSC with ATV augmentation. Elsewhere, attention is focused on the use of an AM/VSB version of the HD-MAC system developed for satellite broadcasting in cable networks and for future applications in terrestrial broadcasting.

Bearing in mind the need *in all cases* to display pictures using current standards on HDTV screens, it would seem clear that it would be advantageous to maintain simple compatibility with the 525/59.94 structure in 60 Hz countries and with the 625/50 structure in 50 Hz countries. This is the case for the HDTV display standards and, in almost all approaches, for HDTV emission.

b) *The Satellite Broadcasting* issue is more complex. As for the terrestrial case, most of the Administrations in Western Europe support the concept that there must be compatibility with the existing broadcasting infrastructure. More specifically, they support the HD-MAC system being developed under the Eureka 95 project. It would seem that the situation in the USA is also such as to constrain HDTV emission of all types to be compatible with the existing terrestrial infrastructure and hence with the 525 line/59.94 system.

The situation elsewhere is less clear, although it could be expected that the arguments for HDTV emission standards to be compatible with the existing infrastructure already accepted by many in the CCIR would prevail in most situations. There is, however, a view expressed by a few Administrations that in the satellite case certain advantages can be gained from the use of a non-compatible emission standard.

Whilst it is recognised that in the longer term advances in digital processing may allow digital emission of HDTV and so reduce the strict requirements for simple compatibility, the problems of economic conversion of HDTV to the existing infrastructure will still need to be addressed.

To summarise: for many administrations, close compatibility with the existing broadcasting infrastructure is the dominant consideration in selecting an HDTV emission format. For satellite or terrestrial broadcasting in its simplest form, this implies *emission standards* based upon 1050/59.94/16:9 in 60 Hz countries and 1250/50/16:9 in 50 Hz countries. It also implies HDTV *display standards* which are closely and simply related to the emission standards and are thus different in 60 Hz and 50 Hz countries.

It should be noted that in their simplest forms these proposed emission and display standards would bear very simple relationships to SMPTE RP 125.

2.2 The Relationship to SMPTE RP 125, Existing Studio Standards and Proposed Emission Standards

Proposed CDR and CIF approaches to the HDTV studio standard have been discussed in some depth[3], and it has been demonstrated that a particular set* of parameters from within the CDR approach arose naturally from considerations based on SMPTE RP 125, and would therefore lead to very simple relationships to proposed HDTV emission standards.

By contrast, in general, the CIF approach diverges from simple relationships to SMPTE RP 125 and hence to proposed HDTV emission standards. This would make some aspects of systems and equipment design and digital processing more difficult, and could lead to a degradation in emission quality.

The possibility of longer-term advantages of a CIF in terms of CCD cameras and LCD displays are discussed further below.

On the studio side there will be a real practical need for the inclusion of picture material originated at 525/59.94 and 625/50 into HDTV productions and vice versa. A simple relation between HDTV and present definition television, as provided by the CDR approach, will make these conversions more straightforward[4].

2.3 Field-rate Standards Conversion

The key issue here is that of achieving good motion compensation performance. To this end (and for other purposes) the European proposal is for a 1250/50 progressively scanned single world-wide HDTV studio standard from which the appropriate 1250 and 1050 interlaced standards could be derived. For international exchange purposes, it is judged that the advantages of progressive scanning would be achievable within the (1250/1050) CDR approach. With the CIF approach, however, progressive scanning has not been seriously addressed in proposals to date, and there is some risk that the higher data rates associated with the 59.94 Hz frame rate would rule out progressive scanning at that rate. If this proved so there would be an additional loss of picture quality and the need for interlace-to-progressive conversion in the field-rate converter.

Assuming a progressively scanned input signal, the CIF approach would allow the elimination of line-standards conversion in the international programme exchange field-rate converter. This must be offset, however, by the need for line standards conversion at all points downstream where compatible transmission or emission is required.

In summary, the balance of argument in this case, mainly owing to the progressive scanning requirement and to the need for downstream conversions in the CIF approach, is in favour of CDR.

2.4 Digital Recording

Following worldwide agreement on the standards to be adopted for use in the digital studio (CCIR Rec. 601/SMPTE RP 125), intensive international work by the SMPTE, the Magnum Group* and the CCIR led to world-wide agreement on the standards to be used for digital television tape recording in 1986 (CCIR Rec. 657/SMPTE 224M-228M). Recognising the operational and economic advantages that a worldwide compatible digital format would realise, the technical and operational criteria to be used in recording the 4:2:2 member of the family were defined, laying the foundation for the development of the D1 recorder — a digital recorder which can be switched to operate on either of the 525/59.94 or 625/50 studio standards. The guiding principle for this important achievement is the use of a common data rate approach based upon SMPTE RP 125/CCIR Rec. 601.

It has been further argued that the principles established for digital recording of the 4:2:2 member of the SMPTE RP 125 family should apply equally to the digital recording of *component HDTV* signals[5].

More specifically, it is argued that the elegant D1 concept of grouping equal numbers of 525-lines and 625-lines, so as to form 5 segments and 6 segments of each respectively, allows common tape transport and data write/recovery systems to be used.

Additionally, the use of a common recorded data rate and sampling frequency means that the bulk of the digital processing in the machine is common. This could provide the basis for reduced development and manufacturing costs and should overcome the problems of delayed introduction of novel features in either of the 50 Hz or 60 Hz markets.

By contrast, it has been demonstrated[5] that the degree of commonality obtainable with the CIF approach is significantly less than with CDR, and that the main effect of adopting such an approach will be the need to develop and manufacture separate machines for the 50 Hz and 59.94 markets. This will result in different trade-offs of critical design parameters such as recorded wavelength, track-width and tape consumption, etc.

* 1250/50/16:9 and 1050/59.94/16:9 in either a common orthogonal or a quincunx sampling structure.

* A joint EBU and world-wide Tape Recorder Manufacturers Group set up to establish a single format for the 4:2:2 DVTR.

In summary: it seems clear that in considering, in isolation, the factors relevant to the manufacturing cost and the performance of HD-DVTRs, it is difficult to identify any rationale to support adoption of the CIF approach.

By contrast, the CDR approach offers the general benefits of commonality of equipment without any identified penality.

Bearing in mind the central importance of digital recording in the HDTV studio operation, it is believed that these provisional conclusions should carry due weight in the overall balance of the discussion in the CCIR.

2.5 CCD Cameras

CCD cameras are widely used in the home video market and in professional broadcasting for ENG. Although developed originally for the consumer market (which is the only market to justify the development effort needed), performance is now improving rapidly and it is to be expected that CCD cameras will eventually replace most of the tube cameras in use today. There is also considerable confidence among television camera manufacturers that high-definition CCD cameras will be developed in the longer-term, and it has been suggested that in this future scenario the CIF approach would produce significant economic benefits in the design and manufacture of HD-CCD cameras.

This suggestion has been studied by industry in West Germany and The Netherlands[6]. These studies have noted that within the CIF approach both the CCD and the camera will have to be able to operate on the two field rates of 59.94 Hz and 50 Hz, using different timing and synchronising arrangements. Thus the camera circuit boards, including such processes as digital contour enhancers and matrixing, will have to work on two standards with considerably different clock rates and timing.

Furthermore, it should be noted that, for 4:2:2 digital TV, several CCD matrixes produced by manufacturers use different numbers of pixels per line. This arises because there is an intermediate analogue process for filtering and control purposes which means that there is no need for a close correlation between CCD line pixels and the 4:2:2 sampling structure. Thus the argument that the adoption of CIF will lead to a single CCD matrix for HDTV is, to say the least, rather questionable. Moreover, it should be noted that even in the CIF approach different CCDs limited to the emission formats to be used in 50 Hz and 60 Hz countries would need to be developed for the consumer market.

On the other hand, the CDR approach requires the manufacture of two CCD sensors for studio cameras, one for each line format. This requirement might be overcome, for example, by the use of anamorphic lenses to apply correction in the vertical direction.

Thus, on balance, neither the CIF nor the CDR approach has a significant advantage.

2.6 CRT Displays

CRT HDTV monitors used in the HDTV studio will fall into two catagories:

(i) Very high quality monitors specifically designed for HDTV studio use

(ii) Lower quality HDTV monitors derived from consumer product designs

In a *CIF approach*, there would either be different monitor designs for 50 Hz and 60 Hz versions of the category (i) monitor, or considerable additional complexity would be needed in a monitor able to work on both field-rates.

Furthermore, it is unlikely that category (ii) monitors could easily be based upon a relatively cheap consumer product.

In a *CDR approach*, category (i) monitors would be of a single design switchable between 50 Hz and 60 Hz field rates.

Furthermore, as is current practice, category (ii) monitors could be based upon relatively cheap consumer product designs.

Thus, on balance, the CDR approach is to be preferred.

2.7 Liquid Crystal Displays

The possible advantages of the use of a common liquid crystal display with the CIF in the consumer environment have been claimed in several contributions to the CCIR, but the reasons not given.

An important parameter for LCDs is the time available for addressing each cell; this governs the designs of the cell, of the active element (if used), and of the row drivers.

The use of CDR directly related to SMPTE RP 125 results in these designs being common for 50 Hz and 59.94 Hz displays; this advantage outweighs the disadvantage of manufacturing displays with different numbers of rows.

The use of CIF imposes a 20% shorter address time for 60 Hz than for 50 Hz displays; this could result in 60 Hz displays having an intrinsically lower performance than 50 Hz displays, as well as possibly requiring different cell, active element and driver characteristics.

In the consumer environment, display resolution, in terms of both the number of cells per line and number of lines per field, is likely to be directly related to the emission format in order to avoid the unnecessary complexity of interpolation in the receiver.

Furthermore, the emission standards currently being developed in 50 and 60 Hz countries already differ in terms of the number of lines per field and the horizontal resolution, because each is designed to be compatible with the respective existing standards and channels. This suggests, therefore, that different LCD formats would be appropriate in 50 and 60 Hz countries. Moreover, the different LCD sizes that will exist require different compromises to be made in design and manufacture. Thus any claimed advantages due to the use of CIF do not arise.

2.8 Relationship to Computer Graphics for Non-Broadcast Applications

Some segments of the personal computer industry see 'interactive video' as the big growth area for tomorrow. Applications in this field would include:

— computer assisted education

— expert systems

— complex diagnostics

where video sequences might be called up, under computer control, as and when required in the interactive learning/diagnostic process.

More importantly perhaps, there are applications in the business world. A foreign exchange dealer, for example, could have on his screen simultaneously Reuters money rates, a spreadsheet and a broadcast 'windowed in' from a leading politician discussing financial policy.

For all of these reasons, leading elements in the computer graphics industry would like to see a convergence of television and computer *display* standards.

Nevertheless, it should be noted that the uses to which HDTV may be used in interactive video will involve picture-in-picture (16:9 windowed-in to a 4:3 picture). Under such circumstances, spatio-temporal filtering will be necessary, thus making the CIF and CDR approaches comparable.

Currently, the *de-facto* display standards used in the computer industry are various and quite different to those used in television, and many are based upon line numbers which are multiples of 128, e.g.

640 pixels × 480 lines progressive 60 Hz
1024 pixels × 768 lines interlaced 43.5 Hz
1152 pixels × 900 lines progressive 66 Hz
1280 pixels × 1024 lines progressive 60 Hz

all on a 4:3 aspect ratio.

The Industry recognises, however, the importance of 'backwards compatibility' with the various standards used in computer graphics, and would not itself wish to change aspect ratio or field rate standards in existing computer displays.

The use of square pixels in the *display* is considered as particularly important in order to be able to create certain geometric shapes (e.g. circles) easily and to preserve compatibility with existing software packages. It is not clear just how precisely square the 'square pixels' need to be for these applications or whether the spread in current HDTV proposals (1.04 - 1.07) would be acceptable.

There is no preference for any of the active line numbers contained in the current HDTV proposals, although it could be noted that 1152-lines is a multiple of 128.

2.9 Electronic Graphics and Digital Effects

Recent years have seen conspicuous advances in the areas of Electronic Graphics and Digital Effects, so that today these have become integral elements in many television programme productions. It has been suggested in the CCIR that a CIF approach employing square pixels will be of benefit in these areas, and that due weight should therefore be given to this aspect in the overall standards debate.

For example, it has been suggested that a CIF with square pixels will facilitate image manipulations and digital effects (e.g. rotations). It may appear that square pixels will facilitate the drawing of circles in graphics devices. However, a graphics device will need also to draw ellipses, and so current SMPTE RP 125 devices simply treat circles as right ellipses (with an eccentricity of around 9% in the case of the 625/50 system); non-square pixels therefore present no real hardship for graphics.

It has also been confirmed by some manufacturers that the design cost and complexity of digital effects equipment (as with other digital video circuitry) is heavily dependent on the maximum sample-rate. In the opinion of major European manufacturers of graphics and post-production equipment, for any standards approach which does not also use a constant data-rate, equipment has to be designed for the highest sample-rate. The additional cost of development and production then penalises the lower field-rate (and sample-rate) user.

In summary: investigations in this area lead to the conclusion that there is no identified benefit of any significance for the CIF approach.

2.10 Electronic Cinematography

The factors of importance for Electronic Movie production include:

Vertical Resolution — where a greater number of lines and progressive scanning have merit

Movement Portrayal — where the relationship between movie and television frame rates is of importance (and where progressive scanning can assist motion compensation in the transfer process)

A progressively scanned standard employing a 50 Hz field-rate has particular merits in terms of obtainable picture quality[7]. However, the question of CDR or CIF with regard to progressive scanning has not yet been addressed.

Further study is required to define the relative advantages of CDR or CIF for this special application. Noting, however, that the number of video-to-film transfers is likely to be small and their costs very high, it seems unlikely that either CIF or CDR will offer major advantages.

2.11 Relationship to the ISDN

Two areas for discussion have been identified:

(i) A heirarchy of services

(ii) A hierarchy of transmission bit rates

A hierarchy of services might include international transmission of services related to:

- still pictures
- teleconferencing
- conventional television distribution
- conventional television contribution
- HDTV distribution
- HDTV contribution

At present, digital *still picture* stores use 525 or 625 line versions of the 4:2:2 standard. Standards for reducing the digital bit-rate in transmission are being progressed internationally in ISO. The use of CIF would remove the need for spatial interpolation when HDTV stills are exchanged between 50 Hz and 60 Hz countries. There is, however, no information on the traffic in stills nor on economic importance of this.

A Common Intermediate Format has been agreed between 50 Hz and 60 Hz countries for *teleconferencing* only.

In simple terms, the source standard is based upon progressive scanning; 575/2 (approx) lines, 59.94 Hz/2 frame-rate, and 352 pels per line orthogonal. In transmission, heavy bit-rate reduction and temporal sub-sampling is required to achieve bit-rates as low as 64 kbit/s. The temporal impairments involved in these processes are such as to make those arising from field rate conversion for display on 50 Hz screens relatively trivial.

Digital standards for the transmission of *conventional television* standards will remain closely linked to a CDR approach. The (32-45 Mbit/s) CODECS and multiplexing equipments under development for digital component television, for example, being switchable from 50 Hz to 60 Hz operation. Enhanced NTSC, PAL and SECAM transmission is almost certain to follow the same CDR pattern.

Noting the divergence that naturally arises in the different approaches to picture transmission for different types of service, it is difficult to see what the relative advantages of either CIF or CDR have in terms of a service hierarchy including HDTV transmission.

A hierarchy of digital transmission bit-rates

Different hierarchies for digital transmission have been adopted in different parts of the world. Some of the main hierarchies in use today are as follows:

North America (Mbit/s)	Japan (Mbit/s)	Europe (Mbit/s)	International Satellite (Mbit/s)
45	32	34	60
274	97	140	120

A rationalisation of standards for international use in the ISDN is being discussed in the CCITT. To date, agreement has been reached on the H2 level which makes 33 Mbit/s available to the user, but agreement is yet to be achieved on the H4 level (in the range 130-140 Mbit/s).

Whilst agreement on a hierarchy of bit rate standards for international use mitigates in favour of a CDR approach to HDTV standards (switchable CODECS and multiplexing), the degree of CDR advantage over CIF is not clear at present.

In summary: on balance there is no clear evidence that the CIF approach will provide concrete advantages in ISDN terms in either a 'hierarchy of service' or a 'hierarchy of transmission bit rate scenario' — the relatively small benefits for CIF that would result in the still picture transmission area being more than offset by the use of CDR in a common bit-rate channel for the appropriate conventional (4:2:2) and HDTV levels.

2.12 Conversion to other Media

One of the reasons for interest in the CIF approach is its claimed easy relationship with other media including print. However, the aspect ratio of printed images is determined by the pictorial subject matter and page layout; there is thus no particular reason for linking it to an aspect ratio chosen by other criteria (e.g. 16:9). If transfer to and from print will require cropping and sizing (i.e. windowing and resampling) then the CIF or CDR aproaches will be equally applicable or problematic.

2.13 Combined CDR and CIF Proposals

Several examples for achieving a CIF with a CDR structure have been put forward for discussion.

The principle adopted to achieve this for both 50 Hz and 60 Hz field rates involves 'padding-out' the line and field blanking times of the 50 Hz system to achieve parity in the *overall data rate*.

While this can provide for common digital HDTV interface, the principle has major disadvantages:

a) The *active-line data rates* are 20% different, thus contradicting the basic principle of a CDR based on SMPTE RP 125 that *both the active-line and overall data rates* should be the same.

b) A major advantage of the CDR approach — commonality of equipment — is lost.

c) In terms of active/total bit rate, the 50 Hz versions have an efficiency of less than 80% which will have to be paid for in terms of unnecessary videotape usage and processing rates.

For all these reasons, it is believed that a truly combined CDR and CIF system with satisfactory parameters is economically unrealistic.

3. CONCLUSIONS

(i) In considering the relative merits of the CIF and CDR approaches to HDTV standards in all of the most relevant areas of television operations, significant practical advantages for the CIF approach have yet to be identified or demonstrated.

By contrast:

(ii) HDTV standards based upon SMPTE RP 125 and a CDR:

   a) offer the simplest relationship to the HDTV emission and display standards under consideration in many countries.

   b) provide for commonality of equipment in almost all areas of the digital HDTV studio (HD-DVTRs/digital mixers and effects/electronics graphics, etc.)

   c) offer the simplest and highest quality interface to existing studio and distribution standards and equipment.

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DEVELOPMENT OF PLANNING FACTORS FOR ADVANCED TELEVISION

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This paper sets forth an approach to constructing a framework for providing assignments for Advanced Television (ATV) System stations which will be capable of coexisting with the existing VHF/UHF television structure. The paper reviews the TV assignment plan used as the basis for awarding licenses to existing VHF and UHF television stations. It then identifies the factors to be considered for an ATV assignment plan including: the status of activity presently underway in the FCC Advisory Committee.

THE EXISTING NTSC TV ASSIGNMENT STRUCTURE

The basis for the existing NTSC Television station assignment structure was adopted by the FCC in July 1952. Thus it is about 40 years old. While the technical assumptions used to establish this plan may be out of date because of increased knowledge and improved technology, it is basically sound. It has allowed the establishing of over 1000 VHF/UHF television broadcast stations which have provided reasonable service to the vast majority of the population in the United States.

NTSC Structure

In order to insure that both the new and the existing television systems will not cause mutual interference to each other, it is important to understand where the boundary between the two is going to be drawn.

The diagram in Figure 1 shows the basis for the existing NTSC system. This system can conceptually be explained in relatively simple terms: i.e. "6 MHz Channels are repeated throughout the country based on maximum transmission power and antenna height, and separated so as to minimize interference to each other."

In fact, the NTSC structure is based on a complex set of technical and service assumptions which represent a series of trade-offs. They were directed toward finding the best solution to providing the most TV channels, to the most people, with the best quality using the technical understanding available at the time.

Figure 1 illustrates, the areas of trade off. They can be broken down into three general areas: Receiver, Transmitter, and Interference. All these are interconnected, and indeed the fundamental challenge to designing a TV service is choosing the best trade-offs. The assumptions chosen are the planning factors. The Planning Factors for NTSC are in Figure 2.

Receiver Model

The receiver model planning factor assumptions used by the FCC are indicated in the first Table of Figure 2. These assumptions are the type of information which must be identified and agreed before a TV Assignment Plan can be constructed.

The relationship of picture quality to interference was done by subjective testing. This testing was key to establishing many of the important planning parameters of the NTSC system. Basically it determined the level ofsignal required in a Grade A service area which would have to be protected. This in turn established the basis for a modified "receiver noise-limited plan" (minimum co-channel interference within a prescribed service area).

D/U Ratios/Grade of Service

The fundamental choice that has been made is the separation between co-channel stations. This was determined by arriving at a value for the minimum acceptable signal which would be required in the grade A (Shaded) and B service areas of station #2 in the presence of a co-channel signal from station #1. This Desired to Undesired (D/U) signal ratio is the level of signal related to certain assumptions about the receiver.

The D/U ratios chosen for the NTSC Plan are shown below in the second Table of Figure 2.

Grade A represents a specific value of ambient median field strength existing 30 feet above ground which is deemed to be sufficiently strong, in the absence of interference from other stations, but with due consideration given to man-made noise typical of urban areas, to provide a picture which the median observer would classify as of "acceptable" quality, assuming a receiving installation (antenna, transmission line, and receiver) considered to be typical of suburban or not too distant areas. This signal level is sufficiently strong to provide such a picture at least 90 percent of the time, at the best 70 percent of receiving locations. The Grade A contour represents the outer geographic limits within which the median field strength equals or exceeds the Grade A value. The specific values for Grade A are 68 dBu (2.5 mV/m) for Channels 2 to 6, 71 dBu (3.5 mV/m) for channels 7 to 13, and 74 dBu (5.0 mV/m) for channels 14 to 83.

Grade B represents a specific value of ambient median field strength existing 30 feet above ground which is deemed to be sufficiently strong, in the absence of man-made noise or interference from other stations, to provide a picture which the median observer would classify as of "acceptable" quality, assuming a receiving installation (antenna, transmission, line, and receiver) considered to be typical of outlying or near-fringe areas. This signal level is sufficiently strong to provide such a picture at least 90 percent of the time, at the best 50 percent of receiving locations. The Grade B contour represents the outer geographic limits within which the median field stregnth equals or exceeds the Grade B value. The specific values for Grade B are 47 dBu (0.22 mV/m) for Channels 2 to 6, 56 dBu (0.63 mV/m) for Channels 7 to 13, and 64 dBu (1.6 mV/m) for Channels 14 to 83.

As a consequence of information obtained from the subjective testing it was learned that when two co-channel stations did not operate on exactly the same frequency, but were slightly "offset" there would be a significant reduction in interference. Thus it was possible to carry out the assignment planning taking this into account.

Transmitter Model

To achieve satisfactory coverage areas with the transmission equipment available, the maximum transmitter antenna heights, and transmitter powers were chosen as indicated in the Table Table in Figure 2.

The vast majority of U.S. territory is in Zone II. This was intended to be the standard in establishing the transmitter powers and antenna heights to achieve the quality of service in the Grade A and B service areas. Zones I and III were established because of special circumstances. Zone I was established because of the higher density of population in the Northeastern part of the United States. As a consequence, instead of the 190 miles for VHF and 175 miles for UHF in Zone II, Zone I was given 170 miles VHF, and 155 miles UHF separations. This was in recognition of the shorter distances between the large population centers. This is also reflected in the lower permitted antenna heights for Zone I. Zone III is the Gulf Coast region where because of unusual propagation conditions the co-channel mileage was increased to 220 miles (VHF), and 205 miles (UHF) respectively.

Taboos/Propagation

These are the basic planning factors that have gone into the design of the NTSC assignment of frequencies for specific geographic areas. However, to complete the picture, and to set the stage for the introduction of an advanced television service within the framework of the NTSC system two additional subjects need to be addressed. These are 1) Propagation assumptions, and 2) UHF Taboos.

At the time when the assignment plan was designed for the UHF channels, separation distances were also necessary for other channels related to the primary channel, n. The separation distances, associated with n ± 1, 2, 3, 4, 7, 8, 14, and 15 were to protect against interference due to IF selectivity limitation, beat, intermodulation, oscillator radiation, sound image and picture image. Together these additional prohibitions against use of certain channels in each geographic area, are known as TABOOS. They were adopted because of the state of UHF receiver technology at the time. As described subsequently, they have also served to make available spectrum, which when taking account of improved technology, data compression processing techniques, and advanced modulation techniques may permit the co-existence of significantly improved TV transmissions.

The second point is the natural occuring sources of noise and interference. To take account of the urban noise and interference, a planning factor of 14 db for low VHF and 7 db for high VHF was used. Secondly, the service quality indicated above assume that tropospherically propagated co-channel interference occurs 10% of the time. Finally, CCIR Recommendation 655 recommends a protection ratio 10 db higher than the tropospheric interference to provide margin for continuous interference. This is the "taboo" interference.

ADVANCED TELEVISION SYSTEM PLANNING

What are the planning factors which must be considered to plan an Advanced Television System in the VHF/UHF Television allocations? The previous section has indicated how a number of interrelated planning factors have had to be defined in order for a framework to be developed.

Some initial work in this area has been done by the FCC's Advisory Committee on Advanced Television. One of its planning committee's working groups has been examining the availability of VHF/UHF spectrum to accommodate an Advanced Television System. This work has revealed that under certain circumstances there would be enough spectrum to accommodate an ATV terrestrial broadcast system. However, to actually plan such a service requires the answering of many questions. Figure 3 shows availability of ATV spectrum if no taboos are assumed.

This part discusses the considerations related to the development of an ATV Assignment Plan in a fashion analogous to those developed for the NTSC Plan. As indicated below, the approach is similar, but the information is different.

Necessary Information

As the discussion on how the existing NTSC system reveals, constructing an assignment plan for an ATV system requires answering similar if not identical questions. The questions that need to be answered require more specific information. They include the following:

Receiver Model

- Which ATV emission schemes which have been proposed will use: 1) Improved 6 MHz, NTSC, 2) NTSC + 3 MHz, or 6 MHz, NTSC, 3) Simulcast. The nature of their characteristics are in Table I. Given the choice of one of these three alternatives, for each ATV system a typical receiver needs to be characterized in order to form the basis for an assumption with respect to the level of input signal which relates to meeting the desired level of quality for receivers in the indicated coverage area.

D/U Ratio

- What is the D/U ratio which will be required by the new system? This is a basic planning factor which will in turn determine the separation distances between co-channel ATV stations, and as a consequence, the amount of spectrum which will be available, particularly in the Northeast high density population areas. The field strength requirements which will determine the numbers are in Figure 4. The separation distances are related to the four interference situations which need to be understood. These are:

I. NTSC to NTSC

II. NTSC to ATV

III. ATV to NTSC

IV. ATV to ATV

NTSC to NTSC is the basis of the present assignment structure. Recent work by the FCC and reported to the Advisory Committee indicates that there will not be a problem for ATV to ATV using some worse case assumptions. The modulation of several ATV schemes suggest that they will cause no "continual" interference to NTSC. It remains to understand the effect of NTSC on the ATV Systems.

As ATV emission systems are likely to be quite different from NTSC, and NTSC, equivalent power level (thus D/U ratio) needs to be defined for each in order to objectively compare them.

Taboos

The analyses to date in the Advisory Committee's Planning Working Group on spectrum also indicates that for the ATV systems not to cause unacceptable interference to the existing NTSC receivers and yet provide every, or nearly every present NTSC broadcaster with extra spectrum, the ATV system has to be capable of operating at much lower equivalent NTSC to ATV D/U ratios. This is due to the expected much lower level of transmission of the ATV signal (compensated for by processing in the receiver).

The principal area of uncertainty is what will the NTSC signal do to the ATV receiver. This question relates to the actual impact of using taboo spectrum for the ATV transmissions. The original spectrum availablility estimates were based on the assumption of being able to use all the taboo spectrum. Now, for the planning of assignments it is necessary to know what the effect of the NTSC taboo transmissions will have on the ATV receivers. This can only actually be known when the ATV receiver is defined and tested. However, it is possible to make some assumptions and perform some computer analyses to get an idea of what to expect. This is presently underway in the FCC Advisory Committee.

Transmitter Model

Once some assumptions are made about the receiver, D/U ratio, separation distance, and actual availability of taboo spectrum, the final piece of information to be decided is the nature of the transmitting system. These decisions concern:

- Size of service area. This relates to maximum antenna height, and, transmitter power. Will the service area be equivalent to that of the NTSC.

- Number of stations in a market. This relates to the probable requirement to have an ATV assignment available for every existing VHF/UHF license in a given market regardless of whether it is intended to be used or not.

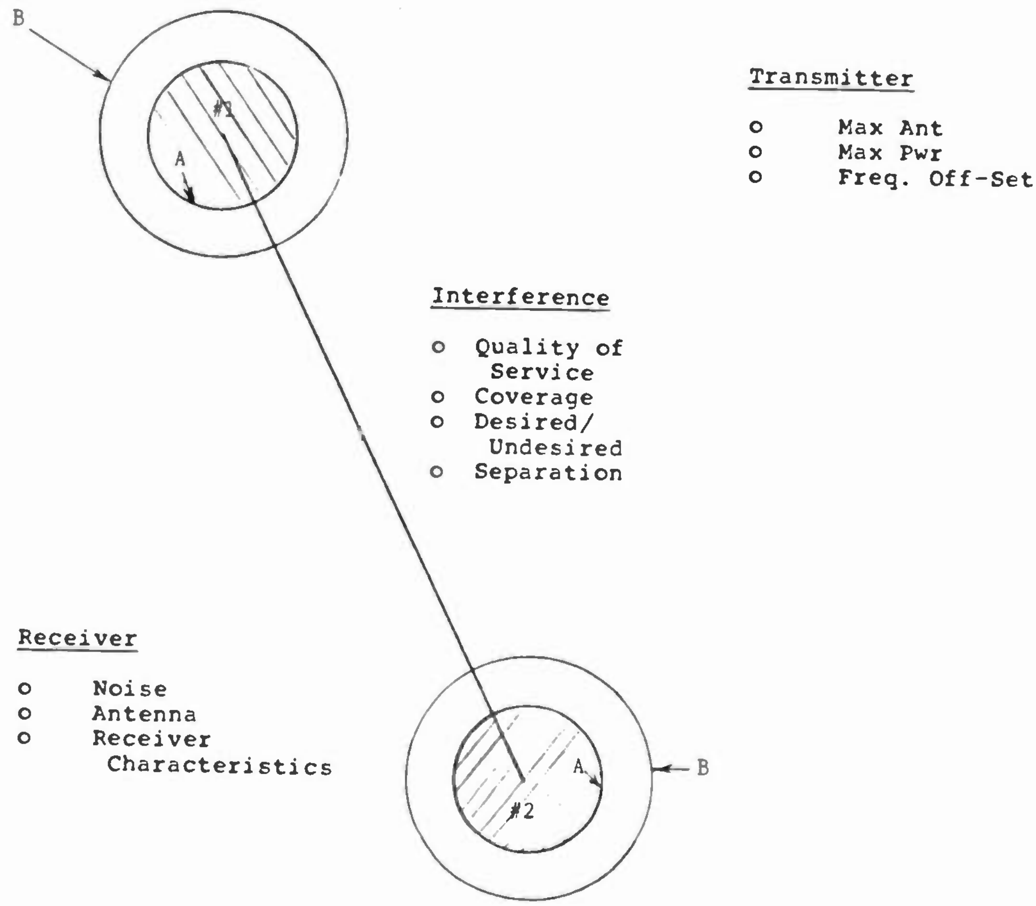
- Co-location of transmitters. This may be required, particularly in major Northeast metropolitan areas to obtain maximum spectrum availability. The nature of the benefit to be obtained from this needs to be examined.

Propagation

Finally, the effect on the ATV signal of the typical radio propagation path needs to be understood, to determine if the previous planning assumptions are valid, when a Plan is being constructed. This should be accomplished by over the air tests.

SUMMARY

The development of the Necessary Planning Factors for an ATV Service in the United States is under way. It is quite possible to proceed with the boundaries of these parameters regardless as to whether the proponent systems have been tested. Doing so will provide a framework for assuring that a particular system will permit the implementation of sufficient stations. The Planning Factors, which similar in concept to those for NTSC, may well take on different characteristics.



NTSC Assumptions

Figure 1

Figure 2

NTSC RECEIVER MODE ASSUMPTIONS

	Receiver Planning Factor	Assumption
1.	Thermal Noise	7 db, ref 1 microvolt
2.	Receiving Antenna Gain	0 db (VHF), 8 db (UHF)
3.	Dipole Factor	2.9 db (Low VHF), -6.0 db (High VHF) -16.5 db (UHF)
4.	Receiver Noise Figure	12 db (VHF); 15 db (UHF)
5.	Antenna to Receiver Losses	1-5 db depending on variables

D/U RATIOS

The D/U ratios chosen for the NTSC Plan are shown below in Table 3.

Table 3

Grade	VHF (db) N	VHF (db) O	UHF (db) N	UHF (db) O
A*	51	34	53	36
B**	45	28	45	28

N = Non OFFSET   O = Offset

TRANSMITTER MODEL

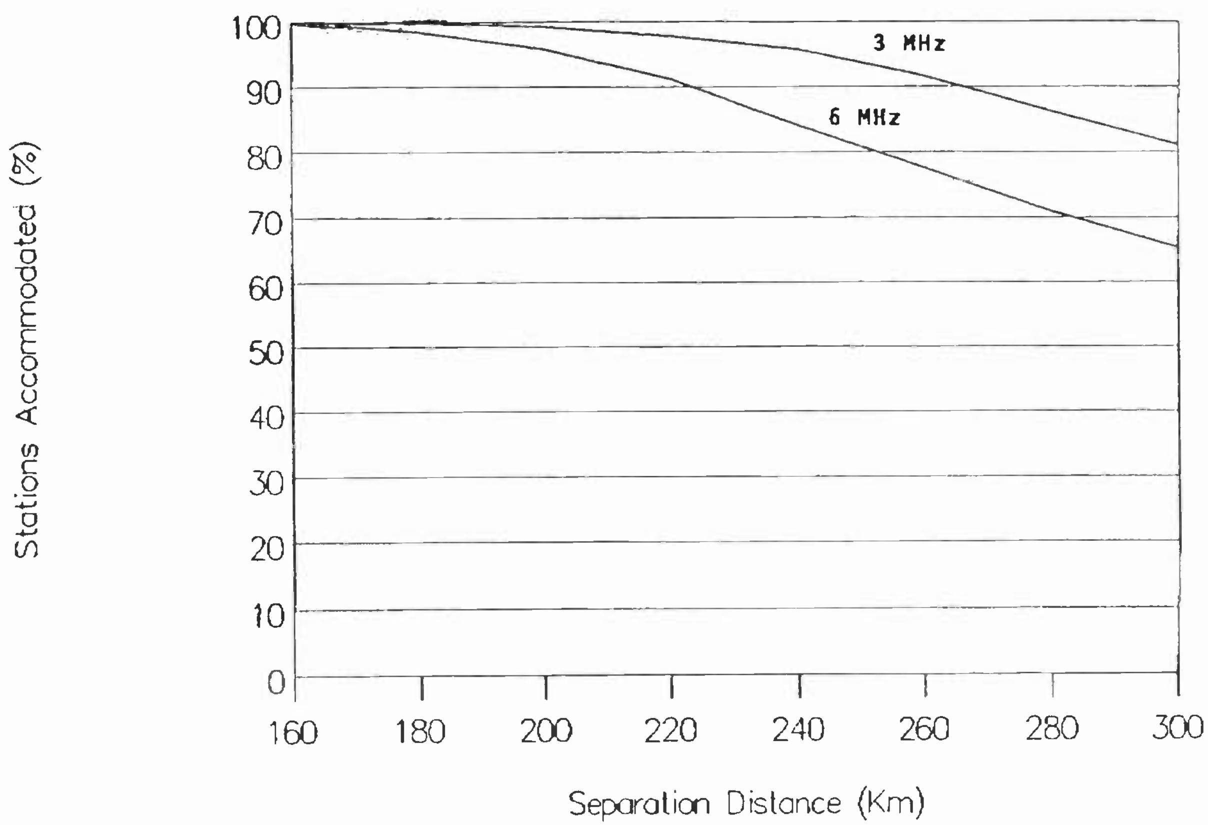
To achieve satisfactory coverage areas with the transmission equipment available, the maximum transmitter antenna heights, and transmitter powers were chosen as indicated in Table 4.

	Zone	Low VHF	Sep Dist (miles)	High VHF	UHF	Sep Dist (miles)
Trans mitter	All*	100 kw		316 kw	5000 kw	
	I	305 M	170	305 M	610 M	155
Antenna Heights	II	610 M	190	610 M	610 M	175
	III	610 M	220	610 M	610 M	205

Figure 3

3 vs. 6 MHz Noncontiguous

(No Adjacent Channel Protection)



APPENDIX WP-3A

Figure 4
FIELD STRENGTH REQUIREMENTS FOR BASIC SERVICE AREA
(In the absence of interference)

PARAMETER	Sign	NTSC GRADE B CONTOUR						ATV ?		
		CHS 2-6		CHS 7-13		CHS 14-7		CHS:		
		O	C	O	C	O	C	2-6	7-13	14-
N_1 (dBu)	+	7		7		7				
N_R (dB)	+	12	6	12	7	15	10			
S/N (dB)	+	30	40	30	40	30	40			
K_d (dB)	-	3		-6		-16				
G (dB)	-	6		6		13				
L (dB)	+	1		2		5				
ΔT 90% (dB)	+	6		5		4				
ΔL 50% (dB)	+	0		0		0				

TOTALS 47dBu 56dBu 64dBu

TYPICAL RADIUS 65 MILES 60 MILES 55 MILES

BASIC EQUATION $E = N_1 + N_R + S/N - K_d - G + \Delta T + \Delta L$

O ORIGINAL 1952 VALUES

C CURRENT ESTIMATE

HDTV OPTICAL DISC PLAYBACK SYSTEM

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ABSTRACT:

A new HDTV optical disc player, HDL-2000, will be described that operates in accordance with the SMPTE 240M 1125/60 production standard. The format is a wideband component analog system. The system introduces a mass media distribution mechanism for HDTV. It is intended as a high performance HDTV playback system within a variety of applications in business and industry.

The HD disc player employs a 12 inch optical disc which is prestamped via a specially developed mastering scheme. There are two playback modes - constant angular velocity (CAV) giving an 8 minute playback and a random access time (worse case) of 3.0 seconds and two, constant linear velocity (CLV) with a random access time (worse case) of 10 second and a play time of 15 minutes.

1.0 Background:

In 1984 Sony introduced a first generation HDVS product line designed to support High Definition program origination and post production. This included camera, VTR, switcher and display (both studio monitors and projection systems). This equipment helped stimulate the first ventures in HDTV production in Japan, Europe, U.S. and Canada and the USSR. At the same time, worldwide interest in HDTV had led to the formation of Study Groups and Working Groups within professional organizations to begin the long process of standards development of a future HDTV production television format.

It was early recognized, however, that the initial distribution of program software, originated in HDTV, would be via existing media - 525/59.94 and 625/50 based broadcast television, video cassette, and film. For this reason, Sony, and other others, undertook as a first priority the development of appropriate conversion systems to allow transformation from the HDTV format to each of these distribution formats (**Figure 1**). By 1986 high quality downconversion from 1125/60 to 625/50 (employing a novel motion vector detection algorithm) had been successfully demonstrated.[1] In 1987 Sony introduced an 1125/60 to 525/59.94 downconverter to the HDVS product line. For more than a decade a major development effort on an HDTV Electron Beam Recorder (EBR)[2] has been underway to support high quality transfers from 1125/60 HDTV to 35mm film. A first EBR service was brought on line in Tokyo in October, 1986. A second EBR service opened outside London in February 1990 and a U.S. EBR operation will begin in Los Angeles in 1990.

While the primary goal of the early thrust in HDTV equipment development centered about those system elements essential to support program production - a considerably broader view of HDTV anticipated important advances in its wider use. The application of television imaging has already branched out in many directions from the original exclusive over-the-air broadcast system intended to bring entertainment programming to the home. Throughout the past three decades, television has been increasingly applied to a vast array of teaching, training, scientific, corporate, industrial applications. Within this same era there also emerged an extensive worldwide infrastructure of independent production and post-production facilities (more than 1000 in the U.S. alone) to support all of these needs.

As the Hollywood film industry became increasingly involved in supplying prime time programming for television (on an international basis) via 35mm film origination, the use of television technologies were harnessed to support the off line editing of these film originals, and to provide creative special effects. Meanwhile, the world of Computer Generated Imaging is growing at an explosive pace and it too is penetrating countless industrial, scientific and corporate applications - including the film production community. Looked at today, therefore, the television industry has splintered into disparate (though at certain levels, overlapping) specialized industries - **Figure 2**. Any one of these television sectors is gigantic in itself.

It was into this environment that HDTV was born. And HDTV brings an entirely new dimension to imaging: where formerly 525/625 television is inadequate (for many applications) in the clarity of the image it portrays, the HDTV image contains almost six times more total electronic information - thus dramatically altering the scope of the potential uses envisaged by industry visionaries.

2.0 HDTV Production Standard:

No single topic in television has preoccupied the attention of standardization committees - worldwide - as the quest to develop a high quality HDTV studio origination standard. This has been much complicated by the attendant high goal of achieving international unanimity on a single worldwide HDTV production standard. The goal to secure a new global flexibility in high performance television production, post-production, and international program exchange is high on the agenda of many farseeing participants within the global television industries. Sadly, the achievement of this high goal is not yielding easily to rational technical examination or business examinations.

However, the pursuit continues. A great deal has been learned, and much accomplished. Specifically, the needs of an HDTV origination standard that contains sufficient technical overhead to support subsequent conversions to: 525 and 625 existing television systems; to future ATV transmission systems; and transfer to 35mm film are far better understood today than half a decade ago. A considerable HDTV experience, at the hands of prominent international program producers and organizations, has added a great deal of vital commentary to the seemingly endless technical discussions. A wide variety of HDTV programs have been made to date; downconversions have allowed programs to be released in 525 and 625; two motion pictures, originated in HDTV, have been theatrically released following transfer to 35mm film.

Meanwhile, the detailed work of SMPTE and ATSC, begun in 1983, continues with a high level of activity today. The entire subject of transfers between film and HDTV are now under intense study, as is the complete digital representation of a production standard.

From all of this work, the parameters of a durable origination standard have been structured to endow the HDTV studio format with more electronic information than is required to satisfy, for example, the psychophysical aspects of large screen viewing of the picture. This additional information translates into the technical overhead that can sustain complex picture processing in post production, multi-generation recording, downconversion, transfer-to-film, etc. The next important work phase is distribution of the HDTV program material itself.

3.0 Business and Industrial Applications of HDTV:

The work described has formed an essential underpinning to a viable, robust and high performance standard. The front end of the system, as it were, has been thoroughly researched and unusually complete standards writing is emerging. However, the practical implementation of this standard calls for the harnessing of high technologies and especially in the case of cameras and recorders, certain key components are pushed to the current limits of physical implementation.[4,5] As a consequence - equipment costs are high in these early days - prohibitively so, for many possible and worthy applications of HDTV.

Nevertheless, the establishment of a strong production standard was essential to the definition of some of the more fundamental parameters of a broad based HDTV system.

The sheer breadth of the television industry today (and that which we can only anticipate in the future) calls for flexibility in the actual broad deployment of an HDTV standard. The concept of a system hierarchy has been much discussed in recent years. And indeed, a carefully structured hierarchy offers an ideal methodology to tailor a basic HDTV system to a wide variety of diverse applications while optimally meeting specific performance requirements, system needs, and, of course, budgets. Within B&I (business and industrial - and indeed, other non-entertainment television) applications - there exists a wide diversity of imaging requirements. The picture quality and system facilities needed are unique to each application. The spearhead of a B&I HDTV evolution, however, was recognized to be dependent on a lower cost method for storing and distributing HDTV images. Central to realization of such systems, lies the challenge of reducing the very large bandwidth of the HDTV studio origination standard without compromising the essence of the HDTV image.

4.0 Business & Industrial HDTV Distribution Format:

An imperative was therefore given to developing a general purpose lower member subset of the recently standardized 1125/60 system (SMPTE-240M) that would allow cost-effective HDTV program distribution within a wide cross-section of B&I applications (**Figure 3**). Such a subset would, in some respects, parallel the concurrent and separate research now actively underway in the U.S. and elsewhere to develop an Advanced Television (ATV) transmission system. The B&I subset would, however, be squarely based (as indeed it is today) on distribution via packaged media - That is, cassette tape and optical disc.

Under the guidance of NHK a format definition was generated in 1987 that substantially lowered the total component video bandwidth compared to the SMPTE-240M production standard. Yet, this retained all of the requisite information to adequately meet the psychophysical visual requirements for HDTV (as originally laid out in the early NHK research program). **Table 1** shows a brief summary comparison of the basic parameters of this signal format.

Table 1

Parameter	SMPTE-240M	B&I Subset
SCANNING PARAMETERS:		
Total No. Lines	1125	1125
Active Lines	1035	1035
Field Rate	60	60
Interlace	2:1	2:1
VIDEO COMPONENT BANDWIDTH:		
Y	30	20
P_b	15	7/7
P_r	15	(Line Sequential P_b/P_r)
TOTAL BANDWIDTH:	60	27

Based on this format definition Sony embarked on the development of a 1/2 inch cassette-based VTR (now supported by nine manufacturers) and an HD optical disc system - **Figure 4**. This paper describes the latter - which is a new product addition to our available HDVS product lineup. We first introduced the HDL-2000 HD disc player in 1988 and it has been selling worldwide since late in that year.

5.0 HD Optical Disc System Applications:

The optical disc is ideal to facilitate repeated playings of a prerecorded program (or portion thereof). The medium is very robust, convenient to handle, and offers the striking advantage of very long life because of the absence of any physical contact with the medium other than a laser scanning beam. The more obvious applications are:

- Teaching training (Scientific, Medical)
- Point of sale (product promotion)
- Theme parks/Recreation Centers
- Library (Image retrieval/storage)
- Museums
- Exhibitions
- Animation
- Simulation

- Special effects
- Public relations

The very high visual quality of the video, accompanied by two channel CD quality audio, makes for a very powerful presentation medium.

6.0 Laser Disc Player - The Disc:

A twelve inch standard disc size was selected. The overall physical dimensions (diameter, thickness, etc.) are the same as today's conventional laser disc. Thus, conventional laser disc production techniques can be employed for HD disc stamping - **Figure 5**. Master disc cutting is accomplished, however, using a newly developed precision optical cutting machine.

The disc employs very tiny pits (of the order of um) whose primary axis is modulated in length by the variable frequency recording signal. **Figure 6** shows the essentials of the disc format. A double spiral track is employed to record the HDTV video in component analog form. The large amount of information (27MHz total component video bandwidth), and the physical constraints on recording density, determined the play time available from the disc. However, there are two modes of operation:

- CAV - constant angular velocity
- CLV - constant linear velocity

which speak to the manner in which the diminishing radii of the spiral tracks (as one scans from outer to inner) are managed. The two modes of operation offer a substantial difference in playing time - attendant with a trade off in certain facilities (some of which can be important in specific applications).

Figure 7 simply explains the difference between the two modes of recording. In CAV the disc rotates at a constant angular speed: the linear velocity of the tracks consequently vary from a maximum of 27.7 meter/sec (at the outer edge of the disc) to a minimum of 18.5 meters/sec at the inner section of the disc. This mode of recording affords the minimum play time -being 8 minutes and some 20 seconds in duration.

Constant linear velocity, on the other hand, varies the angular velocity of the disc (proportional to the progression of the scanning from disc outer to disc inner) in a manner that maintains a constant linear velocity of all tracks. This almost doubles the play time - which now becomes for CLV a 15 minute capability. Both of these recording times can be doubled by the simple expedient of recording on both sides of the disc (the player accommodates this) - in the same manner as an audio phonograph - the operator reverses the disc.

CAV, however, offers maximum flexibility in terms of the facilities offered. **Table 2** summarizes these facilities.

Prominent among the limitations of CLV is the inability to play video in reverse, or to fast scan. Also, the maximum access time to a given video frame is almost 10 seconds in the case of CLV, in contrast with some 3 seconds in the case of CAV. Video and audio technical quality remains constant - regardless of the mode of recording. So, the choice of recording mode becomes the end user's - and is basically a trade off between playtime and specific facilities - a trade off which varies by application.

7.0 HDTV Video - Optical Disc Recording Format:

A double spiral track recording format was elected the optimum methodology to record the three wideband input analog component video signals Y, P_b, P_r. This method, and the bandwidths of the individual video signals involved, determined the recording packing density - and hence the ultimate playtime of the disc recording.

As shown in **Table 1** the luminance (Y) bandwidth prior to modulation for recording is pre-filtered to 20MHz, while that of each of the color difference signals are pre-filtered to 6MHz (C_b and C_r). The waveform structure of these three separately filtered signals are as shown in **Figure 8**. The next task is to multiplex the three video signals into two video signals of essentially equivalent bandwidth preparatory to modulation for recording on the two tracks. Time division multiplexing techniques are employed to structure such signals.

7.1 The two multiplexed video signals have effectively half the original line rate.

7.2 The Luminance Y video is time stretched by a factor 5/3 to reduce its effective bandwidth from 20 to 12MHz. The resultant new television line duration is now twice that of the original input HDTV video.

7.3 The color difference signals C_b and C_r are each time compressed by a factor of 2:1 (which raises their effective frequencies from 6 to 12MHz).

7.4 The Luminance signal is present on every television "line", while the two color difference signals are line sequential (that is, each are present, in turn, on every other television "line").

7.5 There is a significant time dispersion between associated Luminance and Color Difference signals.

7.6 The bipolar nature of the color difference signals are maintained by placing their center axis reference on a 50 percent pedestal.

7.7 A sinusoidal timing burst signal is introduced and multiplexed into the waveform (also riding on the 50 percent pedestal.

The timing pilot signal becomes a reference for the recovery clock used to demultiplex the three HDTV video components during disc playback. **Figure 10** shows the relationship between clock counts and the relative timings of the multiplexed components.

The time dispersion of the multiplexed components constitutes an important part of the dropout compensation scheme. Such dropouts are often unavoidable in disc production, generally manifesting themselves as slight damage to adjacent track pits or some form of microscopic contamination. The time dispersion scheme allows for a degree of predictability in playback and subsequent substitution of damaged data with previous line data to restore a subjectively clean picture.

8.0 Modulation and Playback:

FM modulation of the two multiplexed signals is employed. The modulated carrier signal is then converted to a square wave signal via a zero crossover comparator circuit. This variable frequency signal, in turn, is used to modulate the length of the major axis of the elliptically shaped pits actually recorded on the optical disc. **Figure 11** simplistically illustrates the modulation process.

The center carrier frequency is 16.3MHz and a 1.74MHz deviation is employed. The playback signal-to-noise ratio is typically greater than 42dB.

Figure 12 is a simplified block diagram of the HDL-2000 playback electronics. Two photodiodes are stimulated by the very fine laser beams reflected back from the pits during playback. The output RF from each photodiode is amplified and FM demodulated to reproduce baseband analog replicas of the original two multiplexed video signals.

The digital audio signals are separated at this point, time base corrected and error corrected. Following digital filtering they are digital-to-analog converted to the two separate audio channels. The multiplexed analog video waveforms meanwhile are analog-to-digitally converted for digital time base correction, demultiplexing, time compression (for Y) and time decompression (for C_b and C_r) prior to final video processing. Within this complex digital treatment the line sequential color difference signals are restructured as parallel "every line" signals and the original timing relationship between Y and the two color difference signals is restored.

9.0 Audio Recording Format:

Two audio channels are provided for in the recording format - both are digitally recorded. The incoming analog audio signals are quantized into 16 bits linear at a sampling rate of 48kHz.
CD quality audio playback is ensured by employment of a Reed-Solomon error correction system.

The digital audio data is multiplexed with the two video multiplexed waveforms. The audio data is structured to be carried in the vertical interval of the two video channels in the manner shown in **Figure 13**. A more detailed illustration of the relationship of the digital audio data with the video, sync, and vertical interval time control (VITC) signals is depicted in **Figure 14**. The digital audio is structured into data blocks each of which is carried on one of the channel television lines (which have a duration of 59.26 microseconds, that is, half the line rate of the original 1125/60 HDTV signal). The audio rate for recording is 24.3Mbps and the line block structure is shown in **Figure 15**.

10.0 Optical Pickup System:

A single 5mW laser diode of 780 nmeter wavelength is used as the primary light source in the HDL-2000. This single laser beam is divided into three separate beams by an optical grating - **Figure 16**. The power ratio between the beam spots is, however, different. The two side beams are intended for reproduction of the two-channel multiplexed signals and for provision of tracking-error signals. The center beam is exclusively dedicated to tracking error and focus error sampling. The power ratio of the three beams favors the outer beams according to 3:2:3.

An overall block diagram of the HDL-2000 is shown in **Figure 17** and this illustrates the relationship between the optical disc itself and the optical pickup system. **Figure 18** shows more detail of the optical system, in particular the three incident beams that are reflected back from the disc pits. Via a beam splitter, these reflected beams are directed to a precision semiconductor detector structure.

The physical relationship between the three incident beam spots and the recorded pits on the disc is shown in **Figure 19**. The reflected beams contain a great deal of information - essentially encoded - comprising deviation of the signal FM carrier and critical positional information that speaks to a sought-for precision in the relationship between the tracking optics and the rapidly rotating disc. A special proprietary detection system is employed to ensure a high degree of precision both in the optical focussing of the three beams onto the tiny disc recorded pits, and in the delicate tracking of these pits as the objective lens assembly moves progressively inwards toward the revolving disc center during playback.

The detector itself is the heart of the system. This is a tiny structure of eight photodiodes rigidly positioned with high precision into the physical format depicted in **Figure 20**. The eight photodiodes are identified as blocks A through H. The detection algorithm for recovery of the two RF signal channels are:

E + F = summation of signal outputs from photodiodes E and F produces the RF signal modulated carrier for Channel 1 and

G + H = equivalent summation that produces the RF signal for Channel 2.

Figure 21 (a) is a closeup of the four central photodiode cluster ABCD which exclusively produces information on optical focus error according to the algorithm: **(A+C)-(B+D)**.

Physically, any defocussing will manifest itself as an elliptical image with an orientation of its major axis inclining in a horizontal or vertical direction. The chosen algorithm recognizes this defocussing distortion and produces an error signal which has a relationship with the degree of spot elongation according to **Figure 21 (b)**. The servo system feedback loop physically moves the objective lens system to continually maintain a zero error signal which in turn automatically centers the spot symmetrically above the four photodiodes. This symmetry ensures a precise circularly focussed beam spot.

The novel optical tracking detection system is the most complex of all the servo systems in that it utilizes information picked up by all eight photodiodes which are stimulated by the three separate return beams. The algorithm for this positional detection system is reproduced in **Figure 22** and this servo system has a control loop which ensures a constant incremental repositioning that places the three focussed beam spots in the manner shown in **Figure 19** - that is, the outer spots separately and symmetrically straddle the outer recorded pits of the disc, while the single center spot symmetrically and simultaneously senses two adjacent parallel spots.

The final precision servo system supervising the slide assembly is the radial skew control system. This recognizes the physical fact that the horizontal plane of the flat disc is imperfect in its flatness. The disc can bend due to its own mass, or it may incur other slight warping imperfections due to environment or handling. The effect can cause a small deviation from a perfect normal incidence in the landing of the laser beams on the surface of the disc. The system recognizes this and tilts the lens accordingly to restore an orthoganal relationship between the beam and the recording pits at any given radius of the disc. The sensing system for this is independent of the source laser diode and the eight photodiode sensor previously described. A separate LED light source is mounted on the slide assembly and its light beam is focused and directed at a 45° angle to the point of contact of the normal laser beam. The reflected beam from this new light source (also at 45°) is sensed by a fixed photodiode appropriately mounted -as shown in **Figure 23**. The servo system involved automatically tilts the objective lens system to continually ensure equal angle of incidence and reflectance relative to the plane of the disc - at the specific point of contact with the disc. This inherently maintains orthogonality between the objective lens and the disc regardless of radial deviations in the horizontal plane of the optical disc.

11.0 HDL-2000 - System Interface:

Figure 24 shows a simplified block diagram of the HDL-2000 and its interface capabilities. The HDTV output video can be selected as RGB analog component or as a Y, P_B, P_R Luminance and Color Difference set. The two channels of analog audio are available on standard 3 pin XLR.

Control of the HDL-2000 can be via a dedicated handheld controller RM-2000 which uses a proprietary serial digital control protocol - or an external computer can control all of the basic functions of the player via a standard RS-232C interface.

12.0 Conclusion:

A carefully structured subset of the SMPTE-240M 1125/60 production standard has been developed to support HDTV program distribution by cost effective optical disc or cassette VTR. The technical parameters of this subset fully preserve the attributes of HDTV imaging from the viewpoint of the psychophysical human visual response - while allowing significant bandwidth reduction by sacrificing technical overhead crucial to the post production process (but not to playback for display).

An important first step in providing an HDTV packaged media distribution mechanism based on this 1125/60 subset of the HDL-2000 HDTV optical disk player. This paper described the basic principles of this player, now being sold worldwide by Sony into a broad range of business, industrial, scientific and educational markets that seek to optimally exploit the dramatic prowess of HDTV imaging for diverse applications.

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[3] SMPTE-240M, SMPTE Standard for Television - Signal Parameters - 1125/60 High Definition Production System.

[4] "Second Generation HDTV Camera", Laurence J. Thorpe, Better Video Images, Proceedings of 23rd Annual SMPTE Conference, February, 1989.

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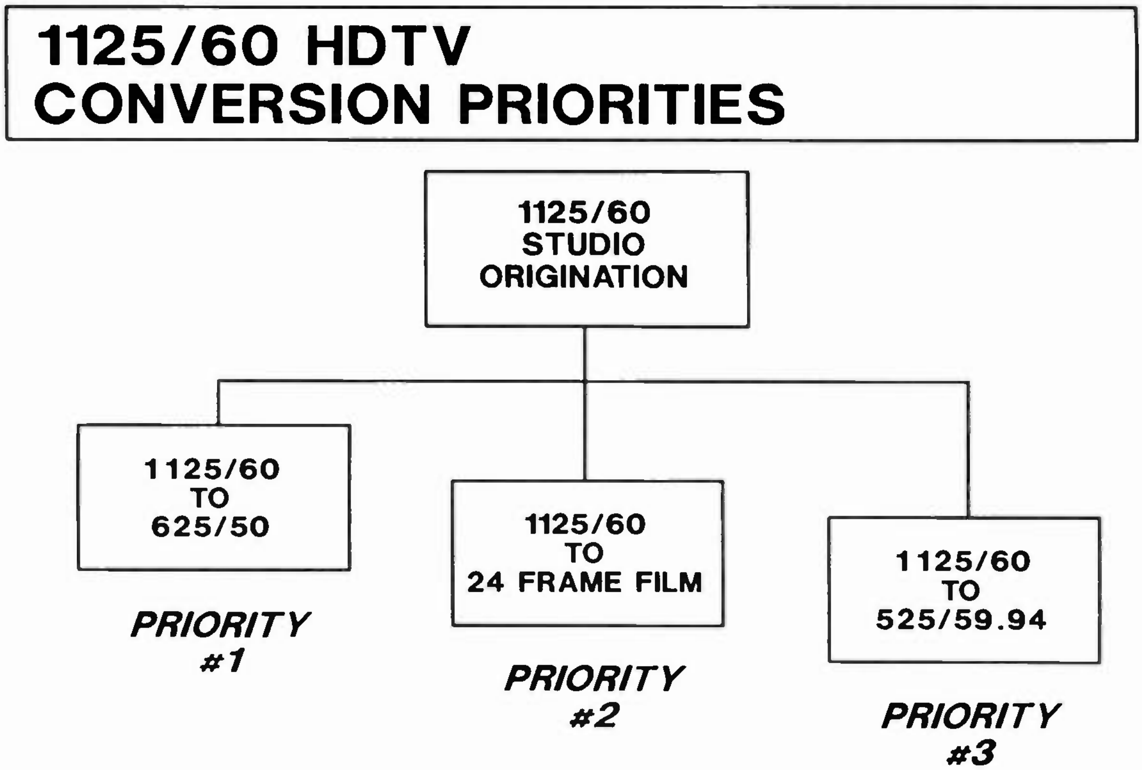


Figure 1 Development priorities in HDTV distribution media.

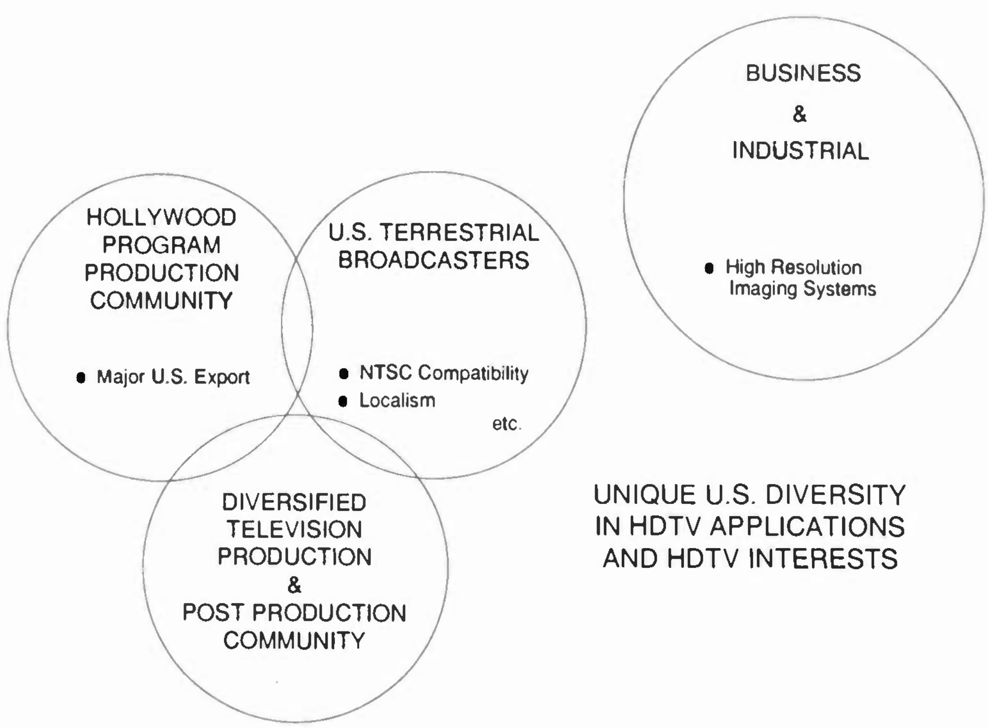


Figure 2 Diversity of present day television applications in the U.S.

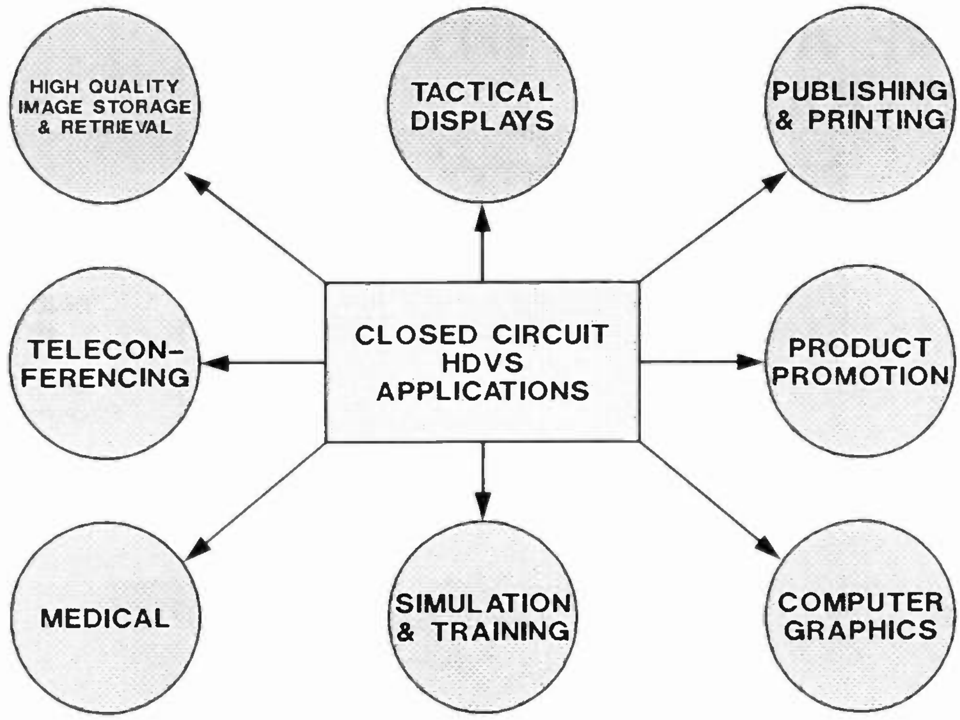


Figure 3 Some of the many business and industrial applications of HDTV currently being explored.

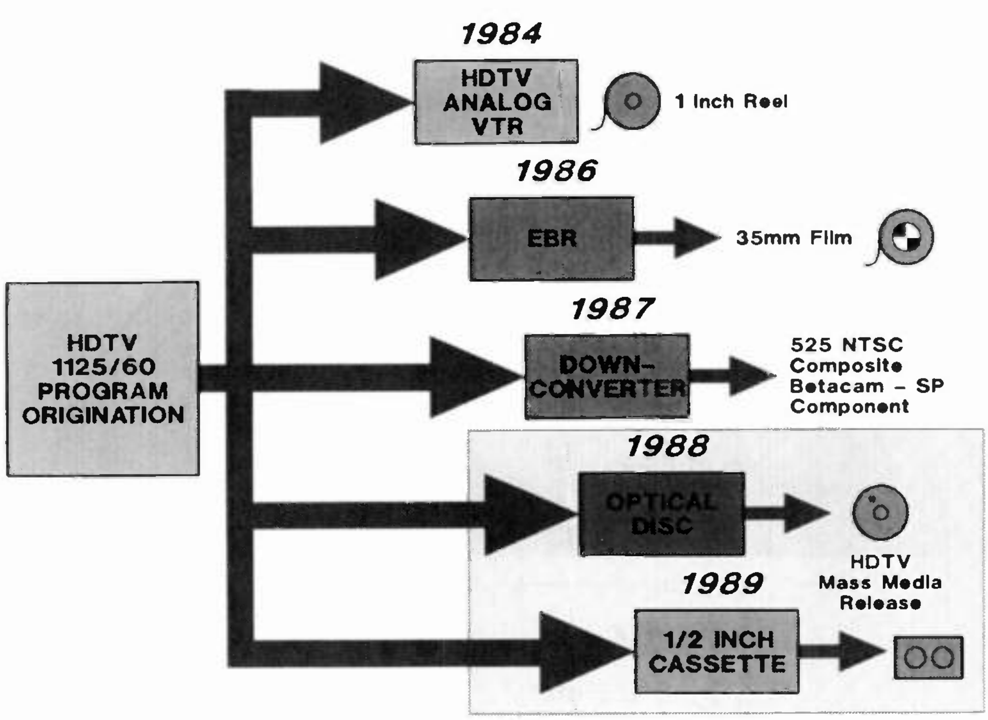


Figure 4 Evolution in HDTV developments of program distribution mechanisms.

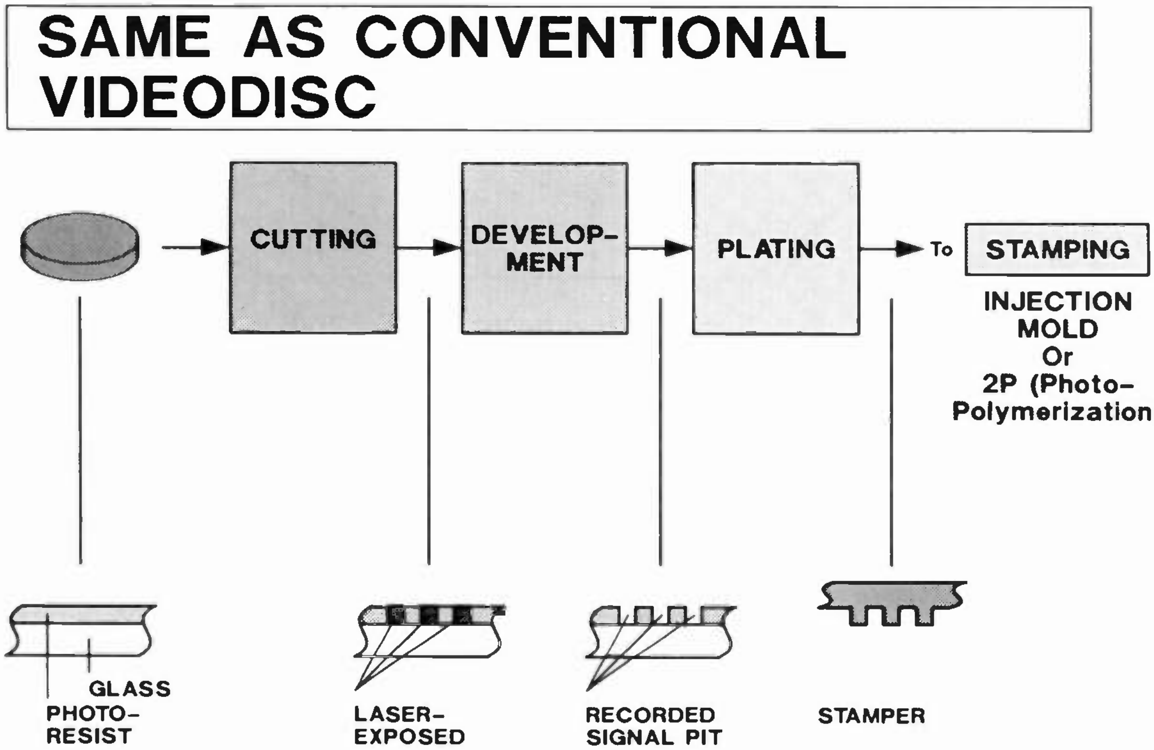


Figure 5 Simplified overview of HD disc production process.

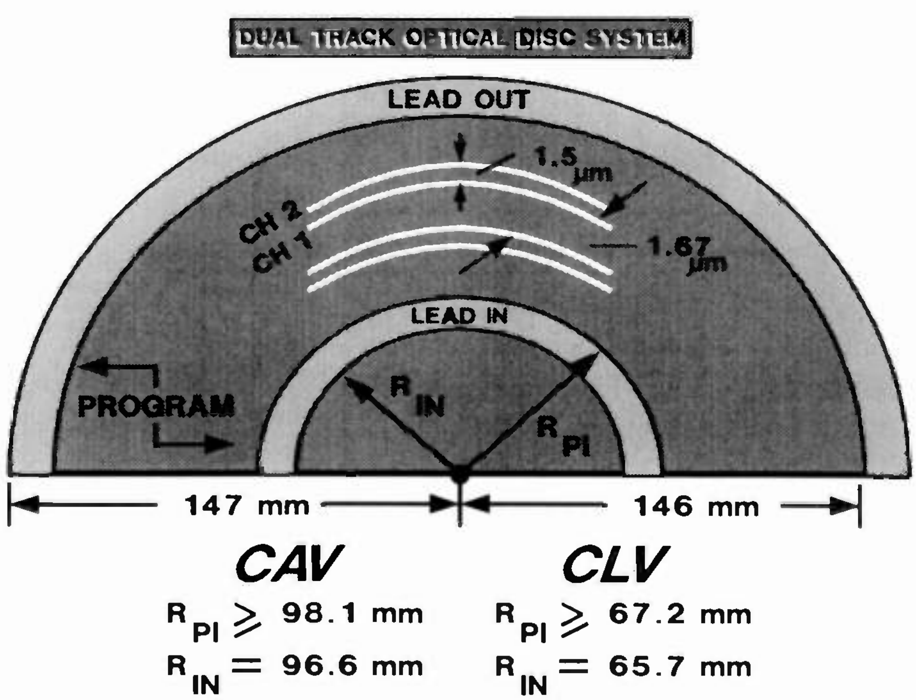


Figure 6 HDTV optical disc format

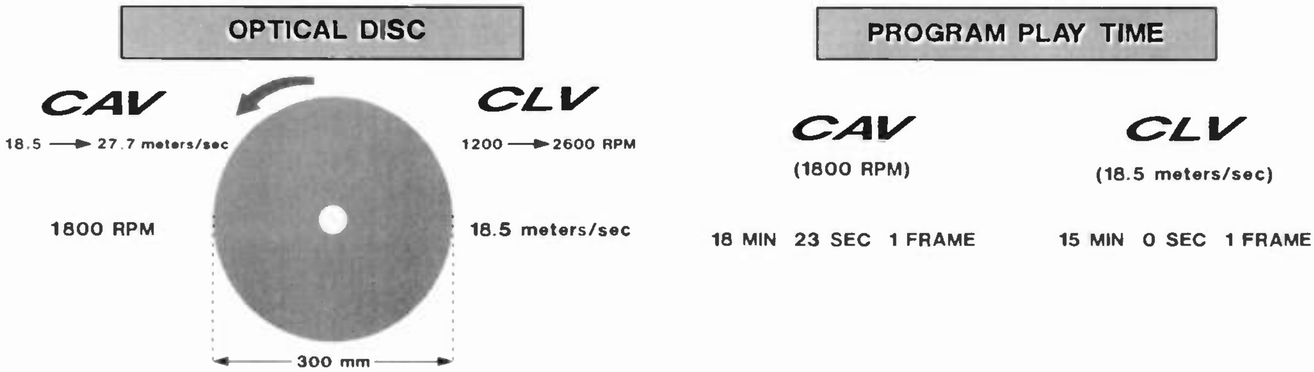


Figure 7 Difference between Constant Angular Velocity (CAV) and Constant Linear Velocity (CLV) modes of recording on HDTV optical disc: **(a)** relative speeds; **(b)** relative play times.

Table 2

COMMAND FEATURE	CAV	CLV
NORMAL PLAY MODE - FORWARD	YES	YES
- REVERSE	YES	NO
SCAN	YES	YES
FAST SCAN (x3)	YES	NO
STILL	YES	YES
STEP (frame by frame)	YES	YES
CHAPTER NUMBER - SEARCH/REPEAT	YES	YES
TIME CODE - SEARCH/REPEAT	YES	YES

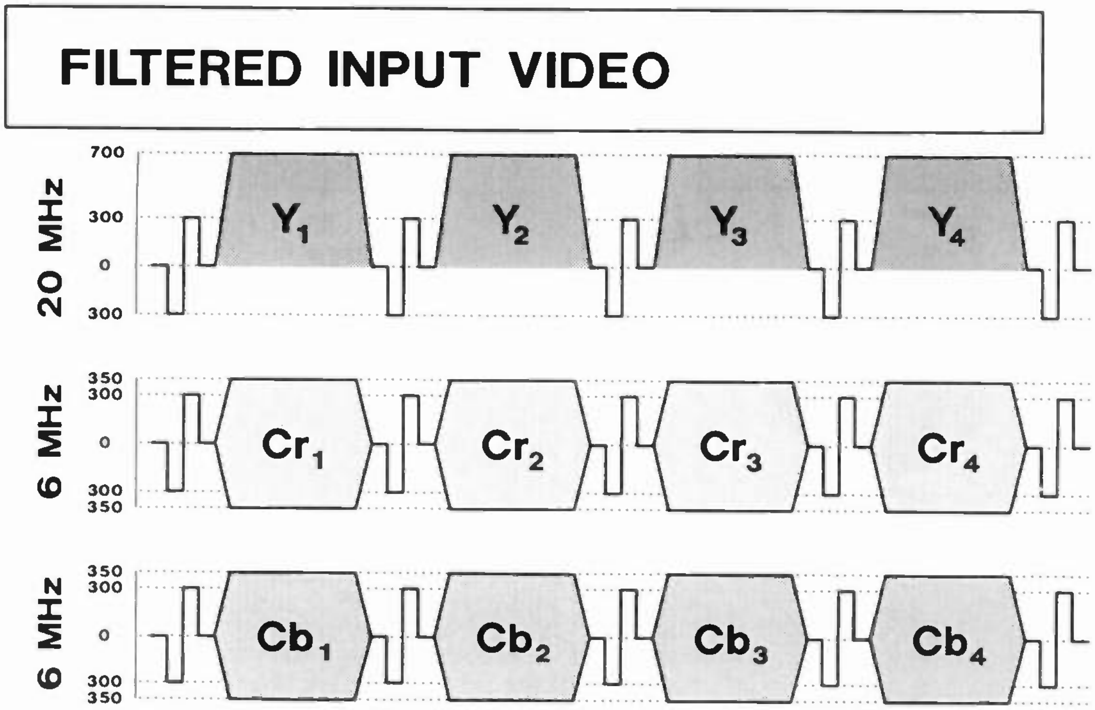


Figure 8 Three HDTV video component signals following low pass filtering prior to recording on HD optical disc.

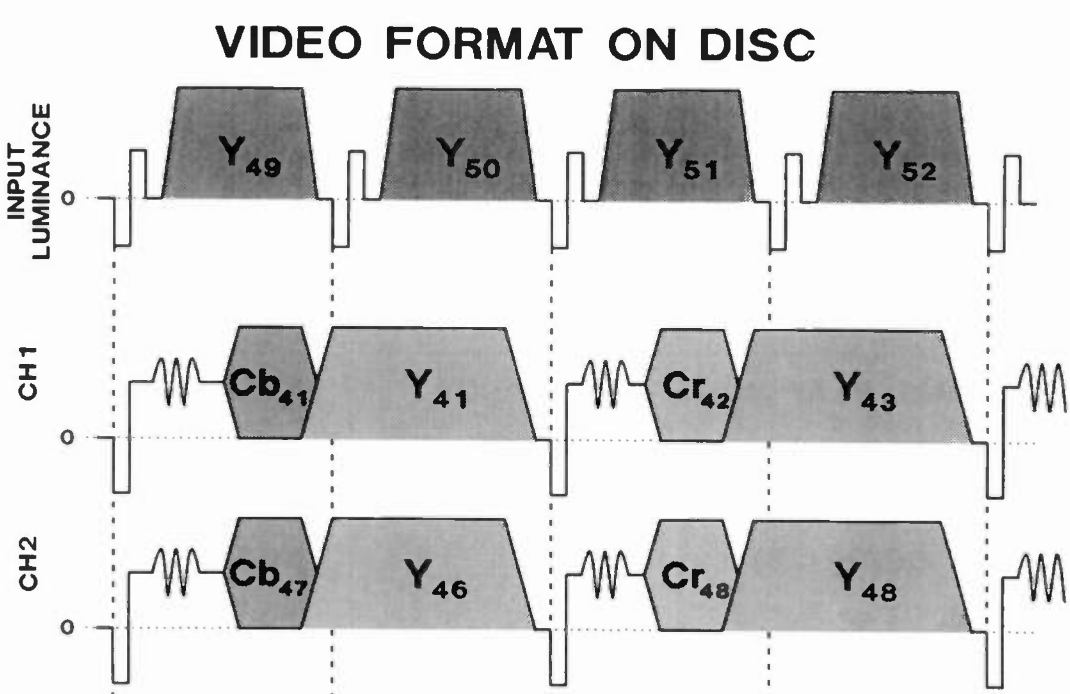


Figure 9 Video format of the two multiplexed signals (CH1 and CH2) recorded on two parallel tracks of disc.

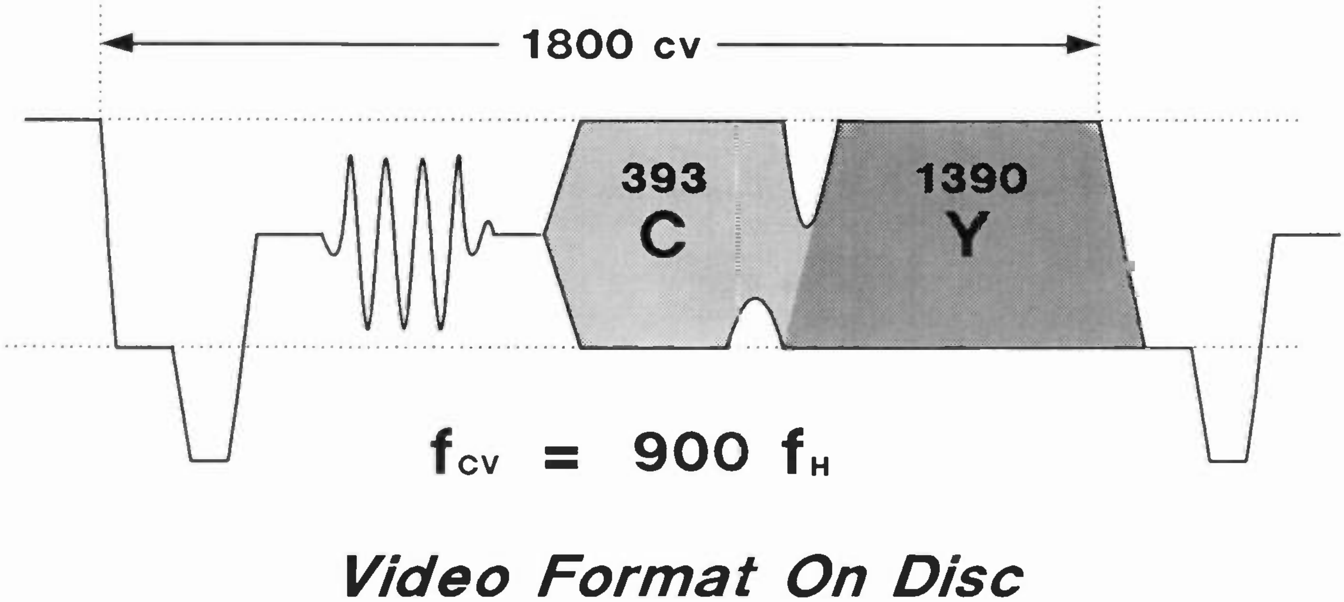


Figure 10 Single television line as recorded on the disc - showing relative clock timing counts

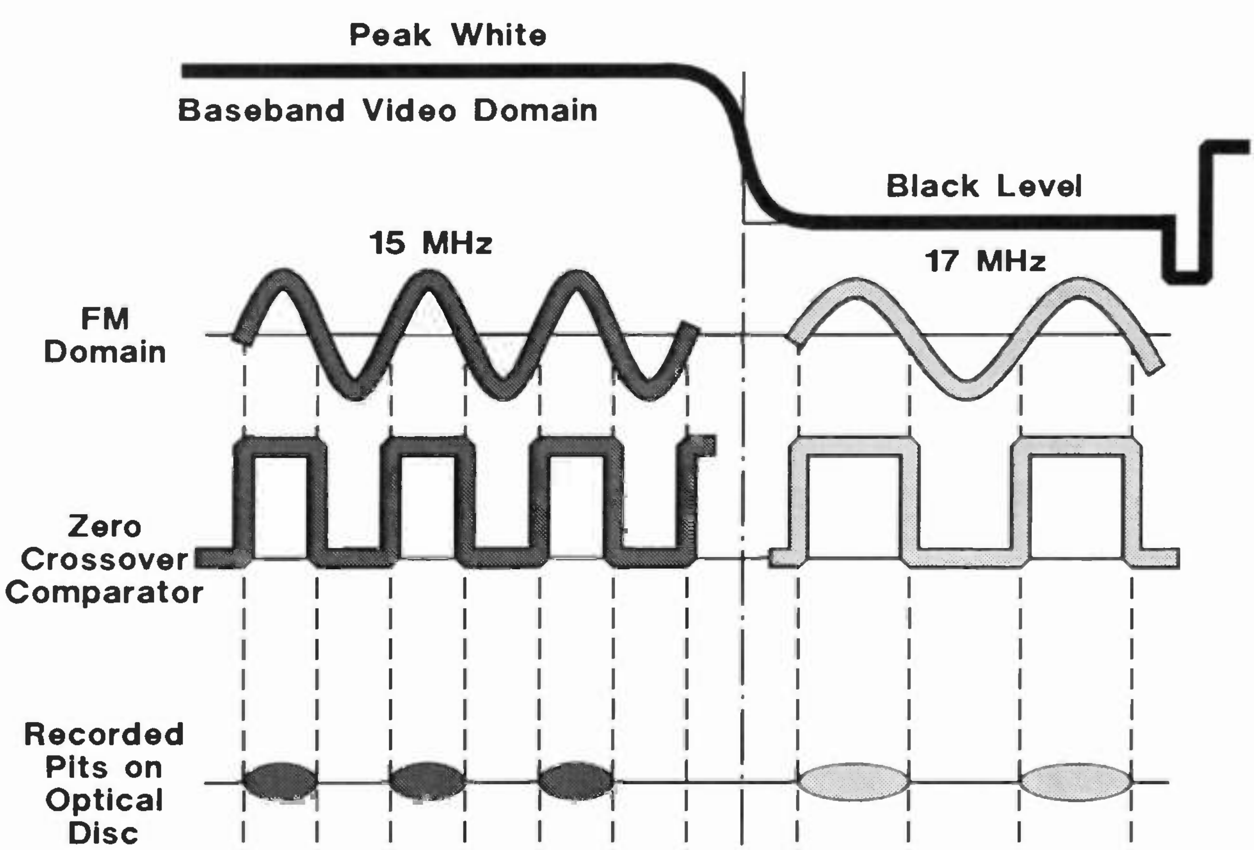


Figure 11 Basic principles of FM recording scheme used to record on HDTV optical disc.

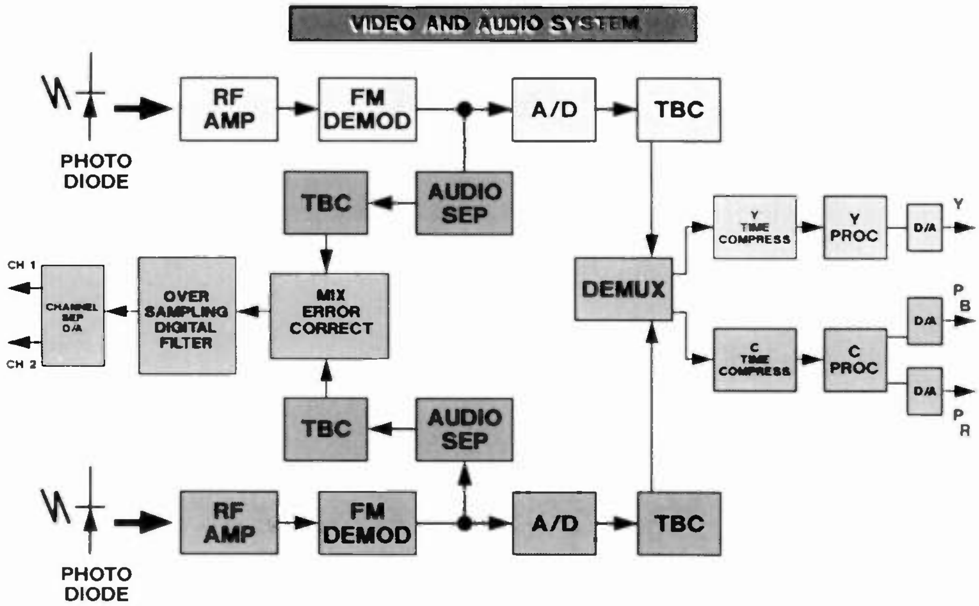


Figure 12 Simplified block diagram of HDL-2000 video and audio playback electronics.

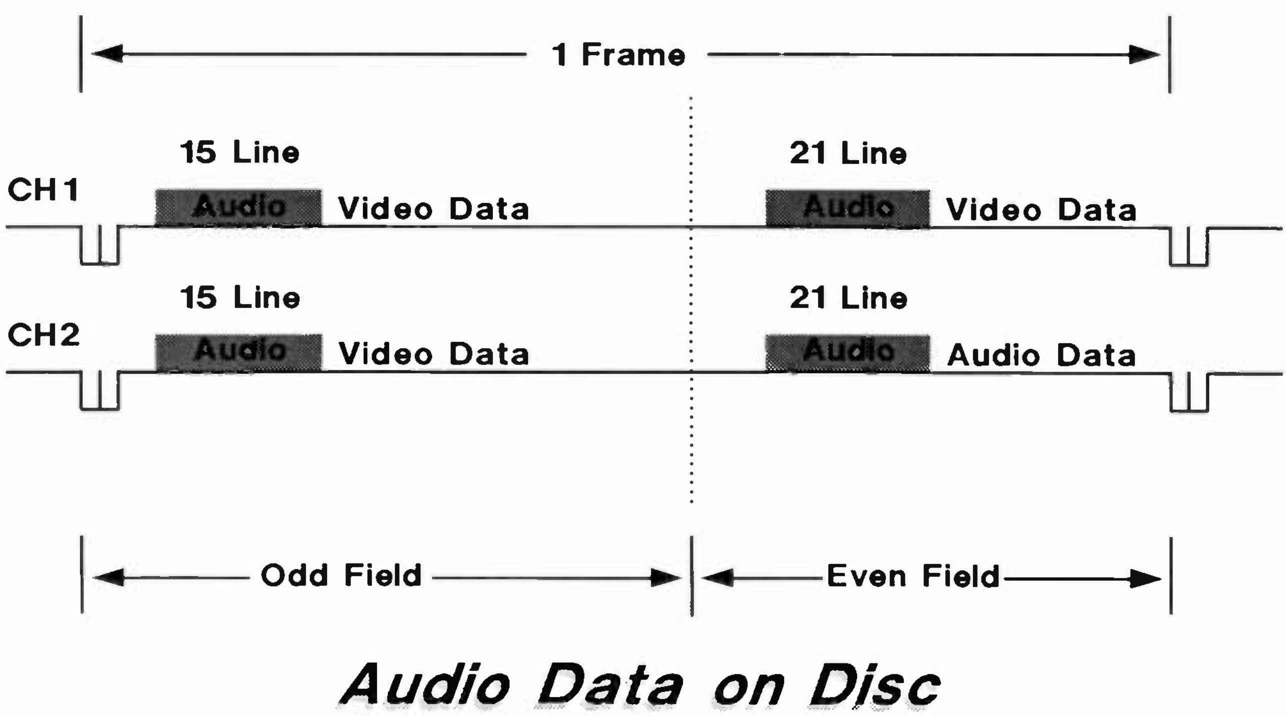


Figure 13 Overview of audio digital data insertion into vertical blanking intervals of the two multiplexed HDTV video signals.

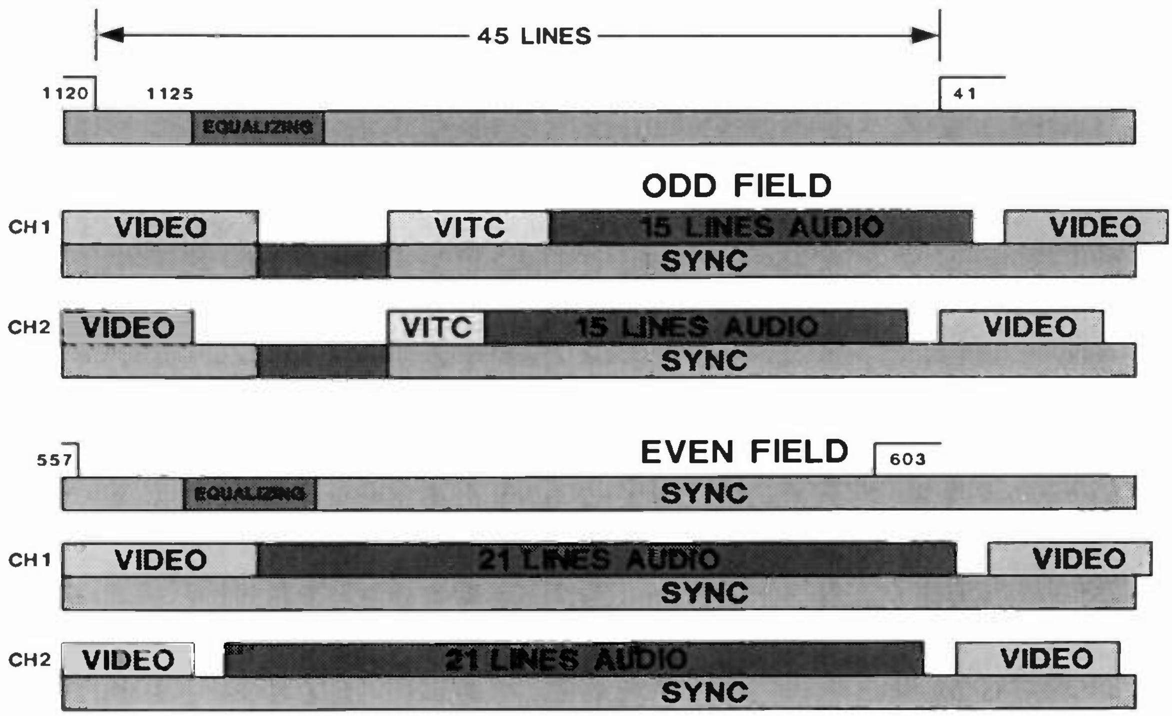


Figure 14 Some details of the composite HDTV analog video, digital audio, VITC and sync signals recorded on the disc.

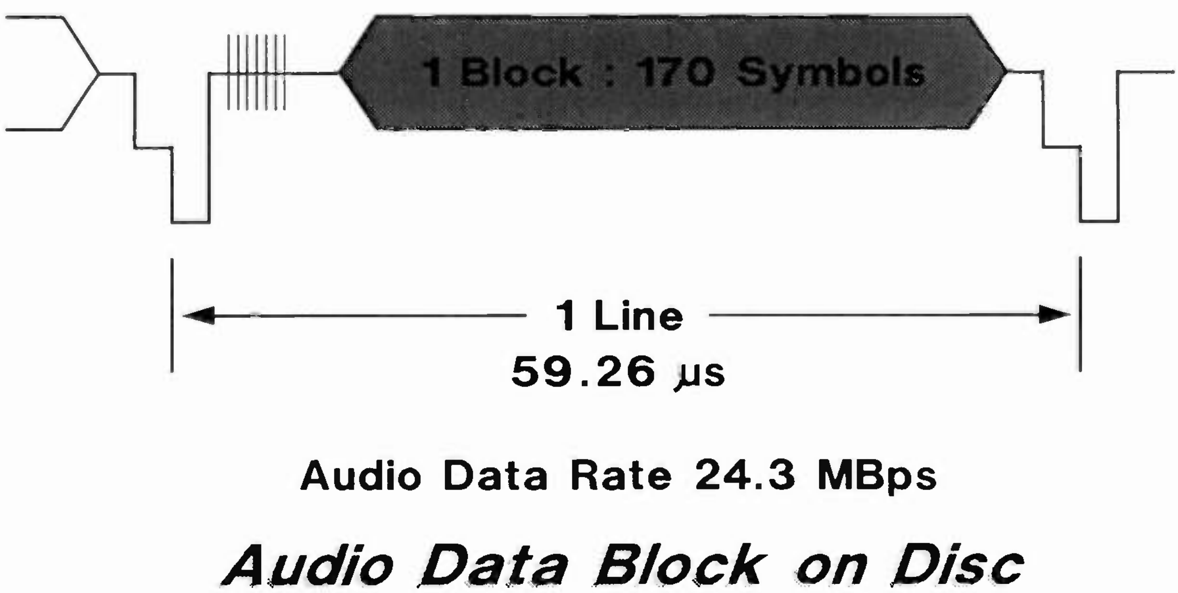


Figure 15 Detail of one television line carrying digital audio block within vertical blanking interval.

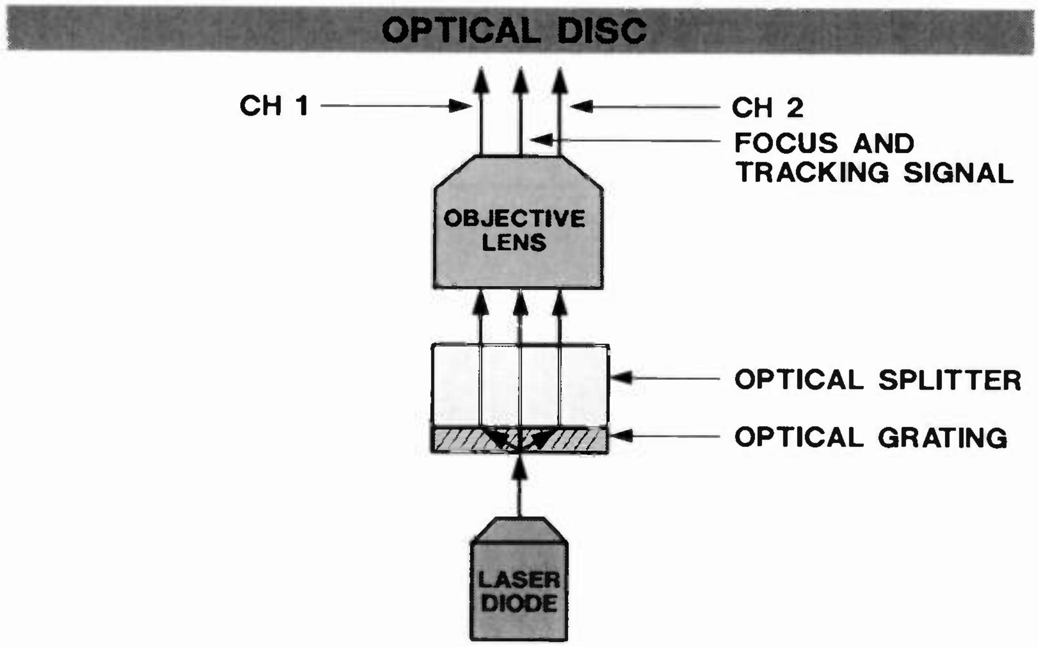


Figure 16 Laser diode source and beam splitting system employed for disc sensing.

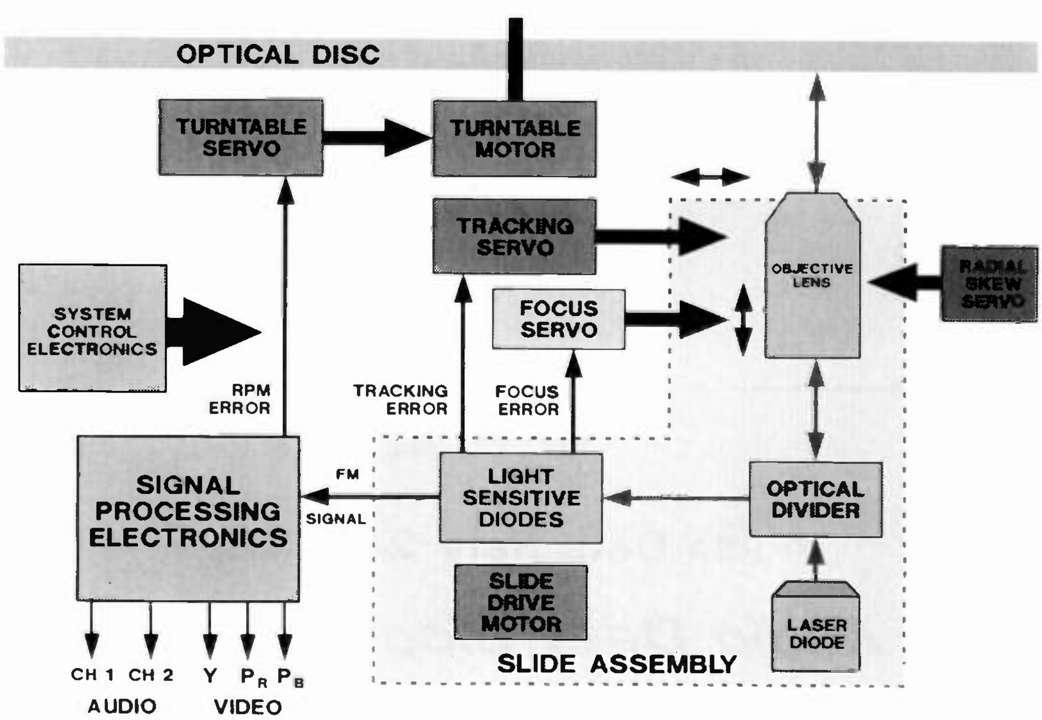


Figure 17 Block schematic of optical slide assembly and associated servo control systems.

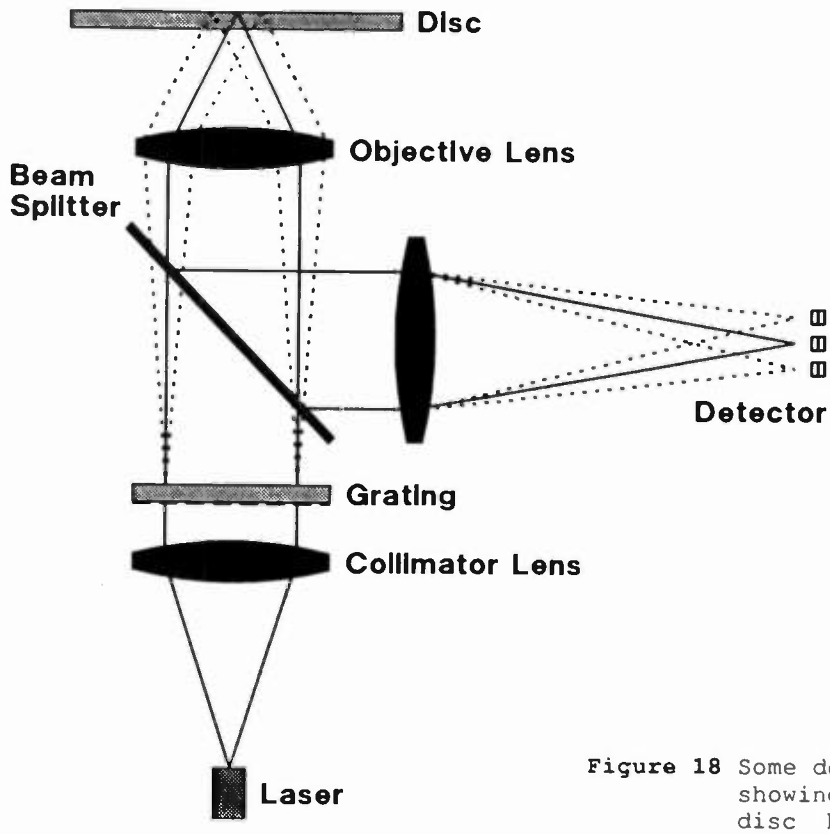


Figure 18 Some details of optical system showing reflected beams off disc being directed to the photodiode detector system.

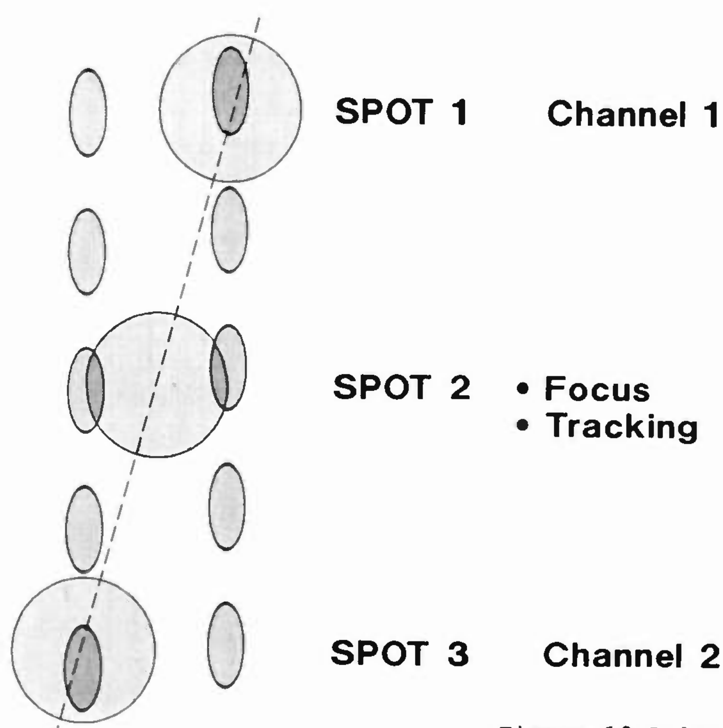


Figure 19 Relationship of three beam spots and recorded pits on disc under conditions of optimal focus and tracking.

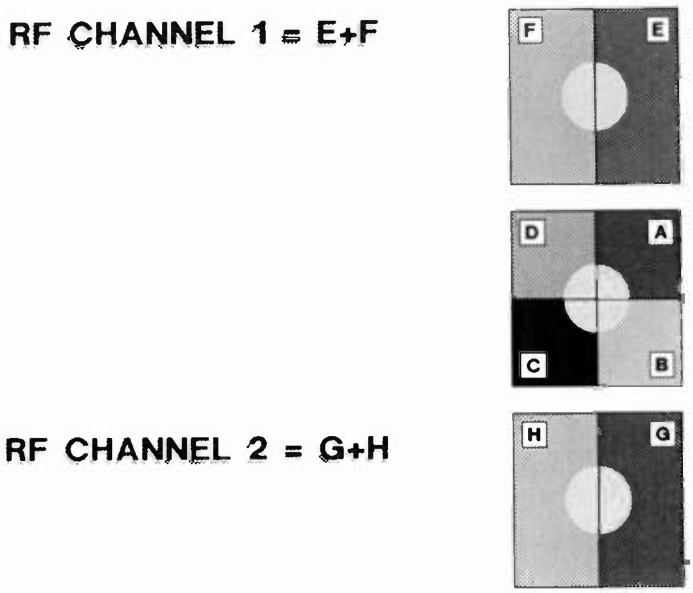


Figure 20 Physical arrangement of the eight separate photodiode sensors comprising the detector system.

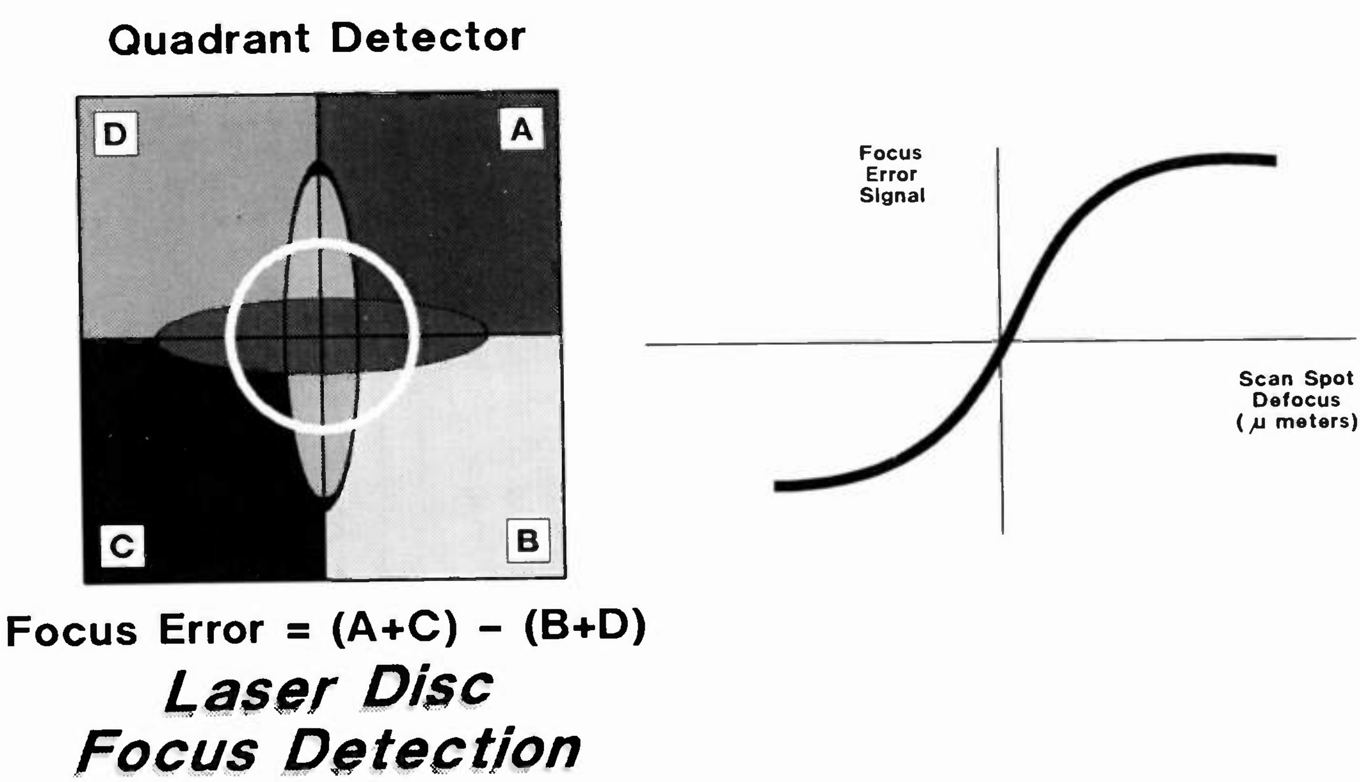
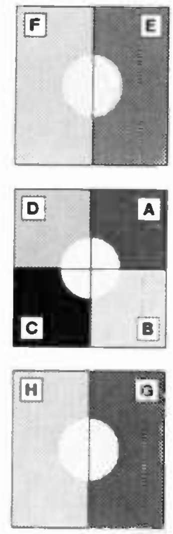


Figure 21 Beam Focus Detection System: **(a)** Closeup of the central four-photodiode quadrant detector; **(b)** Transfer characteristic of focus detector.

LASER DISC TRACKING DETECTION



8 Sensor Detector

Tracking Error=

$$[(A+B)-(C+D)]-K_1[(E-F)+K_2(G-H)]$$

Figure 22 Algorithm used in utilization of all eight photodiodes for beam tracking detection.

RADIAL SKEW CONTROL

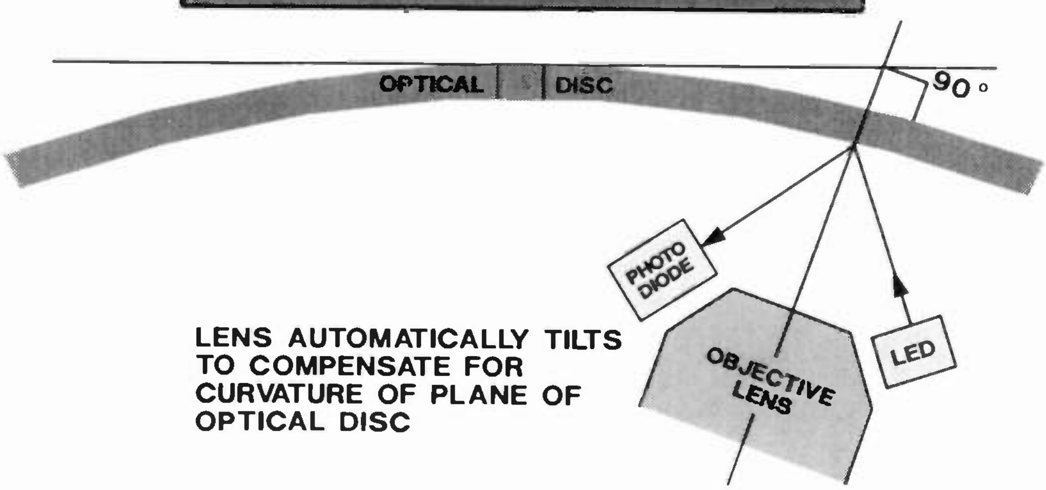


Figure 23 Radial skew detection system for automatic control of incident light beam onto disc surface.

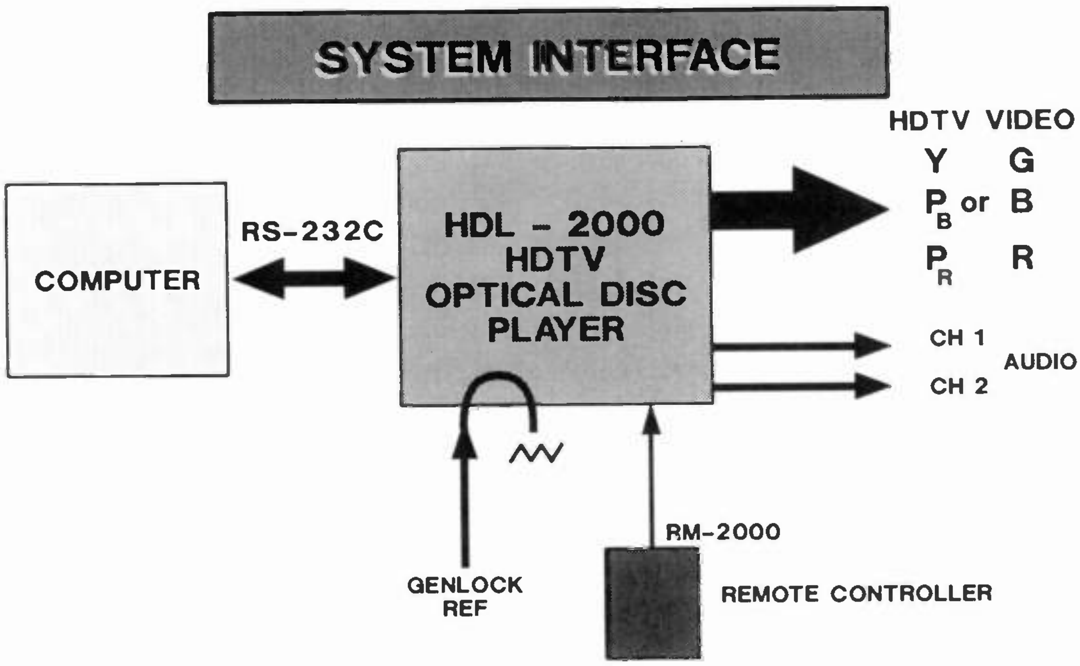


Figure 24 Simplified block schematic summarizing primary HDL-2000 interface with external system.

Figure 25 Photograph showing front view of HDL-2000 optical disc player.

STUDY OF THE METHODS OF SIGNAL PROCESSING APPLICABLE TO THE WIDE ASPECT EDTV COMPATIBLE WITH NTSC

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INTRODUCTION

Last year at this meeting, we proposed the Wide Aspect EDTV System Compatible With NTSC and have been testing Mode 1, the side panel system, and Mode 2,the letter box system, since that time. Also, we have been examining signal processing technorogies applicable to these systems.

For the applicaton of Mode 1 and Mode 2, the signal processing systems using the block thinning-out technorogy, this report explains overview and the result of computerized simulation by still picture.

'Block thinning out' refers to limitting bandwidth and compressing a picture by dividing it into two or more blocks,and thinning out picture elements in each block. It creates high-resolution compensating signals while compressing wide-band video signals into NTSC TV signals, then multiplexes these signals. Finally, high-resolution signals are obtained by reversing the above rocessing at the receiving side.

CONCEPT OF MODE 1 AND MODE 2

It may have to pass further discussion to reach the final conclusion, which of these two,the side panel(pan & scan) system or the letter box system, should be exclusively applied to the Wide Aspect EDTV (ATV) System. The utmost we can do now is to present the overview of these two systems that we have been testing.

The concept is shown in Fig.1,Mode 1,the side panel system,separates the side panel signals into high-frequency and low-frequency components by using the frequency filter,and compresses the low-frequency components. It multiplexes then transmits these components to the invisible part of a screen through over scan of a TV screen,the low-frequency signals to the side edges of the screen and high-frequency components to the top and bottom edges.

The number of vertical scanning lines allowed on the main screen is restricted to 444. Scanning lines exceeding this limit are invisible because of the over scan of the home-use TV receiver.

Mode 2 is a letter box system,in which the whole screen is compressed to three fourth,and the number of vertical scanning lines that are valid is restricted to about 360,so that the high-frequency components of the whole screen are transmitted to the resulting top and bottom spaces equivalent to about 120 scanning lines.

The auxiliary signals transmitted to the top and bottom edges of the screen should be colored black or the like,in order to avoid a visual disturbance.

AN OUTLINE OF THE BLOCK THINNING-OUT TECHNIQUE

The block thinning-out technique is to divide a screen image into multiple blocks in order to limit the bandwidth and to compress the screen signals by thin-out the picture elements within each block. The wide-band image signals are band-compressed into NTSC TV signals as well as generating the high-resolution compensation signals, and the two signals produced are transmitted in multiplexing. The receiver side reverse the signal processing of the transmitter side to resume the original high-resolution signals. We have made the experiments using the block thinning-out technique in the study of EDTV-1 system, and extended the thought to apply the same technique to the wide-aspect EDTV system.

BLOCK SETTING AND INFORMATION SEPARATION

First,an image screen(luster) is divided into blocks, each consisting of a picture elements. (The blocks are basically 2-dimensional but include some 1-dimensional.) Then n picture elements

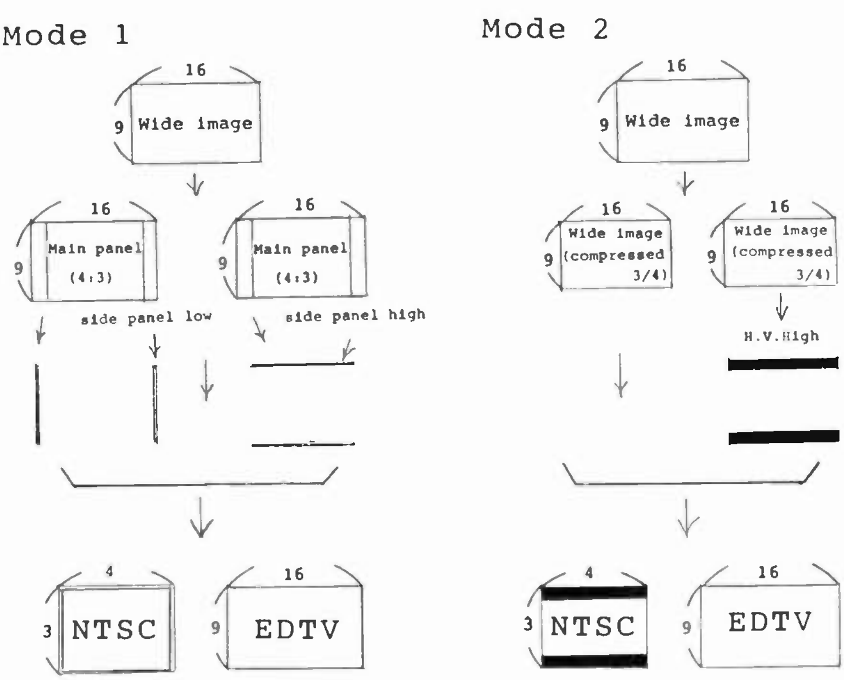


Figure 1. Concept of 2 Modes

(m›n) are extracted within each block and they are transmitted over the conventional channel.

On one hand, among the blocks having much high-resolution components are transmitted over the multiplexed channel, added with the respective addresses on the screen. At this time, if all the blocks are processed in this method, the original information of the transmitter side could be completely transmitted, however, it is generally required to select an appropriate number of blocks to be sent this way depending upon the capacity of multiplexed transmission channel(see Fig.2,Fig.3).

According to our experiments , when the original signals of common images with the horizontal bandwidth of 6MHz and vertical 480 lines are considered, it has been revealed that, if the number of transmission blocks (including the information in the range of about 4-6MHz) is about 10% of total blocks, it is sufficiently enough. Besides, there is a way to use the technique that requires no addressing information. In other words, when the transmission blocks are arranged horizontally and vertically in equal intervals on the screen, the address data are unnecessary. This method is good only if the capacity of multiplexed transmission channel is relatively large, and it gives an advantage that the algorithm of the transmitter and receiver becomes simpler.

WEIGHTING AND DOUBLE WEIGHTING

A) Weighting process

When the thinning-out of structured picture elements is to be made, if a elements are simply thinned-out from m elements, an image distortion at the low-band information occurs. In order to avoid the distortion, the process as described below is applied to all the blocks before the thinning-out. Fig.4 shows the process. For example, when m=4 and n=3, if c picture elements are thinned- out and then b and d picture elements are transmitted as they are, the above- mentioned distortion occurs. Therefore, let's replace b and d with b' and d' as the low-band information. In the relationship between b, c and d, the generation of b' and d' is made by the folowing weighting:

$$b' = 2/3b + 1/3c$$

$$d' = 1/3c + 2/3d.$$

In this case, a, b' and d' include the picture elements of low-band information and c includes that of high-band information. Picture elements c can be transmitted as it is. However, as the redundancy exists in the amplitude direction, it is more efficient to transmit the differences between c and the average values of b' and d' over the multiplexed transmission channel.

By reversing the process at the receiver side, c is generated from the

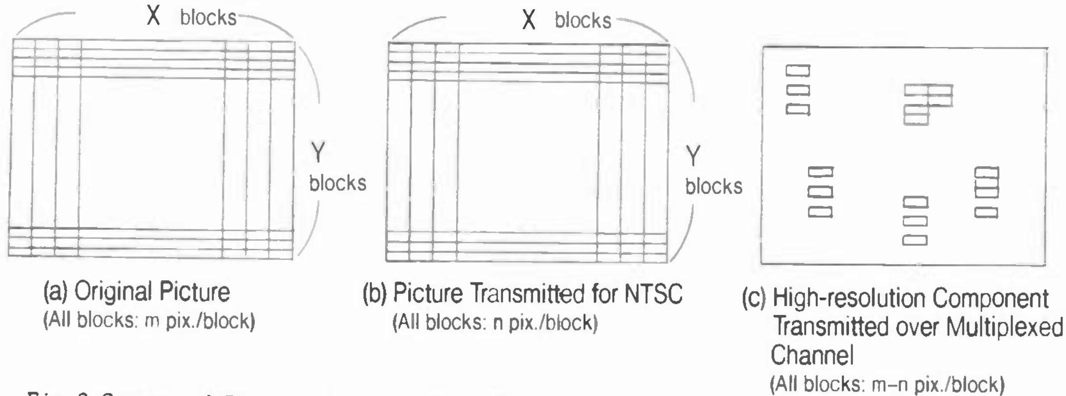


Fig.2 Separated Picture into blocks and Generation of Low and High Components
(Picture indicates horizontal one-dimentional blocks.)

average values of b' and d' and the high-band information, b and d are generated from c, b' and d'. Then the original high-band image can be regenerated.

B) Thinning-out by double weighting

The weighting process method to avoid the image distortion associated with the thin-out of blocks was described above.
A technique of weighting to equalize the signal spectra of each picture element will be discussed now.

A correction of the picture element amplitude value is made by a process

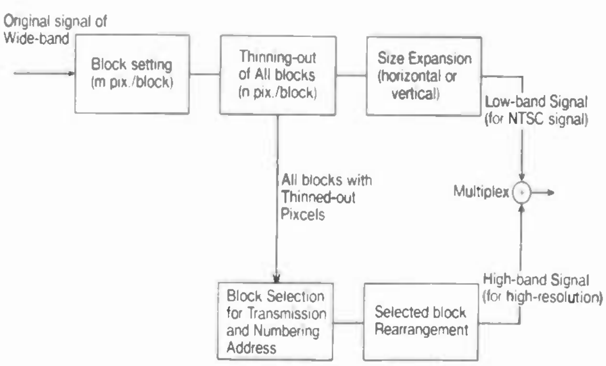


Fig.3 Generation of High and Low Frequency Components

applying a certain cofficient to the adjacent picture element amplitude. Qualitatively, the resultant amplitude value obtained by the above process of each picture element within the block is used for the correction of amplitude. In this case, if the correction values of picture elements in the block are evenly equalized, the purpose of the correction can be achieved.

This means that the correction of picture element amplitude value is applying a digital filter with over sampling against the original signals, and nothing but reduction alias by optimizing the characteristic of the filter.

A concrete example will be described taking a case when this method is applied to a vertical 1-dimensional block consisting of 4 picture elements.

In Fig.5 as previously explained, the weighting for positional correction is made using c and d, and then picture elements b' and d' are generated. Their amplitude values are :

$$b' = 2/3b + 1/3c$$

$$d' = 1/3c + 2/3d.$$

Next, a' is generated from the picture elements d of the immediately preceding block and a and b, in a manner that the spectra processed by the picture element corresponding to the position of a is approximate to that of b' and d'. When this is regarded as a digital filter, the picture element a' has the relationship of 1/6 4/6, and 1/6 with the surrouding picture elements, in order that the LPF's Nyquist characteristics is evenly distributed. Therefore, a' can be generated by:

$$a' = 1/6\,\underline{d} + 4/6\,a + 1/6\,b.$$

The vertical high-band information h is:

$$h = -1/3\,b + 2/3\,c - 1/3\,d,$$

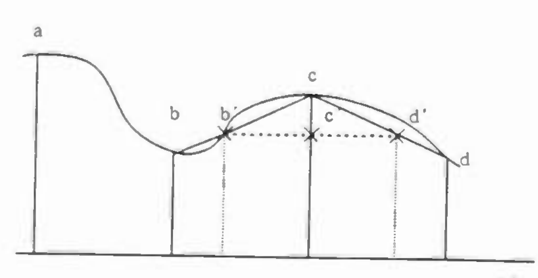


Fig.4 Weighting thinning-out process

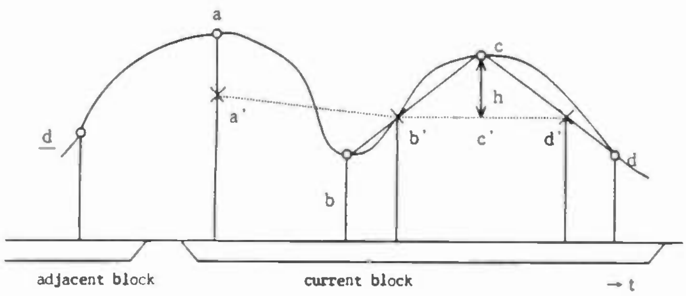


Fig.5 Correction of picture element amplitude at transmitting side.

and this becomes the information to be transmitted over the axiliary channel.

In Fig.6 the operation of process at the receiver is as described below. First, c is resumed from b' and d' and the high-resolution information h for compensation as follows:

$$c = 1/2\ b' + 1/2\ d' + h.$$

Then next, b and d are resumed by weighting the positional distortion correction by the following process:

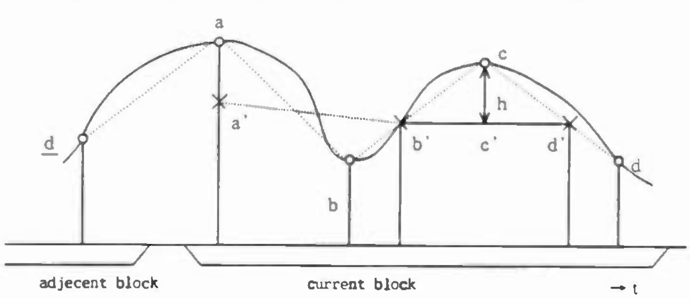


Fig.6 Resuming Process of picture element amplitude at receiving side.

$$b = 5/4\ b' - 1/4\ d' - 1/2\ h$$

$$d = -1/4\ b' + 5/4\ d' - 1/2\ h.$$

Finally, a is resumed from the picture element d at the most nearest position of the adjacent block and also from the picture elements a' and b as follows:

$$a = 3/2\ a' - 1/4(\underline{d} + b)$$

By this time, the adjacent block has been precedingly resumed.

According to the computer simulation, a good result is obtained as shown in Photo 1.

APPLICATION OF THE BLOCK THINNING-OUT TECHNIQUE TO MODE-2

Fig.7 shows a block diagram of the experimental system. Now, the signal processing will be discussed.

A) Signal processing of vertical direction

The TV camera output of progressive scanning of 525 lines is applied to a matrix and to pre-filter to limit the

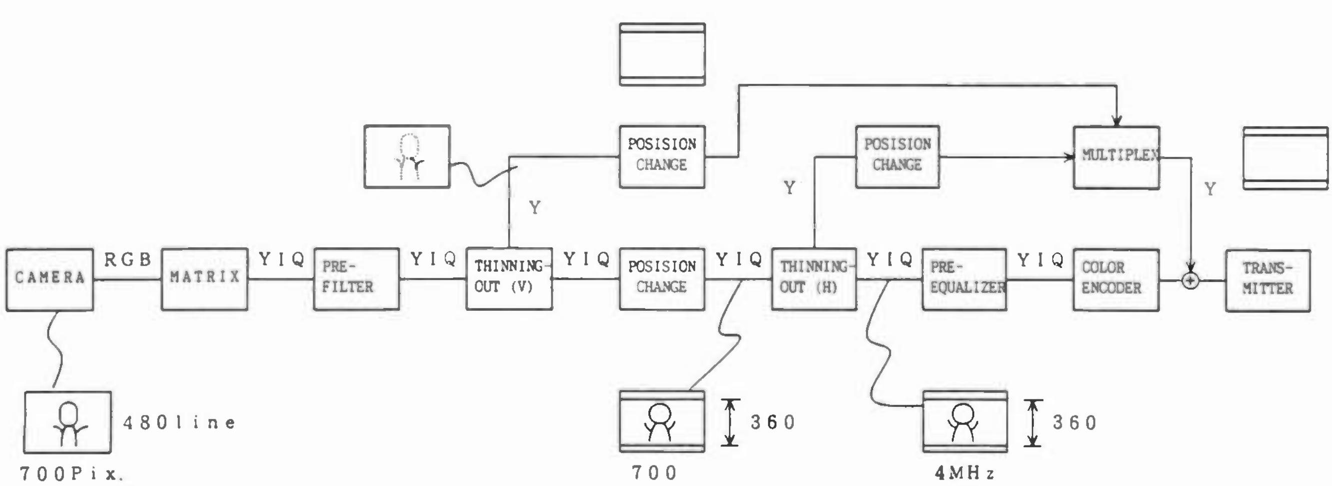
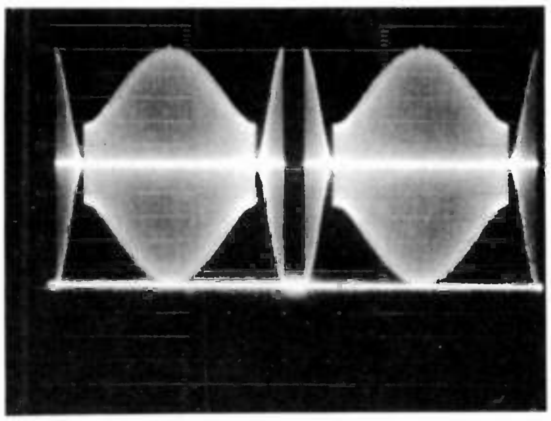
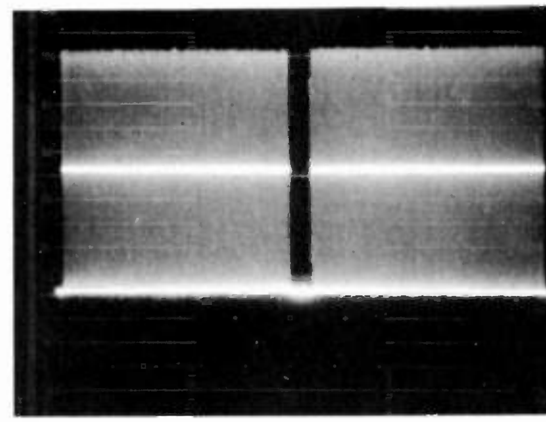


Fig.7 SYSTEM BLOCK DIAGRAM OF MODE-2 TRANSIMISSION





(a) Encoded picture (b) Decoded picture

Photo 1 Waveforms of computer simulated picture (C.Z.P) with Double Weighted Block Thinning-out technique

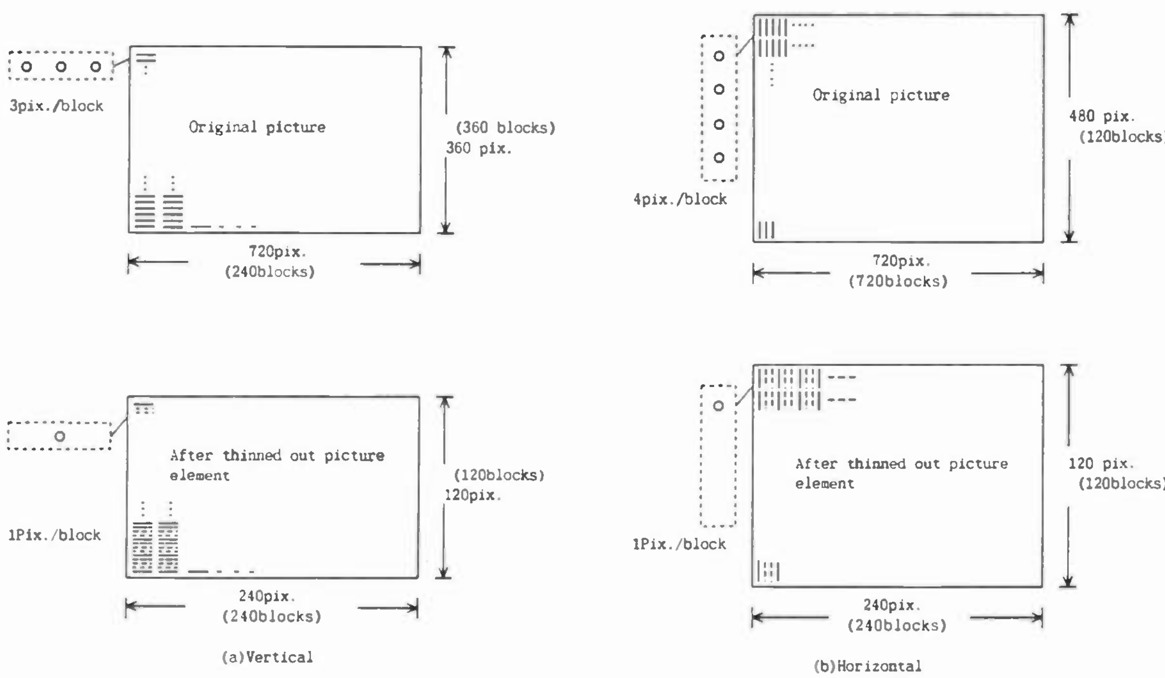


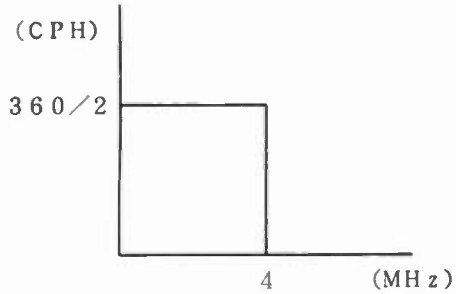
Fig.8 Block size and pixeles

diagonal regions of the horizontal and vertical bandwidths, and then first the thinning-out of vertical direction is made by setting the vertical direction blocks. A block size is 4 picture elements (4 scanning lines) as shown in Fig.8-a. Blocks are corrected with picture element amplitude values in order to prevent the distortion, and after making the weighting for equalizing the frequency spectra of each picture element (scanning line), the thinning-out of picture elements is made.

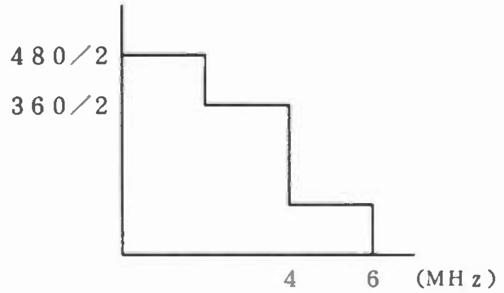
A single picture element is thinned-out per a block, and thus the screen size consisting of blocks after the thinning-out is reduced into a 3/4 in the height. This size is equivalent to 360 lines out of effective NTSC 480 scanning lines, and the reduced-size signals are put into the main signal.

On one hand, the thinned-out picture elements in total include a 1/4 in height of the original signals, and they are put into the upper and lower dark belt portions of letterbox screen.

Out of the thinned-out blocks, 1 of every 3 is selected in equal intervals to the horizontal direction, and the rest is not used.



(a) transmission



(b) receiver

Fig 9. H-V Picture bandwidth

The bandwidth of the signal components to the horizontal direction is reduced into about a 1/3 by this process.

B) Signal processing of horizontal direction

Next, the main signals are compressed to the horizontal direction by the horizontal block setting.

A block size is a unit of 3 picture elements as shown in Fig.8-b . Blocks are applied with the correction of picture element amplitude values and the weighting, and then 1 out of 3 picture elements are thinned-out. The screen signals consisting of blocks after the thinning-out are reduced into a 2/3 of size in horizontal direction. This size is expanded 3 times to resume the original size and put into the main panel.

The screen signals consisting of the thinned-out blocks are reduced into 3 times and put into the upper and lower belt portions of the letterbox screen. Those signals, as in the previously mentioned block processing of vertical direction, 1 out of every 4 is selected in equal intervals to the vertical direction and the rest is not used.

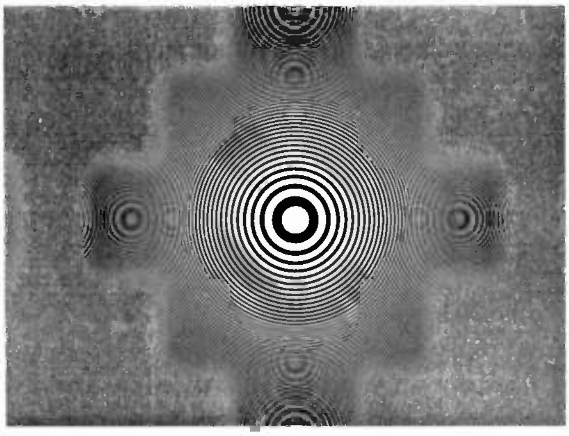
By this processing, the signal elements of the vertical direction now have the bandwidth of about a 1/4 of the original signals.

C) Signal transmission and the receiver's process

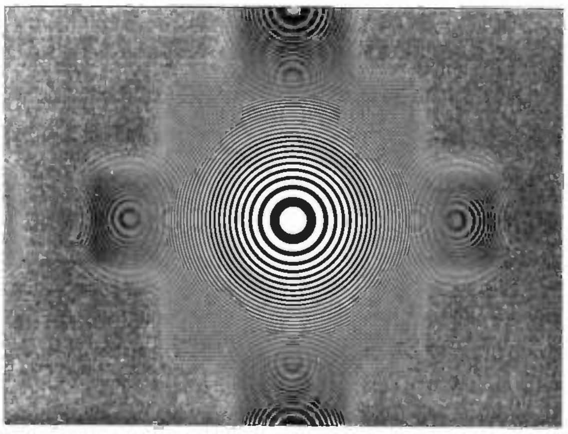
The horizontal and vertical bandwidth of the main panel portion after applying the thin-out of the horizontal direction are, as shown in Fig.9-a , compatible with the NTSC (EDTV-1) system. However, the horizontal frequency characteristics of the signals is slight modified by a pre-equalizer before input to the color encoder. This is to maintain a flatness of the frequency in the range that is compatible with the conventional receivers.

Besides, before the signal resume processing, the receiver make a reverse compensation (post-equalizer) with the complementary characteristics.

The previously-described high-resolution horizontal and vertical signal components are transmitted over the upper and lower belt portions of letterbox screen in the quadrature amplitude modulation or the frequency multiplexed transmission channel.



(a) Original picture



(b) Decoded picture

Photo 2 Vertical characteristics of Mode 2
(Computer simulated using a C.Z.P)

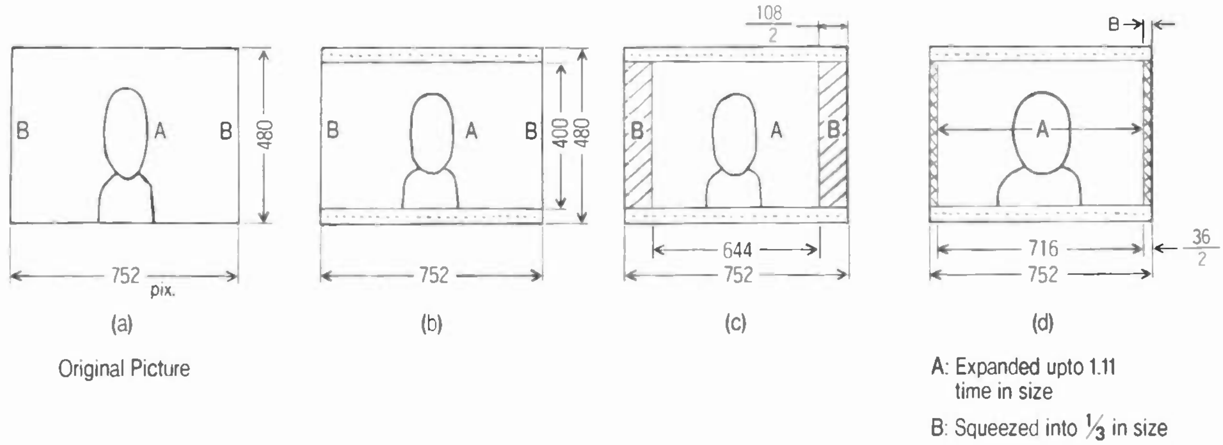


Fig.10 Generation Process of Transmitting Picture of Mode 1 (see Fig.11)

At the receiver side, the signals are resumed in the reverse processing of the transmitter. The horizontal and vertical band-width of an image regenerated by this method is shown in Fig.9-b.

Photo 2 is of a computer simulation result of the system and shows the vertical characteristics using a circular zone plate.

APPLICATION OF THE TECHNIQUE TO MODE-1

In the side panel system, the important subjects of study are signal processing at the splicing points between the main panel and the side panels, and the channel for transmission of the side panel information.

A method we examined this time is setting blanking portions, 80 lines in total at the upper and lower ends of screen (lusters) and about 5% of the horizontal width at the left and right ends, and to transmit the side panel information over the blanking portions(see Fig.10).

A) Signal processing of the transmitter

Fig.11 shows a block diagram of the transmitter. The signal processing of the transmitter side includes first passing the TV camera signals to horizontal and vertical lowpass filters, and then converting scanning lines by the block thinning-out technique to the vertical direction (vertical blocks of n=6 and n=5).

Camera signals include 480 vertical scanning lines, but the number of scanning lines of visible image is reduced into a 5/6 which is 400 lines by the above process.

The information corresponding to the reduced 80 lines is band-limited to horizontal 1.2MHz and put into the upper and lower blanking portions.

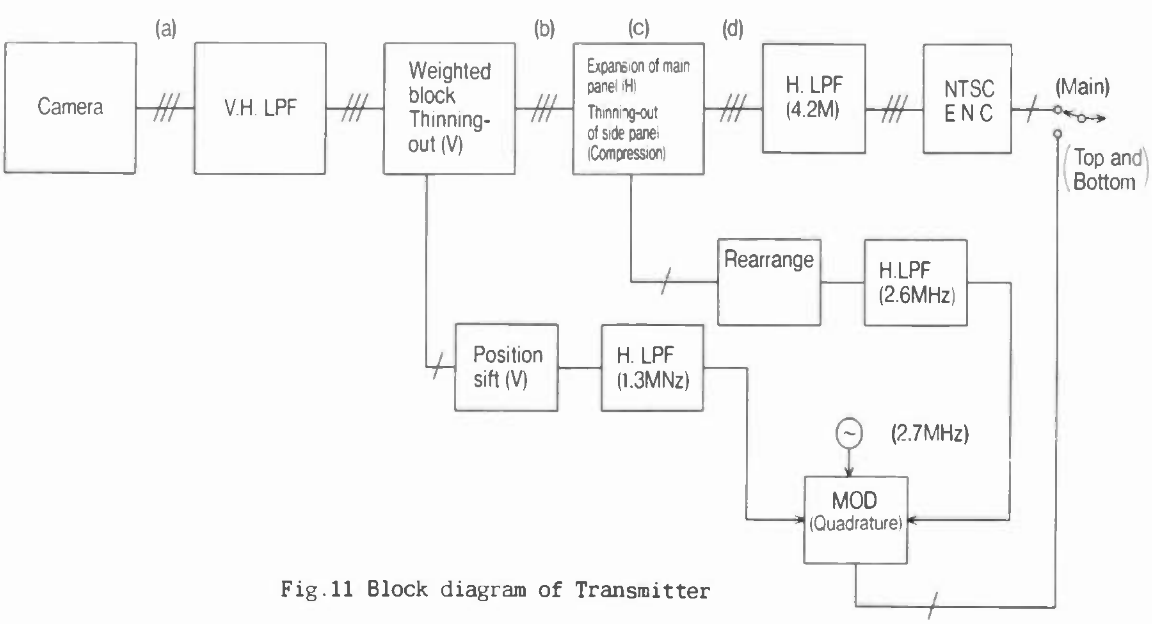


Fig.11 Block diagram of Transmitter

Next, only for the side panel regions of the main panel at the left and right ends are applied with the block thin-out technique. The side panels width in total is about 108 picture elements, and the main panel width is 644 picuture element.

Applying the block thinning-out technique to the side panel specifying m=3 and n=1 results in reducing the side panel width into 1/3. Thus, the low-band components of the side panels can be included in the width of 36 picture elements at the left and right ends of a screen. The picture element signals of the thinned-out blocks are band-limited

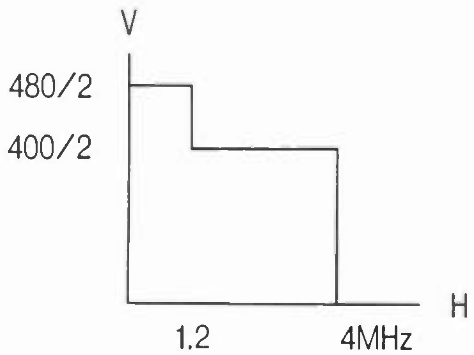


Fig.12 Bandwidth of both Main-panel and Side-panel

the thinned-out blocks are band-limited into horizontal 2.8MHz and put into the upper and lower blancking portions.

There are two kinds of signals contained in the upper and lower blanking portions, one for vertical and the other for horizontal siganls, but their bandwidth are different, 1.2MHz and 2.8MHz, respectively. Those signals are transmitted using the quadrature amplitude modulation (with the carrier frequency of about 2.8MHz).

The above processing is illustrated in Fig.12,Fig.13.

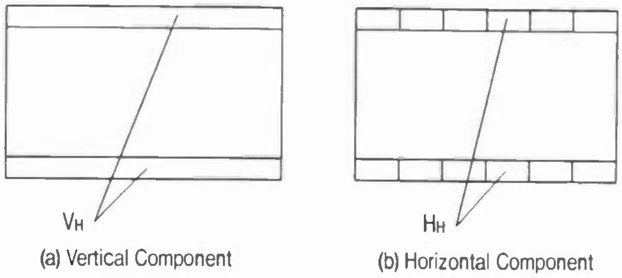


Fig.13 Horizontal and Vertical High-Frequency Components

B) Processing of the receiver

The receiver makes the reverse signal processing of the transmitter.The regenerated bandwidth of the main panel is vertical 480 line and horizontal 4.2MHz.This is the same as the original signals. The regenerated bandwidth of the side panels is the same as the main panel,but the bandwidth of the diagonal direction is slightly narrower.

POSTFACE

Various types of signal processing technorogies are provided for separating side panel signals and high-resolution-oriented signals from the original picture before transmission . But,for actual use, various factors,such as the performance, circuit complexity,cost and feasibility, must be considered. In addition to the block thinning-out technorogy introduced in this report,we have been testimg several other technorogies which we are going to introduce in the separate reports if given a chance.

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(3) S.Takayama, J.Urano, T.Ito, Y.Araki "EXPERIMENT WITH SECOND GENERATION EDTV SYSTEM" SMPTE Preprint No.131-09, Nov.1989.

COMPATIBLE MUSE SYSTEMS FOR TERRESTRIAL BROADCASTING OF HDTV SIGNALS—ADTV

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SUMMARY:

Compatible MUSE (ADTV) systems have been developed by NHK for terrestrial broadcasting of HDTV. They consist of NTSC-Compatible MUSE-6 and Narrow MUSE. The NTSC-Compatible MUSE-6 is compatible with the NTSC system, and The Narrow MUSE system is designed for simulcasting.

Experimental encoders and decoders of these systems were demonstrated at the NAB convention in April 1989.

The RF modulation and interrelated interference issues with the introduction of an ATV service in the existing terrestrial environment are important, especially for a simulcast system. A new modulation method for the Narrow MUSE signal has been developed. The NTSC-Compatible MUSE-6 and the Narrow MUSE have been prepared for testing of ATTC in 1990. This paper describes the technical characteristics of these systems and the modulation schemes.

INTRODUCTION

NHK has developed the MUSE system[1] for the transmission of high definition television via direct broadcast satellite. In Japan, daily one hour experimental HDTV satellite broadcasting in June 1989 by using BS(broadcasting satellite)-2 and practical broadcasting is expected to begin via BS-3, scheduled to be launched in 1990.

In the United States, advanced television (ATV) systems for terrestrial broadcast are under investigation. The FCC's tentative decision for the development of ATV puts emphasis on compatibility with the current channel allocation and the NTSC system. In line with requests by NAB and AMST, NHK has been engaged in the development of Compatible MUSE systems (ADTV) for terrestrial broadcasting using 1125/60 studio signals. Two different systems have been proposed to the FCC advisory committee for testing of ATTC in 1990: NTSC-Compatible MUSE-6 and Narrow MUSE. The NTSC-Compatible MUSE-6 is compatible with the NTSC system, and the Narrow MUSE system is designed for simulcasting.

Experimental encoders and decoders of these systems were demonstrated at the NAB convention in April 1989.[2]

This paper mainly describes the 7.5Hz offset frequency multiplexing method for expanding the horizontal bandwidth of luminance signals used in the NTSC-Compatible MUSE-6 and the modulation schemes solving interference issues with the introduction of the Narrow MUSE system in the existing terrestrial environment.

NTSC-COMPATIBLE MUSE-6

1.Features and System Concept

The NTSC Compatible MUSE-6 has the following features:

* NTSC and 6 MHz channel compatibility,
* Approx. twice NTSC resolution for the static portion of the picture,
* Upward compatibility with NTSC-Compatible MUSE-9, and
* 2-channel digital audio.

The NTSC-Compatible MUSE-6 system uses the letter box method. The reason for our choice is that the difference in picture quality between a center panel picture and a side panel picture might occur in the side panel method, and the letter box method enables the entire 16:9 picture to be viewed on conventional receivers, making for compatibility in program content.

2.System Configuration

Figure 1 shows the block diagram of the encoder of NTSC-Compatible MUSE-6.

1125/60 HDTV signals are digitized and sent through the motion detector and the pre-processor. The motion-detected signal is used in the noise reducer, line-number converter and high-frequency component multiplexing block in the encoder. After noise reduction, the signal is fed to the line-number converter and converted from 1125 lines to 750 lines. Then the field-rate converter changes the field rate from 60Hz to 59.94Hz. Y,I, and Q signals, whose scanning parameters are now 750 lines and 59.94Hz, go into the letter box processor. The signal is divided into vertical low and high frequency components. The low frequency component signal is sent by the 345 lines that correspond to the center part of a 525 line picture. The vertical

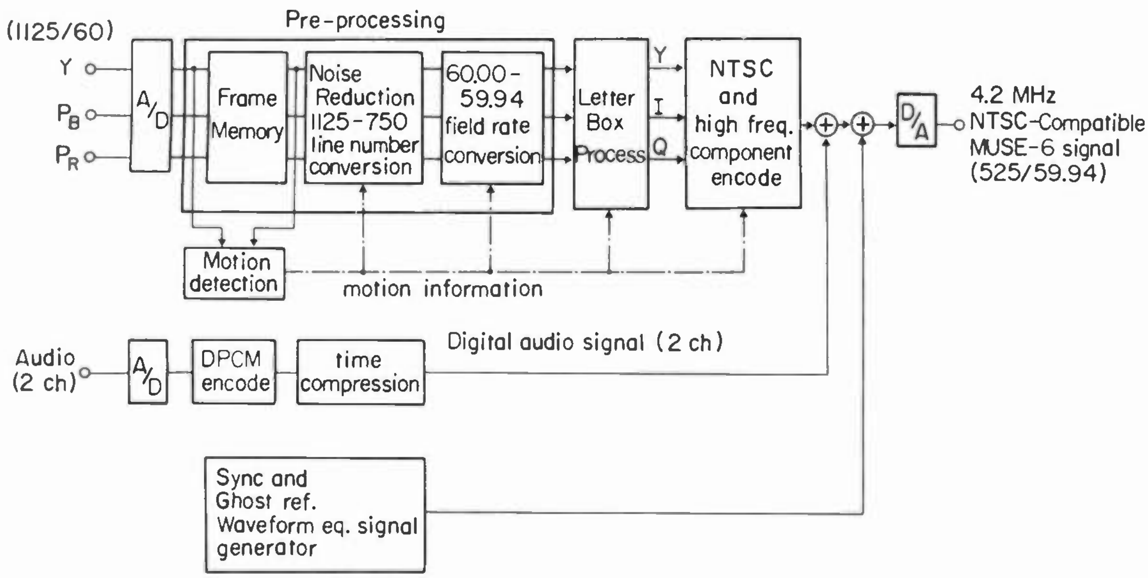

Figure 1. NTSC-Compatible MUSE-6 Encoder

high frequency component signal is band-limited horizontally, time-axis compressed, then multiplexed in the top and bottom portion. The vertically pre-processed Y, I, and Q signals go into the NTSC and high frequency component encoder (modified NTSC encoder) where frequency-multiplexing of the luminance high frequency components is performed, and high frequency components of I and Q signals are frequency-shifted and field-sequentially-multiplexed into the top and bottom portions. Before transmission, a reference signal for ghost cancellation and waveform equalization is added. The decoder performs an approximate inverse.

The two-channel digital audio signal is available as well as the conventional FM audio signal. The input 14 bits and 32KHz sampling digital audio signals are compressed 8 bits in a DPCM encoder called DANCE, and multiplexed in the horizontal blanking intervals of the video signal.[3]

3. Improvement of luminance resolution

The horizontal high frequency components are amplitude-suppressed and frequency-interleaved into the area between 1.9MHz and 3.9MHz.

Two subcarriers are used to shift the horizontal high frequency components into the low frequency area. As shown Figure 2, the area #3 is shifted into the area #2 by the subcarrier fs1, and #4 is shifted into #2 by the subcarrier fs2. The subcarrier phase is shifted 90 degrees by line and by frame, and -90 degrees by field as shown in Figure 3 ('shuffling'), resulting the eight field sequence. Based on this figure, the subcarrier, s(x,y,t), is given as

$$s(x,y,t)=\cos 2\pi \left(x+\frac{1}{8}y-\frac{3}{8}u+\frac{1}{4}v\right) \quad \text{—— (1)}$$

$$\begin{cases} -\frac{3}{8}u+\frac{1}{4}v=\frac{1}{8}t & : u=\text{even} \\ -\frac{3}{8}u+\frac{1}{4}v=\frac{1}{8}t-\frac{1}{2} & : u=\text{odd} \end{cases}$$

where x=1 corresponds to one cycle of fs1 or fs2
y=1 corresponds to one scanning line spacing
u=1 corresponds to field interval (u=0,1)
v=1 corresponds to frame interval (v=0,1,2,3,...)
t=1 corresponds to field interval (t=0,1,2,3,4,...)

Taking the scanning structure (two dimensional sampling) into consideration, Eq.(1) can be written as

$$\begin{aligned} s(x,y,t)=&\sum_m\sum_n \cos 2\pi \left(x+\frac{1}{8}y+\frac{1}{8}t\right)\cdot\delta(y-2n,t-2m) \\ &+\sum_m\sum_n \cos 2\pi\left(x+\frac{1}{8}y+\frac{1}{8}t-\frac{1}{2}\right) \\ &\cdot\delta(y-(2n+1),\ t-(2m+1)) \quad \text{— (2)} \end{aligned}$$

The three dimensional Fourier transform of Eq.(2), S(X,Y,T), gives the arrangement of subcarriers in the three dimensional frequency domain.

$$\begin{aligned} S(X,Y,T)=&\sum_m\sum_n [\delta(X-X_0),\ Y+\frac{8n+1}{8}Y_0,\ T-\frac{8m-3}{8}T_0) \\ &+\delta(X+X_0),\ Y-\frac{8n+1}{8}Y_0,\ T+\frac{8m-3}{8}T_0)] \\ &+\sum_m\sum_n [\delta(X-X_0),\ Y+\frac{8n-3}{8}Y_0,\ T+\frac{8m+1}{8}T_0) \\ &+\delta(X+X_0),\ Y-\frac{8n-3}{8}Y_0,\ T-\frac{8m+1}{8}T_0)] \\ &\text{—— (3)} \end{aligned}$$

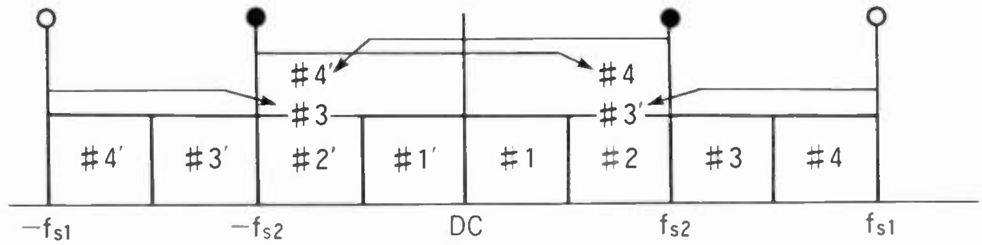

Figure 2. Spectrum of Frequency-Multiplexed Signal

where X0=fs1 or fs2
Y0=2*525 lines/picture height
T0=60 Hz

Figure 4 shows the arrangement of the subcarrier, and Figure 5 shows its projection onto the X-T (horizontal frequency - temporal frequency) plane. Figure 5 contains both subcarriers, fs1 and fs2, and the color subcarrier. The interleaved spectrum in the three dimensional frequency domain is also shown in this Figure 4. As is clear from this Figure 5, high frequency components interleaved in the area #2 can be separated by temporal frequency axis.

By shifting the subcarrier phase by line, frame, and field, the high frequency components, #3 and #4, are interleaved into the relatively high frequency area of the vertical domain in #2 as shown in Figure 4. Therefore, the interleaved high frequency components are less perceptible because of the visual system.

NARROW MUSE

1.Features and System Concept

The Narrow MUSE system, intended for ATV simulcast as set out in the FCC's tentative decision, has the following features:

- 6MHz RF channel compatibility,
- Highest picture quality among the Compatible MUSE systems, and
- 4-channel digital audio.

In order to use a single 6MHz RF channel and cost-effectively to utilize the bandwidth compression technology used for MUSE,[1] the Narrow MUSE system has been developed. In Narrow MUSE, 750 scanning lines, derived from a 1125/60 signal, are used, implying reduced vertical resolution when compared to the original signal. However, vertical resolution of 750 lines/picture height nearly corresponds to that of an 1125 line interlace system with a Kell factor of around 0.65. If the decoded Narrow MUSE signal is up-converted from 750 lines to 1125 lines, fully using the transmitted 750-line information, then a displayed picture of high quality can be expected.

To reduce the signal bandwidth, the same multiple sub-sampling technique as in

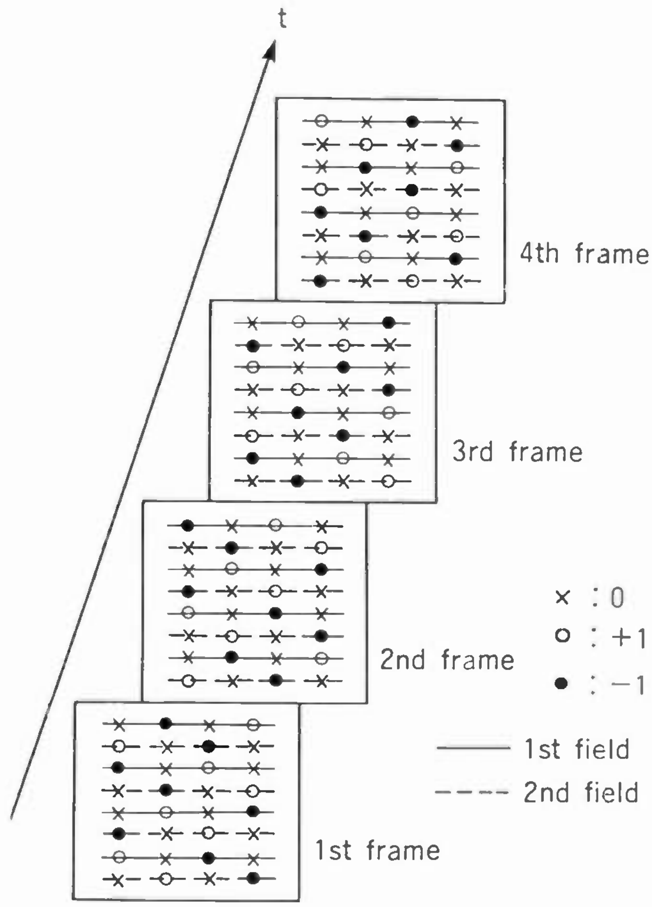

Figure 3. Pattern of Subcarrier Phase

the MUSE system is adopted. The bandwidth of Narrow MUSE signals is 4.86MHz after band-reduction.

2.System Configuration

There are two ways to build the Narrow MUSE system, a specialized encoder/decoder or an adapter. The experimental Narrow MUSE equipment has been developed by using an interface adapter method as shown in Figure 6. Signal processing is done as follows:

1)In the encoder, an input signal is digitized and converted from 1125 lines to

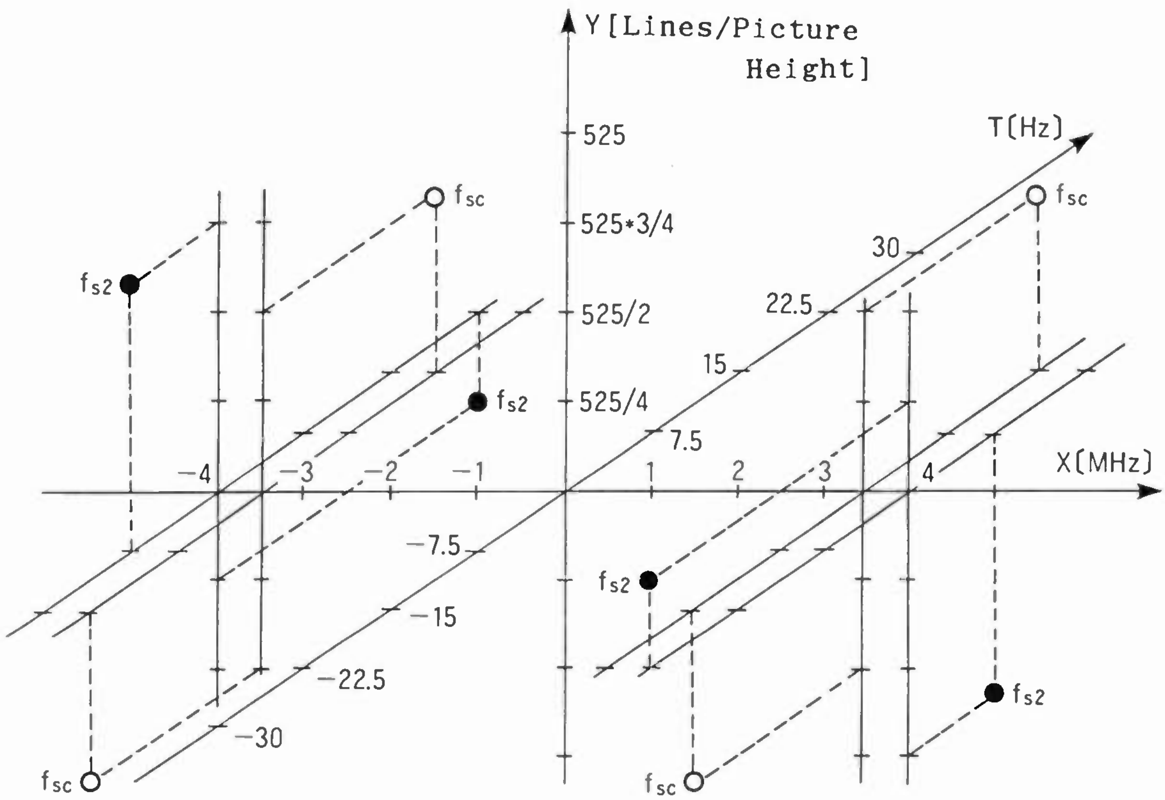

Figure 4. Subcarrier Arrangement in 3-Dimensional Frequency Domain

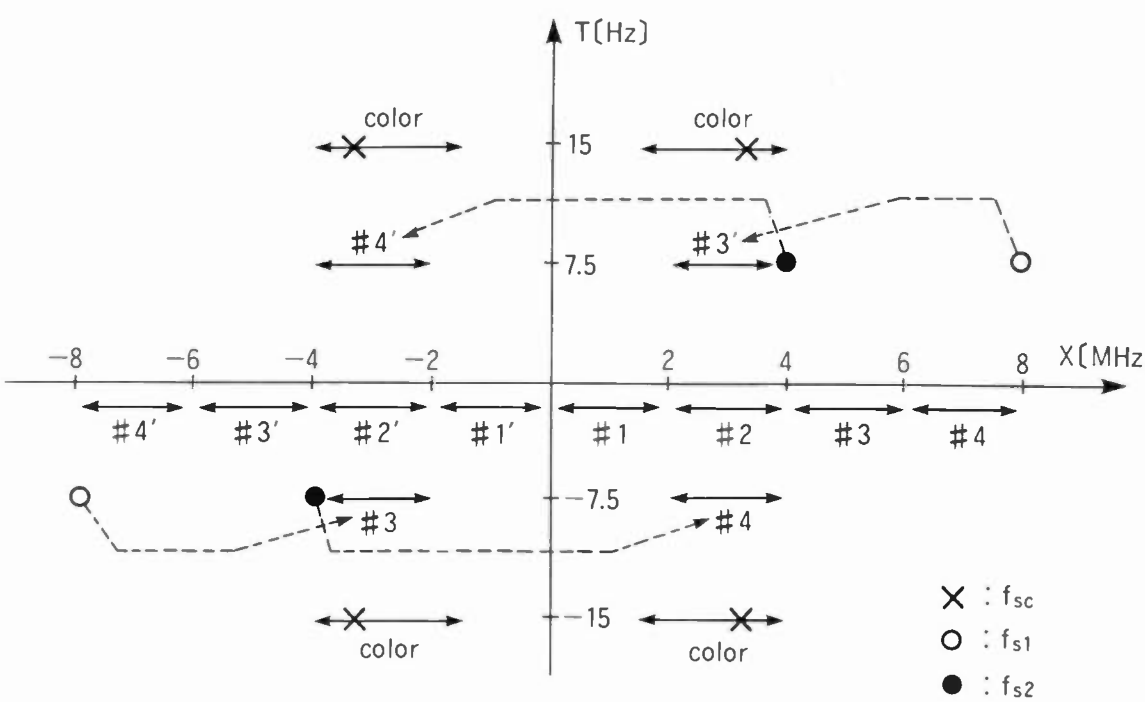

Figure 5. Modulated Signal Spectrum in X-T Domain

750 lines.

2)The 750 lines signal goes to the interface adapter 1 and the data format, 1188 picture elements x 750 lines, is changed to the 1440 picture elements x 1125 lines format of the MUSE input signal by adding dummy picture elements.

3)The signal is processed by the MUSE encoder.

4)The MUSE encoded signal is changed to the Narrow MUSE format in the interface adapter 2. The bandwidth of 4.86MHz is obtained by removing the unnecessary portions of processed signals.

5)The decoder performs the inverse process and the result is displayed on an 1125/60 HDTV display.

The audio DPCM encoder is the same as the MUSE system using DANCE. Digital audio signals, multiplexed in the vertical blanking interval, are decoded by the same decoder as the MUSE system.

3.Modulation Method

In the FCC's tentative decision, it is required that an additional channel for an augmentation channel system or a simulcast system should be assigned only in the current channel allocation.

The recent studies of the FCC advisory committee lead to some conclusions as follows:

1)That a contiguous 3MHz augmentation channel can be assigned to only 75% of the current licensees and only 67% of the licensees could receive a contiguous 6MHz assignment.

2)That an additional 6MHz non-contiguous channel assignment would be available to cover 99.7% of the current licensees if:

-"taboo" restrictions can be eliminated;

-the co-channel transmitter spacing can be reduced to about 160Km, as opposed to the currently required separations which vary from 240Km to 350Km.

3)That to operate with a co-channel spacing of 160Km requires a reduction of effect of the interfering ATV signal approximately 30dB for interference-free reception.

For ATV systems, it is required that they are not casing interference and are not affected by interference in the current channel allocation. Therefore, a modulation method is an important subject to be solved.

The Narrow MUSE system adopts a frequency-multiplexing modulation method as follows.

The modulation scheme of this method is shown in Figure 7. An input Narrow MUSE signal is divided into two components by low-pass and high-pass filters. The spectrum allocation is shown in Figure 8.

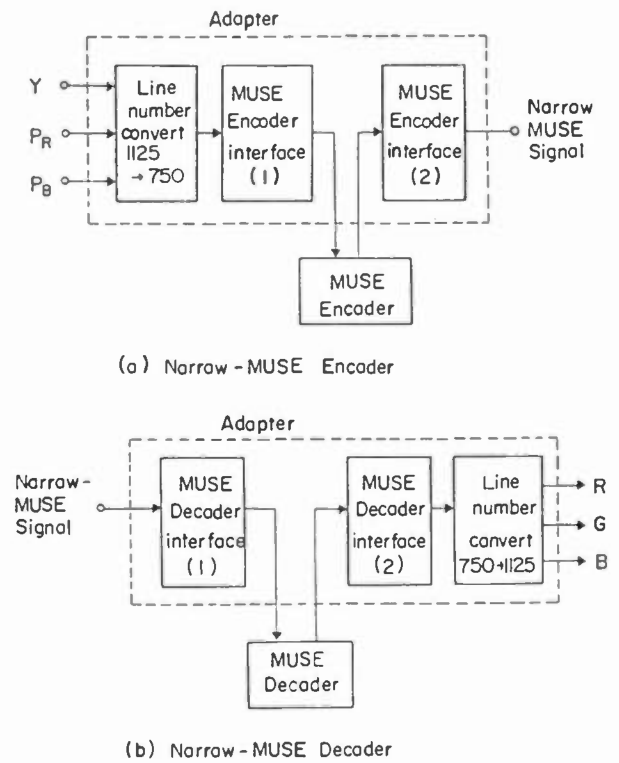

Figure 6. Narrow MUSE System Using MUSE System and Its Adapter

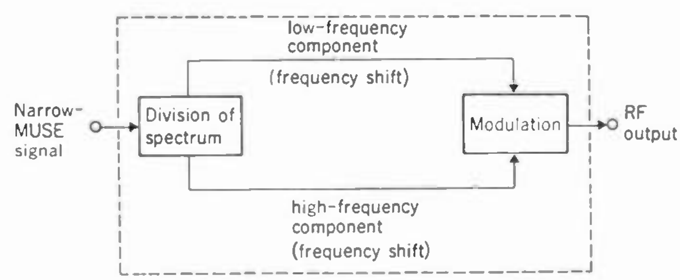

Figure 7. Block Diagram of Frequency Multiplexed modulation Method

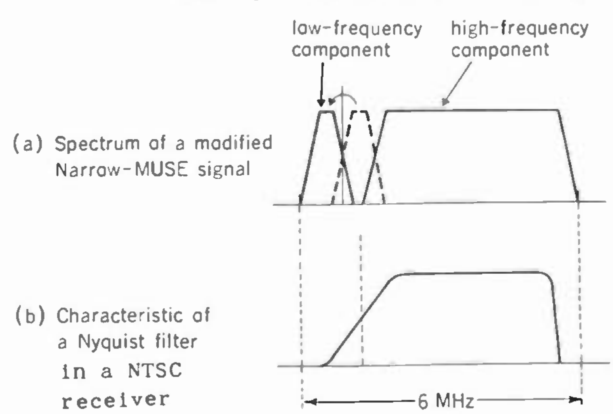

Figure 8. Spectrum of Modulated Narrow MUSE Signal

Modulated Narrow MUSE signals have no spectrum in the area corresponding to the low frequency component of the co-channel NTSC signal. Then high and low frequency components are transmitted in analog format. In a NTSC receiver, co-channel interference caused by the low frequency component of Narrow MUSE signals is less visible because of the reduction of its level through a Nyquist filter.

CONCLUSION

Compatible MUSE systems are for terrestrial broadcasting of HDTV. Among them, NTSC-Compatible MUSE-6 and Narrow MUSE have been prepared for testing of ATTC. In NTSC-Compatible MUSE-6, the technical characteristics have been improved, and the modulation method solving interference issues with simulcasting has been developed for Narrow MUSE.

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A PRACTICAL APPROACH TO APPLYING THE MS STEREO MICROPHONE

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ABSTRACT

The MS (Mid-Side) microphone is an effective and versatile stereo production tool available to the broadcast sound engineer. It provides a convenient, adjustable single-point stereophonic pickup of both a wide sound stage and overall ambience. Stereo imaging can be very accurate, and full mono compatibility is inherent.

The basic concepts of stereo reproduction are presented, especially in relation to the MS microphone. Understanding these concepts helps the sound engineer to properly use the MS microphone and obtain the desired stereo effect. A number of examples illustrate typical applications.

INTRODUCTION

If accurate, realistic stereo reproduction is the objective of the broadcast sound engineer, then the MS (Mid-Side) stereo microphone is an effective tool to use in achieving this objective. As with any other tool, the engineer must develop an understanding of its functions and capabilities to use it most effectively.

This understanding should include a knowledge of how stereo images are obtained and reproduced, the MS microphone characteristics which provide quality stereo and mono pickup, the level of control the MS microphone allows over the stereo effect, and how to apply the MS microphone to stereo production.

BASIC STEREO CONCEPTS

The main goal of stereo reproduction is to create realistic images using two or more loudspeakers. For simplicity, the two-loudspeaker case will be considered, though commercially available Stereosurround™-type decoders (such as Dolby Surround® and Shure HTS®) can improve imaging using additional loudspeakers. A typical listening setup is shown in Figure 1. The listener and the two loudspeakers are at the corners of an equilateral triangle, with the stereo stage reproduced between the loudspeakers. Stereo images can be created when there are intensity (level) or time differences between the signals reproduced by each loudspeaker. Left (L), left-center (LC), center (C), right-center (RC) and right (R) images can be localized in the reproduced stereo stage as shown. It should be noted that stereo requires loudspeaker reproduction. Headphones, on the other hand, provide binaural reproduction, and totally different principles apply. Although headphone monitoring is sometimes necessary for isolation, or in field production, loudspeakers must be used to monitor the stereo perspective.

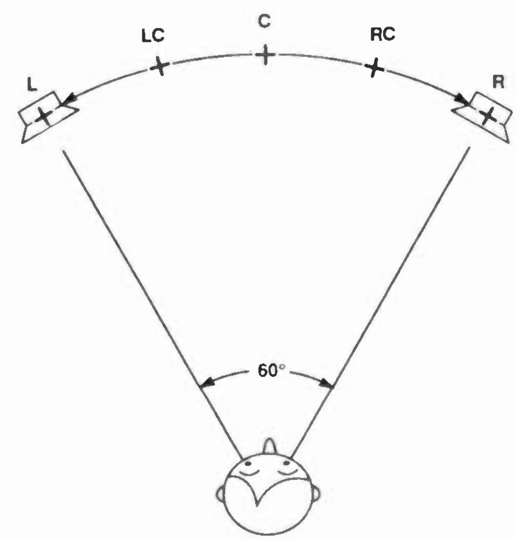


Figure 1
Two-Loudspeaker Stereo Monitoring

Intensity Stereo

Intensity stereo is created when a signal is reproduced at two loudspeakers at different levels, but with no time differences. If an audio signal (representing, for example, an announcer's voice) is delivered to the left and right loudspeakers at equal levels and the same polarity (in-phase), the resulting image is centered between the loudspeakers. If the signal at one loudspeaker is more than 15 dB louder than the signal at the other, the image is localized at the louder loudspeaker. By varying the left-to-right level ratio from +15 dB to -15 dB, the image is moved smoothly and fully between the two loudspeakers. When the signals at each loudspeaker are of opposite polarity (180° out of phase), the image is diffuse and cannot be localized. If the left and right signals are electrically summed to mono, the frequency response of the original signal is maintained; therefore, intensity stereo is mono-compatible.

Panpot stereo (the technique used in almost all popular music recordings) is a familiar example of intensity stereo. Multiple sound sources are picked up (usually by closely-positioned microphones) and individually positioned in the reproduced stereo stage using the panpots on the mixing console. Imaging can be excellent with this approach, but the close microphone positioning does not often allow good ambience pickup. Of course, artificial ambience may be added using time-delay-based effects such as electronic reverberation, but in many instances pickup of the natural environmental ambience is desirable. Besides, the relatively large amount of equipment required may be inconvenient, especially for field production.

In many cases, a coincident microphone technique can conveniently provide good intensity stereo with ambience pickup. A coincident microphone technique uses two directional microphones, with their cartridges as close together as physically possible, to provide stereo signals with intensity differences but without significant time differences. (A directional microphone is one that rejects sound from some incident angle or angles. The cardioid or unidirectional microphone is the most familiar, but supercardioid, bidirectional, and shotgun microphones are other examples.)

Time-Based Stereo

Time-based stereo results when a signal appears at equal levels in both loudspeakers, but with a time difference. If the signal at one loudspeaker leads the signal at the other loudspeaker by 3 milliseconds or more, the image is localized at the leading loudspeaker. By varying the left-to-right time difference from +3 milliseconds to -3 milliseconds, the image can be moved fully between the loudspeakers, though not as smoothly as with intensity stereo. Unfortunately, summing the signals to mono results in comb filtering, with more severe combing for longer time differences. Figure 2 illustrates an idealized comb filter produced when the outputs of two identical flat-response microphones are summed, positioned so that the time-of-arrival difference is 3 milliseconds and equalized in level. Since the comb filter corrupts the original signal, time-based stereo is not fully mono-compatible. For this reason, the classic spaced microphone techniques (which rely mainly on time- of-arrival differences for stereo cues) must be monitored closely for mono reproduction anomalies in broadcast applications. On the other hand, coincident microphones exhibit practically no time-of-arrival difference, and do not generate comb filtering in mono reproduction.

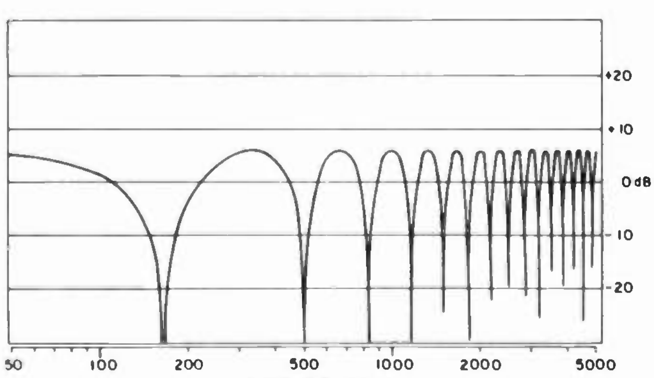


Figure 2
Comb Filter – 3 Millisecond Delay

THE MS MICROPHONE

The ideal coincident microphone technique is realized in the MS microphone technique. This technique employs a mid (M) cartridge which directly picks up the mono sum signal, and a side (S) cartridge which directly picks up the stereo difference signal (analogous to the broadcast stereo subcarrier modulation signal). Although two individual microphones may be used, single- unit MS microphones are more convenient and generally have closer cartridge placement. Figure 3 indicates the pickup patterns for a typical MS microphone configuration. The mid cartridge is oriented with its front (the point of greatest sensitivity) aimed at the center of the incoming sound stage. A cardioid (unidirectional) pattern as shown is often chosen for the mid cartridge, although other patterns may also be used. For symmetrical stereo pickup, the side cartridge must have a side-to-side facing bidirectional pattern (by convention, the lobe with the same polarity as the front mid signal aims 90° to the left, and the opposite polarity lobe to the right).

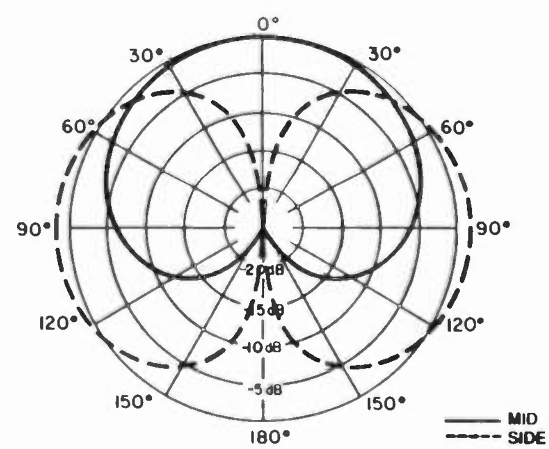


Figure 3
MS Microphone Pickup Patterns

Creating Intensity Stereo with the MS Microphone

In a stereo FM or television receiver, the mono sum baseband signal and the stereo difference subcarrier signal are demodulated and then decoded, using a sum-and-difference matrix, into left and right stereo audio signals. Similarly, the mid (mono) signal and the side (stereo difference) signal of the MS microphone may be decoded into useful left and right stereo signals.

The mid cartridge signal's relation to the mono sum signal, and the side cartridge signal's relation to the stereo difference signal, can be expressed simply by the following equations:

$$M = 1/2\ (L + R) \tag{1}$$

$$S = 1/2\ (L - R) \tag{2}$$

Solving for the left and right signals,

$$L = M + S \tag{3}$$

$$R = M - S \tag{4}$$

Therefore, the left and right stereo signals result from the sum and difference, respectively, of the mid and side signals.

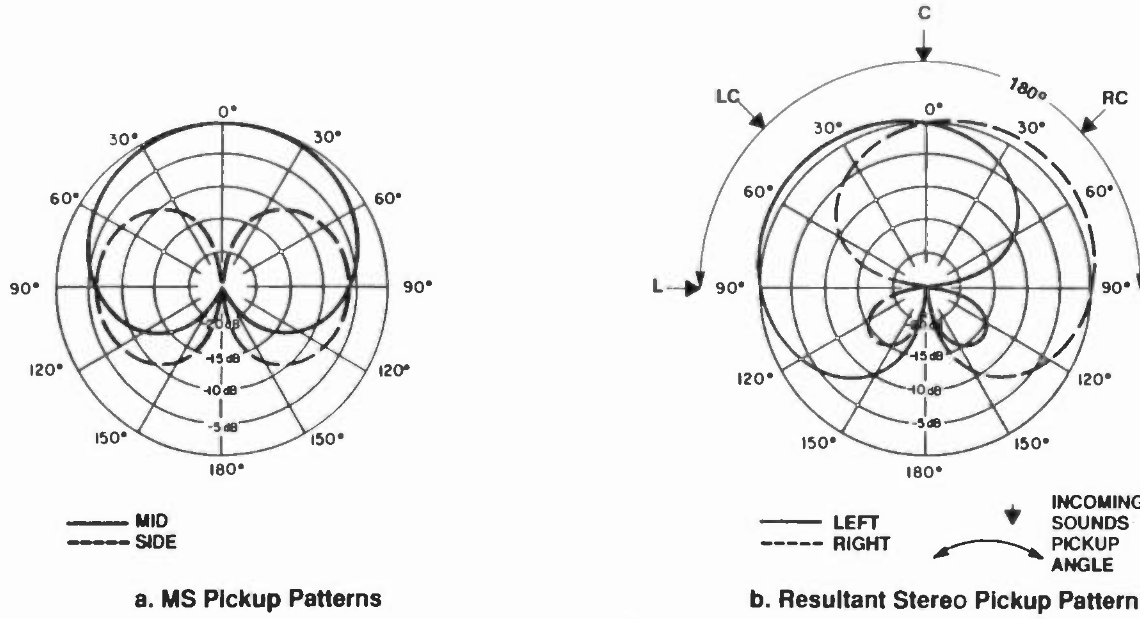


Figure 4
MS Microphone: Side Level 6.0 dB Lower Than Mid Level

These stereo signals can be obtained by processing the mid and side signals through a sum-and- difference matrix, implemented with transformers or active circuitry. This matrix may be external to the MS microphone or built-in.

The left and right stereo signals exhibit their own equivalent pickup patterns corresponding to, respectively, left-forward- and right-forward-facing microphones. Figure 4A illustrates an MS microphone with a cardioid mid pattern, and the required bidirectional side pattern having a maximum sensitivity 6.0 dB lower than the maximum mid sensitivity. The resultant left and right pickup patterns are shown in Figure 4B. The small rear lobes of each pattern in Figure 4B are of opposite polarity with respect to the main front lobes. For sound sources arriving at 0° the left and right output signals are equal, and a center image is reproduced between the loudspeakers. As the sound source is moved off-axis, an amplitude difference between left and right is created, and the loudspeaker image is moved smoothly off-center in the direction of the higher amplitude signal.

At 90° off axis, the sound source reaches a null in one pattern. At this angle the image is reproduced fully left or right, in one loudspeaker only. The angle between the right and left pattern nulls (180° in this example) is the pickup angle. It defines the angular range of incoming sound sources where reproduced images appear between the loudspeakers. As an illustration, left, left-center, center, right-center and right images are created for sounds arriving at the angles shown.

Sounds arriving from outside the pickup angle are of opposite polarity in the left and right outputs. In two-loudspeaker reproduction, their images will be diffuse and without specific localization, adding a pleasant sense of spaciousness and ambience. In Stereosurround™-format reproduction, these images will move towards a rearward directionality.

Mid Pattern Selection

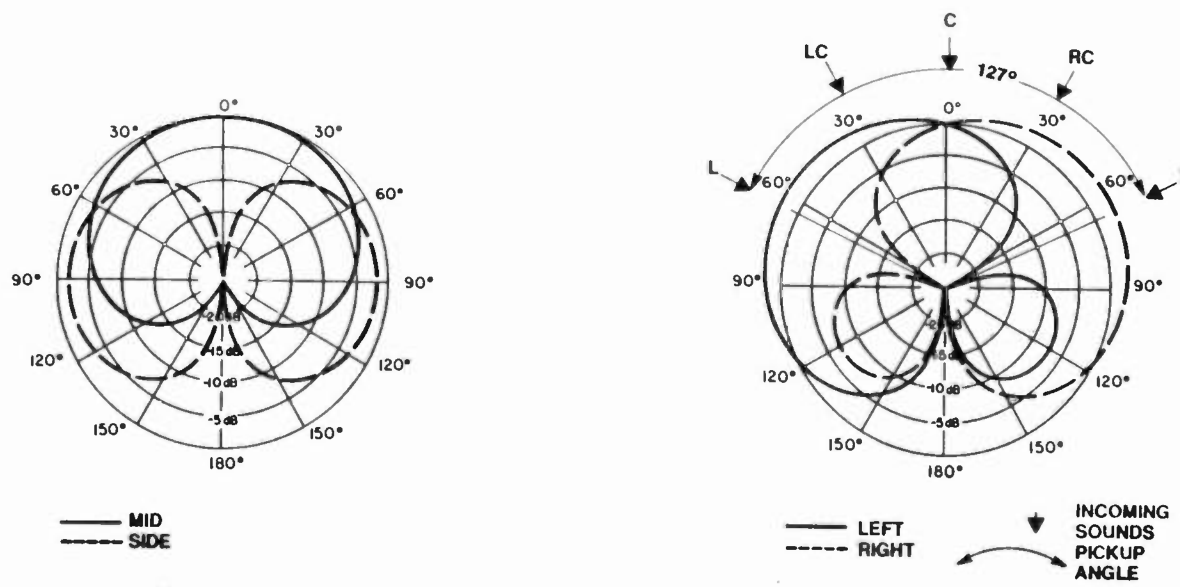
In theory, any microphone pattern may be used for the mid signal pickup. Some studio MS microphones even provide a selectable mid pattern. In practice, however, the cardioid mid pattern is most often preferred in MS microphone broadcast applications. At 180° off-axis, there is a null in both the mid and side patterns. This will provide rejection of unwanted sounds from the rear of the microphone. Other mid patterns, which exhibit pickup at 180°, would allow sounds from the rear to create a solid front-center image. In addition, most engineers are familiar with the basic operation of the cardioid microphone cartridge, so the mono pickup (again, derived entirely from the mid cartridge) is predictable. If other patterns are chosen for the mid cartridge, the resultant stereo patterns will not match those from an MS microphone with a cardioid mid cartridge.

Some commercially available MS microphones use a shotgun microphone for the mid pickup. In mono, the shotgun mid cartridge provides its characteristic "reach" and off-axis rejection. In stereo, however, the frequency-dependent polar pattern variations will cause the images of any off-center source to exhibit frequency-dependent localization, or "smear", reducing the reproduction accuracy.

The polar diagrams of Figures 3-6 are adapted from the data sheet of the Shure VP88 MS microphone, which employs a cardioid mid cartridge. In addition to mid and side outputs, a built-in matrix provides left and right outputs with a selectable level of side signal contribution.

Effect of Side Level

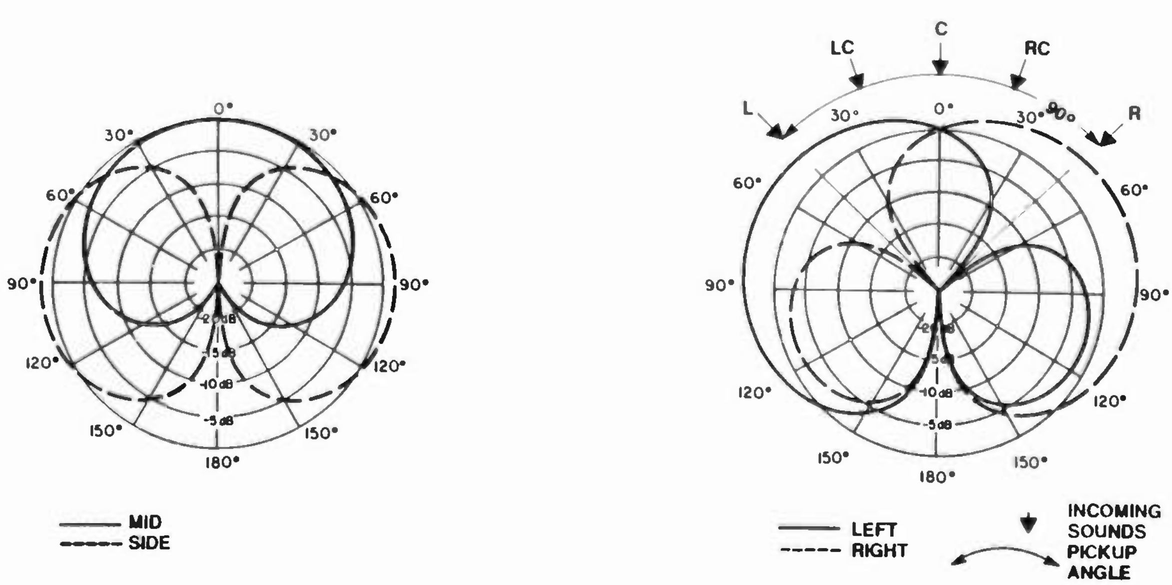
Even when the mid (mono) pattern is fixed as cardioid, the stereo pickup pattern can still be varied by changing the side level relative to the mid level. In contrast to Figure 4, which showed the stereo pickup patterns with the side level 6.0 dB below the mid level, Figure 5 shows an MS pattern with the side level 1.9 dB lower than the mid level. Figure 6 increases the side level to 1.6 dB higher than the mid level. The three resultant stereo patterns exhibit pickup angles of 180°, 127°, and 90°, respectively. The incoming sound angles which will create left, left-center, center, right-center and right images are also shown. Note the changes in the direction of the stereo patterns and the size of their rear lobes.



a. MS Pickup Patterns

b. Resultant Stereo Pickup Patterns

Figure 5
MS Microphone: Side Level 1.9 dB Lower Than Mid Level



a. MS Pickup Patterns

b. Resultant Stereo Pickup Patterns

Figure 6
MS Microphone: Side Level 1.6 dB Higher Than Mid Level

As the side level is increased relative to the mid level, the following interrelated changes occur in the stereo characteristics:

1. Increased stereo ambience pickup—The amount of environmental or room sound pickup increases. Since the stereo pickup patterns have larger opposite rear polarity lobes and are angled somewhat wider apart, ambience pickup comes more from the sides of the microphone and is more spaciously reproduced (mono ambience pickup remains unchanged).

2. Increased stereo spread—Sounds from an incoming sound stage spread wider apart across the reproduced stereo stage.

3. Decreased pickup angle—A greater spread of the incoming sounds across the reproduced stereo stage translates into a narrower pickup angle. Obviously, if the reproduced sound images spread wider apart, then the total angle of incoming sounds which can be reproduced between the loudspeakers must be narrower. For example, if the sound stage is 90° wide relative to the microphone, a 90° pickup angle will spread the sound fully between the loudspeakers. A 180° pickup angle would spread the 90° sound stage only about halfway between the loudspeakers.

All of the changes occurring with increasing side level tend to provide the effect of "more" stereo—a wider, more spacious sound. This is not surprising. The side signal is, in fact, the stereo difference signal, and the side pattern favors the pickup of side ambience over direct front sounds. Of course, excessive side level relative to the mid level should be avoided, since stereo reproduction would become overly spacious, and mono reproduction would become drastically different from stereo when the side signal was cancelled out.

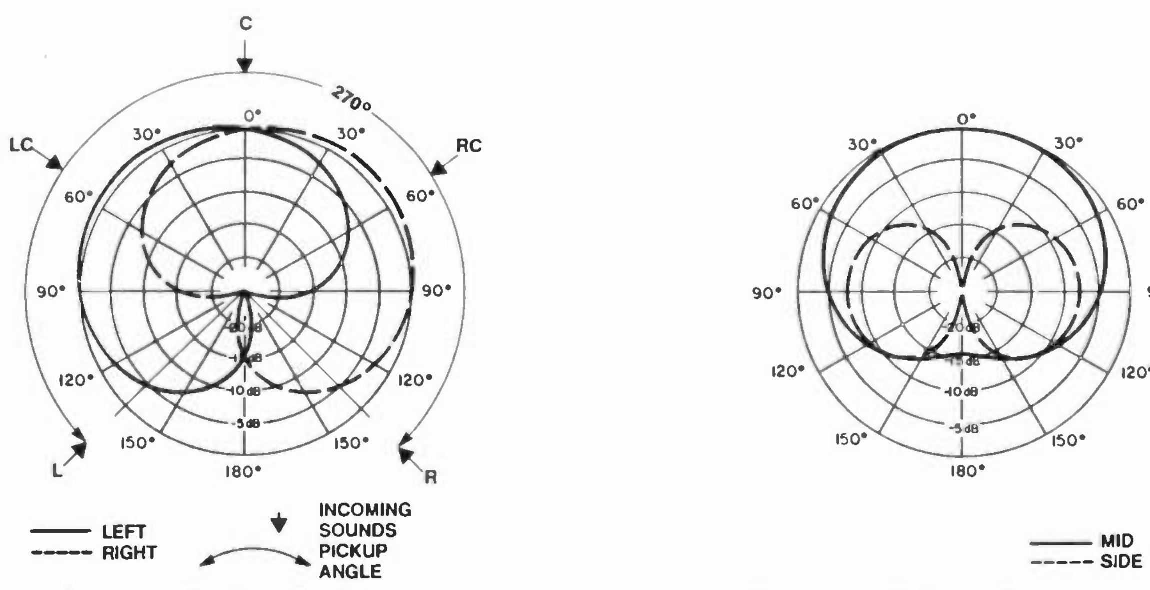
Post-Production Flexibility

If the mid and side outputs of the MS microphone are each recorded on individual channels, they can be matrixed to left and right signals in post- production. The ratio of side level to mid level can be controlled and the stereo perspective varied in mixdown. This flexibility is not as readily available if the matrixed left and right stereo signals are recorded directly. In this case, the stereo effect can be reduced by panning left and right channels toward center (in effect reducing the side, or stereo difference, component). If other signals are premixed with the matrixed stereo signals, then modifications could not be made without affecting the additional signals.

Comparison of MS and XY Techniques

When mid and side signals are matrixed, the resulting left and right patterns are equivalent to matched coincident directional microphones having various polar patterns, and aimed outward at various angles. The XY microphone technique, generating left and right signals directly using two coincident, angled directional microphones, is another closely related coincident stereo microphone technique.

Unfortunately, the relationship between the aiming angles of the XY microphone pair and the MS-derived stereo patterns is not as intuitively straightforward as might be desired. For example, consider cardioid microphones, angled 90° apart, as the XY microphone pair (see Figure 7). Use this array to pick up an incoming sound stage, also covering 90°. This way, the left microphone is aimed at the left edge of the sound stage, and the right microphone at the right side. The casual observer may assume that this microphone array will spread the sound stage between both loudspeakers.



a. Coincident Cardioid Microphones Angled 90° Apart

b. Equivalent MS Pickup Patterns

Figure 7
Comparison of MS and XY Techniques

In reality, this XY approach results in a spread of the reproduced sound stage less than halfway between the loudspeakers, some pickup from the rear in both stereo and mono, and a relatively non-ambient character. A careful look at Figure 7A, and the MS equivalent in Figure 7B, reveals that the pickup angle is 270°, which is much wider than the 90° angle of the microphones (most sounds will bunch up in the middle of the reproduced stereo stage). The mono sum (mid) pattern is somewhat subcardioid with moderate rear pickup. The stereo difference (side) level is fairly low, about 8 dB below the mono sum (mid) level. There are no opposite-polarity rear lobes, reducing spaciousness.

The MS-derived patterns in Figure 4B are also aimed 90° apart, but the stereo characteristics are quite different. In contrast to the coincident cardioids, the patterns of Figure 4B result in a 180° pickup angle, spreading the images wider between the loudspeakers. The opposite-polarity rear lobes add to the spaciousness, and the rear rejection is good.

This comparison illustrates that there is little direct intuitive relationship between the aiming angle of XY microphones and the aiming angle of the general MS-derived pickup patterns. Widening the aiming angle of an XY pair does increase stereo ambience pickup, increase the stereo spread and decrease the pickup angle, but not as quickly as, or with results similar to, widening the aiming angle of the MS-derived stereo patterns. The differences become greater as the microphone aiming angles become wider.

Finally, in the XY technique, the front-center sounds are off-axis to the microphones, where their frequency response may be less accurate. In contrast, the MS technique provides on-axis pickup of front-center sounds by the mid cartridge, ensuring their best reproduction in stereo and mono.

TYPICAL MS MICROPHONE APPLICATIONS

Stereo productions can be easily and effectively accomplished using the MS microphone either as a primary or secondary pickup source, with or without additional microphones. The application examples described here are intended to illustrate guidelines, not hard and fast rules. The ideas embodied in these examples may serve as a starting point; an understanding of the MS principles reinforced by experience will enable the broadcast sound engineer to develop a creative and effective MS microphone technique.

In any stereo production, monitoring the mono sum program is as important as monitoring the stereo mix, since many listeners are still limited to one speaker. The mono listeners miss the ambience picked up in the side signal, but less ambient pickup in mono is usually desirable to maintain clarity. Mono compatibility is not a problem with a single MS microphone, but must still be verified in any multiple microphone situation. As always, the engineer's ears are the best judge of audio quality.

For illustration purposes, the MS microphone shown in Figures 8, 10, and 11 is assumed to have an internal matrix selected to provide left and right signals directly to the stereo mixer. If extra recording channels are available, mid and side signals may be recorded individually for post-production flexibility. For that matter, in a multiple microphone production, each individual microphone signal may be recorded separately onto a recorder with multichannel capability (most video recorders, for example, have more than two channels available). This approach provides the most flexibility, but requires a more elaborate setup.

Musical Ensemble

Stereo pickup of a musical ensemble is one of the most common applications of the MS microphone. This application also serves to illustrate the effects of pickup angle and side level on the reproduced stereo image. Notice that the principles of ensemble pickup apply just as well to the stereo pickup of any wide sound source.

In Figure 8, the MS microphone is set to place the ensemble sound stage across the full pickup angle. In playback, the reproduced direct sound spreads fully between the loudspeakers. The side level may be lowered either to increase the pickup angle for a very wide incoming sound stage or to reduce the stereo spread (minimizing image shifting in hand-held use, for instance). Increasing the side level results in greater reproduced ambience and stereo spread.

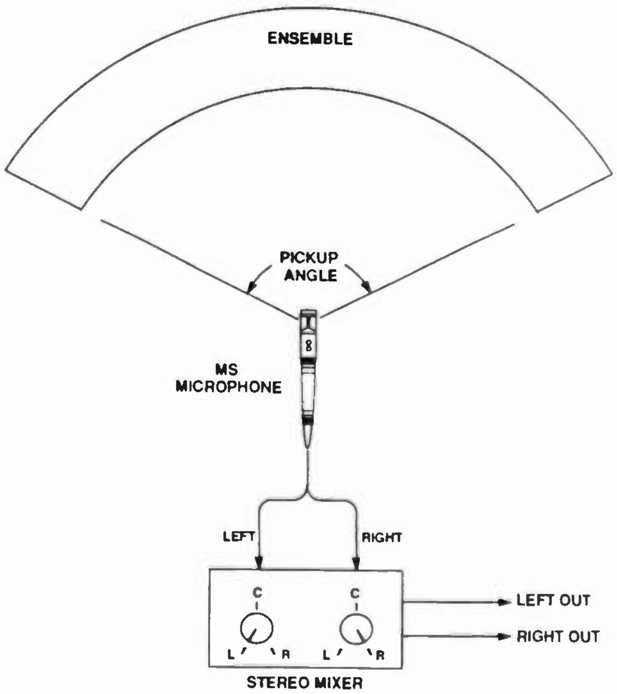


Figure 8
Stereo Pickup of Typical Ensemble

Suppose a small ensemble is being picked up from a considerable distance as shown in Figure 9. Full stereo spread of the reproduced image would require a very high side level, which would probably pick up too much room ambience. However, a side level appropriate for proper ambience pickup results in a very small image spread between the loudspeakers. In this situation, it is impossible to achieve full stereo image spread along with balanced ambience effect. If a compromise is not acceptable, the microphone must be moved closer to the ensemble.

If a large ensemble features a soloist, or if softer instruments and voices are overpowered by louder ones, spot microphones may be required to improve balance. These microphones should be placed appropriately close to the sound source requiring reinforcement, panned to match the corresponding image from the MS microphone in the reproduced stereo stage, and increased in level just enough to provide proper

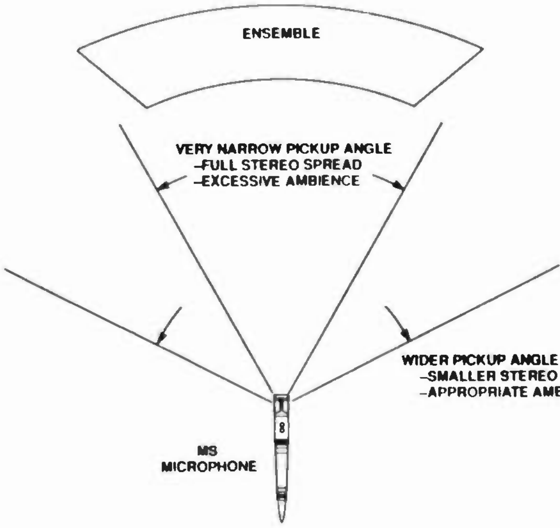


Figure 9
Stereo Pickup of Distant Ensemble

balance without noticeable comb filtering (the MS and spot microphones would probably be spaced far apart). Both mono sum and stereo program should be monitored to verify the effectiveness of the spot microphones.

ENG With Announcer

In the Electronic News Gathering (ENG) situation, unlike the music ensemble example, the MS microphone is generally used for environmental ambience pickup, not for the main pickup (in this case, of the announcer's voice). For convenience, the MS microphone may be mounted directly to the camera, its relative side level set according to the desired stereo ambient effect. A second (mono) microphone is used for the announcer. This microphone should be panned to center, and is typically mixed in roughly 10 dB higher in level than the MS microphone. Mono sum monitoring should be used to

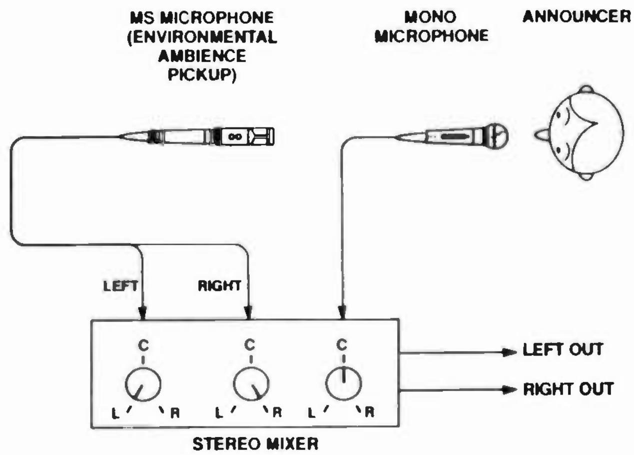


Figure 10
Typical System for ENG Situation

verify that the balance between the announcer level and the MS mid level is as intended and that minimal comb filtering is occurring. A typical stereo ENG system is illustrated in Figure 10.

Sports Event

A sports event is basically an ENG situation, and all of the suggestions listed above apply here as well. However, the unique aspects of sports coverage require special attention. For example, if only the final stereo mix is processed through a limiter, an overly enthusiastic announcer could cause the ambient sounds to pump up and down. This background pumping is noticeable in mono listening, annoying in stereo, and extremely irritating in surround reproduction, where the ambient reproduction from the sides and rear is spatially separated from the announcer's voice. By limiting the announcer's microphone separately from the overall stereo mix, background pumping can be eliminated.

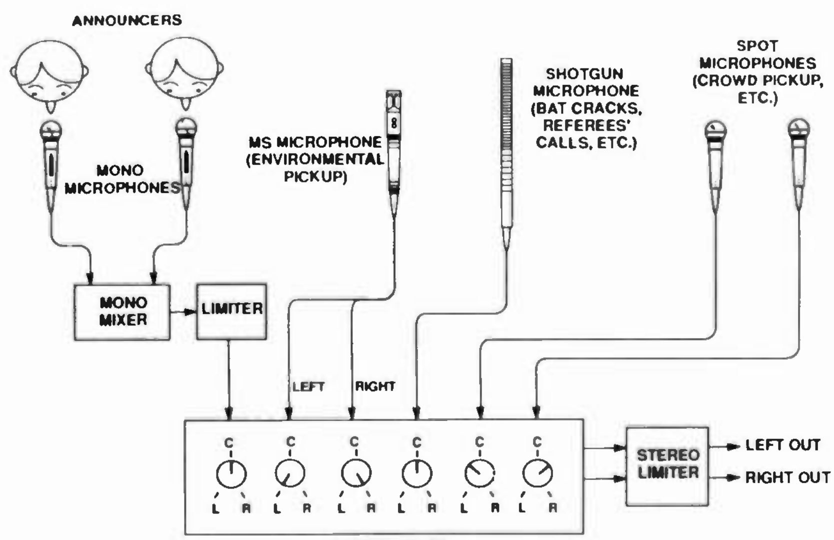


Figure 11
Typical System for Sports Event

Experimentation and creativity are vital elements in microphone techniques for a sports event. A one-to-one correspondence between visual and audio images is not possible, of course, since camera selection and angles are constantly changing. A believable, stable audio perspective should be created using one or more MS microphones, with the announcer's microphone panned to the center. For pickup of certain crowd noises, bat cracks, referees' calls and other distinctive sounds, spot microphones may be used as required to augment the basic stereo environment. See Figure 11 for a typical sports system.

In the end, learning how to use the MS microphone requires the same steps as for any new audio production tool: understand, experiment, listen and trust your ears.

SUMMARY

The MS coincident stereo microphone is a convenient, flexible tool for creating mono-compatible intensity stereo, employing direct pickup of the mono sum and stereo difference components. Control over stereo ambience pickup, stereo spread and pickup angle is possible without affecting mono pickup. Better results are generally obtained, with easier, more intuitive adjustment, than with the related XY microphone techniques. The MS microphone is easily incorporated into a stereo production, with or without additional microphones.

The MS microphone has the capabilities required for effective stereo reproduction. The broadcast sound engineer who makes the effort to learn these capabilities will be able to exploit them fully.

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NTSC COMPATIBLE DIGITAL MODULATION FOR TV SOUND

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0. Abstract

Digital transmission techniques which may be applied to broadcast TV sound are reviewed. A data carrier may be added to the NTSC television signal to deliver stereo digital audio to the home. Specifications for the data carrier are developed which give the best compatibility and lowest cost receiver.

I. Introduction

There is a growing interest in the use of digital data transmission techniques to improve broadcast audio quality. Digital transmission is inherently robust. While the coding of high quality audio into digital form theoretically entails a loss of quality, there is no further loss of quality when the digits are transmitted through an error-free channel. This is in stark contrast to the transmission of analog audio, in which a "perfect" channel is required to avoid degradation. Unfortunately, the intercarrier sound channel used with broadcast television is somewhat less than perfect, and inherently limits the sound quality of the analog BTSC stereo system.

Digital transmission techniques have matured to the point where they can be applied to broadcast audio. Previous work[1,2] led to a proposal for the addition of digital audio to the NTSC television broadcast signal. That proposal borrowed heavily from work performed in Sweden[3] and Finland where a 512 kb/s QPSK carrier was extensively tested with PAL system B. The similarities between B-PAL and M-NTSC indicated that the Scandinavian test results would apply in the U.S. Our original 1987 proposal was:

A. QPSK carrier with alpha=0.7 filtering.
B. Carrier frequency 4.85 MHz above video carrier.
C. Carrier level -20 dB with respect to peak vision carrier level.

Compatibility testing of that system has been performed, and some television sets have been found on which the data carrier causes detectable interference to the upper adjacent video channel in a clean laboratory setting. The interference occurs into luminance, and manifests itself as additive noise between approximately 1 MHz and 1.4 MHz. The noise level in a problem TV is subjectively similar to the noise level resulting from a video carrier-to-noise ratio (CNR) of approximately 47 dB. While this level of interference may not be detectable given the current quality level of signals in consumers' homes, the potential of optical fibre to raise the quality level of cable TV warrants a reduction in the level of interference by approximately 6 dB to be safe.

Considerations of the nature of existing broadcast transmitters also call for some changes so that the signal is more easily transmitted.

II. Digital Modulation Basics

Figure 1 shows the spectrum of a random data stream with a data rate of 250k bits/sec. In a digital transmission the data is always intentionally scrambled so that the transmitted data appears random. Note the spectral nulls at multiples of the data rate (250 kHz, 500 kHz, etc.). From Nyquist sampling theory we know that all information in this signal may be gleaned from the first 125 kHz, i.e. the Nyquist frequency

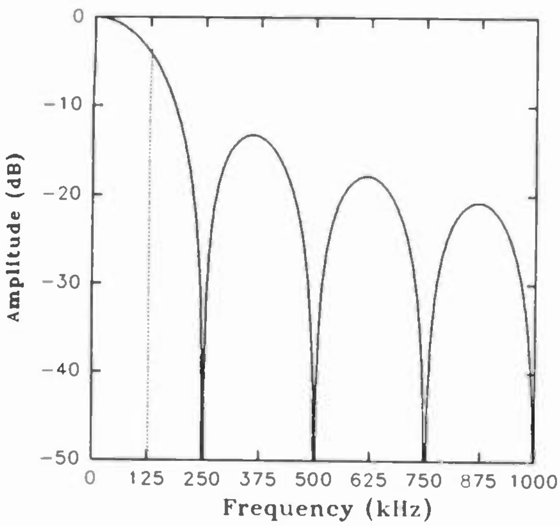

Figure 1 Data Spectrum of 250 kb/s data.

which is ½ of the 250 kHz sample rate. If we brick wall filter this signal at the Nyquist frequency of 125 kHz and use it to modulate a carrier, we will have a double sideband modulated bandwidth of 250 kHz, which will carry 250k bits/sec of data (1 bit/sec/Hz). If we add a second similar channel at the same frequency but in quadrature, we will double the information in the 250 kHz bandwidth RF signal to 500k bits/sec, or 2 bits/sec/Hz. Fig. 2 shows a block diagram of this modulator which is simply a pair of double sideband AM modulators in quadrature. The double sideband modulator can be considered to be a phase modulator, generating phases of 0 and 180 degrees. The second modulator can be considered to generate phases of 90 and -90 degrees. The combination of the two phases will generate one of four phases, ±45 and ±135 degrees. This form of modulation is typically called Quadrature Phase Shift Keying or **QPSK**.

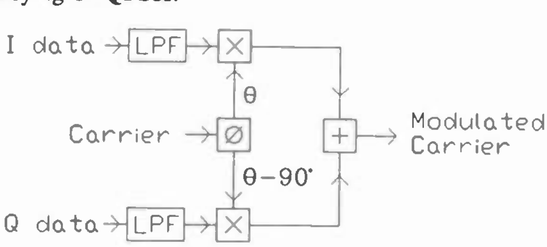

Figure 2 QPSK Modulator

Fig. 3 shows the constellation diagram for QPSK. There are four possible states of the RF carrier phase. We refer to these states as symbols. Since there are four states per symbol we are transmitting two bits per

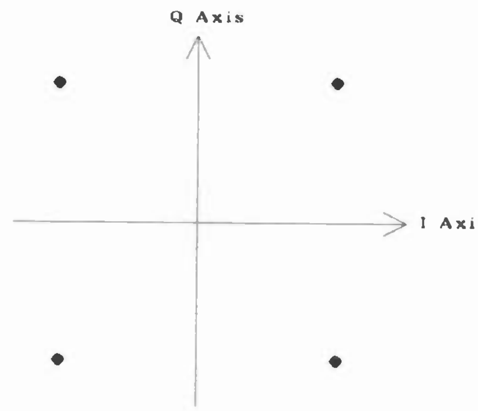

Figure 3 QPSK Constellation

symbol with a symbol rate of 250 k symbols/sec, for a total data rate of 500k bits/sec. If the low pass filters were not present, the phase would be instantaneously jumping between the four states shown, at a rate of 250 kHz. In practice the actual signal will be continuously traversing between these points and the receiver will sample the signal at the time when the signal passes through the constellation points.

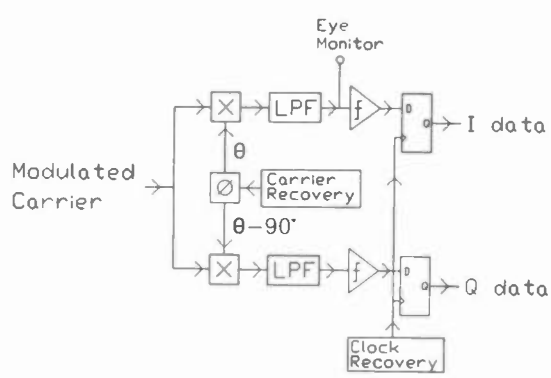

Figure 4 QPSK Demodulator

The QPSK receiver is shown in Fig. 4. A carrier recovery loop coherently regenerates the transmitted carrier, which was suppressed in the modulator. Quadrature mixers demodulate the I and Q signals, low pass filters limit the effective receive bandwidth, and comparators and clocked latches recover the data. The latches are driven by a symbol timing recovery circuit which reconstructs the transmit data clock at the proper phase so that the sampling time is correct.

$$H(j\omega) = \begin{cases} 1, & \left[0 \le \omega \le \frac{\pi}{T_s}(1-\alpha)\right] \\ \cos^2\left\{\frac{T_s}{4\alpha}\left[\omega - \frac{\pi(1-\alpha)}{T_s}\right]\right\}, & \left[\frac{\pi}{T_s}(1-\alpha) \le \omega \le \frac{\pi}{T_s}(1+\alpha)\right] \\ 0, & \left[\omega \ge \frac{\pi}{T_s}(1+\alpha)\right] \end{cases} \qquad \text{Eq. (1)}$$

$$QPSK\ Spectral\ Efficiency = \frac{2}{1+\alpha}\ bits\,/\,Hz \qquad \text{Eq. (2)}$$

While QPSK can theoretically achieve a spectral efficiency of 2 bits/sec/Hz, this is not achievable in practice since a brick wall filter is not realizable. The steepness of the filter determines the spectral efficiency. Steep filters ring, and as the spectral efficiency of QPSK is raised by sharpening the filters, the ringing of the individual data pulses increases. This ringing can cause the pulses to interfere with each other (intersymbol interference) unless the ringing is controlled. If the filtering is done properly, each individual pulse waveform, except for the pulse being detected, will pass through zero at the instant the receiver samples the waveform. Nyquist filters provide the desired characteristics. A commonly used class of Nyquist filters are the "raised cosine" filters which have a frequency response as described in Eq. 1, and shown in Fig. 5. The alpha term determines the fractional excess bandwidth over a perfect brickwall (alpha=0) filter. A filter meeting the amplitude response of Eq. 1, and having constant group delay (linear phase) will have no intersymbol interference.

When we speak of the channel filter, we mean the filter function of the complete channel. This includes the transmit baseband lowpass filter, any transmit RF filtering, receiver preselection and IF filtering, and receive baseband lowpass filtering. It also includes the effect of the transmission path which might have a non-flat response (or selective fade). Typically, all of the intended filtering is designed into the system at IF or baseband, and all other circuitry is made wideband. The most precise control over filtering is achieved with baseband lowpass filtering. It is much easier to make a precision 125 kHz lowpass filter than a precise 250 kHz IF filter.

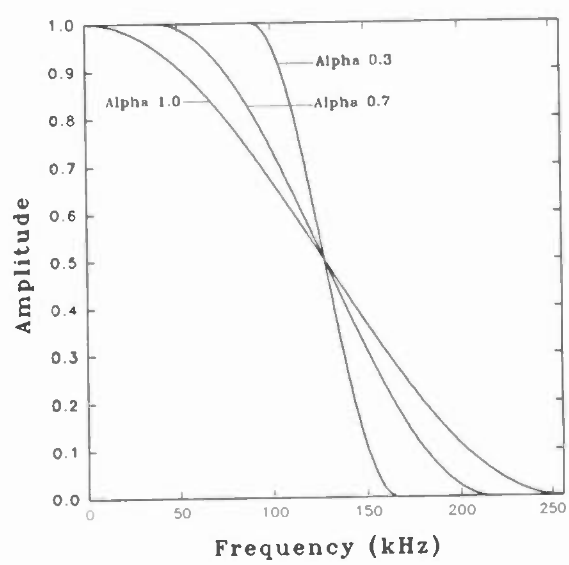

Figure 5 Raised Cosine Nyquist Filters

As alpha is reduced, spectral efficiency of QPSK is improved. Equation 2 shows the relationship between spectral efficiency and alpha. As alpha is reduced, the data waveform will have more ringing and the "eye" pattern will become more complex. The eye pattern is observed at the point where the analog waveform is sampled to recover the data. In the QPSK decoder block diagram (Fig. 4), the eye monitor point is shown for the I channel, and is just after the baseband lowpass filter. In Fig. 6 we see some examples of eye patterns for alpha=0.3, alpha=0.7, and alpha=1.0. The eye patterns become more complex as alpha is lowered and spectral efficiency is improved. More accuracy is required in implementation of low alpha systems. Any error in amplitude or phase linearity will rapidly degrade a complex eye. The eye will appear to close. Fig. 5d shows an alpha = 0.3 eye with a lot of closure due to poor filtering. Partial eye closure due to imperfect filtering leaves less margin against noise and interference.

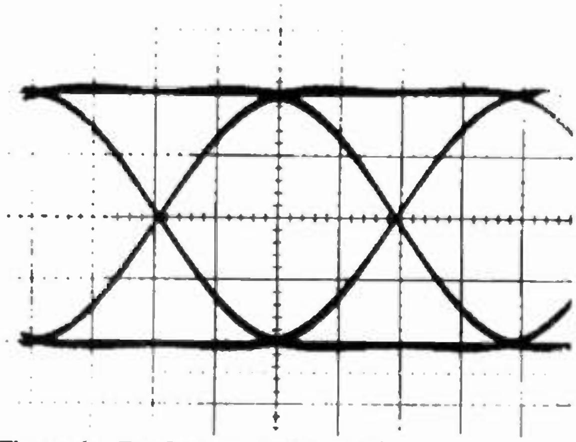

Figure 6a Eye Pattern, Alpha = 1.0

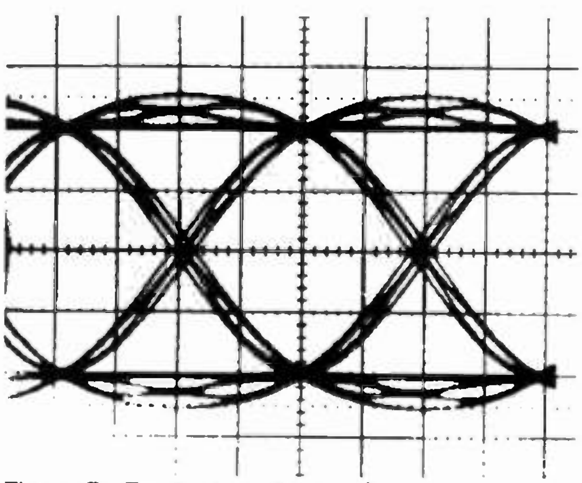

Figure 6b Eye Pattern, Alpha = 0.7

For optimal performance over a noisy channel, the amplitude portion of the filtering should be partitioned in equal portions between the transmitter and receiver i.e. the transmit filter and receive filters should both have an amplitude response equal to the square root of the nyquist filter response. The cascade of the two filter characteristics will then be a Nyquist filter. The combination of the filters should have linear phase (constant group delay). If both the transmit and receive filters individually have linear phase, the combination will as well. However, it is also acceptable for the receive filter to have group delay variations and to compensate for these in the transmit filter.

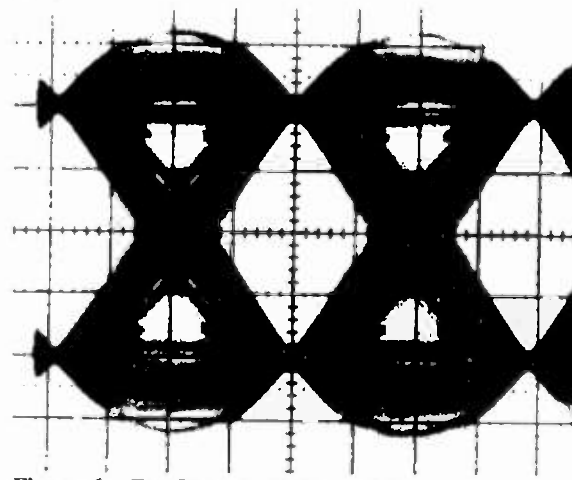

Figure 6c Eye Pattern, Alpha = 0.3

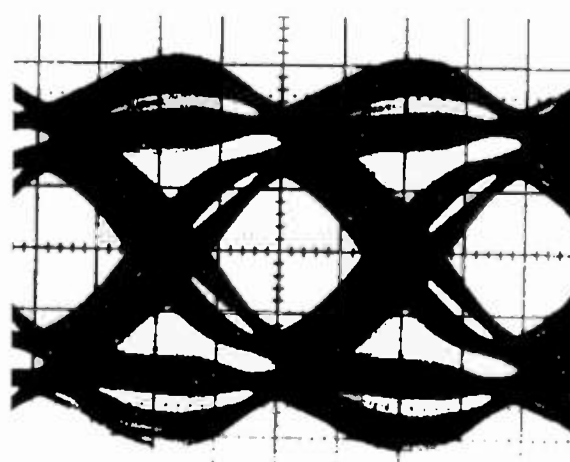

Figure 6d Alpha = 0.3, Poor Filter Accuracy

Attempting to achieve good spectral efficiency with low alpha filtering requires the use of more accurate filters with good phase compensation. Filters may be implemented with either analog or digital circuitry. Since we are starting with digital data, the use of a digital filter in the transmit side is attractive. The receiver is more easily implemented with an analog filter which does not require the use of A-D converters with high sample rates and anti-aliasing pre-filters. In a broadcast application we want to keep the receiver cost low, so we want to use the minimal receive filter (without phase compensation). If the receive filter magnitude response accurately matches the square root of a Nyquist response, then the transmit filter can be implemented as an FIR (finite impulse response) filter with an impulse response which is simply the time reversal of the impulse response of the receive filter. The cascade of the filters will then have a symmetrical impulse response with constant group delay, and will have an amplitude response equal to that of a Nyquist filter. The matched filter criterion is satisfied, so performance in the presence of noise will be optimal. All phase compensation is done in the transmit filter so receiver cost is minimized.

In order to achieve spectral efficiencies of 2 bits/sec/Hz or greater, QPSK must be abandoned and a higher level modulation method used. This can be done by hitting the modulator filters (Fig. 2) with multilevel data symbols. With QPSK, the I and Q data are simple binary symbols with two levels (1 and 0, or +1 volt and -1 volt). If pairs of bits are fed to 2 bit D-A converters, we can form four level data (0, 1, 2, 3; or +3 volts, +1 volts, -1 volts, -3 volts). QPSK extended to 4 level symbols is known as 16 QAM (Quadrature Amplitude Modulation). 16 QAM has a theoretical (with ideal brick wall filtering) spectral efficiency of 4 bits/sec/Hz. Unfortunately, the added spectral efficiency is achieved at the expense of ruggedness. The eye pattern of 16 QAM is similar to that of three QPSK eyes stacked on top of one another. For the same absolute eye opening (which would give the same performance in a noise or interference environment), the peak level is 3 times higher, or about 9.5 dB. For the same carrier level, the 16 QAM signal is 9.5 dB less rugged than QPSK.

III. Partial Response Signaling

Another attractive technique which can achieve improved spectral efficiency is known as "partial response signaling". This method involves the use of a controlled amount of intersymbol interference that is removed by an extra processing step in the receiver. The simplest form of partial response involves performing a running average of adjacent pairs of bits before driving the modulator. If we average a pair of bits which can each be +1 or -1 we get symbols with three levels: +2, 0 or -2. This 3-level signal still only contains 1 bit of information per baseband symbol, but the spectral characteristics are changed. The spectrum of the data is changed from that shown in Fig. 1 to that shown in Fig. 7. The first spectral null has moved from 250 kHz down to 125 kHz. If we filter off all of the energy beyond 125 kHz (which is relatively easy), we will achieve a spectral efficiency of 1 bit/sec/Hz at baseband. The use of a pair of these partial response

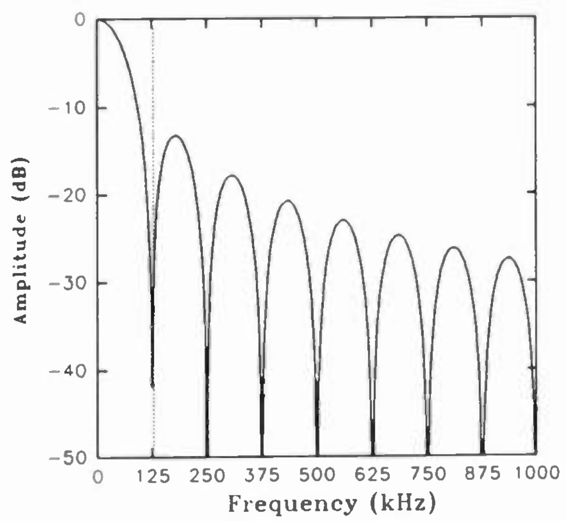

Figure 7 Partial Response Spectrum

system modulators in quadrature is known as QPRS. The constellation diagram for QPRS is shown in Figure 8.

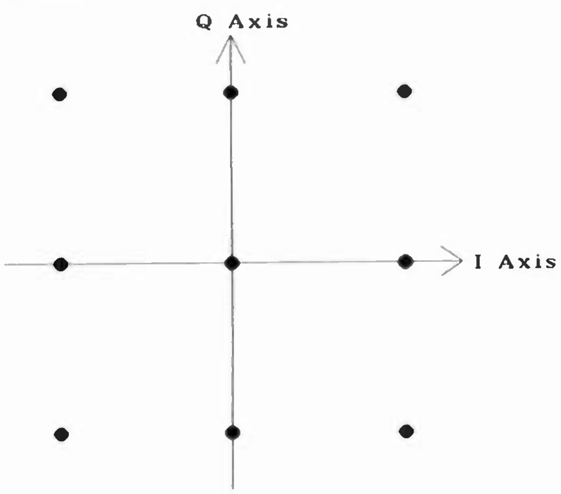

Figure 8 QPRS Constellation

The pairwise bit averaging is equivalent to a filter with an amplitude response equal to a cosine function, as shown in Fig. 9. In order to have matched filtering, the bit averaging should not be done at the transmitter, but the cosine filter function should be divided equally between the transmit and receive filters. That is, both the transmit filter and the receive filter should have an amplitude response equal to the square root of the first

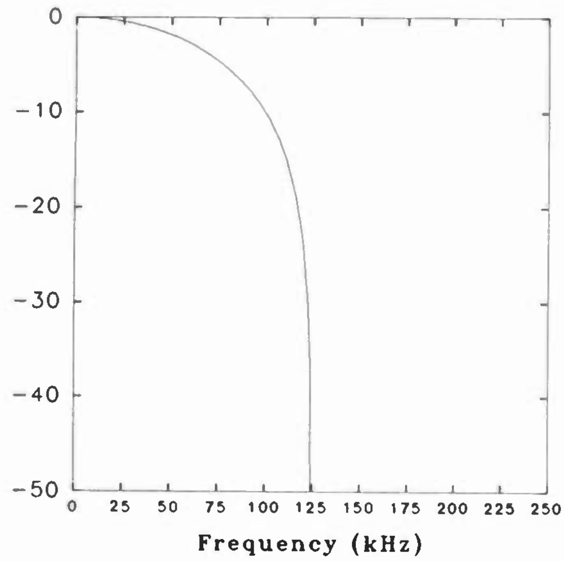

Figure 9 PRS Cosine Filter

quadrant of a cosine. The transmit filter may be a digital FIR filter with an impulse response which is the time reversal of that of the receive filter.

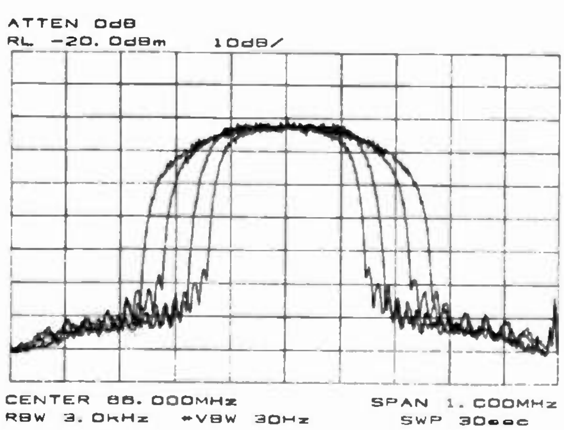

Figure 10 Comparative RF Spectra

Fig. 10 shows the spectrum of the transmitted data carrier for several modulation schemes. The narrowest spectrum is for QPRS, the second narrowest is alpha=0.3 QPSK, and then we have alpha=0.7 and alpha=1.0. The transmit filters used are FIR filters with a magnitude response equal to the square root of the raised cosine curve (for the QPSK cases), or the square root of a cosine curve (the QPRS case). The spectrum of QPRS is very attractive, but its eye pattern (Fig. 11) shows it to be less rugged than QPSK. Since there are two eyes, in order to achieve the same absolute eye opening (and thus the same ruggedness), the peak power must be increased by a factor of two, or 6 dB. For the same peak power, QPRS should be 6 dB less rugged. There are some mitigating factors though.

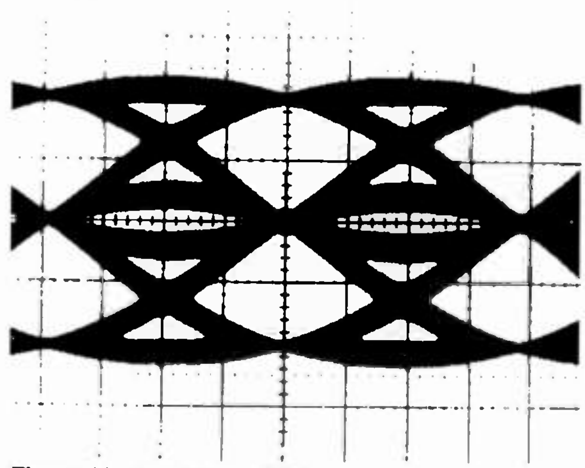

Figure 11 Eye Pattern, PRS

First, the PRS data stream spends half its time at the 0 level (data = 1,0 or 0,1), and half at the +2 or - 2 level (data = 0,0 or 1,1). This means that half of the time the carrier is modulated by zero, and half of the time it is modulated by the peak value. QPRS will have an RMS power level half that of QPSK for the same peak level. For the same RMS power level, QPRS will thus lose only 3 dB of ruggedness compared to QPSK, and not the 6 dB that is lost with equal peak power levels. Second, since there is some redundancy in the PRS 3-level data stream, it is possible to detect and correct some errors without intentionally including redundant FEC data in the data multiplex. Coding gains of 1.5 to 2.0 dB can be achieved economically. The combination of these factors indicate that for comparable carrier-to-noise ratios (CNR), QPRS is only about 1 to 1.5 dB less rugged than QPSK against gaussian noise. QPRS is an attractive candidate if its narrow spectrum would solve our interference problem.

While QPRS requires some added complexity in the receiver, the amount is within reason for a consumer product. If the narrowness of the QPRS signal significantly reduces the potential for interference, QPRS could be adopted for digital TV sound broadcasting without seriously impacting the receiver cost.

IV. Practical Broadcast Considerations

The proposed data carrier is placed between the FM sound carrier and the lower vestigial sideband. Figure 12 shows the RF spectrum of TV Ch 3 and 4, with an Alpha = 0.3 QPSK carrier inserted between the channels. In the case of adjacent channel operation, it makes little difference whether we say that the data signal is assigned to the upper or the lower channel. The BBC[4] choose to allocate the signal to the lower channel. There is an advantage to this choice in receiver complexity. If the first sound IF is widened slightly, the inter-carrier mixer used to recover the 4.5 MHz FM sound will also recover the data sound carrier at 4.85 MHz. This point can be tapped and fed to the QPSK demodulator. If the data carrier is associated with the upper channel, the inter-carrier offset will be -1.15 MHz, and additional IF filtering and another inter-carrier mixer will be required. Filtering will have to be sufficient to reject the image at 1.15 MHz above vision. The choice of placing the carrier above the channel is a good one, and we originally agreed with it.

Unfortunately, there are serious problems with this choice for the U.S. The most obvious one is that in the case of TV Ch. 6, the data carrier would lie at 88.1 MHZ, making both 88.1 and 88.3 unusable for FM broadcasting. The second is that many transmitter installations are unable to easily broadcast the signal due to the notch diplexers in use. Stations which use separate amplification for sound and vision would introduce the signal into the sound transmitter. The sound and vision transmitters are combined in a device known as a "notch diplexer" in which tuned cavities (creating notches) reflect the sound signal into the antenna. There are no cavities tuned to the frequency of the data signal, so the diplexer would need expensive modifications. While this is not a technical problem, it is an economic one and could seriously hamper the acceptance of the digital system by broadcasters.

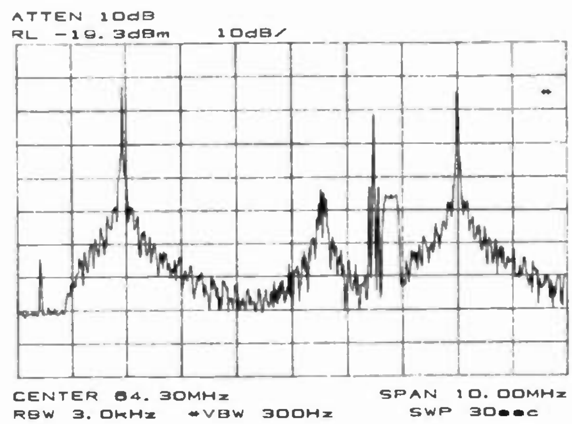

Figure 12 Ch3, Data, Ch4 Spectrum

Choosing to place the data carrier below the vision carrier appears to be the correct choice for the U.S. The signal can be added to the vision transmitter just after the vestigial sideband filter, and will be amplified linearly along with the picture. At this frequency there will be no problem with the notch diplexer.

V. Interference of Data into Picture

Placing the data signal on the lower sideband of our own channel makes it even more critical to avoid interference into video because now we will interfere with ourselves. The mechanism of interference is imperfect VSB Nyquist filtering in the TV receiver. While it is easy to make a TV receiver reject the data signal there are a wide variety of receivers in use, some of which have poor rejection. Fig. 13 shows the ideal receiver IF filter response. Also shown is the kind of response that could be susceptible to interference. The problem is the 'tail' in the rolloff. If the filter goes right down to near zero response at 1 MHz away from the vision carrier the data signal will be fully rejected. If the filter has a tail, the data will not be rejected and may be visible in the picture.

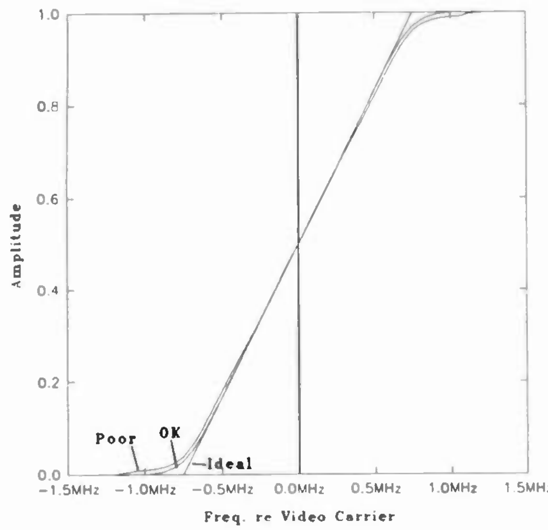

Figure 13 Receiver VSB Nyquist Filters

The only way to reduce the potential interference is to: 1) reduce the data carrier level, and 2) move the data carrier farther away from the video carrier (so as to get farther down on the slope of the receive filter). The data carrier can be moved by either moving the center frequency or narrowing the data spectrum, or both. Moving the upper bandedge of the data spectrum down by 100 kHz reduces the interference into the upper video picture by approximately 6 dB on problem receivers.

VI. Interference of Data into FM Sound

Our initial studies of this system concentrated on the interference of data into the BTSC sound signal. We found that with QPSK and alpha=0.7 filtering, the data carrier could be placed 350 kHz above the sound carrier. With alpha=0.3 filtering, the carrier could be moved down another 50 kHz so that it is only 300 kHz above the FM sound carrier. The mechanism for interference into FM sound is demodulation of the data signal as noise by the FM detector. The frequency of the demodulated noise is equal to the offset between the FM carrier and the data signal. Since the data signal is wideband, the demodulated noise is also wideband.

Fig. 14 shows the demodulated composite BTSC FM signal from DC to 200 kHz with no modulation on any of the BTSC channels. The 15.7 kHz pilot is clearly visible, along with L-R noise (produced by the BTSC

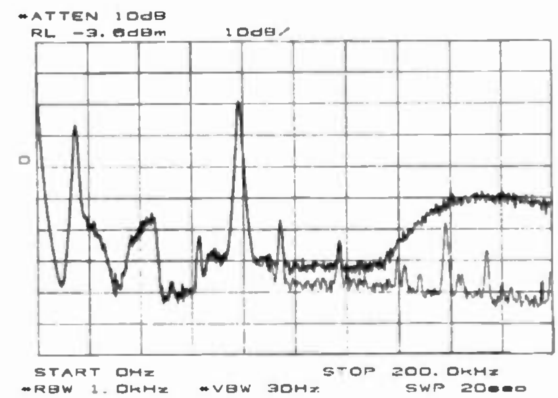

Figure 14 Spectrum from FM Demod, No Channels Modulated.

compressor working at full gain) and the SAP carrier at 78.7 kHz. Overlaid is a plot with the data carrier turned on (using the parameters to be specified in section VIII). The data carrier causes a slightly higher noise level starting at about 90 kHz, and at 140 kHz the noise level rises abruptly. A higher noise level at these frequencies should be of no consequence in a properly designed mono, stereo, or SAP decoder, and of only minor significance to a Pro channel decoder. A poorly designed decoder which inadvertently mixes components above 100 kHz down to baseband might suffer a slight degradation due to the addition of the data carrier, but this type of decoder would also suffer a premature noise degradation as CNR is reduced. Figure 15 shows the same plots, but with the Left and SAP channels fully modulated with a 400 Hz sinewave. This figure confirms that the noise induced into the composite baseband signal is similar whether the FM carrier is fully modulated or not. The presence of non-linear FM modulation does not appear cause the noise to spread down to lower audio frequencies.

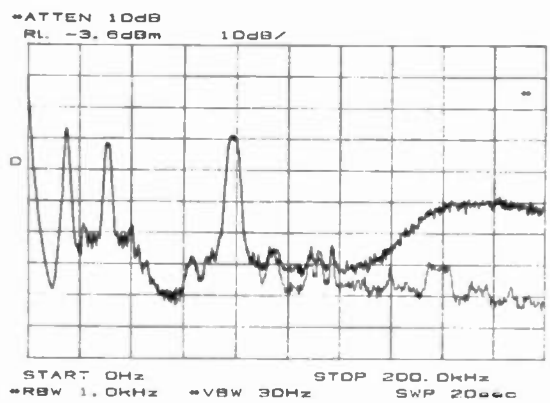

Figure 15 FM Demodulated Spectrum, All channels modulated.

VII. A More Compatible System

Our original proposal of alpha=0.7 QPSK placed 350 kHz above the audio carrier at a -20 dB level is adequately compatible with BTSC audio, but could benefit from about 6 dB of improved compatibility with video. This could be achieved by dropping the carrier frequency by 100 kHz, but that would compromise compatibility with audio. Another way to achieve the 6 dB improvement is to narrow the carrier bandwidth as the frequency is dropped, keeping the lower data bandedge the same distance from the FM sound carrier. In theory, a 50 kHz drop in carrier frequency accompanied by a 100 kHz narrowing in bandwidth should meet our target. The narrowing could be done by staying with QPSK and changing the filtering from alpha=0.7 to alpha=0.3. The bandwidth will narrow by 0.4 x 250 = 100 kHz, and the carrier frequency can be dropped 50 kHz.

Unfortunately, when this is tried, the improvement in video interference is only about 3 dB. This is because the more tightly filtered data has a higher peak level, and the peaks are visible in the picture. The higher peak level has cost us 3 dB of the expected improvement. Changing from QPSK to QPRS, using the same RMS signal level, does not help, because QPRS inherently has a peak level which is 3 dB larger than the average level.

In making these comparisons, we have found it useful to normalize our signal levels to the nominal peak level, ignoring the overshoots. The eyes shown in Fig. 6 are normalized in this way, with a nominal peak level of $\pm$ 2 divisions. We measure the level of the modulated carrier with a static data pattern which produces a pure carrier modulated with this level. This is how we set the level in our original proposal. Using this method of measuring level, we find that we can meet our target by lowering the carrier frequency by 50 kHz and using either alpha=0.3 QPSK with a level of -23 dB, or QPRS with a level of -20 dB. These two signals will have comparable RMS levels, and using non-redundant error correction in the QPRS demodulator, similar performance in the presence of gaussian noise (the QPSK would actually be about 1.3 dB more rugged).

In order to choose which method to pursue, we have to look at the relative costs, and the performance in the presence of interfering signals. QPRS will require a more expensive decoder. Since the eye is multilevel, two comparators instead of one are required to convert from the analog waveform back to digital data. The level of the QPRS waveform at the comparator inputs is also critical, so a better AGC is required.

The data is subject to interference from both the FM audio carrier, and lower sideband visual information. The receive data filtering can adequately reject any interference from the FM carrier. Interference from video cannot be fully filtered out because it can be inband, depending on the VSB filter in the modulator. If the QPRS and QPSK carriers were used with the identical carrier levels as defined above, the QPSK eye would start out twice as open as the QPRS eye. Since we can operate QPRS with 3 dB more level (as defined above, or at the same level on an RMS basis), the QPRS eye in the receiver will be 70% (-3 dB) the size of the QPSK eye. This smaller eye will be more susceptible to interference from lower sideband video information. Careful measurements show the -20 dB QPRS system suffers a 2 dB interference penalty compared to the -23 dB QPSK system. QPRS looses out on both circuit costs and ruggedness. Despite the attractively narrow spectrum of QPRS, QPSK is the superior solution. Therefore, should stay with QPSK.

The new QPSK carrier frequency will be 1.2 MHz below the vision carrier frequency. This value is very close to 1/3 of the NTSC chroma subcarrier frequency, which is 1.193182 MHz. This is a convenient value to choose, because it can allow future receiver circuits to use the chroma oscillator as a reference to demodulate the data, instead of a requiring a separate crystal oscillator to be locked to the incoming carrier.

The data rate may also be locked to video. We need approximately 512 k bits/sec of data to use a low cost digital audio coding method based on adaptive delta-modulation. It is desirable to use the same clock frequencies that are used in satellite implementations of this audio system such as B-MAC and HDB-MAC. Those systems place integral numbers of audio bits on horizontal lines and so have a direct relationship to video. It turns out that using 1/7 of the chroma frequency as a data clock gives us a total bit rate of 511.363 kHz. With the simplest conceivable multiplex structure, the audio clock rate will be 13 times the horizontal scan rate, which is identical to clock rate used by the BMAC systems. Using the chroma oscillator in the receiver as a data clock reference will save another crystal in the data timing recovery circuitry.

VIII. The 1990 Proposal

The following seems to be the optimum set of parameters for an NTSC compatible digital carrier:

A. QPSK carrier with alpha=0.3 filtering.

B. Transmit filtering phase compensates for a specified receive filter.

C. Carrier frequency 1.193182 MHz (locked to 1/3 chroma) below video carrier.

D. Carrier level -23 dB relative to peak vision carrier level.

E. Data rate 511.363 kHz (locked to 1/7 chroma).

IX. Conclusion

Minor modifications have been made to the NTSC compatible system described three years ago. We explored a number of methods of improving compatibility and settled on a new set of specifications. The new parameters improve compatibility, reduce receiver cost, and simplify transmission of the signal.

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USE OF SAP AND PRO CHANNELS AT WGAL-TV, INC.

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For many years, WGAL-TV has used aural subcarriers for various purposes. Before the advent of the BTSC Stereo System, a 39 KHz subcarrier was used for transmitter-to-studio remote control telemetry. With the growth of ENG and remote pickups, the 39 KHz system was used as a means to deliver the IFB audio to crews in the field. Now we are using the SAP and PRO channels of the BTSC System for the delivery of audio for new purposes.

SAP Development

With the advent of stereo television, WGAL-TV installed a stereo generator during May of 1985. The 39 KHz system could no longer be used and our IFB was shifted to a 450 MHz radio system. At that time, the SAP and PRO channels were not activated.

In 1987, we were presented with a problem that the SAP channel seemed to solve. Like many television stations across the country, WGAL-TV purchases time on local radio stations for our promotional efforts. Most of these promos were originally distributed on audio carts, but this system always seemed to be plagued by quality problems. Audio cart standards simply are not adhered to by all stations and poor quality was the result. We then switched to reel-to-reel recordings and our quality problems were partially solved. For many years, we mailed or delivered these tapes to about thirteen (13) radio stations. However, we kept looking for a better means of delivering these promos. Then in 1987, in an effort to provide daily topical news promos, we investigated various other delivery systems, ie., telephone lines, etc. All proved to be very costly. The SAP channel seemed to be the answer to our delivery problem.

Implementation

We purchased thirteen (13) Sony ST-7TV Stereo Receivers (this unit is no longer available; we now use Sony ST-72TV Receivers) and installed one (1) at each radio station. Surprisingly, most stations had cable TV installed and, hence, they did not need antennas. Those few stations without cable TV were provided with antennas. In a matter of a few weeks, our system was up and running. A specific time was set aside each day to feed promos to the selected stations. We playback a tape which gives each station its specific commercial(s) with the necessary instructions. They record the session on their equipment and, in most cases, the quality is exceptionally good. Distortion and frequency response tests on the SAP channel have proven to be very good.

We began to consider what other uses we could find for the SAP channel. Many different possible sources of program material were considered. We decided that weather information would be the most desirable. The National Oceanic and Atmospheric Administration (NOAA) has a transmitter operating in our area with current weather information available. After a brief discussion with our local NOAA Office, we began to rebroadcast, in real time, the NOAA weather on our SAP channel. Of course, this rebroadcast may not "be sponsored" but is offered as a public service by WGAL-TV. The mechanics of the system are extremely simple. We purchased a rack mounted Gorman Redlich weather receiver and installed it in our MCR area. A simple switching system selects either the weather receiver or the audio tape system used to distribute our promos.

SAP Problems

As soon as we began to experiment with the SAP channel, we knew we had problems. Our telephone operators would receive calls from viewers whose TV sound suddenly went away and was replaced with tones, etc. We rapidly learned that many receivers were designed to default to the TV main audio channel when the SAP channel was not active. Many viewers had their receivers set in the SAP mode, but since no one was using SAP, their sets functioned normally. Suddenly, their receivers were receiving SAP and they didn't know what to do. To solve this problem, we prepared an endless tape that advised viewers that they were listening to the SAP channel and asked them to consult their TV receiver's Operating Manual to find the means to return to the main program audio channel. One serious problem should be noted. TV receiver manufacturers call SAP various names - bilingual, second audio channel, second audio, etc. Many viewers were confused and we answered many calls about this problem. I would not recommend turning on your SAP channel for the first time during a rating period! However, after a few weeks, the SAP problem went away and now very few calls come into the station regarding this problem.

Does Anyone Listen?

We do not have any definite answers about who is listening and how often. We do know that on the few occasions when we have unintentionally deleted the weather from the SAP channel for any extended period, it has resulted in calls from our viewers. We believe that any service that can make viewers tune to your TV channel is a definite plus. The fact that they tune to you for weather also has them looking at your video. Perhaps, they will become interested and remain tuned to watch your program! In the future, we may record our own weather and sell it to a sponsor. It could be another profit center.

Pro Channel Use

In January of 1988, we implemented the Pro Channel. Presently we have our IFB audio on this system. We have also used the Pro Channel to send tone actuated control signals to a LPTV station about 85 miles away. In general, the system works well, but we are not having any luck in finding small hand-held receivers for the reception of the signal. We have constructed receivers for our microwave vans using ISS GL1000A Demodulators with Johnson Electronics SCA/880A SCA demodulators.

In general, the system seems to work well. We hope some manufacturers address our receivers needs in the future.

PREPARATION AND IMPLEMENTATION OF SAP AND PRO FOR MULTICHANNEL TELEVISION SOUND

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In the development of stereo audio for television broadcast, two channels are allocated to provide additional program and technical services. The SAP (Second Audio Program) is specified at 2X fH above the aural carrier to be used in conjunction with, or as a separate audio service of the main stereo program channels. The PRO channel is specified at 6.5 fH above the aural carrier to be used as a technical service for the station, either as an analog or digital signal, (fH=15.734 Khz).

INTRODUCTION

"What's that music I hear when I turn on Channel 8?" ..."It's SAP!"

Informing the general public of your new program service will be as important as the technical implementation. In discussion with other TV stations that are operating SAP program services the one consistent problem has been in getting the message across to the viewing audience. A great deal of effort goes into the technical preparation of the facility for SAP, but there never seems to be enough preparation or understanding in the viewing audience.

The following is a compilation of thoughts, ideas, and problems that we and other TV stations have experienced over the past several years in implementing SAP and PRO services.

SYSTEM PREPARATION AND MEASUREMENT

To prepare yourself to better understand the technical parameters of multichannel television sound, there a number of resources to review. The first is the EIA committee report on Multichannel Television Sound, published by the NAB. In addition, there have been many articles written and published in the major technical publications such as Broadcasting, Broadcast Engineering, BME, TV Technology and several others. Papers have also been presented at the NAB, AES and IEEE conferences. Additional references are available from the MTS equipment manufacturers. It is highly recommended that this information is reviewed prior to system testing and equipment selection.

By this time you may have already installed your TV stereo system, widebanded the aural exciter, provided a second audio channel through to the transmitter and are satisfied with your performance. However, with the addition of SAP and PRO you may very well be stretching the limits of the RF System. The bandpass of the aural cavity in most transmitters should be sufficient to accommodate the wideband BTSC signals. However, attention should be given to all of the tuning parameters while maintaining low VSWR on the input match and peak power output, across the bandwidth. Ideally you will have as good of performance at 400 KHZ as you do for the main channels even though you are not going to modulate to this bandwidth. The next area for concern is the combiner for the aural and visual signals. In particular VHF combiners will exhibit more problems than UHF. Single cavity notch diplexers do not have the same performance specifications as a dual cavity. Therefore, SAP and PRO may be compromised by bandpass, VSWR, temperature variations or other non-linearities. The manufacturer may be able to provide data for your combiner, if not the rental of good test equipment to measure these parameters is recommended. Tuning of the notch cavity in the field is not recommended by most manufacturers. A 3db variation may be acceptable for your immediate service area, a 6DB variation will impair the performance, particularly of the PRO channel. Hybrid combiners do not exhibit these problems, to the extent as do the notch diplexers.

The next area for concern is the transmission line and antenna. Hopefully the transmission line was optimized for the system at the time of installation. Some optimizing can be accomplished in the field to improve VSWR and consequently audio separation and crosstalk in the stereo signal between the main and secondary services.

SECONDARY TRANSMISSION SYSTEMS

Cable companies should be made aware of your added service. If the cable company remodulates your signal from baseband on the cable, they should be informed of the 4.5 mhz options available on TV demodulators and the ability to take the composite aural signal directly into their TV modulators. Cable companies that follow this procedure can successfully deliver the SAP signal to the cable users.

Translators will successfully pass the signal provided they have been tuned properly. Several units have been cascaded across the state, with the signal passing through at least three translators.

TV/VCR/RECEIVER MANUFACTURERS

SAP is known by many different names. Contacting the manufacturers for information on all their MTS units is one way to familiarize yourself with the many different combinations of switches and controls for the various TV's and VCR's. When you receive that first telephone call, you will realize there is no consistency in the terminology. The following list is an example of what we have found:

SAP	BILINGUAL
SA	STR/L2
SA SELECT	AUDIO II
SECOND AUDIO	AUDIO B
SECONDARY PROGRAM	LANG
BILINGUAL/TRACK 1/TRACK 2	

"Where is the switch?" It is found in a number of locations, usually under a cover along with many other switches and controls that tend to intimidate many viewers. The SAP switch can also be located on the remote control unit and not on the TV or VCR.

There are also TV tuners available that provide main channel stereo and SAP.

The manner in which SAP is controlled will vary by each manufacturer. There may be only light indicators on the unit or "on screen" indication. Also, SAP may be listened to separately or in combination with the main channels on different models.

After you have had an opportunity to review the manufacturers literature it would be advisable to contact the major retailers in your area. Informing them of your service and encouraging them to work cooperatively in promoting and educating the viewers as to how the system works. Not all sales personnel understand or demonstrate the SAP functions to the customers.

By this time you should have enough information to put together a basic information packet for your staff and one for the general public. This information packet can be used to promote the service, and as important, make it easy for the viewer to use.

STAFF PREPARATION

Most of those first calls will probably come to your office, but if you can inform the rest of staff of the terminology the Public Relations importance of the implementation should go smoother. As the technical staff becomes more familiar with the typical questions asked by the viewer, it will be easier to instruct the receptionists, secretaries, programming and PR staff. It has not proved difficult to instruct the non-technical staff, and has proven to be a very positive feedback and public relations opportunity for the station.

Demonstrations of a typical TV and VCR for all the staff is beneficial. The hands-on experience builds confidence and encourages more people to field the questions, promote the service and satisfy the viewers concerns and needs.

PUBLIC PREPARATION

One of the first goals is to identify target audiences and user groups within the population. Having identified these groups, contact them to set up demonstrations and promote the service.

All department managers should have access to your information packet and be prepared to discuss this service with the viewers.

Setting up demonstrations for the Press has proven to be quite beneficial in gaining public support and awareness. Press releases directed toward specific groups to be included in newsletters has resulted in identifying additional applications for SAP.

The information describing the services can be kept very simple and include names and phone numbers of station contacts for further information. These information packets, newspaper ads and promo's have generated up to 8 to 10 calls a day for more information and ideas.

The information packet should contain information on what the service is, how to receive it, how to manipulate the controls on the TV or VCR, and a phone number to call for additional help. Another piece of information that you may want to distribute is a list of TV's and VCR's that provide SAP. The viewer should be made aware that most, but not all, stereo units provide for SAP. Such terms as stereo "ready" or "capable" have led to a lot of confusion, at least one manufacturer builds an MTS unit without SAP. The viewers should be encouraged to have the salesperson fully demonstrate the unit before it is purchased.

Last but not least, a 30 to 60 second station promo will help considerably in getting the message to the viewer. "Show and tell" has proven beneficial in informing the viewer if it has been run throughout the days prior to implementing the service. Continuing to run the promos catches the newcomers and those needing additional information. The more you repeat the promo the better the chances of exposing the information to as many viewers as possible.

PROGRAMMING

Fulfilling a community need, providing a public service or taking advantage of an economic opportunity are all good reasons for implementing SAP. SAP does not have to be tied directly to the main program service. The FCC has left the application open to a very broad range of services.

Some examples of programming include: Spanish and Asian translations, other foreign language translations, programming for the disabled or impaired, special community projects, classical music, news and weather, and in-school programs.

As more people immigrate to the community, there is a greater need for these individuals to be assimilated into the english-speaking community. Translations of programs are not readily available, but more agencies are becoming aware of the need to accelerate the learning process for these individuals. One method to encourage and reinforce the learning process is through practical application of what they have learned from the textbooks by viewing TV programs of news, public interest and entertainment, through the process of providing a translation on SAP. Closed captions have been available to the hearing impaired for several years, now a new service Descriptive Video Service (DVS) will provide an aural description of the video action on the TV screen. Translations of news, weather, sports and other public events may prove as popular sources of information. In-school projects, classical music and other special projects are being implemented in various communities.

PRO CHANNEL

The PRO channel is intended to be a technical service for the station. The signal is limited to voice grade audio (3KHz) or for digital control circuits. The limiting factor in implementing the PRO channel is the availability of a receiver. As of this date we have not found a manufacturer who can deliver a unit. The solution has been to take a current SCA receiver and modify it for PRO channel specifications. Each day the unit is used in communicating with the news or production crew. Prior to and during the program the PRO channel is used to communicate between the talent and the host. It has performed quite successfully as long as there is a strong signal received from the station. The notch diplexer for the aural is several db down at the PRO channel frequency. Hopefully a manufacturer will come forward soon with a portable receiver that will have sufficient sensitivity to work under less than desirable conditions.

UTILIZING THE SAP CHANNEL TO SERVE VISUALLY IMPAIRED PERSONS

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INTRODUCTION

Multi-channel Television Sound has brought a new dimension to television broadcasting. The addition of the Separate Audio Program, or SAP, channel brings additional opportunity. The SAP channel allows broadcasters to exploit both non-program and program-related activities. Non-program related activities refer to SAP transmissions that are not dependent upon the regular television signal or program being broadcast, while program-related activities are those that are, in someway, a part of the normal television program. For example, non-program related transmission can be data, a local radio program or a special reading service for blind and low vision people. Program-related transmissions might be a second language track, or program descriptions for visually impaired viewers. While complex, transmission issues related to SAP seem relatively straightforward. Consumer understanding of SAP represents an area of great confusion, coupled with a confusing lack of standardization of SAP on receivers and VCR's.

CURRENT USES OF THE SAP CHANNEL

There is an interesting diversity in the use of SAP throughout the country. In the area of non-program-related activities, KAET in Tempe, Arizona, has been offering the equivalent of a classical radio station, because such a regular broadcast service is not available in their community. Maryland Public Television has been using the SAP channel to provide software and data to schools for use in classrooms. WNET in New York provides program schedule information in an audio format on their SAP channel for part of the day and also is exploring the delivery of a wide variety of services, including news. WGBY, Springfield, and WGBX, Boston, Massachusetts, provide a state-wide radio reading service and a local radio reading service on their SAP channels.

Program-related use of the SAP channel has been somewhat less diverse. A number of local television stations have offered Spanish-language translations of the English audio for news and, on an experimental basis, for programs. Home Box Office provides Spanish tracks on a selected number of movies using SAP. The Public Broadcasting Service, in a project originated by WGBH, has recently launched the Descriptive Video Service, which provides narrated descriptions of the visual elements of a television program, so blind and low vision audiences may more fully understand what is happening in a television drama. The new DVS service is carried, as of this writing, on 32 public television stations. HBO and PBS, to our knowledge, provide the only national services that utilize the SAP channel.

Use of SAP for visually impaired people

Of particular interest to WGBH is the use of SAP for serving the visually impaired. We currently provide both types of service, a radio reading service state-wide network and the national Descriptive Video Service.

Radio Reading Service. The state-wide radio reading service program uses SAP as its primary distribution channel, supplemented by FM subcarriers and participating cable systems. Typically, the service is split into a western and eastern component. This allows more localization of information; eastern users of the system can have Boston-based newspapers read while listeners in the western part of the state would hear Springfield newspapers. This is particularly important when local news, weather, and especially shopping information is provided. National newspapers, magazines and books are also read. Two origination centers, funded in part by the state, provide the service using volunteer readers. The eastern part of the state is served by The Talking Information Center, located in Marshfield, Massachusetts. Their studios are linked via microwave to the WGBX transmitter in Boston. The signal is then broadcast on SAP, from 7:00AM to Midnight, every day. Local cable operator are permitted to take the SAP signal from WGBX and rebroadcast it as an audio "wrap around" on any of their public access channels. In the western part of the state, the Western Radio Reading Service microwaves its signal to WGBY; both located in Springfield, Massachusetts. WGBY follows the same pattern as its sister station in Boston, allowing cable outlets in the extreme western part of the state to rebroadcast the signal on their own cable channel. In areas of the state where there is no participating cable service or that are not reached by the station's own coverage area, local FM stations are contracted to carry the service on a subcarrier.

Since WGBX and WGBY are part of the WGBH Educational Foundation, they are directly linked by microwave. This allows for the instant creation of a state-wide network , if needed. Live programming or announcements can be originated at either reading service site and can be switched onto the network, providing full coverage. This is a feature that has not been fully utilized, but is planned to be used in the future.

Descriptive Video Service. Perhaps the most exciting use we have found for the SAP channel has been the development and recent launching of a new service, called the Descriptive Video Service. This is a national service that makes regular television programs accessible to visually impaired viewers. DVS was developed and tested by WGBH over the last five years. In January, 1990, with the beginning of the American Playhouse series, DVS was launched on 32 public television stations, covering one-third of the country's television households.

DVS provides narrated descriptions of the visual elements of a television program. These narrations occur in the pauses between dialogue. Descriptions of the action, locations, characters, costumes, etc. are provided. Describers, working at a specially designed computer work station, create the descriptions, using the program time code to accurately locate the appropriate pauses in the dialogue. A final script is prepared and then recorded by professional narrators. The narrated audio track is then placed on the third audio track of a 1-inch videotape master, maintaining sync with the regular program. The program is distributed by PBS via satellite uplink, including all three audio channels. Participating local stations route the third audio track to their SAP inserters for final transmission.

Initial response to DVS has been overwhelmingly positive. During this first operational year, additional PBS dramatic programs will be added: *Mystery!, Masterpiece Theatre, and Degrassi High.*

Utilizing the SAP channel for both local and national purposes is not without confusion. In our testing and in the current operational phase of DVS, the techniques of syncing the description track with the program and uplinking and retransmitting the SAP signal have been relatively straightforward. Areas of confusion and concern have centered mostly on equipment and usage at the consumer end.

CONSUMER CONFUSION

We have found that most consumers are unaware of the existence of the SAP channel on their stereo receivers and vcr's. Even after running the radio reading service for several months, we get several calls a week from viewers telling us that we have somehow "messed up the audio." Viewers tell us that we, the station, are transmitting the wrong audio. We ask the caller if they have a stereo tv and if they do, to check to see if they have selected SAP instead of stereo. Even after confirming that the viewer has had the SAP button selected, we are sometimes told that it is the station's responsibility to "fix it." We have tried to do so by making regular announcements about the SAP channel on both the SAP signal and our regular air. Our launch of DVS compounded this problem, since viewers this time were able to hear the regular program audio mixed with the descriptions. We expect this to be an ongoing public education issue.

Retailers

We have also found that local retailers do not understand SAP and may advise consumers that they don't need it or that SAP is only for Spanish language broadcasts. Every station using the SAP channel is experiencing this problem. Short-term solutions require each station to direct an education campaign to every retailer in their coverage area. WGBH is trying to implement a long-term solution by reaching receiver manufacturer individually and through organizations such as CES and EIA.

Equipment

Another area causing confusion in the use of SAP is the lack of consistency in the way receivers handle SAP. Some sets assume that viewers would want either SAP or stereo. Therefore, when SAP is selected, the regular program audio disappears, leaving only a SAP signal. In the case of radio reading service or non-program related activities, this is desireable. In the case of program-related audio, such as DVS, this necessitates the preparation of a mix track for SAP. The program audio must be pre-mixed with the descriptions before being synced to the videotape master. This increases the cost of producing DVS. Some receiver anticipate program-related uses and provide for SAP Only, Mix and Stereo. The mix switch usually places the SAP signal on one (left) channel and the program audio (monaural) on the other channel. This is desirable for DVS.

To make this situation even more confusing, SAP is known by different names. Some receivers label the SAP channel as the "Second Language Channel," "Second Audio." "Audio 2," or "SAP."
Owner's manuals do little to clear up the confusion.

Cable

Once a consumers have gotten past the confusion of what SAP is and how it is selected, they may run into another confusing area if the television signal is received on cable. Some cable operators strip stereo completely. Other cable systems may pass enough of the signal to light up the stereo indicator but still strip the stereo. We wonder how many consumers might think they are receiving stereo, because their stereo light is on, but are really receiving monaural sound. We would have to believe that such consumers would not be too impressed with stereo television sound.

Stereo vs. MTS Stereo

There is one final area of confusion to be noted. Many home videotapes are stereo, but a type of stereo that is different from broadcast stereo. Recently I wanted to buy a new VCR. My plan was to buy a stereo VCR so that I could use the built-in tuner to decode broadcast stereo, particularly SAP. I found stereo VCR's that were marked stereo but were designed only for stereo home videos. I found that the retail salesperson was not particularly helpful in explaining the difference between stereo and MTS stereo. Without my own "inside knowledge" I surely would have wound up with the wrong equipment.

CONCLUSION

Many of the problems presented here can be solved by better communication between broadcasters, producers, and equipment manufacturers. All stations broadcasting in stereo need to communicate with the manufacturers and especially with the retailers in their communities. Standardization of SAP nomenclature and SAP selection on receivers and VCR's will surely require station and producer communication, but the effectiveness of communicating these issues would be increased by the participation of key industry organizations such as the NAB. If the Separate Audio Channel is to become a universally useful addition to television broadcasting, then all interested parties must begin communicating with each other now, before misconceptions become too firmly entrenched.

DEVELOPMENT OF A GHOST CANCEL TECHNOLOGY FOR TV BROADCASTING

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Abstract

As buildings in urban areas become higher and more numerous, we are experiencing increasing ghost interferences in receiving TV signals. This interference has now become the leading cause of deterioration in the qualities of received TV pictures.

One way to reduce ghost interference is the use of a "ghost canceler" designed to eliminate ghost signals inside a receiver. However, without proper reference signals which are used to detect ghost signals, there will be a limit to reducing this interferences.

In an effort to eliminate the ghost interference, a group to which we belong in the Broadcasting Technology Association (BTA) has developed a ghost cancel reference (GCR) signal[1]. We have inserted a bar signal, which has a rise wave form obtained from integrating a sin x/x pulse, and a 0IRE pedestal signal into the vertical blanking period. These two signals are sent in a manner of specially developed "8-field sequence" from transmitter. The ghost canceler is able to reduce ghosts with a delay time of about 45 microseconds or less. This new GCR method was adopted as Japan's standard for broadcasting and introduced in the fall of 1989.

Introduction

Many methods were already developed for a ghost elimination. A reception antenna with high directivity is sometimes effective to reduce ghost interference, but it cannot reduce ghost interference signals which come from many directions or follow almost the same direction as the desired TV signals.

As a measure to eliminate the ghost signal, a "ghost canceler" built into a TV set was developed a few years ago in Japan. However, the canceler currently able on the market relies on a vertical synchronizing signal to detect the ghost, and because the frequency characteristics and its waveform are inadequate as the reference signal for the ghost detection, it is difficult maintain and standardize the system. To improve the situation, it was suggested that a reference signal be inserted into part of the image signal. This idea fell through as well because of inadequate frequency characteristics and linearity due to the use of a CCD (charge coupled device) as a transversal filter to eliminate the ghost signal. Thanks to the further digitalization of the circuit in recent years, however, we currently have highly advanced full digitized ghost cancelers. What is needed now is a GCR signal good enough for the canceler.

Research and development into a GCR signal best-suited to ghost detection was carried out by BTA's Ghost Canceller Committee, comprising TV broadcasters and manufacturers of TV receivers2).

The committee first set target functions by studying the conditions that the GCR signal should satisfy. Several GCR signal methods were then tried through indoor simulation and field tests using broadcasting signals were carried out in an effort to find a signal best-suited to meet the required conditions. As a result, the committee has come up with a GCR method in which a bar wave form with the rise characteristics of a sin x/x function and a zero pedestal wave form are inserted into the vertical blanking period with an 8-field cycle. As this method has flat frequency characteristics up to about 4.1MHz and maintains sufficient energy to some 4.2MHz, it can deal with almost all kinds of ghost interference and has an adequate wave form equalization function for EDTV. Further, with its ability of delay time to detect ghost signals to approx. 45 microseconds, this new method can be used to eliminate ghost signals with very long delay time such as called "blanking ghost" (ghost image of a horizontal sync. pulse and color burst are appeared on the screen).

In the field tests, we found that by inserting the GCR signal over the broadcasting signal, this reference signal possessed enough ability to eliminate ghost interference under any conceivable conditions.

Principle of Ghost Canceling

Fig. 1 shows the principle to eliminate the ghost signal. We obtained this diagram showing a series of processes from generation of ghost signals to their elimination by treating as linear distortion all wave form distortion including ghost signals generated in the transmission, radio wave propagation and reception systems, because the high-frequency distortion can be calculated in terms of the baseband.

As all of these systems are connected in series, their frequency characteristics can be expressed as multiplication on the frequency axis. Therefore, the optimal frequency characteristics for ghost elimination can be determined by detecting distortions in the frequency characteristics of the GCR signal on the reception side.

The ghost signal can also be eliminated by detecting distortion on the time axis instead of the frequency axis. Such ghost canceler circuits remove ghost signals by first calculating the difference, or error, between the output of the transversal filter and the reference signal level and then adjusting the tap coefficient so as to minimize the error.

Requirements for the Ghost Cancel Reference Signal and Selection of GCR Wave Forms

The GCR signal should meet the following requirements:

(1) An adequate high-frequency area and the ability to define wave forms with high precision for efficient wave form equalization and color ghost removal.

(2) Few insertion lines and a wide ghost delay time detection area.

(3) The ability to prevent error detection caused by ghost signals beyond the delay time to be cancel.

(4) A high S/N as a wave form.

To find proper GCR signals that satisfy the required conditions, we examined seven types of wave forms as shown in Table 1. This table show the each wave form's specification. We evaluated these waveforms for twenty-four different categories. As a result of the first stage evaluation work, we selected the four wave forms as candidates for the reference signal, shown in Fig. 2.

(1) Bar wave form with sin x/x fall characteristics (Named B)

(2) Pulse wave form with sin x/x characteristics (Named P)

(3) Pseudo random noise signal wave form with a transmission rate of 7.16Mbps (Named PN)

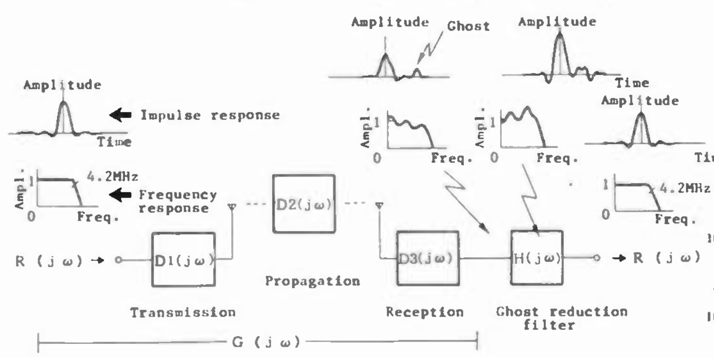
(4) Bar wave form with sin x/x rise characteristics (Named RB)

Final selection of the GCR wave form to be used as a reference signal was made by subjective and objective evaluation tests using a ghost simulator and in real field conditions.

These tests are held under the preconditions of no radical fluctuations in the strength of the electric field, ghost wave levels or phases due to flutter.

Evaluation Methods for performance of the GCR Signal

To evaluate the degree of ghost interference, we relied on two methods: one a subjective



R (jω) : Ideal frequncy response on base band signal
D1 (jω) : Frequency response of transmission
D2 (jω) : Frequency response of propagation with ghost
D3 (jω) : Frequency response of reception
G (jω) : R (jω) × D1 (jω) × D2 (jω) × D3 (jω)
Amount of frequency response in front of ghost reduction filter
H (jω) : Ideal frequency response of ghost reduction filter

$$H(j\omega) = \frac{R(j\omega)}{G(j\omega)} \quad (1)$$

Fig. 1 Principle to eliminate the ghost signal

Table 1 Specification of examined GCR wave forms

Wave form	Frequency charac. 3.58 MHz (dB)	Frequency charac. 4 MHz (dB)	Amplitude level (IRE)
T pulse	-4.7	-6.0	100
T bar	-4.7	-6.0	80
sin x/x pulse	0	0	80
sin x/x bar	0	0	80
1.12 T bar	-6.1	-7.7	80
Pseudo random	-6.0	-15	70
Vertical sync.	-4.7	-6.0	40

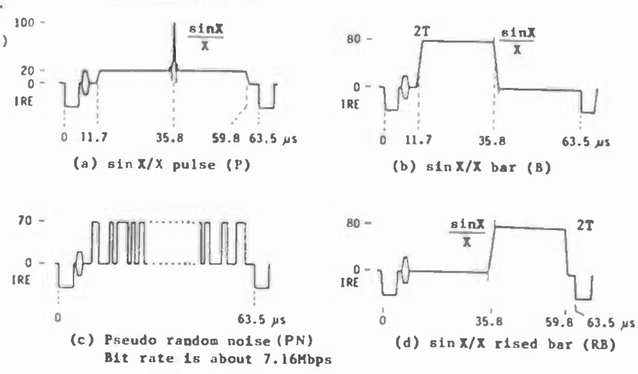


Fig. 2 Four candidate wave forms of GCR

evaluation of 5 point impairment comment scale with CCIR and the other an objective measurement of the video signal wave forms.

(1)Subjective Evaluation

To evaluate the degree of ghost interference on the TV screen, we can use ordinary TV programs and TV pictures especially designed for this purpose. As this is an subjective evaluation, results are often affected by the contents of such TV programs. To eliminate this difficulty, we used a still picture as shown in Fig. 3 in the indoor experiment and the field tests in which test signals were used. This still picture, called the "Woman and Pillar," was created to facilitate detection of the ghost image.

(2)Measurement of Residual S/I and Pseudo PDUR

Residual S/I after the ghost component is removed can be measured using a GCR wave form or its equivalent. However, if such measurement is conducted with the GCR wave form used in the removal of the ghost signal, precise measurement is often difficult due to non-linearity of the ghost canceler and the demodulator.

To overcome this problem, we used GCR wave forms other than used in the removal of the ghost signal in the measurement of S/I[3]. In the actual measurement, we used the rise and fall wave forms of a bar signal and also used a pulse signal as a support. We obtained our results by averaging the Pseudo PDUR values gained by calculating residual components of ghosts measured with the rise and fall wave forms.

The PDUR[4] is calculated from the following formula using the DU ratio of the ghost wave, delay time (τ) and high-frequency phase difference (ϕ). The result of this calculation gives objective evaluation values with high correlation to the results of the subjective evaluation.

$$PDUR=-10\log\sum_{i=1}^{n}10^{-\frac{Gi}{10}} \quad (dB) \qquad (2)$$

There, $Gi = D/U + W_{\tau}i + W_{\phi}i \quad (dB) \qquad (3)$

$W_{\tau}i$ and $W_{\phi}i$ are Phase and delay time weighting value where is a function of τ and ϕ respectively.

With the ghost canceler evaluation, in which video signals after cancellation are used for measurement, information in the high-frequency phase is lost, leaving only in-phase and negative-phase $W\phi i$. It is therefore called the pseudo PDUR. Since phase information is not available, we checked in the experiment correlations with the results of the subjective evaluation. The correlation coefficient was 0.9 after linear regression of the experimental results used a still picture.

Fig. 4 shows the schematic diagram of measurement for the residual S/I and the pseudo PDUR. The measurement waveform of ghost was AD converted and added synchronously for many times, and we process the signals by transferring the data to computer after AD conversion of the video signal. When the bar signal was used, computer differentiated this wave form into a pulse wave form with its process, then after calculation display the in-phase and negative-phase component respectively on the CRT screen as shown in Fig. 5, with (a) before ghost cancel and (b) after. The numeral in the figure is the pseudo PDUR. The horizontal axis represents the delay time and the vertical axis the S/I ratio, with the middle indicating 50dB and the upper and the lower sides each extending up to in-phase 0dB and negative-phase 0dB.

With 256 times as the standard frequency of synchronous additions, we improved the S/N by about 24dB. It is possible to obtain a peak S/N of 50dB (P-P/P-P) after synchronous addition if the S/N of the signal to be measured is 40dB (P-P/rms) or more. If S/N is smaller than 40dB, it can be improved by increasing the frequency of synchronous additions.

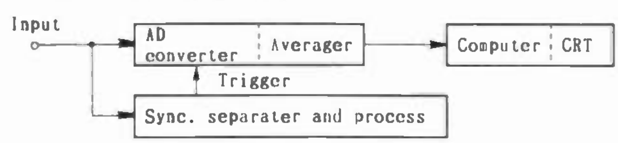
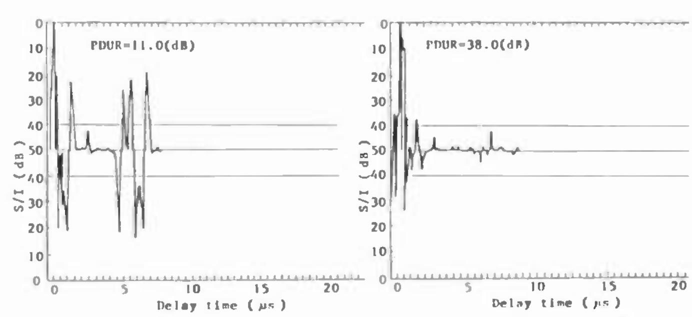


Fig. 4 Ghost measuring equipment for residual S/I and pseudo PDUR



(a)Before cancel (b)After cancel

Fig. 5 Residual S/I and pseudo PDUR

(3)Measurement of Residual Color Ghost S/I

The wave form of pseudo PDUR is measured using sin square 2T characteristics so that it can be compared with an ordinary image. Because of this, evaluation can be imprecise for a picture with high color saturation. To avoid errors, we measured the residual color ghost using a special color burst wave form[3] whose phase does not become inverse every sync.line.

Experiment to Evaluate GCR Wave Form Qualities and Results

(1)Evaluation System

Fig. 6 shows a diagram of the equipments used in the experiment we conducted to study the functions of the GCR wave form[5]. With indoor testing, we modulated the wave form with the image signal of the still picture and added a ghost signal using a ghost generator.

We prepared three ghost generators each capable of generating two ghost waves using the delay lines of SAW(surface acoustic wave) devices. With these devices, we could add six waves and a maximum delay time of 20 microseconds. As a demodulator for detecting TV signals, we used a standard PLL synchronous detection for professional use capable of sampling the position of sync chips and pedestals.

The digital VITS generator used in the experiment has an inserter function with a 10-bit ROM which stores the wave form data. We input the GCR wave form data in the ROM.

We used a 1-inch VTR to reproduce the prerecorded image signals containing detected signals including the ghost wave form. We had already recorded in advance a reference ghost signal on video tape so that we could conduct experiments under the same ghost wave conditions at any time.

We prepared four ghost cancelers which could deal with the candidate GCR wave forms and also one conventional device based on the vertical synchronous signal.

(2)Experiment Results

Table 2 shows the result of indoor function evaluation using the ghost canceler most effective in eliminating ghost interference among the devices used in the experiment.

For each ghost condition, the results of subjective evaluation and the pseudo PDUR before and after the improvement are shown. For the pseudo PDUR, values measured with the rise and fall bar wave forms are indicated on the upper and lower sides in each column. The each column contains a mean of several conditions of ghost. The results with the broadcasting signals are provided as a reference since they were the signals received indoors.

From this data we can see that both rise and fall bar wave forms offer the best results, followed by the sin x/x pulse and then the pseudo random signal. As the evaluation results of the color residual ghost shown in Table 3 suggests, the pseudo random signal column has about 20dB more in residual S/I than other wave forms, supporting the results of the subjective evaluation and the pseudo PDUR measurement. With the pulse wave form, DC offset error occasionally occur depending on ghost conditions, causing the ghost canceler to malfunctioning.

These results clearly indicate that a bar wave form with sin x/x rise or fall characteristics is the best in terms of stability and ability to eliminate ghost interference.

8 Field Sequence Method to Deal with Long Ghost

(1)Ways to Prevent Ghost Detecting Errors Caused by very Long Delay Time Ghost

For both sin x/x bar wave forms for the rise and fall of the signal, bar wave forms measuring about 24 microseconds in width are inserted in one horizontal scanning period. If a ghost signal with a delay time of 24 microseconds or longer is present, there will be overlapping after the standard ghost detection point of such ghost wave forms as rise wave form itself, synchronizing signals before the bar wave form, color burst and previous line signals as shown in Fig. 7. By mistakenly taking these wave forms as "ghost," the ghost canceler will add pseudo ghost. The result will be the same if the pulse wave form is inserted in the middle of a single horizontal scanning period.

To eliminate this inconvenience, we have developed a new method in which only GCR wave forms will be extracted by inserting different GCR wave forms in the vertical blanking period depending on the fields and by subtracting the values between two fields. An example is shown in (a) and (b) of Fig. 8. In Fig. 8(a), we created a field by inserting a pulse wave form with the amplitude of 100IRE and, in Fig. 8(b), another field by inserting a pedestal signal with the amplitude of 0IRE. The delay time of a ghost signal was set at 24 microseconds or more. The ghost element that occurred during the detection period was eliminated by subtracting signals in (b) from those in

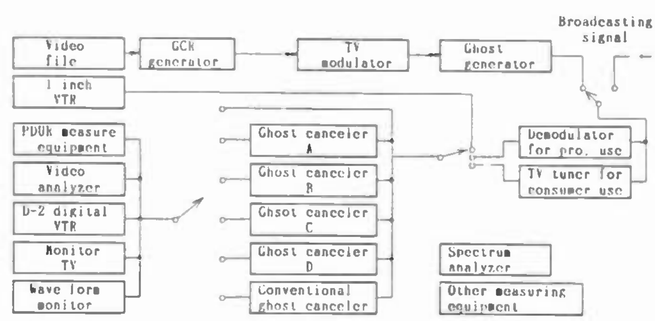


Fig. 6 Block diagram of equipment for GCR test

Table 2 Experimental results used with ghost generator (Mean of several conditions of ghost)

		G C R				
		Before cancell	sin X/X bar	sin X/X pulse	P N	sin X/X Rised B
Subjective evaluation (MOS)		2.3	4.2	4.1	3.6	4.1
PDUR(dB) measured with	RB	12.7	37.2	34.0	29.5	44.4
	B	13.4	43.1	38.5	31.5	36.2

Table 3 Residual S/I of color component (Mean of several conditions of ghost)

	G C R			
	sin X/X bar	sin X/X pulse	P N	sin X/X rised B
S/I(dB)	30.8	35.3	14.6	32.4

(a), leaving only the GCR wave form in (c). We called the method 'Field sequence Method', and this principle can be applied to both rise and fall bar wave forms.

In using the rise of a bar wave form, a bar wave form with wide width is desirable if a long delay time in ghost detection is required. When a bar wave form is inserted in one horizontal period, the delay time is limited by the effective scanning period to a maximum of about 45 microseconds. We call this bar wave form a "Wide Rise Bar (WRB)" as it has a wide width and detected at rised edge for ghost detection.

Detailed Specifications of the GCR Signal and Confirmation Experiment of Improvement with a Broadcasting Signal

After carefully studying the results of the experiments discussed above, we have finally decided on the WRB/8 field sequence as the best GCR method[6]. Fig. 9 shows detailed information on its wave form[7].

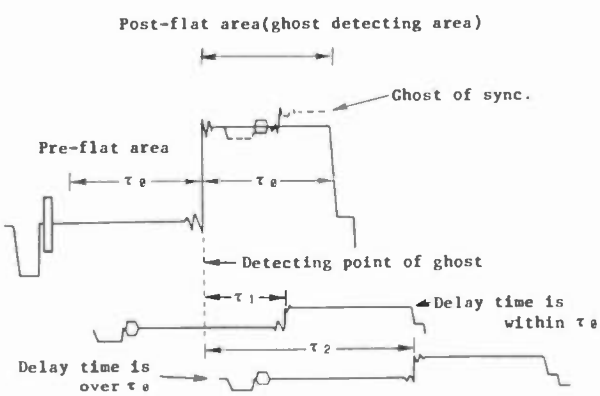
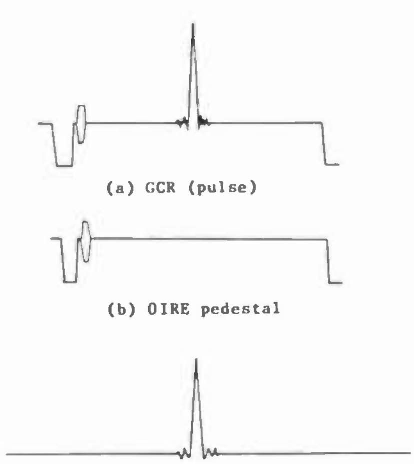


Fig. 7 Mechanism to detect ghost mistakenly by long delay ghost

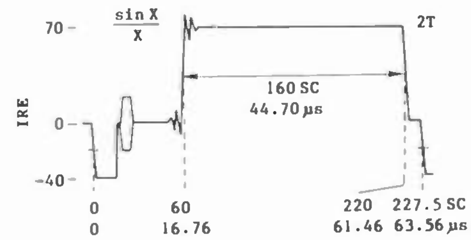


(a) GCR (pulse)

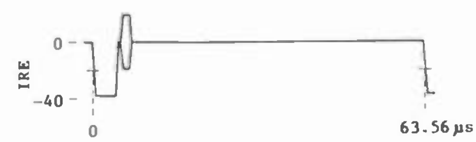
(b) 0IRE pedestal

(c) Subtract (b) from (a)

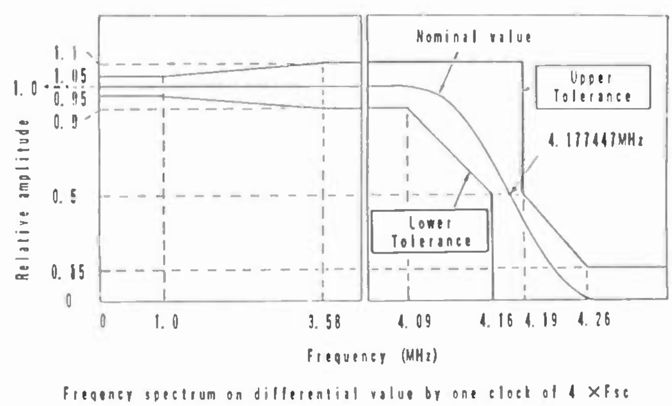
Fig. 8 Subtract method by field sequence of GCR



(a) GCR wave form



(b) 0 pedestal wave form



(c) Frequency response of GCR

	Sequence number of field							
	S1	S2	S3	S4	S5	S6	S7	S8
Frame No.	1		2		1		2	
Field No.	1	2	1	2	1	2	1	2
8 field sequence	GCR	0	GCR	0	0	GCR	0	GCR
C. burst phase	+	-	-	+	+	-	-	+
Previous line signal	X_A^-	X_B^+	X_A^+	X_B^-	X_A^-	X_B^+	X_A^+	X_B^-

GCR: GCR wave form 0: 0IRE pedestal wave form

color phase of chroma signal +: θ degree, -: θ+180 degree

$X_A^- = Y_A - C_A$ $X_A^+ = Y_A + C_A$ $X_B^- = Y_B - C_B$ $X_B^+ = Y_B + C_B$

Y is luminance component. C is chroma component

Caluculation formula of field sequence is as follow:

No. of field sequence	Calculation formula of field sequence GCR=	previous line signal after calculation
8	((S1-S5)+(S2-S6)+(S3-S7)+(S4-S8))/8	0

(d) 8-field sequence

Fig. 9 Final GCR signal

(1)Wave Form, Width and Position

The bar wave form is disadvantageous in terms of S/N as it is made into a pulse by differentiation for ghost detection, lowering its signal levels in the course of differentiation. However, it is less susceptible to DC fluctuation and can therefore perform ghost detection with stability. Further, it is possible to fill the effective horizontal scanning period, the detection range of ghost signals, with the width of the bar wave form because the field sequence method enables us to retain the no-signal part for more than one line before the ghost detection point. Here, we obtained a bar wave form with a width of 44.7 microseconds by subtracting from the effective horizontal scanning period the ringing time before the sin x/x rise.

For ghost detection, the ghost canceler uses a GCR wave form obtained by AD conversion at a clock time of 14.318MHz, four times the frequency of the color subcarrier. Because of this, the position of the GCR wave form should be within jitter of ±40 degrees when seen from the color burst. If jitter is larger than this, frequency characteristics will lose more than 1dB for each 3.58MHz, raising the frequency characteristics after ghost cancellation.

(2)Level

Generally speaking, the higher the signal level is, the better the S/N. However, non-linear distortion of the detection characteristics will appear in the case of quad phase ghost signals, because the time constant to choose the carrier of the PLL synchronous detection IC used in a commercially available TV tuner is not so large[8]. By testing the amplitude levels and ghost improvement effects in the experiment, 70IRE was found to be the best after taking S/N into consideration.

(3)Frequency Characteristics

It is desirable that the level of frequency elements 3.58MHz or higher be as large as possible in order to remove edges of superimposed words and color ghost. It is also necessary to have a flat energy level up to about 4MHz, the maximum frequency in the NTSC system.

The bar wave form is first differentiated into a pulse and then used to detect ghost signals. In a ghost canceler, the bar wave form is differentiated at about 70 ns which is 1 cycle (1 clock) of 14.318MHz in order to extend the frequency characteristics as much as possible. With this 1 clock difference, the Nyquist frequency is about 4.18MHz and the wave form has a sin x/x function with flat frequency characteristics up to about 4.09MHz.

(4)8-Field Sequence Method

Wave forms of the horizontal synchronous pulse and the luminance signal can be the same in any fields, but the phase of the color burst and the color signal elements differs from one field to another, returning to the original phase and therefore the same waveforms every four fields. Also, signals of the previous line inserted in the first and second field often differ.

After studying the various sequences to do away with this difficulty, we have selected an 8 field sequence, which enables us to eliminate the color component of the previous line signal as well as the color burst in one computation6). Fig. 9(d) shows a conceptual diagram of this idea.

(5)Confirmation of Interference Removal Effects with Broadcasting Signals

In Japan, we use 14(277)H - 16(279)H and 21(284)H as a vertical blanking period for teletext and 17(280)H - 20(283)H for VITS. Of these, CCIR's first and second VIT signals or others with similar wave forms are inserted in 17(280)H as a test signal by a broadcasting station. Because of this, 18(281)H is judged appropriate for the 8 field sequence method.

We used the 8 field sequence method to insert 18(281)H GCR signals at six broadcasting stations in Tokyo and five stations in the Kansai area and confirmed the ghost improvement effects using the carrier signals. In the Kansai area, the effects were confirmed at the end of transmission after up to five relay stations.

Fig. 10 shows the results of the experiment conducted in Tokyo. Figures on the horizontal axis represent pre-improvement evaluation and the vertical axis indicate that after improvement. It is clear that the ghost signal with a subjective evaluation value of 2.5 or more was almost improved to 4 or more. Similarly, the pseudo PDUR value of the ghost increased from 20dB or more to 30dB or more after elimination of the ghost. These results testify to the validity of the method.

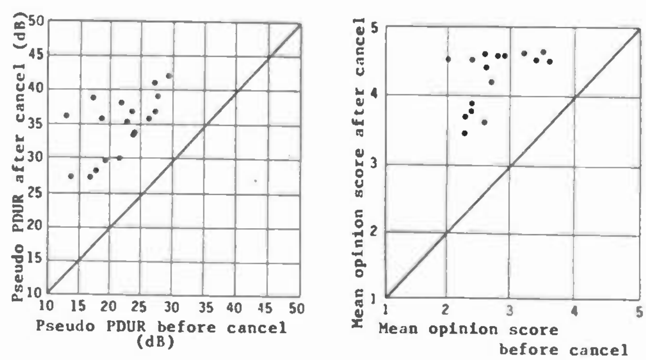


Fig. 10 Improvement results of ghost at Tokyo

General Structure of Ghost Reducer

Fig. 11 shows the general structure of the ghost reducer which contains the near-by ghost canceller, normal ghost canceler, GCR signal receive memory, clock generator, timing generator and CPU.

The near-by ghost canceler is a non-recursive digital filter circuit which eliminates the near-by ghost including the prior ghost. The normal ghost canceler is a recursive digital filter circuit which cancels the ghost after the near-by ghost.

The digital filters for the near-by and normal ghosts requires approximately 650 taps in total (multiplies and adders are about 650 each) to process 44.7 μs of GCR signal.

The ghosts eliminates algorithm specifies the area of the GCR signal receive memory as shown in Fig. 11. In these areas, the processings are performed at a high speed and the dual port memory structure is required to interface with the CPU.

The clock generates a clock which actuates digital signal processor such as a digital filter. To attain clock stability, the generator provides a phase-locked color burst signal as a clock. In general, the four-times color subcarrier (14.32 MHz) is used as a clock.

The timing generator produces a timing signal for sampling the GCR signal from the video signal. It is necessary to guarantee the stability of timing signals, even though ghost is excessively input.

The CPU detects the ghost element from the GCR signal and controls the near-by and normal ghost cancelers to eliminate the ghost. An 8-bit CPU can manage the feedback control algorithm. For the feed-forward control algorithm, however, a 16-bit or more CPU is required to perform the great number of calculations.

Configuration of A Practical Ghost Reducer

There are many algorithm to reduce ghost. Two typical algorithms and hardwares are explained here.

(1) Ghost Reducer Employing A Feed-forward Algorithm

The Fig. 12 shows an overall view of one of the Ghost Reduction Tuners and its remote controller. It uses a ghost elimination algorithm (dividing method)[9] under the particular feed-forward control and a high-performance video IF circuit to realize the high-speed, high-stability and high-performance operation.

This selection describes the main features of the device, the ghost elimination algorithm and the filter structure in a ghost canceler.

Main Feature

1) The keyed APC type PLL synchronous detection circuit and the keyed blanking AGC circuits to installed as video detector circuits to enhanced the ghost eliminating ability.
2) A wide range (-1.5 to +40 μs) of ghost can be eliminated because the device handles the GCR signal.
3) the ghost canceling time is less than 5 seconds, as a result of the feed-forward controls algorithm (dividing method).
4) The ghost level indicator clearly indicates the condition of ghosts that are received.

Fig. 12 Overall view of the ghost reduction tuner and its remote controller (NEC GCT-1000)

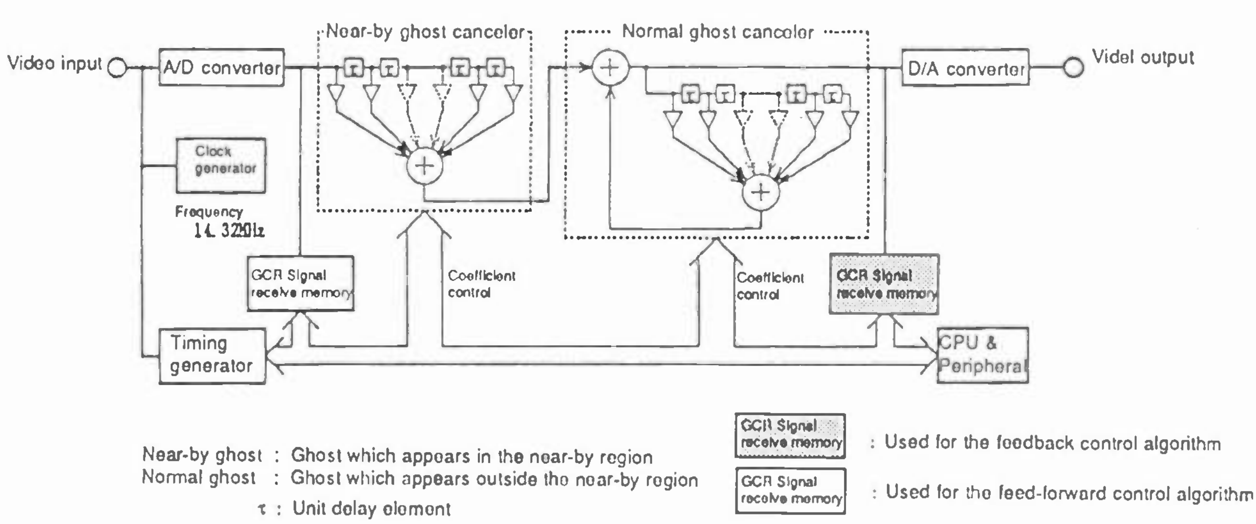


Fig. 11 The general structure of ghost reducer

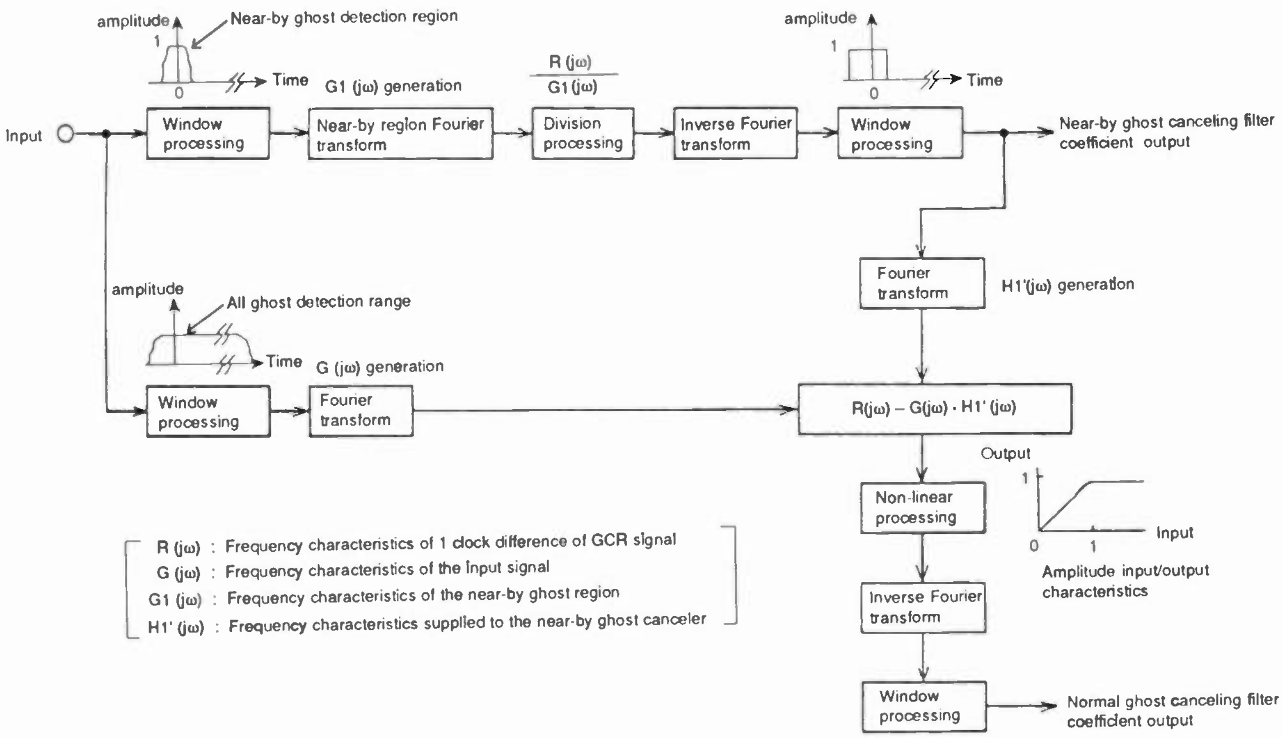


Fig. 13 Ghost elimination algorithm (Dividing method)

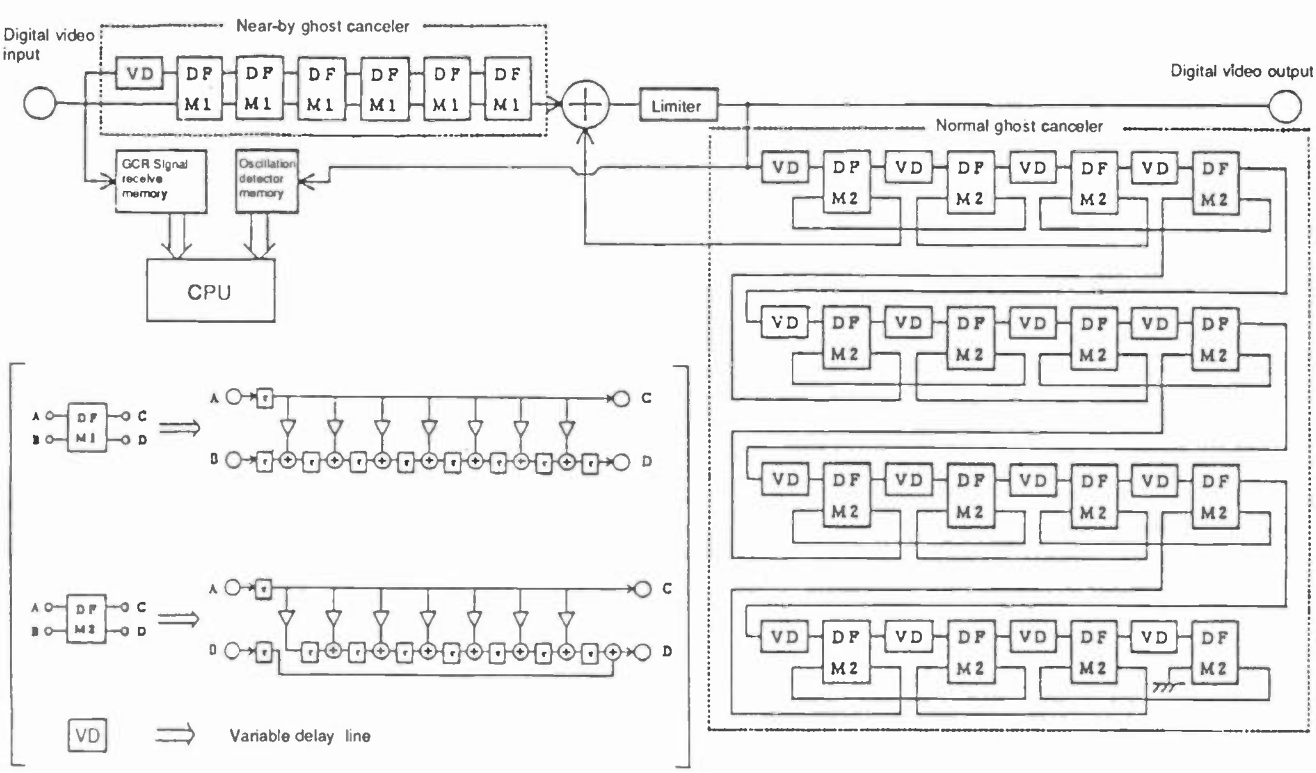


Fig. 14 Filter structure in a ghost canceler

Feed-forward Algorithm

A feed-forward control algorithm is adopted for a ghost elimination. The feed-forward control algorithm such LSE (Least Absolute Error) method, which has been studied so far. Under this algorithm, the device directly analyzes the characteristic of transmitting channel using the received GCR signal, and calculates the formula (1).

In this algorithm, the response characteristic of the sin x/x pulse is obtained using the GCR signal. Refer to Fig. 13 for details.

Filter Structure in A Ghost Canceler

The filter structure in a ghost canceler is shown in Fig. 14. The near-by ghost canceler consists of 42 taps of non-recursive filters (7-tap filter × 6) to remove -1.5 to +1.0 μs of ghost.

The normal ghost canceler is comprised of 112 taps of recursive filters (7-tap filter × 16). These filters can move to the arbitrary position using a variable delay line, eliminating ghosts much more effectively than the general structure of +1 to 40 μs of ghosts.

(2)Ghost Reduction Employing A Feed-back Algorithm

Fig. 15 shows the external view of the ghost reduction tuner. The tuner set is controlled by a 4-bit microprocessor in a tuner unit. The tuner employs the frequency synthesis technics in order to realize stable synchronous detection. A PLL-synchronous detector in the IF stage has an improved phase response so that the ghost interference can be linearly detected as far as possible.

The ghost and the waveform distortion are reduced in the video stage[10]. Newly developed transversal filter (TF) IC and control IC largely cut the total cost. An unique algorithm realizes both minimum residual ghost and stable operation.

Most results obtained from field test shows satisfactory improvement in ranking of between 0.5 and 2 in subjective evaluation as shown in Fig. 16.

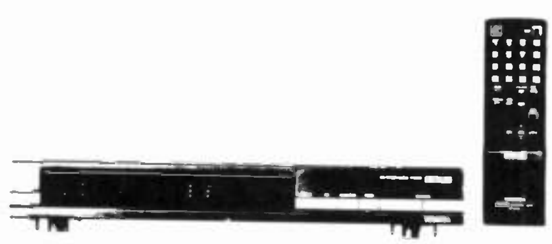


Fig. 15 External view of ghost clean TV tuner and its remote controller (Toshiba TT-GC9)

Configuration of the Ghost Reduction Part

Fig. 17 shows the block diagram of ghost reduction part. Clock signal (4fsc) and synchronizing pulses (H, V) are regenerated from input video signal. The 8-bit digital video signal via an A/D converter is fed to both one transversal filter (TF) IC and the control IC. The TF-IC is assigned to the delay between -2 and +1.7 microsecond.

The control IC (Gate Array) calculates tap coefficients and supplies them to the corresponding TF-ICs under the sequence control of a 8-bit microprocessor. The five TF-ICs in the recursively connected six TF-ICs are assigned to the delay between 1.7 and 24 microsecond. The last one equipped with a delay line is shifted and assigned to the delay between 24 and 40 microsecond according to the delay of ghosts.

Transversal filter (TF) has the bit width for the input video signal and the tap coefficients of 8 and 10 bits, respectively. They contribute stable operation. Highly integrated 64-tap TF-IC largely cuts the cost.

Control IC is the gate array IC with 10.6k gate. It has such functions are calculation, direct memory access (DMA) control, and communication with the microprocessor in the unit. Calculation is most characteristics function among them. Such four modes as correlation, bit shift, 16 bit addition/subtraction, and block transfer are performed using 280 nsec. clock. High speed calculation contributes rapid convergence of tap coefficients.

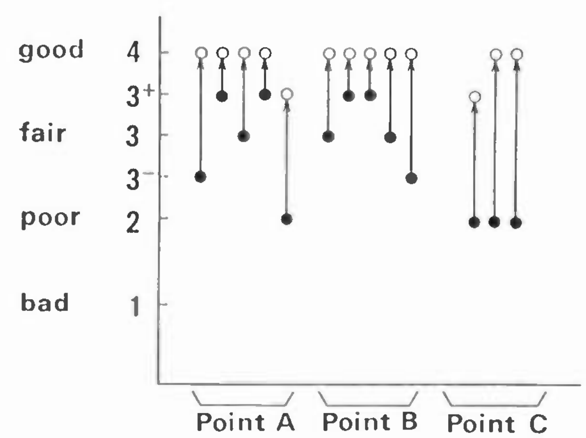
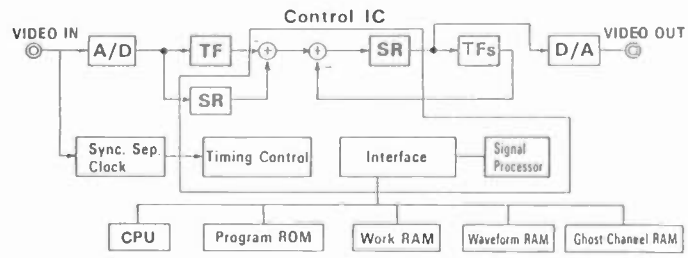


Fig. 16 Results obtained from field tests



TF : Transversal Filter, SR : Shift Register

Fig. 17 Block diagram

Feed-back Algorithm

Fig. 18 shows the operation flow. When the power is switched on or the TV channel is changed, tap coefficients are supplies to the TF-ICs via the corresponding ghost-channel memory. Thus, once the tap coefficients for the corresponding channel are stored in the ghost-channel memory, ghosts are reduced width one second. Then equalization loop is started. Under the sequence of the microprocessor, the control IC mainly performs the following iteration algorithm:

1) to acquire input GCR and output GCR,
2) to get differentiated GCRs (X_k and Y_k),
3) to detect the address (p) of maximum value in X_k,
4) to calculate error (E_k) according to peak address (p)

$$E_k = Y_k - R_k$$

where R_k designates the reference signal,

5) to calculate correlation results (D_k)

$$D_k = \sum_{j=p-m}^{p+n} X_j \bullet E_{j+k}$$

where m and n are small natural numbers,

6) to adjust tap coefficients (C_k)

proportional:

$$C_{k,new} = C_{k,old} + A1 \bullet D_k - B1 \bullet C_{k,old}$$

incremental:

$$C_{k,new} = C_{k,old} + A2 \bullet \mathrm{sgn}(D_k) - B2 \bullet \mathrm{sgn}(C_{k,old})$$

where A1 and A2 are adjustment coefficients, B1 and B2 are leakage coefficients.

Most characteristics parts among them are the correlation and the tap coefficients adjustment. The correlation between the input signal and the error signal detects ghosts and waveform distortion exactly. At first, proportional adjustment with leakage method is used in order to speed up tap coefficients convergence.

Then, incremental adjustment with leakage method is used in order to achieve stable and continuous operation. Thus, wherever ghost condition is slowly changed, the ghost is automatically reduced. Moreover, such abnormal missing states as oscillation, incorrect timing, and missing GCR are detected in order to ensure safely operation.

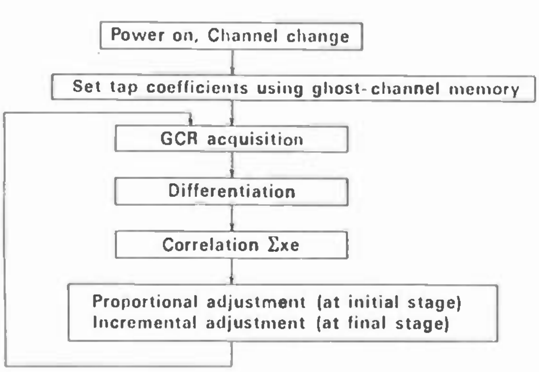


Fig. 18 Operation flow

Conclusion

We have developed a GCR signal method for a ghost canceler in order to reduce ghost interference in VHF and UHF broadcasting. With its special sin x/x characteristics, the GCR signal has enough frequency characteristics and, thanks to the 8 field sequence method, can eliminate long ghost with a delay time of up to 45 microseconds. It can be very useful both for the present NTSC system and the EDTV system. With several manufacturers having introduced their own versions of ghost cancelers on the market, price reduction of those device is expected.

This new GCR signal method are available in not only NTSC but also in PAL an SECAM systems.

Acknowledgement

To conclude this paper, we would like to express our deep appreciation to the GCR project members of BTA, the Ministry of Posts and Telecommunications, and other concerned offices for their cooperation.

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VIDEO AND AUDIO IN POST PRODUCTION SWITCHING SYSTEMS

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INTRODUCTION

This paper is a compilation of the works of three authors; Mr. Craig Birkmaier, Mr. Gary Carter and myself. In order to provide the background for the material presented in this paper, we thought it appropriate to review the historical evolution of today's video switchers as well as Edit Controllers and audio mixers in the following commentary.

BACKGROUND - VIDEO SWITCHERS

For the first few decades of television, video switchers were used in live production whether aired directly or recorded for later airing. Initially mixing only cameras, possibly with film chains in earlier switchers, the evolution of the VTR allowed pre-recorded segments to be used in programming. The architecture of these switchers was relatively simple by today's standards and based on the requirements of live programming.

By the mid-1970's, technology had advanced so rapidly that switchers were forced to change in order to meet the demands of the new products that were based on these technological advances. These newer picture sources included: chroma keyers, character generators, computerized editing controllers and later in the decade, digital picture effects sources. Most of these related to the keying aspects of the switchers. Editing controllers created the need for a computerized control interface and a micro-computer inside to control the switchers inner workings.

This in turn paved the way for effects memory registers built within the switcher control architecture. Still focused on live production applications and more specifically the needs of increasingly sophisticated news and event coverage markets, the keyers played an important role in advancing the switcher capability. Additional keyers in each M/E provided the ability to build a multi-effect without tying up other M/E subsystems. CDL first introduced this concept in their 480 series switcher. Typically the announcer was chroma keyed over background mixes or wipes and keys added over both announcer and background - essentially the DSK. additional M/Es provided the capability to build picture layers, and the more sophisticated switchers pioneered by GVG extended the multi-level capability by allowing priority swapping of two key levels in the M/E and the ability to re-enter effects in the upstream M/E as well as downstream.

The switcher architecture was initially based on a serial structure where mixes, wipes and keys were introduced in series with respect to the signal flow from the input buses to the program or preset outputs. There was an accumulating distortion level as a result of the expansion of switcher functions and capabilities as more serial functions were added. The solution, first introduced by GVG was the use of parallel multiplier stages which were controlled by more complex serial waveforms that literally put the picture together in pieces. By structuring the control signal, based on the picture piece priority, the operator was able to generate pictures as if the signal flow were serial, but most importantly without accumulating distortion.

In the decade of the 1980's, the sophistication and complexity of switchers continued to grow spurred on by the microprocessor. Our ability to control the switcher through its editor interface port allowed a high degree of system integration. The modern video switcher works hand in hand with a computerized editing system and other pictures sources to create complex images that sometimes involve over a hundred pieces carefully layered together to form the final image.

In the future we can definitely foresee further growth as a result of digital signal processing and the expanding role of the computer. Digital switchers are here already and most certainly will be the technology of choice in future production systems. The digital control we already see in post productions systems not only controls the switcher but also effects generators, character and graphics generators, VTR picture sources and the audio mixer.

THE EVOLUTION OF THE EDIT CONTROLLER

The history of video tape editing is difficult to accurately document. As soon as you cite one pioneer, another will be found who preceeded his work! For purposes of this paper, editing after time code editing practices were developed will have to be assumed, even though several significant techniques used today sprang forth much earlier than most of us realize.

Hardware

As video switching evolved through the 1970's, several applications necessitated the use of remote control switching functions from several locations within the plant. While consideration for these early serial/ parallel control ports was far from editing, its significance wasn't lost on the early pioneers in the field.

Once switcher manufacturers like ISI, GVG and Vital had provided the parallel and later serial "links," editor manufacturers like CMX, Datatron, and Convergence were more than able to adapt their machine control systems to be broadbased peripherals controllers.

Conceptually, the basic editor architecture has remained the same over a number of years. A central microprocessor performs the task of command and status reporting linked serially to numerous dedicated micros capable of communicating with a wide range of serial and parallel interface products.

In the early stages of time code edit control development, "cuts only" applications were considered more than enough to keep both customers and developers well occupied. For the engineers, edit control of parallel interface video tape recorders was a formidible challange due to the physical and electronic limitations of the VTR's in use then. Success in those days was more a factor of reliable deck interface than human control elegance.

Software

The limits of technology were being tested by hardware types, new concepts were evolving in software as well. Due to the extreme flexibility of the software based computer, it was easy to see early in the game that without some level of standardization of protocols and commands, the industry would quickly drown in it's own technological "soup." Thanks to a few farsighted product managers in resposible positions, this was not to be the case.

Spurred by demand, not only by operators, but also network plant managers, companies like Sony Corporation were offering their own serially controlled machines at the turn of the decade. Soon to follow in this manner was the Grass Valley Group who, having also seen the wisdom and economy of serial control interfacing, was well on the way to it's now well accepted series of serial command protocols.

Documentation

Aside from the physical task of machine communications, another major consideration in post production is the the human interface. Current editor design has evolved both up and down the scale, as software designers each have added personal touches to their creations. Fortunately for most of us, early in the documentation phase of editor development, certain specific edit protocol elements had undergone "de facto" standardization and the evident result has been the current CMX and Paltex "Tempo" EDL formats.

In both cases, these formats are capable of documenting most of the basic edit functions as they are performed, and storing this information to disk for future modification or reference.

Audio Interfacing

The latest entry into the automation process has been the audio mixer. While traditional audio processing has remained a manual process, the late 1980's has seen this concept undergo an extensive "rethink" in the video world. The early applications of ESAM I and the later ESAM II protocols and their rapid adoption as "sublists" within the EDL structures of current editors attest to the value of treating the audio mixer in the same manner as the video switcher.

A COMPUTER CONTROLLED AUDIO MIXING SYSTEM FOR VIDEO POST PRODUCTION

For years, limitations in the audio recording capabilities of videotape recorders, and the inability to broadcast television stereo sound provided an ample excuse for the television industry to ignore the importance of the audio portion of the programs most Americans have watched.

But now the excuses are gone. Multi-channel Television Sound is a reality, a standard feature on the majority of video receivers being sold in the United States. The number of homes served with stereo broadcast or stereo cable services is becoming too large a number to ignore. But it has only been in the past few years that manufacturers have paid attention to the requirements of automating the audio post production process. It is still quite common, and expensive, for the audio portion of the program to be "sweetened" in a separate post production facility optimized for audio production.

The requirement for a computer controlled audio mixing system that is fully integrated with the video post production process has been recognized for years. And now, it is becoming a reality in a new generation of microprocessor controlled, Audio For Video post production mixing system.

An Historical Perspective on Audio for Video Editing Systems

An historical perspective on Audio for Video editing systems. The focus of computerized video editing systems was centered on two main areas for more than a decade: 1) controlling machines - specifically VTR's but also some ATR's; 2) controlling a production switching system for video transitions and audio transitions. Why control the video switcher for audio transitions? For one thing, audio interfaces were expensive; for another, there were no audio consoles that could be externally controlled; and finally, the editor assumed that transitions occurred between buses for which it could select inputs - a concept for which there was no analogy in audio console design.

So the task of gaining control over audio transitions fell on the manufacturers of video production systems. The first systems provided two audio input buses with the ability to mix from one bus to the other. Typically the inputs selected on the buses followed the video source selections, and the transition followed the video transition, bringing about the classical term - Audio Follow Video. No control was provided over audio input levels, and the simultaneous mixing of multiple sources was not possible. The key feature that made these systems attractive, was the ability to accurately control the timing of the audio transition. Many end users brought sub-master outputs from their audio consoles in as inputs on the AFV buses to take advantage of the computer controlled transitions.

In looking at the market for a new Audio For Video mixing product, several aspects of the prior art stood out, and became an important consideration in the design and implementation of the AFV-500 Audio For Video mixer, developed for FOR-A Corporation.

1. The concept of Stereo Imaging
2. The concept of a Virtual Machine
3. The need for a serial control network to fully integrate the audio mixer with other products in the post production environment.

This paper will deal with the implementation of the AFV-500 design as it relates to these concepts.

The Concepts of Stereo Imaging and the Virtual Machine

One of the more difficult concepts that one encounters in audio for video production is the relationship of multiple audio tracks to the final stereo program. In the music industry, masters may have 24 tracks or more which eventually get mixed down to the final two channel "stereo" product. In audio for video production, the relationships can be equally complex.

The video editing process is focused on video sources as they relate to the controlled source machines; the editor is "configured" to select the appropriate crosspoint on the video mixer for each controlled machine, allowing automated cuts, wipes, and dissolves between video sources. For audio, this one-to-one relationship of machine and crosspoint does not exist. The VTR may have four tracks of audio recorded along with that single video track; in the field, natural sound may be recorded on one or more tracks, while additional tracks are dedicated to voice recording. These tracks must be assigned to the appropriate output channel(s) and mixed with other tracks, such as music and sound effects, to create the final stereo product.

The process of creating the appropriate audio mix to match the video portion of the program is referred to in this paper as "Stereo Imaging." The resulting mixer setup is referred to as a "Virtual Machine."

Let's start by examining the requirements for Stereo Imaging and the resulting design implementation in the AFV-500. In Figure 1, several of the important functional groups of the AFV-500 control panel have been highlighted. The Machine Bus relates to the concept of Virtual Machines which will be covered in a moment. The source assignment buses are where the process of properly imaging the sound begins, while the faders are used to set levels and create panning effects. The mixer has eight monaural inputs each of which is equipped with a seven-band equalizer (refer to figure 2). Each of these eight sources feeds a pair of Voltage Controlled Amplifiers, the outputs of which are summed on two independent busses, with the VCA's from the other seven sources. Thus a single input can be assigned, and its level controlled independently, on the Channel 1 (stereo left) and/or Channel 2 (stereo right) outputs of the mixer. Any combination of sources can be enabled sumultaneously, creating a complete audio mix setup or Virtual Machine.

This architecture provides a wide range of imaging possibilities for each source (refer to Figure 3). Basic imaging can be achieved by assigning the monaural source to Channel 1, Channel 2, or both. When assigned to Channel 1 only, the control panel fader sets the level for only the left channel VCA, while the right channel VCA is fully off. Channel 2 only, controls the right channel VCA while the left channel is off. Assigning the source to both channels causes both VCA's to be set at the same level as the panel fader.

At times it may be desirable to more precisely match the sound source to its location on the screen, or move the source of the sound to match a video effect. This is where the architecture of the AFV-500 provides a unique advantage, allowing the panning of each source under computer control. Panning is achieved by programming a software offset between the levels of the two source VCA's. This is accomplished by allowing the master transition fader to pan the sound from right to left while the source fader maintains the output level: to pan the source left, the right channel VCA is attenuated; attenuating the left channel VCA moves the sound right.

In Figure 4 a programmed panning effect is illustrated. In this scene, a car moves across the screen from left to right. To begin, the mixer is setup with the source panned fully left; the operator learns this setup as the first keyframe (or Virtual Machine setup). The pan control is then moved fully to the right and learned as the second keyframe. The tape is observed to determine the duration of the effect, which is then used as the transition rate between the two setups. Initiating the crossfade transition causes the sound to move from left to right in the number of video frames selected.

The previous discussions provide a significant insight to the concept of a Virtual Machine. As was mentioned previously, the editor must be configured to identify the crosspoints on a video mixer that correspond to each controlled machine. Since the corresponding audio mixer setup typically cannot be represented by a single crosspoint command, a method for keeping track of each element of the audio mix is required. The original ESAM I protocol allows the editing system to control the audio mixer by sending the appropriate crosspoint commands and fader levels for each source that is part of the beginning and ending mix; as well as crossfade transition commands. Unfortunately, this places a great deal of communications burden on the editing system. To help minimize this burden, the ESAM II protocol added the Virtual Machine layer to the protocol. With Virtual Machines implemented as a feature of the audio mixer, the editor is able to send From and To machine numbers, which are translated by the audio mixer into the required selection of crosspoints for the From and To machines.

The AFV-500 goes a step further by using the Virtual Machine concept as the backbone of its memory system. The Machine Bus shown in Figure 1, represents a subset of the available memory setup registers that can be programmed in the system; there are 100 setups organized into ten banks (i.e. 0-9, 10-19, 50-59, etc); only one decade is shown at a time, but the system allows random access between any two setups. In each bank machine 0 is permanently learned with all VCA's off to provide an easily accessible setup for fades to silence. The other nine machines are programmable mixer setups. Thus to configure the audio mixer for an editing session, the operator creates the mix setup that is appropriate for each video source and stores it as a virtual machine, typically using machine numbers that correspond with the crosspoint assignment for each VTR on the video mixer. Several variations of a machine can be stored; for example, different Equalizer settings for each voice on a tape can be learned. When a machine setup is recalled, the entire mixer setup is restored, including: assignment of sources to output buses; fader levels; pan settings; EQ settings; transition rates; and Audio Matrix input assignments.

Crossfade transitions are actually smooth computer controlled analog transitions between the VCA levels stored for two Virtual Machine Setups. If a source is not included in the From or To setup, its VCA values are assumed to be 0. By learning each mixer setup as a Virtual Machine, the editing process can be significantly enhanced, providing major improvements in throughput, enhanced audio effect capabilities, and an archival record that can be used for re-editing the material in the future.

Implementation of the Serial Control Network

To achieve the design goals of complete integration with FOR.A video mixer products and stand alone use in systems using editing systems and video mixers from other manufacturers, an intelligent communications network was developed. The Central Processing Unit of the AFV-500 is a Motorola MC68000 microprocessor with logic level and analog outputs to the audio processing hardware, and four serial ports: Control Panel; Editor Interface; Audio Matrix Interface (AMI); and an External CPU (refer to Figure 5).

The Control Panel is an intelligent keyboard that accepts input from the operator which is converted to serial data. This information is sent to the processing frame at video field rate, over an RS-422 serial link operating at 38.4 Kbaud. The frame processes this data, along with input from the editing system to create the machine state, which is output to hardware at field rate, synchronized to the video system. The frame prepares a display message that is sent to the panel at the same time that the panel is communicating new change information to the frame; the panel displays are updated once a field to reflect operator changes and changes requested by the editing system. The system is fully interactive, allowing the operator to override editor commands, or manually control portions of the mix.

The Editor Interface supports the RS-422 serial interface hardware and addressing standards jointly developed by SMPTE and the EBU, known as ESbus. The nominal communications rate is 38.4 Kbaud, however, the communications can be slowed down to interface with editors and other CPU's that communicates at 9600 baud or 19.2 Kbaud. At the current time, seven different communications protocols are supported to handle a variety of interface situations. The FOR-A Audio Follow Video interface is a dedicated protocol developed to receive serial data output by FOR-A PVM/CVM 500/600 video mixers. This data stream contains video crosspoint selections; auto transition rates and commands to initiate transitions; and manual lever arm transition data.

The data stream is created either through manual operation of the video mixer, or when an editor commands the video mixer to perform a video transition. The AFV-500 interprets this data and translates it into Virtual Machine selections and transitions, allowing the audio mixer to follow the operation of the video mixer.

The remaining protocols are used when the editing system offers a direct serial link to the audio mixer, and with video mixers from other manufacturers. The direct audio mixer interface is the most powerful and desirable, and is now available with many editing system. The AFV-500 supports both the ESAM I and ESAM II protocols as published by Graham-Patten Systems. When interfaced to an editor using the ESAM I protocol, the AFV-500 interprets crosspoint commands as Virtual Machines. When the editor uses the ESAM II protocol, both crosspoint and Virtual Machine commands are supported.

in some cases, the editing system may not offer a dedicated serial link to the audio mixer, or may use parallel control techniques. If the editor is interfaced with a video mixer using one of several popular serial protocols for video mixers, the AFV-500 can derive the required control information for Audio Follow Video operation from the communications between the editor and video mixer. With this type of interface, the transmit lines from the editor to the video mixer are paralleled to the editor port of the audio mixer which listens for video crosspoint and transition commands. The crosspoint commands are interpreted as Virtual Machines and transition commands initiate crossfade transitions that follow the video mixer. This type of interface is available for video mixers that use protocols developed by the Grass Valley Group for the 100, 200, 300, and 1680 series video mixers.

The Audio Matrix Interface is another RS-422 port capable of communications rate of up to 38.4 Kbaud. The AMI system consists of a special control panel for the AFV-500 that provides the necessary buttons and displays to select crosspoints from an external routing switcher matrix. Typical implementation requires the use of ten output buses of the audio routing switcher to feed the eight primary inputs and two Record Return monitoring inputs of the AFV-500. The control panel allows the operator to pre-select crosspoints (in a range from 0 - 99) for each of the primary inputs and the Record Return monitor. Assignment of a crosspoint causes a command to be sent from the AFV-500 to the routing switcher to select the crosspoint on the appropriate output bus. When a Machine setup is learned, the external crosspoint assignments are learned for each active source. When a machine is selected on the Machine Bus as the Next Setup, commands are sent to the routing switcher matrix to assign all required crosspoints for that setup. The current implementation controls a 30 X 10 monaural audio matrix product from Dynair, named Dynamite. The structure of the system will allow protocols to be developed for any serial controlled routing switcher matrix.

Keeping archival records of the audio mixer setups used during a edit session is an important requirement for any post production system. Many editing systems offer the feature of storing the contents of the memory registers of video and audio mixers as files on the Edit Decision List disk. In some cases this data may be stored as part of each edit event, while in other systems the files are maintained separately and downloaded to each device prior to reconstructing an edit.

CONCLUSION

This paper has provided a background for video switchers and edit controllers and has demonstrated a hardware and software architecture for a microprocessor controlled audio mixing system for use in video post production. It is hoped that through the use of this product, and by the adoption of these techniques in other products in the future, that the task of producing stereo audio in the video post production editing environment with be significantly enhanced.

AFV-500 CONTROL PANEL FUNCTIONAL GROUPS

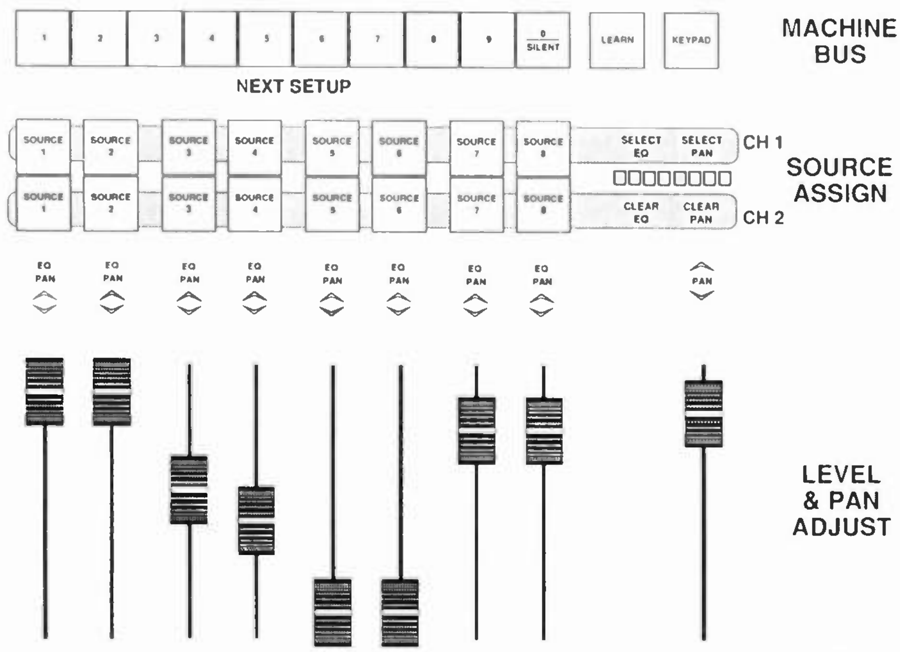


FIGURE 1

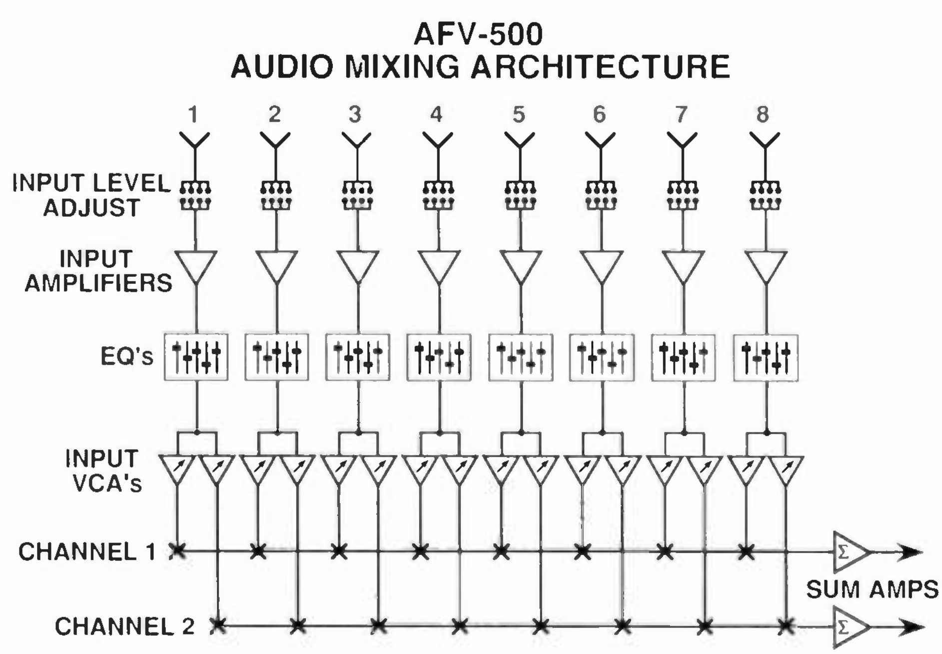


FIGURE 2

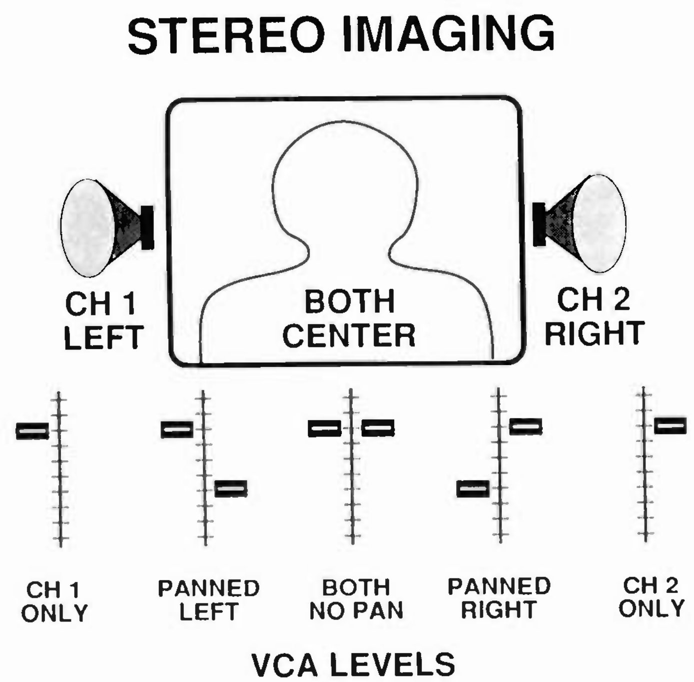


FIGURE 3

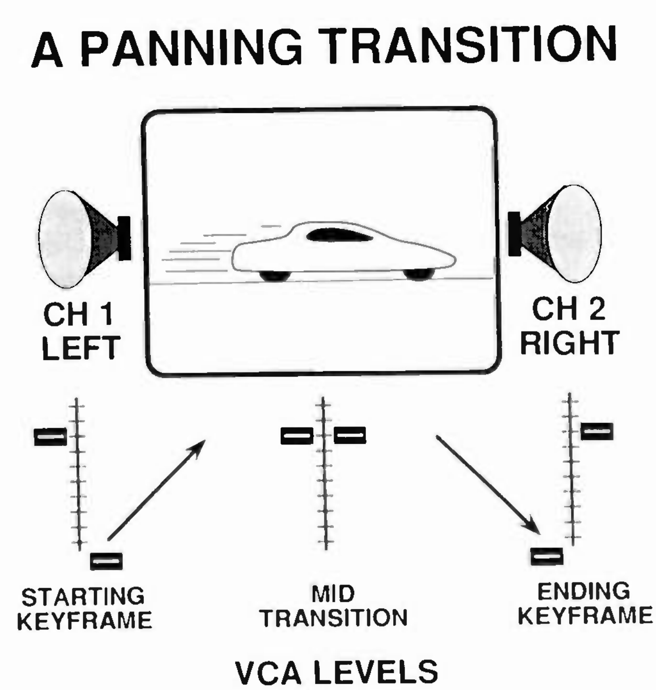


FIGURE 4

SERIAL COMMUNICATIONS NETWORK

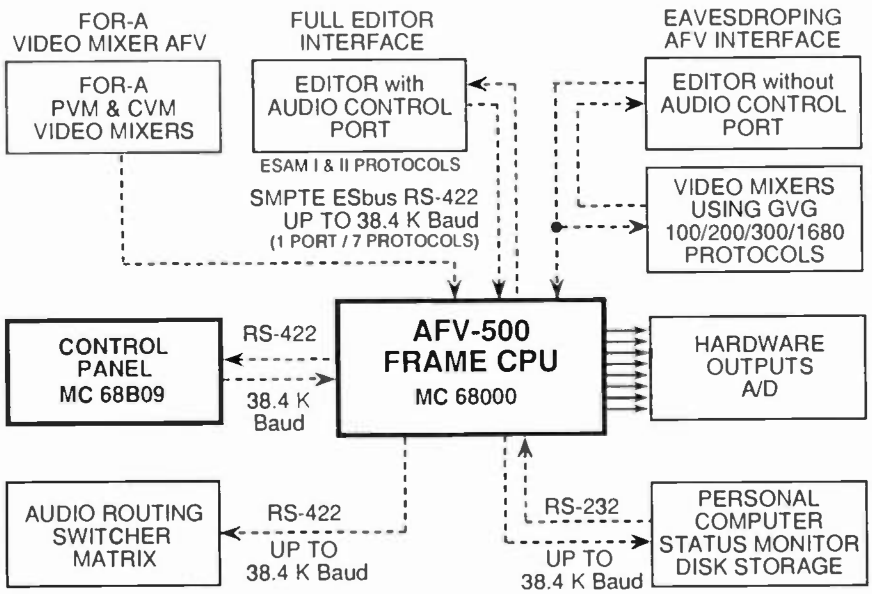
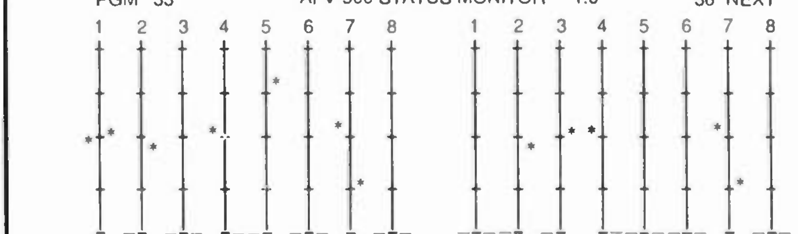


FIGURE 5

AFV-500 STATUS MONITOR

PGM 33 AFV-500 STATUS MONITOR 1.0 36 NEXT



	1	2	3	4	5	6	7	8	1	2	3	4	5	6	7	8
CH 1	48	0	0	56	0	0	58	0	0	0	0	56	0	0	58	0
CH 2	55	40	0	0	75	0	23	0	0	40	53	0	0	0	23	0
160	+3	+5	0	+2	-2	0	+1	0	0	+5	-12	+2	0	0	+1	0
400	+2	+1	0	+1	-1	0	+1	0	0	+1	-3	+1	0	0	+1	0
1k	0	-1	0	-1	+1	0	0	0	0	-1	0	-1	0	0	0	0
2.5k	-3	+5	0	+7	+2	0	-1	0	0	+5	-1	+7	0	0	-1	0
6.3k	0	+3	0	+2	+3	0	+3	0	0	+3	-3	+2	0	0	+3	0
TRIM	-2	-3+	0	-2	-1-	0	0	0	0	-3+	+3	-2	0	0	0	0
AMI	5	3	13	14	17	18	6	7	5	3	13	14	17	18	1	2

12-> 30 MON - 60
1->120 <--> RR1 - 19
2-> 60 <--> RR2 - 20
FILE: FORASMON.SET SOURCE(PC) [001-010] - - - - -> DEST. [010-020]
AFV-500> Enter DESTINATION lower limit: _
F1:LOAD F2:CLEAR F3:SOURCE F4:DEST. F5:DOWNLOAD F6:UPLOAD/SAVE F7:EXIT

FIGURE 6

BROADCAST AND VIDEO PRODUCTION APPLICATIONS OF THE STEREOSURROUND® AUDIO PRODUCTION PROCESS

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Evanston, Illinois

The Stereosurround audio production format is a practical, compatible 4–2–4 matrix encode/decode process that allows the generation of more realistic sound fields using four discrete-like reproduction channels that may be stored or transmitted using two conventional audio channels. A brief technical description of the encoding and decoding process will be presented followed by an analysis of a variety of field and studio production techniques. Specific program examples discussed will include sports, music entertainment, made-for-television dramas, and commercial advertising productions. In addition, the technical robustness of the process will be analyzed with respect to its tolerance to amplitude and interchannel delay errors.

0 INTRODUCTION

The Stereosurround production format has as its roots, techniques developed by the motion picture industry that were designed to add more realism and excitement to the audio-visual entertainment experience. Nearly 50 years ago, the motion picture "Fantasia" was produced using a multichannel sound process (Fantasound) that involved not only three reproduction channels behind the screen but loudspeakers located to the sides and overhead [1]. Since that time, numerous multichannel motion picture formats have been developed with a common objective of creating a more realistic sound field for the listener/viewer. By using at least three front channels, and one rear or side channel (L = Left, C = Center, R = Right, S = Surround) as shown in Figure 1, significant benefits can be gained over conventional two-speaker reproduction.

With the addition of the surround and center channels, stable center images can be created for a large listening space as well as interior or all around ambience images. In addition, sounds can be created that move not only across the front sound stage but between front and back. These additional channels of information contribute significantly to establishing a greater sense of spatial reality resulting in an increased sense of being in a real-world sound field. In spite of the impressive performance of many multichannel systems, practical considerations such as exhibition costs and multiple print inventories have prevented the wide acceptance of any discrete format. What has evolved, however, are two four-channel matrix oriented discrete sounding formats commonly referred to as Dolby Stereo® and Ultra Stereo® based upon the patents of Scheiber [2], [3], [4]. In addition to providing a discrete sounding four-channel production format, these processes have been accepted as both two-speaker stereo and single-speaker mono compatible. Unlike the matrix techniques developed as a part of the consumer quadraphonic products of the early 1970's, these processes use high performance steering logic techniques that allow the creation of unambiguous images over a wide listening/viewing area [5], [6]. These production techniques are in widespread use today in the feature motion picture industry and are commonly used in major film releases.

The matrix techniques associated with the motion picture industry had limited additional applications until the development of new audio and audio/video technologies associated with broadcasting and home entertainment systems. Specifically, the BTSC–MTS Stereo Television System, stereo satellite television services, the VCR with FM (Hi-Fi) audio tracks, the optical video disc with PCM digital audio tracks, and the compact disc all represented new media with robust characteristics capable of accurately processing matrix encoded audio programs. Initially, programming consisted of matrix encoded feature films that were broadcast or transferred to video cassette and video disc for reproduction by consumers. As a result of these productions, a home theater oriented professional and consumer awareness has developed, along with a market for decoding and related audio equipment to accurately reproduce these programs [7]. A particular driving force behind the acceptance and interest in this multichannel audio process has been the fact that a picture is involved and, hence, the listener/viewer is more aware of the spatial realism that can be created. This should be contrasted to the situation that existed with the earlier consumer quadraphonic techniques. These techniques were developed around music-only programs and had a restricted ideal listening space due to the lack of a center channel.

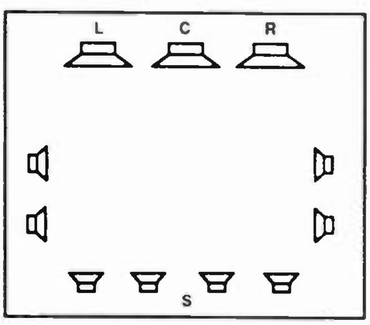

Fig. 1. Loudspeaker locations for a multichannel reproduction system with improved spatial reality

Because of a growing consumer interest in stereo audio with video as well as an increasing number of home theater type playback systems, there is a desire among program producers to create a greater variety of productions with stereo audio in addition to major motion pictures. Specific examples include sports events, musical variety programs, made-for-television dramas, music videos, commercials, and audio-only programs. The Stereosurround production format has been developed for and is currently being used for such productions. It is designed to give maximum performance benefit to a growing number of consumers with sophisticated decoding equipment, while at the same time, providing a compatible production for two-speaker stereo and single-speaker monaural listening. In addition, experience has demonstrated that the Stereosurround production format makes it possible for audio engineers to mix more efficiently and accurately due to the inherent ability of the process to reveal common production errors.

1 THE STEREOSURROUND PRODUCTION FORMAT

1.1 General

The Stereosurround production format uses a 4-2-4 matrix process oriented around a four input, two output encoder, and a two input, four output decoder as shown in Figure 2. In program production, the output of the encoder is recorded or sent to the transmission chain as well as fed to a decoder to monitor the production. Encoded programs are optimized for playback through a decoder with similar characteristics.

The characteristics of the encoder and decoder are based on the matrix portions of the Dolby Stereo and Ultra Stereo production process, but have been refined for modern broadcast and consumer audio storage and transmission systems.

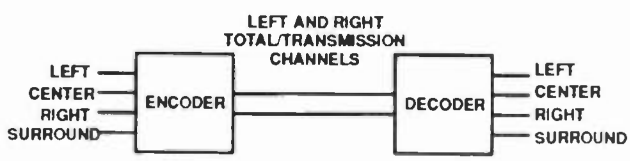

Fig. 2. Stereosurround Encoder/Decoder Block Diagram

The Stereosurround production format has, however, been designed to allow the creation of programs that are highly compatible with existing Shure HTS Acra-Vector logic, Dolby Surround Pro-Logic, and Ultra Stereo® decoding equipment as well as being downward compatible with Dolby Surround® and other decoding systems designed for three front channels and one rear channel. In addition, such productions are compatible with two-speaker stereo and single-speaker monaural reproduction systems.

1.2 Encoding and Decoding—Theory and Performance Capabilities

The matrix portion of the Stereosurround production format is a nonsymmetrical encode/decode process in the sense that the decoding portion involves adaptive matrix modification (steering) logic circuits to make the system perform as if four independent discrete channels are being used. Figure 3 indicates the basic elements of the encoding process in block diagram form.

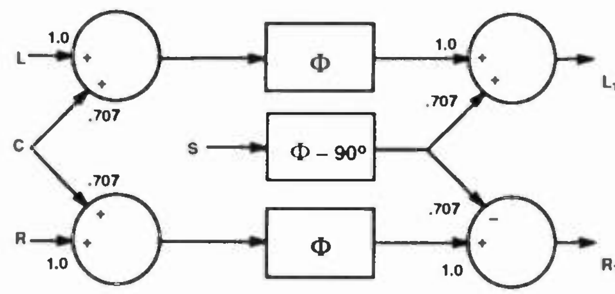

Fig. 3. Simplified Stereosurround Encoder Block Diagram

Mathematically, this process may be described as follows:

$$L_T = L + .707\ (C + jS)$$

$$R_T = R + .707\ (C - jS) \qquad (1)$$

where L_T and R_T represent the left and right total or transmission encoded signals, and L, C, R, and S represent the left, center, right, and surround input signals, respectively. It is important to note that the elements in the block diagram denoted by Φ and Φ - 90° are all-pass networks characterized as having a flat amplitude response, but a frequency dependent phase response of $\Phi(f)$ and $\Phi(f) - 90°$. Analysis of the complete set of encode/decode equations reveals that such phase shift functions are required in order to allow for an interior or all around sound mixing capability. The decoding matrix (before directional enhancement or matrix modification) is defined by the four equations:

$$L' = L_T$$

$$R' = R_T \qquad (2)$$

$$C' = .707\ (L_T + R_T)$$

$$S' = .707\ (L_T - R_T)$$

where the prime notation is used to denote that the decoded signals are only approximations to the original L, C, R, and S inputs. The need for matrix enhancement or modification becomes clear when the decoded signals (2) are represented in terms of the original input signals. This can be seen from equations (3), where equations (1) and (2) have been combined to produce:

$$L' = L + .707\ (C + jS)$$

$$R' = R + .707\ (C - jS) \qquad (3)$$

$$C' = C + .707\ (L + R)$$

$$S' = jS + .707\ (L - R)$$

These equations clearly show that each decoded signal contains the original input signal, as well as its spatially adjacent channel's signal reduced in level by 3 dB. This can also be shown graphically as in Figure 4, where the 3 dB arrows indicate the areas of significant crosstalk, and the infinity arrows indicate channels producing no crosstalk. For example, if L was the only input to the encoder, a portion of L (attenuated by 3 dB) would appear in both the center and surround channels, and none of L would appear in the right channel.

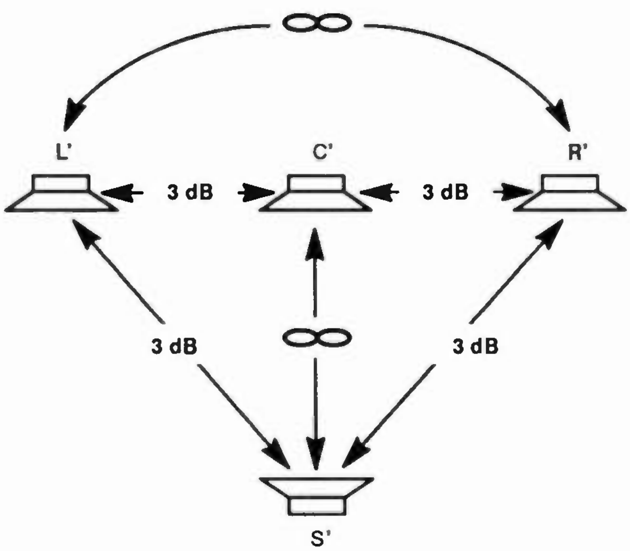

Fig. 4. Graphical representation of 4-2-4 matrix decoding crosstalk prior to directional enhancement

From a perception standpoint, the situation depicted in Figure 4 is a step in the right direction in that all reproduced sounds are most dominant in the intended directions. Subjectively, however, this level of decoding accuracy cannot produce unambiguous localization perception over a wide listening area, particularly with respect to centrally placed images. In order to significantly improve this situation, matrix enhancement circuitry is employed in the decoder. Functionally, this circuitry serves to reduce the undesired crosstalk signal components (3 dB crosstalk elements) as a function of the dominant content of the program. Since it is not possible to achieve full independence of the four channels at all times under all conditions, it is necessary to make system compromises. In the case of the Stereosurround format, compromises have been made with artistic considerations in mind. For example, the integrity of the center channel is most important, and a number of steps are taken to keep center channel information out of the surround channel. On the other hand, the ability of the system to simultaneously enhance a center and right or left discrete image is found not to be as important and, consequently, is not possible without some crosstalk. The decoding action can, however, very quickly alternate between these image locations with a low level of crosstalk giving the perception of very discrete behavior. As a result of these necessary compromises, it is common practice in Stereosurround format productions, to monitor through the encode/decode process so that any localization difficulties resulting from a particular mix can be heard and dealt with. Based on these artistic considerations, the Stereosurround production format can be viewed as a technological art form with the following capabilities:

1. Accurate localization of dialog, music and sound effects across a three-loudspeaker sound stage for listeners seated in a wide listening area.

2. Unambiguous rearward localization of specific dialog, music, and sound effects.

3. The creation of interior sounds which surround the listener with environmental or ambience effects from all directions.

4. Smooth panning capabilities across the front sound stage and between rear, interior, and front locations.

5. Simultaneous reproduction of localized or panned sounds and interior sounds.

From a compatibility standpoint, it can be seen from Equation (1) that a two-speaker reproduction system will accurately decode left and right information with the center channel appearing in both outputs at a reduced level creating a phantom center image. The surround information will also appear in both channels at a reduced level, but with opposite polarity which typically results in a diffuse image. A monophonic system will reproduce the sum of L_T and R_T which will contain all signals with the exception of the surround channel input. Although this may seem to be an unacceptable situation, it is not the case in actual practice. Since hard pans to the surround position are generally not required for artistic reasons, but rather pans to and from the surround position, a great deal of compatibility exists. For such pans, the monophonic listener simply hears the production element get loud then soft, or soft then loud. This generally correlates very well with the screen action in audio/video productions. In practice, a compatible surround image can be created by panning near, but not all the way to the surround position, in which case a portion of the surround element of the mix will appear in the sum or monaural signal. As previously indicated in Equation (3), decoding circuits that do not employ enhancement logic will decode the production, but with reduced separation (location accuracy) between channels.

1.3 Encoding and Decoding—Hardware Considerations

A practical realization of the Stereosurround format encoder described in Figure 3 is shown in block diagram form in Figure 5. Two groups of fixed-position inputs in addition to a panable input have been found to be useful for a variety of productions due to the limitations of many mixing situations. In addition, the panable input allows the creation of a custom mix bus position that might otherwise be difficult to create with available mixing controls. Precision all-pass encoding networks are required in order to guarantee acceptable (greater than 40 dB) encoded channel separation. Of particular importance are center to surround, and left to right crosstalk performance. An 80 Hz low frequency shuffling circuit has been included, on a switchable basis, for those applications where low-frequency out-of-polarity or difference information must be avoided. Functionally, this circuit attenuates $L_T - R_T$ information below 80 Hz without affecting the $L_T + R_T$ component. As an aid to system setup, a metering circuit and input trim controls are provided to allow the calibration of all inputs.

All inputs on the encoder are full bandwidth including the surround channel. In practice, however, the decoded surround channel is often limited to approximately 7 kHz by the characteristics of typical surround channel consumer circuitry. This is due in part to the bandwidth of cost effective time delay circuits employed in the surround channel signal path whose primary function is to put the listener in the acoustic center of the listening space, and to increase perceived center-to-surround channel separation.

In some applications, it may be desirable to use a companded noise reduction system within the encoding and decoding process other than with the L_T and R_T signals. Such noise reduction may be desired due to a particular limitation or characteristic of the storage or transmission medium. This is commonly done in the Dolby Stereo and Dolby Surround processes with the surround channel, due to limitations in the optical film recording process. In this case, a modified B-type noise reduc-

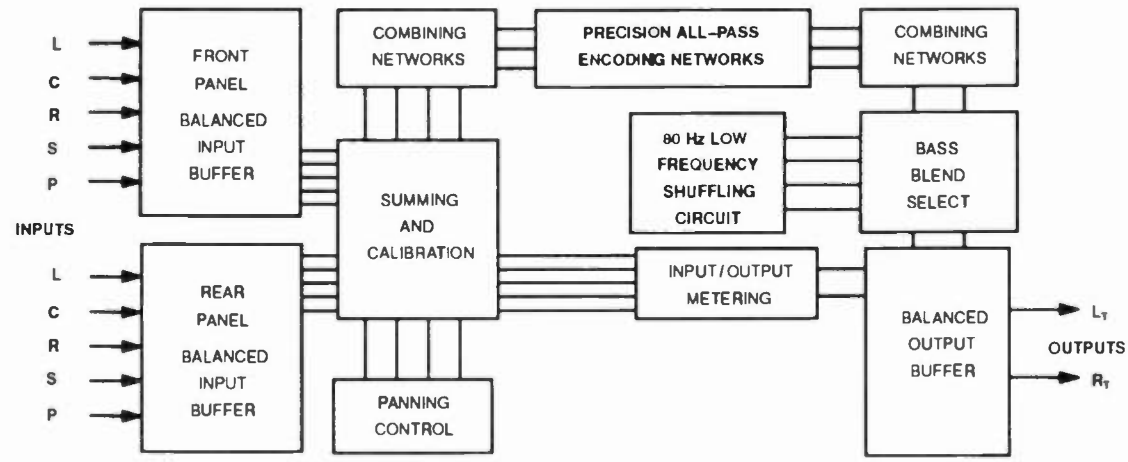

Fig. 5. Stereosurround Encoder System Signal Processing

tion encoder is placed in the surround input of the encoder and a complementary decoder in the surround output of the decoder. If a discrete multichannel system was being used, this would indeed be a logical application of the noise reduction process. Due to the fact, however, that a matrix system is being used, proper companding will not occur because the output of the noise reduction encoder is not what is fed to the decoder. A more technically accurate solution would be to use noise reduction on the difference signal after the encoding process. This could be accomplished by deriving both $L_T + R_T$ and $L_T - R_T$ from the encoder outputs, sending $L_T - R_T$ through the desired noise reduction encoder and then generating a new L_T and R_T for transmission or storage. Similarly, prior to decoding, $L_T + R_T$ and $L_T - R_T$ would be generated and $L_T - R_T$ processed through the complementary noise reduction decoder and finally, L_T and R_T would be regenerated for multichannel decoding. The MTS stereo television system uses a variation of this technique, where sum and difference signals are transmitted instead of left and right, and the difference signal is companded to optimize performance given the restricted dynamic range of its transmission channel.

A system block diagram of a Stereosurround format decoder is shown in Figure 6. The initial sum and difference matrix decoding generates the L', C', R', and S' signals as described in equations (3) prior to processing by the "Directional Enhancement Signal Modification" circuitry. It should be noted that the enhancement process involves a decision-making side chain that does not process any of the actual output audio signals, but rather controls linear sum and difference circuits in the "Directional Enhancement Signal Modification" block where the L", C", R", and S" signals are generated. This approach to decoding serves to significantly reduce the amount of circuitry by which the actual signal is processed. An additional element of the process is the derivation of a low frequency (below 80 Hz) or subwoofer output which is not processed by the directional enhancement circuitry. Since the position of low frequency information below 80 Hz is difficult to localize in rooms, this output serves in effect as a nondirectional fifth channel suitable for creating dramatic impact in productions. A further benefit of this channel is that it allows the use of relatively small loudspeaker enclosures for the L, C, R, and S reproduction locations.

Due to the fact that listeners seldom sit in the acoustic center of the loudspeaker array shown previously in Figure 1, and are typically biased towards the surround loudspeakers, a digital time delay is used to, in effect, move the surround loudspeakers farther back into the listening space. An additional benefit

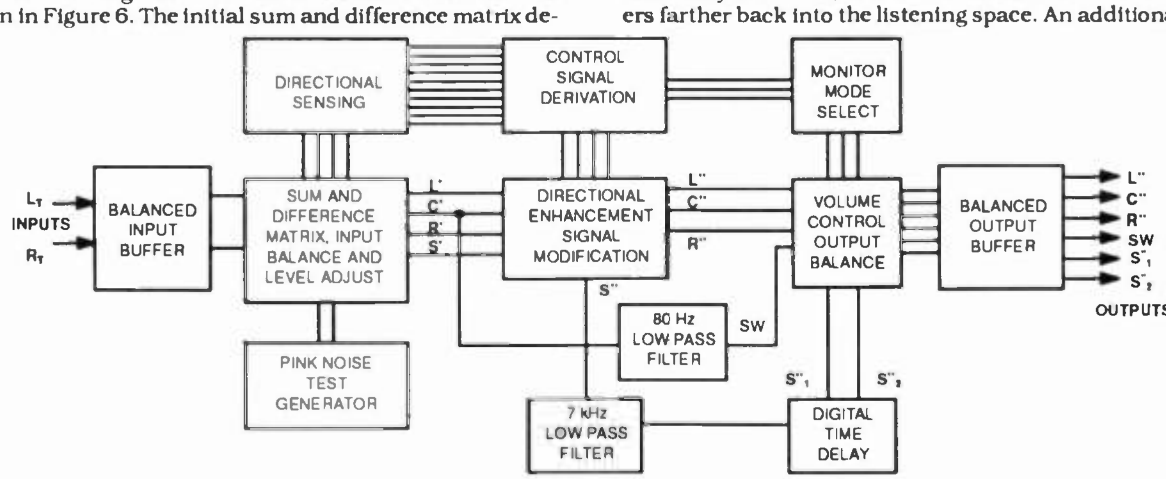

Fig. 6. Stereosurround Decoder System Signal Processing

of this form of signal processing is to further reduce the perception of any crosstalk from the center channel into the surround channel by virtue of the precedence effect. In addition, since some listening environments are long and narrow, two separate surround outputs can be made available with different selectable time delays to more accurately accommodate a larger number of listeners. Finally, a monitor mode selector is provided to allow productions to be monitored using Stereosurround format decoding with and without directional enhancement, as well as using two-speaker stereo and single-speaker monaural reproduction systems. Figure 7 is included to show the design characteristics of an actual encoder (top) and decoder (bottom).

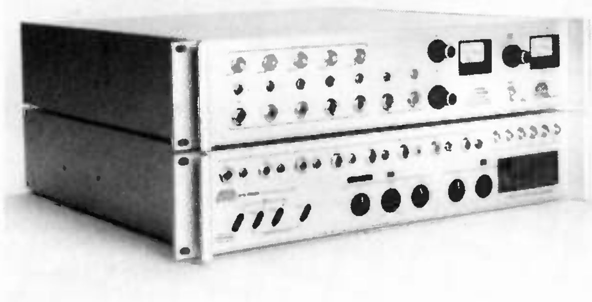

Fig. 7. Stereosurround Encoder (top) and Decoder (bottom)

1.4 Production Considerations

The Stereosurround production format is ideally suited for a post-production environment because it allows the mixing engineer to refine the mix so as to obtain maximum benefit from the process. Excellent performance is also obtainable from a live or live-to-tape production environment provided the audio elements are of a predictable nature, and known mixing compatibility problems are avoided.

A basic element of any production is the establishment of a monitoring environment that will allow the mixing engineer to properly audition the Stereosurround mix. Such a setup is shown in Figure 8. Ideally, all three front loudspeakers should be the same, or at least have very similar tonal characteristics, and should be located at the same listening height. The center loudspeaker should be as close as possible to the video monitor (if used) and located just above or below the screen. Because of the center channel, the left and right loudspeakers can be more widely spaced than for conventional two-speaker stereo productions. As a consequence, it may be desirable to audition the two-speaker stereo compatibility of a Stereosurround production with the left and right loudspeakers moved inward.

Even though a single surround channel is produced in this format, it is highly desirable to reproduce it with a group of loudspeakers so as to create a diffuse sound field. Four loudspeakers are shown in the figure; however, two are frequently used in small production environments. The speakers should be positioned so that the mixer is free to move around with no significant change in the perceived level of the surround chan-

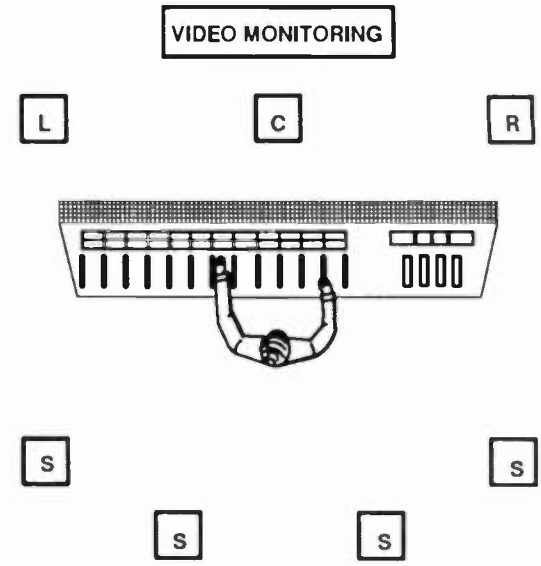

Fig. 8. Stereosurround Loudspeaker Monitoring Configuration

nel. Best results are generally obtained with loudspeakers located high off the floor (5 to 8 feet) so as to reduce the effect of other listeners disrupting the sound field as they move near an individual loudspeaker. A practical check on the surround reproduction channel is to move around the desired listening space, checking for level uniformity while reproducing a pink noise source.

Since productions using the Stereosurround format are capable of generating sound fields that vary between being excessively reverberant to nearly anechoic, an ideal listening environment should be as acoustically absorptive as practical. As an example, excellent results have been obtained in control room environments with midband reverberation times varying between .25 and .3 seconds. Finally, when completing any mix or starting any live production, the monitor system should be objectively or subjectively balanced, using a test generator that alternately cycles a band limited random noise source between the L, C, R, and S decoder outputs.

Figures 9 and 10 indicate typical equipment block diagrams for post-production and live or live-to-tape production environments. A common element of both situations is that of monitoring the encoded signal, as it goes to tape or transmission, through a decoder at the production site. The primary difference between these two production environments is the source of input program material and the need for time code and synchronization between audio and video recording equipment.

Due to the fact that the encoding process is linear, more than one encoder can be used during a production provided that the L_T and R_T outputs of each decoder are summed as a part of the decoding and monitoring process. This capability is of particular value in made-for-television or video productions where there may be a need to alter the production mix at a later time, particularly with respect to additional language releases. In such cases, three encoders are used, one each for dialog, music, and effects (DME), and their outputs recorded on a six-track format. Similarly, in a multi-site live production environment, separate encoders may be used for each location and combined at the production mixer location. In certain live production environments, it may be necessary to delay the L_T/R_T audio mix to properly match the video program. A typi-

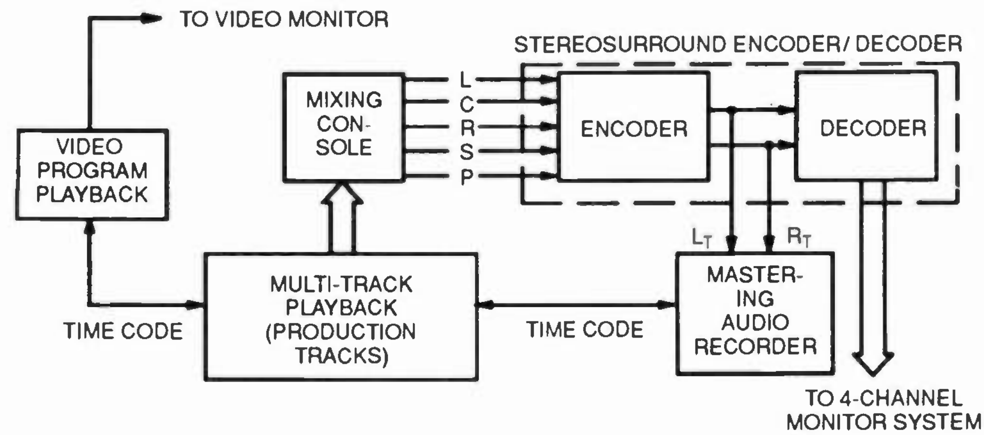

Fig. 9. Stereosurround Post-Production Equipment Block Diagram

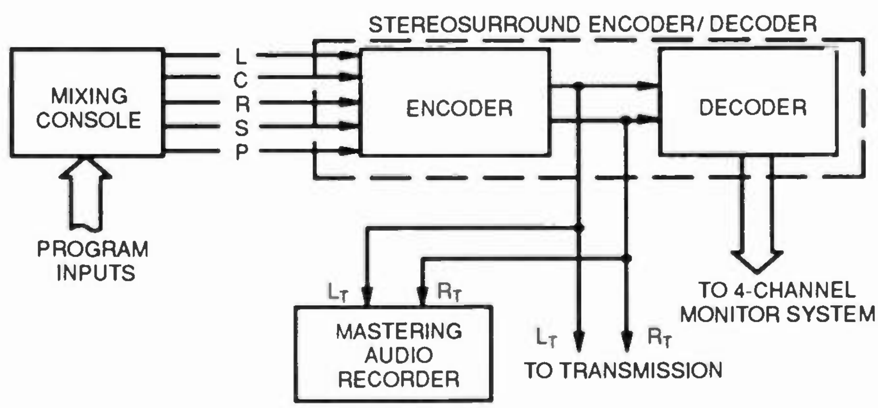

Fig. 10. Stereosurround Live or Live-to-Tape Production Block Diagram

cal situation might involve a pre-recorded audio mix that performers are reacting to as they listen to its playback over a sound reinforcement system. If the acoustic delay due to the distance between the sound system and performers exceeds approximately 40 milliseconds, the audio will appear to lead the video creating an unrealistic result. In order to correct this condition, it is necessary to delay L_T and R_T by precisely the same amount so as to maintain the integrity of the mix. A typical solution is to use a two-channel digital time delay in which a common clock is used to control both channels. Figure 11 is included to show a typical live field-production environment commonly found in mobile audio/video production facilities.

1.5 Mixing Techniques

Developing a mix using the Stereosurround production format is very similar to two-speaker stereo and single-speaker monaural mixing, once microphones have been selected and placed, signal processing has been set, and the configuration of the mixing board and source assignments have been made. In a typical set up, the input buses and outputs of the encoder are connected to the mixer patch bay, such that each input fader can be assigned to or panned between any two input buses. With this condition established, the signal controlled by each fader can then be panned in the "mixing space" to the desired position as the mix is being monitored. Based upon Stereosurround format productions to date, a number of mixing techniques and observations have emerged:

1. Due to the fact that 4-2-4 matrix systems are not discrete, and that localization problems can occur, start by mixing the most dominant sound elements first.

2. When using compression or limiting on dialogue or a vocalist, avoid processing any additional ambient signals through the same compressor/limiter, so as to avoid the dominant sound causing the ambient sounds to pump up and down. This result is particularly bothersome with a multichannel reproduction system.

3. Keep in mind that it is possible to continuously pan signals between the front and surround locations, and that it is not necessary to pan a program element completely to the surround position in order to get a noticeable rearward or surround sense to the mix.

4. When mixing music, start by mixing sound elements directly to the L, C, and R positions and effects to the interior position using separate effects devices for each element. Once the mix has been roughed in, further adjustments should then be considered.

5. If problems occur regarding localization interactions between two or more sound elements, consider panning one

of the signals more toward the interior position, altering its timing, or reducing its amplitude.

6. When using multiple microphones in situations where two or more microphones may pick up the same signal, confirm the polarity of each microphone between its acoustic input and the mixing console output to the encoder. This will greatly reduce the possibility of undesired out-of-polarity information being mixed into the interior or surround position.

7. In many situations where a center oriented vocal element is competing with loud ambience element (sports announcer versus crowd), it is desirable to "widen" the vocal element. This may be accomplished by either panning the element towards the surround position or by using a small amount of stereo synthesis returned to the left and right front positions. Similarly monaural effects associated with sports productions such as baseball "bat cracks" and football player contact sounds can be given more dominance in the final mix.

8. MS and XY stereo microphone techniques can be integrated into Stereosurround format productions provided proper selection of microphone patterns are made. As an example, a popular MS microphone configuration using a cardioid mid pattern and bidirectional side pattern allows the creation of equivalent XY combinations that place sounds properly in the front and rear sound stage. For an in depth discussion of this compatibility, see the reference by Ross [8].

9. Realistic crowd or ambience mixes can be created by using multiple groups of microphones placed so as to sample distinctly different portions of the production environment. Once placed, the outputs of these microphones are then positioned at different points in the "mixing" space by panning between the left or right and surround encoder input buses.

Fig. 11. Stereosurround Live Sports Production Environment

1.6 Technical Requirements for Recording and Transmission

Stereosurround format productions depend upon the relative amplitude and phase integrity of two channel transmission and storage systems in order to maintain a high level of performance. Variations introduced in either of these parameters during recording, playback, or transmission in essence "re-encode" the production. Fortunately, audio recording and transmission technology has been moving in the direction of improved performance in this regard. Consumer audio formats such as the CD and DAT recorder are extremely robust. Similarly, audio/video formats such as VHS-HiFi, S-VHS-HiFi, Beta-HiFi, and Laser Video Disc technology offer extremely stable relative amplitude and phase performance. In addition, professional digital audio recording systems and cassette based audio/video formats such as D1, D2, Betacam SP, and MII provide a high level of performance. From a broadcast perspective, FM Stereo, and the BSTC-MTS stereo television formats are quite good due, in part, to the need for monaural compatibility.

Those systems offering lesser performance typically involve analog recording techniques where recording and playback azimuth calibration is required to provide good program interchangeability. Errors typically introduced by such devices generally involve phase or interchannel delay with relatively minor amplitude discrepancies. Based upon our laboratory studies, interchannel delay errors of 60 microseconds or less are acceptable. Once errors begin to exceed this value, high frequency components of speech sounds, which are typically encoded for center front reproduction, become surround channel information, and are correctly decoded as such to the detriment of the listening experience.

As far as amplitude errors between channels are concerned, transmission requirements are really no different than those associated with two speaker stereo production techniques. Practically speaking, however, a Stereosurround format monitoring system makes it much easier for an audio mixer to detect a channel imbalance due to the ease by which localization errors can be detected. In addition, crosstalk or channel separation requirements are similar to those associated with conventional two speaker stereo. At least 15 dB channel separation is required for good subjective performance, which is generally not a problem. As channel separation is degraded, the program simply becomes "more monaural."

Looking at the transmission and reception aspects of the MTS System has shown that certain factors can impair the proper transmission and reception of encoded programs. Those problems are typically perceived as localization errors involving front channel leakage into the surround channel, program modulated hum and noise in the surround channel, and excessive difference or surround channel information. A number of factors responsible for these problems are described in detail by Gibson [8] and have been reduced by improvements in transmitter and receiver design. In particular, incidental carrier phase modulation (ICPM) of the video signal into the audio during both transmission and reception can be a problem. Many of these problems are significantly reduced when multipath distortion is reduced or these signals are transmitted over a well-designed cable distribution system.

1.7 Future Improvements

A consideration of fundamental importance in developing the Stereosurround production format has been compatibility with existing consumer and professional two channel reproduction systems while at the same time offering a means of creating a significant improvement in spatial reality. The success and acceptability of the format, consequently, depend heavily on maintaining this compatibility with future system improvements. As an example, the bandwidth of the surround channel is not limited by the encoder, but rather by decoder and delivery system technology. Future programming is expected to take advantage of this fact resulting in productions suitable for playback on both wideband and restricted band surround channels, much the same as audio programs are now produced for both FM and AM broadcast.

2.0 Summary

The Stereosurround audio production format has been described from both a theoretical and practical standpoint. The process has been developed to allow the production of both audio only and audio-for-video programs with significantly improved spatial reality. It is compatible with a variety of existing multichannel decoding hardware as well as two-speaker stereo and single speaker monaural playback equipment. Because two-channel audio storage and transmission systems will be the dominant consumer format for many years to come, it is felt that this format offers a much better approach to creating more realistic sound fields with four-channel reproduction than is possible with any two-speaker stereo system, while retaining complete compatibility with two-speaker and single-speaker reproducing systems.

3.0 Acknowledgement

Of the many individuals involved with the development of the Stereosurround production format, the author would like to particularly thank John Bullock, Mark Gilbert, Stephen Julstrom, and John Owens for their dedication and numerous contributions.

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SYNTHEVISION—A NEW CHROMA-KEY IMAGING TECHNIQUE WITH HI-VISION BACKGROUND

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NHK(Japan Broadcasting Corporation) has developed a new image synthesizing technique called "Synthevision". A live NTSC camera foreground picture is synchronously keyed into a background picture which is digitally processed from an original Hi-Vision image. This background derived-picture is controlled by a computer using data from the foreground camera. If the camera image is altered by actions of zooming, panning and focussing, the background picture is altered accordingly. The combined image thus appears more real than that of conventional chroma-key imaging.

Synthevision is in daily use at NHK for the purpose of changing the background for each program.

INTRODUCTION

Synthesized images using the chroma-key technique are often used in TV program production. The main advantages of these images are that they:

(1) provide scenes that do not really exist in the studio,
(2) eliminate locations in order to shorten production time, and
(3) simplify studio sets, resulting in lower production costs.

In conventional chroma-key imaging, the synthesized image separates into two distinct areas of motion: the foreground camera image and the background scene in the case of panning, tilting or zooming of the foreground camera. Accordingly, the camera-person cannot change the direction or viewing angle of the camera during picture synthesis.

NHK has developed a new image synthesizing system which solves previously existing problems involved in conventional chroma-key imaging. With this system, as the foreground image is altered by the panning, tilting, zooming and focussing operations of the camera, the background picture changes accordingly. Thus, the synthesized image appears as real as that is taken with one live camera.

The theory of using digital data for background images has been previously reported[1], but has not been put into practical use until today. The principle feature of our system is the use of Hi-Vision pictures as background ones. Therefore, still and motion pictures can be used as background.

The picture synthesis technique employing this method is called "Synthevision". Currently, NHK uses the Synthevision system in its daily broadcasts of evening news, sports news and other special programs.

CONFIGURATION OF THE SYNTHEVISION SYSTEM

Fig.1 shows the configuration of the Synthevision system.

The background processor and defocus filter shown in Fig.1 are newly developed equipments. The background processor first stores an original Hi-Vision background in a frame store. The processor then selects a part of the Hi-Vision image that will follow camera work - panning, tilting and zooming, - and then converts it to a standard TV image. Undesired aliasing during image compression and judder during zooming and scrolling are eliminated by a digital filter.

The defocus filter is connected behind the background processor. It can blur a background image in accordance with the depth of focus for the object of the camera lens. Input and output images are the component signals of the standard TV format. The digital filter has an impulse response identical to that of the optical transfer function (OTF) of the lens.

Two live standard TV cameras shoot the subject in front of a chroma-key blue panel. Each camera has two rotary encoders on the universal head to measure panning and tilting angles, and three potentiometers within the camera lens to detect zoom, focus and iris positions. The computer illustrated in Fig.1 receives all these data from the cameras every 1/60 second and calculates the start address and magnification ratio to clip a part of the Hi-Vision image and the amount of blur to defocus. Then the computer transfers such data to the background processor as well as to the defocus filter.

As the derived background picture has a three-frame time delay from the time the camera data is received, foreground camera pictures must be delayed the same number of frames in order to be synchronously combined with the background picture. As a result, a three-frame delay is inserted after the video switcher of the two

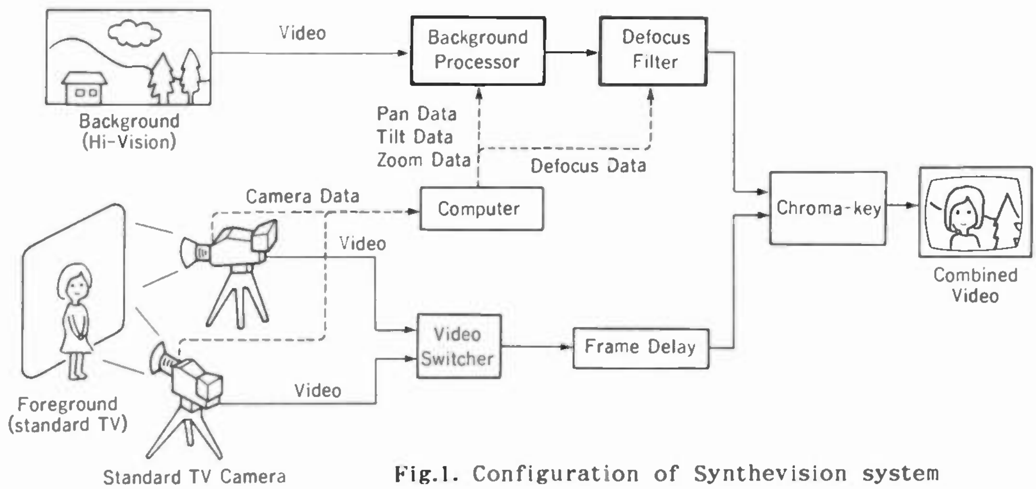

Fig.1. Configuration of Synthevision system

cameras.

To synthesize foreground and background, the conventional chroma-key technique is used. More realistic images can be obtained by using the soft chroma-key technique, rather than the hard chroma-key technique when defocussing the foreground camera picture.

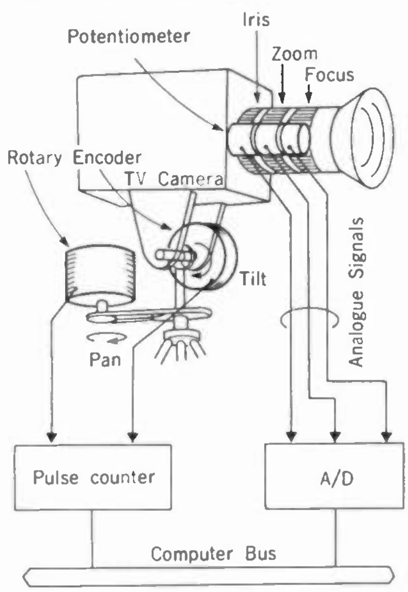

Fig.2. Camera data detection mechanism

Image size		2/3inch
Zoom ratio		10×
Maximum iris ratio		1 : 1.6
Focal length		11～110mm
M. O. D.		1m
Angular field	Wide	47°×36°
	Tele	5°×3.5°

Table 1. Lens specifications

CAMERA DATA DETECTION MECHANISM

Fig.2 shows the pan, tilt and lens data detection mechanism. Panning or tilting is detected by counting the output pulses of each rotary encoder mounted on the camera's universal head. Two rotary encoders are connected to the pan and tilt rotary shafts by a belt with a gear ratio of 1:1.

The camera data accuracy was designed based on a hand-held camera's ability to shoot a foreground with the 2/3 inch standard zoom lens shown in Table-1. The minimum data detection resolution must be equal to that of control amounts in digital processing. The hardware was designed to such required accuracy.

The minimum detection angle ($\Delta\theta$) of the tilt angle was chosen to be equal to 1/2 the scanning line size during maximum zooming (vertical picture angle 3.5°), since the effective number of scanning lines is 490.

$$\Delta\theta = \frac{3.5^\circ}{490} \times \frac{1}{2} = 0.004^\circ \qquad (1)$$

The rotary encoder used in our system generates 81,000 pulses per revolution and almost satisfies the minimum detection angle requirement.

The detection accuracy of pan data can be the same as that of tilt data.

As shown in Fig.2, lens data are detected by A/D, converting the output voltages of the potentiometers for control of zoom, focus and iris by a servo motor mounted on the regular lens.

Digital image processing of expansion and compression equivalent to zoom action requires the fixing of pixel positions in the center of the picture and changing the first read address (top left) of the image in the frame store to be processed in accordance with the magnification ratio. Thus the minimum resolution of zoom data greatly affects the edge of the picture more than the center. In addition, the amount of jitter must be less than 1/2 the scanning line size chosen when discussing the minimum detection angle.

Under the above conditions, we can obtain a result in which the zooming detection accuracy will be less than 0.002.

Fig.3 shows the actual measurement values of the zoom ratio versus output voltage of a 2/3 inch lens. As the zoom ratio increases approximately exponentially with the voltage, the A/D converter requires an accuracy of more than 12 bits.

The focus data are used when defocussing background images in accordance with the depth of focus for the object. Therefore, the problem of calculating focus data detection accuracy is the same as the problem of designing the minimum unit of the coefficient of a filter (explained below) that can defocus images smoothly and continuously, as seen by the human eye. Simulated vision evaluation tests have shown that the minimum sufficient coefficient unit is 1/8. The detection accuracy of focus data at this time is 0.001.

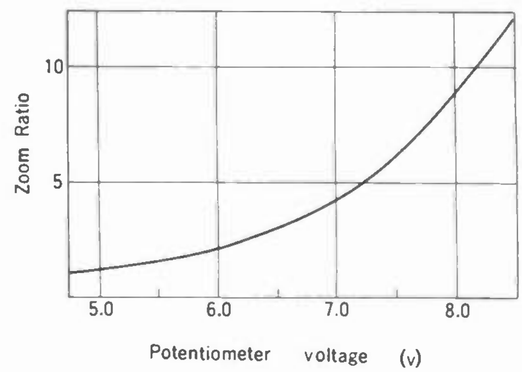

Fig.3. Zoom ratio vs. detecting voltage

BACKGROUND PROCESSOR

Fig.4 shows a diagram of the background processor.

The processor has two Hi-Vision frame stores assigned to the computer's real memory space. The two stores are used as background images for the two cameras. Background images are input into the processor as follows.

(1) Still pictures are transferred from an external computer memory device such as a magnetic disc.

(2) Still pictures are frozen in the frame store by connecting Hi-Vision signals to the input section.

(3) Motion pictures are input by connecting a Hi-Vision VTR or a camera to the input section.

The maximum expansion ratio is 4 times and the minimum compression ratio is 1/4 using pixels as a reference. For example, a magnification ratio of 1 means to output one Hi-Vision pixel to one standard TV pixel. The magnification data are expressed in terms of two bits of an integer part and 12 bits of a fractional part. The minimum number of effective numerals is 10 bits. This value satisfies the required zoom data accuracy.

The scrolling units are 1/16 pixel or 1/16 line of a Hi-Vision image. This corresponds to 1/4 pixel or 1/4 line of an output image, when input image is expanded to the maximum ratio. The data for scrolling are expressed in terms of 12 bits of an integer part and 4 bits of a fractional part in 2's complement.

The digital filter for expansion, compression and system conversion is separate from the horizontal(H) and vertical(V) filter.

DEFOCUS FILTER

A new defocussing technique with a recursive filter was adopted to obtain the effect created by a lens.

In geometric optics, images of a spot light source become a disc of uniform brightness called a circle of confusion. In principle, these characteristics can be expressed using a transversal two-dimensional digital filter. However, several tens of thousands of taps are needed to configure the filter.

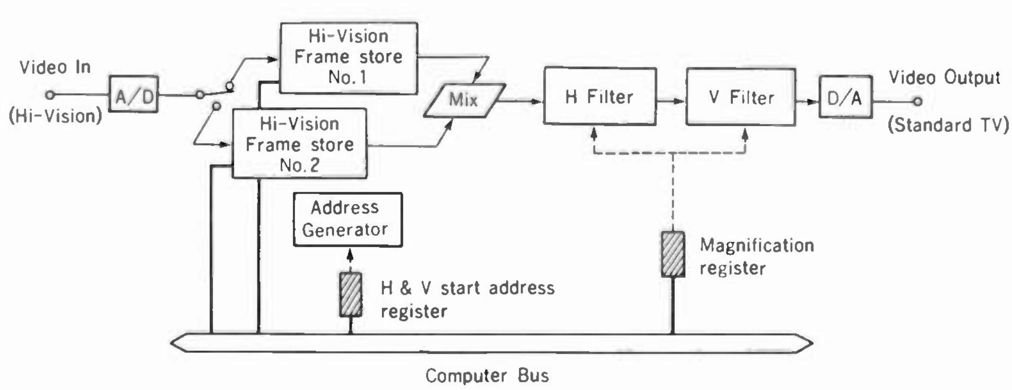

Fig.4. Block diagram of Background Processor

It is for this reason that filtering is performed independently in H and V directions.

Assuming the defocus amount (number of pixels along the radius of a circle of confusion) to be R, the impulse response I(z) into H direction of optical defocussing shown in Fig. 5 can be expressed as follows.

$$I(z) = \frac{1}{2R+1} \cdot \frac{1-z^{-(2m+1)}}{1-z^{-1}} \cdot z^{m} + \frac{\delta}{2R+1}(z^{(m+1)}+z^{-(m+1)}) \tag{2}$$

$$R=m+\delta,\quad m:\text{integer},\quad 0 \leq \delta < 1$$

This equation represents a kind of recursive filter and shows that a filter can be built with a small number of devices rather than employing a transversal filter.

The maximum value of the radius of a circle of confusion equals roughly 60 scanning lines, assuming a 2/3 inch lens is used, and so the maximum value of R was chosen to be 64. The data for defocussing are expressed in terms of 7 bits of an integer part which corresponds to (m) and 4 bits of a fractional part which corresponds to (δ). A fractional part matches the minimum unit of the filter's coefficient which is described above.

The defocus amount caused by the depth of field can be derived from optics calculations. The defocus amount of a background image, if the camera is focussed on a person in the foreground, as shown in Fig. 6, is given by

$$r = \frac{f^2}{2F}\left(\frac{1}{a} - \frac{1}{c}\right) \tag{3}$$

where

r: radius of a circle of confusion,
f: focal length of the zoom lens,
F: F value of a lens,
Assuming, f<<a.

If the defocus amount (R) is denoted as a unit of the scanning line size (ρ) (: 13.5 μm with a 2/3 inch device), the following equation is given.

$$R = \frac{r}{\rho} = \frac{f^2}{2F\rho}\left(\frac{1}{a} - \frac{1}{c}\right) \tag{4}$$

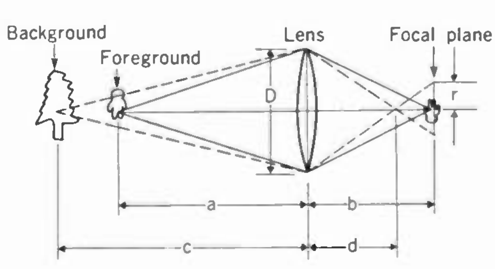

Fig.6. Basic method for calculating the circle of confusion

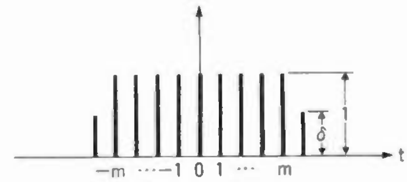

Fig.5. Defocus filter impulse response

The defocus amount of background images is proportional to the square of the focal length (f) and inversely proportional to the F value. The defocus amount surpasses that of electronically expanded background images, shown in Fig.7, indicating that background images must be further defocussed with a defocus filter. The computer employs this equation for calculation based on the focus data and transfers the resulting data to the defocus filter.

APPLICATIONS

Special Effect Scenes

The announcer , while providing commentary, moves across the surface of Mars - a blue curtain spread out over the studio floor - as far as the cyclorama. The bird's eye foreground camera follows the announcer, while the Hi-Vision still image of Mars is subjected to perspective conversion. The camera follows the broadcaster's movement, as shown in Photo.1-(a) and (b). Such a vivid background has become possible only since the advent of Synthevision.

Electronic Studio Settings

All that occupies the studio is a desk placed in front of a chroma-key panel. The newscaster is followed by a commentator, and two different backgrounds are prepared for them. Background scenes are prepared using Hi-Vision computer graphics(CG), recorded in a magnetic disc and transferred to the frame store of the background processor as necessary. Switching from the background for news (Photo.2) to background

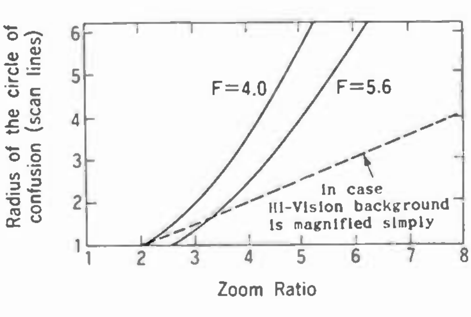

Fig.7. Defocus value vs. zoom ratio where a/c is 0.8 in Fig.6

for the commentator (Photo.3) occurs instantly, as if the programs were broadcast from different studios. In this way, Synthevision enables settings in the studio to change rapidly, creating the possibility of broadcasting a live program which requires multiple settings from a small studio.

CONCLUSION

A new image production technology called Synthevision has been developed by NHK.

Image production by Synthevision not only expands upon the conventional chroma-key ability, but also makes it possible to achieve the following effects.

(1) The flow of conventional image production can be changed. Flexibility of program production can be increased by synthesizing background scenes after shooting people appearing in pictures.

(2) By using CG images in background scenes, studio settings that cannot possibly exist can be easily employed for program production.

(3) By storing background images in databases, background scenes can be changed instantly, allowing production in narrow-space studios.

ACKNOWLEDGMENTS

The authors would like to thank Mr. Kutsuzawa, and Dr.Ninomiya, former and present General Managers, respectively, and Mr. Ishikawa, Chief Researcher, Video Engineering and Data Processing Division, as well as Dr. Sugimoto, Director, Science and Technical Research Laboratories of NHK for their guidance in conducting this research.

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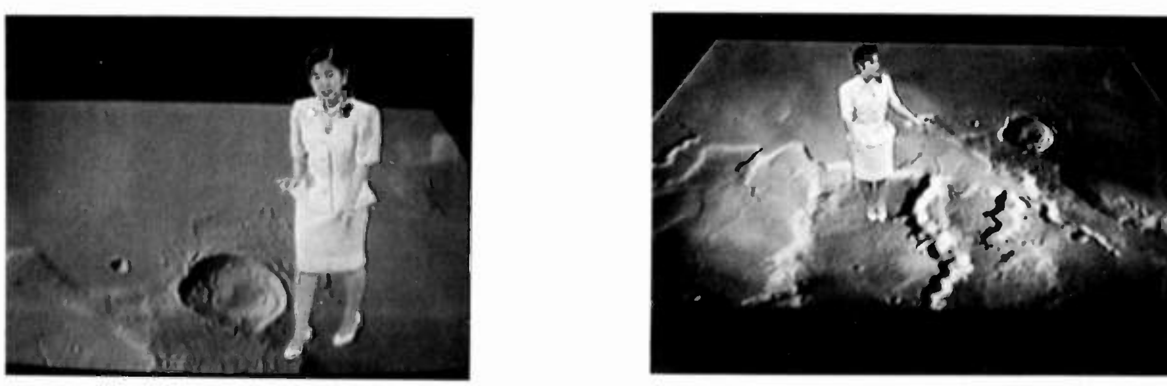

(a) (b)

Photo.1. Examples of sequence of frames of scene

Photo.2. Scene of news

Photo.3. Scene of commentary

A REPORT ON THE NRSC FM SUBCOMMITTEE ACTIVITIES

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BACKGROUND

Late in 1988 in response to continued concerns with the so called "AM-ization" of the FM band, the National Radio Systems Committee formed a working group dedicated to identifying and investigating specific problems in the FM domain. As is the usual case with NRSC sponsored activities, the FM Subcommittee members are composed of interested persons in the broadcast industry, equipment vendors, and receiver manufacturers. It is operated under the patronage of the National Association of Broadcasters and the Electronics Industries Association and follows the committee guidelines established by these two organizations.

At it's initial meeting, the committee considered several different proposals that could warrant investigation. After thorough discussion, committee members decided to concentrate on two fundamental problems, multipath and composite spectrum occupancy, and in February, 1989, working groups were formed to examine these areas of concern.

MULTIPATH STUDIES

Not one engineer breathing today would deny that, aside from audio processing, this is one of the most discussed technical topics in FM broadcasting. And, after a year of work, the multipath studies group has, as of yet, failed to remove this one from the list.

WAEB

The February, 1989, meeting produced a volunteer test bed in the form of radio station, WAEB, Allentown, Pa. Working group chairman, Harry Simons, Chief Engineer of WAEB, was also selected at this meeting. He and consulting engineer, Ted Schoeber would spearhead a multipath investigation project. Receiver and antenna manufacturers were also involved early on.

The WAEB proposal was ambitious. Multipath distortion, synchronous AM noise effects, and combinations of the two were to be primary targets. But, the test criteria was eventually expanded to include group delay, SCA effects, antenna patterns, echo delay by pulsed transmitter, and many other facets of field testing. Multiple waivers of FCC rules were to be obtained to accommodate the tests. At this point, to save working time and problems with committee decisions, the NRSC FM subcommittee decided not to officially sanction this project and allow WAEB to follow its own course. Principals involved would report to the FM subcommittee as they saw fit.

Laboratory Studies

Soon after the WAEB project began, a proposal for a multipath study in the lab was presented by Tom Silliman of Electronics Research, Inc. Since an eight bay end fed antenna represents a worst case situation with fairly dramatic phase error at the top bay, he suggested fabrication of an eight bay in-line feed harness that would mate with an eight bay branch feed power divider. This network would produce a simulated antenna with a 50 ohm input and 50 ohm output that could be used for laboratory measurements. Group delay would be measured and, using an exciter on the input port and various pieces of test equipment on the output port, multipath distortion, AM noise, and baseband distortion measurements could be made. Other antenna parameters could also be analyzed during these tests.

Unfortunately, the real world being what it is, time demands on ERI and the FM subcommittee caused a delay in start up of this project. Since this also represents a departure from the original intent of purely defining multipath, the committee recently voted to proceed with this project under the auspices of an informal "RF Studies" group. The ERI project has been commissioned and will commence within the near future.

COMPOSITE SPECTRUM OCCUPANCY

Running almost neck and neck with multipath, occupied bandwidth is another hot topic of discussion. More stations are coming on the air all the time. Contrary to what some may think, 80-90 upgrades, A+ upgrades, C3s, and directional antennas represent some degree of interference. As the contours get closer we will have to retain some semblance of protection by staying within the allocated spectrum bandwidth and yet maintain competitive loudness.

With this thought in mind, the composite spectrum working group first met in April, 1989. Under the leadership of chairman, Ed Anthony, working group goals were assembled. "Real world" situations were to be considered, including SCA and possible causes of splatter. Receivers were to be included as part of the FM system.

Working group participants agreed to focus on four specific areas: (1) acceptable adjacent channel protection, (2) how much bandwidth for a transmitted signal is too much, (3) necessary 19 kHz protection, and (4) how much SCA/stereo and stereo/SCA crosstalk protection is necessary. A paper concerning composite clipping and overmodulation was also presented for the committee's information.

The second group meeting in November, 1989, produced only limited work on the original goals. A list of action items, developed at the first meeting were dealt with and three items, investigation of overshoot bandwidth, Walsh decoders, and further investigation of SCA requirements were added. Regrettably, the submission date of this paper prevents further reporting on this committee's activity since the third meeting will be in February, 1990.

CONCLUSION

Problems with multipath distortion and adjacent channel interference will not disappear by wishing them away. Controversy has enveloped some of the work accomplished by this committee and that is unfortunate. Many people have contributed their talent and much time, labor, and money has been spent. The people who support the committee are to be congratulated for their efforts. A lot of work remains to be done by the NRSC FM subcommittee and it will continue to move forward with the projects it is involved in.

OPTIMIZING THE PERFORMANCE OF THE FM TRANSMITTER ANTENNA

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This paper reports on the results of a study to determine the optimum configuration for FM transmitter antennas. The study included a survey of antennas in current use for FM broadcast and revealed a significant gap between what could be done, in a practical way, to optimize FM transmission, and what is actually done in practice. The important factors of antenna impedance, radiation pattern, polarization, bandwidth and "on-the-tower" performance are discussed. The original study was conducted on the basis of operation external to the U.S. The impact of FCC regulations, however, is included in this discussion. An optimized antenna configuration, the Linear Vertically Polarized Antenna, is described. this antenna has been tested in operation in both urban and rural areas and has been shown to provide significantly better performance than do antennas commonly used. Both the theory of operation and measured performance data are presented.

1. Introduction

Recently, FM broadcasts using vertically polarized transmissions were initiated in Israel by IDF Radio. The subject of this paper is 1) the design of the vertically polarized antenna system that is being used in Israel, 2) the success of this system in achieving better penetration into service areas than had previously been achieved using either horizontal or dual polarizations and 3) the implications that this may have for FM broadcasting in the U.S.. In order to have a proper reference for this discussion, we will review both the general requirements for FM transmitter antennas and also the characteristics that are typical of the circularly polarized antennas in common usage in the U.S.. Of special importance are the factors of gain, impedance, bandwidth and polarization.

2. FM Transmitter Antenna Requirements

The following is a very general and necessarily brief listing of the requirements that all FM transmitter antennas are required to meet.

2.1 Directional Gain: - all antennas, to a greater or lesser extent, are directional. It is important to maximize gain in the desired directions and to minimize gain in the unwanted directions. For example, it is generally required that radiation on the horizon be maximized in order to maximize signal to the service area; and that it be minimized in directions up to the sky and down to the tower base in order to minimize fading, radiation hazard, RF interference and sheer waste.

2.2 Polarization: - The question of the optimum choice of polarization for FM has historically been troublesome and confusing. The original plans for FM broadcasts, made prior to World War II, called for horizontal polarization. In 1946 in order to overcome the deficiencies of horizontal polarization, the FCC authorized the use of "... supplemental vertically polarized effective radiated power..." [Ref. 1]. In practice, this has come to mean "circular" or "dual" polarization in which presumably equal amounts of horizontally and vertically polarized energy are radiated. Today the vast majority of FM transmitter antennas in use are of this so-called CP type. From an engineering rationale, the polarization used for transmission should be chosen so as to best match the polarizations of receiver antennas, i.e. for car and home stereos. One may therefore ask " How many car or home FM receivers have a circularly polarized receive antenna? ". The answer is - zero to none. If one asks " How many have a horizontally polarized receive antenna ? "; the answer is - a tiny percentage. If one asks " How many have a vertically polarized receive antenna ? "; the answer is - most cars, the bulk of todays market, have a vertical whip receive antenna ! Most home FM receivers have a wire hanging out the back, having equal probability of being vertically or horizontally polarized. The conclusion to be drawn from this is that, for optimum match between the polarizations of transmitted and received signals, vertical polarization is best.
This author is aware that the above is a highly simplified analysis. The conclusion, however, is believed to be

correct. This is supported by the highly successful results achieved by IDF Radio in Israel using vertical polarization. These results showed increases in signal levels that exceeded 5 dB as compared to the levels with equivalent CP antennas in both urban and rural areas. Besides increased signal levels, there are additional bonuses to be gained by the use of vertical polarization: a) a vertical dipole has a null in the tower base direction, thereby minimizing radiation hazard in accordance with OSHA regulations. Also RF interference to nearby studio equipment and telephone lines is minimized, b) vertical polarization provides an extra 20-25 dB interference rejection between Low VHF (Channel 6) and FM, and c) interference by power lines and telephone lines is reduced.

2.3 Broadband Operation: The FM band is 88 to 108 MHz, and the FM channel width is 200 KHz. It is desirable that the FM transmitter antenna have multichannel, broadband impedance performance (i.e. low SWR) in order to minimize frequency sensitive performance degradation due to icing and other perturbations. Ideally, the antenna should cover the entire 88-108 MHz band. (This was, in fact, a requirement of IDF Radio).

2.4 On-the-Tower Performance: It is important that specified radiation pattern and impedance performance be achieved in operation, on-the-tower, and not solely in some idealized free-space environment. This means that the design of the antenna should either account for tower effects or de-couple the antenna from tower effects or both.

3. Conventional CP Antenna

The antenna element that is conventionally used in the U.S. is circularly polarized (CP) and as pictured in Fig 1. is a variation of the helical loop antenna [Ref 2]. We shall list here nominal performance characteristics for a 4-bay array* of such elements: - Peak Gain: 3 dBd (horizontal or vertical polarization), Circular polarization: +2dB, Maximum SWR: 1.1 over ± 200 KHz, Azimuthal Plane Radiation Pattern: omnidirectional + 2dB.

The above characteristics are stated by manufacturers with the proviso that they are valid for free space conditions only. The CP element is an open, high Q, frequency sensitive antenna. Therefore its on-the-tower SWR and radiation pattern performance is uncertain as it is strongly affected by coupling to the tower structure, icing and other perturbations. In fact, manufacturers of CP antennas recommend the use of heating elements and radomes, at added cost and weight, as a means of at least partially overcoming these difficulties.

4. Vertically Polarized Linear Antenna Element

A study to determine the optimum configuration for FM transmitter antennas was undertaken by this author at the request of IDF Radio, Israel in 1986. As a result of this study it was concluded that a broadband half wave dipole with reflector as shown in Fig. 2., was the optimum FM transmitter antenna element. Antennas of this type are manufactured by the Bogner Broadcast Equipment Co. and are designated as the BVDO series. The guaranteed on-the-tower performance characteristics for a 4-bay array of such elements (i.e. BVDO-4) are as follows: Maximum SWR: 1.1 over ± 500 KHz anywhere in the 88-108 MHz band. Changes in the frequency of operation of more than ± 500 KHz. are easily made by means of a variable impedance tuner. This can be done in the field if necessary. But it is often unnecessary to do so as the maximum SWR is 1.4 throughout the entire 88-108 MHz band. Tower effects are minimized due to both the low Q, broad band nature of the dipole and to the isolation provided by the reflector element. The comparison between the SWR/Bandwidth performance of typical CP and Vertical Linear antenna elements is shown in Fig. 3.
Icing effects and other perturbations are negligible due to the dipole feed point being internal, widely spaced and enclosed by an integral radome.

Peak Gain and Radiation Pattern: The peak gain of a 4 bay array* of vertical dipole/reflector elements is 8 dBd, vertical polarization only.
Now 3 db of this difference is due to the dual polarization of the CP antenna. The remaining 2dB difference is due to the presumed omnidirectional azimuth plane radiation pattern. However, for side-mounted installations, tower effects would preclude an omnidirectional azimuth pattern for the CP array. The azimuthal plane radiation pattern of the BVDO antenna is a smooth cardiod, Fig. 4. The smooth cardiod pattern is ideal for directional, side-mounted specifications. For top-mounted installations it is possible to achieve an excellent omnidirectional azimuthal plane pattern using the Bogner BVDT design (Fig. 5). We note that such top-mounted, self supporting antennas are heavy and expensive.

5. Conclusions

A vertically polarized linear antenna for FM transmission has been described. The performance advantages of this antenna in gain and bandwidth has been detailed. It is believed that there are many cases, given proper filing procedures, where this antenna will be the optimum choice for FM transmission.

* The 4-bay array is taken as an arbitrary reference for purposes of our discussion.

References:
[1] FCC (10-1-88 Edition)

§ 73.316 FM antenna systems.

(a) It shall be standard to employ horizontal polarization; however, circular or elliptical polarization may be employed if desired. Clockwise or counterclockwise rotation may be used. The supplemental vertically polarized effective radiated power required for circular or elliptical polarization shall in no event exceed the effective radiated power authorized.

[2] J. Kraus, " Antennas ", 2nd Edition, McGraw-Hill 1988.

Figures:

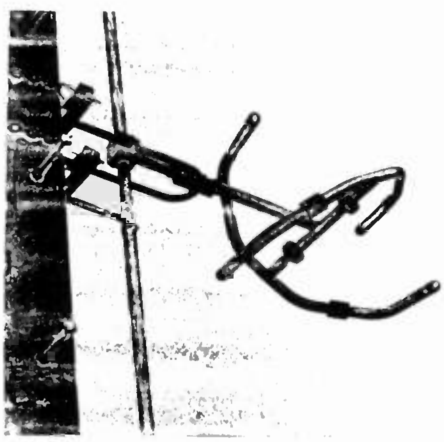


Figure 1: Typical CP antenna element.

Figure 2: Vertical dipole/reflector antenna element.

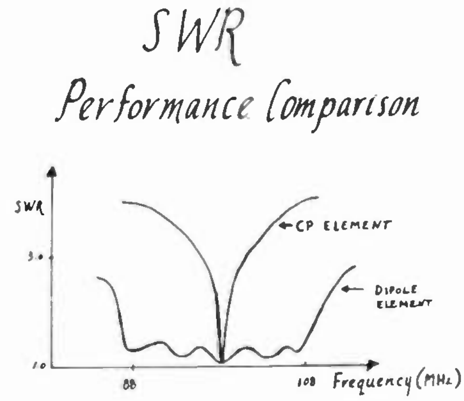
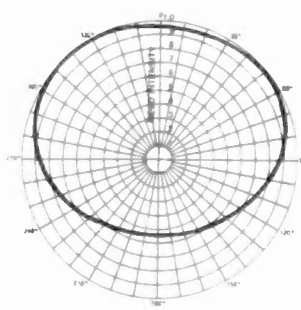


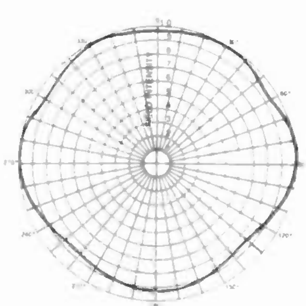
Figure 3: SWR performance comparison of typical dipole/reflector and CP antennas.



Side-Mount Cardioid (220°) Mounted Off 5" dia. Leg
Gains (above a dipole)

BVDO2 = 5dB
BVDO4 = 8dB
BVDO6 = 10dB

Figure 4: Azimuthal plane pattern of side mounted BVDO antenna.



Top-Mount Omnidirectional*
Gains (above a dipole)

BVDT2 = 3dB
BVDT4 = 6dB
BVDT6 = 8dB

*Self-supporting

Figure 5: Azimuthal plane pattern of top-mounted BVDT antenna.

Acknowledgment: The author is grateful to Gabriel Koerner of IDF Radio and to Carol Hamilton of Bogner Broadcast Equipment Co. for their very helpful support of this work.

CUSTOMIZED PATTERN APPLICATIONS OF THE FM CBR ANTENNA

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Abstract: The CBR (cavity backed radiator) panel element has been proven to have high versatility and predictability in multiplexed FM antenna array radiation patterns. In this paper special customized patterns with deep nulls (notch patterns) are discussed. Additionally, the importance of predicting the radiation patterns by using both amplitude and phase information from a single cavity on the corresponding tower is also stressed.

INTRODUCTION

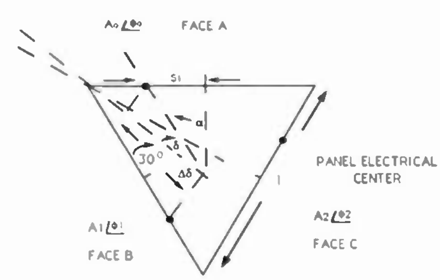
The request for special patterns from the broadcasters is becoming more common every day. In some of these patterns, protection needs to be provided in certain directions (notch pattern) in order to avoid interference between close stations. These patterns, as well as others, can be accomplished by skewing the corresponding CBR elements off center by an appropriate amount, and by adjusting the electrical phase of the corresponding feedline in order to obtain the necessary phase difference between the adjacent cavities.

As the cavity antenna is an aperture radiating element (1), both the radiation pattern changes in amplitude and phase throughout the azimuth plane. Therefore, it is of vital importance to take both into account to predict accurately the shape and nulls in "special" patterns.

MATHEMATICAL BACKGROUND

From antenna array theory it is well known that by adjusting the electrical phase, as well as the distance between point sources an addition or cancellation of the E-field pattern results in a certain direction.

Figure 1 depicts a planar array of three point



α = skew angle
s = cavity displacement
A = relative feeding current amplitude
ϕ_i = electrical current phase
L = tower section
δ = pattern direction control

GEOMETRICAL CBR ARRANGEMENT
AND CORRESPONDING PARAMETERS
FIGURE 1

sources (cavities) around a triangular tower. If the desired pattern control is in the direction δ (relative to a tower leg) the following equation has to be applied:

$$\Delta\phi + K\Delta\delta = \Psi \qquad (1)$$

Where $\Delta\phi$ is the current and radiating element phase difference between face A and B:

κ = is the wave number $2\Pi/\lambda$

$\Delta\delta$ = is the projected path difference between source A and B respectively.

Ψ = is the desired angle for addition or cancellation $2N\Pi$ or $(2N+1)\Pi$ respectively where $N = 0, \pm1, \pm2$.

s = mechanical panel shift

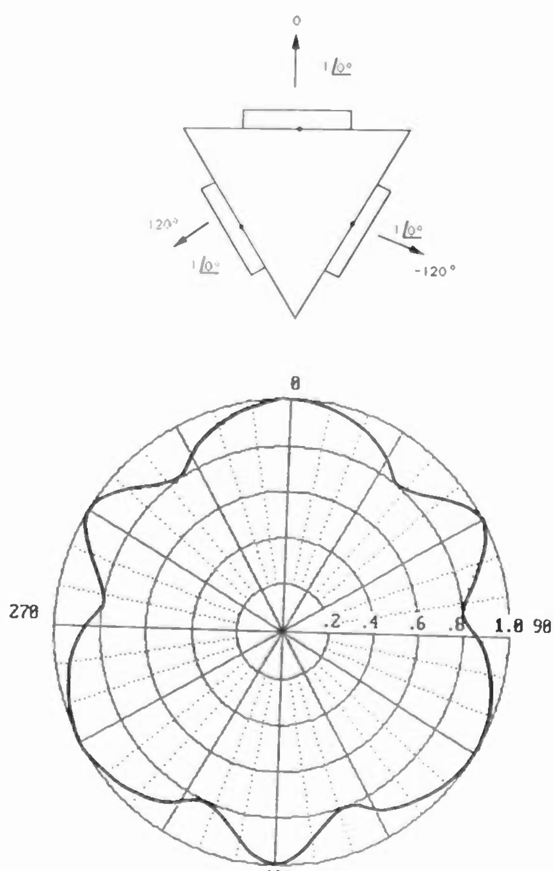
l = tower section

δ = direction of pattern control

From equation, (1) by replacement, it can be shown that the required mechanical panel shift is given by the following expression:

$$s = \frac{1}{2\sqrt{3}\cos\delta}\left(\frac{\Psi - \Delta\phi}{\Pi} + L\sin\delta\right) \qquad (2)$$

It can be noted that for $\delta = 0$, the mechanical shift is independent from the tower dimension [2], [3]. Additionally, for this case, $\Delta\Phi$ is equal to the phase excitation difference between the adjacent panels.

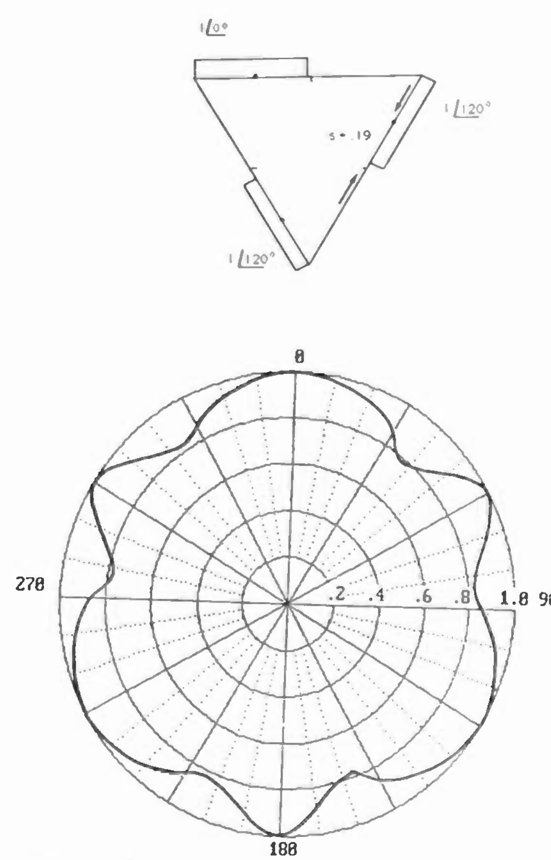


OMNIDIRECTIONAL PATTERN WITH EQUAL ELEMENT EXCITATION AND NO SKEW
FIGURE 2a

PATTERN EXHIBITS

By taking the element H-pol and V-pol E-field amplitude and phase far field measurement data and feeding them in a PC computer (using an azimuth pattern program called AZIPAT), different new patterns have been generated.

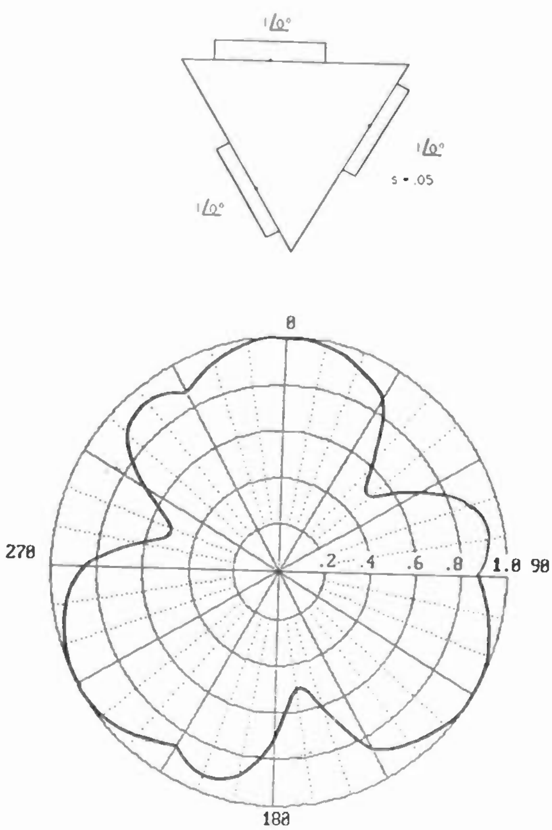
In Figures 2a and 2b, omnidirectional H-pol patterns of three cavities around a 6'6" tower, are shown. Both patterns look very similar but were obtained under different conditions. While in Figure 2a pattern the amplitude and phase excitation of all the three elements are the same, in Figure 2b the elements are excited in progressive phase (0, +120, -120°) and shifted .19 wavelengths. This value of displacement can be obtained directly by using expression (2) and in order



OMNICIRECTIONAL PATTERN WITH PROGRESSIVE PHASE AND SKEW
AMPLITUDE: 1, 1, 1
PHASE: 120, -120, 0
SKEW: .19, .19, .19
FIGURE 2b

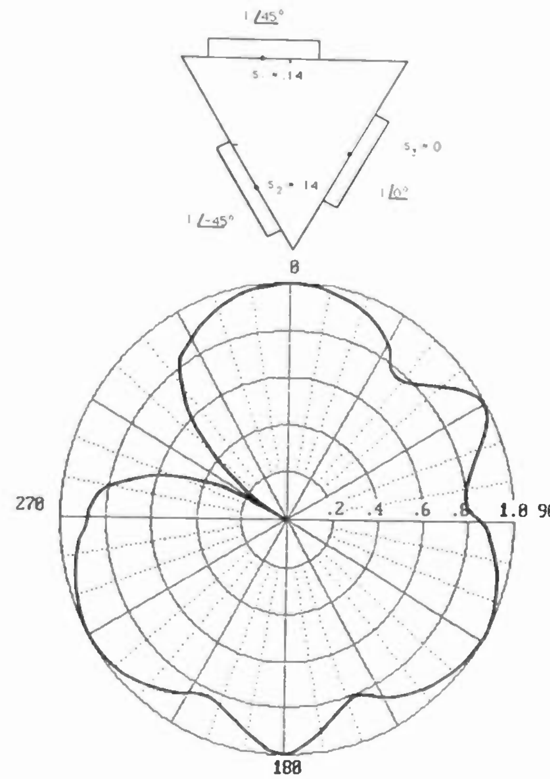
to accomplish an omni pattern field addition has to occur for 0, 120, and -120° (Ψ = 0) respectively.

By reducing the mechanical phase shift to s = .05 wavelengths, a trilobe pattern emerges with three nulls at approximately 50% as shown in Figure 3.

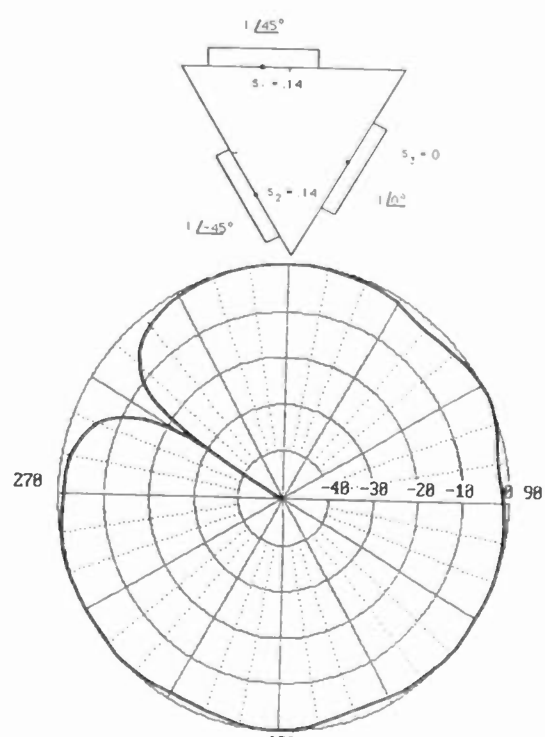


TRILOBE PATTERN WITH PROGRESSIVE SKEW
AMPLITUDE: 1, 1, 1
PHASE: 120, -120, 0
SKEW: .05, .05, .05
FIGURE 3

In Figure 4a and 4b a pattern with a very deep null in relative and dB field plot, respectively, is depicted. These patterns are useful in case an omnidirectional radiation is required with the exception of one direction which needs to be protected (notch). It can be accomplished by shifting two cavities .14 wavelengths and exciting them with equal amplitude and different phase (+45°, -45°) which leads to $\Psi = \Pi$. The third cavity has zero shift and phase and same amplitude. A



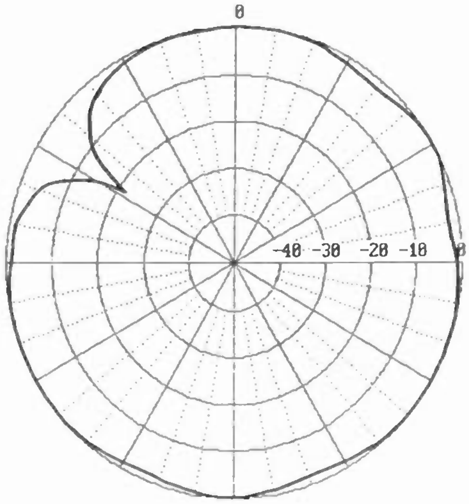
NOTCH PATTERN (AZ: 300°) IN RELATIVE FIELD
AMPLITUDE: 1, 1, 1
PHASE: 45, -45, 0
SKEW: .14, .14, 0
FIGURE 4a



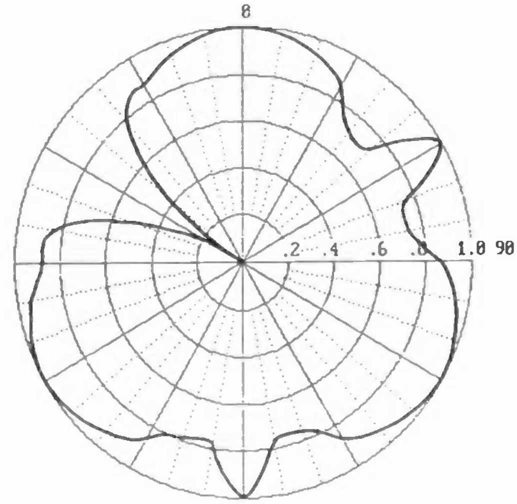
NOTCH PATTERN (AZ: 300°) in dB
FIGURE 4b

total signal cancellation in the direction of the tower leg adjacent to the shifted cavities (Azimuth: 300o) results and in the other two directions perfect signal addition occurs.

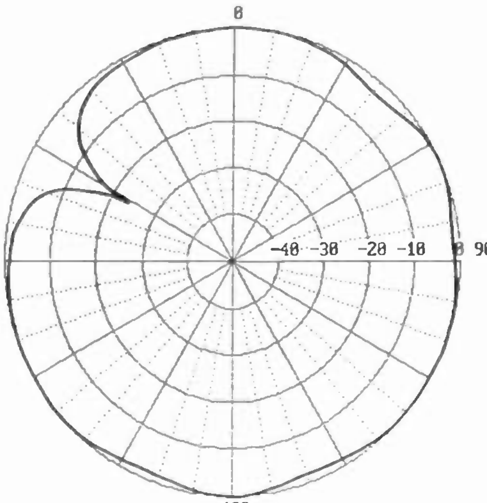
For a pattern corresponding to a different frequency (within ±5% in the FM bandwidth) a small shift in the position of the null results, but on the other hand, a drastic reduction in attenuation is obtained. Figures 5a and 5b show how the null is shifted and changed in attenuation (less attenuation) respectively in case of a change in frequency of ±5% respectively, happens. As can be pointed out, cnly a minor shift is produced (approximately ±1o) but attenuation, still appreciable, decreases (from α to -25dB). For the case of V-pol, from Figures 6a and 6b, a slightly different pattern appears with lower circularity in the omnidirectional region and a deeper null appears in case of 5% change in frequency. This is basically due to the fact that the element V-pol phase radiation



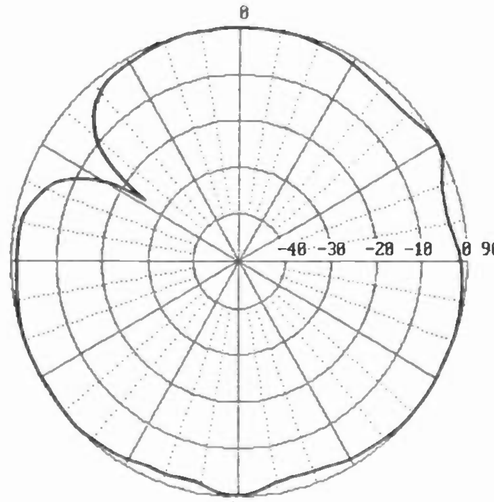
SHIFTED NOTCH PATTERN IN dB
(5% IN FREQUENCY CHANGE)
FIGURE 5a



V-POL NOTCH PATTERN
FIGURE 6a

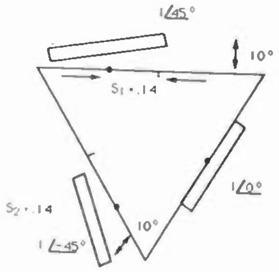


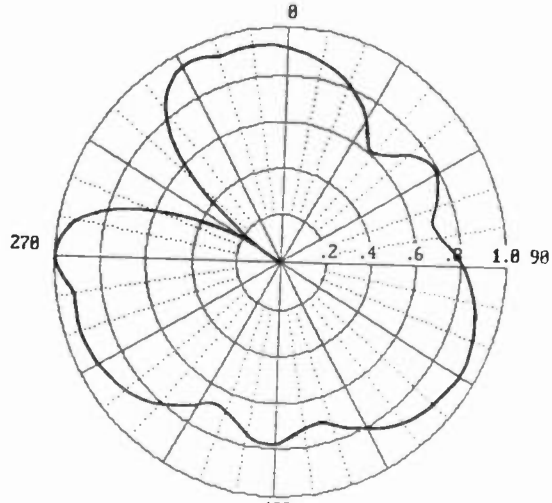
SHIFTED NOTCH PATTERN IN dB
(-5% FREQUENCY CHANGE)
FIGURE 5b



SHIFTED V-POL NOTCH PATTERN IN dB
(5% FREQUENCY CHANGE)
FIGURE 6b

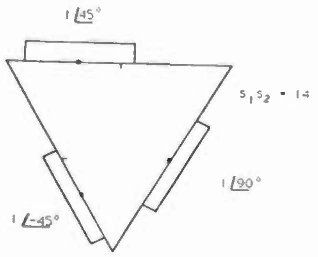
pattern is different from H-pol. By using in addition a skew, phase change and panel rotation (±10°), the rate of the notch drop off can be increased, and a more directional pattern results as shown in Figure 7.

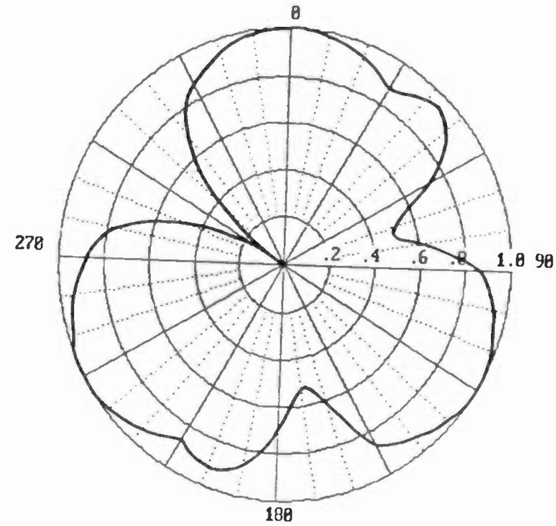




NOTCH "HEART" PATTERN
AMPLITUDE: 1, 1, 1
PHASE: 120, -120, 0
SKEW: .19, .19, 0
ROD: 10, -10, 0
FIGURE 7

A trilobe pattern with a deep null can be achieved by keeping the same arrangement set up as in Figure 4a, but increasing the phase on the cavity of face 3 to 90° as observed in Figure 8.

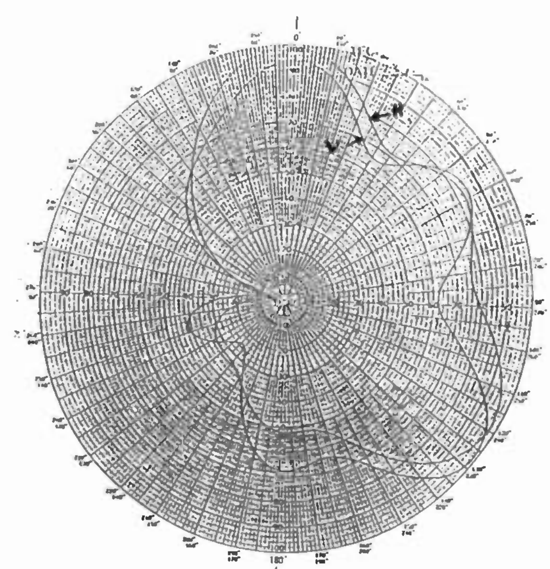




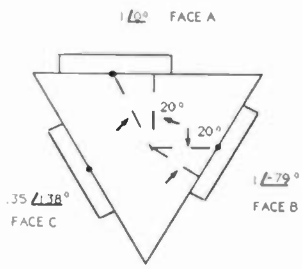
TRILOBE NOTCH PATTERN
AMPLITUDE: 1, 1, 1
PHASE: 45, -45, 90
SKEW: ,19, .19, 0
FIGURE 8

MEASURED VS. CALCULATED NOTCH PATTERN

Figure 9 shows the H-pol and V-pol radiation pattern of an actual 1 bay x 3 cavities around tested at Harris Test Range in Palmyra, Missouri. The corresponding arrangement is depicted in Figure 10.



MEASURED H AND V-POL PATTERN
FIGURE 9

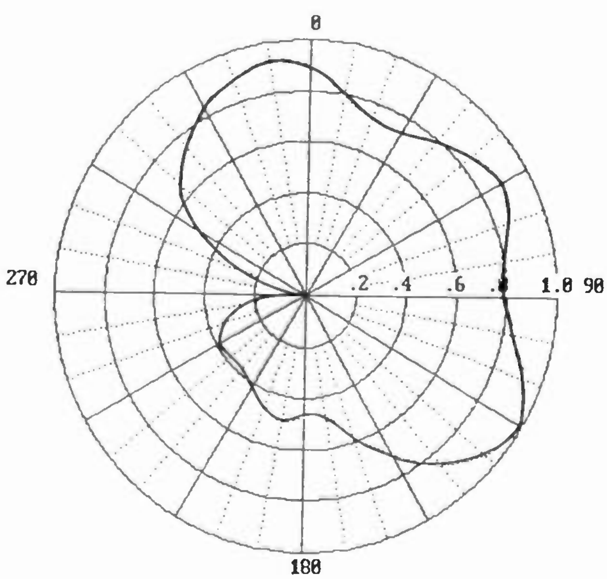


TESTED CAVITY ARRANGEMENT

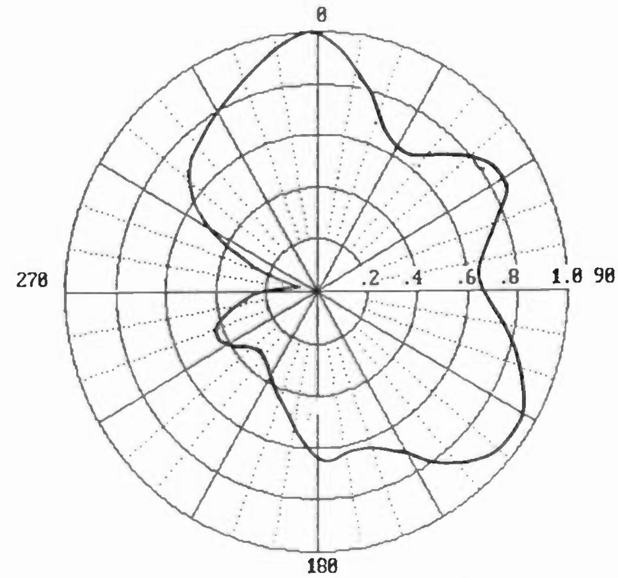
Amplitude: 1, 1, .35
Phase: 0, -79 and 138°
Skew: 20, 20, 0

FIGURE 10

By taking into account both amplitude and phase information of the radiation pattern of a single cavity, the patterns shown in Figures 11a and 11b were calculated. Close agreement can be observed by comparing with Figure 9 which shows the actual measured V and H pol patterns. Small differences in the pattern can be observed at, especially in the V-pol pattern due primarily to the tower scattering and diffraction, and also to mutual coupling between the cavities.



CALCULATED H-POL PATTERN
FIGURE 11a



CALCULATED V-POL PATTERN
FIGURE 11b

COMMENTS

Different notch patterns have been obtained by using phase change, skew, and cavity rotation. Additional patterns with null depth control at a certain direction can be obtained by using different amplitude excitation in adjacent cavities.

Also, notch steering can be accomplished by adjusting appropriately the forementioned variables.

CONCLUSIONS

1) The CBR panel element has been proven to be very versatile to produce comes to getting different patterns. By using skew and element rotation and difference in amplitude and phase excitation, non "standard" patterns may be achieved.

2) In the notch patterns, small deviations have been observed by changing frequency, which makes the cavity antenna a potential building block for multiplexed FM array antennas that need to provide protection in the pattern in a certain direction.

3) It is important to take into account both amplitude and phase of the radiated E-field of a single element for the calculation of patterns has been highlighted, in order to predict more accurately the coverage.

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2) C. Pantsios and F. Pantsios, "Smoothing of Omnidirectional Horizontal Radiation Patterns of TFD Panel Antenna Arrays with Progressive Current Phase Shift Feeding", IEEE Transactions on Broadcasting, Vol. BC-31, No. 2, June 1985.

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THE SIGNIFICANCE OF RF POWER AMPLIFIER CIRCUIT TOPOLOGY ON FM MODULATION PERFORMANCE

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ABSTRACT

The power amplifier bandwidth affects not only the FM modulation performance but also the transmitter's immunity to RF intermodulation. The trade-offs involved are discussed with recommendations on the choice of bandwidth. The design of RF power amplifiers for high quality FM performance requires careful considerations in the selection of input matching and output coupling circuit topologies due to their effects on the transmitter amplitude and group delay responses. Results of computer circuit analysis and measured amplitude and group delay responses are compared for different circuit topologies. Modulation performance data of a typical 20 kW single-tube FM transmitter is also presented to illustrate the effects of RF power amplifier input and output circuits. The effects of RF power amplifier tuning for symmetrical amplitude response (minimum synchronous AM) versus symmetrical group delay response on FM modulation performance is also discussed.

INTRODUCTION

Limited bandwidth RF power amplifiers are widely used in frequency modulation (FM) broadcast transmitters to increase the level of the FM signal at the exciter output to the kilowatt or higher power output levels. The power amplifier (PA) is typically a high gain, single-tube, class "C", tuned, radio frequency (RF) amplifier. The PA design goal is to deliver the authorized power output to the antenna with high efficiency and reliability while providing excellent modulation performance.

This paper discusses various topologies of the input and output circuits of a vacuum tube RF power amplifier and analyzes their effects on the transmitter bandwidth. Results of computer circuit analysis and actual measured bandwidth data of two different topologies are compared. Design considerations for high quality transmitter performance and desired level of transparency to a wideband FM broadcast signal is discussed including the effects of RF power amplifier tuning on the FM modulation performance. The contents of the paper is divided into the following main headings:

1. Frequency Modulated Signal And Effects Of Bandwidth Limitation On The Transmitter Performance.
2. RF Power Amplifier Design Considerations
   - Primary Design Factors
   - Input Circuit Configurations And Their Effects On The Transmitter Amplitude And Group Delay Responses
   - Output Circuit Configurations And Their Effects On The Transmitter Amplitude And Group Delay Responses
   - Computed/Measured Amplitude And Group Delay Responses.
3. Modulation Performance Of A Typical 20 kW Single-Tube FM Transmitter.
4. Effects Of Power Amplifier Tuning On FM Modulation Performance.

FREQUENCY MODULATED SIGNAL AND EFFECTS OF BANDWIDTH LIMITATION ON THE TRANSMITTER PERFORMANCE

Frequency Modulated Signal

A Frequency Modulated RF Signal with modulation index "m", carrier frequency "fc", and single-tone modulation frequency "fm" can be represented by the following mathematical expression [1-6]:

$$E(t) = Ec \cdot Cos[\omega c \cdot t + m \cdot Sin(\omega m \cdot t)],$$

Where:

Ec = The unmodulated carrier amplitude constant
$\omega c = 2\pi \cdot fc$ (carrier frequency)
$\omega m = 2\pi \cdot fm$ (modulating frequency)
$m = \Delta f/fm$ = frequency deviation/modulating frequency

In an FM signal, the deviation "Δf" of the instantaneous frequency from the average (or the carrier frequency) is directly proportional to the instantaneous amplitude of the modulating signal. The rate of frequency deviation is the modulating signal frequency.

The above FM Signal can be expressed as an infinite series of discrete spectral components using trigonometric expansions and series representations of Bessel functions [1-5].

$$E(t) = Ec \cdot \sum_{n=-\infty}^{\infty} Jn(m) \cdot Cos[(\omega c + n \cdot \omega m) \cdot t],$$

Where Jn(m) are Bessel functions of the first kind and nth order with argument "m". The numeric values of the Bessel functions Jn(m) for different "n" express the amplitudes of the various frequency components relative to the unmodulated carrier amplitude. The values of Jn(m) depend on the argument "m" and the order "n". These can be found from the mathematical tables.

For example:

If n = 0, Jn(m) = Jo(m), which is the amplitude of the carrier component.

If n = ±1, Jn(m) = J1(m) and J-1(m), which are the amplitudes of the first order sideband components.

If n = ±2, Jn(m) = J2(m) and J-2(m), which are the amplitudes of the second order sideband components.

Figure 1 shows a graphical representation of how the Bessel function values for the carrier and the first twelve pairs of sidebands vary with the modulation index [1-4]. Bessel function value diminishes as the order of the function "n" gets larger. As the modulation index varies, the carrier or the sideband frequency pair may vanish completely. Representative spectrum plots for six different values of modulation index are shown in Figures 2 and 3 [4]. In FM the energy that goes into the sideband frequencies is taken from the carrier; the total power in the overall composite signal remains the same regardless of the modulation index.

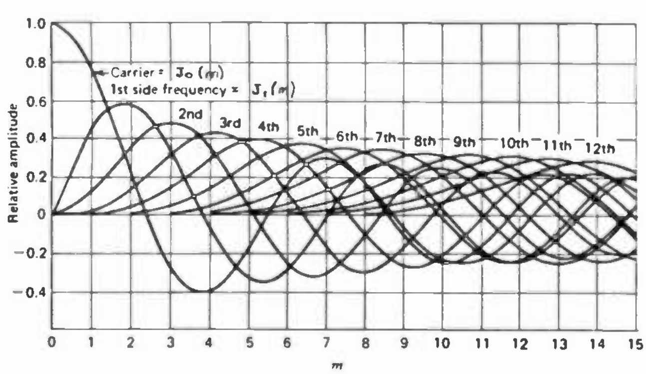


FIGURE 1. PLOT OF BESSEL FUNCTION OF FIRST KIND AS A FUNCTION OF ARGUMENT m.

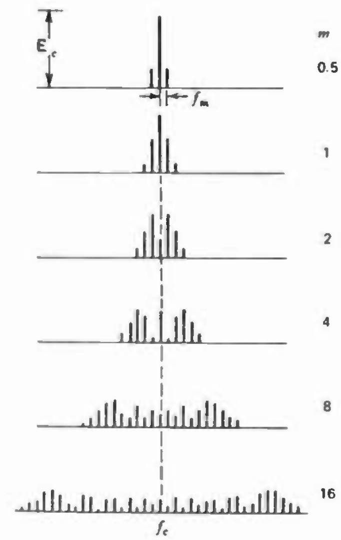


FIGURE 2. FREQUENCY SPECTRUM PLOTS FOR SINUSOIDAL MODULATION, WITH VARIOUS VALUES OF "m" ("fm" IS CONSTANT AND "Δf" IS VARIED).

Occupied Signal Bandwidth

Occupied Signal Bandwidth "BW" of an FM signal can be calculated for a single tone modulation by the following formula:

$$BW = 2 \cdot n \cdot fm.$$

Where "n" is the number of significant sideband components which depends on the value of Jn(m) and changes with the modulation index "m". "fm" is the modulating frequency.

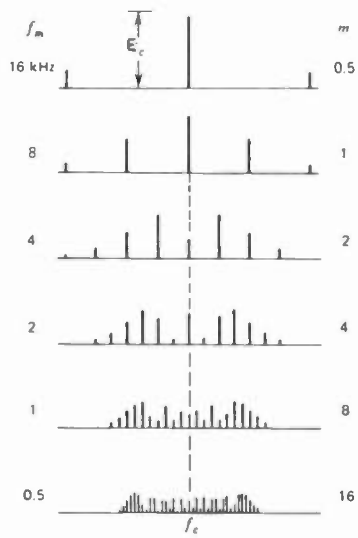


FIGURE 3. FREQUENCY SPECTRUM PLOTS FOR SINUSOIDAL MODULATION, WITH VARIOUS VALUES OF "m" ("Δf" IS CONSTANT AND "fm" IS VARIED).

The value of "n" can be accurately found from the mathematical tables by ignoring sideband components with amplitudes Jn(m) less than a certain desired number. The maximum value of "n" which need be considered for a given "m" may be found from the following empirical expression [5]:

$$n = m + k \cdot (m)^{0.27}$$

where "k" is 2.4 for Jn(m) = 0.01, and 3.5 for Jn(m) = 0.001.

For a single tone 15 kHz modulation with 75 kHz deviation, the modulation index is "5". If we ignore components with amplitudes less than 1% (Jn(5)<0.01), the number "n" is 9. The bandwidth required is 270 kHz.

The signal bandwidth for stereo Left or Right only single tone 15 kHz modulation is typically less than that for monaural modulation. This is due to the reduction in modulation index. The frequency deviation is held constant at 75 kHz but the composite baseband spectral components comprise of modulation frequencies at 15 kHz ("L + R" Main Channel), 19 kHz (Pilot), 23 kHz and 53 kHz ("L - R" Subcarrier Channel). The bandwidth calculation for multiple tone or stereo FM modulation is quite complex because several combination frequencies must be accounted for. This is caused by the nonlinear process inherent to frequency modulation.

Effects Of Bandwidth Limitations

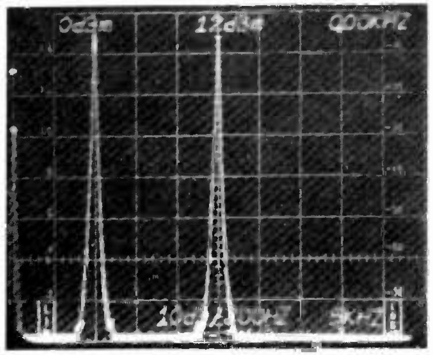
Typically, bandwidth limitations occur when the FM signal passes through the RF path comprising the transmitter, filterplexer/combiner, antenna system and the particularly in the receiver.

Figure 4 shows the effect of passing a twin-tone (10 kHz/25 kHz) modulated FM signal through a wideband and a narrowband (-3 dB bandwidth of 400 kHz) RF path [7]. The top three pictures show that the wideband RF path has negligible effect on the demodulated audio signal. But the bottom three pictures show the distortion products caused by bandwidth restriction.

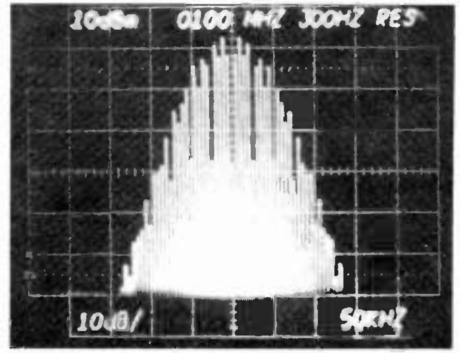
Effects On The Transmitter Modulation Performance

The following is a summary of results based on a study done by Broadcast Electronics to determine how much bandwidth is required for low distortion FM transmission, and at what bandwidth the point of diminishing return regarding the audio performance improvement is reached. This information has been published by Broadcast Electronics [8].

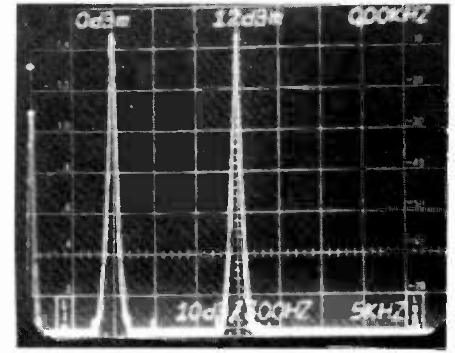
TWO TONE (10KHz & 25KHz) MODULATION THRU WIDEBAND RF PATH



BASEBAND SPECTRUM TO BEI MODEL FX-30 MODULATOR

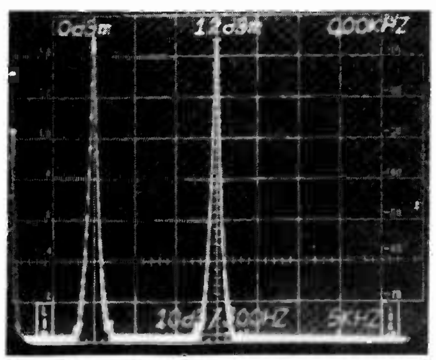


RF SPECTRUM TO DEMODULATOR

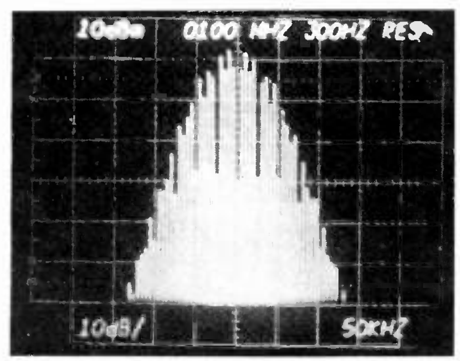


DEMODULATED BASEBAND SPECTRUM FROM BOONTON MODEL 82AD

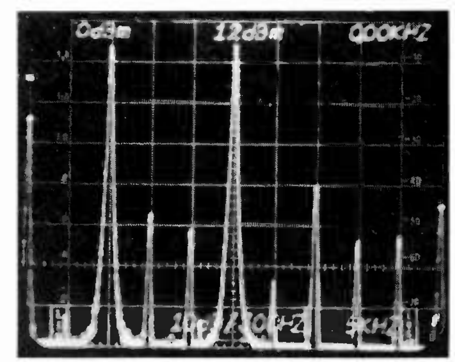
TWO TONE (10KHz & 25KHz) MODULATION THRU NARROWBAND RF PATH



BASEBAND SPECTRUM TO BEI MODEL FX-30 MODULATOR



BANDWIDTH LIMITED RF SPECTRUM TO DEMODULATOR



DEMODULATED BASEBAND SPECTRUM FROM BOONTON MODEL 82AD

FIGURE 4. THE EFFECT OF PASSING A TWIN-TONE (10 kHz/25 kHz) MODULATED FM SIGNAL THROUGH A WIDEBAND AND A NARROWBAND RF PATH.

Composite Total Harmonic Distortion And Noise (THD+N) At 15 kHz

THD is better than 0.1% with more than 600 kHz bandwidth. There is no improvement in performance above 1.5 MHz as shown in Figure 5.

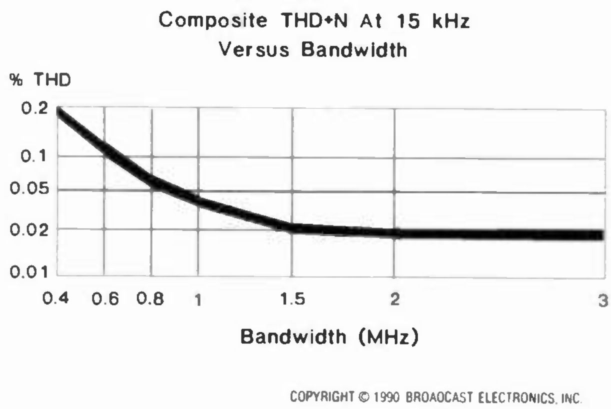


FIGURE 5. COMPOSITE THD+N AT 15 kHz VERSUS BANDWIDTH

Composite SMPTE Intermodulation Distortion

SMPTE IMD (60Hz/7kHz 1:1 Ratio) is better than 0.1% with 400 kHz bandwidth and crosses 0.01% with 1 MHz bandwidth as shown in Figure 6.

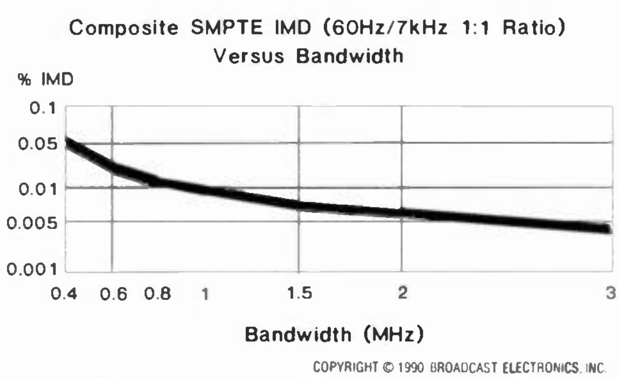


FIGURE 6. COMPOSITE SMPTE IMD (60 Hz/7 kHz 1:1 RATIO) VERSUS BANDWIDTH

Stereophonic Total Harmonic Distortion And Noise (THD+N) At 15 kHz

Stereophonic THD at 15 kHz modulation is better than 0.1% with 600 kHz bandwidth. There is no improvement above 1 MHz as shown in Figure 7.

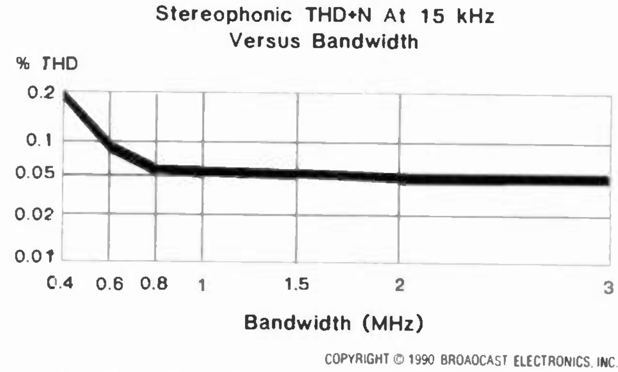


FIGURE 7. STEREOPHONIC THD+N AT 15 kHz VERSUS BANDWIDTH

Stereophonic SMPTE Intermodulation Distortion

Figure 8 shows that the Stereophonic IMD is better than 0.05% with 600 kHz bandwidth. There is very little improvement above 1 MHz.

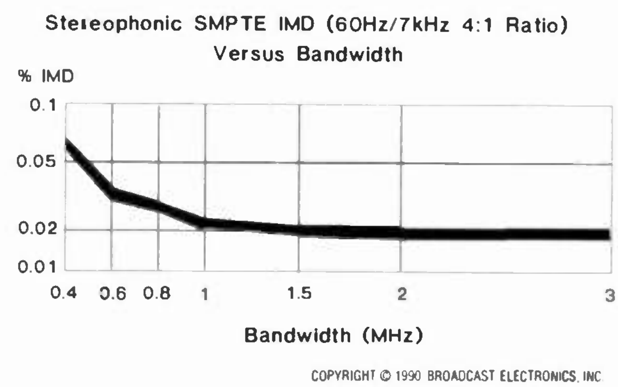


FIGURE 8. STEREOPHONIC SMPTE IMD (60 Hz/7 kHz 4:1 RATIO) VERSUS BANDWIDTH

Stereo Separation

Stereo Separation is about 40 dB with 400 kHz bandwidth, 50 dB with 700 kHz and 60 dB with 1.5 MHz as shown in Figure 9.

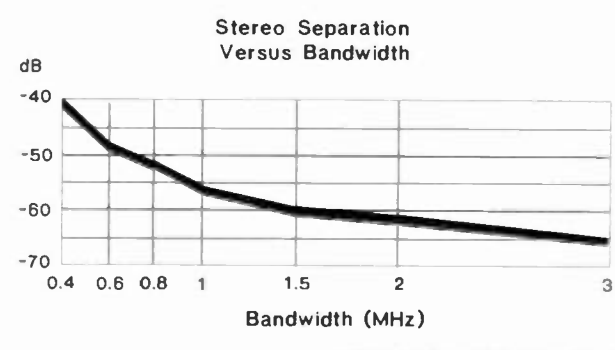


FIGURE 9. STEREO SEPARATION VERSUS BANDWIDTH

Synchronous Amplitude Modulation

Figure 10 shows that Synchronous AM is better than 40 dB with 800 kHz bandwidth and better than 50 dB with 1.5 MHz.

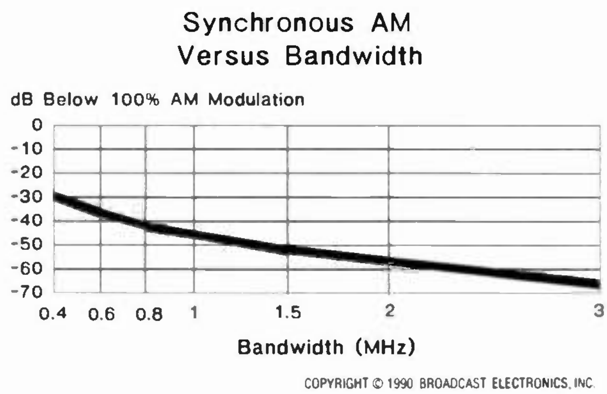


FIGURE 10. SYNCHRONOUS AM VERSUS BANDWIDTH

The conclusion drawn from the above study is that a minimum -3 dB bandwidth of 800 kHz is required for good audio performance and that excellent performance can be achieved with 1 to 1.5 MHz bandwidth [8].

Effects On The Transmitter RF Intermodulation

Frequency Spectrum is a very limited natural resource. The FM broadcast band is shared by several users at the same location. When multiple signals are present, any non-linear device such as tube in the transmitter power amplifier will generate RF intermodulation products due to mixing of these multiple signals. This mixing will have some conversion loss called "turn-around-loss". The degree of intermodulation interference generated within a given system can be accurately predicted when the turn-around-loss of the transmitter is available.

Turn-Around-Loss depends on three factors [9].

- In-band Conversion Loss
- Interfering Signal Attenuation due to PA Output Selectivity
- Attenuation of Resulting IM products due to PA Output Selectivity.

The transmitter with a narrower bandwidth will have higher selectivity thereby making it more immune to RF intermodulation. Therefore, there is certainly a trade-off between modulation performance and immunity from RF intermodulation.

RF POWER AMPLIFIER DESIGN CONSIDERATIONS

Primary Design Factors

The primary factors which should be considered in Power Amplifier design are:

- Desired Power Output
- Optimum Modulation Performance
- High Efficiency and Reliability
- Best Value for the Cost

This paper will focus its discussion on the second item - the design considerations necessary to achieve the optimum modulation performance.

RF Power Amplifier Bandwidth

The transmitter power amplifier bandwidth affects the modulation performance. Available bandwidth determines the amplitude response, phase response, and group delay response. There is a trade-off involved between the bandwidth, gain, and efficiency in the design of a power amplifier.

RF power amplifier bandwidth is restricted by the equivalent load resistance across parallel tuned circuits. Tuned circuits are necessary to cancel low reactive impedance presented by relatively high input and output capacitances of the amplifying device such as a vacuum tube.

The bandwidth for a single tuned circuit is proportional to the ratio of capacitive reactance, Xc to load resistance, R_L (appearing across the tuned circuit) [10]:

$$BW \simeq \frac{K}{2\pi \cdot fc \cdot C \cdot R_L} \simeq \frac{K(Xc)}{R_L}$$

Where: BW = bandwidth between half-power (or -3 dB) points
K = proportionality constant
R_L = load resistance (appearing across tuned circuit)
C = total capacitance of tuned circuit (includes stray capacitances and output or input capacitances of the tube)
Xc = capacitive reactance of C
fc = carrier frequency

The RF voltage swing across the tuned circuit also depends on the load resistance. For the same power and efficiency, the bandwidth can be increased if the capacitance is reduced.

Input Circuit Configurations And Their Effects On The Transmitter Amplitude And Group Delay Responses

Newer transmitter designs utilize solid-state intermediate power amplifiers to provide necessary RF drive level to operate the tube in the class C mode. The output load impedance is typically 50 Ohms. It is, therefore, necessary to design a matching network to transform a high grid input impedance to 50 Ohms at the PA input. The following three types of input matching circuit configurations are used:

- Single Element Capacitive (C) Input Match
- Single Element Inductive (L) Input Match
- Broadband (L-C) Input Match.

The first two are the popular ones. These matching circuits have different effects on the PA amplitude and group delay responses.

Capacitive Input Match

Single Element Capacitive Input Matching Circuit is shown in Figure 11 [6,10]. This is the simplest in design as well as implementation. Variable inductor "Lin" tunes out the tube input capacitance past parallel resonance to make the input impedance slightly inductive. This is necessary to transform the high grid impedance to an equivalent series 50 Ohms resistance and some inductive reactance. The reactance is then tuned out with a series variable capacitor, "Cs". Interactive adjustment of "Lin" and "Cs" is required. This configuration has the characteristics of a high pass filter.

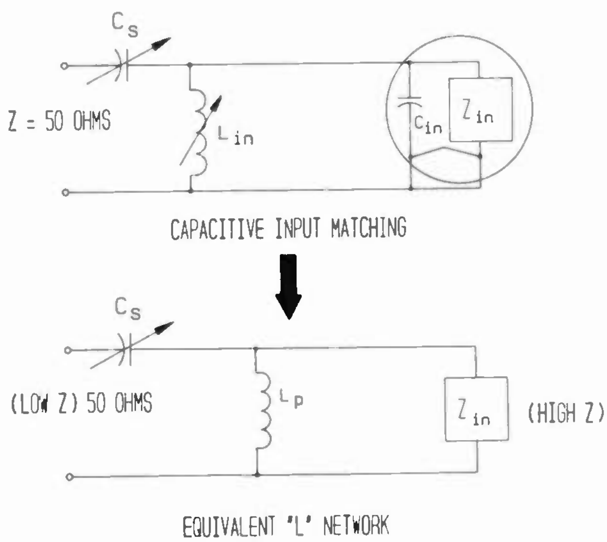


FIGURE 11. CAPACITIVE INPUT MATCHING CIRCUIT

Inductive Input Match

Single Element Inductive Input Matching Circuit is shown in Figure 12 [6,10]. This is the next most popular method for input matching. The variable inductor "Lin" is used to tune out the input capacitance slightly before the parallel resonance is reached, on the capacitive side. This is necessary to transform the high grid impedance to an equivalent series 50 Ohms resistance with some capacitive reactance. The reactance is then tuned out with a series variable inductor "Ls". Interactive adjustment of "Lin" and "Ls" is required. In this circuit configuration, a part of the input capacitance is used for impedance transformation and the equivalent capacitance across the tuned circuit becomes less, thereby increasing the bandwidth. This configuration has the characteristics of a low pass filter.

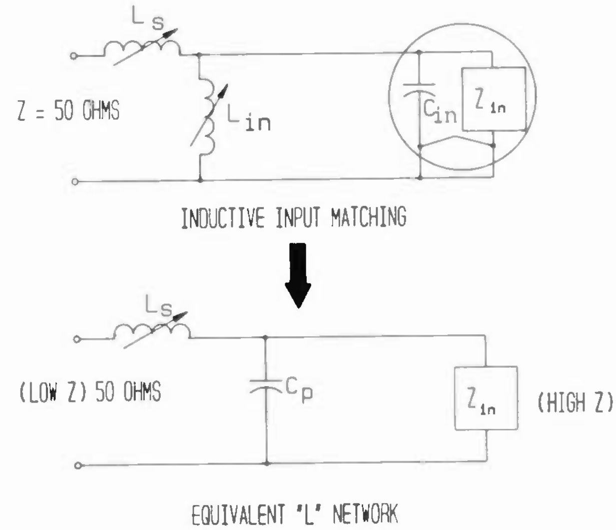


FIGURE 12. INDUCTIVE INPUT MATCHING CIRCUIT

Broadband L-C Input Match

Broadband L-C Input Matching Circuit is shown in Figure 13 [10]. This is an extension of L-Match circuit. It utilizes multiple L-C sections with each section providing a small step in the total impedance transformation. This technique provides a broadband impedance match without interactive adjustment and improves the transmitter operation. This configuration also utilizes a part of the input capacitance for impedance transformation and has the characteristics of a multiple section low pass filter.

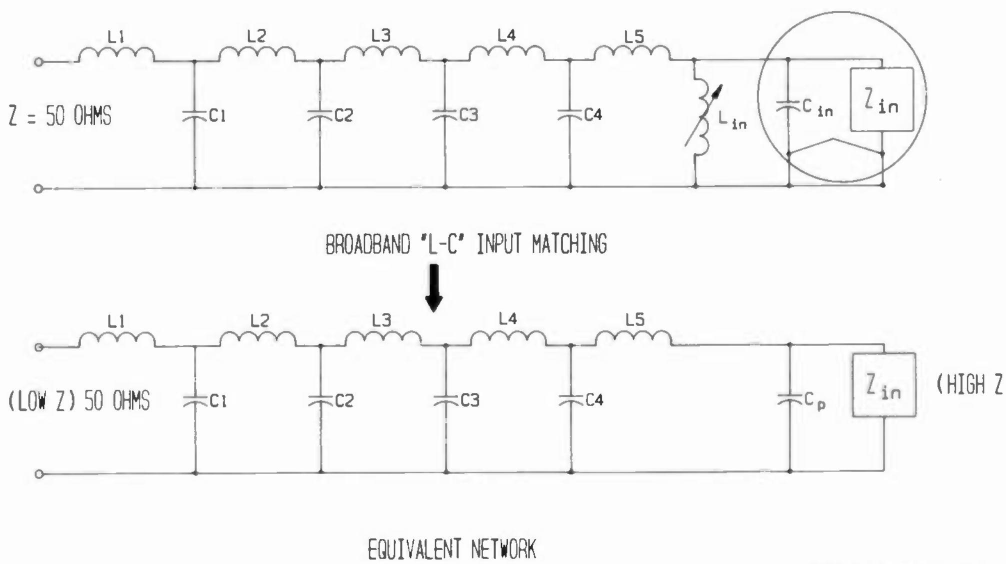


FIGURE 13. BROADBAND L-C INPUT MATCHING CIRCUIT

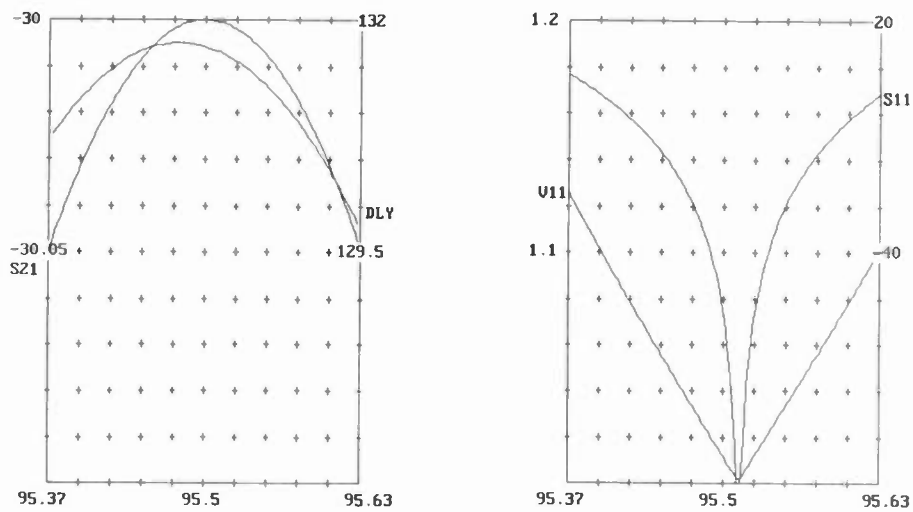
Computed And Measured Amplitude/Group Delay Responses Of Capacitive And Broadband Input Matching Circuits

The results of computer analysis and actual measurements made with two different input circuit configurations, C-Match and Broadband L-C Match, in a real transmitter operating at 20 kW RF power are presented below.

Two input matching circuits were designed to transform the high impedance of PA input tuned circuit to 50 Ohms resistive impedance. For the purpose of illustration it was assumed that a total capacitance of 165 picofarads in parallel with 375 Ohms load resistance appeared across the tuned circuit. This represents the input impedance of a typical 20 kW FM transmitter power amplifier input circuit utilizing an Eimac 8989 tetrode.

Computer software programs ECA [11] and =SuperStar= [12] were used to analyze the circuit designs and obtain the response plots. Hewlett Packard Network Analyzer Model 3577A was used to obtain the measured response plots.

Figure 14A shows the computed amplitude response (S21) and group delay (DLY) response plots for capacitive input matching circuit. Figure 14B shows the input VSWR (V11) and input return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots do not coincide in the case of capacitive input matching circuit. Figure 15 shows the measured amplitude response, group delay response and input return loss plots of a typical 20 kW FM transmitter at the tube grid ring for capacitive input matching circuit.



14A. AMPLITUDE (S21) AND GROUP DELAY (DLY) RESPONSES

14B. INPUT VSWR (V11) AND RETURN LOSS (S11) RESPONSES

FIGURE 14. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A CAPACITIVE INPUT MATCHING CIRCUIT

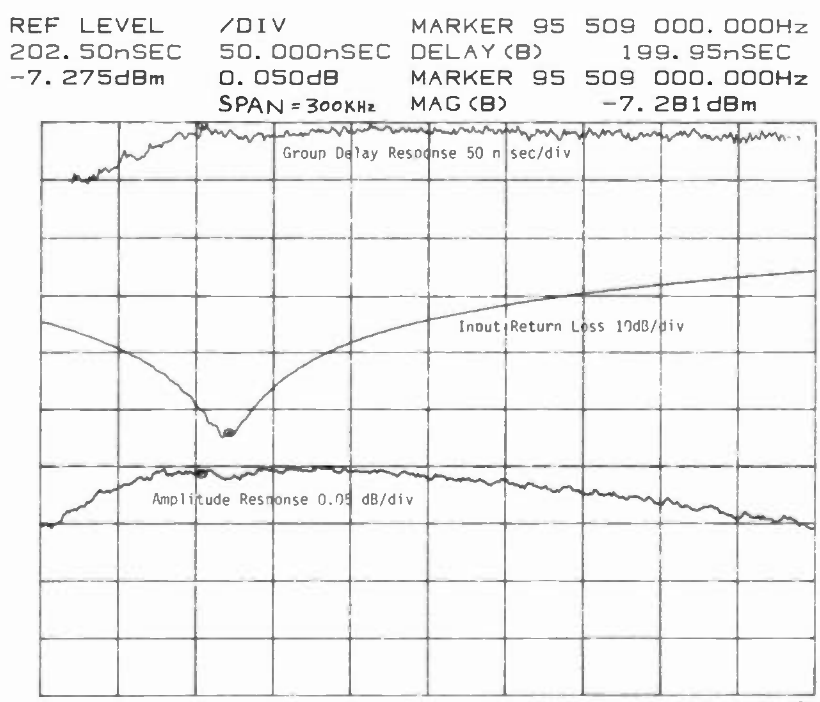
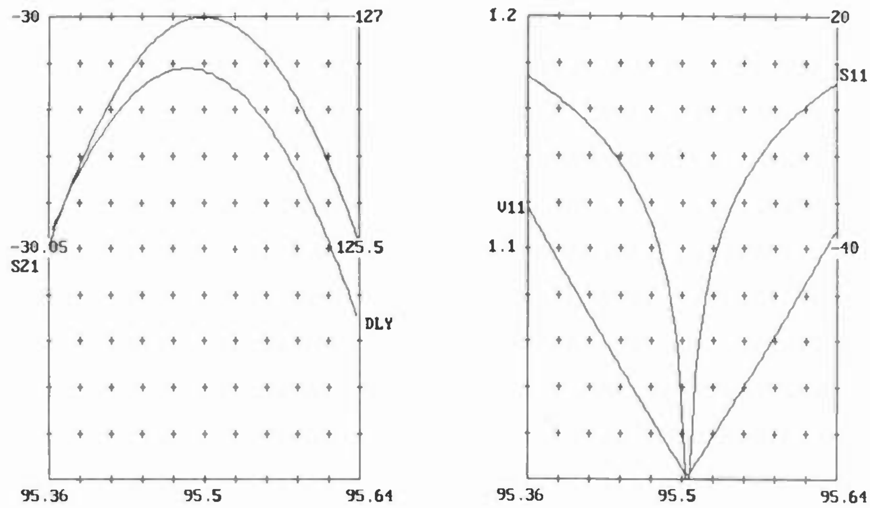


FIGURE 15. MEASURED AMPLITUDE, GROUP DELAY, AND INPUT RETURN LOSS RESPONSES OF A CAPACITIVE INPUT MATCHING CIRCUIT IN A 20 kW TRANSMITTER

Figure 16A shows the computed amplitude response (S21) and group delay (DLY) response plots for broadband input matching circuit. Figure 16B shows the input VSWR (V11) and input return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots coincide closely to provide a symmetrical group delay response.

Figure 17 shows the measured amplitude response, group delay response and input return loss plots of a typical 20 kW FM transmitter at the tube grid ring for broadband input matching circuit.



16A. AMPLITUDE (S21) AND GROUP DELAY (DLY) RESPONSES

16B. INPUT VSWR (V11) AND RETURN LOSS (S11) RESPONSES

FIGURE 16. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A BROADBAND L-C INPUT MATCHING CIRCUIT

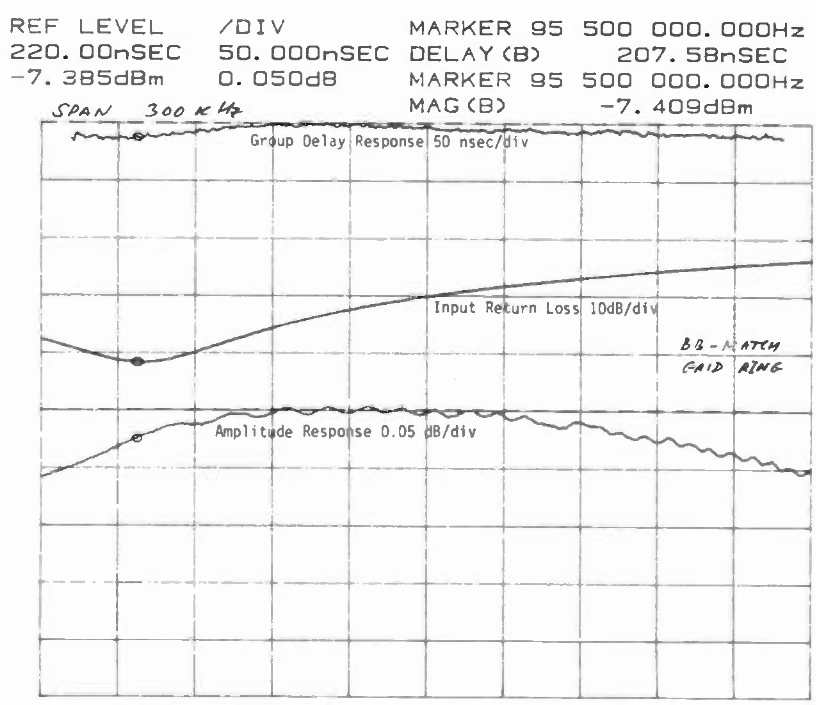


FIGURE 17. MEASURED AMPLITUDE, GROUP DELAY, AND INPUT RETURN LOSS RESPONSES OF A BROADBAND INPUT MATCHING CIRCUIT IN A 20 kW TRANSMITTER

Figure 18 shows the comparison of computed and measured bandwidth of a typical 20 kW FM transmitter at tube grid ring. Broadband L-C input matching circuit has a higher bandwidth than the capacitive input matching circuit.

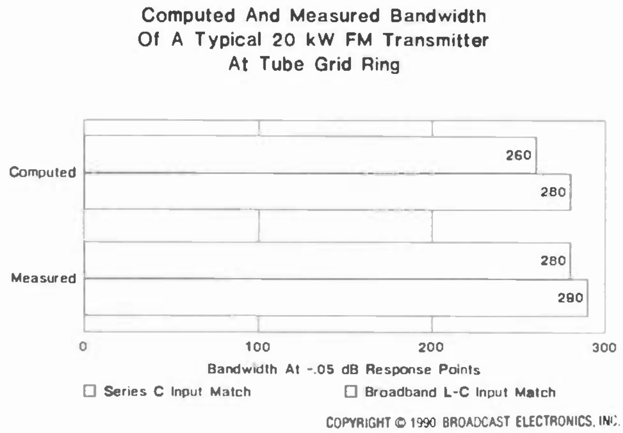


FIGURE 18. COMPUTED AND MEASURED BANDWIDTH OF A TYPICAL 20 kW FM TRANSMITTER AT TUBE GRID RING

Figure 19 shows the comparison of computed and measured asymmetry of amplitude/delay response peaks of a typical 20 kW FM transmitter at tube grid ring. Capacitive input matching circuit has a greater group delay asymmetry than either the inductive or broadband L-C matching circuit.

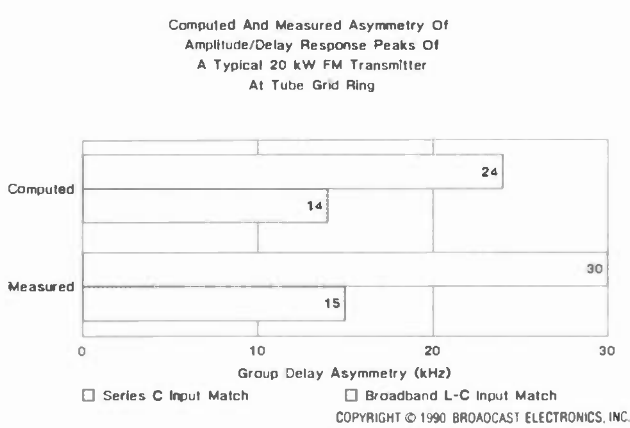


FIGURE 19. COMPUTED AND MEASURED ASYMMETRY OF AMPLITUDE/DELAY RESPONSE PEAKS OF A TYPICAL 20 kW FM TRANSMITTER AT TUBE GRID RING

Output Circuit Configurations And Their Effects On The Transmitter Amplitude And Group Delay Responses

Tuned circuits are required to resonate the output capacitance of the vacuum tube and to present a fairly high impedance between anode and cathode (screen grid in case of tetrodes) to the fundamental carrier frequency component and present a low impedance to the harmonics.

The output circuit may utilize a lumped inductor, a strip line, or higher "Q" (low loss) transmission line section to resonate the tube capacitance.

In newer design transmitters, the tube power amplifier is typically constructed in a cavity enclosure utilizing (larger physical size) coaxial transmission line section of either quarter-wavelength, or half-wavelength to increase the unloaded "Q" and minimize losses.

The Power Amplifier efficiency depends on the RF plate voltage swing developed across the load resistance, the plate current conduction angle and the cavity efficiency. The PA cavity efficiency is related to the ratio of loaded to unloaded "Q" as follows [10]:

$$N = 1 - \frac{QL}{QU} \times 100$$

Where: N = efficiency in percent
QL = loaded "Q" of cavity
QU = unloaded "Q" of cavity

The loaded "Q" is dependent on the equivalent plate load resistance presented across the tuned circuit and output circuit capacitance. Unloaded "Q" depends on the cavity volume and the RF resistivity of the conductors due to skin effects. A high unloaded "Q" is desirable, as is a low loaded "Q", for best efficiency. As the loaded "Q" goes up the bandwidth decreases. For a given tube output capacitance and power level, loaded "Q" decreases with decreasing plate voltage or with increasing plate current. The increase in bandwidth at reduced plate voltage occurs because the load resistance is directly related to the RF voltage swing on the tube element.

For the same power and efficiency, the bandwidth can also be increased if the output capacitance is reduced. Power tube selection and minimization of stray capacitance are areas of particular concern in PA design for maximum bandwidth. Bandwidth can be further improved by minimizing added tuning capacitance and by properly selecting the output coupling method.

The following are the two popular methods of output coupling circuits used:

- Series Capacitive Output Coupling
- Magnetic Output Coupling Loop.

These circuits have different effects on the PA amplitude and group delay responses.

Series Capacitive Output Coupling

Figure 20 shows a schematic of a tetrode power amplifier with a capacitive output tuning and capacitive output coupling circuit [6,10].

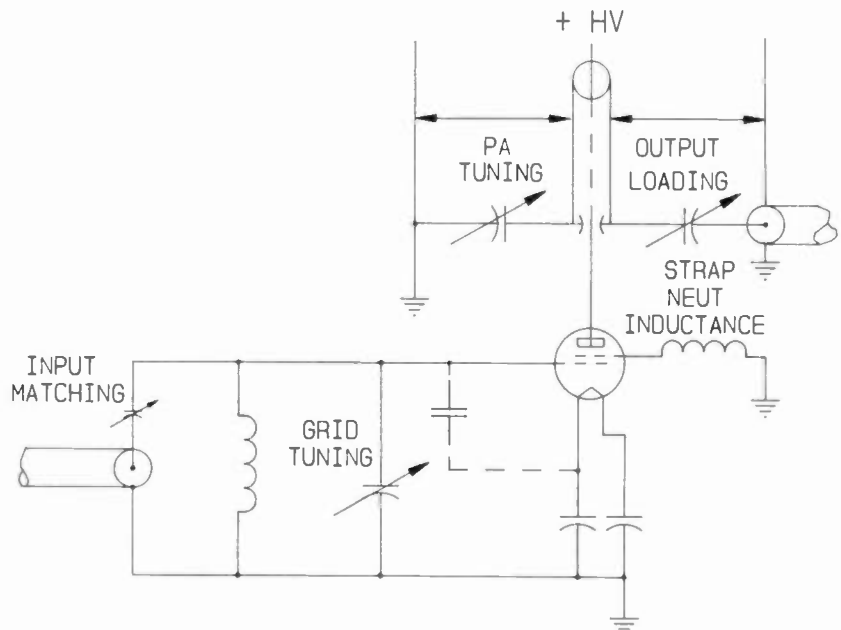


FIGURE 20. TETRODE POWER AMPLIFIER CIRCUIT WITH CAPACITIVE OUTPUT TUNING AND CAPACITIVE OUTPUT COUPLING

Figure 21 is a simplified diagram showing the construction of this type of output circuit in a quarter-wave length coaxial cavity [6,10]. The tube anode is coupled through a DC blocking capacitor to the transmission line. The tube's output capacitance is brought to resonance by the inductive component of the transmission line that is physically less than a quarter-wave length long. The coarse output tuning is done by moving the ground plane at low impedance end of the line. Variable capacitors are used to fine-tune the output resonant circuit and to couple the power from the high RF voltage point located at the anode end of the quarter-wave line to the load. An interactive adjustment of tuning and loading is required in this type of power amplifier cavity design. This type of output coupling is similar to a series capacitive input matching circuit discussed above.

A capacitive output coupling circuit was designed to transform the high impedance of the PA output tuned circuit (with total capacitance of 20 picofarads and 1590 Ohms load resistance across it) to 50 Ohms. This illustrates the output circuit of a typical 20 kW FM transmitter PA utilizing an Eimac 8989 tetrode. Computer software programs ECA [11] and =SuperStar= [12] were used to analyze the circuit design and obtain the response plots.

Figure 22A shows the computed amplitude response (S21) and group delay (DLY) response plots. Figure 22B shows the output VSWR (V11) and output return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots do not coincide in the case of capacitive output coupling circuit.

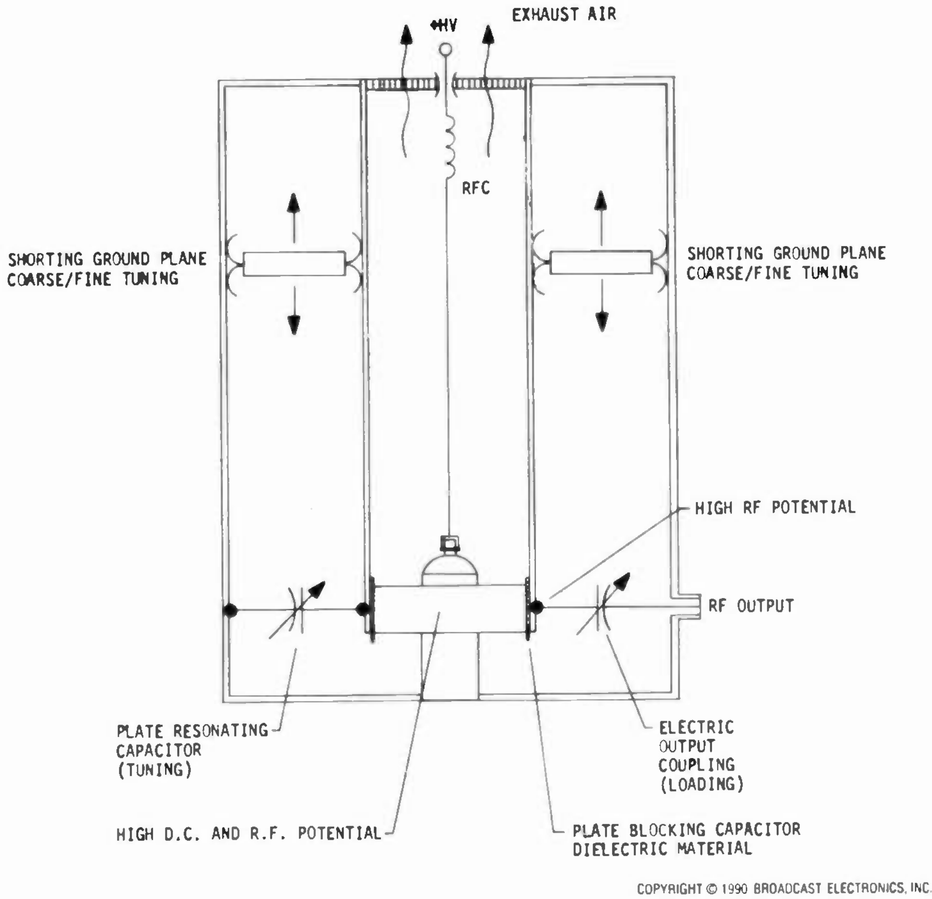
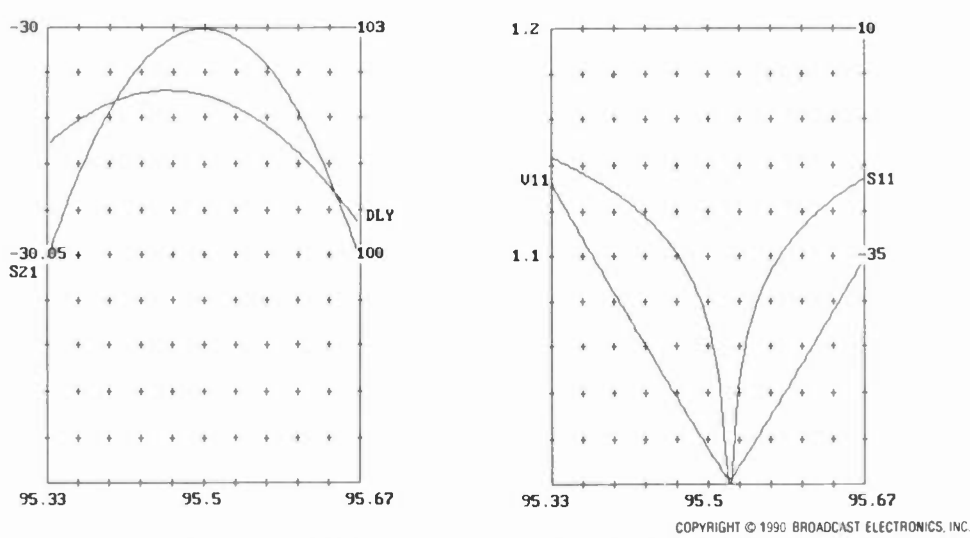


FIGURE 21. QUARTER-WAVE COAXIAL CAVITY WITH CAPACITIVE TUNING AND CAPACITIVE OUTPUT COUPLING



22A. AMPLITUDE (S21) AND GROUP DELAY (DLY) RESPONSES

22B. OUTPUT VSWR (V11) AND RETURN LOSS (S11) RESPONSES

FIGURE 22. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A QUARTER-WAVE CAVITY WITH CAPACITIVE OUTPUT COUPLING CIRCUIT

Magnetic Output Coupling Loop

Figure 23 shows the simplified diagram showing the construction of a magnetic output coupling loop circuit in a folded half-wave coaxial cavity with a secondary re-entrant line section [5,9]. The tube's output capacitance is brought to resonance by the inductive component of the transmission line that is physically less than a half-wave length long. The coarse output tuning is done by presetting the depth of the secondary line into the cavity. The PA output fine tuning is accomplished by varying the physical length of a flexible extension (bellows) on the end of the secondary transmission line. This type of output circuit does not require a plate blocking capacitor in the anode. The output power is coupled to the load by means of a magnetic loop positioned near the RF voltage null point located near the center of the primary transmission line where the RF current is maximum.

The coupling to the cavity varies as the square of the effective loop area and inversely as the square of the distance of the loop center from the cavity center axis. There is a minimal interaction between the tuning and loading adjustments in this type of power amplifier cavity design.

Furthermore, this type of cavity has a greater turn-around-loss due to its higher selectivity and provides better protection against RF interference. An additional 10 dB protection is available for an intermodulation product due to an interfering signal frequency separated from the carrier by 4 MHz.

A magnetic output coupling loop circuit was designed to transform the high impedance of the PA output tuned circuit (with total capacitance of 20 picofarads and 1590 Ohms load resistance across it) to 50 Ohms. This illustrates the output circuit of a typical 20 kW FM transmitter PA utilizing an Eimac 8989 tetrode. Computer software programs ECA [11] and =SuperStar= [12] were used to analyze the circuit design and obtain the response plots.

Figure 24A shows the computed amplitude response (S21) and group delay (DLY) response plots. Figure 24B shows the output VSWR (V11) and output return loss (S11) plots at -0.05 dB response points. The peaks of amplitude and group delay plots coincide in this case providing a symmetrical group delay response.

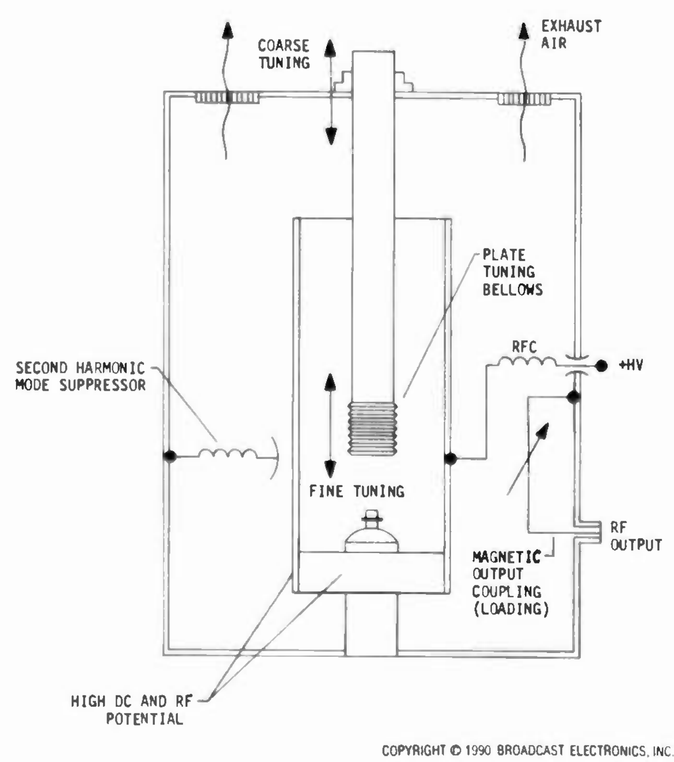
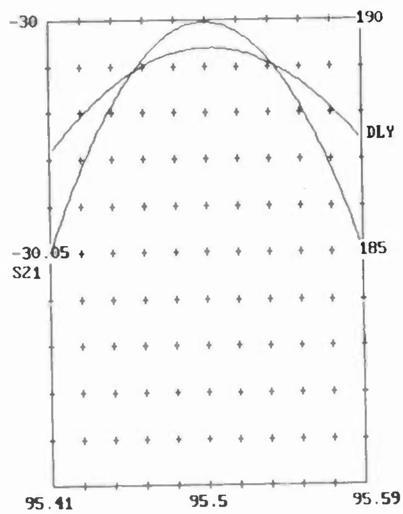
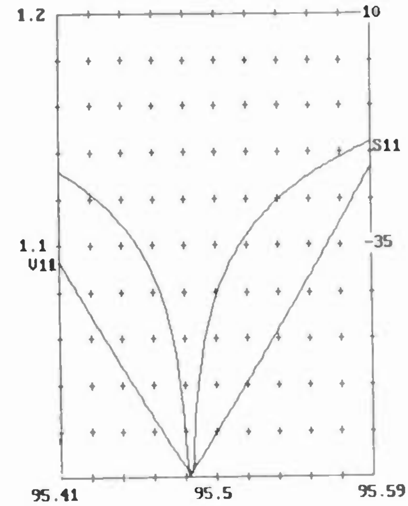


FIGURE 23. FOLDED HALF-WAVE COAXIAL CAVITY WITH INDUCTIVE TUNING AND MAGNETIC OUTPUT COUPLING LOOP



24A. AMPLITUDE (S21) AND GROUP DELAY (DLY) RESPONSES



24B. OUTPUT VSWR (V11) AND RETURN LOSS (S11) RESPONSES

FIGURE 24. COMPUTED AMPLITUDE, GROUP DELAY, VSWR, AND RETURN LOSS RESPONSES OF A FOLDED HALF-WAVE CAVITY WITH INDUCTIVE TUNING AND MAGNETIC OUTPUT COUPLING LOOP

MODULATION PERFORMANCE DATA OF A TYPICAL 20 KW SINGLE-TUBE FM TRANSMITTER

The following is the measured modulation performance data of Broadcast Electronics Model FM-20B 20 kW FM transmitter taken at the input and output circuits of the RF power amplifier. Three RF signal samples, one each from Broadcast Electronics Model FX-50 exciter output, the FM-20B PA tube grid ring, and the FM-20B PA output were taken to measure the modulation performance and to see the effects of input and output circuits on the transmitter performance.

The FM-20B transmitter was used to illustrate an example of a real world limited bandwidth RF tube power amplifier, designed for high quality performance. The RF power amplifier uses an Eimac 8989/4CX12,000A high gain tetrode tube in the folded half-wave cavity output circuit with magnetic coupling loop. The FM-20B also uses a broadband L-C matching circuit at the tube grid input to minimize signal degradation.

An RF sample from either the exciter output, or the PA tube grid ring, or the PA output was connected to the FMM-2. The deemphasized audio and wideband composite outputs were used for composite tests. The composite baseband was used to drive the FMS-2. The decoded left and right outputs of the FMS-2 were used for the stereo performance tests.

Figure 25 shows the test setup used to measure the modulation performance of the FM-20B transmitter. The Audio Precision System One test set audio oscillator fed either the FS-30 stereo generator for stereo performance tests, or the FX-50 composite input directly for baseband composite performance tests.

Figure 26 shows the MVDS (Microprocessor Video Diagnostics System) screen printout of FM-20B transmitter meter readings operating at 20 kW RF output power.

Measured Bandwidth Of FM-20B Transmitter

Figure 27 shows a plot of the measured overall amplitude response, group delay response and input return loss response of FM-20B transmitter. It shows a -3 dB bandwidth of 1.45 MHz.

FM-20B MODULATION PERFORMANCE TEST SETUP

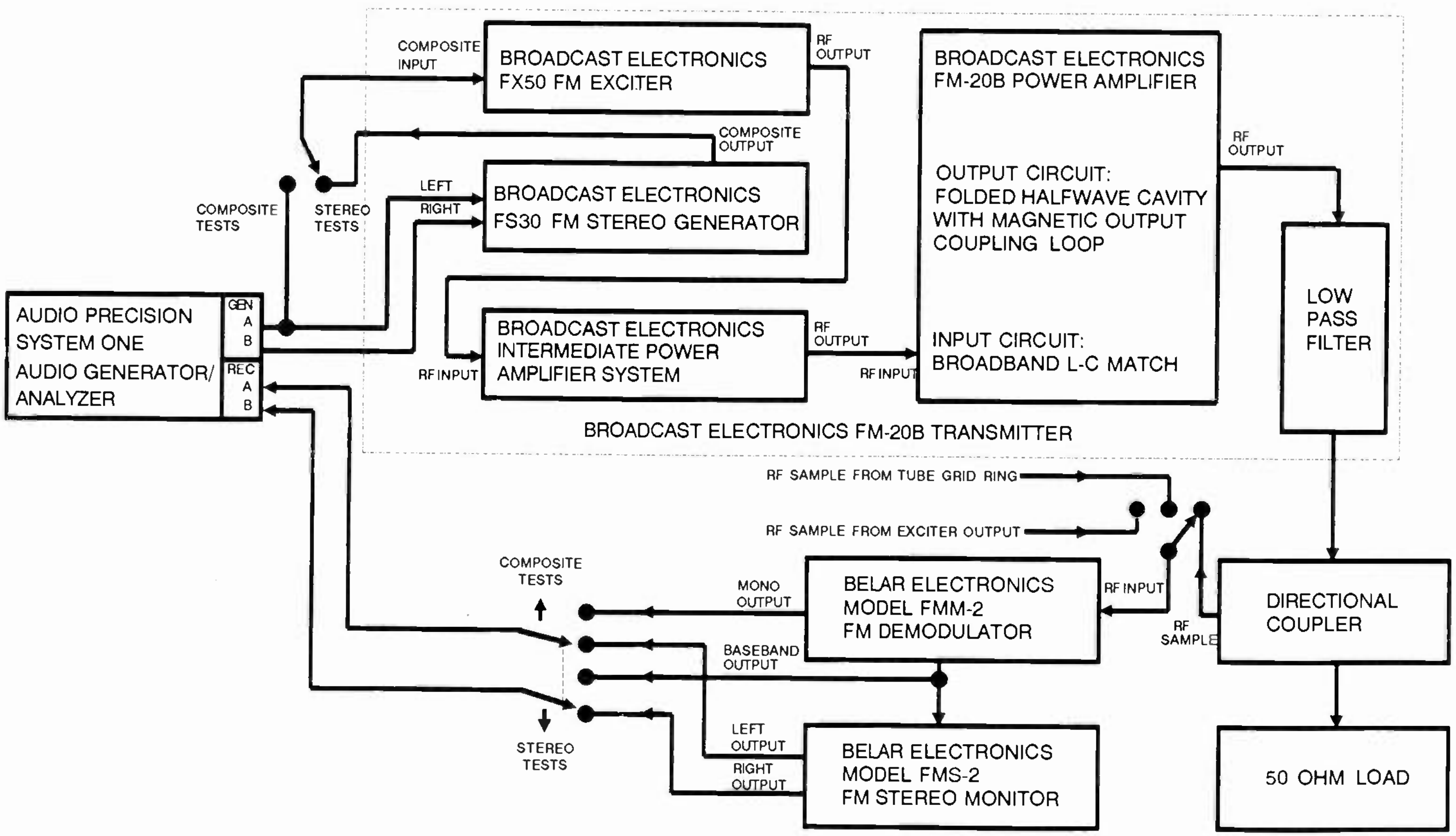


FIGURE 25. FM-20B MODULATION PERFORMANCE TEST SETUP

```
    FM-20B XMTR TEST 95.5 MHZ AT 20KW              TUE 09/05/89    09:45:37AM
    BROADCAST ELECTRONICS INCORPORATED    MODEL FM-20B      S/N 0000    VR2.2
--------------------------------------------------------------------------------
CONDITION: NORMAL OPERATION
--------------------------------------------------------------------------------
POWER AMPLIFIER (PA)    EFFICIENCY= 80%  | TRANSMITTER POWER OUTPUT
                                         |
                 PLATE    SCREEN    GRID | AUTHORIZED 20.00KW=100%       0.0KW ERP
VOLTAGE         8.94KV     710V    -298V | ACTUAL      20.00KW=100%       0.0KW ERP
CURRENT          2.81A    149MA     31MA |
POWER OUTPUT   20.00KW                   | REFLECTED    0.00KW=  0%
DISSIPATION     5.12KW     106W          | VSWR         1.0:1
--------------------------------------------------------------------------------
INTERMEDIATE POWER AMPLIFIER (IPA)       |
                    -1-        -2-       |
VOLTAGE            27.9V      28.0V      |
CURRENT            11.4A      11.3A      |
FORWARD POWER       204W       203W      |
REFLECTED POWER       2W         1W      |
DISSIPATION         114W       113W      |
TOTAL POWER        FWD= 367W  RFL=    3W |
---------------------------------------- | EXHAUST AIR TEMP=  46°C
EXCITER FORWARD POWER       36W          | APC ON
EXCITER REFLECTED POWER      1W          |
```


FIGURE 26. MVDS SCREEN PRINTOUT SHOWING METER READINGS OF FM-20B TRANSMITTER

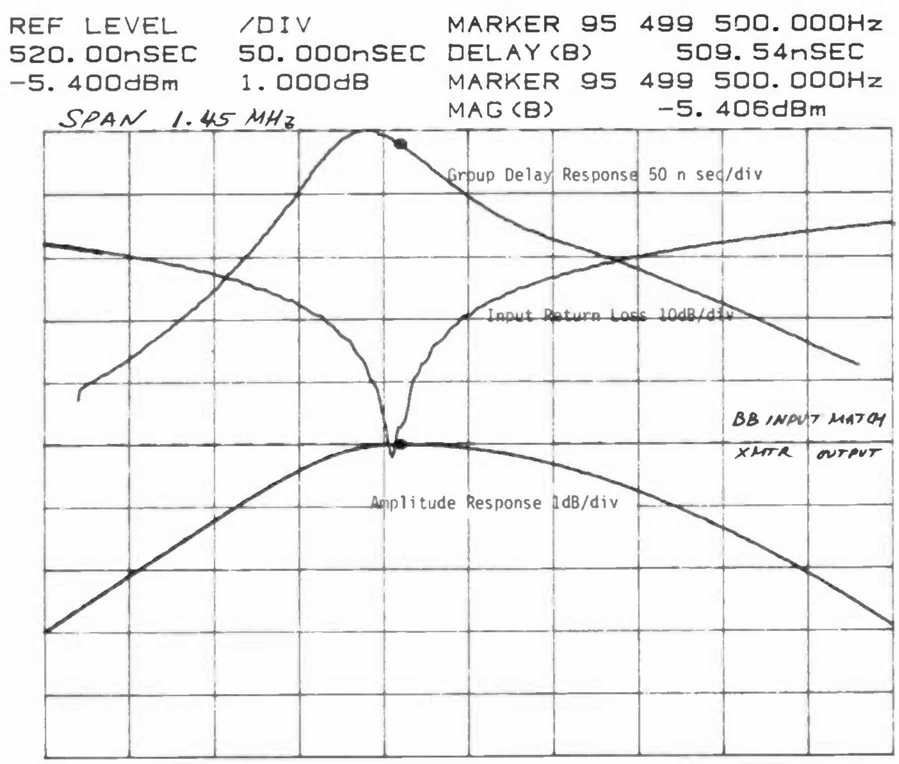


FIGURE 27. MEASURED AMPLITUDE, GROUP DELAY, AND INPUT RETURN LOSS RESPONSES OF FM-20B TRANSMITTER

FM-20B Synchronous AM

Figure 28 shows the measured synchronous AM (amplitude modulation) performance of the transmitter. The performance degradation is 11 dB from the exciter to the tube grid ring and 2 dB from the grid ring to the transmitter output.

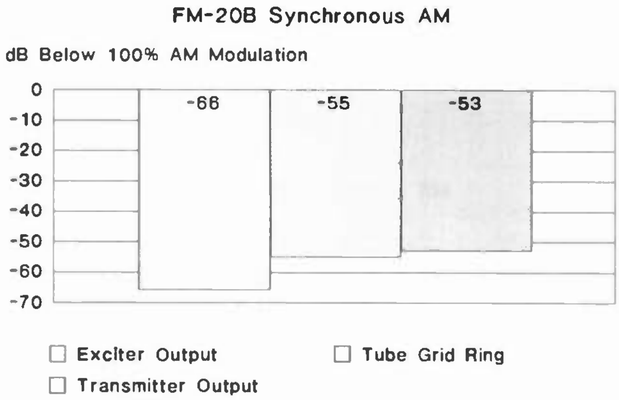


FIGURE 28. FM-20B SYNCHRONOUS AM

FM-20B Composite Amplitude Response At 100 kHz

Figure 29 shows the measured composite amplitude response performance degradation of 0.08 dB from the exciter to the tube grid ring and a further 0.08 dB from the grid ring to the transmitter output.

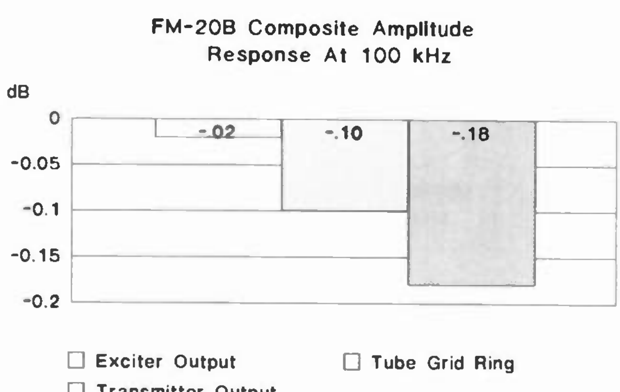


FIGURE 29. FM-20B COMPOSITE AMPLITUDE RESPONSE AT 100 kHz

FM-20B Composite THD+N At 15 kHz

Figure 30 shows an increase in THD+N at 15 kHz of 0.006% from the exciter to the tube grid ring and 0.052% from the grid to the transmitter output.

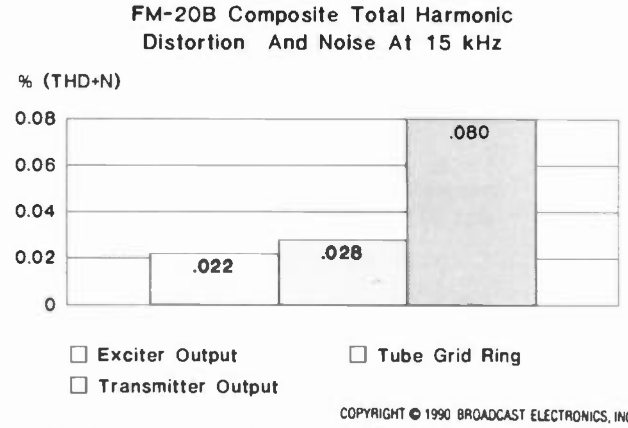


FIGURE 30. FM-20B COMPOSITE TOTAL HARMONIC DISTORTION AND NOISE AT 15 kHz

FM-20B Composite THD+N At 400 Hz

Figure 31 shows 0.001% increase in THD+N at 400 Hz from the exciter to the tube grid ring and no performance degradation from the grid ring to the transmitter output.

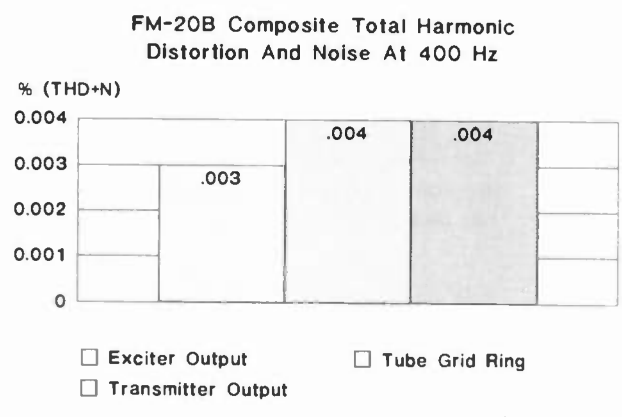


FIGURE 31. FM-20B COMPOSITE TOTAL HARMONIC DISTORTION AND NOISE AT 400 Hz

FM-20B Composite SMPTE IMD (60Hz/7kHz 1:1 Ratio)

Figure 32 shows the performance degradation of 0.002% IMD from the exciter to the tube grid ring but no change from the grid ring to the transmitter output.

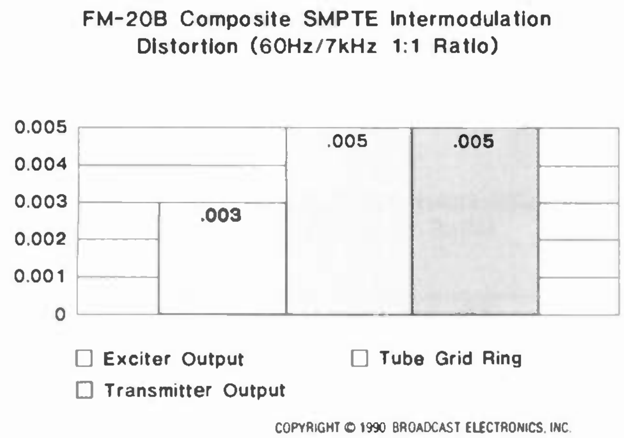


FIGURE 32. FM-20B COMPOSITE SMPTE INTERMODULATION DISTORTION (60 Hz/7 kHz 1:1 RATIO)

FM-20B Composite FM Signal-To-Noise Ratio

Figure 33 shows 1 dB drop in FM Signal-to-Noise ratio from the exciter to the tube grid ring and no degradation from the grid ring to the transmitter output. The small degradation may be caused by the changes in the input circuit and output circuit reactance due to mechanical vibration.

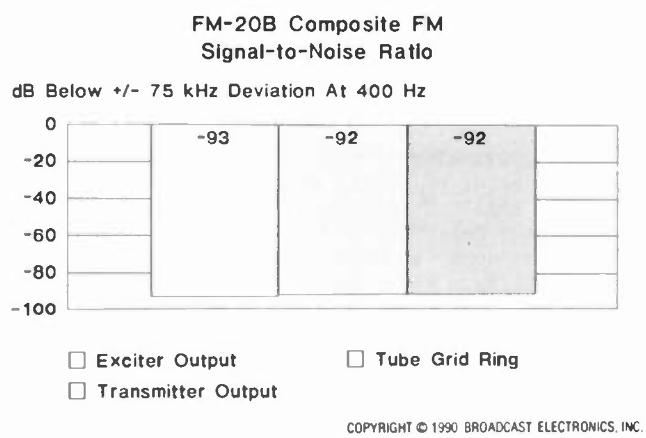


FIGURE 33. FM-20B COMPOSITE FM SIGNAL-TO-NOISE RATIO

FM-20B Stereo THD+N At 15 kHz

Figure 34 shows an increase in THD+N at 15 kHz of 0.011% from the exciter to the tube grid ring and 0.030% from the grid to the transmitter output.

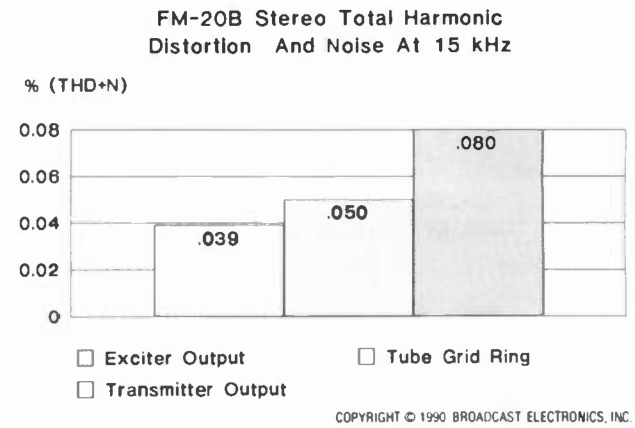


FIGURE 34. FM-20B STEREO TOTAL HARMONIC DISTORTION AND NOISE AT 15 kHz

FM-20B Stereo THD+N At 400 Hz

Figure 35 shows no performance degradation of THD+N at 400 Hz from the exciter to the tube grid ring as well as from the grid ring to the transmitter output.

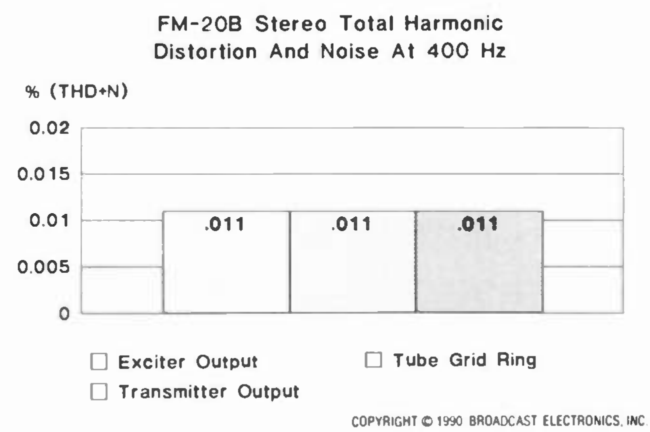


FIGURE 35. FM-20B STEREO TOTAL HARMONIC DISTORTION AND NOISE AT 400 Hz

FM-20B Stereo SMPTE IMD (60 Hz/7 kHz 4:1 Ratio)

Figure 36 shows the performance degradation of 0.012% stereo IMD from the exciter to the tube grid ring. There is actually an improvement of 0.009% stereo IMD performance from the grid ring to the transmitter output which may be caused by partial cancellation of distortion products in the left and right channels.

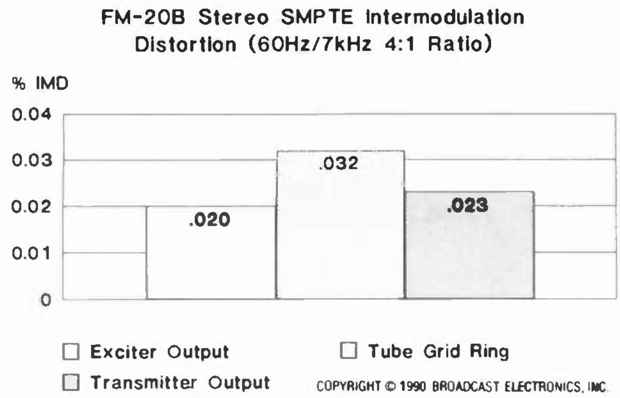


FIGURE 36. FM-20B STEREO SMPTE INTERMODULATION DISTORTION (60 Hz/7 kHz 4:1 RATIO)

FM-20B Stereo Separation At 15 kHz

Figure 37 shows a 6 dB drop in stereo separation from the exciter to the tube grid ring and a further 3 dB drop from the grid ring to the transmitter output.

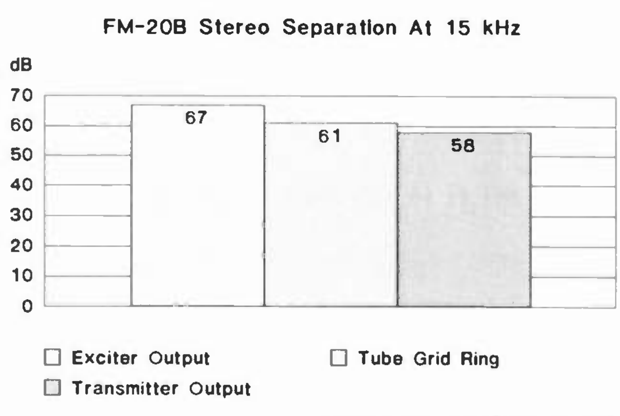


FIGURE 37. FM-20B STEREO SEPARATION AT 15 kHz

FM-20B Stereo Separation At 400 Hz

Figure 38 shows a 6 dB drop in stereo separation from the exciter to the tube grid ring and a further 3 dB drop from the grid ring to the transmitter output.

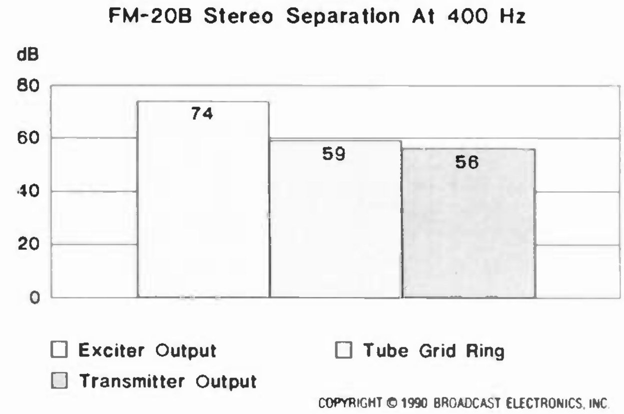


FIGURE 38. FM-20B STEREO SEPARATION AT 400 Hz

EFFECTS OF RF POWER AMPLIFIER TUNING ON FM MODULATION PERFORMANCE

In an FM broadcast transmitter, the RF power amplifier is typically adjusted for minimum synchronous AM (incidental amplitude modulation) which results in symmetrical amplitude response. This will assure that the transmitter's amplitude passband is properly centered on the FM channel. The upper and lower sidebands will be attenuated equally or symmetrically which is assumed to result in reduced modulation distortion. This will be true if the RF power amplifier circuit topology results in simultaneous symmetry of amplitude and group delay responses.

Actually, symmetry of the group delay response has a much greater effect on FM modulation distortion than the amplitude response. Tuning for symmetrical group delay will cause the phase errors to affect the upper and lower sidebands equally or symmetrically. Group delay response is constant if the phase shift versus frequency is linear. The signal is delayed in time, but no phase distortion occurs.

In most FM transmitters, the tuning points for symmetrical amplitude response and symmetrical group delay response do not coincide. Therefore, simply tuning for minimum synchronous AM (symmetrical amplitude response) does not necessarily result in best FM modulation performance.

Measurements taken on a typical 20 kW FM transmitter showed that tuning the RF power amplifier for symmetrical group delay response resulted in minimum distortion and crosstalk. It confirms that group delay response asymmetry causes higher FM modulation distortion and crosstalk than the amplitude response asymmetry. Therefore, RF power amplifier circuit topologies that exhibit coincidence of symmetrical amplitude and group delay responses will result in a better overall FM modulation performance.

Figure 39 shows a plot of the measured overall amplitude and group delay responses of a 20 kW FM transmitter at -0.05 dB response points with the power amplifier tuned for minimum synchronous AM.

Figure 40 shows a plot of demodulated composite baseband spectrum with 4.5 kHz single channel stereo modulation (plot obtained using Hewlett Packard Model 3585B spectrum analyzer).

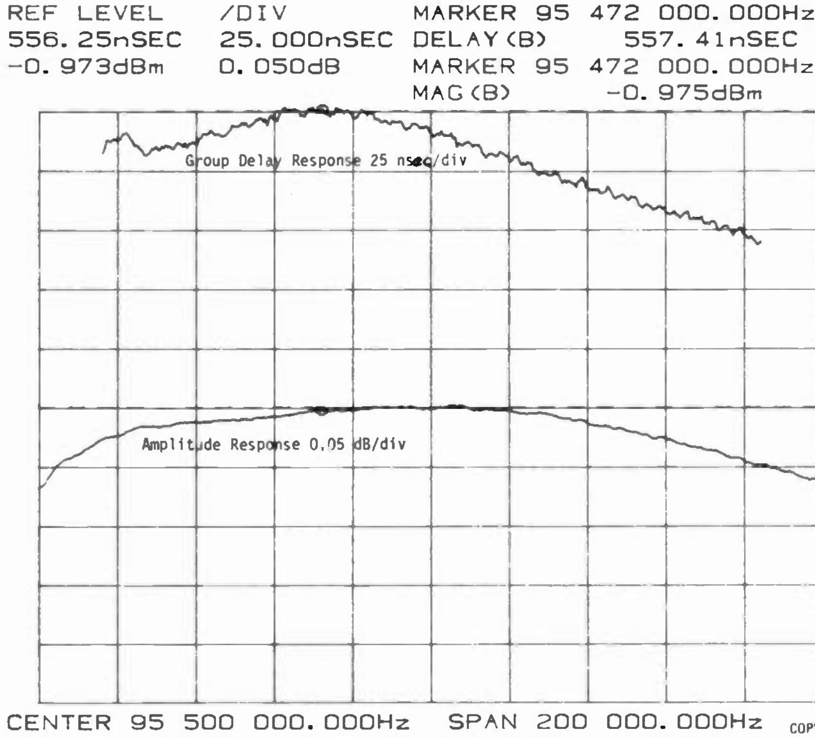


FIGURE 39. MEASURED AMPLITUDE AND GROUP DELAY RESPONSES OF A 20 kW FM TRANSMITTER WITH POWER AMPLIFIER TUNED FOR MINIMUM SYNCHRONOUS AM

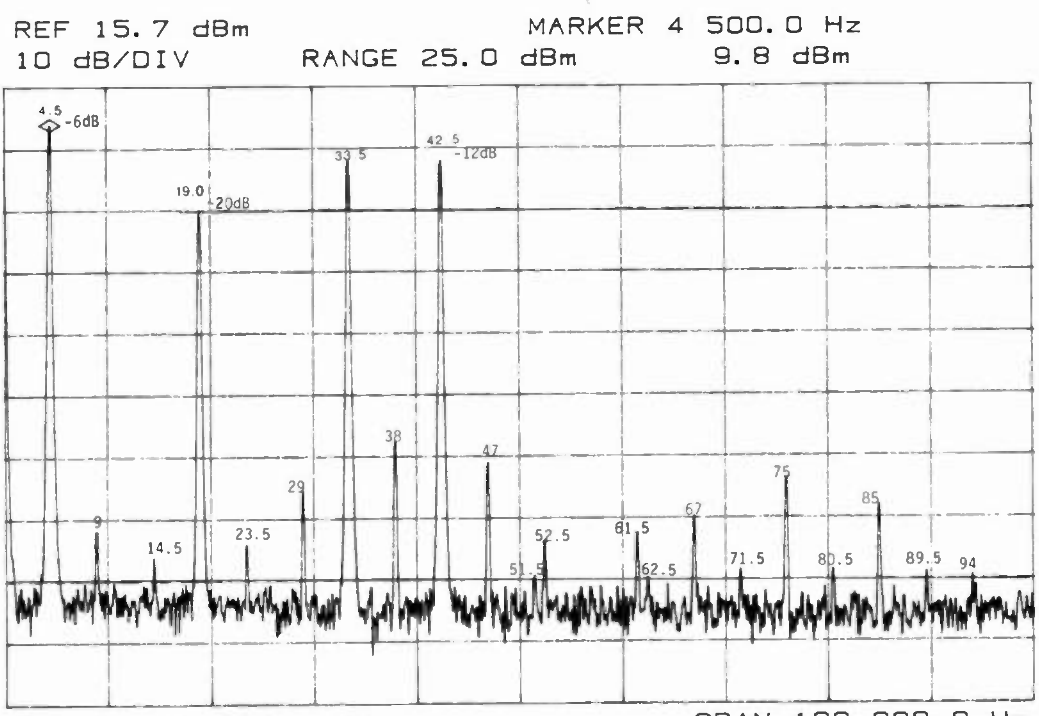


FIGURE 40. DEMODULATED COMPOSITE BASEBAND SPECTRUM OF A 20 kW FM TRANSMITTER WITH 4.5 kHz SINGLE CHANNEL STEREO MODULATION AND RF POWER AMPLIFIER TUNED FOR MINIMUM SYNCHRONOUS AM

Figure 41 shows a plot of the measured overall amplitude and group delay responses of a 20 kW FM transmitter with the power amplifier tuned for symmetrical group delay.

Figure 42 shows a plot of demodulated composite baseband spectrum with 4.5 kHz single channel stereo modulation.

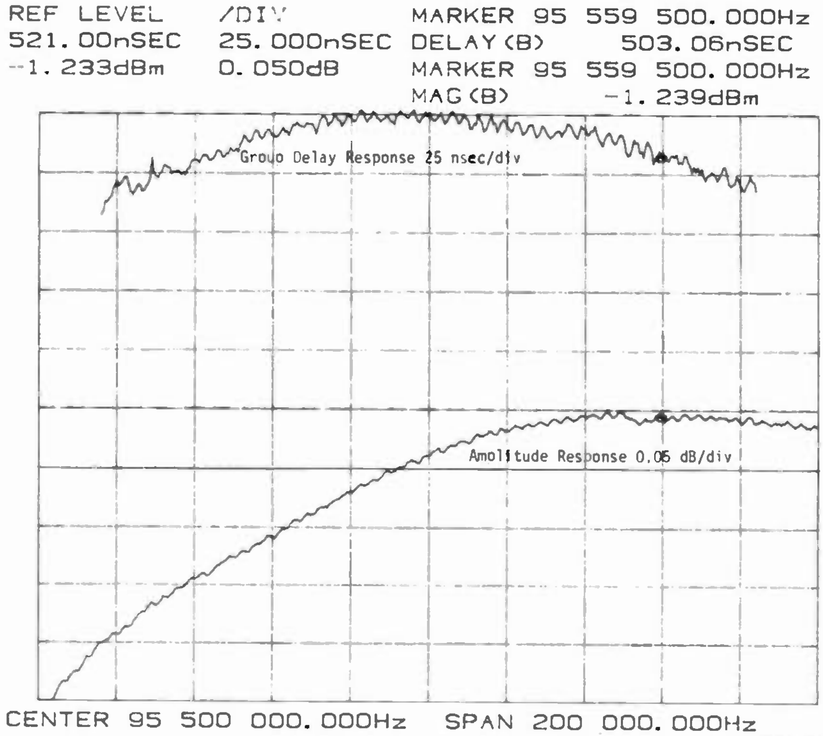


FIGURE 41. MEASURED AMPLITUDE AND GROUP DELAY RESPONSES OF A 20 kW FM TRANSMITTER WITH RF POWER AMPLIFIER TUNED FOR SYMMETRICAL GROUP DELAY

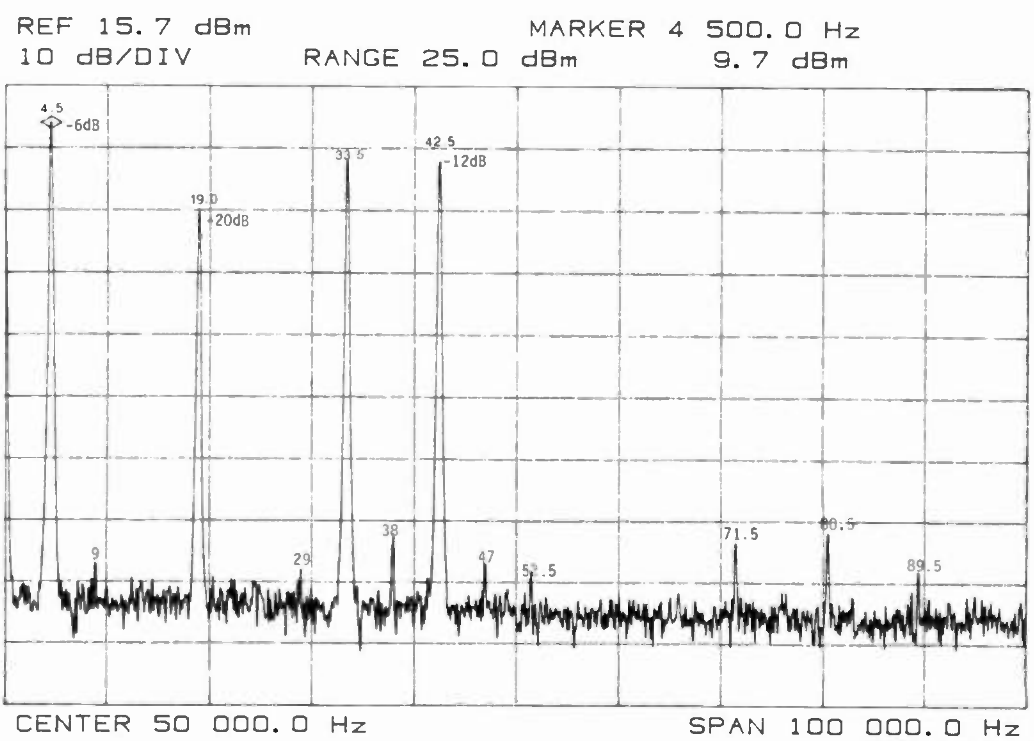


FIGURE 42. DEMODULATED COMPOSITE BASEBAND SPECTRUM OF A 20 kW FM TRANSMITTER WITH 4.5 kHz SINGLE CHANNEL STEREO MODULATION AND RF POWER AMPLIFIER TUNED FOR SYMMETRICAL GROUP DELAY

The modulation distortion and crosstalk components were significantly reduced (see Figure 42) when the power amplifier was tuned for symmetrical group delay (see Figure 41). The modulation distortion and crosstalk components in Figure 40, when the power amplifier was tuned for minimum synchronous AM, are identified and tabulated in Appendix 1 including the combination frequencies which caused these distortion products. The levels of modulation distortion and crosstalk components in dB relative to 100% modulation for the composite baseband spectrum plots of Figure 40 and Figure 42 are also tabulated in Appendix 1.

Figure 43 shows a comparison of composite THD+N and stereo THD+N at 15 kHz in a 20 kW FM transmitter tuned for minimum synchronous AM versus group delay symmetry.

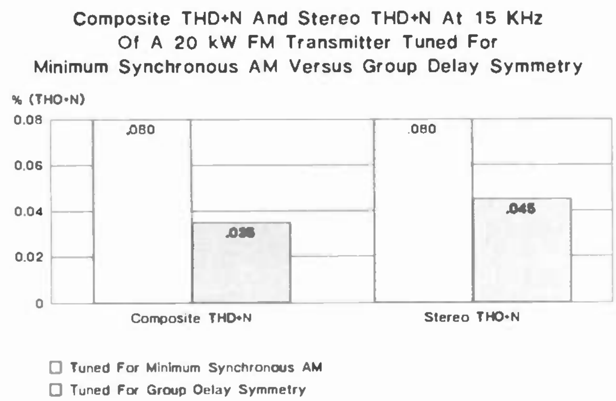


FIGURE 43. COMPOSITE THD+N AND STEREO THD+N AT 15 kHz OF A 20 kW FM TRANSMITTER TUNED FOR MINIMUM SYNCHRONOUS AM VERSUS GROUP DELAY SYMMETRY

Figure 44 shows a comparison of stereo separation at 15 kHz and Figure 45 shows a comparison of synchronous AM for the two different tuning conditions.

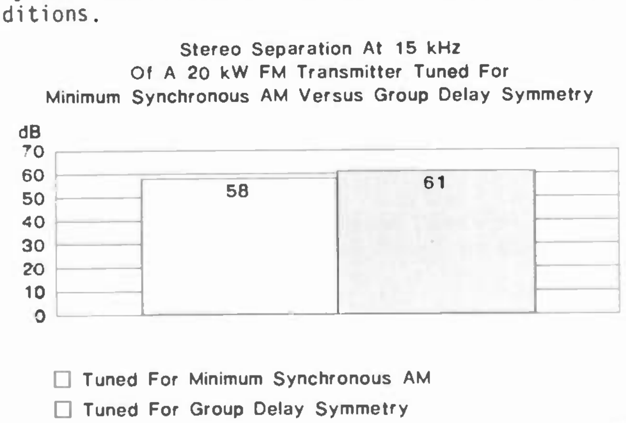


FIGURE 44. STEREO SEPARATION AT 15 kHz OF A 20 kW FM TRANSMITTER TUNED FOR MINIMUM SYNCHRONOUS AM VERSUS GROUP DELAY SYMMETRY

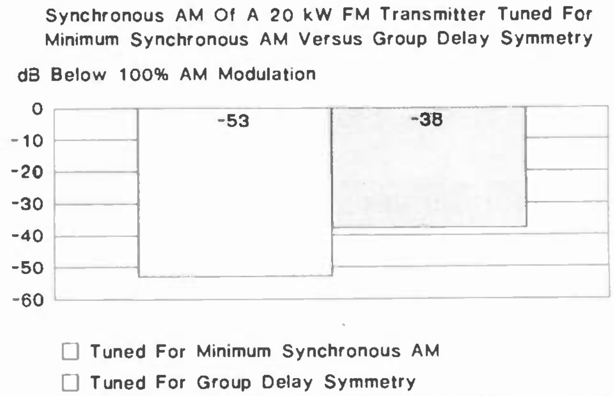


FIGURE 45. SYNCHRONOUS AM OF A 20 kW FM TRANSMITTER TUNED FOR MINIMUM SYNCHRONOUS AM VERSUS GROUP DELAY SYMMETRY

CONCLUSIONS

The significance of RF power amplifier circuit topology on FM modulation performance has been identified. The conclusions reached are as follows:

1. RF bandwidth affects audio performance. It is, therefore, necessary to minimize bandwidth limiting components in the RF path to reduce performance degradation.

2. Good engineering judgement is called for to balance the trade-offs between modulation performance and immunity to RF intermodulation. A bandwidth of 1.0 to 1.5 MHz is adequate for excellent modulation performance while providing protection from RF intermodulation.

3. The design of RF power amplifiers for high quality FM broadcast transmitters requires careful consideration in the choice of input and output circuit topology due to their combined effects on the amplitude and group delay responses. Certain topologies such as broadband input matching and magnetic output coupling result in better overall transmitter performance because their amplitude and group delay responses coincide closely.

4. Measurements taken on a 20 kW FM transmitter showed that tuning the transmitter power amplifier for minimum synchronous AM did not necessarily result in best modulation performance. Tuning the power amplifier for symmetrical group delay response results in minimum distortion and crosstalk.

5. Equipment manufacturers may provide information on the amplitude and group delay responses of transmitters. This would allow broadcast system engineering to be tailored to particular needs. It would also create opportunities for designing delay equalization networks in the system to compensate for transmitter caused group delays as well as delays caused by the filterplexer and combining system.

APPENDIX 1

MODULATION DISTORTION AND CROSSTALK COMPONENTS IN THE DEMODULATED COMPOSITE BASEBAND SPECTRUM FOR A SINGLE CHANNEL 4.5 kHz STEREO MODULATION

(ORIGINAL FM SIGNAL MODULATION COMPONENTS ARE 4.5 kHz, 19 kHz, 33.5 kHz, and 42.5 kHz)

MODULATION DISTORTION AND CROSS-TALK COMPONENTS (kHz)	COMBINATION FREQUENCIES CAUSING SECOND AND THIRD ORDER PRODUCTS (kHz)	LEVEL IN dB RELATIVE TO 100% MODULATION	
		RF AMPLIFIER TUNED FOR MINIMUM SYN-CHRONOUS AM	RF AMPLIFIER TUNED FOR SYMMETRICAL GROUP DELAY
9.0	42.5-33.5 2x4.5 (2x19)+4.5-33.5 42.5+4.5-(2x19)	-72	-77
14.5	19-4.5 33.5-19 19-(2x4.5)	-76	-80
23.5	4.5+19 42.5-19 (2x4.5)+33.5-19	-74	-80
29.0	33.5-4.5 (2x19)-(2x4.5) (2x33.5)-(2x19) (2x19)+33.5-42.5	-65	-77
38.0	33.5+4.5 42.5-4.5 2x19 33.5+42.5-(2x19) (2x4.5)+(2x33.5)-(2x19)	-57	-72
47.0	42.5+4.5 (2x4.5)+(2x19) (2x42.5)-(2x19) (2x19)+42.5-33.5	-62	-77
51.5	42.5+(2x4.5) (2x42.5)-33.5 (2x42.5)+4.5-(2x19)	-79	-82
52.5	19+33.5 19+42.5-(2x4.5) (2x33.5)+4.5-19	-73	-78
61.5	19+42.5	-72	-83
62.5	(2x33.5)-4.5 (2x19)+(2x33.5)-42.5	-79	-83
67.0	2x33.5 33.5+42.5-(2x4.5) 33.5+(2x19)-4.5	-70	-83
71.5	42.5+33.5-4.5 (2x19)+33.5 (2x33.5)+4.5 (2x33.5)+42.5-(2x19) (2x19)+42.5-(2x4.5)	-78	-74
76.0	33.5+42.5 (2x19)+42.5-4.5 (2x19)+33.5+4.5 (2x42.5)-(2x4.5) (2x33.5)+(2x4.5)	-64	-82
80.5	33.5+42.5+4.5 (2x19)+42.5 (2x42.5)-4.5 (2x42.5)+33.5-(2x19)	-78	-73
85.0	2x42.5 (2x19)+42.5+4.5	-68	-84
89.5	(2x42.5)+4.5 (2x42.5)+(2x19)-33.5	-79	-78
94.0	(2x42.5)+(2x4.5)	-79	-83

ACKNOWLEDGEMENTS

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The author holds a U.S. Patent for electronic design utilized in broadcast equipment and is a member of the Institute of Electrical and Electronics Engineers. He is also a member of Tau Beta Pi and Phi Kappa Phi honor societies.

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DIGITAL AUDIO TECHNIQUES FOR REMOTE BROADCASTS

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Gannett Radio Division
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ABSTRACT

Several technologies including digital audio bit rate reduction systems, digital telco networks, VSAT satellite systems, and high speed modems have recently converged to permit high quality digital audio remote broadcasting. These technologies can now be cost effectively implemented by stations for 7.5khz to 15khz bi-directional remote broadcasts.

INTRODUCTION

Remote broadcasting has been with the broadcast industry since it's inception as a business. Remotes have encompassed everything from sporting events, to wars, to human interest stories and are truly a major force in having made the world a smaller place for all of us. Radio as an industry has had a renewed interest in remotes during the 1980's and early 1990's and is "hitting the streets" with greater frequency in an effort to get closer to it's listeners. Technology is also improving and this can permit us to deliver better and more entertaining live remote broadcast events. This paper endeavors to explore the options that new technology delivers to add entertainment value in remote broadcast events.

Remote broadcasts which extended beyond the range of a stations local signal coverage were once the exclusive territory of the O & O news stations. Today these needs are changing rapidly with the demand for both news/talk and music stations to deliver more local, regional, national, and international remote broadcasts. This renewed demand for remotes comes from several different sources; radio personalities taking listeners on a vacation to ski resorts or beach resorts, resorts involved in promotional activity, and news events which emphasize a stations commitment to it's music format and listeners. Remotes are becoming a method for radio stations to extend the scope of entertaining their audience beyond simply delivering music and jokes on- air. Locations range from a sporting bar across the street, to the Beatles' Abbey road studios in London, England, to the beach in Puerto Vallarta, Mexico. These "Radio Remote Events" are occurring much more often and, judging from their success, they will be with us for some time to come.

Consistent with the changes in stations and formats doing remote broadcasts from regional, national, and international locations, we see a change in technical requirements. When an AM news personality does a remote broadcast or story from the field, it has always been acceptable to deliver a limited bandwidth feed to the studio and to recieve a limited bandwidth cue signal in return. This is usually not acceptable when a music station's morning team takes a group of 200 to 300 listeners to a beach party in Mexico or the Carribean. The team must be able to relate with the audience at home and also to entertain the listeners on the trip. The technical requirements which are required to accomplish this type of "Remote Broadcast Event" are more comprehensive than ever before and include: Bi-directional 15 Khz audio, cue channel, and a telefax machine hookup. A block diagram of an idealized setup is shown in fig. 1.

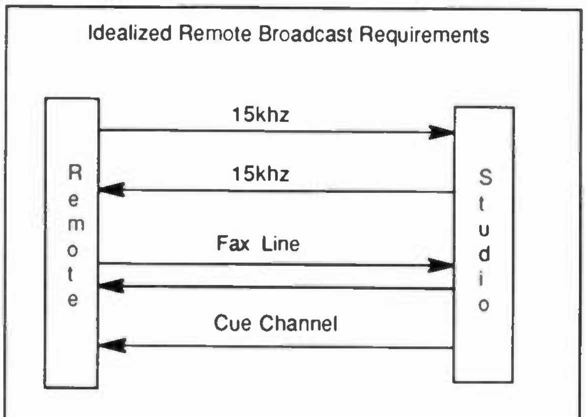


Fig. 1

Radio broadcasters have used many different technologies to deliver remote broadcast feeds from the field with varying success. The two broad categories of signal delivery technologies used are RF or Telco lines. The RF systems typically operate under part 74 of the FCC rules and are limited in noise/ interference free, high bandwidth coverage to roughly 45 miles. Both RF and Telco systems currently are used to deliver analog audio. This imposes a limitation with reference to noise immunity on the RF systems. Satellite portable earth stations have also been used when the remote broadcast is geographically beyond the limitations of the station's RF and the telco system. However, untill recently, satellite earth stations have not been widely utilized because of their cost to operate. The hardware has been beyond the scope of ownership by a radio station or group of stations, which required using outside contractors to provide satellite-delivered remotes.

Multiple telephone line frequency multiplexed systems have also begun to gain acceptance but they are cumbersome and still fairly costly for regional or national remotes. They also suffer from a limitation of delivering high quality bi-directional audio which is desireable for staging "Remote Broadcast Events". International broadcast remotes from underdeveloped countries can also render these devices useless because of poor-quality telephone systems.

Digital audio for remote broadcasting has offered great promise, primarily for it's inherent immunity to noise and it's inherent quality. Digital audio systems have the ability to deliver a high quality recovered audio signal, even at the threshold of RF detection. The main obstacle to implementing digital audio transmission systems for remotes is from the bandwidth requirement. Digital audio is a bandwidth hog and simply will not conveniently or cost effectively fit into the transmission systems which are traditionally used by radio broadcasters. We need to explore new transmission systems to begin using digital audio remotes.

The formula for determining the bandwidth requirement of a digital audio system is simple and an example is included in Fig. 2 for a system with 90 db S/n ratio and 15 Khz bandwidth. We therefore required 15 bits in each sample at a sample frequency rate of 32Khz. Because we use serial transmission systems we will have a bandwidth requirement equal to number of samples times the sample frequency or in our example 480kbits/sec per channel. For the example we therefore require a digital transmission system which is capable of 960kbits/sec for the two main audio feeds illustrated in Fig. 1. These 15Khz digitized signals simply cannot fit into an FCC part 74 - 450mhz channel or down a simple telephone line. Satellite time for a T-1 carrier (1.544 mbit/sec) is an expensive option. As shown in the Fig. 2 example and stated previously , digital audio for live radio remote broadcasting can only be viable with a reduced bandwidth requirement. To achieve a bandwidth reduction in digital audio, we need to explore using digital audio encoding systems which are more bandwidth efficient than simple PCM encoding.

DIGITIZED AUDIO PCM BANDWIDTH REQMTS

RESOLUTION

Desired S/n = 90db @ 6db/bit = 15 bit resolution

SAMPLE RATE

Max audio freq. = 15khz 15khz x 2 = 30khz
nearest std = 32khz

32khz x 15bits = 480kbits/sec/channel

Fig 2.

For digitized PCM audio to be transmitted through a traditional RF system or telco system we must first use a modem. A block diagram of a digital audio PCM encoded RF transmission system is shown in Fig. 3.

The broadcaster requirements for a modem using the fig. 2 example are for 960 Kbits/sec capacity and minimum bandwidth in the modulating process. High speed T-1 and fractional T-1 modems are available which utilize bandwidth reduction techniques over the bandwidth normally required by FSK. More common techniques used are BPSK(bi phase shift keying) and QPSK (quad phase shift keying). New RF modulating schemes which involve multiple phase shifts per cycle are also currently in limited use.

Digital Audio PCM Transmission Block Diagram

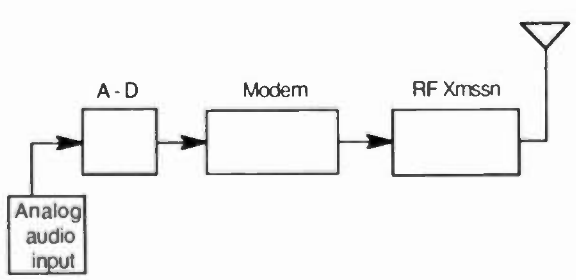


Fig. 3

For our purposes, the cost of using modulating schemes beyond the sophistication of QPSK are not yet cost effective. Fig. 4 shows a method of applying BPSK and QPSK to an RF sinewave.

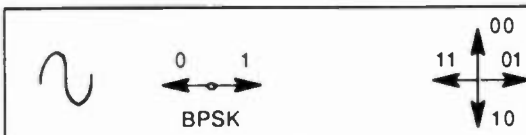


Fig. 4

DIGITAL COMPRESSION

Because a digitized PCM audio transmission is essentially a serial data transmission, we can consider bit rate reduction techniques to make digital audio more attractive to broadcasters. Digital bit rate compression systems have been around since the 1950's. Early systems used the "Huffman Squeezing" method which took advantage of the redundancy of the letters E,R,T,N, and O in text. In the late 70's the Ziv-Lempel compression method gained favor by sending a single binary digit code for each repetitive data string. The goal of any digital bit reduction or compression algorithm is to identify similarities or redundencies in data and to minimize the bandwidth or storage requiments of the transmission or storage medium. Digitized audio bit stream reduction becomes more difficult to implement than simple text reduction or data streams because mistakes become audible easily and errors in data cannot be retransmitted in real time.

The international telephony companies early recognized the cost savings inherent to bit rate reduction of PCM ditgitized audio and pushed for a standard. CCITT standard G.721 addressed the need and used ADPCM (adaptive pulse code modulation) to reduce the digitized audio in the spectrum of 300 to 3400 hz. The G.721 reduced 3400hz PCM telephone data into 32Kbits/second. This resulted in a 2:1 data compression scheme. A good reference for digital audio speech bit rate reduction systems such as employed in CCITT G. 721 and G.722 is by Crochiere, Webber, and Flanagan. "Digital coding of speech in sub-bands" (Bell System Technical Journal. 1976)

Beginning in 1983 a need for "high quality" audio transmission bit reduction schemes was recognized and CCITT standard G.722 was developed to address this need. G.722 was finished in 1986 and describes a method of delivering a 50hz to 7500hz PCM encoded audio signal with 14 bit resolution into transmission systems with either 48 Kbit/sec, 56 Kbit/sec, or 64 Kbit/sec. The CCITT G.722 standard uses two bands of ADPCM and sub band coding to deliver a very cost effective bit rate reduction system. Using equipment designed around the CCITT G.722 standard the United States, broadcasters can now use the in-place 56Kbit dial up data network to deliver bi-directional 7.5Khz remote broadcasts. When ISDN is implemented, the G.722 standard will provide an even greater opportunity to eliminate reliance on analog systems for live remote radio remote broadcasts.

Recent developments by several companies and engineers worldwide also provide cost effective bit rate reduction systems for the broadcaster in the 1990's. SSL has demonstrated the APT-X system of digital audio bit rate reduction at the 4:1 compression ratio with no "audible" audio degradation. Dolby is also working on a system which uses transform coding to implement 4:1 bit rate reductions. For our earlier example in Fig. 2, the 4:1 system would occupy 128kbits/sec. These systems use the psychoacoustic theories of Temporal masking and Spectral masking (critical band theory) under the umbrella term of "perceptual coding" to develop the PCM bit rate reduction algorithms. A new system, the MUSICAM (**M**asking-pattern **U**niversal **S**ub-band **I**ntegrated **C**oding **A**nd **M**ultiplexing) system was also introduced at the CCIR convention in Oct. 1989. This system was developed jointly by French, Dutch, and German engineers and delivers CD quality in 100kbits/sec and "near CD quality" in 64kbits/sec. All described systems offer potential to the U.S. broadcaster for remote broadcast applications, particularly when used in conjunction with currently available VSAT (very small apperature terminal) Ku satellite hardware.

VSAT

VSAT Ku satellite terminals were developed to cost effectively deliver digital data between large national and international users with multiple locations such as K-mart, automobile dealer networks, and similar business users. This hardware typically uses cost effective Ku band transceivers, 0.75mtr to 3.5mtr Ku band dishes, and high speed digital modems. Because of the large quantity of Ku band digital VSAT hardware now in use by business, the cost has dropped dramatically and is in the range of "reasonable" for some broadcasters. By "marrying" the VSAT hardware technology to digital audio bit reduction systems, radio stations can begin to to deliver the remote broadcast events which require the Fig. 1 signal paths. VSAT hardware also lends itself to extreme portability which is a perfect fit for radio remote broadcast applications.

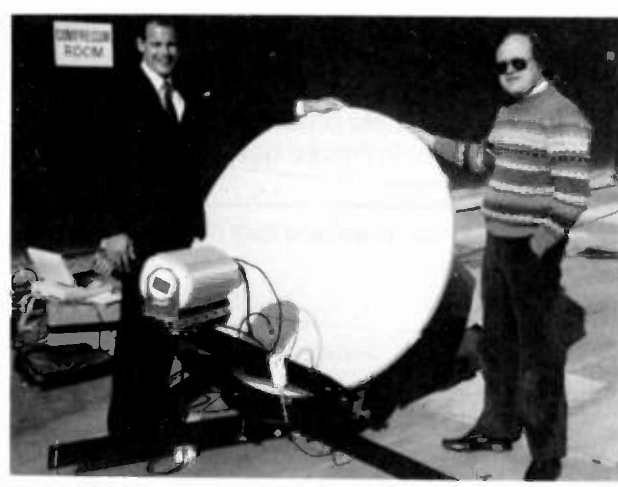


FIG. 5
"Flyaway-Portable Digital Audio System"

TEST RESULTS

At Gannett Radio, we have explored the various technologies described herein for the past year and have run substantial tests on Ku band VSAT systems using digital audio bit rate reduction techniques. The results of the tests are quite encouraging, with no "perceptual" degradation of the digital audio after bit rate reduction. In the tests, we used a 1.8mtr fixed installation VSAT and a 1.2mtr flyaway-portable VSAT. Fig. 5 shows the flyaway 1.2mtr digital audio earthstation which was used for the tests.

Frequency response was flat to 15khz and wide band noise was 85db below clipping levels. The graphs of the recovered frequency response and wide band noise are attached as Fig. 6 and Fig. 7 respectively.

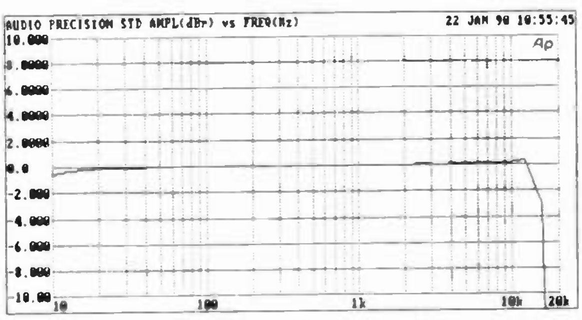


Fig. 6

The typical RF fade margin for the Ku band VSATs used was in the 8-9db range, which would result in availability 96.6% of the time.

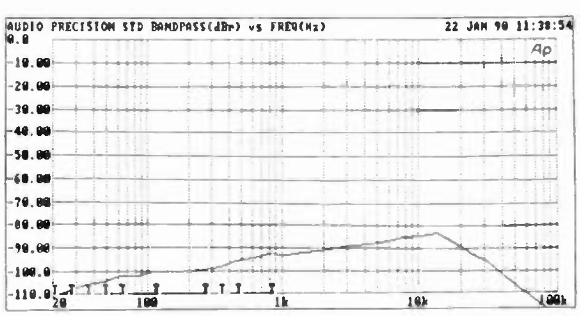


Fig. 7

Based on our experience with these technologies, I believe that live digital audio remote broadcasts are within the reach of many radio stations. . .today! ! ! Current cost levels for a fixed wideband digital audio VSAT teleport system is in the range of $25,000.00 and under $35,000.00 for a flyaway portable system.

We currently have no direct experience using the CCITT G.722 systems and the 56kbit dial up data network, but systems are now available and we will begin testing in the near future. The G.722 standard, when used with the installed 56kbit network, should provide a more cost effective method of delivering bi-directional regional remotes than existing analog methods.

SUMMARY

Remote broadcasting is entering a new dimension with the merging of digital audio, telco data lines, and Ku band VSATs. This is an exciting era for Radio broadcasting and holds the promise of bringing greater entertainment value to the medium.

A HYBRID SATELLITE/TERRESTRIAL APPROACH FOR DIGITAL AUDIO BROADCASTING WITH MOBILE AND PORTABLE RECEIVERS

D. Pommier, CCETT, France
P.A. Ratliff, BBC, United Kingdom
E. Meier-Engelen, DLR, RFA

Abstract -
The Digital Audio Broadcasting technique offers many advantages, especially in «hostile» environments which portable and mobile reception meets in dense urban or mountainous areas. It also provides for a complete digital sound programme chain from the digital studio to consumer receivers.

The European Broadcasting Union (EBU) and Eureka DAB Project have designed and implemented a new advanced digital broadcasting technique which offers the opportunity of developing new radio networks which are particularly power and spectrum efficient by means of terrestrial repeater stations operating on the same frequency. In a further application of the technique, a new hybrid satellite/terrestrial network approach is proposed, having the following advantages :

- minimum satellite power demand,
- minimum power flux density at earth's surface,
- high spectrum efficiency,
- complete coverage of the service area.

This paper outlines the new approach and includes examples of systems, in addition to preliminary experimental evidence. It also demonstrates the advantages of a hybrid approach for satellite sound broadcasting systems, which make highly efficient use of the radio spectrum and improve possibilities for sharing with other services.

1- Introduction

Technological evolution and the ever-increasing demand for higher-quality sound gives broadcasters a strong incentive to completely digitize their audio broadcasting networks. This digitization, which is already well advanced in many programme production areas and transmission links, now has to be extended to complete the last link in the broadcast chain; i.e. from broadcast transmitter to consumer receivers.

It is therefore necessary to develop a wholly new technique for the broadcasting and direct recording of very high-quality digitally-coded sound programmes. Thus, an efficient baseband digital coding, which has to be well-adapted to recording purposes, must be combined with an efficient digital modulation and channel coding scheme that can meet the requirements of every mode of broadcast reception ; i.e. mobile, portable and fixed.

The sound broadcasting service would then evolve alongside the development of recording media such as compact discs and consumer digital audio recorders.

However, the application of digital techniques to sound broadcasting is a viable proposition only if it includes the effective use of the radio-frequency spectrum, a vital condition for these new techniques in times of increasing competition for the use of a precious natural resource.

It is for this reason that a large number of European research centers are currently engaged in defining the basic principles of this future system, within the framework of the Eureka DAB Project and the EBU. [1], [2].

Both satellite and terrestrial applications are under consideration, depending on the size of the coverage areas, the number of national or local programme channels, the available operating frequencies and the associated economic factors. However, if the modulation scheme allows for a receiver able to cope with long delay multi-path signals, the latter does not have to know where the signals have originated from. This novel situation may be reproduced by a network of spaced transmitters broadcasting the same signal at the same frequency. Based on this idea, a terrestrial VHF single-frequency network is already being looked into by the Eureka DAB project members [3], but the same feature can be used to fill any gaps in the coverage area of a UHF satellite broadcast network.

D. POMMIER is Deputy Director of the CCETT.
P.A. RATLIFF is Chairman of EBU Specialist Group V4/RSM studying Digital Audio Broadcasting system.
E. MEIER-ENGELEN is Project Manager of the Eureka DAB project 147.

This paper deals with a new approach which combines a satellite's capability to cover a large area and the flexibility of terrestrial co-channel retransmission in completing coverage in the service area where shadowing prevents mobile or portable receivers in the home from working reliably.

2 - Basic requirements

Digital audio techniques are expanding into the domestic consumer market, leading to wide public appreciation of high-quality stereophonic sound.

Nowadays the present VHF/FM terrestrial network provides excellent service to fixed home receivers equipped with a suitable antenna ; in contrast, reception of FM signals in moving-vehicles and on portables in the home is often impaired by multipath propagation, obstruction and/or man-made interference. But, as a result of the growing demand for high quality sound and of the extent to which mobile in car and portable home radio receivers are now used, the fundamental requirements for a new sound broadcasting system are to :

- provide as good quality as consumer digital recordings (compact discs),
- ensure service continuity for mobile reception (car radios and for portables around the home),
- provide simple programme selection and clear service information.

With regard to the definition of the service area for both mobile and portable reception, a new sound broadcasting system should aim to provide good reception in at least 99% of locations for at least 99% of the time.

3 - Satellite or terrestrial network ?

Depending upon the size of the service area, whether it is local, regional or nationwide, and programme content flexibility in time and location, many trade-offs must be considered between the choice of a satellite or terrestrial broadcasting networks.

In the future, when digital audio broadcasting matures it seems likely that the most cost-effective and radio spectrum efficient solution will be provided by a mix of UHF satellite-based services and VHF/UHF terrestrial services. But new spectrum must first be found to establish these new services before in the longer term, the full advantages of increased spectrum capacity and utilisation can be exploited by re-engineering existing FM radio bands.

Whilst keeping in mind these practical and operational aspects, people in charge of studying a reliable system are doing research into three types of networks :

- a terrestrial network,
- a satellite network,
- a hybrid satellite/terrestrial network.

With regard to multipath and obstruction by obstacles located on the earth, terrestrial transmission has the drawback of a low elevation angle of wave propagation. But in many cases a low elevation angle is perfectly adequate to provide favourable propagation conditions, (e.g. in tunnels, in the case of mobile reception, or in blocks of flats and homes in the case of portable receivers).

In contrast, satellite transmission may be endowed with the advantage of a high elevation angle and hence a clear path in open sky receiving conditions, but it is probably not adequate for reception under highly-enclosed conditions, e.g. tunnels, flats, etc... Obviously, in the case of satellite broadcasting, the propagation margin trade-off between open sky and non-open sky receiving conditions depends on the elevation angle, but it is clear that if one margin decreases, the other will increase. So, bearing in mind that a satellite broadcasting system must be designed in such a way that the transmitted power is kept as low as possible, different system architectures and concepts must be carefully compared on the basis of providing the same required service availability.

Not withstanding the value of terrestrial networks, [1], [3], the following will examine two different system architectures: satellite and hybrid satellite/terrestrial.

- **Satellite**

Depending on the elevation angle, this configuration is subject to shadowing for mobile or portable home reception. According to the CCIR [4] the key figures of a typical link budget for advanced digital systems are :

• Carrier frequency	:	1 GHz
• Propagation margin	:	10 dB
• Beam width	:	1°
• Receiving antenna gain	:	5 dB
• Transmitted power per stereophonic channel	:	20 W

Even if the carrier frequency does not exceed 1 GHz, two problems still arise from the above link budget :

- a beam width of 1° may be too narrow, given the size of some coverage areas,
- the relatively low propagation margin, which is unlikely to prove adequate in many shadow areas, i.e. : streets at the base of high building, tunnels, flats, ...

Moreover, even a large increase in the propagation margin will not ensure flawless reception in heavily shielded areas. For instance, by applying the Annex 2 method of CCIR Report 955 for 99% and 90% of locations, the propagation margins given in Table 1 are found to be necessary :

Table 1. Calculated propagation margins for a simple satellite broadcast service

f = 1 GHz, a ≈ 30°	Propagation margin		
	Percentage of locations	99%	90%
Urban	Obstructed visibility*	≈20dB	≈16dB
	Direct visibility*	≈10dB	≈7dB
Rural	Obstructed visibility*	≈10dB	≈8d
	Direct visibility*	≈3dB	≈2dB
	* As defined in [4]		

- **Hybrid satellite/terrestrial**

This new concept can only be made possible if the modulation and channel coding scheme is able to cope with long-delay multipath.

If the above condition is fulfilled, the satellite can be designed just to provide sufficient signal strength in open sky receiving conditions, since terrestrial repeater stations working at the same frequency will boost the signal in areas where an extra propagation margin is required, and only in those areas.

This technique is termed gap filling or an active echo technique, the principle of which is illustrated in figure 1. A very simple and small terrestrial station receives the signal from the satellite in an open area and retransmits it, at the same frequency, towards the shadow areas.

It will be shown in sections 5 and 6 that the retransmitted EIRP(1) per stereophonic sound channel can be as low as 0.25 W or less, depending on the size of the space to be filled, and, of course, the degree of isolation between the receiving and the retransmitting antennas which can be achieved.

(1) *Equivalent Isopic Radiated Power)*

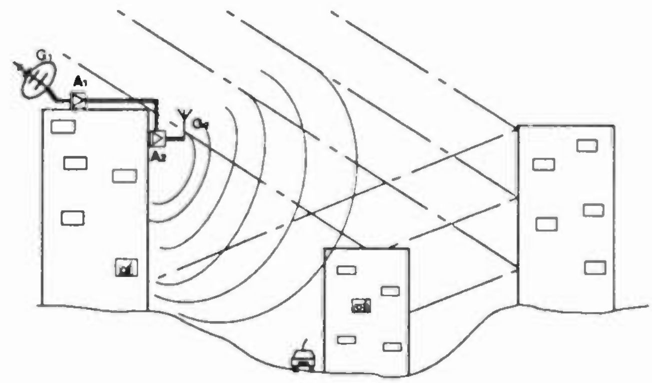


Figure 1 -Gap filling technique. The EIRP of the retransmitted signal is given by :

$$PFD^{(2)} + 10\ Log\left(\frac{\lambda^2}{4\pi}\right) + G1 + A1 + A2 + G2$$

Figure 2 gives a comparison of the required PFD(2) for 99% of locations in urban area when considering :

- a simple satellite system (no gap filling technique)
- use of one retransmitting station
- use of three retransmitting stations

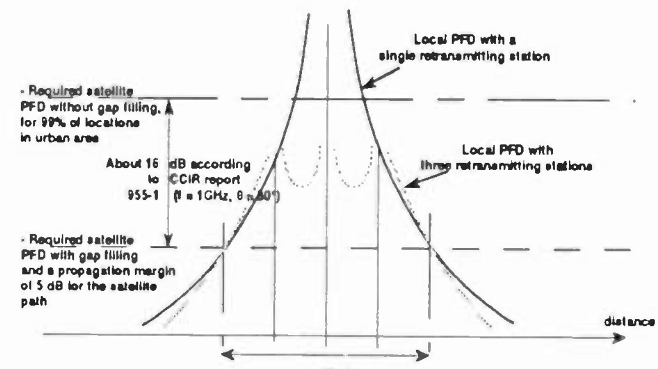


Figure 2 - Required PFD according to the architecture of the system used.

For a large gap to be filled, it is worth noting that several small retransmitting stations may be used with the benefit of a lower local PFD than with a single retransmitting station.

According to this new approach, the satellite propagation margin could be assessed as low as 5 dB, leaving some 5 dB available for trade-off between the transmitted power, the beam width, or possibly the carrier frequency. Such a hybrid system can provide complete low-cost coverage, by minimizing the satellite power demand as well as the potential interference at the earth's surface.

(2) *Power Flux Density*

4 - Typical channel model of a hybrid satellite/terrestrial system

The broadcasting channel model resulting from a hybrid satellite/ terrestrial system is a combination of the two channel models. The terrestrial channel is delayed by an extra path length , the maximum value of which is shown in figure 3, where «d» is the distance between the receiver and the terrestrial station and «α» the elevation angle of the satellite direct path.

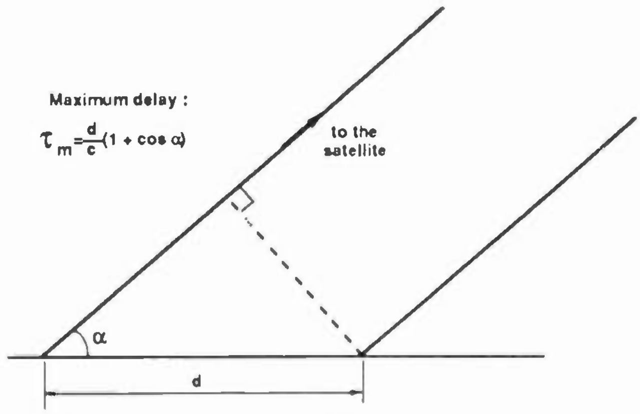


Figure 3 - Maximum delay between the satellite and the terrestrial paths.

These channels may be considered to be both of the independent selective Rayleigh type. The typical reference delay profile is shown in figure 4. The first set of delayed paths representents the spread of the satellite channel, whereas the second set originates from the terrestrial channel spread.

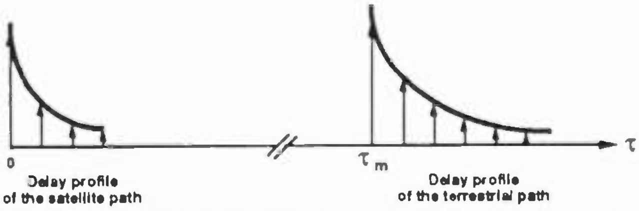


Figure 4 - Typical delay profile resulting from the combination of the satellite and the terrestrial paths.

On the basis of this broadcast channel, a modulation and channel coding scheme capable of positively combining the signals from all satellite and terrestrial paths must be used.

The modulation methods generally proposed to solve this type of problem involve the use of M-orthogonal alphabets [12], which increases the bandwidth of the signal, allowing for path discrimination and their positive combination. But this technique prosesses relatively poor spectral efficiency, which is totally incompatible with the broadcasting application.

Fortunately, a new approach combining high spectrum and power efficiency has been specially designed for mobile reception in a long-delay, multipath environment. This new approach lies in the conjunction of an orthogonal frequency division multiplex technique and a powerful coding strategy [5],[6]. The combined technique is termed COFDM, which stands for : «Coded Orthogonal Frequency Division Multiplex».

5 - The COFDM technique

The fast fading and selective Rayleigh channel model, of which the delay profile is described in figure 4, can be explained as a two-dimensional function of frequency and time. At a given time (t), the frequency response is the Fourier Transform of the channel impulse response ("h" [τ, t]). In the time domain, at a given frequency (f), the field strength follows a random function defined by the well-known Rayleigh distribution. Figure 5 combines the frequency and time response of the channel and illustrates the two-dimensional channel concept.

The COFDM technique operates within the area of this function, where it is locally invariant and statistically independent from other areas [5], [6].

To achieve this, the information belonging to the same programme channel is first split into a large number of orthogonal and overlapping carriers (fig. 5). A Fast Fourier Transform is then used to process the large number of carriers. In order to maintain the orthogonality between carriers in a time-dispersive channel (i.e. with multipath) a guard interval before each symbol prevents

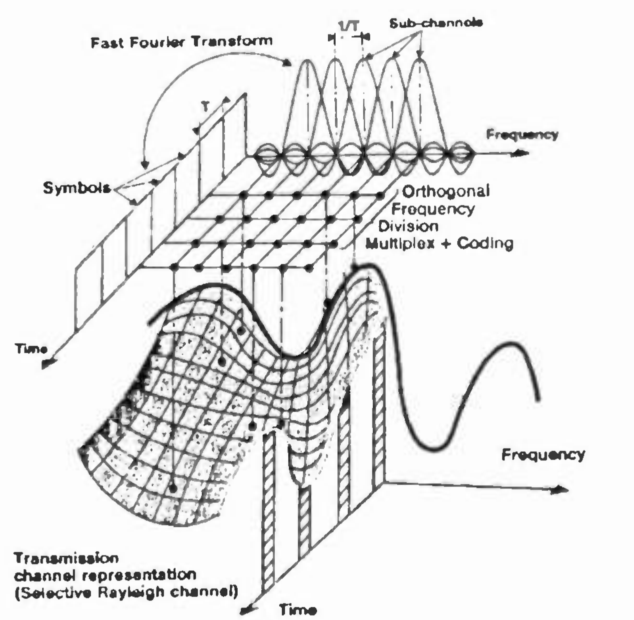


Figure 5 - Two dimensional representation of the COFDM scheme associated to the transmission channel representation.

intersymbol interference [5]. The guard interval consists of a symbol extension in the time domain leading to symbol transmission with a slightly longer period than the inverse of the frequency separation of two successive carriers.

The other key element of the system is the use of powerful convolutional coding in conjunction with a Viterbi maximum likelihood decoding algorithm.

Figure 6 gives the BER performance of a COFDM system in a fast-fading and selective Rayleigh channel. This results refers to a system with spectrum efficiency of 0.8 useful bit per Hertz and a channel coding and modulation scheme featuring :

- a Fast Fourier Transform (for both modulation and demodulation processes) for overcoming frequency selectivity, using many simultaneous narrow band carriers (e.g. 512 in a bandwidth of 4 MHz),

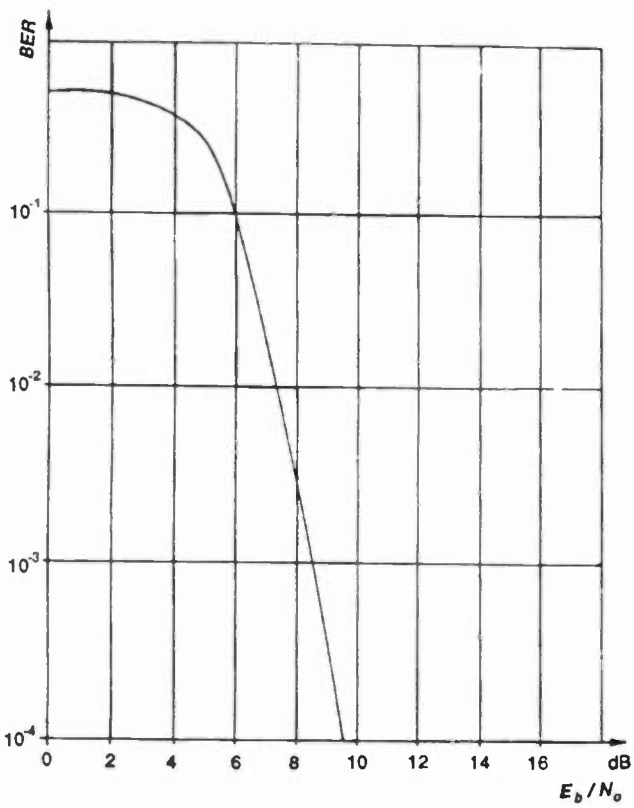


Figure 6 -BER performance of the COFDM system in a fast fading and selective Rayleigh channel.

- frequency and time interleaving to ensure independence between successive symbols at the Viterbi decoder input of the channel decoder.

- a temporal guard interval, with a period of 20% of the total symbol period,

- a total symbol period compatible with the time coherence of the channel (e.g. 320 us for a carrier frequency in the region of 1 GHz and a vehicle speed of 150 km/h),

- a 4 PSK modulation of each carrier,

- a convolutional code of which the rate and free distance are 1/2 and 10, respectively.

Note - Because of the COFDM technique, the multipath contributes positively to the received signal, rather than impairing reception. Thus, the bit error ratio only depends on the mean received power, as with direct wave reception; accordingly, Eb/No denotes the mean energy received by useful bit over the noise power density.

6 - An example of a COFDM system suited to any UHF network, including a hybrid satellite/terrestrial network

In the light of a comprehensive study, including consideration of the temporal and frequency coherence of the channel with respect to the choice of the COFDM key parameters [7], a system (see figure 6 for bit error performance) can be readily implemented with the following constraints :

-carrier frequency	: up to 2GHz
-vehicle speed at which a reduction in performance of about 1 dB occurs in the worst multipath situation	: 150 km/h
-maximum delay for which two signals are still combined constructively	: 12 km.

For one RF channel, the bandwidth is a function of the total useful bit rate conveyed. The following figures can be similarly considered :

Bandwidth	7 MHz	3.5 MHz	1.75 MHz
Total useful bit-rate	5.6 Mbits/s	2.8	1.4

The number of sound channels will depend upon source coding efficiency which is now expected to be high, i.e., 110 Kbits/s, or even less, per monophonic high quality sound channel [8], [9], [10], [11].

As an example, a system using a 3.5 MHz channel, and well suited to a hybrid satellite/terrestrial network would feature the following :

- 3 dB bandwidth : 3.5 MHz
- total number of useful carriers : 448
- symbol period (useful) : 128 µs
- guard interval : 32 µs
- total symbol period : 160 µs
- convolutional code :
  - rate : 1/2
  - free distance : 10
- total useful bit rate : 2.8 Mbits/s
- total stereophonic sound channels :
  1) with 110 Kbits/s per monophonic sound channel (+ 150 Kbits/s for additional data) : 12
  2) with 90 Kbits/s per monophonic sound channel (+100 Kbits/s for additional data) : 15
- minimum Eb/No : 8.5 dB
- minimum C/N measured in 3.5 MHz of noise bandwidth : 7 dB

7 - Architecture of a hybrid satellite/terrestrial network

The key feature of this approach is that the signal can be locally retransmitted at the same frequency into any area (a cell) where shadowing prevents reception of the satellite signal.

However, the size of one cell fed by a terrestrial station is limited by the necessary continuity of service between the cell and the neighbouring area where the signal can be directly received from the satellite. So, the maximum size of one cell is then limited by the delay, expressed in distance, at which signals travelling the two different principal paths are no longer constructively combined.

In the COFDM system example of section 6, the maximum distance between two paths is shown to be 12 km ; Fig.7 shows the limit at which the ratio between the direct signal received from the satellite and the signal received from the terrestrial station (C/I) must be higher than the minimum carrier to noise ratio (C/N). Assuming a free space attenuation for the terrestrial path, and defining the cell by the equal power of the two signals, a useful cell of 3 km radius can be achieved with a C/I of at least 18 dB at the fringe where the two signals are no longer constructively combined. Such a C/I exceeds by about 10 dB the minimum required C/N value, reducing in very few locations the theoretical performance of the system by less than 0.4 dB.

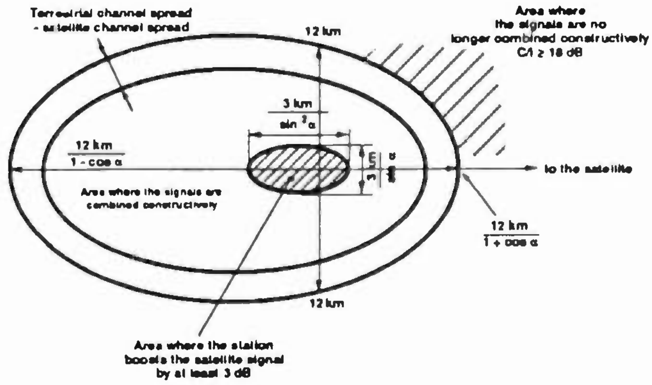


Figure 7 - Areas around a terrestrial station, when assuming free space attenuation.

More comprehensive theoretical studies (including those investigating building effects via statistical model propagation in urban areas) are currently being undertaken. To accompany this theoretical work, a comprehensive experiment has been successfully conducted in the city of Rennes.

8 - Experimental evidence

In June 1988, the CCETT installed the first COFDM UHF transmitter having the following characteristics :

-transmission frequency : 794 MHz
-transmitting antenna height : 140 m
-number of stereophonic sound channels [1] : : 16
-antenna gain in the main direction of the service area : 12 dBi
-power per stereophonic sound channel at the input of the transmitting antenna : 1 W
-total bandwidth: 7 MHz
-total number of carriers processed by FFT : 512
-useful symbol period : 64 µs
-guard interval : 16 µs
-maximum difference at which two signals are still combined constructively : 6 km

The broadcast signal was received in a car equipped for mobile tests. The first successful trial runs under real conditions were conducted in cooperation with the IRT in July 1988 under the time constraint of the EBU first public demonstration of the so-called

[1] *Only one channel was operated with a sound programme, the 15 remaining channels were loaded by a fixed pattern configuration. Today, using the up to date source coding technique, a total number of 24 stereophonic sound channels may be transmitted at the same total useful bit-rate of 5.6 Mbits/s.*

COFDM/MASCAM experimental system in September 1988 at the WARC-ORB 88 in Geneva [2].

These trial runs showed that despite a quite large service area where the reception was perfect, some locations in the urban area were impaired by heavy shadowing and included gaps with signal attenuation greater than 30 dB.

At that time, the idea of using a gap-filling technique materialized, but two questions were brought up by several experts :

- How much separation in dB is feasible between a directive receiving antenna and a transmitting antenna installed in a building environment when the geographical separation is in the range of 50 to 100m ?

- How will the COFDM receiver behave when receiving a signal between the zone served by the main transmitter and that served by the retransmitting station ?

Since that date, two retransmitting stations have been installed (see figure 8) which have the following characteristics :

Station 1

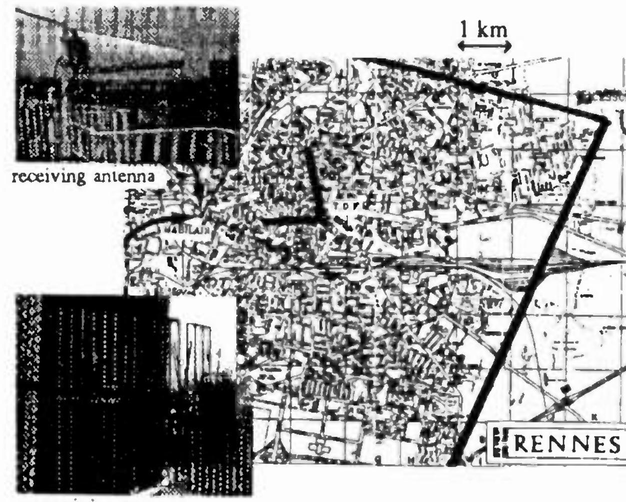
- Receiving antenna gain : 14 dBi
- Transmitting antenna height : 50m
- Retransmitting antenna gain : 9 dBi
- Isolation between the input of the retransmitting antenna and the output of the receiving antenna : 86 dB
- Overal gain of the amplifier : 55 dB
- Cable losses : 5 dB
- Power per stereophonic sound channel at the input of the retransmitting antenna : 2 mW
- total EIRP : 250 mW
- EIRP per stereophonic sound channel : 16 mW

Station 2

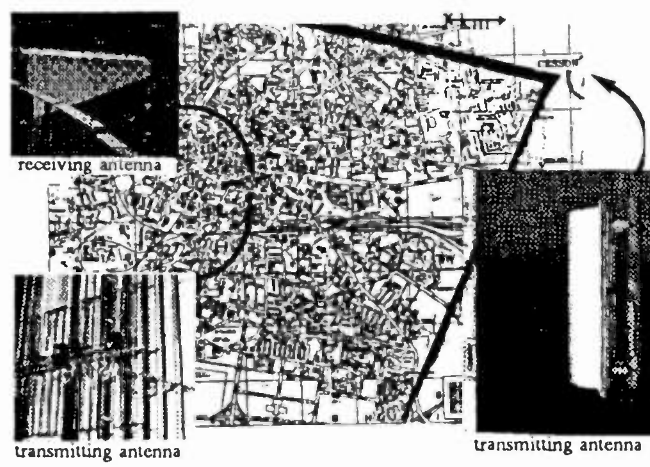
- Receiving antenna gain : 14 dBi
- Transmitting antenna height : 60 m
- Transmitting antenna gain : 9 dBi
- Isolation between the input of the transmitting antenna and the output of the receiving antenna : 90 dB
- Overall gain of the amplifier : 70 dB
- Cable losses : 5 dB
- Power per stereophonic sound channel and the input of the retransmitting antenna : 40 mW

- Total EIRP : 5 W

EIRP per stereophonic sound channel : 313 mW



a) First retransmitting station



b) Second retransmitting station

Figure 8 - Experimental site based on a main UHF transmitter augmented by two cochannel retransmitters (gap filling).

Using this complete single UHF frequency network (including the main transmitter and the two retransmitting stations), numerous test runs and measurements have been made, leading to the preliminary conclusions :

- at UHF, very simple and cheap equipment can be used for a retransmitting station with an available amplification gain of at least 70 dB.
- despite the relatively short guard interval used in that first experimental system (16µs), the behaviour of the COFDM receiver remains excellent even in some exacting situations where two signals of equal power are received with a delay difference exceeding the guard interval of the COFDM symbols by a few µs (see figure 9).

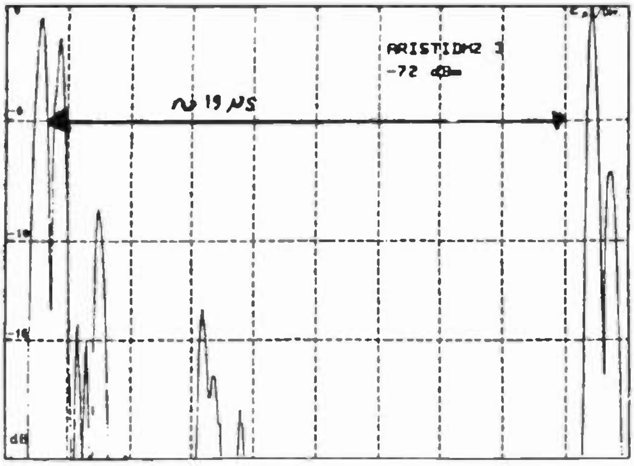


Figure 9 -Example of impulse channel response showing two different originated signals.

Today, apart from a few very small areas, the whole city of Rennes and a wide periphery in the countryside around the city are perfectly served, with a total transmitted power per stereophonic sound channel of1.1 W corresponding to an EIRP of about 11 W.

Since it is more than a simple experiment, this first UHF network has demonstrated the viability of using the gap filling technique rather than increasing, by a factor of 100 or more, the power of the main transmitter . The same result also applies to any satellite sound broadcasting approach of the geostationary or inclined orbit types. Indeed, as explained in section 3, some areas with heavy shadowing will exist whatever the satellite orbit position with respect to the service area.

Conclusion

A good sound broadcasting service to portable and mobile receivers requires an uninterrupted reception in all environments, whether in heavily built-up urban areas, tunnels or blocks of flats, while using a low-gain receiving antenna. In practice, such a condition can be met only if the satellite signal is boosted by small terrestrial co-channel retransmitters. Moreover, the use of small terrestrial co-channel retransmitters makes it possible to significantly reduce transmitted power from satellites. As a matter of fact, a continual trade-off exists between the amount of satellite power and the number of terrestrial stations. Typically, with a reasonable satellite propagation margin of about 5 dB, all open or near open sky locations will be directly served by the satellite signal, whilst the number of retransmitting stations required for shadowed areas will be low enough to insure a global cost effective solution.

The main advantage of the hybrid satellite/terrestrial gap-filling approach is that it minimises the required satellite power and hence also the power flux density at the earth's surface, bringing about a further easement in the geographical separation required outside the broadcast service area for other terrestrial services to operate. The advanced COFDM signal itself, having a noise-like spectrum, is benign in terms of the interference it can cause, and is coded such that it is also highly resistant to interference (especially narrow-band) from other services. The terrestrial gap-filling transmitters are of necessity of very low power and confined to the highly-cluttered environments, as they must not create mutual destructive interference to the satellite signal in the distant, rural parts of the service area. They should, therefore, not give rise to any additional sharing constraints.

Moreover, whilst a satellite provides an instant level of service over the whole coverage area, the terrestrial gap-filling part of the broadcast system may be installed step by step, progressively improving receiving conditions and service continuity for mobile and portable receivers.

Despite the very promising new modulation and channel coding scheme which has undoubtedly proved the suitability of this new approach for digital audio broadcasting, optimum system parameters and a preliminary outline of standard still need to be determined. This work is planned to be completed by the end of 1990.

One of the most important factors affecting the development of such digital sound satellite broadcasting networks is the need for the International Telecommunications Union (ITU) to allocate a sufficient band of UHF spectrum, between the limits of 0.5 and 2 GHz, to the Broadcasting Satellite Service (sound) at WARC 1992[1] preferably on a worldwide basis. In the future, these services can be expected to be in demand in all regions of the world where reasonably large area coverage is required ; this may be where existing terrestrial radio bands become overloaded or where there is currently little widespread development of reliable terrestrial broadcast area coverage.

Everyone involved in this significant step into the 21st century for sound broadcasting is pinning all hope on the ITU WARC 1992 finding a satisfactory solution.

Acknowledgements

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NARROWBAND DIGITAL AUDIO

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Recent developments in digital audio conversion methods have resulted in astounding reductions in data rates required for high-fidelity recording and transmission. These systems employ psychoacoustic masking principles to allow lowered resolution from the typical 16-bit PCM coding to codes of less than 2 bits, while maintaining acceptable fidelity. An examination of the current state of technology among these systems is presented. (Reprinted in part from BME Magazine, November 1989, with permission. Expanded and updated, January 1990.)

INTRODUCTION

The concept of putting digital audio signals on the air for broadcast or point-to-point uses is nothing new. But wideband, high-fidelity stereo audio such as that found in the CD format requires extremely high data rates, on the order of 1.5 Mb/sec, given its two-channel configuration of 16-bit resolution at a sampling frequency of 44.1 kHz.

$$16(44.1 \times 10^3) \times 2 = 1.411 \times 10^6 \quad (1)$$

Putting this kind of signal on the air with standard frequency modulation (i.e., data transmission efficiency of no better than 1 bit/cycle of occupied bandwidth) means that channel widths approaching 2 MHz for digital stereo audio signals are required. This sort of spectrum inefficiency, relative to traditional analog audio transmission, has done little to ingratiate digital audio into the RF domain.

Nevertheless, some point-to-point digital audio transmissions have been utilized by broadcasters for STL and RPU uses. In most cases, these systems require even wider bandwidths (typically a 6 MHz video channel), due to their output formatting of the digital audio datastream as a monochrome video signal. Such "pseudovideo digital" devices were originally designed for use with video recorders, but off-the-shelf video transmission equipment and channels made the move to RF application fairly straightforward, albeit spectrally gluttonous.

More recent products in this arena have utilized the T-1 transmission standard, developed in the telecommunications industry for wideband digital signal distribution. This format's 1.544 Mbit/sec data rate places it in the right area for such an application, but the T-1 system is primarily intended for multiple voice-grade channels (24 per carrier), so there has been a motivation toward lower data rates, to keep the channel/carrier rate high. As a result, most of the telco T-1 hardware and service offerings for wideband audio have been of a 12-bit, 32 kHz PCM variety (although some 8-bit non-linear PCM products are also in use). Within the last two years, however, some new broadcast industry products have been introduced that provide 44.1 or 48 kHz, 16-bit stereo for T-1 applications, with correspondingly higher occupation of the carrier, leaving little room on it for much else. Besides standard telco T-1 paths, transmission of such links for STL, TSL, and RPU uses can be accomplished on wideband video subcarriers, direct or T-1 fiber-optic links, or 18 and 23 GHz microwave channels. Hardware for such systems is now available from Graham-Patten Systems (VAMP series) and QEI (CAT\Link), but occupied bandwidth is still 1 MHz or more for a single stereo pair.

The Key Role of the Listener

The laws of physics governing such data rates and their transmission are clear and unequivocal on these points, and that might be the end of the story, were it not for the fact that the end-user of such transmissions is not some piece of hardware, but rather the human listener, and thus psychoacoustics (the study of the human hearing sense) is involved. Research in this realm provides information that tempers the abovementioned limits to a great extent, and liberates the users of digital audio systems to employ far higher transmission efficiencies by drastically reducing data rates.

To understand the workings of these reductions, some basic concepts of psychoacoustics must be discussed, most notably the so-called Critical Band Theory, and the subject of Temporal Masking. Both refer to the perceptual threshold of the hearing sense, and their study allows a designer to understand what he or she can "get away with" when designing a data compression scheme along these lines. Critical Band Theory tells us that the ear works very much like a 24-band real-time analyzer, and that the perception of noise in each of the bands is subject to a high degree of masking by signals in that band. Thus, a data compression system that analyzes the presence and level of signals in each of these bands against its masking threshold can decide how much the resolution (number of bits) can be reduced in each band without listeners detecting the increased quantization noise that would result. Temporal masking tells us that the ear is insensitive to a sound that occurs within a certain time proximity to another sound roughly equal in level or louder. This phenomenon, also known as "premasking", allows error signals -- even relatively high-level ones -- to be spread within a given time window preceding the actual onset of a desired signal, and remain inaudible.

Current Results

Recent work has shown that systems employing these masking techniques (often referred to as "perceptual coding") can reduce data rates by dropping the resolution of the audio data, with little or no audible alteration of the audio signals. Quantization noise is certainly added to the signal, but these systems shape the noise in time and frequency such that it is masked by the original signal. Now, data rates of under 200 kb/s are conceivable for a digital stereo signal. This obviously puts the application of digital audio in a whole new light for the broadcaster, allowing a move to digital modulation techniques without a corresponding penalty in occupied bandwidth.

Typically, these compression schemes start with a well-converted 32, 44.1 or 48 kHz, 16-bit PCM signal, and reduce its resolution to some lower effective bit-rate. Notice that the operation reduces performance in the amplitude domain (i.e., dynamic range) only, and that the sampling frequency is unchanged. Thus the audio bandwidth and frequency response of the originally sampled signal is unchanged, as is its temporal resolution (speed/timing accuracy), both being desirable attributes of digital audio systems, since Nyquist limits still apply regarding sampled spectrum, and crystal-controlled clocks still operate the sampling and reconstruction of the audio waveform during conversion to and from the analog domain. For this reason, these systems are often referred to as performing data compression down to "n bits per sample"; among current devices, $1.45<n\leq4$ bits/sample (see below). Sampling rates also range from 32 to 48 kHz.

In order for these "bit-rate reduction" systems to operate practically, however, several specific criteria must be satisfied besides pure audio quality retention. First, the complexity of the algorithms for such data compression must be minimized; the several systems under development in this genre vary widely in this respect. For actual future broadcast use (i.e., digital transmission to the listener), low-cost receiver hardware is critical to any system's success. For today's point-to-point "first-mile" or network applications, this is somewhat less critical, although still desireable. Next, robustness of the signal must be considered. The error probability of some transmission paths may be higher than any type of current digital recording media, so error correction schemes and coding algorithms must take this into account. Finally, the coding delay that these systems introduce due to their often rather long buffers (100 ms or more) may present a problem for broadcast synchronization, real-time off-air monitoring, and related timing issues.

A BRIEF HISTORY

The first attempts at such data-rate reductions were made independently by several organizations in the U.S., Japan, and Europe. Most share one common trait: original analog audio signals are sampled and quantized by a high-quality digital (usually 14 or 16-bit PCM) converter, then processed through the data compression algorithm to provide the final, reduced bit-rate output. This output is in turn expanded back to 16-bit resolution after reception, and finally converted back to an analog audio signal for reproduction.

Early methods used variations on the standard PCM process, such as floating-point PCM, block companding, or adaptive differential PCM. All of these systems reduce data overhead by modifying the fixed quantization levels of true PCM, and adapting the dynamic range of the system to the needs of the moment. The first, floating-point PCM, applies the companding instantaneously, allowing the quantization levels to vary by a small amount constantly; this sort of system is in widespread use by telephone companies, and typically provides a bit reduction of 16 to 14, or 14 to 11. The other systems reassess the amplitude range of the signal over a given block of data (typically every 32 samples), redefining quantization levels in this "near-instantaneous" (NI) fashion, based on the maximum level encountered in the block. Standard PCM using block companding (as in the BBC's NICAM system) can provide a bit reduction of 14 to 10. Differential PCM was added to this approach by the Japanese (the NI-DPCM system), allowing reduction from 15 to 8 bits. Actual data output rates here range from around 700 kb/s to 1 Mb/s for stereo audio.

Delta Modulation

Next came the Adaptive Delta Modulation system (ADM) from Dolby Labs, the first to base its algorithm on serious psychoacoustic research. Unlike those before it, and most since, this system does not start with a PCM quantization step first. Delta modulation is a one-bit system that carries information about which way the sampled waveform is changing at that instant -- up or down. There are no discretely coded level addresses as in PCM. A form of DM had been used in a commercial system developed by dbx in the mid-80s as a professional pseudovideo digital processor, still in use by several stations as a recording system (with a VCR) or as an STL (on a video link). Dbx's so-called Companded Predictive Delta Modulation (CPDM) used analog companding and predictive algorithms to reduce the noise inherent in all DM systems, but was not intending the format to be a method of data rate reduction; rather their interest was in raising the sampling rate (CPDM used 700 kHz), to avoid the need for the brick-wall filters required by the Nyquist limits of competing pseudovideo PCM systems using 44.1 or 48 kHz sampling rates. In fact, they were largely successful at this, and developed a good-sounding, rather robust system, but one which still required a 6 MHz video channel for transmission. Dolby's version of DM takes a wholly different tack, and was designed as a method of low data-rate transmission with high audio quality and low decoder cost.

Starting with a 204 kHz sampling rate, Dolby ADM varies from standard delta-modulation by adapting the step-size of the DM data from simply up/down to a variable step size, and adds variable pre/deemphasis to the audio. Both the step-size and emphasis schemes are based on the noise-masking principles of psychoacoustics, and are also optimized for low distortion. (Larger step size increases quantization noise, and smaller step size can increase distortion; therefore, step size is kept as large as possible for a given audio level change [the slope of the input waveform], thus minimizing distortion, and variable, sliding-band preemphasis is added to compensate for the additional noise.) In addition to the delta-modulation audio data, the ADM datastream contains step-size data and emphasis control data as well. The pre- and de-emphasis occurs in the analog domain, but under digital control. Dolby ADM is already in use for television audio in an Australian B-MAC DBS system, and is also implemented in the Super-NTSC proposal for domestic ATV. It provides high-quality stereo audio with a data rate of 512 kb/s. The datastream is relatively insensitive to bit errors, since in a DM system, all bits are equal -- there are no most- or least-significant bits (MSBs or LSBs) as in PCM. Nevertheless, a simple error concealment system is included, to further desensitize the format from audible errors.

Frequency Domain Converters

More recently, frequency domain converters have been developed, again strongly based on critical band theory. Three basic types of converters are in current use: Subband Coders (SBC), Adaptive Transform Coders (ATC), and Optimum Coders in the Frequency Domain (OCF). These systems have been able to achieve the most effective data companding to date, and show the most promise for the future of this technology.

With subband coders, audio is first split into many bands, generally imitating the ear's critical bands, using cascaded quadrature mirror filters (QMF). Within these bands, quantisation levels are assigned as the signal energy in each band warrants, such that critical band masking will assure no noise is audible. In early systems, this assignment was a fixed process; current systems used a soft function here, with the adaptive bit allocation per band. (This technique was actually originated in the early 1970s at M.I.T., but seemingly forgotten until recently.) Solid State Logic recently introduced a subband ADPCM coding system called APT-X 100. The coder is available on the AT&T DSP16 digital signal processing chip, and is now also being delivered in card level (including A/D and D/A converters), and full hardware form with power supply and clocking; the latter is a 1-rack unit, XLR in/out device. The APT-X 100 system provides high audio quality at a data rate of 4 bits/sample/channel (256 kb/s for stereo). SSL is currently pursuing a 2 bit/sample system that may be included on the AT&T DSP16A chip. Decoders are relatively equal in complexity to encoders in these systems, and processing delay is relatively low.

The other two forms of frequency domain converters apply transform functions to the signal. This process may be thought of as taking 16-bit standard PCM ("time domain", i.e., quantized in amplitude vs. time) data, and every so often shooting a snapshot (transform "window") of the instantaneous "frequency response" (i.e., spectrum) of the signal. The amplitude of signal in each of the critical bands is then assessed, quantized and coded in various ways to produce a low data rate output, which can be reconverted to its original time domain (16-bit PCM) data form, and subsequently reconverted to analog audio. Due to the processing involved, delay through the system is longer here, ranging up to 100 ms. Generally, in these systems, processing delay is directly proportional to the amount of data compression accomplished.

Adaptive transform coders begin with a transformation (FFT or otherwise) of a block of data into the frequency domain, followed by assessment of the overall spectrum of the signal, bit assignment within frequency bands, and adaptive coding of those spectral values. Simpler and more robust "Low-complexity adaptive transform coders" (LC-ATC) are being developed in Europe at present. Stereo data rates here range from 300 - 500 kb/sec (3.5 to 5 bits/sample/channel), although lower rates are predicted to be possible. In fact, Dolby Labs introduced the first commercial product in this area late in 1989, in its Model 500 system, providing high-quality stereo audio (32 kHz fs) in a 256 kb/s output datastream.

Optimum coders transform blocks of data into the frequency domain as well (using a Modified Discrete Cosine Transform [MDCT]), quantizing and entropy coding the spectral coefficients produced by the transform through two fixed iteration loops. The loops insure that quantization noise will be masked but that available bits are not exceeded for any band. If necessary, one of the loops allows quantizer step size to be increased in order to reduce the bit count. In order to reconstruct the data, only the iteration count through the loops and the coded coefficients are required, making the decoder complexity far lower than the encoder. The data rates in these systems are the most promising yet, with high quality already possible at 2 bits/sample. Quite acceptable fidelity is has been demonstrated at 1.45 bits/sample with such a system, allowing monophonic transmission at 64 kb/s (the data used rate in ISDN paths, which will undoubtedly figure in future transmission strategies). Most notorious among this style of converter is the so-called "Brandenburg" system, after its developer (see References below).

SOURCE VS. CHANNEL CODING

All of the abovementioned systems are referred to in the jargon of digital broadcasting as "source codes," and are designed to function as data compressors only. The format in which a source code is actually transmitted is called its "channel code." While the source coding is intended to provide a relatively robust and low-data-rate audio signal, the channel coding furnishes a datastream format suitable for the RF domain, well-prepared for the additional adversity that a signal might encounter there. An additional consideration here is the form of modulation used, and the further efficiencies that it can offer, since some of the more exotic types (QPSK, high-order QAM, etc.) allow encoding of 3 or more bits/cycle, which compares quite favorably to standard FM's practical limit of 1 b/c mentioned earlier. (Digital technologies may make these forms of modulation more practical, reliable and cost-effective as well--see References below.)

It is worth discussing terminology at this point, since a variety of terms are coming into use in this area. The term bit/Hz has been employed to describe the data efficiency of a given channel code and modulation scheme. Although the term has become fairly widely understood in its intended meaning, the unit would be more accurately expressed as bits/sec/Hertz, since it is a measure of the transmission system's ability to apply a given data rate to a given carrier. Since it is the comparison of two RATES, however (Hertz = cycles/sec), it is proposed here that

$$\frac{\text{bits/sec}}{\text{cycles/sec}}$$

be reduced to

$$\frac{\text{bits}}{\text{cycle}},$$

and the term b/c be employed as a measure of such digital transmission effieciency. It is further proposed that a Digital Transmission Index (DTI) be implemented, wherein standard FM (using no special channel coding), with its generally assumed 1 b/c capability, be used as a reference (DTI=1), and other systems' nominal capabilities be compared to it. For example, 64-QAM can be considered to have a DTI of 4.5.

It should be noted that the attraction of such high-efficiency modulation techniques is countered by their lower power efficiency. Depending on the application, diminishing returns may be quickly reached.

Waveform Modulation

Another recent innovation from Tuscon, Arizona-based Genesys Technologies may hold some promise here as well. A clever and efficient direct source coding algorithm that is a hybrid of ADM and binary coding is combined with a unique channel coding system called "waveform modulation." The latter puts a "bump" in the sinusoidal waveform of a carrier wave, somewhere in the linear portion of the waveform between the peak and the trough. Depending on the exact location of the bump, a discrete data value is recovered, corresponding to an address in the source coding scheme. The channel coding is well-suited for use on an unrelated signal's carrier, making it possible for new signals to be added to existing formats without loss of compatibility. For this reason, Genesys is being looked at by developers of HDTV and other video applications. Psychoacoustic and psychovisual elements are incorporated into the Genesys data compression systems. The industry is not without skepticism regarding this system, along with the equally exotic "Slip-Code" system from Pegasus Data Systems of Middlesex, New Jersey, although both seem intriguing enough to warrant further exploration and testing.

Satellite Systems

In the area of satellite audio distribution, LNR Communications of Hauppage, NY has recently introduced an SCPC (single channel per carrier) digital system capable of transmitting four high-quality channels into 800 kHz of transponder bandwidth, comparing favorably with the channel space required by existing analog SCPC systems. Other systems are under development for network and direct-to-the-listener audio broadcasting with satellites, including some that intend to provide reliable mobile reception.

WHAT THE FUTURE HOLDS

Digital channel coding and modulation forms are being explored primarily in Europe at present, and one system employing Orthogonal Frequency-Division Multiplexing and Coding (COFDM) has been designed and tested favorably there by the French broadcast authority CCETT, in conjunction with the German IRT and the European Broadcast Union (EBU), under both fixed and mobile reception conditions. It will be tested in Canada in Spring 1990. COFDM's most notable advantage is its nearly total immunity to multipath, which also allows fairly free use of boosters for improving coverage without worry of destructive interference in overlap areas. Hardware cost and spectrum usage may be difficult to justify domestically in the very near term, but the advantages in reception quality seem to be significant enough to warrant serious consideration by several European countries at the state level presently.

The system employs a robust subband-style source coder (called MASCAM), then performs transforms and interleaving on the output data in both the frequency and time domains, using multiple carriers with QPSK modulation. Due primarily to the time interleaving of redundant data across these multiple carriers, time delay through the system approaches half a second. Current implementations have been set up for multiplexing of many audio services into a single wideband transmission system; such an approach may be critical to both the reception reliability and the spectrum efficiency of such a system, so our future may indeed hold some sort of enlightened "master community transmitter/antenna(s)" plan. The first COFDM implementations in Geneva, Switzerland and Rennes, France, put 16 stereo signals (plus one auxiliary datastream for 33 channels in all, each of 144 kb/s) on 7 MHz of occupied bandwidth in the UHF band. This works out to about 300 kHz per stereo signal. The total number of RF carriers used is 448, spaced at about 15 kHz. Each carrier runs 25 kb/s. Existing VLSI chips are used. Power efficiency of the system is astounding, quoted as about 50 times greater than current FM broadcasting. This might allow usage in the UHF TV band on frequencies used by (or first-adjacent to) neighboring-market TV channels or allocations. See references below for full explanation of this important and possibly influential system.

Another item to watch for is the possible revisitation of space-diversity reception techniques. These are beginning to find favor among high-end automotive FM systems (although not all current applications are adequately spaced), but may be an integral part of future digital receivers. Given the potentially higher carrier frequencies used, diversity reception could be easily implemented in both auto and other portable systems, since required spacing between antennas would be small.

Finally, an important caveat should be observed regarding the data compression algorithms (source codes) mentioned above, regarding their performance in a multigenerational application. How does an audio signal behave after it has passed through one of these data compression systems more than once? A typical example might be the use of such a system on a multi-hop STL, where the signal must be converted to analog between hops; or one in which it is used to feed a distant transmitter location as well as a remote off-air monitoring return path. (Thanks to Charles Lubrecht, C.E. at KUFM, Missoula, MT for bringing this issue to our attention.) Further, how do competing systems interact when used as successive parts of a single path?

It now seems certain that digital audio systems are about to become practical for broadcast applications. This is both appropriate and timely, if for no other reason than to compete with the higher audio quality already available to listeners in digital prerecorded media. Once established, it is hoped that broadcasters will view the digital age as a period of cease-fire and armistice in the audio processing wars, thereby restoring the original definition to the term "broadcast quality audio."

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CD PLAYER MAINTENANCE

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ABSTRACT

CD Players are quickly taking the prominent position in the radio studio. The radio engineer should keep in mind a few survival tips when relying on a CD player for the station's primary audio source. This paper will serve as a "crash course" on CD player maintenance, beginning with a general overview of a typical CD player, and a discussion of block diagrams and schematics. Also discussed is a general maintenance schedule, the correct procedure for cleaning the lens of the optical pick-up, and the Servo Section alignment procedure. This paper is not concerned with the theories behind CD player design (i.e. quantization levels, sampling rates) but rather with the real life situations involved in using a CD player in a professional environment.

INTRODUCTION

By the end of 1988, 42 percent of all radio stations were using CD players in one form or another. By the end of 1990, this figure could be as high as 75 percent. Now, radio stations are not only competing with their market stations, but also with audiophiles using sophisticated home and car audio systems with CD players. As more car manufacturers offer a CD Player as optional equipment, America's ears are getting fine tuned to superb quality audio. Listeners are more critical of audio quality than ever before. Why should they listen to mediocre quality broadcast audio when they have the option of compact disc in their homes and cars? The only way for broadcasters to compete with this format is to offer it, as well.

Today, radio stations are beyond the era of the disposable CD player. With the use of a permanent CD player in the Air Studio, maintenance and upkeep must become a regular routine. Engineers must learn about CD players and how they work. They must become as comfortable with CD players as they are wit tape cart machines. Because it is still a new format to radio, broadcasters are still learning what the requirements are. The CD player which will work perfectly well in the living room, playing perhaps 50 to 100 different CD titles, three hours a day, may totally collapse in the air studio. Radio stations are the ultimate testing ground for any piece of equipment. A factory environment cannot simulate the type of use, wide range of CD's, and handling which a CD player will get in the studio. Manufacturers are constantly learning what makes CD players work in the broadcast environment. After all, radio is the establishment for reliability and longevity.

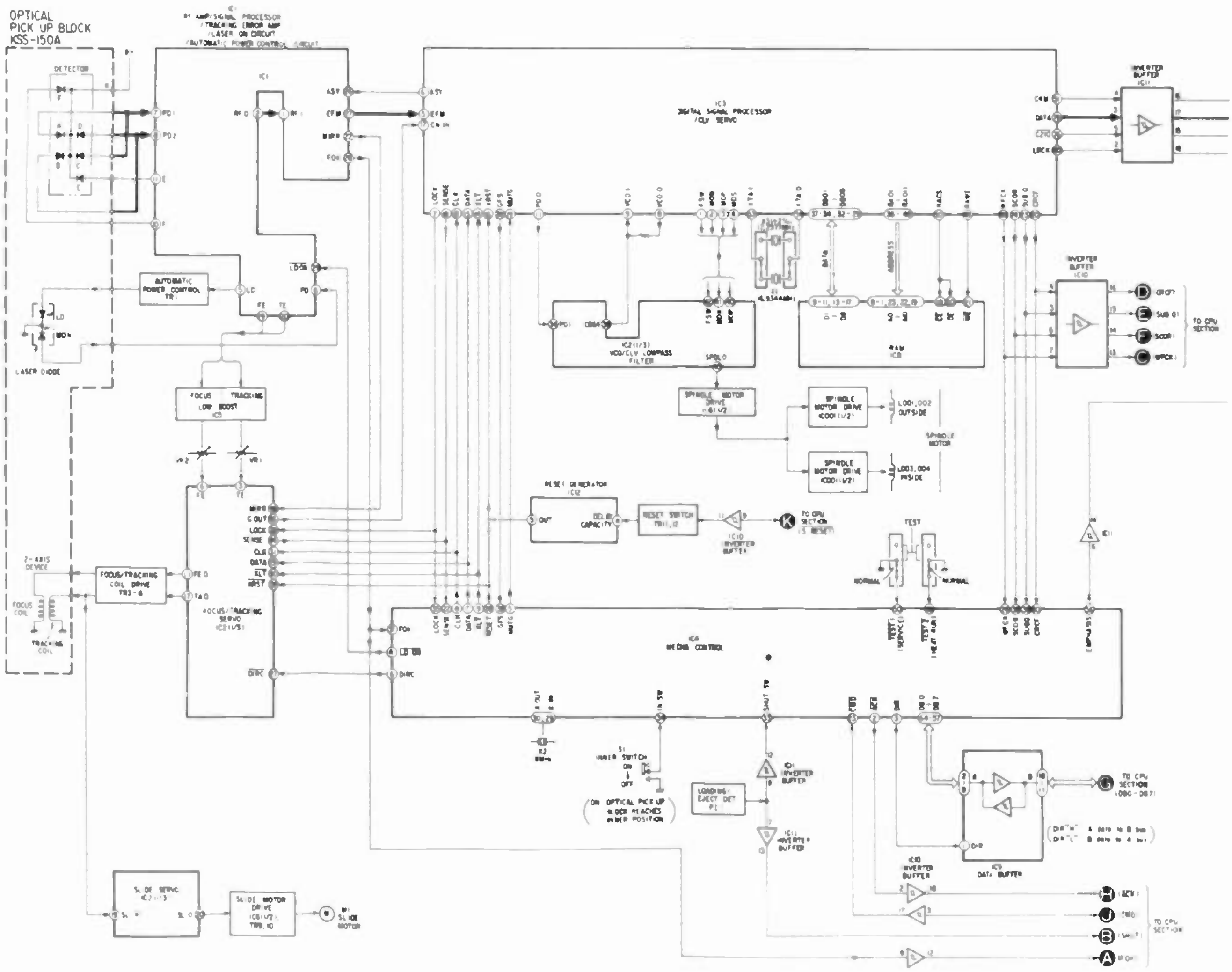
GENERAL OVERVIEW OF A CD PLAYER

To simplify things, consider a CD player as three main sections: The Audio, Servo and CPU sections. Figures 1, 2, and 3 illustrate block diagrams for each of these sections.

The Servo section of a CD player is the most critical area with respect to tracking, cueing, and general playback performance. The Servo section includes the optical pickup assembly, tracking and focusing circuits, slide motor and turntable motor controls. It also includes digital signal processing, and error correction. The optical pickup gathers the raw data, including subcode, error correction, audio, and timing information, and sends it to the Digital Signal Processor. The audio signal is separated in its digital form and sent to the Audio section. Subcode, timing information, and error flags are routed to the CPU section.

The Audio section usually contains the digital filters, D to A converters, low pass analog filters, and buffer amps. The audio response of the CD player mostly depends on the Audio circuit from the D to A converters on out. Any defect on the disc large enough to cause audible muting will also affect the audio quality. The CPU acts as the master controller between the Servo and Audio sections. It contains the master clock, data buffers, display information, and muting commands. Other than any software updates offered by the manufacturer, the CPU section requires virtually no maintenance.

Figure 1. Block Diagram of Servo Section



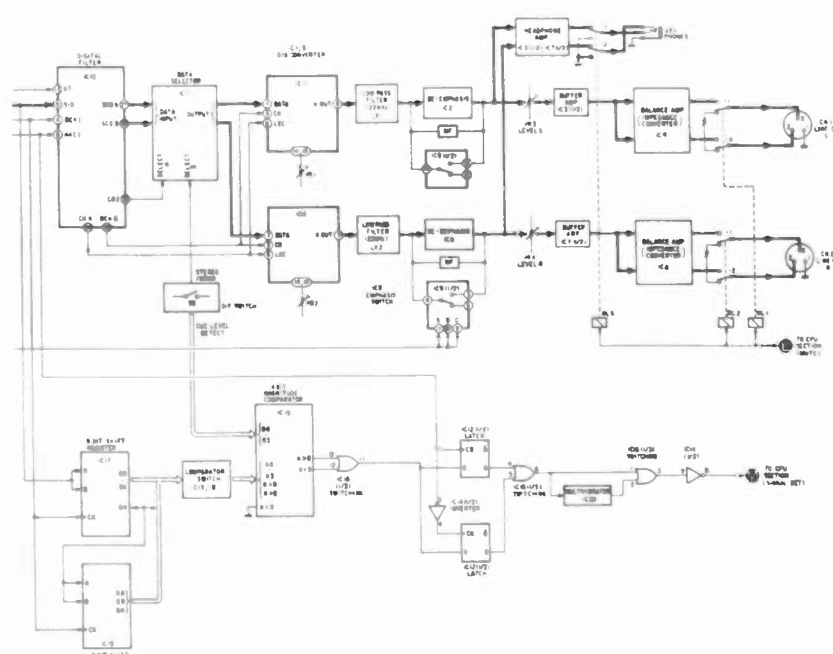


Figure 2. Block Diagram of Audio Section

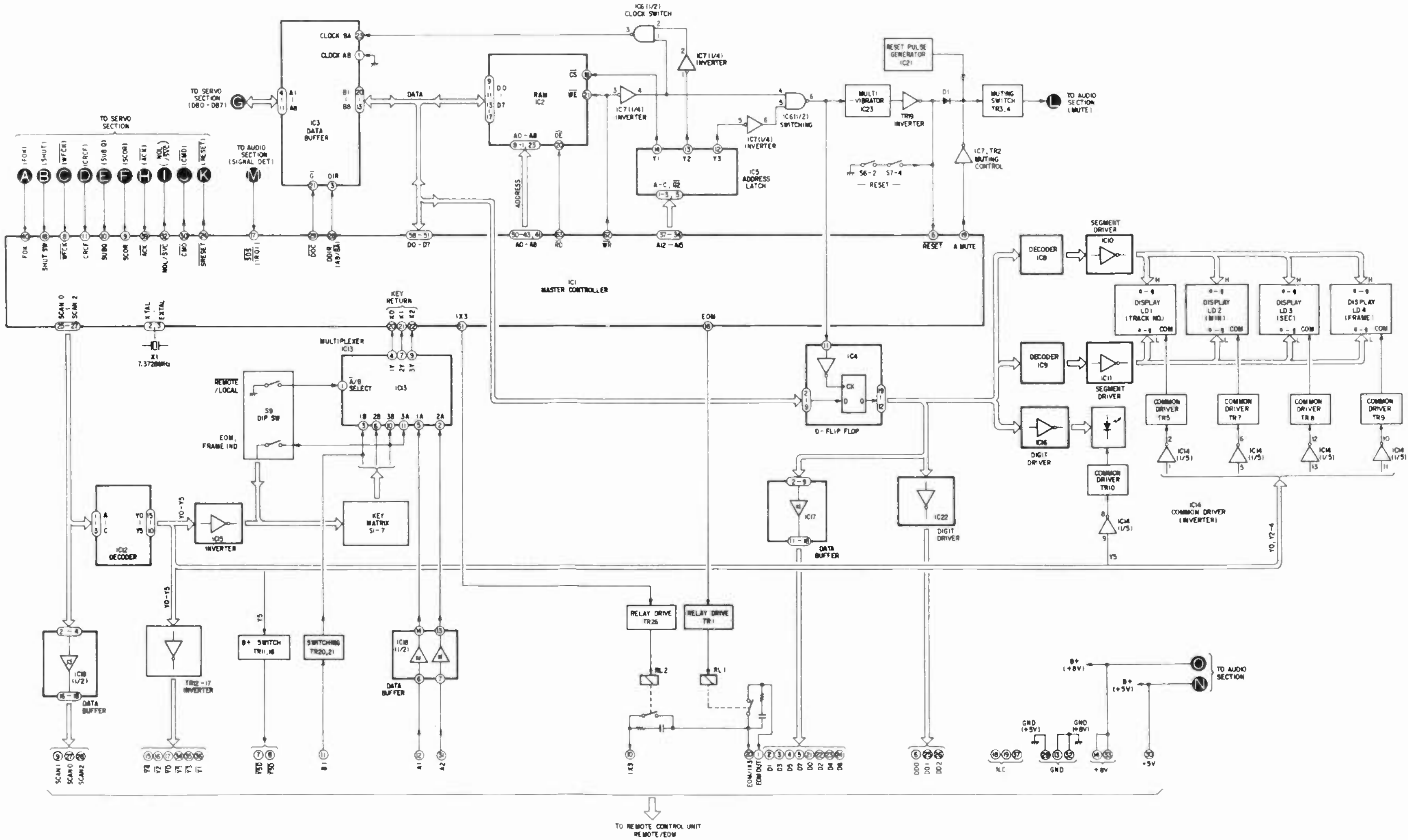
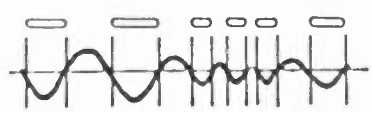


Figure 3. Block Diagram of CPU Secion

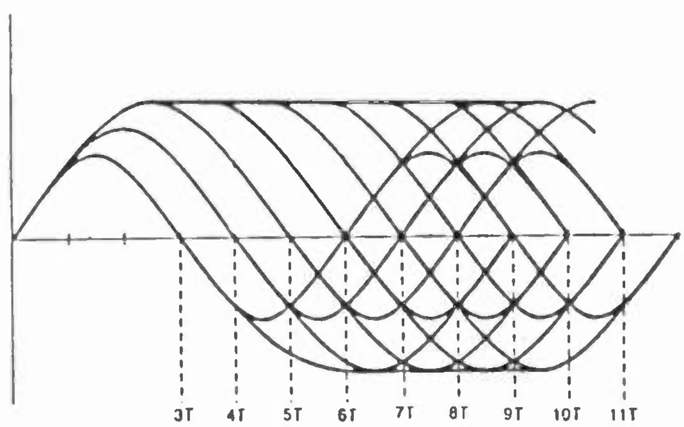
THE OPTICAL PICKUP

The optical pickup is the most important element of any CD player. (It is also the most expensive.) The CD player relies on the accuracy of its information gathering ability to reproduce audio correctly.

The optical pickup functions as an opto-mechanical device. A tiny laser beam emits from the source diode, passes through a series of lenses, hits the disc and reflects back. A quadrant of photodiodes located within the optical pickup assembly reads the reflection from the disc as a stream of 0's and 1's. The photodiodes interpret no change in the surface of the disc as a 0. They interpret a change between land and pit, or pit and land as a 1. The steady stream of all 0's represents no change in voltage. A switch from a 0 to a 1, or a 1 to a 0 represents a change in voltage. The seemingly random swings between 1's and 0's creates a sine wave of irregular frequency. When viewed in a longer time domain, these sine waves layer up to produce the Eye Pattern.



Data pits on the disc create a represenative RF signal



A collection of RF signals forms the Eye Pattern

Figure 4. How Pits on the Disc Create the Eye Pattern

Imagine driving up and down large hills in your car with the cruise control set at a comfortable 40 miles per hour. A magnetic monitor watches the rotation speed of the wheel of the car, and sends the appropriate "speed up" or "slow down" information to the accelerator. As the car goes uphill, the wheel slows down. The monitor notices this, and tells the accelerator to speed up. As the car reaches the top of the hill and starts down, the wheel rotation increases. Again, the monitor notices, and tells the accelerator to slow down. Of course, this is a very crude example, but if you can understand it, then you can understand how the optical pickup performs tracking and focusing.

The auto focus and auto tracking of the CD player is a quite a feat. Since the focusing and tracking servo commands are derived from processed information obtained optically, the pickup does not need to make any physical contact with the disc. The lens of the pickup, however, will physically move from side to side and up and down in order to maintain correct focusing and tracking.

In focusing, the reflecting beam from the disc passes through an astigmatic lens, and creates a circle on the quadrant of photodiodes. If the lens gets too close or too far away from the disc, the circle becomes skewed one way or the other. The compared signal from the photodiodes results in the Focus Error signal. The Focus Error signal commands the Focus Servo to magnetically pull the lens back into position, to maintain proper focus. While the difference signal of the photodiodes creates the Focus Error signal, the sum of all four photodiodes creates the HF signal. The HF signal contains all the audio data.

Most CD players use a three beam system to maintain tracking. A three beam pickup really has only one laser beam, which passes through a beam splitter, resulting in a primary beam, and two secondary, or side beams. The primary (center) beam reads the data track on the disc. The two side beams carefully monitor the edge of the data track. In addition to the quadrant of photodiodes, the optical pickup contains two other photodiodes dedicated to measuring tracking error. As the pickup drifts to one side or the other, the tracking photodiodes measure a difference in returned light intensity. This difference signal becomes the Tracking Error signal. The TE signal commands the Tracking Servo to magnetically pull the lens of the pickup back in line.

A set of magnets and coils controls the physical movement of the optical pickup lens. The lens has attached to it a set of Tracking coils and a set of Focus coils. The strength of the voltage passing through either set of coils determines the movement of the lens around the two magnets.

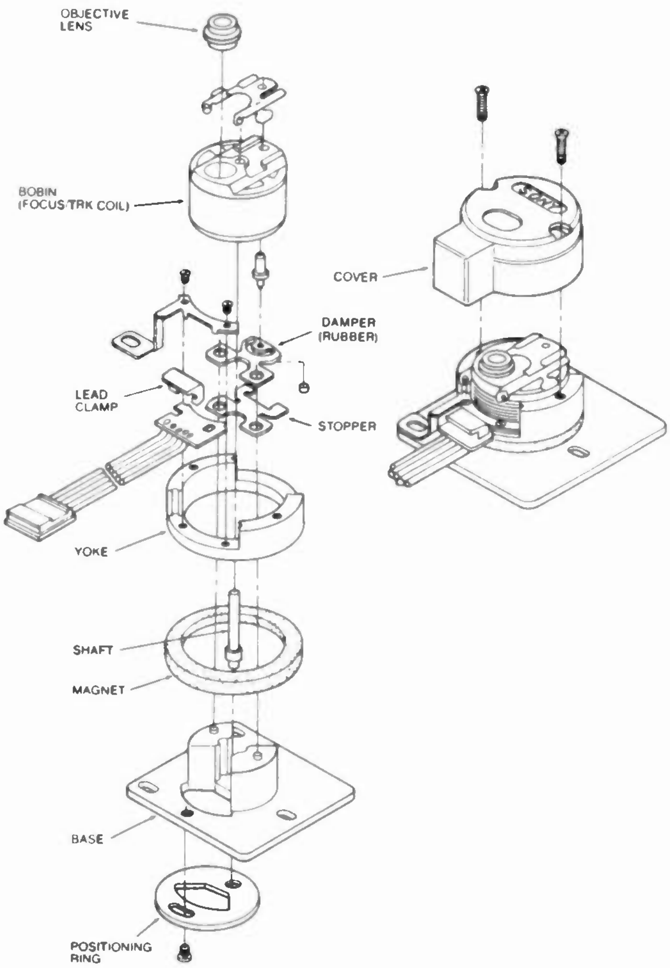


Figure 5. Exploded view of an Optical Pickup

Understanding the complexity of the optical pickup structure will enable you to understand how CD players work, and why it is important to follow maintenance guidelines. The *theory* of how the pickup works is relatively simple, however the pickup itself is extremely complex and must adhere to tight tolerances.

MAINTENANCE SCHEDULE

Consider the CD players in the air studio as important elements in the overall sound of the station, and treat them with respect. This means following standard maintenance routines. Most manufacturers do suggest cleaning of the lens of the optical pickup on a periodic basis. Others outline elaborate alignment procedures of the Servo and Audio sections of the player. Follow the manufacturers reccomendations as to when and how these alignments should be done. If in doubt, check with the manufacturer. They should be able to provide you with basic service suggestions.

CLEANING THE LENS OF THE PICKUP

To minimize tracking and focus errors, the player must gather optical information as efficiently as possible. Any dirt or scratch on the disc will make this task more difficult. A scratched disc will exercise the threshold of the player's capability. It may play, or it may skip or stutter. Keep disc surfaces as clean and scratch free as possible. Also, clean the lens of the optical pickup on a regular basis. Most CD players have the lens facing up, a target for dust accumulation. When a thin layer of dust covers the lens, the basic stream of 0's and 1's from the disc becomes jarbled. When the laser beam passes through the lens the first time, the dust reduces its intensity. The beam then reflects off the disc, passes back through the lens, and is reduced by the dust again. Finally, a quadrant of photodiodes measures the intensity. By this time, the resulting signal will be low and difficult to process. The likelihood of an error increases. The CD player may begin skipping intermittently.

The environmental quality of the air studio will dictate how frequently the pickup lens should be cleaned. Do the jocks smoke in the air studio? Is the studio located in a dry, dusty building, in the middle of Arizona? Is an air filter or humidifier used? Is the studio in a city of high smog levels?

In general, clean the lens of the optical pickup on a monthly basis. The best way to clean the lens is as follows:

1) Always turn the power to the player off before removing the top cover. Disconnect the power cord to protect yourself from electrical shock. As long as you remove the power cord, there is no danger of damaging your eye from the laser beam. Most CD players have an interlock switch, which means the laser beam is only on when a disc is in place.

2) Slightly moisten a cotton swab with either alcohol or camera lens cleaning fluid. Standard isopropyl alcohol is fine. Denatured alcohol is fine, too. The lens will, however, probably experience a natural death long before the accumulation of impurities left over from isopropyl alcohol damages it. Water will work to some extent, but alcohol will evaporate much quicker. Alcohol will also cut through sticky nicotine and pollutants floating around in the air.

3) Press the cotton swab on the back of your wrist to eliminate excess fluid. Gently rub the lens in a circular motion. You may notice very little difference in the appearance of the lens before and after cleaning, unless the lens is extremely dirty. In the worst case, the lens will appear slightly dull or cloudy before cleaning. After cleaning, it should have a near liquid appearance. The reflection of the overhead light showing up in the lens should appear sharp and clear.

Periodically blowing off the lens of the pickup with compressed air won't hurt, but it doesn't actively cut through the greasy film which may have coated the lens. To be sure, clean the lens with a solvent cleaner on a periodic basis.

One way of measuring the effectiveness of the lens cleaning is to monitor the HF (High Frequency) signal of the player during playback. This resulting wave form is frequently referred to as the Eye Pattern. Use the NAB Test CD, Track 1, as the source. Connect an oscilloscope probe to the point in the servo section circuit board labeled HF or RF. This wave form should show up somewhere just following the optical pickup in the schematic. Set the scope to 50 mV, 0.5 microseconds per division, and use the probe set at 10:1. You should then see the Eye Pattern.

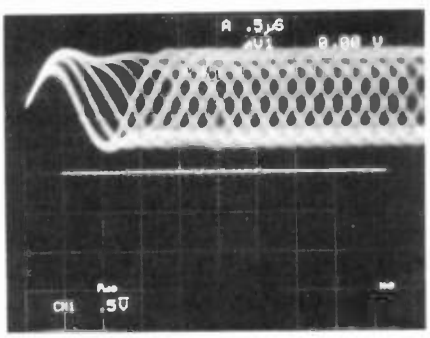


Figure 6, Eye Pattern

Measure the amplitude of this waveform. Clean the lens, and measure the amplitude again. You can consider the increase in amplitude as factor of effectiveness. If you noticed no increase in the amplitude of the Eye Pattern after cleaning the lens, then it probably was clean in the first place. If the amplitude increased by more than 0.30 volts, then the lens was quite dirty, and should have been cleaned sooner. After following this procedure on a monthly schedule, you should get an idea of how frequently the lens needs to be cleaned, according to your specific environment.

REPLACEMENT OF THE OPTICAL PICKUP

The optical pickup has a limited life span, which is rated by the manufacturer. Typical ratings are 5,000 to 7,000 hours of actual use. Experience shows the pickup will probably last about a year in a station using 3 On Air CD players for constant playback of their music. If more CD players are used, the pickup will probably last longer, as a result of the decreased work load. Factors which decrease the life of the optical pickup include changes in ambient temperature, air pollution, and smoke. The laser diodes within the pickup assembly will normally deteriorate with time. As a result, the laser intensity will decrease.

Upon close inspection of the optical pickup assembly, you may notice a tiny variable resistor. The manufacturer of the pickup assembly uses this VR to set the laser intensity prior to installation in the CD player. Many manufacturers do not encourage readjusting the laser intensity in the field. If you suspect the pickup intensity has decreased, it probably needs to be replaced. Also, the pickup may undergo physical fatigue, beyond any improvement by alignments. In that case, it should be replaced. Has it been in use for more than a year? Do the jocks smoke in the air studio? Has the CD player begun skipping intermittently for no apparent reason? Does the player not respond to any alignments? If the answers are yes, the pickup probably needs replacing.

THE ALIGNMENT PROCEDURE

AUDIO ADJUSTMENTS AND MEASUREMENTS

The measurements of the audio section of a CD player require a dual mode oscilloscope and a test disc. Use the NAB Test CD for accurate audio performance measurements. Use this disc to measure audio output level, channel phase error, harmonic distortion, frequency response, high frequency separation, signal to noise ratio, and D to A converter linearity. With the NAB Disc, the player becomes a self contained signal generator. With a given input (the source material contained in the disc), you can calibrate the output of the CD player to exceptional accuracy. Then, use the CD player with the NAB Test CD as a signal source to measure the response of any other piece of gear in your studio.

The pure audio sound quality of a CD player depends mostly on the D to A converters and everything following in the circuit. Many D to A converters have an adjustable Most Significant Bit. Adjusting the MSB removes the distortion at the zero voltage cross point. As a result, D to A converter distortion during low level audio passages is reduced. This type of distortion, totally unlike its analog counterpart, is only present during extremely low level audio passages, and sounds similar to static. This is the only adjustment in the Audio section which will truly improve the audio sound quality. Of course, you can adjust the output level of the audio, but this will not affect the sound quality. Keep in mind the adjustment of the D to A converters, or any other improvements in the audio circuit, will not have any effect on tracking or cueing problems.

SERVO ADJUSTMENTS

The audio performance, i.e. lack of skipping and stuttering sounds, depends largely on the condition of the optical pickup, and the Servo section alignment.

The Servo Alignment is probably the second most important maintenance procedure for the CD player (behind cleaning the lens of the pickup.) It should be done after the first 4 to 6 months of a CD players life. Think of it as a 3,000 mile tune up. During the first 6 months of the players use, the optical pickup will go through a breaking-in and settling period. It will physically shift. It may droop down a bit lower than before. With constant use, the Focus and Tracking coils within the pickup will change slightly. The coils will loosen up. It is extremely difficult and painstaking to recalibrate the height of the disc turntable to maintain lens-to-disc distance integrity. It is just as difficult to get a precise mechanical adjustment for tracking.

A much easier method is to electrically coax the lens back to its original position. The bias current to the focus and tracking coils can be increased just slightly, to pull the lens back to its original position. The Servo Alignment allows this. Don't wait for your CD players to start skipping intermittently before doing the alignment. Instead, plan on doing the alignment about once every 6 months of the player's life.

Also, always perform the entire servo alignment after replacing the optical pickup in a CD player. The Servo Section alignment compensates electrically for minute physical differences from one optical pickup to another. Remember, the optical pickup is a mechanical device, and requires the appropriate treatment.

PREPARING FOR THE SERVO ALIGNMENT

Most of the Service Manuals for broadcast CD players outline the Servo Alignment in detail. Follow the directions closely for the CD player you are working on. If for some reason you do not have the Service Manual for the specific CD player you are using in the air studio, then by all means get it.

Each manufacturer calls for a specific disc for the Servo alignment. Use this disc. Do not use your own disc. The reference disc represents a disc of average thickness, reflectivity, error rate, etc. It is not especially good quality. It is not especially poor quality. It's just average. When doing the alignment with the reference disc, you can be sure your results will be similar to the results as achieved by the manufacturer. If the manufacturer does not specify an alignment disc, then use the NAB test disc. Every radio station should have it on hand, and it suits the purpose.

Other equipment required for the Servo alignment include a Dual Mode oscilloscope, a Frequency Generator, and a Frequency Counter. Some manufacturers offer test jigs, interconnecting cables, or other equipment which makes the alignment procedure easier. Check with the Service Manual, and your dealer to see if you really need these.

HELPFUL HINTS

Rather than learn by trial and error, here are a few pointers to keep in mind when first attempting the Servo alignment:

1) Allow at least 45 minutes the first time you attempt the Servo alignment. Allow time to learn your way around the Servo Board. The next time will be easier.

2) Always do the Servo alignment with the CD player at normal operating temperature. Don't let it sit and get cold overnight. You may find the PLL frequency seems especially sensitive to heat.

3) It it helpful to have the phone number of the local factory rep or service department on hand. Perhaps you will find something slightly unusual, and need to get a second opinion. It always helps to talk to someone who has been through the alignment many times, and knows the particulars.

4) While doing the procedure, try to eliminate any distractions. Have your receptionist take phone messages. Try not to stop halfway through the procedure. You will find it tricky to pick up again, and might as well begin from the start.

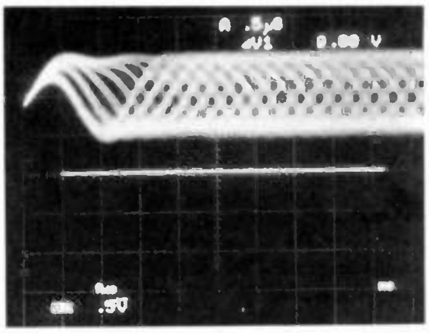
5) Always clean the lens of the optical pickup before starting the alignment, no matter what! One engineer unknowingly put a greasy smudge on the optical pickup lens before starting the alignment, and was horrified at the quality of the waveforms. After cleaning the lens, everything was back to normal.

THE SERVO ALIGNMENT

The servo alignment typically begins with the Phase Lock Loop Frequency Adjustment. The PLL adjustment is the most fundamental adjustment of a CD player. It sets the master timing for the entire player. In theory, the PLL frequency is a multiple of, or closely related to 4.3218 MHz, which is the bit rate of a compact disc. In practice, the PLL is usually specified somewhere between 4.25 and 4.33 MHz, depending on the CD player. Each model is different. Check the Service Manual to be sure. The PLL circuit should be indicated on the schematic, with the appropriate variable resistor. Set the PLL frequency when the player is at its normal operating temperature. Use a plastic screwdriver for best results.

After setting the PLL frequency, make coarse adjustments of the Eye Pattern and Tracking Error. For the Eye Pattern adjustment, set up your scope to monitor the HF as if you were cleaning the lens of the pickup. Find the test point labeled HF, RF, or EFM. EFM stands for Eight to Fourteen Modulation, the code process in which digital information is stored on the disc. Set your scope to 50 mV, 0.5 microseconds per division, with a 10:1 probe. Adjust the Eye Pattern for best clarity. Adjust for cleanest X intersections in the waveform. This may end up as highest peak to peak amplitude, or close to it.

Incorrect Waveform



Correct Waveform

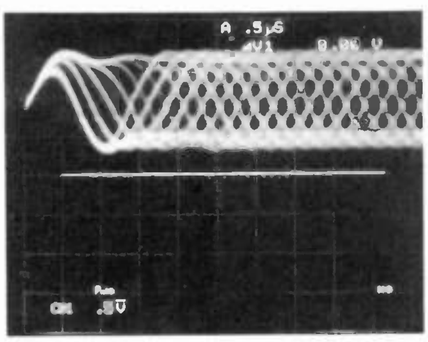
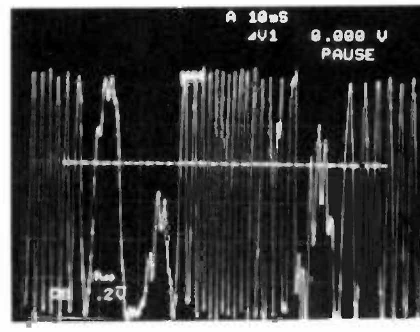


Figure 7, Eye Pattern Adjustment

After completing the Eye Pattern adjustment, set the Tracking Offset or Tracking Balance. Many people have performed a similar type of adjustment to the Tracking Offset. Most consumer VCRs have a Tracking adjustment knob. This allows the operator to align the playback head according to a specific tape. This adjustment is comparable to the Tracking Offset or Balance adjustment in a CD player. Fortunately, the Tracking Offset does not need to be adjusted for every disc. The Tracking Offset adjustment involves monitoring the TEO (Tracking Error Out) waveform, and adjusting for zero DC bias. Before connecting the scope probe, set the input of channel to Ground, and align the beam to the center of the scope display. Then reset the channel for DC, and connect to the TE signal, occuring before the Tracking Servo IC. Set the variable resistor so the bottom half of the TE is equal in amplitude to the top half.

Incorrect Waveform



Correct Waveform

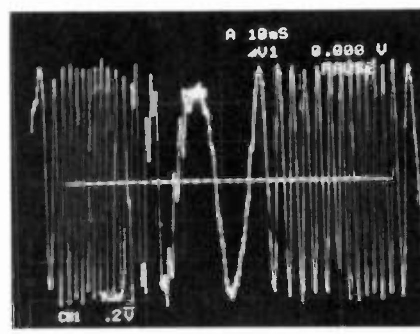
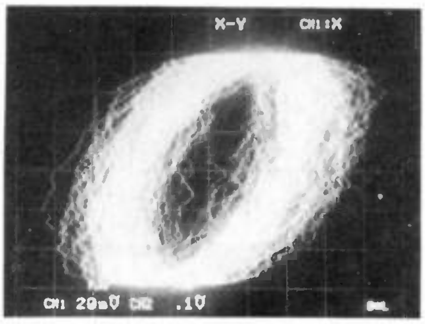


Figure 8, Tracking Offset Adjustment

The Focus Gain adjustment controls the gain of the Focus Error signal. The adjustment involves the phase comparison of the focus error signal just after the low boost amp, and the focus error just prior to the focus servo circuit. Compare the phase difference of these two signals, using sine wave as an reference input. You should detect a 90 degree phase difference between the two resulting sine waves. If checking the adjustment with the oscilloscope set in the X-Y mode, you should see a Lissajous figure.

Adjust the variable resistor for 90 degrees phase difference between the X and Y channels of the scope. Watch for distortions such as vertical or horizontal movement or shuddering of the waveform. Watch for a "breathing" of the Lissajous figure. In theory, you should see a moderately noisy Lissajous figure, which does not jump around in the scope. Any deviations could indicate a mechanical problem with the CD player or the optical pickup.

Incorrect Waveform



Correct Waveform

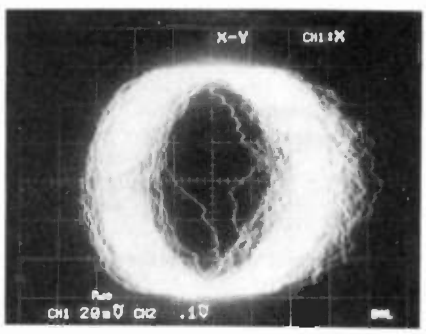
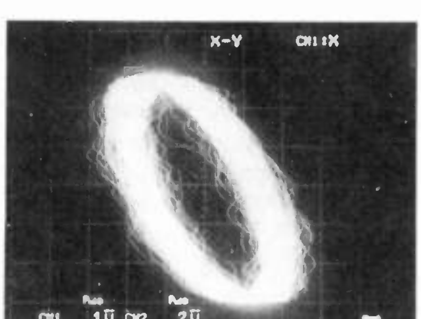


Figure 9, Focus Gain Adjustment

The Tracking Gain adjustment is similar to the Focus Gain adjustment, except it affects the tracking servo. Compare the TE signal just after the signal boost amp and before the Tracking servo circuit, using a reference sine wave as an input to the Tracking servo loop. All of the suggestions of the Focus Gain adjustment apply to the Tracking Gain adjustment, as well.

Incorrect Waveform



Correct Waveform

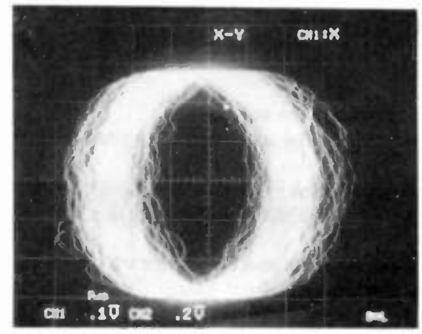


Figure 10, Tracking Gain Adjustment

Keep an eye out for any distortions of the Lissajous figure. You should be able to make an adjustment to the VR to obtain 90 degree phase difference. In the case of Focus and Tracking Gain, it is important to follow the specific instructions for performing this adjustment as suggested by the manufacturer.

MORE HELPFUL HINTS

The proper sequence to perform each adjustment should be described in the Service Manual. The PLL frequency adjustment will have no affect on the results of the other adjustments. It can be done first. As far as the other adjustments, try setting up course adjustments of the Eye Pattern and Tracking Offset. Then, do the Focus and Tracking Gain. Finally, check the Eye Pattern and Tracking Offset once more. Typically, they will not change after the Gain adjustments are made, but it doesn't hurt to check.

After doing these adjustments several times, you should get an idea of what is normal, and what is not. Look at the waveforms on a player which is functioning normally. Note how they appear. Measure them.

Take pictures of the waveforms. Then, if a player starts acting up, go through the alignment again, checking the waveforms very carefully. You may be able to determine the exact problem with the player just by studying the waveforms. This method becomes extremely handy in troubleshooting a player with intermittent skipping problems.

Always try to have an arsenal of test and torture discs on hand. A few of the better ones include the Pierre Verany Test Disc #2, Abex #713 and #712, and the Technics Black Band disc. Each of these have a built in physical defect, which acts as a challenge to the tracking servos of any CD player. While these "factory" made tests disc work well, sometimes they are not quite difficult enough. It is always a good idea to collect a few "natural" test discs, as well. Ask the

Music Director to give you any reject discs, or any discs manhandled by the jocks. Record company promos make especially good torture discs, as they tend to be slightly less than perfect, anyway. These are the CD singles placed on the full, 5" disc size, containing only 1 or 2 cuts. For some reason, radio stations seem to report more problems with these discs than store bought discs. Look at their waveforms on a known calibrated CD player, and you may see some deviation.

After becoming familiar with the Servo alignment, you might want to try a few variations. To troubleshoot a difficult player, set up a Before and After scenario with the alignment. Before starting the alignment, test the player with your torture discs, and note the results. Then go through the alignment, noting carefully the condition of the "before" waveforms. Make the adjustments as required. Then go back to your torture discs, and run the same tests. Look at the wave forms from those discs. How does the player respond?

Also, go through the alignment first with the factory approved reference disc, as suggested by the Service Manual. (If the Service Manual does not specify a disc, then use the NAB Test disc. It seems to be a good average.) Note the settings. Then, grab a handful of discs out of the music library, and do the alignment using them. How severely do the settings deviate from the reference position? You will probably notice the most differences in the two Gain adjustments, and perhaps in the Eye Pattern.

OTHER MAINTENANCE

Besides the lens cleaning and servo alignment, CD players require very little maintenance. You may want to inspect the slide rails which the pickup moves on. Clean these with a cotton swab, and remove any lint or hairs. It is amazing how much stuff makes it into a piece of audio gear in a broadcast studio! Remove any paperclips, chewing gum, and donut crumbs. Make sure the slide path is clear from debris. A single drop of oil (not grease) on the slide rails once a year will keep things running smooth.

Call the manufacturers service representative for other hints. Surely he can tell you all sorts of insider information about the CD player you are using. He will know exactly what the maintenance requirements are, and may point out specific hints. Perhaps he will know a quick solution to the tiny problem that's been haunting you for the last month.

RECORD KEEPING

Keep a thorough discrepancy log on each CD player in the Air Studio. Have the jocks record any problems they notice. Make it easy for them. Provide them with a form which they can simply check boxes off to save time. Assume (unless they are very well trained) the jocks will not record any discrepancies unless it is painless for them to do so. Use a chart similar to figure 11.

```
Date__________  Time______________  Initials_______  CD Player
#___________
Disc #______  Track #______  Time on Display when problem
occurred__________
What Happened?   Check all which apply:
_____  Wouldn't cue to track
_____  Made stuttering sound  ("I wanna h-h-hold your hand")
_____  Skipped forward                   _____  Skipped backward

_____  Jumped to another track           _____  Stopped in middle

_____  Noise in (Right, Left) Speaker    _____  Disc wouldn't unload

_____  Other (Please describe in detail):
___________________________________
____________________________________________________________
____________________________________________________________
____________________________________________________________
```

Figure 11. Discrepancy Log

When a tracking problem occurs, there is equal chance the problem could be the player or the disc. Log all discrepancies onto a chart format which allows comparison by player serial number, disc number, and track number. For even more efficient troubleshooting, enter the logs into a data base, then sort by serial number, disc number, and track number. Perhaps you will learn CD player #3 only skips on high track numbers. Also, a problem disc will quickly surface if it's a repeat offender. The only way to eliminate the disc as the problem source is to try it in the same player, with the same track number. Does it skip again? If so, the disc is probably flawed. Try it in a different player for a second opinion.

Not all CD players respond the same way with a bad disc, however. Try playing Track Number 40 of the Pierre Verany Disc #2 in your CD player. Some will track it, others not. Just as a bad CD player may not fail with every disc, a bad disc may not cause a failure with every player. A consistent, informative log is critical to solving intermittent skipping problems. Not all CD players are created equal. Not all discs are the same, either. Go back to the servo alignment, and use the NAB Test CD. Then do the alignment on the same player with a disc from your music library. Note the differences in the waveforms. Try another disc. Each one "looks" slightly different to the CD player.

SPECIAL CONSIDERATIONS

CD players are very susceptible to smoke. The nicotine from the smoke gets inside the delicate optical pickup assembly. Then, microscopic dust particles adhere to the surface of the lens, along with everything else. The lens surface can be cleaned once a month, but there is no way to clean the rest of the pickup. The focus and tracking coils get clogged. The pickup will experience a shortened life. The nicotine will also turn the slide gear lubricant into a sticky, brown mess. Things will go wrong. Prohibit smoking in any room containing CD players. Otherwise, count on replacing your optical pickups about every 6 months.

Avoid placing your CD players near a heat source. Do not mount them directly above tape cart machines. If you do, plan on using fans to keep the players cool. CD players will not tolerate high temperatures like tape cart machines will. The optical pickup is sensitive to heat, and especially, to changes in ambient temperature.

SUMMARY

This paper has explained some of the care and maintenance requirements of CD players. Perhaps at this point, some broadcast engineers are thinking, "I shouldn't have to go through all this!" Keep in mind CD technology is relatively new, and CD technology applied to the rigors of broadcast is even newer. It may take a few years for CD players to have the reliability tape cart machines boast today. But remember, were tape cart machines as reliable when they first came out 20 years ago as they are today? Both CD player and disc manufacturers are constantly learning new things to make their players and discs better. The format will only improve with age. What was inconceivable a year ago is a reality today. When handled with the right amount of respect, CD's on the air will lure new listeners and keep ratings impressive.

ACKNOWLEDGEMENTS

The author wishes to thank Ken Pohlmann, University of Miami; Ron Young, Century 21 Programming; Martin Ledford, Denon Digital Industries; Jim McGuinness, Denon America; Shigeo Todoroki, Nippon Columbia; and Jim Mercer, JMCS; for their help and shared knowledge.

DISCLAIMER

This paper encompasses all of the experience and knowledge I have gathered in over two years of working in Technical Support at Denon America. I have learned by both formal training as well as trial and error. I have learned what works, and what doesn't. I have looked at many different CD players from many different environments, and noted results engineers have reported to me.

I tried to make this paper as generic as possible. After studying the Service Manuals for several broadcast quality CD players, I have attempted to explain this information in a logical way, regardless of which brand CD player you use. Unfortunately, it seems each manufacturer has a different name for the same adjustment. In some cases, test points are labeled in the schematic, in others, not. The main point of this paper, however, is not to tell you *how* to adjust your CD players, but to let you know you *can* adjust your CD players. I have found many radio engineers consider CD players as totally non-field serviceable. They seem to think CD players can only be repaired or serviced by highly trained, specialty technicians with names like Ishihara or Watanabe. This is not the case. I hope you are encouraged to learn more about your CD players after reading this paper.

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Teac Professional Division, Service Manual for Tascam CD-501

BUILDING AND OPERATING A MULTI-PURPOSE REMOTE STUDIO VEHICLE

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This paper embodies the design considerations and the construction of a multipurpose remote and promotional vehicle.

It documents how a motor home was converted into a fully self sufficient broadcast facility, with the ability to keep an AM and an FM station on the air following a natural disaster.

The vehicle uses a variety of transmission paths to return programming to the studio in both stereo and monophonic modes.

A 'narrow-deviation' microwave link is used as the primary transmission system. Detailed herein are a series of novel steps involved in aiming the parabolic antennas at both ends of the transmission path.

Included are Block diagrams of signal paths, the floor layout and primary power wiring.

Today's remote broadcasts consist of more than just taking a mike mixer out on the road. Stations are expected to offer a 'presence' -- and to take advantage of being away from the main studios to mix and interact with our listeners, and to reinforce their loyalty as an audience.

Remotes have changed. A number of alternatives to the traditional card table and mike mixer have become popular in recent years.

The fiberglass 'boom-box' which resembles a giant radio on wheels, certainly establishes an identity. It also lacks comfort and may present an appearance not in keeping with your format.

Further, trying to do an extended radio show in a space that small could affect program flow and on-air performance.

The KIIS Starcruiser was designed and built with both the convenience of the air personalities and the quality of the transmitted audio in mind. We needed a vehicle that would provide not just the comfort and room we needed, but which would also offer a strong platform for a forthcoming satellite up/downlink system. And we wanted it to make a 'positive' statement no matter where it went.

We chose a 33 foot long Fleetwood 'Eleganza' motor home, and refinished the vehicle, inside and out. To help KIIS have the 'image' we sought, the interior cabinets were all made with solid oak and grey formica. The outside was coated in Imron enamel - a special finish that allows fiberglass to take a high gloss and which stays flexible throughout its lifetime. It was applied by a company that specializes in finishing aircraft to provide the quality we wanted.

A major part of the cost justification for this project was the ability to keep KIIS on the air after an earthquake. The main studios are in a tall office structure - and the building will be inaccessible after an earthquake, until it can be checked for structural integrity. This could take quite some time, and if we weren't allowed in the building, we would have no place to originate programming.

Addressing this, the Starcruiser has the capability of sleeping three people. It stores drinking water, has a refrigerator and a microwave oven and includes a bathroom and living facilities.

To keep us on the air, it has its own power plant, as do the KIIS transmitter sites. The Starcruiser power distribution systems are shown in Figure A. Each of the three main systems is fully protected, and each load is fed from its own breaker. When we use outside (shore) power, we tap into two separate circuits. One of them operates the technical and studio equipment, and the other powers the air conditioning, 120 volt lighting and other non-technical loads. This provides maximum protection if the air conditioners prove more than the outside source can carry and we cause an overload.

The 12 volt power supply that came in the motor home was incorporated to keep the DC system charged. The studio lighting is a combination of 12 and 120 volt systems. Either provides adequate lighting to do a show if needed.

Of major importance in the design of the Starcruiser was the way it would be used for our different types of programming; the morning show involves three different personalities, while the programming the rest of the day has the more traditional single- personality more music/less talk emphasis.

During the morning show, the major personality, Rick Dees, needs to have visual contact with both his sportscaster and the newsperson. He interacts more frequently and broadly with the newscaster, so she sits in a chair close to him. The sportscaster may have his own guests in for interviews, so he was placed at the other end of the vehicle, with a small worktable and two sets of microphones and headphones for his use.

There is room behind Rick for his producer to pull carts from a rack that runs the width of the vehicle, and to take and set up phone calls to go on the air. The producer also faces the other talents to use hand cues and other visual aids to keep the show moving.

The physical distance between the staff members allows enough isolation for them to write and work without being in the way of Dees, and without having mike bleed-through cause hollow sounds and boominess when more than one mike is open.

The console is a 10-channel LPB unit, with rotary pots. This has the advantage of requiring less maintenance than a slide pot board, and the steep slope of the front panel both conserves space and allows a work area in the front of the console for papers and copy.

During remotes other than the morning show, the studio operates conventionally. It provides 6 cart decks, a reel-to-reel transport, a cassette machine, and separate compressors on each of the mike channels. Interfaces have been included for the 450 RPU radio, the cellular phone and the regular phone all to feed the console and go on the air.

A wireless mike allows the personality to leave the studio and go outside to 'work' the crowd while the producer runs the board, feeds the cart machines, and takes phone calls.

The floor plan of the vehicle is diagrammed in Figure B. This shows the separation of the talent, and locates the kitchen and work areas. The studio and drivers areas are carpeted, while the mid-ship 'technical' portion has non-slip rubber tile for the greatest utility. The equipment racks are placed against one of the walls. Doors are provided in the sides of the racks, and a removable work table is placed between them. The table lifts out and the doors open to provide access to the rear of the rack mounted equipment.

In the studio area, access hatches are provided on the outside of the Starcruiser to gain rack access. Other weatherproof hatches open to reveal the P.A., telephone, audio, and wireless mike connectors. A number of interior patch bay jacks appear on the outside for unforseen needs.

The Starcruiser accommodates two different PA systems... The on-board system uses two E/V Model 200 speakers up on collapsible tripods. These are fed with a rack-mounted dual 75 watt/channel amplifier.

For special occasions, we use a pair of huge Cerwin-Vega concert speakers powered by a 1200 watt Carver amplifier. This is carried in a separate support van that also carries promotional prizes and give-away merchandise.

Custom cabinets inside the vehicle store tools and spare parts.
Two cellular phones keep us in touch with the outside, and one of them is equipped with a FAX jack so we can get copy, music lists, and contest sheets while on the air or on the road.

Programming is presently carried back to the studios by two means:

The most commonly used is a narrow-deviation microwave system. This operates at a frequency of 951.7875 Mhz., a channel sandwiched between KPCC, licensed at 951.5, and the upper edge of the STL band. The frequency was carefully researched and coordinated with the Southern California Frequency Coordinating Committee before it was placed into use.

An Orban 8000 limiter/stereo generator is fed with the console program output. It's composite output feeds a specially modified Moseley PCL-606/C transmitter. This goes through low-loss flexible coax to the transmitting antenna, a Scala Miniflector.

This antenna is mounted on a 40 foot pneumatic mast which is extended when the Starcruiser arrives on location for a remote.

At Mt. Wilson, we use a Scala Paraflector mounted on top of a HAMM-M Ham Radio Rotator. This rotator includes a potentiometer that tracks the azimuth exactly.

Using this as a voltage divider makes it simple to return a sample that reads to within one degree on a digital remote control.

To help in aiming the antennas, the AGC output of the PCL-606/C receiver is used to drive a Voltage Controlled Audio Oscillator.
The higher the signal strength, the higher the pitch this produces. During the aiming process, the output of this oscillator feeds the broadcast loops back to the studio. The oscillator is set to provide about a 400 Hz. tone even when so signal is present. This keeps a tone on the line, and gives us a way to make sure the lines stay balanced and operational even when the remote system is idle. Figure C shows a diagram of the Mt. Wilson equipment interface. The composite output from the narrow deviation receiver can either feed the main exciter directly, (as would be the case after an earthquake), but it is usually demodulated into left and right channels for return to the studio.

For a field test, we send the Starcruiser out on location, and the antenna is extended. The studio is called, and the Mt. Wilson antenna is turned to an approximate azimuth.

The transmitter is turned on, and the antenna is peaked for the highest pitch from the oscillator. Then the tone is sent back to the Starcruiser, either by cellular phone or on the SCA. The operator at the remote location turns his antenna for a peak. Once both antennas are peaked, the azimuths at both ends are noted and the oscillator is turned off. A signal to noise test verifies that we have enough signal strength, and the test is complete.

Just prior to the actual remote, the operator at KIIS simply turns the antenna to the noted position, and the Starcruiser usually just has to aim in the same general direction to get the required result.

When the microwave system is in use, the communication to and from the studio is done using cellular phones, or a 450 Mhz radio channel, or, as a last resort, the SCA channel of the station.

Most of the time, the music and commercials originate from the main studios, and we just use a mono feed from the remote site. This improves the Signal-To-Noise ratio, and increases the range of the Starcruiser. We've successfully used the narrow-deviation system from a distance of over 50 miles. The stereo system has worked from a range of 20 miles.

In the absence of a good microwave shot, we use the typical phone lines and order them up in advance of the remote. If all else fails. we can use the RPU transmitter to get the programming back to Mt. Wilson and from there to go on the air.

We will shortly be implementing a dedicated satellite system - it will use digital compression to allow wide-band audio to be carried on a low-cost, small bandwidth digital channel.

This vehicle has averaged two events per week since it was finished. The programming it produces is fully studio quality. It allows our listeners to see and meet us, and has increased our visibility in a very competitive radio market. For the cost, it has been one of the best investments we have made. The forthcoming satellite system will increase our operating range well beyond our listening area. And we're well protected and prepared in the event of a disaster....

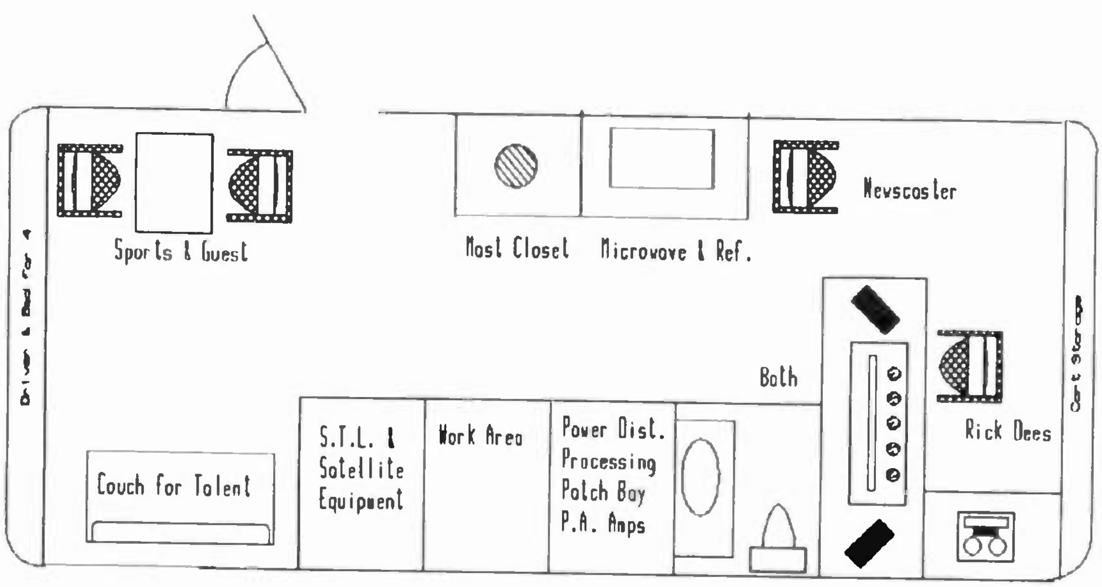

Figure B
Starcruiser Floor Plan

KIIS Mobile Studio Power Distribution

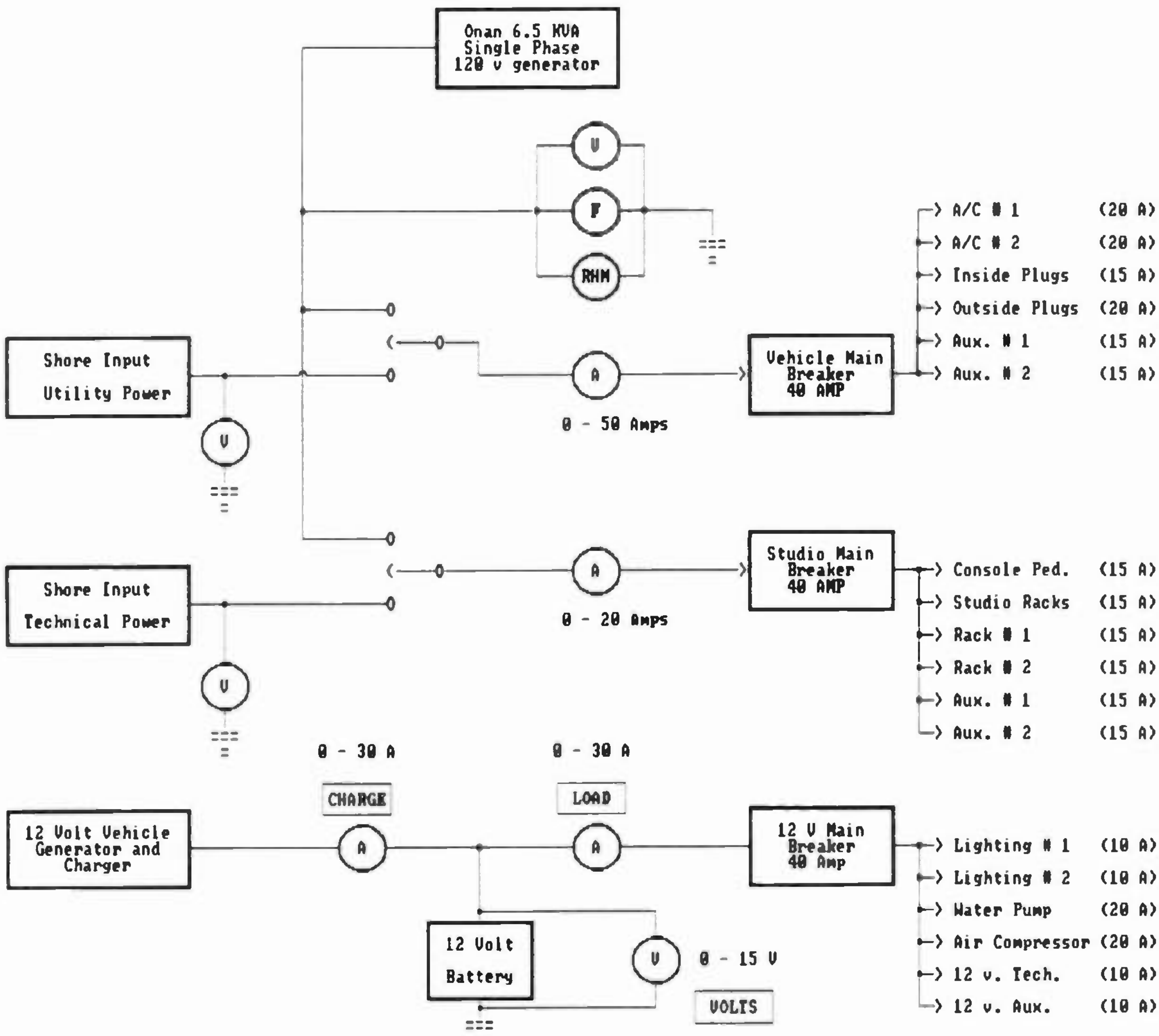

Figure A
Starcruiser Power Wiring

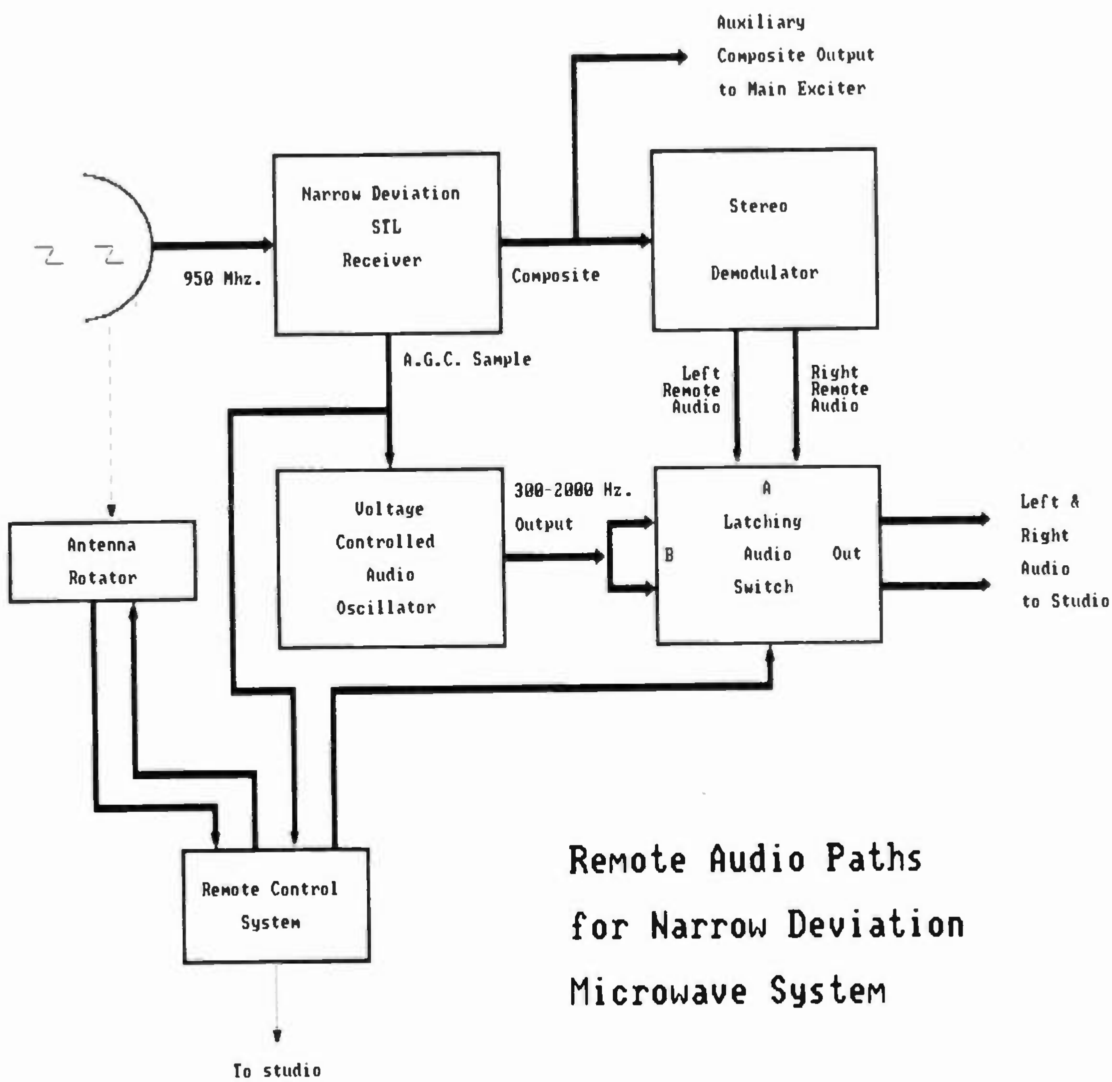

Figure C
FM Transmitter Site Signal Routing

DESIGN AND DEVELOPMENT OF A COMPUTER CONTROLLED ON-AIR AUTOMATIC MUSIC MANAGEMENT SYSTEM

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The purpose of the project was to design a system capable of automatic play of pre-recorded music, commercials and announcements. It was also desired that the commercial and announcement information could be updated locally by non-technical personnel. The DigiTotal™ system accomplished these goals. Music and commercials are stored at 20- 20KHz bandwidth with 90 dB dynamic range using commercial high fidelity VHS recorders accessed via computer control through a custom interface.

Introduction

This paper describes the development of DigiTotal™, a computer controlled music management system for the delivery of high-fidelity syndicated music. The DigiTotal™ system controls the automatic playback of broadcast-quality audio from VHS-HiFi cassettes provided by Drake-Chenault.

System Function

The DigiTotal™ system was designed as the next generation of syndicated music distribution, as an alternative to the 30-year old reel-to-reel automation systems. DigiTotal™ incorporates the latest advances in computer control of audio equipment. Currently, delivery of syndicated music is provided in the following ways.

1. Distribution of a "playlist". Perhaps the most simple and flexible of all methods, the programming company creates a list of events to be played and the sequence and time to play each event.
2. Distribution of playlist and music. Virtually the same as above, but the programming company also provides all or a subset of the music to be played.
3. Distribution of music on reel-to-reel tape.
4. Live satellite feed. The music is fed directly from the programming company's studio for immediate airplay.

The DigiTotal™ system extends these services by providing a mechanism by which the programming

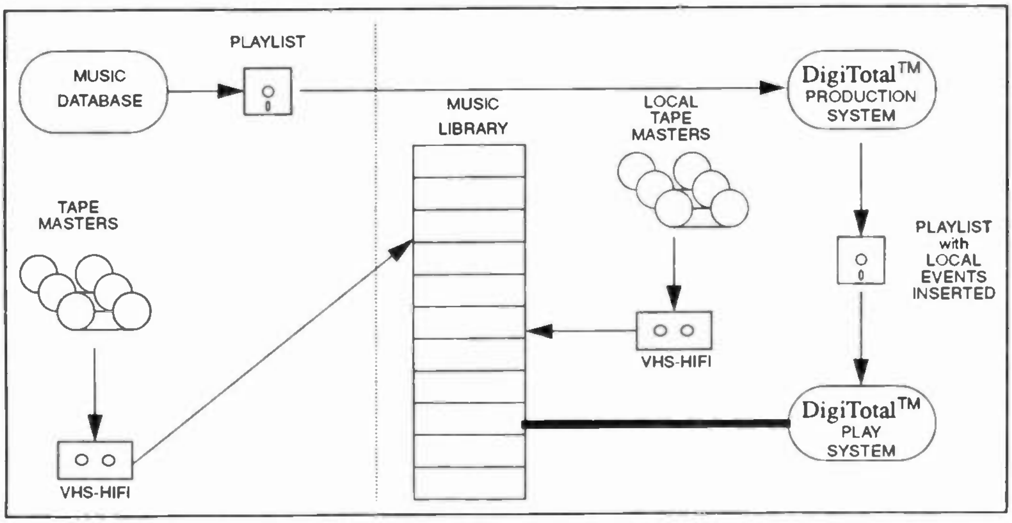

Figure 1: The DigiTotal™ System

company can provide the flexibility and local control of the first two methods, but also the automatic operation of the last two.

At its most basic level the DigiTotal™ system provides control over the play of randomly accessed music from a large library of pre-recorded material, and provides the means by which events (commercials, news, etc.) may be locally added to the program.

As shown in Figure 1, music is mastered at Drake-Chenault from the sources onto VHS cassettes. These tapes are indexed using SMPTE time code. Information identifying every piece of music and its location is kept in a master database, from which individual playlists are created. The playlist and cassettes are then sent to individual radio stations. At the local stations the users may, if necessary, record new or local material onto tapes, add identifying and locating information to the database, and add events to the playlist. Once the events are scheduled as desired, the operation of the station is totally automatic. The station could now run indefinitely, with a library of up to 1536 hours of music and other events.

Design Requirements

The overall design of the computer controlled radio station was governed by the requirement that the system perform in a manner consistent with current radio practice. It was important to the designers that the system's operation closely match existing procedures and not force any fundamental change. Specifically, the system must meet the following criteria.

1. The audio quality of the pre-recorded music must be as high as possible, approaching that of CD-quality sound. (20-20KHz, 90 dB)
2. The system must be able to precisely locate any event in the library. Precise cueing is required so that at no time is any event started early or late.
3. Location of any event must be quick enough to minimize scheduling conflicts. When accessing a specific point on any medium there is some delay incurred, and as one increases the size of a media that is accessed sequentially, that delay goes up.
4. The system must be able to cross-fade from any input into to any other input. Additionally, one must be able to control the duration and precise timing of the fade.
5. The system must be able to detect anomalous events. For example, if for sopome reason an event is not ready at the specified time, the system must be able to recover and start another event.
6. The system must be able to maintain a log of what is actually played.
7. The user must be able to perform a limited number of control operations. For example, he must be able to pause and restart the system.
8. There must be a mechanism for the creation of local events. One must be able to pre-record and schedule local events that do not come from the programming company.

System Design

The DigiTotal™ system consists of two separate systems: the Production system and the Play system. The Play system reads the playlist, or schedule, and controls the mixer, fader and decks to produce a continuous program for airplay. The Production system controls the recording of local events and local editing of the playlist sent from the programming company. Since the Production system runs totally separately from the Play system, it is possible to create local tapes and edit playlists without interrupting play. The new material is then moved to the Play system and is incorporated into the music already playing.

Both systems consist of an industry-standard AT-class computer, a custom-designed interface system, and a number of commercial grade VHS-HiFi videocassette recorders.

After several years of evaluation and comparison to other technologies, VHS-HiFi was selected. Other media investigated included 8-mm videocassette, Beta videocassette, and compact discs. By using commercially available VHS-HiFi videocassette recorders, several advantages were realized.

1. Audio quality. Certain VHS-HiFi recorders were found to match CD fidelity. These recorders have a bandwidth of 20-20KHz, and a dynamic range of 90 dB.
2. Recording length vs. access time. In extended play mode, each deck can hold up to 6 hours of audio, yet the maximum time to access any point on a tape is only five minutes.
3. Number of audio tracks. In addition to the high fidelity stereo tracks, each tape has a monaural track. This makes it easy to add SMPTE time code indexing to a tape.
4. Wide availability. Unlike several of the other possible media, VHS-HiFi is widely available from many different vendors. This insulates the system from the whims of any one vendor and drives prices down.

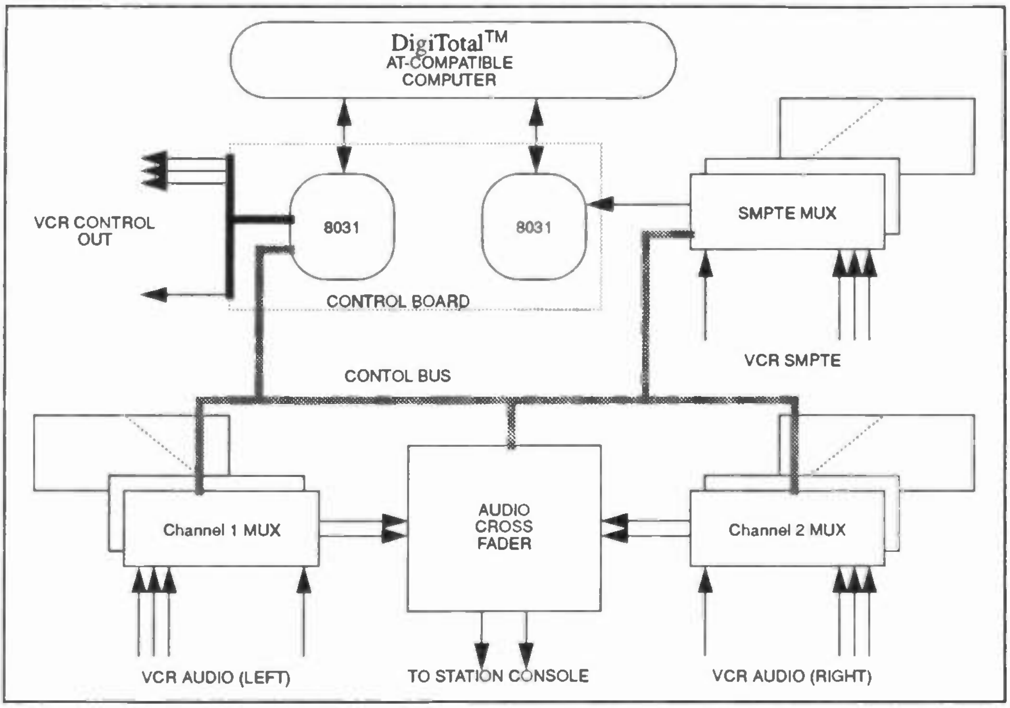

Figure 2: DigiTotal™ Interface Components

5. Capability of videocassette recorders. The videocassette market is highly competitive, and vendors must differentiate their product with special features. Some of the features greatly simplify system design.
6. Strong international acceptance. Through discussions with programming customers overseas, it was found that VHS-HiFi was not only well-accepted but the preferred standard.

Once a suitable media was chosen, it was necessary to select the specific product. It was found that "professional" recorders were far more costly and included many features that would never be used in the DigiTotal™ system. "Commercial" recorders did not have the capability of external computer control, yet were cost-effective and had features that would simplify program design.

The final product chosen was the Mitsubishi U-50, which is a commercially available VHS-HiFi videocassette recorder. The deck was modified to provide access to the videocassette recorder's microcontroller, and an interface was built into the DigiTotal™ control system.

In its most basic configuration, the control system will operate as many as 16 videocassette recorders. However, the system was designed for easy expendability, and additional modules may be added to control an additional 240 decks.

Figure 2 shows the components of the DigiTotal™ system. As shown in the diagram, the system consists of an AT-class PC-compatible computer which performs overall control of the system and maintains the playlist and music database. In the videocassette interface there is a control board, audio and SMPTE multiplexors and a digitally controlled audio cross-fader.

Control Board: The control board incorporates two microcontrollers to oversee the operation of the interface portion of the system, including all multiplexors and the cross-fader. It also provides direct control for the first sixteen videocassette recorders.

The first microcontroller continuously monitors SMPTE signals coming from the multiplexor and returns a value to the AT when queried. If the SMPTE is unreadable, as it is when the deck being read is in fast-forward or rewind, the microcontroller returns a special code. These communications take place over a serial link at 19.2K baud.

The second microcontroller performs all switching and control translation functions. This controller receives relatively high-level commands from the AT at 9600 baud, and performs operations on the multiplexors, fader and each recorder.

Multiplexors: The multiplexors each have 16 inputs and two outputs, and are designed with expandability in mind. As more inputs become necessary, one adds another board and the output of the high fidelity solid-state switches from that board is fed into the summing node of the master board for that channel. Of course, if there is only one board (and therefore only 16 channels) that board is the master board. For simplicity in design, all audio multiplexor boards are the same. A board will behave as the master board if its summing node is enabled by a jumper. Each board is also given a unique address; that address is selected by a DIP switch.

The SMPTE multiplexor board is much simpler in design, as there is no requirement to maintain the audio fidelity. This board has only one output, which goes to the SMPTE reader on the control board.

Cross-fader: The cross-fader takes its inputs from the multiplexors. Both channels may be faded in/out in any duration from one tenth of a second to 25 seconds. The fade occurs in steps of .2 dB.

Aquila Bus: All of the boards described above are mounted into a custom-designed bus. This bus provides power and signal routing to make system expansion easy. The boards are physically mounted in a 19" RF shielded card cage. This design provides for easy serviceability and a smooth upgrade path.

Software: The system software for DigiTotal™ consists of several parts. On both the Play and Production systems the software interface is completely menu-driven and makes use of an external keypad through which all functions are accessible. The user interface also makes use of a color monitor to improve readability.

Using the Production system the user records events on a videocassette recorder that is operated by the control board which is installed in the computer. As the events are recorded a database is generated which describes these local tapes. The user may then edit an existing playlist, adding events from the local tapes.

The software in the Play station cues and plays decks as prescribed by the playlist, controlling the decks, fader and audio and SMPTE multiplexors as necessary. This software also detects problems that would prevent continuous play from occurring. For example, if one deck should not become cued in time for the scheduled event to play the software recovers and plays another event that has been cued.

The software allows the operator to enter commands on the special keypad. Commands available allow the altering of scheduled events. For example, the operator may pause the system after the current event fades out and insert live material. At the moment the user strikes "GO" the next event in the queue begins.

Future Directions

The DigiTotal™ project is young. Although all of the design goals have been met and the system is in commercial production (and in fact on the air in the South Pacific), every new installation suggests new possibilities. Some of the ideas under development include:

1. Hard disk subsystem. For playback of locally recorded events, a hard disk subsystem would allow much greater scheduling flexibility, as there would be no cueing delays. Any event can be accessed for immediate play. It is now possible to digitize full stereo audio at 20KHz, but at those data rates each minute of audio requires 10 megabytes of disk storage. There are disks readily available for the PC market which would allow storage of up to 1 hour of this high-fidelity audio.
2. Control of external devices. In every radio station there exist a wide variety of audio playback devices, including cart machines, satellite receivers, and microphones. Each of these devices could require as little a relay closure for activation. A control board for the DigiTotal™ system is now in the testing phase.
3. Increase software capabilities. At the current level of operation, a single-tasking system has been more than sufficient. However, as we add more capabilities such as improved error detection/ correction, control for external devices and digitized audio from a hard disk, and networking Production and Play systems, a more capable operating environment will become necessary. The current software is designed for transportability to such an environment.
4. Video. At the current time the video track on the videocassettes is not used in any way. The only change required to the current system is the addition of a video multiplexor. There would be no change necessary to the software.

Summary

By combining the industry standard and well -accepted VHS HiFi videocassette technology with the standard AT architecture, a reliable computer controlled turnkey system was developed that provides playback of high fidelity audio in either an automated or interactive configuration.

DIGITAL PROGRAM CONTROL TECHNOLOGY: EXTENDING THE ENVIRONMENT

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With the advent of high performance Personal Computers, manufacturers of Program Control equipment no longer have to concern themselves with designing, building, and supporting computer hardware. Instead, the vendor can now concentrate on such hardware aspects as the proper audio switching and mixing, as well as the software to provide a meaningful "environment" within which the operators can work.

Creating such an environment involves the integration of a variety of tasks historically performed manually, but which lend themselves well to inclusion in the overall development of a station's programming strategy and execution. Music and commercial scheduling processes can be combined with overall format design to provide a framework equally suitable to unattended performance in a totally automated capacity; or to a live-assist environment in which the on-air talent is interactively involved with the final result.

Of equal importance is the ability of the system hardware to accommodate a variety of source equipment; including new devices that promise unprecedented capacity and utility (such as DAT, multi-disk CD "jukeboxes", and Mass Digital Audio Storage). Such versatility requires an "agile" control architecture that manages not only the variances between source types, but also between different manufacturers of the same sources; and finally insulates the operator from all of the physical complexities of the system.

This paper outlines the concepts of a system that addresses all of these issues in an attempt to provide a useful "Computer Oriented Radio Environment".

INTRODUCTION

The word "environment" gets a lot of play these days; primarily in the context of the physical world. It suggests the sky above us; the ground below us; the place in which we live and work. When we narrow the scope of the word to cover the radio broadcasting world in which we (many times quite literally) live and work, "environment" still means very much the same thing. It is a framework, a structure that contains and constrains us; imposing frustrating limitations on us and yet often challenging us to push those limits and explore new territory.

I propose that the radio "environment" exists in two planes or dimensions: the physical and the conceptual. The physical environment concerns itself with the actual materials and equipment that we have to work with; the tools we use to do our jobs. The conceptual environment is how we choose to put those tools to work for us; the "model" we build in our minds that defines the framework we must work under. Changing the physical environment is rather easy: technology is advancing exactly fast enough to obsolete a piece of equipment just at the point we are able to afford it. But bringing about a change in the conceptual environment if a much bigger challenge. Just providing a new tool or way of doing something does not necessarily make our lives better. But when a change to the physical environment is introduced that generates an instinctive feeling of "the right way to do things"; the course of history is altered.

Let me provide a good example of how broadcasting history was changed virtually overnight by a single tool. Imagine if you can (or remember, if you are old enough) what life in the radio industry was like before the invention of the cart machine. While the technical merits of magnetic tape slipping on itself are debated today, can anyone argue that life as we know it changed back 30 years ago? The physical environment was changed from reading commercials live and playing from transcription disk into "push a button and go!"; and bulk erasers and 1 kHz cue tones became as much a fixture in radio station life as the microphone and the Station ID. Why? Because it was a Good Thing, and everyone knew it; and our physical environment and our conceptual environment has never been the same.

A more recent example of how our radio environment can evolve is the introduction of Satellite Broadcasting. Can you imagine someone twenty years ago saying, "Yep, Bob, I sure would like to lock my doors at night and run my radio station from Space"! Yet, today, in only ten years no less a broadcasting force than ABC now controls a piece of the satellite industry; and over a thousand stations now transmit a significant portion of their broadcast day from a source thousands of miles from their studio site. It is hardly arguable that this change is a significant extension in our radio environment.

In each of the above cases, a technological advance was forwarded as an improvement or enhancement to the "physical" environment. The key to success, however, was the acceptance of those changes into the "conceptual" environment. Broadcasters decided to change their world, so to speak, not just because a new tool was proposed to them; but only when this physical change prompted a new way of thinking; a fundamental difference in the ways we chose to view and interact with our surroundings. The environment was extended not because of a new way of doing things, but because of a better way.

Perhaps you may have noted that I have not included any of the multitude of recent changes and advance in broadcasting technology as an example of a significant extension of our working environment. You may ask, "What about Compact Disks? What about DAT? Or any of the proposed digital replacements for cart machines and audio recorders? Wouldn't any of these technologies qualify as having changed the face of broadcasting?". I submit that although these may have changed the *face* of broadcasting, they have not yet changed the *environment* of broadcasting. They are simply new tools that perhaps may do the things we want to do in a different and better way; but they have not impacted on the conceptual environment. To this date, while they may have altered life in the radio station, they have not necessarily improved life to any great extent. None of these advancements have shown the earth-shattering impact that really true breakthroughs such as the cart machine have demonstrated.

I believe that there is a simple reason why these worthy technological advances have not had the effect that they surely deserve, and why we have yet to realize their full potential. There is one underlying obstacle to the success of these and even future technologies that propose to benefit the everyday operation of a radio station. The reason can be summed up in a single word - CONTROL.

CONTROL IS THE KEY

It is undeniable that the introduction of the Compact Disk has had a significant impact on radio station operations. Where once you had to drop a vinyl record onto a turntable, count by eye to the cut you wanted, carefully drop the tonearm onto the record, and back-cue it just enough to give you the start up time you needed (a genuine skill); now you simply have to slide in a CD, push a couple of buttons to get to the cut you want, put it in PAUSE and go! The sound quality is inarguably better (although whether the listener hears the difference continues to be debatable), and never suffers from cue burn or day to day deterioration. But do you ever look at a CD player and its TRACK UP and TRACK DOWN buttons and its ten key pad and wonder if there must be a better way? Well, stare into the multicolored reflections of your CD and come with me on an imaginary journey where we attempt to realize the full potential of this remarkable new medium:

Your entire music library of 10,000 selections is completely on-line; stored in less than 10 self-contained jukebox units completely under computer control. The units are locked and you have the key, so you can't remember when the last disk "walked away" and you had to buy another one retail at Musicland.

The disks in the jukeboxes consist of a mix of commercial CD's and special library CD's that are packed with a full 70 minutes of material from various artists. Many of these disks have 25 cuts or more. Each cut is electronically encoded with end cue signals that inform the computer controller that the song is over.

About 45,000 other selections in the CD collection you have built up over the years are stored in the same amount of wall space you used to devote to a couple of thousand vinyl singles. You freshen up the on-line libraries once a month or so from this off-line stock just to keep things interesting.

The computer that controls these juke boxes communicates with you by its multicolored, pleasantly laid out screen. When you want a given selection, you pull up a database of all of the selections conveniently filtered to include only those currently on-line at the moment; although you can easily browse the off-line library as well. Switching your search from an alphabetical listing of song titles to a similar listing of artists takes only a fraction of a second, and a "Speed Search" function allows you to locate the song you want by typing in only the first few characters of the name; narrowing down the search with each letter you press.

Once you have located the song you want, it is added to the system's play list at your command. The true duration of the song is drawn from the database and is instantly reflected in the expected playing time on the screen. Perhaps the song you have selected comes from a commercial CD that has extra intro and outro material that you would just as soon skip (Dire Strait's "Money for Nothing" comes to mind); so you select from a set of Alternate In Cues and Alternate Out Cues to give you only the portion of the cut you want. This, of course, affects the cut timing; and the computer adjusts all references to it automatically.

Your commercials, promos, ID's and jingles are all stored on a single mass storage random access device, offering instant access to any of hundreds of individual cuts. The production studio is connected *to this device as well, and can be recording a new spot into the storage system at the same time you are playing one back. Selection of spots or jingles is done by use of an English description; not a number.*

At the heart of all of these various audio sources is a compact System Controller that resembles more a remote controlled console than an audio switcher. Individual VCA's on each channel allow for level control on command from the computer; allowing for true crossfades between sources. When paired with information stored in the computer database, the controller can even automatically compensate for level differences between CD's in a single source on a selection to selection basis.

Perhaps the most important aspect of the entire system is that at your option it can be run in either a live-assist mode, with the on-air announcer interactively controlling every aspect of the system's operation; or in a totally automated fashion. So, for overnights, weekends, or whenever a jock is late or you simply want to take a 10 minute break you can hand over control confidently to the computer.

Does all this sound like a dream? Two years ago, it would be; but with hardware available in the very near future an operation exactly as described above is within yours or any stations reach. The key, again, is *control*. The audio sources (CD jukeboxes, DAT, and Mass Digital Audio Storage) have been around for some time. But can you imagine an on-air operator trying to interact with each of these devices through a ten-key pad? The ability to be able to command each of these devices on its own terms is a necessary prerequisite to the to the achievement of this *control*. So is the ability to collect all these varying sources into a system, with a centralized *control point* from which the operator may interact with the system. But all this physical control still only addresses part of the story. Unless this actual hardware interfacing is paired with a new conceptual environment to totally integrate the source equipment, the control equipment, *and the operator*; the entire system becomes simply another tool. Not revolutionary; not even necessarily better; but simply different.

To provide an example of how such an integration can be achieved, let me focus with a little more detail on a system architecture oriented towards solving a number of problems in dealing with the physical world of broadcasting today: the variety of types of source equipment, the different means of interfacing to equivalent source types between manufacturers, and even attempting to provide a suitable platform allowing interface to sources not even invented yet. However, recognizing that this physical control (as complex and as necessary as it is) addresses only part of the overall story, a system such as described here must also attempt to provide a practical conceptual model under which the operator works. Perhaps the key word to use to describe the overall system is *perspective*.

THE VIEW FROM THE TOP

Consider for a moment how much the technology of weather monitoring and forecasting has changed in just the past few years. A hundred years ago, your ability to predict what the weather would be like even 24 hours in the future was extremely limited. Perhaps a look at cloud formations might give you an insight into a storm ahead; but the ability to predict coming temperatures was inconceivable.

As communications across distances became better, the ability to collect information and infer coming events became possible. If, over a matter of days, a cold spell was reported from a number of neighboring locations and appeared to be moving toward you; you could reasonably predict when it might reach you and the temperatures you could expect. Over a longer term, patterns might emerge; allowing such phenomena as high pressure and low pressure systems to be identified and even to discover such things as the Jet Stream.

But now, consider the quantum leap that was achieved when weather satellites were launched; allowing us in real time to visualize the weather over an entire hemisphere! While there is still room for error, fairly accurate predictions can be made for a given area simply by observing the actual weather conditions as they happen elsewhere.

If you examine the evolution in weather forecasting, you see these advances have come by changing our point of view, or perspective. Obviously, when our perspective was on the ground looking up, our ability to make judgements based on the weather was severely limited. However, by changing our perspective to a point that allows total visibility of the entire area of concern, we may predict accurately events many days into the future. With the proper perspective, the important elements come into view; and judgements can be made based on the overall picture.

Consider now carrying the concept of "total visibility" to a broadcast operation. In most cases, the announcer or engineer in charge of operating the various types of audio equipment must work at the level of the machine. For example, if an announcer decides that he wants to play a particular selection from a CD, he must go through each of the following steps:

- Find out what disk holds his desired selection; and locate that disk from the library.

- Read the CD label to determine the cut number he wants.

- Load the CD into a player; opening and closing a drawer as necessary.

- Using the numeric keypad on the player (or the TRACK + or TRACK - buttons if it doesn't have a keypad), locate the song on the disk.

- If he's conscientious, the operator will play the first few seconds of the cut to determine that he is at the correct place on the disk. This may involve secondary operations such as setting the console channel to CUE and adjusting monitor volume.

- Place the player in PAUSE mode on the desired cut. If necessary (and the player supports it), this may also involve cueing the disk to actual start of audio.

- When the time comes to play the song, the console fader is raised and the player starts.

As painful as this process sounds, we all can recognize this series of steps as one of the basic operations of on-air broadcasting. Similar, if not more cumbersome procedures are required for playing vinyl disks, carts, cassettes, reel to reels, and virtually every other type of audio source currently in existence in a radio station today. Technological advances such as CD's and DAT may help the sound of the station; but they do little to uncomplicate the life of the jock!

Consider now if you throw the variable of multiple access into the pot; such as with a CD jukebox or a means of mass storage for commercials. Does it really matter that with the simple entry of 4 or 5 numbers into a ten key pad that any one of a thousand or ten thousand selections can be addressed? What method must be used to determine the 4 or 5 digit number to be entered? A three ring binder or a 3x5 card file?

Perhaps now the concept of *perspective* becomes apparent. When an announcer deals with his equipment in the narrowly focused manner described above, he loses visibility of his overall objective. If he is concentrating on the details of the requirements of the machinery he is using, his tools; he is obviously not concentrating on his primary job which is to entertain the radio audience. When he is limiting his attention to equipment concerns, he is *detail oriented*. Yet, what we want and expect from him is to be *result oriented*; to separate the end from the means.

Just as in the case of the weather satellite, the key to perspective is *distance*. Obviously, not physical distance but rather a conceptual separation from the minute details that go into operating his tools. Some of this separation has been accomplished by means of his console; giving him centralized control of the audio levels from each of his sources. In some cases, the console may contain machine control capability; so he is freed from having to turn the console channel on and push a START button at the same time. But to achieve the kind of perspective that should be viewed as the ultimate goal, the operator should be distanced even further; allowing him freedom from the physical handling of sound media. He should never have to worry how to physically access the item he wants; whether it is a song on a CD, a commercial on a mass storage device, or a newscast from another room. And he should have sufficient control from the system to give him the ability to easily select what he wants to do; do it; and then verify that what was done was actually what he wanted.

What kind of technology does it take to create this kind of an environment for the announcer to work in? Amazingly enough, it takes a technology that has been in place for a number of years: Program Automation. To some, this may be considered a sensitive topic; perhaps even a dirty word. Automation means a number of different things to different people. To some, it means a robot that takes the place of live DJ's and makes the station sound like a jukebox. To others, it means a cranky piece of machinery that is nearly always broken down. There are those who believe automation is an economic godsend to small market stations that can't seem to find or hold enough talent to fill up the hours in a day. And there are also those who believe, correctly, that automation is simply another tool to use in the appropriate circumstances to do the job of radio.

Why, then, do I dare to suggest that this much-maligned technology holds the keys to the future of broadcasting? Once again, the answer lies in the concepts of *control* and *perspective*. The ability of an automation system to control the equipment connected to it is the very key to its existence. The entire premise of automation is to provide a machine to control other machines. Because automation must work without the benefit of a human operator; it must be given the means to play music, commercials, promos, and jingles just as an announcer would. This involves controlling a variety of machines; some specially suited to the requirements of automation (such as Carousels or Go-Carts); to those considered part of the mainstream (such as cart machines and reel to reels); as well as some new and exotic technologies (such as CD jukeboxes). Automation has been doing this kind of thing for well over 30 years now; and that kind of experience is exactly what is needed to provide the level of control necessary for the "radio station of the future".

Another aspect of the control question is the fact that in many cases, an automation system can take quicker action in the case of a failure than a human could in the same situation. When an automation controller is capable of "Ready Sensing" of the machines connected to it; it can tell immediately if something wrong occurs such as the loss of power resulting from a blown fuse. Systems of this type don't even bother trying to start a machine in a "Not Ready" condition; rather, they skip on to another song or commercial without any dead air. A human, on the other hand, takes a second or two to realize that nothing is happening as a result of his attempt to start a machine; a few more as he keeps trying to start it until he realizes it is "dead", and a couple more while he frantically loads another machine so that he has something to play.

We have concentrated to now about the benefits of control of individual source machines. However, the idea of system control also has its advantages. The ability to simply switch the system to AUTOMATIC when necessary allows all the freedom you choose to make use of. Automation systems used to be measured in terms of "walk-away time"; although as this specification has approached hours and days with current technology it begins to matter less. However, even *minutes* of walk-away time must apparently be desirable; as is evident by the popularity of sequencing capability in high-end audio consoles. This is simply a very crude implementation of automation.

At the heart of the control matter, however, is the delegation of the physical operation of the sources, as well as the switching, fading, and other manipulations of the audio from them, to a central *control system*. As the details of operating a source become more complex, the more the benefit of letting a system much more adept handle them. This becomes even more important as the kinds of equipment become more diverse. It is becoming quite apparent that a machine that is perfectly suitable for the storage of music fails miserably at handling commercials; and vice versa. Different types of equipment are evolving for different uses; and commonly, the details of operating them are optimized for the particular use at hand. Mixing these kinds of specialized equipment and placing them in the hands of a human operator is beginning to approach overload conditions.

The control aspect of the situation is rather easily explained. But where does the concept of *perspective* fit in? Consider again the benefits that the weather satellite derives from its perspective - the ability to view the entire situation as a whole; not just a collection of parts. An automation system provides much the same benefits: consolidating every relevant factor of the broadcast operation - audio control, machine control, format and music control - into a single point to be monitored. Instead of a variety of *pieces* of information, the operator can be presented with a *collection* of information; giving him the ability to see what he has access to and what he is doing in the most concise and efficient fashion.

Another element of perspective is the *orientation* of information. Remember that our goal is for the operator to be *results oriented* and not *detail oriented*. The job of the system is to collect the information from the physical world of the radio station, which is seldom in the form that communicates best with the operator, and then translate it and display it in a format that helps him achieve his end result.

It might appear that the more information that is presented to the operator, the better. To a certain extent, this is true; but there is a great deal of information that could be displayed that is debatable whether is should be or not. Part of the desirability of the new environment that the operator works in is that the information can be *filtered* to include only that which assists him in getting results. Excess information is really just so much "noise"; and in radio we can all appreciate the benefits of noise reduction!

The extraneous information that has historically been associated with automation is primarily there not for the benefit of the operator; but for the benefit of the machine! For example, almost all automation systems from the very beginning associated each audio source connected to it with a number. This was part of the overall architecture of the system; making it easy for the system designer and the internal logic. This number, however, held no particular meaning for the operator except that it was the only way of identifying the source that he wanted to play to the system! Perhaps you've seen an older vintage automation system; if so, you've likely noticed every machine mounted in the system had a number prominently displayed on it. For the benefit of the system, the operator was forced to begin thinking in machines and audio sources in terms of numbers.

If the source machine had the ability to hold multiple "shelves" or "trays", such as a random access cart player like a Carousel or a Go-Cart; the situation became even more complex. Each cart or shelf was assigned another number; resulting in 4 digits being necessary to uniquely identify a given cart holding the desired selection. It really didn't matter much if the cart was labeled or not - to the system, only the numerical information mattered, and somewhere in the station HAD to be a cross-reference list that associated commercials with their numerical positions.

Even if somehow you managed to learn to live with source and shelf numbers; there was an even more insidious bunch of numbers lurking inside the automation system - Event Numbers. In order to uniquely specify the position of every "event" or step the system was called upon to perform, another group of 3 or 4 digit numbers was used. Programming typically began with low numbers and progressed sequentially through the higher numbers as the step-by-step instructions of the system were entered. In some fashion, event numbers usually corresponded with a rough time position; for example, Event 2300 would occur somewhere around 3 PM on Thursday. No one knew this except the system, however, unless he looked it up on another cross-reference list!

So consider the job of the operator when the time came to program the system: First, he'd look up where in time he wanted to begin programming and find out what event number corresponded to that approximate time. Then, he'd search out that event position only to have to key in 2 or more numbers to identify what he wanted to play at that position. Smart operators figured out all the numbers in advance; and indeed, a "ten-key dancer" could load up thousands of events in an hour.

Even when programming was completed, monitoring the system was a chore; because the only information displayed was the numerical information that was programmed. Granted, the job of the system was to perform while no one was watching; so who should complain about the displayed information. But it cannot be argued that the information available to the operator was there for his benefit; but rather for the convenience of the system. In this type of environment, it makes one wonder who was the master, and who was the slave?

Now, would systems such as I have just described be considered *results oriented* today? Obviously, if you consider the point of view of the automation system, the information did offer the appropriate level of information to reach the desired end. As with most any computerized application, a number provides the most compact means of specifying exact location and identification (does your Social Security Number come to mind here?). But who are we attempting to serve in this situation: the machine or the operator?

Is the situation any better today? Not really; remembering our discussion of what it takes to command a CD, a DAT player, or virtually any other random access device. These machines are inherently numerical in their orientation. So the physical interface to these and most other forms of audio source equipment is likely to remain numerical as well. Does this, however, condemn us to another generation of numerically oriented control devices? Not necessarily; as long as the system is designed intelligently and the appropriate considerations are made for who is more important: the system or the operator!

The situation can be solved best by using a series of *layers* to insulate the operator at one end from the physical variety of source equipment at the other. Remember the old trick of wrapping a present in successively larger boxes; making it absolutely impossible to guess what the identity of the gift might be? The same concept can be used, and very effectively, in the software environment of a Program Control system to hide the various levels of complexity involved in performing an action at the command of the operator. The idea here is to hide the details of the operation; not the result.

Let's examine a situation in which we have a control system connected to a pair of CD jukeboxes; a couple of DAT machines for current or local music, and a mass storage digital audio device for commercials. Physically, the connections to each of these devices are radically different: The CD jukeboxes may have a proprietary parallel interface; the DAT machines might have only a consumer-type infra-red remote control capability; and the mass storage commercial device might interface via RS-232. Older types of automation systems would have a very difficult time handling such a variety of interfacing tasks; in fact, most systems could not address any of these sources without the addition of expensive and exotic interfaces. It is altogether possible that some of these sources may even exceed the architectural limits of programmability of some systems; because the level of detail needed to address down to the lowest level (such as the track level in a CD jukebox) cannot be contained in the space allowed for source and shelf numerical information.

Given that a new control system is required to handle such a diverse interfacing requirement; what is the best way to structure such a system so that 1) the physical needs of the sources can be addressed; 2) the conceptual requirements of the operator can be met; and 3) that the system is "agile" enough not to be rendered obsolete at the next turn of events in source equipment technology? The answer is to separate each of the elements of the interface into a series of layers; each doing a specific job of translating information to the layers above and below. If the architecture is open enough, changes in technology that require modification in one or another layer can be accomplished without affecting any others.

For example, an Operator Interface works directly with the operator; providing him with an English language method of specifying what he wants the system to do. Below this is a System Layer that breaks his requests down into information the system can understand; such as looking up a song title in a database to determine its physical (probably numerical) location in the system. Once found, an Execution Layer is invoked to give the system the necessary instructions of how to perform the desired task.

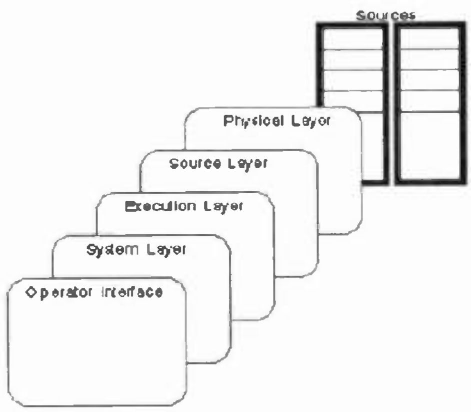

Figure 1. A layered architecture allows a system to adapt to changing conditions without massive software re-writes.

However, up to this point, all sources look virtually alike; sharing such common traits as START and CUE commands. Individual source machines have different means of specifying how to perform such commands; such as whether it involves a single line to be pulled to ground, or a string of ASCII characters to be sent through an RS-232 channel. So, a Source Layer is invoked that determines what kind of source machine is attached and the actual details of how to perform the given command. As new source machines are invented and added to the system, new "source drivers" are added to this layer to accomodate them; much like a "printer driver" is used in a word processing system to define the actual characteristics of a given printer.

Even after this many translations, the final command has yet to be transferred to the machine itself. This is accomplished by the Physical Layer; which defines the electrical interface to the machines and actually communicates the command directly to the source.

In a layered architecture such as this, the operator is able to view the system at a level that is appropriate for his best understanding. His commands are then interpreted through the various layers underneath in an invisible fashion. In fact, if the operator specifies that a given song or commercial is to play at a given time, he need not be concerned about the actual location of the source of that material; only that it is executed properly according to his instructions.

WHERE THE FUTURE LIES

Before you dismiss this proposition that the future of broadcasting rests with technology derived from Program Automation; let me qualify that statement somewhat. The future of broadcasting will be dependent not on old automation technology that you might have seen in the past; but rather with new technology that you will be seeing in the very near future. Some of the concepts that were specific to automation will carry on; but many will be dropped in favor of new and better ways of visualizing what the system is intended to do. In short, don't let any past prejudices blind you to the many advantages and benefits to be derived from the new technologies that are and will be appearing.

There are several reasons why the technology of Program Automation today deserves to be considered on its own merits and not in light of previous successes or failures. Many things have come together now that make for an opportune time for a radio revolution:

- The Personal Computer

Probably no one reading this paper is unaware of the Personal Computer Revolution. In only 10 years, computers have advanced in capability and declined in price to such a degree that the computing power and storage capacity needed to provide the kind of environment we have been discussing are affordable to even the smallest of stations. In the past, manufacturers of Program Control equipment have had to be computer manufacturers as well (or try to provide such capabilities without a computer, which is a lost cause). This artificially inflated the cost of automation because of the small quantities involved; making the technology out of reach of many of the stations that could benefit from it most.

But now, a PC provides all the processing power needed to run the most ambitious of control systems. Since they are mass produced, the economies of scale allow this power to be purchased at a fraction of the cost of earlier systems. Local sales and service outlets are readily available; and a station can even trade out for a computer if they want to!

With the automation manufacturers no longer having the responsibility to expend their resources in the design and building of computers; they are now free to spend their time enhancing the balance of the product: the audio switching, machine control, and software aspects of the system. In short, you can now get far more of an automation system for less money than ever before.

- The Source Equipment situation

As we have discussed previously, many of the new types of source equipment such as CD Jukeboxes, DAT, and mass digital audio storage simply do not fit into the radio environment without some type of automated control equipment. In fact, many of these devices could not conceivably be used without an entire new generation of control technology. It is safe to expect that this condition will continue into the future, perhaps even at an accelerated pace; so any system used to control these types of sources must be adaptable to change.

- Technological Awareness in the Station

You can thank yourselves collectively for many of the advances you will be seeing in the area of Program Control. As you have progressed in your own fields of endeavor, you have seen the strengths and weaknesses of the current generation of equipment. If you have had the occasion to become computer literate, you have watched the evolution of technologies such as Word Processing and DataBase Management into extremely powerful tools. In turn, you have begun to demand similar advances in the areas that matter to you. We, as manufacturers, have to respond to these demands with products that meet your expectations, or face being replaced by others who do a better job.

- The Quest for Perfection

Perhaps the single reason why anything changes is because there seems to always be this nagging feeling that "there must be a better way". For as long as radio is around, there will always some who are dissatisfied with things as they are. Count in this number the inventors of the cart machine and the CD; and don't be surprised in the future when other advances just as revolutionary are introduced. Evaluate new technologies not on how different they are from the way "it has always been done"; but from the standpoint of whether they truly offer a more efficient and enjoyable way of doing the job. "Extending the Environment" involves constantly testing the limits for a better way of doing things; and then accepting new ideas when they make sense. We may never reach a state of perfection; but we can always keep trying.

ACKNOWLEDGEMENTS

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THE AUTHOR

David J. Evers received an AASET Degree from DeVry Institute of Technology in Phoenix, Arizona. A native of Nevada, he returned to his home town of Fallon where he was employed for 3 years at station KVLV. During his experience there, he found special interest in Program Automation; serving roles in Programming, Operations, Sales, and Engineering.

In 1979, Mr. Evers joined Broadcast Electronics, Inc. in Quincy, IL. For the past 10 years, he has held positions in Field Service and Training, Sales and Sales Management; and in his present position of Manager of Digital Product Management, he directs company efforts in the development of new products and management of existing products.

Mr. Evers is computer enthusiast and has authored several computer programs; many of which have been contributed to the Public Domain.

RDS: THE EUROPEAN EXPERIENCE AND A PROPOSAL FOR NORTH AMERICA

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ABSTRACT

As of 1989, the Radio Data System (RDS) technology has been implemented in 16 European countries.

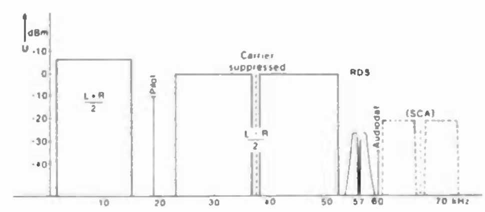
In the United States, RDS will bring a range of features and functions to broadcasters and radio listeners. A proposal for North American RDS standards is suggested, and various RDS benefits are described herein.

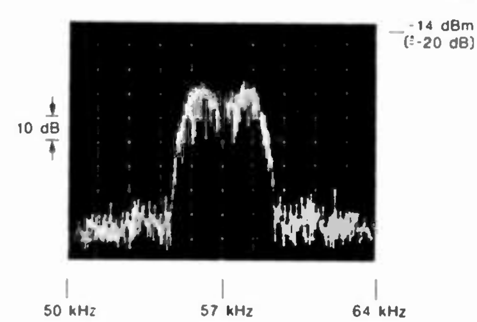
INTRODUCTION

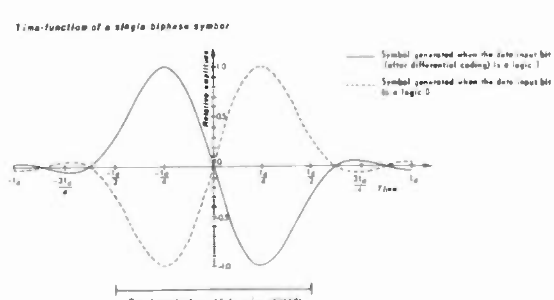
The RDS (Radio Data System) technology was developed in the early 1980's in response to a number of technological requirements and desires of European broadcasters. In particular, European broadcasting is largely a network configuration wherein the same program is transmitted on numerous frequencies throughout a country. It is important that the listener be able to track the program automatically from frequency to frequency as he or she drives. In addition, there was the desire for visual station identification by call letters or slogan on the radio display, the ability to select stations by format, a high capacity paging system, a road traffic information system, an in-house data channel, and, especially, a national and local emergency alerting system.

The RDS technology is an 1187.5 bit per second double side band suppressed carrier PSK data channel centered at 57 kHz. 57 kHz was chosen because it is the third harmonic of the 19 KHz pilot, and is the best subcarrier frequency with minimum artifacts and disturbances during multipath. Details of the RDS transmission technology are widely distributed, and are available as standards from the European Broadcasting Union and CCIR.

MODULATION TECHNIQUE







The first commercial European RDS transmissions began at the end of 1987, and by the end of 1989 16 countries, including England, France, Germany, Sweden, Norway, Italy, etc. had installed RDS on virutally all of their stations. In response to this meteoric growth of RDS, virtually every car radio manufacturer began to deliver RDS radios in 1989. Among the 24 manufacurers are Pioneer, Sanyo, Sony, Sharp, Blaupunkt, Grundig, Denon, Philips, Kenwood, Hitachi, Yamaha, etc., and Philco on various of their Ford models. Unquestionably, General Motors and Chrysler will incorporate the RDS technology in their European products shortly in response to the market demand. The decoder technology required for RDS signals and emergency alerting has been reduced to one integrated circuit. Presently, Philips, Sanyo, Grundig and Blaupunkt are manufacturing these IC chips, and more will be coming. The cost of the decoder IC should get down to less than $5 per unit making RDS affordable and attractive to receiver manufacturers in all price ranges of receivers.

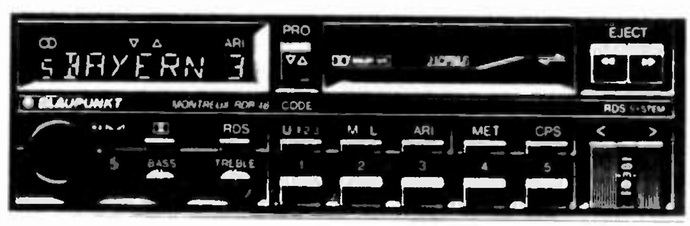
List of RDS features to be implemented in various countries

Features: Country	PI	PS	AF	TP[1]	TA[1]	PTY	DI	MS	PIN	EON	CT	RT	TDC[2]	IH[2]	RP
Austria	I	I	I	I	I	-	-	-	-	-	-	-	-	-	-
Belgium	I	I	I	I	I	-	-	-	-	A	-	A	-	-	-
Denmark	I	I	I	I	L	-	-	-	-	L	-	-	-	-	-
Finland[6]	A_{90}	A_{90}	A_{90}	T	T	L	T	T	L	-	-	L	-	-	-
France	I	I	I	I	L	-	-	-	-	-	-	-	-	-	I[3]
Germany, FR	I	I	I	I	I	-	-	-	-	-	-	T	-	T	-
Ireland	I	I	I	I	I	T	-	-	-	A_{91}	I	-	-	I	I[4]
Italy	I	I	I	I	I	-	-	-	-	-	-	-	T	-	-
Netherlands	I	I	I	L	L	-	-	-	-	-	I	-	-	-	-
Norway	I	I	I	T	T_{90}	-	-	-	-	-	I	-	L	-	A_{90}
Portugal	I	I	I	I	A_{89}	-	T	T	-	A_{89}	T	A_{89}	L	-	-
Spain	A_{89}	A_{89}	A_{89}	A_{89}	A_{89}	-	-	-	-	-	-	-	-	-	-
Sweden	I	I	I	A_{89}	A_{89}	A_{89}	L	A_{89}	A_{89}	A_{89}	I	A_{89}	-	-	I
Switzerland	I	I	I	I	I	-	L	-	-	L	-	-	-	T	-
United Kingdom															
BBC	I	I	I	A_{89}	A_{89}	T	-	-	T	I	I	T	-	-	-
ILR	I	I	I	T	T	-	-	-	-	L	L	T	-	-	-
Yugoslavia[5]	I	I	I	I	T_{89}	-	-	-	-	T_{89}	I	-	-	-	A_{90}

RDS AS A PROMOTIONAL AND RESEARCH TOOL

So, we come to the United States and Canada today. Why RDS? As we all know, a radio station's revenue is largely a function of its ratings. The methodologies used by Arbitron and Birch leave a great deal to be desired, yet they're the best we have. It is interesting to note that as an industry, we spend approximately $20 million per year on television and outdoor advertising to promote our radio stations. Why? We try to get our listeners to remember our call letters, our dial position, or our slogan, in the hope that the ratings services will give us our fair share of the audience. Wouldn't it be nice if every car and home radio had your call letters, your slogan, and your frequency displayed directly on the radio? With RDS it is not only possible, but it is already being done. RDS can display a station's call letters, its slogan, and its frequency each up to 8 characters directly on the radio display. I'm sure you will agree that it makes more sense to have your listeners see your call letters, slogan and frequency on their car or home radio than it does on a television or a billboard. RDS should not be confused with ROM based memories such as PRS where radio stations and their call letters are stored in a static memory. If you change call letters, frequency or format these radios may be wrong forever. RDS is a dynamic system which you can modify at the station when things change.

With RDS we can now use the radio receiver as a research tool. RDS transmissions contain their own time clock. It is simple to take a data output from a car or home receiver and bring it to a micro cassette recorder. Whenever the receiver is in use, the data of the call letters, frequency, and time are recorded on the micro cassette which can then be analyzed for listening habits.



EFFECTIVE USE OF TRANSLATORS

There are over 1700 FM translators on the air today. If and when the Commission opens up translators for applications again, many more can be put on. The marketing problem with translators is explaining to your listener the alternate frequencies to use in various areas. With RDS its automatic. Each station sends out a list of alternate frequencies on which the station can be heard. The RDS receiver constantly tests those frequencies to determine if there is one of better quality, and if there is, the receiver automatically switches to that alternate frequency. This action occurs in a few milliseconds, almost unnoticeable allowing the listener to use the full coverage area of the station from both its primary and translator sites. In the case of one station in New York, a translator in New Jersey extended the reach of the station by almost 5 miles in a high density population area. In the future, a more intricate system of handoffs may be developed wherein networks such as Satellite Music Network, ABC or CBS, would send alternate frequencies of other affiliates so that a listener could drive across country and, for instance, always be listening to the Oldies channel.

In order to to allow receivers to automatically find stations and their alternate frequencies, each station must transmit a hexadecimal 4 digit PI, or Program Identification code. In Europe, these codes carry information about the country and region from which the station emanates, as well as uniquely identify the station. In the United States, we want to establish PI codes based on the location of transmitters. In order to have non-duplicated PI codes to cover all of the FM stations in existence today, and 50% more in the future, it is necessary to develop a matrix whereby PI codes are reused in states where the signal from one state cannot possibly be heard in another. For example, the PI code used in New Jersey can be used again in Colorado and Louisiana. The PI codes in North America would also distinguish between American, Canadian and Mexican stations. Whenever a station goes on the air with RDS, its PI code would be assigned in advance. This is the only way to prevent duplication of PI codes which would invalidate the system. A detailed proposal for PI codes in North America is contained in Appendix I.

RDS PAGING

RDS in the U.S. will have paging capacity capable of sustaining tens of thousands of paging customers on each radio station. These can be networked to form regional, statewide or nationwide paging systems. In the United States and Canada the MBS paging technolgy, a derivitive of RDS, is operational in some areas. It will be necessary to convert the MBS pagers to RDS to insure that all stations can partake in the benefits of RDS. In Ireland, a scheme for time sharing between MBS and RDS is being tested, and may lead to an alternate solution of this issue.

FORMAT SELECTION

Automatic program format selection is interesting for the United States. RDS allows stations to transmit codes indicating their format. RDS radios would let listeners select only stations of the format they wanted to hear. The format definitions which are applicable in Europe are not complete enough to cover the wide spectrum of music and other programs presented on North American radio. We have proposed a new program type, or PTY code for United States and Canada in Appendix 2. It may be useful to have a committee of broadcasters review the Program Type proposal for North America to ensure that all relevant formats are properly represented and described. Program Types 16-30 are spare and could be defined as needed. Program Type 31 as in the European System, is dedicated to emergency broadcasts and alerts.

IN-HOUSE DATA CHANNEL

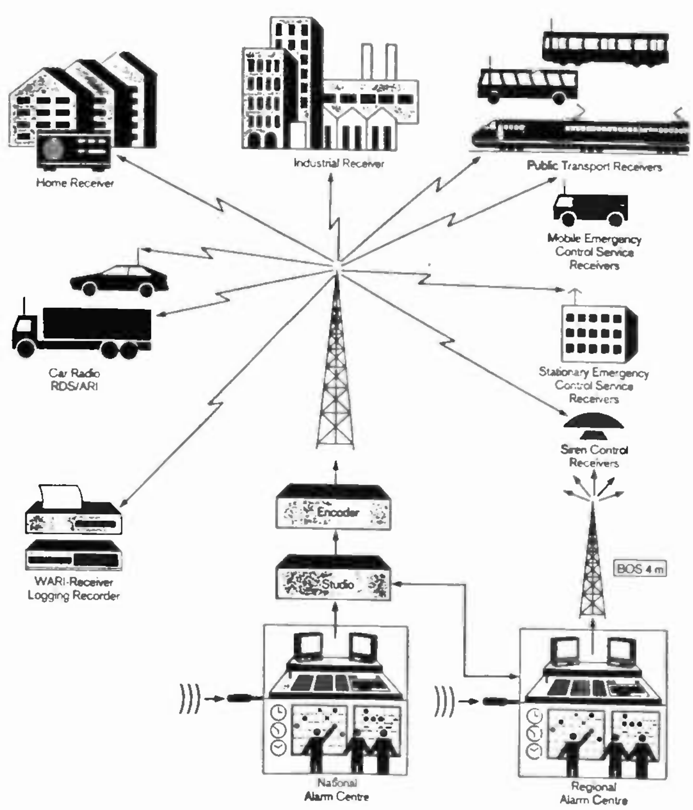
With RDS, broadcasters have a data channel that they can use for the radio station operation. Depending on the location of the encoder, and whether the station sends composite or L and R, the RDS data stream can be used between the studio and the transmitter or transmitter and studio. This could eliminate the need for a costly TSL telemetry link or phone line from the transmitter to the studio or visa versa.

EMERGENCY ALERTING

Emergency alerting and notification using the RDS technology has been fully developed and is called WARI in Europe. It is a comprehensive alerting and notification system for the general public as well as emergency workers and broadcasters. The system has the capability of actuating sirens automatically overriding a radio station's main channel audio, sending closed circuit digital messages to emergency workers and radio stations, paging emergency workers, and even interfacing with cable TV and television stations to provide audio and visual data on the impending emergency. If required, this technology can even turn radios and TVs on automatically in an emergency situation. In the United States, Sage Alerting Systems, Inc., the licensee for this technology, is proposing to use this technology in parallel with EBS as an enhancement for local, regional, statewide and national alerting and notification.

For the first time, a total systems approach has been employed in emergency communications and emergency management for emergencies such as tornadoes, hurricanes, earthquakes, floods, chemical spills, and other natural and manmade disasters. As is the case with the 100 plus nuclear power plants in the United States, it is possible and likely that more than 10,000 hazardous process chemical plants in the United States will be required, by federal law, to have their own emergency alerting and notification systems for the general public. This RDS based technology will permit relatively inexpensive and quick implementation because the decoding circuitry will already be contained in RDS radios for cars and homes. Small, battery operated, frequency agile home receivers will be available for approximately $50 in the near future. These, and other emergency alerting receivers will use 2 new subcodes of RDS, Group 1-A and Group 9-A, to identify the emergency alerting station in an area and automatically tune the receivers to that station. Should the key alerting station go off the air, the receiver would automatically retune to one of several secondary stations. The Group 9-A data would also actuate sirens and send closed circuit digital messages.

This RDS based alerting and notification technology has created interest at FEMA and FCC in the United States and DOC and EPC in Canada, as well as at various state emergency preparedness committees and LEPC's (Local Emergency Planning Councils). This system is being proposed as the alerting system of choice for the hazardous process chemical industry and could potentially be a nationwide system within the next five years.



RDS ON THE AIR IN THE U.S.

The first full time RDS service in the United States began on January 5, 1990 when an RDS encoder was installed at WHTZ-FM, 100.3 MHz Z100, New York City's top rated CHR station. WHTZ-FM has 2 translators, one on 96.7 in Asbury Park, New Jersey, and 94.3 MHz in Rockland County, North of New York City. WHTZ provides a good opportunity to demonstrate the alternate frequency capabilities of RDS, as well as its compatability in a highly competitive RF and audio processing environment. WHTZ-FM is PI code 8201, 8 for New Jersey, 2 for having transmitters in 2 different states (New York and New Jersey, and 01 as the first station on in this area it is transmitting PTY 01 for CHR, TP for traffic program and transmitting Radio Text saying "Hello, New York".

WHTZ-FM uses a very sharp, low pass filter to remove unwanted artifacts above 53 kHz. The filter reduced the upper sideband of the RDS signal by more than 40 dB. Despite this, the data channel worked flawlessly even to the point where the main channel was unlistenable. This is a real test of the inate ruggedness of the RDS signal.

Plans for 1990 call for installing RDS in Atlanta (before NAB), Los Angeles, Chicago, Detroit, Orlando and one or two cities in Canada. Delco Electronics, the world's largest car radio manufacturer (they make 7,000,000 car radios a year) has expressed interest in RDS, and is currently working on several RDS products. A number of companies including R&E Instruments, Rohde and Schwarz, SKAT and Telefunken produce RDS encoders. These units are available today in the United States and Canada.

CONCLUSION

While RDS was designed in the European environment, its uses and applications are wide and varied for the United States and Canada. By adopting the RDS technology, radio stations gain significant new exposure to their listeners by the visual displays of their call letters, slogan and frequency, as well as the potential for additional revenue from paging and the expense saving on telemetry channels. Stations can use RDS to provide traffic information to their listeners, a full scrolling radio text system, format identification, and emergency alerting and notification capabilities.

ACKNOWLEDGEMENTS

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No. 10 - December 1989

APPENDIX 2

Proposed North American Program Type (PTY) Code

1.	News
2.	Current affairs
3.	Magazine
4.	Sport
5.	Education
6.	For children
7.	For young people
8.	Religious
9.	Drama, literature and feature
10.	Pop and rock music
11.	Light music
12.	Serious music
13.	Jazz
14.	Folk music
15.	Variety
16.	Country
17.	Top 40/CHR
18.	AC/Oldies/Soft Rock
19.	Black/Urban
20.	Beautiful/Easy
21.	Spanish
22.	MOR/Variety
23.	Nostalgia
24.	Weather
25.	25-30 Spare
31.	Emergency

Proposal for USA
Maintaining EBU Specification
Second Element

	1st	2nd	3rd	4th	Element
Bits	1 … 4	5 … 8	9 … 12	13 … 16	
	State	Type of Program	1 to	255	

0 = Class A stations w/o translators
1 = International program e.g. Voice of America
2 = Stations with towers in se veral states
3 = Stations with program for another state
or program source in another state (= supraregional)
4 = Stations with translators except state capital
C = Stations with translators, main tower within
bounds of state capital

Complementary for California (if needed)
A = Stations in Los Angeles area
D = Stations in San Diego area
F = Stations in San Francisco area

each case
max. 255

New York and Texas could be made in a similar way.

I. North America

A reserved for (future) national networks USA
C reserved for Canada and Mexico

1	Alabama	Arkansas	Minnesota	Pennsylvania
2	Alaska	Hawaii	Mississippi	Rhode Island
3	Arizona	N. Carolina	Missouri	Wyoming
4	Georgia	Illinois	Montana	Texas
5	California	Indiana	Nebraska	Utah
6	Tennessee	Iowa	Nevada	Vermont
7	S. Carolina	Kansas	N. Hampshire	Virginia
8	Colorado	Lousiana	N. Jersey	West Virginia
9	Connecticut	Kentucky	N. Mexico	Washington
B	N. Dakota	Maine	N. York	Wisconsin
D	S. Dakota	Maryland + DC	Ohio	Idaho
E	Delaware	Massachusetts	Oklahoma	
F	Florida	Michigan	Oregon	

APPENDIX 1


AK
2
WA
9
MO
4
ND
B
MI
1
ME
B
OR
F
ID
D
WI
B
MI
F
NY
B
VT
6
NH
7
MA
E
CT
9
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OFF-PREMISE CONTROL OF BROADCAST FACILITIES VIA SATELLITE

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A technical operations summary of the National Supervisory Network - the nation's first interactive, satellite network specifically serving the broadcast industry. Through dedicated, Ku-Band VSAT links, NSN provides stations across the country with FCC legal off-premise control of transmitters and EBS, complete operations logs, trend analysis, and equipment performance reports. Because the system is addressable, private sub-networks can be added to the service for easy transfers of data, CD quality audio spots, and NTSC video frames between members of station groups.

Introduction

Since the implementation of the National Supervisory Network in 1989, off-premise control of broadcast facilities via satellite is now a reality. Offering an economical, FCC legal, quality alternative to local control for broadcasters across the country, the NSN system is built upon the proven reliability of packet switching and interactive Ku-Band technology. Ku-Band "Very Small Aperture Terminals" (called VSAT's) are the backbone of over 200 private corporate communications networks, providing high speed data interconnections between thousands of stores and offices nationwide.

Your hometown skyline probably sports several Ku-Band VSATs right now: they are on the roof tops of your local discount store, supermarket, and distributor warehouse performing such varied tasks as on-line credit card verification, connecting all the cash registers in a chain of stores to a home office point-of-sale system, warehouse inventory tracking, supplying voice circuits for communications between branch offices, and industrial television for training program distribution or tele-conferencing. Oil producers, gas pipelines, electric utilities, and some state emergency districts routinely use VSATs in "SCADA" (Systems Control And Data Acquisition) applications for telemetry of remote equipment sites and locations.

Because Ku-Band VSATs are virtually free from terrestrial interference, frequency coordination problems, and cumbersome hardware installations, they are a natural choice for use in broadcast applications. While broadcasters have long recognized the "one-way" aspects of satellite services in electronic news gathering and television program distribution, the National Supervisory Network is the first broadcast concern utilizing the interactive, Ku-Band VSAT networking available through current satellite technology.

The Dedicated Satellite Link

An error-free, dedicated satellite link is the foundation of the National Supervisory Network's FCC legal off-premise control and data services. This link is created by combining VSATs and computers across an X.25 Packet Switching Network to establish virtual circuits with client stations. (Figure 1) A virtual circuit is a bi-directional association between two locations across a Packet Switching Network (PSN). In a

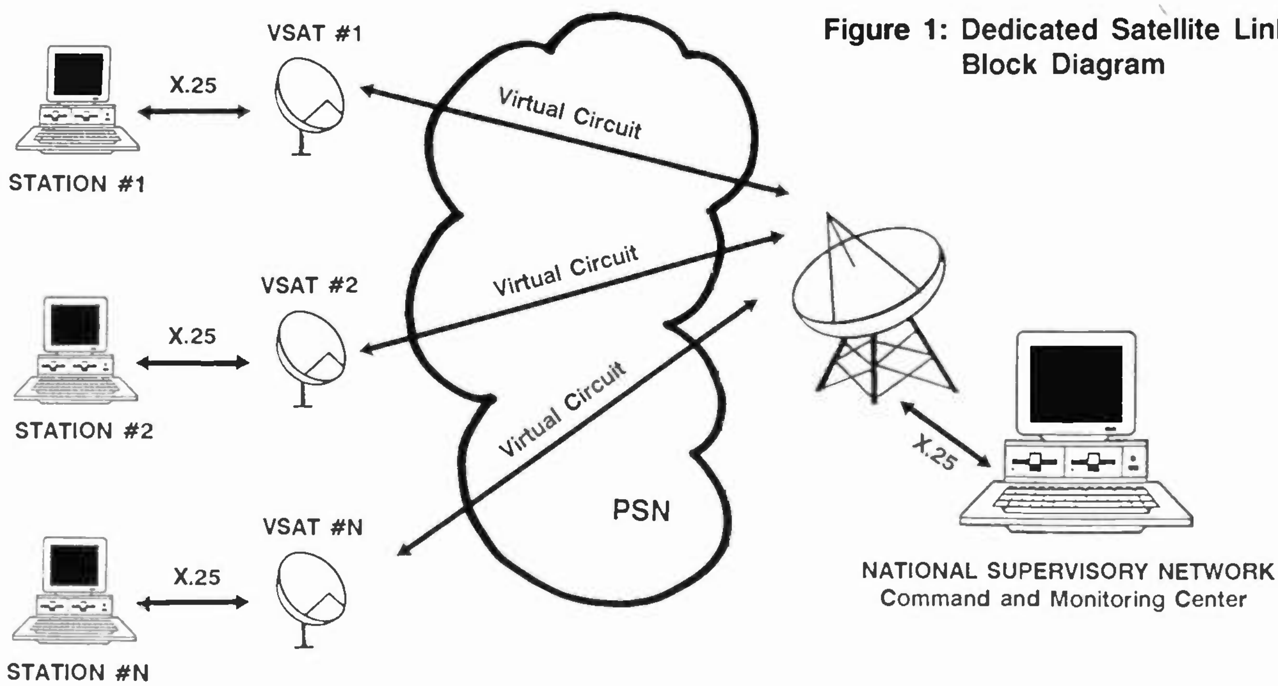

Figure 1: Dedicated Satellite Link Block Diagram

PSN, client data is broken into discreet blocks, or "packets", which contain user information plus source and destination addresses. By their nature, packets are much easier to handle in a network than intermittent streams of data. Ease of handling offers PSN users many benefits including:

> Flexibility PSN's are extremely flexible and can be reconfigured to meet changing communication requirements without difficulty.

> Cost Sharing The address information each packet contains allows packets from many sources and destinations to be interleaved in time on the same PSN circuit, so that all users share the costs of the network.

> Reliability The reliability of Ku-Band satellite PSN's is designed to be 99.995% or better.

> Connectivity The use of X.25 in a PSN encourages connectivity. Private, secure, bi-directional connections can be made from user-to-host, and from user-to-user concurrently. Because of this feature, PSN's are often represented graphically as fluffy clouds.

The Recommendation X.25 International Packet Data Standard is an internationally defined method of connecting a computer system to a PSN. The X.25 protocol combines the best features of other packet protocols with continuous error correction. X.25 was selected by the National Supervisory Network because it offers exceptional security and the confidence of a data link that is as error-free as technology allows.

Through the use of multiple levels, X.25 provides its own very strict built-in error correction, protecting the client from the possibility of corrupted data interfering with station operations. There are seven levels of X.25 in each transmission across the network. Levels 1 & 2 contain low-level call and address information while levels 3 through 7 contain the actual user data. The network is only able to interpret the two lowest levels. Because of this, PSN operators and other users are unable to read or tamper with user data contained in packets.

Sending data across the PSN in the X.25 packet format also eliminates the possibility of transmission problems that differences in remote control units and other local internal hardware may cause. As long as the packets arrive error-free and in the order guaranteed by levels 1 & 2, the data contained in them makes contact with the command center through the virtual circuit. Unlike dial-up or leased lines, a virtual circuit does not represent a physical connection between client stations and the NSN command center, but is instead a logical communications path. The use of X.25 allows NSN to maintain virtual circuits with thousands of client stations simultaneously, giving the equivalent of a dedicated line to each at all times.

The Station Interface

The National Supervisory Network has developed a comprehensive hardware and software interface to gather, process, and store client data as well as maintain the satellite link. (Figure 2) The interface is installed at the studio site of each client station and consists of an 80286 based computer, an X.25 Packet Assembler/Dis-assembler, an NEC Nextar series VSAT, and a proprietary software package. The computer runs NSN software and the X.25 Packet Assembler/Dis-assembler (PAD) to act as an interpreter between the station's equipment and the satellite system. The PAD establishes the virtual circuit with NSN, assembles outbound serial data into packets and dis-assembles inbound packets into serial data.

The site computer obtains and processes transmitter telemetry readings, commands and other parameters directly through the serial data ports on most major brands of digital remote control units. These include: the Moseley MRC-1600 CRT Option Port, the Burk Technology (Advanced MicroDynamics) TC-8 and ARC-16 Computer Interface Options, the Gentner VRC line data port option, and other manufacturers'

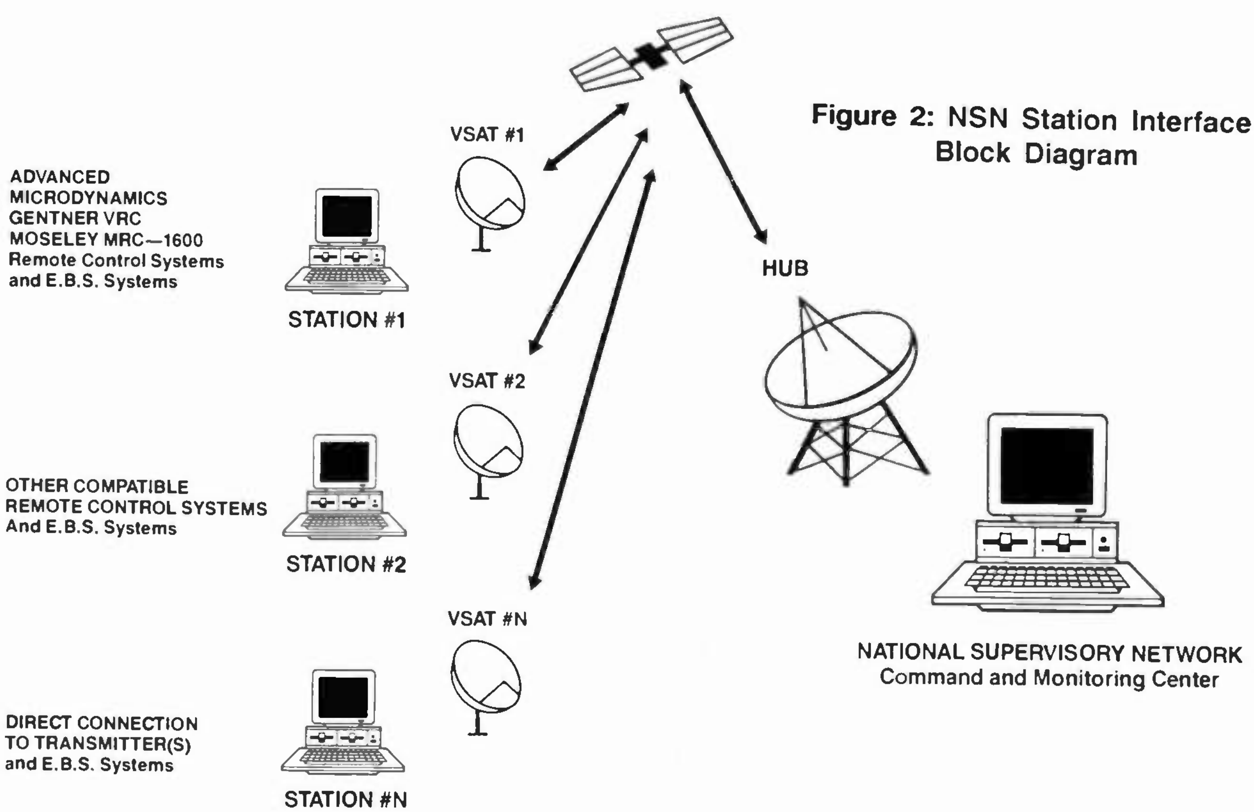

Figure 2: NSN Station Interface Block Diagram

serial data ports. If transmitter equipment is located at the studio, NSN offers a 16-channel analog-to-digital card to connect the computer with on-site station equipment. A separate computer card and relay panel are used to connect to the Emergency Broadcast System and studio fire and security alarm systems.

The Ku-Band VSAT is used to link the computer to the NSN command center. This VSAT transmits and receives data at 56.0 kilobits per second. Data from the terminal travels to the satellite (Satcom K-1, in most cases) and is transponded to the NSN command center in Colorado by the PSN. Response times are efficient; no matter where a client station is located, operators at the NSN command center receive client data within 1 second of generation and can respond immediately.

In addition to interpreting remote control equipment data, NSN software permits each network affiliate to have its own unique configuration. Using the set-up program, the local engineer dictates which telemetry channels he or she wants monitored or controlled, what the ranges or alarm states of each channel will be, alarm state priorities, and command options and functions. Explicit instructions for routine and alarm operation can be entered for each channel and all commands. Whenever a change is made in station configuration, the engineer can modify the set-up. Current set-up information is stored in the client computer and up-linked to the command center.

NSN software also functions as the local access point between station personnel and network operations personnel. The station computer displays present equipment readings, handles message traffic to and from the station, and stores and displays transmitter log information, news, weather, messages, features, and operations log summaries for review and printout by the local designated chief operator.

The NSN Command Center

The National Supervisory Network command center is a custom designed facility staffed with competent technicians familiar with broadcast operations. All NSN computer, satellite, and power systems are state-of-the-art and fully redundant. Command center personnel log on as the fixed control point duty operator for each station on the network. Transmitter and other readings from client stations are automatically logged in both the NSN computers and the client site computer every thirty minutes. Client station alarms are sent to NSN immediately regardless of the regular reading transmission cycle. In the event of multiple or simultaneous alarms, the command center computer allocates incoming data to its numerous operator terminals using a priority system designated by the client stations themselves.

Emergency Broadcast System alarms are always assigned the highest priority the system allows. Local station engineers cannot change this assignment. In the event of an EBS alarm, the NSN duty operator is notified immediately through the satellite, while the computer dials the command center via a redundant telephone line. The operator has the following menu of functions for each site:

1. Listen to local CPCS EBS station audio

2. Place a station's program audio on our monitor line

3. Activate a local EBS Tone Encoder

4. Start a local EBS Alert or Test Cart

5. Place local CPCS Audio feed on a station's air feed

6. Go on the air live from our Command Center (for live announcements or to re-play recorded CPCS information)

7. Reset normal program audio feed to air

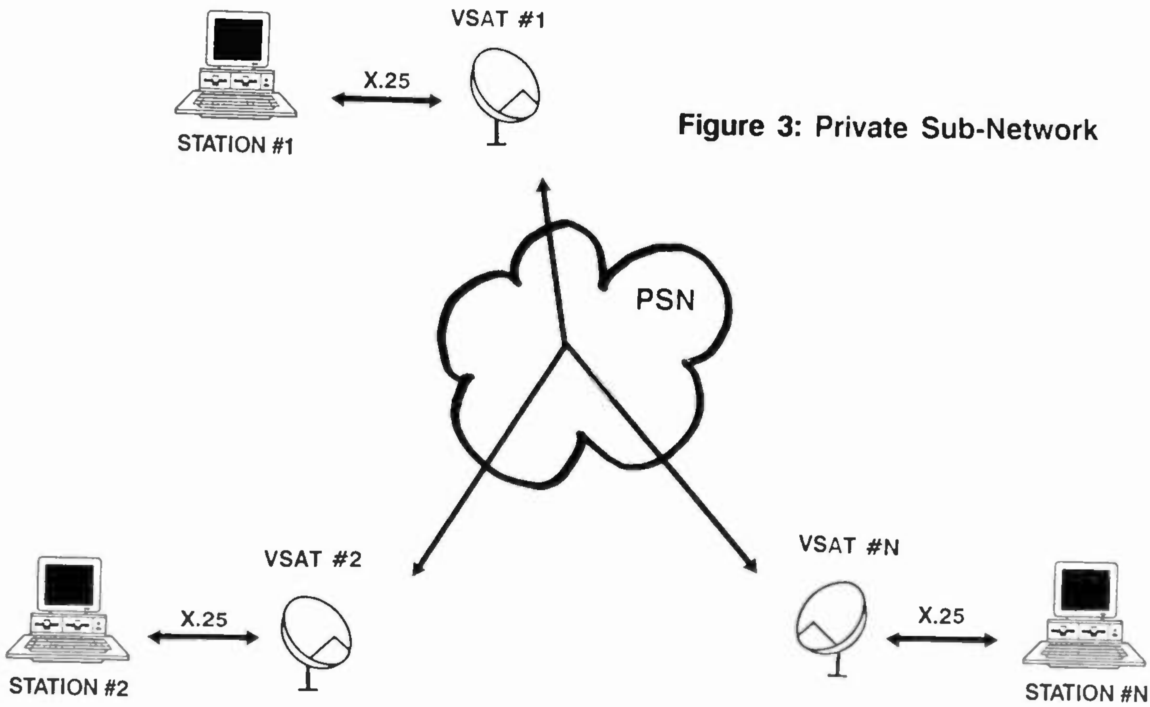

Figure 3: Private Sub-Network

8. Reset the EBS Receiver/Decoder after an alert or test

9. Test the dial-out system to "spot check" local audio or any portion of the system

All corrective actions taken for any alarm are recorded in both the command center and site computers. Readings are taken before and after each action. Statistical analysis is performed on alarm log data and cumulative log data on a regular basis. The resulting graphic reports clearly show trends in equipment and systems performance. Local engineers use these reports for maintenance planning, troubleshooting, and reference. If a station is having problems with a particular piece of gear, custom reports for specific telemetry channels can be created.

Command center operators use trend reports to determine possible causes of multiple or recurring anomalies in client site data. Command center data undergoes analysis to reveal seasonal and national trends. To help forecast alarm loads for geographic regions, NSN duty operators constantly monitor national weather conditions with satellite data that includes radar, jet stream activity, and real-time lightning strike displays. With this information the NSN staff can often anticipate the number one cause of unexpected problems for broadcasters - bad weather.

Satellite Sub-Networks

The inherent connectivity of X.25 PSNs make terminal-to-terminal communications via sub-network a simple matter. For station groups affiliated with the National Supervisory Network, a sub-network is an inexpensive add-on to off-premise control services.

Private sub-networks give groups of stations the ability to interchange information directly between members via the same satellite circuits and equipment used for NSN's off-premise control. (Figure 3) Typical exchanges include: financial spreadsheets, memos, program logs, sales proposals, digitally stored CD quality audio programming, commercials, news actualities, even public service spots. Transmission is faster than any overnight delivery service, much more secure than facsimile, and less costly than tape dubbing and shipping.

Each site in the sub-network must have as basic equipment an IBM-compatible computer with minimum 80286 architecture, 512k RAM, 3Mb free hard-disk space, and an X.25 PAD. For sites wishing to exchange audio or video files, an audio storage/playback card or video storage/playback card is also necessary. In many cases, stations do not need to purchase new computer equipment for sub-network access, as any computer available at a site that meets the minimum requirements may be used after installing the X.25 PAD. It is also possible to run an X.25 PAD in a LAN environment, giving designated users on the LAN access to the satellite.

As in NSN's monitoring and control system, the X.25 PAD establishes the virtual circuit, assembles outbound serial data into packets and disassembles inbound packets into serial data. However, the sub-network virtual circuit (and so the sub-network computer) need not be dedicated. Users may initiate or terminate transfer sessions at will.

Any member of the group can begin a transfer session. At each site, terminal emulation software is used with the PAD to select the VSAT address of the destination, and which files should be sent or collected. Audio spots and video frames are first digitized by storage/playback cards and stored in files before selection. To assure security, sub-network member addresses can only be accessed by the individual user group. Password protection safeguards confidential files. As with any communications program, several types of transfers are offered. These include: ASCII, XMODEM, XMODEM Batch, YMODEM, and YMODEM Batch.

The transmission travels from the call initiator to the National Supervisory Network hub for re-direction. This intermediary step is invisible to end-users. X.25 levels 1 & 2 are read at the hub by the PSN and the data routed to its destination.

So as not to interrupt monitoring, packet data from each computer at a site is sent through a separate port on the VSAT. Thus, an FM-TV combo might use three ports: one for FM off-premise control, another for TV off-premise control, and a third to participate in a sub-network. To accommodate expansion, all standard NSN VSATs are equipped with two data ports.

Conclusion

In this paper, we have presented a broad overview of remote control of broadcast facilities and sub-networking through a Packet Switching Network. The operations of underlying VSAT communications protocols (ie: Slotted Aloha, TDMA, and Reserved Aloha) which work transparently within the described applications, have not been covered here, but are addressed in other NSN papers.

The National Supervisory Network has been conceived and developed by broadcasters to offer the best engineering operations support to the broadcaster at a reasonable cost. This is possible through today's interactive satellite network technology, packet switching techniques, and by the use of computers to augment the years of experience available on the NSN staff. The NSN blend of technology and human experience, working together to anticipate, rather than react; to forecast, rather than just respond; to operate, not babysit each client's station is proving to be the perfect operations solution for many of the remote control problems today's broadcasters face.

Further Reading:

X.25: The PSN Connection
Hewlett Packard, 1985
Part Number: 5958-3402

Data Communications - A User's Guide
Ken Sherman
Reston Publishing Co. 1981, 1985

CORRELATING AM TRANSMITTER PERFORMANCE WITH THE ABILITY TO COMPLY WITH THE NRSC-2 "RF MASK"

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Atlanta, Georgia

ABSTRACT

In order to reduce interference in the AM Broadcast Band, the FCC has modified its Rules regarding off-channel emissions. This paper explores the ability of existing transmitters to comply with these changes. Computer modeling of the RF transmission system is employed.

INTRODUCTION

Background

Recognizing that many AM broadcasters were, due to competitive pressures, transmitting signals which exceeded their assigned channels, the National Radio Systems Committee (NRSC) adopted a resolution on September 5, 1985 to study the establishment of an industry-wide standard for pre-emphasis/de-emphasis in the AM band. (1)

On January 10, 1987, the NRSC adopted a proposal which rigorously defined the audio signal which could be delivered to a transmitter's audio input terminals. This audio standard bore the informal name "NRSC-1" and soon became EIA/IS-40. Shortly thereafter, the Committee also adopted proposed limits on the RF signal which could be radiated by an AM station. This "RF mask" bore the informal name "NRSC-2" and was soon assigned EIA/IS-51. (2)

Responding to a Petition by the National Association of Broadcasters (NAB), the FCC modified Part 73.44 of the Rules to reflect the spirit of the NRSC proposals. (3) Stations must comply with the radiation limitation no later than June 30, 1990. However, stations are waived from having to make the periodic emission measurements required by Part 73.1590(a)(6) until June 30, 1994. Between thoses dates, and "... absent any reason for the Commission to believe otherwise", implementation by each station of NRSC-1 shall be assumed to insure compliance with NRSC-2.

Real-world transmitters

The FCC recognized in paragraph 14 of the Notice of Proposed Rulemaking (NPRM) in Docket 88-376 that delivery to the transmitter of an NRSC-compliant audio signal did not necessarily guarantee an RF output which is also NRSC-compliant. Real-world transmitters have minimal non-linearities... some more than others. This non-linear performance will cause RF sidebands in the transmitter's output which may fall outside of the limits of the standard.

By example: A six-kilohertz fundamental (which is within the passband of the emission limitation) will generate harmonics at 12, 18 and 24 kilohertz and higher frequencies (which are outside of the passband). The greater the transmitter's non-linearities, the greater the amplitude of these sidebands will be. In the extreme, compliance with the standard may be impossible for some broadcasters, due totally to the non-ideal behavior of the transmitter.

Real-world monitoring

Throughout this proceeding, a primary concern of both the NRSC and the FCC was the ability to implement the standard without placing unreasonable costs on

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(1) The NRSC is sponsored jointly by the National Association of Broadcasters (representing radio stations) and the Electronic Industries Association (representing receiver manufacturers).

(2) Copies of the complete NRSC standards are available through the NAB's Department of Science and Technology and the EIA's Engineering Department.

(3) Readers desiring detailed information on this Rulemaking proceeding are invited to explore the Commission's record for RM-6174 and MM Docket 88-376.

the broadcaster for monitoring equipment. The equipment which can directly measure station emissions is not cheap.

The FCC has mandated that measurements made with a "swept-frequency RF spectrum analyzer" shall be the ultimate determination of compliance or non-compliance. Although, prices on this type of equipment have dropped dramatically within the last five years, one should be prepared to pay between $5,000 and $10,000 for its purchase.

Recognizing that ownership of this type of equipment would be a prohibitive burden for all but the most affluent AM stations, the FCC has allowed broadcasters considerable latitude in their measurements: 1) It is not necessary for stations to own measurement equipment; it can be rented, shared or borrowed. 2) When there is no dispute as to their accuracy, indirect measurements will be acceptable. (See Part 73.44(a).)

One such indirect monitor is manufactured by Delta Electronics of Alexandria, Virginia. Specifics on this device can be found in the Record for RM-6714 or can be obtained from the manufacturer.

It is the purpose of this paper to explore additional "indirect" measurements which can be performed with equipment already owned by many stations. Specifically, correlations are developed between the total harmonic distortion (THD) performance of a transmitter and its propensity for creating additional sidebands.

Given the amplitude of all significant sidebands, one could numerically calculate the harmonic distortion percentage. There is a unique distortion value which corresponds to that sideband spectra.

However, it is impossible to work the problem in the other direction. For a given percentage of distortion, there are infinite combinations of sideband amplitudes. As it is therefore impossible to determine with certainty, the amplitude of any given sideband, it is impossible to (rigorously) prove NRSC compliance given nothing more than a (nonzero) distortion level.

However, it is also true that sidebands do not rise and fall in a completely random manner. (e.g. it is unlikely a given imperfection will generate a 7th harmonic and only a 7th harmonic.) A given type of system imperfection will generate harmonic spectra lines which follow repeatable patterns. If the degree of a given imperfection is increased, each of the significant sideband amplitudes will increase (roughly) proportionally.

By knowing the spectra patterns produced by common equipment imperfections, one can make cautious assumptions concerning NRSC compliance.

While these indirect inferences are not definitive (nor can they be), they can be useful in much the same manner as a general physical is useful to a medical doctor. If a patient has an unusual response to a portion of the general physical, it does not necessarily indicate that he (or she) is sick. However, additional, diagnostic tests are indicated which will resolve that question. If the patient has normal responses to all portions of the general physical, even absent the more detailed tests, one can be reasonably assured that the patient is healthy.

RECREATING THE REAL WORLD IN SOFTWARE

The normal arrangement of equipment in an AM broadcast station is shown in Figure 1. The signal travels from the studio, through the NRSC-compliant limiter and the transmitter, into the antenna. A spectrum analyzer is shown receiving an off-air signal. Part 73.44(e) requires that measurements made to insure compliance be performed at a distance of approximately one kilometer from the center of the AM antenna.

Additionally, Figure 1 shows audio being recovered by a precision AM detector (usually in the form of a modulation monitor) and delivered to an auto-nulling, audio, distortion analyzer.

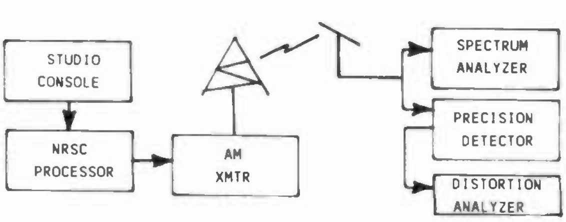


Figure 1
Typical AM monitoring arrangement

Non-ideal transmitter behavior is attributable to distinct and identifiable phenomena, among them are: simple tube non-linearity, PA emission limitation and crossover distortion. Most will recognize the latter as being indigenous to transmitters employing plate-modulation. And all modulation schemes are susceptible to operator misadjustment.

One method to investigate these phenomena might be to simply acquire measurements on a large number of AM transmitters. In addition to the scheduling of downtime and other obvious, logistical problems, this method has two additional disadvantages. First, as each transmitter will have a combination of several non-linearities, each mechanism cannot be observed individually; they can only be observed in the aggregate. Secondly, as many effects are fixed (e.g. iron component saturation), one cannot vary the degree of the anomaly to study the reaction of the system.

One would likely be faced with a number of like-valued data points, clustered about the norm, providing little insight and giving little indication of the behavior of the system with parameters at extreme values.

A simpler (and more revealing) solution is to create, in software, analogues of the transmitter, spectrum analyzer, detector and distortion analyzer. In such a computer model, each anomalous effect can be studied individually. Additionally, the degree of each effect can be varied at will to observe the system response. Such an analogue is shown in Figure 2.

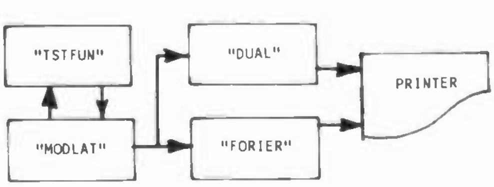


Figure 2
Software Analogue of AM System

The transmitter is replaced by a subroutine called "MODLAT". The subroutine "FORIER" (short for Fourier) replaces the spectrum analyzer. The function of an auto-nulling audio analyzer is emulated by the code segment "DUAL" and the transmitter anomalies are created in the subroutine "TSTFUN". When considering different types of imperfection, only "TSTFUN" is changed. The remainder of the program is unchanged.

This model was written in the FORTRAN programming language and executed on a SUN workstation. Although slightly cryptic, program names are limited to six characters, as is the convention in FORTRAN.

The model user has individual control of the carrier frequency, the modulating frequency and the type and degree of transmitter imperfection. For each set of parameters selected, the model will compute the THD and the sideband amplitudes. To eliminate redundant computation, only sidebands above the carrier are considered. Because there is no phase-modulation present, the sideband spectra will be symmetric about the carrier.

Caveats of the model

To obtain meaningful data from any instrument requires understanding its limitations. Several items pertinent to the performance of this model are relevant.

The definition of harmonic distortion is not as "fixed" as some might believe it to be. The official IEEE definition of harmonic distortion is 100 times the ratio of the root-sum-square (RSS) of the sideband voltages divided by the root-mean-square (RMS) voltage of the fundamental. In practice, an exact implementation in an audio analyzer of this definition has been (until recently) prohibitively complex. For that reason, a second, less rigorous, but easier to implement, definition has evolved into common usage.

This unwritten standard computes harmonic distortion to be 100 times the ratio of the average voltage of the "residual" to the average voltage of the original wave. The typical hardware implementation consists simply of a high-sensitivity audio voltmeter and a sharp notch filter, which can be switched in and out. While reading the signal under evaluation, one adjusts the "set level" control of the analyzer for a reference reading on the meter (usually 100%). The notch filter is then adjusted to remove the fundamental. The average value of any residual signal is then read on the audio voltmeter, expressed as a part of the original. Most modern analyzers have automated the "set-level" and "null" functions.

As the purpose of this paper is not to produce absolute data, but rather to correlate measurements taken with commonly-available instruments, the model was coded to use this second definition of distortion.

The noise floor of the model is approximately -84 dB. This is a function of the finite-length representation of real numbers within the computer. Any values in the following tables which show supression values greater than this amount are anomalous. Declaration of certain variables within the FORTRAN program to be DOUBLE PRECISION would significantly lower the noise floor at the expense of increased computation time. As the greatest supression required by NRSC-2 is -80 dB, this appears to be a needless refinement.

No audio feedback was incorporated in any of the models, as its inclusion would have been counter-productive. Whether a device initially has a minor imperfection, or it has a gross imperfection which is later minimized with feedback, the test results will be nearly identical.

SOME DESCRIPTIONS OF IMPERFECT WAVES

Figures 3 through 6 show exaggerations of the distortion which result from four types of transmitter imperfections. In each case, the idealized wave is defined as:

$$y(t) = \sin(t) \qquad \text{(Eq. 1)}$$

The section of FORTRAN code which models each of the imperfections is shown with the corresponding wave. For ease of comparing results, the code was written so that the maximum excursion of each wave would be 1.0 units in both the positive and the negative direction.

It must be recognized that both NRSC-1 and NRSC-2 are dynamic standards; they are defined for input signals which have broadband spectra and varying amplitudes. The steady-state, sinusoidal tests performed by the model are less than exhaustive. However, the need to correlate to measurements which can be performed with commonly-available equipment makes this a necessary compromise.

This is actually not a fatal limitation. As a member of the NRSC who contributed to the wording of the standards, and as one who has designed a commercially-available NRSC processor, the author is able to draw, not only on the text of the standards, but also on an understanding of why the Committee made certain choices. The primary purpose for the dynamic or "pulsed" nature of the NRSC-1 test signal was to "exercise" the audio processor. Specifically, it was to require that the integration of the limiter/clipper/filter be effective over widely-varying conditions, such as would be encountered in human speech.

Steady-state, sinusoidal analysis could not be used to test this subsystem because both the limiter and the clipper are markedly non-linear devices (both in amplitude and in time). Therefore, superposition would not hold and any steady-state analysis would be meaningless. However, the transmitter itself is, despite the anomalies which are the subject of this paper, a reasonably linear device. Therefore, superposition is valid and one may cautiously extrapolate from the sinusoidal case to infer the more complex cases.

Simple tube non-linearity

The most straightforward imperfection considered is that of simple tube non-linearity. When designing a plate-modulated transmitter, or when employing a linear power amplifier stage, the tube (or transistor) manufacturer's data sheet is a key tool. This sheet contains a family of transfer function curves relating the cathode current to the grid voltage (or the collector current to the base current).

Trading higher idling current for linearity, the engineer attempts to find an area on the chosen characteristic curve which is reasonably straight. In practice, some curvature will always be present. As a result, a perfectly sinusoidal input does not map into a perfectly sinusoidal output. Figure 3 shows the "bowed" wave which results and the code used to model this condition.

Crossover distortion

Closely related to simple non-linearity is crossover distortion. To minimize power supply loading and heat, and to increase tube life, most plate-modulation circuits bias the final, audio amplifers at or near cutoff. Because there is not a sharp knee in the transfer curve, there will be a nonlinearity in the output wave as one tube turns on and the other turns off. Its exaggerated appearance and the code to model it are shown as Figure 4.

Tube Saturation

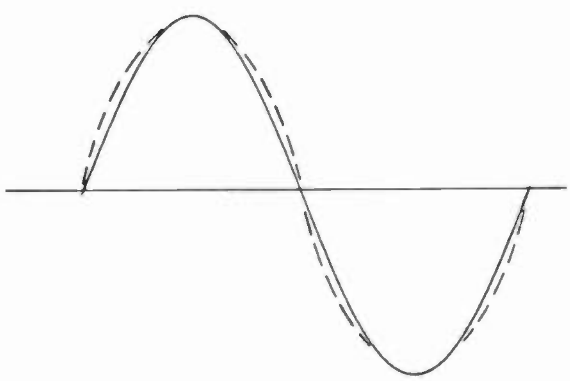
Tubes are less prone to catastrophic failure than they are to simply wearing out. Gas or a falloff in cathode emission will cause a flattening of the transfer curve. In the extreme, cathode current will have an internal limit which is irrespective of the driving condition or plate voltage.

Figure 5 shows the flattening of the positive peak which results from a soft tube in the final, power amplifier. This variation of the model has two programmable variables. One is the threshold where the tube enters the non-linear region. The second is the degree of compression experienced after that threshold is crossed. They are coded as THRESH and SOFT respectively.

This effect applies to schemes other than plate-modulation. A soft tube in the PA of a PDM transmitter will react identically.

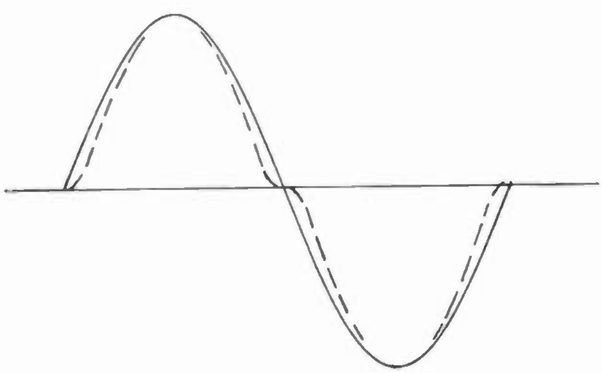
Misadjusted Vector Modulation

Although no longer in production, there are still a number of transmitters in place employing vector modulation. Originally called "outphasing" by the French inventor who developed it in the 1930's, this system was marketed by RCA under the Ampliphase tradename.



```
v=1.-thresh
tstfun=(thresh*sin(radaud*t))+v*sin(radaud*3.*t)
return
end
```

Figure 3
Simple Tube or Transistor Non-linearity



```
tstfun=sin(t*radaud)
amult=1.0
if(tstfun.lt.0.) amult=amult*(-1.)
temp=abs(tstfun)
tstfun=temp**thresh
tstfun=tstfun*amult
return
end
```

Figure 4
Crossover distortion

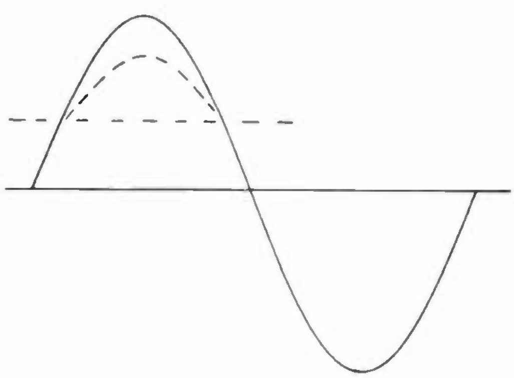
The static phase angle between the currents in the two power amplifiers should be adjusted by the user to 135 degrees for symmetrical modulation, and to some greater value for higher positive peaks. One frequent tuning error is for the static angle to be set at less than 135 degrees. In that instance, the positive peak reaches 100% before it should and then "folds back" on itself at exactly the time it should be at its highest value. The resultant waveshape and the code which emulates it are shown in Figure 6.

VERIFICATION OF THE MODEL

All models should be verified. There are an alarming number of people employing models in an effort to prove things that are, on their face, untrue.

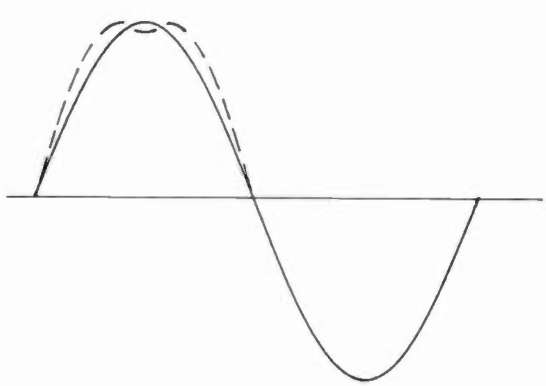
If the reader will permit a crude analogy, computer models are more like transistors than they are like tubes. Usually, the performance of a tube will degrade gradually and progressively, perhaps unnoticeably at first. By contrast, the failure of a transistor is usually immediate, total and quite obvious. Model disfunctions also tend to be gross and obvious, usually the result of flawed assumptions or defective code.

One perennial, favorite example concerns the growth of a bacteria culture. By observing its growth over several hours and projecting that rate into the future, one is led to believe that, by the end of the second day, the culture will weigh more than a car. Clearly, there is an errant assumption somewhere.



```
tstfun=(f1*sin(radaud*t+phi))
excess=tstfun-thresh
if(excess.lt.0.) return
tstfun=thresh+(excess*soft)
return
end
```

Figure 5
Tube Saturation



```
    zz=abs(t)*radaud
 50 if(zz.lt.twopi) goto 100
    zz=zz-twopi
    goto 50
100 if(zz.gt.pi) goto 200
    if(zz.lt.halfpi) goto 150
    zz=pi-zz
150 zz=thresh*zz
200 tstfun=sin(zz)
    return
    end
```

Figure 6
Misadjusted Vector Modulation

Generally, there are two steps to model verification. The first is to compare the model with known data. Obviously, the purpose of the model is to explore unknown situations, but data usually does exist for some instances. Secondly, when the model must be used in uncharted territory, inspect its predictions for plausibility. Are the answers believable? If they show the power output of a motor increasing when the fuel flow is reduced, the model is not yet ready for use.

Table 1 shows the predicted sideband levels and distortion for the "soft tube" model. However, in this instance, the saturation threshold is set to a level greater than 1. In other words, this is an ideal transmitter. One observes that the first sideband (at 1001 kHz) is 6.02 dB below the carrier. All other sidebands above the carrier frequency are at or below the noise floor. This conforms to theory, as the symmetrical sidebands for a sinusoidally-modulated AM transmitter will each have one-quarter the power of the carrier. All other sidebands in this ideal case should be zero.

```
****************************
**                        **
**     PROGRAM NAB90      **
**    TUBE SATURATION     **
**                        **
****************************

Carrier frequency  (kHz) -  1000.0
Modulating frequency (Hz) -  1000.0
Number of sidebands       -  10

Threshold - 1.100
Softness  - 1.300

Harmonic distortion   -    .000%
Fundamental amplitude -   1.000
Fundamental phase     -  45.000 degrees.

Sideband level     .00 dB at 1000.000 kHz
Sideband level   -6.02 dB at 1001.000 kHz
Sideband level  -89.34 dB at 1002.000 kHz
Sideband level  -85.51 dB at 1003.000 kHz
Sideband level  -85.21 dB at 1004.000 kHz
Sideband level  -88.36 dB at 1005.000 kHz
Sideband level  -97.90 dB at 1006.000 kHz
Sideband level  -83.03 dB at 1007.000 kHz
Sideband level  -85.87 dB at 1008.000 kHz
Sideband level  -91.51 dB at 1009.000 kHz
Sideband level  -87.08 dB at 1010.000 kHz
```

Table 1

Table 2 shows the predicted performance of the same saturated power amplifier tube. This time the model is not ideal. The positive peaks are limited to 58%. (Threshold = 0.400, Softness = 0.300) Notice the amplitude of the 1001 kHz sideband has decreased by 1.7 dB. Notice the model has also picked up the fact that the soft tube has reduced the carrier power. The carrier amplitude is now -0.92 dB.

```
****************************
**                        **
**     PROGRAM NAB90      **
**    TUBE SATURATION     **
**                        **
****************************

Carrier frequency  (kHz) -  1000.0
Modulating frequency (Hz) -  1000.0
Number of sidebands       -  10

Threshold - .400
Softness  - .300

Harmonic distortion   -   5.035%
Fundamental amplitude -    .823
Fundamental phase     -  45.000 degrees.

Sideband level    -.92 dB at 1000.000 kHz
Sideband level   -7.72 dB at 1001.000 kHz
Sideband level  -24.87 dB at 1002.000 kHz
Sideband level  -32.83 dB at 1003.000 kHz
Sideband level  -66.65 dB at 1004.000 kHz
Sideband level  -42.19 dB at 1005.000 kHz
Sideband level  -47.31 dB at 1006.000 kHz
Sideband level  -57.87 dB at 1007.000 kHz
Sideband level  -50.02 dB at 1008.000 kHz
Sideband level  -57.80 dB at 1009.000 kHz
Sideband level  -58.80 dB at 1010.000 kHz
```

Table 2

While these cursory observations do not constitute a rigorous verification of the model (which has been done), they remind us of this critical phase of simulation.

EXAMINATION OF THE DATA

Two Critical Understandings

Before proceeding, two key principles shculd be enumerated:

1) The amplitude of the sidebands and the calculated harmonic distortion are not affected in any way by the carrier frequency, and

2) The modulating frequency affects the frequency of the sidebands but not their amplitude or the calculated distortion. The sideband amplitude is determined by the ORDER of the sideband.

Compare the data in Table 3 with that in Table 2. Both are simulations of the "soft tube". The carrier frequency in Table 7 has been changed from 1000 kHz to 540 kHz. It is the same in all other respects. Notice that the harmonic distortion of the two simulations agree within one one-thousandth of one percent and that the sidebands agree within the roundoff error of the machine.

```
****************************
**                        **
**      PROGRAM NAB90     **
**     TUBE SATURATION    **
**                        **
****************************

Carrier frequency   (kHz) -   540.0
Modulating frequency (Hz) -  1000.0
Number of sidebands       -   10

Threshold -  .400
Softness  -  .300

Harmonic distortion    -  5.035%
Fundamental amplitude  -   .823
Fundamental phase      -  45.000 degrees.

Sideband level    -.92 dB at  540.000 kHz
Sideband level   -7.71 dB at  541.000 kHz
Sideband level  -24.86 dB at  542.000 kHz
Sideband level  -32.80 dB at  543.000 kHz
Sideband level  -70.12 dB at  544.000 kHz
Sideband level  -42.07 dB at  545.000 kHz
Sideband level  -47.26 dB at  546.000 kHz
Sideband level  -57.52 dB at  547.000 kHz
Sideband level  -49.79 dB at  548.000 kHz
Sideband level  -57.86 dB at  549.000 kHz
Sideband level  -58.26 dB at  550.000 kHz
```

TABLE 3

Similarly, the data in Table 4 mirrors the data in Table 2 except that the modulating frequency has been increased from 1 kHz to 5 kHz. Again, the results agree within the limits of the computing machine.

Clearly, for a given transmitter imperfection, the amplitude of a sideband is a function only of the degree of the imperfection and the order of the sideband. For that reason, the data in Tables 5 through 8 and Figures 7 through 10 are catalogued by sideband order.

```
****************************
**                        **
**      PROGRAM NAB90     **
**     TUBE SATURATION    **
**                        **
****************************

Carrier frequency   (kHz) -  1000.0
Modulating frequency (Hz) -  5000.0
Number of sidebands       -   10

Threshold -  .400
Softness  -  .300

Harmonic distortion    -  5.035%
Fundamental amplitude  -   .823
Fundamental phase      -  44.999 degrees.

Sideband level    -.93 dB at 1000.000 kHz
Sideband level   -7.72 dB at 1005.000 kHz
Sideband level  -24.89 dB at 1010.000 kHz
Sideband level  -32.99 dB at 1015.000 kHz
Sideband level  -67.45 dB at 1020.000 kHz
Sideband level  -42.40 dB at 1025.000 kHz
Sideband level  -47.98 dB at 1030.000 kHz
Sideband level  -57.43 dB at 1035.000 kHz
Sideband level  -50.54 dB at 1040.000 kHz
Sideband level  -59.41 dB at 1045.000 kHz
Sideband level  -58.77 dB at 1050.000 kHz
```

TABLE 4

Comparing The Mechanisms

One goal of the simulation was to provide varying degrees of each type of imperfection, covering the range between zero and approximately five percent THD.

Data from each model is presented in tabular form and then graphed in a companion figure.

Tabular data for the simple device non-linearity is contained in Table 5. Figure 7 presents this data graphically. The data shows simple non-linearity to be a rather innocuous affliction, increasing the amplitude of the third order sidebands significantly; the first order sidebands, slightly. All other sidebands remain at or below the noise floor. This is to be expected from the form of the code in Figure 3.

Table 6 contains the tabular data from the model of an imperfect crossover. It is graphed in Figure 8. While not quite as simplistic as the data in the previous case, the distortion products are still well-behaved. This is, each iso-distortion line trends smoothly downward.

DATA TABULATION FOR SIMPLE NON-LINEAR DEVICE

Coefficient	1.01	1.02	1.03	1.04	1.05
THD	0.993%	1.974%	2.941%	3.896%	4.839%
Level (dB)					
Carrier	- 0.00	- 0.00	- 0.00	- 0.00	- 0.00
1st sideband	- 5.93	- 5.85	- 5.76	- 5.68	- 5.60
2nd sideband	-90.72	-90.81	-90.90	-90.98	-91.05
3rd sideband	-46.05	-40.02	-36.49	-33.99	-32.05
4th sideband	-85.70	-85.69	-85.69	-85.69	-85.69
5th sideband	-92.41	-92.53	-92.65	-92.78	-92.89
6th sideband	-101.36	-100.98	-100.60	-100.23	-99.88
7th sideband	-84.73	-84.73	-84.72	-84.71	-84.70
8th sideband	-84.60	-84.59	-84.58	-84.57	-84.56
9th sideband	-92.73	-92.55	-92.36	-92.16	-91.99
10th sideband	-88.49	-88.45	-88.42	-88.40	-88.36

TABLE 5

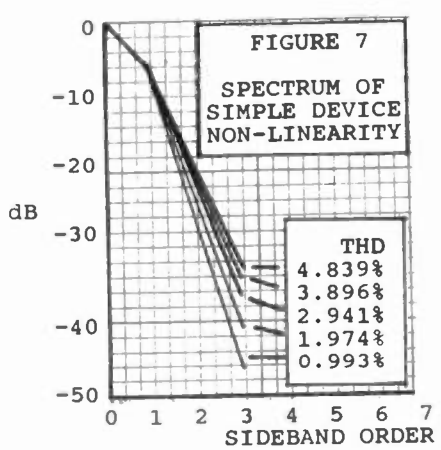


FIGURE 7

SPECTRUM OF SIMPLE DEVICE NON-LINEARITY

DATA TABULATION FOR CROSSOVER DISTORTION MODEL

Exponent	1.050	1.100	1.150	1.200
THD	1.295%	2.563%	3.807%	5.027%
Level (dB)				
Carrier	- 0.00	- 0.00	- 0.00	- 0.00
1st sideband	- 6.10	- 6.19	- 6.27	- 6.34
2nd sideband	-90.58	-90.54	-90.48	-90.43
3rd sideband	-44.25	-38.42	-35.09	-32.77
4th sideband	-85.74	-85.77	-85.81	-85.84
5th sideband	-54.10	-48.51	-45.48	-43.48
6th sideband	-101.66	-101.56	-101.48	-101.40
7th sideband	-60.24	-54.79	-51.92	-50.09
8th sideband	-84.68	-84.75	-84.82	-84.88
9th sideband	-64.56	-59.33	-56.56	-54.86
10th sideband	-88.44	-88.36	-88.30	-88.24

TABLE 6

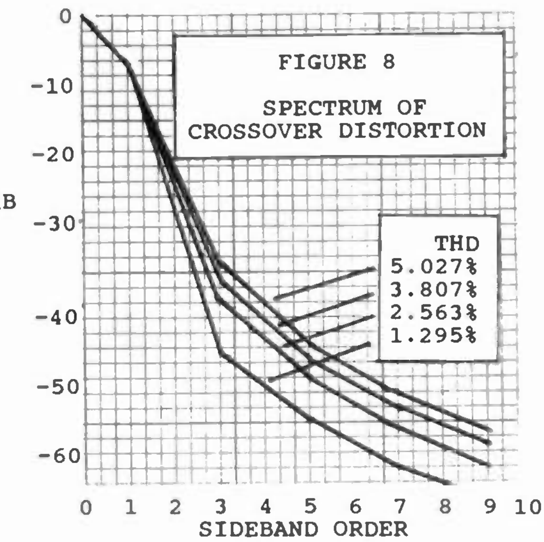


FIGURE 8

SPECTRUM OF CROSSOVER DISTORTION

Effects of the saturated device are shown in Table 7 in tabular form. The first three columns of this data are graphed in Figure 9. The remaining columnns, if also graphed, would appear to be similar. The spectrum is not as well-behaved as the two previous cases. While the amplitude decreases as the order increases, the slope is more gradual.

Predicted performance of the mistuned vector modulator is shown in Table 8 and Figure 10. Despite a pronounced accentuation of the seventh and eighth order products, the trend by the 10th order product is steadily downward.

One Significant Difference

To reduce clutter, the second-order products were not plotted in Figures 7, 8 and 10. They would plot below the lower edge of the drawing in Figures 7 and 8 and below -50 dB for all cases in Figure 10. Figure 9, by comparison, has second-order products which plot higher than the third-order products.

CONCLUSIONS

Aside from the difference in second-order performance, there are more similarities than there are contrasts. Each simulation produces a family of amplitude curves which eventually fall as order increases. Excepting the 10th order product of the vector modulator, the curves do not cross.

DATA TABULATION FOR TUBE SATURATION CASE

Threshold	0.400	0.400	0.400	0.600	0.600	0.600	0.800	0.800	0.800
Softness	0.300	0.500	0.700	0.300	0.500	0.700	0.300	0.500	0.700
THD	5.035%	2.345%	0.769%	1.724%	0.834%	0.284%	0.310%	0.155%	0.055%
Level (dB)									
Carrier	- 0.92	- 0.65	- 0.38	- 0.49	- 0.34	- 0.20	- 0.17	- 0.12	- 0.07
1st sideband	- 7.72	- 7.20	- 6.71	- 6.93	- 6.66	- 6.40	- 6.34	- 6.25	- 6.16
2nd sideband	-24.87	-27.80	-32.24	-28.40	-31.33	-35.77	-35.92	-38.85	-43.31
3rd sideband	-32.83	-35.75	-40.19	-32.85	-35.78	-40.22	-37.86	-40.79	-45.23
4th sideband	-66.65	-69.17	-72.74	-40.87	-43.78	-48.21	-40.87	-43.80	-47.99
5th sideband	-42.19	-45.12	-51.01	-64.73	-67.73	-72.37	-45.34	-48.22	-52.62
6th sideband	-47.31	-50.23	-54.65	-47.44	-50.36	-54.79	-52.76	-55.69	-60.14
7th sideband	-57.87	-60.80	-65.26	-49.18	-52.10	-56.52	-81.53	-84.59	-88.89
8th sideband	-50.02	-52.95	-57.29	-61.09	-64.00	-69.01	-58.67	-61.59	-66.44
9th sideband	-57.80	-62.28	-66.75	-58.29	-61.12	-65.60	-56.34	-58.62	-63.03
10th sideband	-58.80	-61.67	-66.00	-55.30	-58.26	-62.77	-57.92	-60.79	-65.13

TABLE 7

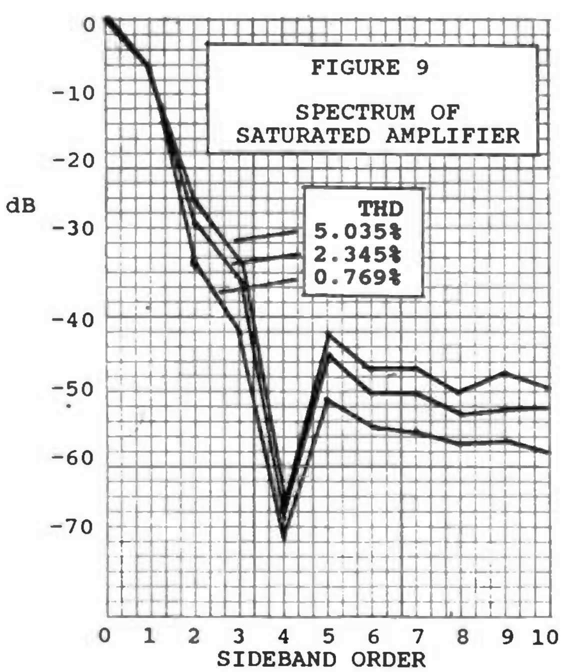


FIGURE 9

SPECTRUM OF SATURATED AMPLIFIER

These similarities are helpful, as they allow us to make generalizations with slightly more confidence than we could if there were marked differences among the models.

Additionally, for a large number of permutations, the spurious performance of the saturated amplifier is the worst of the lot. If one creates guidelines which properly protect this model, it is then logical to assume that they would be at least sufficient for the other cases as well.

To satisfy the NRSC-2 maximums depicted in Figure 11, the following derivations are made from Table 7:

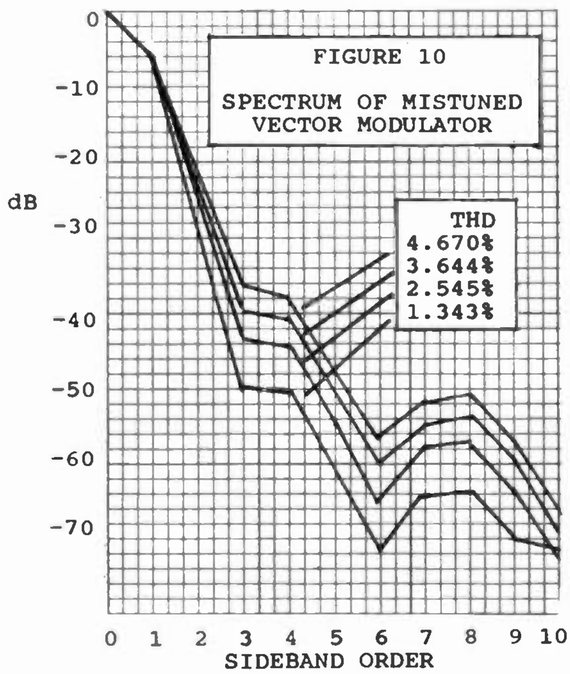
To prevent the second harmonic of a fundamental between 5,000 and 10,000 Hz from exceeding the NRSC-2 -25 dB maximum between 10,000 and 20,000 Hz from the carrier, the system distortion for that fundamental can be no greater than 5.0%.

To prevent the third harmonic of a fundamental between 6,666 and 10,000 Hz from exceeding the mandated -35 dB limit between 20,000 and 30,000 Hz from the carrier, the system distortion for that fundamental can be no greater than 2.5%.

DATA TABULATION FOR VECTOR MODULATION MODEL

Angle ratio	1.050	1.100	1.150	1.200
THD	1.343%	2.545%	3.644%	4.670%
Level (dB)				
Carrier	+ 0.08	+ 0.14	+ 0.20	+ 0.25
1st sideband	- 5.92	- 5.83	- 5.76	- 5.71
2nd sideband	-73.13	-74.58	-59.07	-50.48
3rd sideband	-49.68	-43.08	-39.10	-36.16
4th sideband	-50.56	-44.25	-40.45	-37.68
5th sideband	-60.58	-54.11	-50.19	-47.34
6th sideband	-73.34	-64.46	-60.09	-56.85
7th sideband	-65.73	-58.92	-55.07	-52.28
8th sideband	-63.80	-57.27	-53.63	-50.91
9th sideband	-70.23	-63.62	-59.75	-56.93
10th sideband	-72.55	-73.57	-69.17	-65.94

TABLE 8



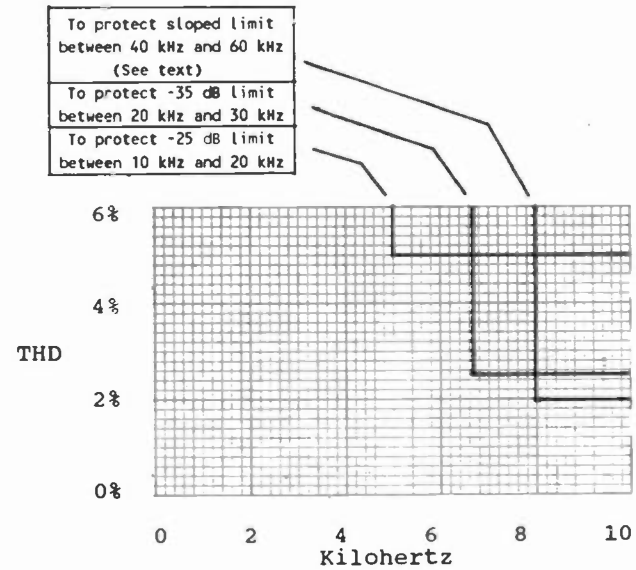


Figure 12

Approximate Minimum THD Performance Necessary to Meet NRSC-2 RF Mask

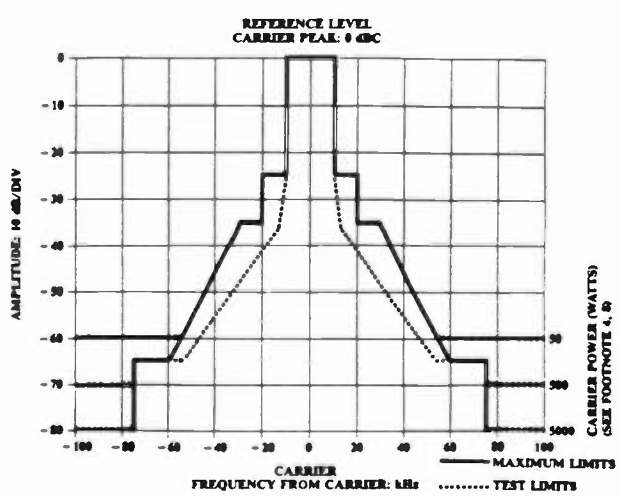


FIGURE 11

NRSC-2 RF MASK
(FROM EIA/IS-51)

To prevent fifth and higher order products of fundamentals between 8,000 Hz and 10,000 Hz from exceeding the sloped maximum between 40,000 and 60,000 kHz, the system distortion for those fundamentals can be no greater than 2.0%. This limitation may be superfluous if selectivity of the RF tank circuit provides appreciable attenuation at frequencies removed from the carrier by more than 40,000 Hz.

Although mathematical conflicts also exist at frequencies greater than 60 kHz from carrier, due to the selectivity of the RF tank circuit, it is unlikely they are real.

These conditions have been been drafted together to form a template in Figure 12.

This template is not exact. It is based on interpolation and a number of assumptions, which may not hold in all cases.

However, it does serve as a starting point, a threshold, beyond which additional equipment should be brought on-site to perform more comprehensive tests.

* * *

AN ANALYSIS OF POTENTIAL INTERFERENCE BETWEEN AM STATIONS SEPARATED IN FREQUENCY BY TWO OR THREE CHANNELS

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ABSTRACT - This report discusses the potential for substantial amounts of harmful interference which may occur to or between AM broadcast stations legally operating on frequencies separated by two or three channels. Due to a limitation in the FCC's rules involving separation of second or third adjacent channel stations, such interference may occur from a station complying with the recently-adopted rules regarding occupied bandwidth of AM stations.

The quality of AM radio reception has experienced a decline for a variety of reasons, but steps are being taken to alleviate some forms of interference which have contributed to the decline. For example, the FCC took a valiant step with the adoption of rules defining and limiting the out-of-band emissions by AM stations, as developed by the National Radio Systems Committee.[1] Due to a long-standing deficiency in the FCC's allocation rules, however, a surprising AM stations meeting the current bandwidth requirements but which are separated two or three channels from a neighboring AM station may cause or receive significant interference with another station. The cause of this deficiency is discussed herein, with examples from an actual case of interference.

From the following description, it is evident that the FCC has considered protections for AM stations separated by two or three channels as contained in §73.37 of the FCC rules. Paragraph (a) of that section states, in part, that "...no application will be accepted...if the proposed operation would involve the overlap of signal strength contours...between the stations involved:

Frequency Separation	Contour of Proposed Station (mV/m)	Contour of other station (mV/m)
Co-channel	0.005	0.1 (Class I)
	0.025	0.5 (Other Class)
	0.5	0.25 (All Class)
10 kHz	0.5	0.5 (All Class)
20 kHz	2	25 (All Class)
	25	2 (All Class)
30 kHz	25	25 (All Class)."

The shortcoming with the above rule is that AM stations usually provide a service area far beyond the allocations contours specified for stations with 20 kHz or 30 kHz separation. For example, solid AM coverage is expected at least to the 5 mV/m contour, and in suburban or rural areas even to the 0.5 mV/m contour. In some cases, then, AM stations operating with two- or three-channel separation may actually be located within the usable service area of their neighbor; when the out-of-band products fall in the neighbor's channel, interference may result.

Television and FM stations are protected to a greater degree, since potentially interfering stations are proscribed from operating within their "protected contour". The Grade B contour of TV stations and the 1 mV/m (60 dBu) contour of FM stations serve as both a demarcation of service and a protected contour for interfering signals.

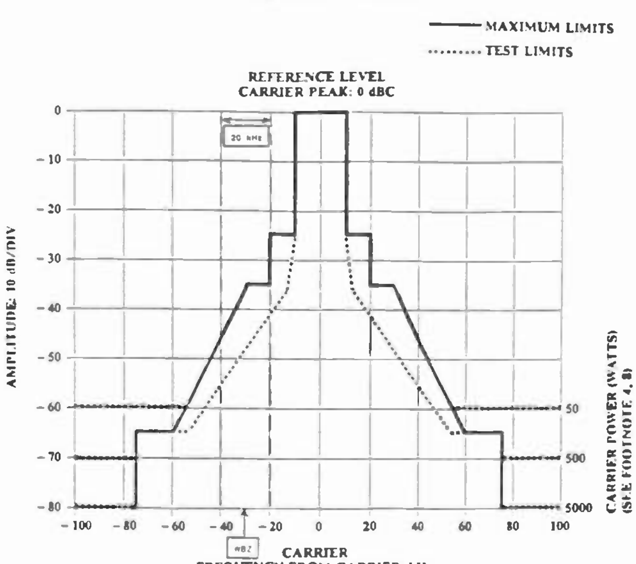
To illustrate the effect of interference to AM stations operating on third adjacent channels, the following case is presented. The situation involves WBZ-AM, 1030 kHz, Boston, Massachusetts, and station WBIV(AM), 1060 kHz, Natick, Massachusetts, located 26.1 miles from the WBZ-AM site.

Initially, in order to depict the extent of harmful interference, it is necessary to determine the carrier ratio for a third-adjacent channel station that results in spurious sideband interference within the desired channel. This ratio is not a simple determination, since it depends on a number of factors, including the peak and average levels of out-of-band energy within the third-adjacent channel for a particular station, the bandwidth and frequency response of the receiver, etc.

For this illustration, the level of permissible third-adjacent channel interference is determined to be equivalent to the FCC's interference standard for co-channel stations. This study also assumes that the interfering station is in minimal compliance with the spurious emissions mask recently adopted by the FCC.

FIGURE 1

AM BROADCAST RF EMISSION LIMITS



The purpose of an RF maximum occupied bandwidth specification, such as FCC §73.44 and NRSC-2, is to control modulation and spurious products that fall outside the necessary RF bandwidth. An example of the AM Broadcast Emission Limits for NRSC-2 are shown in Figure 1. (Figure 1 was adapted from Figure 1 of the NRSC-2 document to indicate the "maximum limits" for a hypothetical station.)

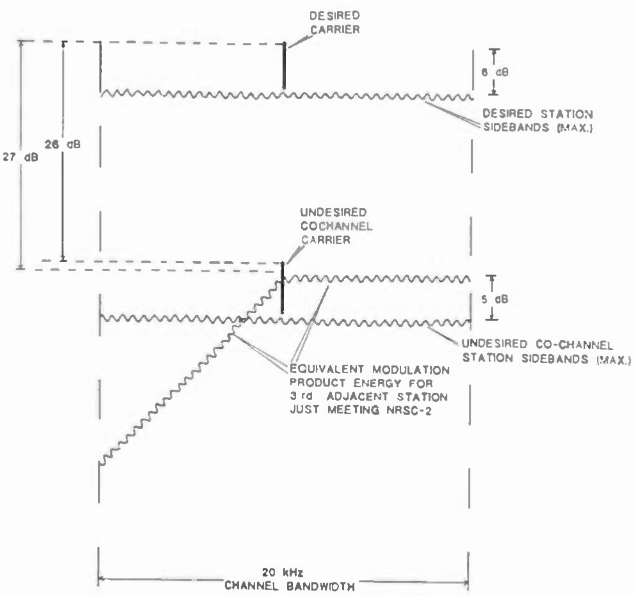
It is evident from Figure 1 that these standards limit the amplitude of out-of-band (unwanted) products at specified frequencies bands above and below the carrier. Therefore, a certain amount of reception interference may be expected for other stations in the vicinity of a given station.

The out-of-band products of any undesired station which result in perceptible interference are essentially co-channel spurious products within the pass-band of the receiver tuned to another station. (That is, as long as the products are not generated within the receiver, such as through front-end overload in blanketing areas.) In other words, the interfering energy 30 kilohertz below a station operating on 1060 kilohertz appears to be in band modulation products to a receiver tuned to 1030 kilohertz.

There are, of course, some obvious differences in the way that linear envelope detectors respond to an AM signal on 1030 kHz and to spurious energy from a station on 1060 kHz. For example, the 1030 kHz station is a double sideband AM signal, whereas the noise and spurious products within the receiver pass-band are uncorrelated relative to the center frequency (1030 kHz) and are not necessarily uniform in amplitude. (The reader will note that the NRSC-2 RF emission limit changes from the center of WBZ's pass-band at 1030 kHz, to the lower side of the pass-band 1020 kHz.)

Figure 2 depicts the level of desired and interfering carriers for co-channel conditions (per §73.37(a)), where the minimum interference suppression is 26 dB (20:1 field ratio). Note that for 100 percent modulation, the sidebands of each station would be approximately 6 dB below the respective carriers.

FIGURE 2



RELEVANT CARRIER, SIDEBAND AND MODULATION PRODUCTS LEVELS DEPICTING EQUIVALENT COCHANNEL INTERFERENCE

It is suggested that the presence of the interfering station's carrier contributes little to the overall interference in this situation; a required carrier frequency tolerance of ± 10 Hz would result in a heterodyne product of 20 Hz or less. A frequency this low would be unlikely to cause an audible effect. (Slight modulation of the desired station's audio may occur due to AGC gain fluctuations.) The principal source of interference are the sidebands of the interfering station which are converted by the linear detector into audible signals 26 dB below the program level of the desired station.

Following on the above, if the interference consisted of noise on the upper side of the desired carrier only, the potential interference would be 6 dB less, relative to full-bandwidth interference. This would apply to the level step of the RF emission mask from -30 kHz to -20 kHz, shown in Figure 1.

Below the desired carrier frequency, the RF emission mask restricts modulation products to a slope of 1 dB per kHz, dropping 10 dB from -30 kHz to -40 kHz. As an approximation, this lower side energy produces an audible interference about 12 dB below a full-bandwidth interference.

By summation of non-coherent noise powers, the combination of both signals results in an interference level approximately 5 dB below the full-bandwidth interference. By comparison with the lowest permissible interference level from a co-channel interferer, then, the energy from the specified sloped-plus-flat spectrum may be 27 dB below the WBZ carrier to produce interference equivalent to maximum permissible co-channel station interference.

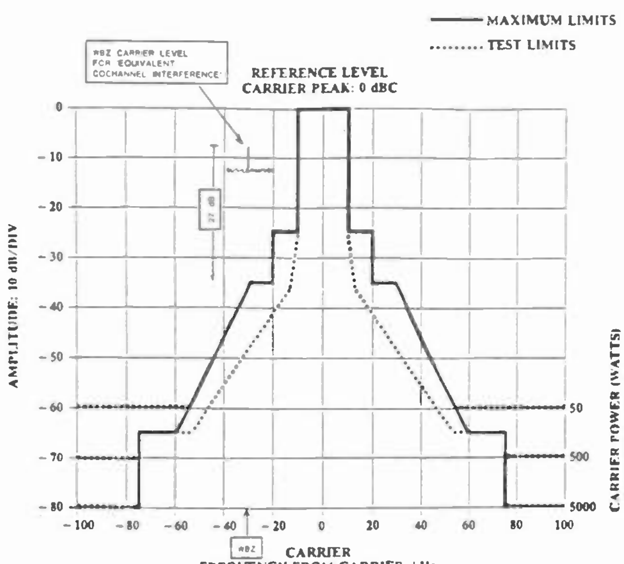
The emission limit of Figure 1 shows the suppression of modulation products to be at least 35 dB at 30 kHz from the carrier. Therefore, an interfering station to WBZ which is in minimal compliance with the emission limit would cause equivalent co-channel interference where the interfering station carrier is approximately 8 db above WBZ's carrier (-27 dB + 35 dB = 8 dB). This situation is depicted in the graph of Figure 3.

With a third-adjacent channel carrier interference ratio expressed in dB, an interference area may be depicted showing where the carrier ratio is exceeded by the interfering station. The map of Figure 4 is based on the methodology discussed herein. Contours of interference to WBZ are the locus of points for which 8 dB and 25 dB carrier ratios are met, corresponding to NRSC-2-required suppression. Contours are based on the predicted field of WBIV for both its presently authorized operation (solid lines) and proposed operation (dashed lines).

Actual spurious energy from WBIV's present operation has been measured with a digital-storage spectrum analyzer of the type specified in NRSC-2. The measurement shown in Figure 5 was made on September 29, 1988, by WBZ's Chief Engineer, and is typical of numerous measurements around the WBIV site. The spectrum photos show the principal sideband energy of WBIV in the center and WBZ offset 30 kHz below WBIV. The out-of-band products of WBIV are evident across the full spectrum display (1010 kHz to 1110 kHz). The data indicate that a typical ratio for WBIV's carrier to its spurious energy interpolated within the 1020 to 1040 kHz band is 25.6 dB. With adjustment for analyzer bandwidth, the spurious level is approximately 40.8 dB.[2]

FIGURE 3

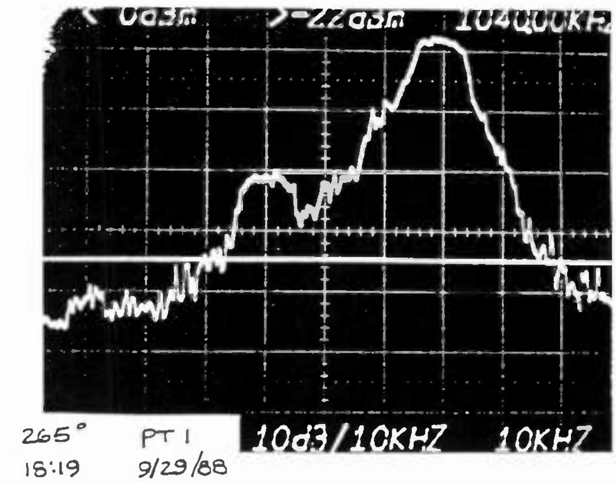
AM BROADCAST
RF EMISSION LIMITS



From Figure 2, the co-channel interference ratio of desired to undesired station sidebands is 32 dB (26 dB carrier ratio + 6 dB carrier-to-sideband ratio). For equivalent co-channel interference, then, WBIV's carrier must exceed WBZ's carrier by 8.8 dB (40.8 db - 32 dB).

The map of Figure 4 includes a depiction of the equivalent co-channel interference resulting from an 8 dB undesired-to-desired carrier ratio; due to the similarity of the 8.8 dB ratio, the 8 dB contour depiction of Figure 4 is assumed for measured interference

FIGURE 5



RF SPECTRUM OF WBIV(AM) NATICK, MA

DATE: SEPTEMBER 29, 1988, 6:19 PM

LOCATION: PINE HILL CEMETARY, 25.7 MILES AT 265°T FROM WBZ-AM

FIELD: 62.5 mV/m (WBIV), 6.3 mV/m (WBZ-AM) (MEASURED WITH FIELD STRENGTH METER)

from WBIV's operation. Figure 4 also depicts the equivalent co-channel interference assuming minimum NRSC-2 compliance, as well as an arbitrary supression requirement of 52 dB.

Based on the contours shown in Figure 4, the areas and population receiving interference from WBIV under present conditions, and NRSC-2 compliance are summarized below.

Interference Conditions	Area(sq.mi.)	Population
Measured WBIV interference (8 dB assumed)	231	209,884
WBIV-CP, NRSC-2 compliance (8 dB ratio)	231	209,884
WBIV-APP, NRSC-2 compliance (8 db ratio)	214	201,704

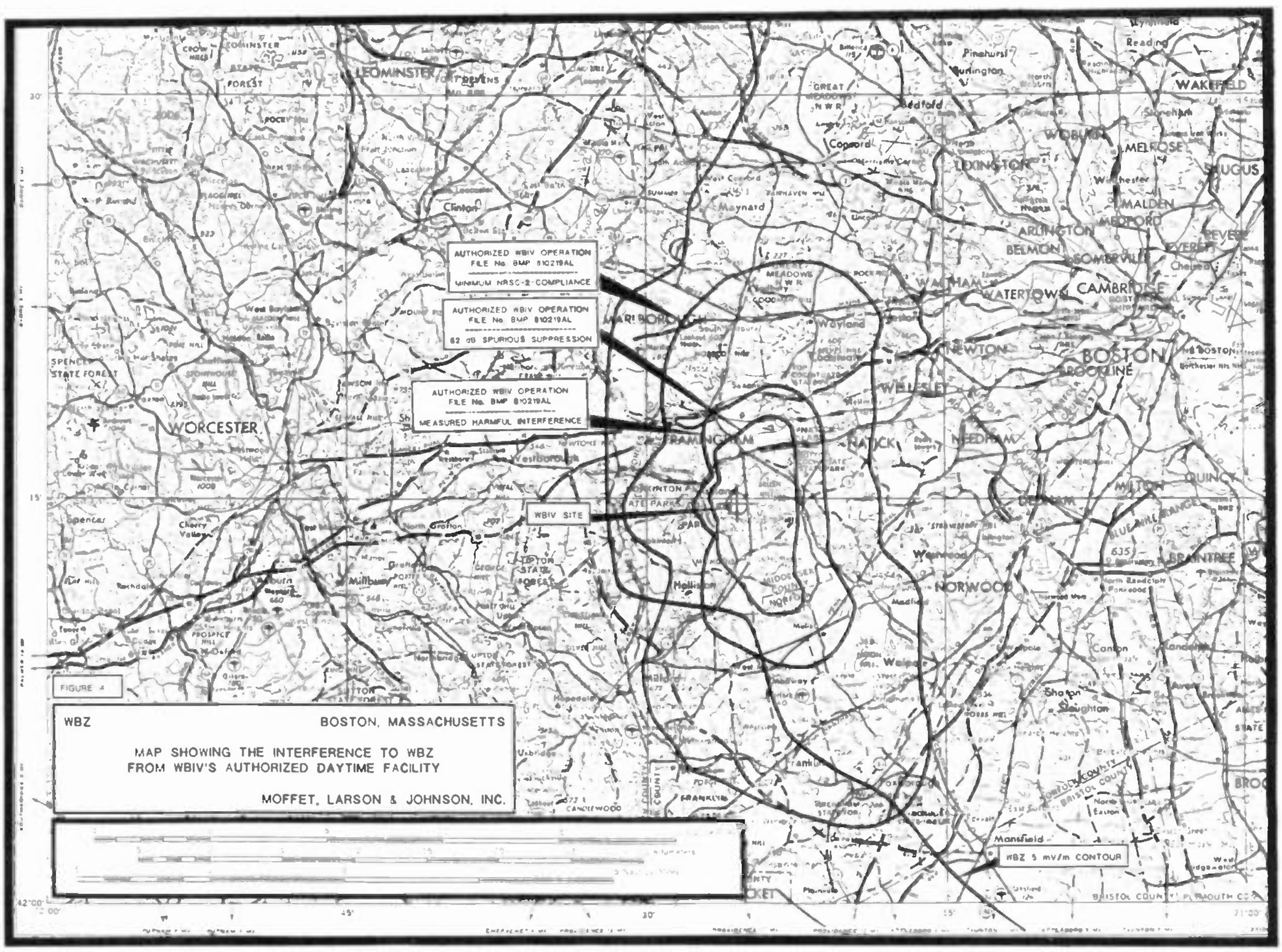
It is apparent from Figure 4 that a substantial area of third-adjacent channel interference may result, even from a station in compliance with the FCC's rules for out-of-band products. However, the NRSC-2 Standard, on which the rule is based, was developed as a minimum standard for AM transmitters now in the field, spanning a broad range in out-of-band emissions performance. While the NRSC-2 Standard encompasses most existing stations, many transmitters can do significantly better than the standard requires.

When two high-powered third-adjacent channel stations are located in a densely populated area, as are WBIV and WBZ, special consideration should be given to the potential for interference to large numbers of people.

The authors wish to thank Norm Avery, formerly Chief Engineer of Group W station WBZ-AM, for his assistance in this project.

1. Incorporated into the FCC Rules as §73.44, occupied bandwidth limits and measurement procedures were based on Voluntary Standard No. NRSC-2, developed by the National Radio Systems Committee, June 1, 1988.

2. A resolution bandwidth of 10 kHz was employed for the measurements; the NRSC-2 procedure specifies 300 Hz bandwidth. The amplitude correction for this bandwidth would be approximately $10^{*}\log B_1/B_2$ or 15.2 dB.



A NEW LOOK AT THE COUNTERPOISE

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Nokomis, Florida

I. INTRODUCTION

When the four wire counterpoise was proposed by Christman and Radcliff at the 1987 IEEE Broadcast Symposium, and with the concurrence of Tom Keller then V.P. of Science and Technology for the N.A.B., I offered to test the concept during the construction of the antenna facility for the test of my VH-Antenna. Extensive data have been gathered in the studies of the counterpoise and numerous calculations have been made including the use of Mininec, some of those data will be useful in explaining the actual operation of a counterpoise system.

On first appearance one might think that a vertical antenna and a counterpoise system is a two terminal network just as we view the typical vertical antenna and ground. When analyzed this way one may get a mistaken opinion of its performance. The vertical antenna and counterpoise over ground is in reality a three terminal network just like a typical two element medium frequency directional antenna and its performance must be analyzed with that in mind. The reasons for this will become obvious in what follows.

The test antenna system was built for the test of the VH-Antenna. The tower was made to be one half wavelength at the proposed operating frequency (1660 kHz) and to permit operation as a receiving antenna at lower frequencies so as to test many aspects of the VH-Antenna operation. Therefore it may not be optimum for the counterpoise test but still leads to a valid interpretation of the counterpoise operation.

II. THE TEST ANTENNA AND COUNTERPOISE TEST SYSTEM

The system consists of a vertical base insulated mast which is a 90 meters in overall height above ground. The base pier and base insulator together are about one meter in overall height. The counterpoise consists of four wires 45.7 meters long oriented at 90 degree intervals and is at an elevation of 4.3 meters. The wires are #10 copper and are insulated from the support posts at the outer end and from the mast at the center. In an attempt to make a symmetrical feed system, the wires were connected by pairs at the mast end and then the pairs were connected together to make a single connection point.

A temporary ground connection consisting of four 8 foot ground rods driven into the earth at the corners of the base pier was installed. After the first set of measurements were made , a 7.2 meter square ground screen was installed and further measurements were made. Finally a conventional ground system of 120 buried radial wires was installed. New measurements on the counterpoise system have not yet been made with the full ground system.

A typical construction job trailer has been located about 150 feet from the base of the mast. Transmission lines for system feed and sampling have been installed and the system is ready for full testing.

III. BEHAVIOR OF COUNTERPOISE

When the counterpoise system testing was begun the writer was considering the system as a two terminal system as it had been discussed by Christman. Since impedance measurements and feed systems for balanced loads are difficult to measure, match and feed, impedance measurements were made from each terminal to ground and in all combinations so that balance to unbalance calculations could be made from the data.

A. THE BEHAVIOR OF ONE WIRE OVER GROUND

Consider first the behavior of a single horizontal wire over a conducting earth. One of the counterpoise wires was examined for impedance and radiation as a function of frequency. The wire is 45.7 meters long, 4.3 meters above earth and the input lead comes down to earth at the point of the mast and the four ground rods. We know that the image of a horizontal wire carrying a current over ground will have an opposing current (boundary condition, tangential E field must be zero). Since the wire is close to ground (in terms of wavelength) the wire and its image behave as a horizontal two wire transmission line with very little radiation. Therefore our model will radiate only a vertically polarized wave due to the vertical current on the feed part of the wire which behaves as a short vertical antenna of 4.3 meters in height. Further its input impedance will consist of a resistance term which is the sum of the radiation resistance and a resistance term which is due to ground resistivity and a reactive term which is due to the input impedance of the open circuit transmission line.

Since the vertical height of that feed wire is small compared to the wavelength the vertical radiation pattern from any current in the feed wire is a cosine function of the angle of elevation and neglecting losses is independent of frequency so long as it is still short.

The radiation resistance R_r of that system is found from the effective height He, the wavelength W and the following equation.

$$R_r = 160x\,^2(He/W)^2$$

The total length of the single wire is 45.7 +4.3 = 50 meters and with the height of 4.3 meters the wire as a transmission line with ground return should have a calculated characteristic impedance of 529 ohms. The quarter wavelength resonance for the open ended transmission line (one wire to ground) should be approximately 1.5 mHz. Then the reactance at 1.66 mHz should be a small positive value. The measured resonant frequency is 1.5 mHz.

Above we indicated that there would very little or no radiation from the horizontal wire so it acts only as top loading for the vertical feed wire which is a short antenna and will radiate as such.

Since the measured impedance of the single horizontal wire and vertical feed wire, acting as a short top loaded vertical antenna has a positive reactance at 1.66 mHz the base impedance has just gone through resonance at that frequency and the current distribution on the vertical wire must be approximately uniform and the effective height is the same as the physical height. Therefore the radiation resistance should be approximately 0.9 ohms.

Table I shows the input impedance of the one wire as measured and the theoretical radiation resistance and expected reactance as calculated from the characteristic impedance.

TABLE I ONE WIRE IMPEDANCE

FREQ.	MEASURED		CALCULATED	
KHZ	R	X	R	X
0.4	17.5	-1060	.052	-1188
1.0	14.8	-283	.33	-305
1.49	20.0	-5.9	.73	-5.5
1.5			.74	0.0
1.51	20	+5.1	.75	+5.1
1.55	26	+26	.79	+27.7

Inspection of the impedance data on Table I indicates that the one wire above ground behaves as lossy transmission line. The loss being due to radiation from the vertical wire and from ground losses associated with the image of the wire. It will be interesting to see if this measured loss can be correlated with ground conductivity which may provide a good method of establishing conductivity, an interesting future project. The loss indicated by the measurements when compared with the theoretical radiation resistance suggest that this would be a poor radiator.

B. THE FOUR WIRE COUNTERPOISE SYSTEM

Since there is more capacitance to earth from the four wire system than from the single wire, the resonant frequency will be lower. From the measurements the resonant frequency appears to be about 1.29 mHz. The exact frequency of course depends lead length and dress at the measurement point.

Tables II and III show the vertically (theta) and horizontally (psi) polarized components of radiation in volts per meter at one meter from one watt (CMF) from the four wire system as calculated by Mininec for 1.66 mHz and 0.4 mHz, (there is no significant horizontally polarized component).

TABLE II COUNTERPOISE
VERTICAL PATTERN
FREQUENCY 1.66 MHZ

THETA DEGREES	E(TH) CMF	E(PSI) CMF
0	.003	.003
15	2.94	.003
30	5.44	.002
45	7.23	.002
60	8.31	.001
75	8.83	.0007
90	8.98	.0000

TABLE III COUNTERPOISE
VERTICAL PATTERN
FREQUENCY 0.4 MHZ

THETA DEGREES	E(TH) CMF	E(PSI) CMF
0	-	-
15	2.47	-
30	4.77	-
45	6.73	-
60	8.21	-
75	9.14	-
90	9.45	-

Past experience with the impedance behavior of short antennas at LF has demonstrated that the reactance as a function of frequency can accurately be represented by the low frequency capacitive reactance of the antenna and the cotangent of the ratio of the desired frequency to the resonant frequency.

$$X = X_0/\mathrm{Tan}(90\ f/f_0)$$

From the measured data the low frequency capacitance should be approximately 1292 picofarads.

From logarithmic potential theory we can calculate the capacitance of one 90 meter long wire (one half of our counterpoise) to ground and by doubling that value we obtain a capacitance of 1135 picofarads. A reasonable correlation.

Table IV shows the measured reactance, the reactance calculated by Mininec and that calculated from the above equation.

TABLE IV COUNTERPOISE REACTANCE

FREQ MHZ	MEASURED OHMS	MININEC OHMS	CALCULATED OHMS
0.4	-283	-317	-274
0.6	-167		-179
0.8	-100		- 96
1.0	- 50		- 49.6
1.2	- 8.3		- 8.5
1.41	+ 21.3		+ 29.5
1.61	+ 59		+ 71.4
1.66		+65.7	
1.8	+ 88.4		+125

While taking the impedance measurements on the four wire system it was discovered that there was a small resonance loop in the Smith Chart plot of the data in the region of 1.66 mHz. Since this was our intended operating frequency its cause was worthy of investigation. Although it was believed that the crew erecting the counterpoise wires carefully measured the horizontal runs, the horizontal lengths being 45.7 meters suggested that slight differences in their length could cause such a loop. Since this was the only speculation available a test was made with the aid of Mininec. By staggering the lengths by 0.5 meter we verified that the loop was in fact a result of slight differences in length.

IV. VERTICAL ANTENNA OVER CONDUCTING EARTH

The vertical antenna was constructed for the VH-Antenna tests and its base insulator and feed point are near the ground. The base of the steel mast is approximately one meter above the earth. The base impedance was measured over the frequency range of 0.4 to 2.0 mHz. For comparison, Mininec was used to calculate the impedance of a 90 meter vertical of radius 0.15 meters.

The impedance of the vertical antenna was measured under a variety of conditions. With the ground system limited to the four ground rods measurements were made without the counterpoise, with the four counterpoise wires up but not interconnected, with the four wires interconnected and with the counterpoise grounded. Except for the case of the grounded counterpoise there was little difference between the various conditions. This is in a large measure due to the high base impedance of the vertical antenna at 1.66 mHz. In spite of the apparent small effect of the counterpoise on the impedance of the vertical antenna there is significant mutual impedance between them and it can not be ignored.

V. VERTICAL ANTENNA WITH COUNTERPOISE

Attempts at measuring the impedance of the vertical antenna with the counterpoise led to confusing results. It was found that the impedance was significantly effected by the measuring set up. Baluns, transmission line transformers and direct bridge measurements were tried. The best data were obtained with the aid of a transmission line transformer that had very large inductance between the ends of the two wire transmission line which had a characteristic impedance of approximately 107 ohms and was wound on a ferrite core.

A. FEED SYSTEM

An adjustable network was made which first tuned the balanced load to near resonance, then used a balun to convert to an unbalanced (coaxial) system and then a transformer to obtain a match to the transmission line. This network was pre-tuned by the use of a matched load at its input terminals and then adjusting the network to exhibit at its load terminals a conjugate impedance match to the target values. In spit of the care in pre-tuning the network required considerable adjustment when placed in the circuit.

A second network was then made with similar but not identical components so as to release the adjustable network for other use. It was also pre-tuned to match the adjustable with concern for only the input and output impedances. When placed in the system it was a complete failure.

Measurements of the longitudinal impedances indicated that there were significant differences between the two networks in terms of the impedance between the antenna terminal and ground and the counterpoise terminal and ground. Recognition of this problem was the clinching argument that the counterpoise system was really a three terminal network. A concept that had developed as a result of several unexpected impedance behaviors observed during the impedance measurements.

B. OPERATION

Under the operating condition it was observed that the counterpoise current was 0.95 amperes and the antenna current was 1.6 amperes, in spite of the care used in making the network the currents to the antenna and counterpoise were not balanced.

Since in normal operation the counterpoise wires will be carrying current, at one point in the phase of the RF cycle a large charge will accumulate at the end of the wires resulting in high electric fields near the wire ends. It seemed that there could well be concern about high levels of non-ionizing radiation. A calculation of the E fields under a counterpoise wire was made at 2 meters above the ground with the counterpoise carrying one ampere. Tables V, VI and VII provide these data.

Sample measurements with a Holliday meter confirmed these values with a current of approximately one ampere in the counterpoise.

TABLE V E FIELD
UNDER ONE RADIAL WIRE

DISTANCE FROM MAST METERS	E FIELD V/M
1	3.65
5	0.687
10	2.63
15	4.20
20	5.56
25	6.71
30	7.62
35	8.25
40	8.38
45	5.27
50	0.98

TABLE VI E FIELD
TRANSVERSE UNDER END OF WIRE

DISTANCE FROM WIRE METERS	E FIELD V/M
10	0.62
8	0.86
6	1.30
4	2.12
2	3.72
0	5.26

TABLE VII E FIELD
VERTICAL BEYOND WIRE END

HEIGHT METERS	E FIELD V/M
1	3.10
2	3.75
3	5.11
4	7.06
5	5.89
6	3.68

VI. THE ANTENNA COUNTERPOISE AS A THREE TERMINAL SYSTEM

Consider the three terminals as being located at the ground level and at or near the base of the tower. For the counterpoise we have short heavily top loaded vertical antenna whose vertical height is the elevation of the counterpoise. For the vertical antenna we have a vertical antenna whose height is the overall height above ground of the tower.

These two vertical antennas are co-linear, or essentially so since the horizontal spacing between the counterpoise lead and the tower is very very small compared to the wavelength. Therefore the net radiation from the aperture from the ground up to the level of the counterpoise is the vector sum of the counterpoise and tower radiations from that aperture.

Now by separately feeding each against ground we can control the amplitude and phase of the currents in the vertical antenna and the counterpoise and we have a pair of stacked vertical antennas.

With the foregoing background consider the problems of feeding the antenna and counterpoise with currents of equal amplitude and 180 degrees out of phase. This is the condition that should obtain if there are to be no currents to earth at the base of the system. In the experimental case the vertical antenna had an impedance of approximately 120 -j300 ohms and the counterpoise to ground had an impedance of 20 +j65 ohms. Further as we well know there is mutual impedance between them and therefore a change in the drive to one alters the operating impedance of the other as in any directional antenna system.

At this point we are entitled to speculate on what may happen to the radiation efficiency of the system as a function of the relative feeds to the vertical and the counterpoise. Note that Carl Smith (Short Low Loss AM Antenna IEEE Trans. on Broadcasting June 1989) has placed an inductance between the counterpoise and ground and tuned it for maximum radiation.

Consider the range of control Smith had with only an inductance between the counterpoise and ground. First visualize the current which will flow in the counterpoise when it is tuned for maximum current and the vertical antenna is driven. Visualize a vector diagram of appropriate scale with the vertical drive current plotted on it at reference phase i.e. 0 degrees. Now plot the counterpoise current on the same vector diagram. Draw the circle whose diameter is the line joining the origin (zero current) and the current with the counterpoise tuned for maximum current. That circle is the locus of tip of the current vector for the counterpoise when tuned with any possible reactance, inductive or capacitive. Further, the area inside the circle is the locus for the tip of the current vector for all conditions where the resistance of the counterpoise is negative.

A quarter wavelength vertical antenna and a quarter wavelength four wire counterpoise were modeled in Mininec to evaluate the changes in radiation efficiency that would result from tuning between the counterpoise and the ground. The model was made with symmetry for the counterpoise wires and as a result they had to be loaded individually. Consequently, the effective load between counterpoise and ground is 1/4 of that indicated.
Table VIII shows the results of this study.

TABLE VIII RADIATION EFFICIENCY

TUNE LOAD OHMS	RADIATION CMF
1E6 - J1E6	9.93
1.0 - J300	9.88
0.3 - J100	9.73
0.1 - J 30	10.88
0.0 + J 0	10.25
0.1 + J 30	10.10
0.3 + J100	10.01
1.0 + J300	9.96

These data show a that a significant control over radiation efficiency is available by adjusting the impedance between the counterpoise and ground. There must also be a region where driving the counterpoise separately will be result in even more radiation efficiency than just tuning with a reactance. There is, but the real choice is now clearly a function of ground system losses as the counterpoise current becomes very large for the maximum gain. This dictates a good ground system in the area below the counterpoise.

VII. CONCLUSIONS

This analysis of the behavior of a vertical antenna with a counterpoise over ground is not complete, but enough has been discovered to indicate to the author that the apparent increase in efficiency, or reduction in ground losses is really a result of antenna gain due to the change in vertical radiation pattern as a result of the use of a sectionalized antenna.

DESIGN CONSIDERATIONS FOR AM DIRECTIONAL PHASING EQUIPMENT

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Pattern bandwidth and common point impedance bandwidth are two major factors considered in the design of a phasing system. In some directional arrays, good specifications can be achieved in a single design.

More often, there is a tradeoff between pattern and impedance bandwidth. For example, one design may provide excellent pattern bandwidth, but poor common point impedance bandwidth. The common point impedance bandwidth can be fixed, but only at the expense of the pattern bandwidth.

In this paper, four phasing systems are analyzed. It is shown how phasor design affects the performance specifications of the system. Some guidelines are presented that will assist in making the best choice for a phasor design.

Introduction

Pattern bandwidth and common point impedance bandwidth are two important considerations in the design of a phasing system. Pattern bandwidth refers to the variation in the patten of a directional array as the frequency is varied from the carrier. Common point impedance bandwidth refers to the variation in the impedance at the common point of the phasing system as the frequency is varied from the carrier. For this study the 10 kHz sideband frequencies of each design will be compared.

There are an infinite number of phasing system designs that will meet the carrier pattern specifications of the station. Each design has its own characteristics. To be able to make a decision as to which design will give the station the best overall performance, many designs should be analyzed and compared. More important is the experience of the design engineer who can eliminate design possibilities. He can also correct designs which can be improved and come up with unique solutions to problems. To assist the engineer software should be available as an aid in the designing and analyzing of a phasing system in a minimum of time. The software should include information about cost, pattern bandwidth, common point impedance bandwidth, and transmitter matching. The software must also be flexible enough to explore many different design schemes.

Each phasing system designed will have its own set of distinct qualities. These qualities must be compared to determine the best design for the station involved. Many times, the specifications for each design vary greatly. In the interest of finding a design that will meet the broadcast needs of the station, different design possibilities should be explored. To show the design process and to demonstrate the fact that different designs can produce widely varying specifications, four different designs will be analyzed. Each design will be done for the night mode of the three tower array of station CHLN.

Good Common Point Bandwidth - Poor Pattern Bandwidth (Design One)

CHLN operates at a frequency of 550 kHz. It should be noted that the 10 kHz pattern or impedance bandwidth that can be achieved at 550 kHz is not as good as can be obtained with the same pattern at a higher frequency. This is because 10 kHz is a greater percentage from the carrier at 550 kHz than it would be at 1550 kHz. The design criteria is then to obtain the best pattern and common point bandwidth for the given pattern at the given frequency.

The field ratios, spacings, orientations, and tower heights are found in Table 1. These parameters are a function of the stations protections and remain the same for all four designs.

Table 1
Array Parameters

Tower	Field	Phase (degrees)	Spacing (Degrees)	Orientation (degrees true)	Height (degrees)
1	1.00	+110	0	0	60
2	1.57	0	100	0	60
3	1.00	-110	200	0	60

The results of the analysis of design one showed an excellent common point bandwidth. The VSWRs at the 10 kHz sideband frequencies are 1.09:1 and 1.11:1. The common point impedance sweep is found in Table 2 below.

Table 2
Impedances - Design One

Frequency (kHz)	Impedance (Ohms)	VSWR
540	52.1 + j 4.0	1.09:1
550	50.0 + j 0.0	1.00:1
560	49.9 + j 5.0	1.11:1

For many phasor designs, if the common point impedance bandwidth is good, the pattern bandwidth is also good. This, however, is not always the case. Figure 1 shows a plot of the 10 kHz sideband patterns produced by this design, compared to the carrier pattern.

Although the sideband patterns resemble the carrier pattern, there are considerable difference even in the main lobe. The next design shows an improvement in pattern bandwidth but the common point bandwidth is not as good.

Good Pattern Bandwidth - Poor Common Point Bandwidth (Design Two)

The results of the analysis of design two showed a decrease in common point bandwidth. The VSWRs at the 10 kHz sideband frequencies are 1.37:1 and 1:33:1. This is considerably poorer than the common point impedance bandwidth of design one. The common point impedance sweep is found in Table 3 below.

Table 3
Impedances - Design Two

Frequency (kHz)	Impedance (Ohms)	VSWR
540	68.2 + j 4.3	1.37:1
550	50.0 + j 0.0	1.00:1
560	37.8 + j 0.8	1.33:1

Although the common point impedance bandwidth is not as good as design one, design two does offer improved pattern bandwidth. Figure 2 shows a plot of the 10 kHz sideband patterns compared to the carrier pattern.

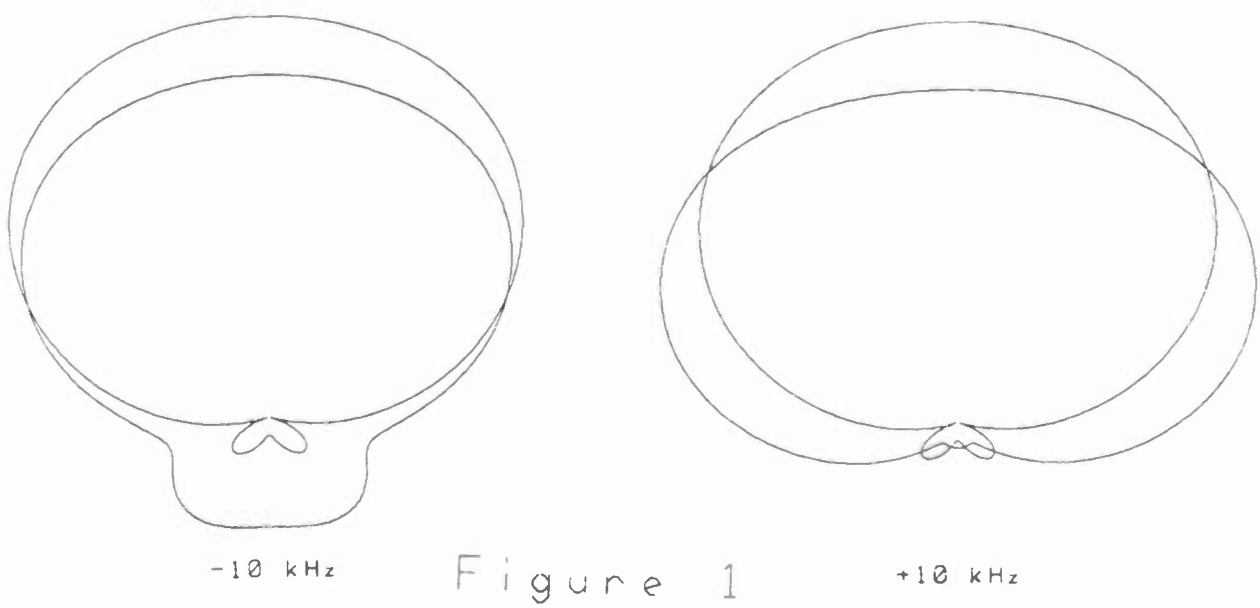

Figure 1

Design one produced a system with excellent common point bandwidth, but had poor pattern bandwidth. Design two showed a marked improvement in pattern bandwidth but a decrease in common point impedance bandwidth. Is it possible to design a phasing system that will yield acceptable common point impedance and pattern bandwidth? Designs three and four do just that.

Good Pattern Bandwidth - Better Common Point Bandwidth (Design Three)

The results of the analysis of design three showed an increase in common point bandwidth. The VSWRs at the 10 kHz sideband frequencies are 1.07:1 and 1.26:1. The common point impedance sweep is found in Table 4.

Figure 3 shows a plot of the 10kHz sideband patterns, produced by this design, compared to the carrier pattern.

The pattern bandwidth for design three is as good, or maybe slightly better than that of design two.

In the next design the common point impedance bandwidth was made more symmetric with only a slight alteration of the pattern bandwidth. This is the design proposed to CHLN.

Good Pattern Bandwidth - Good Common Point Bandwidth (Design Four)

The results of the analysis of design four showed an increase in symmetry of the common point impedances. The VSWRs

-10 kHz **+10 kHz**

Figure 2

Table 4
Impedances - Design Three

Frequency (kHz)	Impedance (Ohms)	VSWR
540	56.8 + j 0.3	1.07:1
550	50.0 + j 0.0	1.00:1
560	42.9 + j 7.8	1.26:1

Although the average VSWR of the 10kHz sideband impedances are lower than the previous design, the impedances are not symmetric. If a transmitter is presented with an uncorrected asymmetric load there may be distortion introduced. It is believed that presenting the transmitter, or more importantly, the transmitter drive with a symmetric load is more important to sound quality than a lower VSWR asymmetric load.

at the 10 kHz sideband frequencies are 1.26:1 and 1.25:1. The common point impedance sweep is found in Table 5 below.

Table 5
Impedances - Design Four

Frequency (kHz)	Impedance (Ohms)	VSWR
540	43.1 + j 8.3	1.26:1
550	50.0 + j 0.0	1.00;1
560	49.9 + j11.1	1.25:1

Figure 4 is a plot of the 10 kHz sideband patterns, produced by this design, compared to the carrier pattern.

-10 kHz **+10 kHz**

Figure 3

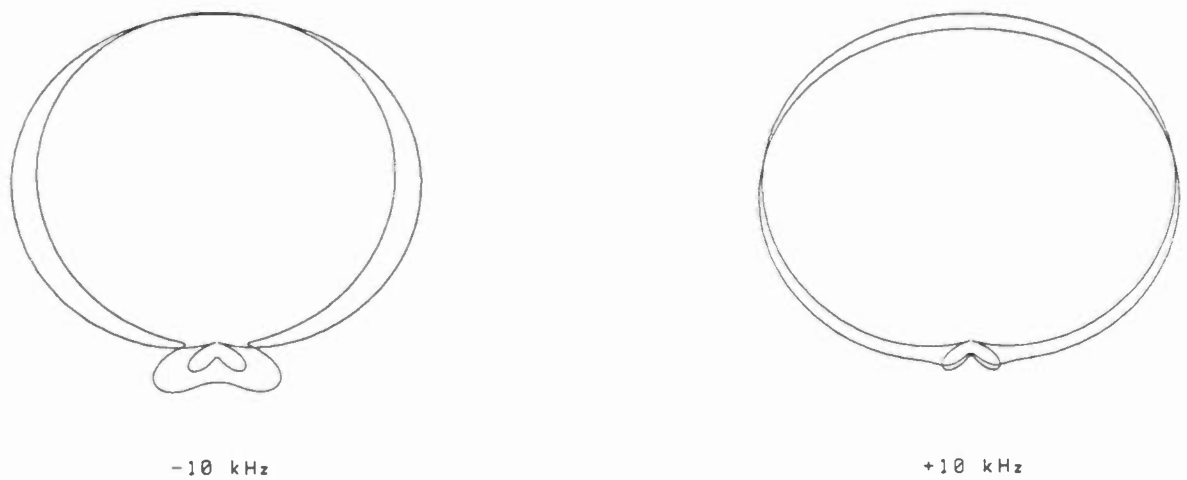

Figure 4

It can be seen that the pattern bandwidth was only changed slightly with a significant improvement in common point impedance symmetry.

Impedance Bandwidth vs Pattern bandwidth

At this point the question might come up "Which is more important to sound quality, impedance bandwidth or pattern bandwidth?". To answer this question a discussion on what each affects is in order.

Common point impedance bandwidth affects how well or efficiently the transmitter can deliver high frequency signals into the antenna system. If a system has extremely poor common point bandwidth the transmitter may not be able to fully modulate. For less poor systems, the high frequency components of the audio signal will be attenuated, causing a bass sounding received signal.

Pattern bandwidth affects the ratio of the sideband frequencies to the carrier frequency signal detected by the receiver. If this ratio is not the same as that transmitted, the received signal will be distorted. Extreme cases of this distortion can be heard in the sharp null area of some directional arrays where the carrier is suppressed but the sideband frequencies are present. The received signal is unlistenable.

Although both common point and pattern bandwidth are important to the sound quality of the received signal, pattern bandwidth is more important when designing a phasing system. Pattern bandwidth is a function of the phasing scheme and networks chosen for the system. Once the system is installed, little can be done to change the pattern bandwidth without considerable rework or at least realignment of the system. These changes to the system also affect the impedance bandwidth of the system. This process can be time consuming and expensive.

Common point impedance bandwidth can be adjusted after the phasing system is installed and tuned without affecting the pattern bandwidth. There have been numerous papers written on this subject. Networks can be installed between the common point tee network and the transmitter to improve the match to the transmitter.

Impedance Bandwidth and Pattern Bandwidth

Of course the ideal situation is to achieve acceptable pattern and impedance bandwidth in the same design. With the use of digital computers and software to easily design and analyze phasing systems, it is now possible to know and make adjustments on phasing systems in the design stage. This is much more cost effective than field work performed after the system has been completed.

Acknowledgement

I wish to thank Kintronic Laboratories, Bristol Tennessee, and Pierre LaBarre and Associates, Montreal, Canada, for their assistance. A special thanks to radio station CHLN.

A HELICAL ANTENNA FOR FM MULTISTATION BROADCASTING

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INTRODUCTION

With tightening restrictions for buildings and mountain tops on the number of antennas, coupled with much more demanding operational specifications on FM antennas and transmission systems to accommodate maximized channel usage, a truly broadband high-power circularly polarized antenna is called for. A truly novel approach to this technical challenge, an axial-mode helix exhibits pseudo-frequency-independent antenna characteristics.

Broadband High-Power Antennas

In the face of tightening construction restrictions for new towers, particularly in urban areas; and with the downtown intermodulation problems large cities growing as more stations locate around a metro market, there is an ever increasing need for a truly broadband, circularly-polarized high-power antenna for FM multistation use. Presently, several antennas are available for this purpose. However, most of these networks have unacceptable pattern and impedance frequency responses, too narrow to qualify as a truly broadband antenna. Our definition of a broadband antenna is a network that exhibits a voltage standing wave ratio (VSWR) of 1.10:1 or better throughout the entire FM broadcast band, 88 - 108 MHz, while at the same time, providing true circular polarization everywhere in the band. Obviously, this is not a trivial design task.

In single or close-spaced applications able to use a narrower-band CP antenna, the physical geometry of the network, including the relative magnitudes and phases of its horizontal and vertical signal radiating centers, is responsible for setting up the space-quadrature polarization relationship between H and V fields, necessary for circular polarization.

An example of this type of network is shown in Figure 1 below:

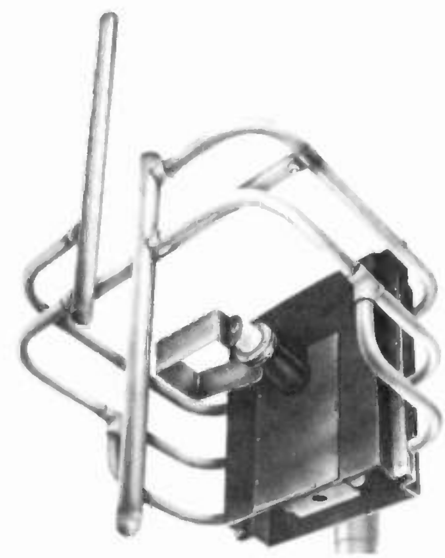


Figure 1

Basically a section of the element functions as a magnetic, or electric dipole, and is responsible for the radiation of the vertically polarized segment of a CP wave. Another dipole in the same element will function as the horizontal. These two radiating centers are set up so they are 90 degrees phase-related at the same electrical radiating point. The result is a circularly polarized wave, propagating away from the element. However, you can immediately see that for any appreciable bandwidth, this approach will degrade the CP signals, due to changing wavelengths operating on a fixed-geometry structure. So, as the excitation frequency is varied, this all-important space-phase relationship is changed, and the once circularly polarized signals become more elliptical because of this change. On an azimuth pattern range, the effect would manifest itself as a changing horizontal to vertical quotient, or axial ratio. A second undesirable characteristic of this type of antenna, when used over a large bandwidth, is its impedance profile over frequency. In any distributed parameter network such as this, the physical dimensions of the network are specific portions of a wavelength at the operating frequency. The ratio of the physical dimensions of the antenna to its input frequency determines other important operating parameters, such as its input impedance. This impedance is greatly affected as frequency is varied above and below the design frequency, causing an increase in VSWR.

Another important parameter when designing a multistation antenna, is its power rating. Obviously, when several 100 kilowatt stations share the same antenna, extreme concern arises about the ability of the system to handle the enormous currents and voltages that will be present. Due to the different frequencies of the input signals in any linear frequency-domain multiplexed system, there will be worse-case events where the equivalent instantaneous peak powers are many orders of magnitude greater than the sum total of all transmitters using the system. This is due to direct voltage vector addition from each transmitter. The worse-case instantaneous power may be calculated using the relation:

$$P_i = N^2 \times P_t$$

Where "Pi" is the instantaneous peak power
"N" is the number of transmitters in the system, and
"Pt" is the transmitter power output

For example, five 30 kW transmitters would produce an instantaneous peak power on an antenna and feed system of 750 kW (5^2 x 30 kW).

It should now be apparent that great care must be used when designing the antenna and its feed system. It must be able to withstand the extremely high peak power without breakdown.

Over the years, many broadband circularly polarized high power FM antennas have been built, intended for use in a master system. Most of these designs are based on pairs of horizontal and vertical half-wave dipoles. Usually, these dipoles are placed over a reflective panel, mounted in a crossed configuration. Their centers are roughly positioned at the same point over the panel, with the dipole axes perpendicular to one another. Each dipole is fed with one port of a quadrature hybrid coupler. One is fed from the reference -3 dB @ zero degree phase port, and the other from the reference -3 dB @ -90 degree port. An important property of quarter-wave hybrid couplers is that over a wide (greater than 20%) band, the equal power division and phase balance remains constant to within better than ± 0.25 dB in relative magnitude, and ± 2 degrees in relative phase. Unlike some antennas that are designed using junction power dividers with a quarter-wavelength line section, this approach ensures true CP over the band.

The crossed-dipole design is quite straight-forward and cost-effective, but the desired continuous broadband impedance characteristics are difficult to achieve over the entire FM band. This is because, as a rule, dipole antenna elements themselves have a relatively narrow impedance bandwidth. By placing these dipoles in a half-wave spaced array, the bandwidth increases, however, this approach does not lend itself readily to producing an omni-directional pattern (± 2.5 - 3 dB) when installed on a three-sided tower. The azimuth beamwidth of each of the three panels per level is much too narrow. The resultant azimuth array pattern contains deep nulls, especially on large facewidth towers. However, this type of panel antenna is good for use on four-sided towers, much more common in Europe.

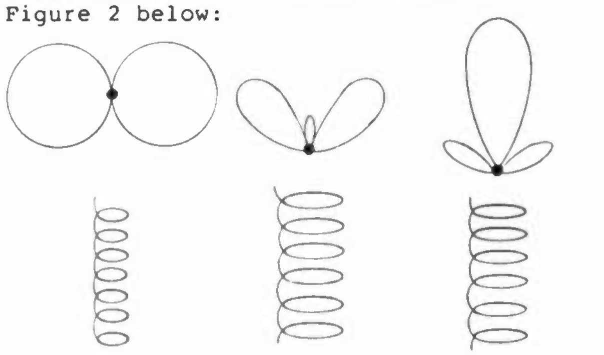
Another style of broadband CP antenna is the so-called cavity-back radiator. This design produces acceptable results, but it is also very large, heavy, complex to build and substantially more expensive.

The Possible Alternative

As a result, we began a research effort to investigate new styles of antennas that would be quite broadband (in both impedance and pattern), circularly polarized, relatively inexpensive, and would not present undo weight and windload characteristics. Many models were researched, among them, archimedean spirals, conical spirals and arrays of log periodic structures. All were members of the so-called frequency-independent antenna group. The axial mode helical-beam antenna was selected as the most practical. Though much more research can be done, some interesting and conclusive results are presented here.

The helical-beam antenna is essentially a simple conductor, wound into a helical shape and placed over a ground plane. This interesting antenna was developed and mainly characterized by J. D. Kraus in 1947. Generally speaking, the helix can radiate in several different modes, however, only three of these are of particular interest for this application. One is the so-called normal mode, which radiates in a direction perpendicular to the helix axis. The second is the axial mode, yielding radiation with major lobes along the helix axis. The third is the conical mode, which produces two major lobes of radiation that are approximately 40 degrees to each side of the helical axis.

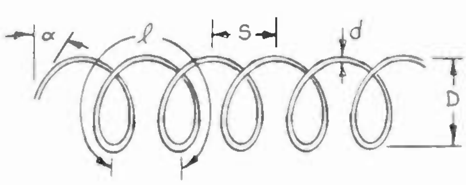
These three radiation modes are shown in Figure 2 below:



Normal Mode   Conical Mode   Axial Mode

Figure 2

Next, general properties of helical antennas are defined and discussed in terms of the unit's geometry. Figure 3 below is a diagram of a basic helix, with its major parameters identified.



D = diameter of helix
C = circumference - pi*D
S = spacing between turns
α = pitch angle = arctan(S/(pi*D))
N = number of turns
L = axial length of helix
d = diameter of helix conductor
l = length of one turn = $\sqrt{(\pi D)^2 + S^2}$

Figure 3

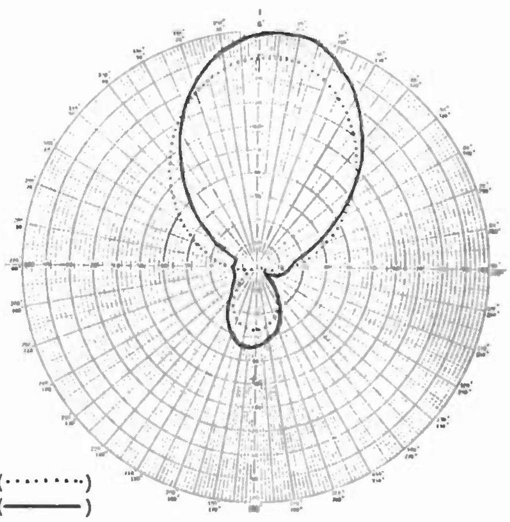
In the helical-beam antenna, the radiation mode is determined mainly by the circumference of the element. Obviously, this dimension is controlled by the diameter, D. When D is adjusted so that the circumference is small compared to a wavelength, a normal mode helix will radiate energy perpendicular to the helical axis. The impedance characteristics of this type of helix are extremely frequency sensitive and therefore not acceptable for our application. Alternately, if D is increased so that the circumference of the element is larger than a wavelength, then the conical mode will be excited. This creates a unit elevation pattern with almost all energy directed at the ground and sky, and very little at the horizon. Finally, when D is made such that the circumference is about one wavelength long at the band center, a so-called axial mode helical-beam antenna results. This is the helix style that is of greatest interest for our application.

The axial mode helical-beam antenna possesses a number of interesting properties. It has both wideband impedance characteristics and circularly polarized radiation. These two specific properties were exactly what attracted our attention when an alternative antenna design was sought.

Generally, in these antennas, both the axial beamwidth and horizontal and vertical (H-V) field ratio characteristics vary with the number of turns in the element. Paradoxically, they are opposite. The greater the number of turns, the better the circular polarization. Unfortunately, a greater number of turns will also result in a narrower axial beamwidth.

Initially, several model axial mode helices were built for pattern testing on our range. The pitch was determined according to the design rules outlined earlier (Fig. 3). Due to practical limitations imposed by a three-around tower mounting, and because initial research into the ratio of turns number versus axial beamwidth, "N" was initially selected to be two and one half. The first order of business was to measure the 3 dB beamwidth and H-V field ratio of this model over a range of frequencies covering the FM broadcast band. Since a three-around configuration was chosen, the azimuth beamwidth must be wide enough to allow relatively even and smooth illumination over the entire 360 degree range when the elements are placed at 120 degree intervals. As seen in the field patterns below, initial results indicated two characteristics to be of concern. First, for the elements to array smoothly and form an even, omni-directional pattern, the 3 dB beamwidth of each had to be a minimum of 120 degrees.

However, as seen in Figure 4, the beamwidth of the initial test unit was much less.



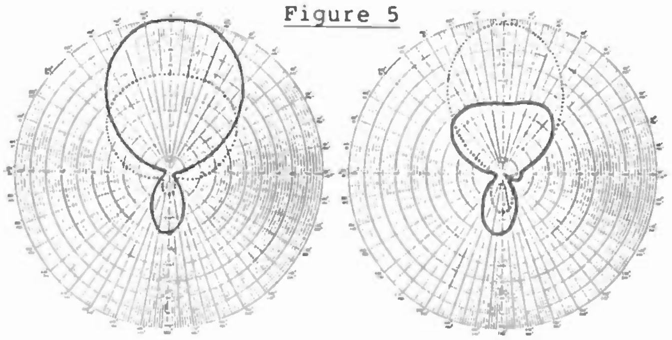
V (........)
H (———)

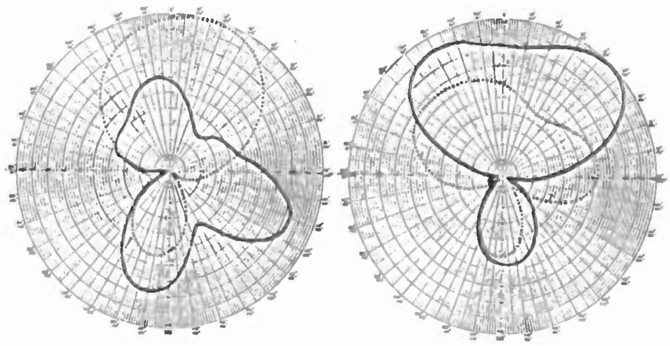
Figure 4

The second concern, as we continued in the development of this system, was its H-V field ratio. Generally, axial-mode helical-beam antennas that are relatively short (the number of turns is less than five) tend to radiate signals whose polarizations are slightly elliptical in nature. So since the test helix had only half that number, we expected a slightly elliptically polarized signal to result, which it did. There is a slight variation in H-V field ratio with frequency, as seen in Figure 5 below. This was disturbing, because in order to improve the azimuth beamwidth, the element had to be shortened even more, and that would surely result in H-V field ratio frequency sensitivities that were even worse.

In order to evaluate these effects, several helices of different dimensions were constructed and pattern tested. As illustrated by the data in Figure 5 below, these altered designs produced a variety of results, some good, some bad.

Figure 5





Subsequently, different grounding schemes were tried and produced small, but successive improvements.

Additional variations of the antenna were then tried. Diameters were varied, and the numbers of turns were altered. Also, various sizes and styles of reflectors were tried. All of these produced results which provided a greater understanding of the characteristics and behavior of the fascinating network emerging.

One particular overriding concern was that the radiation center of each helical-beam element in the three-around array would move further out, away from the reflecting ground plane while operating at lower frequencies, and move further in at higher frequencies. This movement, though small when compared to the corresponding change in wavelength as the frequencies are varied, still imposed another layer of frequency sensitivity on the azimuth array pattern.

A complete set of azimuth array patterns confirmed this exaggerated frequency sensitivity. Due to the changing position of the radiation centers over frequency, the resulting azimuth array pattern is not uniform. As shown in Figure 6 below, a degradation from an optimum array pattern (± 2.5 dB), is displayed as a result of this movement.

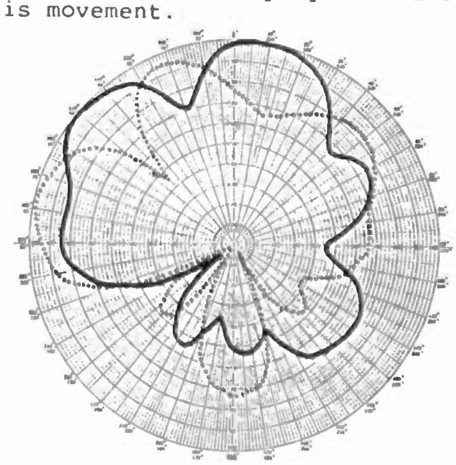


Figure 6

We next investigated the impedance bandwidth of this family of antennas.

Some of these results are illustrated in Figure 7 below:

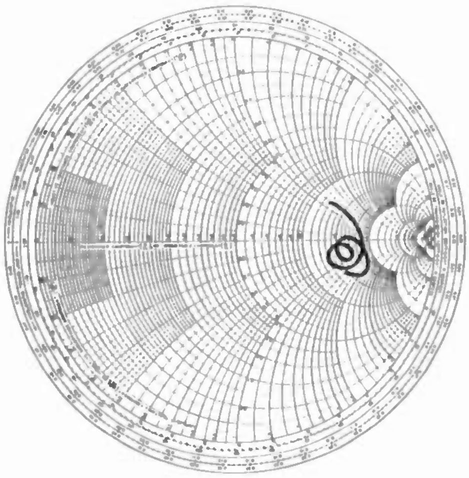


Figure 7

As can be seen, the impedance bandwidth of the axial mode helical-beam antenna has good broadband characteristics. Even though the actual value of the driving-point impedance seems high, about 135 ohms, the important characteristic is that the impedance coordinate does not change much over a wide range of frequencies. This means that with the appropriate matching networks, this element would exhibit vary favorable impedance bandwidth characteristics.

Some Promising Results

We have seen a number of patterns, impedance diagrams and discussions that demonstrate the general behavior of the axial mode helical-beam antenna under a variety of design rules, and help our understanding of this type of antenna. Still unanswered are questions of real-world practicality, and the resolution of design parameter trade-offs, particularly axial beamwidth versus circular polarization.

Of course, ongoing efforts involving other configurations of helical-beam antennas continue. Some of these include nonresonant hybrid-fed bifilar helices with design variations different than those presented in this paper. But by keeping in mind the information gathered in this project, we can move on and apply the insight gained to the next.

Bibliography

Antennas, J.D. Kraus, copyright 1950, McGraw Hill Book Co.

Antenna Engineering Handbook, Richard C. Johnson and Henry Jasik, copyright 1961, McGraw Hill Book Co.

FM STEREO, WHAT CAN BE EXPECTED FROM THE NEW TECHNOLOGIES

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ABSTRACT

The introduction of FMX® Stereo, noise controlled blend, Walsh function decoders, diversity automobile receivers, and reduced transmitter synchronous AM can be expected to improve stereo FM broadcasting. FMX Stereo will significantly extend the stereo coverage of FM stations. With the few stations where the first adjacent channel mileage separations are not met, interference may be encountered. Receivers combining the Walsh function decoder with FMX Stereo noise reduction will reduce the first adjacent interference, allowing for an even greater increase in station stereo coverage. The introduction of noise-actuated blend, a system that adjusts blend as a function of signal-to-noise ratio, will minimize high RF level multipath problems. This paper describes how these new technologies and broadcaster controlled synchronous AM combine to improve the FM stereo broadcast system.

INTRODUCTION

To prevent erosion of the FM audience to new audio delivery systems, the FM broadcaster must continuously upgrade the quality of the transmission system. The objective is to upgrade the existing technology to a level that will allow the local terrestrial broadcaster to compete with the new delivery systems on a level playing field. The implementation of new technologies that improve the quality of the existing transmission system should be expedited.

TECHNICAL OBJECTIVES

The FM broadcast should be capable of delivering full bandwidth 15 kHz stereo sound, with minimum distortion and noise. The broadcaster should generate a signal that can be received by the largest audience for the class of station, with the least distortion, and cause the least amount of interference to other stations' signals. The receivers should be capable of reproducing the full quality of the signal the broadcaster transmits and be the least susceptible to transmission conditions. The system also should be capable of competing with the new technologies like compact disc or digital audio tape.

Conventional Blend

Stereo transmissions have been accompanied by a significant increase in received noise that has resulted in a decrease in stereo coverage. With the stereo baseband signal extending to 53 kHz, the FM noise level increases 6 dB per octave with increasing frequency, the so-called triangular noise (Fig. 1). To compensate for the stereo noise, receiver manufacturers have incorporated an electronic circuit called blend. This circuit uses the received radio frequency signal level to determine the amount of blend needed to reduce stereo noise and multipath distortion. Receiver blend circuits reduce the stereo difference signal at the expense of stereo separation. The weaker the signal, the greater the blend. Figure 2 is a composite plot of several auto radios' stereo separation versus received signal level. The measurements were made on contemporary receivers. The vertical axis represents separation and the horizontal axis the signal level in dBf.

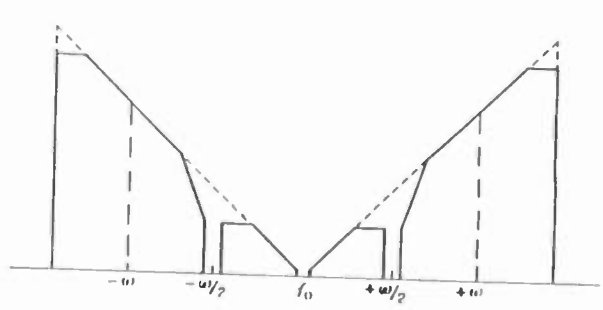

Figure 1

FM NOISE SPECTRUM

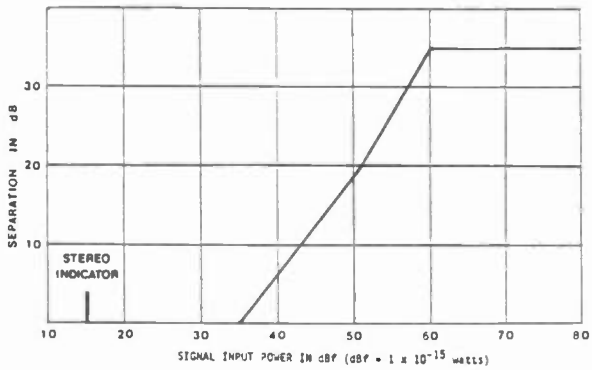

Figure 2

STEREO SEPARATION vs. SIGNAL LEVEL

High Cut

In addition to blend the automotive receiver uses the high cut function as a final resort in reducing noise. This circuit also operates on received signal level and will reduce the high frequency response of the audio channel. Figure 3 shows the audio frequency response of a representative receiver with varying levels of radio frequency input. The high cut is activated only when the receiver is in blend.

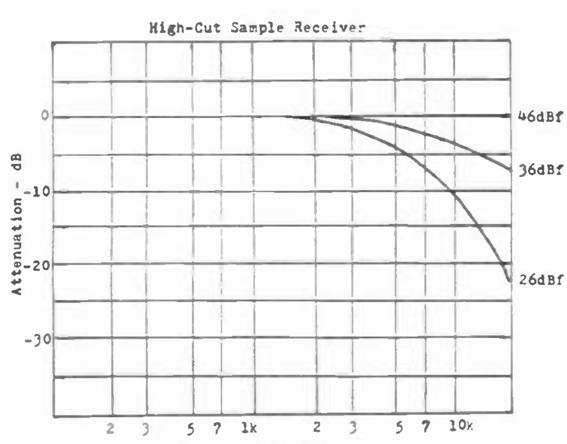

Figure 3

HIGH-CUT SAMPLE RECEIVER

Stereo Indicator

Tests show the activation of the receiver stereo indicator has little to do with the reception of stereo, but it does confirm that the station is broadcasting in stereo. Stereo indicator thresholds vary from receiver to receiver. Non-FMX Stereo receivers averaged about 20 dB below the 100% blend point (Figure 2). Stereo reproduction for this 20 dB of signal range is a figment of the receiver's imagination.

New Blend Systems

Sprague in their FMX Stereo decoder IC has developed a noise operated blend circuit that is an improvement over the conventional signal level system. The circuit design assumes that the spectrum around the 19 kHz pilot is clear of program signals and is usable for measuring the received noise (Figure 4). Sprague uses a quadrature detector to measure the noise and distortion in the pilot area. With the quadrature detector, the system blends on distortion or noise. This means that the circuit will blend on distorted or noisy signals at any RF level.

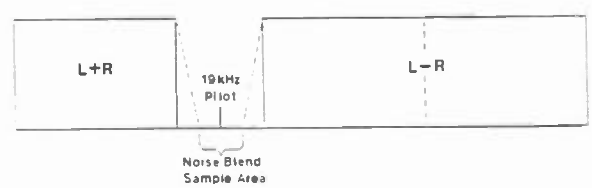

Figure 4

SPECTRUM FOR NOISE CONTROLLED BLEND

FCC 73.322 states that "An FM broadcast station shall not use 19 kHz +/- 20 Hz, except as the stereophonic pilot frequency." Sprague has monitored many stations and found that the spectrum 2 kHz above and below the stereo pilot is free of modulation and has concluded that this space is usable for signal noise and distortion sensing.

INTRODUCTION OF FMX STEREO

Blend has generally increased the coverage of FM stations at the price of stereo coverage. The problem is that a good portion of stations' listeners are located in areas where the signal level results in a blended receiver. This has resulted in an increase in station coverage with the loss of stereo, seriously limiting the ability of FM broadcasters to compete with new technologies. The introduction of FMX Stereo, a system that reduces noise in stereo reception by compressing a new quadrature difference signal (Figure 5) before transmission and expanding it to its original dynamic range at the receiver, allows the receiver designer to reduce significantly the threshold level for blend

[1]. FMX Stereo with its 14 dB of stereo noise reduction allows radios to be built with full stereo separation down to input levels of 40 dBf and with usable separation to 32 dBf. Figure 6 shows the stereo blend curve and the FMX Stereo blend curve for a JVC Model KS-RX5500 automobile receiver. One can see that stereo coverage is greatly improved with the new system.

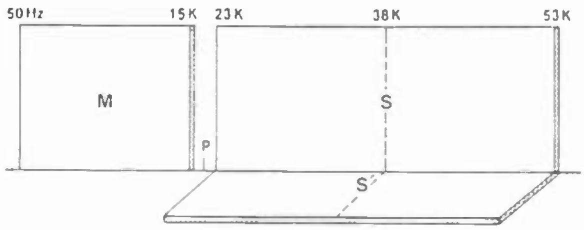

Figure 5

FMX STEREO BASEBAND SPECTRUM

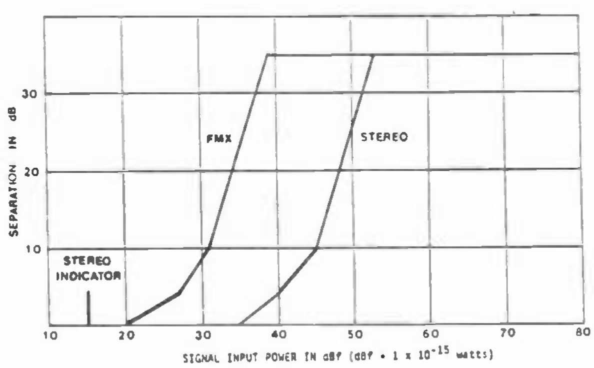

Figure 6

FMX STEREO SEPARATION vs. SIGNAL LEVEL

FMX Stereo is designed to reduce FM stereo noise, thereby extending the stereo coverage of the broadcast station. It was originally thought that stations that used moderate or no processing would be the only stations to experience the benefits of FMX Stereo. However, blend has changed this situation. The stereo receiver blend circuit responds to radio frequency signal levels and blends all weaker signals to mono, regardless of station format or the amount of processing used by the broadcaster. Therefore, the new system extends the stereo coverage lost by blend to all FM stations.

Walsh Function Decoders

The stereo decoder in the FM receiver is the device designed to convert the composite stereo signal to the left and right component signals [2]. Two types of stereo decoders, frequency division and "phase-locked loop," are shown in Figures 7 & 8. The PLL decoder is the only type commonly used in today's receivers. For comparison purposes, the following is a review of frequency division and PLL stereo decoders.

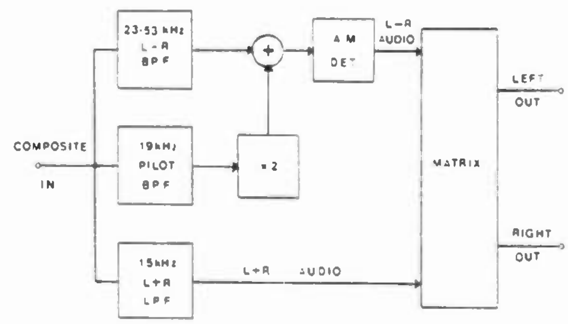

Figure 7

FILTER STEREO DECODER

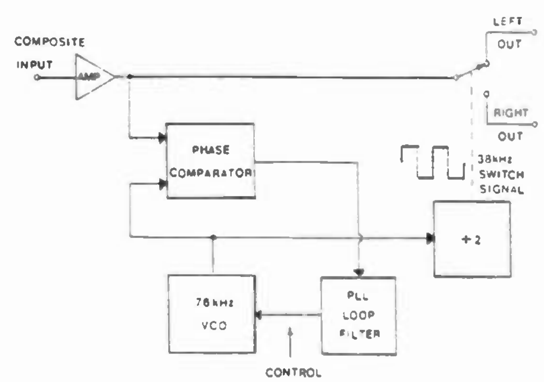

Figure 8

CONVENTIONAL PLL STEREO DECODER

In the frequency division decoder the composite signal is filtered to L+R and L-R and the pilot is doubled to 38 kHz (Figure 7). The regenerated 38 kHz carrier is then reinserted into the 38 kHz double sideband stereo difference signal and AM detected to the L-R. The L-R and L+R signals are then matrixed to produce the L and R signals. Frequency division decoders have been used in modulation monitors when individual metering of the components is required. The high cost of the filters has ruled out the use of the frequency division decoders in consumer equipment.

The PLL stereo decoder in Figure 8 is really a time division demultiplexer. The pilot is phase compared with the 76 kHz oscillator, and a control voltage locks the 76 kHz oscillator to four times the pilot. The VCO signal is divided by two and the 38 kHz square wave drives the decoder switch. The switch operates in synchronism with the switch in the broadcast encoder (stereo generator). The PLL decoder, because of its comparing square waves with sign waves, has the disadvantage of being sensitive to the odd multiples of 38 kHz. This results in decoding the odd multiple signals at 114 and 190 kHz, making the receiver sensitive to the first adjacent channel (Figure 9).

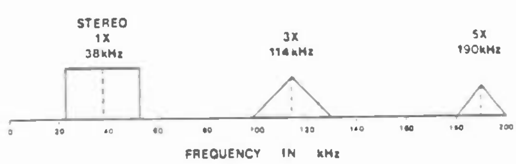

Figure 9

SENSITIVITY SPECTRUM FOR PLL DECODER

With the introduction of the Walsh function stereo decoders, the PLL decoder is no longer sensitive to the odd multiples of 38 kHz. This reduces the decoder's sensitivity to 190 kHz, resulting in an increased first adjacent channel rejection of 20 to 25 db. Therefore, the use of PLL stereo decoders with Walsh functions has a major effect on the coverage of FM stereo stations.

Receiver RF Dynamic Range

Receiver design has been constrained for many reasons, but most usually to compensate for stereo noise. The RF gain of a receiver may be set higher than good intermodulation performance will allow. With the introduction of FMX Stereo, the need for a super-sensitive receiver to overcome triangular noise is eased. This will give future designs greater latitude in minimizing the effects of blanketing and adjacent channel interference.

Diversity Antennas

Torick and Finger describe the advantages of diversity automobile receiving antennas in a paper published in the IEEE transactions on Consumer Electronics [3]. Several other papers have been written on the subject [4].

A limited supply of diversity receivers has been on the market for about four years. The receivers' diversity performance has been good to excellent. But the advantages of the technology has been one of the best kept secrets from the American motoring public.

Figure 10 shows a representative plot of relative carrier level (field strength) along a line in space. Fluctuations may or may not result in audible multipath events. Moving a single receiving antenna only a few feet usually will improve reception. This has made possible space diversity reception for the automobile. Figure 10 shows the improvement that could be realized if a 9 foot antenna space diversity system were used. It can be seen that a more practical antenna spacing of less than 9 feet will also give significant improvements.

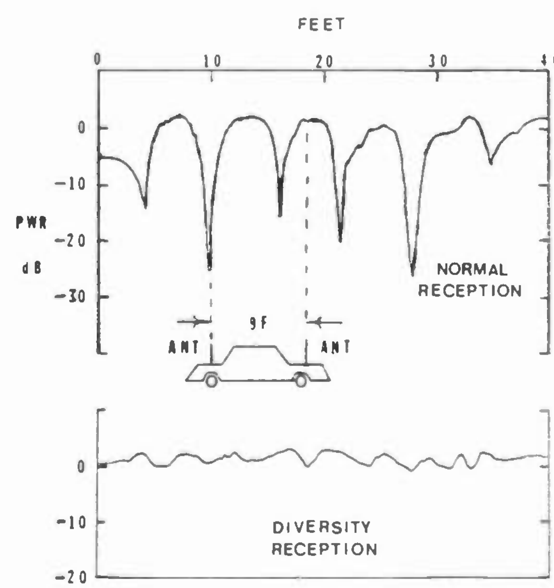

Figure 10

NORMAL vs. DIVERSITY AUTO RECEPTION

SYNCHRONOUS AMPLITUDE MODULATION

Several technical papers and articles have been written on the subject of Synchronous Amplitude Modulation noise in FM broadcast transmitter systems [5]. Several authors have described the causes and effect of Synchronous AM noise. This section summarizes SAM as it relates to transmitter aggravated multipath.

The FM broadcast engineer should be concerned with two types of AM modulation distortion, asynchronous and synchronous. Asynchronous AM is the distortion that is caused by poor power supply filtering or rectification and results in AM modulation that is a harmonic of the 60 Hz power line frequency. Asynchronous AM distortion is normally measured without modulation. This distortion is well understood by broadcast engineers and has been an FCC required measurement.

Transmitter SAM, on the other hand, is the amplitude modulation caused by the frequency swing of a limited bandwidth transmitter and is synchronous with the transmitter's modulation (Figure 11). Standards for this distortion have never been a part of the FCC transmitter rules. SAM is related to the tuning and bandwidth of the transmitter system. SAM modulation affects the stereo signal by increasing the distortion during a multipath event. The effect is to cause an additional chopping sound. Reducing SAM will in most cases reduce the loudness of the multipath distortion.

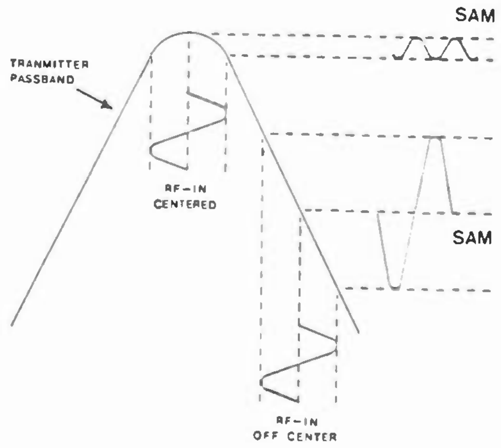

Figure 11

SYNCHRONOUS AMPLITUDE MODULATION

Broadcaster field experience has demonstrated that transmitters with SAM of -40 dB or more below 100% AM modulation are acceptable. Tests reported by engineers experienced in this field have shown that improvements in excess of -40 dB cannot be heard by the listeners. Calculations have shown that in order to achieve -40 dB, the transmitter amplifier's bandwidth will have to be at least 730 kHz. The table shows the approximate bandwidth required for varying SAM levels.

APPROXIMATE SYSTEM BANDWIDTH AS RELATED TO SYNCHRONOUS AM

SYNCHRONOUS AM	APPROXIMATE BANDWIDTH
-30 dB	410 kHz
-35 dB	550 kHz
-40 dB	730 kHz
-45 dB	1.0 MHz
-50 dB	1.4 MHz

The SAM stability of some transmitters may leave a lot to be desired. It has been found that changes in power line voltage, room temperature, screen voltage, and corrections by the automatic power control system may affect the SAM stability of the system. To determine the SAM stability of the transmitter system, continuous monitoring may be necessary. SAM monitors capable of measuring the noise during programing are now commercially available.

In many cases source aggravated multipath is caused by the limited frequency response of the transmitter. Components beyond the transmitter, filters, combiners, transmission lines, and antennas may also cause excessive SAM [6]. Filters used in community antenna combining systems should be designed to have sufficient bandpass not to cause an increase in SAM. Transmission line VSWR should be kept within good engineering practice. The FM engineer should be aware that narrow bandwidth antennas can also increase SAM.

The development of new devices like the super power isolators described in a paper at the 1989 NAB Convention will greatly reduce the effects of VSWR by absorbing the reflections from the antenna system and transmission line [7]. This device can also be used for the construction of high power combiners that will reduce the need for the use of frequency response limiting filters. The development of the super power isolators and the use of wideband antenna systems will give the broadcaster the tools to reduce significantly SAM caused by the RF system.

Broadcast engineers now have the tools and information to control source aggravated multipath. In many cases this will reduce or eliminate the multipath event. In most cases the solutions are not costly for the benefits received.

CONCLUSION

FM broadcasters can be assured of a bright future for the industry if they aggressively work to implement the technologies that will improve the quality of the transmission system. A lesson can be learned from AM broadcasting, not to wait until the system is in trouble to implement quality improvements in the service. With the present success of the FM system, the broadcaster has a strong influence over the receiver manufacturing industry.

* FMX is a trademark of Broadcast Technology Partners.

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AN ALL DIGITAL APPROACH TO FM LIMITING AND COMPOSITE BASEBAND GENERATION

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DSP (digital signal processing) technology was developed over 20 years ago and has resulted in the introduction of a number of practical products for science and major industries. It is only recently that this technology has become economical enough to allow for its implementation in devices designed for the radio broadcaster. In particular, this technology makes it possible to carry out sophisticated processing of digital audio signals which heretofore was either impossible or difficult to perform in the analog domain.

INTRODUCTION

Recent product introductions from the major manufacturers of DSP hardware have allowed for an economical implementation of an all digital FM limiter and stereo generator. This paper discusses the advantages of such a device relative to existing analog systems and establishes the important design considerations for such a digital device.

INITIAL CONSIDERATIONS

The conservative broadcast engineer might ask the question: "Why do we need digital signal processing for the analog medium of FM broadcasting?"

There are limitations in fidelity in the FM medium due to the analog nature of the broadcast medium. The entire potential audience for a radio station does not live within the 70 dBu contour. Not every listener is located in an area free of multipath reflections. Most of the FM receivers in the field have only adequate specifications for noise and distortion. The medium is further compromised by the multiplexing method employed to broadcast stereophonic and subsidiary programming. Additionally, the introduction of high quality analog audio processing in the late 1970's and its continued development have given us 95% of all we could ever hope for in the compromises we must make between modulation control, loudness, and fidelity.

The broadcasting industry can benefit from another entry in audio processing technology for the reasons described below.

An audio processor that utilizes a 100% numerical processing method through a hardware and software implementation offers the following important operational advantages over analog devices:

a) Stability

Microprocessors make use of thousands of transistors operating in a non-linear mode. Though currents that flow in these transistors may vary with time and temperature, this variance does not effect the devices'ability to operate as a switch representing two binary conditions. Software parameters stored in non-volatile memory are not subject to aging or temperature effects. As a result of these factors, the critical operating parameters of an audio processor implemented with DSP technology will not drift with time or changes in temperature. The parameters for the amount of limiting, pilot injection, or absolute modulation limit will be exactly the same in two years as they were on the day these parameters were set. These parameters for all analog counterparts will drift with time and temperature. This is particularly true in those devices in which limiting thresholds are derived from the junction voltages of silicon devices.

b) More Accurate Control of Modulation

The use of special algorithms and digital memory allows us to store several consecutive samples of audio, evaluate them as a whole and then make decisions on what will be done to improve the density or conform to a modulation limit. Careful algorithm design can greatly improve limiter distortion performance under all limiting conditions. These functions cannot be efficiently implemented employing analog technology.

c) Product Flexibility through Software

A digital approach to audio processing could provide the customer with increased variety and flexibility in his processing applications. For example, crossover frequencies in multi-band sections could be programmed by the user from the front panel. This would allow the user to easily tailor the processor to a wide variety of formats. Classical stations could program a system to preserve dynamic range and only protect the modulation limits; rock formats could adjust for desired loudness and density requirements. Analog counterparts might require the user to remove a circuit board and change precision capacitors and resistors to produce the same result.

Since all processing functions are derived from software in the digital system, future improvement in algorithms or product features can be easily delivered to customers in programmable read-only memory (PROM) chips. This protects customers from product obsolescence as technology advances.

d) Repeatability

Digital implementations in audio processors allow for absolute repeatability of their adjustments. A broadcast group owner can determine the processing parameters that yield a competitive edge and easily duplicate them at all of his stations. This eliminates the "setting the pot a little bit past the five or until the limiter grunge goes away" syndrome.

Precise modulation control is achieved through software in the digital system, not from hand selected high tolerance components potted in black epoxy modules. This eliminates audible differences between two identically adjusted analog units due to out of tolerance components.

e) More Potential for Creative Solutions Through Software

A manufacturer of an all digital audio processing platform could license third parties to author software algorithms for a digital device. This would allow more people to work together on the industry's needs and apply their knowledge to reducing the distortions generated by the amplitude limiting process, the control of modulation overshoot, and so on.

f) Higher Immunity to Noise and RFI

A digital audio processor uses non-linear components and is not subject to induced noise and other distortions when operated in high RF field environments. Analog processors' linear components tend to be subject to induced noise and distortion in the RF fields.

g) Reliability and Factory Support

The digital implementation of an audio processor is free from problems resulting from small drifts in component values and does not require factory tweaking on a regular basis. In addition, unique customer requirements can be handled with custom software. Changes in analog equipment often involve cutting traces, changing resistor values, or adding modified circuit boards in order to effectively meet custom requirements.

h) Cost

As of the date of this paper a digital implementation would be fractionally more expensive than its analog counterpart; however, the cost of digital signal processing components is falling rapidly. Some of the additional cost at the present can be offset in the manufacturing process. All digital architectures can to a large degree test and verify themselves through software routines that are run during the manufacturing process. This assures a higher quality, more thoroughly tested product at a lower manufacturing cost than an analog version. Unlike their current analog counterparts, digital units do not require the hand selection of high tolerance components or the adjustments of trimmer potentiometers in order for the product to conform to its design specifications.

A digital audio processor's better performance and greater flexibility can justify its slightly higher cost. Additionally, the product's stability eliminates monthly tweaks and occasional returns to the factory for re-calibration, thereby lowering the user's operational and maintenance costs.

PRACTICAL IMPLEMENTATION

This section will examine some of the technical criteria that must be established and met in a digital audio processor.

Sampling Rates

If we are to generate the complete FM composite baseband, including SCA's, it is necessary to digitally represent signals up 100 kHz in frequency. Applying the Nyquist sampling rule to the problem we can see that we have a theoretical minimum sampling frequency of 200 kHz. This sampling rate is not appropriate, however, due to the fact that we would need a perfect output smoothing filter that would apply 4 dB of boost at 100 kHz in order to correct for (sin x)/x attenuation and then yield 80 dB or better of attenuation at 100.001 kHz. Therefore we require an output sampling rate that is somewhat higher than 200 kHz.

It should also be noted that the digital implementation will be made easier if the output sampling rate is a binary multiple of either of the two steady state vectors generated in the composite baseband: the pilot and the unmodulated L-R subcarrier. In the proposed architecture presented here an output sampling rate of 304 kHz is proposed. This rate satisfies our requirements in that 304 kHz is 8X the subcarrier frequency (16X pilot). In addition, lower sideband alias frequencies of this rate will extend down to 204 kHz (sample rate minus highest baseband frequency). This allows a more reasonable smoothing filter that only requires 1.6 dB of (sin x)/x correction at 100 kHz and then has a transition band of slightly greater than one octave to reach the desired attenuation.

The input sampling rate is also an important consideration. After careful consideration and study of the problem the reader would reason that to digitally generate the FM stereo multiplex properly the audio sampling rate would have to be 38 kHz (or a integer multiple of 38 kHz). Due to the physical limitations of the current multiplexing standard, a higher sampling rate than 38 kHz would not yield any additional improvement in fidelity.

This 38 kHz sampling rate might seem strange to those with a background in digital audio. The most popular sampling rates are 44.1 kHz for Compact Discs and 48.0 kHz in professional digital recording equipment. The proposed 38 kHz sampling rate will not pose a significant problem, since almost all broadcast facilities have analog mixing consoles in their on air positions. This analog output can be digitized at a 38 kHz rate prior to application to the proposed processor architecture.

In the event that broadcast studios convert to an entirely digital format, a DSP-based sampling rate converter can be placed just ahead of the digital limiter/stereo generator for the purpose of decimating the sampling rate from either 44.1 kHz or 48.0 kHz to the FM broadcast rate of 38 kHz. This decimation process can be achieved entirely in the digital domain without any degradation in audio quality, with the exception of audio bandwidth which is in any case restricted by the stereo generation process.

Quantization Precision

The most probable digital source material to be used in broadcast during the years to come is the Compact Disc. Since the Compact Disc is limited to 16 bit quantized samples it would be a natural assumption that 16 bit quantization of analog audio would be adequate. The 16 bit analog to digital conversion devices that are currently available do not perform to the theoretical limits of 16 bits in linearity and noise; 18 bit (or greater) devices should be employed to insure theoretical performance to at least 16 bits. The architecture presented in this document is designed around 18 bit devices but allows for future expansion to 20 bits in the serial data path formats.

Processing Precision

FM limiting and stereo generation requires the use of several filter functions. When these filters are implemented digitally, the mathematical precision of the processing architecture must be considered. Sufficient precision is required so that the desired filter functions can be realized from their static numerical coefficients. Many of these filters will be implemented as iterative convolutions, and insufficient digital word widths can produce significant rounding and other mathematical errors that contribute noise and sampled artifacts to the processed signal. The architecture presented utilizes 24 bit data paths and up to 56 bit accumulator precision. This design guarantees that all digitally generated artifacts and noise will be several dB below the theoretical limits of the input and output data precision.

Audible Digital Delay

Audible delay is an important concern in digital processors. Announcers who listen to off air signals could be subjected to problems if their monitor audio is significantly delayed. Research in this area reveals that 10 milliseconds of delay becomes noticeable to the average person. Careful filter and processing algorithm design maintains processor input to output delay time at under 5 milliseconds in the limiting and stereo generation processes.

Hardware versus Software Implementation

The digital device previously described could be implemented entirely in hardware as a dedicated algorithmic state machine. If the manufacturer could bear the non-recurring engineering costs associated with custom integrated circuit design the design could be implemented employing several custom ASIC's (application specific integrated circuits). The overall cost of goods in the product would be low, but the up front engineering costs would be very high. The product would also be somewhat restricted in its flexibility. All filter and processing algorithms would have to be determined and computer modeled in advance and then committed to custom silicon chips.

A better approach is to employ readily available digital signal processing components in a more general purpose arrayed architecture and construct the majority of the signal processing implementation in software. This approach has a very high degree of flexibility allowing complete signal flow paths and processing blocks to be moved, changed, and manipulated at will by the software architect. Although this method costs more in terms of component cost, the up front design costs (which also have to be recovered from unit sales) are much less. This method would provide the ultimate in long term growth of the design through software and protect the customer to a higher degree from product obsolescence.

HARDWARE DESCRIPTION

Optical Encoder

Figure 1 shows the block diagram of the Optical Encoder.

The purpose of the Optical Encoder is to accept balanced stereophonic audio and convert it to individual 18 bit left and right digital samples at a 38 kHz rate. The single integrated analog to digital converter IC is of the oversampling/decimation type. As is typical in this type of converter, the device contains linear phase digital anti-alias filters. These digital filters exhibit a near perfect phase response, perfect channel to channel frequency response matching, and they suppress alias components greater than 87 dB without any external tolerance sensitive components.

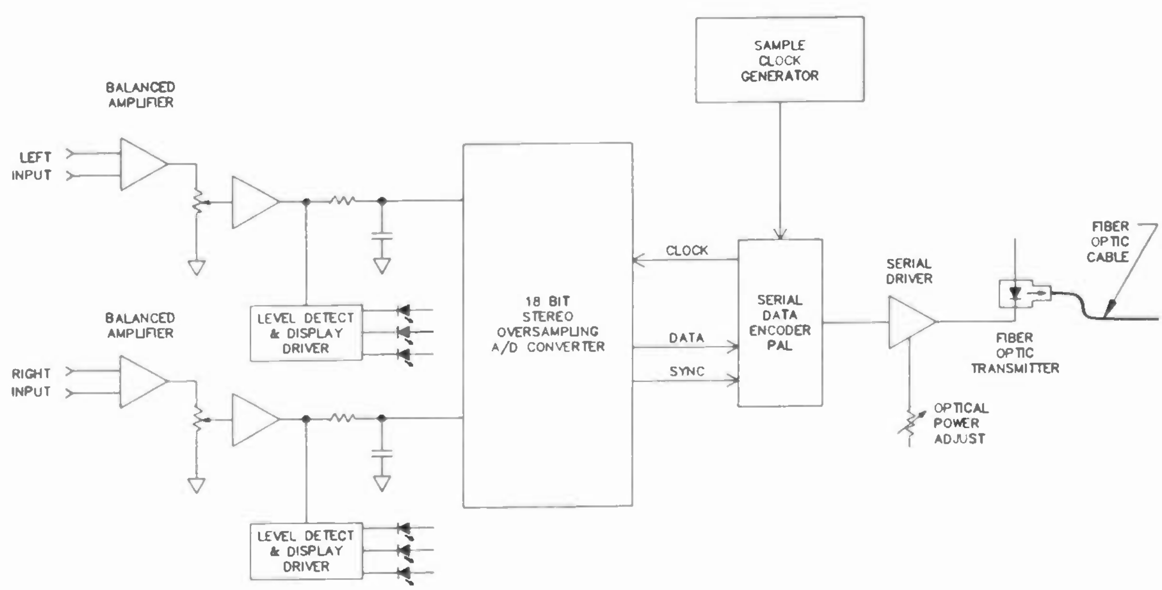
Digital data from the converter IC is then combined with control and synchronizing data bits into a serial data stream in a single PAL (programmable arrayed logic) device. The serial data is then fed to a fiber optic transmitter. Digital audio is conveyed to the limiter/stereo generator unit via fiber-optic cable. This allows the highest immunity to RFI and eliminates data errors and analog input noise due to ground return current loops. A series of analog level detectors and display components allow the user to set the audio input levels for the optimum headroom and clipping levels.

FM Limiter and Stereo Generator

Figure 2 presents the block diagram for the FM Limiter and Stereo Generator.

Digital audio to be processed arrives at the fiber-optic receiver where it is amplified, sliced, and fed to a serial data decoder circuit. The serial decoder disassembles the incoming data stream and directs left and right audio samples to the DSP array processor. The serial data decoder also provides digital error signals to phase lock the sample clock generator to the incoming sampling rate.

The DSP array processor was designed for maximum software flexibility and data processing throughput. This array consists of four individual general purpose DSP's, a common block of 24 bit random access memory, and bus arbitration logic. A block of eight dedicated Finite Impulse Response (FIR) filter DSP's function as peripheral co-processors to the DSP array.



OPTICAL ENCODER ™
FIGURE 1

Digitally processed, limited, and pre-emphasized audio is fed to a digital oversampling filter to raise the sampling rate of the left and right channels to the 304 kHz sampling rate required for stereo generation. The left and right oversampled audio along with SCA data is passed to a final dedicated DSP where the composite baseband is assembled and corrected for any addition overshoots. This final DSP also monitors and counts modulation data that exceeds user specified thresholds and supplies the peak information to the managing microcontroller.

The final DSP sends the completed FM baseband signal to a 24 bit digital output connector in anticipation of the development of a numerically modulated FM exciter. This baseband output data is also converted in an 18 bit high speed digital to analog converter to an analog composite signal for application to an existing analog FM exciter or composite STL transmitter.

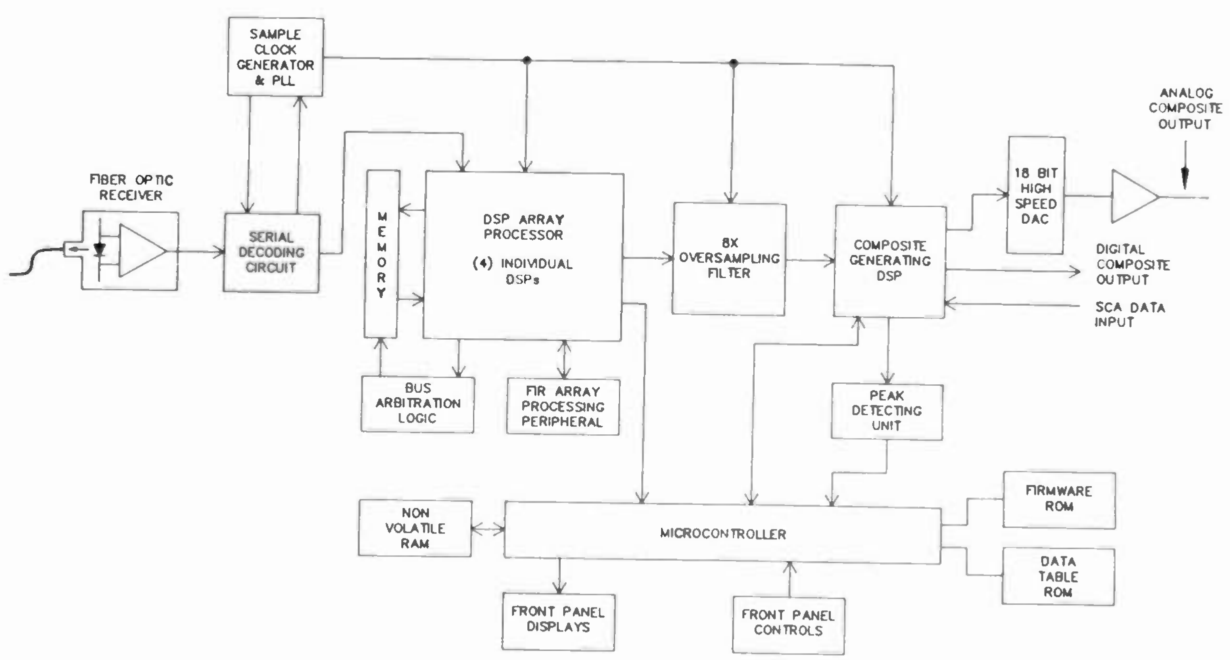
A general purpose microcontroller provides general management functions, front panel display, and user interface. The microcontroller utilizes a non-volatile random access memory for storage of user selected parameters, allowing for a graceful return from possible power failure. Microcontroller firmware, DSP machine code, and filter coefficient data tables are stored in two associated PROM's that are socketed for easy user access for future software upgrades.

PROCESSING SIGNAL FLOW

In a software oriented application the signal flow diagram can be radically different from the hardware block diagram. In this section we will differentiate the signal processing flow from the previously described hardware diagram.

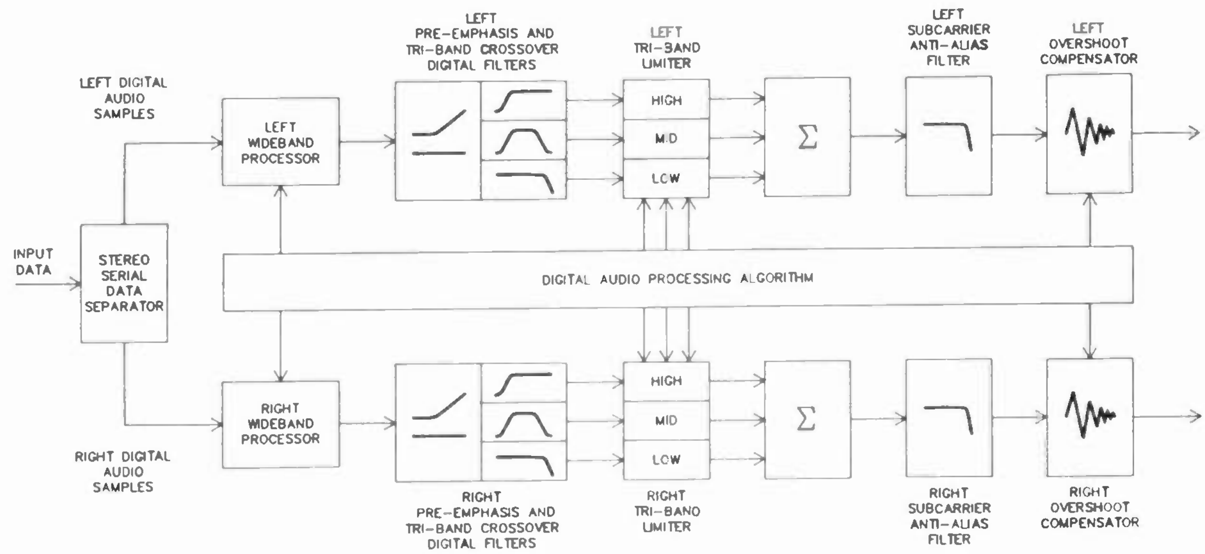
Figure 3 presents the signal flow diagram. The incoming digital audio stream is separated into left and right channel data and applied to the wideband processing functions. This software implementation yields the equivalent of a slow moving automatic gain control. Gain is adjusted by the manipulation of a fractional multiplication constant. Audio from the wideband processors is then pre-emphasized and separated into the three limiting bands by a series of carefully designed, time coherent, Finite Impulse Response filters. This design was optimized for minimum delay by mathematically combining different filters together into a series of single filter functions.

Audio from the crossover filters is applied to the tri-band limiter stage. The algorithm that drives the limiters adjusts a single band's gain coefficient only after considering the coefficients of the adjacent band or bands.



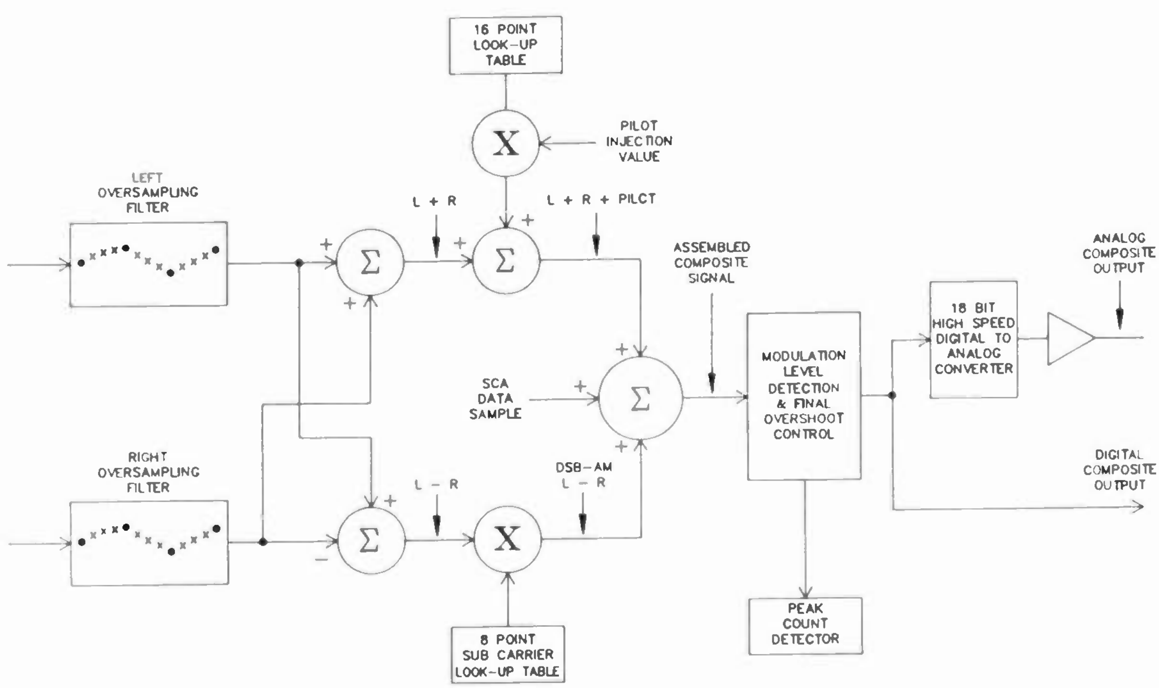
LAZER™ FM LIMITER/STEREO GENERATOR

FIGURE 2



LAZER ™ SOFTWARE SIGNAL FLOW DIAGRAM

FIGURE 3



LAZER ™ SOFTWARE SIGNAL FLOW DIAGRAM

FIGURE 3 (CONTINUED)

Digital audio from the limiter section is summed back to a single wideband source. Amplitude data from the summing block output is supplied to the limiting algorithm to provide for an additional safety margin in the limiting process, allowing all three limiters to be adjusted in common should the arithmetic sum of the three bands exceed the specified limits.

The wideband audio is then lowpass filtered to protect the pilot and the L + R bands from L-R alias. This lowpass filter was carefully designed with the connecting overshoot compensation algorithm to provide a coherent system of deriving maximum filter benefits while minimizing the effects of filter amplitude overshoots. It is very doubtful that any analog implementation of this stage, no matter complex, could derive the benefits attained in this software implementation.

The processed, pre-emphasized audio is then applied to an 8X digital oversampling filter to raise the sampling rate to 304 kHz. This oversampling is accomplished by employing an FIR lowpass filter that performs its convolution at 8X its input sample rate. The filter very accurately produces 8 interpolated output samples for every single input sample supplied.

Left and right audio signals are then numerically matrixed using simple addition into L + R and L-R channels. The pilot vector is generated from a 16 point, 24 bit look up table. This vector is then multiplied by a user selectable pilot injection coefficient and added to the L + R channel.

The 38 kHz subcarrier vector is derived from an 8 point, 24 bit look up table and multiplied by the L-R channel to produce the 38 kHz, double sideband, suppressed carrier subchannel. The subcarrier suppression is perfect because of its inherently balanced multiplier. The pilot and 38 kHz subcarrier look up tables are locked together via control software so that their phasing is perfect at all times.

A final summing block adds L + R, pilot, L-R subchannel, and the externally supplied SCA sample together to form the complete assembled composite baseband spectrum. A final stage compares instantaneous composite amplitude against a user specified threshold to derive modulation peak data for display on the front panel. This stage also provides final overshoot control of the signal.

The output data is split into two paths. One 24 bit path plus word clock is supplied to the rear panel as a digital composite output. The other path is truncated to 18 bits and applied to a high speed digital to analog converter for derivation of a traditional analog composite output to drive existing exciters and composite STL transmitters.

CONCLUSION

An all digital approach to FM limiting and stereo generation is within the grasp of current signal processing technology. The application of this technology can yield many cost effective and audibly competitive advantages for the broadcaster. Proper design of these devices includes a software driven architecture which yields a good amount of operational flexibility and product longevity. The full embrace of these technologies can become a driving force for further digitization of broadcast audio and transmission plants.

A NEW APPROACH TO FM COMPOSITE BASEBAND OVERSHOOT CONTROL

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INTRODUCTION

It is well known, that certain program material, processed aggressively, causes certain composite STL systems and FM exciters to produce overshoot in the FM composite baseband signal. This overshoot is produced even in systems that are properly band-limited and which employ STL paths that have no multi-path. Accordingly, loudness is compromised because modulation levels must be reduced to prevent illegal over-modulation.

Previous attempts to eliminate this overshoot have compromised other system performance aspects. We have developed a new approach to minimizing this overshoot without any other system compromises.

THE FM STEREO SYSTEM

The world-standard FM stereo "pilot-tone" system encodes the sum of the channels (L+R) in the frequency range of 30-15,000Hz in the stereo baseband -- the "stereo main channel." It encodes the difference between the channels (L-R) on a double-sideband suppressed-carrier sub-channel centered at 38kHz, and occupying 23kHz to 53kHz in the stereo baseband -- the "stereo sub-channel." A pilot tone at 19kHz tells the receiver that a stereophonic transmission is being received, and provides a phase and frequency reference to permit the receiver to regenerate the 38kHz sub-carrier to use in its stereo de-modulator.

Any energy that appears in the frequency range from 30 to 19,000Hz caused by a signal in the stereo sub-channel is termed "sub-channel-to-main channel crosstalk." Any energy that appears in the frequency range from 19 to 57kHz caused by a signal in the stereo main channel is termed "main channel-to-sub-channel crosstalk."

When the stereo encoder is driven by a pure right-only or left-only signal, "stereophonic separation" can be measured at the stereo decoder as the ratio between the desired and undesired signal levels, where the "desired" signal is the signal appearing in the decoder output channel corresponding to the channel driven at the encoder, and the "undesired" signal is the signal caused by the desired signal that appears at the remaining output.

Ideally, crosstalk is non-existent and stereophonic separation is infinite. In practice, both linear and non-linear errors cause these characteristics to deteriorate.

In the linear domain, separation and crosstalk are mathematically orthogonal. Phase and frequency errors that cause one to deteriorate will generally not affect the other. For example, phase or frequency errors in the composite signal channel will cause separation to deteriorate, but cannot affect crosstalk, since the stereo main and sub-channels are already separated in frequency and changes in phase or amplitude response in the composite channel cannot affect this frequency separation. Conversely, mismatches between the linear response of the left and right signal paths prior to the stereo encoder will cause crosstalk, but cannot affect separation.

Non-linear errors in the composite channel can cause both separation and crosstalk to deteriorate because such errors cause harmonic and intermodulation distortion that introduce new frequencies into the baseband. These new frequencies are likely to add energy into a part of the baseband spectrum that will be decoded by the stereo decoder in spatial locations different than the locations of the original sound sources. Further, these new frequencies are perceived by the ear not as changes in spatial localization, but as highly offensive distortion.

This is somewhat analogous to "aliasing distortion" in a sample-data audio system. In such a system, any input frequencies greater than one-half of the sampling frequency (the "Nyquist frequency") are encoded with the wrong frequency: they "fold around" the Nyquist frequency and appear at the decoder as frequencies unrelated to the program material that produced them. The ear perceives this "aliasing" as offensive distortion.

PREVIOUS NON-LINEAR SOLUTION

The most common technique for reducing FM composite baseband signal overshoot has been composite baseband clipping. Over the years, composite clipping has been very controversial because it causes signal degradation and because early implementations that clipped the pilot could violate the FCC rules. No implementation prevents dynamic signal degradation.

The composite baseband clipper is a non-linear device. Thus, it generates distortion and aliasing products that contaminate the composite baseband signal, degrading dynamic stereo separation and causing dynamic audible distortion. It also produces distortion products in the sub-carrier region. This limits, or even prevents sub-carrier use, destroying the sub-carrier's market value and reducing revenue potential.

While it is possible to low-pass filter the clipped baseband (thus protecting the sub-carriers), such filtering does nothing to eliminate intermodulation distortion in the stereo baseband region below 57kHz, and will also tend to increase peak modulation, partially negating the peak control produced by the composite clipper.

If the FM exciter, as opposed to the STL, is the source of overshoot, the composite baseband clipper cannot control overshoot; likewise, if the clipper precedes the STL. In fact, it can actually increase overshoot because the clipping process produces increased amounts of sub-sonic inter-modulation distortion products.

Some have argued that in addition to reducing composite overshoot, composite baseband clipping achieves more loudness than audio clipping in the left and right channels. This loudness increase can only be achieved by degrading dynamic stereo separation and crosstalk. To preserve dynamic stereo performance, the spectra of the stereo main channel and sub-channel must be completely separate: the main channel must not have any energy above 19kHz, and the sub-channel must not have any energy below 19kHz. One consequence of this frequency separation is this: in a system that achieves high dynamic separation and low crosstalk, it must be impossible for the system's final filter/limiter to reproduce a square-wave above one-third the cut-off frequency of the low-pass filter prior to stereo encoding -- above approximately 5000Hz, assuming 15kHz low-pass filters. Because they generate spurious harmonic and intermodulation products, composite baseband clippers do not meet this criterion, and thus compromise dynamic stereo performance.

Composite baseband clippers are a very easy method of increasing apparent loudness of the broadcast signal. However, the old adage "cheap and dirty", has never been so true.

There is another interesting aspect of the FM stereo system. It is usually stated that the peak modulation of the composite baseband is the greater of the left or right channel levels, plus the pilot. It turns out that this is only an approximation, because the pilot is correlated in phase with the 38kHz suppressed sub-carrier. This causes the total composite modulation to decrease slightly under channel imbalance and single channel conditions. Assuming 10% pilot injection and holding the left channel at 100% modulation, decreasing the right channel from 100% to 0% modulation will cause the composite modulation to decrease by 2.8%. Thus even perfectly accurate peak limiting in the audio domain, prior to stereo encoding, can only control the composite modulation to an accuracy of -2.8%/+0%. Since composite baseband clipping controls the peak deviation of the composite precisely, it can in fact be somewhat louder than peak limiting in the audio domain. But the "advantage" is an imperceptible 0.25dB at best!

In these digital times, the consumer is becoming increasingly aware of audio quality issues. It is preposterous to think that composite baseband clippers are being used to degrade system performance below that of some of the least expensive receivers!

EXACT INTERLEAVING FAILURE
IN THE FM STEREO SYSTEM

Left channel modulation 100%.
Right channel modulation from 0% to 100%.
10% pilot injection.
Peak deviation of composite shown with 100% normalized to L = R with pilot present.

Right % Modulation	Composite % Modulation
100.0%	100.00%
99.5%	99.79%
99.0%	99.59%
98.0%	99.26%
97.0%	98.99%
96.0%	98.77%
95.0%	98.59%
94.0%	98.45%
93.0%	98.33%
92.0%	98.22%
91.0%	98.14%
90.0%	98.06%
80.0%	97.65%
70.0%	97.48%
60.0%	97.39%
50.0%	97.33%
40.0%	97.29%
30.0%	97.26%
20.0%	97.24%
10.0%	97.22%
0.0%	97.20%

Fig. 1 INTERLEAVING FAILURE IN THE FM STEREO SYSTEM

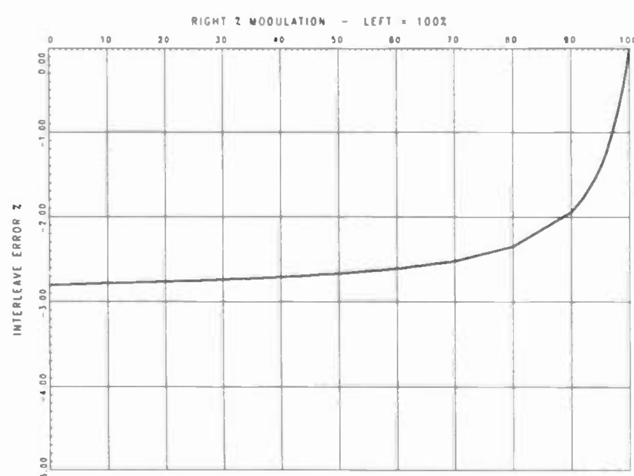

Fig. 2 INTERLEAVING ERROR IN THE FM STEREO SYSTEM

Fig. 1 indicates the exact interleaving failure due to pilot summation in the FM stereo system*. Fig. 2 is a plot of the interleaving error in % of modulation versus 100% left channel modulation and 0% to 100% right channel modulation.

* Robert Orban, private communication.

OVERSHOOT SOURCE

After extensive computer modeling and analysis of several popular composite STL systems and FM exciters, it became clear that the overshoot problem was not in the high frequency domain, as previously assumed, but in the sub-audible, low-frequency domain. Several of the systems modeled have sub-audible peaks in their frequency response, and have insufficient low frequency response to properly reproduce an aggressively processed composite baseband signal. This poor low frequency transient response is the element that causes composite baseband overshoot. It can be caused by incorrectly designed AFC loops and/or insufficient low frequency response of the composite baseband amplifiers. The system response of two popular composite STL systems are shown in Figs. 3 and 4. The system response of two popular FM exciters are shown in Figs. 5 and 6. Note the radical difference in performance, and frequency dependent response.

NEW MONITOR PARAMETER SOLUTION

Recently, a device has been introduced to change the way in which modulation is measured. By delaying the peak indicator, the modulation monitor ignores peaks of less than 1 millisecond duration, assumed to be overshoots, and does not indicate over-modulation under these conditions. Although this concept of modulation measurement is under close investigation by the industry for reasons not relevant here, it does not ignore the composite baseband overshoot described above. In this case, it correctly measures modulation because the duration of this overshoot is far longer than the delay of the of the peak indicator. This means that this overshoot must still be eliminated to achieve maximum loudness with aggressively processed composite baseband signals, because this new monitoring device will still indicate overmodulation induced by this kind of overshoot.

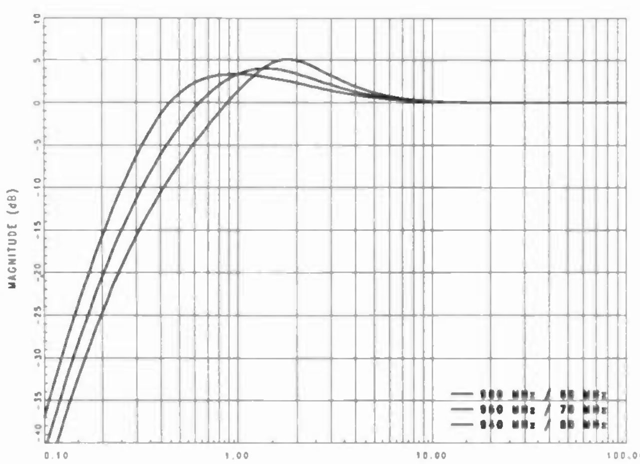

Fig. 3 STL 1 SUB-AUDIBLE FREQUENCY RESPONSE

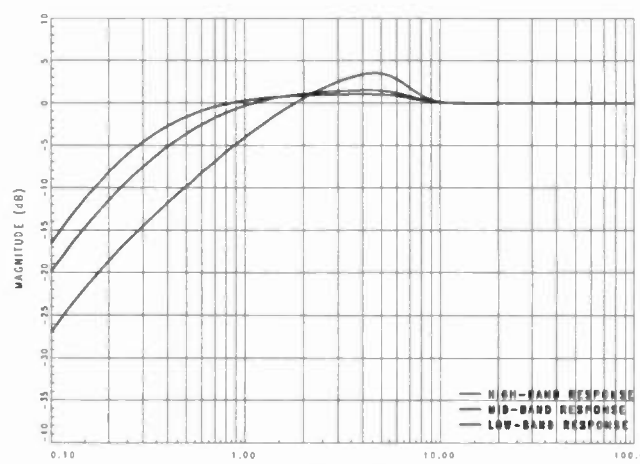

Fig. 5 FM Exciter 1 SUB-AUDIBLE FREQUENCY RESPONSE

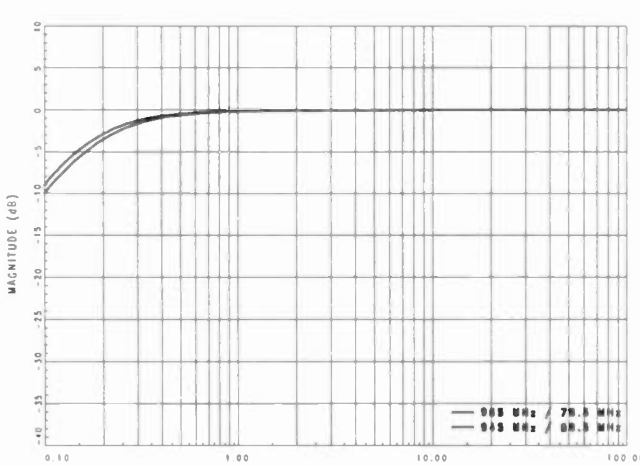

Fig. 4 STL 2 SUB-AUDIBLE FREQUENCY RESPONSE

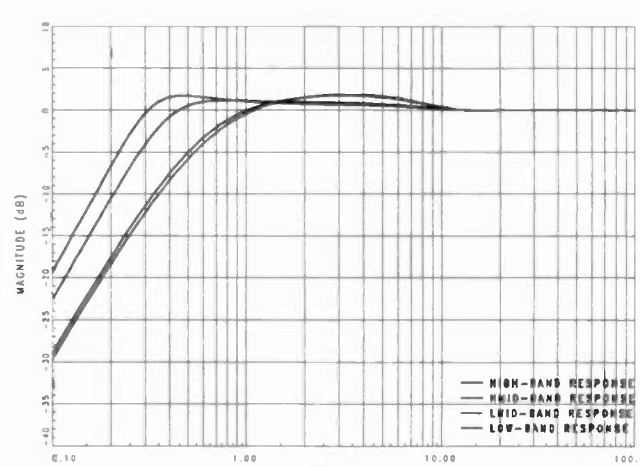

Fig. 6 FM Exciter 2 SUB-AUDIBLE FREQUENCY RESPONSE

RESPONSE REQUIREMENTS

If the system is modeled as a high-pass filter with a single dominant pole, the equation in Fig. 7 can be used to compute the percentage of overshoot when a square-wave of frequency F (Hz) is applied to the system input. This overshoot will occur regardless of whether the low-frequency roll-off is in the composite channel, or in the left and right audio channels after peak limiting. In the former case, the roll-off can also compromise low-frequency separation. To achieve less than 1% overshoot with a 50Hz square-wave (a reasonable criterion for good peak control), the dominant pole must be located at 0.16Hz or lower! If more than one low-frequency roll-off element is cascaded in the composite path, each element's cut-off frequency must be substantially below 0.16Hz. This minimum low-frequency response requirement is shown in Fig. 8.

$$\text{Overshoot \%} = 100\left(1 - e^{\frac{P}{2F}}\right)$$

$$\text{where } P = -\frac{1}{RC}$$

$$\text{where } F = \text{Frequency in Hz}$$

Fig. 7 OVERSHOOT EXPRESSION

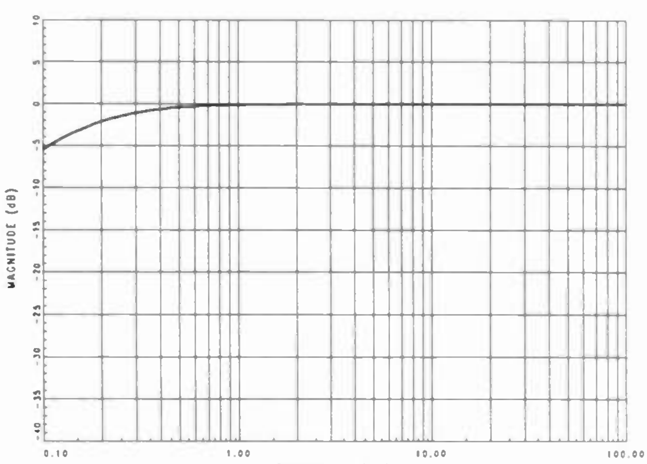

Fig. 8 MINIMUM LOW-FREQUENCY RESPONSE REQUIREMENT

NEW LINEAR SOLUTION

Since the composite baseband overshoot problem is caused by a frequency response discrepancy, it can be corrected to a nearest approximation, with an equalizer. An example of the system response in Fig. 3 is shown equalized in Fig. 9. This is a linear solution to a linear problem. Unlike the composite baseband clipper, NO distortion and aliasing products are produced, thereby making it a far more desirable solution to the overshoot problem of composite baseband signals.

The block diagram of a device available to produce the necessary sub-sonic equalization is shown in Fig. 10. The composite baseband signal passes through accurate high-speed amplifier circuitry, with only the additional equalization signal passing through the high gain stages to maintain maximum stereo separation. Two high-current output amplifiers, capable of driving long, terminated transmission lines, are included with individual level controls to provide two independent composite sources. Composite baseband signal metering, provided by a precision peak detector and a 91% to 110% bar graph display in 1% increments, is switchable between input and output signals for a precise comparison indication of overshoot control.

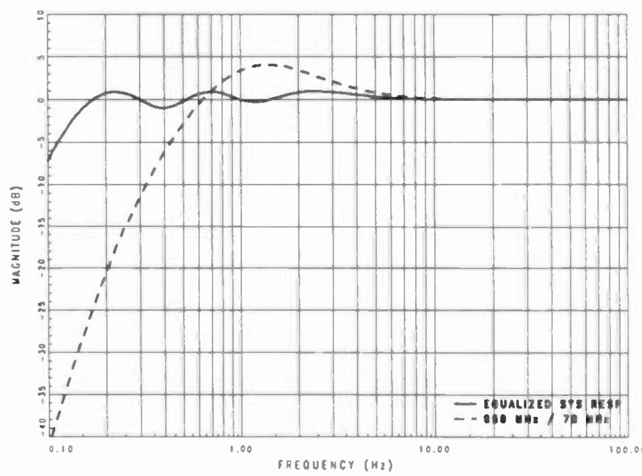

Fig. 9 EQUALIZED STL 1 SYSTEM RESPONSE

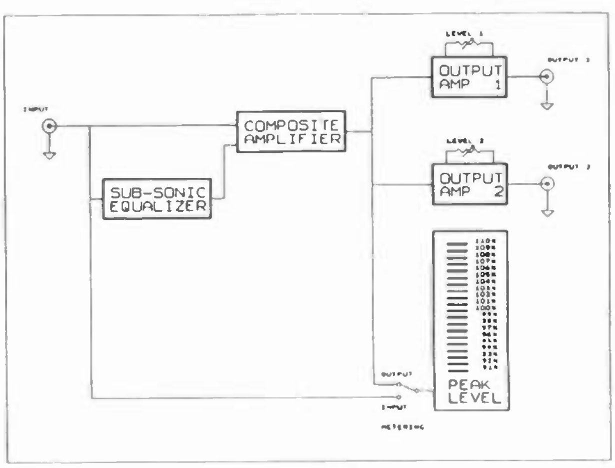

Fig. 10 FM COMPOSITE BASEBAND EQUALIZER BLOCK DIAGRAM

NEW FIRE PROTECTION REQUIREMENTS FOR INDOOR INSTALLATION OF COAXIAL CABLE AND WAVEGUIDE

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Abstract

Beginning in 1986, tighter requirements for fire resistance and fire gas toxicity for indoor installation of coaxial cable and elliptical waveguide have been adopted by many regulatory bodies. Six fire resistance categories of broadcast type coaxial cable have been defined by the National Electrical Code® (NEC®), and all cable and waveguide is regulated by New York State. Also, new editions of model building codes prohibit cable from being installed in ducts and plenums unless it meets the highest standards of fire resistance and smoke emission. This paper explains the relevant categories of cable and their applications, presents several design considerations, and describes a method of assuring compliance with all regulations.

Definitions

Before the regulations can be clearly understood, some terms used in building construction must first be defined:

(a) Conduit - A tube or duct for enclosing electrical wires and cable. Conduit may be metallic or nonmetallic.

(b) Duct - A closed channel, tube, or pipe used to transport air, dust, vapors, etc. Electrical ducts include metallic and nonmetallic conduit.

(c) Plenum - A compartment or chamber to which one or more air ducts are connected and which forms part of the air distribution system of a building (1990 NEC Article 100, p.70-13).

(d) Raceway - An enclosed channel designed expressly for holding wires, cables, or busbars, with additional functions as permitted in the National Electrical Code (1990 NEC Article 100, p.70-13). Raceways may be of metal or insulating material, and the term includes rigid metallic and nonmetallic conduit, intermediate metal conduit, liquid-tight flexible conduit, flexible metallic tubing and conduit, electrical metallic and nonmetallic tubing, underfloor raceways, cellular concrete and metal floor raceways, surface raceways, wireways, and busways.

(e) Riser - A vertical shaft passing from floor to floor. Risers may or may not be fireproof or have firestops at each floor.

Fire-Retardance Requirements

In the United States, fire retardance requirements for coaxial cable used within buildings are set by the National Electrical Code. Since the NEC forms the basis for most local electrical codes in the U.S., in whole or in part, its requirements often have the force of law. In addition, model building codes now have requirements for cable installed in ducts and plenums, and requirements may be set by local authorities independently of the NEC, such as the Fire Gas Toxicity requirements of the Office of Fire Prevention and Control of the New York State Department of State. As a further complication, many states and cities have added amendments to the model electrical and building codes.

In the past, Underwriters Laboratories Inc.® (UL®) also set fire retardancy standards for coaxial cables under the Wires, Miscellaneous category. Coaxial Cable was in the RF Coaxial Cable category. While these listings are still in effect, they have been largely superseded by listings required for the NEC.

National Electrical Code Requirements for Coaxial Cable

In the National Electrical Code, coaxial cable for broadcast systems falls under the Community Antenna Television Systems (CATV) category. The 1987 edition of the NEC established strict requirements for cables used inside buildings, effective 1 July 1988. The requirements were placed in a new section, 820-15, Fire Resistance of Coaxial Cables, and stated that coaxial cables installed as wiring within buildings, in a vertical run in a shaft, or in ducts or plenums or other spaces used for environmental air shall be appropriately listed as being resistant to the spread of fire (and, in the case of plenum installation, having low-smoke-producing characteristics).

These requirements were modified somewhat in the 1990 edition of the NEC. Section 820-15 was replaced by a new section, 820-50. The requirements for coaxial cables have been broadened to include being (1) listed as being suitable for the purpose, (2) listed as being resistant to the spread of fire, (3) installed in accordance with the requirements, and (4) marked in accordance with the requirements. The 1987 category of CATV drop cable was discontinued and several other categories were modified.

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While the 1987 version of the NEC is still in effect in California and probably other jurisdictions as well, the 1990 NEC supersedes it and is expected to be adopted everywhere eventually. Consequently, the requirements of the 1990 NEC are used throughout the remainder of this paper, with differences from the 1987 NEC noted where applicable.

Coaxial Cable Applications Defined by the NEC

NEC Section 820-50 provides the requirement for coaxial cables installed as wiring within buildings, for all applications except where the length of cable within the building does not exceed 50 feet (15.2 m) and the cable enters the building from the outside and is terminated at a grounding block. (In the 1987 NEC, the limit for unlisted coaxial cable was 10 feet (3.05 m), in non-concealed spaces with the cable exposed. There was no separate limitation for drop cable. Consequently, the 50 foot limit provides a relaxation of the 1987 NEC.)

The requirements of Section 820-50 state that coaxial cables used within buildings shall be listed as being suitable for the purpose, listed as being resistant to the spread of fire in accordance with the requirements of Section 820-51, installed in accordance with Section 820-52, and marked in accordance with Table 820-50.

Section 820-51 defines four categories of listed coaxial cable, given in descending order of fire-resistance rating: Type CATVP, CATV plenum cable; Type CATVR, CATV riser cable; Type CATV (General Purpose) CATV cable; and Type CATVX, CATV cable, limited use. The requirements for these cables are provided in paragraphs (a) through (d).

Wiring In Ducts, Plenums and Other Air-Handling Spaces

Type CATVP cable, Section 820-51 (a), is required to be listed as being suitable for use in ducts, plenums, and other spaces used for environmental air, such as above a false ceiling. In addition, it is required to be listed as having adequate fire-resistant and low-smoke-producing characteristics. (The requirements for Type CATVP cable are found in Section 820-15 (c) of the 1987 NEC.)

To achieve listing as Type CATVP, cables must pass the most stringent of all fire-retardance tests: Test for Fire and Smoke Characteristics of Wires and Cables, NFPA 262-1985. This test is identical to UL 910, Standard for Test Method for Fire and Smoke Characteristics of Electrical and Optical-Fiber Cables Used in Air-Handling Spaces. In addition, U.M.C. (Uniform Mechanical Code) Standard No. 10-3 is also based on UL 910. It uses the 25-ft Steiner Tunnel test furnace specified in ASTM E 84, Surface Burning Characteristics of Building Materials. The cables are supported by a cable rack a foot wide. A single layer of cable cut into 24-ft lengths fills the rack (Figure 1).

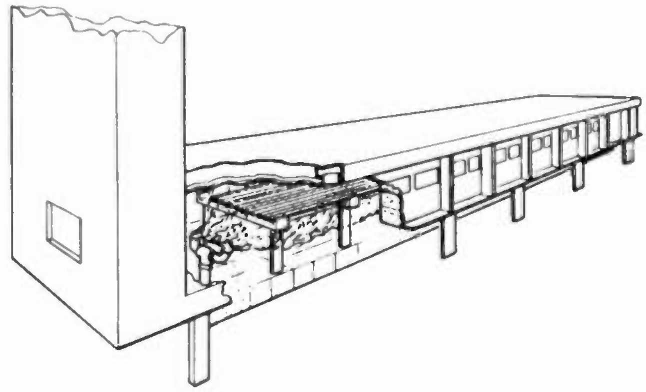

Figure 1. NFPA 262-1985 Test (Same as UL 910)

The 20-minute test begins with ignition of the cable lengths by an 88-kW (300,000 Btu/hr) flame 4-1/2 feet long. Flame spread is aided by a 240-ft/min draft. During the test, the progress of combustion is observed through small glass windows spaced 1 ft apart, and smoke is measured by a photocell installed in the exhaust duct. This is a very severe test because of the large ignition source and the tunnel effect, which contains the heat of combustion. To pass, cables must have (1) a flame spread of no more than 5 ft (1.52 m) beyond the end of the ignition flame; and (2) a maximum peak optical density of 0.5, or one-third light transmission, and a maximum average optical density of 0.15, or 70% light transmission.

Wiring In Vertical Runs (Risers)

Type CATVR cable, Section 820-51 (b), is required to be listed as being suitable for use in a vertical run in a shaft or from floor to floor. In addition, it must be listed as having fire-resistant characteristics capable of preventing the carrying of fire from floor to floor. (The requirements for Type CATVR cable are found in Section 820-15 (b) of the 1987 NEC. They applied only to cables "in a vertical run in a shaft." Since open shafts are no longer used in modern construction, the requirement was broadened in the 1990 NEC to treat any floor penetration connecting more than one floor as a riser.)

In order to be listed as Type CATVR, cables must pass the Standard Test for Flame Propagation Height of Electrical and Optical-Fiber Cables Installed Vertically in Shafts, ANSI/UL 1666-1986. This test is conducted in a facility that simulates a cable installation in a building's riser shaft (Figure 2). It uses a 150-kW (512,000 Btu/hr) burner to ignite the cables. The test duration is 30 minutes. To pass, cables must not propagate flame to the top of the 12-ft high compartment.

General Purpose Wiring Within Buildings

Type CATV cable, Section 820-51 (c), is required to be listed as being suitable for general purpose CATV use within buildings, with the exception of risers and plenums.

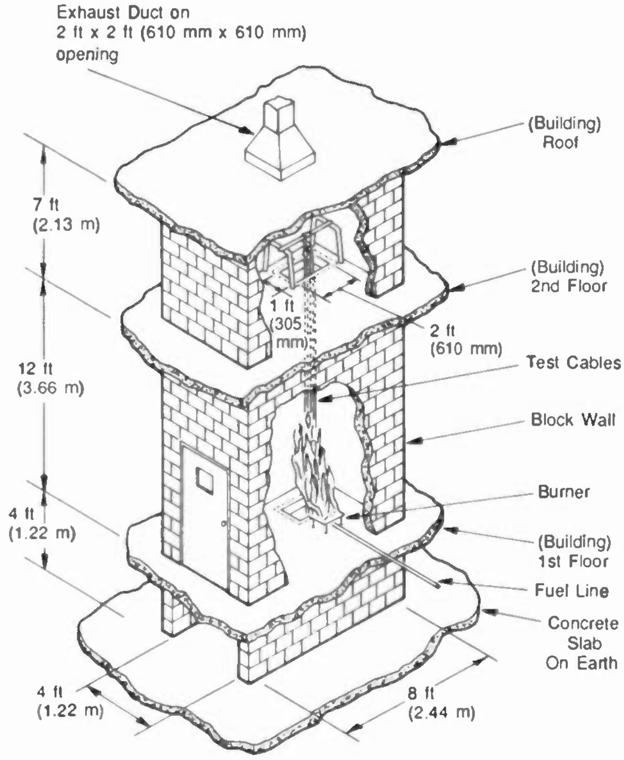

Figure 2. ANSI/UL 1666-1986 Test

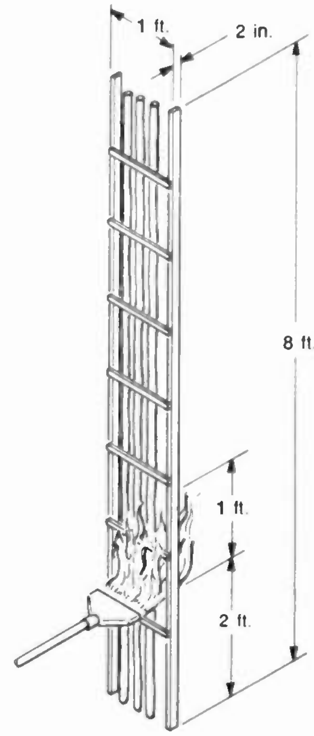

Figure 3. ANSI/UL 1581 Vertical Tray Flame Test

In addition, it must be listed as being resistant to the spread of fire. (The requirements for Type CATV cable are contained in Section 820-15 (a) of the 1987 NEC.)

To receive listing as Type CATV, cables must pass either the "Vertical-Tray Flame Test" in the Reference Standard for Electrical Wires, Cables, and Flexible Cords, ANSI/UL 1581-1985, or the "Vertical Flame Test -- Cables in Cable trays" of Test Methods for Electrical Wires and Cables, CSA C22.2 No. 0.3-M-1985.

In the ANSI/UL 1581 Vertical Tray Flame Test, the cables under test are supported by a vertical cable rack 12 in wide and 8 ft (2.4 m) long. The center 6 in of the rack is filled with 8-ft lengths of cable spaced half a cable diameter apart. A 21-kW (70,000 Btu/hr) flame from a 10-in ribbon burner is used to ignite the cables, as shown in Figure 3. The burner is positioned two feet above the floor, and the flame contacts 9 to 12 in of cable. It is kept on for 20 minutes, and the cables are allowed to continue burning after it is turned off. Cables which are blistered or charred at the top of the 8 ft rack fail the test.

The CSA C22.2 No. 0.3-M-1985 test is somewhat similar, but the requirement for being resistant to the spread of fire is for the damage (char length) not to exceed 4 feet 11 inches (1.5 m) at the conclusion of the test.

Limited Use Wiring Within Buildings

Type CATVX cable, Section 820-51 (d), is required to be listed as being suitable for use in dwellings and in raceway. Also, it is required to be listed as being flame-retardant. (The requirements for Type CATVX cable are contained in Section 820-15 (a), Exception No. 3, of the 1987 NEC.)

To receive listing as Type CATVX, cables must pass the "VW-1 (Vertical-Wire) Flame Test" in the Reference Standard for Electrical Wires, Cables, and Flexible Cords, ANSI/UL 1581-1985.

In the ANSI/UL VW-1 Flame Test, an 18-in cable sample is mounted vertically over a cotton mat. An 816°C flame, directed at the center of the sample from a distance of 1-1/2 in (Figure 4) is applied for 15 seconds. The flame is applied again after either 15 seconds or the cessation of burning, whichever is longer. Cables which (1) support flame longer than 60 seconds or (2) emit flaming drops which ignite the cotton mat fail the test.

Coaxial Cable Installation Requirements

Where coaxial cable is exposed to lightning or to accidental contact with conductors operating a voltage of over 300 volts with respect to ground, the outer conductor

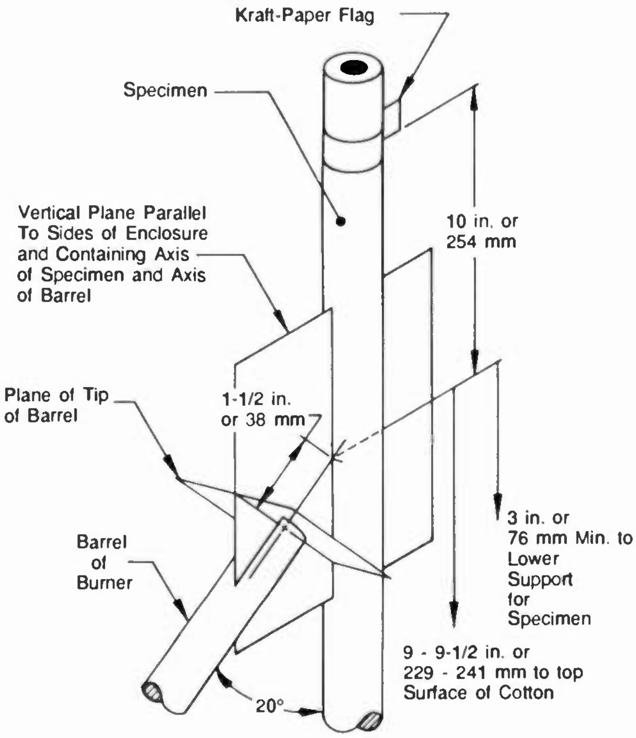

Figure 4. ANSI/UL 1581-1985 VW-1 Vertical Wire Flame Test

is required to be grounded at the building premises as close to the point of cable entry as practicable. Details of this requirement are given in Section 820-33 of the 1990 NEC (Section 820-7 of the 1987 NEC).

Beyond the point of grounding, the coaxial cable installation must meet requirements for (1) separation from other conductors; (2) spread of fire or combustion products in ducts and shafts; (3) equipment in space used for environmental air; and (4) listed hybrid power and coaxial cable in certain applications. These requirements are spelled out in Section 820-52 of the 1990 NEC (Section 820-13 of the 1987 NEC).

Coaxial Cable Application Requirements

The 1990 NEC defines three broad application categories for coaxial cable: plenum, riser, and other wiring within buildings. In certain cases, substitutions or other alternative installations are permitted. The requirements for these categories are as follows:

Plenum

All coaxial cable installed in ducts, plenums, and other spaces used for environmental air is required to be listed as Type CATVP (Section 820-53 (a)). One important exception is given: cables listed as Type CATVR, CATV, or CATVX -- but not unlisted cable -- contained within electrical metallic tubing, flexible metallic tubing, intermediate metal conduit, or rigid metal conduit and installed in compliance with NEC Section 300-22 are also permitted. This represents a strengthening of the 1987 NEC requirements, contained in Section 820-15 (c).

Riser

Coaxial cables installed in vertical runs and penetrating more than one floor, or cables installed in vertical runs in a shaft, are required to be listed as Type CATVR or CATVP (Section 820-53 (b)). Two exceptions to this requirement are given: (1) where the cables are encased in metal raceway or are located in a fireproof shaft having firestops at each floor and (2) in one- and two-family dwellings, in which case Type CATV and CATVX cables are permitted. This requirement is also strengthened from the 1987 NEC requirements, given in Section 820-15 (b), except for residential installations.

Other Wiring Within Buildings

Coaxial cables installed within buildings in locations other than plenums and risers, as defined above, are required to be listed as Type CATV, CATVR, or CATVP (Section 820-53 (c)). To this requirement, three exceptions are permitted: (1) where the cables are enclosed in raceway; (2) in non-concealed spaces where the exposed length of cable does not exceed 10 feet (3.05 m), in which case Type CATVX cable is permitted; and (3) in one- and two-family or multifamily dwellings, in which case Type CATVX cables that are less than .375 inch (9.52 mm) in diameter are acceptable. This requirement is likewise somewhat tighter than the equivalent requirement in the 1987 NEC, found in Section 820-15 (a).

Cable Substitutions

To summarize the preceding requirements, the following listed coaxial cables are permitted to be substituted for other types (Section 820-53 (d)):

Cable Type	Permitted Substitutions
CATVP	None
CATVR	CATVP
CATV	CATVP, CATVR
CATVX	CATVP, CATVR, CATV

Summary of NEC Requirements

As they apply to coaxial cables used in broadcast transmission, these NEC requirements may be summarized as shown on following page:

Application Within Buildings	Type CATVP	Type CATVR	Type CATVD	Type CATV	Type CATVX	Unlisted	NEC 1990 Section	NEC 1987 Section
In Ducts, Plenums, and Other Environmental Air Spaces, Exposed	X						820-53 (a)	820-15(c), Exception
In Ducts, Plenums, and Other Environmental Air Spaces, Contained in Metal Tubing	X	X		X	X		820-53 (a), Exception	820-15 (c), 300-22
In Vertical Runs Penetrating >1 Floor or in a Shaft, Nonresidential, Exposed	X	X					820-53 (b)	820-15 (b)
In Vertical Runs Penetrating >1 Floor or in a Shaft, Encased in Metal Raceway, or in a Fireproof Shaft Having Firestops at Each Floor	X	X	X	X	X	X	820-53 (b), Exception No. 1	820-15 (b), Exception
In Vertical Runs Penetrating >1 Floor or in a Shaft, in 1- and 2-Family Dwellings	X	X		X	X		820-53 (b), Exception No. 2	Not Covered
All, Except Plenums and Risers, Nonresidential Exposed >10 feet	X	X		X			820-53 (c)	820-15 (a)
All Except Plenums and Risers, Enclosed in Raceway	X	X	X	X	X	X	820-53 (c), Exception No. 1	820-15 (a), Exception No. 1
All, Except Plenums and Risers, Non-concealed, Exposed ≤10 feet	X	X		X	X		820-53 (c), Exception No. 2	820-15 (a), Exception No. 2
All, Except Plenums and Risers, in Single or Multifamily Dwellings	X	X		X	.375" max. dia.		820-53 (c), Exception No. 3	820-15 (a), Exception No. 3
All, Length Within Building ≤ 50 feet, Cable Enters from Outside, Grounded	X	X	X	X	X	X	820-50, Exception No. 2	Not Permitted
Lead-in or Aerial-Drop Cables Attached to a Building			X				Discont. Category	820-4, 820-11 (b)

Model Building Code Requirements for Coaxial Cable

Recent editions of model building codes and related mechanical codes have begun to include fire-retardance requirements for coaxial cable. Two of these are discussed below.

Uniform Building Code

The Uniform Building Code is published triennially by the International Conference of Building Officials, Whittier, California. It is the most widely adopted model building code in the world. Harmonized with the Uniform Building Code is the Uniform Mechanical Code, sponsored jointly by the International Conference of Building Officials and the International Association of Plumbing and Mechanical Officials.

The 1988 edition of the Uniform Building Code, Section 4305 (e), states that wiring in plenums shall comply with the Uniform Mechanical Code. Section 1002 (a), Exception 6, of the 1988 edition of the Uniform Mechanical Code states that electrical wiring in plenums shall comply with the NEC. However, it then goes on to give flame propagation and smoke production requirements for exposed electric cables installed in concealed spaces above suspended ceilings used as air plenums, based on U.M.C. Standard No. 10-3. These requirements are identical to those for Type CATVP cable. The cable is also required to be "listed and labeled as plenum cable as required by the electrical code."

Standard Building Code

The Standard Building Code is published at approximate three-year intervals by the Southern Building Code Congress International, Inc., Birmingham, Alabama. It has been adopted by a number of municipalities throughout the Southern United States. A related Standard Mechanical Code is also published by this association.

The 1988 edition of the Standard Building Code, Section 704.10, requires that wiring in concealed spaces such as spaces over suspended ceilings, plenums, ducts, and other spaces used for environmental air handling purposes shall meet the requirements of UL 910, i.e., those required for listing as Type CATVP. In the 1989 Revisions to the Standard Building Code, the requirement is added that this wiring shall be listed and labeled as plenum cable as required by the NEC.

Essentially the same requirements are repeated in the 1988 edition of the Standard Mechanical Code, Section 512.2, and its 1989 Revisions, Section 509.1.1.

Conflict Between Building Codes and the NEC

While the Uniform Building Code and Standard Building Code do not add any requirements for coaxial cable beyond those of the NEC, the fact that they repeat the requirements for plenum cable instead of incorporating them by reference opens up the possibility of conflict between a building code and the NEC -- particularly in the area of definition of a plenum. That this is not merely hypothetical is shown by an article by John "Gus" Degenkolb in the November/December, 1989 issue of Southern Building.

In his article, Mr. Degenkolb, a Fire Protection Engineer/Code Consultant in Carson City, Nevada, argues that the cable space under the raised floor of a computer room meets the definition of a plenum used by the model building codes, and so all cables used there should be listed and labeled as plenum cable. The 1990 edition of the NEC does not require this. Consequently, the model building codes have more restrictive requirements than the NEC in this instance.

Fire Gas Toxicity

Fire-Retardant Jacketing Characteristics

Some coaxial cables use a halogenated polymeric as a jacketing material to impart flame resistance to the cable. (Halogens are chemically related elements such as fluorine, chlorine, and bromine.) The drawback to such a material is increased levels of smoke and toxic gases under fire.

Other coaxial cables achieve fire retardance by using a non-halogenated jacketing material. While such a material has low toxicity characteristics when burned, it is somewhat less effective than halogenated jacketing. At present, Andrew Corporation has not found it possible to achieve the highest fire retardance rating, CATVP, without either employing halogenated jacketing or omitting the jacket entirely. For lesser degrees of fire retardance, non-halogenated jacketing is usable.

Fire Gas Toxicity Standards

Fire gas toxicity is often specified by the results of two international test methods, Naval Engineering Standard (NES) 711, Determination of the Smoke Index of the Products of Combustion from Small Specimens of Materials, and NES 713, Determination of the Toxicity Index of the Products of Combustion from Small Specimens of Materials. Both of these tests are useful for comparing the combustion characteristics of one material to another, but are not suitable for assessing the total fire hazard of a product. They are performed on the jacketing material itself, not the actual cable.

For non-halogenated fire-retardant cable jacketing materials presently used by Andrew, the Smoke Index as per NES 711 is 50-55 and the Toxicity Index as per NES 713 is 0.5.

New York State Requirements

On December 4, 1980, virtually the entire executive staff of Arrow Electronics was killed when a fiery inferno roared through the Stouffers Inn in Westchester, New York. Shocked by the rapid spread of this fire and the intense heat and toxic gases evolved, the New York State Legislature formed a fire safety task force. This led to the creation of a statewide uniform fire prevention and building code and, effective 23 October 1986, regulations requiring representative product testing and a fire gas toxicity data bank, to be maintained by the Department of State (DOS).

As a result, the New York State DOS Office of Fire Prevention and Control publishes a Fire Gas Toxicity Data File for products covered by Article 15, Part 1120 of the New York State Uniform Fire Prevention and Building Code. Only products listed in this directory are permitted to be used in the construction of some buildings in New York State.

Combustion toxicity testing is performed using a method developed at the University of Pittsburgh. This test measures not only the maximal concentration of carbon monoxide and carbon dioxide in the exposure chamber, the minimal concentration of oxygen in the chamber, the heat given off and the temperature at which spontaneous flame occurs, but also the effects of exposure of the products of combustion on small animals. Animal effects are reported as an LC-50 index, the estimated concentration of smoke needed to kill 50% of the test animals.

At present, however, no performance standards or pass/fail criteria have been established; the state is simply creating a data base to be referred to in the event of a fatal fire. Unfortunately, the utility of this data base is compromised somewhat by the necessity for applying a single test series to an entire class of products in order for the laboratories involved to complete the testing in the allotted time, which means that most building products have not been tested individually. Consequently, it is likely to be many years before any standards for fire gas toxicity for coaxial cables are developed.

The current edition of the Fire Gas Toxicity Data File is dated October 1988. Manufacturers are listed alphabetically. For each class of products, a DOS file number is assigned; following this number is given the market name or names for the product class. Both coaxial cable and elliptical waveguide are covered.

Future Trends

There is a clear trend toward increasing the restrictiveness of electrical, building, and fire prevention codes concerning the fire retardance, smoke production, and fire gas toxicity of coaxial cable, regardless of cost. Each new edition of a code adds fresh requirements, yet continuing ambiguities and exceptions suggest that further restrictions are to come. The treatment of coaxial cable in the codes is not likely to stabilize until the mid-1990's.

Another trend is the creeping expansion of the definition of a plenum, in the form of "other space used for environmental air." This is so open-ended a term that it could readily be extended to virtually any space within a building that is heated or cooled and that vents to another part of the building.

Should these trends continue, it is possible to foresee the several fire-retardance categories of coaxial cable collapsing into two: plenum listed and completely unlisted.

Design Considerations

Fire-retardant coaxial cable is significantly more costly than general purpose cable. But it is usually even more expensive to install unlisted cable in noncombustible tubing.

To minimize cost, the hierarchy of cable installations given below should generally be followed (listed in order of increasing cost):

1) No coaxial cable runs inside a building, or 50 ft or less of cable within a building, entering from outside and terminated at a grounding block. Unlisted cable is acceptable.

2) No coaxial cables installed in risers or plenums (or other spaces used for environmental air). Type CATV listed cable is acceptable.

3) No cables run in plenums or other air-handling spaces. Type CATVR listed cable is acceptable.

4) Unrestricted cable runs using Type CATVP listed cable.

5) Unrestricted cable runs using cable installed in approved noncombustible tubing. Type CATVX cable is required in plenums and air-handling spaces.

Of course, cables with different fire retardance ratings could be connected together to adjust the rating to the installation conditions, but this is unlikely to be practical. Consequently, if even one short section of a coaxial cable passes through an air plenum, it is probably best to use Type CATVP listed cable for the entire run.

How to Assure Compliance With All Regulations

The following is intended to be a foolproof checklist for avoiding problems due to noncompliance with fire protection regulations. As a result, some steps may seem obvious and unworthy of remark. However, they are included because otherwise competent engineers have made costly mistakes through overlooking them.

1) Check state and local building and fire protection codes. Determine what is required.

2) If in any doubt as to any requirements, definitions, etc., seek rulings from building and electrical inspectors in advance.

3) Specify cable with correct NEC listing or higher.

4) When cable arrives at the site, verify that the required NEC listing (CATVP, etc.) is labeled on the jacket (or on the reel, if the cable is unjacketed). This will assure the building and electrical inspectors that the proper cable is being used.

5) For New York State or similar requirements, ask the manufacturer or distributor to provide you the New York State DOS File No. for the cable so that you can cite it to the inspector. (DOS File Nos. are generally not marked on the cable.)

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RF HOT SPOT FIELDS: THE PROBLEM OF DETERMINING COMPLIANCE WITH THE ANSI RADIOFREQUENCY PROTECTION GUIDE

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ABSTRACT

Determining compliance with radiofrequency (RF) radiation guidelines of the Federal Communications Commission often requires measurements to assess the strength of the RF fields. A major issue in determining compliance with RF protection guides has been the treatment of so-called RF hot spots. This paper addresses RF hot spots by examining the peak specific absorption rate (SAR) which results from individuals contacting objects which are the source of hot-spot fields. Peak SAR was assesssed by measuring the contact current which flows between the hot-spot source and the individual touching the source and calculating the current density in the wrist which represents the area of smallest geometrical cross-section. SAR is proportional to the local current density and tissue conductivity. Several contrived situations including a resonant dipole, a metallic mail box, a filing cabinet, a simulated guy wire and an aluminum frame window in a wooden building were exposed to known RF fields for which surface fields and contact currents were measured. It was shown that the field strengths are not good indicators of either whole-body or local SAR in exposed individuals. In many cases where ambient fields are less than the RF guide but local, hot-spot fields exceed the guide, based on field strength, an evaluation of contact currents reveals that the underlying intent of the protection guide is not, in fact, exceeded.

BACKGROUND

Rarely is there ever an occasion when performing environmental surveys of radiofrequency (RF) fields that the issue of so-called RF hot spots does not come up or does not become a problem in making a decision about whether the site is in compliance with the RF protection guide (RFPG) recommended by the American National Standards Institute (ANSI). Actually, whether the exposure protection guide be the ANSI RFPG or some other guide, promulgated by a state, county or city, makes little difference; commonly, the individual performing the RF survey is ultimately confronted with the problem of interpreting the meaning of very localized regions in which the RF field strengths are excessive while the predominant field values in the general area are well below the RFPG. Figure 1 illustrates a common occurrence during field surveys where a metallic object can enhance the ambient fields in the general region. To what extent do these small regions of space count for declaring the site "off limits"? Can these local regions with high fields actually be considered hazardous despite the fact that most of the space surrounding these points shows field strengths no where near the exposure limits? How do these enhanced fields commonly found on the surface and very near to reradiating objects relate to absorption of RF energy in the body if a person is exposed? The subject of this paper evolved from an indepth consideration of these issues; this paper is an attempt to summarize the some of the key findings of this earlier study (Tell, 1989).

Hot spots occur anytime that ambient RF fields expose objects which are conductive; the RF fields induce currents in the object and

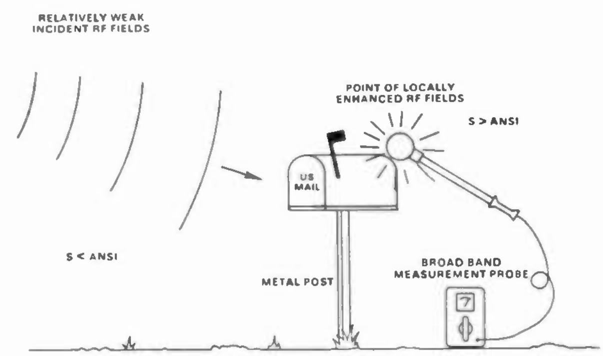

Figure 1. Metal objects exposed to weak ambient RF fields can produce enhanced field strengths near their surfaces, leading to apparent exceedances of applicable RFPGs.

these currents, in turn, lead to electric and magnetic fields being produced by the object. The fact that the strength of these surface fields is typically substantially greater than the fields in the surrounding area has led to the common name, RF hot spot. Hot-spot production is a natural and not unexpected consequence of the action of RF fields. That metallic objects like a curtain rod might act like a kind of antenna is not difficult to understand; but anything that is conductive, not just metal objects, will tend to do the same thing. The human body, for example, when illuminated with an RF field, will also exhibit hot-spot production; with outstretched arms in a very-high-frequency (VHF) field, one can easily verify this phenomenon by simply placing a RF sensing probe near the tip of the fingers and observing a substantially elevated field strength when compared to the value without the body present. So, the real issue is one of developing some insight to what these hot spots mean when compared to our understanding of antennas and other common experiences.

The term hot spot, as used in this paper, refers to limited volumes of space in which RF field strengths are substantially greater than in the surrounding space. Reradiators are sometimes referred to as secondary radiators or scatterers; some form of passive object, a metal handrail, a curtain rod, a metal door handle, a mail box or a child's swing set can all meet the criteria for becoming a reradiator if ambient RF fields exist in its vicinity. Reradiating objects produce enhanced (elevated) RF fields which typically diminish to the ambient levels in the surrounding areas within very short distances of the source. Field strength reduction is generally exponential with the highest strengths on the surface of the reradiating object. Very often, reradiating fields have very high field impedances; i.e., the ratio of the electric and magnetic field strengths at the surface of the object is much larger than the free space impedance of 377 ohms. Reradiator hot spots are also characteristically highly non-uniform in their distribution on the reradiating object and are generally not amenable to calculation.

THE ANSI RFPG AND RELEVANT DOSIMETRY CONSIDERATIONS

Introduction

Interpreting human exposure to RF fields usually reduces to a comparison of measured field strengths with the limits specified in applicable RF protection guides (RFPGs). For purposes of this study, the relevant RFPG is that recommended by the American National Standards Institute (ANSI, 1982). The ANSI RFPG, as well as several other recommended limits on human exposure to RF fields including those of the International Radiation Protection Association (IRPA, 1988) and the National Council on Radiation Protection and Measurements (NCRP, 1986), is based on the observation that the vast majority of biological effects caused by RF fields in experimental animals are directly related to the rate at which RF energy is absorbed from the exposure field. The quantity used to specify energy absorption is called the specific absorption rate (SAR). The SAR is expressed in units of watts per kilogram of body mass (W/kg) and is related to the internal electric field strength in the tissue by the relationship:

$$SAR = sE^2/p \qquad [1]$$

where

SAR = specific absorption rate (W/kg);
s = conductivity (S/m);
E = electric field strength in tissue (V/m);
p = density of tissue (kg/m^3).

In practice, experimental studies of SAR in exposed objects rely on either a measurement of the field strength in the exposed tissue or the increase in temperature caused by absorption of the RF energy.

Through a critical examination of the technical literature, ANSI elected to use the SAR, determined as an average over the mass of the entire body, as a fundamental basis for the RFPG. But because direct measurements of whole-body averaged SAR in individuals is not practical, ANSI chose to establish guides on maximum values for the electric and magnetic field strengths or plane-wave equivalent power density. In actuality, the external field limits are in terms of the squares of the electric and magnetic field strengths (E^2 and H^2). For a plane wave in free space, the power density can be related to the squares of the electric and magnetic fields as follows:

$$S\ (mW/cm^2) = E^2/3770 = 37.7\ H^2 \qquad [2]$$

where

S = plane wave equivalent power density (mW/cm^2)
E = electric field strength (V/m)
H = magnetic field strength (A/m)

and the factors 3770 and 37.7 are factors associated with the impedance of free space (377 ohms) and conversion to appropriate units of milliwatts per square centimeter.

For convenience, ANSI chose to round the value for the impedance of free space from 377 ohms to 400 ohms. So, the actual ANSI limits are expressed in slightly different form from equation [2] as follows:

$$S\ (mW/cm^2) = E^2/4000 = 40\ H^2 \qquad [3]$$

Frequency dependence

To arrive at practical limits on the external exposure fields, both theoretical and experimental data have been developed to estimate the SARs associated with various field strengths and frequencies. Perhaps the first effort to examine the ability of the human body to act as an antenna was that of Andersen and Balling (1972). In this classic series of measurements, individuals were stood on a ground plane and the admittance and radiation efficiency of the human body was measured. Radiation efficiency was determined by comparing the apparent gain of the body to that of a metallic whip antenna. Thus, as early as 1972, the concept of body gain was introduced; Andersen and Balling reported that at a frequency of 60 MHz, the gain of the body was about -3 dB relative to that of a vertically polarized, perfect monopole.

More recently, extensive studies, notably by Gandhi and colleagues at the University of Utah and by Guy at the University of Washington, have been conducted. These studies have included experimental dosimetry studies based on calorimetric measurements and theoretical electromagnetic field calculations. Based on modeling the human body as the simplistic shape of a prolate spheroid having a uniform internal constituency, it was found that the body exhibited properties much like a radio antenna. These important findings described how the SAR varied as a function of frequency throughout the radio spectrum. Durney and co-workers have assembled several comprehensive treatments of RF dosimetry data during the past fourteen years (Johnson et al., 1976; Durney et al., 1978; 1980; 1986). The so-called standard man model consists of 70 kg of a muscle equivalent material having dielectric properties similar to muscle tissue and shaped as a prolate spheroid.

Since the SAR of the body will depend on the body dimensions, mainly the length of the body which is parallel to the electric field, there is a band of frequencies over which the SAR is at its maximum value depending on size. Hence, the ANSI RFPG is frequency dependent as shown in Figure 2. The region within the range of 30 to 300 MHz takes into account the fact that different size individuals are in the potentially exposed population, including very tall persons and small children. To compensate for this, the ANSI RFPG is most stringent in this so-called body resonance range of frequencies. At frequencies below body resonance, the SAR decreases typically inversely with the square of the frequency.

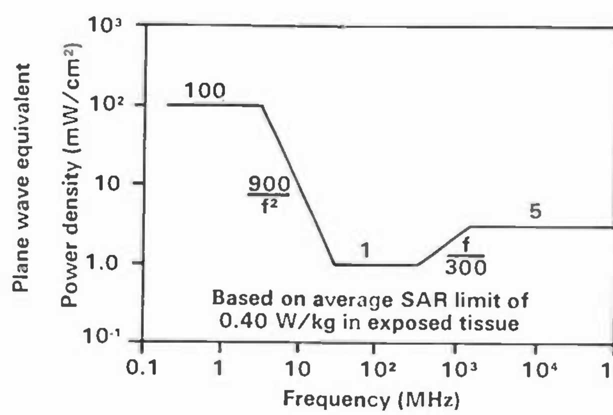

Figure 2. Frequency dependence of the ANSI RFPG (ANSI, 1982) for plane-wave equivalent power density. Values are to be time-averaged over any 6-minute period.

The ANSI RFPG field limits are based, however, not on free-space SARs but rather the SAR which occurs when an individual is standing on the ground. Under this circumstance, the body acts more like a monopole antenna above a ground plane and its gain is increased over that of the simple dipole suspended in space, far removed from ground and the resonance frequency is decreased. Results from Gandhi et al. (1977) have verified that the SAR of the grounded body is very close to about 0.4 W/kg/(mW/cm^2), being approximately 3 dB greater than for the free space situation. It is on the basis of grounded SARs that the field limits in the ANSI RFPG were derived, this being the more conservative case for exposure. Thus, in the body resonance range, the recommended power density limit is set at 1 mW/cm^2 which will result in limiting the whole-body SAR to 0.4 W/kg.

Local SAR Limitations

The underlying basis for the ANSI RFPG is limiting the whole-body averaged SAR to no more than 0.4 W/kg. This specific value is related to the observation in experimental animals exposed to strong RF fields that thermally mediated alterations in trained behavior were relatively reliably induced when the whole-body SAR reached a value of about 4 W/kg. ANSI incorporated a safety factor of ten in arriving at a working level of 0.4 W/kg for application to human exposure. It was clear from numerous dosimetry studies, however, that the actual distribution of SAR within the body of the experimental animals was considerably nonuniform with local, spatial peak values as great as about 20 times the whole-body average value. So, while the average SAR over the body mass associated with behavioral disruption was about 4 W/kg, there were some areas within the body that could be absorbing energy from the external exposure fields at a rate as high as 80 W/kg. Hence, the present ANSI RFPG limits the whole-body averaged SAR to 0.4 W/kg but permits the spatial peak SAR to rise to 8 W/kg to account for the fact that, in the experimental studies on animals, such nonuniformity in the energy absorption rates undoubtedly existed.

A very important note is in order, however, relative to the 8 W/kg spatial peak SAR limit of the ANSI RFPG. The limiting value of SAR was derived from the nonuniformity of SAR within free-space irradiated experimental animals. Virtually no experimental results reported in the literature have been for the case of animals situated on a ground plane. Experimental studies of human models exposed on a ground plane and theoretical studies have shown, nevertheless, that in the grounded condition, much higher ratios of the local to average SAR over the body can occur. In some of these studies, local, spatial peak SARs up to 60 times the body average have been observed for the grounded condition and far-field exposure. But because the vast majority of the data base that existed prior to the issuance of the 1982 ANSI RFPG was for free space, ungrounded exposure conditions, ANSI elected to maintain a local SAR limit of 8 W/kg. More recently, numerous studies have been conducted which have explored the issue of local SARs resulting from near-field exposures. These studies have also been primarily restricted to models of the human body, continuing to leave the question open as to the biological impact of near-field exposures and high level, local SARs. Thus, on a dosimetry basis, the ANSI RFPG restricts whole-body SAR including the condition of an individual standing on earth but restricts the elevation in spatial peak SAR to a factor of 20 times the average, consistent with free-space, ungrounded exposures.

SAR and Current Density

The SAR in the body can be expressed in terms of the local current density according to the following relationship:

$$SAR = J^2 rho/p \qquad [4]$$

where
SAR = specific absorption rate (W/kg);
J = current density in the tissue (A/m^2);
rho = tissue resistivity (ohm-meters) = 1/s;
p = density (kg/m^3).

The resistivity is the reciprocal of the conductivity. Equation [4], when practically applied at 30 MHz for muscle tissue, can be simplified to:

$$SAR = 0.0013J^2 \text{ (W/kg)} \qquad [5]$$

The practical significance of this expression is that if the induced currents can be determined in the body, the SAR can be estimated. This becomes particularly relevant when considering currents that flow through the legs, ankles and feet or through the hands and wrists when touching objects which have RF currents flowing on them such as reradiating objects. It is in this context that equation [4] will be found to provide a meaningful way to interpret the significance of human exposure to RF hot spots.

Since the local SAR is directly related to the current density, determining the effective cross-sectional areas through which the current flows becomes the critical issue. In general, the two prime areas of potential high SAR are the ankle and the wrist, those two parts of the anatomy having the smallest cross-sectional areas. But even within the ankle and wrist, considerable variation in conductivity exists since there is a variety of tissue types involved including not only muscle tissue but large amounts of bone which is relatively nonconductive. Thus, a major concern becomes the determination of how much high conductivity tissue exists in these regions. One approach to this problem is examining the cross-sectional anatomy of the structure through anatomy text books and assigning physical areas to each of the major tissue types. Gandhi et al. (1986), for example, reported an effective cross-section area of only 9.5 cm^2 for the average adult ankle. This small area of muscle equivalent tissue, having high conductivity, is due to the large amount of bone in the ankle. If the RF current must now flow through this small area, the local current density becomes approximately 300 A/m^2 with a resulting SAR in the smallest area of the ankle of 117 W/kg! This value obviously far exceeds the spatial peak SAR limit of the ANSI RFPG yet the incident electric field does not exceed the RFPG at 30 MHz.

Chatterjee et al. (1986) have also examined the wrist from an anatomical cross section viewpoint and report that the effective conductive cross section is 11.1 cm^2. Their finding that the conductive cross section is greater than their estimate of the conductive cross section for the ankle, despite the ankle's larger gross cross-sectional area, is presumably due to the disproportionately smaller amount of bone in the wrist.

Recently, the Accredited Standards Committee C95 on Non-ionizing Radiation Hazards has recommended the use of a 20 W/kg SAR limit for use in the wrists, ankles, hands and feet (ASC, 1989). The draft rationale states "The 20 W/kg limit for the wrist and ankles allows higher absorptions in the soft tissues produced by the induced currents specified in Table 2 flowing in these bony cross-sectional areas. Relatively high surface area to tissue volume ratios for these parts of the body, the common experience of high-temperature excursions on these parts without apparent adverse effects and the lack of critical function when compared to other organs are considerations which mitigate these higher permitted local SARs." While this statement is still in draft form, it closely parallels the provisions of both the IRPA and the National Radiological Protection Board in the United Kingdom (NRPB, 1989) which allow local SARs up to 20 W/kg as averaged over tissue volumes of 100 g in the extremities.

In summary, the ANSI RFPG (ANSI, 1982) contains limits on both the whole-body average and spatial peak SARs which are not to be exceeded when averaged over any 6-minute period of exposure. For exposure to highly nonuniform RF fields, spatial peak SARs generally become the limiting exposure criterion before the whole-body SAR is reached. This is especially likely when evaluating exposure to RF hot spots where enhanced absorption in the wrists and ankles may become the controlling factors for practical assessments of compliance with the RFPG in the field. Induced RF currents provide one indirect measure of local SAR in these regions which has practical utility during the conduct of routine environmental surveys of RF fields.

TECHNICAL APPROACH USED IN STUDY

In approaching the problem of evaluating exposure to RF hot spots, an underlying concept was envisioned. This concept is straightforward and intuitively apparent. It can be stated simply as whole-body energy absorption rates should be related to the ambient field strengths in a region occupied by the body rather than local fields which may be significantly elevated by comparison with the ambient fields but present only partial body exposure. Stated in another way, there is only so much energy which can be absorbed from the field in a given region of space, regardless of how it may be redistributed by the presence of reradiating objects. Yet another alternative expression of the concept is that the local field strengths, when exposure occurs near or at the surface of a reradiating object, are not necessarily good indicators of absorbed energy in the body and thus not necessarily good measures for determining compliance with the RFPG. Kucia-Korniewicz (1974) discussed long ago the issue of coupling between biological objects and field sources, indicating that the electric field strength alone was not always sufficient to assess the biological action of the fields on exposed specimens.

This underlying concept influenced one of the measurements carried out in the project; a resonant dipole antenna was constructed which was connected to a sensitive power meter as illustrated in Figure 3. The dipole was exposed to RF fields at its resonant frequency which were a small fraction of the permitted limits recommended by ANSI. Then, the power absorbed by the dipole was measured and the local RF fields near the dipole's surface were measured. In this fashion, the strongly enhanced surface fields, which easily exceed the ANSI RFPG, suggest one conclusion while the total power absorbed by the dipole suggests quite another conclusion.

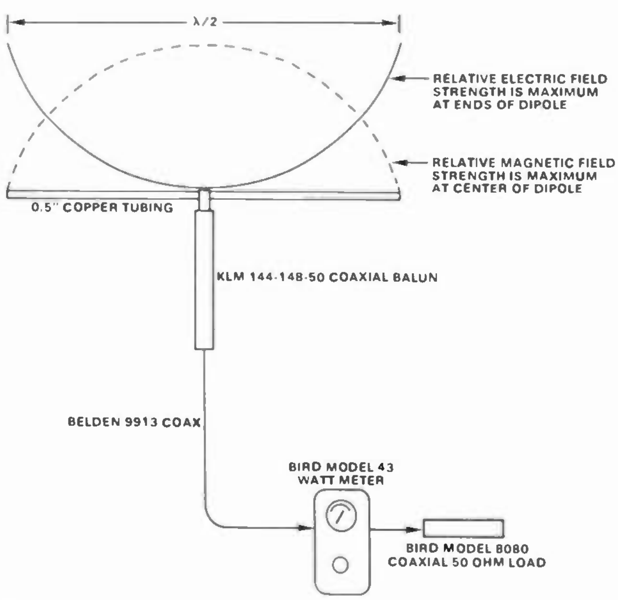

Figure 3. Distribution of relative electric and magnetic field strengths near a half-wave length dipole exposed to a uniform electric field. A resonant dipole at 144.05 MHz was used as a passive receiving antenna and reradiating source in some of the measurements.

Although the resonant dipole is interesting to study since its characteristics can be easily computed and measured, reality is such that the objects encountered in field surveys are more usually not resonant structures, i.e., they are not good antennas. There are innumerable objects which, each in its own characteristic way, can become reradiators. To address the generic category of reradiating objects, it was decided that a limited sample of such objects would be included for measurements in the project. A driving concern was to develop insight to the hot spots that could be caused by reradiating objects exposed to RF fields in general rather than developing a comprehensive catalog of such objects. For this project, in addition to the half-wave dipole, a metallic mail box, a filing cabinet, an aluminum window frame in a wooden frame building and a simulated guy wire were all investigated for their ability to produce enhanced RF fields near their surfaces.

While measuring the absorbed power of a resonant dipole is one matter, it was not feasible to consider a similar absorbed power measurement from the other objects studied. So, in addition to the measurement of ambient fields and surface hot-spot fields, a measurement technique was developed to measure the magnitude of induced current which would be conducted into the body upon contacting the reradiating object. It was reasoned that since the current flowing in the tissues of the body could be related to the local value of SAR, a current measurement represented the most practical approach to the SAR determination and consequently the most efficient approach to evaluating whether exposure to the reradiated fields would lead to exceedances relative to the ANSI RFPG. A further step was taken in the thinking about this approach; virtually any situation in which RF fields caused by reradiating objects call to question compliance with the RFPG implies the possibility of physical contact with the reradiating object. Thus, the issue of simple exposure to the RF fields becomes of less importance since an individual may not just be exposed by virtue of being positioned very close to the object but also will probably have access to the object in which case they will form contact. The physical contact situation substantially increases the possibility that RF currents can flow between the object and the body thereby raising the probability of exposure to increased SAR. With this rationale, the reradiating objects used in this project were evaluated on the basis of contact currents when an individual forms contact by touching the object.

INSTRUMENTATION AND MEASUREMENT METHODS

RF Fields

The primary instruments used in this project for measuring electric and magnetic fields consisted of broadband, isotropic survey probes similar to those commonly used in conducting RF field surveys. For electric field measurements, a Holaday Industries, Inc. Model HI-3001 broadband field strength meter (S/N 39586) equipped with a Model GRE (S/N 297A) green electric field probe (0.5 MHz - 6 GHz) was used. Measurements of magnetic fields were performed with The Narda Microwave Corporation Model 8616 (S/N 24332) meter equipped with a Model 8631 (S/N 17174) magnetic field probe (10 - 300 MHz).

Both of these instruments were calibrated by inserting them in a transverse electromagnetic (TEM) cell (Instruments for Industry, Inc. Model IFI 105) using a method described by Mantiply (1984). By propagating a known power through the TEM cell and knowing the characteristic impedance of the cell, the electric and magnetic field strengths developed inside the cell can be known very accurately. Both calibrations were carried out at a frequency of 144.0 MHz. In each case, the probe was inserted through a door in the side of the TEM cell and slowly rotated, noting the maximum and minimum readings. The average of the two readings was compared to the known, applied electric or magnetic field strengths in the TEM cell to obtain a correction factor which would be applied to all readings obtained with either probe during the project.

Contact Currents

Measurement of contact currents was conducted by using an assembly consisting of a metal probe, RF current transformer and tuned receiver. The current probe assembly consisted of a 6 inch long piece of 7/8 inch diameter copper tube (type M) soldered to another 6 inch long piece of 5/8 inch diameter copper tube with the use of an adapter to accommodate the two different diameters. A commercially available RF current transformer (Ailtech Model 94111-1 [S/N 451]) (NOTE: Ailtech is now known as Eaton Corportion) was affixed to the smaller diameter copper tube with the use of a high density, closed cell type of polyurethane foam cut in a cylindrical form which surrounded the smaller diameter copper tube and formed a tight friction fit with the inside of the current transformer circular aperture (1.25 inches diameter). The current transformer was positioned approximately 3 inches from the end of the small copper tube. To facilitate making connection with various objects, the end of the smaller tube was cut across two diameters for a length of 7/8 inch and then flared. This arrangement proved to be effective in making good electrical contact with different objects for which contact currents were to be measured. Figure 4 is a photograph of the current probe assembly as used in the project. The operator, by holding the larger diameter end of the copper tubing, formed the other part of the circuit through which the current would flow. In each measurement of contact current, the current probe assembly was pressed into contact with the object under evaluation and the surface of the object was explored until a maximum indicated RF voltage was observed on the field strength receiver. The maximum voltage was used in the conversion to contact current. All contact current measurements were performed with the individual wearing rubber sole tennis shoes. Measurements of contact currents while barefooted were performed to evaluate the effect of the shoes.

Figure 4. Photograph of the current probe assembly used to measure contact currents.

The contact currents measured with this system were those actually flowing between the RF hot-spot source and the body; others (Stuchly and Lecuyer, 1989) have employed various forms of equivalent circuits to represent the impedance that the body represents upon contacting objects. In this fashion, should excessive currents be present, they would not flow into the body of the person conducting the tests.

The RF output voltage from the current transformer was connected to the input of a Potomac Instruments Model FIM-71 field strength receiver (S/N 888) through a 1.5 m long piece of RG-58/U type coaxial cable (Belden 9310) equipped with type BNC connectors. A type N to BNC adapter was used to connect the coaxial cable to the current transformer. The field strength receiver was tuned to 144.05 MHz, the frequency used for the experimentation performed in this investigation, and the measured RF voltage was then related to the current flowing through the current probe assembly. Figure 5 shows the current probe assembly along with the Model FIM-71 receiver.

The current flowing through the aperture of a current transformer is related to the transfer impedance according to the relationship:

$$I = V/Z_T \text{ where} \qquad [6]$$

I = the contact current obtained by touching an object exposed to a RF field that may generate local RF hot spots (mA);

V = the rms RF voltage delivered by the current transformer (mV);

Z_T = the current transformer transfer impedance (ohms).

As the experimental data began to develop in the project, the question of how the contact currents which were being measured related to currents elsewhere in the body arose. To investigate this curiosity, a 3/4 inch inside diameter PVC pipe was cut to a length slightly greater than 2 m, capped on one end and then filled with a saline solution (0.9 percent by weight sodium chloride - table salt). At the top of the pipe, a metal electrode was fashioned which made electrical contact with the saline solution and provided a forked contact for contacting the dipole antenna used in some of the testing program. The PVC pipe assembly was brought into contact with the dipole antenna near its driven point (center). The PVC pipe permitted the current transformer, used for making contact current measurements, to be fitted around the pipe and moved along its axis to study the spatial variation of conducted current.

Figure 5. Photograph of the current probe assembly and field strength receiver used to measure contact currents. The receiver is used as a sensitive tuned voltmeter for detecting the RF output voltage from the current transformer.

Field Generation

The equipment configuration used to generate RF fields is outlined in Figure 6. A four element yagi antenna (Cushcraft Model 124WB) designed for use in the two-meter amateur radio band was mounted two meters above ground on a fiberglass, telescoping mast inserted in a five gallon plastic pail filled with concrete for ballast. The yagi has a nominal gain of 10 dB.

The signal source used was an ICOM Model IC-275A two meter transceiver which is capable of approximately 25 W output power. The transceiver output was used to drive a Mirage Model B1016 amplifier which could produce up to 160 W into the yagi transmitting antenna. A Bird Model 43 power sensor was used to monitor the input power to the yagi antenna.

Reradiating dipole

A resonant half-wave dipole antenna was constructed from 5/8 inch diameter copper tubing with the arms of the dipole physically held in

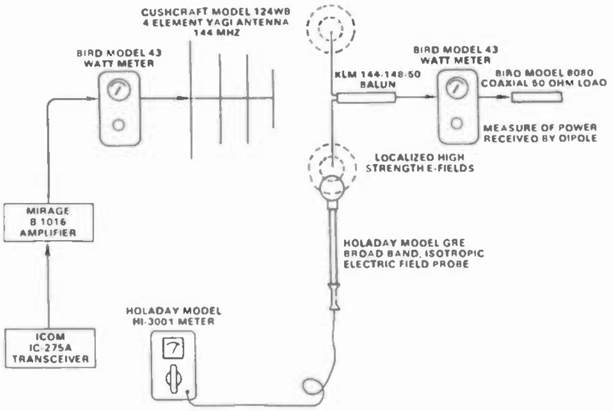

Figure 6. Equipment configuration for generating and measuring localized electric fields near a reradiating, passive dipole antenna.

place with a wooden (pine) structure. A KLM Model 144-148-50 coaxial balun was used to minimize currents flowing on the outside conductor of the coaxial cable which was connected to a Bird Model 43 power sensor. The dipole was constructed with the arms too long, approximately 40 cm each, and the antenna was brought into resonance on a frequency of 144.05 MHz by successively trimming equal lengths from both sides of the dipole until the voltage standing wave ratio (VSWR) was minimized. Through this process the VSWR was minimized to a value of 1.12. Under this condition, only 0.3 percent of the power delivered to the dipole will not be absorbed.

The dipole assembly was mounted 2 m above ground approximately 10 m from the transmitting dipole and was supported by three nylon guy ropes. Another nylon rope was stretched horizontal to the ground and parallel to the dipole, immediately next to the elements. This rope served the purpose of identifying the position in space that the dipole occupied for subsequent field strength measurements. The area used for the tests was a grassy, flat area with no intervening vegetation. The power sensor used to measure absorbed power by the dipole was equipped with a sensitive 0-1 W sensor element. The output of the power sensor was terminated with a 50 ohm load, a Bird Model 8080. With this system, absorbed powers as low as 0.02 W could be readily measured.

The general approach used in the project was to generate an arbitrary exposure field at 144.05 MHz and then introduce various objects into this field and then examine them for RF hot-spot production and contact currents. This is diagrammatically illustrated in Figure 7 for a standard filing cabinet, similar to that found in many offices, and Figure 8 for the simulation of a guy wire that might be found at FM broadcast station antenna tower sites. The guy wire was configured to evaluate the effect of a guy wire rigged near an FM or TV broadcast antenna which would be exposed to high intensity RF fields near the antenna but might conduct RF currents down the guy wire, leading to the generation of RF hot spots near the wire.

MEASURED AND CALCULATED RESULTS

Half-wave Passive Dipole

Fields - Using the nylon rope mentioned above, the exact location of the half-wave dipole axis was determined and marked on the reference rope using a marking pen with the dipole in place. The dipole was then removed, leaving the nylon rope in place, and an arbitrary RF field was radiated toward the rope with the yagi transmitting system. Electric and magnetic fields were measured along the axis of where the dipole would subsequently be located by slowly moving first the electric field probe and then the magnetic field probe along the nylon rope. The

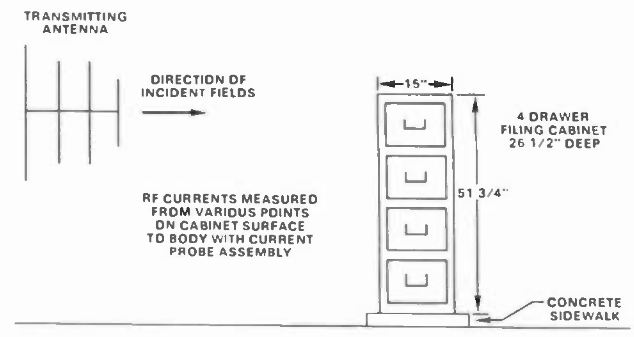

Figure 7. Measurement setup for determining induced RF currents when contacting a filing cabinet exposed to ambient RF fields.

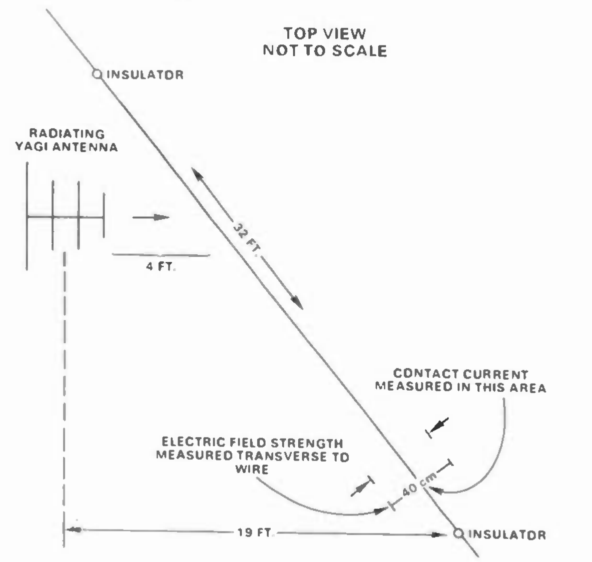

Figure 8. Measurement setup for determining local electric field strength and contact current from a simulated antenna tower guy wire.

measurements were accomplished by kneeling on the ground and holding the measurement probe above the head, keeping the probe as close to the nylon rope as possible as it was moved uniformly between the two marks representing the ends of the dipole. The resulting data, in terms of the square of the electric and magnetic field strengths, is presented in Figures 9 and 10 respectively. These data have been corrected for probe response at the test frequency of 144.05 MHz and indicate that the mean field strengths over the length of the dipole were 341 +/- 23 V^2/m^2 (+/- 0.29 dB) and 0.00207 +/- 0.00015 A^2/m^2 (+/- 0.31 dB) for electric and magnetic fields respectively. The magnetic field strengths were obtained by using relation [2] to convert from units of power density as indicated on the Narda Model 8616 meter.

This reference field represents an electric field strength that is approximately 8.5 percent of the ANSI RFPG and a magnetic field strength that is approximately 12 percent of the RFPG. The fields were found to be quite uniform along the length of the dipole, typically +/- 0.3 dB, and to represent a field impedance equal to 406 ohms. This is near the free-space value of 377 ohms but the difference is likely due to the presence of ground reflections between the transmitting yagi antenna and the dipole's location.

Next, the dipole was re-positioned adjacent to the nylon rope and connected to the high sensitivity power sensor. Electric and magnetic fields were measured along its axis using the same technique described above and the absorbed power was recorded. The results are given in

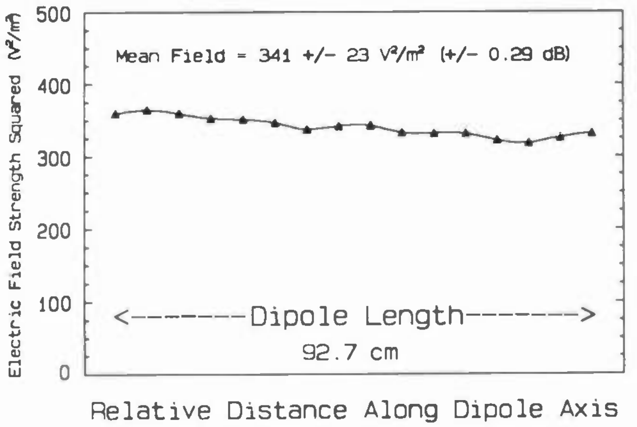

Figure 9. Variation of free-space electric field strength along axis of half-wave dipole antenna.

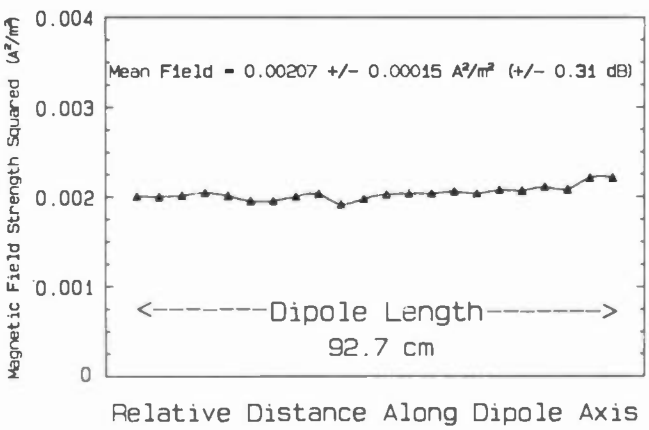

Figure 10. Variation of free-space magnetic field strength along axis of half-wave dipole antenna.

Figures 11 and 12 for electric and magnetic field strengths with a separation distance between the probe protective covering and the surface of the dipole of approximately 5 cm. The peak field strengths measured are indicated as 6770 V^2/m^2 and 0.0424 A^2/m^2 for electric and magnetic fields respectively. These peak fields are representative of RF hot spots encountered in routine field surveys and are substantially greater than the field limits given in the ANSI RFPG which are indicated by horizontal dashed lines in Figures 11 and 12. The location of the peak fields is also expected since the voltage distribution on a half-wave dipole produces a maximum at the ends of the dipole while the current will be a maximum at the feed point. While the surface peak field strengths were relatively high, the power absorbed by the dipole was measured as 0.46 W (460 mW).

Plane-wave equivalent power densities were calculated based on the measured squares of the electric and magnetic field strengths using relation [2]. The resulting power densities, S_E and S_H, corresponding to the electric and magnetic fields, were used to calculate expected absorbed power from the field by the dipole using the relation between gain and effective area of an antenna. The effective area of the half-wave dipole was calculated to be 0.566 m^2 and a power gain of 1.64 was assumed, based on the performance of a perfect, lossless dipole. Applying the power densities S_E and S_H and the effective area, absorbed powers of 0.512 W and 0.442 W were obtained. The mean value of these two powers is 0.477 which is in very good agreement with the measured value of 0.46 W. This difference is assumed to represent the difference between a perfect dipole and the losses and less than perfect impedance match inherent to the dipole constructed for the tests.

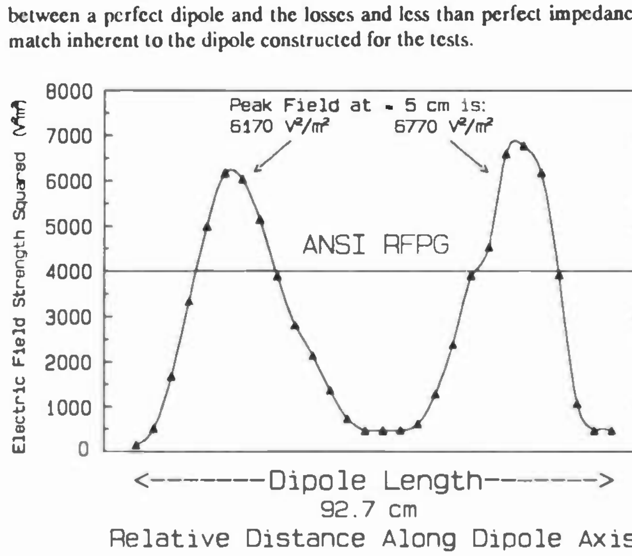

Figure 11. Measured electric field strength (E^2) distribution near surface of reradiating dipole in RF field comparable to approximately 10 % of the ANSI RFPG. The ANSI RFPG is shown as the dotted line.

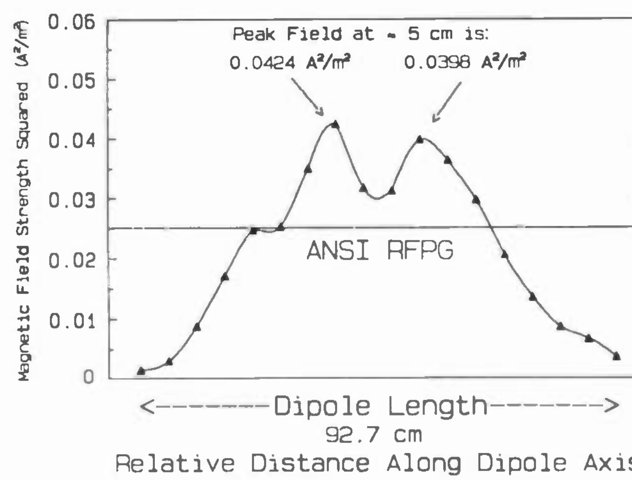

Figure 12. Measured magnetic field strength (H^2) distribution near surface of reradiating dipole in RF field comparable to approximately 10% of the ANSI RFPG. The ANSI RFPG is shown as the dotted line.

The field distributions shown are representative of many repetitions of the measurements and show that, for the real world dipole, the magnetic field strength exhibited a minimum directly adjacent to the feed point of the dipole, even though the peak in the magnetic field was very near the center of the dipole. This is presumed to be due to the physical separation of the feed point gap or the actual current distribution that existed on the dipole at resonance.

An antenna analysis program, MININEC (Rockway et al., 1988), was used to model the passive receiving dipole used in the tests. A resonant dipole having the same diameter as the test dipole was modeled with an input power of 0.46 W. The near fields were computed for a distance of 5 cm from the surface of the dipole along its length. Interestingly, the calculated peak field strengths of 7056 V^2/m^2 and 0.046 A^2/m^2 are in remarkable agreement with the measured values, being no more than 0.35 dB different than the experimentally determined values. The theoretical results are seen to track very well the measured values and indicate that such high surface fields are to be expected from a dipole either radiating or receiving low powers.

Currents - Contact currents were measured with several different incident field strengths to illuminate the passive dipole. The results are summarized below:

Dipole Contact Current Summary

Power absorbed by dipole (W)	Contact current (mA)	I-max dipole (mA)
0.12	27.7	40.5
0.50	60.9	82.8
0.79	73.8	104.0
1.0	83.0	117.0

The third column lists, for reference purposes, the maximum RF current which would be expected to flow in the feed terminals of the half-wave dipole when absorbing the indicated power in column one.

These data demonstrate that the contact current increases as the square root of the power absorbed by the dipole, as would be expected. The point at which the maximum contact current was observed was near the center of the dipole. Clearly, the current was reduced when contacting the end of the dipole where the maximum surface electric fields were measured. The maximum contact current was found to be approximately $83(W^{1/2})$ mA where W is the power absorbed by the dipole in watts.

The contact current was also measured when the dipole terminals were shorted, i.e., when no or very little power is actually being absorbed by the dipole but is being essentially completely reradiated. It was found that the contact current increased, for example, from 79.3 mA to 99.6 mA when the center gap was shorted and the dipole had been delivering 0.73 W into the matched load and power sensor. When the short was effected at the terminals, the power sensor indication reduced to 0.02 W. This observation indicates that the contact current increased by approximately 25 percent over the un-shorted condition.

A brief test of the effect of wearing shoes revealed that very little, and some times unmeasureable, increases in the measured contact currents occurred when bare-footed.

RF Hot Spots Caused by Other Objects

To examine the local fields and contact currents associated with RF hot spots, several situations were arranged in which objects were exposed to measured RF field strengths. These objects included a rural mailbox supported by a metal post, a filing cabinet, and an aluminum frame window. It was commonly found that the locations of the peak surface electric and magnetic fields were not the same. The highest electric field was found near one corner while the maximum magnetic field was found along the vertical member between the two panes of glass. This implies that the current and voltage distributions that existed on the window frame at the time of the measurements were highly complex. In general, surface field measurements performed on the various objects evaluated for hot spots revealed the same inconsistent pattern of local fields, suggesting that analytic solution of the expected fields would be very difficult, if at all possible.

Mail box test - Data representing the mail box test are summarized as follows:

Ambient fields		Surface fields	
E^2 (V^2/m^2)	H^2 (A^2/m^2)	E^2 (V^2/m^2)	H^2 (A^2/m^2)
620	0.0037	9800	0.022

Maximum contact current = 88.6 mA
E^2 field enhancement factor = E^2(surface)/E^2(ambient) = 15.8
H^2 field enhancement factor = H^2(surface)/H^2(ambient) = 5.9

Window frame test - Data representing the aluminum window frame test are summarized as follows:

Ambient fields		Surface fields	
E^2 (V^2/m^2)	H^2 (A^2/m^2)	E^2 (V^2/m^2)	H^2 (A^2/m^2)
1500	0.0094	9800	0.23

Maximum contact current = 133 mA
E^2 field enhancement factor = E^2(surface)/E^2(ambient) = 6.5
H^2 field enhancement factor = H^2(surface)/H^2(ambient) = 24.5

It is interesting to note the considerable difference in surface field enhancement factors between the window frame and the mail box; in the case of the mail box, the RF hot spot was characterized by a very enhanced electric field but in the case of the window frame, the opposite was true, the hot spot exhibited a greatly enhanced surface magnetic field. This may have been due to the loop-type structure of the window frame which may have tended to produce much higher currents relative to incident, exposure fields than could exist in the mail box.

Filing cabinet test - Data representing the filing cabinet test are summarized as follows:

Ambient fields		Surface fields	
E^2 (V^2/m^2)	H^2 (A^2/m^2)	E^2 (V^2/m^2)	H^2 (A^2/m^2)
600	--*	9430	0.0032

* Not determined
Maximum contact current = 42.4 mA
E^2 field enhancement factor = E^2(surface)/E^2(ambient) = 15.7

Guy wire test - Data representing the results obtained in the simulated guy wire test are presented graphically in Figure 13 in which the electric field strength transverse to one end of the wire from 10 cm on one side of the wire to 10 cm on the opposite side is plotted. Figure 13 dramatically illustrates the high degree of spatial variability of surface fields encountered at RF hot spots; the surface field measured with the electric field probe situated in contact with the wire is approximately 19 times greater than the ambient field just 10 cm away from the wire. Thus, even though the ambient field represented a plane-wave equivalent power density of only 0.08 mW/cm^2, the surface field was determined to represent an equivalent power density of 1.5 mW/cm^2. An important difference in the guy wire test is that one end of the wire was exposed to a rather intense (unmeasured field strength) field while the other end, near which the local fields were mapped, was actually in a very weak RF field; thus, the use of the term ambient in this case is not the same as for all of the other exposure tests performed. Using the value of surface fields as a determining factor in assessing RF fields at a broadcast site for compliance with RFPGs may easily lead to apparent exceedances; in many cases, these surface fields exist only in the immediate region of a conducting structure and are not representative of whole-body exposure.

Guy wire test - Data representing the guy wire test are summarized as follows:

Ambient fields		Surface fields	
E^2 (V^2/m^2)	H^2 (A^2/m^2)	E^2 (V^2/m^2)	H^2 (A^2/m^2)
300	--*	5800	--*

* Not determined
Maximum contact current = 94.1 mA
E^2 field enhancement factor = E^2(surface)/E^2(ambient) = 19.3

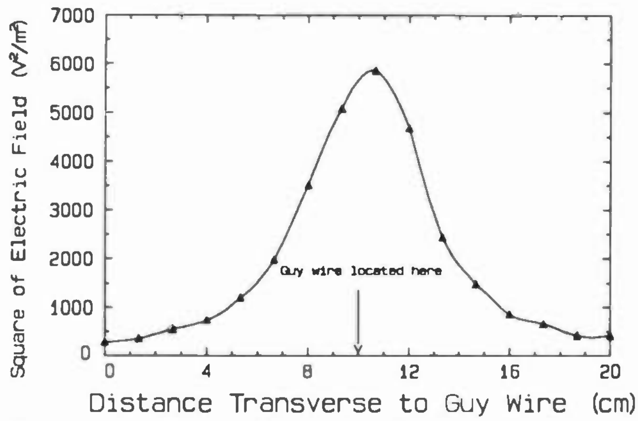

Figure 13. Spatial variability of electric field strength (E^2) near a simulated guy wire exposed to RF fields.

Contact Current Distribution

Empirically, it was observed that as the current probe assembly was brought within 10 cm of the half-wave dipole antenna that served as a reradiating object, there was a very slight increase in the induced current but upon contact, the measured current abruptly and very substantially increased above the non-contact condition. This observation led to the insight that contact currents were, for the situations explored in this study of RF hot spots, always greater than any other current which might be induced in the body. A question remained, however, as to whether the contact current as measured with the current probe assembly constructed for the project, might be less than the current at some other point within the body. Though such might be the case, intuition argued that the local current densities, which the SAR is related to, would generally be substantially less due to the much large tissue cross sections involved in other parts of the body. Since the wrist represents the smallest cross-sectional area in which the contact current can practically flow, the wrist SAR appears to be the determining factor for analyzing contact currents. To investigate the possibility that the current that flows on contact might not be maximum at the contact area, the spatial distribution of the current was examined by sliding the current transformer up and down the saline filled pipe described earlier. The finding was that the measured current was a maximum at the very top of the fluid filled pipe and decreased smoothly as the distance from the contact point was increased with no indication of relative peaks in the measured current. It was concluded, therefore, that in the contact situation of a body contacting a reradiating object, the measurement of current at the point of contact represents the greatest current that can be experienced. This finding is not what one would expect for the distribution of body current induced without physical contact with an energized object since the body will then tend to exhibit antenna-like characteristics resulting in possible current maxima at other points on the body.

DISCUSSION OF RESULTS AND INSIGHTS OBTAINED

RF Hot-spot Fields

The data presented in Figures 11 and 12 clearly reinforce common experiences of detecting enhanced RF fields near the surface of objects which act as reradiators. Both Figures 11 and 12 show localized fields that are well in excess of the ANSI RFPG while the nearby ambient field is only about 10 percent of the recommended limits in terms of field strengths. The obvious question one might encounter upon evaluating this exposure situation is whether the fields mean that human exposure would violate the restrictions imposed by the RFPG. This example case shows that the power absorbed by the dipole, under the conditions which led to the high surface fields, was far less than one watt. Hence, <u>on the basis of whole-body SAR, actual contact with the reradiating antenna can not lead to excessive energy absorption rates.</u> To exceed the whole-body SAR value specified in the ANSI RFPG, an average individual of 70 kg mass would have to have available 28 watts to reach the whole-body SAR limit of 0.4 W/kg, assuming that the power were actually all absorbed by the body.

The data for the resonant dipole show that the incident electric field strength would have to be 20756 V^2/m^2, or the equivalent of a plane wave having a power density of 5.5 mW/cm^2, before the absorbed power would reach 28 W; in this case, the whole body SAR would have already exceeded the limits of the RFPG on the basis of field exposure over the whole body, let alone touching the reradiating object. The point to be made here, however, is that the experimental results suggest that typical RF hot-spot situations encountered in RF surveys, i.e., the existence of very localized high fields well above the surrounding values, unless the ambient values are already above the RFPG field strength limits, cannot result in <u>whole-body</u> SARs exceeding the RFPG. This observation has a strong bearing on interpreting partial body exposure conditions caused by reradiating objects; if typical exposure levels are found to be less than the RFPG, high fields caused by reradiating structures do not appear to have the capacity to produce <u>whole-body SARs</u> in excess of the RFPG. Thus, commonly encountered localized high fields caused by such objects as clothes lines, curtain rods, guy wires, metal window frames, metallic office furniture, etc., are generally only sources of near field, partial body exposure and consequently are not good indicators of <u>whole-body</u> energy absorption rates.

The more pertinent issue, then, is <u>local SAR</u> in the body tissues and whether these highly localized fields can lead to spatial peak SARs in excess of the limits in the RFPG. Except for a very limited number of possible reradiators, for which their interactions with RF fields are well understood and for which absorbed powers can be readily measured, some means of determining local SAR is required. Lacking commercially available and socially acceptable methods for directly measuring SAR in exposed humans, surrogate measures must be implemented. Experimental approaches, making use of phantom, tissue equivalent models of the human body, are generally not easily adaptable to the rugged field environment in which such assessments are required.

The measurement of contact currents was deemed to represent a more practical method of determining compliance with the intent of the ANSI RFPG. This is so since the contact currents are greatest at the point of contact, the current flowing at the hand contact is very close to the anatomical area of the body in which the highest local SAR will likely occur (the wrist because of its small cross-sectional area) and the instrumentation required is portable and permits relatively straight forward measurements in the field, even in relatively awkward situations, such as on antenna towers.

Contact Currents and Local SAR

As described earlier, the local SAR can be determined by relation [4]. For purposes of discussing the contact current data developed in this project, the literature was reviewed for information on conductivity of muscle tissue in humans. Data compiled by Durney et al. (1986) were ultimately used the analysis presented here. Muscle conductivities for excised human muscle tissue were used to develop a least squares polynomial fit of the data of the form:

$$s = A_0 + A_1F + A_2F^2 + A_3F^3 \qquad [7]$$

where

s = tissue conductivity (S/m);
F = frequency (MHz);
A_0 = 0.564;
A_1 = 0.00266;
A_2 = -6.79×10^{-6};
A_3 = 6.31×10^{-9};

Relation [7] should be valid for the frequency range of 1 MHz to 800 MHz. Relation [7] was used to generate values of muscle conductivity for selected broadcast bands. In addition, relation [4] was manipulated to obtain expressions for calculating values for the peak SAR that would be expected to occur in the wrist of an adult touching an RF exposed object which delivered a contact current. These simplified expressions are of the form:

$$SAR = K I^2 \qquad [8]$$

where

SAR = spatial peak SAR in the wrist (W/kg), with assumption of an effective conductive cross section of 11.1 cm^2;

K = a conversion factor specific to the frequency of interest;

I = contact current (mA);

Table 1 lists the conversion factors appropriate to various broadcast frequencies to use when applying relation [8]. For broadcast band designations, the expressions obtained from the listed values of K will result in a conservative estimate of the wrist SAR over the particular broadcast band since the lower edge of each band was used to find the conductivity, i.e., the expression will not underestimate the expected SAR.

Table 1. Proportionality factors for use in relation [8] used to compute the peak SAR in the wrist based on the measured contact current for selected broadcast bands and limiting contact currents to control peak SAR in the wrist to less than 8 W/kg and 20 W/kg.

Broadcast band	K used in [8]	Limiting current to control SAR (mA) $\leq$8 W/kg	$\leq$20 W/kg
AM	0.00142	75.1	119
Low VHF (54-88 MHz)	0.00113	84.1	133
FM (88-108 MHz)	0.00105	87.3	138
High VHF (176-216 MHz)	0.000913	93.6	148
Channel 14	0.000805	99.7	158
Channel 20	0.000789	100	159
Channel 66	0.000517	124	197

As an example, at 144 MHz, the expression for local, wrist SAR, based on an interpolated value of conductivity of 0.824 S/m, becomes

$$SAR = 0.000950 I^2. \qquad [9]$$

The expressions given by [8] must be understood to not be precise limits since there is considerable variability in the measured data on tissue conductivities. The conductivity data used represent mean values with a spread in the conductivity values varying from about 2 percent to as much as 20 percent.

Applying [9] to the data on contact currents obtained in the project permits immediate estimates of wrist SARs. Another approach, however, is to determine from expression [8] the contact current required such that the wrist SAR is equal to the peak SAR limit in the ANSI RFPG of 8 W/kg. This procedure was performed for the same frequencies and/or bands used in Table 1 to find the contact currents equivalent to a wrist SAR of 8 W/kg and 20 W/kg. The higher value of local SAR has been suggested by the IRPA as a more appropriate limit for the ankles and wrists. ANSI (ASC, 1989) has included this higher limit in a recent draft of its revised RFPG. Table 1 lists the limiting contact currents which would control the wrist SAR to values of 8 and 20 W/kg. The general trend observed that lower currents are necessary to deliver the same SARs at lower frequencies is due to the variation of muscle conductivity with frequency; the conductivity decreases at lower frequencies. At 144 MHz, a contact current of 91.8 mA is projected to result in a wrist SAR of 8 W/kg while a current of 145 mA would be required to result in a SAR of 20 W/kg.

An important observation to make from the limiting currents in Table 1 for wrist SARs of 20 W/kg is that these values exceed those currents sometimes associated with RF burns. Rogers (1981) has recommended a current threshold of approximately 200 mA for the onset of RF burns. More recently, a value of 100 mA has been assumed to represent the threshold for RF burns for small contact areas during the grasping of an energized source (ASC, 1989). Thus, while maintaining wrist SARs within the 20 W/kg value, the contact currents could lead to RF burns upon touching the object and for practical purposes, unless protective gloves are used, the upper limit becomes essentially 100 mA rather than the listed higher currents.

The contact currents measured in the project were analyzed in terms of these threshold currents for SARs of 8 and 20 W/kg in the wrist. For the resonant dipole, the mailbox, the window frame and the filing cabinet, the ratio of contact current to ambient E^2 was calculated and the incident electric field strength (E^2) required to produce a contact current equivalent to 8 or 20 W/kg was calculated. The results are given in Table 2. These results show, not surprisingly, that the resonant dipole requires the least strong incident field strength to produce contact currents sufficient to deliver the limiting values of SAR. But an interesting observation is that even apparently non-resonant objects, like the filing cabinet, can produce contact currents associated with wrist SARs of 8 W/kg, or even 20 W/kg, in ambient fields considerably less than the field strengths given in the RFPG. For example, in a field of 1300 V^2/m^2 the filing cabinet would produce a contact current of 92 mA, sufficient to result in a wrist SAR of 8 W/kg. Thus, even though the ambient field strengths may be less than those prescribed to limit whole-body SAR to less than 0.4 W/kg, contact currents in some exposed objects can exceed the level that would result in local SARs greater than 8 W/kg.

Table 2. Summary of contact currents and required ambient fields to cause peak SARs of 8 W/kg and 20 W/kg in the wrist for several reradiating objects.

Object	Ambient electric field strength squared (V^2/m^2)	Contact current (mA)	Ambient field (V^2/m^2) required to produce peak SAR in wrist = 8 W/kg	= 20 W/kg
Dipole	370	60.9	556	878
Mail box	620	88.6	642	1014
Window frame	1500	133	1035	1635
Filing cabinet	600	42.4	1298	2051
Guy wire*				

* Data are not given for the simulated guy wire since it is difficult to quantify the actual exposure field. While one end of the wire was exposed to a rather intense RF field, the other end was in a very low intensity ambient field; thus, the exposure conditions were substantially different in comparison to the other listed objects.

Most RF surveys designed to determine compliance with the ANSI RFPG do not investigate the issue of contact currents and in many circumstances such an omission is likely not important since the SAR limits of the RFPG are to be time-averaged over any 6-minute period. In a practical sense, it is unlikely, except for certain extenuating circumstances, that individuals will be continuously contacting objects which are linked to RF hot spots. For example, even in the case of a metal stair way hand rail, the hand is making and breaking contact periodically with the rail, thereby reducing the time-averaged contact currents and thereby reducing the local SAR. In a similar sense, the enhanced RF fields that might exist near the end of a metal curtain rod and associated contact currents do not necessarily represent real problems since people do not normally remain in direct contact with curtain rods, other than momentarily. Nevertheless, consideration should be given to the possibility that individuals may come into contact with RF hot-spot objects, if such is practically feasible.

The time-averaging provision of the ANSI RFPG can be stated, for the VHF frequency band, as:

$$\text{Peak SAR} \leq 48 \text{ W-min/kg} \qquad [10]$$

Relation [10] simply says that the product of the peak SAR and the exposure time in minutes should be no more than 48 W-min/kg in any 6-minute period. Thus, limited duration exposures for less than six minutes may exceed the 8 W/kg value, if, on the average, the SAR does not exceed 8 W/kg. This provision of the RFPG provides for limited flexibility in managing exposures where RF hot spots may be present. For example, it is conceivable that the time averaged contact currents could be determined either by direct measurement using data logger type instruments or by timing of repetitive contacts as when climbing a tower ladder. In any event, excursions of the peak SAR above 8 W/kg should not be interpreted as necessarily exceeding the basis for the ANSI RFPG; careful assessment of the time-averaged peak SAR is called for.

The observation that wearing of shoes by the person contacting a reradiating object made little, if any, difference in the measured contact current is presumed to be related to two factors: (1) for contact situations, the current magnitude dissipates as the distance from the point of contact increases and (2) the capacitive reactance of the shoes becomes significantly less at higher frequencies, thereby apparently reducing the impedance to current afforded by the shoes. Thus any current escaping from the body via the feet, aside from that leaking off the body surface due to capacitive coupling to the nearby environment, meets with little change in impedance, whether the person is wearing shoes or not.

CONCLUSIONS

RF hot spots are a natural consequence any time that RF fields exist in the presence of conductive objects. RF hot spots are not unique, nor mysterious sources of RF fields. This study of RF hot spots has focused on developing information which may be helpful in interpreting the significance of hot spots when determining compliance with the ANSI RFPG (ANSI, 1982) at broadcast sites. The measurements and calculations have been directed to applications in the VHF and UHF broadcast bands but the concepts are applicable to assessing RF hot spots near AM stations as well.

A RF hot spot may be defined as a point or small area in which the local values of electric and/or magnetic field strengths are significantly elevated above the typical ambient field levels and often are confined near the surface of a conductive object. RF hot spots usually complicate the process of evaluating compliance with the RFPG because it is often only at the sources of the hot spots that fields exceed the RFPG.

RF hot spots may be produced by an intersection of narrow beams of RF energy (directional antennas), by the reflection of fields from conductive surfaces (standing waves) or by induced currents flowing in conductive objects exposed to ambient RF fields (reradiation). RF hot spots are characterized by very rapid spatial variation of the fields and typically result in partial body exposures of individuals near the hot spots. Uniform exposure of the body is essentially impossible due to the high spatial gradient of the fields associated with RF hot spots.

Several conclusions relevant to the RFPG compliance issue have been drawn from the results and experiences of this investigation:

(1) The ANSI RFPG (ANSI, 1982) is based on limiting the time-averaged SAR in the body. Field strength (or power density) limits specified in the RFPG represent working levels for determining compliance with the underlying basis of the standard. In lieu of more definitive measurements or analysis which relate to the SARs caused by RF exposure, the field limits in the RFPG should be used for determining compliance.

(2) In the typical RF hot-spot scenario, involving reradiating objects, the high, localized fields at the hot spot do not generally have the capacity to deliver whole-body SARs to exposed individuals in excess of the RFPG, regardless of the enhanced field magnitude. When the ambient RF field strengths are already at or exceeding the RFPG, then the partial body exposure which accompanies proximity of the body to the object will increase the whole-body SAR, generally only slightly.

(3) The high-intensity electric and magnetic fields accompanying RF hot spots are not good indicators of whole-body or spatial peak SARs in the body due to the high variability in coupling between the body of an exposed person and the hot-spot source.

(4) A measurement of the contact current which flows between the exposed person and a reradiating object provides a meaningful alternative to field measurements and permits the evaluation of the peak SAR which may exist in a person touching the hot-spot source.

(5) For most practical exposure situations, when hand contact is made with a RF hot-spot source, the greatest RF current will flow in the body, resulting in the worst case scenario for peak SAR. The contact case will result in significantly greater local SARs than for the non-contact condition and should be assumed to be the exposure of possible concern. This maximum SAR will be in the wrist, the anatomical structure with the smallest cross-sectional area through which the contact current can flow.

(6) Determining the wrist SAR for contact conditions requires a measurement of the contact current, knowledge of the conductivity of the tissues and knowledge of the effective conductive cross-sectional area.

(7) For most practical field evaluations, simple approaches to the measurement problem and necessary calculations to obtain the peak SAR are available and have been outlined in this paper. This conclusion is at variance with section 6.11 of the present ANSI RFPG (ANSI, 1982) in which ANSI, in referring to the 8 W/kg peak SAR limit, declared "It was also recognized by the subcommittee that to determine whether a particular RF source would meet these absorption criteria would be difficult and could be done only by a properly qualified laboratory or by an appropriate scientific body for a general class of equipment. In no case could a routine field survey determine conformance with the criteria of this part of the exclusion." These cautionary words were prepared in 1979, prior to the extensive work on induced and contact current measurements which has developed since 1979. At that time, recognition of whole-body resonance was the major new change in the RFPG over previous versions of the guide and the statement in section 6.11 is presumed to have referred principally to the issue of deep heating of internal body tissues. Even today, for such tissues, the dosimetric procedures required for accurate SAR assessments remain complex and are relegated, for many cases, to the laboratory setting. In the case of peak SARs incurred from exposure to RF hot spots, however, it must be concluded that section 6.11 of the RFPG was written in an overly narrow context relative to the ability for in-the-field contact current measurements and assessments of wrist SARs.

(8) Results obtained with those objects investigated in this study showed that when ambient RF electric fields were in the range of 16 to

32 percent of the RFPG (for field strength squared) for the frequency range 30-300 MHz, contact currents in unprotected individuals could reach a level that is associated with a wrist SAR in adults equal to 8 W/kg. Electric fields that would induce contact currents sufficient to deliver wrist SARs of 20 W/kg, the peak SAR limit recommended by IRPA and the NRPB, range from 22 to 51 percent of the RFPG field limits.

(9) Application of the 8 W/kg peak SAR limit to the extremities of the body by ANSI (ANSI, 1982) appears to have been a matter of convenience and simplicity when the RFPG (ANSI, 1982) was developed. Current revision of the RFPG (ASC, 1989) reflects acceptance of the higher peak SAR value of 20 W/kg for the extremities (ankles, feet, wrists, hands).

(10) In many RF hot-spot cases, an ambient field limit comparable to approximately one-fifth of the present field limits at VHF of the ANSI RFPG would likely protect against contact currents capable of inducing a peak SAR in the wrist of 20 W/kg.

(11) Complex exposure environments, such as the interior of antenna towers, that present highly localized RF fields on climbing structures (ladders and tower members) are candidate locations where contact current measurements may prove effective in evaluating compliance with the RFPG.

(12) Contact current measurements appear the only practical avenue of evaluating RF hot spots found in public environments where ambient field levels are usually well within the RFPG but local fields are apparently excessive.

(13) Maximum contact currents are associated with those points on a conducting object which generally exhibit the greatest surface magnetic field strengths. This is apparently because such points represent relatively low impedance points on the structure and current transfer is most effective when contacted by the relatively low impedance of the human body.

(14) Use of protective clothing in the form of insulative gloves will likely provide limited protection in the case of worker exposures but this means of controlling peak SARs in the body needs further evaluation in the VHF band.

(15) When evaluating an area for compliance with the RFPG, field measurements on the surface of objects are useful for locating the presence of hot spots and determining points at which subsequent contact currents should be measured; the indicated field strengths (power density) at these surfaces provide no useful information on the potential for high SARs in exposed persons. Separation distances between the survey probe and the RF source should be selected on the basis of providing good indications of exposure over the entire body; 20 cm is a more meaningful distance than the minimum distance specified in the present ANSI RFPG (ANSI, 1982). Where exposure may exist at distances closer than 20 cm to the apparent source of RF fields, other dosimetric methods should be employed to evaluate the exposure for compliance with the RFPG. Contact current measurements appear to represent one good surrogate dosimetric measure.

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FIELD TESTING OF A SHORTENED EBS ALERT TONE

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ABSTRACT

This paper reports on the design and implementation of a field test authorized by the FCC to determine the effects of utilization of a shortened EBS tone length on the reliability of the present EBS system.

A BRIEF EBS HISTORY

The Emergency Broadcast System (EBS) has been designed to provide the President and other Federal Officials a means of communication to the general public during periods of emergency using the nation's broadcast stations. As an improvement of the CONELRAD system of the 1950's which designated only selected stations operating at 640 and 1240 kilohertz, EBS now allows dissemination of emergency information on about 95% of the TV and radio stations in the country that elect to participate in the system.

Inherent in the design of the system is a star network monitoring scheme, in which stations either receive Emergency Action Notification messages via radio or TV networks, news wire services, or via monitoring of another broadcast facility. Since continuous aural monitoring of such broadcast signals is impractical, an alert system was needed. Until 1974, this was a simple carrier interruption scheme. Loss of carrier from the monitored station caused a receiver to demute, and thus the EBS message was monitored. Although the carrier interruption method was simple, it's reliability was less than optimum. A brief loss of received carrier, such as might occur during an electrical storm or a power loss at the monitored station, resulted in false demuting of the monitoring stations' receivers. In 1975, the FCC adopted a two-tone attention signal to replace the carrier interruption system. Per 47 CFR 73.906 (a) two precisely controlled tones, 853 hertz and 960 hertz, ±0.5 hz are to be used. The two tones, combined within 1.0 dB of each other, shall be transmitted for a period of not less than 20 seconds or more than 25 seconds. Each shall modulate the transmitter at not less than 40%. (CFR 47 73.906 (c)(d)). The two-tone system has proven to be an effective alert signal to the listener as well as a reliable means of activation of EBS decoders.

USE OF EBS TODAY

National

The use of EBS has expanded greatly since it's inception in the 1950's. At the national level, it's primary function remains as a means of delivery of emergency communication to the general public for live presidential messages, and other official information sources. In fact, the system is currently undergoing restructuring to make it even more effective.

State

The important role that EBS can play at the state level has been developed more recently. Operational plans are now in effect for all fifty states which break each state into operational areas, each served by one or more common program control stations (CPCS). Emergency situations that affect an entire state or portion of a state and which may extend across several operational areas are coordinated through the State Emergency Operating Center. Therefore, an entry point to EBS must be available in each state so that dissemination of emergency information to the state's residents can be handled quickly and efficiently. The great importance of EBS at this level has been amply demonstrated recently in areas that have suffered trauma from earthquakes, hurricanes, nuclear plant mishaps or other natural or man-made disasters. Depending upon the structure of the plan for the state, EBS transmission paths may mirror those of the national plan or may be developed separately using a different set of stations.

Local

At the local level, the use of EBS has greatly expanded. It is used most commonly used for emergencies that threaten life or property and that affect localities, counties, or multiple counties. Emergencies such as severe storms, tornadoes, toxic waste accidents, ice storms, blizzards, and other localized emergencies affecting smaller geographic areas are its common use.

THE EBS CHALLENGE

The effective use of EBS for any level of activation requires:

- public awareness of what the system is intended to do
- knowing what the alert tones signify
- having effective plans in place in each EBS operational area and state as well as the nation
- harmonious cooperation among broadcasters to promote the system, use it, maintain its integrity and to take it seriously.

However, while desiring to provide service to the public, broadcasters are simultaneously faced with continual, and in some markets, relentless competition. In such a competitive environment, programmers must place each element of programming under the microscope. In the case of radio, their decisions must consider the flow and selection of music, the right air talent, results of focus group testing and research, the proper promotion, and even extend to the selection and adjustment of the audio processing. All this effort is aimed at one goal: eliminating tuneouts, increasing cumes, quarter hours, and hopefully, profits.

With this increased emphasis on providing the listener with no reason to tune away from their station, programmers have voiced their belief that the transmission of 20-25 seconds of atonal EBS alert tones provides some listeners a reason to "hit the button" and find another station. Whether this concern is real or imagined or to what extent it actually occurs, would be very difficult to accurately measure. What really matters is that programmers feel it does exist. Therefore, they may make the decision that the EBS will not be used except for the required weekly EBS test and for that fateful national level activation, should it ever be required.

To be sure, the length of the alert tones is not the only challenge facing the continued effectiveness of the Emergency Broadcast System. With the implementation of more music intensive formats, reduction of the time that many stations wish to commit to news and public service programming and increasing reliance on formats delivered from off premise sources such as tape or satellite, the effective use of EBS may be on the decline. Additionally, with some stations facing tighter budgets, frequent employee turnover and reduction in staff sizes, maintaining staff awareness and proper operating procedures related to EBS may become more difficult.

Therefore, if the system can be made more "user friendly" while maintaining it's purposes, it would seem advisable to try.

THE MICHIGAN FIELD TEST

In response to numerous contacts from broadcasters in Michigan, it became apparent to the State chairman that the condition of EBS in the State was in need of review and perhaps some revision. The Michigan Association of Broadcasters was contacted and requested to assist in providing funding for research, clerical support and other support so that the concerns of Michigan broadcasters could be gauged. A telephone survey of all Michigan radio and television stations was conducted in the Spring of 1989. Results of the survey provided much useful input, and one common concern was heard quite often. Broadcasters felt that EBS served a definite public need but at times they were hesitant to use the system, especially with the alert tones, feeling that it could provide a reason for audience tuneout. In light of the results of this survey, a field test was proposed to the FCC for a six month period to determine the possible negative effects of reducing the length of the present EBS alert tones from the present 20-25 second length to a reduced length of time. The test would occur in the Lansing and Central Michigan EBS Operational Area and would include all

AM, FM and television stations that presently participate in that area. It was proposed that a post card verification system would be used to check the stations' receipt of the normal weekly test. It was proposed to the commission that the following stations reduce the length of the alert tones to approximately 8 seconds:

1. WILS (1320 Khz)
   Designation: CPCS-1
2. WKKP (101.7 Mhz) Designation: State Originating Primary Relay
3. WITL-FM (100.7 Mhz)
   Designation: Primary

Further requested was permission be granted to change receiver demute times for all stations in the operational area as well as the following stations outside the operational area:

1. WOOD-FM (105.7 Mhz) in Grand Rapids, Michigan. Designation: State Primary Relay & CPCS-1

2. WJR (760 khz) Detroit, Michigan. Designation: State Primary Relay & CPCS-1

No changes in tone frequencies, tone level balance, or minimum modulation levels were proposed. Permission to conduct the test was granted on August 8, 1989 to commence October 1, 1989 and to conclude March 31, 1990. Results of the test to be presented to the commission in May, 1990.

Operational Area Profile

The Lansing and Central Michigan Operational Area consists of 7 contiguous counties, located in the south central lower portion of Michigan including the State capital of Lansing. 32 stations operate in this area, 14 AM, 14 FM and 4 television. Since the State Emergency Operating Center is also located in Lansing, it was decided it would be prudent to include 2 more distant relay stations who will monitor a newly designated Originating Primary Relay station (WITL-FM) upon completion of a revision of the State plan anticipated to be implemented mid to late 1990. Hence, WOOD-FM in Grand Rapids was included in the test as was WJR in Detroit. It was felt that inclusion of these stations would provide additional test data on the viability of such shortened tones under distant reception conditions. Of the three stations who were authorized to broadcast shortened EBS length tones, WILS is an AM (5 kw DA-D & 1 kw DA-N), WKKP is a Class A FM, and WITL-FM is a Class B FM. Additionally, the three stations vary in music formats. One is an Urban station, while another is a AC/Gold format and another is Country. All three stations use varying degrees of audio processing.

Equipment Modifications

An equipment survey was conducted in the Operational Area to determine the makes and models of EBS encoding and decoding in use at all stations. The following makes and models are presently in use:

1. Time and Frequency Technology (TFT), model 760
2. McMartin, model EBS-2
3. Gorman Redlich, model CEB

Through cooperation with SBE Chapter #91 (Central Michigan), several engineers investigated the feasibility of circuit modifications to the units in service in the operational area to reduce the decoder demute time and encoder tone length time.

Selection of modification components were based on achievement of the following times:

DEMUTE = NO GREATER THAN 4 SECONDS
ENCODE = NO GREATER THAN 9 SECONDS OR LESS THAN 8

Selection of these times were based upon the following criteria:

A shortened tone length should still provide a noticeable interruption and distinct change from regular programming even to a listener who may not be actively listening or viewing.

Additionally, in response to the results of the survey, it was the intent to reduce the tone length by a factor significant enough that the temptation to change the station, via pre-set tuning buttons in autos or by readily accessible TV remote controls, was reduced. A reduction of the tone length by 2/3 was felt to be a fair compromise.

Selection of an appropriate time length for demute was based upon a balance between a time long enough to prevent false receiver demuting from program material versus a time short enough to provide an ample attention signal to the receiving station's operating staff. Additionally, if external alarm systems, warning lights or automatic recording equipment were employed at the receiving station, such equipment would have time to function and alert station staff before the actual transmission of the emergency message.

Decoder Modification

Investigation of schematics of all three manufacturers' units in use in the area revealed each used a fairly similar circuit design. How this is accomplished is shown in Figure 1.

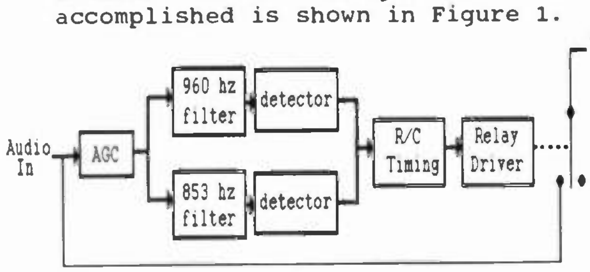

Figure 1

To change each decoder's demute time to the desired four second delay, required changing a capacitor in the R/C circuit.

	Comp ID	Orig	New
TFT 760	C13	100uf	33uf
McMartin	C24	220uf	68uf
GR CEB	D3&D4	47uf	22uf

Encoder Modification

Circuit designs for tone encoders varied among the three manufacturers' units. The Gorman Redlich and the McMartin encoders both use a simple R/C timing circuit. Length of the tone encode period was able to be adjusted to the desired length by the replacement of a single capacitor. Figure 2 shows a block diagram of the circuit configuration for these two units.

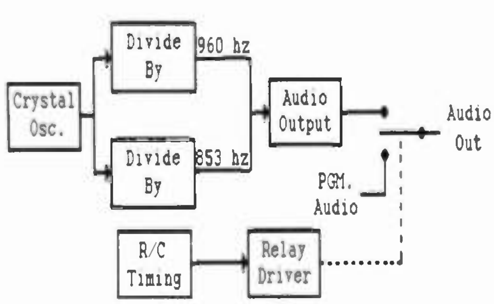

Figure 2

The encoder timing circuit of the TFT 760 utilizes a digital countdown configuration. As can be seen in Figure 3, using the already available frequency of 853 hertz as its reference, an array of four 7490 and one 7476 ICs are used to divide the reference by 20,000. This generates the needed 23.4 second pulse for activation of the tone output relay.

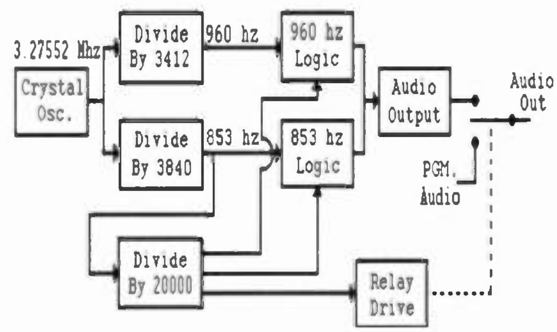

Figure 3

To achieve the desired tone length of approximately 8 seconds, it was necessary to change the divide-by array to a new value of 7000. Figure 4 reflects this change.

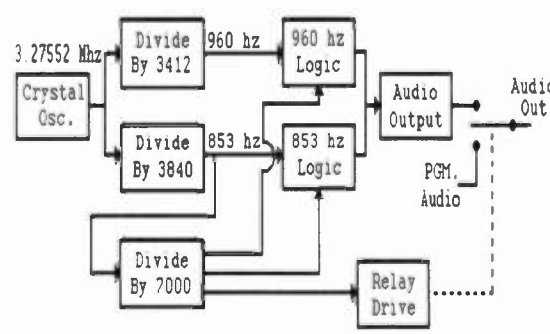

Figure 4

This modification can be accomplished without changing any components, but rather simply by cutting six circuit board traces and adding five jumpers to two 7490 ICs (Z6 and Z7). This modification is shown in Figure 5.

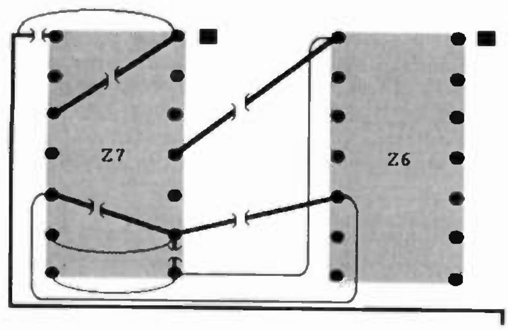

Figure 5

Equipment Modification Workshop
Through a cooperative effort with the SBE, an equipment modification workshop was held. Chief engineers of all Operational Area television and radio stations were requested to bring in their equipment for modification during the workshop. The purpose and scope of all modifications were outlined to the engineers. For those few stations who were not able to attend, modification schematics and required parts were mailed. All parts were supplied at no charge to the stations. A test bench was set up to handle the modifications and to then verify the performance of each station's decoder using the test setup shown in Figure 6.

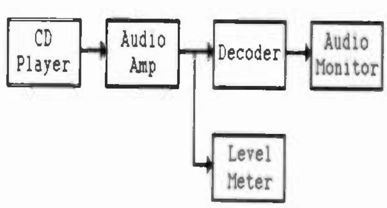

Figure 6

Proper decoder performance was verified using the appropriate track from the "NAB BROADCAST AND AUDIO SYSTEM TEST CD" and track timing indication from the CD player. All decoders so modified displayed demute times of not less than 2.6 seconds nor more than 4.0 seconds. Variations from the 4.0 second projected demute time were attributed to component tolerance variations in the time constant portion of the circuits in the various units. A demute time of less than 4.0 seconds was not deemed a serious problem for purposes of the field test. However, should a time of exactly 4.0 seconds be desired, aged components which may have drifted out of tolerance might have to be replaced with new and tighter tolerance components in this portion of the circuit.

To verify the performance of each station's encoder, the test setup as shown in Figure 7 was used. Center frequency of the two tones were checked as well as the level balance between the two tones to a tolerance of +/- 0.1 DB was verified. For the encoders that were modified to a shortened tone length for purposes of this test, the length of the tone was verified using an electronic timer.

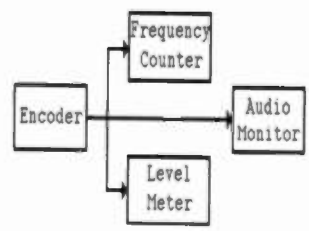

Figure 7

RESULTS OF THE FIELD TEST
A test evaluation system was designed to provide weekly feedback by each station utilizing a pre-posted and addressed postcard. Weekly performance of each station was tallied by the author noting day and time the test was received as well as any equipment problems that may have occurred or false trips of the station's decoder. While the field test is ongoing at the time of submission of this paper, results to date are presented in Figure 8.

Test Variations
As indicated on Figure 8, most stations indicated successful reception of the weekly test. In a few cases, equipment problems were discovered early in the test period. Feedback from the stations indicated such problems were not related to the modified demute circuits, but with external antennas, decoder input overload due to improper level setting, and other component failures.

No station has reported false demuting of their EBS receiver during regular programming by their monitored station.

LANSING AND CENTRAL MICHIGAN OPERATIONAL AREA - EBS FIELD TEST - DATA SUMMARY INTERIM REPORT AS OF 1/25/90

	October 1989					November 1989				December 1989					January 1990				February 1990				March 1990				
Call	1	8	15	22	29	5	12	19	26	3	10	17	24	31	7	14	21	28	4	11	18	25	4	11	18	25	MNTR
WDBM	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK											A
WFMK	OK		OK	OK	OK	OK	OK	OK	OK	OK		OK	OK	OK	OK	OK	OK										B
WFYC	OK	OK	1	1	1	1	1	1	1	1	1	OK	OK	OK	OK	OK	OK										A
WGOR	OK	OK	OK	OK		OK	OK	OK	OK	OK	OK	OK	OK	OK	OK		OK										A
WHMI	NO	NO	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK		OK	OK	OK										A
WION	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										B
WILX	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WITL	OK	OK	OK	OK	OK	OK	NO	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WJR	2	2	2	2	2	2	2	2	2	2	2	2	2	2	2	2	2										C
WJIM	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WKAR	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WKAR	OK	OK	NO	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WLNS	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WMLM	NO	OK	NO	OK	NO	NO	NO	OK	OK	OK	NO	OK	NO	OK	OK	OK	OK										A
WMMQ	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WNLF	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WOAP	OK	OK	OK	OK	OK	OK	OK	OK	NO	OK	1	1	1	OK	OK	OK	OK										C
WOES	1	1	1	OK	NO	NO	OK	3	OK	NO	OK	OK	3	3	OK	OK	OK										A
WOOD	OK	OK	OK	NO	OK	OK	OK	NO	OK	OK	OK	OK	OK	OK	OK	OK	OK										C
WSYM	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WUNN	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WVIC	OK	OK	NO	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WWSJ	OK	OK	OK	OK	OK	OK	OK	NO	OK	OK	OK	OK	OK	OK	OK	OK	OK										A
WXLA	OK	OK		OK		OK	OK		OK			OK	OK		OK												A

OK= Test Received
NO= Test Not Received or Reported
blank= No Verification Card Received

1= Equipment Problem/Under Repair
2= Equipment Ordered, Not Installed
3= Station off air (Educational)

EBS Monitor: A= WILS (1320 Khz)
B= WKKP (101.7 Mhz)
C= WITL (100.7 Mhz)

Figure 8

It should be noted that due to delays in equipment delivery and inhospitable conditions for outside antenna installation, WJR, Detroit was not able to participate in the test until January 28, 1990.

CONCLUSION

1. The preliminary results of this field test indicate adoption of a shortened EBS tone length of approximately 8 seconds and a receiver demute time of approximately 4 seconds seems to have no detrimental effect on the system's reliability despite variations in musical formats, whether the monitored station is AM or FM, or reception is local versus distant.

2. Modification of existing FCC rules allowing stations to select a length from not less than 8 seconds to the present 25 seconds limit would seem to be a prudent and flexible enhancement to the EBS system. With adoption of such new standards, the FCC, NAB and other interested parties should continue to encourage the use of EBS for local and state emergencies as well as its continuing national role.

3. For the purpose of these tests, relatively simple and low-cost component changes were made to accomplish the shortened tone generation and demute times. Should a shortened tone and demute specification be adopted, requests should be made to manufacturers of current units in the field to supply their specific recommendations for circuit modifications to accomplish these new standards on their units.

4. Manufacturers of new models of EBS units might wish to make the length of the encode signal and demute time able to be field selected by the user within the FCC mandated limits.

ACKNOWLEDGEMENTS

The author wishes to thank the Michigan Association of Broadcasters, its Executive Board and Executive Director, Ms. Karole White, for their assistance in funding this test. Additionally, he gratefully thanks Mr. Mike Winsky, Chairman of SBE Chapter 91, for his work on the circuit modification designs and his many contributions to the preparation of this paper. He also wishes to thank Mr. James White, General Manager of WOOD/WOOD-FM, Grand Rapids, Michigan, for his continuing encouragement of EBS improvements.

AUTHOR

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NEW TOWER CONSTRUCTION TECHNIQUES

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INTRODUCTION

This paper presents the design, fabrication and erection sequence for the new 600-foot KGON Radio Tower, located in Portland, Oregon. The 600-foot, freestanding tower consists of a composite, high-strength concrete and steel tube space frame. The tower configuration is a basic three-dimensional verandeel truss space frame with 54 members. The triangular tower face is 75 feet at the base and 20 feet at the top of the tower. The antenna on top of the tower is approximately 91 feet long, weighing 13 tons for multiple (6) FM radio stations. The tower has eight platforms for various tenant antenna and operational equipment.

This paper will discuss the design report, design development, public involvement, tower layout, fabrication, erection, operational equipment, material specifications, and quality control procedures followed.

DESIGN REPORT

During the spring of 1987, a wind and icing study was performed for the owner, KGON Radio, of Portland, Oregon to establish the design parameters for a 400, 500 or 600-foot freestanding structure. The structure site is located in the neighborhood of Healy Heights, Portland, Oregon with an average mean sea level elevation of 1,020 feet. The design report included determination of wind speeds for the greater Portland area, using the National Weather Service hourly records for wind speed, temperature, humidity and wind direction. A minimum of 250,000 hours or 25 years of data were used for the statistical analysis. The fastest mile wind speed of 90 miles per hour was recommended for the tower design. The statistical analysis showed that a 3-1/2-inch radial ice build-up would be used for the tower members and equipment, along with an associated fastest mile wind speed of 80 miles per hour.

Along with the wind and icing criteria, a mathematical model of the tower site was made to determine the hill speed-up effects of wind approaching the tower from any direction. The results showed that a hill speed-up of 30% at the tower base and 10% at the antenna would be used to increase the design wind speeds. The preliminary report recommended that a full, static and dynamic analysis be performed for the tower design.

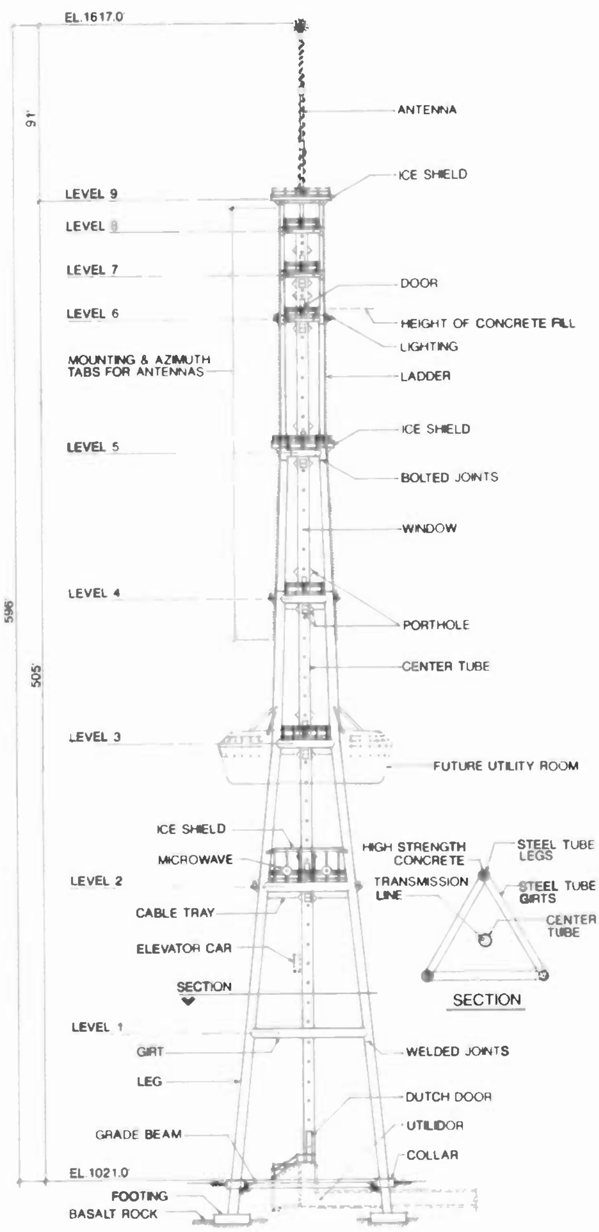

FIGURE 1

The preliminary design report recommended that a tower with adequate stiffness be provided to support the 13-ton, 91-foot antenna at the top of the tower for a maximum rotation of 1/2 degree with a fastest mile wind speed of 55 miles per hour.

DESIGN DEVELOPMENT

The design development of the new KGON tower was started in the spring of 1988, using the above design criteria to determine the final configuration of the tower structure. The design procedure involved the use of Skilling Ward Magnusson Barkshire Inc.'s (SWMB) computer STRUDL program and wind tunnel criteria to obtain the static and dynamic loads. The dynamic periods of vibration of the tower are 2.1, 1.8, and 1.2 seconds for the first three modes of vibration. The associated damping factors for each mode of vibration are 2%, 1% and 2%. The dynamic analysis includes the root mean square of the primary modes within the 20 STRUDL modes included for analysis.

The resulting structure configuration consisted of a high-strength, concrete-filled steel tube space frame. The final design of the tower required high-strength concrete in the steel tube leg and girts of at least 12,000 psi with a 14,000 psi laboratory design mix. The steel plate for the tube legs and girts consisted of 5/16" thick ASTM A588 Grade 50 material. After erection, the tower legs and girts are pumped with high-strength concrete to 450 feet above the base. The erection of steel and pumping of concrete was performed in two level (150 feet) increments to allow for curing and setting of the concrete. An initial set of 40 percent or 4,800 psi was established for sequencing of erection.

PUBLIC INVOLVEMENT

After completion of the preliminary design, a public involvement document was furnished to the owner for presentation of the proposed structure. The public involvement process included meetings with a local neighborhood association to inform the neighbors of the various issues involving the proposed new tower. The issues of radiation, interference, aesthetics, construction and operation were highlighted with use of presentation material at evening meetings in the neighborhood.

The neighborhood was given the opportunity to make a choice of which lighting system would be most appropriate on the tower. The result of this process is that the tower will be painted a conventional orange and white with lighting in accordance with FAA regulations. The public involvement process also included meetings with the City of Portland Planning Department to outline the scope and permit process of the new project.

After several community meetings, the owner was able to obtain the FCC Construction Permit and City of Portland Conditional Use Permit. This paved the way to proceed with the final design, construction documents and submittal for a building permit.

TOWER FOUNDATIONS

The foundations for the new tower were designed in accordance with the geotechnical study and recommendations, using a spread footing and grade beam system for each of the three legs of the tower. The spread footings are founded in a minimum of 3 feet of the basalt rock with an allowable bearing pressure of 5 tons per square foot. The design of the foundations resulted in spread footings of 18 feet square and 5 feet thick, located 20 feet below the finished grade of the site. The minimum concrete strength required was 8,000 psi. All reinforcement for the foundation is ASTM A615 Grade 60. The construction of the foundation included the use of steel cofferdams and backfilled with well compacted structural backfill (see Figure 1).

UTILIDOR

The tower structure is connected to the transmitter building with a 150-foot long underground utilidor. The utilidor is 9 feet wide and 8 feet high, built of reinforced concrete. Utilidor ceiling and walls have mounting inserts at every 4'-4'-1/2" to provide easy installation and access for transmission lines and conduit (see Figure 1).

TOWER FRAME LAYOUT

Final plans (see Figure 1) for the tower space frame included basic steel plate rolled tubes with steel studs, filled with high-strength concrete to form a composite structure. The steel tubes vary in diameter for the legs from 5'-0" to 2'-0". The horizontal members or girts located at each 75 foot level vary from 4'-6" to 1'-3" in diameter. The sizes are incrementally changed from the bottom to the top of the radio tower. The tower consists of nine levels with platforms at the top eight levels. The only level that does not contain a platform for operational equipment is at the bottom which is 75 feet from the base.

The connections for the tower legs and girts are welded in the bottom 300 feet and bolted using high-strength bolts in the remaining portion of the tower. The bolts are in accordance with ASTM A325 and vary in diameter from 5/8" to 1-1/2" in diameter. The use of load indicator washers was specified to assure that the tensions in the connections would be obtained in the shop and field.

Corrosion protection for the structure utilized A588 grade steel for the steel tubes, along with sandblasting to white metal and painting with three coats of polyurethane paint with a total minimum thickness of 8 mils. The miscellaneous steel for the tower is galvanized and painted with two coats and a minimum of 6 mils.

PLATFORMS

The owner requested that as much platform space as possible be provided on the tower to facilitate tenants, maintenance and operation. The platforms were framed between the girts with light structural steel beams of 10" to 12" deep and weighing from 10 to 50 lbs/linear ft. All platforms are furnished with railing and kickplates in accordance with OSHA requirements.

Also, the owner wanted the option of using platform grating of steel or dielectric material. After study during the design development, it was determined that the dielectric grating would be used in the majority of the platforms and that the steel grating would be used at the very top of the tower, just below the antenna. The platform railing posts were designed to mount TSL microwave antennas up to 6 feet in diameter. The platforms will have access by installation of an elevator on the outside of the center tube and interior fixed ladder system to facilitate maintenance and moving minor tools and equipment. A hoist beam system will be provided in the center tube of the tower to facilitate the erection, installation or modification of operational equipment.

CENTER TUBE

A 6-foot diameter steel center tube up the center of the tower was designed to enclose all transmission lines and electrical conduits between the transmitter building to various levels of the tower. The major feedline in the center tube is the 9-3/16" rigid coax that feeds the antenna at the top of the tower. A vertical ladder was included in the center tube to provide access and installation of various lines. The center tube includes brackets for mounting of transmission lines along with cable trays for conduits and cables. The center tube provides 8-inch diameter windows at approximately 12-foot spacing and sliding 2'-6" x 7'-0" door openings at each platform level.

Design of the center tube included the use of portholes below each platform that would serve the midpoint of each girt framing between the tower legs. The porthole lines will feed onto cable trays suspended under the platform framing to connect with the various antenna locations on the tower. In addition, six portholes were provided above the platforms for access to lines to the northeast and south legs of the tower.

The center tube is designed to be suspended in tension from two fixed points of the tower. The fixed points are located at the third level from the top of the tower and the mid level of the tower structure. The expansion and contraction of the center tube due to temperature changes will be in harmony with the transmission lines.

ELEVATOR SYSTEM

The tower will have an elevator track system mounted to the center tube to provide two-person car access from the underground utilidor tunnel to the top of the tower. The elevator will consist of a rack-and-pinion car system with controls to start and stop at any platform level. The elevator car will be parked underground in a separate utilidor room when not in use.

ANTENNA SYSTEMS

The main antenna on top of the tower is a Jampro Circular Polarized Antenna with a capacity for six FM radio stations. The tower will have space allocation for backup antennas, two-way radio, microwave and cellular antennas.

QUALITY CONTROL

Structural Steel: All fabrication was in accordance with the AISC specification. All structural rolled steel plates for the tower legs and girts were in accordance with ASTM A588 Grade 50. Mill test certificates and shop drawings were submitted to the engineer prior to fabrication. All miscellaneous framing steel for platforms, ice shields, etc., were in accordance with ASTM A36.

Welding: All welding of structural steel was in accordance with the American Welding Society (AWS) Specification D1.1-1986. All welds were prequalified with welders certified by AWS and the City of Portland.

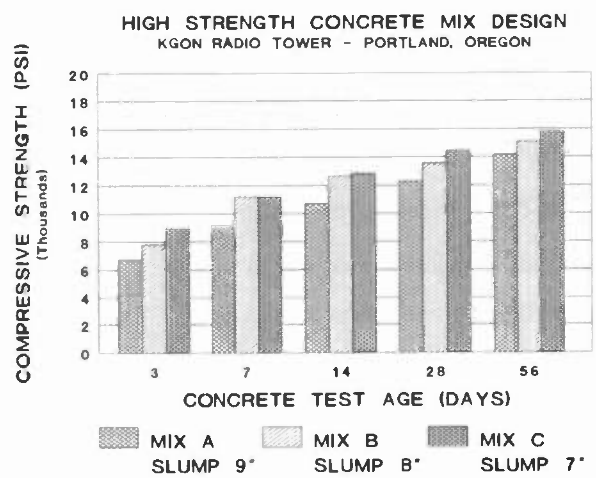

FIGURE 2

High-Strength Concrete: The high-strength concrete for the tower required a design mix for determination of proportions by the three point method (see Figure 2). A minimum strength of 12,000 psi and test mix of 14,000 psi to satisfy design and stiffness was required. A concrete test correlation for 6 x 12 and 4 x 8 test cylinders to obtain a relationship of concrete test using 4 x 8 cylinders on the project (see Figure 3). The test equipment in Portland did not have the capacity for concrete strengths in the range of 10,000 to 20,000 psi.

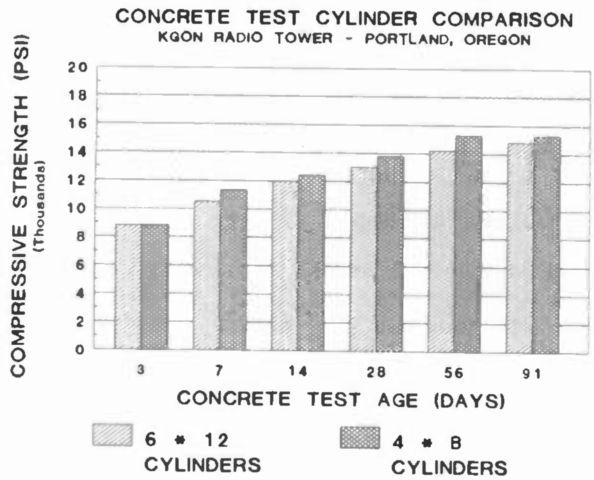

FIGURE 3

Corrosion Protection: All fasteners for the tower are ASTM A325 galvanized bolts and safety nuts. The legs and girts of the tower are painted with three coats (8 mils) of urethane enamel. All miscellaneous steel fabrication is galvanized and painted with two coats (6 mils) of urethane enamel.

Adhesion Test: A checkerboard scribe and tape pull-off test was specified to check the adhesion of the finish paint system.

Tower Survey: The contractor provided a professional land surveyor to establish survey control for each stage of the tower erection.

SITE WORK

After the tower structure is complete, the owner will provide parking and landscaping on the tower site in accordance with the concepts that were coordinated with the community association to provide screening and aesthetics.

CONCLUSION

We believe the KGON 600-foot radio tower provides:

- A convenient, multi-use broadcasting and communications facility.
- A stiff tower that will exhibit very little drift and rotation due to wind and icing loads.
- A simple, aesthetic structure suitable for the urban environment and acceptable to the neighborhood.
- A cost effective structure for joint tenancy and multi-use purposes.

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Professor Clifford Mass

MULTI-CHANNEL TV COMBINERS TECHNOLOGY FOR THE 90'S

James Stenberg
Micro-Communications
Manchester, New Hampshire

ABSTRACT

Increasing antenna, tower and real estate costs along with the benefits of co-location make multi-channel operation economical and efficient. In some cases, it is the only method available to get two or more stations on the air.

Recent advances in multi-channel combiners provide these benefits to almost any combination of stations. High power, low power, VHF and UHF stations both in the U.S. and internationally have found this to be true.

This paper presents the basic design considerations for VHF and UHF multi-channel systems. Criteria are given for choosing the proper type of combiner system for a given set of conditions. Examples of measured performance are given for each combiner type, along with details of installations now in operation.

COMBINED SYSTEM BENEFITS

There are, as mentioned above, two major reasons to install a multi-channel system. The first reason, lower installation and operating costs, results from combining several stations into a common transmission line and antenna. The second, increased viewership due to co-location, results when two stations together can afford a taller tower or more effective antenna system than either could have obtained alone.

The savings in installation costs can be significant even when two stations are to be combined. This is especially true when maximum power and antenna height installations are considered. For example, if we assume two 5 megawatt ERP UHF stations on the same 1500 foot tower each with it's own waveguide transmission line and antenna, the savings afforded by using a dual channel combiner and single line/ antenna reduce each stations initial investment by almost 30%.[1] Of course if the two stations were originally to operate on separate towers the savings increases to approximately 45%per station. These figures do not include any of the "little extras" like land, lights, buildings, etc.

In many cases the only method available for adding additional stations onto crowded antenna farms is to combine the new station with an existing station already in place. The Sears Tower, channel 50/60 installation, in Chicago is a fine example of this.

COMBINER BASICS

Multi-channel combiners, also known as multiplexers and diplexers, are of two basic types.

1. Constant impedance types use two identical filters placed between quadrature hybrids.

2. Starpoint types consist of single filters joined in common at a tee junction.

Constant Impedance Combiners

To understand the constant impedance combiner, we must first examine the operation of the quadrature hybrid.

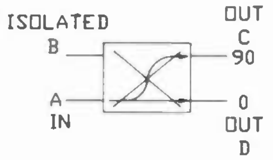


FIGURE 1
QUADRATURE HYBRID

When a signal is applied to Port A it splits equally between Ports C and D. No power appears at Port B since it is isolated. If we now put equally spaced short circuits on Ports C and D all input power will be reflected and appear at Port B with very low insertion loss.

If two hybrids are put back to back and a device that passes one channel and appears as a short circuit at another channel is placed between them,a combiner is created.

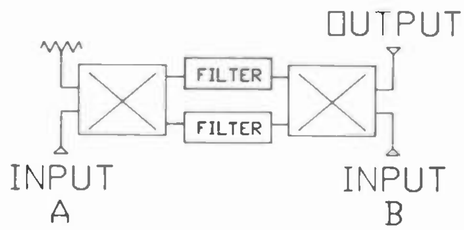


FIGURE 2
CONSTANT IMPEDANCE COMBINER

Channel A is split in the first hybrid and passes through the filters easily. The second hybrid then works in reverse to combine the split signal at the antenna port. If the two filters have identical performance the hybrid will remain isolated and Channel A power will not appear at the Channel B input. The value of this isolation dependends on the hybrid's isolation and balance properties. The Channel B input is split in the second hybrid and because the filter has an effective short circuit at Channel B it is reflected and appears at the antenna port. The tiny amount of Channel B power which slips by the filter is combined in hybrid #1 and appears at the reject load. Channel B is thus isolated from Channel A input by the filter and hybrid isolation combined.

The performance of this combiner type thus depends on two things:

(1) The filter's passband VSWR, insertion loss and reject band isolation.

(2) The hybrids isolation, split,and VSWR performance.

Each of these elements is equally important to system performance.

Starpoint Combiners

The starpoint combiner consists of individual channel filters connected to an output tee. Each channel input has an associated filter which passes only that channel and rejects all other channels. Once again if the rejection of the filter is high the filter will appear as a short circuit.

The filters on the other inputs pass only that channel and reject the others. To combine together the outputs, we must correctly design the lines between the filters and tee. At the reject channel a perfect open circuit must appear at the tee. (FIGURE 3) When lengths l_1 and l_2 are correct, both channels combine at the output with the insertion loss and VSWR of the filter.

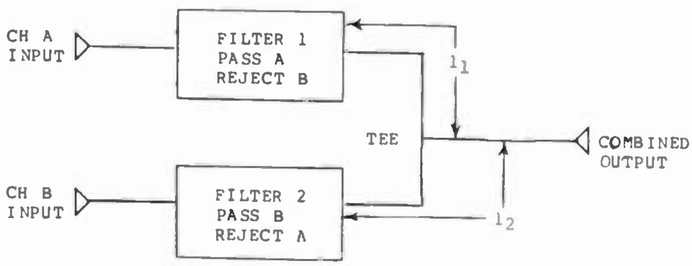


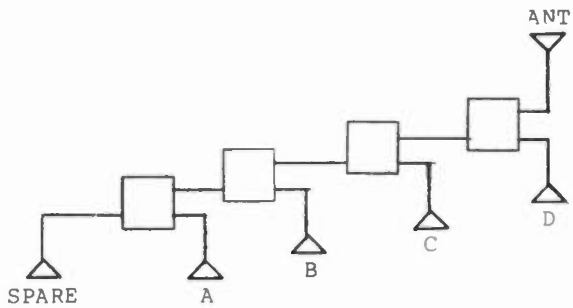
FIGURE 3
STARPOINT COMBINER

No additional VSWR or loss results from the other channel's operation. Isolation between the two inputs is that of the filter plus approximately 3dB which results from the tee power split. Thus this design relies completely on the filter performance.

Multi-Station Considerations

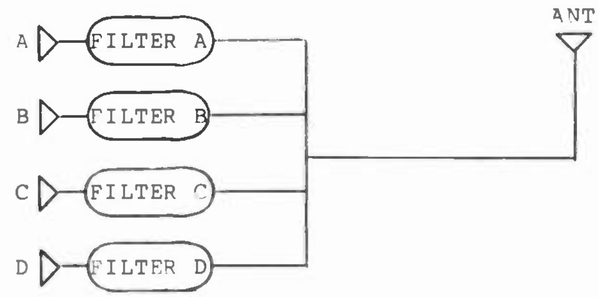
Constant Impedance combiners must be linked together in an in line or series chain in order to combine more than two channels. Starpoint combiners can accomodate many channels in one assembly since each has its own filter and a parallel arrangement is used. Parallel assemblies of up to 8 channels have been demonstrated. In, the past, most multi-station (i.e. more than 2) installations have utilized the constant impedance series approach because it offers modular design and can be expanded. Figure 4 lists the advantages and disadvantages of both types.

SERIES CHAIN OF CONSTANT IMPEDANCE COMBINERS



ADVANTAGES	DISADVANTAGES
EXPANDABLE BY ADDING MODULES	HYBRIDS EFFECT PERFORMANCE
EACH CHANNEL HAS SEPERATE MODULES	MANY INTERCONNECTS NEEDED
LOWER POWER MODULES CAN BE USED AT THE INPUT END	INSERTION LOSS INCREASES AT EACH LINK, FIRST STATION HAS TO GO BY ALL OTHERS
ISOLATION HELPED BY HYBRIDS	COSTLY TO BUILD
PERFORMANCE INDEPENDENT OF LINE LENGHTS	POWER HANDLING MUST INCREASE AT EACH LINK
	ISOLATION VARIES AT EACH INPUT

PARALLEL OR STARPOINT FILTER COMBINERS



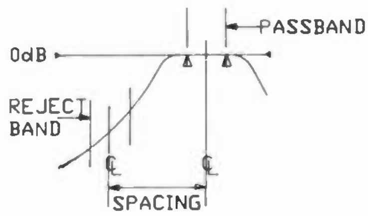
ADVANTAGES	DISADVANTAGES
LESS COSTLY	NOT EXPANDABLE
FILTER DETERMINES LOSS AND VSWR	ISOLATION ONLY THAT OF FILTER
ALL INPUTS HAVE SIMILAR PERFORMANCE	ADDITIONAL LOSS ADDED BY FILTER
CONSTANT POWER HANDLING REQUIRED	JUNCTION MUST HANDLE FULL POWER
MUCH SMALLER PHYSICALLY	PERFORMANCE DEPENDENT ON TEE AND LINE LENGTHS

FIGURE 4
MULTI-CHANNEL COMBINER TYPES

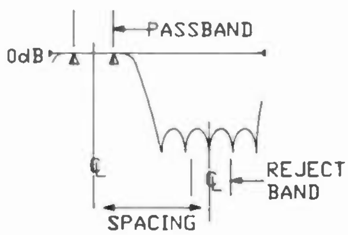
FILTERS

The above paragraphs demonstrate that combiner performance is controlled primarily by the filter used. There are many types of filters available for use in combiners,however only two major types are used, band-pass, and band-stop (or, band-reject/notch as it is sometimes called).

Selection and specification of filters is done with respect to two parameters. The first parameter,passband bandwidth,defines the amount of bandwidth available to pass the through channel. The second parameter defines the reject band spacing from pass-band center and reject insertion loss. Design of a filter consists of choosing a wide enough passband to provide low insertion loss and VSWR while at the same time maintaining the necessary rejection. To increase the passband bandwidth while maintaining the same rejection, the number of resonant sections must be increased.



a) BAND-PASS FILTER



b) BAND-STOP FILTER

FIGURE 5
FILTER RESPONSES

Band-pass

When resonant sections are arranged so that RF is coupled through them in a series connection, each section having an input and output terminal, a band-pass filter is created. If the tuning and spacing of these sections is carefully controlled, specific responses can be obtained. For narrow bandwidths such as those required here band-pass filter performance is quite predictable and designs can be readily calculated. Figure 6 shows the effect of adding more resonant sections while holding the passband bandwidth constant. Band-pass filters used in combiners consist of three different forms.

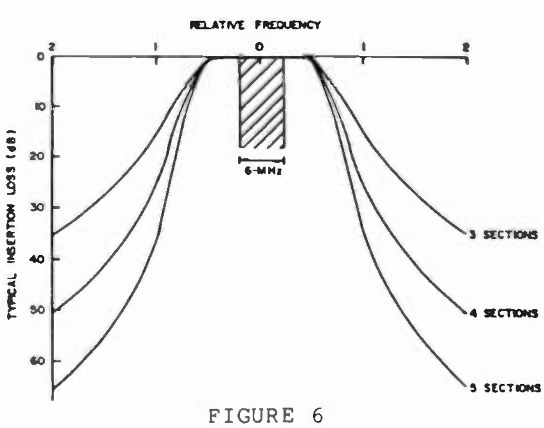


FIGURE 6
REJECTION VERSUS # OF SECTIONS

In waveguide, iris coupled resonant length sections are joined to form a multisection filter. These are used in waveguide channel combiners. This design was chosen because of its ruggedness and stability as well as excellent power handlng capability. Since waveguide cavities are used extremely low insertion loss can be achieved. A typical unit is shown in Figure 7.

FIGURE 7
UHF WAVEGUIDE BAND-PASS FILTER

Coax band-pass filters are built in two different forms, interdigital and loop coupled cavities. For VHF and FM combiners loop coupled coaxial cavities can be joined together to form a filter. MCI VHF TV combiners do not use band-pass filters since their cost and complexity is greater than an equivalent band-stop filter.

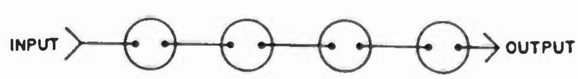


FIGURE 8
COAX BAND-PASS CAVITIES

MCI UHF low power band-pass filters utilize a unique interdigital design. Quarter wavelength rods are fixed inside a rectangular housing and the coupling is varied by adjusting the rod spacing. The design results in a small rugged unit which is easily integrated into a small package with filters for other channels. The resulting filter is approximately 1/10 the size of a similar waveguide filter. It does, however, have a higher passband insertion loss since coaxial rods are used.

FIGURE 9
INTERDIGITAL FILTER

The benefits of band-pass filters are that they reject all energy except that near the narrow passband. A negative aspect of this filter is that the passband is dependent on the resonances remaining stable. If any one section changes, the passband insertion loss and VSWR will be disturbed. The reject band is generally stable and independent of section changes.

Band-stop

Band-stop filters are formed by joining together resonant sections which are coupled across the line in a shunt connection. Each resonance has only one connection and thus passes all frequences except that of the resonance. Frequencies at resonance are shunted to ground. Here again if the sections are tuned and spaced correctly specific responses can be obtained. In a band-stop filter the response is determined by the number of sections and the amount of coupling or loaded Q of each section.

Figure 10 shows the effect on isolation of adding more sections.

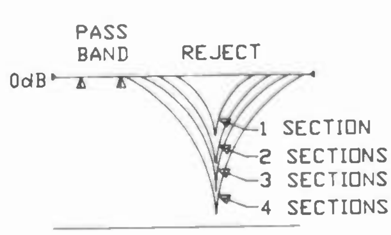


FIGURE 10

The steps in designing band-stop filters consist of adding sections to provide the desired reject bandwidth and isolation while maintaining low insertion loss at the passband. If the passband loss is too high the sections must be narrowed and/or more added. Band-stop filters have the advantage that passband performance is relatively independent of the tuning. Unlike band-pass filters the reject isolation is dependent on cavity tuning. If one cavity drifts the rejection will change. The band-stop filter is very effective when a narrow reject band is needed. When wide reject bands with close spaced passbands are needed a large number of narrow cavities must be used.

Two forms of band-stop filters are used in combiners, waveguide notches and coaxial notches. Waveguide notches are used at UHF on visual/aural diplexers and have not been used on channel combiners since they are much larger and more expensive than band-pass waveguide filters. Coax band-stop filters are used in VHF combiners. Large diameter cavities are joined together to form the necessary response. Cavity coupling is adjusted by varying the input loop length inside the cavity. These filters utilize invar thermal compensators to reduce drift. Figure 11 shows a typical VHF filter.

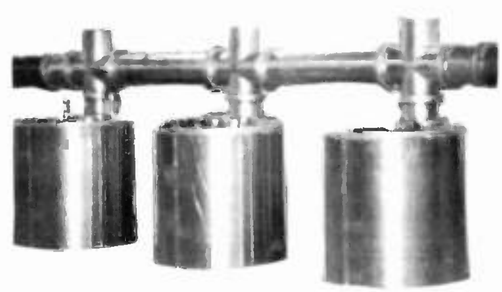


FIGURE 11

COMBINER PERFORMANCE COMPARISON

If one reads much of the currently available combiner literature it becomes apparent that there is a lot of controversy over which type of combiner is best. Each claiming that their's is the best. Since the filter design is the most important part of the combiner, if equivalent filters are compared the differences become much less significant. To examine and compare the three types of combiners we will assume that for all three cases equivalent filters are used. The three filters have the performance shown below.

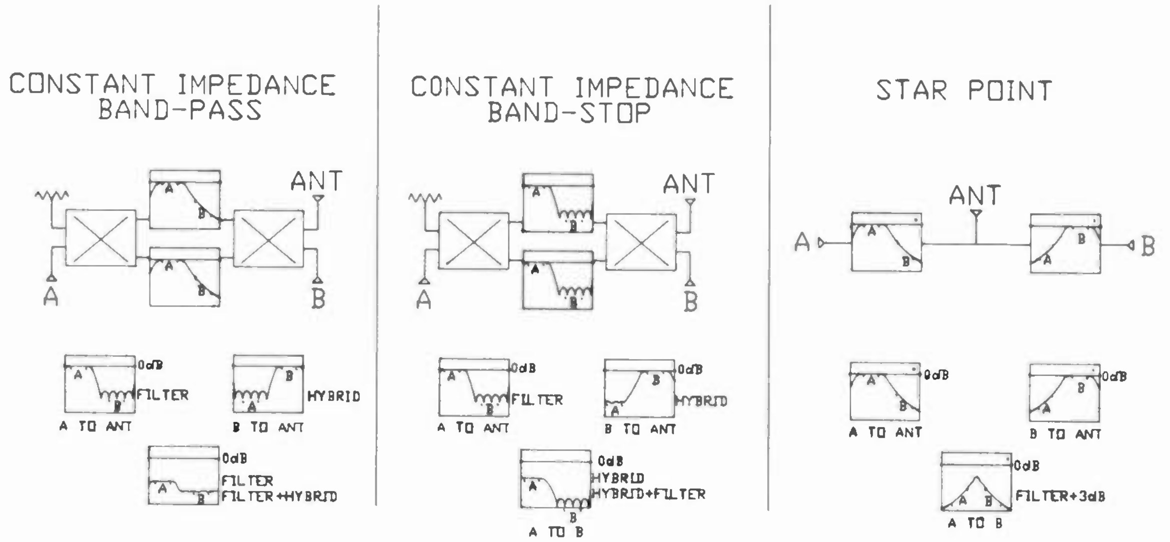
1. Passband (CH A)

   a. VSWR = 1.05:1
   b. I.L. = .25dB
   c. Group Delay ±20ns

2. Rejectband (CH B)

   a. Spacing from CH A = 40MHz
   b. I.L. = 35dB

If these specifications are transposed to filters for the three combiners at the same frequency, the values given in Table I result. The corresponding combiners are also shown.



	CONSTANT IMPEDANCE		
PARAMETER	BAND-PASS	BAND-STOP	STARPOINT
CH A	THROUGH	THROUGH	FILTER 1
VSWR	1.05	1.05	1.07
INSERTION LOSS	.27dB	.27dB	.27dB
GROUP DELAY	±20ns	±20ns	±20ns
ISOLATION A INTO B	35dB (HYBRID)	35dB (HYBRID)	38dB (FILTER+3)
CH B	REJECTED	REJECTED	FILTER 2
VSWR	1.04	1.05	1.07
INSERTION LOSS	.08dB	.08dB	.27dB
GROUP DELAY	±10ns	±20ns	±20ns
ISOLATION B INTO A	>60dB (HYBRID+FILTER)	>60dB (HYBRID+FILTER)	38dB (FILTER+3)
FILTER SENSITIVITY	A=HIGH B=LOW	A=LOW B=HIGH	A=HIGH B=HIGH
IMPEDANCE VARIATION	A=MID B=LOW	A=MID B=LOW	A=MID B=MID
ISOLATION AT OTHER CHANNELS	A=HIGH B=MID	A=LOW B=HIGH	A=MID B=MID

TABLE I
COMBINER PERFORMANCE COMPARISON

Table I shows that the differences between band-pass and band-stop combiners are really very small. In fact the performance is essentially the same. The only significant difference relates to the isolation between inputs outside the channels of interest. This is a consideration when more than one multiplexer is used in line, as discussed in the preceding section. In this case the band-pass combiner provides input isolation of the filter response plus the hybrid isolation at all frequencies. The band-stop combiner on the other hand has only the isolation of the hybrids at frequencies outside Channel B.

The starpoint obviously has greater loss on Channel B since it must pass through a filter unlike the other two which rely on the hybrids.

MCI'S COMBINERS

MCI has developed products based on all three types of these combiners. The band-pass combiner is used to combine high power UHF stations. These combiners utilize waveguide throughout. Band-stop combiners are used in VHF and FM applications. Here coaxial hybrids and multiple cavities are used. Starpoint combiners are used for low power UHF stations. Low power combiners with up to 4 channels have been manufactured. Table II lists the recent combiners built using these designs.

INSTALLATION	CHANNELS	TYPE	LINE	DESIGN POWER/CH	LOSS/CH	NOTES
MUTLAA, KUWAIT	E45/E47	BAND-PASS	W/G	240kW	.08dB .14dB	SINGLE CHANNEL SPACING
FAILAKA, KUWAIT	E24/E39	BAND-PASS	W/G	120kW	.08dB .15dB	EXTREMELY BROADBAND 40%
SEARS TOWER CHICAGO, IL	50/60	BAND-PASS	W/G	100kW	.1dB .2dB	FIRST HIGH POWER UHF COMBINER IN U.S.
KHAFJI, S.A.	E51/E53	BAND-PASS	W/G	40kW	.08dB .24dB	SINGLE CHANNEL SPACING
MUTLAA, KUWAIT	E5/E12	BAND-STOP	COAX	40kW	.06dB .08dB	BROADBAND
FAILAKA, KUWAIT	E8/E10	BAND-STOP	COAX	40kW	.05dB .07dB	SINGLE CHANNEL SPACING
THUNDER BAY, ONTARIO	9/13	BAND-STOP	COAX	15kW	.20dB .30dB	
JEDDAH, S.A.	E50/E53	INTER-DIGITAL STARPOINT	COAX	1kW	.43dB .49dB	CLOSED SPACED
BARBADOS, B.W.I.	14/18/22/26	INTER-DIGITAL STARPOINT	COAX	1kW	.37dB .36dB .38dB .48dB	4 CHANNELS TO ONE OUTPUT
DENVER, CO	49/63	INTER-DIGITAL STARPOINT	COAX	1kW	.24dB .28dB	BROADBAND

TABLE II
MCI MULTI-CHANNEL TV COMBINERS

Table II shows that the performance of all four high power waveguide combiners is essentially the same. Since the channel spacing varies from 16MHz to 120MHz the four filter designs had to be optimized seperately. The narrow spaced channels required a five, section narrow passband design while the wide spaced channels used four sections and a wide passband. All efforts were made to minimize the insertion loss on both channels. Figure 12 shows a typical combiner.

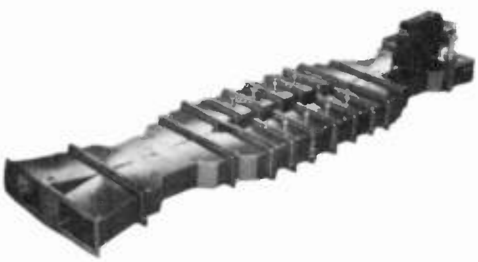


FIGURE 12
WAVEGUIDE CHANNEL COMBINER

Systems

In most high power installations,switching splitting systems, and waveguide transmission lines are used along with the combiner itself. The design of a multichannel system consists not only of the combiner but all other components in the systems. Unlike single channel operation where standard designs can be used, each component in a multi-channel system must be evaluated relative to the necessary bandwidth and power handling capacity. It must first be determined whether waveguide should be used[2], and if yes, then what size is needed for the channel combiner. The bandwidth and power limitations of the components are examined next and necessary new component designs or system reorganizations are made.

The Mutlaa, Kuwait system (CH 45/47) combined two 240kW stations for a total system output of 480kW, the highest system power level in the world. This system, although relatively narrowband, required development of new super power switches. The close channel spacing also made the combiner optimization much more difficult. On the other hand the Failaka, Kuwait system (CH 24/39) operates over an extremely broad bandwidth (144MHz) and required extensive redesign of many components to maintain the low VSWR and loss needed. New elbows, hybrids, and switches were developed to allow the complicated switching layout shown in Figure 13. It was also required that the dual waveguide tower run, output splitter and antenna be operable over the full bandwidth so that rechanneling could be done without tower/antenna modifications. Novel tuning and length selection techniques were utilized to meet these requirements.

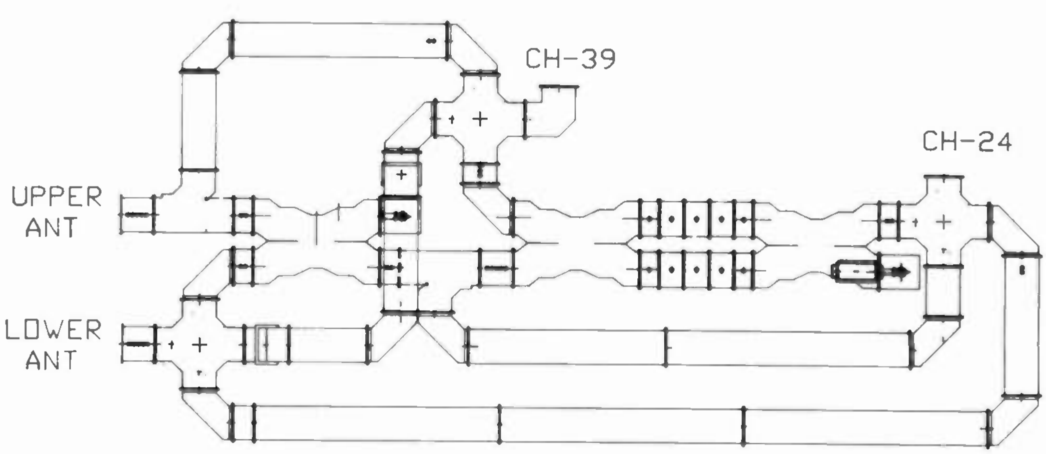


FIGURE 13
BROADBAND WAVEGUIDE CHANNEL
COMBINER, SWITCHING, AND SPLITTING SYSTEM

VHF band-stop combiners have also been built with both close, and wide spaced channels. The wide spaced channels used three broadly coupled cavities to produce a wide reject band. The close spaced unit used four narrow coupled cavities to maintain low passband loss. From the table it can also be seen that the loss of these combiners is very close to that of the UHF models since similar filter design goals were used.

VHF coax system design considerations are similar to those for waveguide systems. The bandwidth limitations of coax are however much less restrictive than for waveguide since coax is non-dispersive. The limited number of channel allotments at VHF frequencies also helps reduce the bandwidth requirements. Thus VHF systems seldom require component redesign, and layouts can be easily made using standard components. Figure 14 shows the output switcher/splitter system used on the Failaka, Kuwait 80kW installation. Inputs for each channel can be routed to the combiner/splitter or bypassed to either of the dual transmission lines allowing full system redundancy.

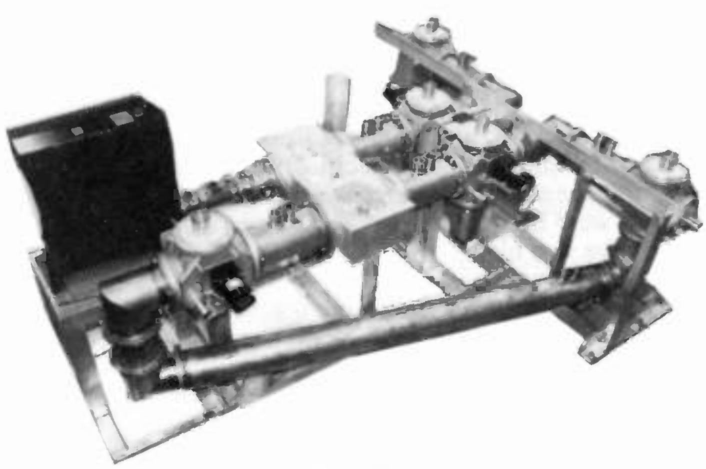


FIGURE 14
VHF OUTPUT SPLITTER/SWITCHING SYSTEM

Low power UHF combiners have been built for two, three, and four channels. Each channel has a 1 5/8" EIA input connection and either a 1 5/8" or 3 1/8" EIA output connection. Figure 15 shows the four channel unit and it's associated support frame. The frame allows mounting in any convenient location.

The insertion loss of these combiners is higher than the high power models due to the reduced size of the filters and the lower input power used. At 1kW the amount of heat dissipated in the filers is very low.

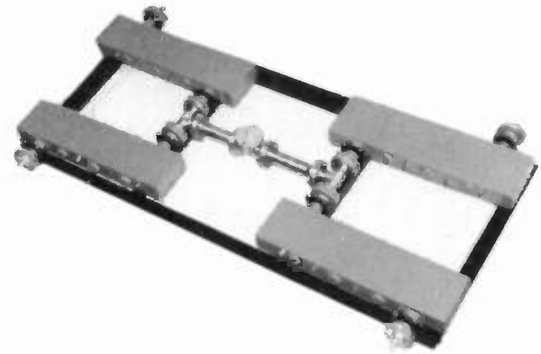


FIGURE 15
LPTV FOUR CHANNEL COMBINER

CONCLUSION

Multi-channel combiner technology has evolved to the point where widespread use is now possible. Significant economic and technical benefits result from the combination of two or more stations into a common antenna.

The various types of combiners have been reviewed and an introduction to their operation and design was presented. We have demonstrated that the performance of any combiner is highly dependent on the filter used. The advantages and disadvantages of each type have been discussed.

Details of combiners currently in operation were given and several system configurations were shown. It is hoped that from this paper prospective multi-channel system installations can be examined and manufacturers quotes can be evaluated so that a proper system selection can be made.

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DISTRIBUTION OF BROADCAST QUALITY VIDEO USING DIGITAL TELEPHONY TRANSMISSION

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Abstract- Advances in signal processing allow the distribution of broadcast quality video signals using the transmission facilities of the public telephone network. Increasing use of fiber optic transmission systems is dramatically lowering the cost of digital transport compared to analog systems. The implications to broadcasters, affiliates and other users of video services will be discussed.

INTRODUCTION

The nationwide field trial of fiber optic networks for broadcast television was activated by the American Broadcasting Company (ABC–TV) in December of 1989. Eight cities from coast-to-coast, 100% linked with fiber optic transmission systems, will be on line for 15 months of trials. The network is fully interconnected using a telephone standard transmission system called DS–3, which operates at a data rate of 44.73 Million bits per second. Following ABC's use of the network, CBS, NBC, Fox and PBS will have their turn for about three months each. A separate trial in Canada is planned for the Canadian Broadcasting Company (CBC).

Of course, the use of the public telephone network to carry television signals is nothing new. While satellites have taken over the majority of the distribution task, terrestrial systems are still common. But the telephone network is changing. Today, as the first generations of satellites are reaching the end of their service life, the national networks are looking at substantial investments in new ones.[1] And even the most ardent proponent of terrestrial systems cannot dispute the satellite's unsurpassed ability to provide point-to-multipoint distribution of program material.

The present nationwide field trial is a reflection on the potential of expanded network services. We will not dwell on the technical details of the present field trial. Another author, Mr. Bob Blackburn of Bellcore, has a complete explanation of the field trial system in his paper, "Eight City Digital Video Trial - What Makes It Work", printed in another part of these Proceedings. Bellcore has taken the lead in the planning, implementation and development of the specifications for both the structure and the operations of the network. All the participating carriers are very interested in the trial and have been following its progress carefully. We will see why in a moment.

THE PUBLIC TELECOMMUNICATIONS NETWORK – PAST AND PRESENT

The Network of the Past

The telephone industry was simpler in concept, if not in implementation, a generation ago. For television services, the large Bell Operating Companies set up special organizations, usually called "Television Operating Centers," or TOCs. Point-to-point networks were built in a long and involved process...it worked, but had limitations. The circuits installed by AT&T were treated as a special case due to their unique characteristics and used specifically developed equipment. The ability to change and modify the configurations were limited at best. The cost of supplying the services was high.

Eventually, the need for a more sophisticated network forced the broadcasters to look for an alternative. Some of the desired operational features were real time control of the routing of program material, backup networks, higher standards for transmission quality and consistency, the ability to conduct two-way communications, rapid installation of facilities....the list goes on. Satellites overcame many of the disadvantages of the landline network and soon became the standard means for program material distribution.

The Network Now

Today's telephone operating company has discovered the valuable nature of its franchise. The regulatory climate has changed dramatically. At the same time, a strong push to provide new services has become a strategic thrust for the companies. On the demand side, today's telecommunications users are sophisticated professionals who demand the same level of service from the public network that is now possible with a private network.

The telephone companies now consider themselves telecommunications service providers, and are working on either introducing new services or greatly expanding the capabilities of existing products. But the main point to consider is that the term "telecommunications services providers" now covers a lot of ground. Before, it was the telephone company. Now, a service provider could be one of a number of companies, each one anxious to establish a position in the market. One example of today's expanded market is the emerging bypass companies that are running fiber optic cables around a city, connecting end-users with each other, as well as local and interexchange carriers with a privately owned voice, data and image communications network. In Washington, D.C., a bypass carrier called Metropolitan Fiber Systems Inc. (MFS) is planning to install over 650 miles of fiber cable.[2] It is the regional companies and their competitors, such as MFS, that are shaping the industry.

Special Services, Not Special Networks

Under the threat of competition from new, highly focused and determined competitors, the telephone operating companies have responded. The biggest success stories have come when they used their greatest competitive advantage: putting new technology into their sophisticated infrastructure and thereby introducing new services without building special networks. Remember the TOC concept? A special group, doing nothing but handling the special requirements of just one set of customers. Today, the idea of special networks for special services is no longer workable.

Instead, the telephone operating companies have a long-range goal of using the same digital switching and transmission system for any service, no matter if it is voice, data...or image. And that network has to be flexible and robust, of superior quality, high reliability, and cheap. The alternative is that their customers will continue to build and operate their own networks, or move to competitive suppliers. That is because, up until recently, the public network has provided transport: simple links between two points. Put in video at one end, out it comes at the other. If you wanted it to go somewhere else, it meant a new circuit, new construction, delay and expense. But not any more.

CUSTOMER-CONTROLLED VIDEO NETWORKS

The idea of a customer-controlled network is nothing new. Before 1990 is over, the industry may see an agreement on the coding method used to translate analog video and aural information into a standard telephone company format. These devices, called codecs (short for coders-decoders), utilize a sophisticated compression technique to convert baseband video into high speed data symbols. The format, called DS–3, operates at about 45 Million bits of information per second; the payload is just over 44 Million bits per second, and the rest is overhead. The advantages of having an industry standard coding process is that, for the first time, video codecs from different vendors will be able to exchange program material. No true network can exist if there is no certainty that the endpoints can communicate.

The North American telecommunications network is built on the DS–3 system: commonly used, well understood and extensively deployed. Figure 1 shows how a typical DS–3 video network system could be designed. Note that once the signal is in the DS–3 format, it becomes part of the larger network. And that carries with it substantial advantages.

The Case for Digital

What are the advantages of using the public network for video transmission? There are several. Some are a

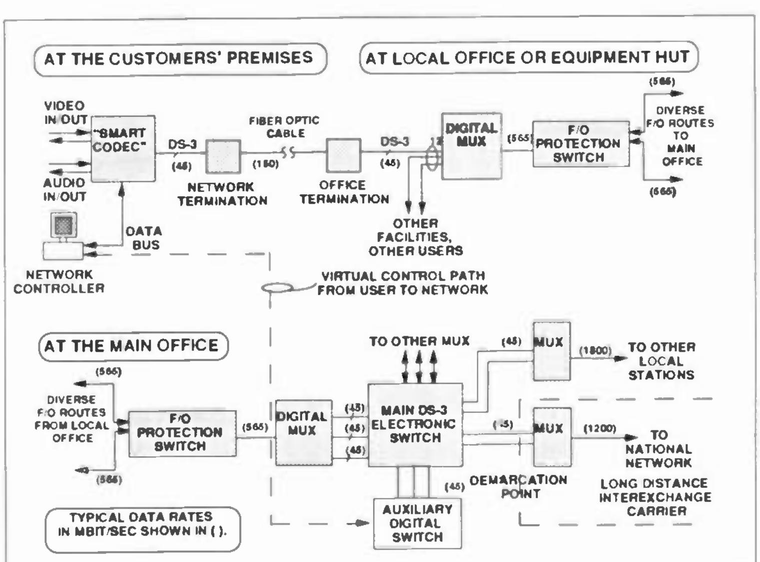

Figure 1.
A Representative Digital Video Network

function of the digital format, while others are related to the way the public network functions. A further advantage is the control offered by the networking standards issued by Bellcore and implemented in the nationwide video field trial. A summary chart showing the primary points for digital networks is shown in Figure 2; let's discuss each item in turn.

(1) Distance independence. An analog signal must be reamplified at intermediate points, degrading the signal due to the inevitable addition of noise. The standard for analog transmission (EIA RS–250B) has distance-specific performance specifications depending on the distance. Digital transmission systems are different: a digital transport system does not carry video...it is handling data. Bits and bytes. An eight-bit word is an eight-bit word whether it travels ten miles or ten thousand miles. Consistent quality independent of the route is a critical advantage. The user no longer cares about the path the signal has taken. A signal originating from the White House and destined for New York could travel via Los Angeles. Such a situation can and does occur with conventional telephone calls in the present long distance networks. And, in today's Bellcore field trial network, that would be a distinct possibility for video as well. The distance may change, the routing may be different, but the quality remains the same.

(2) Reliable. Once the digital signal enters the network, it is grouped with other signals in higher order systems that carry the payload from source to destination using any available facility. If one path is disrupted, automatic systems take over and route the signal through another. Since the DS–3 video signal looks like any other, it receives equal treatment.

Conventional analog video systems are built for that specific purpose. Any backup routes need the same special facilities. In today's telecommunications networks, special handling is getting far too expensive. But DS–3 circuits are the building block of the modern digital network. High-speed digital fiber optic circuits carry up to 36 DS–3 systems. In terms of telephone voice circuits, that amounts to over 24,000 conversations riding on a single fiber or digital microwave system. With that type of load, a carrier has a strong incentive to keep those circuits up and running no matter what. The network providers will put in hot-standby equipment, install fiber optic switches for route selection and other operational precautions. The use of DS–3 standard signals means that the network may be engineered, built and maintained using standard equipment and standard procedures. Special video expertise is no longer required. All the above adds to the integrity of the digital network and bestows a much higher degree of reliability on the transport of video signals through the public network.

PARAMETER	ADVANTAGE
DIGITAL TRANSMISSION	• Distance / route insensitive • Stable signal quality • Two-way
NETWORK COMPATIBLE	• Reliability • Security • Cost
CUSTOMER CONTROL	• Rapid changing • Full network performance • Trouble isolation

Figure 2. Digital Video Networks: Summary of Features

(3) Security. Once encoded to DS–3 and into the network, the video signal becomes one of thousands of circuits. The signal has been coded using equipment that is not required to be present in the network. Even for testing, there is no need to bring the signal down to baseband. Thus, there is little danger of interception. The use of clear broadcasts over satellites has a long history of trouble. One example that springs to mind was the controversy over an actuality of the recent tragic plane crash in Iowa, where a signal was picked off a satellite feed and used by others. Security for the commercial broadcasters is particularly important for internal news feeds and in the transmission of commercial material prior to airing. The use of terrestrial DS–3 networks could handle the security issue with ease.

(4) Cost. No discussion on the use of the network would be complete without some mention of the costs. Consider an analog video network: special services, special equipment, special staff. All are factors that the operating companies are striving to reduce or eliminate. Indeed, most interexchange carriers, outside of AT&T, are not likely to have analog circuits. With standard DS–3 transmission systems, the users have a choice of many carriers. What are the cost implications of buying services which only few suppliers can provide? Witness the recent reaction of the broadcasters when AT&T tried to adjust its charges.[5] The price of digital transport continues to decline. Recently, AT&T proposed to lower prices by 42% for DS–3 circuits over 2000 miles in length.[3] While analog systems may still retain a cost advantage in short haul applications, the cost of digital transport will continue to decline.

Of course, the circuit cost will depend on the rates charged by the long-haul companies and the local carrier's tariffs. While the bypass companies are aggressively seeking business, most local access will still be through the local telephone company. And the tariffs for high capacity services such as DS–3 vary from city to city and state to state. Unfortunately, there is no uniform method for handling such high capacity services such as DS–3 video, and there may not be in the immediate future.[4] Broadcasters interested in DS–3 networks will have to contact their local operating companies and look to alternative bypass carriers as well.

(5) Control. Customer-controlled networks for low-speed data have been available for some time. The operating company provides a high quality, reliable, secure, and cost effective transport network. The end-user is provided an interface or data communications link that is used for actually controlling the configuration of the network. In some cases, network management tools are provided so that the user can determine the source of a problem and inform the network provider if corrective action is needed.

Recently, service providers have introduced DS–1 (a lower speed digital format at 1.544 MBit/sec) and DS–3 customer-controlled networks. Called Switched Multimegabit Data Services, or SMDS, it is available today under tariff in some areas.[6] The network experts of Bellcore have developed requirements for a special system optimized for use by broadcasters and others interested in network distribution and collection of video signals. While the commercial SMDS system provides basic networking features, the Bellcore network has operational and maintenance features designed for video services. Figure 3 represents some of the more important networking aspects of the system. Again, the reader is referred to Mr. Blackburn's paper elsewhere in the Proceedings for details.

USERS OF DIGITAL VIDEO NETWORKS

Who will use DS–3 video? And why? Of course, the first candidates are the national broadcast networks. While a terrestrial network is not likely to replace the satellite network for general program distribution, the advantages of digital networks make it an attractive alternative for special applications. For short and simple point-to-point links, analog transmission over fiber may be the technology of choice, but such systems come with limited added utility. It gets you there, but little more. For a true network, where a group of stations are required to be interconnected with two-way video transmission for extended periods, the digital terrestrial network is definitely worth serious consideration.

But there are only a handful of national television networks in North America. Who else would consider a DS–3 switched system? Local broadcast stations may want to have a flexible and always-available network for selecting various points in an urban area for program origination or distribution. Such a system has been used in Cincinnati, where WKRC–TV aired a live remote over digital fiber; an analog fiber system is in operation within the District of Columbia.[8,9]

Another area of growth is in interactive situations.

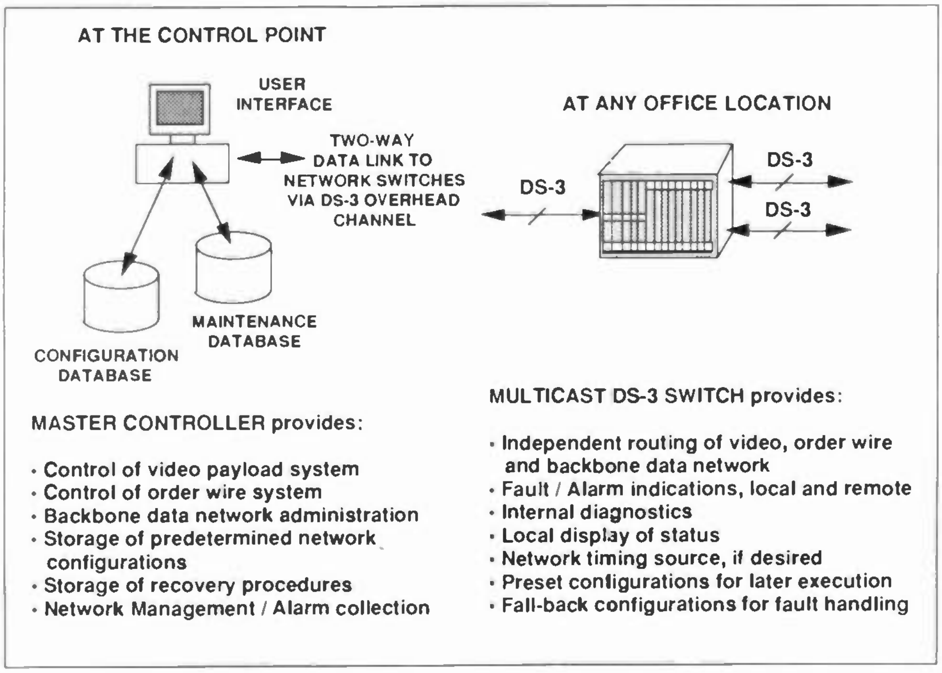

Figure 3. Network Aspects of Digital Video Systems

Video conferencing networks are expected to expand dramatically in the 1990s.[10] Commercial and educational video conference users typically do not require the high quality and superior features of DS–3 networks for workaday situations. However, telecommunications managers are specifying the high-speed networks where required. One market survey of top corporations found about eight percent of the companies and several educational institutions using DS–3 video products, usually as a complementary service to lower performance systems.[11] Interest is high at all levels of government in the potential of digital video services. Traditionally, most governments acquire telecommunications services through the public networks, and the advantages of using public network video systems are readily apparent to state and national telecommunications directors.[12]

In all these cases, the same advantages keep coming up: distance insensitivity, reliability and security, cost, and the ability to control the network.

What's Next?

The present systems, sophisticated as they are, are only the leading indications of the public networks to come. New technologies are being developed now for possible commercial inplementation. One example is bandwidth on demand. Need a low-speed signal for simple communications? Order what is needed. When full quality becomes important, the bandwidth can be increased accordingly. You pay by the megabit, not by the month.

The marketplace for telecommunications will drive the development of goods and services. When customers have a need and are willing to pay for it, a source of supply will develop. And the providers of digital television networks need not be the traditional telephone operating companies. Smaller and aggressive carriers are in place now: look to see them expand.

Summary

The evolution of digital technology offers the television industry new opportunities for networks based on new technologies. Telecommunications providers, in particular the public carriers, will use their competitive advantages of performance, reliability, security, and control as a way to offer new services such as customer-controlled digital video networks. Overcoming the disadvantages of the traditional analog networks, new technologies deserve a close look by television networks, local affiliates, industrial and commercial users of video services and educational systems.

Acknowledgements

The author would like to emphasize the role that Bell Communications Research has played in the development of the concepts of DS–3 video networks. Without the planning and vision of the staff at Bellcore, in particular Bob Blackburn, the District Manager in charge of the field trial coordination, the field trial would never have occurred. Mr. Jay Reynolds of Telettra U.S.A., supplier of DS–3 video codecs used in the field trial, also provided additional assistance.

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8-CITY DS3 DIGITAL VIDEO TRIAL—WHAT MAKES IT WORK

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ABSTRACT

The nationwide, 8-city, digital video trial is now a reality. More than 40 companies have been involved in putting this trial together. It is only through the good will of the participants that this historic undertaking has been able to succeed. The two primary challenges of 1) high quality picture and stereo sound and 2) flexible customer-controlled networking, have been met. For example, the trial network could be used to collect television from Los Angeles to New York for distribution to affiliates via satellite in order to demonstrate a successful marriage between the two technologies. This paper describes how this all works. Papers presented at the 1987, 1988 and 1989 NAB Conferences provide further background.

INTRODUCTION

Since early 1987, Bellcore has been generating the concept, publishing the technical requirements for the customer controlled networking features of Multicast Switches, Video Terminals and Audio Terminals, and encouraging the proposal of standards for digital coding algorithms for transmission of broadcast quality NTSC television over standard 45 Mb/s telephone company fiber optic facilities. A major goal has been to incorporate powerful customer controlled networking capabilities for distributing and collecting program material among a large number of locations. An 8-city trial network (see Figure 1) has now been constructed and has been in operation at ABC since November 1989. In the coming months, CBS, NBC, FOX, PBS and CBC will each have the opportunity to experience this pioneering system.

In addition to the major television broadcast networks and their affiliates in the eight cities, trial participants include Bellcore, the seven Regional Telephone Companies, several long distance carriers, and numerous equipment suppliers (see Figure 2).

One of the most significant and innovative aspects of this new network capability is its provision for multipoint networking with distributed switching and the ability to create large nationwide, regional, statewide, or local networks. In addition, the network includes an embedded control channel and private telephone channel (orderwire), each of which can be separately controlled to create multilocation, national, regional, statewide or local control and private telephone sub-networks independent of how the television network is arranged.

In addition to privacy and program security, digital transmission on fiber optics offers the significant advantage of a consistent level of picture and sound quality regardless of weather conditions and regardless of transmission distance. There is also considerably less transmission delay than satellite networks.

This new nationwide compatible networking capability comes none too soon - commercial satellite launch capability has been extremely scarce. The networking features are intended to be flexible enough to supplement or replace satellite networks for transmission to and from fixed locations.

HOW IT WORKS

First, let us talk about how the high picture quality is achieved. This gets into a discussion of coding algorithms and this will be described in simple terms although this is a highly complex subject.

Second, we will talk about how the network is created and controlled. If there are troubles, these can be readily detected, dealt with and cured.

PICTURE QUALITY

First, let's discuss picture quality.

There have been several attempts at coding pictures into a DS3 (45 Mb/s) signal over the years. Not all of these have been good enough for broadcast television quality. Some people remember these and are not aware that there are good and bad DS3-rate coders.

Without some form of bit compression, it is not possible to transmit television at 45 Mb/s. But compression does not necessarily mean loss in picture quality, and compression to 45 Mb/s is not very severe. Still there are good ways and bad ways to do it.

In the United States, several attempts at making broadcast quality coders in earlier years had failed to yield satisfactory results. Meanwhile, in Europe, a company called Telettra was working to solve a more difficult problem than existed in the United States. This was the problem of compressing a larger bandwidth (625 line) picture into a smaller digital channel (34 Mb/s) - compared to the US's 525 line picture and 45 Mb/s channel. Therefore they had to turn to a more complex technology than was being used at that time in the United States - a technology called Discrete Cosine Transforms. It turned out that this technology was more than good enough for the US situation.

Telettra was one of the many companies invited to participate in the DS3 digital video trial. By August of 1987, they had disclosed the details of their coding algorithm and proposed it to the American National Standards Institute for consideration as a standard. In June 1988, the trial participants agreed that it would be suitable for use in the trial and Telettra agreed to provide the codecs and to incorporate the features, described in published Bellcore Technical Advisories, needed for networking control as well. Other suppliers were invited to build to these fully disclosed requirements for compatibility.

How does the coding algorithm work? Most important of all, it does not produce any motion artifacts. Therefore it is suitable for sporting events with difficult scene changes and audience panning - to wit - a diver coming off a platform moving rapidly past a detailed audience background. The reason it produces no motion artifacts is that it is powerful enough to encode each individual picture frame clearly without resorting to interframe coding. Since motion comes from changes between picture frames, there is therefore no possibility for motion artifacts (things that are not really there).

Compression does not mean loss of picture information. Most pictures have at least some degree of correlation between picture elements. If they didn't, they would be to the eye what noise is to the ear.

Discrete Cosine Transforms is a fancy technical word for a process that takes and 8 line by 8 picture element sample (64 points) of a picture, and transforms the samples into measures of how much correlation there is between sample elements. This is not a simple process and it takes VLSI technology to do it in real time.

If there is correlation, as there always is, then the bit rate required to transmit these measures (called transform coefficients) is considerabley less than the bit rate required to transmit the samples themselves. Actually there is a series of coefficients or measures that need to be transmitted, but depending upon the degree of correlation in the picture many of these coefficients are zero. This information is therefore transmitted in a sort of shorthand notation - for example the notation "ten zeros" is transmitted, rather than transmitting zeros ten times. This is called "zero run-length coding".

Of the coefficients that are not zero, another type of shorthand is used. This is called "variable word length coding". This is directly analogous to the English language, where frequently used words have a short spelling -to wit- the word "a", a frequently used word and therefore very efficiently spelled by a single letter. Frequently used coefficients are assigned short codes and less frequently used ones are assigned longer codes - again, to reduce the bit rate required to transmit the picture.

There is more to the complexity than can readily be described here but you should get the idea by now. The bit rate needed to transmit the picture while still preserving high quality and without introducing motion problems, is reduced by a combination of "transform coding", "zero run-length coding" and "variable word length coding". And this has been shown to produce excellent quality, far superior to some of the earlier attempts at bit compression.

NETWORKING

Now let us discuss networking, specifically, the ABC trial network. Earlier papers have discussed how this type of network can be used for creating large tree-like, or more complex nationwide, or regional networks. But to show how the trial network works lets refer to Figure 3.

This figure shows the details of the network. It shows how the Video Terminals are connected to the Multicast Switches. It also shows the port number assignments (the switch is an 8-port device). Each line represents a two-way path.

Also shown are the unit address numbers assigned to each Video Terminal, Multicast Switch and Master Controller. In addition, a short hand name that can be used instead of the number assigned to the switch is shown. The name can be whatever is desired - these names were chosen by ABC for convenience.

For example, the address of the Minneapolis switch is 27. It is also called MINN for convenience. The 2-way port #4 is a DS3 channel headed towards Indianapolis. The 2-way port #2 heads toward the ABC affiliated station number 17 - which is also named, for convenience, by its call letters "KSTP".

The 2- way DS3 port # 5 is headed toward Los Angeles.

The Master Controller address is number 1.

Initialization

The first thing that has to be done is to initialize the network by sending commands from the Master Controller, one at a time to each Multicast Switch and Video Terminal in the proper sequence.

Initially, each Multicast Switch and Video Terminal is defaulted to think that the Master Controller address is #1 and that its own address is FFFE. This is hexadecimal for 15 ones followed by one zero - a 16-bit address. So the first thing that is done is to send a command to change the address of the New York switch from FFFE to 22. No other part of the network (except the New York Video Terminal which is treated separately) sees this command because, initially, the New York switch has no control channel connections to the rest of the network.

The next step is to send an establish network command to the New York switch, now known as #22, or NY. This command does several things. 1) It allows changing the address of the Master Controller, if desired, but we have not done that here. 2) It tells the NY switch to be the Master Timing Source for the network - that is, it is told to free run its DS3, Stratum 3 clock. This means it is a very precise clock, accurate to +/-4.6 parts per million. 3) It tells the NY switch to enable the ports numbered 0, 1, 2 and 3 toward NY, Boston, Washington and Indianapolis, respectively.

Now the NY switch has been set. The next step is to set another Multicast Switch, say Boston. This is done by first sending a command to the Multicast Switch at NY to connect the control channel from NY toward Boston and back again (the back response is on the input to a control channel gate at the NY switch and the output of the gate is connected back toward port 0, the Master Controller side - this will be explained later.)

Now a command can be sent to the Boston switch to change its address from FFFE to #21. The other switches and Video Terminals are still not connected up to the control channel so they remain as FFFE. Once the address has been changed, now the Master Controller sends a commmand to the Boston switch to establish its network. The establish network command to the Boston switch tells that switch to slave its timing to port #0, so that it will be synchronized from the NY switch's DS3 clock. It also tells it to enable its ports #0 and #1.

By now you should get the idea how the boot strap initialization takes place. Eventually all the addresses of the Video Terminals and Multicast Switches can be set in this way, and the entire network can be synchronized from the NY Multicast Switch. The Video Terminals are all arranged for loop timing - that is, the incoming DS3 clock to each Video Terminal is extracted from the bit stream and used to time the outgoing clock. The Audio Terminals, which are DS1 rate (1.544 Mb/s) terminals are also loop-timed.

You might ask, how can you send a command through a network to tell it how to synchronize itself if the network has not yet been synchronized? The answer is that the individual DS3 clocks are very precise, and although they are initially free-running, a high speed squirt of a command is very likely to get through error free.

This is also helped by a process called DS3 frame-slip buffering. That means that the network operates synchronously until its timing is off by an entire DS3 frame (actually it has been implemented as two DS3 frames).

If the incoming DS3 signal to a Multicast Switch is slightly faster than the free running clock, then, only after an excess of about 9000 bits is accumulated (two DS3 frames), a single DS3 frame will be discarded. If, the incoming signal is running slower, then instead of being discarded it is repeated.

The rate at which this happens is so slow that it is a simple matter to get a command through to the Multicast Switch even before it has been synchronized since the command is just a short spurt. If you happen to be unlucky enough to be sending a command during a frame slip then you will know that because you will not receive a response. So simply repeat the command.

Gated Operation of Control Channel

Now that you have the general idea of how to initialize the network, let me explain how the control channel network is arranged.

The idea is to create a directed tree subnetwork of the overall network. You don't want commands to arrive at the same Multicast Switch or Video Terminal by more than one path. A directed tree accomplishes this. That is, beginning at the Master Controller (the tree trunk) the control channel is connected at each Multicast Switch to branch out to another Multicast Switch, and so on and so on, without closing back on itself, for example, by excluding the two CTI links from the control channel network. No matter how complex the DS3 network is, the control channel network must always be a directed tree subnetwork. If it were to close on itself, then commands or responses could collide and be in error and therefore be ignored.

So that responses will come back to the Master Controller, the Multicast Switch includes a control channel gate and all returning control channels (from the directed tree subnetwork) are commanded to be connected to the input to the gate, while the output of the gate is commanded to be connected to the port that is headed back toward the Master Controller. Also, the internal microprocessor inside of each Multicast Switch is connected to the control channel gate input so that its response to a command will come back to the Master Controller.

Synchronization Network

The synchronization network must also be a directed tree subnetwork of the overall network. This can take the same path as the control channel directed tree subnetwork for convenience. Basically, the reason for the directed tree is the same as why two conductors cannot conduct the same orchestra at the same time.

The Multicast Switch has the provision that if there is a failure on the primary timing port it will automatically revert to a secondary timing port that it is given in the establish network command. If one is not available, then it will revert to free running. This will cause all down stream Multicast Switches to possibly free run as well, in which case there is the possibility for frame slipping.

The effect of a DS3 frame slip on a television picture is very minor. It creates an 8 line stripe in the picture, but very rarely. Once the network is resynchronized, the frame slipping disappears. Currently, it is necessary to send a command to reset the timing to the primary port; consideration is being given to automatically restore the timing to the primary port once the failure condition goes away. The process of resetting, causes a quick picture roll - which is not desirable. However, these are trouble conditions and only occur if there is no secondary port to take timing from. In an actual network such an event would be rare and hardly noticed.

Orderwire Network

Via the control channel network, a private telephone network can be established. The Video Terminal converts the human voice into a digital signal that is synchronously multiplexed into the DS3 bit stream (a small portion of it).

The Multicast Switch contains a switching fabric for the orderwire channel and so a directed tree subnetwork can be arranged for the orderwire channel. This enables all affiliated stations to communicate directly to the Master Station by a private telephone. The telephone contains a Touch-tone pad so that individual stations can be rung or by dialing an appropriate code all stations can be rung, or by just pushing a single button, all stations can be alerted while the button is depressed.

Like the control channel, the Multicast Switch contains a gate for the orderwire channel, so that whichever affiliated station is talking, his/her voice will go through to the Master Station to be heard.

An option is available in the Video Terminal to loop/through the voice at the Master Video Terminal so that affiliates can talk to affiliates via the directed tree subnetwork while still talking to the Master Station.

Because of the gated operation, the affiliate station speaking person must squeeze a push-to-talk button on the handset while talking so that any room noise will not interfere with another affiliate person talking through the gate.

Squeezing the push-to-talk button, causes the Video Terminal at that location to remove an idle code generator from the orderwire channel and instead allow the coded bits to move toward the Multicast Switch and work their way back to the Master Station. This has been found to be only a minor inconvenience - quickly learned.

The person at the Master Station does not have to push-to-talk because his coded voice does not go into the input of a gate. So he is not inconvenienced in any way.

Each telephone has been assigned a unique three-digit number for dialing between telephones.

Television Network

Remember, what this is all about is distributing and collecting television programs.

Once the network has been initialized, and the control channel, orderwire and timing tree subnetworks have been established, now the hard part is over. This won't ever have to be done again. If a new station is added, or a new Multicast Switch, this will not interfere with the already established network. The new station can be brought on-line without affecting any of the other portions of the network.

Moving television around the network is the simple part. The picture and the sound and the forward error correction bits that protect the picture from DS3 bit errors are all switched together as a single unit.

For example, to collect television from Los Angeles and bring it back to New York, ABC personnel would simply type in the following commands one at a time:

```
CP LA 0-5;
CP MINN 5-4;
CP IND 4-6;
CP ATL 4-3;
CP WASH 3-2;
CP NY 2-0;
```

Get the idea? Each command would beget a response from each Multicast Switch to confirm that the command had been received and properly acted upon.

CP means "connect payload". The letters represent the shorthand name of the Multicast Switch being addressed. The first number is an input port and the second number is the output port that that input is to be connected to. The semi-colon ends the message and the return button causes the message to be sent. Typing errors are all forgiven, you can simply back space and type over any mistakes. If you still make a mistake in format the Master Controller gives you a "syntax error" message and you just start over again.

If ABC wanted to not only receive television back at New York but wanted to send it to say Boston and Washington but without decoding it first to preserve its locked in high digital quality, then the following additional commands could be sent:

CP NY 2-0,1,2;
CP BOST 0-1;
CP WASH 2-7;

All of these commands could be sent ahead of time as programmed commands (PCP instead of CP) not to be acted upon right away but to be acted upon simultaneously when an "all stations execute" command is sent. Any Multicast Switch that has not been pre-programmed to change will not be affected.

This would cause an almost immediate complete reconfiguration of the television network to take place, limited in speed only by the speed of light and some minor processing time at each switch, but all of which takes place basically simultaneously, not one at a time.

So a complete network can be preset to change its configuration entirely in about 20 or so milliseconds (thousandths of a second), much faster than anything else that exists today. And this is not because the trial network is only eight cities. Even a large several hundred city network can change configuration in a very brief time.

One of the advantages of the two way nature of the network is the ability to do loopbacks both at Multicast Switches and at Video Terminals.

For example, while television is being sent from Los Angeles to New York, a command could be sent to the NY Video Terminal to loopback /through the DS3 payload on the line side of the NY Video Terminal and a path through the Multicast Switches could be set up from New York back to Los Angeles.

This would enable a television crew in Los Angeles to monitor the picture and sound after a round trip from LA to NY and back again so they could be sure it got there OK without affecting the quality of the picture being received in New York. In fact the quality of the picture received back in LA would be identical to the quality received in NY even though one had gone about 5000 miles and the other had travelled about 10,000 miles! - because both had gone through only a single digital encoding and decoding process.

CONCLUSION

The Digital Video Trial has been underway since November 1989. ABC has been the first television network to use it. Soon, CBS, NBC, FOX, PBS and CBC will get their turn.

More than forty companies have contributed to make this a phenomenal success story!

Not only has the technical capability been demonstrated, but all the equipment used in the trial is now commercially available from the independent manufacturing community, making this capability a potential commercial reality. And non-too-soon, considering the scarcity of launches of commercial satellites.

We encourage all broadcasters to learn more about the networking capabilities that have resulted from the success of this national Digital Video Trial which could be easily fitted into regional, statewide, or even local television networks with full user control.

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Bellcore
Document Registrar
Room 2J-125
435 South Street
Morristown, NJ 07960-1910

1. T1Y1.1/89-059 "DCT Algorithm for Broadcast Quality Encoding of NTSC Television at 44.736 Megabits/second (DS3)", and
T1Y1.1/89-058 "Proposal for a Multiplexing Structure of a Digital Video Signal into the DS3 Format" -Source for both documents: Telettra, Date: July 1, 1989

2. TA-TSY-000195 Broadcast Quality Digital Television Terminals
Appendix 1: Broadcast Quality Algorithm and DS3 Payload Structure for 45 Mb/s (DS3)
Appendix 2: Complete Video Terminal with Network Control and Order Wire Features
(Preliminary - October 14, 1988)

3. TA-TSY-000907 Customer-Controllable Multicast Switch - Generic Requirements
(Preliminary - October 19, 1988)

DS3 Digital Video Trial

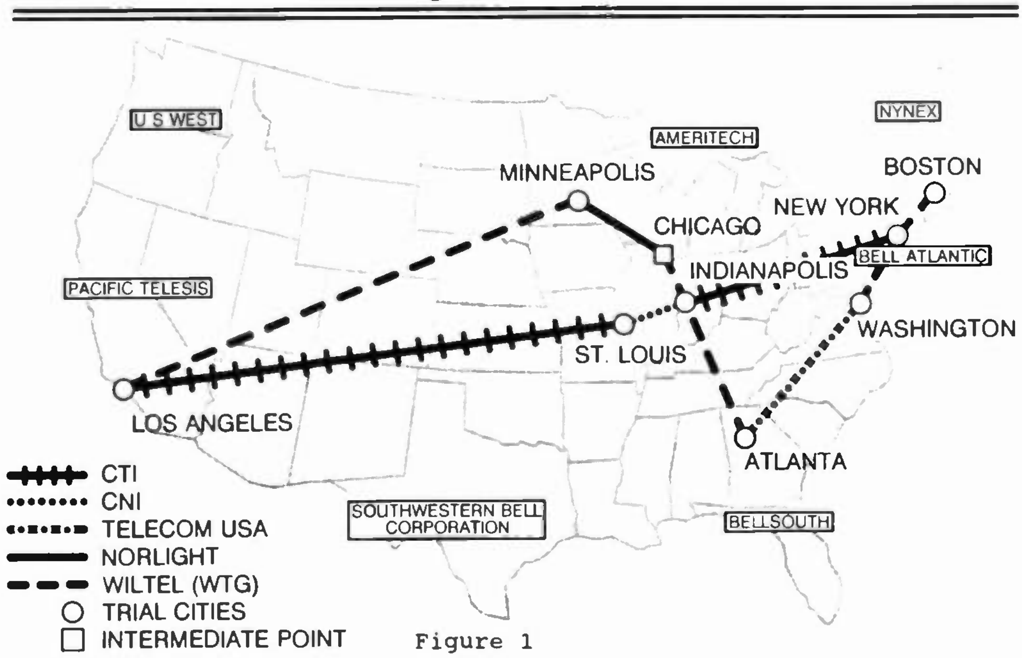

Figure 1

Trial Suppliers:

Video Terminals:
- Telettra

Audio Terminals:
- Intraplex
- RE Instruments
- NEC
- A.E.G. Bayly
- Coastcom

MulticastSwitches:
- Telinq/ADC

Master Controllers:
- public domain/ADC
- IBM-AT w/Linear I/F

Test Sets (control code observability):
- telinq
- tau-tron

Figure 2

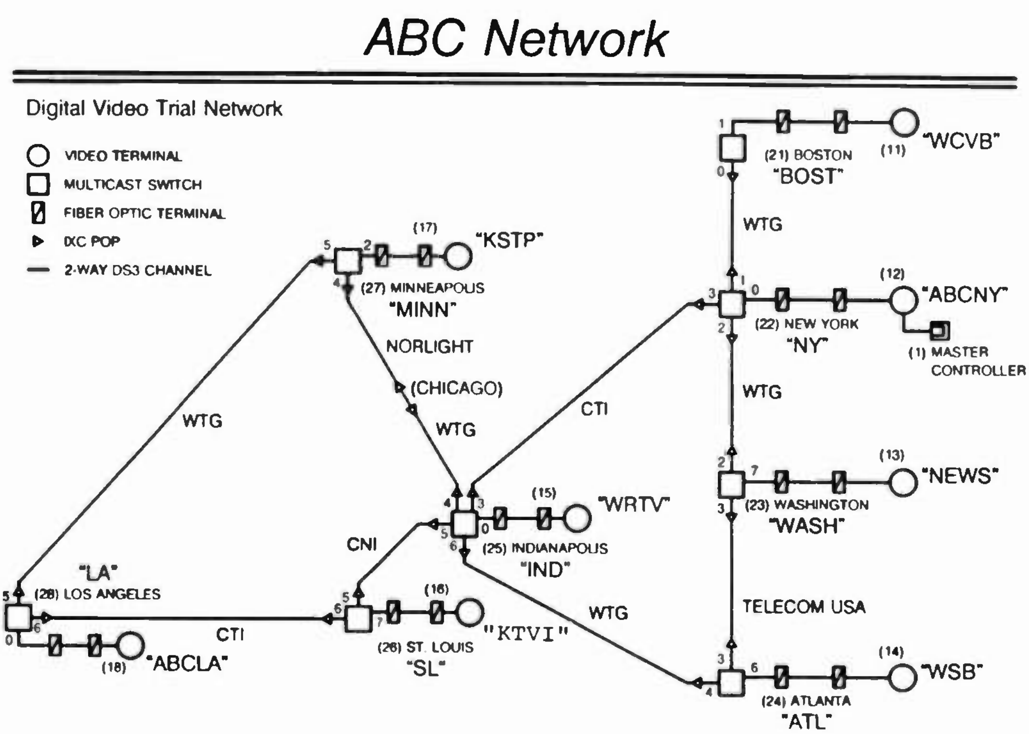

Figure 3

DYNAMIC PRE-CORRECTION OF COMPONENT VIDEO SIGNALS FOR IMPROVED NTSC COLOR ENCODED PICTURES

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A new technique to compensate for the luminance matrixing errors associated with conventional color encoders is presented. The compensation is applied to the component signals, R, B, G (which could also be derived from Y, R-Y, B-Y) prior to the color encoding process. The technique provides means to reduce errors due to the violation of the constant luminance principle, and to dynamically enhance the luminance signals of high frequency color details. The level of enhancement is a function of color and color saturation. Certain colors (e.g. blues) can be high frequency boosted by more than 18 dBs without affecting other high frequency color details and without affecting black and white details. The technology can be used with any television standard, e.g. NTSC, PAL, SECAM.

Introduction

Significant picture quality improvements to the present NTSC color system can be realized by optimizing the color encoding process. Conventional color encoders, be it NTSC, PAL or SECAM, can introduce very large luminance errors when encoding high frequency color details resulting in a loss of picture sharpness. The major defect is the violation of the constant luminance principle. The violation of the constant luminance principle is so pervasive (in fact we are aware of only one NTSC Encoder on the market that addresses this defect) that most engineers are either not familiar with this problem or have been lead to believe that it is insignificant.

In order to understand how and why the constant luminance principle is violated, it necessary to understand the principle itself. At this point it is useful to briefly review the make-up of NTSC color signals.

NTSC Color Encoding

At the source the television signal normally consists of wideband red, blue and green signals (R_w B_w G_w). For simplicity's sake we will assume that the color difference signals carried by the NTSC color signal are $(R-Y)_{\ell f}$ and $(B-Y)_{\ell f}$, (normally they are I and Q). The NTSC color signal then consists of:

$$\begin{aligned} Y_w &= Y_{\ell f} + Y_{hf} \\ (R-Y)_{\ell f} &= R_{\ell f} - Y_{\ell f} \\ (B-Y)_{\ell f} &= B_{\ell f} - Y_{\ell f} \end{aligned}$$

At the display, if one tries to recover the wideband red, blue and green signals, the best one can obtain is:

$$\begin{aligned} R_w &= R_{\ell f} + Y_{hf} \\ B_w &= B_{\ell f} + Y_{hf} \\ G_w &= G_{\ell f} + Y_{hf} \end{aligned}$$

Note that the high frequency signal is the same for all three colors, therefore high frequency details can only be displayed in black and white. Any error in Y_{hf} will directly affect picture resolution and overall picture sharpness. The violation of the constant luminance principle causes Y_{hf} to be in error by more than 10 dB in certain cases. The error always takes on the characteristic of a reduction in amplitude.

The Constant Luminance Principle

What is the constant luminance principle?[1] Very briefly, when displaying color pictures in black and white, each color is perceived as having a specific brightness or luminance value. It was experimentally determined for example that a red picture having the same signal level as as blue picture appears to be about three times brighter than the blue picture, but only about half as bright as a green picture having the same signal level. More precisely, equal levels of red, blue and green signals contribute to picture brightness in the ratios of 30/100, 11/100, and 59/100 respectively. Hence the luminance signal Y was defined as:

$$Y = .30R + .11B + .59G$$

That should guarantee constant luminance, i.e. if one were to look at a color picture on a black and white monitor, a given color scene would appear to have the same brightness or luminance level as on a color monitor.

In addition, when the present TV standards were defined, the characteristics of the luminance signal was designed to complement the characteristics of CRT displays. For a linear brightness response on a CRT, the signal applied to the CRT must be pre-distorted, (gamma corrected). Since CRT's normally had a gamma of 2.2, it was required that the luminance signal have the inverse gamma of 1/2.2. Thus, for example, if picture brightness and luminance signal are normalized to unity, to obtain a brightness of .5 on a CRT requires a signal level of .5 exp(1/2.2) = .73. For linear brightness response on a CRT with a gamma of 2.2, the luminance signal level should have the characteristic shown by the upper curve of Figure 1. Note that the lower, unity slope straight line in Figure 1 represent unity gamma.

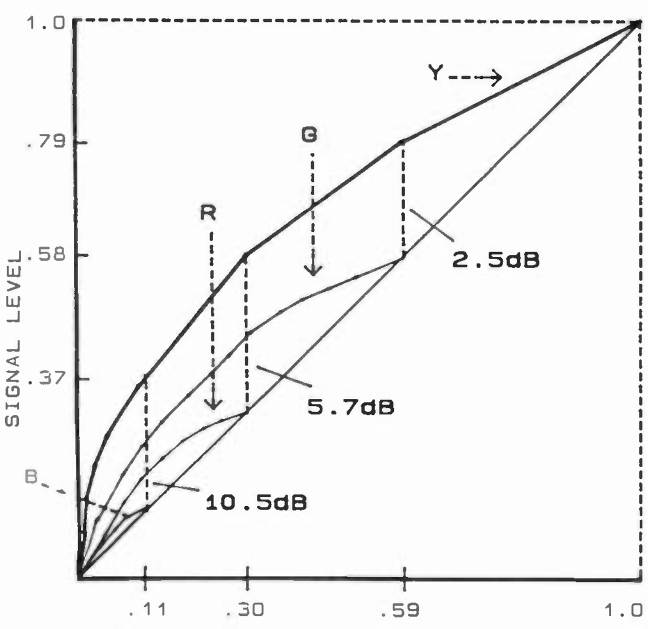

FIGURE 1. Luminance signal level characteristic for linear brightness on a CRT. Also contribution to the luminance of saturated red, blue, green signals, illustrating the violation of the constant luminance principle.

When one deals with a color signal, the red, blue and green signals at the CRT should also have the gamma of 1/2.2. Thus the red, blue and green signals are gamma corrected right at the source. The picture then will look correct when viewed on a RGB monitor, but as soon as the RGB signal is matrixed into a luminance signal the correct gamma is no longer retained.

As stated, each color is gamma corrected at the source, i.e. prior to the luminance matrix. The contribution of each saturated color to luminance then becomes as shown in Figure 1. Note, for example, that a saturate green signal at 100% level, which should have a brightness contribution of .59 according to the constant luminance principle, contributes a maximum luminance signal level of .59. But to display a brightness level of .59, the luminance signal should be .79. The error in this case is about 2.5dB. A fully saturated red at 100% level will contribute a maximum luminance signal level of .31. Yet for correct brightness, the luminance signal should have been .58. Here the error is 5.7dB. A saturated blue at 100% level will contribute 11% to the luminance level. But for proper brightness display the luminance signal should have been .37. The error is more than 10dB. Note that as the amplitude of the saturated colors is reduced, the ratio of correct signal level vs. actual signal level becomes higher, i.e. the error gets worse. Add to the above errors the fact that today's CRT's have gamma's greater than 2.2 and the result is an even larger violation of the constant luminance principle. Incidentally, it should be clear, that if one deals with black and white information, i.e. with equal red, blue and green signals, the luminance matrix will work perfectly well, and assuming the CRT gamma is 2.2, the constant luminance principle is satisfied. Note also that at low frequencies, the original R_{lf} B_{lf} and G_{lf} signals are always available and gamma is correct.

In addition to the incorrect matrixing of color information resulting in high frequency amplitude errors, higher than 10dB with some colors, the human eye's response for high frequencies exhibits a worse roll-off in darker areas. Thus the reds and the blues, which have low luminosity, take a double beating at high frequencies: first because of the violation of the constant luminance principle, and second because of the loss of perception by the eye of low luminosity details. It should not be surprising then that NTSC encoded pictures look so much softer than the original RGB source pictures.

Techniques For Compensating the Violation of the Constant Luminance Principle

A number of techniques have been proposed to compensate for the violation of the constant luminance principle, and in fact some of those techniques are an integral part of the Improved and Enhanced TV systems being proposed for the future.

A straightfoward technique to avoid the violation of the constant luminance principle is to generate linear color signals for high frequency color details, obtain a linear high frequency luminance signal using the standard luminance matrix, than gamma correct the high frequency luminance signal.

Another technique to correct for the violation of the constant luminance principle is to decode the NTSC color signal at the color encoder output, compare the decoded signal to the original wideband red, blue, green signals, obtain an error signal and compensate the encoded signal using the inverse of the error signal. These techniques are quite effective, but they are costly and difficult to implement properly. In addition they cannot provide significant high frequency saturated color enhancement. There is however a relatively new, patented technique[2], that permits one to compensate for the violation of the constant luminance principle and, at the operator's discretion, apply as much as 19dB of of enhancement to certain color details as a function of color and saturation. It is called Color Intelcomp (for intelligent color compensation).

Color Intelcomp

Color Intelcomp is a detail extraction and insertion technique that permits one to obtain high level luminance boost for low luminosity colors without significantly affecting black and white details. Figure 2 is a simplified block diagram of Color Intelcomp.[3] High frequency R,G,B signals, as well as luminance, are inputted into a mixing circuit. The mixer output is such that if only one input contains information (details), the output is equal to that input. If details appear in more than one input, the output is a

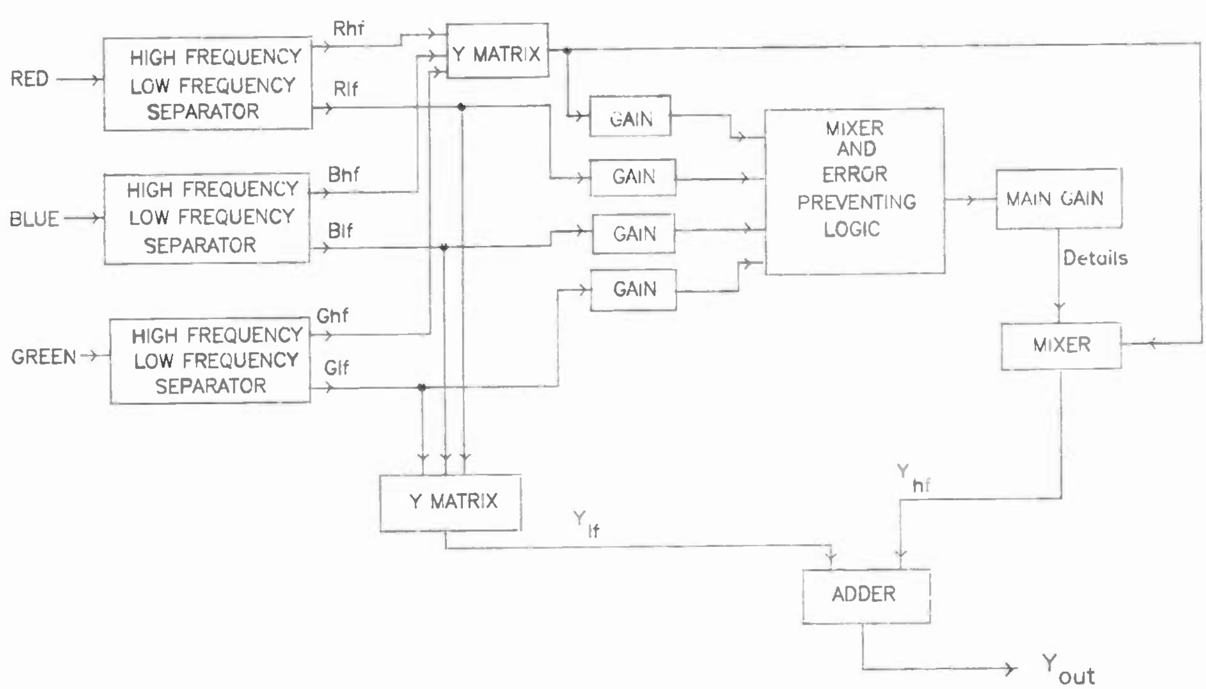

FIGURE 2. COLOR INTELCOMP

combination of those inputs, but it never exceeds the highest of the inputs. Thus we see that the output dynamic range never exceeds the input signals dynamic range (a very important feature when dealing with enhancement). The first case, that of a single input, is approximately the case of saturated high frequency blue details, where the blue signal contribution to luminance is almost insignificant. We can see that, in that case, we can obtain an output detail signal that can be as high as the maximum dynamic range of luminance. Because, in the luminance matrix, blue only contributes 11%, we see that the detail signal generated in this system can be as much as nine times higher than it would normally be (a 19dB boost!), and yet be within the system's dynamic range. A similar analysis can be applied to the other colors. Obviously, additional precautions need to be taken to prevent generation of details with an incorrect phase. Here the availability of the conventional luminance signal permits one to determine and prevent the generation of details signals that are out of phase with the luminance signal.

Color Intelcomp generates luminance details signal that can be used to compensate for the violation of the constant luminance principle as well as to enhance high frequencies from individual colors. The detail signal is mixed (not added!) into the conventional high frequency luminance. This implies that the dynamic range of luminance is not changed, and most important, when luminance is the predominant signal, practically no noise contribution from the detail signal is added to the luminance. This is of outmost importance and we wish to emphasize it. While the system can deliver, as we have seen, up to 19dB of high frequency boost under certain conditions, this can be achieved with practically no noise penalty.

Compensation and enhancement of saturated color details is a process that needs to be done at the encoder, but it is not practical to retrofit presently used encoders to compensate for the violation of the constant luminance principle. Fortunately, the Color Intelcomp technique can be separated from the encoder and applied to the component signal prior to NTSC color encoding.

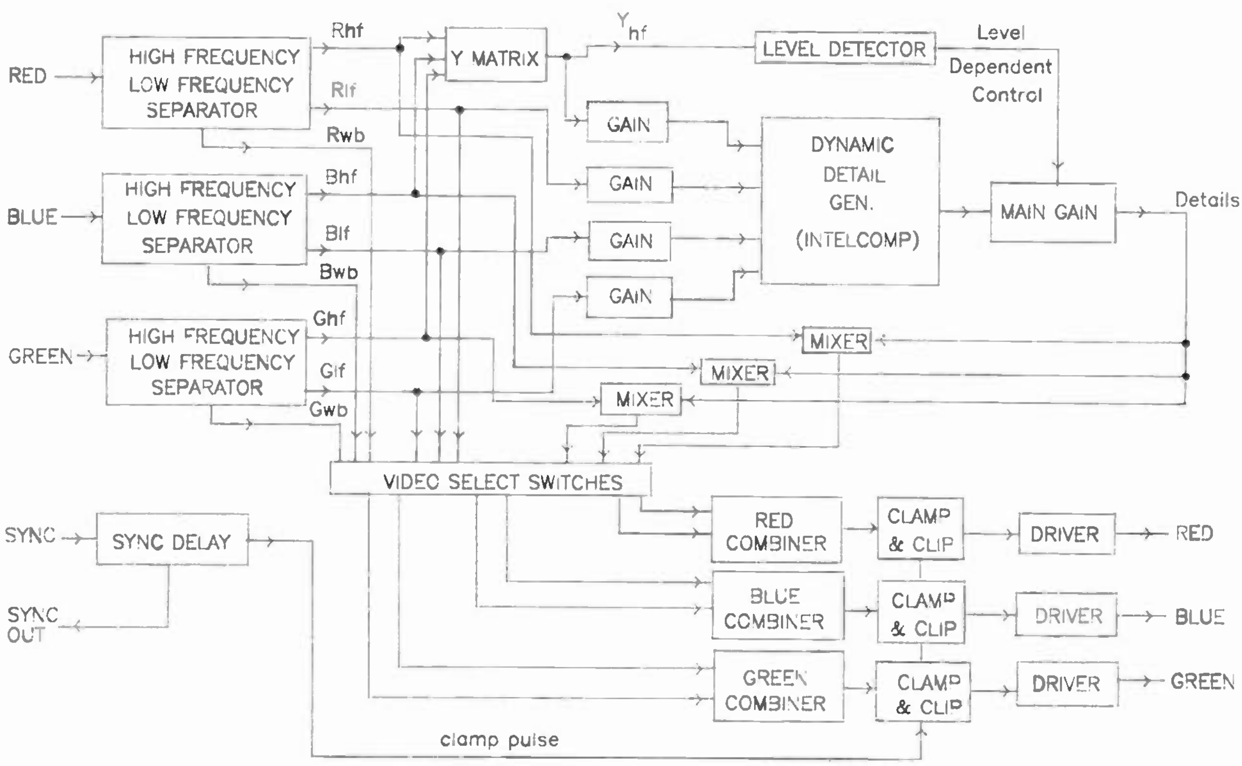

FIGURE 3. PRE-CODER BLOCK DIAGRAM

Compensation/Enhancement Using a Pre-Coder

As we have seen from the equations for the color encoding process, the high frequency details are never color signals, but they are the result of matrixing high frequency R,G,B details. Thus, if we wish to compensate or enhance the high frequency details generated in a color encoder, we can feed those details to the encoder via its R,G,B inputs. This, in effect is how Intelvideo's Pre-Coder, using Color Intelcomp, delivers the compensation and enhancement signal to the encoder.

Figure 3 is a complete block diagram of the Pre-Coder. Note in particular how the detail signal from Color Intelcomp is inserted into the R,G,B channels. The details are "inserted" or mixed into all three color channels and thus even after the color encoding process, their full luminance amplitude is retained. It should be emphasized once more that details are not being added to the color signals. This has a very significant implication on signal-to-noise. The mixing process used is similar to an OR gating process; it does not change the dynamic range of the signal and it prevents additive noise contributions.

The use of the Pre-Coder for dynamic compensation/enhancement results in significant NTSC color picture improvement. Incidentally, it should be pointed out that, as the Pre-Coder signal processing is essentially independent of the TV standard (it simply processes high frequency details), it is equally applicable to NTSC, PAL, SECAM, as well as to other non-standard systems.

The Pre-Coder we have described is an exclusive Intelvideo product. It was introduced in the Fall of 1989 and it has become an instant success. Several units have already been installed in video production facilities. The Pre-Coder also provides means to independently and remotely control the level of enhancement used for each primary color as well as luminance. That permits operation under computer control, as a function of source signal.

The Pre-Coder can be seen in operation on the Exhibit floor at Booth 6509.

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TRANSMISSION LINE MAINTENANCE USING A HIGH POWER REFLECTOMETER

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The changing complexion of today's radio and television facilities require new and innovative technology to meet new demands. The evaluation of transmission lines is not excused from this requirement. Longer line runs and the clustering of several transmitting antennas at one location have severely taxed the capabilities of the time domain reflectometer. A new instrument, manufactured by Delta Electronics, offers an inexpensive solution to these problems. It gives both radio and television broadcast engineers the ability to assess transmission line conditions, taking corrective action before a catastrophic flashover occurs. Dubbed the "Heart Monitor", this instrument permits routine monitoring of the heart of the broadcasting facility - its transmission system.

Time domain reflectometry has been with us for many years, helping engineers locate faults on transmission lines. The traditional time domain reflectometer suffers from several drawbacks brought on by a change in broadcast facility operation. A relatively weak pulse of energy is coupled into the line to be measured. As broadcast facilities improve their market position by transmitting from taller tower structures, the length of line this pulse must traverse approaches limits that raise questions as to the validity of the displayed response. In order for the instrument to display these weak echoes, the receiving front end must be made very sensitive. Herein lies another drawback for today's broadcasters - co-located antennas from several stations. When the system under test is being measured, that station's transmitter is not exciting the line. However, RF from other transmitting antennas using the same tower structure can still be coupled into the line under test. The result of any appreciable RF coupled into the line under test is damage to the sensitive front end of the time domain reflectometer.

Both of these problems have been addressed in an innovative High Power Pulse Reflectometer, developed by Delta Electronics, Inc. of Alexandria, Virginia.

The PRH-1 was developed as a solution to a problem that was experienced by the Voice of America. Their HF transmission sites consist of very long runs of transmission line, fed by million watt transmitters. The RF field coupled into adjacent antennas made the use of traditional TDR impossible.

THE PRH-1 HIGH POWER PULSE REFLECTOMETER

Figure 1

As radio and television broadcasters have clustered transmitting antennas together in recent years, the need to perform routine evaluation of the station's transmission line - but without everyone on the tower turning their transmitters off - has become more apparent, and in many markets, nearly impractical.

Since the PRH-1 was designed to operate in the high field of antenna farms or co-located transmitter sites, the transmission lines and antennas measured for this paper occur in such locations. Though the instrument will give an adequate picture of a transmission line's condition when no interfering signals are present, more impressive are the pictures obtained when the pulsed line is located amidst high power FM, VHF, and UHF antennas.

Before we investigate the lines measured, a brief explanation of the pulse reflectometer is in order. This instrument is designed to measure unbalanced transmission line systems. Accessories permit coupling of the unit to balanced-line systems as well. A high voltage, short duration pulse is coupled to the line under test. An oscilloscope permits examination of the initial pulse and any echoes in a time versus amplitude display. These echoes originate from faults, line discontinuities, and from the antenna or load that terminates the line.

Evaluation of the echoes offer two benefits to the engineer. First, specific faults or discontinuities display distinct characteristics. Also beneficial is the ability to measure the distance to the fault using the time calibration of the oscilloscope display.

The PRH-1 generates a pulse by charging a ceramic capacitor to a high voltage, and then discharging it through the transmission line under investigation using a pair of ball gaps. At the point of discharge, the spark gap acts as a low impedance switch, transferring the energy to the line. As this pulse of energy travels down the line, any irregularity or imperfect termination will cause a portion of the energy to be reflected back to the PRH-1. This reflected energy is absorbed by two high power terminating resistors. An RC sampling circuit connected across the termination resistors provides a sample of the initial pulse and any echoes (or reflected pulses) for viewing on an oscilloscope. The length of time required for pulses to propagate down the line and back to the sampling circuit is used as a measure of the range. By using suitable horizontal deflection rates, the line characteristics will be displayed linearly along the horizontal scale of the oscilloscope with the range of each separate discontinuity of fault measured by its horizontal displacement from the original pulse.

A front panel DC voltmeter measures the amount of voltage used to charge the capacitor. The full scale sensitivity of this meter is 5 kilovolts. The very short pulse duration makes for a very low average power level so the output is not hazardous. The current level is so low that placing one's finger across the output connector will result in little or no sensation. 5 kV pulses may be applied to small transmission lines without damage. Before testing very small diameter coaxial cables, however, verification of the cable's breakdown voltage is advised. For cables used in routine radio or television operation, however, the PRH-1 should have no adverse effect. The front panel control that varies the voltage works in conjunction with the ball gap spacing to determine the pulse height, repetition rate of the pulse, and the intensity of the oscilloscope trace.

The display range is adjusted by selecting different horizontal deflection rates on the oscilloscope. A sweep rate of 0.1 microseconds per centimeter translates into a range scale of approximately 50 feet per centimeter for air dielectric line. If the velocity factor for the line under investigation is known, the range would be 50 feet times the velocity factor per centimeter of display. Most commercial air insulated lines have velocity factors approaching unity, therefore no correction is required. The range data for other types of line and other oscilloscope sweep rate settings is displayed below.

DISPLAY RANGES FOR VARIOUS
TYPES OF COAXIAL CABLE

Sweep Rate us/cm	Display Range - Ft/cm Open Wire	Rigid Coax	"Heliax"* Coax	Foam Coax
1.0	492	485	458	389
0.5	246	242	229	194
0.2	98	97	92	78
0.1	49	49	46	39
0.05	24	24	23	19

*Heliax is a trademark of Andrew Corp.

Figure 2

The oscilloscope will display the initial pulse on the left side of the CRT. Echo pulses will be seen along the horizontal axis. The range to a specific fault echo is determined by measuring the distance in centimeters from the oscilloscope graticule from the leading edge of the initial pulse to the leading edge of the echo pulse and applying the appropriate factor from the figure above.

Specific types of faults have distinct characteristics, as can be seen in Figure 3 below. By storing photographs or plots of the scope displays for specific transmission lines, a history of the transmission system can be maintained. In addition to cataloging the pictures, a record of the oscilloscope settings and pulse height setting for each photograph is useful. Each antenna will have a characteristic signature as the pulse resonates the termination. Through comparison of the echoes viewed during a line inspection with those in the catalog, the engineer can quickly ascertain the transmission system performance, and in many cases, spot developing faults before they destroy the line.

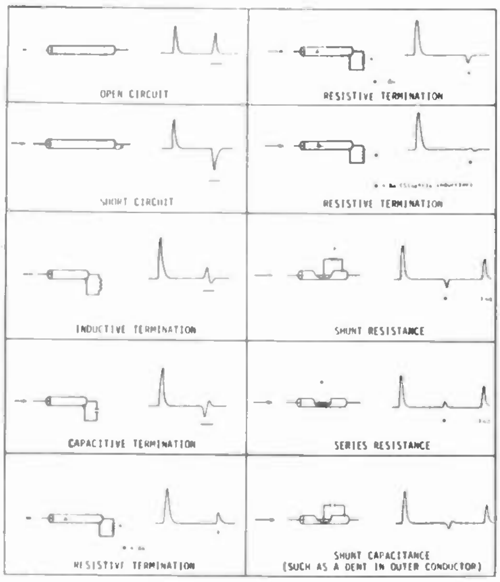

Figure 3

It is important to remember that the oscilloscope display is a time record of the transmission line system under investigation. The pulse generated by the instrument is of short duration, typically 30 nanoseconds, and of a fast risetime - 5 to 10 nanoseconds. Such a pulse therefore contains energy over a wide bandwidth. The system bandwidth is only limited by the high frequency response of the sampling circuit and the oscilloscope used. It is therefore not practical to quantitatively interpret the display in terms of performance over a specific band of frequencies, such as VSWR versus frequency. Several additional corollaries must be considered when evaluating the oscilloscope displays. An open circuit or short circuit on a transmission line will cause very large echo returns. Inductive and capacitive faults can be discovered by remembering that the highest frequencies of the pulse are on the leading edge. An inductive termination would appear to be an open circuit at the leading edge of the pulse since its reactance is high for high frequencies, and a short circuit later since its reactance is low for low frequencies. A capacitive termination would display a short circuit at the leading edge and an open circuit at the trailing portion of the pulse due to its frequency sensitive characteristics. When an element is placed on a line at a position other than at the end of the line, the element is shunted by the characteristic impedance of the remainder of the line. Such would be the case of a shunt capacitance, where the outer conductor of the transmission line has been dented.

Because of the "frequency insensitivity" of the PRH-1, the instrument can be used in a variety of communications settings. Starting with the AM broadcast band, the PRH-1 may be used to not only evaluate the main feedlines to each tower in a directional array, but also in determining the performance of the sampling system. In a time where broadcast facilities are bought and sold in a relatively short period of time, it becomes more difficult for the Director of Engineering of the purchasing company to adequately ascertain the condition of the "new" facility. Up-to-date notes on the spliced cable or crushed transmission line may no longer exist, and what's buried underground or run 500 to 1000 feet in the air is difficult to examine.

Such was the case for this non-type approved AM Sampling line shown below. A shift of directional parameters after the spring "thaw" led the engineer to suspect the sampling system. Evaluation of the sampling lines disclosed a trouble spot.

Approximately 25 feet from the transmitter building the transmission line had been spliced. Water had contaminated the splice, resulting in the apparent parameter shift.

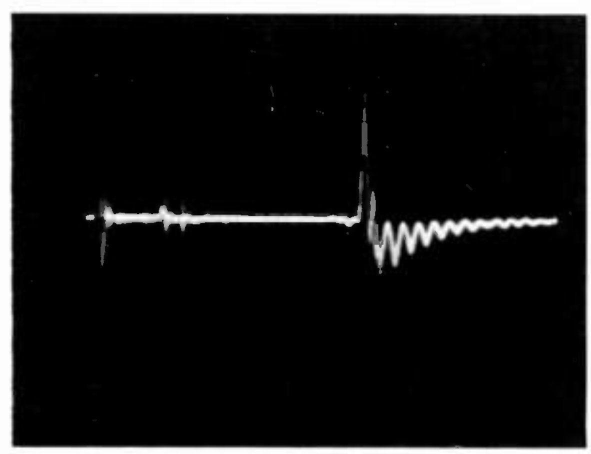

Figure 4

The sounding of this sampling line is explained as follows. The initial pulse is located at the left of the photograph. The relatively slow horizontal scan rate of the oscilloscope makes the short duration of the initial pulse hardly visible. For convenience purposes, a 100-foot test cable was used to connect the PRH-1 to the input of the sampling line, after removing it from the antenna monitor. The contaminated splice can be seen approximately 25 feet beyond the end of the test cable. The pulse continues down the transmission sampling line until it reaches the termination, a sampling loop. The damped, "oscillating" characteristic is the result of the pulse exciting the termination - in this case, a loop antenna. Similar to striking a bell, once struck, the bell will continue to ring - or resonate. The sampling loop exhibits the same characteristic as the bell. It has been struck, or "thumped" with a pulse of energy. It returns a picture showing its resonant characteristics.

Another example of antenna characteristics, which is similar to the sampling loop is the picture below of an FM antenna. Using a time base of 0.2 microseconds per division and a 1 volt per division vertical input, this sounding demonstrates a normal performing antenna connected to 3-1/8 inch Heliax.

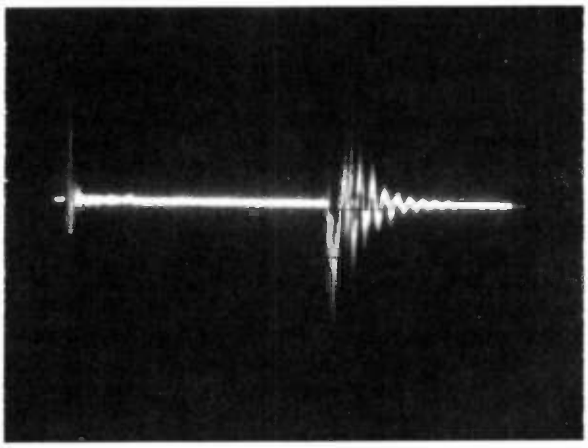

Figure 5

Again, the input pulse is very faint, due to its fast risetime. The 100-foot test cable has again been used to couple the signal through a reducing cone and into the 3-1/8 inch cable. The smoother line demonstrates there are no dents or faults. As with the sampling line, the pulse strikes the termination and the antenna begins to resonate.

Not all antennas demonstrate such graceful damped "textbook-type" waves. Figure 6 describes the main antenna used by this FM station.

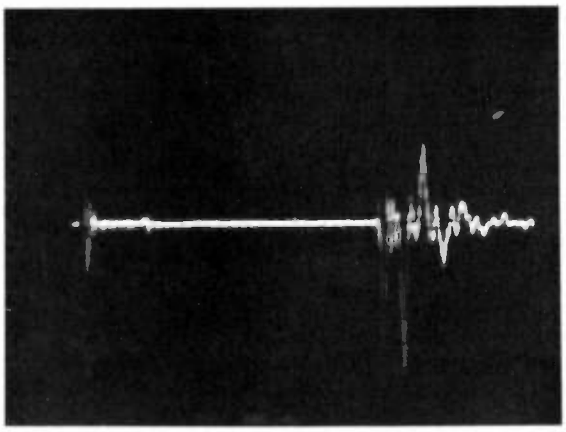

Figure 6

Again, the test cable can be seen following the initial pulse to the left. From the test cable, the pulse enters the transmission line and travels to the antenna input. As with the auxiliary antenna, a time base of 0.2 microseconds and vertical input of 1 volt per division were used.

The measurements we have seen thus far were taken at an AM/FM site. When the FM lines were pulsed, both the 5 kW AM and 50 kW FM transmitters were operating. Of course measurements cannot be made while the transmitter feeds the transmission line of interest. A coaxial transfer switch permitted pulsing of the main FM line while the station operated on its auxiliary antenna. The auxiliary line was viewed while the station operated with full power into the main antenna. No degradation of the echoes or damage to the PRH-1 was noted. Furthermore, operation of the PRH-1 had no listenable effect on the transmitted signal as we monitored in the field.

The ringing of television transmission lines is equally enlightening. Our first sounding took place at a site where two 50 kW FM's, a 5 MW UHF TV, and two 2.25 MW UHF TV transmitters operated. In addition, this tower supported numerous two-way and microwave links. We viewed the transmission line of one of the 2.25 MW UHF TV stations. Measurements were made during a scheduled maintenance period, but when the other stations were still transmitting.

Though this station had experienced no problems with their line, a peculiar "hump", not seen in other line soundings, was discovered.

As can be seen, this echo does not represent the echoes caused by elbows or back-to-back elbows. Using the transmission line installation drawing and the range measuring capability of the PRH-1, it was determined that the fault was located at a junction of two pieces of transmission line. Subsequent disassembly by a tower crew disclosed a bullet that had heated and was in the process of deteriorating. It was interesting to note that the station had not experienced any indication of high VSWR or poor transmitter performance prior to discovery of this fault.

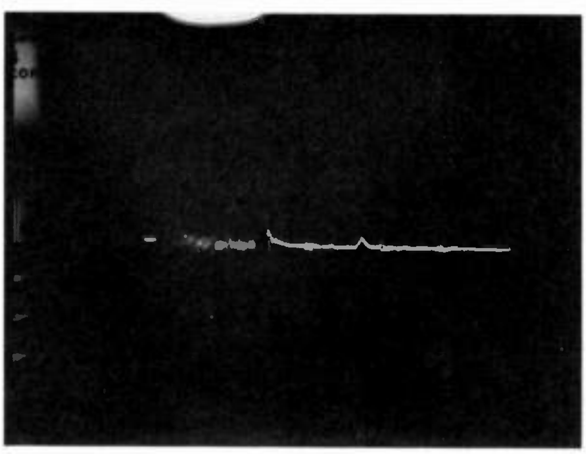

Figure 7

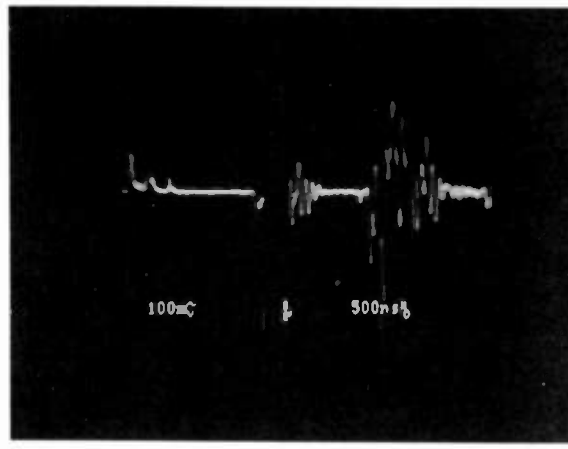

Figure 8A
VHF LEFT LINE

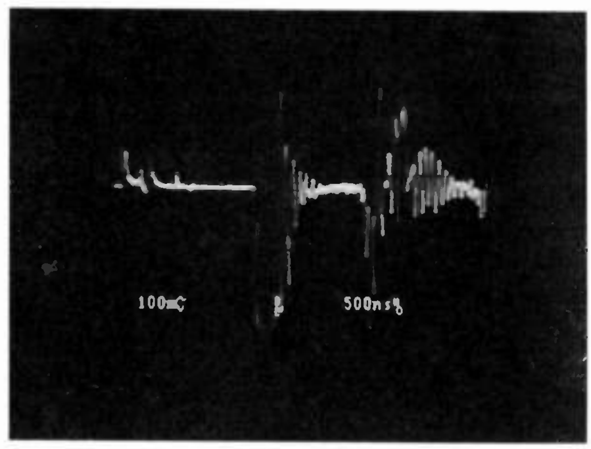

Figure 8B
VHF RIGHT LINE

Stations utilizing parallel transmission line runs to the antenna allow the engineer a better frame of reference, since each line can be compared to the other. The photos below represent a VHF station. This compressed view takes us from the input to the patch panel - on the left - out to the tower and then up to the antenna. The first "ringing" pattern is the antenna itself. The termination can produce multiple echoes, and the "second" ringing pattern is just that. For purposes of analysis, these multiple termination echoes can be ignored.

What is interesting, however, is the small bump which appears to be a shunt capacitance. It was found that at the base of the tower, the right transmission line was dented. Though the left line was only slightly deformed, the right line was definitely depressed. Falling ice or perhaps a tool accidentally dropped from the tower were two explanations.

These measurements were made with another VHF station and three FM stations all transmitting from the same tower. Other field measurements were made at a VHF station which was diplexed with another VHF station into the same antenna. Attempts to view "station A" while "station B" was still on the air were only moderately successful. Beyond the reject filter, the amount of energy coupled into the oscilloscope made line assessment difficult. However, measuring "station A's" auxiliary feedline, mounted on a tower next to the main antenna, proved to be no problem. When "station B" signed off, measurement of the main feedline was straight-forward. At no time did the absorbed energy pose a problem for the PRH-1, which can withstand up to 1 kW of power on an intermittent basis.

An alternate means of diagnosing transmission line performance has been presented in this paper. Several of the major drawbacks of traditional time domain reflectometers have been overcome. These include the high cost of the instrument, its weak pulse, and inability to withstand high induced RF fields. Through systematic documentation of a station's transmission line systems, problems can be discovered before they develop into catastrophes. The transmission line is perhaps the most neglected component of a broadcast transmission facility. That the engineer has been unable to "see" inside the line - except through costly disassembly - probably accounts for most of the neglect. The PRH-1 opens the door to a new type of routine transmission line maintenance, and one that can be economically adopted by broadcast stations, consultant and contract engineers, as well as tower crews.

ACKNOWLEDGEMENTS

The author wishes to ackowledge the support of the stations comprising W.E.B.E. - the Washington Executive Broadcast Engineers - for their assistance in providing radio and television facilities where many of the measurements discussed in this paper were made. Additional thanks given to Chris Wilk for preparing the photographs used in this paper, and Michelle Wilk for completing the finished manuscript.

TRUE 3-DIMENSIONAL BROADCAST TELEVISION WITHOUT GLASSES

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ABSTRACT

The VISION III (TM) Autostereoscopic processes for texture and depth enhancement make possible the broadcast and viewing of video and motion picture images in full color with a true three-dimensional illusion without the use of glasses. The images can be recorded on standard video tape or motion picture film (in any format) using a VISION III camera or camera attachment, transmitted from any broadcast or cable television station, received in the home and viewed on a standard TV set without modification or the need for additional equipment. A VISION III video signal is identical to a standard video signal and meets all FCC requirements.

INTRODUCTION

To most people the term 3-D means exaggerated depth as displayed in a motion picture theater using hyper- or super-stereoscopic methods. This is accomplished by increasing the disparity between cameras to a distance that exceeds the human inter-ocular distance. This effect produces a pseudo-three-dimensional effect beyond what the eyes naturally perceive. VISION III is a true three-dimensional process that produces a natural and realistic depth and for that reason is referred to as "texture and depth enhancement" rather than 3-D.

True three-dimensional display methods provide the viewer with parallax as well as monocular depth information. They can be divided into two categories, stereoscopic (binocular) and autostereoscopic.[1] For the viewer, conventional stereoscopic television systems usually require the use of special glasses, special display equipment, or both. Examples are Anaglyphic (red/green glasses), Pulfrich and shuttering methods. The necessity for special screens or glasses makes these methods either non-broadcast compatible or awkward to use. These methods also can cause eye strain or headaches if used for extended lengths of time. Autostereoscopic systems do not require special viewing aids (glasses) or special viewing screens. VISION III is the only commercial quality system for autostereoscopic television broadcast. Other systems have been suggested and patented but the author is not aware of any demonstrations of commercial quality imaging.

A great deal of effort has been devoted to developing stereoscopic television display hardware. The approach has been mostly mechanical -- an apparatus is used to force each eye to see a different perspective of the image. While the depth effect from these methods can be spectacular, they are impractical for broadcast television with home viewing in mind. Surprisingly, little effort has been devoted to understanding and developing "Gestalt" broadcast and projection methods of depth perception, i.e. it is not enough simply to increase image resolution; the brain relies upon other cues for image interpolation. What is needed to attain a Gestalt approach is to incorporate into current broadcasting techniques, other visio-psychological perceptual information. This is the basis of VISION III imaging and the subject of this paper.

HISTORICAL BACKGROUND

Ever since early man scratched pictures on a cave wall, man has sought to preserve images of the world in which we live. As early as the Tenth century Ibn Al-Haytham (Alhazen), an arabic astronomer, experimented with the perceptual issues of binocular vision and image recognition in his book Optics. The Greeks used the depth cue of perspective in paintings to portray a three-dimensional scene on a two-dimensional surface. Over the centuries artists experimented and developed chiaroscuro techniques of light and shadow to add a greater sense of depth and reality to their paintings.

In the early part of the nineteenth century artistic skills were augmented with mechanical and chemical advancements. Charles Wheatstone, who invented the Stereoscope in 1832, realized that each eye viewed an object from a slightly different horizontal point of view. His device used a pair of drawings of right and left views (stereograms) and presented them to each appropriate eye.

The discovery of the physiological phenomenon of persistence of vision lead to the invention of parlor devices like William Horner's Zoetrope (patented 1834) that allowed the viewing of images with the illusion of movement.

The invention of Daguerreotype photography in 1839 by Louis Daguerre, as well as the subsequent development by William Henry Fox Talbot of a system of negative recording and positive reproduction, allowed for the accurate documenting of real images on a two-dimensional surface. In 1849 the Scottish physicist David Brewster developed a convenient device for viewing stereoscopic photographs. Today's Viewmaster can be traced to Brewster's stereoscope.

In 1884 the German engineer Paul Nipkow patented the first attempt at a television system (Fig. 1). His device proposed the use of a selenium cell with a mechanical, spirally-apertured, rotating disc at both the sending and displaying units. Selenium proved to be too slow in its response to light intensity for image transmission.

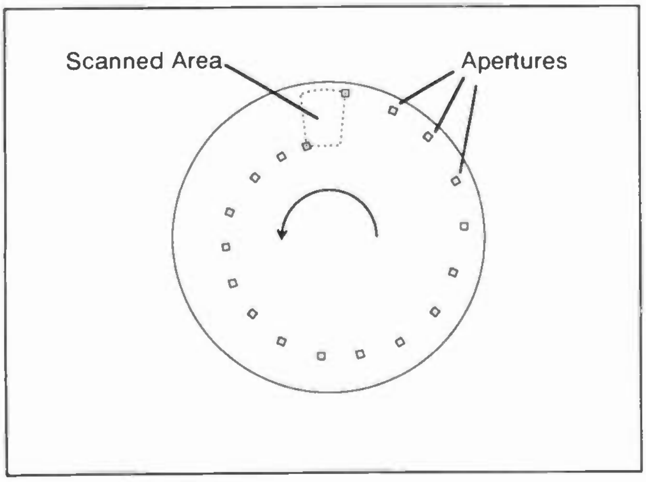

Figure 1. Nipkow disk

The later part of the 1800's saw the invention of flexible photographic film by George Eastman and a workable motion picture camera/projection system by Thomas Edison's New Jersey laboratories. On December 28, 1895, the Lumieres brothers held the first public screening of Cinematographe films at the Grand Cafe, Boulevard des Capucines, Paris, and the Movies were born.

A. A. Campbell Swinton, a Scottish electrical engineer, outlined in 1906 a method that fundamentally is the basis of modern television. He proposed the use of cathode-ray tubes, magnetically deflected, at both the camera and receiver. Unfortunately for Swinton, the technology of 1906 was inadequate for applying his ideas.

In 1912 Max Wertheimer published a paper describing what he called the phi-phenomenon. The phi-phenomenon is a psychological visual illusion in which static images shown in rapid succession appear to have movement by exceeding the threshold in which they can be perceived separately. The phi-phenomenon is experienced in viewing motion pictures and television. Wertheimer was of the Gestalt school of psychology. Gestalt psychology, a 20th century German/Austrian movement with a holistic approach to the scientific study of mental processes, provided the foundations for the modern study of perception.

By 1920, Hollywood dominated the world film industry and the movies had become a popular form of entertainment. In 1923 a Russian immigrant to the U.S., Vladimir K. Zworykin, filed a U.S. Patent for the "Iconoscope" camera tube, and by 1932 the Radio Corporation of America (RCA) demonstrated a 120-line resolution, all-electronic television system. By 1935 Germany began regular television broadcasts with 180 lines of resolution.

A patent was issued to Zworykin in 1938 for reproducing stereoscopic television images. The system incorporates a vertical parallax grating that allows each eye to see only its appropriate view and was the first attempt at three-dimensional television. Over the following years numerous methods were developed for displaying stereoscopic images in motion pictures and television. They all had one thing in common -- special screens and/or glasses.

AUTOSTEREOSCOPIC BACKGROUND

In 1967, K. N. Ogle observed that left and right eye depth information can be presented alternately to the left and right eye, resulting in depth perception as long as the time interval did not exceed 100ms (Fig. 2A & 2B).[2] This is the basis of current shuttering technologies which are stereoscopic and not autostereoscopic. Ogle's discovery is important because it proved that the brain did not need simultaneous left and right images to perceive depth.

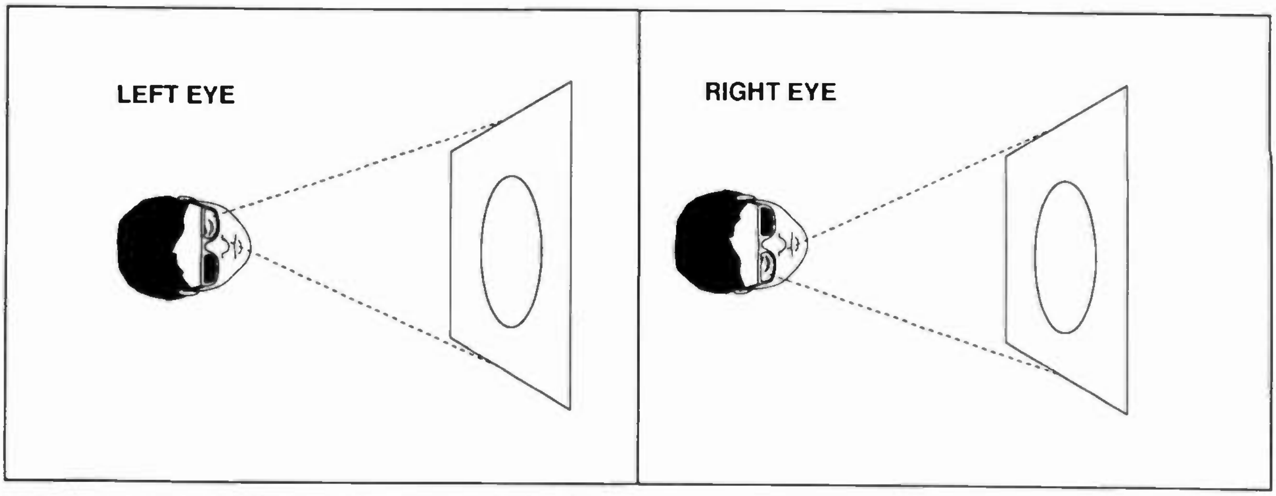

Figure 2a & 2b. Alternating Stereoscopic Information

Donald Imsand was granted a patent in 1974, for a three-dimensional display system that was glassesless (the term autostereoscopic was not coined until 1980).[3] The Imsand method used two cameras to record two different views. Camera one provided an extended display of a particular view, which was intermittently interrupted by camera two's other view for a brief duration resulting in some sense of depth.

In 1982, CJM Associates announced the development of the VISIDEP system for glassesless, three-dimensional broadcast television.[4] This announcement was widely covered by the media, and examples of VISIDEP were broadcast by the networks. The VISIDEP images did have depth when displayed on an ordinary television set, but the images were unstable and possessed a distracting, rocking motion. CJM attempted to control the rocking by introducing a mixing technique to the system.[5] This did little to control rocking, and compounded the problem with intermittent image softening.

By early 1985, the author began experimenting with the CJM methods and other autostereoscopic methods using stop-motion film animation that allowed total control of the image, frame-by-frame. After hundreds of film and video tests over a two-year period, methods were developed that resulted in stable texture and depth-enhanced autostereoscopic images.[6]

HUMAN DEPTH PERCEPTION

Depth perception is not always dependent on seeing three-dimensional scenes. It is possible to perceive depth in a two-dimensional space without parallax depth information. This monocular depth is perceived when viewing a still photograph, a painting or watching a standard movie or television show. The depth perceived on the screen is read into the picture by the viewer through a processes learned from childhood and is thought to be dependent on their cultural and sociological background.[7] In such cases monocular or two-dimensional depth cues are used by the viewer to determine depth. Some examples of two-dimensional depth cues are:

> Relative size: the fact that a recognizable object is larger in the picture and therefore closer than another known object of equal size which is smaller and therefore farther away (Fig. 3).

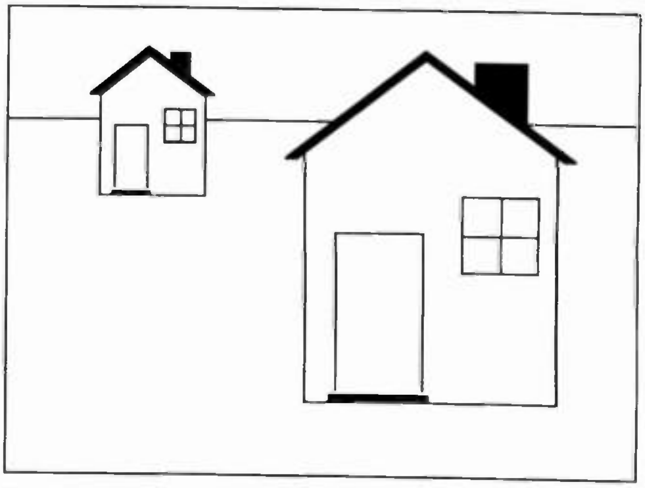

Figure 3. Relative Size

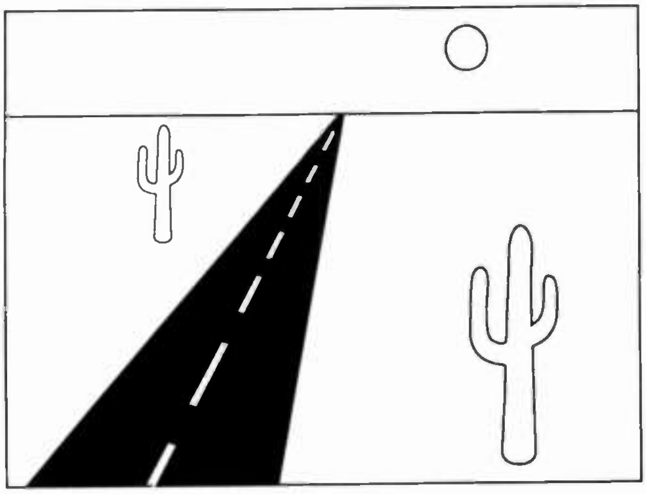

Figure 4. Linear Perspective

Linear perspective: is when lines appear to converge in the distance like railroad tracks seeming to come together as they get further away (Fig. 4).

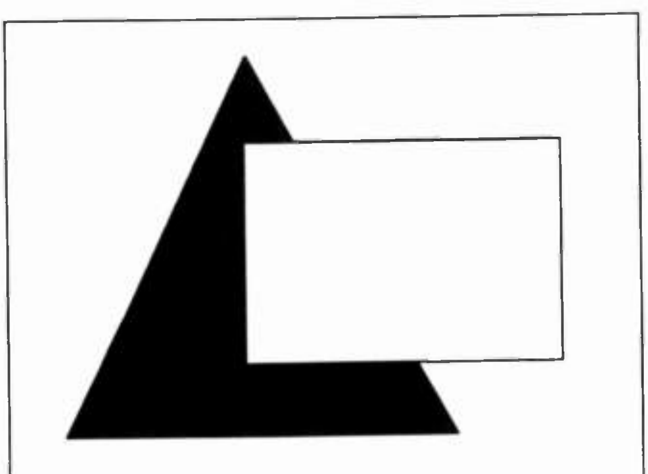

Figure 5. Overlapping

Overlapping: is when two objects are along the same line of sight, and one object (front) partially conceals another object (behind) (Fig. 5).

Shading: provides information about the shape and texture of a surface (Fig. 6).

These and other two-dimensional depth cues, along with depth information provided by parallax views, produce a true three-dimensional image.

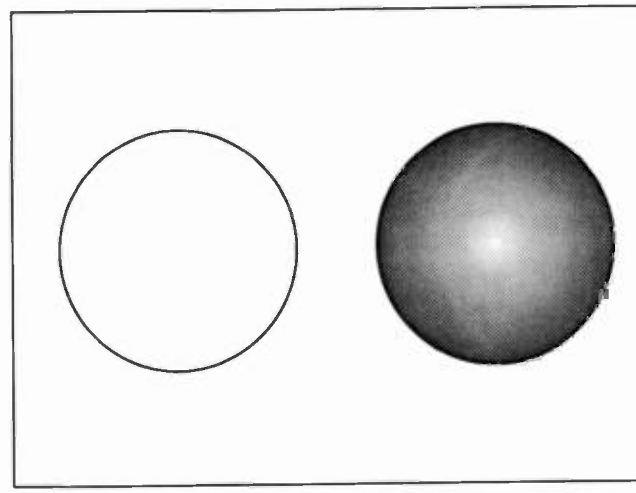

Figure 6. Shading (circle becomes Sphere)

Humans have binocular vision -- two eyes which look in the same direction and whose visual fields overlap.[8] The eyes view the scene from two slightly different angles. Each eye focuses its two-dimensional image onto the retina, a concave surface at the back of the eye (Fig. 7). Each two-dimensional image is transmitted along the optic nerve to the brain's visual cortex (Fig. 8). In the visual cortex the two images are combined, through stereopsis, to form a three-dimensional image. There is some question as to whether stereopsis is created by simultaneous processing or, as evidence suggests, a replacement process in which short-term memory is used to compare views differing in parallax.[9] Even though human eyes are horizontally opposed, the brain is capable of processing parallax information from <u>any</u> direction.[10]

Stereoscopic vision differs among individuals.[11] It is a process we learn and practice from birth. Some people with binocular vision, possibly as high as 10

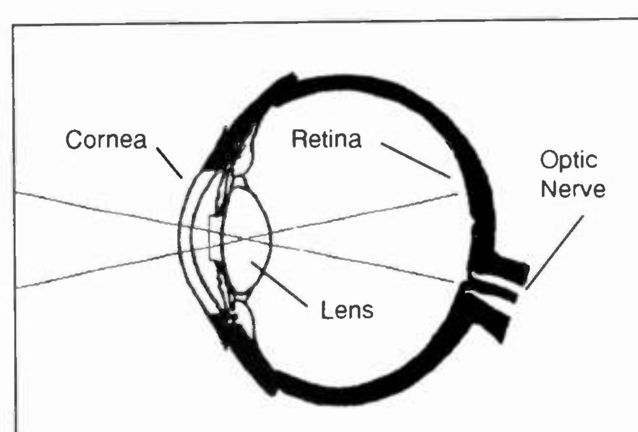

Figure 7. The eye

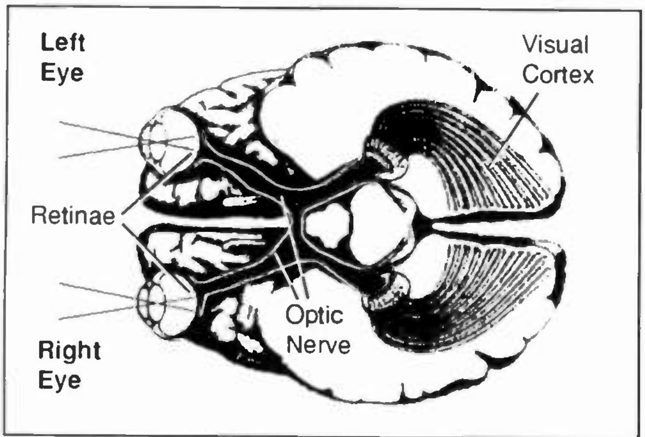

Figure 8. Brain's visual pathway (bottom view)

percent, do not perceive three-dimensional images, and another 10 percent may only have limited depth perception (Fig. 9).

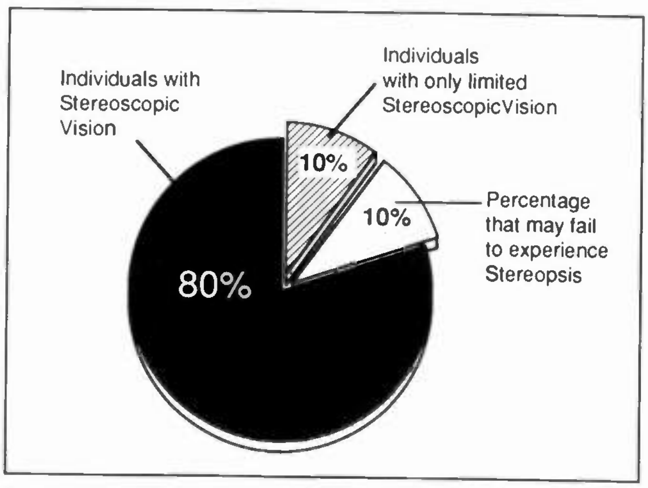

Figure 9. Percentage of Population Without Stereoscopic Vision

VISION III METHODS

VISION III was described technically by the author in a paper presented at the 131st SMPTE Technical Conference in Los Angeles.[12] In short, the VISION III process "tricks" the brain by exploiting the short-term memory believed to be used by the human brain in depth perception. By simultaneously providing both eyes with consecutive images, which differ in parallax and are time displaced to coincide with the persistence of the visio-psychological memory rate, the processes can present depth information to the brain in a form that can be translated into a stable three-dimensional illusion on a two-dimensional surface. This process also takes full advantage of two-dimensional depth cues.

VISION III APPLICATIONS

VISION III images can be produced for nearly any display device. In live-action video production, they can be used in advertising, programming, sports coverage, ENG production and promotions, to name a few. In the case of sports coverage, VISION III images cannot only be used for action coverage, but also for slo-mo, depth-enhanced instant replays. This would be accomplished by a dedicated camera (or cameras) whose time displacement rate is doubled and then played back on a VTR at half speed.

The current VISION III systems are comprised of a two-camera, broadcast-quality video system that employs a special switcher to record images on a VCR or VTR, and a stop-motion film (16mm or 35mm) animation stand. Both systems are prototypes, and were fundamental in developing the VISION III process.

Since 80 percent of all television production is shot on motion picture film, VISION III is currently developing a 35mm motion picture camera and camera attachment. A prototype 35mm film camera is being manufactured for short-term experimentation with the goal of developing a VISION III single-camera attachment (SCA) system capable of using any standard 35mm motion picture or broadcast-video camera with any lens (zoom or prime), thus enabling such cameras to shoot VISION III texture and depth-enhanced images.

With software modifications, the VISION III methods can also be applied to computer graphics and animation. This allows the production of texture and depth-enhanced computer imaging using existing workstations with software that is augmented with a VISION III software module. Production of depth-enhanced, computer-generated station IDs, program titles and weather graphics are also possible. It is also possible to use VISION III technology in cel animation.[13]

Other possible applications for this technology are defense, medical, industrial and scientific imaging and research. Using existing equipment, computer/ video games can be developed with VISION III processes. Even neon signs can be developed which display a VISION III depth illusion.

AUTOSTEREOSCOPIC ADVANTAGES

The obvious advantage of autostereoscopic technology is that it is compatible with motion picture projection and broadcast television. However, there are other advantages. John Merritt in his 1988 SPSE paper, Often-Overlooked Advantages of 3-D Displays, states that stereoscopic images have a greater effective image quality.[14] Stereoscopic displays can improve the quality of images suffering from low resolution, limited gray scale, defocusing and motion blur. Merritt further states that stereoscopic displays allow the viewer to determine what objects are and where they are in an unfamiliar or complex scene. Focus-group studies suggest that, like stereoscopic systems, image resolution and object recognition are greatly improved with autostereoscopic imaging techniques when viewed on an ordinary TV set. Unlike stereoscopic systems, they do not cause eye strain or headaches when viewed for extended periods of time.

CONCLUSION

3-D movies have not been a success in the market place. One reason cited for this failure is addressed by James Monaco in his 1977 book, How To Read A Film. He states that stereoscopic motion pictures were a short-lived gimmick because the systems were cumbersome and inflexible, and they distorted the viewer's perception of depth by producing pseudoscopic-stereoscopic images.[15] This is not, however, a problem with VISION III. VISION III produces an image with a natural and realistic "feel."

Television, as we know it, has evolved dramatically over the last sixty-seven years. Most of this development has been in the area of improving the equipment, i.e. cameras, recording systems and transmitters. However, with the exception of color, very little research has been done to make television images more naturally acceptable to the viewer.

In a recent article for the SMPTE Journal, W. F. Schreiber et al. suggests that there is no evidence that the television viewing audience will pay more money to receive the same programs they now get with NTSC, simply because they would have a higher image quality.[16] Schreiber believes that audiences are attracted to TV by programming, and not by image quality; he sights the slow growth of color television after its introduction.

The television industry today seems to be hung up on what to do with the plethora of proposed advanced television systems (ATV). Do we really need to go through all the trouble and expense of bringing ATV to market just to get a sharp, flat picture?

SUMMARY

Autostereoscopic television and motion pictures will soon be a reality with clearer and sharper texture and depth-enhanced images. Broadcasters, theater owners and home viewers will not have to invest in new equipment to receive the benefits of autostereoscopic technology. Hardware and production techniques are now being developed and will soon be available for program production. Autostereoscopic processes like VISION III can be implemented with only a slight change in the production process, i.e. use of a VISION III camera or camera attachment and autostereoscopic production techniques. All the rest remains the same -- from the "can" all the way to the viewer's home TV set.

ACKNOWLEDGMENTS

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CABLE'S APPLICATION OF FIBER OPTICS FOR IMPROVED VIDEO QUALITY AND BANDWIDTH

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The fiber revolution has come to cable. But it has done so in an interesting and efficient manner. To understand cable's strategy with fiber, it's important to review cable technology and some of its short comings.

The principal shortcoming of cable is its high loss. Half inch aluminum cable has a loss of 1 dB per 100 feet at 181 MHz. This loss is proportional to the square root of frequency. Thus if there is 10 dB of loss at Channel 2 (54 MHz) in a given length of cable, there will be 20 dB of loss at Channel 13 (214 MHz).

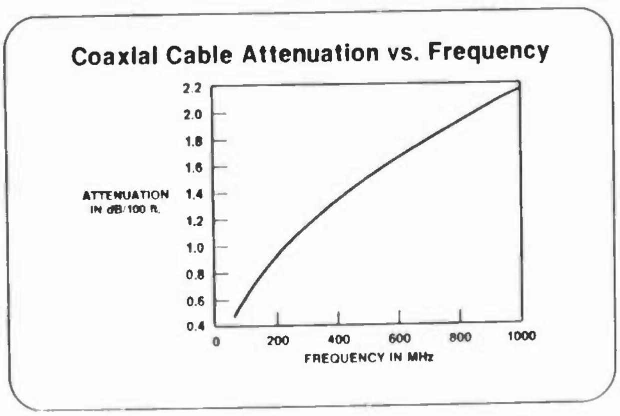

Fig. 1

This loss must be made up with amplifiers. The amplifiers are the real source of nearly all of cable's headaches.

Cable Topology

If we start at the home television set and work backwards we find a VCR in most cable homes. About 25% to 30% of cable homes have an addressable descrambler. Many more have a cable converter which tunes channels the television set isn't capable of providing. Flexible in-home wiring leads to a grounding block just outside the house. A flexible drop cable leads to a tap which splits off energy from the solid distribution cable. This cable has a copper clad aluminum center conductor, foam insulation, and an extruded aluminum tube outer conductor. The cable is usually half inch in diameter. Signal levels in the distribution cable are relatively high, 48 dBmV. This signal measure is the dB ratio compared to 1 mV in 75 ohm cable. Line extender amplifiers make up for cable losses and splitting losses. Because of the high signal levels, no more than two line extenders are cascaded. Line extender amplifiers are spaced 300 feet to 900 feet depending on housing densities.

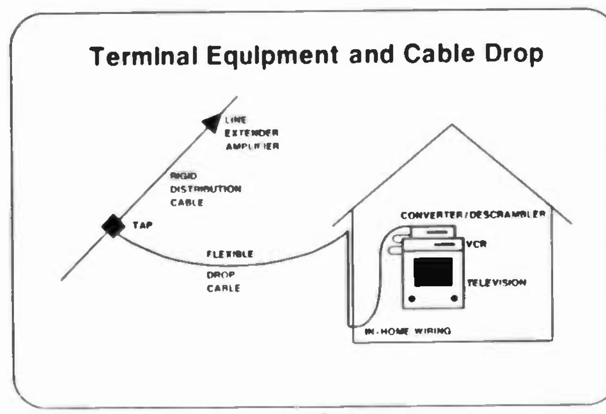

Fig 2

Moving farther up the cable, we find a large number of taps which split off energy to serve subscribers. These taps are a source of minor impedance discontinuities which cause signal reflections referred to as micro-reflections. The Bridger amplifier is the point of demarcation between the distribution system and the trunk system. Trunk is intended to carry signals long distances with minimum distortion. This cable is typically 3/4" to 1" in diameter. Trunk cable is not tapped. Signal levels in the trunk plant are 30 dBmV to 32 dBmV in order to stay well within the linear range of the trunk amplifiers. Trunk amplifiers are spaced roughly every 2,000 feet.

Moving still farther back into the cable system, we arrive at the cable headend. Here satellite reception, distant signal pick up and local signal pick up takes place. Local origination also is provided at this site.

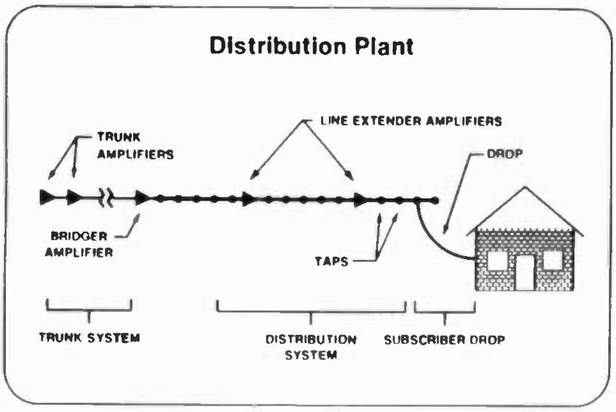

Fig 3

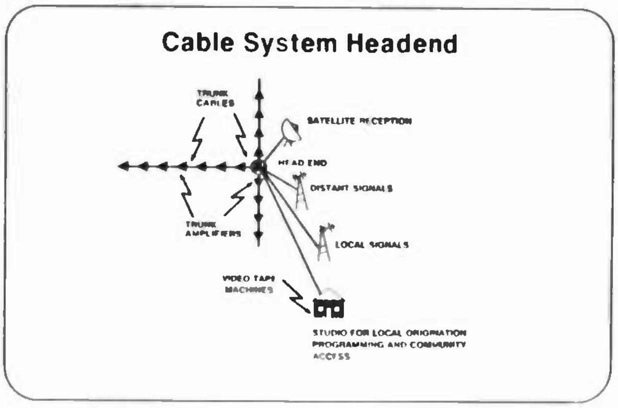

Fig 4

Putting the whole picture together, we arrive at the classical "tree and branch" topology of the modern U.S. cable system. Another of cable technology's compromises is apparent here. Trunk amplifiers are strung in long cascades. Older cable systems with fewer channels cascade 40, 50, even 60 amplifiers. More modern, high bandwidth systems limit the cascades to twenty or so.

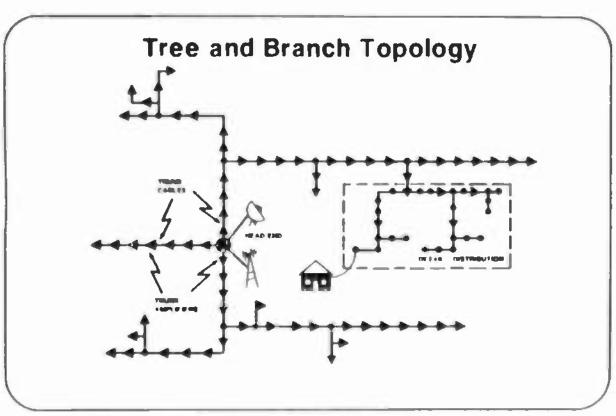

Fig 5

Signal quality parameter target values are listed in figure 6 while the relative distribution of cable plant costs are shown in figure 7.

Signal Quality Parameter Target Values

Parameter	Symbol	Value
Carrier/Noise	C/N	46 dB
Composite Second Order	CSO	-53 dB
Composite Triple Beat	CTB	-53 dB
Signal Level at TV		0 dBmV

Fig 6

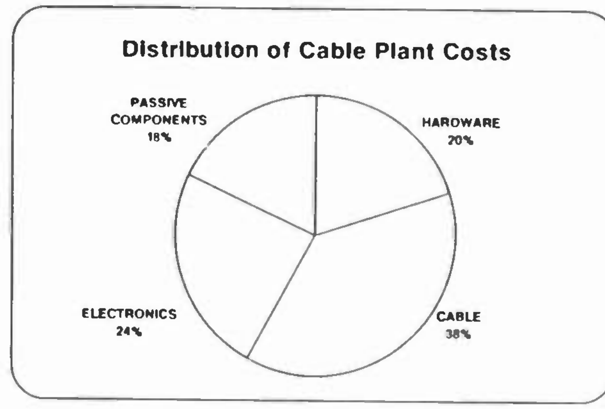

Fig 7

About fifty percent of cable footage in a typical cable system is in the flexible drops which lead to the home. Forty percent of the footage is in the tapped distribution cable which runs through the neighborhood. Only ten percent of the footage is in the trunk cable. This is an important fact for our fiber strategy.

Capacity Limitations

Coaxial cable has surprisingly high bandwidth. We find that our drop cable is capable of passing more than 1 GHz. The same is true of the vast majority of our aluminum cable. The capacity limitation is not the cable, it is the amplifiers. More to the point, it is the cascade of amplifiers that limits bandwidth in cable systems.

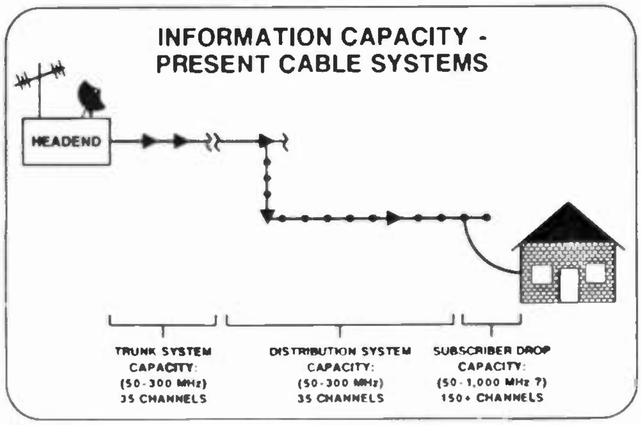

Fig 8

The fiber backbone principle invented at ATC recognizes that the principal limitation of cable technology is not the cable, but the amplifier cascades. Fiber backbone overlashes fiber on the trunk reaching into the neighborhood. The cable system is broken into a multitude of little cable systems each with very short cascades. Some amplifiers are turned around to feed in the opposite direction.

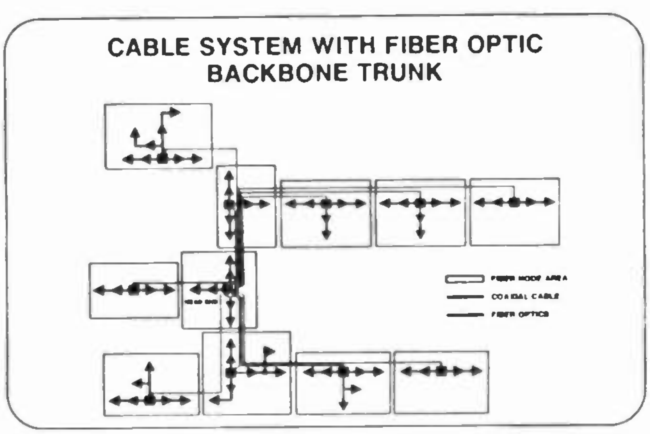

Fig 9

An important feature of this configuration is the reduced failure points between any subscriber and the headend. Much fewer subscribers are impacted by an amplifier failure. A secondary benefit is that each small cable system may be fed with different signals. If the demographics of the neighborhoods are different, appropriate programing can be supplied. This topology also facilitates two way services because ingress does not accumulate as in Figure 5.

Now if we look at the information capacity of a hybrid fiber / coax system, we find that the capacities of the elements are more nearly matched. At present it is practical to put 80 or so channels through the cable system. This does not quite fully exploit the 150 channel capacity of the raw cable. But improved electronics should afford cost effective upgrades in the not too distant future.

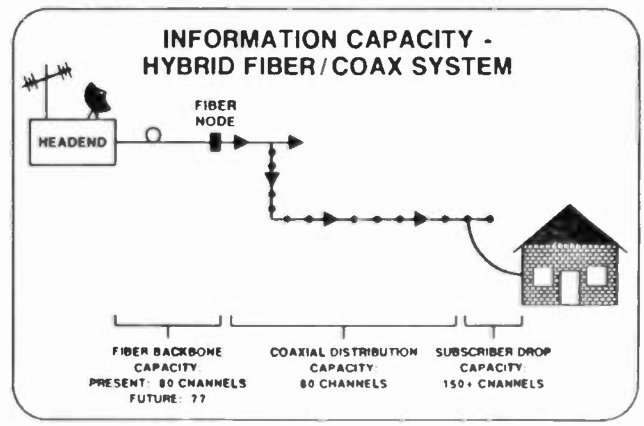

Fig 10

<u>Fiber to the Home???</u>

If we analyze what we have done in Fig.9, we can see a continuum between present cable practice and fiber all the way to the home. If the little boxes are made huge so only one encompasses the entire cable system, we're where we're at now. If we go to the opposite limit and make the boxes so small there are no amplifiers in them, we bring fiber all the way to the home.

If we plot the relative improvement in signal quality and reliability increase against the number of amplifiers in cascade and simultaneously plot cost, we obtain Figure 11.

At the right hand side we obtain maximum improvement with zero length cascades, i.e. fiber to the home. But the cost is huge. At the far left, we have no improvement at no cost. Good engineering practice looks for compromise. It appears that we can obtain 75% to 80% of the benefit at about 4 or 5 amplifiers in cascade for a cost of less than $50 per subscriber. If this is true why would we go all the way to the home at costs of more than $1,600 just to get another 20% to 25% improvement? It doesn't make good engineering sense! Two important points to remember are: 1) Cable, like fiber, is a broadband medium. 2) If we fiber the trunk, we're dealing is less than 10% of the total cable footage. Most of the plant remains the same, reused, untouched. But the major limiting factor, the long amplifier cascades, are eliminated.

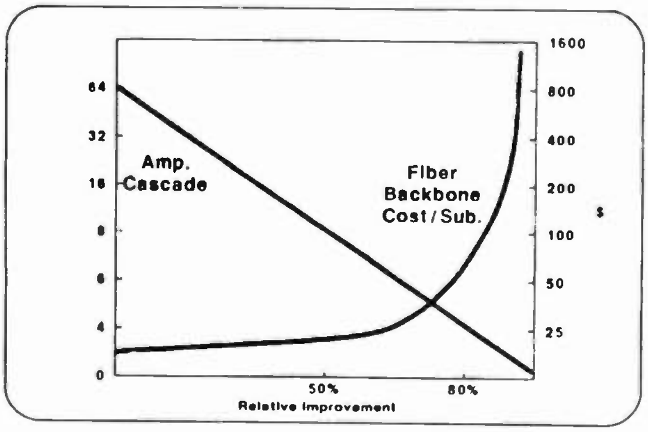

Fig 11

<u>Fiber Signal Transmission</u>

Signal transmission on the fiber backbone is not very different from modulation of radio carriers. A laser is amplitude modulated by the composite broadband television signal. The whole cable spectrum is impressed on to the light signal in the fiber. At the receiving end, an optical detector directly converts the light signal to a broadband RF signal ready to feed the remainder of the cable plant without any additional processing.

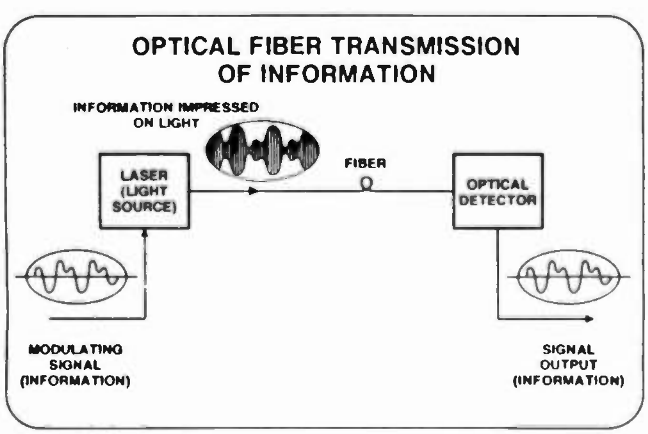

Fig 12

This can be portrayed in another way:

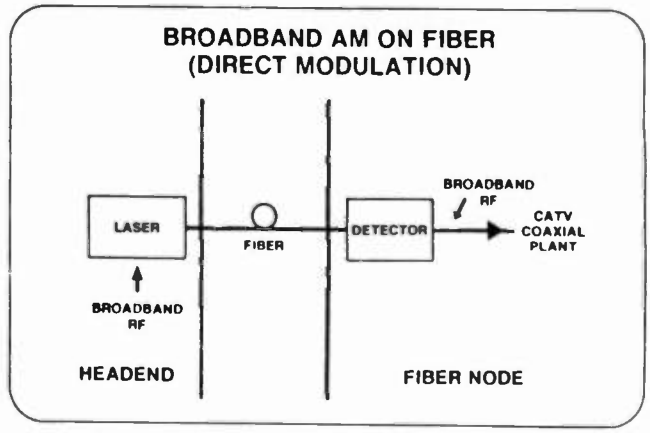

Fig 13

Some research is under way using a powerful continuous wave laser feeding an external modulator. Longer distances may be achievable this way.

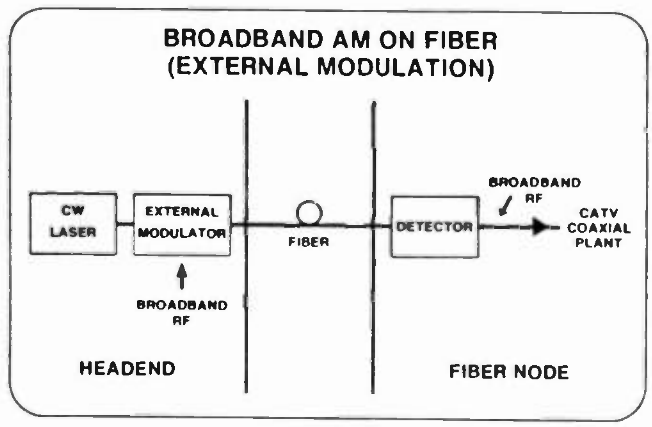

Fig 14

Yet another variation is shown in Figure 15. Several lasers are fed each with only a portion of the spectrum..The spectrum components are carried on separate fibers and separately converted by detectors at the receive site. The RF energy is then combined to provide the broadband feed for the cable. This approach relaxes the linearity requirements on the lasers.

Fig. 16 demonstrates the disadvantages of FM modulation. At the receive site, each FM signal must be separately demodulated. Then it must be separately remodulated in the Vestigial Sideband Amplitude Modulation format demanded by the subscribers' television or VCR. This system usually requires a small building while the approach of Figure 13 neatly fits into an ordinary amplifier housing.

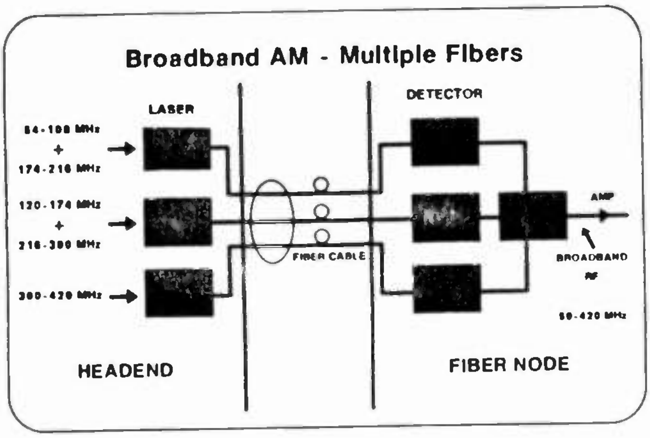

Fig 15

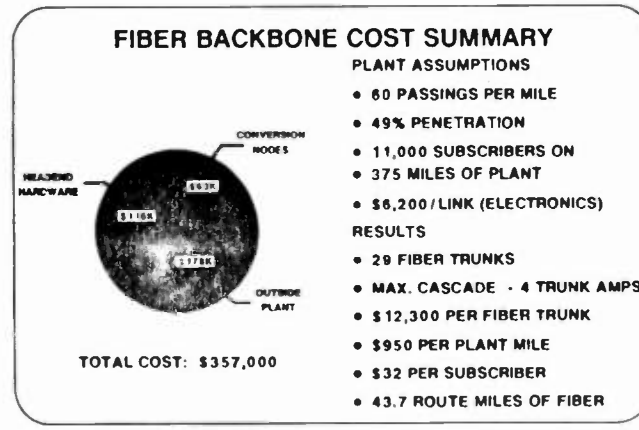

Fig 17

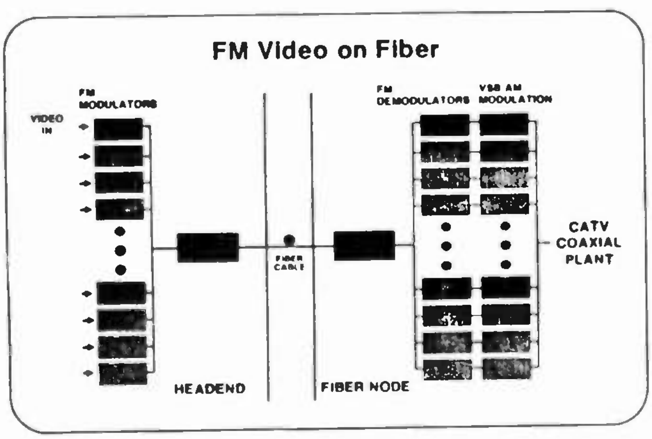

Fig 16

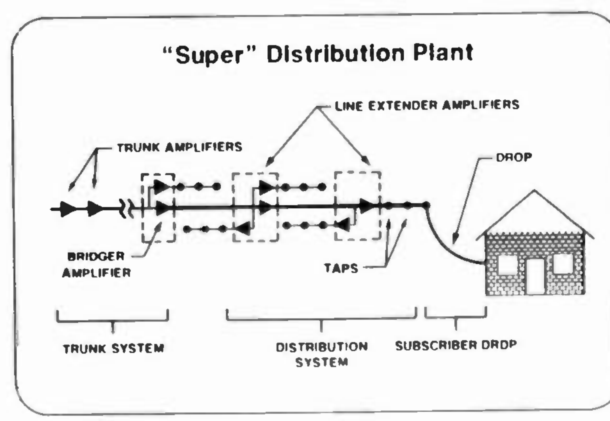

Fig. 18

The FM modulators of Figure 16 can be replaced with analog to digital converters, and a multiplexer at the origination point and a demultiplexer followed by a bank of digital to analog converters at the receive site. The hard part is still the Vestigial Sideband Amplitude Modulation process required on each separate channel.

Figure 17 displays the cost of applying this principle to an actual portion of ATC cable plant. This was not a particularly advantageous example, simply one that was conveniently in the computer. The dominant number is the $32 per subscriber required to install a fiber back bone in this portion of the system.

Super Distribution

Back when we were discussing Figure 3, we noted that the taps caused nasty micro-reflections. Figure 18 shows a technique championed by Rogers Cable of Canada. They call it "Super Distribution" because it minimizes micro-reflections and allows lower operating levels.

The lower levels are possible because fewer taps are fed from any one of the small amplifiers. These lower levels mean that the amplifiers are operated more linearly. Most of the micro-reflections are eliminated and most of the non-linear distortions are avoided by this technique. Rogers estimates the cost of implementing this technique at well under $50 per subscriber.

Conclusion

The fiber backbone affords us lower noise and much higher reliability. Signals can be customized for each neighborhood. Two way cable is greatly simplified. All for under 50$ per subscriber. For another $50 or so per subscriber, the amplifiers can be upgraded to higher bandwidths. The short cascades makes this possible. Super distribution techniques reduce micro-reflections and minimize non-linear distortions. Again for $50 or so. These three techniques make possible high capacity, low noise, low distortion cable upgrades at very rational costs. These upgrades can be done in a evolutionary manner.

Finally, the practicality of this approach calls into question the proposition of fiber to the home. This extravagant approach yields almost no further benefits at tremendous increased in cost.

The Author

Dr. Ciciora is Vice President of Technology at American Television &Communications, ATC, in Stamford Connecticut. Walt joined ATC in December of 1982 as Vice President of Research and Development. Prior to that he was with Zenith Electronics Corporation since 1965, He was Director of Sales and Marketing, Cable Products, from 1981 to 1982.

Earlier at Zenith he was Manager, Electronic System Research and Development specializing in Teletext, Videotext and Video Signal Processing with emphasis on digital television technology and ghost canceling for television systems.

He has nine patents issued. He has presented over seventy papers and published about thirty, two of which have received awards from the IEEE. Walt writes a monthly column titled "Ciciora's Page" for Communications Engineering and Design magazine.

He is currently chairman of the National Cable Television Association, NCTA Engineering Committee, Chairman of the Technical Advisory Committee of Cable Labs, and President of the IEEE Consumer Electronics Society. He is a past chairman of the IEEE International Conference on Consumer Electronics. Walt is a Fellow of the IEEE, a Fellow of the Society of Motion Picture and Television Engineers, and a senior member of the Society of Cable Television Engineers. Other memberships include Tau Beta Pi, Eta Kappa Nu, and Beta Gamma Sigma. He served on several industry standard-setting committees. Current interests center on competitive technology, the consumer electronic interface with cable, and HDTV.

Walt received the 1987 NCTA Vanguard Award for Science and Technology .

Walt has a Ph.D. in Electrical Engineering from Illinois Institute of Technology dated 1969. The BSEE and MSEE are also from IIT. He received an MBA fro the University of Chicago in 1979. He has taught Electrical Engineering in the evening division of IIT for seven years.

Hobbies include reading, wood working, photography, skiing, and a hope to someday become more active in amateur radio (WB9FPW).

HOW TO IMPLEMENT A COMPUTERIZED SYSTEM FOR SCHEDULING TECHNICIANS AND ENGINEERS

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ABSTRACT

Scheduling the work duties of broadcast technicians is a demanding and complex task requiring that the scheduler have both the degree of exactitude normally expected of a bookkeeper and the creative problem solving ability of an experienced manager. Once a schedule is made, scheduling changes are made repeatedly and intermittently, usually under conditions of stress. In designing and implementing an automated system to address the solution problems, what should be the design goals of such a system? What functions are necessary for such a system to succeed, and what are merely "nice to have"?

THE SCHEDULING PROBLEM

In theory scheduling is not a very difficult operation (see diagram entitled THE SCHEDULING PROBLEM). To begin, one must gather together a list of their requirements, i.e. the work that must be accomplished. Once the requirements are defined, a process begins of matching resources, i.e. technicians, to satisfy these requirements. A requirement is only satisfied if the applied resource is not in conflict with any implicit or explicit constraints in the requirements definition.

But as in many activities which are theoretically easy, the devil is in the details. Typically, we find that highly skilled people are spending on the order of ten to twenty hours a week scheduling and rescheduling their technicians in moderately sized stations. If the task is given to less skilled people, considerably more time is spent doing the schduling and conferring with their superiors. Larger stations have a full time scheduler or schedulers, as the problem grows exponentially with additional personnel to schedule. Why is this seemingly trivial task consuming so many resources?

Defining Constraints

In automating the scheduling of technicians, the most difficult area lies in the domain of constraint definition. There are four kind of constraints to cope with in ascending order of difficulty. These are:

1. Common Sense Constraints

   These are constraints that the ordinary person should grasp with little or no training. You don't have to be a rocket scientist to realize that if you schedule the same technician to be in two places at once, you have a problem.

2. Skills Constraints

   The technician(s) scheduled to do the job must have the skill(s) to accomplish it and at the required proficiency level.

3. Work Rule Constraints

   Within each station, there can exist an elaborate set of work rules and penalties for violations of those rules. A scheduler may require considerable experience to minimize these penalties, having to refer not only to the current schedule, but to historical schedules.

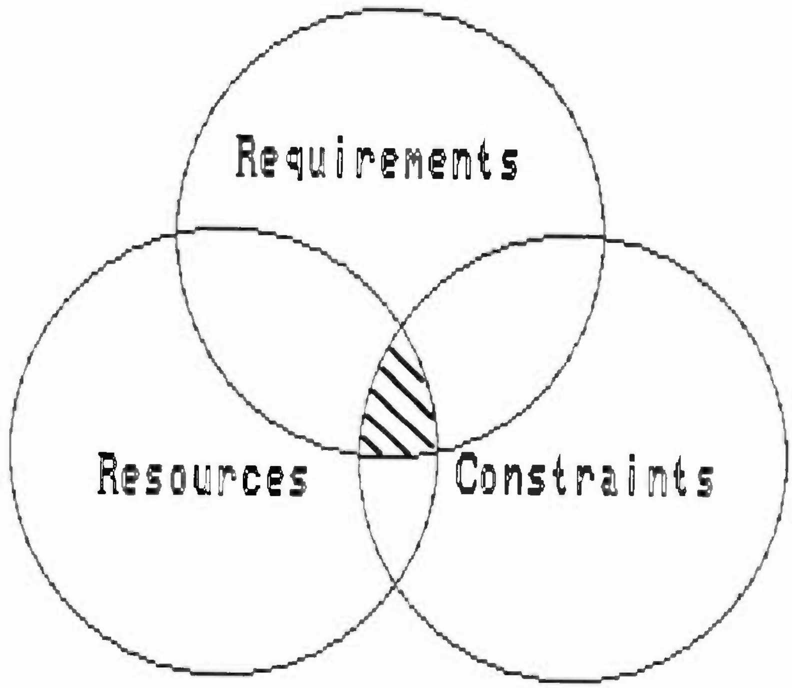

Figure 1: The Scheduling Problem - Schedules are valid only at the intersection of Requirements, Resources, and Constraints

4. Personnel Constraints

   There are times when two people do not get along, and you should not schedule them for the same task. Likewise, some people have bad backs and should not be placed on a task requiring heavy lifting. For a scheduler to master these details may require years of experience.

Decision Support not Decision Making

As a practical matter, it is exceedingly difficult to exhaustively define and quantify all the constraints that an experienced scheduler will apply when making a scheduling decision and prohibitively expensive in time an money to implement such a solution in an automated system. Experience has shown that trying to do so is like hitting a fast moving target.

As a consequence, the first and most important principle in designing an automated scheduling system is that the system should **not** attempt to provide a 100% solution to the scheduling problem. The underlying purpose of such a system must be to provide the scheduler with decision support, checking that the decisions made are correct, and aiding in making those decisions quickly and efficiently, but not actually making them.

The goal is to make the scheduling task easier for less skilled personnel with only occasional supervision and/or intervention from more skilled personnel.

Requirements Driven

The second design principle that emerges from the foregoing is that an automated system should be requirements driven. By this, I mean that the system determines that all requirements are satisfied for a schedule to be valid, rather than determining that all resources are utilized. If, over a reasonable period of time, all requirements are satisfied (i.e. all work is scheduled to be completed) and resources are still available, it is an indication that you are overstaffed! a resource driven process, i.e. one that assures you that each technician is doing 40 hours work will help to conceal an excess resource problem.

THE UNDERLYING DATABASE

The accompanying database diagram is a simplified schematic representation of the relational database model that PROMPT Corporation used to solve the automated scheduling problem.

Interpreting the Diagram

Enclosed in the rectangular boxes are each of the major files in the database. Each file contains zero, one, or many records. Each record contains information that describes and differentiates attributes of interest belonging to the entities described by the file. Connecting the boxes are arrows indicating the underlying relationships in the database. A file at the tail of an arrow has a "one-to-many" relation with a file at the head of the same arrow. For example, The TECHS file has a one-to-many relation with the ABSENCES file (read as "for each record specifying a technician in the TECHS file, there may be zero, one, or many absences recorded for him in the ABSENCE file.

Many-to-many relations are represented by combining two one-to-many relations with the "many" file being the same. For example, look at the files TECHS, SKILLS and SKILLEE. The SKILLEE file acts a a cross reference between technicians and their skills. With this structure, we can interrogate the database with the question "What skills does technician Jones have?", or conversely "Which technicians have the skill Master Control Board Operator?".

In keeping with our underlying model of Requirements, Resources and Constraints, we will show how these have been implemented in this structure.

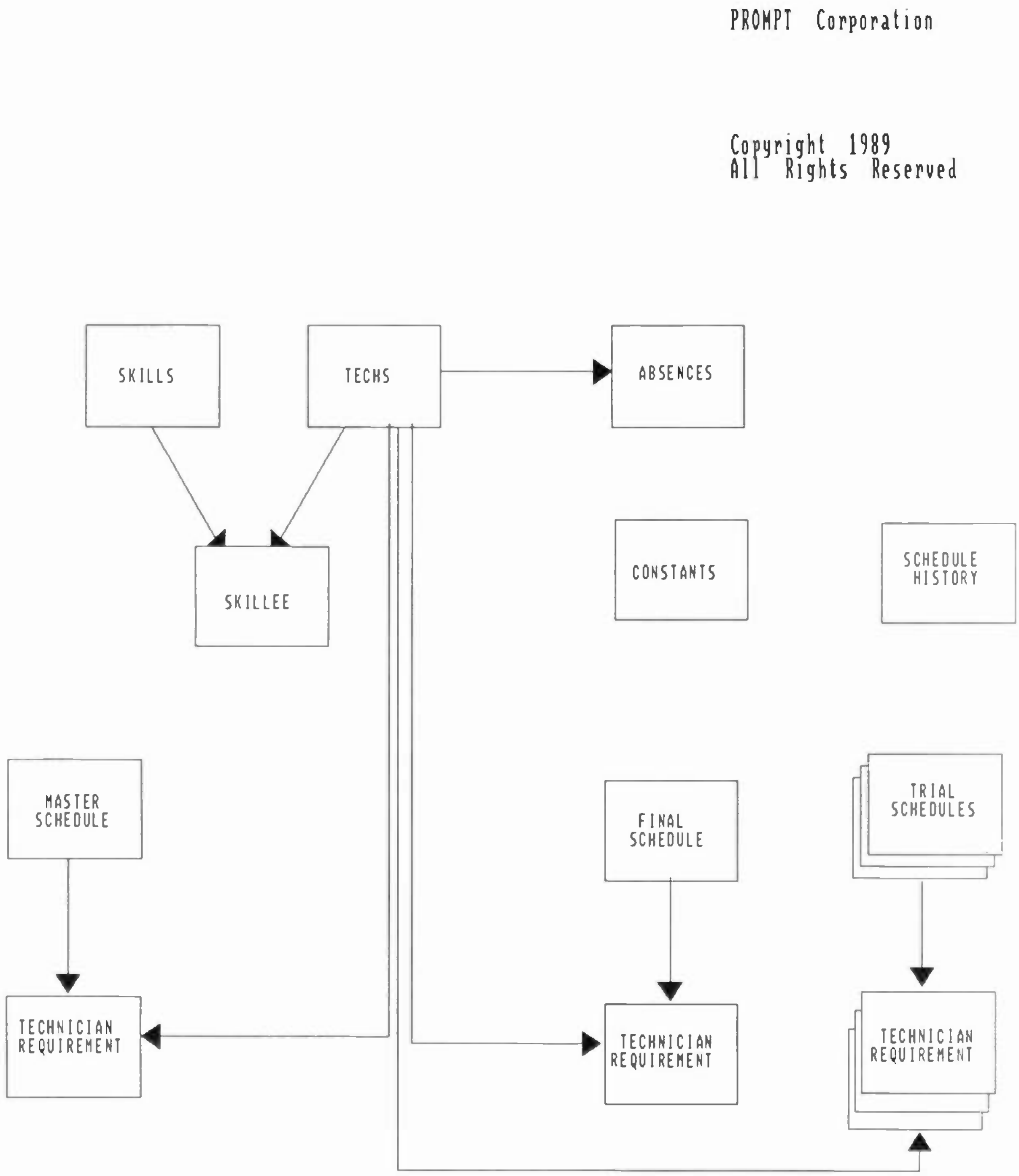

Figure 2: The PROMPT Corporation database

Requirements

Requirements are identified in the files identified as MASTER SCHEDULE, FINAL SCHEDULE, and TRIAL SCHEDULE(S) and their respective TECHNICIAN REQUIREMENTS. The structure of these files is identical. The contents varies depending upon the use.

1. SCHEDULE FILES (MASTER, FINAL, AND TRIAL) contain a record for each task to be completed during the work week. Each record details where and when the task is to be executed.

2. TECHNICIAN REQUIREMENT FILES detail the number, skills, and proficiency level of technicians required to complete the associated task. Once assignments are made, these files will contain the assigned technician and/or his replacement (in case of absence or reassignment).

PROMPT Corporation has chosen to track three levels of scheduling to ease the scheduler's task. A Master schedule is created only once per shift pick, probably by the Chief Engineer, certainly by the most knowledgeable scheduler. The master schedule serves as a template for the TRIAL and FINAL schedule, while reducing data entry chores to a minimum.

Each week, the scheduler creates one (or more) trial schedules from the master schedule or from a trial schedule already created for that week. The trial schedule is means of recording events specific to that week with out disturbing the master schedule. On any week, we know that some employees will be absent due to sickness, vacation, etc. but we do not know whom. The trial schedule allows us to copy standard technician assignments from the master schedule and then modify only those that need to be modified.

PROMPT allows for multiple trial schedules so that contingent schedules can be drawn up in advance for difficult situations. for example separate trial schedules can be prepared in the event both the Celtics and the Bruins win their respective playoffs and in the event only one team wins and finally in the event neither wins.

The final schedule is simply the trial schedule designated by the scheduler as the one to be executed.

Resources

Resources are identified in three files in the database. These are:

1. The TECHS file which has a record for each technician of interest identifying him/her by name, employee number, date of hire, rate of pay, as well as various status flags which indicate whether the technician is full-time, part-time, a crew-chief, free-lance, and active/inactive (retired or terminated technicians are never removed from the database as they would then not be available for historical analysis).

   The TECHS file also records each technicians shift pick.

2. The ABSENCES file which documents each absence the technician takes as to dates and reason. This file accumulates absence history.

3. The SKILLS file which lists each of the skills a technician might (or might not) have.

Constraints

Constraints are implemented in three ways.

1. Validity checks are applied to all data entered into the system to

test for correctness and consistency.

2. Each time a technician is assigned to a task, the system will verify that that assignment will not put him in conflict with any other outstanding assignment.

3. The CONSTANT file contains the work rules of station. In it are detailed standard items, e.g. the number of hours in a work day, the number of days in a work week, the number of hours in a standard rest period, etc. It also includes the penalties to be applied when violations of these rules occur.

RESCHEDULING - THE CRUCIAL ACTIVITY

If scheduling simply involved the creation and distribution of schedules on a weekly basis, life would be much simpler than it actually is. The problem is that the world does not stick to our schedule.

PROMPT Corporation recognizes that the problem of scheduling is overshadowed by the problem of rescheduling - making changes to allow for last minute events that could not be predicted at the time the schedule was drawn up.

The "Cascade"

The most frequently encountered major scheduling problem is the cascade. This often occurs when a technician who has a relatively rare skill is absent due to illness. In seeking a replacement for the absentee, often the only choice is a technician who is currently assigned to another task. If we assign this technician to replace the absentee, we create a second order absence (or "cascade") that must be filled. Subsequent choices may create third, and forth level absences that also must be filled. PROMPT warns you when a cascade is about to take place, provides a special tool to help you navigate through a cascade with minimum disturbance to your current schedule, and if a cascade is created, provides a simple method of shuffling technicians to fill the holes in the schedule.

FINISHING TOUCHES

There are a variety of features that a scheduling system might have. Among these are:

1. Recognition that tasks that begin on one day, cross over midnight, and end on the next calendar day are part of the same working day.

2. Facility to vary the day on which the schedule begins and ends.

3. The ability to "cost" a schedule, extending the hours worked for each technician by his wage rate, and applying appropriate penalties when necessary.

4. The ability to keep a history of each final schedule actually executed. This is useful for reconciliation of payroll, and for determining the actual impact of various wage concessions when bargaining.

5. The ability to graphically represent a technician's work schedule by use of a time line or Gantt Chart. This can be a great help when trying to see if one technician can substitute for another.

6. The ability to indicate that a task does not have a specified ending time (e.g. a ball game).

7. Consistent and easy to use screen interface.

8. Provision for scheduling replacements during lunch and rest breaks.

9. Online "help" may or may not be required according to your needs.

Each of these features while providing additional information or functionality has a corresponding cost in increased system complexity, difficulty of use, and cost.

DOCUMENTATION AND TRAINING

Merely having a working computer program is not the end of the implementation task. In order to consider the implementation complete, an exhaustive documentation of the system and its use has to be undertaken. Failure to do this can result in a total loss of investment if personnel who know how to use the system leave.

Likewise, provision must be made to train new personnel who will have to use the system.

MAKE OR BUY

Perhaps the key decision in implementing an automated scheduling system is that of whether to undertake building a custom made system or purchasing one off the shelf.

Scheduling systems appear seductively simple to build, but can quickly become a sink for resources. For example, The PROMPT scheduling system represents three man year's effort by a programming team with decades of programming experience and specific experience with implementation of scheduling systems. If you are going to attempt to implement a customized system you must expect to spend a similar order of magnitude of resources if you wish to produce a similar system.

For this reason if you can find an off the shelf package that meets your needs at a price you can afford then by all means purchase the system.

FILE TRANSFER PROTOCOLS

Chuck Forsberg
Omen Technology, Inc.
Portland, Oregon

NO MICRO IS AN ISLAND

The spread of personal microcomputers in broadcasting has heightened interest in reliable, fast, and convenient ways to transfer files over dialup phone lines. This paper discusses microcomputer file transfer protocols, how they evolved, and how they can affect your operations. You will learn how to transfer data with a minimum of cost and a maximum of confidence.

Since last year, several new protocols have appeared, if only to grace the footnotes of reference documents. ZMODEM's performance has been improved with recent extensions.

XMODEM

Since its development in the late 1970s, the Ward Christensen **MODEM** protocol has allowed a wide variety of computers to interchange data. There is hardly a communications program that doesn't at least claim to support this protocol, now called **XMODEM**. More refined protocols were available, but microcomputer hackers with limited resources flocked to XMODEM because it was simple and easy to implement. XMODEM worked well transferring files at 300 "baud" over dial-up phone lines.

Changes in computing, modems and networking have spread the XMODEM protocol far beyond the environment for which it was designed.

XMODEM sends files in 128 byte blocks, waiting for an acknowledgement after each and every block. XMODEM's 90% efficiency on 300 bps (bits per second) calls falls to 20% or less on timesharing systems, high speed modems and packet switched networks. XMODEM's stop-and-go operation

XMODEM-1k

XMODEM was extended to 1024 byte data blocks using a technique introduced with the *YAM* (Yet Another Modem) communications program.

XMODEM-1k (sometimes erroneously called YMODEM) sends eight times as much data in a block, reducing XMODEM's protocol overhead by 87 per cent. **JMODEM** and Super8K extend this to 8192 byte blocks, further reducing overhead under ideal conditions.

Unfortunately, these longer data blocks exact a heavy penalty when transmission errors from line noise or other problems require a retransmission. Some programs "fall back" to shorter block lengths when excessive block retries are required, but XMODEM technology does not allow the most efficient "fall back" to shorter blocks without unduly compromising the safety of the file transfer.

XMODEM-CRC

The original XMODEM protocol used an 8 bit checksum to detect transmission errors. This proved inadequate, and today all but the most primitive communications programs support **XMODEM-CRC.** XMODEM-CRC replaces the 8 bit checksum with a 16 bit number which in not "fooled" into accepting corrupted data as often.

XMODEM Reliability

The CRC polynomial chosen for XMODEM is not the best (most reliable at error detection) possible 16 bit CRC. Unfortunately, this is not the weakest link in XMODEM's chain of operation. XMODEM uses single ASCII control characters to supervise the file transfer. These control characters were designed to be unique in the presence of the single bit errors typical of 300 bps modems, but are easily spoofed by errors made by today's modems. Proprietary Cybernetic Data Recovery™, logic enhancements in *Professional-YAM*, *ZCOMM*, and *DSZ* correct for some but not all of these weaknesses.

Kermit

Another problem with XMODEM arises from the way XMODEM transfers data using all 256 combinations of 8 bits. Some networks and mainframe computers cannot transmit or receive all 8 bit codes. Many have trouble with arbitrary control characters. Some require each input record to end in a carriage return or other special character. The **Kermit** protocol was developed at Columbia University to allow file transfers with balky mainframes that do not support XMODEM. Kermit uses a **quoting** technique to represent control characters and characters with the 8th bit set as multiple printing characters. For example, a hex 9D (¥ on IBM PCs) might be sent as #'[expressing one byte as three printable characters.

Unfortunately, Kermit has evolved into incompatible **dialects** that don't work with each other. One dialect uses 8th bit quoting to transfer data over 7 bit systems. Another Kermit dialect uses 8 bit transmission (as does XMODEM) instead of quoting the 8th bit. Users of other programs sometimes fail to select the proper Kermit dialect. When Kermit cannot transfer files the most common cause is a disagreement between the sender and receiver over the choice of Kermit dialect.

The design of Kermit packets allows Omen Technology's *Professional-YAM* and *ZCOMM* programs to automatically recognize Kermit, select the proper dialect, and download files without user intervention.

Properly implemented with Kermit's optional reliability enhancements, Kermit transfers files reliably. Compression, Sliding Windows, and recently developed Long Packets have reduced Kermit's high overhead somewhat. Kermit is still twenty five per cent or more slower than ZMODEM when transferring archives[1] made by the popular ARC and ZIP programs.

Kermit has two advantages over XMODEM besides higher reliability. Kermit transfers both the file name and the file contents, so the user does not have to type the file name twice, once to the sending program and once to the receiving program. Protocols that transfer file names are called **BATCH PROTOCOLS.**

1. An archive is a collection of related files grouped together in a single file for convenience. The popular ZIP and ARC archivers reduce storage requirements and modem transmission time by compressing their archives. File compression saves space by eliminating redundant information.

Another advantage of Kermit over XMODEM is the ability to transfer file contents exactly, without adulterating the data with garbage added at the end of the file.

YMODEM

The **YMODEM** protocol was developed in the early 1980s as part of Chuck Forsberg's *YAM* program. YMODEM also provides batch transfers, allowing many files to be sent with one command. Minimal YMODEM sends the file name with each file. Full YMODEM sends the file length and date with the file name, allowing the receiver to reconstruct the file exactly. Full YMODEM also informs the receiver of the total number of files and the total length remaining to be sent, allowing the receiver to display an estimate of the remaining transmission time. Programs fully implementing the YMODEM specification may be certified by YMODEM's author as supporting **True YMODEM™** [2].

YMODEM uses XMODEM-CRC and XMODEM-1k technology. YMODEM shares XMODEM's reliability and performance limitations.

When completely error free links are available, **YMODEM-g** provides excellent throughput by eliminating the per block acknowledgements required to permit error recovery. YMODEM-g reduces protocol overhead to one half per cent, but a single transmission error will ruin an entire transfer.

Users of YMODEM and YMODEM-g have been confused and hampered by some programs (Qmodem, Procomm, Crosstalk) whose authors did not care to follow the YMODEM protocol specification published in 1985.

X/YMODEM Variants

IMODEM, JMODEM, WXMODEM, Telink,

2. True YMODEM™, ZMODEM-90™, and MobyTurbo™ are Omen Technology Trademarks.

SEAlink, and Megalink are derived from XMODEM. Each improved on one or more aspects of XMODEM. Most claimed to be the fastest available. Some of these protocols remain in use to this day.

ZMODEM

In late 1985 Telenet was concerned by unsatisfactory results network users were having with XMODEM and Kermit file transfers. They funded the initial development of the **ZMODEM** file transfer protocol. ZMODEM is a language for implementing file transfer protocols, not just a single protocol good for a small set of applications. Since then, ZMODEM's speed, reliability and ease of use have made it the protocol of choice for more and more applications.

ZMODEM transfers files efficiently with buffered (error correcting) modems, timesharing systems, satellite relays, and packet switched networks such as Telenet (PC-Pursuit) and Tymnet (Starlink).

In normal operation, ZMODEM transmits data non-stop. The receiver does not transmit to the sender unless an error is detected. This "no news is good news" technique optimizes performance with high speed modems (Robotics HST) and saves kilopacket charges for network users.

Optional ZMODEM compression provides even faster transfers on suitable files. This speedup varies from 20 per cent for executable files to an incredible 1000 per cent for the *Personal Computing Magazine* ASCII test benchmark.

In 1990 Omen Technology introduced ZMODEM-90™ with the MobyTurbo™ accelerator to maximize speed transferring compressed files. ZMODEM-90 with MobyTurbo provides the speed advantages of YMODEM-g without compromising ZMODEM's ease of use, flexibility and reliability.

ZMODEM's 32 bit CRC protects against errors that continue to sneak into even the most advanced networks. Even more important than the number of bits in the CRC, ZMODEM safeguards all data and supervisory information with effective error detection. Since XMODEM, YMODEM, JMODEM, et al. do not protect all of their transmissions with CRC, they cannot be trusted under difficult conditions.

User Friendliness is an important ZMODEM feature. ZMODEM's design allows automatic file downloads initiated without user intervention. AutoDownload is a popular feature that greatly simplifies file transfers compared to traditional protocols. AutoDownload and security verified command download allow programs with integrated ZMODEM to operate in a server mode for reliable remote access.

Crash Recovery is one of ZMODEM's most popular features, allowing a transfer interrupted by a disconnect to be continued from were it left off. Crash Recovery is especially welcome when one trips over the phone cord just as a long file transfer is nearly finished.

Some ZMODEM programs provide security verified downloading and flexible control of selective file transfers. For example, one can transfer all files in a directory, skipping over files missing from the destination directory and skipping over files where the destination's copy is up to date. Some ZMODEM programs can maintain an entire directory tree with a single command.

CHOOSING A PROTOCOL

When attempting to transfer a file between two computers, one or the other of the communications programs may limit your choice to a poorly written XMODEM or Kermit implementation.

When a choice is available, a properly written ZMODEM provides unmatched reliability, convenience, and network compatibility while closely approaching the raw speed of YMODEM-g. MobyTurbo™ closes this speed gap, without sacrificing reliability or robustness.

Many Kermit implementations provide reliable transfers for those willing to suffer Kermit's higher overhead.

The performance and reliability of Kermit and XMODEM family protocol transfers varies greatly with the quality of the programs involved. An enthusiastic review in a magazine is no assurance of reliability because reviewers do not test file transfers under adverse conditions.

In continuation of a tradition made necessary by poor phone lines, Omen Technology conducts Protocol Stress tests from time to time. A special setup [3] generates moderately severe transmission errors under controlled conditions. In these tests ZMODEM and Kermit[4] are the most reliable of the protocols tested. XMODEM and YMODEM with *Professional-YAM's* Cybernetic Data Recovery™, logic enhancements almost but not quite always succeed. Many XMODEM and YMODEM programs failed often; a few failed almost every time. Some even locked up one or both computers.

Typical File Transfer Speeds
(2400 baud, no transmission errors)

Prot	ch/sec	Notes
X	53	CIS/Tymnet PTL 5/3/87
X	56	The Source PTL 5/29/87
SK	170	The Source PTL 5/29/87
ZM	221	CIS/Tymnet PTL
ZM	231	BIX/Tymnet 6/88
ZM	227	CIS PTL node
ZM	229	TeleGodzilla (local)
ZM	534	MNP Level 5 (text)
ZM	2561	pcbench.txt

Abbreviations:
K Kermit
SK Sliding Window Kermit
X XMODEM
ZM ZMODEM

3. Described in PTEST.DOC, available on TeleGodzilla BBS, 503-621-3746

4. As implemented by Omen Technology Software

TRACKING SYSTEM FOR INCLINED ORBIT SATELLITES

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Richardson, Texas

ABSTRACT

The imminent plans of satellite communications suppliers to extend the life of their satellites by allowing them to operate in an inclined orbit are discussed. This requires earth stations to have more sophisticated tracking mechanisms. Both domestic and international satellites will be utilizing the inclined orbit maneuver.

The tracking systems in use today are investigated, including both steptrack and program track options. The advantages and disadvantages of each are weighed.

Hardware requirements for program tracking are defined. Software methods of controlling antenna movement are identified. Existing and proposed methods of disseminating satellite positional parameters are explored.

A unique Andrew solution using existing standard motor drive systems that meet the physical tracking requirements is offered. Use of the APC100 Antenna Controller in conjunction with the ASC2000 System Controller or an IBM-compatible PC (Personal Computer) is supported by software that Andrew supplies. In addition, Andrew is willing to offer solutions to users of non-Andrew antenna systems that utilize elevation over azimuth mounts with motor drives.

INTRODUCTION

The space shuttle disaster in January 1986 has had significant repercussions for the commercial satellite industry. Primary among these has been increased cost for new satellites and delay of launch schedules. It has caused satellite operators to adopt methods of extending existing satellite service life to support the capacity demand for space segment.

One method of extending the life expectancy of an existing geosynchronous satellite is by permitting a controlled decay in satellite orbit, to preserve stationkeeping fuel. This approach is known as the Inclined Orbit Maneuver (IOM).

SATELLITE ORBIT

A satellite is essentially a repeater located at an approximate altitude[1] of 22,236 statute miles above the Earth's equator. Commercial Communications satellites are most commonly placed in a geostationary orbit. This orbit is circular. It is selected because:

a. The period of rotation for the satellite matches the period of rotation of earth
b. The plane of rotation for this circular orbit coincides with the equatorial plane of the earth

Simply stated, the satellite <u>appears</u> to be stationary to an observer on the earth.

The mechanics of a geosynchronous orbit are derived from Kepler's first and third laws. (See appendix A for additional detail). Geosynchronous satellites are subjected to forces such as:

a. The gravitational attraction of the sun and moon.

b. The rotational force from the Sun.
c. The earth's gravitational field.

Because of these forces, the satellite tends to sway away from its track in East-West and North-South directions. In order to maintain the satellite in its allocated slot, some error correction is needed. The procedure used to correct this problem is called stationkeeping. Typical stationkeeping and maneuvering requirements for a spin-stabilized spacecraft are listed below (in feet per second):

260 fps = Initial error correction
2 fps = Attitude per year
175 fps = N-S Stationkeeping per year
6 fps = E-W Stationkeeping per year

In order to keep the satellite at the assigned latitude and longitude approximately 200 pounds of fuel, called monopropellant, are needed for 7 years. It is not generally economical to allocate more than 200 pounds of fuel for a commercial satellite.

The apparent stability of a geosynchronous satellite is defined by its 'stationkeeping window'. This window represents the East-West (E-W) and North-South (N-S) travel limits that apply to a satellite. Most domestic geosynchronous satellites are maintained in a stationkeeping window of 0.1 degree.

The stationkeeping activity propels the satellite to remain in its defined window which requires the consumption of fuel for the satellite's attitude thrusters. The amount of fuel available on a satellite and the amount of fuel required for stationkeeping is the primary (non-catastrophic) determinant to satellite life expectancy. Other useful life expectancy determinants of a satellite include:

a. battery life expectancy
b. solar panel durability
c. active component reliability
d. life expectancy of the mechanical bearings for spin stabilizers (gyroscope flywheel on satellite body)

The N-S stationkeeping activity requires approximately 96% of the total fuel allotment of the geosynchronous satellite. If the N-S stationkeeping activity is stopped in the 7th year for conserving fuel, the effective life of a satellite can be significantly increased. At that point a satellite with 95-98% of its initial fuel allotment consumed could have its N-S stationkeeping terminated. This would extend the satellite's E-W stationkeeping ability from a total of 7 years to 10-12 years. The resulting orbit is referred to as the 'inclined orbit'.

INCLINED ORBIT

The inclined orbit is an outgrowth of the geosynchronous orbit. It becomes a 'nursing home' for the aged satellites.

An inclined orbit satellite will have its N-S movement or 'eccentricity' increase from an initial value 0 at the center of the stationkeeping window. The rate increase in eccentricity is approximately 0.85 degrees per year. Thus, the inclined orbit tends towards an elliptical shape over a period of time and the major axis of this orbit is the north-south axis. Figure 1 illustrates diurnal movement of an inclined orbit satellite. The path[2] of a satellite in the inclined orbit is provided in Figure 2. This 'figure 8' shaped curve shows the motion of a satellite in the inclined orbit over a 7-day period.

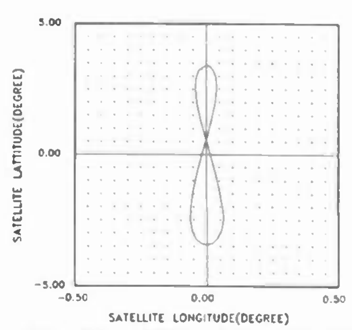

FIGURE 1: DIURNAL MOVEMENT OF AN INCLINED ORBIT SATELLITE OVER A 24 HOUR PERIOD

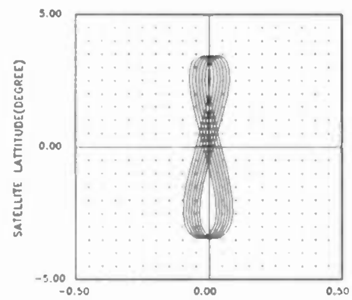

FIGURE 2: DIURNAL MOVEMENT OF AN INCLINED ORBIT SATELLITE OVER A 7 DAY PERIOD

The approximate eccentricity of an inclined orbit satellite is tabulated below for a 5 year period.

At end of Year	Eccentricity
1	0.85°
2	1.70°
3	2.55°
4	3.40°
5	4.25°

An earth station must track a satellite for efficient utilization of its power and bandwidth. Thus, the satellite owner' decision to use inclined orbits has motivated tracking manufacturers to design and produce an efficient and accurate tracking system capable of receiving signals from inclined orbit satellites.

Some of the following international and domestic satellites allowed to go into the inclined orbit are listed below:

a. INTELSAT satellites at 338.5° E, 60° E and 177° E
b. Domestic satellites, such as GSTAR III at 93° W, as well as SBS I & SBS II, both at 99° W
c. EUTELSAT F1 at 16° E

Motivation for Using Inclined Orbits

Some of the key arguments of satellite owners to keep their satellites in this orbit are:

a. As a storage orbit
b. To defray the expense of satellite replacement
c. To provide back-up to primary satellites
d. To provide additional service and revenue

However, the driving motivation for using an inclined orbit satellite is financial. To illustrate this, consider the following example:

A 12 transponder satellite (that has both horizontal and vertical polarity - 24 channels), is assumed to be launched for $240M. With a life expectancy of 7 years and an annual operating cost of $2.4M, this equates to an **annual fixed cost of $1.43M plus an annual operating cost of $0.1M per transponder per year**, in a very simplified model.

If, after 6 years of geosynchronous service, this satellite is permitted to enter an inclined orbit with an additional life expectancy of 6 years (12 years total), then the 'Fixed cost' of the last year of service can be amortized over 6 years. This results in an effective **annual fixed cost of $0.24M plus annual operating cost of $0.1M per year**, all things remaining constant.

Therefore, the motivation is the prospect of saving $7.14M of annual fixed cost. If the population of earth stations in network is small, then the cost associated with tracking an inclined orbit satellite can be readily justified. If the population is large, then the threshold for considering operation on an IOM satellite is:

($/station for tracking equipment) * (Number of stations) + (Nuisance factor $) <= Total $ saved on space segment.

When one considers that new satellites are projected to cost approximately twice as much as satellites launched 5 years ago, the financial motivation for considering more complex ground communications equipment becomes even more evident.

METHODS OF SATELLITE TRACKING

Tracking was not a requirement for C-band earth stations accessing satellites during the mid '60s and mid '70s. The decision to incorporate tracking into an earth station must include the following considerations:

a. Half-power (-3dB) beamwidth of the earth station antenna
b. Satellite orbit
c. Cost

As an antenna deviates off the peak beam, signal strength decreases. The solid angle between the -3dB points on the main beam of the antenna is called the half-power beamwidth[3] (see Figure 3.). Defined in degrees, the half-power beamwidth identifies the angle at which one-half the maximum transmit/receive signal is lost.

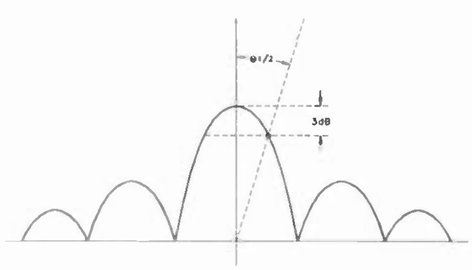

FIGURE 3: HALF POWER BEAM WIDTH

As antennas get larger and/or frequency increases the beamwidth gets smaller. Therefore, it is important to equip a larger or higher frequency antenna with a tracking system. A few significant types of tracking systems are discussed in the following paragraphs. Some of the important features for each of the significant tracking systems are compared[4] in Table 1.0 contained in Appendix C.

Monopulse Tracking

Monopulse tracking is a sophisticated tracking method. It is based on direct feedback of both satellite signal strength and direction. It is very accurate and operates on a real time nulling process, with AGC (Automatic Gain Control) and closed bandwidth beyond wind gust frequency components. This technique is not very common for commercial communications applications because of its high cost.

Step Tracking

This method is less complex than monopulse tracking. It requires moving the antenna's mechanical axis to cause the RF axis to execute excursions about a nominal position. Signal strength is sampled at the extremes of the excursion and the antenna is made to move in the direction of increasing signal strength. The scanning process continues until the antenna finds the peak signal.

Implementation of steptracking requires the following additional equipment:

a. Beacon receiver
b. Down converter
c. Antenna controller that has steptrack capabilities

The beacon receiver and down converter allow the earth station to use the beacon signal of a satellite. This beacon signal is constantly being transmitted by the satellite. When a reduction in signal strength is detected, the earth station 'knows' when to search for the new satellite position.

Steptracking is still an expensive solution for the smaller antennas.

Program Tracking

Program tracking is an accurate way to follow a satellite in the fog and rain for communications above 10 GHz. It is also reliable during the brief black-out periods that occur several times during the year. This method requires pre-determined satellite position data. It is a solution well-suited to smaller antennas. Larger antennas often utilize program tracking in the situations where receiving a signal has become difficult.

The satellite position information required to program track can be entered into the program track controller in several ways. Two of the methods are:

a. Using the Intelsat algorithm with

satellite location data provided by TT&C (Telemetry Tracking and Control) station. See appendix B for a description of the data provided and an example of the pointing angles generated.

b. Using the SBS I and SBS II algorithm currently in use by NBC. The data is supplied by an RS422 data link to the local station.

In each case, once the satellite ephemeris information, a set of time-based positional parameters, is received it is combined with station location data. The algorithm then generates antenna pointing angles. The resulting azimuth and elevation angles are transferred to the antenna controller which in turn commands the antenna to move to position.

HARDWARE REQUIREMENTS

This paper addresses the **Broadcast Market** requirements. Antennas are assumed to be in the 2.4 metre (SNG) to 9.1 metre size range.

Such antennas will require a two- or three-axis motorized mount. The antenna must have sufficient elevation and azimuth travel to view the satellite as it moves through its orbit. An elevation over azimuth mount is preferred, because it eliminates the need for coordinate transformations of the satellite antenna pointing predictions. These pointing angles are derived from such proprietary methods as the INTELSAT IOM algorithm, or from more conventional predictions of elevation and azimuth as a function of time. The use of other mount geometries, such as Az/EL mounts, Polar mounts, and Modified Polar mounts are more complex.

The motor drive systems used on antennas for tracking satellites impose a different set of concerns than those considered for rapid (within 60 seconds) repointing of antennas over the US domestic geosynchronous orbit arc. The broadcast market has focused its interest on such high speed motor drives, configured as follows:

a. High Speed Gear motors... Typically dual speed with failsafe electrical brakes. In the case of Andrew motor drives, these operate at a 10:1 gear ratio, resulting in a High Speed of ~2°/second, and a Low Speed of ~0.2°/second.

b. Dual Winding/Dual Speed motors provided with failsafe electrical brakes are available with a 4:1 drive speed ratio. In the case of Andrew motor drives, these are available in a High speed configuration with drive speeds of ~1.5°/second and ~0.4°/second. A Medium Speed drive system with drive speeds of ~0.1°/second and ~0.03°/second is also available.

c. Variable speed motor drives... Typically Variable frequency drives or DC motor drives are available with drive speed ratios up to 6:1.

For tracking applications, the antennas are moved in small increments to minimize the gain degradation between steps. The suitability of existing motor drive systems for use with inclined orbit satellites will be determined by **the low speed of the motor drive and the antenna's 3 dB beamwidth.**

We have introduced the antenna's 3 dB beamwith, because this defines the angle off of boresight resulting in a 3 dB reduction of receive signal. Three (3) dB reduction of signal is impractical between tracking steps, therefore more common tracking recommendations are based on 0.2-0.5dB signal variation between tracking steps. This degree of signal variation corresponds to angular movement off of boresight approximating

0.1 * (the 3 dB beamwidth)

Consequently the drive speeds for low cost satellite tracking systems are typically equal to a rate of

[0.1 * (3 dB beamwidth)] / second.

For some C-Band and Ku-Band earth station antennas, the drive speeds for satellite tracking are:

Antenna Size (m)	Frequency Band	-3dB Beam-width (°)	Nominal Motor Drive Speed (°/sec)
9.1	C	0.51	0.050
7.3	C	0.64	0.064
7.3	Ku(Hybrid)	0.28	0.028
5.6	Ku	0.28	0.028
4.6	Ku	0.34	0.034
2.4 (SNG)	Ku	0.65	0.065

In the case of the 9.1, 7.3 and 5.6 metre antennas, Andrew can provide High (Dual) Speed gear motors with 100:1 gear ratios, resulting in nominal drive speeds of 2°/second in high speed and 0.02°/second for low speed. Low/Medium speed motor drives are available for all Andrew motorizable antennas.

ANDREW SOLUTIONS

Andrew solutions encompass various tracking systems. Individual customer requirements are evaluated and a specific solution is recommended. Such recommendations are based upon antenna size, application, utilization, feasibility, and tracking implementation costs.

Three program tracking concepts are proposed:

1. PC-Based Software Option

The least expensive option provides for a PC interface to the antenna control unit. It allows tracking of a single satellite. Ephemeris data can be entered via a keyboard, or automatically transferred through an RS232 port. The calculations are performed using the appropriate algorithm. Antenna movement commands are sent to the antenna control unit at the times calculated. The operation is single-task and the PC is dedicated to program tracking functions. See Figure 4. The antenna control unit continues to provide local control with 40 programmable satellite positions, and remote control via an RS232 port.

2. System Controller Option

A more flexible solution involves the use of a system controller. The system controller provides program tracking as well as other general earth station monitor and control

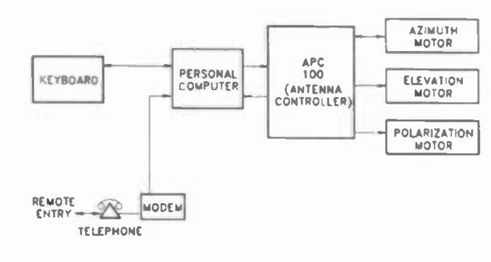

FIGURE 4: ANDREW PROGRAM TRACKING

activities. Because the system controller can support more than one antenna control unit, more than one satellite could be tracked at a time. In fact, it is proposed that more than one tracking algorithm be supported. The ephemeris data provided by Intelsat, or its equivalent, could be entered into the algorithm via keyboard, modem, computer, or satellite. Because of the many options for acquiring and disseminating data, each site would be evaluated for the best possible data entry method. See Figure 5.

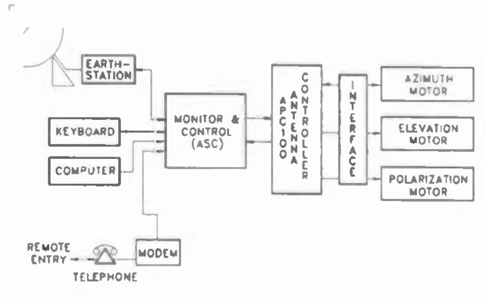

FIGURE 5: ANDREW HYBRID TRACKING (OPTION)

3. Hybrid Options

Options such as a Hybrid or Steptrack can be provided. The hybrid option includes an Andrew supplied PC (or monitor and control system), software, and the customer furnished antenna system. The steptrack option is equipped with TEA (Trajectory Estimation Algorithm)[4], married to an

Andrew interface and hardware which eventually drives the antenna.

CONCLUSION

As a satellite approaches the end of its fuel supply, the supplier may seek to extend its useful life by shutting off the North-South stationkeeping. The satellite then goes into an inclined orbit, requiring earth stations to have more sophisticated tracking mechanisms to transmit or receive a signal. Earth station owners able to track these inclined orbit satellites may take advantage of the significantly lower cost.

Although tracking has been available for some time, it has traditionally been a dedicated application. Recent technological advances allow economical online communications that provide program tracking as a part of total earth station control.

ACKNOWLEDGEMENT

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APPENDIX

APPENDIX A. GEOSYNCHRONOUS ORBIT

Satellites are being placed into a variety of space orbits. Molniya and geosynchronous orbits are very well known. Molniya orbit is a very elliptical orbit and has an inclination of 63.4° ($5 \cos^2 i - 1 = 0$, where i = inclination angle). This inclination permits the apogee to remain over the northern hemisphere and its period is synchronized at half a sideral day. This orbit is being utilized by Warsaw Pact countries. Most of the remaining world satellites utilize the geosynchronous orbit.

Three important parameters for geosynchronous orbit are: the nature of the curve (orbit), the revolution period, and the height of the geosynchronous orbit are derived from Kepler's first and third laws.

The first law[5] states that the orbit of a satellite is a curve of the second order, one focus of which is the center of the earth. It is represented by the following equation[5]:

$$r = \frac{P}{1 + e \cos \theta} \qquad (1)$$

where:
- r = motion of the curve
- p = parameter of the second order
- e = eccentricity
- θ = central angle

If e = 0, then the curve is a circle
If e < 1, then the curve is a parabola
If e = 1, then the curve is a ellipse
If e > 1, then the curve is a hyperbola

Based on this law, the curve of the geosynchronous orbit 'r' is equal to P when eccentricity 'e' is equal to zero.

The revolution period of a satellite moving in a geosynchronous orbit is based upon Kepler's third law which states that the square of the revolution period of the satellite is proportional to the third power of the semi-major axis of the orbit .

$$T^2 = \frac{4\pi^2}{\mu} a^3 \qquad (2)$$

where:
T = revolution period
μ = Kepler's constant
= 398,513.52 km^3/sec^2
and
μ = GM
where:
G = Universal gravitational constant
= 6.67×10^{-20} $km^3/kg/sec^2$
M = Mass of planet Earth
= 5.977×10^{24} kg
and
a = R + h
where:
a = Semi-major axis of elliptical orbit
R = radius of the earth
= 6,370 km
h = Altitude of a satellite moving in a circular orbit

Thus, equation (2) reduces to:

$$T^2 = \frac{4\pi^2}{\mu} (R + h)^3 \qquad (3)$$

Equation (3) explains that the revolution period, T, is proportional to the height of the satellite orbit from the planet Earth's orbit. This conclusion is depicted in the graph contained in Figure 6.

The graph also shows that the revolution period decreases if the satellite orbital altitude is reduced. Thus, the altitude of the geosynchronous orbit, h, should be approximately 35,786 km higher from the earth orbit in order to maintain a period of 24 hours, which is equivalent to the rotational period of the earth. The velocity (V) of a satellite in the geosynchronous orbit is given by:

$$V = GM/a \qquad (4)$$

APPENDIX B. EPHEMERIS DATA EXAMPLE

The eleven ephemeris parameters provided by the INTELSAT TT&C station operator are:

1. Mean Longitude (East of Greenwich), in degrees

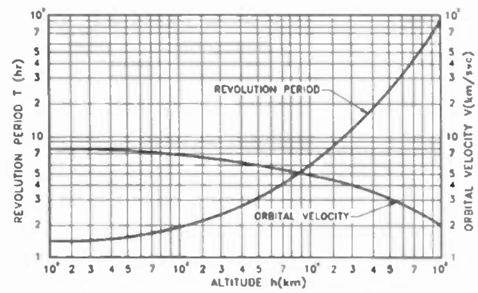

FIGURE 6: RELATION BETWEEN ALTITUDE, REVOLUTION PERIOD AND ORBITAL VELOCITY OF A SATELLITE

2. Drift Rate, in degrees per day
3. Drift Acceleration, in degrees per day per day.
4. Amplitude of the cosine component of the longitude oscillation, in degrees.
5. The rate of change in the longitude oscillation cosine component, in degrees per day.
6. Amplitude of the sine component of the longitude oscillation, in degrees.
7. The rate of change in the longitude oscillation sine component, in degrees per day.
8. Amplitude of the cosine component of the latitude oscillation, in degrees.
9. The rate of change in the latitude oscillation cosine component, in degrees per day.
10. Amplitude of the sine component of the latitude oscillation, in degrees.
11. The rate of change in the latitude oscillation sine component, in degrees per day.

In addition, the following data are provided to help in determining when to send changes of position to the antenna control unit:

1. Identification of the Satellite for which parameters are being sent
2. The Nominal Center of Box position
3. The Date and Time of Epoch
4. The estimated satellite position at epoch + 170 hours

Example of pointing angles over a 24 hour period

```
ZUR-3E1 ZURICH   EARTH STATION DESIGNATOR
      47.393600 EARTH STATION NORTH LATITUDE (DEGREES)
     108.501100 EARTH STATION EAST LONGITUDE (DEGREES)
        .500000 EARTH STATION HEIGHT ABOVE IAU-1977
                ELLIPSOID (KM)
      60.00     TIME INTERVAL BETWEEN AZ/EL POINTS  (MINUTES)
       7.0      DURATION OF POINTING INFORMATION GENERATION
                (DAYS)
     177.00     CENTER OF BOX .02 LONGITUDE BOX .02 LATITUDE BOX

 EPHEMERIS VALUES
 LM0    =      176.91560         176.9156000
 LM1    =      .70000E-03          .0007000
 LM2    =      .18700E-03          .0001870
 LONC   =     -.67000E-02         -.0067000
 LONC1  =     -.80000E-03         -.0008000
 LONS   =      .50800E-01          .0508000
 LONS1  =     -.50000E-03         -.0005000
 LATC   =     -.10194E+01        -1.0194000
 LATC1  =     -.13000E-02         -.0013000
 LATS   =     -.14084E+01        -1.4084000
 LATS1  =     -.37000E-02         -.0037000

 EPOC DATE 20/ 1/90     OF EPHEMERIS  (DAY/MONTH/YEAR)
 EPOC TIME  0: 0: 0     OF EPHEMERIS  (HOUR:MINUTES:SECONDS)

 PREDICTED  SATELLITE POSITION AT +170 HOURS = 176.941       .022
 CALCULATED SATELLITE POSITION AT +170 HOURS = 176.941     -1.685
```

ELAPSED HOURS	YR:MO:DA:HR:MN:SC	AZ(DEG)	EL(DEG)	SAT LONG (DEG)	SAT LAT (DEG)
.0	90: 1:20: 0: 0: 0	106.904	5.163	176.8964	-1.0194
1.0	90: 1:20: 1: 0: 0	107.114	4.906	176.9094	-1.3501
2.0	90: 1:20: 2: 0: 0	107.260	4.717	176.9254	-1.5883
3.0	90: 1:20: 3: 0: 0	107.332	4.609	176.9428	-1.7177
4.0	90: 1:20: 4: 0: 0	107.327	4.590	176.9591	-1.7295
5.0	90: 1:20: 5: 0: 0	107.246	4.663	176.9719	-1.6228
6.0	90: 1:20: 6: 0: 0	107.096	4.823	176.9791	-1.4049
7.0	90: 1:20: 7: 0: 0	106.886	5.061	176.9794	-1.0907
8.0	90: 1:20: 8: 0: 0	106.632	5.359	176.9728	-.7017
9.0	90: 1:20: 9: 0: 0	106.349	5.698	176.9604	-.2646
10.0	90: 1:20:10: 0: 0	106.056	6.054	176.9441	.1907
11.0	90: 1:20:11: 0: 0	105.772	6.401	176.9266	.6331
12.0	90: 1:20:12: 0: 0	105.516	6.714	176.9101	1.0322
13.0	90: 1:20:13: 0: 0	105.304	6.972	176.8965	1.3606
14.0	90: 1:20:14: 0: 0	105.152	7.157	176.8869	1.5958
15.0	90: 1:20:15: 0: 0	105.071	7.256	176.8812	1.7218
16.0	90: 1:20:16: 0: 0	105.067	7.263	176.8787	1.7298
17.0	90: 1:20:17: 0: 0	105.143	7.178	176.8781	1.6193
18.0	90: 1:20:18: 0: 0	105.293	7.009	176.8783	1.3979
19.0	90: 1:20:19: 0: 0	105.508	6.767	176.8786	1.0806
20.0	90: 1:20:20: 0: 0	105.772	6.469	176.8788	.6893
21.0	90: 1:20:21: 0: 0	106.066	6.135	176.8797	.2506
22.0	90: 1:20:22: 0: 0	106.370	5.788	176.8823	-.2053
23.0	90: 1:20:23: 0: 0	106.662	5.450	176.8879	-.6472
24.0	90: 1:21: 0: 0: 0	106.920	5.143	176.8972	-1.0449

APPENDIX C. COMPARISON OF TRACKING TECHNIQUES

Table 1.0

Tracking Technique Comparison

Evaluation Parameters	Monopulse Tracking	Step Tracking	Program Tracking Other	Program Tracking Andrew
A prior data requirement	No	No	Yes	Yes
Minimal scanning	Yes	No	Yes	Yes
Handle signal loss	No	Yes	Yes	Yes
Handle fade	Yes	Yes	Yes	Yes
Transducer accuracy req'd	No	No	Yes	Yes
Transducer, fine resolution	No	No	Yes	Yes
Rapid acquire	Yes	Yes	Yes	Yes
Steady wind	Yes	Yes	No	Yes
Gusty wind	Yes	No	No	No
Portable	Yes	Yes	Yes	Yes*
High cost	Yes	Yes	No	No
Inclined Orbit Satellite	Yes	Yes	No	Yes

* PC-based APT unit makes it extremely compact.

MULTICHANNEL AUDIO MULTIPOINT DISTRIBUTION SERVICE: 2 GHZ BACKGROUND MUSIC

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Recent tests indicate that the 2 GHz frequency range can be effectively utilized for multichannel background music distribution. Rule changes regarding the use of the Multipoint Distribution Service (MDS) band (2150-2162 MHz) allow the use of this band for this purpose. Coupled with this regulatory change is a diminishing use of this band for television transmission and the concurrent availability of multiple background audio channel satellite feeds. The ability of a background music supplier to own and operate a system and avoid the need to lease subcarriers is an incentive to adopt this new technology.

Hardware developed originally for television transmission and reception has been adapted for multichannel audio service and achieves a degree of performance that can improve the quality of the background music signal normally delivered. This paper deals with the technical aspects of utilizing this technology and reports on the results obtained in field tests. It also provides the potential user with guidelines to follow to pursue licensing this mode of operation.

BACKGROUND MUSIC DELIVERY

As background music evolves from a single-channel service to a multiple channel array of music formats, it becomes desirable to find a more effective delivery means. While direct satellite-to-user delivery is possible, the cost effectiveness and technical dependability of this approach limits its use. Methods of adding extra audio channels to existing SCA systems produce trade-offs in signal quality and generally limit the service to two audio channels. Present feeds from the suppliers of this service, such as MUZAK, include up to six channels of music.

Multipoint Distribution Service (MDS) was established as a common carrier service by the FCC in the early 70's. Twelve MHz were allocated to this service, in the band from 2150 to 2162 MHz. The presence of other services in the band above 2160 MHz caused the FCC to allocate only 10 MHz of available spectrum in certain geographical areas. This band was further divided into Channel 1, 2, and 2A. Channel 1 and 2 are each 6 MHz allocations. In areas that have only 10 MHz total spectrum available, the 10 MHz is divided into a 6 MHz Channel 1 and a 4 MHz Channel 2A. As a common carrier service it was intended that licensees lease the available channels to programmers to deliver television and/or other information.

While premium television evolved as the prime use of this service, it was always the case that the regulations allowed for the delivery of other types of data or entertainment programming. Hundreds of MDS channel 1 and a modest number of channel 2 stations were built and operated. In a few cases, channel 2A was used for a black and white television service or a specialized subcarrier delivery service.

MULTICHANNEL MULTIPOINT DISTRIBUTION SERVICE

As the demand for multiple channel television service increased and as it became apparent that the 2.5 GHz Instructional Television band was under-utilized, a proposal was made to authorize a new group of channels to MDS. In the early 80's, 8 channels were set aside in the band from 2600 to 2648 MHz. This band allowed for 8 television channels and was designated the Multichannel Multipoint Distribution Service (MMDS). MMDS systems are currently in use or being constructed in numerous cities. In many of these cities the original MDS operation remains in use but in numerous cities the single channel MDS band has been abandoned. Growth in terms of the addition of channel 2 or 2A in most areas has been curtailed. In most areas of the US, channel 1, 2, or 2A lies dormant. These 6 or 4 MHz bands are a valuable resource that can potentially carry a great deal of information.

The 2 GHz frequency band is similar to UHF in its propagation characteristics. Line-of-site propagation, taking into account refraction and attenuation caused by minor obstructions, allows the service to have moderate coverage areas for television purposes. Equipment for these frequency ranges are relatively inexpensive because of the low operating frequency. Rain attenuation and other microwave propagation anomalies are not significant factors.

MICROWAVE BACKGROUND MUSIC

The transmission of multiple channels of audio is a relatively straightforward process. In addition to quality audio performance, it is desireable that inexpensive receive hardware be utilized. Conversion of the 2.15 GHz band to VHF can be done inexpensively using converters originally intended for television reception. Reception and demodulation of VHF signals can readily and inexpensively be accomplished with crystal controlled or synthesized receivers originally intended for background music operation. The use of standard receive hardware leads to a simple system design, employing frequency modulation of carriers in a frequency division multiplexed channel. A complete system is illustrated in Figure 1.

Transmitters originally intended for television transmission employ linear wideband amplifier stages that can handle multiple carriers without significant distortion as long as the peak envelope power of the combined spectrum is held within the capability of the transmitter amplifier. The limitation in transmitter power is similar to the limitation imposed on television carriers in a cable television system. The limiting distortion is known as composite triple beat and is a result of the carriers beating together producing a plethora of low-level intermodulation products. By judicious selection of the individual carrier frequencies it is possible to operate a 100 watt television transmitter at a level that produces useful FM carrier powers with up to a dozen FM carriers present. Figure 2

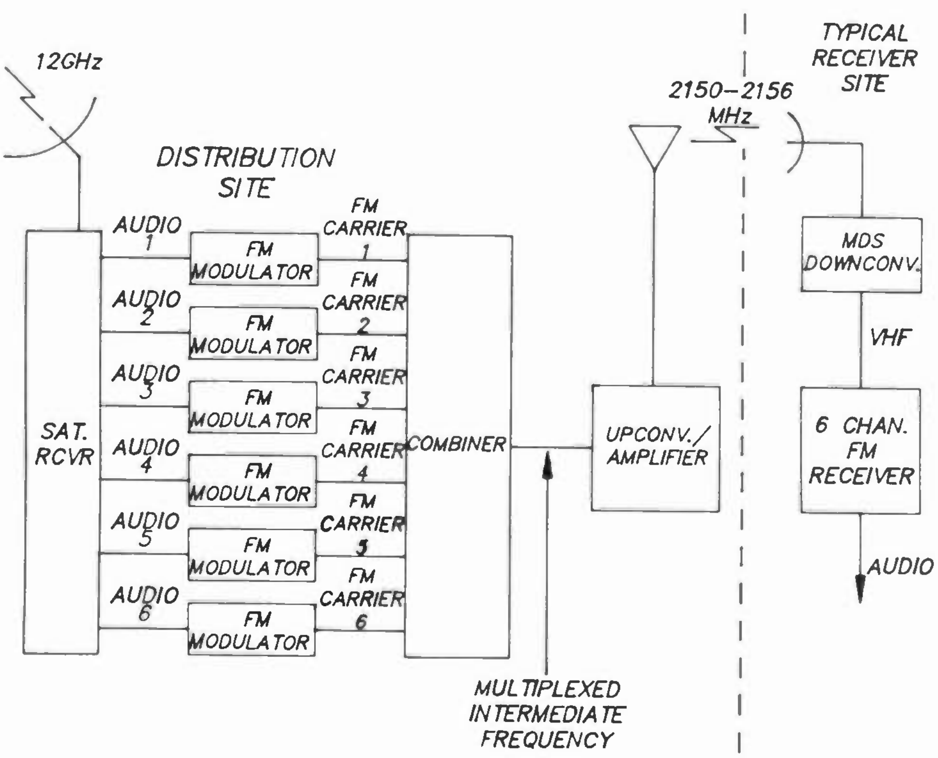

Figure 1 - MDS MULTIPLE CHANNEL BACKGROUND MUSIC SYSTEM BLOCK DIAGRAM

illustrates a spectrum display of 8 FM carriers operating in a 4 MHz band. Note the level of composite triple beat distortion products below the carriers.

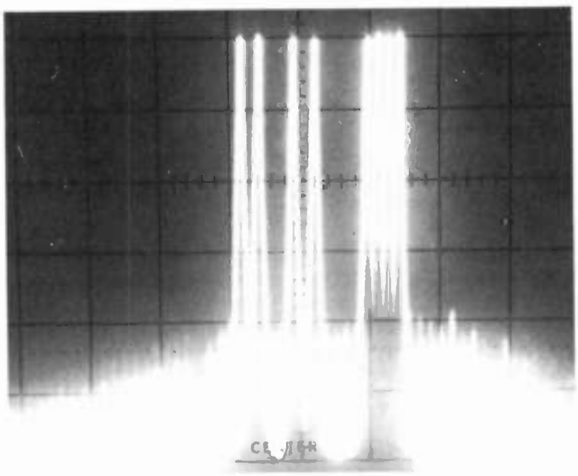

Figure 2 - Spectrum Display

MDS service requires that spurious components be held more than 40 dB below operating carrier levels. Using this criterion of performance, a 100 watt television amplifier can operate 6 channels at approximately 10 watts per channel.

SERVICE AREA

Coverage area depends on receive field strength and the sensitivity of the receive hardware. Inexpensive 2 GHz downconverters are available with noise figures better than 5 dB. FM receivers typically are constructed with approximately 200 KHz of IF bandwidth. It is a straightforward process to determine receive system sensitivity. In practice, 3 microvolts into the downconverter produces an acceptable audio performance. With 10 watts of output power per carrier the coverage achieved in a multiple channel FM system greatly exceeds the coverage area achieved by a 10 watt television transmitter. Television requires several hundred microvolts into the same downconverter to achieve acceptable picture performance. FM service requires 100th as much voltage (40 dB less) to achieve acceptable performance. This significant advantage is primarily due to the noise bandwidth difference between FM and television. Field tests in Reno, Nevada indicate that 30 to 50 miles is a reasonable expected coverage radius.

In general, a degree of path obstruction can be tolerated. Direct radio line-of-site is not required since even in the diffraction region of coverage, behind an obstruction, it is possible to receive the required several microvolts of signal per carrier. There is activity in process to license "boosters" for use in filling in shadowed areas. This should allow this service to achieve coverage superior to typical SCA/FM operation. This service has the potential to provide multiple channels of coverage in a wide service area with quality approaching that of main FM carrier performance.

LICENSING

As the MDS and MMDS services evolved in the 80's, the FCC ruled that this service would be allowed to operate as either a common carrier or originator of signals. This allowed the owner of the transmission hardware to be the supplier of signal services. Given these options, it is possible that a licensee could choose to build an MDS station and lease the available spectrum to a background music supplier. It is also possible that the background music supplier could purchase and licence an MDS system directly. This service can be equal or superior to SCA service and in addition provide a multiple channel format with ease.

MDS is authorized under Part 21, Subpart K of the rules. Form 435 is the required license application. The application requires an analysis of the interference potential and a complete engineering description of the proposed transmitting facility and service area. The emission designator for this service is 4M00F8E or 6M00F8E, depending on whether it is a 4 or 6 MHz channel respectively.

SUMMARY

The evolution of microwave television into a multiple channel service has generated available spectrum that can readily be applied to a multiple channel audio service. The available spectrum is efficient and predictable and the hardware is moderately priced. Persons interested in supplying multiple channels of audio should take note of this technology. They will find it compares favorably to the other alternatives available.

PBS AND THE NEXT GENERATION OF SATELLITES

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The Public Broadcasting Service (PBS) is a private, nonprofit corporation whose members are the nation's public television stations operating some 337 television transmitters.

PBS established American television's first satellite program distribution system in 1978. Westar IV, the currently used satellite, will reach its end of life in 1991, presenting opportunities for PBS to help specify and to use technologies for the next generation of satellites to be launched in the 1992/1993 time frame.

This paper provides a brief historical review of the PBS satellite system, and describes the plans for new space and ground segments that will take public television into the next century.

Introduction

The Public Broadcasting Service (PBS) is a private, nonprofit corporation whose members are the nation's public television stations operating some 337 television transmitters. Founded in 1969, PBS provides quality television programming and related services to noncommercial stations serving the United States, Puerto Rico, the Virgin Islands, Guam, and American Samoa.

PBS established American television's first satellite program distribution system in 1978. Westar IV, the currently used satellite, will reach its end of life in 1991, presenting opportunities for PBS to utilize, and help specify, technologies for the next generation of satellites to be launched in the 1992/1993 time frame.

This paper provides a brief historical review of the PBS satellite system, and describes the plans for new space and ground segments that will take public television into the next century.

PBS Background

PBS initially distributed television programming across the United States through a national terrestrial network with the Bell System and other common carriers. PBS began its initial study of satellites in February 1974[1]. There were no U.S. domestic commercial communications satellites in service; Canada, however, had Anik I and Anik II satellites in orbit providing the Canadian Broadcasting Corporation with television distribution to remote northern provinces. Western Union, RCA, and AT&T/COMSAT/GTE subsequently launched satellites beginning in 1974. A feasibility study was performed using a TelePrompTer Corporation transportable receive-only earth station with Anik satellites during early 1974. Success in the study resulted in a contract with COMSAT for a separate paper study of all sites. In mid-74, a system definition phase began leading to a number of decisions: each public television station should own and operate its own earth station, three to four transponders in the existing fully developed C-band (5.925-6.425 GHz uplink and 3.700-4.200 GHz downlink) should be leased, all stations should simultaneously receive two program feeds, and uplinking facilities should be provided by PBS and at least one station in each of the four public television regions (eastern, southern, central and western). Such a system appeared to be economically feasible for public television.

During 1975, negotiations led to the selection of Western Union for the space segment and Collins Telecommunications Group of Rockwell International for the ground segment.

In 1976, a Satellite Interconnection System Project Ofice (SISPO) was established by the Corporation for Public Broadcasting (CPB) and PBS. Satellite service was leased from Western Union on the Westar I satellite at 99 degrees West Longitude, at a cost of $800,000 per transponder per year.

Construction of receive-only earth stations to serve about 165 public television stations cost about $25 million, regional uplinks, $1.25 million, and the main origination terminal near Washington DC, $5 million. Other costs including project management and interest costs brought the capitalized cost to about $39.5 million.

All earth stations used 10 meter Andrew antennas, except Puerto Rico and the Virgin Islands where 15.5 meter E-Systems antennas were installed. The figure of merit or G/T ranged from 25 to 27 dB/K on the mainland, 30 in Alaska, 31 in Hawaii to 35 in Puerto Rico and the Virgin Islands. All sites were equipped with two Collins video receivers. C/N for each site was a minimum of 14 dB and receiver threshhold was 11.5 dB. All terminals were designed for a 15 year lifetime.

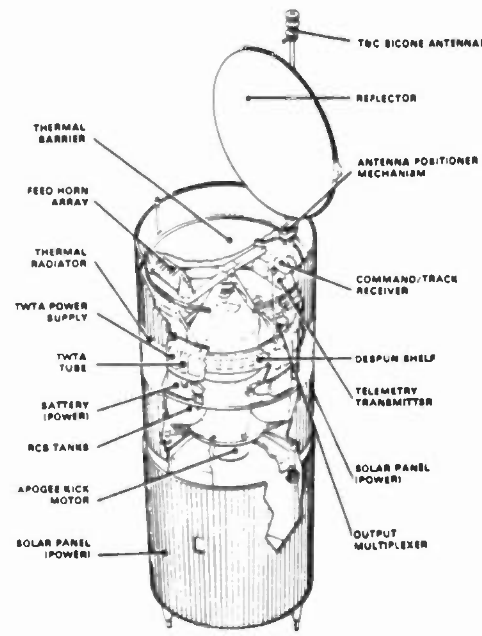
The system was completed and became operational in September 1978, and has served public television well for some 11 years. It provided an improved television distribution system for the national program service, eduational services, and groups of stations exchanging programs, while achieving long range cost savings.

AT&T Background

AT&T Bell Laboratories in 1954 presented formal papers on satellite communications three years before the USSR launched its Sputnik satellite[2]. Echo I, a passive balloon 100 feet in diameter was launched in 1960 and reflected signals between Holmdel, NJ and Jet Propulsion Laboratory in Goldstone, CA. Telstar I, the first active communications satellite, was launched in 1962. It was a 34 inch sphere weighing 170 pounds, and permitted trans-Atlantic communications. Telstar II was launched in 1963 and permitted the reporting on events on both sides of the Atlantic. During the 1970s, AT&T with GTE and COMSAT launched the Comstar series, D1 through D4. In the 1980s, AT&T launched the Telstar 3 series, 301, 302 and 303, with the ABC and CBS television networks as major clients. The 1990s will bring the Telstar 4 series, with ABC and PBS as major clients.

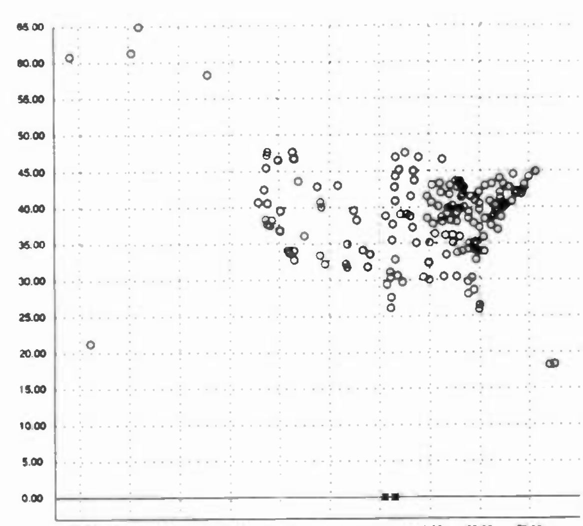
PBS Today

Westar I was replaced by Westar IV at 99 Degrees West Longitude in 1982. PBS presently owns four transponders on Westar IV. A drawing of Westar IV, a spin-stabilized satellite, is shown in figure 1. The system has expanded to 178 earth station sites as shown in figure 2, a computer plot of sites by latitude and longitude. Satellite positions at 97 and 99 degrees West Longitude are shown also in figure 2.



HS 376G Westar IV, V, IV(S)
General Layout

Figure 1.



PBS EARTH STATION SITES
Degrees North Latitude by Degrees West Longitude

Figure 2.

Taped programs originate at the PBS Technical Operations Center in Alexandria VA, and are fed by microwave to the PBS Satellite Operations Center uplinks in Bren Mar VA some 7 miles distant.

Four regional uplinks are located in Albuquerque NM, Columbia SC, Hartford CT and Lincoln NE. Another 14 uplinks are located throughout the nation, and provide uplinking of additional programs including **LIVE FROM LINCOLN CENTER** and **LIVE FROM THE MET.**

PBS provides a national program service with programs such as the **MACNEIL/LEHRER NEWSHOUR, AMERICAN PLAYHOUSE, NATURE, GREAT PERFORMANCES, MASTERPIECE THEATRE** and **SESAME STREET.** Educational services include the Elementary/Secondary Service, Adult Learning Service, Adult Learning Satellite Service and the Business Channel.

PBS has taken engineering leadership roles in satellites, high efficiency UHF television transmitters, vertical blanking interval data services, closed captioning for the hearing impaired and now, descriptive video services for the visually impaired.

Westar IV is expected to expend its stationkeeping fuel in 1991. PBS issued a request for proposals for replacement space segment in 1987. Evaluations and final negotiations led to a contract with AT&T Communications, Inc. for next generation transponder service on Telstar 401 to be launched in February 1993. Interim service from 1991 to 1993 will be provided on GTE Spacenet.

C-band/Ku-band Considerations

Telstar 401 will be a hybrid satellite with 24 C-band and 20 Ku-band transponders. (C-band includes 5.925-6.425 GHz uplinks and 3.700-4.200 GHz downlinks; Ku-band includes 14.0-14.5 GHz uplinks and 11.7-12.2 GHz downlinks). A drawing of Telstar 401 is shown in figure 3. PBS will purchase up to six transponders with the final mix of C-band and Ku-band to be decided by December 1990. C-band has served public television well for 11 years, but Ku-band offers some exciting additional opportunities. Included are wider bandwidth 54 Mhz transponders (compared to 36 MHz at C-band), which provide additional information carrying capacity for dual video per transponder for some services, digital television, digital data services, and future digital High Definition Television. The digital data services can provide increased telecommunications capabilities among the stations plus interactive communications for educational services. The higher frequencies and higher powered satellites permit smaller cost-effective antennas. Ku-band is free from terrestrial microwave interference and uplinks and downlinks can be more easily collocated with studio sites avoiding the cost and spectrum use for microwave interconnection systems.

Ku-band has one significant problem in increased rain attenuation compared to C-band, especially during very heavy rain activity in the southeastern and Gulf coast regions of the United States. PBS is studying this problem in several areas. A study is underway at Virginia Polytechnic Institute and State University in Blacksburg VA, for a site by site analysis to predict rain fade margins and availabilities. Our goal is 99.99 percent availability or about 53 minutes per year of outage. This is better than our present C-band performance when sun outages in the Spring and Fall are considered. Sun outages are predictable, however, and rain is not. Help is also being received from NASA Lewis Research Center, Jet Propulsion Laboratory, NBC Television Network and COMSAT. Public television stations will be conducting rain fade measurements at ten sites around the country, including Miami FL and Alexandria VA.

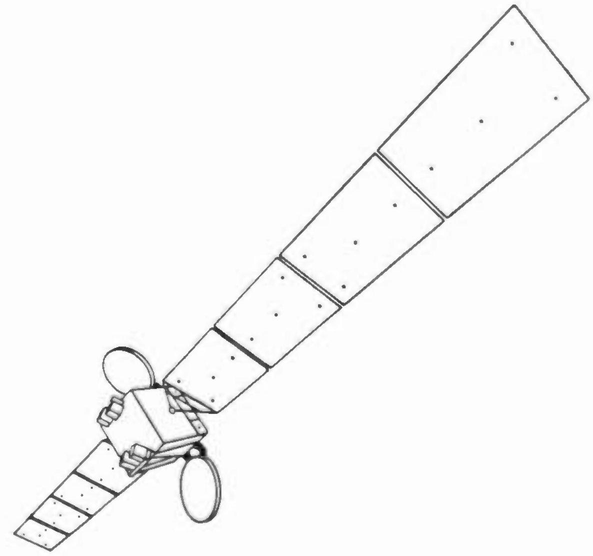


FIGURE 3 TELSTAR 4 ON-ORBIT CONFIGURATION

Telstar 4

AT&T's next generation of satellites, the Telstar 4 series, will be launched beginning February 1993, as replacements for the Telstar 301 and 302 satellites which will be reaching the end of their lives. PBS will purchase up to six transponders on the first of the series, Telstar 401, to be located at 97 degrees West Longitude. The series includes two satellites plus a ground spare[3]. The hybrid design includes 24 C-band transponders with variable power from 10 to 20 watts, 36 MHz bandwidth, and 20 Ku-band transponders with variable power up to 120 watts, and bandwidths of 36 MHz and 54 MHz. PBS is purchasing C-band transponders with 16 watts and Ku-band transponders with 60 watts and 54 MHz. Coverage will be provided to all 50 states, Puerto Rico and the Virgin Islands. Power levels will be adjustable by ground control.

Effective isotropic radiated power (EIRP) levels will be a minimum at CONUS edge of coverage of 35.0 dBW for C-band and 44.0 dBW for Ku-band. Saturation flux densities (SFD) for CONUS range from -74 to -92 dBW/m^2 at C-band and -78 to -96 dBW/m^2 at Ku-band. Input attenuators will be ground controllable in 3 dB steps to a maximum of 18 dB.

PBS will have at least one C-band/Ku-band transponder pair cross-strappable on ground command. This will permit uplinking on C-band and downlinking on Ku-band or both Ku and C-band. Alternately, the uplink can be Ku-band with downlinking on C-band or both C and Ku-band.

Telstar 4 incorporates an anti-intrusion capability to defeat and locate intruders. The system has four uplink spot beams, each covering one-quarter of CONUS, and a special switch on board the spacecraft operated by the ground control center. If a particular transponder experiences interference in the normal receive beam mode, the spot beam mode can be implemented to avoid and/or locate the interferor.

The Ku-band receivers on Telstar 4 also have ground controllable limiters to provide up to 15 dB of overdrive protection.

AT&T has selected General Electric Astro Space Division to design and build the new GE-7000 series spacecraft to Bell Laboratories specifications. The three-axis stabilized design will include state-of-the-art payload, bus and ground-station technology.

The spacecraft will be compatible with launch and geostationary transfer orbit injection from the Guiana Space Center by an Ariane 4 or from Cape Canaveral by an Atlas Centaur IIAS launch vehicle. A liquid bipropellent apogee engine will be used to move the satellite to geostationary orbit.

Stationkeeping fuel will be provided for a twelve year lifetime, maintaining tolerances of +/- 0.05 degrees.

Thermal control will use conventional passive techniques plus heat pipes and heaters. Other reliability features include redundant electrical units, redundant attitude control sensors, cross-strapping of electronic equipment, redundant propellent valve drivers and seats, seven-for-six redundancy in C-band SSPAs, 18-for-10 redundancy in Ku-band TWTAs, and four-for-two redundancy in both C-band and Ku-band receivers.

Telemetry, Tracking and Command (TT&C) will be provided from AT&T facilities at Hawley PA and Three Peaks CA.

Ground segment

The existing C-band ground equipment is 11 to 12 years old. The 10 meter antennas are in generally good condition and will be refurbished as required. Electronics are being replaced and new receivers are being introduced into the system. Uplinks require new antennas or reflectors to meet FCC two degree satellite spacing requirements.

The new Ku-band facilities will require a complete new ground segment. System planning and design is in process. New Ku-band uplinks will be provided for the PBS Satellite Operations Center in Alexandria and the four regional sites in Albuquerque, Columbia, Hartford and Lincoln. New Ku-band downlinks will be provided for all other sites.

Federal Funding

The Public Telecommunications Act of 1988, P.L. 100-626, November 7, 1988, established a Public Broadcasting Satellite Interconnection Fund to finance the capital costs of the replacement, refurbishment, or upgrading of the public television and public radio national satellite interconnection systems and authorized an appropriation of $200 million for fiscal years 1991 to 1993. These funds are to be distributed through the Corporation for Public Broadcasting

(CPB) to the licensees of noncommercial educational broadcast stations or to the national entity they designate. The public television stations have designated PBS as that national entity.

Summary

The existing C-band satellite system has provided a flexible cost-effective television distribution system for some 11 years. The features of this system will be retained and expanded with the new technologies available on the next generation satellites to serve public television and the American people well into the next century.

Acknowledgements

The author acknowledges the assistance and guidance of the PBS Board of Directors, the Public Television Interconnection Committee, PBS Management including Bruce Christensen - President, Neil Mahrer - Executive Vice President, Howard Miller - Senior Vice President, Broadcast Operations and Engineering, and the work of colleagues on the PBS Satellite Replacement Team including Peter Downey, Nancy Hendry, Paula Jameson, Louise Lynch, Shelley Mitek, David Sillman, Beth Wolfe and Gwen Wood and also Jim Hargreaves, Ralph Mlaska and Ralph Schuetz. Also acknowledged are the teams at AT&T Communications, Inc. and AT&T Bell Laboratories.

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PROGRAM QUALITY AUDIO OVER DIAL-UP TELEPHONE LINES

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Engineers are often required to setup remote broadcasts from across town or across the country. The technical considerations involved in providing a high quality audio feed are often complicated by restrictions imposed by station budgets and/or the time required to arrange for the best quality feed. Telephone lines can be used as the transmission medium in many cases, but their inherent limitations make the use of external equipment necessary in order to improve the fidelity of such remote broadcasts.

INTRODUCTION

Since the deregulation of the telephone industry in the 1970's, the costs associated with equalized program service have increased as much as 2,000 percent in some cities.

A five mile 15 kHz equalized loop used to cost about $400 including a one time service charge; after that it was about $300 per month. Broadcasters now pay $1,200 or more per month for that same service.

The primary factors to consider in setting-up a remote program feed over telephone lines today are:

a) Cost. Traditional equalized telephone lines from the local telephone company cost considerably more than in the past.

b) Lead-time. Telephone companies typically require a one month advance notice in order to install an equalized loop. This is a luxury given the demand for immediacy in today's competitive broadcast market.

c) Qualified personnel. Many of the telephone company personnel who were most qualified to install equalized loops have retired or quit, and they have not been replaced.

d) Accessibility. With the advent of cellular telephone service, new possibilities are opened up permitting more flexibility than ever before.

Alternatives exist to equalized telephone loops. Remote Pickup Units (RPU's) offer an excellent alternative for some applications; however, they are quite limited by distance and frequency coordination factors. Satellite audio feeds can be setup at great expense and with sufficient time.

For most broadcasters, standard dial-up telephone lines offer the easiest and least expensive alternative to equalized loops. Standard telephone lines are readily available or can be installed quickly in most cases.

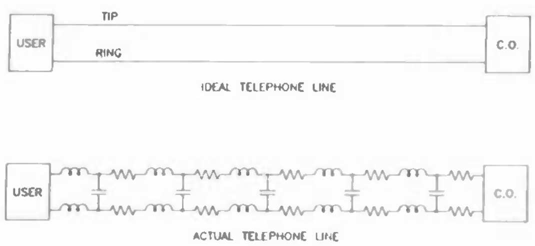
WHY TELEPHONE LINES HAVE LIMITED BANDWIDTH

The biggest objection voiced by broadcasters using regular dial lines for doing remotes is the low frequency response cutoff around 300 Hertz. This characteristic of standard dial-up telephone lines causes the sound of male announcers' voices to lack warmth, which tends to make remote broadcasts over such lines sound "tinny" and may cause listeners to "tune-out".

Untreated telephone lines also roll off high frequencies at about 3,300 Hz, further limiting the audio fidelity of the remote broadcast.

The direct dial telephone network is bandwidth limited because of two deficiencies in the system. Small transformers used at every telephone system central office limit low-frequency response while the twisted wire pair introduces distributed resistance and reactances forming a complex impedance which varies with telephone circuit routing. These two factors work together to limit the frequency response of the telephone line to a band that extends from about 300 to 3,300 Hz.

Figure 1 shows what the ideal telephone line looks like (simply two conductors between the user and the central office) compared to the actual telephone line.

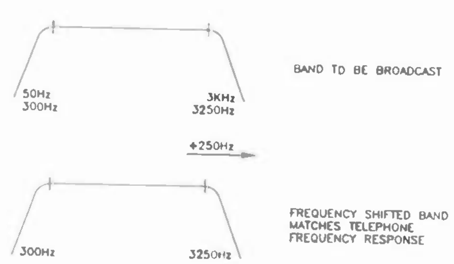


The Ideal Telephone Line versus the
Actual Telephone Line
Figure 1

Noise, inconsistent levels, and unpredictability are also major deterrents when one considers using dial-up telephone lines for remote broadcasts.

SINGLE-LINE FREQUENCY EXTENSION

In 1972 Comrex Corporation introduced a system that shifted all audio frequencies up 250 Hz on the transmit side of a telephone line. A decoder on the receive end of the same telephone line shifted the audio information down 250 Hz. This enables the passage of the important frequencies below 300 Hz which communicate the richness and quality in an announcer's voice. See Figure 2.

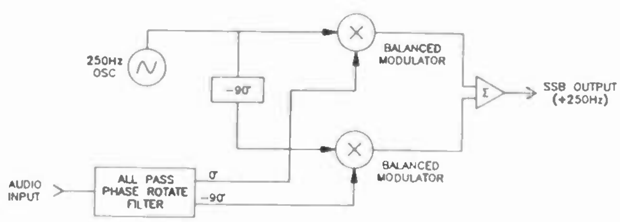


Single-Line Frequency Shifting Process
Figure 2

The original application of frequency shifting dates back more than sixty years.

The process of shifting audio frequencies by single sideband generation was developed by Western Electric in 1926 as a means of multiplexing several telephone conversations on a single channel. They generated single sideband signals and generated multiple carriers in order to create multiple communication capabilities on each channel.

In 1985 Gentner Electronics Corporation introduced a single-line frequency extension system that utilized a multiplying analog to digital converter to achieve the necessary shifting. Two quadrature carriers were generated, one being a sine wave and the other a cosine wave. The quadrature carriers are contained in read-only memory (ROM) chips. The frequency accuracy and phase relationships using this method are very precise and stable. Full duplex operation was also incorporated into this device. Figure 3 presents a block diagram of this method of single-line frequency extension.



Single-Line Frequency Extension
Figure 3

Single-line frequency shifting works due to the mathematical progression of musical octaves. Each octave represents a doubling of frequency. Therefore, by means of this frequency shifting, 2.5 octaves of low frequency information are retained at the expense of 0.143 octave of high frequency information. The small loss in high frequency information is more than made up for in the recovery of the important low frequencies.

There is a problem with single-line frequency extension. This system violates the "500,000 Rule" that was originated by Harvey Fletcher, an important researcher into the phenomenon of human hearing. The 500,000 Rule states that the product of the high frequency limit times the low frequency limit should be near the number 500,000. Maintaining this proportion between upper and lower frequency limits results in a less fatiguing, more pleasant sound.

Due to the violation of the 500,000 Rule, some listeners describe the audio quality of single-line extenders as "boomy". Additional equalization or high frequency synthesis techniques can be used in conjunction with single-line frequency extenders to help ameliorate the spectral balance.

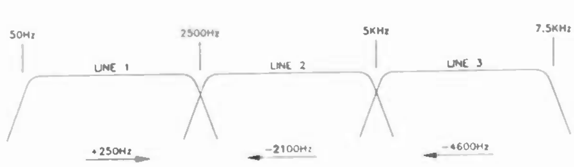
MULTI-LINE FREQUENCY EXTENSION

Frequency extenders that provide even greater audio bandwidth than the single-line extender came into demand as the costs associated with equalized telephone loops increased. This was necessary for use on remote broadcasts, or for any application where full fidelity audio transmission is required. The equalized loops offered by the telephone company had a bandwidth of 5 kHz, 8 kHz, or 15 kHz, and broadcasters needed an alternative.

Several systems were developed that offered response from 50 Hz to 5 kHz using two regular dial-up telephone lines. In general, low frequencies were upshifted and high frequencies were downshifted so that they could be passed through the telephone system.

The obvious next step was a three-line frequency extension system that would divide the audio spectrum into three bands. An important benefit to increased high-frequency response was that the band pass would conform more closely to Fletcher's 500,000 Rule.

Gentner introduced such a system in 1988. Each audio band is transmitted over a separate standard dial-up telephone line. Each band is centered by means of frequency shifting within the pass band, or frequency response "window", of the individual dial-up telephone lines. Refer to Figure 4.



Three-Line Frequency Shifting Process
Figure 4

FOURIER SERIES MATHEMATICS APPLIED TO FREQUENCY EXTENSION

During the research and development process, it was determined that simple band splitting and frequency shifting would not result in a feasible system. The three audio paths must be equalized, their levels must be matched, and some time delay compensation factor must be employed.

The complexity required in order to perform all of these functions in a three-line frequency extension system indiciated that the manipulation of the audio signals involved must be done in the frequency domain, as opposed to the time domain.

What is meant by the "frequency domain"? A spectrum analyzer is used to view signals in the frequency domain. The amplitude is displayed vertically and frequency is displayed horizontally. This is in contrast to the "time domain" display of an oscilloscope, where the amplitude is displayed vertically versus time horizontally.

Digital signal processing (DSP) technology provides the technical means for most efficiently converting time domain digital audio signals into the frequency domain. The basic algorithm for making this conversion was developed in the 18th century by a French mathematician named (Jean-Baptiste-) Joseph Fourier. His work focused on the thermal wave transference of heat through metal, and suggested that all such signals may be represented mathematically by combinations of sine and cosine functions. The infinite equation of harmonically related sine and cosines he authored became known as the "Fourier Series". Applying integral calculus to the Fourier Series and truncating it to evaluate a portion of the time domain yields the "Discrete Fourier Transform".

In 1968 the cumbersome Discrete Fourier Transform was truncated and streamlined by two engineers at IBM. They recognized certain redundancies after it was factored into various sets of matricies. The Fast Fourier Transform (FFT) enabled quick computation of the Discrete Fourier Transform without measurable sacrifice in accuracy.

The FFT is a very efficient algorithm to translate signals from the time domain into the frequency domain. Fourier mathematics had not been previously applied in the processing of real time digital audio signals, and provided the best method available for accomplishing the complex processing tasks required in a three-line frequency extender.

THE EFT-3000 THREE-LINE FREQUENCY EXTENDER

Figure 5 presents the block diagram of the EFT-3000 three-line frequency extender.

Starting at the upper left, audio enters via the two transformer coupled inputs, which are switchable from microphone to line level. After a selectable RMS limiting stage, and metering driver, the audio enters the analog to digital converter.

Fourier transforms and all other processing takes place in the DSP chip.

The analog link to the outside telephone world is made through three hybrid transformers in connection with relays to seize the lines and ring detect circuits.

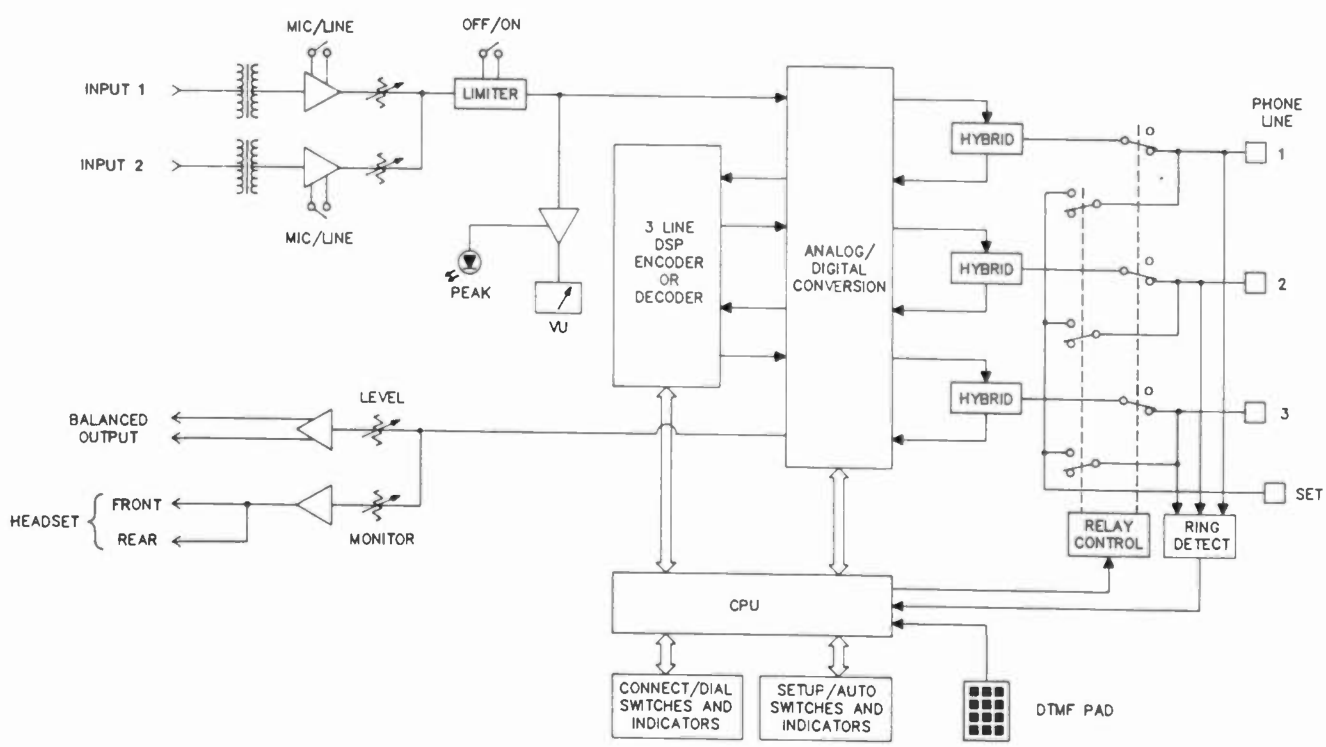
The central processing unit (CPU) presides over all of the functions in response to external commands from the DTMF pad, front panel switches, and telephone signals.

Since the audio processing instructions for the system exist exclusively in software, a unit can be programmed to be either a transmitter or receiver.

DIGITAL SIGNAL PROCESSING IN THE EFT-3000

Most digital audio systems use numbers to represent discrete steps of amplitude information. This is audio information represented digitally in the time domain. The EFT-3000, however, uses numbers to represent audio signals in the frequency domain.

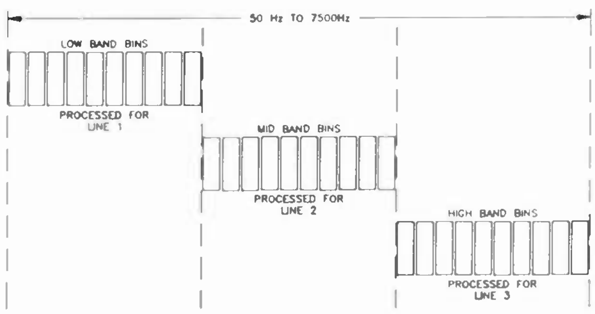
In the EFT-3000, the audio bandwidth of the system is broken up into 512 parts, or "bins", and the audio information contained in each of these bins is represented in terms of a combination of sine waves and cosine waves. At the EFT-3000's sampling rate of 20 kHz, each bin becomes 19.53 Hz wide. The first bin represents frequencies from DC (or zero) to 19.53 Hz. The following bin represents frequencies from 19.53 to 39.06 Hz, and so on.



Gentner EFT-3000 Block Diagram
Figure 5

Representing these signals in terms of sines and cosines involves a complex, or orthogonal, mathematical system in which the "X" axis in a Cartesian coordinate plane contains information relating to the real, or cosinusoidal, component of a signal, and the "Y" axis contains information relating to the imaginary, or sinusoidal, component of the same signal. The information gathered through a mathematical combination of the "X" and "Y" components represents the audio information.

Figure 6 shows the division of frequencies in the frequency domain.



Division of Frequency Bands
in the Frequency Domain
Figure 6

Once the information is divided and processed, three separate Inverse Fourier Frequency Transformations, or IFFT's, are performed in order to convert the digital information back into the analog domain for transmission on three dial-up telephone lines.

Each telephone line carries 2500 Hz of audio bandwidth. The full pass band of each line is not utilized because the worst envelope delay distortion (group delay) occurs at or near the band edges in the telephone system.

One of the telephone lines might be routed via satellite, and another one routed via a microwave link. The audio quality will suffer when any of the bands arrive earlier or later than the other bands.

At the receive EFT-3000 unit, each of the bands of analog audio is taken from the three telephone lines and digitized.

The delay differences between the three telephone lines are analyzed while the information exists in the time domain. These samples are stored in memory in adaptable buffers, the duration of which is automatically adjusted to get the timing of all three bands to coincide. The EFT-3000 compensates for as much as 750 milliseconds of differential delay.

If the differential line delay exceeds the compensation capability of the EFT-3000 system, or if the amplitude variation between any two lines is greater than 6 dB, the system reverts to a default time delay and amplitude correction configuration. In this instance the Setup LED on both units flashes to indicate the system was not able to fully achieve the setup performance goals.

After delay and amplitude correction, the three bands of information are transferred into the frequency domain via an FFT. The three segments of numbers can then be combined to represent the complete audio pass band of the system.

A digital equalization algorithm compensates for up to 6 dB of frequency response anomalies in the telephone lines. An Inverse Fast Fourier Transform (IFFT) converts the digital signals back into the time domain. This process is the inverse of that presented in Figure 6. A digital to analog converter combined with a smoothing filter produces the analog audio output of the receive EFT-3000 unit.

The EFT-3000 system also provides a single-line frequency extended half-duplex cue channel. This requires additional FFT and IFFT functions on both ends, as well as separate digital to analog and analog to digital conversions. The frequency response of this return audio path is 50 Hz to 2500 Hz and appears on one telephone line only. Half-duplex operation is necessary due to the delay inherent in digital signal processing. The cue channel behaves like a speakerphone; the channel switches from receive to send when a predetermined threshold is exceeded.

DELAY ASSOCIATED WITH DIGITAL SIGNAL PROCESSING

Digital signal processing techniques and hardware provide the computing power required to implement the EFT-3000's elaborate filtering schemes, line delay differential compensation, and equalization algorithms. Although the DSP's in the EFT-3000 operate at a 20 Mhz clock frequency, an audible delay is unavoidable given the amount of processing required.

A processing delay of 20 milliseconds is typical in the EFT-3000. Most of this time is required to perform FFT and IFFT algorithms.

ISDN AND ITS USES IN THE BROADCAST ENVIRONMENT

The telecommunications network was originally designed for narrow-band voice communication. There are ever increasing needs for networking capabilities that will handle digital data communications of every sort.

ISDN stands for the Integrated Services Digital Network. An all digital network will be able to combine all forms of digitized information over a common medium such as fiber optic cable.

Presently 214 telephone central offices can offer operational ISDN services on the AT&T 5ESS Switch, with 772 central offices upgraded with ISDN software. The latest figures from AT&T indicate that there are 321,000 ISDN lines presently installed and/or in use in the United States.

Cities with ISDN service include Washington, D.C., Baltimore, New York, Fresno, Atlanta, Austin, Houston, Dallas, Denver, Minneapolis-St.Paul, Omaha, and Phoenix. Plans are underway to eventually offer worldwide ISDN service.

ISDN technology may prove to be very useful to broadcasters. It may eliminate the need for frequency extenders in some cases, and provide capabilities to broadcasters which are completely unavailable today. In particular, ISDN may provide full bandwidth audio channels, combined with high speed data transmission, to broadcasters for program transmission and remote equipment control.

While ISDN is being implemented in some areas, many telephone companies continue to install and maintain traditional switching systems. These traditional telephone central offices will remain in service for many years in order to be cost-effective. For the next five to 10 years, most broadcasters will continue to be faced with the necessity of using frequency extenders, RPU's, and other types of equipment to transmit high quality audio material from one location to another.

CONCLUSION

The relative unavailability and/or high cost of equalized telephone loops from the local telephone company has produced a need among broadcasters for alternatives. Frequency extension equipment provides broadcasters with a flexible, low-cost method for doing remote broadcasts, and will continue to be a useful component in most broadcast facilities for many years. The installation of ISDN systems in many parts of the country will provide a superior alternative to frequency extension techniques, although the costs and flexibility of such ISDN installations is undetermined.

WORLDCOM—A PERSONAL SATELLITE COMMUNICATIONS SYSTEM FOR VOICE AND DATA

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Abstract: CBS News is a world wide news gathering organization that requires reliable communications between its field personnel and CBS News headquarters in order to insure the accurate and timely reporting of major events. Often this communications is not available when most required, such as times of natural disasters, political upheavals or from remote areas where communication is nonexistent.

CBS Engineering & Development has been working with SkyWave Electronics, Ltd. to develop a lightweight, portable personal satellite communications terminal for global voice and data communications. This paper will describe the system, terminal, and initial field trials.

Background: CBS News has bureaus in many cities worldwide. In these bureaus, crews and correspondents are deployed to report on news events in a specific area.

Some of the events covered by CBS News are planned well in advance, such as Presidential trips and elections here and abroad. However, the majority of coverage is breaking news, the timeliness of which is critical.

CBS Engineering has worked with CBS News and many vendors in the past to develop techniques such as Electronic News Gathering (ENG) to assist in this effort. The development of Rapid Deployment Earth Terminals (RADET or Flyaway Terminals) extended Satellite News Gathering (SNG) worldwide. A news crew can now cover an event, anywhere a geosynchronous Ku-band satellite is visible, live within minutes of arriving on the scene. There was no longer a need to return video tapes to a central location for playback. Additionally, CBS News is in full control of the satellite uplink. This system although relatively portable, still requires a crew of several people.

In an effort to extend CBS News' ability to gather information from all corners of the globe and improve communications with our personnel in the field, Worldcom was developed.

Worldcom: Worldcom was conceived of as a system providing voice and data communications, via satellite, from a lightweight briefcase size terminal. Existing portable terminals were heavy and used large power sources. CBS Engineering & Development with CBS News outlined the following requirements for a Worldcom system:

1. Full-duplex voice operation
2. Dial-up operation to and from the Public Switched Telephone Network (PSTN)
3. Sufficient link margin for good voice quality
4. Weight with power supply not to exceed 32 lbs.
5. Data transmission capability
6. Full-time commercial service.
7. 1 Hour operation on internal batteries

In 1986 CBS Engineering & Development produced a specification outlining the Worldcom system. In 1988 CBS contracted with Skywave Electronic Ltd. to develop a proof-of-concept prototype for such a system. Teleglobe Canada provided the space segment as a signatory to INMARSAT (See Figure 1) and The Communications Research Center (CRC) of Canada provided use of its 9 Meter earth station. In order to achieve the 32 lb weight requirement, the prototype was restricted to simplex operation. At the time, commercially available diplexers would have put the weight of the briefcase over the 32 lb weight limit. As the development of these lighter weight diplexers progressed, Skywave was then contracted to develop the final, duplex, product. The final hub station will be provided by Teleglobe through their 18 Meter earth station at Weir, Quebec. Teleglobe will also continue to supply the space segment.

Worldcom System: The Worldcom system consists of three components (See Figure 2). These components are:

1. The LBT-1 L-band Briefcase Terminal
2. The Space Segment on the INMARSAT Satellites MARECS B2 (Atlantic Ocean Region), INTELSAT MCS D (Future Pacific Ocean Region), INTELSAT V MCS A (Future Indian Ocean Region).
3. The SCU-1R ACSSB/DMSK Hub Channel Unit

The links both to and from the briefcase terminal are at L-band (1.5 and 1.6GHz). The satellite converts these to C-band (4 and 6GHz) for the links to and from the hub earth station.

Briefcase Terminal: The Worldcom terminal package consists of the briefcase terminal which contains modulation and demodulation electronics, Solid State Power Amplifier (SSPA), Low Noise Amplifier (LNA), and diplexer, two flat patch array microstrip antennas, NiCad batteries, external power supply and a carrying case (See Figure 3). The total weight excluding the external power supply is 32 lbs. Details on these components are as follows;

Antennas:	Gain 17dBi, G/T -9dB/K, Circular Polarization
Modulation:	Voice; Amplitude Companded Single Sideband (ACSSB) Full-Duplex, Data; Differential Minimum Shift Keying (DMSK) @ 2400 bps synchronous
RF Emission:	Effective Isotropic Radiated Power (EIRP) 27dBW, RF Bandwidth 5kHz, Frequency Accuracy +2.5kHz
Power:	12 volts @ 7 amps or 115/230 VAC ± 10% @ 50/60Hz
Batteries:	Internal NiCad providing one hour operation, external charger from 115/230 VAC ± 10% @ 50/60Hz
Dimensions:	Briefcase 18"X13"X4.5", Case 19"X14"X8.5"
Environment:	Temperature Range from 0°C to +40°C
C/N_0 (db-Hz)	47.0 with 18 Meter Hub Earth Station

Space Segment: The space segment for one briefcase terminal and one control channel occupies only one 5kHz segment of a 25kHz channel (See Figure 4) on INMARSAT's MARECS B2 Atlantic Operating Region (AOR) satellite. The full three channels permit multiple units to call, two at a time, through the hub station. In addition a personal computer is used to control the channel assignment and provide billing information.

THE INMARSAT SATELLITE SYSTEM

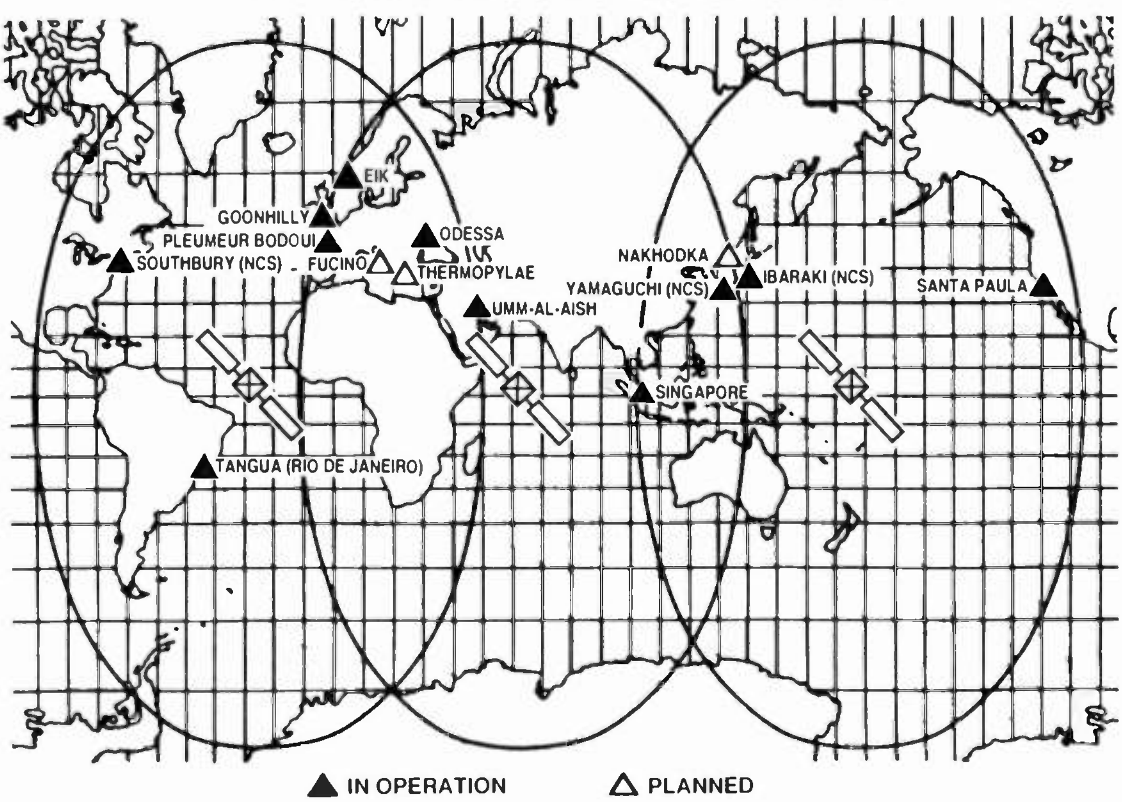

FIGURE 1

Coverage of the three INMARSAT Satellite

Hub Channel Unit: The Hub Channel Unit is a rack mounted chassis that provides the interface between the RF environment and the Public Switched Telephone Network (PSTN) (See Figure 5). Two units can be controlled by a personnel computer running a proprietary application that controls assignment of channels to multiple briefcase terminals. This system permits two simultaneous telephone calls on a single 25kHz satellite channel. Some details on the Hub Channel Unit are as follows;

ACSSB Modulator:	Companding law 3:1
PSTN Interface:	E & M Type 1, 600 ohms, -16dBm to 0dBm
IF Interface:	52 to 88MHz, -20dBm, 50 ohms
Data:	Rates 300, 600, 1200, 2400 bps, RS-232
ACSSB Demodulator:	AGC range -20dBm to -40dBm
DMSK Demodulator:	Detection is differential with single non-redundant error correction
C/N_0 (db-Hz)	47.0 with 18 Meter Hub Earth Station

System Operation: The briefcase unit is completely self contained and is operated through the telephone handset. For ease of operation this handset resembles the ones used on cellular telephones.

The briefcase must first be unpacked and assembled. The briefcase itself is opened and placed on a surface in a location with a clear line-of-site to the appropriate satellite. The unit can be operated from the internal batteries or from the external AC power supply if AC power is available. The two antenna panels are attached to the briefcase lid with Velcro fasteners and then interconnected to the diplexer with miniature coaxial cable. The antenna system is connected to the diplexer with a push-on connector.

The unit can now be powered on. Next the briefcase is oriented in the general direction of the satellite. A code is entered on the telephone handset that causes the unit to detect a beacon from the satellite. The resulting tone will change frequency as the briefcase is moved to acquire the satellite. Azimuth is changed by rotating the entire briefcase on its base. Elevation is changed by raising or lowering the hinged briefcase lid. The lid is then locked in place by a hand tightened knurled nut. The tone will reach the highest frequency when properly oriented and peaked on the satellite.

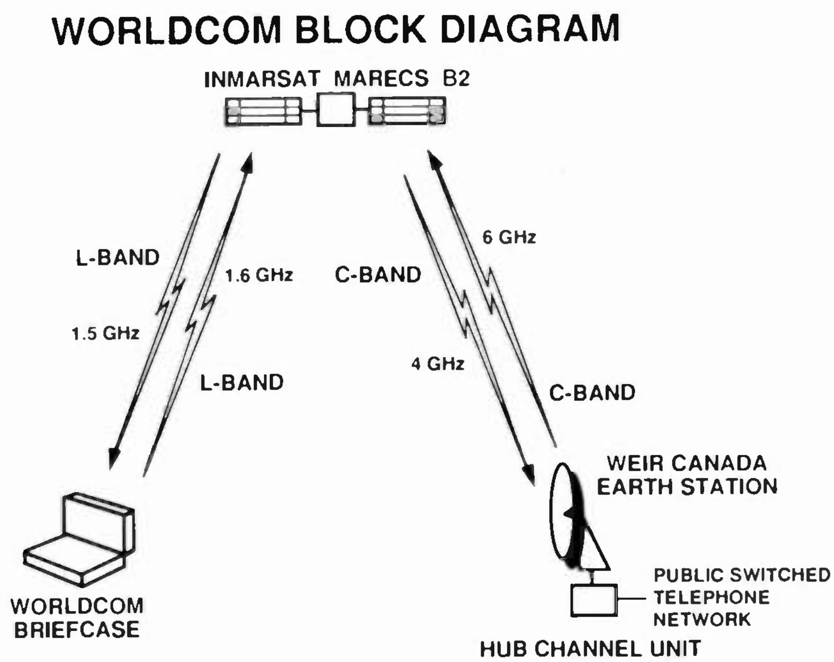

FIGURE 2

Worldcom system block diagram

A code is then entered to set the system up for an Amplitude Companded Single Sideband (ACSSB) voice telephone call. Next an access code is entered for the desired Hub Unit Group which identifies the user to the Hub station. Access is granted and the desired area code and phone number is dialed as with any other phone. At the conclusion of the call each party simply hangs up.

The procedure is similar for a data call. The briefcase contains a "Hayes compatible" modem that can connect to a terminal or personal computer running a communications program. That device is then used to enter all codes and telephone numbers.

Once the LBT-1 unit is peaked on the satellite it is ready to receive calls. The telephone number of the Hub Channel Unit is dialed from within the PSTN. A code for the specific briefcase is entered. If the unit is setup and operating, a ring will be generated at the briefcase. The unit is answered and the call proceeds as a normal telephone. Data calls can also be placed to the briefcase.

Initial Field Trials: The initial field trials currently under way at CBS focus on two areas, technical performance and operational issues.

Technical performance tests will verify conformance with all technical specifications. Both the ACSSB voice mode and the DMSK data mode will be tested and stressed to determine the performance margins of the system.

The unit will also be tested operationally. A wide variety of personnel of varied technical competence will be asked to setup and operate the unit, place and receive calls, and repack the system for transport to another location.

At the conclusion of these tests the briefcase will be placed into actual service by CBS News.

Future Plans: Skywave will continue to develop and manufacture this system. Teleglobe has submitted the system to INMARSAT for consideration as a non-standard land based satellite communications service. The Communications Research Center (CRC) of Canada is also conducting tests of a similar system in support of Telesat Mobile, Inc's planned MSAT service. This service will provide land, marine and air mobile service for Canada and will permit travel throughout otherwise remote areas without a loss of communications. In addition, Teleglobe has plans to expand the coverage of these units to the INMARSAT Pacific Operating Region (POR) in late 1990 and the Indian Operating Region (IOR) in 1991. CBS will be participating in this expansion in order to insure complete world coverage of the Worldcom system for CBS News.

CBS Engineering & Development will also be active in the design and deployment of even smaller, lighter weight devices. As newer higher power satellites are placed in orbit, the size of the Worldcom device will be no bigger than todays state-of-the-art cellular telephones.

Another device recently developed by the CBS Engineering & Development Department allows a single correspondent to provide still pictures, via telephone, of events where the use of a flyaway video terminal (eg. RADET) is precluded by technical and/or political reasons. Quikpix was first deployed in Beijing, China this past year as the only method of transmitting pictures out of the country after all earth stations were cut off by the government. Small electronic still video cameras coupled with Quikpix makes for extremely rapid coverage. Quikpix coupled with Worldcom will give CBS News the ability to transmit voice and still video from any location on the globe.

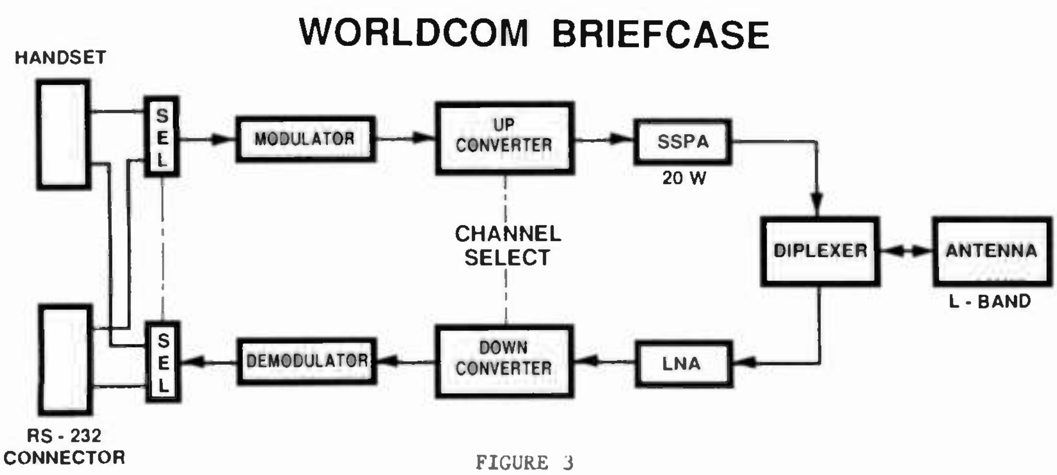

FIGURE 3

Worldcom briefcase block diagram

Conclusions: CBS Engineering & Development has advanced the state-of-the-art in news gathering by supplying CBS News with another tool in their arsenal for making the Global Village even smaller. Worldcom has proven to be a global communication system that provides lightweight, reliable, easy to use personal voice and data communications. The Worldcom briefcase represents the first step of the satellite cellular telephone revolution of the 1990's.

Acknowledgments: This paper would not have been possible without the forward thinking of the entire CBS Engineering & Development department under the leadership of Joseph Flaherty. In addition I would like to thank Peter Rossiter and Les Tibbo of Skywave Electronics Ltd. and Michel Gelinas of Teleglobe Canada for their assistance.

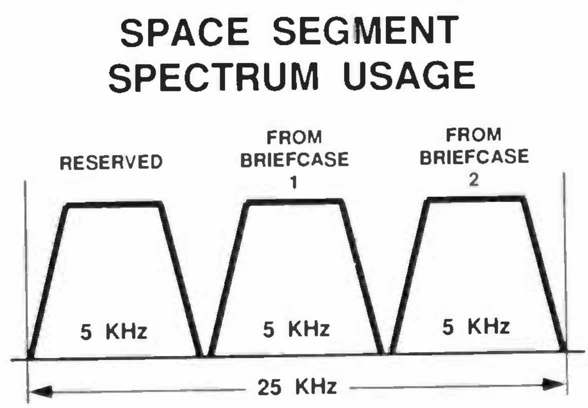

FIGURE 4

Worldcom space segment spectrum usage. The reserved channel is currently used by the Canadian Air Ambulance Service.

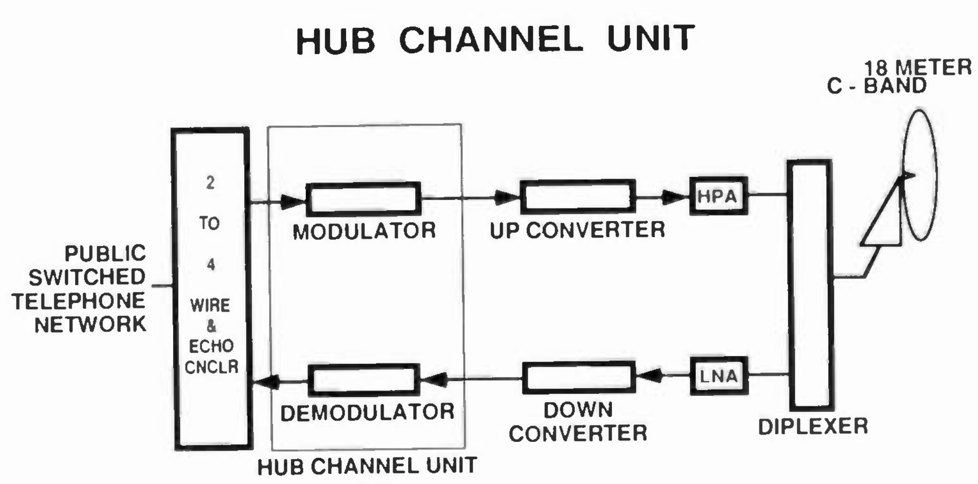

FIGURE 5

Hub Channel Unit interconnections.

RATE-REDUCED DIGITAL AUDIO IN THE BROADCAST ENVIRONMENT

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INTRODUCTION

A bit - reduced audio transmission codec, Fraunhofer OCF, has been introduced on the Boston to Saipan path of the Christian Science Monitor's radio network. Substantial cost savings are gained, and audio fidelity is improved. Bit rate reductions of between 9.1 to 1 and 16 to 1 have been used.

Applications include high fidelity audio transmission over the dialup network and greatly reduced bandwidth on microwave and satellite audio services.

THEORY OF AUDIO BIT RATE REDUCTION.

Historical Antecedents

The inception of FM broadcasting in the 1930's was a successful attempt to improve fidelity, compared to AM, at the expense of greatly increased bandwidth of the transmitted signal. In the same time period there were attempts made to greatly lower the bandwidth required to transmit audio, (i. e., single sideband modulation, the Vocoder of Shannon), but these were low fidelity systems that could not be used in broadcasting. It was not until the 1970's that advances in computer science, signal theory and psychoacoustics made a truly high fidelity reduced bandwidth transmission system seem feasible. [1], [2], [3].

ADPCM

This relatively simple and inexpensive technique, adaptive differential pulse code modulation, stems from work done at AT&T and has been adopted by the CCIT as the G.722 standard for transmissions at 56 kb/s. The frequency response of G.722 is limited to 7.5 kHz and it offers neither the low distortion or high coding efficiency of more advanced methods.

High Coding Efficiency Methods

Depending on the sampling rate, a channel capacity of 64 kbits/s translates to a reduction factor of about 8 to 12 (10.25 for 44.1 kHz compact disc input). To obtain compression ratios of this magnitude the codec relies not only on the statistical properties of the audio signal but also on the psychoacoustic properties of the human ear[4].

The first category of subband coders uses filters adapted to the critical bands known from psychoacoustics [1]. It is usually implemented as a filter tree (e.g. QMF) where each node of the tree splits the remaining frequency band in further subbands. Due to this strategy the bandwidth of each band can be easily adapted to the properties of the human ear (critical bands). Unfortunately this results in rather complex filters with a large delay. Furthermore the frequency resolution is hardly adequate for calculation of a concise masking level[5]. Tonality cannot be calculated with this design.

A second approach for subband coders is the use of polyphase filter banks with the advantage of short delays, low to moderate complexity and an inherently good resolution in the time domain. On a more negative side the frequency resolution is rather low and, what is worse, linearly spaced. Therefore the frequency bands do not correspond well to the critical bands. This is especially true for the low frequency range. Consequently perceptual criteria cannot be easily used with this coder type. Modifications for better frequency resolution dilute the original advantages resulting in higher complexity and lower time resolution.

Transform coders have a low to moderate complexity, an excellent frequency resolution (important for psychoacoustics) and the potential for high redundancy reduction. Two classes of algorithms can be distinguished. Those that use TDAC (Time Domain Aliasing Cancellation) [8] and those that do not. The second category (e.g.[6],[7]) suffers from blocking effects and low time resolution (pre echoes). The blocking effects can be reduced by using overlapping windows but only at the cost of an increased data rate. Through the use of the TDAC algorithm with fully overlapping windows this problem can be overcome. This practically eliminates the blocking effects while retaining the high frequency resolution. The comparatively poor resolution in the time domain remains with its associated pre echo problems. For a fully transparent audio codec additional methods are needed to address this problem.

PRACTICE

Organizations known to have implemented high quality, high efficiency reduced rate digital audio codecs in hardware are: The Fraunhofer Institute (OCF; transform coding),[7];Dolby Laboratories (transform coding),[9]; the Musicam group, (sub - band coding),[10]; SSL - Audio Processing Technology, Ltd. (apt - X; sub-band ADPCM coding)[11],

IMPLEMENTATION IN THE PROFESSIONAL ENVIRONMENT.

OCF In the Experimental Stage

Fraunhofer OCF was first implemented in hardware and demonstrated over a simulated (wire) transmission link in 1988[12].

Initial demonstrations of OCF, developed in the laboratories of the Fraunhofer Institute by a team under the leadership of Dr. H. Gerhaeuser, were made, at this time, before several professional gatherings in this country and abroad. Input for the initial demonstrations was the digital audio bit stream from a compact disc player.

With the demonstration of OCF, it became evident that this new technology could have considerable impact in the broadcast industry.

Transmission Requirements of the Christian Science Monitor

The Christian Science Monitor operates an international network of shortwave stations, as well as a domestic radio network for the transmission of news public affairs, and religious programming. These programs are transmitted by satellite. Single carrier per channel, analog (FM) techniques are used to distribute the domestic programming. Programs to the short wave transmitters in Maine, South Carolina and Saipan are transmitted by Intraplex digitally multiplexed 8 - bit digitally companded PCM codecs over QPSK modulators (Fig. 1)

A bit rate of 128 kb/s is used by the Intraplex PCM equipment at the Monitor to transmit a single 5 kHz audio bandwidth program for use on short wave. Two such programs are sent from the Monitor's Boston studios to Saipan, Northern Marianas. An immediate application of OCF was suggested by the possibility of halving the bandwidth used for program transmission on this trans - Pacific link. (Fig. 2)

It was hoped that OCF, operating at 64 kb/s at an audio bandwidth of up to 20 kHz, would provide savings in transmission costs with no degradation in quality. While the potential for superior quality was there, the decision to pursue OCF was driven by economics; the potential savings of several thousand dollars a month to be gained by bandwidth reduction.

Initial Attempts to Use Laboratory Apparatus Over Satellite

Could this potential be realized? Arrangements were made to bring the OCF laboratory demonstration apparatus to Boston for tests in a functioning broadcast environment. Intraplex, providers of the Monitor's T1 multiplex and audio PCM transmission equipment, provided facilities at their Littleton, Massachusetts site for bench testing, after which satellite tests were undertaken at the Monitor's Boston uplink site.

The results of satellite testing showed that OCF would have to become considerably more robust, in both its hardware and software implementations, to be usable in the professional environment. There were, however, encouraging indications. In particular, the OCF encoded bit stream was shown to be relatively immune to errors when errors were introduced from a transmission error - generating test set. Although, at this early stage, no DSP code had been written to correct transmission errors, it was discovered that, under most circumstances, no audible artifacts could be detected at bit error rates of 1 in 10 exp 6 and quality was acceptable at an error rate of one in 10 exp 5

Results of Initial Tests

In its initial form, the input to the OCF coder was AES/EBU format, 16 bit PCM, 44.1 ksamples/s stereo digital audio. The output of the coder was unbalanced consumer level audio. The digital interface between coder and decoder was non - standard. A considerable amount of interface equipment was required to get standard signals in and out of OCF.

Some other problems with OCF that showed up in tests:

1. Loss of synchronization by the decoder. The decoder could get into illegal modes from which it could not recover.

2. Audio frequency response too high. 20 kHz audio bandwidth is too much for the broadcast environment, where bandwidth is restricted to 15 kHz by pre - transmitter filtering. The audio components between 15 kHz and 20 kHz, besides taking processor time to encode, were likely to cause intermodulation problems in broadcast equipment. Resampling by coders with Nyquist limits too low for 20 kHz audio was likely to cause aliasing. Quality was therefore likely to be improved by limiting audio bandwidth to 15 kHz. At the same time, by correct choice of anti-aliasing and reconstructive filters, the audio sampling frequency could be lowered to 32 kHz, leaving more processing time for the OCF algorithms.

3. Differences in artifacts between speech and music. OCF had been initially implemented to encode music. Artifacts, while always present, should be uniformly (un)objectionable regardless of program material.

4. Use of the AES/EBU signal as input.

5. Algorithms not optimized for transmission at 56 kb/s. Both the fifteen zeros constraint associated with T1 and the use of bit - robbing in N X 64kb/s multiplexing (Fig. 3) may prevent the use of 64 kb/s for certain channels.

Specifications; CSM Installation

Out of the March, 1988 tests came specifications for OCF codecs to be permanently installed in broadcast installations. These specifications can be summarized as follows:

1. Transmission rate of 24 to 128 kb/s in increments of 8 kb/s. (switch selectable)

2. Audio sampling at 48, 44.1 or 32 kHz, jumper selectable; 10th order 15 kHz elliptical low pass encoding and reconstructive active filters.

3. Clipping at audio input of +18 dBm, relative to OVU = +4 dBm.

4. Audio bandwidth 15 kHz (music PROMS); 12.5 kHz (speech PROMS)

5. Bit error rate (threshold of audible errors) > 10 exp -6.

6. RS - 449 interface.

7. 2- channel enclosed Eurocard construction.

8. DSP (Motorola DSP56001) and high speed SRAM - based technology.

It is evident from the above that the problem of differing artifacts with musical versus speech programs has not been completely solved. However, the quality of OCF is good, both for speech and music, compared to conventional transmission methods.

OCF equipment built to the above specifications has been in use on the Boston - Saipan path since November, 1989.

Savings in satellite costs should result in the OCF equipment paying for itself in less than a year from the date of installation and consequent bandwidth reduction.

OCF produces artifacts which are new and thus not directly comparable to what has gone before. Particularly instructive are A/B comparisons between 32 kb/s OCF and 128 kb/s 5 kHz PCM as implemented by Intraplex, and between 64 kb/s OCF and analog companded SCPC, as used, for example, by National Public Radio.

Results - OCF at the Monitor.

OCF exhibits about the same vulnerability to transmission errors as more conventional transmission equipment. Typical burst errors have resulted in occasional dropouts of audio lasting up to a second.

15 or 12.5 kHz audio bandwidth is unusable by short wave stations, which are bandwidth limited at the inputs of the transmitters to (typically) 5 kHz. The relatively high fidelity audio transmitted to the Saipan transmitters never makes it on to the air.

It is possible to use OCF at a bit rate of 32 kb/s and an audio bandwidth of 6 - 7 kHz. In such a configuration, the audio quality, as established by listener tests, is not inferior to 5 kHz companded PCM. The Monitor is therefore soon moving to a bit rate of 32 kb/s per channel for the Boston to Saipan link.

The bit rate reduction of a 32 kb/s signal, compared to 32 kHz 16 bit PCM is 16 to 1. Compared to 5 kHz audio bandwidth (16 khz sampling rate) 8 bit PCM the bit rate reduction (and the reduction in channel capacity required on the satellite transponder) is 4 to 1.

Future Development

For its domestic radio programming the Monitor has been reliant on the daily transmissions of high -fidelity program material from its bureaus to the Boston studios. These transmissions can be largely, if not entirely, replaced by use of OCF over the dialup network, at a substantial savings, and with quality equivalent to that currently obtained by satellite. The cost of a 15 kHz satellite transmission is about $3.00/minute for a ten minute feed. The dialup network costs about fifteen cents per minute. Switched 56 kb/s digital service is now available in most U. S. cities. Switched 64 kb/s service is available in western Europe. The Monitor is currently attempting to acquire bit - rate reduction equipment solely for the purpose of transmitting 15 kHz audio over the dialup network.

A comparison of the gain - power - bandwidth advantages of rate - reduced 15 kHz digital audio, compared to wideband analog SCPC, indicates that: 1. The digital signal requires a 15 dB carrier to noise ratio (25% cyclic redundancy coded QPSK modulated). At a bit rate of 64 kb/s, such a signal occupies a bandwidth of 60 kHz. 2. A 25 dB carrier to noise ratio is required to yield acceptable quality in a 3:1 companded SCPC FM analog system. With 75 kHz deviation, the bandwidth of such a signal is 250 kHz. The rate - reduced digital audio signal, therefore, promises performance equivalent to or better than companded analog SCPC at one tenth the power and one quarter the bandwidth of the analog system. It is therefore possible to envision input power limited audio transmission systems (i. e., satellite transponders) which provide for forty times more audio channels than can be provided by analog SCPC.

The U. S. radio broadcast industry is currently engaged in a search for spectrum space to supplement the 950 MHz band used for studio - transmitter links. The application of rate - reduced audio techniques to the 950 MHz band could easily increase the capacity of the band to accommodate many more licensees and transmission paths than at present. Similarly, the use of rate - reduced audio on satellite transponders promises to make cheaper and more numerous high quality audio channels available to the broadcast industry. Rate reduced audio has the potential to make news gathering over the telephone equivalent in fidelity to studio audio. The near term use potential of bit - reduced digital audio is summarized in Table 1.

Rate - reduced audio equipment should be configurable to data rates of 32, 56, and 64 kb/s. 32 kb/s provides voice -quality audio at the lowest currently practical data rate. 56 kb/s provides high quality audio at a rate compatible with the U. S. dialup network and with bit - robbing and T1 based satellite transmission methods.

TABLE 1

bit rate, kb/s	audio sampling rate, kHz	service	overall audio bandwidth/ fidelity
32	32	voice;satellite	5 kHz
56	32	music/speech satellite and dialup	12.5 - 14.5 kHz
64	32	music/speech low bandwidth STL, satellite and ISDN dialup	15 kHz high fidelity
128	44.1	music/speech low bandwidth STL, fractional T1 satellite and HDTV	20 kHz, CD quality

I would would like to give special thanks to Dr. Gerhaeuser and his research team of the Fraunhofer Society, Erlangen, West Germany, for their helpful support and excellent cooperation. Thanks are also due H. Schott for his assistance during the installation of the codec.

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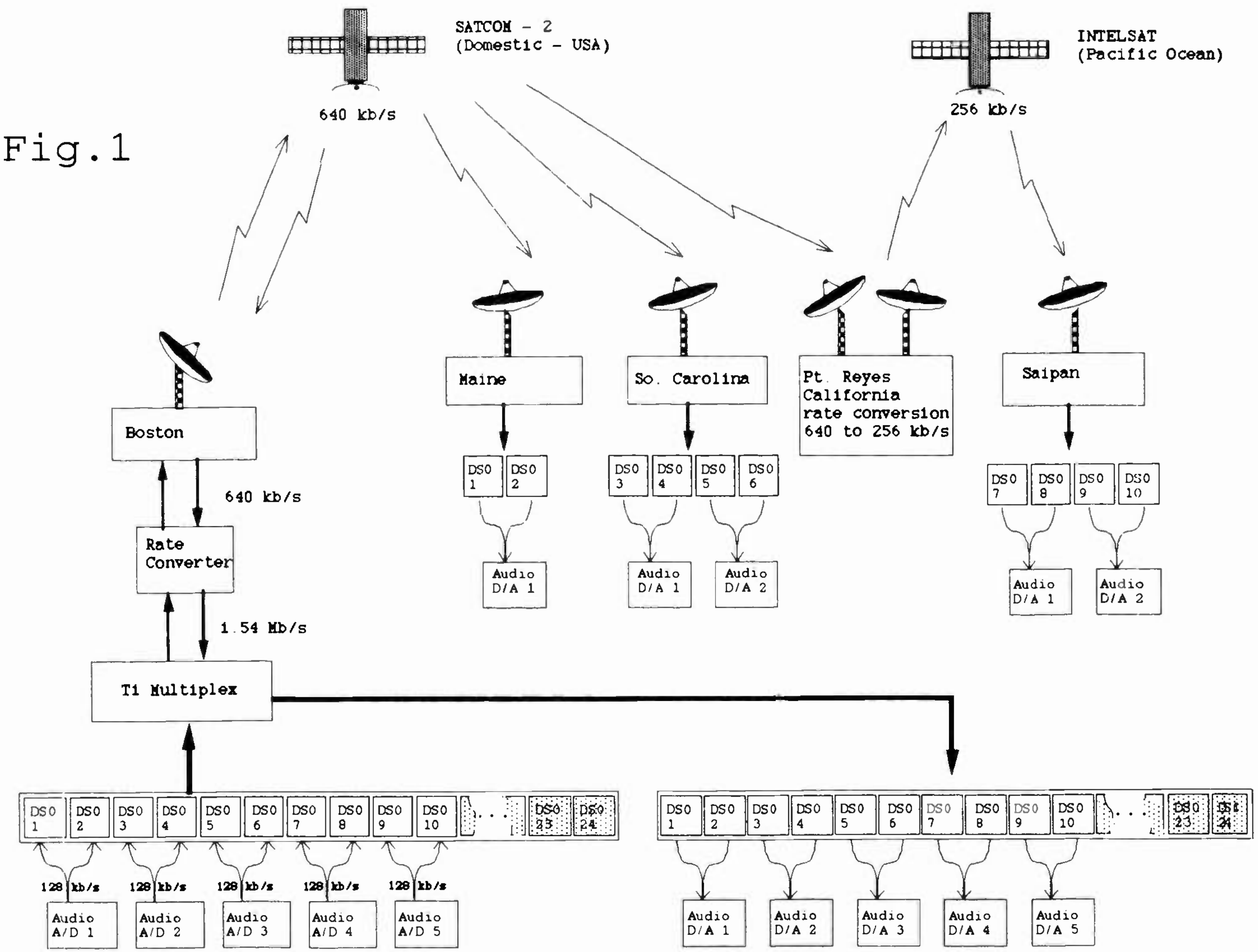
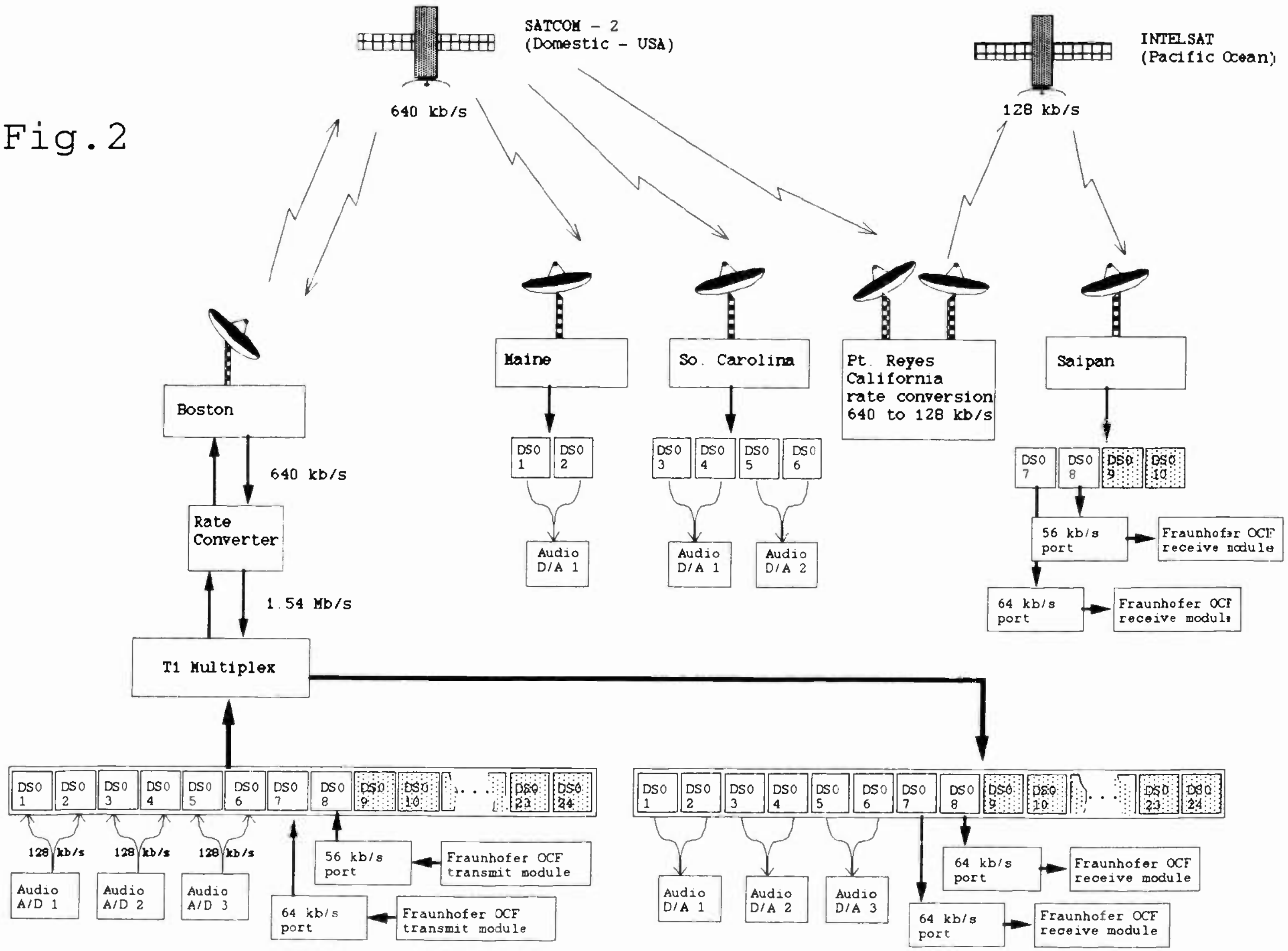
Fig.1



SATCOM - 2
(Domestic - USA)
640 kb/s
INTELSAT
(Pacific Ocean)
256 kb/s
Boston
640 kb/s
Rate Converter
1.54 Mb/s
T1 Multiplex
Maine
So. Carolina
Pt. Reyes
California
rate conversion
640 to 256 kb/s
Saipan
DSO 1
DSO 2
DSO 3
DSO 4
DSO 5
DSO 6
DSO 7
DSO 8
DSO 9
DSO 10
DSO 23
DSO 24
128 kb/s
Audio A/D 1
Audio A/D 2
Audio A/D 3
Audio A/D 4
Audio A/D 5
Audio D/A 1
Audio D/A 2
Audio D/A 3
Audio D/A 4
Audio D/A 5

Fig.2


SATCOM - 2
(Domestic - USA)
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640 kb/s
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Rate
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T1 Multiplex
Maine
So. Carolina
Pt. Reyes
California
rate conversion
640 to 128 kb/s
Saipan
DS0 1
DS0 2
DS0 3
DS0 4
DS0 5
DS0 6
DS0 7
DS0 8
DS0 9
DS0 10
DS0 23
DS0 24
Audio D/A 1
Audio D/A 2
Audio D/A 3
Audio A/D 1
Audio A/D 2
Audio A/D 3
128 kb/s
56 kb/s port
64 kb/s port
Fraunhofer OCF transmit module
Fraunhofer OCF receive module

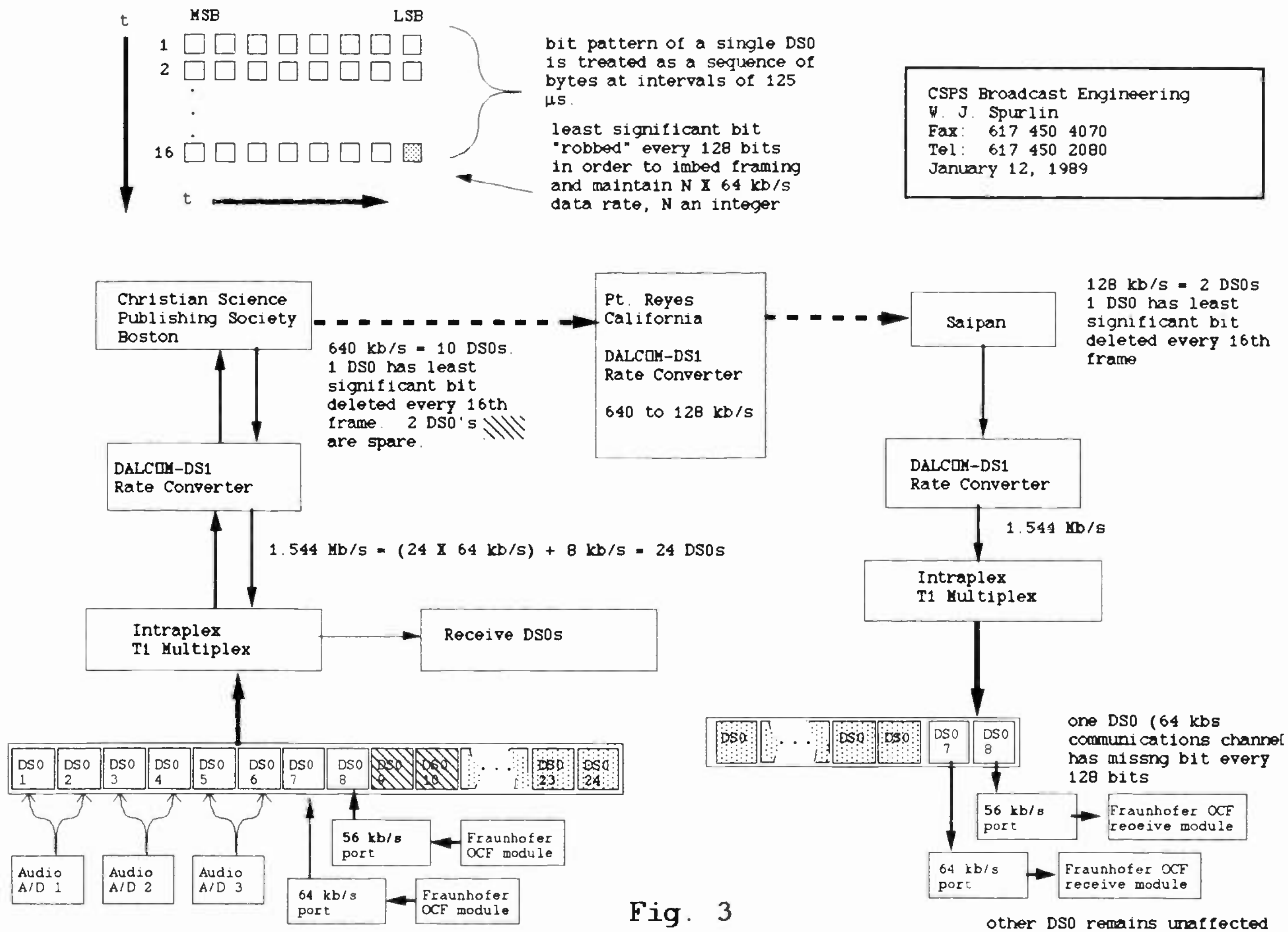
t
MSB
LSB
1
2
16
t
bit pattern of a single DS0 is treated as a sequence of bytes at intervals of 125 µs.
least significant bit "robbed" every 128 bits in order to imbed framing and maintain N X 64 kb/s data rate, N an integer
CSPS Broadcast Engineering
W. J. Spurlin
Fax: 617 450 4070
Tel: 617 450 2080
January 12, 1989
Christian Science Publishing Society Boston
640 kb/s = 10 DS0s. 1 DS0 has least significant bit deleted every 16th frame. 2 DS0's are spare.
Pt. Reyes California
DALCOM-DS1 Rate Converter
640 to 128 kb/s
Saipan
128 kb/s = 2 DS0s 1 DS0 has least significant bit deleted every 16th frame
DALCOM-DS1 Rate Converter
DALCOM-DS1 Rate Converter
1.544 Mb/s = (24 X 64 kb/s) + 8 kb/s = 24 DS0s
1.544 Mb/s
Intraplex T1 Multiplex
Receive DS0s
Intraplex T1 Multiplex
DS0 1
DS0 2
DS0 3
DS0 4
DS0 5
DS0 6
DS0 7
DS0 8
DS0 9
DS0 10
DS0 23
DS0 24
Audio A/D 1
Audio A/D 2
Audio A/D 3
56 kb/s port
Fraunhofer OCF module
64 kb/s port
Fraunhofer OCF module
DS0
DS0
DS0
DS0 7
DS0 8
one DS0 (64 kbs communications channel) has missng bit every 128 bits
56 kb/s port
Fraunhofer OCF receive module
64 kb/s port
Fraunhofer OCF receive module
other DS0 remains unaffected

Fig. 3

STARTING RIGHT, KEEPING CURRENT: EDUCATION AND TRAINING FOR BROADCAST ENGINEERS

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ABSTRACT

Few professionals need to know as much about as many divergent fields as do broadcast engineers. Broadcasting changes rapidly. Keeping pace is difficult. Further, the broadcast industry continually adopts techniques from allied fields such as telecommunications and computer science. Engineers are thus forced not only to keep abreast of broadcasting proper, but must follow developments in other disciplines that may soon become part of the broadcaster's domain.

The best way to cope with this onslaught of information is to enter the industry prepared for it. A sound technical education is a sure foundation. The first part of this paper addresses education for entry level personnel. The authors discuss means to build liaisons with local colleges, such as cooperative education programs and internships. Broadcasters are urged to develop a plan of action to insure that there will be entry level broadcast engineers available to meet the future "high-tech" needs of the industry.

The paper also surveys continuing educational opportunities for experienced broadcast engineers seeking to keep pace, or to advance. Enabling engineers to participate in continuing education activities is a classic way to encourage, foster, and reward technical excellence, all at the same time.

I- INTRODUCTION

The technical skills of broadcast engineers should be substantial upon entering the industry, and must be updated periodically to keep pace with the advance of broadcast technology.

The size and complexity of broadcast facilities varies widely. The technical needs of a 250 watt local AM daytimer can hardly be compared with the needs of large market stations, but technical excellence should be a goal at every facility. If the role of the engineering staff at a particular station is well understood by management, technical excellence will be encouraged, fostered, and rewarded.

The technicians who will be responsible for building, maintaining, and modernizing the broadcast facilities of tomorrow must be well prepared for their tasks. It is important that the broadcast industry develop a strategy to insure an adequate future supply of well qualified entry level broadcast engineers.

II- ENTRY LEVEL TECHNICIANS

If we assume that persons entering the broadcast field are bringing technical electronics skills with them, instead of coming to the industry to learn about electronics, we must consider where those skills were acquired. In most cases we are talking about a recent graduate of a college level electronics program.

Excellent electronics technicians know a lot about electronics. It is experience in the broadcast industry that distinguishes an excellent broadcast engineer from an excellent electronics technician employed in other industrial sectors.

Knowing something about electronics is certainly not the same as knowing something about broadcasting. There is a necessary, unavoidable, "pay your dues" training period. Entry level engineers need some opportunity to gain this experience. This will allow them to become productive sooner, and rapidly realize the rewards of full participation in the industry.

Finally, we must remember that "once an entry level worker" does not mean "always an entry level worker". Many career paths exist within the broadcast industry, and entry level engineers should be prepared to travel them. This implies that in addition to technical skills, technicians need written and oral communication skills and interpersonal skills, so that they can qualify for advancement opportunities. This underscores the importance of college level technical training programs, in which these important, but not necessarily technical, skills are provided.

It has been argued elsewhere that programs in Electronic Engineering Technology (EET), accredited by the Technical Accreditation Commission of the Accreditation Board for Engineering and Technology (TAC/ABET) provide excellent preparation for careers in all areas of electronics. Other kinds of electronics programs exist at the post secondary level, and a few even concentrate on broadcast technology.

On the Job Training

An excellent way to groom new technical talent is to participate in the cooperative education program of your local technical college. Co-op students are generally in the top third of their class, and have had at least one full year of classes. They then alternate semesters of class with semesters of full time employment until they graduate. The students usually return to the same employer for each employment semester. There is never an obligation to hire the student after graduation, but in many cases the student has become so valuable to the employer that an offer of employment is made.

A different option is to provide internships. In this case, students are given a work schedule that can accommodate their scholastic obligations, but still provide meaningful work experience. The most advantageous arrangement for the students is to offer just enough work hours that they qualify for important parts of the station compensation package, say medical insurance and perhaps reduced portions of sick leave and vacation. This often makes interns extremely loyal, especially if they have families or are in a period of retraining from a previous career.

Take the Initiative

It is the broadcaster's responsibility to provide these training opportunities, not the schools. Since less than three percent of the next decade's new jobs in electronics will be broadcast related, it is easy to see why colleges are reluctant to give up instruction in advanced digital and analog electronics in order to make room for courses in broadcast technology.

Attracting and satisfying these entry level workers may be vital, because broadcasters will face stiff competition for them. Two hundred thousand entry level jobs in electronics will be created during the next ten years, far more than the projected number of prepared workers.

There are some things we can do to help attract students to broadcast careers.

First, we can get their instructors interested. Many faculty members at the technical colleges would welcome a summer job or a consulting opportunity at a broadcast facility, and would then be better able to inform their students about broadcasting careers.

If you serve on the industrial advisory committee of your local technical college, you may be able to influence the quality and emphasis of the curriculum, and may be able to encourage more students to study electronics.

Finally, when you replace equipment of any kind, send the old stuff to the technical school. They may later throw it out, but it might be useful. In any event, you will have made your name and industry known.

Upward Mobility

Graduates of TAC/ABET accredited electronics programs will be computer literate, will have well developed communication skills, and will have a solid knowledge of electronics on a theoretical and practical level. The graduates with the best grades will typically have four or five offers of employment at excellent salary levels, and even the marginal students will have found suitable employment within nine months after graduation.

The graduates of modern electrical engineering (EE) programs are not often interested in the "hands on" kind of career that broadcast engineering offers. Graduates of two year accredited electronics programs command typical starting salaries from \$18,000 to \$23,000, while graduates of four year technology and engineering programs can expect starting salaries from \$28,000 to over \$36,000.

These graduates expect to be productive workers when hired, expect to study to keep up with changing technology, and expect to bear more and more job responsibility as they become experienced. They have not been educated for dead end jobs, and will usually ask about career paths at their employment interview. Many times advancement in broadcast engineering comes by taking a more responsible job at a different station. At larger stations, or in group situations, advancement may be possible without changing employers.

Accomplishments and excellence in job performance should be rewarded with professional satisfaction and appropriate financial compensation. The situation at your station should be reviewed so that a clear understanding of the opportunities can be given to prospective entry level technicians.

State of the industry

If we expect to attract well qualified young men and women to the technical side of the industry, we must offer attractive rewards. If broadcast engineering is regarded as a "necessary evil" by management, the reward system is defective. Entry level technicians with knowledge and savvy will be needed to meet the broadcast engineering challenges of the future. The well trained and highly motivated technicians we should seek have lots of options. The broadcast engineering option must be competitive.

III- CONTINUING EDUCATION FOR BROADCAST ENGINEERS

Few spurs can provide as much motivation to the seasoned broadcast engineer as realizing that an entry level worker with the right education and a few years of experience may suddenly shoot ahead and move into management. Although supervisory positions in broadcasting sometimes mean little more than giving up overtime pay, there is something disconcerting about having to take orders from someone who may be educated, but not experienced.

The preventive measure against this is continuing education. There is some evidence that this an industry trend. The Broadcast Engineering Magazine 1989 salary survey shows that tuition refund plans are a fringe benefit offered to 31 percent of radio and TV engineers across all markets, up from 26.7 percent in 1988, and from 24.1 percent in 1984.

Career paths

Working engineers who decide to get more education face two choices. They can seek further training in electronics, which would qualify them to do a better, presumably faster job of what they are doing already, or they can seek training in leadership and management.

Management programs are available at almost any two- or four-year college. Check with the business departments. Some schools also have programs in which credit is granted for demonstrated proficiency in skill areas such as mathematics or physics. The College Level Evaluation Program (CLEP) tests and their equivalents are designed to help translate skills gained on the job or in military service into college credits.

Credit for life experience can also be arranged. In these situations, students present evidence to show that a given life experience, say raising a child, should count for credit in a college course on the same subject. A faculty team evaluates the student's petition, and if it fulfills the requirements, credit is issued.

Of course, nothing is free. The fees charged for taking CLEP tests or evaluating life experiences are usually non-refundable. Once credit is "earned", it must be purchased, often at the same rate per credit hour the institution charges for normally matriculated students. This may seem irritating, since the institution is selling credits for which it has had to provide neither physical facilities nor instructors. However, it provides a certification that the student has at some point completed work equivalent to what would have been required in class, and is a lot easier to fit into one's schedule.

There is one other continuing education option for the working engineer--non-college courses, seminars and independent studies designed to strengthen skills in a specific area.

The authors have identified 13 areas in which these opportunities exist. These are shown in table 1. In many cases, Continuing Education Units, (CEU's) are awarded according to the number of hours of participation in such programs.

Table 1:

Non-college training opportunities for broadcast engineers.

1-Manufacturer's equipment training schools
Many manufacturers offer training sessions for maintenance engineers. These can often be negotiated into the equipment purchase price. These courses provide valuable information about servicing a specific piece of equipment. Best results are usually obtained in a hands-on lab instead of a lecture setting.

2-Manufacturer's "generic" training schools
Several manufacturers offer training in basic technical areas. Their equipment may be used in the class, but is not central to it.

3-Certification programs
Certification programs are an excellent way to gauge your grasp of technology. The Society of Broadcast Engineers (SBE), NARTE, ISCET (CET) and cable TV industry all have certification programs. And although the First Class Radiotelephone Operators License is gone, one can learn plenty on the way to Amateur Extra Class.

4-Meetings of professional societies
Attend meetings such as this. Buy the proceedings. Make or buy tapes (check the rules first). Then read and listen to them. Mail questions to the authors. Maybe give a paper yourself someday.

5-Short seminars
Dozens of companies and universities offer seminars on topics from LANs to fiber optics to diesel generators. You get their flyers by subscribing to trade journals. The most reputable of these companies guarantee your education--if you don't learn, you don't pay.

6-Correspondence courses
Many correspondence schools offer useful training programs. But shop carefully. Ask for references from former students who have gone into broadcasting.

7-Trade Magazines
Trade magazine editors are not in it for their health! Pick magazines you like and read them monthly. Learn what is carried in your public library, and read or scan a dozen or so magazines each month.

8-Books
There are precious few good books on broadcast subjects. A list of recommended texts is published by the Society of Broadcast Engineer's as part of their study guide for the SBE certification exams.

9-Industrial correspondence courses.
Several firms offer industry specific correspondence courses in areas such as lasers, cable television & building engineering (HVAC).

10-Instructional audio & video tapes
Several publishers offer series of tapes on various technical subjects. Listen while commuting. Watch while on your exercycle.

11-Computer-based courses
While interactive videodisks may still be rare, once can buy today several courses that work from floppy disk. The computer asks questions, you type your response. A computer study guide from the SBE is in the works.

12-Computer bulletin boards
A good source of both information and discussion. Several exist for broadcasters, dozens for electronics and computers.

13-Other engineers
Don't try to learn in a vacuum. Webster, the dictionary guy, defined the secret of learning as "Converse, converse, converse". But don't forget to listen more than you speak.

IV- CONCLUSIONS

Graduates of TAC/ABET accredited EET programs have an excellent education for entry level jobs in broadcast engineering.

Broadcasters should provide cooperative education or intern opportunities for technical school students, and employment opportunities for faculty. Equipment donations to technical colleges and service on their industrial advisory boards will benefit the industry.

Well qualified candidates for entry level broadcast engineering jobs have other options. Career opportunities and financial rewards in broadcasting must be appealing.

Broadcast engineers seeking continuing education opportunities face a choice--strengthen their position in electronics, or seek training in management.

There are many non-credit seminars and courses available in which broadcast engineers can strengthen their backgrounds in selected areas.

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PBS TECHNICAL OPERATIONS' TOTAL QUALITY PROGRAM

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Philip Crosby, a quality manager at General Motors years ago, once remarked that quality is a lot like sex. "Everyone is for it" Mr. Crosby wrote. "Everyone feels they understand it. (Even though they wouldn't want to explain it.) Everyone thinks execution is only a matter of following natural inclinations. And, of course, most people feel that all problems in these areas are caused by other people."

At PBS' Technical Operations, we have learned that quality is not an indefinable, abstract concept. But before I continue, it would probably be helpful if I explained a little bit about the departments with whom I work most closely on quality issues.

The Technical Operations Center is responsible for quality and continuity control for videotape origination of most of PBS' prime time, children's, and educational television programs. Other services provided to both internal and external customers include videotape dubbing, program technical evaluation and correction, studio origination for teleconferencing, as well as off-line and on-line videotape editing.

Technical Maintenance provides high-quality service to Tech Center and Edit Suite equipment and systems for both immediate repair as well as preventive maintenance designed to reduce instances and impact of equipment failure. Maintenance technicians also support member stations with advice on installation and service of broadcast equipment with which they are familiar.

Technical Operations, comprised of the Technical Operations Center and Technical Maintenance, is part of the division within PBS known as Broadcast Operations, Engineering, and Computer Services (BOE&CS).

The Public Broadcasting Service, by the way, is a private, nonprofit corporation whose members are the nation's 337 public TV stations.

Now that we have the "lay of the land" so to speak, let's get back to discussing just what we mean by the term "quality".

- First, quality is something that is determined by the <u>customer</u>, not proclaimed by the provider of goods or services.
- Secondly, we must be able to <u>measure results.</u>
- Third, one <u>cannot inspect</u> quality into a product or service.
- And fourth, the <u>commitment</u> to quality must come from the <u>top down.</u>

The <u>customer must</u> be the one to <u>determine</u> whether or not goods or services conform to requirements. After all, they are the <u>customers</u> specifications, and no one will know better whether or not those requirements were met. Therefore, PBS cannot tell its member stations that the services which we deliver are high quality unless we know for certain that those services conform closely to the stations' requirements.

One of the best ways of finding out whether or not your work meets the needs of your customers is to ask them directly. One of the ways we try to do this is through the PBS Engineering Committee, a group consisting of station chief engineers who represent member stations and regional networks. Usually at least once a year each station chief engineer is contacted by a member of the Engineering

Committee in order to obtain information on a variety of subjects including questions relating to the quality of services provided by BOE&CS. It is the collated responses that we get some valuable feedback.

For instance, last Spring several stations voiced concern over inaccurate program length timings they received from us. This was wreaking havoc with stations who employ automated master control switchers.

We attacked this challenge on a variety of fronts. First, the Quality Control Team, which is composed of Tech Center management, supervisors, and technicians, worked on the problem from a procedural point-of-view. Among other things, they came up with a new Standard Operating Procedure in which program videotapes are fast-forwarded just before air in order to verify the running times listed on the logs and within our own automation playlists. These times are transmitted via slate to member stations immediately prior to the program feed. In this way, our techs are able to prevent chopping off the end of a feed and technicians at the stations get a quick length verification of the upcoming event. Management also formed a special interdepartmental, intradivisional team to look at program length inaccuracies from a broader perspective. We discovered that there were a number of contributing factors. The most significant culprit was program re-packaging, which occurs when elements of the show, most often book or transcript offers, are added or deleted. The procedures for repackaging and resulting communications between BOE&CS personnel and members of the Programming division, who initiate the process, were somewhat muddled and ill-defined. This resulted in late and sometimes inaccurate notifications to the stations of program length changes. After working out intradivisional communications problems, we went to Programming and established mutually agreeable techniques for initiating repackaging requests which automatically trigger other systems necessary for communicating the changes to the stations.

The results were gratifying. Tech Center discrepancies relating to timing errors dropped, the stations are getting better program data from us (and on a more timely basis).

BOE&CS management staff have recently begun to supplement the data acquired by the Engineering Committee calls by making our own periodic phone contact to member stations. In this way we may be able to obtain more timely and detailed feedback regarding our services (in a relaxed, informal and personal manner) as well as to discover what new initiatives, projects, and problems each station is facing.

Our stations also determine the technical parameters for PBS programs through participation in the Technical Operating Specifications (TOS) Subcommittee, which reports to the Engineering Committee. The TOS Subcommittee, chaired by Nevton Dunn of Oregon Public Broadcasting, maintains and updates the specs, which are listed in a manual known simply as the <u>Technical Operating Specifications</u> (or <u>TOS</u>), which is distributed to all PBS program producers. Nearly every parameter imaginable is covered- quality of tape stock a producer may use, pulsewidth requirements, multichannel audio specifications, standards for tape leader and trailer, etc. In this way the stations define what level of technical quality is acceptable.

Quality must be <u>quantifiable.</u> We have to be able to measure the results of our work. And we have to be able to display relevant data in a fashion that tells us where we should focus corrective activity in order to get the most bang for the buck.

For example, one measure of quality that we can track is air discrepancies. While we have recorded and reported air discrepancies for years, we never really properly employed techniques which lent themselves to meaningful analysis of the data. Although each discrepancy received some investigative attention the following day, there seldom was a concerted effort to remedy certain recurring problems because we didn't have the data organized in such a way that would indicate the correct focus of pro-active management attention.

Then we began to list discrepancies into categories, including equipment failures, operator errors, incoming problems on live feeds, and others. Through a statistical process based on Pareto analysis we have looked at the magnitude of each specific problem in order to identify initial targets. (In its simplest form, Pareto analysis involves displaying data on a bar chart, such that the different categories appear in descending order.)

A group known as the "QC" or Quality Control Team, (mentioned earlier) consisting of Technical Center management, supervisors, and technicians was formed in September of '88. Using air discrepancy reports, the team initially looked at all operator errors committed during the month of the team's formation. The errors were matched with their sub-groups and showed what elements were most common. These problems were further examined to see if procedures could be written in order to reduce the amount of errors in those specific areas.

As a result of that first meeting four new procedures were initiated as well as proposals for changes in automation software. The QC Team continues to meet on a regular, usually monthly basis to review discrepancies and action item assignments. Team members also report developments, ideas, and work in progress to the entire Technical Operations Center at periodic informal get togethers. At the January '90 QC Team meeting members set a goal to reduce operator errors from the current average of 8 per month to 6.5.

Technical Maintenance, under Associate Director Larry Jefferson, is also using Pareto analysis to help them focus their quality improvement efforts, specifically to reduce the number of equipment failures.

Initially, during the period between July '88 and April '89, Maintenance analyzed the different categories of equipment failures that resulted in air discrepancies. The resulting Pareto chart indicated that attention should be focused on control room automation and VTR problems, which together accounted for almost 75% of equipment related air discrepancies.

Further investigation of automation indicated that software problems, rather than failure of the computer hardware, was responsible for over 95% of automation discrepancies. Problems experienced with earlier software version outnumbered instances noted with the later version by more than 2 to 1, even though we converted to the newer version in November of '88. Maintenance, Tech Center, and Engineering personnel worked closely with the software vendor, which ultimately resulted in Version 5, to which we converted in October of last year.

The second largest source of equipment-related discrepancies were our tape machines. Further study indicated that the highest VTR subcategory failure rate occurred within video subsystems. Examination of the various subsystem elements denoted that head failure accounts for the largest proportion of on-air problems. Average head life on our machines is about 1,700 hours, well above the industry average. Nonetheless, Maintenance did note 2 or 3 instances of premature head death probably attributable to rough handling. As a result, the QC Team investigated and ultimately modified procedures to indicate that head covers should be closed at all times (except while threading) as well as reminders to clean the tape path regularly. Maintenance personnel modified preventive routines and were successful in campaigning for budget increases for head replacement.

When Larry Jefferson graphed our equipment related discrepancy data for April through September '89, we noted a new number 1 problem: The Automated Program Delay Units (APDU) that went on-line last July to provide us with automatic time-zone delays of prime-time and children's programming, which rely upon internally mounted component VCRs, have replaced automation for first place among our list of equipment concerns. Maintenance, working with the Technical Center and Engineering personnel, focused attention in this area. (This same group was successful in concentrating on automation problems, resulting in significant reduction of errors for this category.) I'm happy to report that as a result of their work APDU failure-related problems were down for the last three months of 1989. Now VTRs are again at the top of the list (although the number of on-air VTR failures are less than before). The point here is that your quality targets are moving ones. As you achieve success concentrating on your major problems, the previously lesser categories become your new prime targets.

One of the measurements that must also be tracked is the cost of not doing the job correctly the first time. Take, for instance, the cost of refeeding a program due to a problem during the original transmission.

First, Broadcast Operations must find available satellite time, notify the stations, notify the tape library, and amend electronic and paper logs. On a good day, this can be accomplished in about 30 minutes.

Technical Operations management must create the Technical Work Request and amend personnel schedules. If nothing else needs to be done, such as resheduling other work to make room for the refeed, then only 15 minutes is required.

Technical personnel involved with the refeed include a control room operator, a technician for videotape, one supervisor, and the satellite uplink operator.

Also to be factored into costs are satellite time and equipment wear-and-tear. The total cost (best case) for a one hour refeed is $426. Anything can increase this number, and as Murphy rules in the world of broadcasting, there are seldom situations where "best case" is the reality. This figure, by the way, is for the origination end, and does not take into account the costs at the station for re-recording the program.

The real lesson here is that we can't afford to be without quality. Doing it right the first time is the only thing that makes economic sense.

The third principle is that we cannot inspect quality into a product or service.

For years we have tracked the percentage of programs that arrived with technical impairments. An average of 60% of the shows left the producer with some kind of defect. This figure has not varied much since I began tracking this data eight years ago. Despite the fact that each PBS program master was technically evaluated, these inspections did not vary the percentages of incoming problems.

Nevton Dunn, who chairs the previously mentioned TOS Subcommittee, tried to apply pareto analysis in order to determine which types of technical problems were occurring most often (and therefore provide us some indication as to where to focus attention). Unfortunately, Nev found that the orders of impairments were so diverse that none stood out for special attention.

Different approaches were necessary. The TOS subcommittee and Technical Operations management have come up with some ideas which appear to be bearing fruit.

The first initiative is one we call the Technical Quality Report Card. John Prager, Associate Director-Technical Operations Center, is tracking the performance of our program producers on a quarterly basis by indicating the number of technical impairments (TINS) per unit of program duration. All producers who have submitted at least 2.5 hours of programming during the quarter are listed with the impairment/time ratio data in the order of superior performance.

The Report Cards are sent to all producers listed. This gives each producer a sense of their performance relative to their peers. Accompanying the Report Card is a letter from John Prager which supplies the producer with more detailed information on a program-by-program basis. John also provides the producer with a synopsis in the cover letter, pointing out major trouble areas exhibited, such as 3/4" stock problems, audio level variances, or impairments caused during post-production. As a result of collecting and reporting data in this fashion, we have discovered not only that producers tend to repeat their mistakes, but that by working with producers one-on-one with technical impairments on a quarterly basis we can provide the producers with specific major targets for improvement that might not have been obvious to them before.

The data from the first Report Card, covering the first quarter of CY '89, indicated an overall average of 41 minutes between impairments submitted for the 22 listed producers. That number improved to 43 for the second quarter and to 59 by the time the third quarter was computed! Producers apparently were making beneficial use of the information with which we were supplying them, and it is having a positive effect on the quality of the product.

Feedback systems aren't the only means of helping our program producers improve technical integrity. Both John and I encourage producers to call us with technical questions, no matter what stage of production they may be in. We would rather help them fix or avoid a problem before the videotape is sent to PBS than to affect repairs or reject the program after it has been technically evaluated.

Our eventual goal is to help producers improve the technical quality of program submissions to the point that the expensive technical evaluation/correction process will no longer be necessary.

PBS is also beginning to explore the production of training tapes in order to improve quality both in-house and at our stations. Last year Technical Operations created a tutorial videotape which explained the evaluation process and gave examples of many common types of impairments. The reaction to the tape from program producers and member stations has been highly favorable. We are now working on a training videotape for MTS audio.

Technical Maintenance provides a service that can verify whether or not a type-C 1" videotape was recorded within

SMPTE spec. A producer can send in a piece of tapestock to the Videotape Evaluation Laboratory where it will be developed and then analyzed under microscope in order to determine several parameters, such as track straightness and location, scanner guide alignment, and other factors that can effect tape interchange.

The final tenet I've listed necessary for an effective quality program indicates that the commitment to quality must come from the top down. Without the visible support from senior management, lower levels will not effectively participate in the processes or take new initiatives seriously.

Fortunately for us at PBS, we do have that commitment. BOE&CS's Senior VP, Howard Miller, had years of experience implementing quality programs while with Westinghouse in Japan. Bruce Christensen, President of PBS, and Neil Mahrer, Chief Operating Officer, are also known for their sincere commitment to quality. Earlier in the year Bruce and Neil attended an affair called "skunk camp", which was, in effect, an intensive seminar on quality and customer service given by Tom Peters of "In search of Excellence" fame. Last year a two day seminar entitled "Creating Value for the Customer", also given by the Tom Peters group, was voluntarily attended by most of the management and professional staff at PBS. Also last year, the "Customer Service Committee" was formed within PBS to promote quality service and to publicly recognize individuals, especially those holding lower level "front line" positions, who have demonstrated "above and beyond" attitudes. In short, quality is becoming ingrained within the corporate culture.

Another example of this is the suggestion system recently initiated within the Tech Center. Though rewards are small, participation so far has been beyond expectation. Although the system has only been in place for two months, over a dozen improvement suggestions have been received, most of which have already been implemented!

Upper management asked the Customer Service Committee to come up with recommendations for a company-wide suggestion system. I have formed a subcommittee to do this and we are very close to completing our task.

Because of this support from the top, it is much easier for those of us in the middle and on the front lines to "do the right thing", and to know what "the right thing" is. Without this, the questions of credibility and long-term commitment could ultimately defeat any positive efforts.

Fortunately for us at PBS, a sense of duty and pride is prevalent in nearly every corner. Many of us came from PBS stations, and therefore not only have a pretty good idea what is required, but also some recollection of what happens at a station when feeds from PBS are interrupted or otherwise impaired.

Our objective is to have the best air quality in this country. And our quest for improved quality will never be done.

MAINTENANCE PROCEDURES SIMPLIFIED USING STEREO NOISE

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Noise has long been considered an undesirable entity to be eliminated. However, many of the tests and measurements in a broadcast facility are more easily performed using stereo noise as a test signal. In addition, pulsed noise can be used to measure the dynamic response of audio processing equipment under real-world operating conditions.

This paper defines noise characteristics and describes how various noise test signals can be used to simplify and speed common broadcast tests and measurements.

INTRODUCTION

Noise can be used as a test signal containing all frequencies of interest simultaneously, eliminating the need to repeatedly adjust individual tones. Starting with the flat frequency response of white noise, weighting filters can provide many useful response curves, tailored to the specific measurements being made.

A special pulsed noise signal has been adopted by the NRSC as a standard test signal. Consisting of pulsed USASI noise with partial channel blending, this signal accurately simulates program audio and forms a dependable, repeatable baseline for broadcast measurements.

DEFINITIONS OF NOISE

Several noise concepts should be defined before proceeding to specific applications. White noise is the basis from which noise generators form their various output weightings. True random white noise may be defined as having no periodic spectral components: Its future value cannot be predicted except in statistical terms.

A further measure of the quality of white noise is its statistical amplitude distribution. True random noise should conform to the Gaussian amplitude distribution. This is the familiar bell-shaped normal distribution curve shown in Figure 1. The horizontal axis shows noise amplitude in standard deviations (σ), where one standard deviation is equal to the RMS voltage of the noise. For any voltage segment along the horizontal axis, the corresponding area under the distribution curve is the probability that the noise voltage will lie within that voltage segment at any instant and is equal to the fraction of the time that the noise voltage is in the voltage segment. The area under the curve from -4σ to $+4\sigma$ is very nearly 1. In other words, the instantaneous noise voltage almost never exceeds the value equal to plus or minus four times the RMS voltage. A crest factor of four is therefore sufficient to pass Gaussian noise.

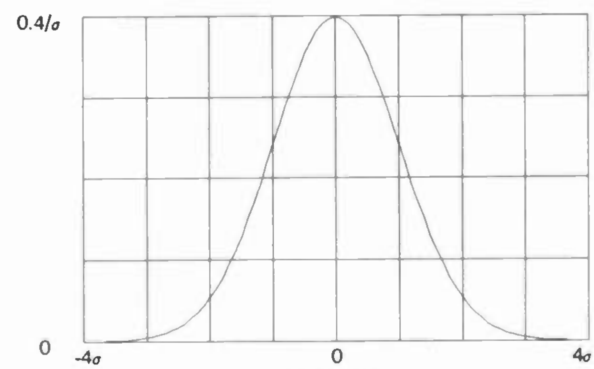

FIGURE 1
PROBABILITY DENSITY FUNCTION

Three noise weightings are most commonly used: White, Pink and USASI (United States of America Standards Institute). White noise contains equal spectral energy per unit of bandwidth, as evidenced by its flat response over frequency seen on a spectrum analyzer or constant bandwidth wave analyzer. Pink noise is formed by filtering white noise with a 3 dB per octave lowpass function. It contains equal spectral energy per octave of frequency. This provides a flat response when measured with an octave or third-octave real time analyzer (RTA). USASI noise is formed by band-limiting white noise at the low and high ends. The filters used are both first-order, 6 dB per octave. The highpass corner frequency is 100 Hz and the lowpass corner is 320 Hz. Figure 2 shows the spectrum curves of these three types of noise.

The NRSC-1 AM Broadcasting Standard, ANSI/EIA-549-1988, defines pulsed USASI-weighted noise as a standard test signal. The USASI weighting curve closely simulates the spectral distribution of typical program audio, having primarily midrange content with rolled-off highs and lows. Dynamic pulsing simulates the repetitive amplitude

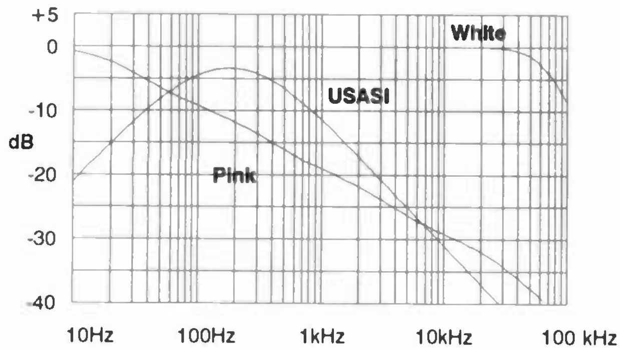

FIGURE 2
RESPONSE VS FREQUENCY

variations of music and speech. The pulsed noise is defined to have a 20 dB peak-to-average ratio, with the peaks occuring at a 2.5 Hz repetition rate, with a 1/8 duty cycle.

This test signal further simulates real-world program audio by partially blending the left and right channels. This creates a sub-channel signal (L-R) which is 3 dB below the main channel (L+R). A block diagram of the circuitry required to create this test signal is shown in Figure 3 below.

STEREO NOISE GENERATOR REQUIREMENTS

A stereo noise generator should provide a minimum level of performance and functions to be useful for broadcast tests and measurements. It must generate two independent, uncorrelated channels of gain-matched noise. This noise should be available in white, pink and USASI weightings. Other more specialized weightings such as CCIR, inverse RIAA, and A- and B-weightings might be available for specific tests.

Many broadcast measurements require crosstalk and separation tests. Left and right channel output combinations like LEFT only, RIGHT only, MONO, and LEFT=-RIGHT can greatly speed these tests, as will be discussed below.

The ability to modulate the noise amplitude provides a much more useful test signal which can simulate the dynamics of program audio. If a noise generator combines USASI weighted noise with the NRSC channel blending and amplitude pulsing, then the special NRSC noise test signal discussed above can be generated.

NOISE APPLICATIONS

Response Measurement

Frequency response measurements are probably the best-known use of noise. Noise is useful for measuring the response of consoles, amplifiers, studio-transmitter links, or even equalized phone lines. System response can be quickly ascertained by feeding the noise test signal into the input and monitoring the output with a real time analyzer (RTA) or spectrum analyzer. Pink noise provides a flat response when measured with a third-octave or octave RTA, whereas white noise is appropriate when using a spectrum analyzer or wave analyzer. Exercise care when using white noise to prevent loudspeaker damage from the strong high frequency content of this signal.

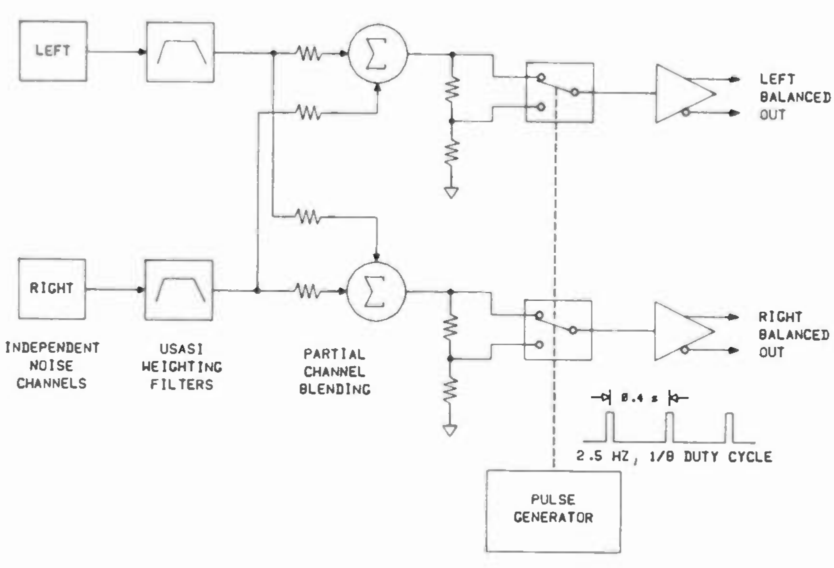

FIGURE 3: NRSC TEST SIGNAL BLOCK DIAGRAM

Tape Machine Head Alignment

Tape head azimuth adjustment using stereo noise makes a complicated process easier. First align the playback head using a standard pink noise alignment tape. Then connect MONO pink noise to the tape deck inputs, and monitor the output while recording. While adjusting the record head azimuth, the operator will clearly hear the increase in high frequency "hiss" when the head is optimally aligned. This measurement can also be monitored on an RTA or an oscilloscope.

After aligning your master cartridge recorder, generate a series of pink noise alignment tapes. These tapes can be used to phase-align all the recorders in the studio to the master recorder.

Filter Response and Alignment

Filters are employed in many broadcast products. They can often be aligned much faster using noise than audio tones. For example, to align the AM Stereo Exciter, start with all delay and EQ switches disabled. Apply LEFT-channel USASI noise and adjust the output level to reach a 50% indication on the modulation monitor left channel. Connect the mod monitor's left and right unbalanced outputs to an oscilloscope's A and B inputs, respectively, and set the scope to read in X-Y mode. The scope display will show a vertical line with horizontal noise extrusions. Now adjust the Left Bulk Delay to minimize the horizontal noise traces. Finally, adjust the Left Low Frequency EQ to produce a near-perfect straight line. Repeat the above steps with RIGHT-channel noise, observing a horizontal line on the scope display.

Separation and Crosstalk

Separation and crosstalk tests measure the level of signal intrusion from one channel into another. Separation refers to the discrete left and right channels; crosstalk refers to the main channel (L+R) and subchannel (L-R) signals. Separation can be measured by using left-channel only and right-channel only noise, and measuring the level of noise in the opposite channel. Crosstalk measurements require MONO (main channel only) and L=-R (subchannel only) noise signals.

White noise, with equal amplitudes over the audio bandwidth, can be used to measure the worst-case intrusion of full-level high frequencies from one channel into another. Use USASI noise for numbers that reflect the real-world separation and crosstalk levels that would be present during program audio.

The dynamic separation of an entire stereo transmission system can easily be measured by applying left- or right-channel noise to the system input and reading the separation on the modulation monitor. This test will show the effects of every device in the system, and can be sobering.

Audio Processor Setup and Alignment

Adjustment and setup of audio processors using USASI noise will give very close correlation between the test adjustment and actual program material. The "quality" of the audio processing can be measured by comparing the peak-to-average ratios of the pulsed noise at the processor input and output. The difference in dB corresponds to the amount of compression taking place. An oscilloscope and an AC voltmeter with a long time constant can provide a rough measure of the peak and average levels. The Potomac Instruments QA-100 QuantAural™ is particularly well-suited to this type of measurement, with Peak and Average settings and four measurement bands that correspond to an "inverse-USASI" function.

The audio processor's frequency response <u>during active processing</u> (peak limiting, compression, etc) can be measured using pulsed pink noise and a real-time analyzer in peak-hold mode. This provides an accurate picture of "real-world" processor frequency response - which may not be as flat as response curves using steady-state tones or noise.

System Cabling and Polarity

By clipping the positive peaks of the LEFT channel and the negative peaks of the RIGHT channel noise, a unique stereo test signal is created. When fed into stereo cabling and viewed using an oscilloscope, channel reversals (LEFT and RIGHT) and phasing errors can be quickly spotted. Switch first to LEFT-channel only noise, and verify on the scope that only the left wire has signal present. Then switch to the CLIPPED noise to verify that channel phasing is correct. If the phasing on the left channel is reversed, the scope will show the <u>negative</u> peaks clipped, rather than the positive. The same procedure can then be repeated for the right channel. In facilities with long runs of wires, multiple runs through patch bays and daisy-chained audio processors, this method of channel verification speeds testing of the equipment wiring and quickly points out potential faults.

Amplifier Dynamic Range

A test signal with a high peak-to-RMS factor which limits the total RMS power is useful for this test. The peaks will test the dynamic range capability of an amplifier, while the lower RMS value won't work it too hard. This ratio is typical of program audio. A noise generator that produces a symmetrical signal with a crest factor of 4:1 or more is useful for this purpose. The peaks of white noise occur often enough to be visible on an oscilloscope. The broadband signal insures that the amplifier will be tested throughout the audio frequency range.

Meter Measurements

Many level meters are polarity sensitive because they use internal half-wave rectification. Two of these meters which are wired across the same line but with opposite polarity will read differently and appear to have different ballistics, even though they read the same with line-up tones. This effect is especially pronounced on asymmetrical waveforms like the male human voice. The clipped noise signal discussed above is inherently asymmetrical, and can be used to quickly pinpoint such meters.

Meter response ballistics can be quickly measured by switching USASI noise on and off into the meter.

Transmitter Tests

White noise is an excellent signal for making transmitter adjustments while monitoring the AM noise component of FM or TV transmitters. Since a broad band of frequencies is applied to the transmitter under test, the transmitter tuning and loading can be more accurately adjusted. Furthermore, when used in conjunction with a spectrum analyzer, the effect of tuning and loading adjustments on system bandwidth can be seen.

Pulsed noise can be used for checking the positive peak capability of AM transmitters without overworking them. It can also be used to verify that the lightning gap on the mast base won't flash over on peaks, without keeping the power there long enough to weld anything.

NRSC Compliance Measurements

A stereo pulsed noise source provides a means for making repeatable measurements to determine compliance with the new FCC emission standard. The repeatability of a pulsed noise test signal provides more accurate readings than can be obtained using regular program material. Pulsed USASI noise combined with NRSC channel blending simulates program audio very closely, yet offers a test that - though it is random in content - has the same distribution of frequencies in each burst.

CONCLUSION

Stereo noise is a useful test signal and an improvement over discrete tone testing in many applications. A stereo noise generator which can amplitude-modulate the noise, and which provides additional output capabilities discussed above such as MONO and L=-R, NRSC channel blending, and clipped noise provide the user with enough flexibility for a myriad of useful measurements.

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The author is a Design Engineer for Delta Electronics, Inc. in Alexandria, Virginia. While at Delta he has worked on the designs for the Splatter Monitor and the Stereo Noise Generator, in addition to numerous smaller projects. He graduated with a BSEE from University of Virginia in 1981, and worked for six years in avionics instrumentation and military aircraft radar signal processing before coming to Delta.

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