PROCEEDINGS

36th ANNUAL BROADCAST ENGINEERING CONFERENCE

NATIONAL ASSOCIATION OF BROADCASTERS

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National Association of Broadcasters

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--TABLE OF CONTENTS--

SELECTED PAPERS FROM PRIOR ENGINEERING CONFERENCES

SMPTE AND THE FUTURE OF TELEVISION

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There are four general factors that influence the Society's role in future television:

- Technological progress - its rate of growth and direction
- Organization of our engineering effort to meet new needs with appropriate concern for due process procedures
- Equitable documentation processes
- An acceptance of and delination of documentation and engineering responsibilities through cooperation with other national and international technical and trade organizations.

As we look to our role, we note that there is the opportunity to fill a need by documenting high technology broadcaster requirements. There is also a parallel need to document the requirements of other video users. The concept of a new technology committee introduced just six years ago has broadened our base. This concept of documentation has also caused us to have prudent concern for the government's attitude toward the regulation of standards activity. To maintain a non-regulated status quo we must strive to maintain an apolitical scientific and technical posture for the national and international individual experts that participate in our technology committees amd working groups.

There are many new subjects that are being addressed and can be addressed by our technology committees. We achieve responsibility for these activities by having the technical expertise to develop valid documentation. We assume the responsibility by agreement of the delineation of its scope through cooperation and effective interface with the many organizations representing technological progress and application. Now let's examine these points in detail:

Last year was particularly significant. The Society concluded the forum for anlayzing the merits of various digital sampling frequencies in the NTSC system, and structured a demonstration to compare sampling frequencies and hierarchial levels. We compared and compromised our goals with other international organizations. This multi-million dollar, multi-thousand man-hour effort has been documented in the special October 1981 issue of the Journal and will be updated by Kerns Powers in the next presentation. I appreciate this opportunity to make you aware of some of the issues affecting our administration, the actions we have implemented, and those pending as they impact on our future role in TV. The government has expressed regulatory interest in standards developing organizations and we were concerned.

Two years ago the Society took affirmative action in responding to proposed government regulation of standards activity in a formal reply to the Federal Trade Commission. This action by the FTC was undertaken even though Congress softened its position and sought <u>not</u> to regulate the standards community by setting aside two earlier bills. The issue angered Congress -- and the FTC was directed to withdraw, and has compiled in all instances except where standards actions may impact or cause restraint of trade. The FTC has announced that it will make public the disposition of its rule next month.

There have been decades of discussions about how to increase the effectiveness of the government's participation in and use of voluntary developed consensus standards. The Office of Management and Budget (in OMB 119) tackled the problem in an attempt to provide guidelines for federal employee participation in voluntary standards development. Their approach was to certify the organizations acceptable to Federal employees. Detailed guidelines for the standards organizations method of standards development, due-process procedures, and participation requirements had to be met before government employee participation was permitted, lest the employee be identified with special interests. Simply a classification system for volunteer standards development bodies.

Although the rule was modified, and was in a form the Society was willing to comply with to achieve the highest listing, a change in national administrative attitude toward regulatory government has set aside the regulation's implementation.

Thus after several years of examination and reexamination of our standards procedures and practices to determine potential compliance with federal rule-making, we are left with only two issues of snigificance.

The first, to encourage federal employee participation in volunteer standards activity because of governments role as contractor, user and regulator. The second is to respond to a document of international significance -- the General Agreement on Tariff and Trade, known as the GATT Code. This agreement recognizes the importance of voluntary standards in international trade and cautions their adoption as potential non-negotiated barriers in restraint of trade. It also provides a basis for encouragement for broad international participation or liaison in our standards documentation process.

Thus as we look to the Society's role in the future of television, we do so looking back over our shoulder at a decade of proposed government regulation of the standardization process. A point I wish to emphasize is that the Society did not require the government to detail equitable standards development procedures, equitable and open committee participation, industry-wide awareness of our standards activity, and due process with the right to appeal standards actions. We have always behaved in an open and equitable manner, and shall continue to behave in such a manner.

The Society's role in the present and future television technology has its roots in motion picture technology documentation. In motion pictures we address the elements of the technology from the scene to the screen. Although this may seem rather straight-forward, the delineation of the motion-picture discipline evolved through the challenges of elemental specification by mechanical, chemical, optical, illuminating and finally, electrical technical societies. It was the question of who takes the responsibility for the audio accompanying motion pictures that finally caused the Society (in the early 1930's) to encourage the ASA to form a Photographic Standards Management Board to allocate standards tasks.

In the early 1950's, the question of who standardizes which elements of television caused concern among the SMPTE, the RETMA (now the EIA), and the IRE (now part of the IEEE). Because the ASA Electrical Standards Management Board declined to become involved in jurisdictional questions of standards development, the three organizations formed a committee to handle these questions -- the IRS. With subsequent participation by the broadcasters through the NAB and cable through the NCTA, the organization has become the Joint Committee for Inter-Society Coordination (JCIC).

Why an intersociety coordination? Well what we are trying to accomplish is basically a communications bus that will interface - with the many centers of technological progress and applications -- simply that's it.

Now, how has it worked? Practically, it has been through the sanction and/or encouragement of the JCIC that the Society has expanded its participation in television standards. Our natural involvement in television evolved from motion-picture technology, telecine, and kine-recording because motion pictures were the sole method of image-recording, distribution, and reproduction of early TV program material. Subsequently, the NAB and IEEE encouraged our acceptance of responsibility for all recording and reproduction technology. Developments since were highlighted by the documentation of 2-inch quad, time-code, and 1-inch helical specifications.

When the Society reorganized its engineering efforts in 1975, we were aware that documentation of common commercial practice was inadequate (although this method represents more than 90 percent of national standards efforts). We recognized it was essential to move into the study and documentation of new technology, and structured our engineering committees to allow this activity. We are prudently aware of the risks attendant to documenting new -- rather than existing -- technology, and require our chairmen to carefully administer due-process standards procedures.

Once organized, the Society looked to the future. With the acceptance of the JCIC six years ago, we began to study the digital video domain.

Five years ago we recognized the significance and potential of high-definition television and brought the issue to the JCIC with the offer to study the technology. Our committee effort for three years was theoretical. Since February 1981, however, the real and significant potential of higher-quality video is an unparalled standards challenge -- not just through higher line rates, but through improved quality within existing radiated signal constraints. Soon we shall return to the JCIC to report our progress and readdress the potential jurisdictional issues.

The new technology concept also allowed us to address the challenge of specifications for video discs. Here we did not find straightforward problems but rather became involved in new technology addressed by a plurality of long-term non-compatible approaches to a consumer need. Some have questioned the Society's strength in affecting a standard and have found us wanting.

The constraint upon us and on other organizations is that we must work within the voluntary consensus concept. Both the needs of the user and the willingness of the supplier to fulfill those needs must be in harmony. Documentation without compromise seldom occurs. Our strength, therefore, comes from the dedication of the committee members to identify and develop a standards needs that serves both user and supplier and encourages response to that need. Past experience has shown that the persistence of user requirements and the need for compatibility have and will bring forth more than one standards compromise. Their interest is high, and we hope that our committees can direct their documentation efforts to fruitful conclusions.

More than two years ago we delved into an area of committee work that some have questioned. One of our members found great inconsistencies in the electronic display devices or monitors for medical equipment used in nuclear, ultra-sound and CAT-scan diagnostics. Here was a life-saving need for consistency, but investigation revealed "no takers" to attack needed standards questions. Thus, after notifying many organizations of its intent, the Society offered its services and expertise for the development of monitor setup specifications. Now, two years later, Mr. Lisks' committee of international scope (SMPTE Subcommittee on Recommended Practices for Medical (Electronic/Photographic) Diagnostic Display Devices) is highly endorsed by the medical community as being the best possible combination of experts to specify the display characteristics of monitors depicting the output of medical diagnostic equipment and their reproduction on photographic and electronic recording materials.

When digital audio was demonstrated to the Audio Engineering Society (AES) in November 1977, the Society recognized the potential impact digital audio could have on sound accompanying film and tape images. Through the JCIC the Society and the NAB offered organizational and secretarial support to the Audio Engineering Society's embryonic standards effort. However, the AES, recognizing the long-term need for technical society support in audio, chose to undertake the task as its initial standardization effort.

The Society is in full support of the AES effort, we have reviewed motion-picture and television needs and last year offered guidelines of suggested sampling frequencies of 48 or 60 kHz. We now have joined AES to formalize standards proceedings directed toward a single compatible digital audio sampling frequency with present discussions centering on 48 kilohertz.

Later you shall hear how editing systems and production innovations created a need to coordinate the operation of many types of equipment. A three year effort has yielded digital specifications for the control of video equipment with world wide impact.

We have also had to touch upon the characteristics of computer language and protocol. Bob Lund's working group has been studing the problem of edit decision list interchange (EDL). The format documented is consistent with familiar industry practice with additional "special data" that might be transmitted to special effects devices, variable speed motion controllers, audio consoles and the like.

We are challenged by the concept of imagery versus television as multi-functional self-luminescent display devices expand on the display role of present receivers. Soon electronic still imagery shall reach the market and we anticipate that it shall be the Society's background and expertise that will fulfill those unique standards needs.

From the preceeding comments on our activity, you can see that our television horizons have broadened significantly in the past six years from one of simply telecine, test materials, and reproduction. How much further should we go, and how should we focus our efforts? Shall we embrace video from scene to screen, or what is our rightful share? How can we best serve the categories of users from professional to consumer, and industrial to government? I have posed these questions to our Standards Committee, to a Special Ad Hoc Committee, and to our Board. Input from you would be welcome as well.

Under new regulations by ANSI, we plan to return the JCIC to a formal status. The Society is now requesting that its roles and goals be reexamined with the intention of full cooperation with our sister organizations to serve the needs of the motion picture and television disciplines. Any look to the future mandates careful examination of international questions.

One of the strengths in our work and the effectiveness of our documentation has been that committee participation is open internationally. Broad-based international input in the early drafting of documents precludes costly compromise in subsequent ISO, IEC or CCIR meetings.

ANSI recognizes and endorses this approach, and has developed procedures whereby companies that do not manufacture products in this country or non-profit techncial organizations not based in the United States may become international members of ANSI.

In recognizing our responsibilities, the Society is looking to improve the effectiveness of its role. We reported to you in the October issue of the Journal the significant liaison established with the European Broadcasting Union by Frank Davidoff's Task Force.

The potential benefits of close-working international relationships are under review. The questions posed span basic communications through correspondence to the potential of international or oversees SMPTE Engineering Committees or to an International Council of Societies of the all-embracing discipline of imaging. As you can see, we must face potential changes, and we are receiving international encouragement to the need to judicially address the issues.

As we look to the future, there is one vital thing we must do -- we must refrain from quenching the spark of creativity and initiative, which resolves seemingly impossible problems or conditions. I do not believe our Society should submit itself to the bondage of status quo -- and its stifling of innovation and advance. The pressure and changes abroad today require our adaptability to new concepts -- logically correlating advances with requirements into our evolving program. The industry needs may be reached by diverse routes, but with the cooperation of other organizations, we hope to encourage a thrust to defined goals.

Standardization is the amalgamation of many approaches, methods, disciplines, and requirements to a worthy conclusion or product. All indications point toward achieving this goal by doing it once -- doing it right -- doing it internationally.

THE SMPTE CONTROL NETWORK

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FORWARD

The cost of modern television production facilities requires high utilization of each unit of equipment. This need can be satisfied by sharing the use of equipment among several functions within a plant.

Editing systems and production innovations create the need to coordinate the operation of many types of equipment from different manufacturers.

These requirements, to share and coordinate the operation of production equipment, create the need for a common, flexible control system which can accomodate a wide variety of equipment.

Many items of present day equipment include provisions for remote control. Unfortunately, the control schemes are quite varied and not given to orderly inter-connections between equipment from different manufacturers, and, occasionally, between equipment from the same manufacturer.

SMPTE WORKING GROUP FOR DIGITAL CONTROL

The SMPTE Working Group for Digital Control of Television Equipment, composed of members from user and manufacturer environments was formed in 1978 to develope standards for a control network to be a commmon interface for all types of equipment from all manufacturers. The group established the following guidelines for its work:

1. The standards must be developed rapidly to prevent proliferation of non-compatible control systems.

2. The system must be reliable. Electrical and mechanical components and the operating specifications must all combine into systems which function reliably in the environments normally encountered.

3. The system must use existing technology instead of requiring development of new elements for system implementation as happened with the IEEE-488 bus.

4. The system would, where possible, utilize established, proven specifications and techniques.

5. The standards must be simple to implement. They must be economical for small systems, yet flexible enough for large systems and future requirements.

6. The system's communication rate must be adequate to provide a response time within one TV frame.

7. System configuration should be left to the user. The standards must permit each user to configure a specific system to meet his needs, and rapidly reconfigure the system to meet changing operational requirements.

The working group has produced initial standards for the control network. Two field tests have been conducted to validate the initial standards and several manufacturers are now incorporating them into new equipment, and producing interfaces to marry older equipment to the network.

SYSTEM OVERVIEW

The elements of the control network are: (fig.1)

1. Bus Controller. One bus controller connects to a network. It supervises all other devices which are connected to the network through the use of a Supervisory Protocol. The bus controller may be incorporated into a device which performs additional functions, such as an edit controller, but it is a distinct system entity whose function is delivery of control messages and the management of the control network. An operational plant will contain as many controllers as there are interface systems.

2. Tributaries. Each operational device in a system connects to the network through a tributary. A tributary transfers the messages to and from an operational device as specified by the system supervisory protocol. The tributary may be a distinct unit of equipment, or incorporated into an operational device, but it is a distinct entity with the function of managing network interface and delivering control messages to and from the operational device.

3. Interface Bus. The interface bus is the communication channel which carries the messages between tributaries and the controller.

The network configuration can be either point-to-point or multi-point as required by the user.

In the point-to-point configuration, also referred to as "star" and "radial," more than one interface bus radiates from the controller and only one tributary connects to each bus. The principal disadavantage of this implementation is the large number of cables required. Since the controller can communicate with tributaries at a rate largely limited by the controller's internal speed a very fast system response time may be realized. This configuration is best suited for dedicated systems such as an edit facility.

A multi-point configuration is one in which one controller and more than one tributary share a common bus. Multi-point has the advantage of reduced cabling costs and ease of reconfiguration. Different tributaries which must be used at different times can be readily connected to or disconnected from the interface bus. This configuration is well suited to a news or production studio. The main disadvantage of multi-point operation is that messages must queue up and be sent serially, resulting in a system response time that is slower than can be realized by the point-to-point configuration.

NETWORK ARCHITECTURE

The SMPTE standards define parts of a local data network. Implementation of this type of network is accomplished most easily using a structured design which groups related services such as control messages, network control and electrical characteristics in independent layers. Each layer performs a specified function within the network in a manner which is transparent to other levels.

The International Standards Organization has developed a reference model which defines seven layers or "levels" for data networks (ISO/TC97/SC16-1977). Local networks, such as the SMPTE control network, typically use five of the seven levels:

The APPLICATIONS level performs specified functions such as sending start/stop signals to a machine and receiving and interpreting status signals from the machine;

The PRESENTATION level is a "virtual machine" which responds to defined data - the "control language" - in a defined manner regardless of the characteristics of the physical machine operated by the applications level;

The SESSION level establishes the logical connection between two presentation levels and provides such services as identifying the characteristics of the virtual machines which have been connected to insure that messages suitable for the controlled device are exchanged.

The NETWORK and TRANSPORTATION levels are not used for typical local networks. They are required, however, when data is exchanged between units connected to different local networks.

The DATA LINK level establishes communication between physical units connected to the network and provides data transfer and error recovery services.

The PHYSICAL level is the electrical/mechanical facilities which form the actual communication channel.

The levels are designed in a manner which permits direct connection betweeen two equivalent levels located in different nodes of the network. Use of this layered approach permits change of one level, such as the control message "language", without impacting the design of other levels.

SMPTE STANDARDS

The initial standards presently specify the physical and data link levels for the SMPTE network.

<u>Physical</u> <u>Link</u> (fig. 2)

The physical link is a four wire, balanced, full duplex (simultaneous transmit and receive) bus. RS-422 transmission standards are used as the base specifications for the bus. Standard LSI transmitter and receiver devices can be used with the bus, minimizing component cost. The following modifications have been made to the RS-422 specifications:

1. Rise times for data have been slowed to the specifications of RS-170A to minimize crosstalk from the control circuits to video and audio circuits in typical installations.

2. Common mode voltages specified are higher than RS-422 providing required protection in typical broadcast plants.

3. Data receivers are required to maintain a marking condition in the absense of signals on the bus to prevent spurious control actions.

Data is transmitted asynchronously at a standard rate of 38.4 kilobits per second. This data rate provides the required system response times for most uses. Higher data rates can be used for special applications but all system devices must initially establish communications at the standard rate. The standard data word permits transmission of eight bit binary information.

System synchronization is accomplished through the use of a special "break" character which is easy to detect and cannot be duplicated by standard data patterns. The use of the break character eliminates the requirement for additional wires in the bus.

Data transmission is accomplished using standard 24 guage shielded pairs. The nominal maximum bus length is specified as 4000 feet, however field tests have shown that reliable operation is possible at cable distances exceeding 15,000 feet.

The standard bus connector is a nine pin "D" subminiature type. This is a rugged connector, well proven in many commercial and military applications.

<u>Data</u> <u>Link</u> (fig. 3)

The data link is the supervisory protocol which specifies the control actions and responses of the bus controller and tributaries. The protocol assures simple, non-ambiguous operation of the network. Significant features of the protocol are:

1. Tributaries enter an idle condition on power-up, when communication is established between the controller and one designated tributary and on encountering specified error conditions. This requirement prevents spurious responses by the tributaries.

2. Tributaries are required to enter a designated state whenever the break synchronization character is received.

3. Tributaries can be polled for status on a rapid basis. Status of up to thirty-two tributaries can be checked in one frame period.

4. Tributaries can be assigned to groups for simultaneous control actions by several devices.

5. Facilities are provided for transmission of standard and non-standard ("user defined") control messages.

<u>Standard</u> <u>Control</u> <u>Messages</u>

The working group is presently developing standards for the session and presentation levels. We have been joined in this effort by representatives of the European Broadcast Union. Since these levels have not as yet been standardized, equipment using the standards for the physical and data link levels transmit control messages via the "user defined" message facilities of the level 2 protocol. This prevents later confusion of the non-standard level 5 and 6 data with standardized messages.

<u>Application</u> <u>Level</u>

The application level is device specific and will vary according to the characteristics of the equipment being controlled.

TYPICAL SYSTEM

The elements of a typical control system, shown in figure 4, illustrate the function of the various architectural levels.

The control panel includes machine access controls, function controls, and function indicators. Discrete signals between the control panel and the control interface are level 7 or applications data.

The control interface, which includes the bus controller function in this simple system, includes five service levels:

Level 7 converts the discrete control panel signals to a common level 6 language for transmission to the controlled machine and converts received level 6 data to discrete signals to operate control panel indicators.

Level 6 is a virtual control panel which has fixed characteristics regardless of the physical properties of the actual control panel.

Level 5 establishes a logical connection between the virtual control panel and a virtual machine. This level determines the type of virtual machine connected, its status and selects appropriate control commands.

Level 2 establishes communication between the control interface and the controlled machine, manages data transmission, and provides recovery from error conditions.

Level 1 hardware provides the physical communication channel, generates and detects system synchronizing signals, and detects basic transmission errors.

The Machine interface connects an existing machine to the SMPTE control network. Five service levels are provided:

Level 7 converts received level 6 data to discrete control signals to operate the machine and converts status signals from the machine to level 6 data for transmission;

Level 6 is a virtual, generic, machine such as a VTR or film island which responds in a fixed manner regardless of the characteristics of the attached physical machine.

Level 5 software connects the virtual machine to a virtual control panel and provides the same general facilities as level 5 software at the control interface.

Levels 2 and 1 provide services identical to those in the control interface.

SUMMARY

Efforts to date by the SMPTE Working Group for Digital Control of Television Equipment have produced the initial standards required to allow users and manufacturers to begin planning and working toward a common control system.

The electrical standard has been approved by the SMPTE and will be published in the near future as a proposed ANSI standard.

The control protocol has received initial approvals and is now being considered by the SMPTE standards committee. It will be published in the near future as a proposed recommended practice.

Initial uses of the standards have been successful and we look forward to the day when we will all be on the same bus together.

SMPTE CONTROL NETWORK

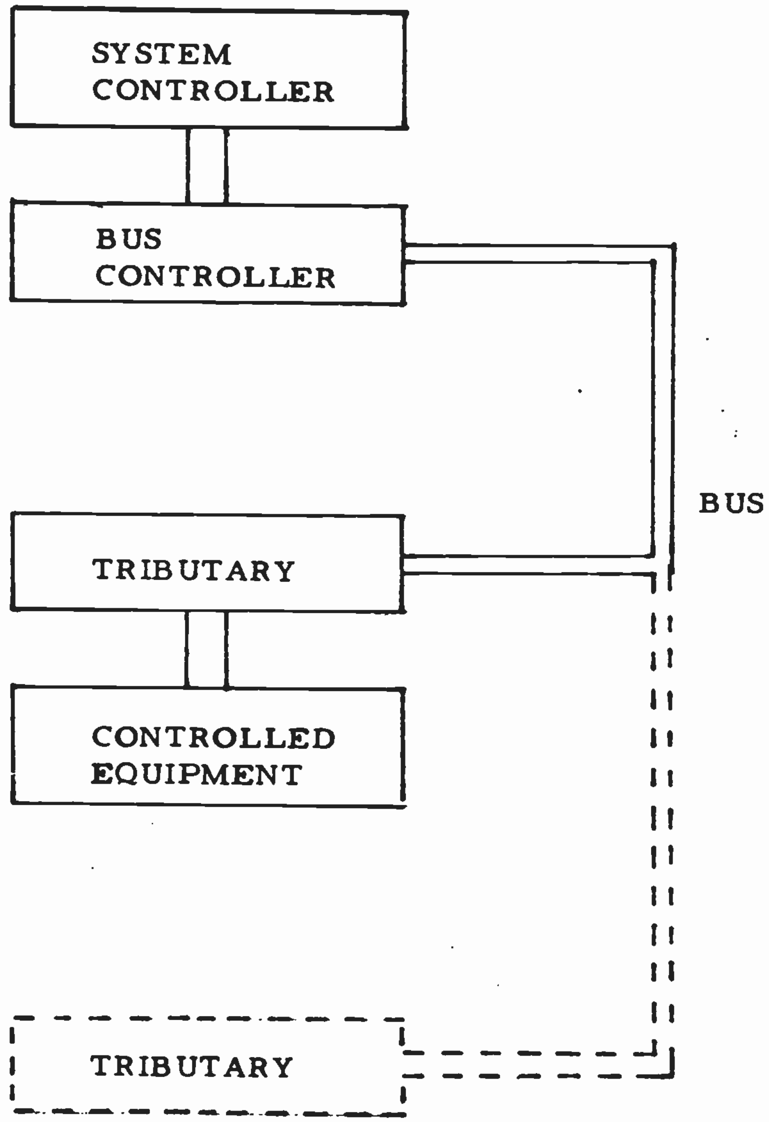
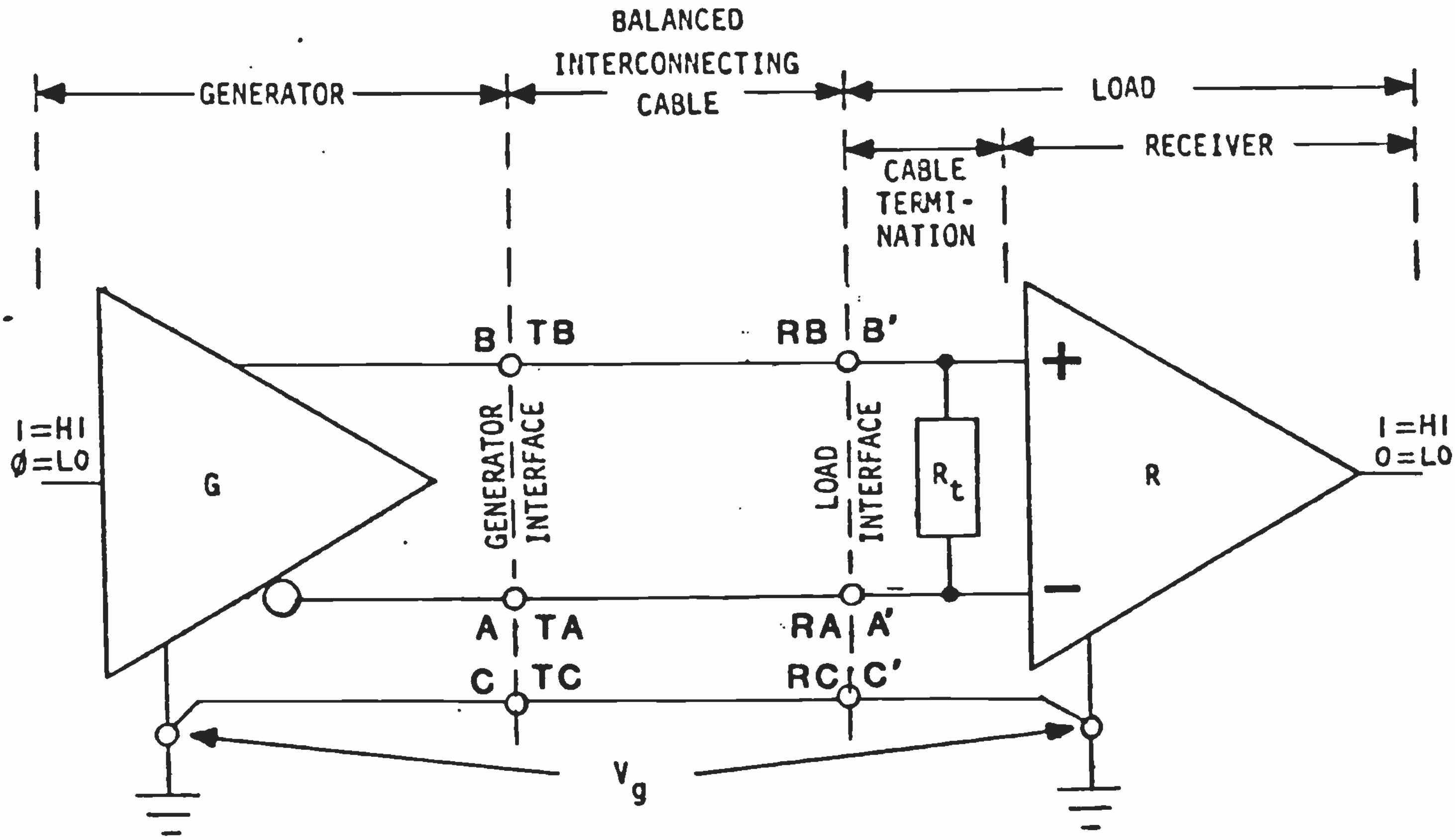


Figure 1



Legend:

R_t = Cable Termination Resistance

V_g = Ground Potential Difference

A,B = Generator Interface Points

A', B' = Load Interface Points

C = Generator Circuit Ground

C' = Load Circuit Ground

Figure 2

Balanced Digital Interface Circuit

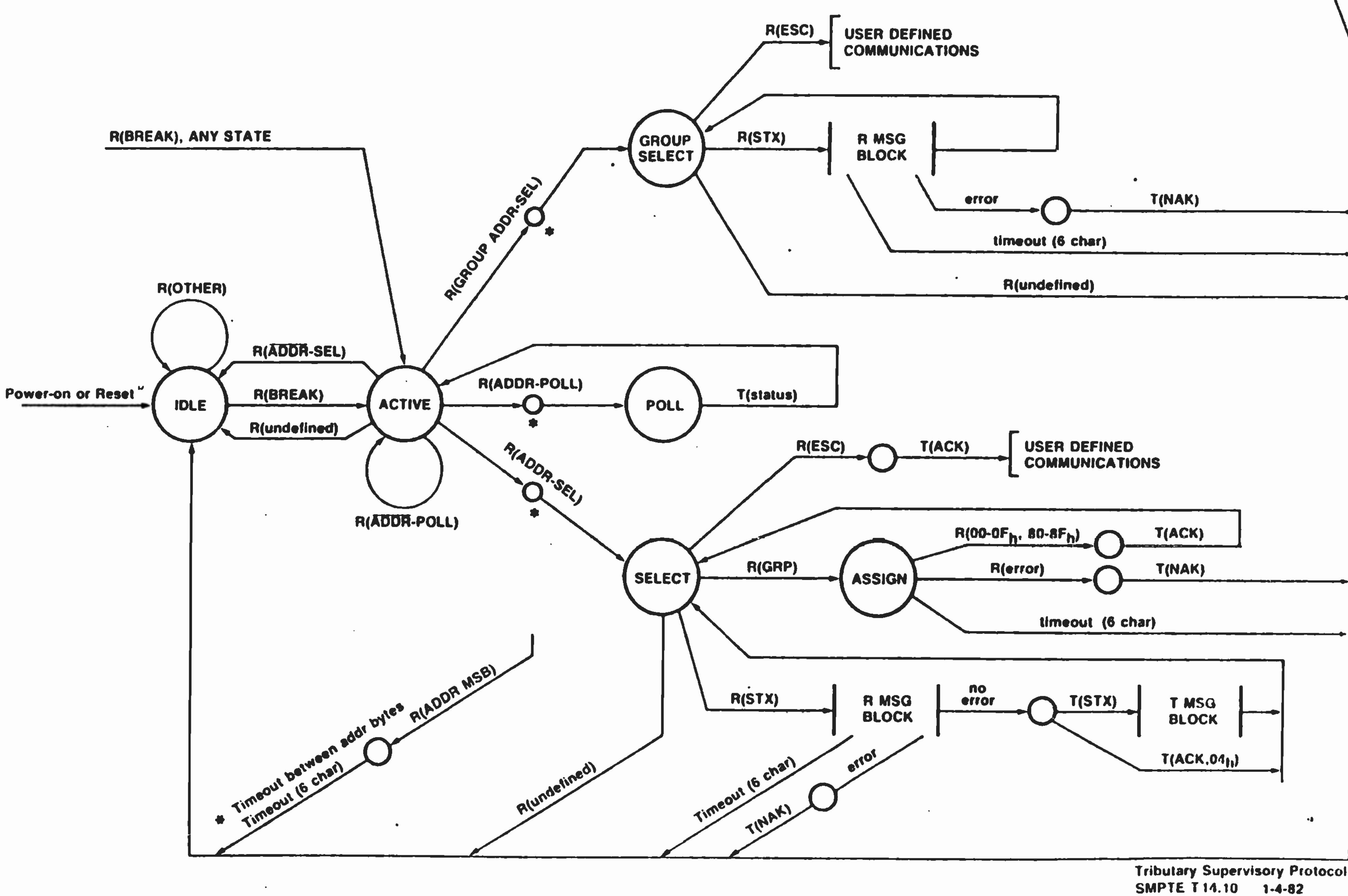

R(BREAK), ANY STATE
R(OTHER)
Power-on or Reset
IDLE
R(ADDR-SEL)
R(BREAK)
R(undefined)
ACTIVE
R(ADDR-POLL)
R(GROUP ADDR-SEL)
GROUP SELECT
R(ESC)
USER DEFINED COMMUNICATIONS
R(STX)
R MSG BLOCK
error
T(NAK)
timeout (6 char)
R(undefined)
R(ADDR-POLL)
POLL
T(status)
R(ADDR-SEL)
SELECT
R(ESC)
T(ACK)
USER DEFINED COMMUNICATIONS
R(GRP)
ASSIGN
R(00-0Fh, 80-8Fh)
T(ACK)
R(error)
T(NAK)
timeout (6 char)
R(STX)
R MSG BLOCK
no error
T(STX)
T MSG BLOCK
T(ACK,04h)
R(ADDR MSB)
Timeout between addr bytes
Timeout (6 char)
R(undefined)
Timeout (6 char)
error
T(NAK)
Tributary Supervisory Protocol
SMPTE T 14.10 1-4-82

Figure 3

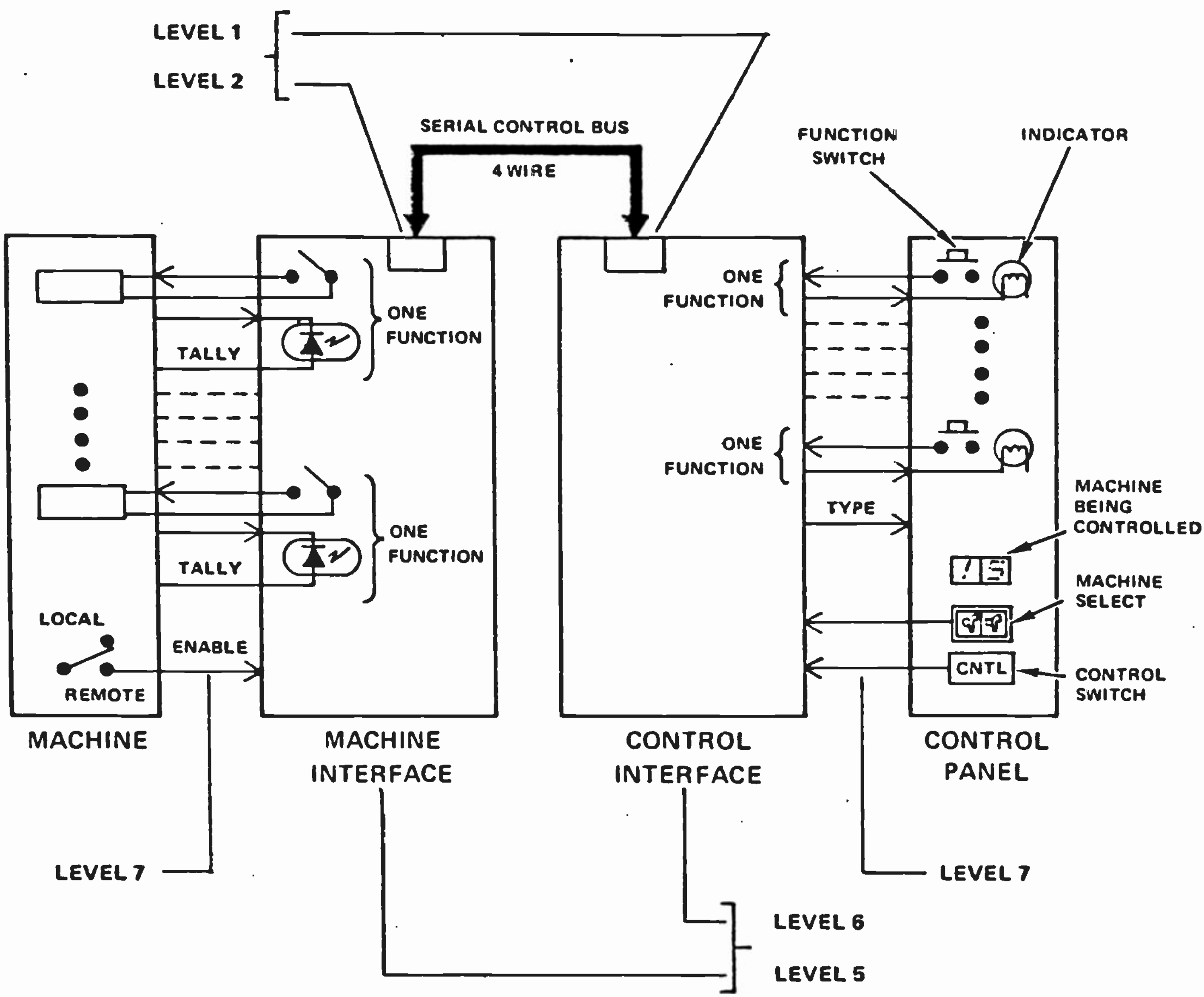


FIGURE 4

BASIC CONTROL SYSTEM

DIGITS AND BEYOND

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In an historic meeting that took place at Geneva in the last two weeks of February 1982, the 155 nations of the ITU, convening the XVth Plenary Assembly of the CCIR, unanimously approved CCIR Recommendation AA/11 leading to a single world standard for digital television studios. The parameters established in that Recommendation are shown in Table 1. The CCIR specification calls for a system based on component rather than composite signals, specifying that those components shall be Y, R-Y, and B-Y. It further states that the specification should allow the evolution of an extensible family of digital codes with luminance-chrominance bandwidth ratios that are simply related one to another. The primary standard to be used as the interface between main studio equipments and to facilitate international program exchange has luminance and color-difference sampling frequencies in the ratio of 4:2:2. The sampling frequency for luminance is 13.5 MHz and that of the color difference signals is 6.75 MHz. Extensions of the 4:2:2 member (which have not yet been standardized) would include, for example, the 4:4:4 member as might be desired for an RGB standard, the 4:1:1, as might be an acceptable quality level for common carrier transmission and perhaps even lower levels that might be established in the future for electronic news gathering applications.

A common sampling frequency for the component signals in both 525- and 625-line systems will permit maximum commonality of digital equipments designed for those two markets. The choice of 13.5 MHz, as most of you know, is a unique frequency that permits an exact integral number of samples in the scan line in both 525- and 625-line systems, i.e., 858 samples per line in 525- and 864 samples per line in 625-line systems. These sampling frequencies will permit a luminance base bandwidth of greater than 5.5 MHz and chrominance bandwidth of 2.75 MHz, more than adequate for post-production processes such as picture manipulation and chroma key.

Another very important common parameter that has been set in this world agreement is a common number of samples in the active portion of the line in both 525- and 625-systems. This unique possibility derives from the fact that the horizontal blanking interval for 625-line systems is slightly greater than

TABLE 1

CCIR RECOMMENDATION AA/11

WORLDWIDE DIGITAL PARAMETER VALUES

	525 Line 60 f/s Systems	625 Line 50 f/s Systems
1. Signals Coded	Y, R-Y, B-Y 4:2:2 Extensible Family	
2. Sampling Structure	Orthogonal with C (R-Y, B-Y) Cosited with Odd Y	
3. Sampling Frequency	Y: 13.5 MHz C: 6,75, 6.75 MHz	
4. Number of Samples per Total Line	Y: 858 C: 429, 429	Y: 864 C: 432, 432
5. Number of Samples per Digital Active Line	Y: 720 C: 360, 360	
6. Form of Coding	Uniformly-quantized PCM 8 bits/sample for Y and C	
7. Video Signal/Quantization Level Correspondence	Y: black at level 16 white at level 235 C: 224 levels with zero at level 128	

that of 525-line systems permitting the difference in the number of samples of the total line to be absorbed in the horizontal blanking interval. The common number agreed to by the CCIR is 720 samples in the active line. At 13.5 MHz, 720 active samples implies 53.333 microseconds which is slightly longer than the active picture portion in both 525- and 625-line systems. However, the use of a number slightly larger than the average actual avoids the possibility of cropping edges of the picture during the digitization process. It also is sufficiently large to allow for the tolerances in the horizontal blanking intervals in both the systems. A common specification for the active line duration permits line-oriented processes such as line-stores, vertical filters, error concealment and line-scanning sensors to be identical in digital equipments designed for world markets.

The sixth specification is that in the digitization of the component signals, eight bits per sample shall be employed. That is, there are a possible 256 quantization levels in both luminance and color difference signals. However, to allow for overshoot and avoid clipping, only 220-224 of these quantum levels are assigned to the dynamic range between black and white levels. Specifically, for the luminance signal, black level is defined to be at level 16 and white level at level 235, giving 16 levels of footroom and 20 levels of headroom taking into consideration the fact that black level clamping will reduce the requirement for footroom. For color-difference signals, 16 levels are reserved at both ends for head- and footroom.

There are a number of additional specifications that must be developed before the final definition of the world digital standard is completed. Studies will be performed by the administrations of the ITU over the next four-year study period of the CCIR. Included in these studies will be a specification for the chrominance bandwidth and, especially, the rate of roll-off of filters associated with the color difference signals. Also, since it is expected that in the digital domain, one will not need to have the full overhead associated with the analog horizontal and vertical blanking intervals, there is a need for a new specification associated with digital blanking or digital synchronization. This would be especially true in the case of the standard format of the digital VTR wherein horizontal blanking can be compressed and ultimately expanded in a digital line store. Probably the most important and most difficult specification to be established over the next few years is the digital format for the first generation component digital VTRs. Finally, additional specifications are required on parallel and/or serial formats at the interface of equipments for interconnection and local distribution.

I am here to report to you today on the activities of the SMPTE in the digital as well as other television areas. As the new Chairman of the *Committee on New Technology* for the SMPTE, I would like to congratulate the outgoing Chairman, Bob Hopkins, for the success of his digital committees and to say just a few words about the future activities of the several study groups and working groups that are currently active under the auspices of the *New Technology Committee*.

The *Committee on New Technology* was formed by the SMPTE in 1976 on the recognition that special attention must be given to the budding new technologies to determine that critical time during development at which standardization activities should be commenced. The establishment of a digital video interface standard on a worldwide basis prior to the extensive introduction by manufacturers of equipments utilizing the new digital technology has proved to be a much easier task than the negotiation of a compromise standard from two or more competing noncompatible equipments. Hopefully, we can continue this trend.

At the time of its formation, the *Committee on New Technology* assumed cognizance for the *Study Group on Digital Television* which had been established in 1974 and is chaired by Charles Ginsberg of Ampex. It was this study group that led to the formation of a *Working Group on Digital Video Standards* to commence work on a digital video interface stanJard for the future all-digital studio. That working group was initially chaired by Robert Hopkins of RCA, and in 1980 was transferred to the very able leadership of Kenneth Davies of the CBC. Work of Ken Davies' committee over the past two years has led directly to the development of specifications for the world digital standard. When it became clear that there was an advantage in the use of component rather than composite signals in future digital systems and also when it was recognized that the possibility existed for cooperative agreements between the 525- and the 625-line countries of the world, a special *Task Force on Component-Coded Television* was formed in 1980 under the chairmanship of Frank Davidoff (then of CBS) to seek out the possibility of a compatible world standard. It was the leadership of Mr. Davidoff's Task Force that ultimately led to the SMPTE's cooperative efforts with the European Broadcasting Union on agreeing to an approach for a single world standard. That agreement was reached at an important joint meeting between the SMPTE Task Force and Working Party V of the EBU at Brussels in March of 1981. From that historic Brussels meeting, accelerated activities by the SMPTE in the U.S. and Canada, by the EBU in Europe, and by an AdHoc Group in Japan led finally to the proposals for a CCIR recommendation at the CCIR Final Meeting in Geneva in September/October of 1981. There the CCIR hammered out the detailed parameter specifications of CCIR Recommendation AA/11 and these were then adopted and approved by the CCIR XVth Plenary Assembly in late February of 1982. Frank Davidoff's Task Force activities have now been terminated but he is continuing to provide active leadership as the Chairman of a *Subgroup on Digital Video Implementation* with the objective of planning the evolution of digital technology in television studios over the next several years to identify potential interface problems for which specific studies and additional standards specifications must be established. That group is making good progress toward their objective.

The *Study Group on Digital Television Tape Recording*, chaired by Bill Connolly of CBS, has compiled and published the results from a user's questionnaire that will be the basis for a specification on a digital tape format. A major activity over the past year has been a study of sampling frequencies for digital audio as they impact the DVTR. The conclusion of that study was that only two frequencies, 60 kHz and 48 kHz, could be supported by the SMPTE as a world standard sampling frequency for television, motion picture, and professional audio applications.

Let me turn now to the *Study Group on Video Disc* chaired by Bob Paulson. The objectives of that working group are to study the status of technology and to assimilate the formats used in various disc systems. A survey of the three existing consumer video disc systems, the CED, the VHD, and the LOR and LOT optical systems, was published by Mike Doyle in the February 1982 SMPTE Journal. The committee is now turning its attention to a study of the applications for professional use of optical video disc technology particularly with the re-recordable media that have been developed over the past two years. Obviously, some of the potential applications for optical disc technology in teleproduction editing and short-segment broadcasting will be investigated by this committee.

The next committee is the *Working Group on Image and Sound Presentation* chaired by Mr. Ed Efron. This committee is newly formed with the extremely important long-term objective of establishing standards and methods of measurement of the presentation quality of both television images and their accompanying program sound.

The last committee I would like to discuss today is the *Study Group on High Definition Television* chaired by Mr. Donald Fink. HDTV has seen a tremendous surge of activity by television and motion picture interests over the past year. The objectives of this study group are to survey the status of technology and to determine the requirements of potential applications for high definition technology. This committee which has been operative since 1977 had completed a final report in 1979 but was reconvened in 1981 because of the increased intensity of activity in HDTV, especially in Japan. Over the past year, additional progress has been made in studying the possible requirements for HDTV systems for not only cinematography or post-production applications but also for a new higher fidelity consumer television service. The committee has now reached the point where it must subdivide into smaller groups to tackle specific problem areas. Four subgroups are being formed in the areas of *Production, Psychophysical Studies, Transmission and Distribution,* and *Component Hardware.* These groups are in the process of getting organized and will be expected to have their first rounds of meetings in late summer to fall of this year. By the subdivision of this study group into the four subgroups, an attempt is being made to partition the study by skill and experience of the participating members of the study group. We would hope that after about a year of studies in these four special areas, a new final report will be issued by the study group and we are looking to an earlier recommendation relative to the formation of a working group to commence discussions on standards. There is a great hope here for the establishment of single world standard for high definition television that does not differ even in detail between the 525-line and the 625-line countries of the world.

THE BETACAM SYSTEM

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INTRODUCTION

Since the Camera/VTR combined packages have been introduced by several manufactures at the last NAB, the users' desire toward the possible format standardization has grown so rapidly which has resulted the formulation of SMPTE and EBU working groups on the 1/2" VTR format. It has been mutually recognized and agreed throughout the SMPTE meetings that the Camera/VTR combined package is the common goal for the ENG and some of the EFP operations. Although the U-matic has contributed tremendously to the growth and expansion of ENG and it will surely continue to serve the Broadcasters' needs, today we are starting the 2nd generation of ENG with the entirely new format of Camera/VTR combined package.

The paper describes the criteria and development of the 2nd generation ENG equipment reviewing the previous paper 'Betacam - A VTR in Camera" presented at the 123rd SMPTE TECHNICAL CONFERENCE, and urges both users and manufactures to seriously re-examine all the possibilities before starting this very important new era of ENG.

CRITERIA OF THE SECOND GENERATION ENG

Based on users' desires and the experience that the manufactures have gained to date, the following three factors have been recognized as the key criteria for the 2nd generation ENG Operation.

1) Film style mobility
2) Improved picture quality
3) System implications with respect to economy situation

MATERIALIZATION OF THE CRITERIA

In order to meet the first and the most important criterion, the 'Film Style Mobility', the camera and VTR must be combined into one package. Furthermore, the package must be as small and light as possible. Along with the technological improvements, the camera portion should become better in quality, smaller and lighter as far as it is possible.

The VTR portion, however, must be very carefully deliberated in the beginning, considering present and future technology in conjunction with the joint usage of existing hardware. The VTR must be smoothly integrated into the existing system without excessive capital investment. In addition to the common requirements of the ENG operation, the following design requirements are considered to be essential for such a VTR;

1. Small cassette size
2. Adequately small drum diameter
3. Small circuitry
4. High picture/audio quality

CHOICE OF CASSETTE SIZE

Selection of cassette size creates a contradiction between 'High picture/audio quality' and 'Small and light weight mechanism'. Although one inch or wider tape is desirable to achieve the high picture quality, it is almost impossible to develop a small and light weight mechanism.

Although 1/4 inch or narrower tape is ideal as far as the size is concerned, it is extremely difficult and impractical to reserve a sufficient video track width to meet the required level of picture quality after the assignment of necessary Two Audio/Time Code/CTL Tracks. It is technically possible to develop the tape width anywhere between 1/4 inch and one inch, however one has to realize that the cassette tape for ENG must be reasonably inexpensive and easily accessible almost anywhere in the world, due to the nature of ENG operation. Therefore the selection of a particular cassette size has been narrowed down to either 1/2 inch or 3/4 inch cassette tape.

Table 1 shows the comparison of the cassette size of 1/2 inch and 3/4 inch tapes commercially popular throughout the world.

As shown in Figure 1, it is not too much to say that the size of the VTR mechanism is determined by the projected area of the cassette, drum diameter and the loading mechanism. The most important factor is the projected area of the cassette.

Therefore a decision has been made to use the Beta cassette, the projected area of which is recognized to be the smallest among all the commercially and widely available video cassettes.

CHOICE OF COLOR RECORDING METHOD

As mentioned earlier, a contradiction exists between the 'Small and light weight mechanism' and the 'High picture/audio quality'. Since the Beta cassette has been chosen, a big challenge has been given toward the development of an innovative color recording method in order to achieve the high quality picture. An objective has been set to produce the superior picture quality to the U-matic with Beta cassette and the drum diameter of commercially available Betamax recorder.

The criteria for the new color recording method should include the following:

1. Superior picture quality to the U-matic

2. Minimum number of heads to minimize the volume of circuitry and rotary transformers, and to incorporate the future possibility such as dynamic tracking capability.
3. Ease of interface with existing external equipment such as switchers, VTR's, editing facilities, and other communication channels.

There are number of color recording techniques to be considered such as direct, 2-component and 3-component methods. It has been experienced at the laboratory that 1/2" tape VTR can reproduce extremely high quality of the picture equal or better than that of 1" type C VTR.

As mentioned in the beginning of this paper, the industry is now at the start point of the new generation ENG and thus anything can be done without being tied up with the previously fixed recording formats. Therefore, it is strongly required to examine every possibility of the technology based on the practical basis.

The examination should include electronics packaging technology to meet the basic criterion-the 'Film Style Mobility' and the power consumption, besides the above criteria. Deep consideration should also be given toward the multi-generation quality of each method including the possible repetition of decode/encode process within the required communication loop.

CONCLUSION ----- THE DEVELOPMENT OF BETACAM SYSTEM

With the result of the consolidation of user's desire, manufactures' past experience and the new developments, the Betacam system has been materialized as:

BVW (VTR in camera) ---Shown in Figs. 2 & 3

and

BVW-10 (Video cassette player) ---Shown in Fig. 4

1. Basic performance of BVW-1

In order to minimize the weight, size and power consumption, the basic concept of High Band Trinicon camera with single Saticon™ tube has been initially chosen as the camera part. In addition to this basic scheme, three-tube configuration with 2/3 inch tubes is also considered as the family.

1) Tape speed : 11.865 cm/s
2) Video writing speed : 6.90 m/s
3) Superior picture/audio quality to the U-matic
4) Max. 20-minute recording time with new HG-20 or L-500 Beta cassette tape.
5) Small, light weight and low power consumption
   The achievement of 7.9 kg in total weight including lens, view finder, battery, cassette and microphone, has been made.
6) Complete wireless system
   For better reliability and operational flexibility, no cable is used on the Betacam. Furthermore, a wireless mic receiver may be attached the back of the VTR portion for the complete freedom from the cable.
7) Camera portion and VTR portion can be easily separated for emergency replacement of the component.
8) With convenient and compact adaptors, each camera and VTR may be operated together with existing hardware.
9) Both video and audio confidence heads are incorporated.

10. Built-in Time code generator provides SMPTE/EBU time code into the assigned track apart from two audio tracks.
11. Complete warning and status indicator on the view finder.

2. Basic performance of BVW-10

The primary design criterion of BVW-10 is to provide a smooth interface with the existing ENG/EFP equipment.

1) Superior video/audio quality to the U-matic
2) Small size, front cassette loading and 19-inch rack mountable
3) Complete Jog and picture search capability
   ±3 times normal speed: Color picture Jog and search
   Fast forward and rewind: B/W picture search

3. System implications

Everytime prior to the new system introduction, manufacturers must consider the system interface to be as smooth as possible with the existing equipment in which users have already invested. The changes, if required, must be minimal. Some typical examples of the ENG system interface are shown in Figures 5, 6, 7, and 8.

ACKNOWLEDGEMENT

The authors would like to extend their special thanks to the NHK (Japan Broadcasting Corp.) for the important advice and support given especially in the development of High Band Trinicon Camera with single Saticon tube, and also to the authors' colleagues involved in the Betacam system development.

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By M. TAKANO and I. Segawa
SONY Corporation,
presented at the 123rd SMPTE
TECHNICAL CONFERENCE IN Oct., 1981.

		WIDTH (mm)	HEIGHT (mm)	DEPTH (mm)	VOLUME (mm^3)	PROJECTED (mm^2) AREA
1/2 INCH	BETAMAX	156	96	25	374	150
3/4 INCH	KCS-20	186	123	32	732	229
	KCA-60	221	140	32	990	309

TABLE 1 CASSETTE SIZE COMPARISON

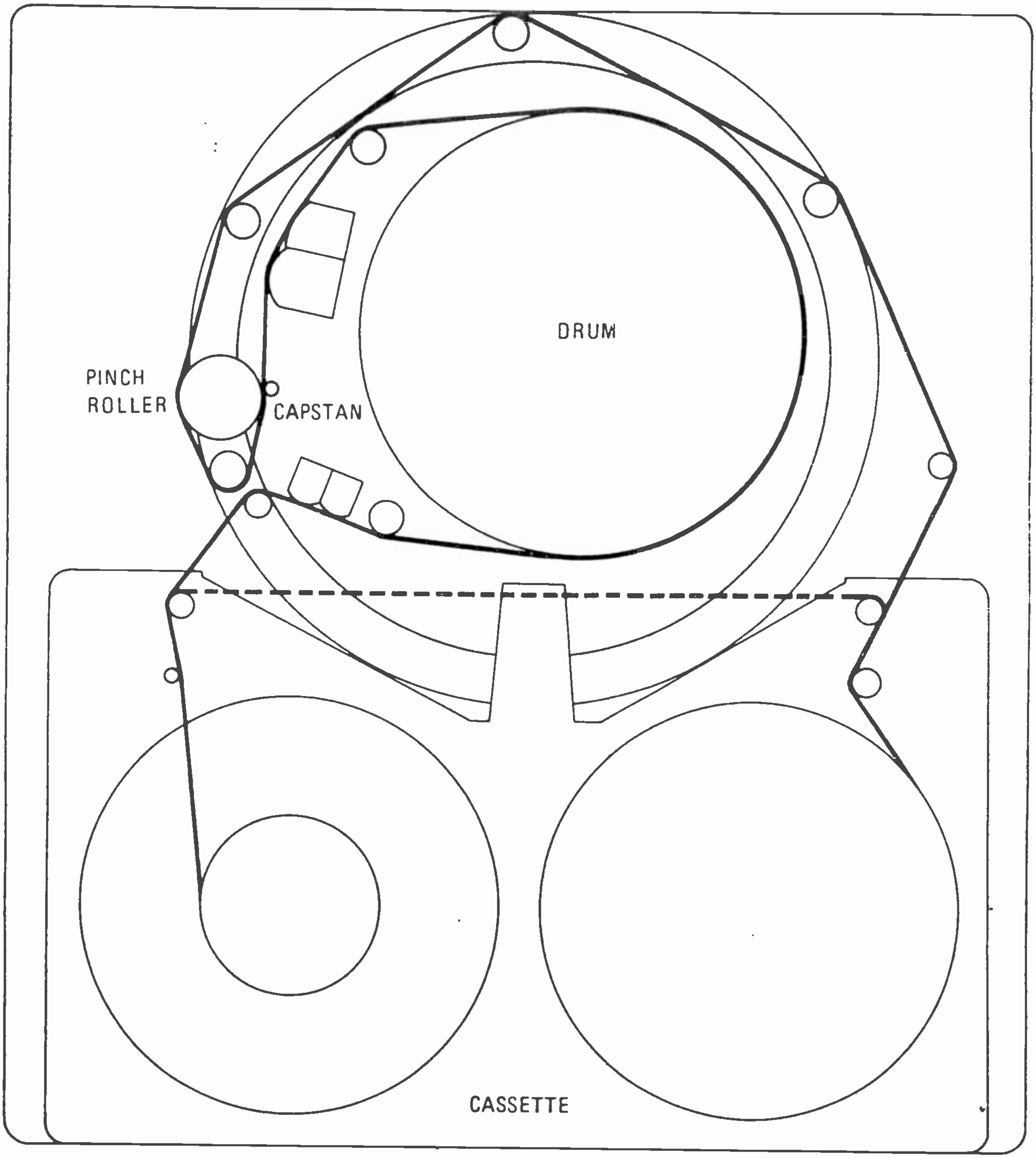

Fig. 1. RELATIONSHIP OF CASSETTE, DRUM, LOADING MECHANISM AND THE SIZE OF VTR MECHANISM

Fig. 2   BVW-1

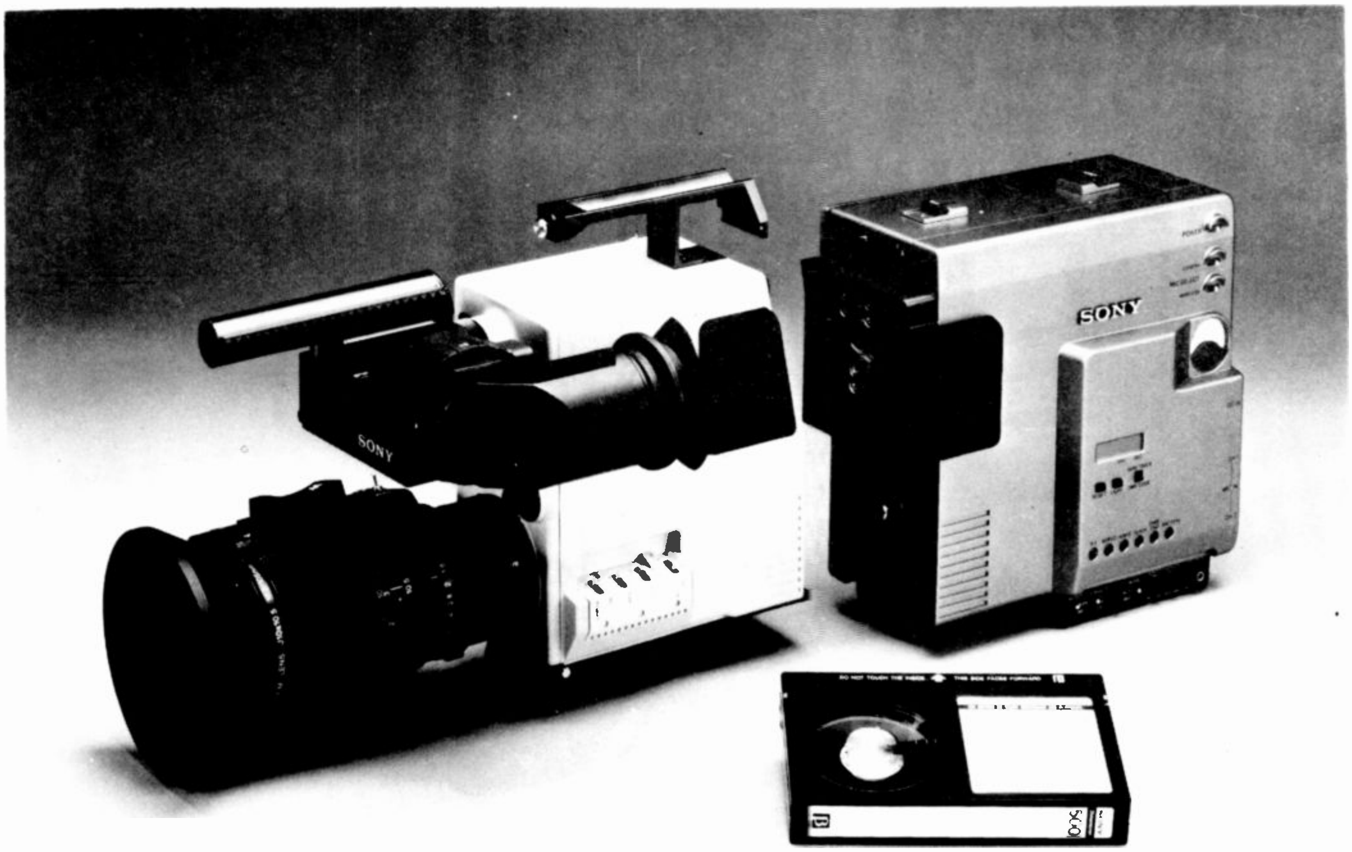

Fig. 3 Separated BVW-1


SONY
POWER
BETACAM
VIDEOCASSETTE PLAYER BVW-10
AUDIO MONITOR
REMOTE
HOURS
MINUTES
SECONDS
FRAMES

Fig. 4 BVW-10

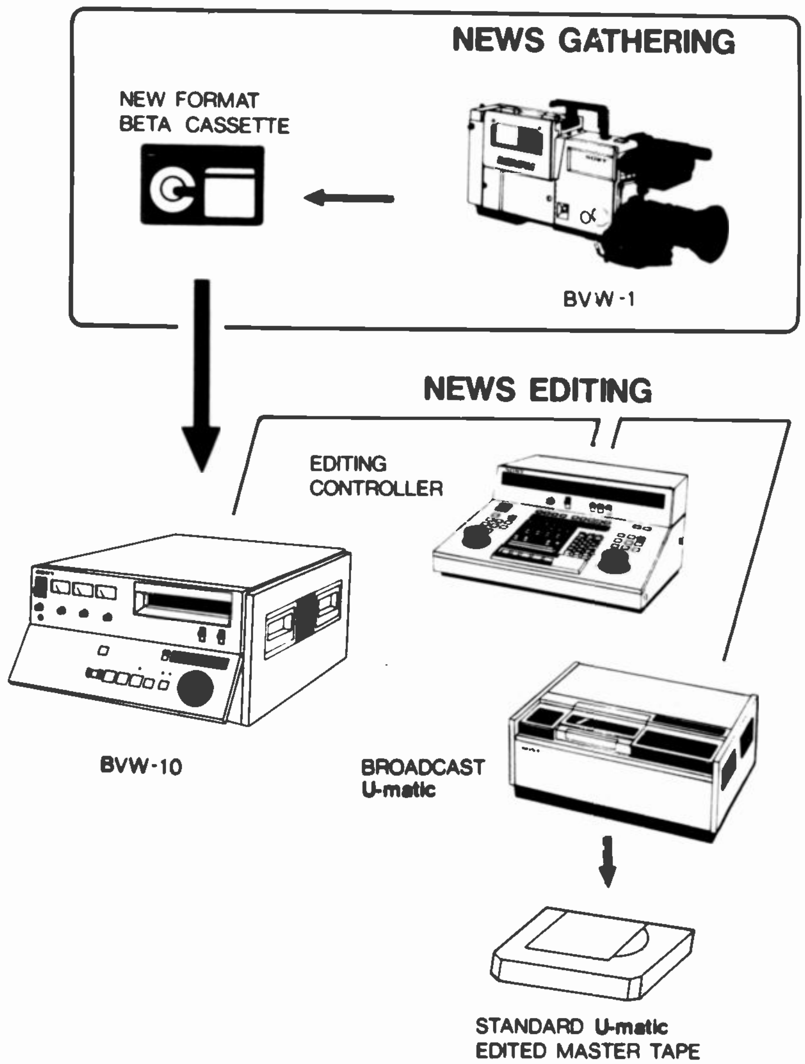

Fig. 5 ENG Operation Example 1

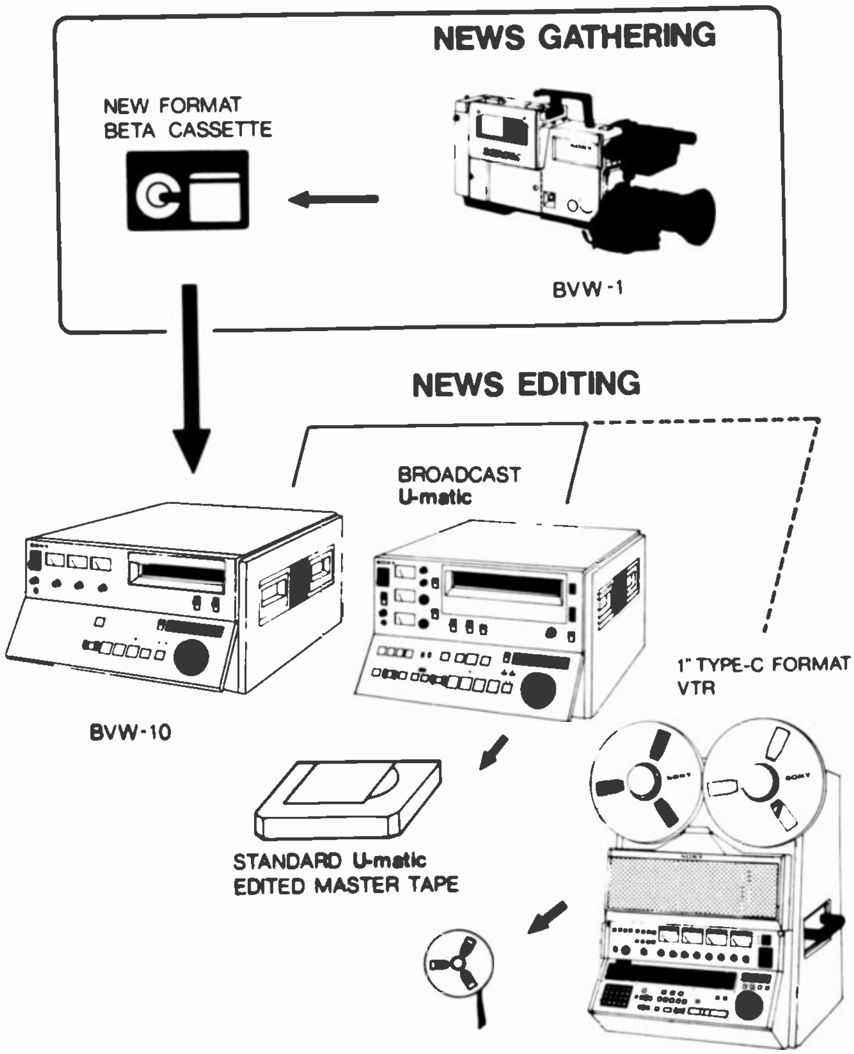

Fig. 6 ENG Operation Example 2

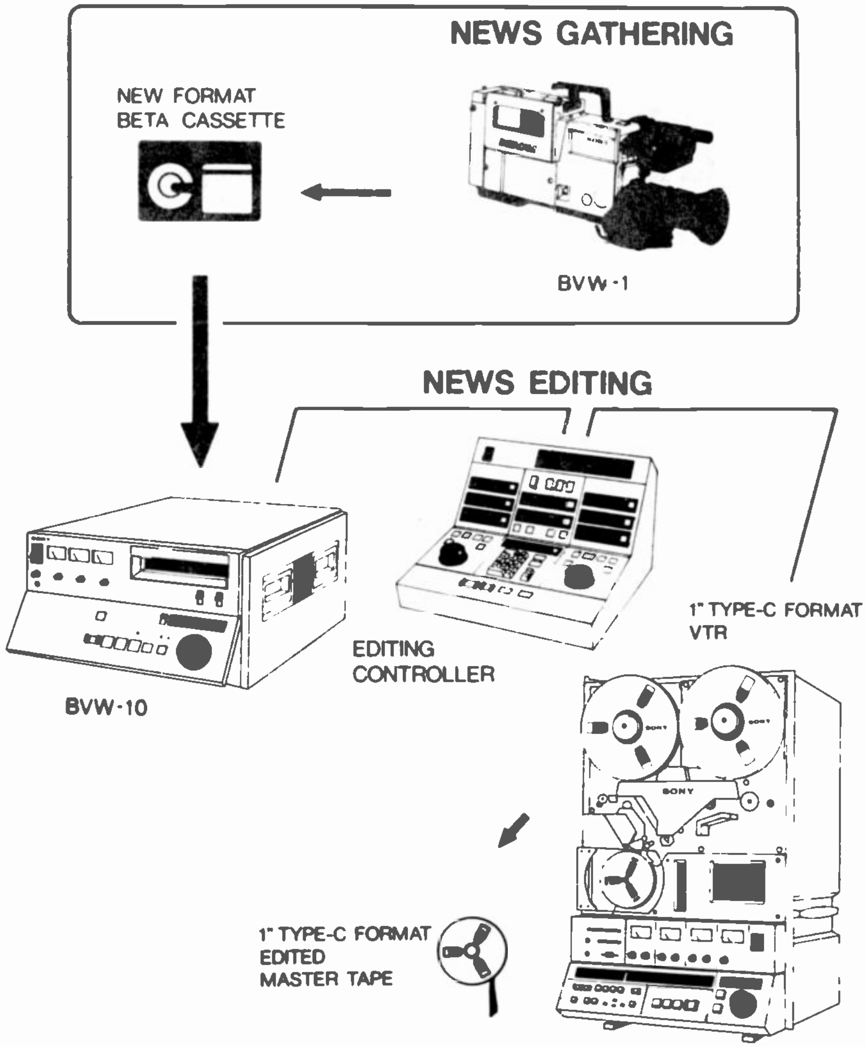

Fig. 7 ENG Operation Example 3

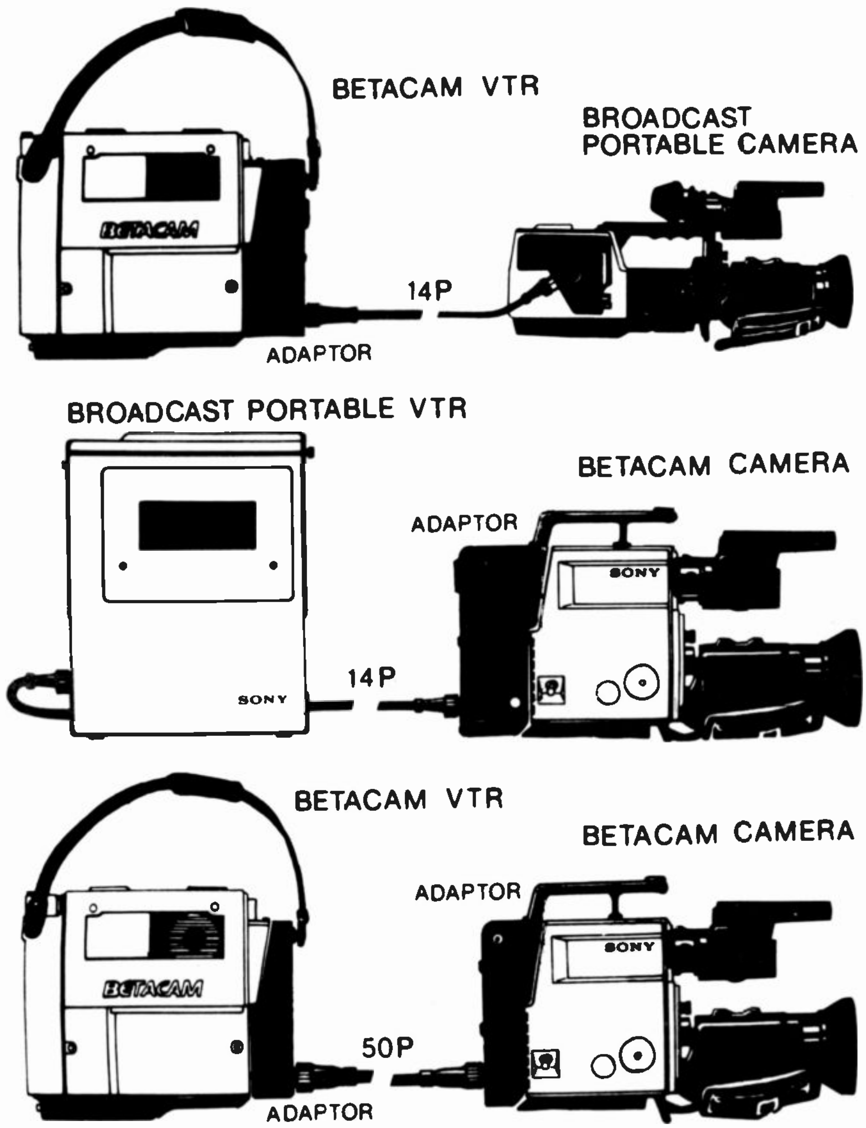

Fig. 8 ENG Operation Example 4

HIGH DEFINITION TELEVISION - AN OVERVIEW

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All over the world and especially in Japan, TV engineers are working on experimental high definition TV systems. Results to date have shown that extraordinarily good pictures can be obtained even with today's technology much less with the developments just emerging from the laboratory. Therefore, the primary problem facing the industry is how to deliver these pictures to the home with a system which is both financially viable and also sufficiently economical of spectrum space to be acceptable by the licensing authorities throughout the world. A second but almost equally important problem is how to achieve a worldwide standard which can support a hierarchy of compatible substandards which can be used for production, post production, home and theatre display at various levels of quality. Without such a unified standard it is likely that competition between various proposed systems, each with its own strong advocates and quite probably each of more or less the same picture quality but incompatible with each other will so fragment the market that no one will succeed. Quad sound came to such a fate. Inasmuch as I am against government intervention in our lives there are some places where it is essential. Driving on the right side of the road is one example. Broadcast High Definition TV, I believe, is another. Let the competition lie within the design of the equipment not the standard used to transmit it to the home.

Now is probably a good time for me to define just what I mean by a high definition picture. I mean a spectacular picture - one that grabs the viewer like the original Cinerama did when first seen by audiences accustomed to small theatre screens - poor sound and fuzzy pictures. I don't mean just a better, sharper 525 line picture - after all most of the American public, myself included, could at least double our resolution merely by putting up a proper antenna and turning down the room lights; since we don't, it's obvious that small improvements won't be enough to launch a new, very expensive service. The early work of John Lowry at Image Transform showed that 655 lines - 24 frames could almost equal 35 mm - more recently Ken Holland of Compact Video has improved the system and has shown, to me at least, that with frame store comb filters the chroma information can be included in the luminance without any deleterious effects, i.e. no subcarrier dots visible - no cross color - no diagonal resolution loss from the filtering process. Using this work as a base line we can scale up to my spectacular picture - first we need more than 24 frames per second. The motion capability of the movies is not good enough. Even 30 frames, 60 field is just barely good enough but since we don't want to flagrantly waste bandwidth, I feel we can settle on 30 frames/second, possibly with progressive scanning in the camera since we can interlace later in a frame store. Next we must increase the aspect ratio to give a wide screen effect - perhaps 2/1 and third, we want to increase the sharpness to even better than 35 mm. Calculate all this out as Dr. Fujeo has done and gets 20-25 MHz of video bandwidth needed for a spectacular picture. If we reduce or eliminate horizontal and vertical blanking - requiring a line or two of storage in the receiver, this can be reduced to 15-20 MHz - at best. What to most of us would appear to be spectacular pictures require spectacular bandwidths and spectacular bandwidths require unrealistic enormous transponder power if a DBS satellite is to be used. Dr. Shannon gave us the rules to calculate this in the late 40's and nature hasn't changed. If we have a video bandwidth of 4 or 5 times the present bandwidth, we need 4 or 5 times the satellite power to cover the same ground area. Such powers are just possibly becoming feasible for the transponder but they may exceed the practical solar cell battery storage capacity of the satelliteif several such bandwidths are used on a single bird, so a new and expensive satellite design may be required if reasonably sized dishes are to be put on the rooftops of America. However, the broadcaster does not have this problem, he can put up a huge dish to catch the signal.

Now FM will almost certainly be used for transmission - quite possibly vestigial sideband FM such as is used in video tape recorders to keep within a channel bandwidth of 25-30 MHz which is all the licensing authorities are likely to give us. This bandwidth does not allow a large enough deviation ratio to achieve much noise improvement so even more transponder power is needed - but so what - this is straightforward engineering and does not require invention - just engineering development which can surely be done just as quickly as someone is willing to pay for it.

You notice I haven't yet mentioned any exotic bandwidth reduction methods. Well, I had the good fortune to start my career under Major Armstrong's tutelage while he was introducing FM to the world. Many of his opponents felt it was wasteful to devote 200 KC channels to 15 KC of audio particularly when the public had not demonstrated any great desire for high fidelity. Those of us in his lab, therefore, proposed all sorts of dynamic commanding systems using subcarrier controls so that we could get FM noise immunity within a reduced 30 KC bandwidth the same as High Fidelity AM. Major Armstrong had the wisdom to say no let's first get the bandwidth allocated by the government to do the job correctly then later we can put in the tricks and make it even better. This wisdom is equally true for High Definition TV - let's allocate the 20-30 MHz per channel to do the job right. In the future we can put in improvements-so many of which have not yet been invented but let's get started.

The standards, must, however, allow for Quasi compatibility with present day broadcast standards since much of what is produced for the new system will also be transmitted at 525 or 625 lines. This would seem to be an easy problem for present day digital standard converters but it is important consideration and should not be neglected, obviously convenient line numbers would be twice 525 i.e. 1050 line or twice 625 = 1250 lines. This is an oversimplication since we will not send vertical blanking in a new system. Our Japanese colleagues have shown themselves to be diplomats by choosing 1125 lines for their tests - right in the middle of these numbers - and a number that both the Americans and Europeans should easily be able to compromise on without the long drawn out battle which led to the 13.5 MHz digital sampling standard.

I have not yet quite eliminated digital transmission for high definition TV from my thinking because it seems just barely possible although without high bandwidth compression schemes the numbers seem too enormous to contemplate. For example, above Nyquist sampling of a base band video signal of 20 MHz requires that the sampling be done at something better than 40 MHz and with 8 bits that results in a bit stream of 320 Megahertz/second. A just possible digital bandwidth compression scheme of 4/1 would get that down to 80 Megabits. If we allow a conservative 1/2 cycle of bandwidth per bit it would require 40 Megahertz of channel capacity - greater than we probably will be permitted - but if we are less conservative and allow only .25 cycle of bandwidth per bit - something the facsimile people do all the time over the telephone line, i.e. they do 9600 bits per second over a 2400 cycle noisy line. Such transmission usually involves multi-phase, multi-amplitude coding and if extended to our postulated system would allow a 20 Megahertz transponder channel with a few megabits extra for several stereo sound channels, teletext, protection, etc. Unfortunately, although such a digital transmission system would probably work, it would require de-scrambling the bandwidth compression at the receiver which would burden each receiver with extra - probably expensive, even with VLSI chips signal processing circuitry.

My vote is, therefore, still for FM since it is simple and straightforward. We don't have to wait for new chip developments, we can start immediately and again according to Shannon it can give us the same signal to noise ratio for the same channel bandwidth and power. Fiber optic or video disc delivery systems will chose some other method of encoding but their output could readily modulate an FM oscillator for entrance into any receiver.

Well, we have looked a bit at the transmission so let's move to the source - the camera and studio equipment. Here our work is easy - the tube designers who have given us Plumbicons and Saticons have shown they can give us sensors with the necessary resolution with margin to spare. 20 MHz, even 60 MHz, video amplifiers are no problem but the preamplifier is. Here again the tube designers have helped with low capacity, pin through the faceplate, targets, but nature is against us on sensitivity. The phototargets already have a quantum efficiency close to the theoretical limit of 100% so we can't hope for much improvement in that area. The camera designer can help by putting most of the light into a green or Y modified green tube and putting minified images into the red and blue tubes at the cost of a very slight reduction in color resolution. This will help but it won't completely solve the problem. Unless someone comes up with a better preamplified approach - the ones we now have are also nearing theoretical limits - we shall probably have to live with a bit less sensitivity. I estimate that 200 foot candles would be plenty for normal studio scenes and 30-50 for low light situations. I believe the CBS experience bears this out.

The signal distribution within the studio will be nonencoded, i.e. in R, B, G or Y R-Y B-Y format. Most likely on three cables. Now, 20 years ago this would have been unthinkable but modern IC amplifiers are so stable that it is no longer a problem. What little drift remains - in this age when even home audio cassette recorders are setup by microprocessors - could easily be taken out at the input of the switcher by a microprocessor looking at some form of AGC pulse during the blanking interval. This is straightforward engineering - if someone places the order any number of suppliers will build the equipment. The switcher itself will be more complex and expensive but will be very straightforward and will produce far better results than today's switchers because it will be using component signals and will not have dot crawl problems on keys and circular wipes, etc.

The tape recorder at first seems like a major problem - but even today commercial recorders can be bought with 12 MHz bandwidth and military machines have been built with 30 MHz bandwidth. You may have heard about the NHK and Sony machines with, I believe, one 20 and two 10 MHz channels so they have proven that the job can be done - obviously with editing and all other features we now expect from a video recorder can be added. It's easy for me to say this since I don't have to do it - by straightforward engineering.

We have now progressed to the point where we see how to generate and transmit a High Definition picture - but how do we display it. For the Hollywood producers using the systems to make movies, the answer is easy - put in on film.

The laser color film recorder originally developed at CBS Labs and now refined by NHK, and the Electron beam recorder used by Image Transform can easily do the job. Each has its advantages but both have demonstrated movie theatre quality. The movie producers can literally save millions by going to electronic production so I think it's safe to say they will be first to buy equipment. They don't have to wait for a new display. The rest of us do. High resolution wide aspect pictures cry out for a large screen. A person can't use the resolution on a small screen unless he sits close to it and this is unnatural for scenes involving peripheral visions. We want to focus on infinity - a large enough CRT to overcome this whould dominate most people's living room. Surely a huge CRT might get the service started - probably in barrooms showing sports the way TV started but I don't believe it will be viable in the home. Projection TV can do the job - but an Eidopher is $400,000 too expensive. The GE Telavia approach pioneered by Bill Glen and Bill Good could also do the job for 1/10th the cost but that is still ten times too much for all but the most affluent. At least one CRT projector is now good enough but still too large and I've yet to see one equal to a phosphor. CCD light valves using either liquid crystals or thermoplastics are being worked on and show great promise but so far are a long way from 1100 lines.

Flat tubes such as the RCA strip approach might work but so far are just a gleam in the researchers eye with feasibility demonstrations showing only a small part of the picture at one time.

One of these ideas will surely work but only after millions are invested. I hope we have a David Sarnoff of the 1980's to put up the money. Display, gentlemen, is one of the biggest problems we face and it must be solved or we don't have a system. We can go ahead with the displays already available but widespread acceptance demands either the picture on the wall using some form of phosphor or fiber optic screen or else a projector the size of a Kodak Carousel - not one the size of a coffee table. It can be done and it will be done once the signals are on the air and once the barroom customers demonstrate their desire to have the spectacular pictures at home where their wives and children can see them. I believe barrooms and sports - not drama - are what will get High Definition TV out of the laboratory.

People keep asking how soon - I say immediately after we have a frequency allocation. Luckily, we have a government which is receptive to new competitive ideas. I'm hopeful that the F.C.C. will see fit to allocate adjacent channels in the 12 GHz band which can be combined to give us the bandwidth needed. Also, we need some terrestrial 12 GHz channels so the Broadcaster can put up his dish and reradiate with high power. Then the public can look and decide if they want to pay for such service. I believe one look and the manufacturers of TV sets will have trouble making them fast enough.

In closing, I would like to say that when television was introduced in the late thirties and early forties the usable spectrum extended to perhaps 500 MHz and 10% of that spectrum was allocated to TV. TV has probably done more to change our lives - throughout the world - than any invention since the electric light bulb. Today the usable spectrum probably extends to 20 GHz while the allocation to TV is now roughly 1%. Surely society can afford to give another couple of percents of precious spectrum to such a promising new service.

Terrestrial Broadcasting of High Definition Television

-- Some Considerations

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I. INTRODUCTION

The rapidly accelerating advances in television technology, and in the related technologies of Very Large Scale Integrated (VLSI) Circuits and digital techniques, have created a great current interest in High Definition Television (HDTV). HDTV as a consumer service will probably be introduced initially in non-spectrum-sensitive media such as cable television and home video players. With the advent of new broadcast spectrum in the 12 GHz satellite band there are proposals for the introduction of HDTV via direct-to-home broadcasts from satellites. Related to this, there is great current interest in finding a way to broadcast HDTV via local terrestrial broadcasting stations.

There is as yet no standard for HDTV, although efforts are currently underway to develop such standards. One of the several features that characterize an HDTV system is an improvement (over the current 525-line system) of approximately two-to-one in both vertical and horizontal resolution, which implies a scanning rate in the order of 1000 lines per frame and a corresponding video baseband in the order of 24 MHz. For distribution direct-to-home via satellite, where some signal-to-noise enhancement is essential, frequency modulation (FM) or digital modulation would be required -- with digital modulation, along with bit-rate-reduction, representing the more efficient use of bandwidth. For terrestrial broadcasting, however, amplitude modulation, or the variants single-side-band (SSB), or vestigial side-band (VSB) modulation would probably be required because of even greater bandwidth restraints. However, other modulation schemes such as low-index FM with over-deviation may also be feasible. In any event, for the purpose of this paper it will be assumed that an RF channel bandwidth of 24 MHz could accommodate an HDTV signal.

The portions of the spectrum most frequently mentioned as possibilities for accommodating such a terrestrial HDTV service are: the UHF broadcast band; the current Instructional Televisional Fixed Service (ITFS) band; or the 12 GHz

Broadcast Satellite Service (BSS) band, which was also allocated to the (terrestrial) Broadcasting Service at the 1979 World Administrative Radio Conference (WARC-79). At the present time studies are in progress to investigate the HDTV potential in all three areas. This report will cover some of the considerations related to these studies.

II. UHF BROADCAST BAND (470-806 MHz)

It is generally acknowledged that the UHF broadcast band is under-utilized for two basic reasons:

1- The band has not been developed to the full extent envisaged by the Table of Assignments.
2- A far greater number of assignments could be made, than is indicated by the Table, if it were not for the 18 allocation "taboos" that severely limit assignments.

Figure 1 represents an indication of the number of assignments on each channel from the current Table of Assignments as well as the number of operating stations, construction permits that have been granted, and pending applications. Also shown is the number of assignments from the original Table of Assignments, showing, in general, an even greater number of assignments. It may be recalled that the original Table was "de-saturated" as one of the measures to foster the development of UHF, in order to permit a more flexible implementation of UHF stations. It can be seen by the shaded area in this figure that the potential exists for much greater implementation -- even in compliance with the present allocation restrictions. This analysis does not include translators or Low Power Television Station applications. Such stations exist on a secondary, non-interference, basis. What impact these stations would have on any proposal to create new HDTV channels would be an added consideration for study.

Figure 2, by way of review, indicates the current taboo restrictions. The restrictions relate to receiver considerations and might conceivably be reduced in the future as receiver improvements are introduced - such as more linear front-ends to ease the intermodulation problem. Note, for instance, that the intermodulation taboo requires the current separation of six channels, or 36 MHz, for two stations in the same city.

Studies are currently underway to consider possible methods of exploiting this under-utilization of the UHF band and the possibility of creating HDTV channels of 24 MHz bandwidth.

III. ITFS BAND (2500 - 2690 MHz)

The ITFS band consists of 31 channels configured in a similar six MHz-wide VSB structure as broadcast television. Twenty-eight channels are allocated for ITFS use and three to private Operational Fixed Service (OFS) use. (Each channel is also assigned a narrow-band voice or data "response" channel). A major re-structuring of this band is the subject of a current FCC proposal which would provide eleven ITFS, ten Multi-Point Distribution Service (MDS) and ten

OFS channels. The proposal has been the subject of much controversy and no FCC decision has been announced as yet. The spectrum involved could provide eight channels, of approximately 24 MHz width, which could probably be designed to provide a nationwide HDTV service. This would indeed be a bold plan, which has not been advanced to-date, but conceivably the FCC, in its role as spectrum manager, might be asked to weigh such a proposal vis-a-vis the other proposals. (Such a proposal would not necessarily have a negative impact, in the long run, on prospective users. The tremendous potential of HDTV for instructional use, as well as for a unique information and entertainment service, might have considerable appeal.)

Another proposal for this band, which has been presented to the FCC just recently, would allow a single operator to use a block of five MDS channels in a given city. If adopted this presents the intriguing possibility for the introduction of HDTV using such a block of four or five channels. In any event, although there are obviously many problems and considerations involved, this matter is the subject of a study currently in progress.

IV. BROADCAST SATELLITE BAND (12.2 - 12.7 GHz)

The appeal of this band is that spectrum has been allocated, at WARC-79, for the (terrestrial) Broadcasting Service, as well as for the Broadcast Satellite Service. In the U.S. preparatory work for the WARC-79 the Television Broadcast Service Working Group recommended that there be spectrum allocated in the 12 GHz band for new and innovative television services such as HDTV. The U.S. did support this position and such an allocation was adopted. Figure 3 shows the allocation of the 11.7 - 12.7 GHz band that was allocated at WARC-79, which will be re-affirmed at the special Region 2 (the Americas) Regional Administrative Radio Conference to be held next year. Although, there is some dispute over the decision on a small segment of this band it is generally believed that there will be an even split between the Fixed Satellite Service (FSS) and the Broadcast Satellite Services (BSS) with each service allocated 500 MHz, as indicated in this Figure. This allocation refers to the Space-to-Earth or downlink, allocation, there being an equal allocation, in the 18 GHz band for the BSS, for the Earth-to-Space, or uplink -- or feederlink -- segment.

The "all capital letters" indicate co-equal and primary status -- subject to the various footnotes that are indicated. With respect to FS there are over 2000 private microwave operations in existence which will have to be dealt with. With respect to the MOBILE allocation there has been no interest whatever in any potential mobile operations. As for BROADCASTING there is a strong position developing that a portion of the BSS band should be set aside for terrestrial services, including broadcasting. Separate specific frequency allocations are necessary because of the extreme sensitivity of BSS receiving systems, which precludes BSS from operation on the same frequency as terrestrial services. (The WARC-79 treaty has not as yet been ratified by the U.S. Senate and the FCC has not as yet completed its Rule Making process with respect to specific implementation of the treaty's provisions).

Most broadcasters express shock at the first suggestion of the 12 GHz band for broadcasting, being familiar with that band only in terms of point-to-point microwave applications, such as television (remote) pickup and studio-to-transmitter links. Following World War II many expressed similar shock at the thought of broadcasting in the area of 500 MHz, the start of the UHF band. History has taught the industry how to use higher and higher frequencies as the ability to generate significant amounts of power and suitable methods of modulation were developed. Considering that WARC-79 has allocated broadcast frequencies up to 86 GHz, it may be that by the year 2000, 12 GHz will be thought of as "prime" spectrum.

One of the factors that makes terrestrial broadcasting at 12 GHz feasible is the fact that the receiving system used would be the same as the very sensitive Direct Broadcast Satellite (DBS) receiving system. As a very crude comparison, compare a fringe area UHF receiving system with a DBS receiving system. The UHF system might consist of a two-bay, 10-element Yagi antenna, with a gain of 12 dB; 5 dB of line loss; and a receiver with a noise figure of 12 dB. In contrast a DBS system would consist of a 3-foot parabolic ("dish") antenna with a gain of 39 dB; about a 1 dB line loss (since the low noise amplifier, which determines the signal-to-noise ratio, is located at the antenna) and a receiver with a noise figure of 3 dB. This 40 dB net advantage would go a long way in compensating for the greater propagation loss; the system margin needed for rain attenuation; and the relatively lower transmitter power.

Terrestrial broadcasting at 12 GHz has been the subject of study in many other countries throughout the world, including Germany, Switzerland, the Netherlands, and Japan. (In Japan, systems are in actual operation on a limited basis.) The most elaborate development work is currently underway in France, where several experimental systems have been in operation for over four years.

1. Cellular concept

One approach to 12 GHz band terrestrial service is the "cellular" concept which involves many low power, low antenna height, transmitters serving small "cells", with many such cells allocated on a grid basis to cover an entire country. The plan provides saturated coverage and permits extensive frequency re-use. Such a system involves the planning and operation by a single entity -- that is, the national administration -- with a mandate to serve the entire country.

Figure 4 is an example of such a system that at one time was considered in the Federal Republic of Germany, which could have provided eight national networks. Such an approach would not be generally applicable in the U.S. but some of the features might be.

2. Wide-area approach

A different approach has been pursued in France where a more conventional system is being used experimentally, with relatively high power transmitters (up to 2 kW), located at high elevation points, using 180° or 360° (omni-directional) transmitting antennas. Experimental systems have been in operation for over four years with the goal of establishing a new national service. The first such experimental installation is located at Puy-de-Dome in central France, using a 200 W transmitter and a 180° horn antenna. This installation is depicted in Figure 5.

As an example of the progress being made in France one manufacturer includes in its catalog a 2 kW transmitter and an omni-directional antenna with a gain of 15 dB. These are the facilities to be used at a proposed new site at Romaineville, near Paris.

The signals being transmitted are conventional 625-line SECAM signals, with two audio channels, using FM and a 27 MHz wide RF channel. As noted earlier such a channel might accommodate an HDTV signal, using another modulation technique, such as VSB or low-index FM with overdeviation.

3. Possible combination approach

For the U.S. it would appear that the best approach might be a combination of: a wide coverage area station, supplemented by low power facilities to reach shadowed areas. It must be recognized that service would be provided essentially on a line-of-sight basis, (or via dependable reflections), and that there would be shadowed areas. Because of the highly directional receiving antennas that would be used such fill-in transmitters might conceivably be operated as "boosters", operating on the same channel as the main station. Booster operation has always posed a problem, because of feedback and co-channel interference from two sources. If this objection could be overcome, however, it would make for a more efficient use of the spectrum. (It is noted that the FCC is currently re-visiting the microwave booster situation and is proposing, in BC Doc. 82-20, to permit such operation by television auxiliary broadcast stations).

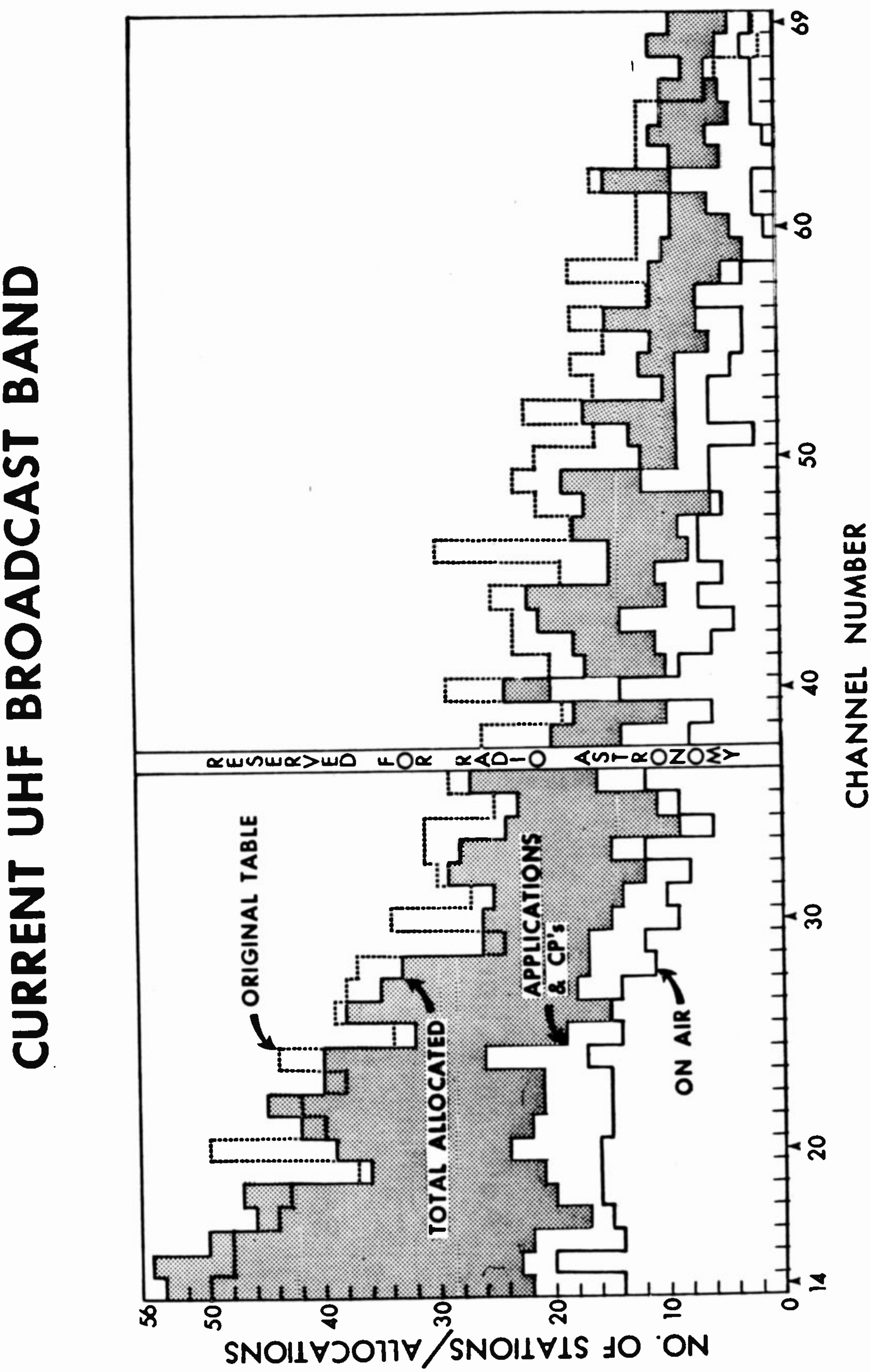
V. CBS 12 GHz PROPAGATION TEST

In order to obtain first-hand experience with terrestrial broadcasting at 12 GHz, CBS, in cooperation with KPIX, the Westinghouse station in San Francisco, is currently conducting a propagation test, using equipment obtained from TeleDiffusion de France (TDF). The transmission equipment consists of a 20 W transmitter installed on the fourth level of the Mt. Sutro tower, 550 feet above ground, feeding a 180° horn antenna identical to the type pictured for the Puy-de-Dome site in France. This combination results in a modest effective radiated power (eirp) but given the relatively high antenna elevation, about 1380 feet above mean sea level, the recorded line-of-site field strength can be extrapolated to predict what would happen with eirp's typical of actual operating stations. CBS has received an Experimental Special Temporary Authority from the FCC to operate on 12.29 GHz. This frequency, which is within the proposed DBS band, was selected on the basis of a frequency search which indicated that such an operation would pose no potential interference problem to existing private microwave operations. Fixed monitoring locations have been established where chart recorders will record the variation of field strength with time and weather. Receiving sites involve a three-foot parabolic antenna (39 dB gain) and a receiver with a noise figure of about 6.5 dB. Additionally, a mobile unit, equipped with a 30 foot mast and a horn antenna, will be used to make mobile measurements.

Figure 6 shows a block diagram of the system. Figure 7 shows the horizontal and vertical patterns of the transmitting antenna, which is oriented at N 74° E, towards downtown San Francisco. Figures 8 through 16 are photographs of the site and the equipment being used. The system is now on the air, and measurements are being made. The results of the tests will be made known in a future report after all the data has been processed and analyzed.

VI. CONCLUSION

In this report several possibilities have been suggested for further study with respect to frequencies that might be used for terrestrial broadcasting of HDTV. CBS has commissioned studies to investigate the potential of the UHF and ITFS bands. Additionally studies are underway to determine the potential for the use of the 12 GHz DBS band for this application. With respect to this band, the FCC has not as yet formulated a final position as to how to deal with the issue. However, one strong position has developed within the FCC's DBS Advisory Committee that implementation of DBS in the U.S. should also provide for terrestrial services, including broadcasting -- and specifically HDTV broadcasting.



CURRENT UHF BROADCAST BAND
ORIGINAL TABLE
TOTAL ALLOCATED
APPLICATIONS & CP's
ON AIR
RESERVED FOR RADIO ASTRONOMY
NO. OF STATIONS/ALLOCATIONS
CHANNEL NUMBER
0
10
20
30
40
50
56
14
20
30
40
50
60
69

Figure I

UHF ALLOCATION TABOOS

Taboo	No. of Channels Separation Required	Corresponding Mileage Separation (Miles)
Intermodulation	± 2, 3, 4, 5	20
I.F. Beat	± 8	20
Adjacent Channel	± 1	55
Oscillator	± 7	60
Sound Image	± 14	60
Picture Image	± 15	75

Figure 2

Probable 12 GHz Satellite Band Allocation Following RARC-83

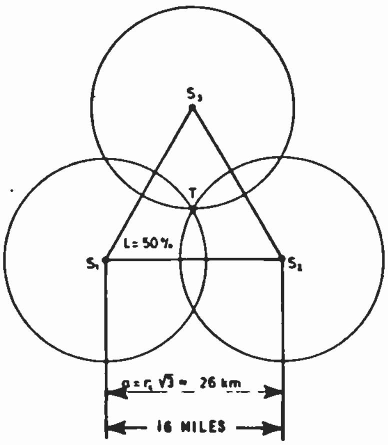
Band	Allocation
11.7 – 12.2 GHz	FIXED FIXED - SATELLITE Mobile (except aero mobile)
12.2 – 12.7 GHz	FIXED MOBILE (except aero mobile) BROADCASTING BROADCASTING - SATELLITE

Subject to footnotes:
3785H, 3787 405BC, and 3787A, C, D, E, F.

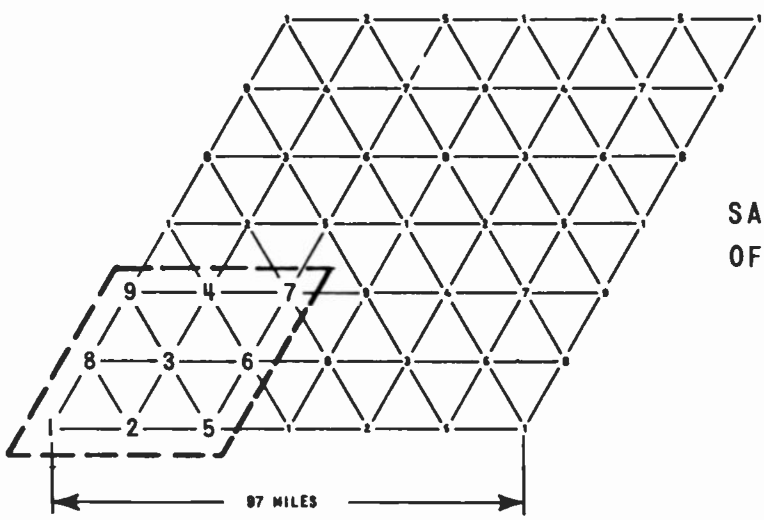
Figure 3

EXAMPLE OF PLAN FOR 12 GHz TERRESTRIAL BROADCAST SERVICE

(Federal Republic of Germany, 1968)



BASIC "SINGLE COVERAGE" UNIT (SATISFACTORY RECEPTION FROM ONE SITE CAN BE EXPECTED AT LEAST 50% OF LOCATIONS)



SAMPLE GRID BASED ON REPETITION OF BASIC 9-SITE RHOMBUS

ONE POSSIBLE NATIONAL PLAN PROVIDING EIGHT NETWORKS; PATTERN REPEATED TO COVER COUNTRY (75 EIGHT-MHz WIDE CHANNELS INCLUDING GUARD BANDS)

Site	Channels						
1	1,	4,	7,	10,	13,	16,	19,	22
2	26,	29,	32,	35,	38,	41,	44,	47
3	2,	5,	8,	11,	14,	17,	20,	23
4	51,	54,	57,	60,	63,	66,	69,	72
5	52,	55,	58,	61,	64,	67,	70,	73
6	28,	31,	34,	37,	40,	43,	46,	49
7	3,	6,	9,	12,	15,	18,	21,	24
8	53,	56,	59,	62,	65,	68,	71,	74
9	27,	30,	33,	36,	39,	42,	45,	48

Figure 4

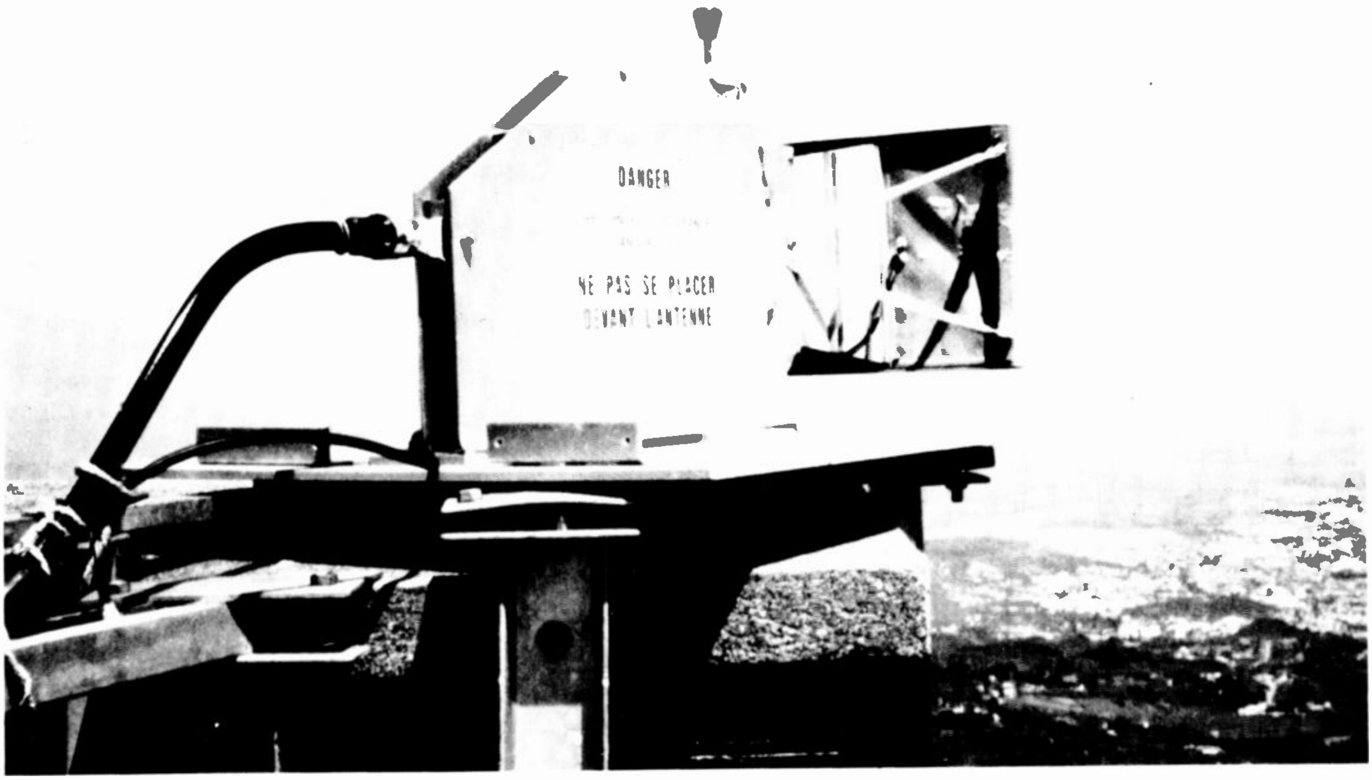


Figure 5

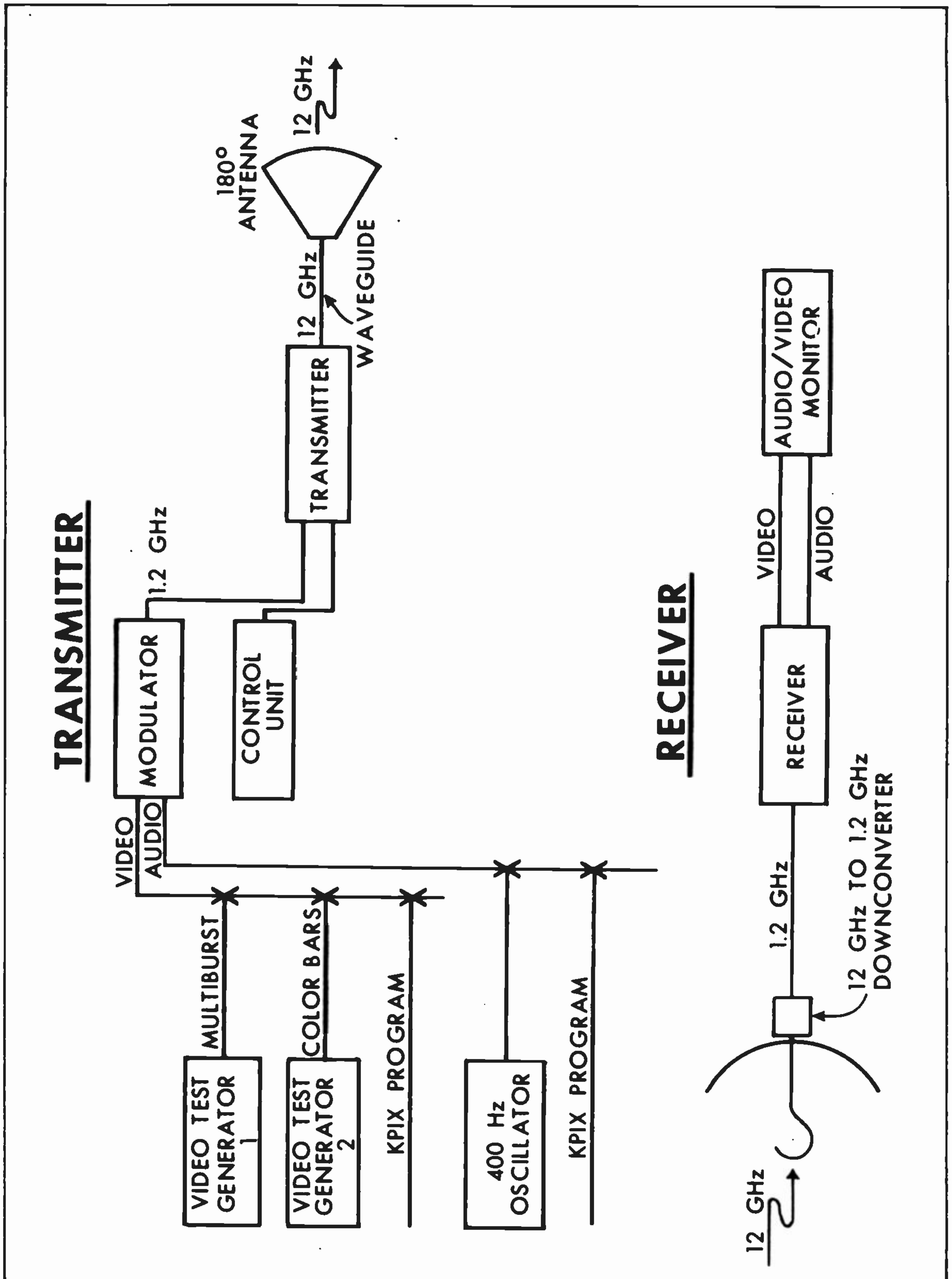
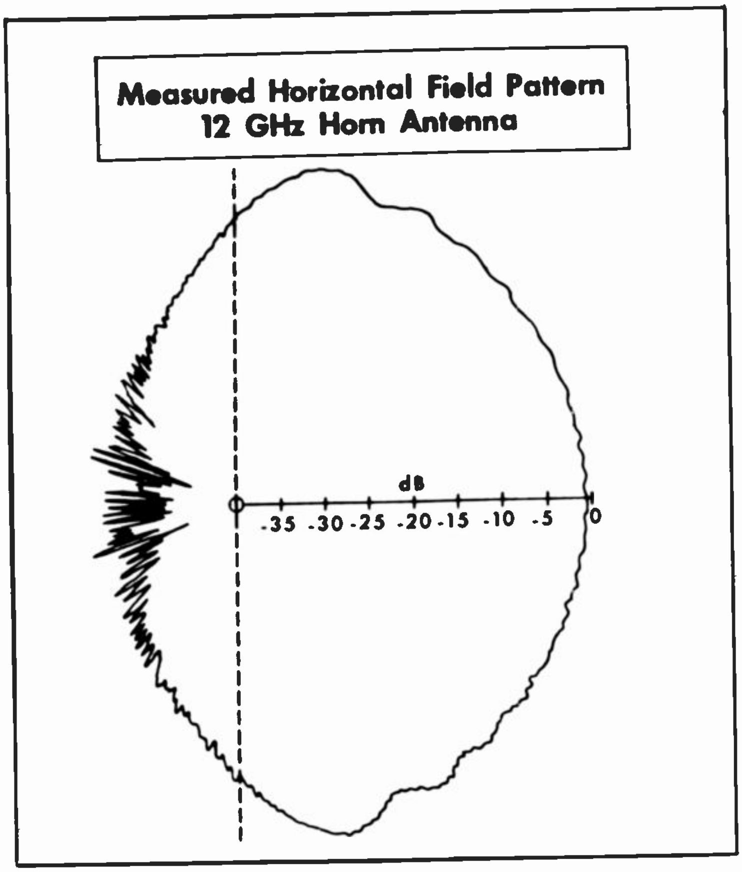

TRANSMITTER
MODULATOR
1.2 GHz
VIDEO
AUDIO
CONTROL UNIT
TRANSMITTER
12 GHz
WAVEGUIDE
180° ANTENNA
12 GHz
MULTIBURST
COLOR BARS
VIDEO TEST GENERATOR 1
VIDEO TEST GENERATOR 2
KPIX PROGRAM
400 Hz OSCILLATOR
KPIX PROGRAM
RECEIVER
12 GHz
12 GHz TO 1.2 GHz DOWNCONVERTER
1.2 GHz
RECEIVER
VIDEO
AUDIO
AUDIO/VIDEO MONITOR

Figure 6


Measured Horizontal Field Pattern
12 GHz Horn Antenna
dB
-35 -30 -25 -20 -15 -10 -5 0

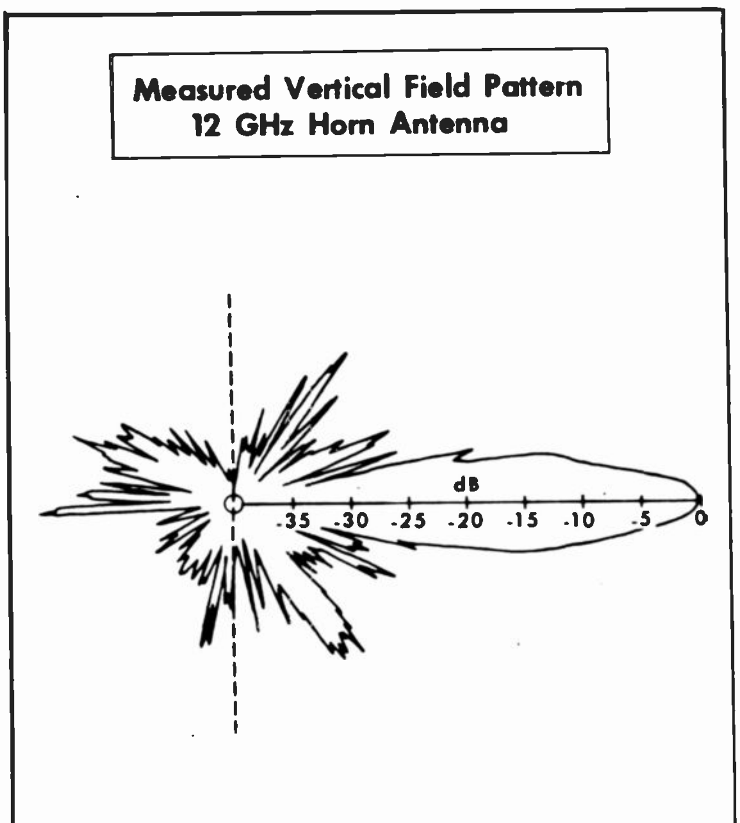

Measured Vertical Field Pattern
12 GHz Horn Antenna
dB
-35 -30 -25 -20 -15 -10 -5 0

FIG. 7

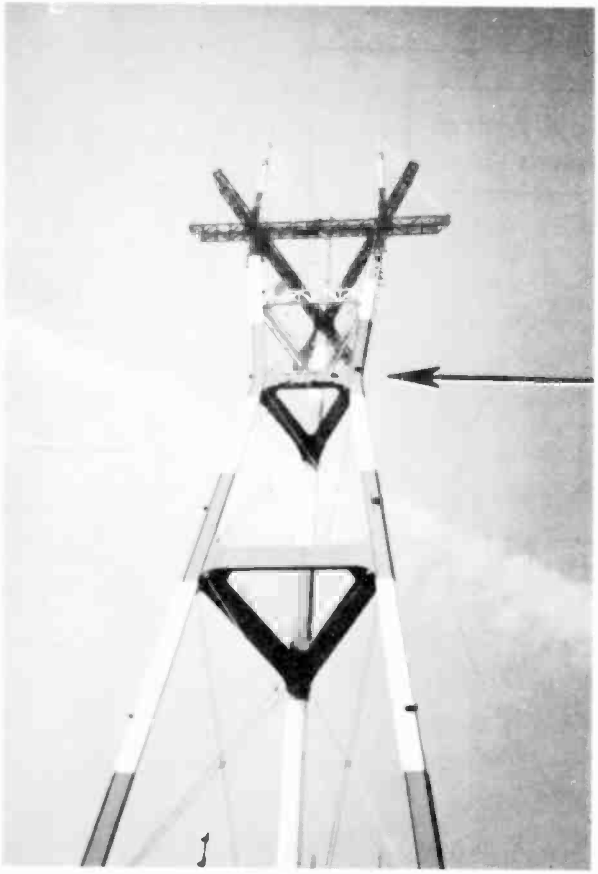


FIG. 8
LOCATION OF TRANSMITTER
ON MT. SUTRO TOWER

FIG. 9
TRANSMITTER BEFORE INSTALLATION

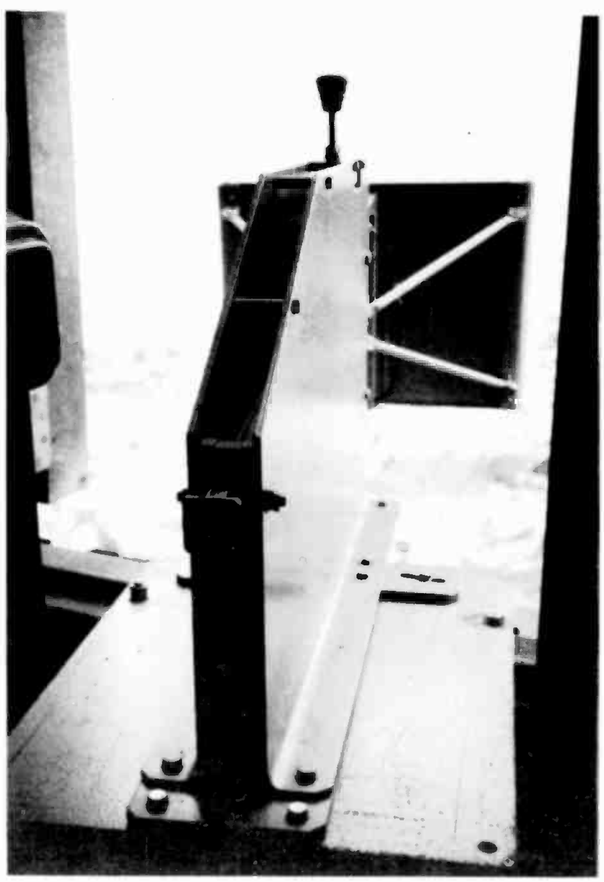


FIG. 10
TRANSMITTING ANTENNA INSTALLED

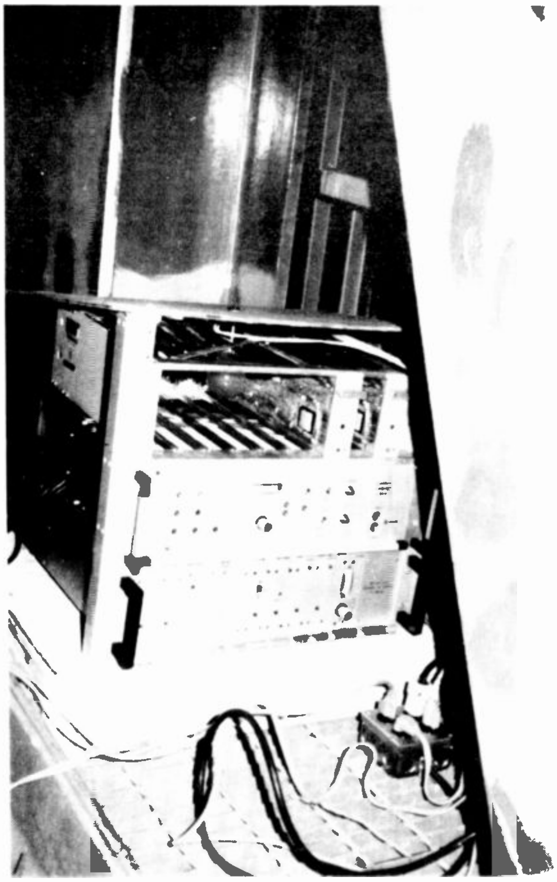


FIG. 11
MODULATOR AND CONTROL UNIT

FIG. 12
MOBILE UNIT WITH SMALL HORN ANTENNA

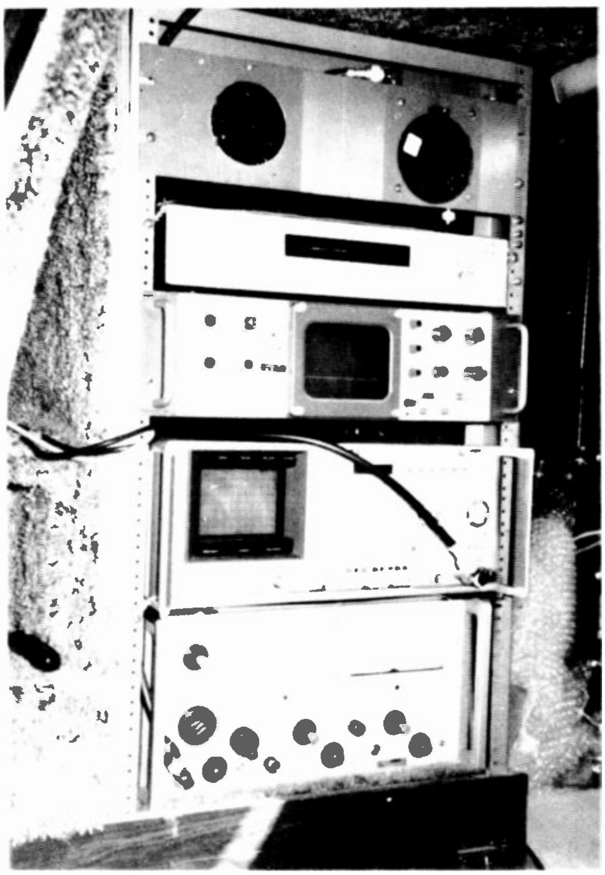


FIG. 13
MEASURING EQUIPMENT IN VAN

FIG. 14
TYPICAL MEASUREMENT SITE

FIG. 15
FIXED SITE AT KCBS, SAN FRANCISCO

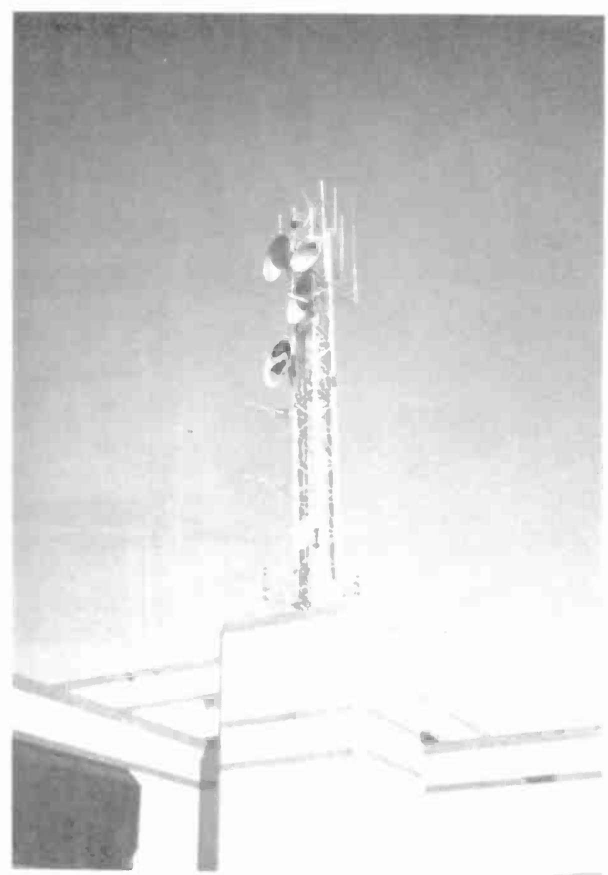


FIG. 16
FIXED SITE AT KTVU, OAKLAND

MAXIMIZING SPECTRUM UTILIZATION FOR TELEVISION SIGNAL TRANSMISSION

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I. Introduction

It is well known that due to its repetition characteristics, a TV signal does not fully utilize the frequency bandwidth used for its transmission. The scanning process and the spatial and temporal correlation that generally exists for picture material implies a high degree of redundancy.

Efficient spectrum utilization requires the elimination of redundancies from the transmitted signal. Bandwidth compression techniques can greatly increase transmission efficiencies so that it becomes possible to distribute existing quality TV signals tnrough lower bandwidth TV channels or, more important, to provide a much higher quality picture through a practical, limited bandwidth channel.

Bandwidth compression for transmission purposes implies decompression at the receiver for picture display. That represents a sharp departure from present TV systems which were designed to maximize receiver's simplicity in order to keep cost low. As a consequence, the present television transmission standards are highly inefficient. With the tremendous advances in LSI and VLSI technology we can now design receivers with memory and intelligent adaptive processing without severe cost penalties. For example, we have consumer receivers already employing comb filters for chrominance/luminance separation and vertical detail processing. This is an application of video processing formerly confined to professional equipment and requiring expensive line stores.

Projections for the 1980's indicate that CCD memory chips of 2 to 4 megabits will be available, leading to a prediction of a $10 frame store. A considerable amount of processing then can be relegated to the receiver, thus allowing adaptive compatibility and spectrum efficient transmission of a much higher quality television signal.

II. Fundamental Approaches to Bandwidth Compression

Bandwidth compression, in general, implies redundancy removal. In fact, however, there are two fundamental ways to achieve bandwidth compression. One is to maximize the entropy of the transmitted signal. The other is to discard picture information which has very little significance with regard to subjective picture quality.

Bandwidth compression generally implies digital processing. Converting an analog signal to a digital signal provides access to every individual picture element of a television program. Access to individual picture elements is essential for recognizing and removing redundancies in a TV picture, hence, achieve more efficient transmission. Bandwidth compression, then, is accomplished by reducing the bit rate of a digital television signal.

Redundancy removal, to achieve bit rate reduction, should be a completely transparent process. Most predictive type coding systems can be completely transparent. Run length coding and entropy coding can also be made fully transparent and still provide significant bit rate reductions.

The other general category of bandwidth compression techniques are based on global picture statistics and psychovisual human responses that permit the removal of unimportant picture information. In this category one can include all forms of subsampling, most filtering processes that reject or attenuate some frequencies within the original signal's baseband, most transform coding techniques, and any requantizing technique that results in coarser quantization. These processes may provide good picture quality under most conditions, however, one can expect to find situations when noticeable impairment will result.

III. Bit Rate Reduction Techniques

Many source coding or bandwidth compression techniques have been explored by researchers throughout the world. A literature search on this subject reveals the existance of more than 500 publications. Many variations of the different techniques have been analyzed. Here we will briefly describe the more common bit rate reduction means.

A. Transform Coding

In transform coding one translates the picture samples to a new set of coordinates. The transformation results in a redistribution of picture information into a new set of variables, some of which have very little significance with regard to subjective picture quality and can, therefore, be eliminated or be transmitted with lower accuracy.

B. DPCM Techniques

Differential PCM coding is one of the most common bit rate reduction techniques. The simplest form of DPCM coding is to transmit the difference between adjacent or nearby samples rather than the actual value of each sample. Because nearby samples have a high degree of correlation, the differences will normally be a much smaller signal than the absolute value of each sample.

DPCM techniques are generally associated with predictive coding. In predictive coding the value of any sample is given by a weighted combination of previous samples (the predictor). This predicted value is then compared with the true value of that sample. The difference or error between the predicted value and the true value is then transmitted. At the receiver a similar predictor uses the received "error" to correct the predicted sample in order to obtain the actual value.

C. Interframe Coding

Interframe coding utilizes picture redundancies from adjacent frames to predict the value of future samples with a high degree of reliability. DPCM techniques are normally used in this type of coding. In the absence of motion, of course, it is obvious that corresponding picture elements on every frame are the same. Therefore, using interframe coding, a still picture can be transmitted at bit rates approaching zero.

D. Frequency Interleaved Coding

Frequency interleaved encoding utilizes the empty frequency gaps present in the energy spectrum of a typical TV signal. The concept is best illustrated by the NTSC color TV system where the color information is transmitted on a subcarrier within the luminance signal's frequency band.

So-called sub-Nyquist encoding schemes are forms of frequency interleaved encoding. Sub-sampling of picture samples either along a TV line, or in the vertical picture axis, and even along the temporal picture axis can be viewed as different forms of frequency interleaved encoding.

E. Polynomial Data Compression

Polynomial data compression techniques are normally used in conjunction with DPCM systems to perform future sample predictions. These techniques are based on the fact that a high degree of correlation exists among neighboring picture samples. One then could predict the value of a future sample by combining previous known samples in a polynomial expression. Transmission is needed only when the predicted value falls outside a pre-selected tolerance level.

F. Entropy Coding

Entropy coding takes advantage of the fact that certain codewords describing picture samples occur much more often than others. In entropy coding the more common source symbols are assigned shorter word lengths. Source symbols that occur infrequently are assigned the longer work lengths. The Morse code is an example of entropy coding. Then the average number of bits required to transmit the signal are reduced.

IV. Bandwidth Compression for High Quality TV Signals

For entertainment type television systems, picture quality has to be assigned top priority whenever bit rate reduction schemes are being contemplated to achieve more efficient bandwidth utilization. Ideally, it should be a completely transparent process. That can be done by removing only picture redundancies.

A television signal can be viewed as a spatial array of picture samples defining a TV frame. This large array of picture samples is transmitted every one thirtieth of a second i.e. at the frame rate. In a fixed picture the number of samples must be high enough to be able to carry the highest video frequencies. For example, if a TV signal is expected to have frequencies as high as 4.2 MHz, the sampling rate must be at least 8.4 M samples/sec. However, we know that television signals typically have very few high frequency details. Often less than 10% of a television frame requires frequencies higher than 1.5 to 2 MHz. In such cases we would need to provide the full resolution array of picture elements for only 10% of that frame. If 10% of the picture has only 36% to 48% of the maximum vertical and horizontal resolution, the number of picture samples required for picture reproduction can be reduced by at least a factor of four. Therefore, simply eliminating redundancies within a TV frame (intraframe horizontal and vertical redundancies), should result in a 4 to 1 bit rate reduction without noticeable picture impairments.

If we now analyze the average television program we will find that, on a frame by frame basis, only 10% to 20% of the total picture elements defining a TV frame are being modified. Yet all picture elements of each frame are transmitted whether they changed or not. Interframe changes imply motion or scene changes, therefore, they will be highly dependent on the type of television program. Some preliminary measurements of picture statistics however disclose that television pictures are 80% to 90% redundant on a frame by frame basis. In addition, even when interframe differences between corresponding picture elements do occur, those differences are much smaller than the full dynamic range of the video signal. It is reasonable to expect then, that by removing interframe redundancies, one could achieve bit rate reduction factors of 5 to 10. Therefore, a bandwidth compression system based on the removal of interframe redundancies is expected to provide large compression ratios, under most conditions, with no visible impairments. Obviously, there will be cases when scene changes and sustained motion will affect much more than 10% to 20% of the picture elements. Under those conditions, unless there is an extremely large buffer to store many bits over long periods, one must rely on intraframe redundancies to keep the bit rate down or to resort to other techniques based on psychovisual characteristics.

In the presence of motion, sub-sampling techniques should provide sufficiently good picture quality with bit reduction factors of 2 to 4. One must be careful, however, when employing sub-sampling techniques even when a picture contains a high amount of interframe motion. For example, consider a scene in which the camera is tracking a runner. One can expect many interframe differences because of the background pan. The subject of interest however will appear stationary in picture space. Using sub-sampling on stationary images can cause objectionable resolution losses. A bit reduction system should prevent sub-sampling of stationary parts of a scene. The number of interframe redundancies in any picture area can provide the information necessary to determine whether any portion of a scene should not be sub-sampled.

Transform coding techniques can also provide significant bandwidth compression but, for equivalent complexity, they are outperformed by predictive DPCM techniques. Therefore, there seems to be no major advantage in using transform coding.

In conclusion, it should be possible to design broadcast type television systems with bit rate reduction factors of 5 to 10 with no performance penalties. It may utilize adaptive interframe and intraframe DPCM techniques and adaptive sub-sampling in picture areas containing violent motion or abrupt scene changes.

V. Preliminary Experimental Results

In order to evaluate bit rate reduction techniques, we have implemented an intraframe bit rate reduction system that can provide a 5 to 1 bit rate compression of a digital NTSC color television signal. The system uses frequency interleaved encoding and DPCM encoding with a simple polynomial for the predictor. The input NTSC color TV signal is sampled at four times color subcarrier, 8 bits/sample, having a total bit rate of 114 M bit/sec. Sub-sampling of the signal on each scan line reduces the bit rate by a factor of two. DPCM encoding provides an additional 50%

reduction. Blanking removal, although not implemented, would provide a 20% reduction. Therefore, the digital signal has a bit rate of 23 M bits/sec.

It is obviously a very simple system and it causes some picture impairment. One of the major reasons for the less-than-broadcast quality performance is due to the composite color television signal placing severe restrictions on the sub-sampling scheme that can be used, and on the DPCM predictor. Nevertheless, when viewing the pictures from 4 to 6 times picture height, the impairments are barely visible.

A signal-to-noise ratio of approximately 43 dB is achieved with this bit reduction system. Furthermore, since it uses only intraframe processes, the resulting quantizing noise is uncorrelated from frame to frame allowing the use of a digital noise reducer to improve the S/N by 8 to 12 dB.

This system represents quite an achievement if one considers that the bit reduction coder utilizes just a handful of parts, less than 30 off-the-shelf IC's. It leads us to believe that with more sophisticated adaptive interframe techniques one can achieve much higher compression ratios with excellent picture quality.

VI. Conclusion

We have examined several bandwidth compression techniques suitable for high quality television signals. If picture quality is assigned first priority, it is our conclusion that the most desirable approach to bandwidth compression is to use adaptive DPCM techniques designed to utilize only picture redundancies to achieve bit rate reduction. In situations where the amount of redundancy is insufficient to achieve a predetermined compression ratio, the use of sub-sampling in areas of motion or areas of fast changes is recommended. By use of these techniques it should be possible to transmit a high definition television signal through acceptable channel bandwidths.

Visual Performance Characteristics Which Affect Multichannel Sound

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BACKGROUND:

Multichannel sound systems for TV are not entirely new. Here in the United States aural subcarrier transmission has been used, with Harris transmitters, for remote transmitter data monitoring since 1972. Subscription television systems employing subcarrier audio have been in wide use since 1978. Stereo has been broadcast by PBS stations since the mid 1960's using FM and AM simulcasting. Recently PBS has introduced second language programs using simulcasting. At present stereo programs can be distributed by satelite and telephone systems. Multilanguage programs are available on tape and have been distributed by satelitte. There are existing means of programming and distributing multichannel sound for television. Even some TV studios are already equipped for stereo.

The EIA proposals have the following multichannel design goals for providing a wide range of services:[1]

1. Stereo
2. Separate audio program
3. Telemetry

Subjective evaluation of the proposed systems is complicated by the difficulty of sorting out both receiver and transmitter limitations. Some transmitter designs produce high levels of interference. On the other hand receivers using common visual - aural IFs and intercarrier detection produce distortion products as high as three to five percent.

The principal sources of visual interference to the aural channel are:

1. Imperfect visual bandwidth limiting.
2. Distortion products produced in high power amplifiers.

Although it is extremely difficult to evaluate the affect of each type of interference because of the various demodulation schemes, it is relatively straightforward to describe the predetected aural carrier to interference ratio. Thus, practical and theoretical performance limitations caused by visual transmitter characteristics can be more readily indentified. Technically, choosing the right multichannel system is not a question of which system performs best in the lab, but which one can produce acceptable quality in the noisy environment of a TV channel.

Poor signal to noise ratio and high distortion is not inherent. Hopefully the industry will adopt a modulation system which fits TV requirements and utilizes the allocated spectrum to the fullest.

INTRODUCTION:

This engineering note describes visual to aural interference sources, recent transmitter improvements incorporated in the new Harris exciter, MCP-2, shown in Figure 8, and area of possible further improvements. Pertinent TV transmission regulations are reviewed to put in perspective the additional visual transmission quality requirements with respect to the implementation of multichannel sound. Specific video tests are suggested for evaluating potential interference levels.

TV CHANNEL TRANSMISSION CHARACTERISTICS

The FCC television RF channel allocation is for a six megahertz bandwidth. Within the channel the visual carrier is 1.25 MHz above the lower band edge. The visual passband is defined over a vestigial sideband including -.75MHz below to 4.18 MHz above visual carrier. Color subcarrier is at +3.58 MHz and aural carrier is at +4.5 MHz. The transmission characteristic is plotted in Figure 1. Also plotted in Figure 1 are maximum permitted modulation sideband levels as a function of frequency. Note the rules do not require visual attenuation at 4.5MHz. Furthermore, there are no minimum requirements for controlling visual to aural crosstalk. For a sinusoidal varying luminance signal, the maximum sideband can be calculated from the following expression:

$$S\ (W) = (r + m\cos w_m t)\cos w_c t \qquad (1)$$

Where $S\ (W)$ = modulated visual signal

r = luminance level

m = video signal amplitude on the luminance level

w_m = angular frequency of the video signal

w_c = angular frequency of the carrier with respect to sync peak

The maximum sideband level at $w_c + w_m$ occurs for $r = .414$, $m = .289$ from (1).

$$S(w_c + w_m) = (.289/2) = .145$$

Or expressed in dB

$$S\ (w_c + w_m)_{dB} = -16.8 \text{ dB below sync peak}$$

The video signal and modulated spectra is shown in Figure (2). A similar development can be shown for an AM modulated chroma signal. Maximum chroma sidebands spaced at 920 kHz can be as large as 28.8dB below sync peak. Figure (3) shows the chroma test signal and its spectrum.

Over the last decade substantial improvement in vestigial sideband filtering techniques have been made. Figure 4 illustrates these advances. Note the performance of the new MCP-2 Harris excite. The vestigial sideband filter is of a new design and rejects the interference at aural by 15dB. The SAW filter used in the MCP-2 is unique in another way as well. It incorporates the FCC receiver equalizer delay correction. Thus an entire subassembly along with its many components and adjustments are eliminated. Now, the FCC curve no longer needs to be periodically adjusted, it is built in.

The improvement in filter attenuation was made possible by increasing the number of acoustic transducer fingers. The number of fingers in an acoustic filter increases the rate of attenuation versus frequency much the same way the number of resonators does in a conventional filter.The frequency band between minimum and maximum attenuation is called the transistion bandwidth. The number of fingers and the transition bandwidth has the following approximate relationship:[2]

$$N \cong 2f_o/f_t$$

WHERE N = Number of acoustic transducer fingers

f_o = FILTER CENTER FREQUENCY

f_t = BANDWIDTH OF MIN TO MAX ATTENUATION

For the IF filter used in the Harris exciter:

$$f_o = [\ (37 + .75) + (37 - 4.18)\ ]\ 1/2 \text{ MHz}$$

$$f_o = 35.28\text{MHz}$$

$$f_t = f_{aural} - f_{visual}$$

$$f_t = 4.5 \text{ MHz} - 4.18 \text{ MHz}$$

$$f_t = 320 \text{ kHz}$$

Increasing the number of fingers makes the filter longer and controling spurious reflection more difficult. In contrast to a transmitter, a demodulator cuts off at 4 MHz. A 4 MHz cutoff results in a 500 kHz transition bandwidth, requiring nearly half as many transducer fingers.

Further improvement to aural interference can be obtained by changing the rules to allow transmitters to roll off at 4 MHz rather than 4.18 Mhz. Another approach would be to use a video filter. A video notch using a crystal monolithic filter at 4.5 MHz can reduce visual crosstalk without causing group delay distortion. Such a filter has been tried in the lab and has obtained 10 dB of attenuation over a 60 kHz bandwidth.

NON LINEAR EFFECTS:

The output of non linear RF amplifier can be expressed by the following power series:[3]

(2) $$e_o = a_1e_1 + a_2(e_1)^2 + a_3(e_1)^3 \ldots.$$

WHERE

$a_1, a_2, \ldots.$.are non linearity coefficients

e_o = output signal

e_1 = input signal

For a video input consisting of modulated luminance in the vestigal sideband frequency range the input signal is of the form :

From (1) $e_1 = r \cos w_c t + (m/2) \cos (w_c + w_m)t$

expanding each term of (2) the result for the first three terms neglecting harmonics and frequencies far from the TV channel.

(3) $a_1 e_1 = a_1 r(\cos w_c t)(m/2) \cos (w_c + w_m)t$

(4) $a_2 (e_1)^3 = a_3 r^2 (m/2) \cos (w_c - w_m)t + r(m/2)^2 \cos (w_c + 2w_m)t$

It can be shown that only odd powers cause interference in the aural channel. This type of interference is called intermodulation, IM. From a visual standpoint all powers can cause visual distortion such as compression or expansion. In other words an amplifier which has a square law non-linearity would have visual distortion but not IM. A modulated signal in the vestigial sideband frequency range can produce IM products which fall above and below visual carrier. The sideband below visual carrier is called the reinserted lower sideband because it can not be distinguished from the ideal double sideband signal. The lower sideband can cause interference to an adjacent lower channel aural. Since the lower sideband visual offset is the same frequency as the modulation it can be easily measured with a tracking sideband generator and spectrum analyzer. The upper sideband can not be measured with a tracking generator because the resultant spectra is at 2 fm and therefore not frequency coincident with the spectrum analyzer sweep signal. A variable frequency test generator should be used. Although IM products of any subharmonic of 4.5 MHz can produce interference it has been found that 2.25 MHz is the most sensitive.

The test signal, modulated luminance, and resulting spectra is shown in figure (5). Note that the upper sideband level is much smaller than the lower sideband. The reason for this can be seen from the following:

From EQ (4)

$IM_1 = a_3 r^2 (m/2)$

$IM_2 = a_3 r (m/2)^2$

THEN $IM_1/IM_2 = r/(m/2)$

WHERE r = Visual Carrier (Average luminance level)

$m/2$ = Video modulation sideband level

IM_1 = lower sideband intermodulation product at $(w_c - w_m)$

IM_2 = upper sideband intermodulation product at $(w_c + 2w_m)$

For example an amplifier with a lower sideband 26dB and a carrier to video sideband level to ratio of 8dB would have an uppersideband level expressed in dB of:

$IM_2 = -8dB + 26dB = 34dB$

Although the derivation is much more complicated a similar development can be shown for incidental phase modulation distortion, ICPM. ICPM produced by some solid state and klystron amplifiers can produce sideband levels comparable to IM product levels. In addition to generating products which fall in the aural channel, ICPM of the visual carrier is transferred to the aural in receivers using intercarrier detection.[4] Testing intercarrier distortion of a subcarrier channel can not be accurately measured with a demodulator employing a Nyquist IF filter. The Nyquist filter causes AM to PM distortion to occur which is quite discernible in the frequency range of subcarrier channels The Nyquist slope attenuates signals below visual carrier and boosts signals above visual carrier. This slope characteristic causes a quadrature distortion which is frequency dependent. In monaural recievers a deemphasis filter limits the increasing distortion level with frequency. For subcarrier detection, the AM to PM distortion of the Nyquist slope severly limits performance. Older precision demodulators could guarantee measuring monaural intercarrier noise to - 46 dB. More recent demodulators can measure monaural intercarrier noise to - 52 dB. To date there is no demod commercially available for measuring intercarrier noise on subcarrier. An effective test signal for evaluating intercarrier noise is a sinusoidal luminance signal with a frequency between 30 Hz to 100 KHz.

ICPM & IM CORRECTION CIRCUITRY:

Controlling incidental phase modulation, ICPM, and intermodulation, IM, is one of the latest challenges to broadcast transmitter designs. In the past IM distortion caused only visual signal impairment such as compression and adjacent channel interference. ICPM caused visual impairment such as spiking and sync buzz in audio. With the advent of multichannel sound such distortions can severly limit the audio performance of subcarriers which are more susceptible to interference.

A correction scheme which reduces amplitude and ICPM distortion by 10 to 15 dB is included in the new MCP-2 exciter. A block diagram of the corrector is shown in figure (6). The correction scheme splits the signal into inphase, amplitude correction, and quadrature, phase correction. The phase corrector is aptly called a quadrature corrector. The phase corrector involves direct precorrection without the need for a separate detector and remodulation. The correction tends to track power output by changing the amount of correction with the envelope level of the visual IF signal.

TEST SIGNALS:

Present visual test signals are intended to check various visual performance characteristics. These signals are not adequate for evaluating visual to aural interference Using a spectrum analyzer with 30kHz IF bandwidth, the new Harris TV exciter, model MCP-2 was checked for disturbance level at 4.5 MHz.

Standard and specialized test signals were used for comparison. Signals are referenced to sync peak, dbs.

	VIDEO INPUT	RESPONSE AT VISUAL CARRIER +4.5MHz -dBs
	EIA Color Bars	63
STANADARD	FCC Composite	63
	Multiburst	55
	4.5 MHz Luminance	32
SPECIALIZED	920 kHz Modulated Chromance	44
	2.25 MHz Luminance (IM,ICPM)	52

The difference between standard and the specialized test signals is substantial, 20 to 30dB. The specialized test signals can be used to identify sources of interference and system sensitivity to IM and ICPM. Since the specialized test signals represent the highest possible modulation levels, the resultant crosstalk level is worst case. The levels represented by the specialized test signals can be generated by modern cameras, and graphic character generatacrs.

HIGH LEVEL FILTERS:

Transmitters using visual-aural diplexers modify the transmission characteristics of both aural and visual. Diplexers using a notch filter is a standard part of all Harris high power UHF transmitters and some VHF transmitters.

Figure (7) is a notch diplexer block diagram. The visual signal is split by HY1 the 3dB 90^{o} hybrid and passes through two identical notches which reflect energy at visual carrier +45 MHz. The signal is combined in the output hybrid HY2. An aural signal enters HY1 and is split into two signals, it is reflected by the notches and the aural signal is recombined in HY1. The notch diplexer attenuates visual interferance at 4.5 MHz by 25 dB. The advantage of this type of filter is that it will attenuate interference not only from linear sources but also IM and ICPM as well. A level aural notch could be added to a VHF transmitter which may use spatial or hybrid multiplexing. The disadvantage of high level filters is that they cause visual signal impairment due to amplitude response and group delay errors. This distortion must be compensated at IF or video.

Typical transmission characteristics of Harris Transmitters using specialized test signals is shown in Table 1. The signals are referenced to sync peak, dBs. The results for transmitters with diplexers indicate the existing line of transmitters provide a high level of aural protection. Transmitters without high level filters may need additional video or IF filtering for adequate multichannel sound protection. If a second aural carrier system is adopted an additional notch will be required for nothch diplexers. A second aural carrier could be placed between +4.5 MHz, the present aural channel, and $4.75Mhz, the upper TV channel limit.[5]

SUMMARY AND CONCLUSIONS:

The Harris transmitters using the new filter and correction circuits in the MCP-2 exciter provide a high level of aural signal protection expecially when a notch diplexer is used. Specific test signals such as a variable frequency signal generator should be used to test visual transmitter and multichannel sound systems to demonstrate compatibility and performance limitations.

[1] Thomas B. Keller Jr., EIA Multichannel Sound Subcommitte, Task Force B Final Report",(to be published).

[2] Anderson Labs, Handbook of Acoustical Signal Processing Vol. 1 SAW Bandpass Fjilters,(1980).

[3] Franz C. McVay, Don't Guess The Spurious Level, (Electronic Design, February 1, 1967).

[4] Pieter Fockews, Carl G. Eilens,Intercarrier Buzz Phenomena Analysis and Cures, (IEEE Transactions on Consumer Electronics, Vol. CE-27, No. 3, August 1981).

[5] James R. Simanton, Added Services for the Telesonics Multichannel sound System,(EIA Multichannel Sound Committee Minutes, September 4, 1980).

TABLE I

TRANSMITTER RESPONSE AT VISUAL CARRIER +4.5 MHz

TEST SIGNAL	TRANSMITTER WITHOUT DIPLEXER (-dBs)	TRANSMITTER WITH DIPLEXER (-dBs)
4.5 MHz Luminance	32	57
920 kHz Modulated chrominance	44	69
2.25 MHz Luminance (IM, ICPM)	42	67

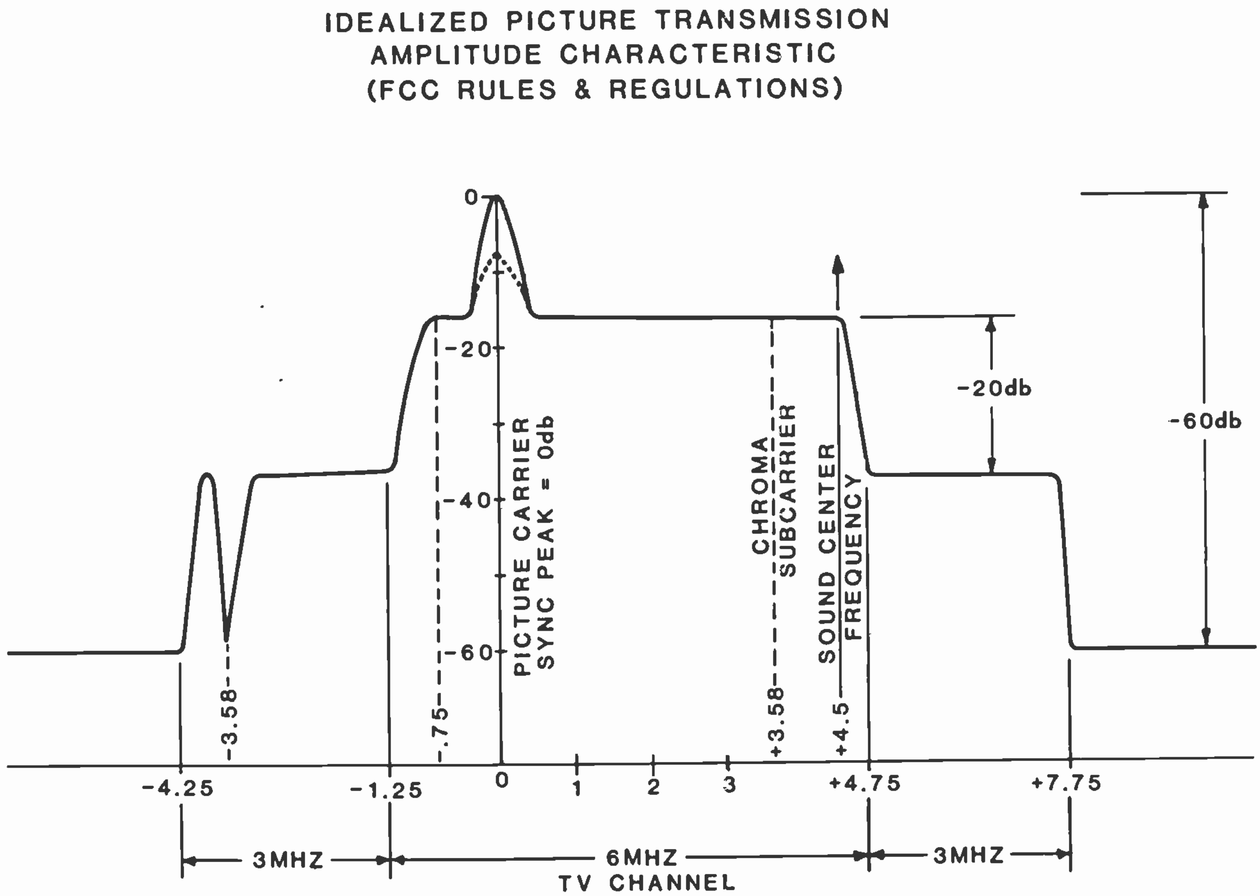


Figure 1

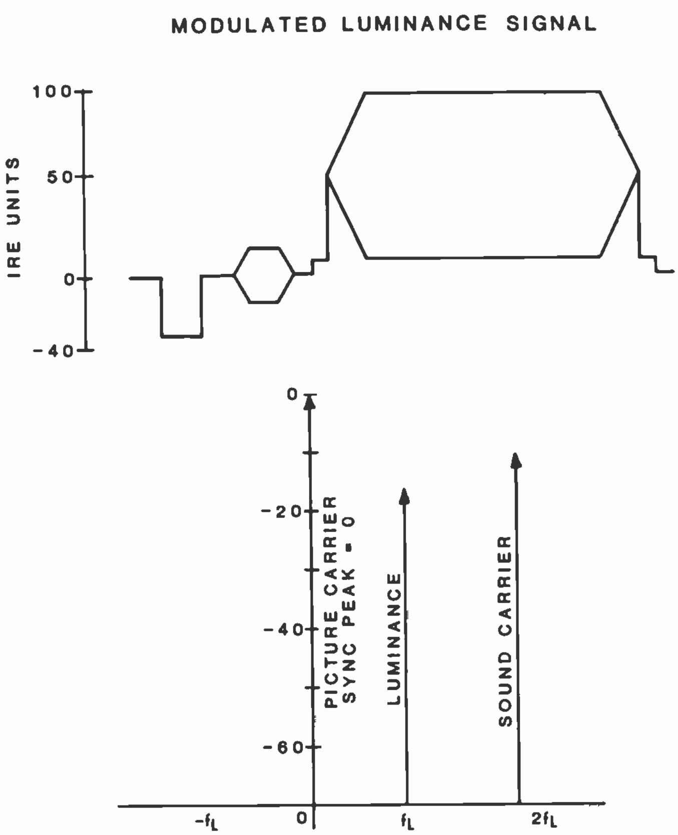


Figure 2

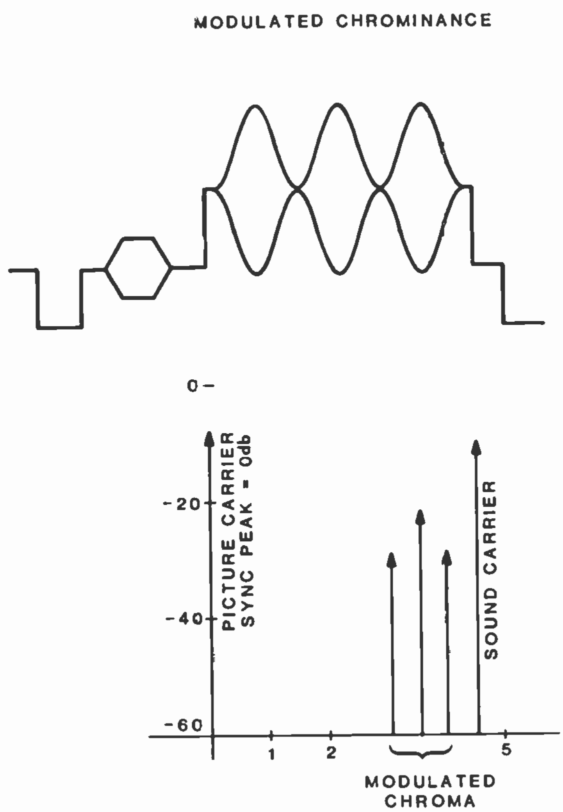


Figure 3

TYPICAL TRANSMISSION
AMPLITUDE CHARACTERISTIC
NEAR AURAL CARRIER

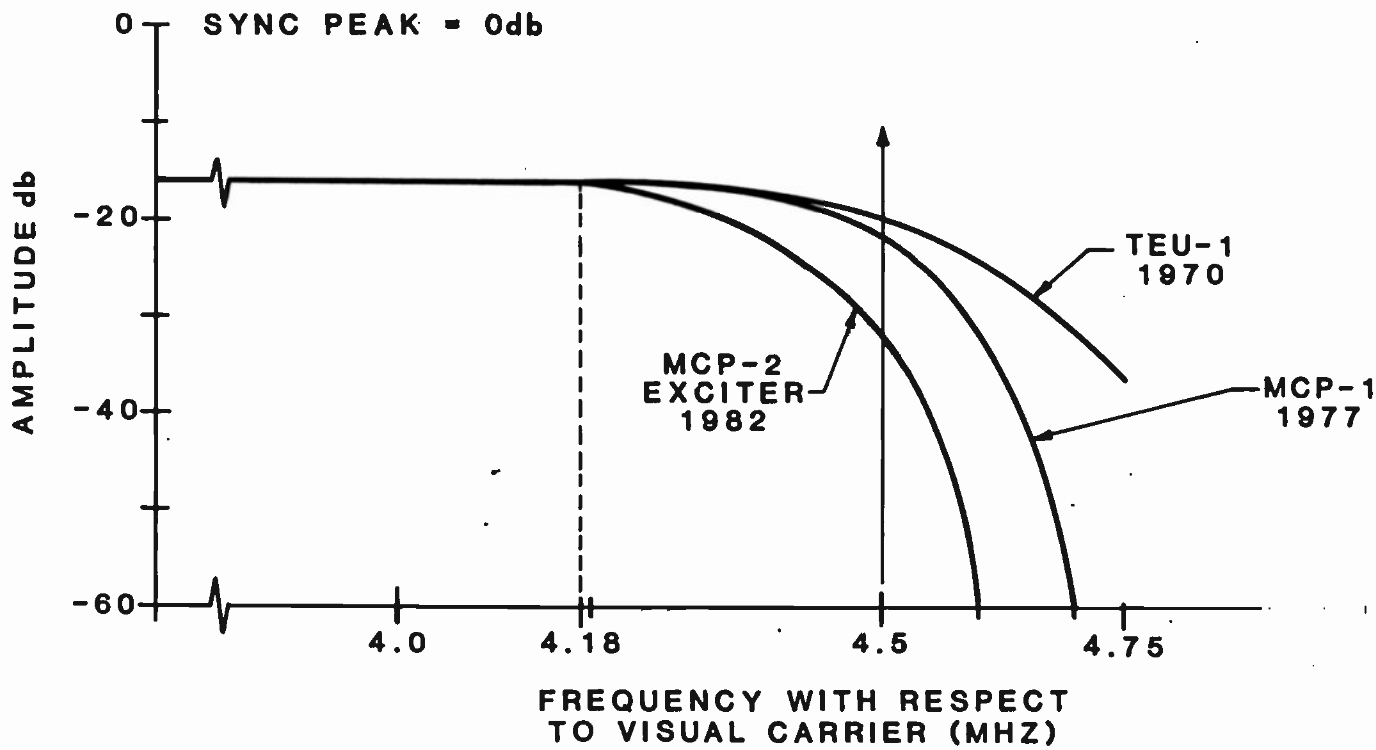
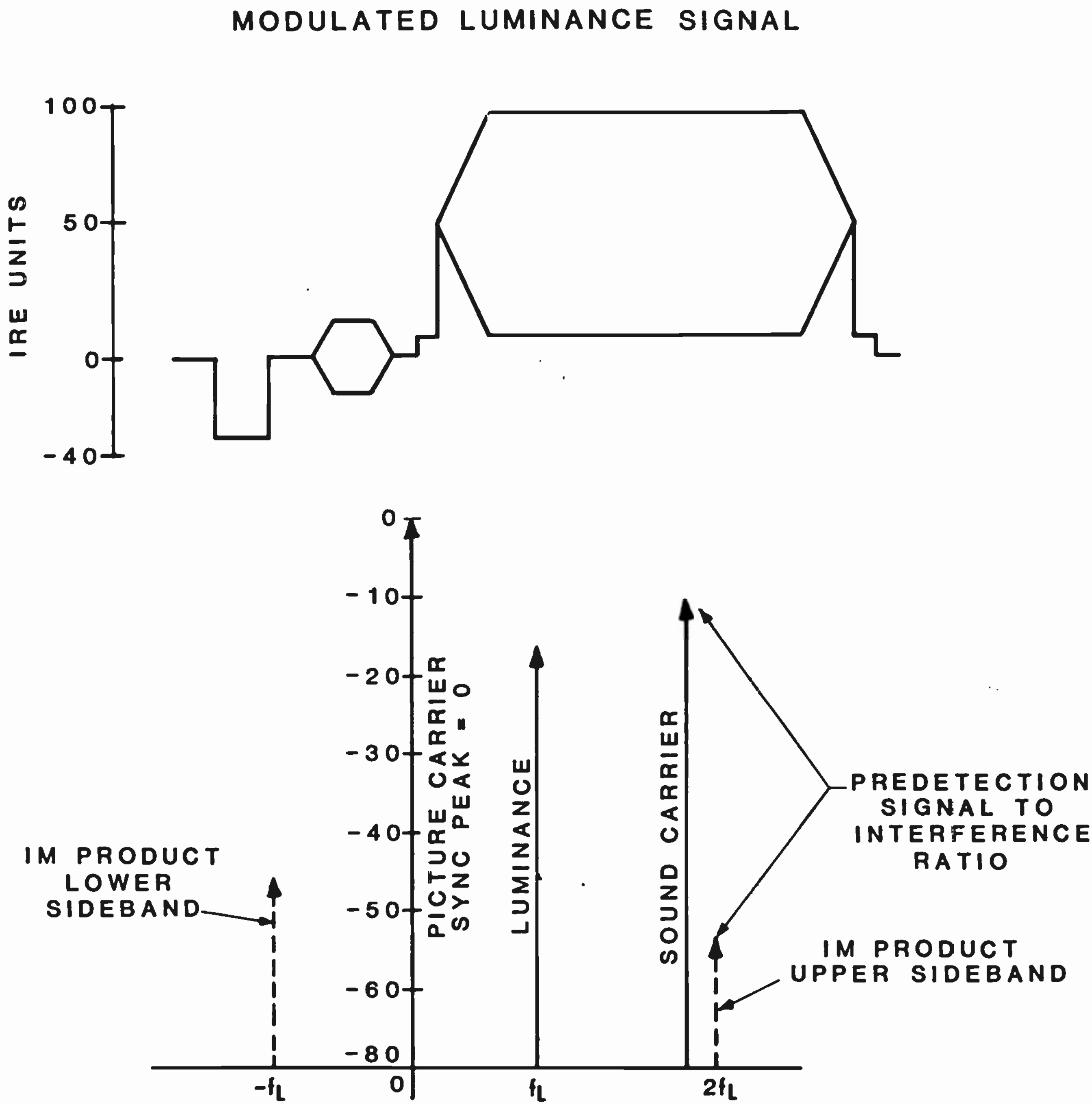


Figure 4



MODULATED LUMINANCE SIGNAL
100
50
0
-40
IRE UNITS
0
-10
-20
-30
-40
-50
-60
-80
0
PICTURE CARRIER
SYNC PEAK = 0
LUMINANCE
SOUND CARRIER
IM PRODUCT
LOWER
SIDEBAND
PREDETECTION
SIGNAL TO
INTERFERENCE
RATIO
IM PRODUCT
UPPER SIDEBAND
$-f_L$
f_L
$2f_L$

FIGURE 5

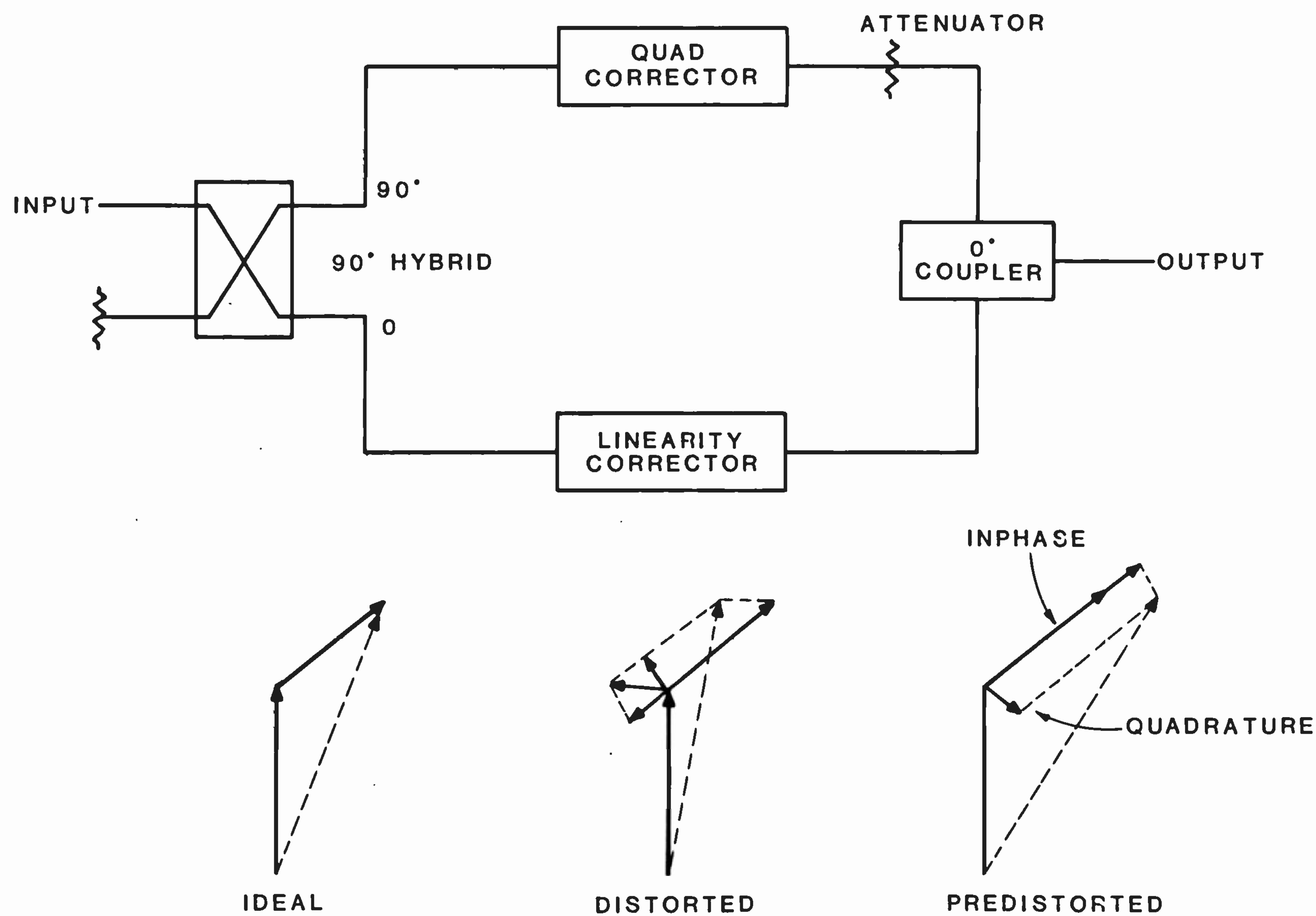

QUAD
CORRECTOR
ATTENUATOR
INPUT
90°
90° HYBRID
0
0°
COUPLER
OUTPUT
LINEARITY
CORRECTOR
INPHASE
QUADRATURE
IDEAL
DISTORTED
PREDISTORTED

FIGURE 6

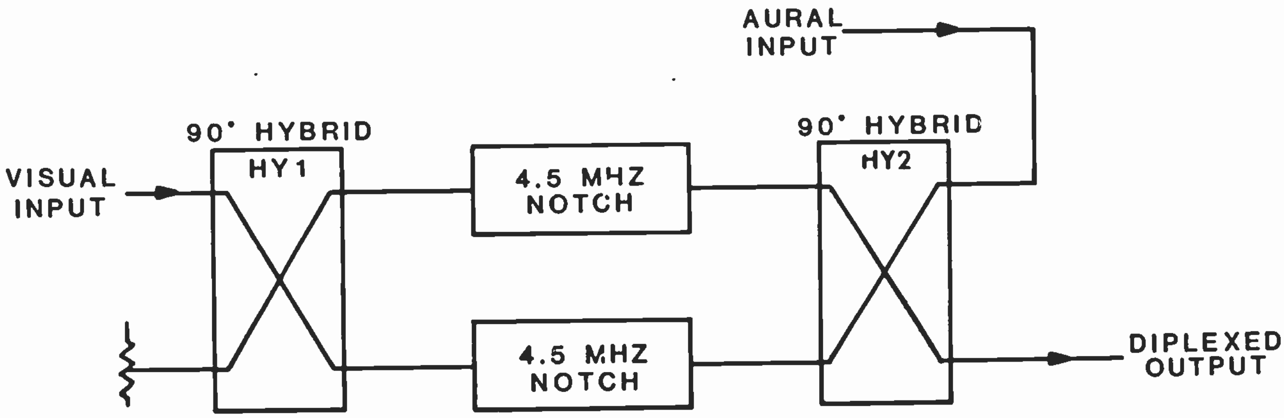


Figure 7

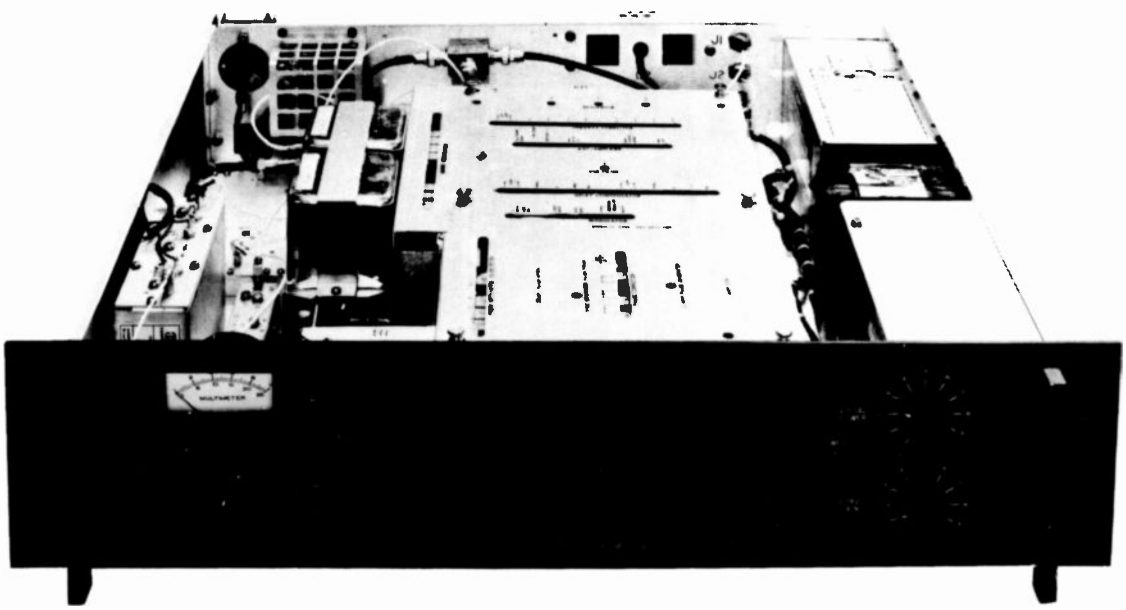


Figure 8

Post-Production - The Key to Stereophonic Audio for Television

Dr. Richard R. Green

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ABSTRACT

There is currently an increasing interest on the part of program producers in multichannel or stereo audio service for television. The program production technology and techniques to economically achieve this service are available today.

This paper is a discussion of the techniques which can be employed to produce television programs with stereophonic sound without inordinate increases in production budgets. The paper examines various types of production for television including musical variety, dramatic, electronic cinematography, sports and special events programming. The requirements for stereophonic audio for each of these program formats is examined. In the final analysis, it can be shown that the majority of the stereo audio requirements can be met during post-production. The incremental cost incurred in providing a stereo mix in post-production is small. Therefore, when possible, the cost of stereophonic sound can be minimized by using conventional audio production in the studio followed by a stereo post-production mixdown and layback to videotape. In addition, some experiments in the coverage of live events and sports in stereophonic sound are described. Some examples of cost reducing methods for these programs are also given.

OVERVIEW

The viewing public as well as program producers are expressing interest in multichannel or stereo audio services for television. Some of the important reasons for this trend are:

1. The wide popularity of stereo programs as broadcast by the majority of FM stations: These programs have acquainted listeners with the superiority of high fidelity stereo broadcasting. The current consumer concept of high fidelity sound is dual channel stereophonic reproduction.

2. The recent innovation into the market of video disc and video cassette recordings that provide high quality stereo audio: Such equipment will lead eventually to a public demand for high fidelity stereo equipped television services. Also, cable television operators add high quality stereophonic sound to TV programs by cable casting the accompanying audio in the FM band.

3. The rapid growth in the number of major motion picture releases with stereophonic sound tracks: When these features are broadcast on television, the medium should be capable of transmitting as far as possible the original effect of the sound track.

4. The successful operation of a TV multiplex sound system by a number of major Japanese broadcasting stations that now provide stereophonic programs. The rapid growth and popularity of stereophonic television in Japan is indicative of the potential for such service in the United States.

In Japan the stereophonic television multiplex system has been in operation for several years. TV stereo receivers are now in quantity production. In these integrated receivers, stereo speakers of the high fidelity type are located at opposite sides of the screen. Approximately 200,000 fully integrated stereo receivers have been sold to date. Monophonic TV receivers are equipped with terminals which permit an external stereo adapter to be connected. The adapter contains stereo amplifiers and two speakers. Independent, TV audio tuners are also available to enable the viewer to listen to the audio from a TV multiplex broadcast program. In short, the Japanese have established a successful service which has generated substantial viewer interest and attention and, consequently, manufacturer response.

In recent years, there has been an increasing interest on the part of the electronics industry and television broadcast organizations in stereophonic television. Issues related to the upgrading of our present television audio service are under consideration by the Broadcast Transmission Standard (BTS) Committee of the Electronics Industry Association (EIA). The committee is currently investigating three multiplex systems and conducting field tests in the Chicago area. Three proposed noise reduction techniques are also being tested and a final report on these studies is expected in the next few months.

THE TECHNICAL SITUATION

In the past several years, many aspects of production and distribution of television sound have been the beneficiaries of outstanding technical advances. For example, transmission facilities have improved and new studios have been equipped with the highest quality audio facilities including multichannel mixing consoles and multitrack recording facilities. Modern TV audio post-production facilities have grown in number and sophistication and now rival record industry production facilities in complexity and capability.

Network transmission quality has improved with the introduction of the diplex 15 kilohertz distribution system introduced by all networks in January 1978. In addition, digital transmission channels and satellite transmission of audio are adding significant improvements to the quality of audio. New one-inch video tape recorders provide high-quality stereo recording capability and are quickly becoming the standard in the television production industry. Specifically, the facilities for generation and distribution of high-quality stereophonic audio are now widely available within the television production industry and can be employed to produce high-quality stereophonic television programs as soon as broadcast standards are approved.

PRODUCTION GOALS OF STEREOPHONIC TELEVISION

The addition of stereophonic sound to television programs should accomplish two major objectives -- realism and controlled auditory localization.

1. Realism. Entertainment value of a television program can be increased by enhancing the feeling of realism and presence inherent in the addition of stereophonic sound. Stereophonic sound adds a new creative dimension to television production with which viewers' attention and interest can be heightened and sound effects properly localized.

2. Controlled Auditory Localization. While this is an ingredient of realism, it deserves special attention because it has a special meaning with respect to current television programming. It is important that the aural stimulus correspond to the visual stimulus at the presentation (viewer) level. If there is a distortion of this correlation, it may create a counter-productive form of conflict in the mind of the viewer. Sound effects which appear to be introduced far off the side of the screen may be confusing and distracting to the viewer. Similarly, panning of sound effects from extreme left to extreme right can provide an undesirably large audio vista without the corresponding visual stimulus.

Due to the geometric limitations of current video displays, stereo audio for TV should be simple (i.e., minimum audio panning and localization of dialogue and effects).

There are now available to us techniques which can be employed to produce television programs with stereophonic sound and achieve the two described objectives. In addition, it is possible to accomplish these ends without an inordinate increase in production budgets.

GENERIC CLASSES OF TELEVISION PRODUCTION

Let us now examine several types of television production and the requirements for adding stereophonic audio to these program formats. Major types of programs we will consider here are: a) musical variety; b) dramatic; c) sports; and d) special events.

Beginning with musical variety production, the original sound for such a production is usually prepared as a "multichannel prerecord". That is, the musical elements are prerecorded in a sound studio on multitrack audio recorders in an environment optimized for sound recording. The multichannel audio tape is then reproduced during the television production so that artists can "lip sync" vocal performances to the recorded music. Television choreography is, of course, performed to the prerecorded multitrack audio. The final program audio is not prepared until the post-production phase in which the final mix from the prerecorded tapes is produced, synchronized to the video recording and transferred to the video tape in the final "layback".

The significance of this process is that the decision to produce either a mono mix or a stereo mix is not made until the post-production phase of program preparation. Post-production facilities for television program audio are very sophisticated and mixing consoles are as elaborate as record industry production facilities. It is, therefore, possible to provide at a very minimal incremental cost a stereo track derived from the multichannel prerecorded audio used in most musical variety programming. Production audio or studio audio recorded during the television taping need only be recorded in monophonic sound and simply introduced as an equally balanced element in the stereo mix. In this way no additional costs are encountered in the studio production. This approach permits all the time-consuming mixing to be done in the post-production facilities at minimum cost rather than during the studio production in which production time is very expensive. The cost of stereophonic programs can be controlled by production planning which emphasizes the post-production techniques and processing.

In our experience at CBS, we have found that dramatic production employing stereo audio must vary only slightly from monophonic production in order to avoid non-productive forms of viewer audio conflict. Specifically, it appears that for television it is wise to retain the dialogue and studio production audio in monophonic form and to introduce music and effects in a final stereo mixdown. Again, studio production remains unchanged from present techniques and employs recording of a single channel of mixed audio from microphones on the sound stage. Music and effects are then added as part of the post-production process. Stereophonic music which accompanies dramatic action can be mixed into the final stereo track and sound effects can be added by panning the effect for appropriate positioning in the action. Some off-screen effects or dialogue can be used as a dramatic device but in general most sound effects introduced in the center of the audio field are quite acceptable. Large amounts of panning of effects or dialogue in most production situations prove to be distractive.

CBS has experimented with sports programming using stereophonic sound. Two types of programs, a football game and a golf tournament, were produced. In both cases, very pleasing stereophonic programs resulted using standard television audio coverage provided by the commentators. Stereophonic crowd reactions were obtained by placing microphones separated from each other by a small distance, and these microphones provided the major contribution to stereo effect in the final mix. Some special audio pickups such as parabolic microphones to record specific sounds (e.g., crashing helmets, or striking of a golf ball) were added to the mix as a monophonic element. In both programs, very successful stereophonic production was achieved at very low incremental costs.

The case of special events programming (i.e., special events that would normally by telecasted live), CBS has experimented using normal monophonic coverage enhanced with specific live stereophonic effects. In one test, the coverage of the fall 1980 Kennedy Center Awards Presentation in Washington, D. C., the final stereophonic track was developed from three sources:

1. A monoaural track which featured the principal performing artist.
2. A stereo mixdown of the orchestra during the live performance.
3. Stereophonic ambiance produced by audience reaction and hall reverberation.

The program presented striking stereophonic impressions without the distraction of the artist moving in the stereo field. It seemed to provide the best auditory localization for the viewer as the artist was always where the viewer expected him to be even though camera angles varied widely. The incremental cost in producing this program in stereo was very small as microphone placements, except for audience separation, were identical to those customarily used for a monophonic broadcast. A live mixdown to two stereophonic channels was accomplished with no significant additional complexity or cost.

CONCLUSIONS

Three major conclusions can be drawn from analyzing the requirements for stereo audio for major types of television production. They are:

1. Most prime time programs are high value productions involving a post-production phase. Economic factors will tend to favor production of a stereophonic mix during the post-production process. Studio time is very expensive and the original audio from the sound stages must be recorded in the least expensive way. The time-consuming subtleties and quality control for the final stereo product can best be achieved in the less-expensive post-production facility.

2. Most dramatic series, including afternoon dramas, now include a post-production phase. In this case, a high-quality stereophonic production can be obtained by using the present monophonic production or dialogue track combined with stereophonic music and effects in a two-channel mixdown in post-production. If necessary, some dialogue elements can be panned to selected positions on the screen. However, it is apparent from initial tests that for television a minimum movement of the dialogue in the stereo field is preferable. Localization of sound effects can be used as a dramatic device but effects centered in the stereo field are generally accepted by the viewer as appropriately placed.

3. In programs which do not normally employ a post-production phase, e.g., sports and special events, high-quality, and yet, economical, stereo audio can be developed using monaural channels featuring the artists or commentators and stereo channels capturing the ambiance from audience reaction or musical accompaniments and the environment in which the event is being performed.

In the final analysis, stereophonic audio can be added with minimum incremental production costs for most types of television programs.

INTEGRAL CAVITY KLYSTRON

VISUAL COUPLER PERFORMANCE UPDATE

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RCA Corporation

Meadow Lands, PA 15347

ABSTRACT

The Varian Visual Coupler introduced at the 1981 NAB will have an impact on the efficiency of today's UHF TV transmitters. Production units of the visual coupler have been delivered and tested in RCA UHF transmitters. This paper describes the results of these tests and outlines the performance expectations in terms of signal quality and efficiency. In addition, the mechanical and thermal design considerations required to incorporate the visual coupler into a UHF Television Transmitter are explained.

Since the introduction of the Varian Visual Coupler and the initial performance test data, the final design of the coupler has been completed and the initial production units have been tested in UHF Television Transmitters. This paper presents the latest test results using one of these production couplers and describes the mechanical interface of the visual coupler into a standard RCA UHF Television Transmitter.

To better appreciate this new data, a brief review of the Varian visual coupler should be helpful.

One method by which klystron efficiency can be improved is by reducing the DC beam current and raising the RF load resistance across the output gap. Although this affects the klystron's frequency response, the resulting response is normally acceptable.

In external cavity klystrons, the RF load resistance presented across the output gap can be changed by withdrawing or rotating the output loop. Since the coupling loop in integral cavity klystron is generally fixed, this adjustment is not possible. However, it is possible to raise the output load resistance by transforming the RF system load impedance (usually 50 ohms) to a higher value by using a visual coupler (transmission line transformer), as illustrated in Figure 1.

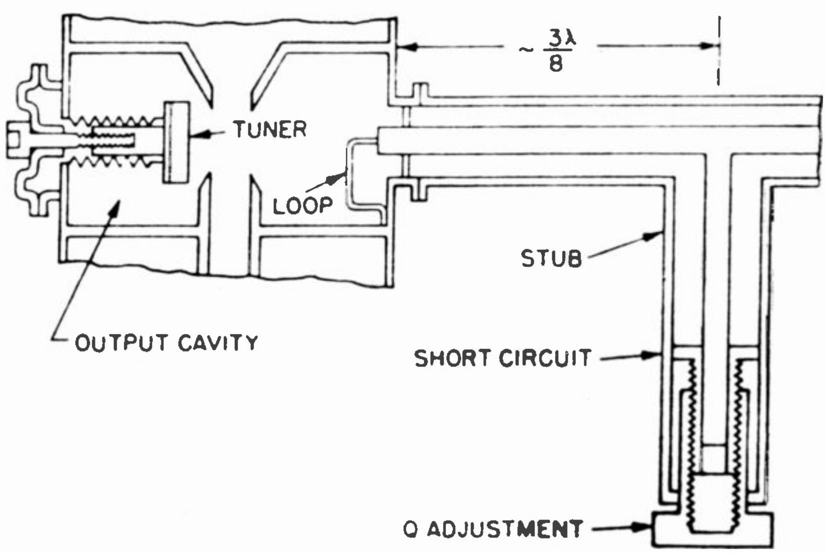


Figure 1. Schematic Diagram of Klystron Output Cavity and Variable Coupler Transformer

The shorted stub length is normally set slightly longer than 1/4λ . This produces a capacitive susceptance in shunt with the transmission line which transforms to a nearly resistive impedance 3/8λ away at the klystron's output port. This resistance value depends on how much the stub has been lengthened beyond 1/4λ . This is shown on the Smith Chart in Figure 2.

The production visual coupler was mounted on . 55 kW klystron as shown in Figure 3. The mounting brackets used to support the coupler are shown in Figure 4. Mounting kits are available for all 30 and 55 kW high efficiency Varian integral cavity klystrons. Note the water connections that provide cooling to the coupler's center conductor. The water is supplied by placing the coupler's cooling lines in series with the magnet cooling water.

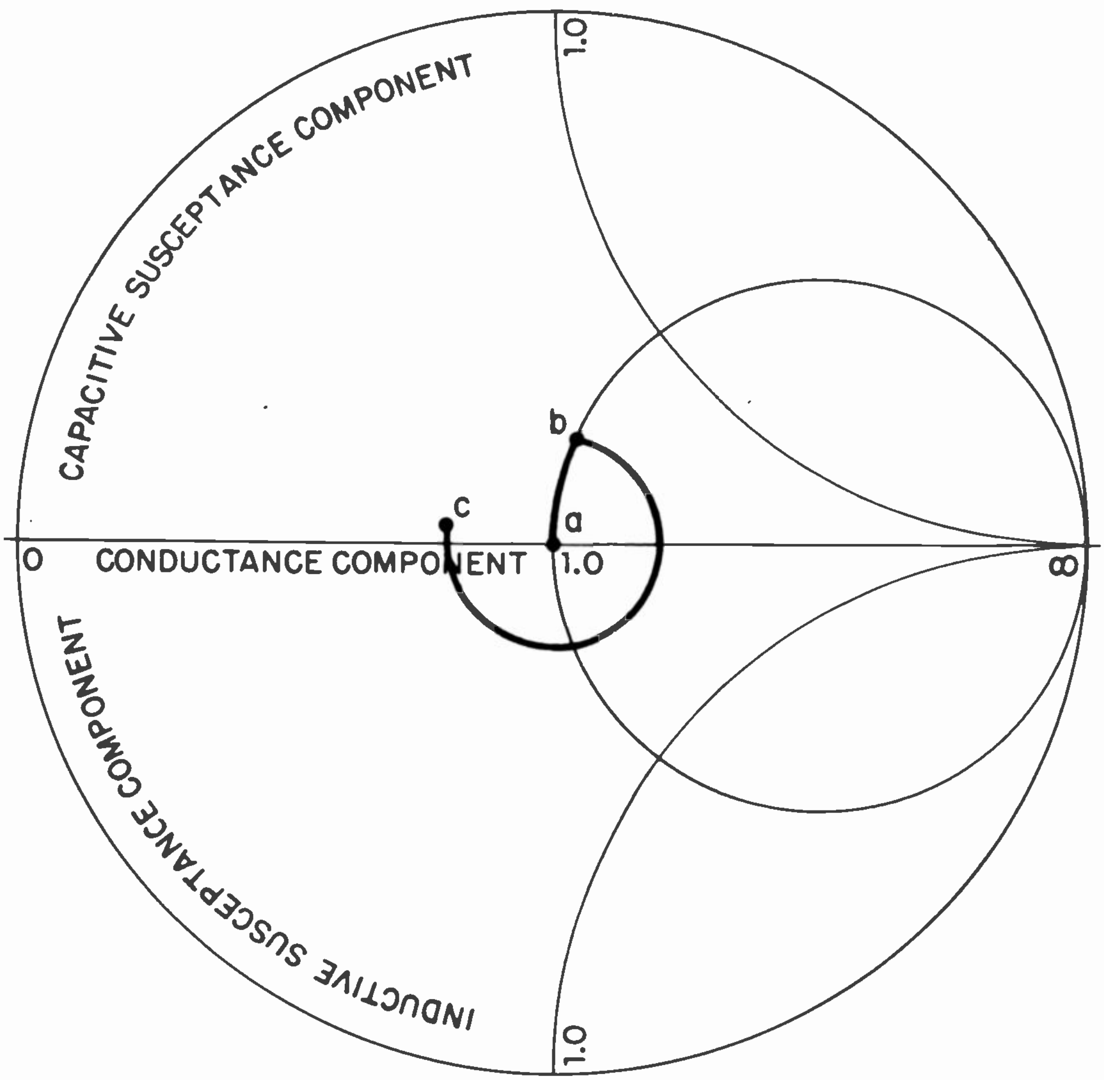


Figure 2. Admittance Chart for Decreased Loading:
(a) Conductance of Output Transmission Line
(b) Conductance A Shunted by Susceptance of Capacitive Stub
(c) Admittance Transformed by $\frac{3\lambda}{8}$ Section as Seen by Klystron

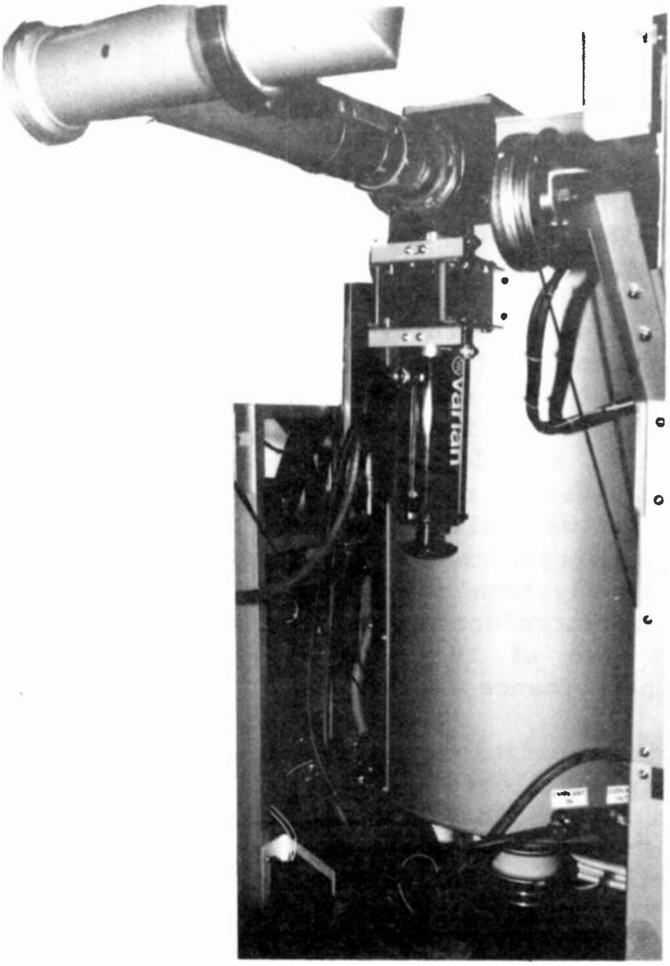


Figure 3.

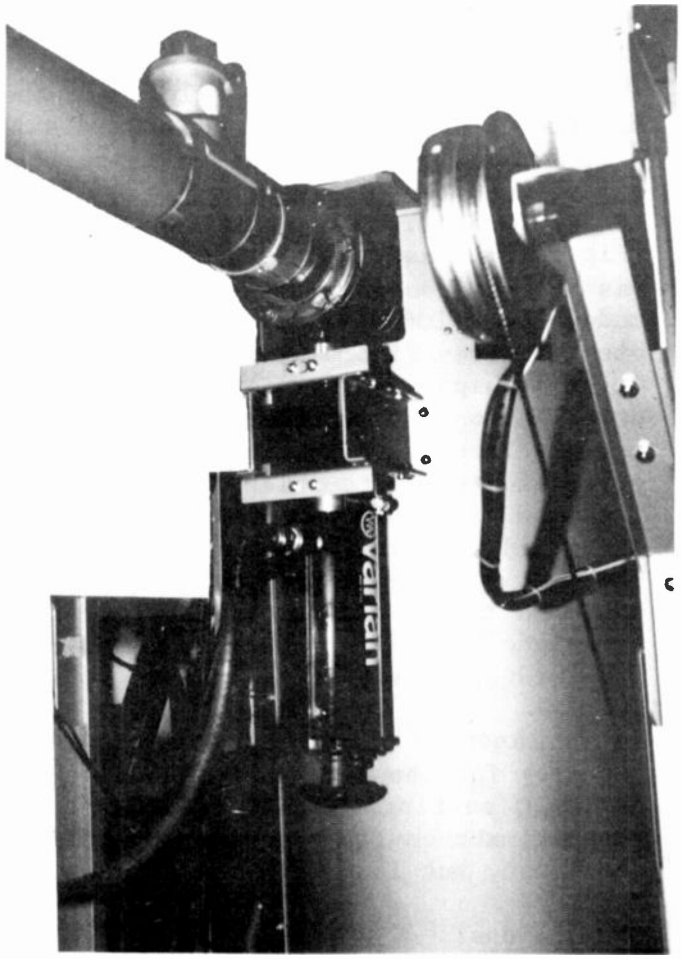


Figure 4.

Since the dimensions from the flange to the stub are frequency dependent, the exact configuration of the coupler varies across the UHF band. However, its overall length from input to output changes very little across the full band.

The resultant test data was taken from an RCA TTU-55C transmitter while in final system test on Channel 29. The transmitter was first tuned without the visual coupler and a standard set of performance data was obtained. Next, the coupler was installed and the system tuning and beam current were optimized for best results.

It should be noted that video performance was adjusted and measured to meet the basic RCA transmitter specifications, with headroom. Figures 5 through 7 illustrate the performance achieved. These levels of performance are readily achievable results rather than best possible efficiency and performance and should be typical of the results UHF stations would get in the field.

Figures 8 through 10 illustrate the results obtained when the coupler was adjusted past the optimum point, raising the output load impedance and cavity Q to values that resulted in instability and poor signal quality. The first effect that becomes apparent is a lack of saturated power capability. In the RCA transmitter, this results in excessive leading and trailing sync spiking (Figure 8) caused by broadband incidental phase modulation. In these tests, it was always possible to correct the lower video frequency incidental phase modulation (ICPM) errors with the transmitter's correctors. However, it was observed (see Figure 9) that there were broadband ICPM products that had phase relationships to the correction products that did not result in adequate cancellation. The frequency response of the relatively high Q input and intermediate cavities introduces an amplitude and delay distortion to the correction products which exist in the lower sideband region (see Figure 5,9). The non-linear phase and amplitude distortion mechanism delay and amplitude distorted correction products reaching this section cannot completely cancel the corresponding products generated by the klystron non-linearity. It is believed that a circuit could be designed to introduce the complement of the amplitude and delay versus frequency response of the klystron input and intermediate cavities to achieve a further performance improvement.

Another limitation in the increase in efficiency attainable is the tendency for the klystron to develop sync tip oscillations at very high output cavity Q settings. This is illustrated in Figure 10. Careful tuning of the pentultimate cavity and fine adjustment of the magnet current can improve this condition, but long term operation near this limit is not practical.

Conclusions

Adding the Variable Visual Coupler to a UHF-TV transmitter using integral cavity klystrons will typically enhance efficiency by about 6 percentage points. This applies to both pulsed and non-pulsed systems. The signal quality obtained meets normal broadcast standards and with the escalating costs of electricity, the incorporation of this new device can normally be easily justified by the efficiency improvement. Improvements in the design of ICPM correctors may further increase the efficiency performance obtained.

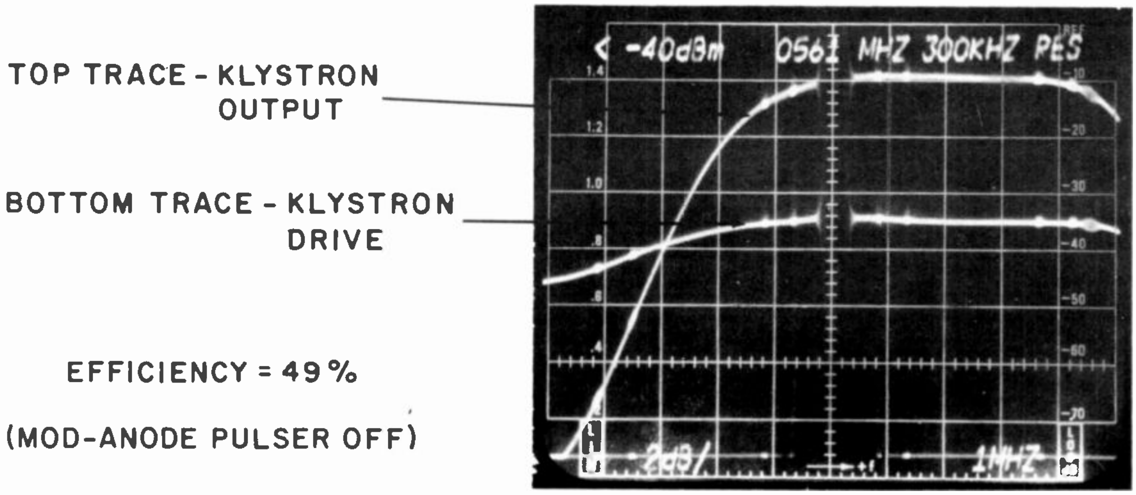


Figure 5. Frequency Response

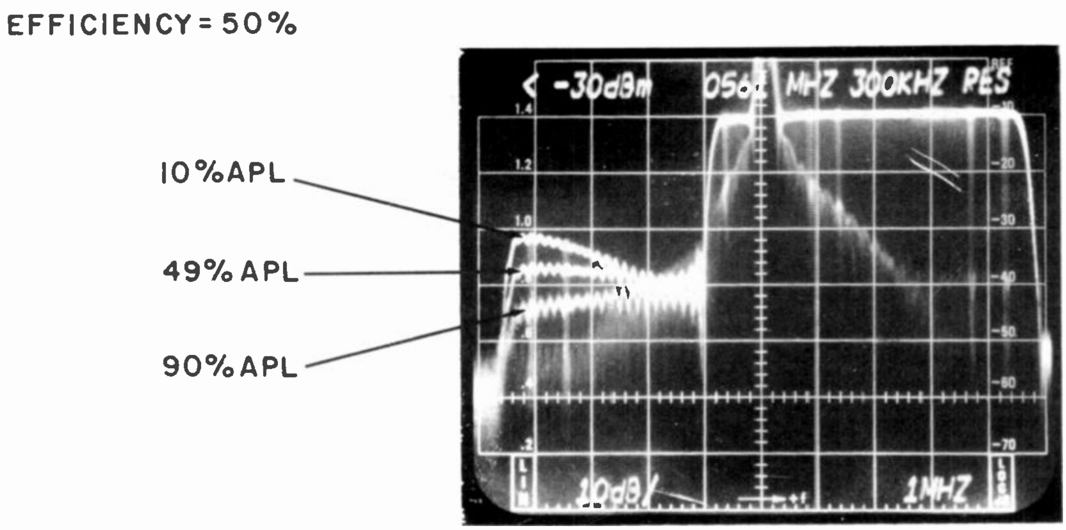
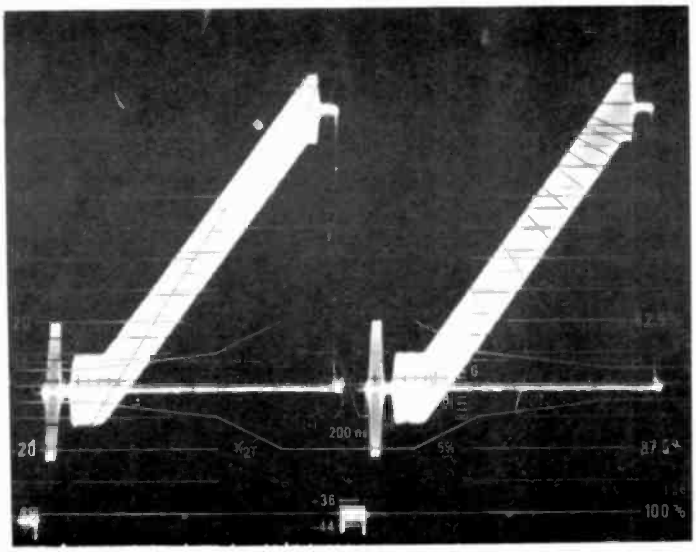
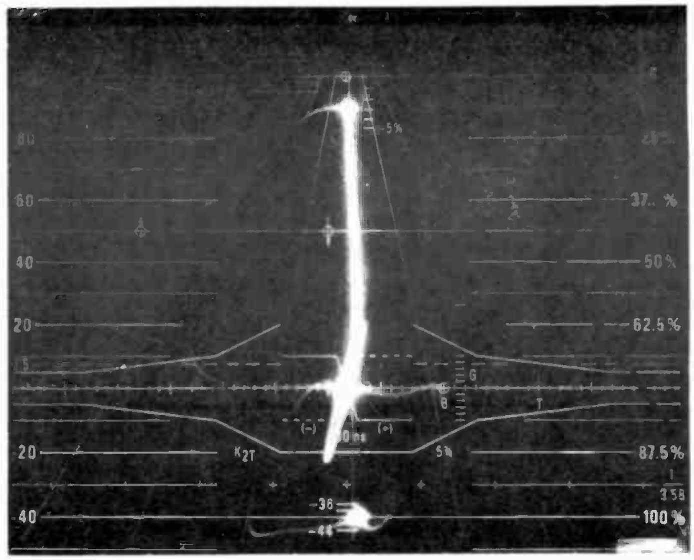


Figure 6. Lower Sideband Reinsertion

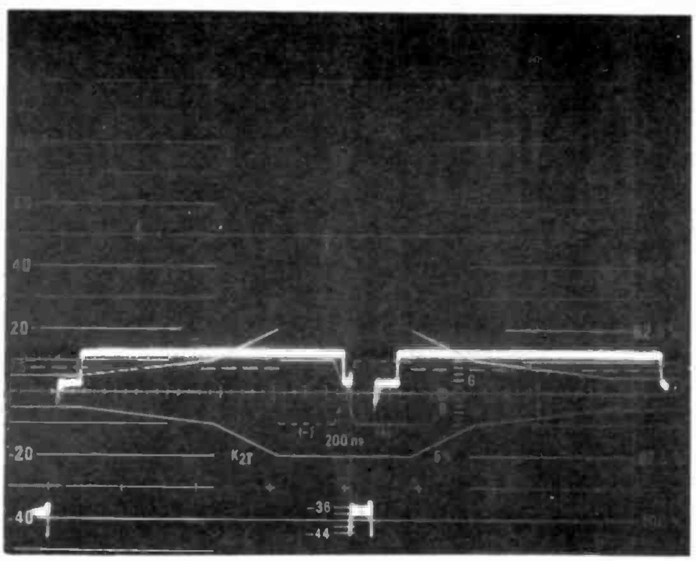


Waveform
(49% Eff.; Non-pulsed)



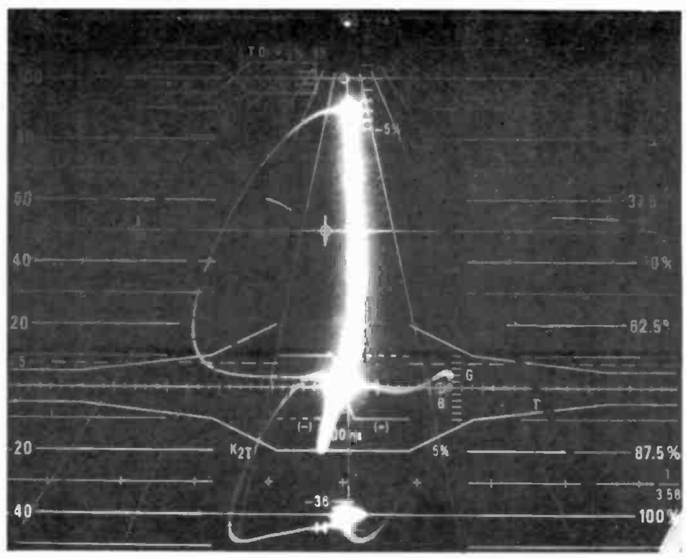
Incidental Phase

Figure 7. Linearity



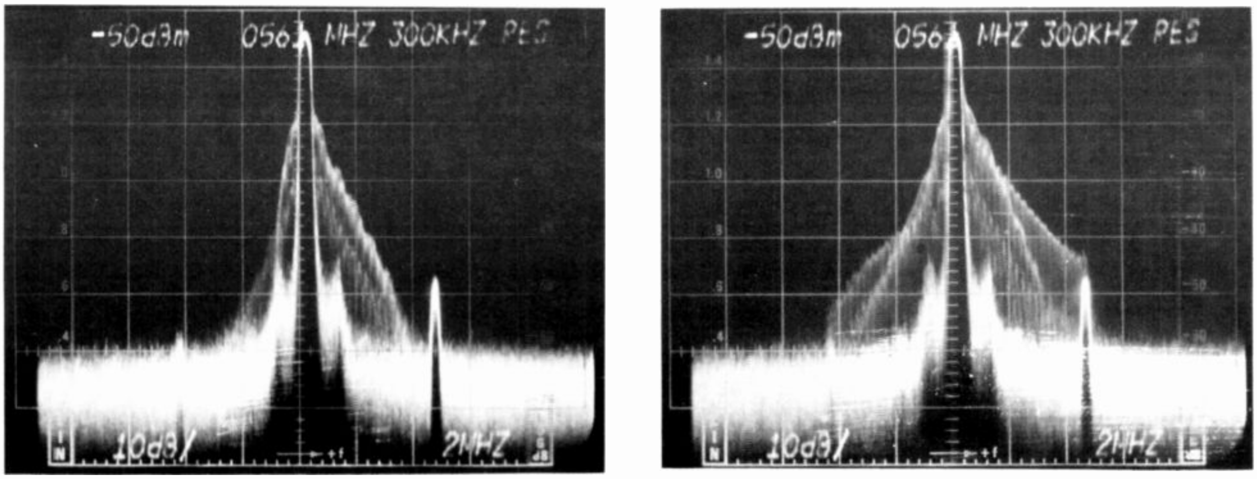
Waveform

(54% Eff.; Non-pulsed)



Incidental Phase

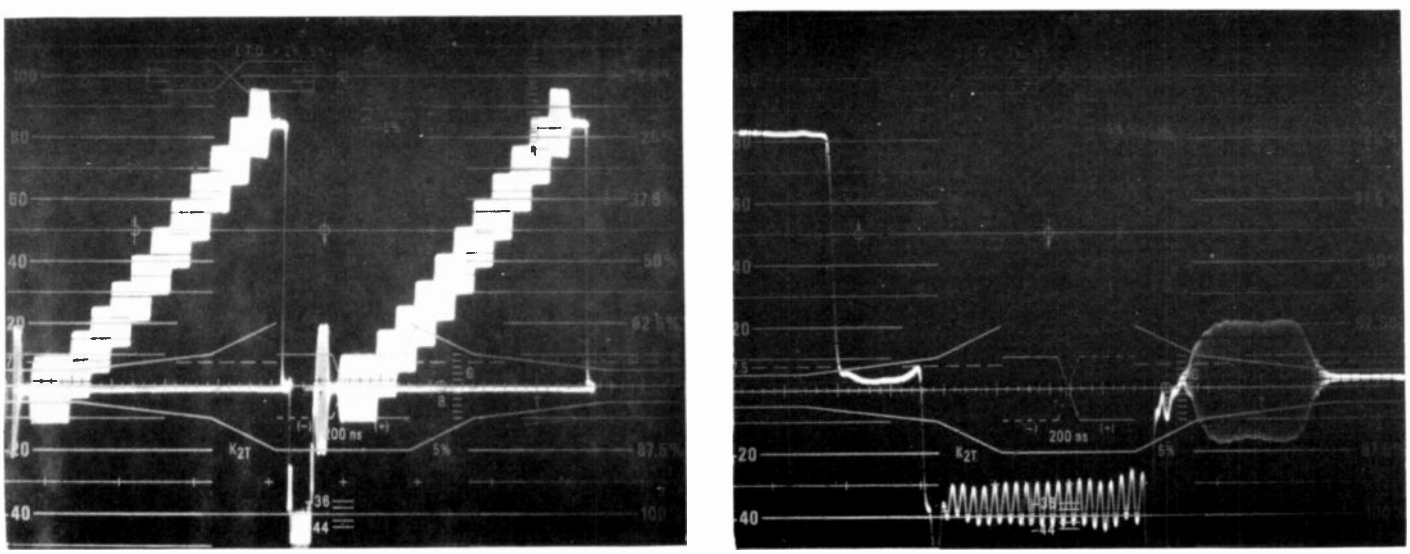
Figure 8. Linearity



Normal VSB Spectrum

VSB Spectrum With Broadband Incidental Phase Modulation

Figure 9



Waveform

Sync Pulse

Figure 10. Sync Oscillation

DEPRESSED COLLECTOR KLYSTRONS FOR

HIGH-EFFICIENCY UHF TELEVISION

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Many of the advantages of high power klystron amplifiers derive from the fact that the electrons are bunched by velocity modulation after they are traveling with a high velocity and have a large amount of kinetic energy. This permits the use of large rugged structures, big cathodes and big low-power-density surfaces for collecting the spent electron beam, and leads to reliable long-life tubes which will be tolerant of mishandling. Conventional tubes must use very small, delicate structures at UHF and microwave frequencies. Unfortunately, in UHF television final output tube service, a substantial fraction of the energy in the klystron electron beam is converted into rf only when the tube is delivering the maximum output power during a sync pulse. When lower output powers are required, for example when white level is being transmitted and the voltage is only 12.5%, or the power about 1.5%, of peak-of-sync, electrons are not slowed down by a retarding rf voltage in the output gap but proceed into the collector and dissipate almost the full beam power in the collector. Cutler [1] has calculated that a klystron operating with 40% peak-of-sync efficiency will actually have a true efficiency, that is, average output power divided by average input power, of only 14% averaged over a typical television signal with an average gray raster.

The maximum efficiency of klystrons can be improved by perveance reduction, use of inductively-tuned third and fourth cavities, and drift length optimization. By using all of these possibilities for improving the interaction of the electron beam with the radio frequency field in the cavities, one can increase the klystron efficiency at peak-of-sync to about 60%. The resulting average efficiency with visual signal modulation is then increased to about 21%. Tubes with this kind of performance will soon be available. Still, 79% of the input power to the transmitter will continue to be wasted.

There is something that can be done about this. For a number of years the klystrons and traveling wave tubes have been built using multistage depressed collectors. In these, each stage or electrode is connected to one of a series of potentials between ground and the cathode potential of the klystron. For example, in a five stage depressed collector the first collector electrode is connected to ground potential, the second electrode to a potential below ground by an amount equal to 25% of the cathode potential; the third stage to 50%; the fourth stage to 75%; and the fifth stage is tied back to the cathode supply. Each collection electrode is supplied from a rectifier and filter capacitor connected to an appropriate tap on the beam supply transformer as shown in Figure 1. All electrons enter the output gap of the klystron with the same energy, corresponding to full beam voltage of the klystron. Depending upon the rf drive furnished to the klystron, the output rf gap voltage will be either large or small and will slow down electrons during part of the rf voltage cycle and speed them up during the other part. When the klystron output power is low there will be only a small change, either positive or negative, in the energy of the electrons. When the output voltage is large, the energy spread will be large. For example when the rf output gap voltage is half of the beam voltage, beam electron energies from half the beam voltage to one and one half times the beam voltage will be present in the beam as it exits from the output gap. When the rf output gap voltage is almost equal to the beam voltage, which is the situation at peak-of-sync, electron energies from zero to up to twice the beam voltage will be present as the beam exits from the gap. There are usually only a small number of electrons with energies greater than the beam voltage because the bunch of electrons which contains most of the electrons passes through the gap during the time when the field is retarding. Conversely, most of the electrons have energy between the beam voltage and zero. If the geometry of the multistage-depressed collector is designed very carefully so that an electron will reach the electrode which has a potential just sufficient to collect it but will not return to an electrode with a higher potential, the saving of power is equal to the summation of products of each electrode current and the potential depression of that electrode, or

$$(\text{Power Saving}) = \sum_{i=1}^{n} I_i \, (V_o - V_i)$$

in which V_o is the beam voltage and V_i is the potential of the ith collector electrode with respect to the cathode. For example, in the case of the five-stage collector mentioned previously, electrons which arrived at the output gap when the electric field is accelerating will be able to reach the fifth-stage electrode at cathode potential. The dissipation on this electrode, when compared to the dissipation on a grounded electrode, will be less by the product of the beam voltage and the intercepted current. For the fourth-stage electrode which is depressed below ground by 75% of the beam voltage the power saving will be the product of the current arriving at that electrode and 75% of the beam voltage, and so forth, so that

$$(\text{Power Saving}) = 0.25\, V_o \cdot I_2 + 0.5 V_o\, I_3 + 0.75\, V_o\, I_4 + V_o\, I_5 .$$

Note that because I_1 arrives at an electrode which is not depressed there is just as much dissipation due to this current component as there would be in a conventional klystron.

As the rf drive level to the klystron is changed, the fraction of the beam current arriving at each collector electrode will change, but the potential of each electrode can and should remain constant. This is why integrating filter capacitors are included in the circuit of Figure 1.

The currents arriving at each electrode can be estimated by assuming that the instantaneous beam current is equal to the dc beam current plus a sinusoidal component proportional to the input cavity voltage and that the output gap voltage is sinusoidal and proportional to the rf beam current.

Table 1 shows such a theoretical calculation of the power that would be dissipated in the collector of a klystron as a function of the number of stages in that collector (see Appendix for details). The efficiency with which the klystron converts beam power into rf power is assumed to be 55% at the maximum power output, and the probability that the power output is any value between zero and maximum is assumed to be constant. This is not too different from the situation in television broadcasting. Also shown as a function of the number of stages in the collector is the average output power divided by the average input power (or the true efficiency) and the peak output power divided by the average input power which is analogous to what is sometimes referred to as peak-of-sync efficiency of a television transmitter. Notice that most of the saving in power has occurred by the time a five-stage collector is used. So you see, it is not entirely accidental that a five-stage collector example was referred to earlier. Both the true efficiency and the peak-of-sync efficiency are becoming quite impressive and can exceed those available from any other kind of amplifier which might be used in television service. By the use of beam current pulsing in conjunction with depressed collectors, power input can be reduced by another 20% with a further increase in the theoretical efficiency attainable. In practice these theoretical efficiencies have been approached, but usually some electrons will go to electrodes with higher than minimum necessary potential (earlier stages) and primary beam electrons will generate secondary electrons which will be accelerated to higher potential electrodes and will waste energy.

Table I

No. of Collector Stages	$\frac{\text{Collector Power}}{\text{Beam Power}}$	True Efficiency	Peak-of-Sync Efficiency
1	0.79	21%	55%
2	0.48	30%	80%
3	0.26	44%	117%
4	0.19	52%	138%
5	0.16	56%	148%

A number of traveling wave tubes and klystrons have been built using depressed collectors [2-4], but these have always been rather low power devices, typically used in space applications where there has always been a premium on power conservation. Some of these tubes have been developed for direct satellite TV although one 10 kW TV experiment has been conducted with some success. Only recently has the cost of energy increased sufficiently in the United States to make worthwhile the serious consideration of the development of a high-power television final amplifier tube using this technology. It is not altogether trivial to scale depressed collectors for use on large, very-high-power television tubes. As mentioned previously, it is necessary to shape the collection electrodes so that electrons are collected by the electrode with the lowest possible potential capable of collecting these electrons. In addition, secondary electrons which can be accelerated to higher potential collection electrodes must be dealt with. While we have impressive capability for simulating problems in electron optics such as these on digital computers, a great deal of experimentation has gone into the development of small depressed collectors. And such experimentation becomes more and more expensive as the size of the tube is increased. For example, the problem of insulating and cooling the depressed collection electrodes in large, high-power tubes has not yet been addressed fully. In other words, even though the feasibility of depressed collector tubes has been demonstrated, there is considerable technology that must be developed in order to apply the principles to television tubes. It is not yet possible to write a specification for a depressed collector tube without a considerable development program first. However, it is reasonable to assert that a large part of the theoretical savings listed in Table I can be achieved.

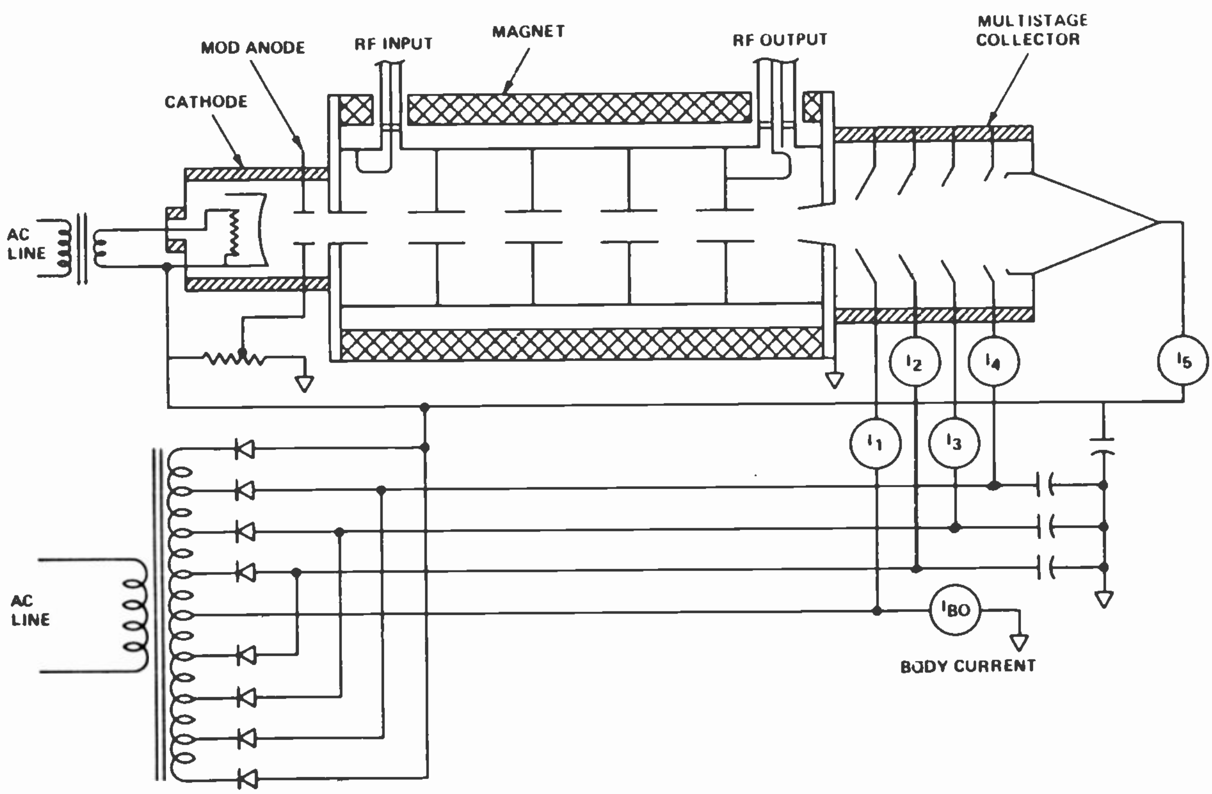

FIGURE 1. SIMPLIFIED SCHEMATIC FOR HIGH-EFFICIENCY, DEPRESSED-COLLECTOR-KLYSTRON UHF-TV FINAL AMPLIFIER

APPENDIX

Most of the time a television klystron is operating well below saturated output power, so the beam current entering the output gap can be represented by

$$I = I_o \,(1 + 2\,A\eta \cos \omega t) \tag{1}$$

in which I_o is the dc beam current, η is the klystron efficiency when $A = 1$ and A is the fraction of full rf output voltage the klystron is producing. V, the change in electron voltage resulting from the electron interacting with the output gap fields will be

$$V = -\,AV_o \cos \omega t \tag{2}$$

If we solve equation (2) for ωt and differentiate we obtain

$$\omega t \cong \cos^{-1}\left(-\frac{V}{A\,V_o}\right) = \cos^{-1}\left(-\frac{\bar{V}}{A}\right) \tag{3}$$

$$\frac{d\,(\omega t)}{d\bar{V}} = [A^2 - \bar{V}^2]^{-1/2} \tag{4}$$

in which $\bar{V} = V/V_o$ is the difference between the voltage corresponding to the energy of an electron entering the collector and the beam voltage. Combining equations (1) and (2) yields

$$I = I_o\,[1 - 2\eta\bar{V}] \tag{5}$$

Equation (5) states that the magnitude of the current decreases linearly with electron energy. Equation (4) tells us for how large an angular increment the electron energy dwells in any energy increment. If these two equations are multiplied together and divided by π which is the angle in which the energy assumes all possible values, we obtain an expression for the part of the current in each energy increment.

$$dI = \frac{Io\,[1 - 2\eta\bar{V}]}{\pi\,[A^2 - \bar{V}^2]^{1/2}}\,d\bar{V} \tag{6}$$

This current can reach an electrode with any normalized potential $\bar{V}_i = V_i/V_o$ with respect to the cathode greater than $-\bar{V}$ and will strike it with an energy corresponding to the normalized voltage $\bar{V} + \bar{V}_i$. The increment of power delivered to an electrode by this current, normalized to the beam power, will be

$$d\overline{P} = \frac{dP}{I_o V_o} = \frac{[1 - 2\eta \overline{V}] [\overline{V} + \overline{V}_i]}{\pi [A^2 - \overline{V}^2]^{1/2}} dV \tag{7}$$

the total power dissipated on the ith electrode will be given by the integral of (7)

$$\overline{P}_i = \int_{V_a}^{V_b} d\overline{P} = \frac{1}{\pi} \left\{ (\overline{V}_i - \eta A^2) \sin^{-1}\left(\frac{\overline{V}}{A}\right) - [1 - \eta(2\overline{V}_i + \overline{V})](A^2 - \overline{V}^2)^{1/2} \right\} \Bigg]_{V_a}^{V_b} \tag{8}$$

In an ideal multistage depressed collector the ith electrode will collect all the electrons which have enough energy to reach it but not so much they can reach the jth electrode at a lower potential. Hence the integration for the ith electrode is performed from $V_a = -A$ or $V_a = -\overline{V}_i$ which is even less negative to $V_b = \overline{V}_j - \overline{V}_i$ except where i is the last electrode of lowest potential in which case the upper limit of integration is $V_b = +A$.

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4. H. G. Kosmahl and P. Ramins "Small-Size 81 to 83.5 - Percent Efficient 2- and 4-Stage Depressed Collectors for Octave Bandwidth High Performance TWTs," IEEE Trans on Electron Devices, Vol. ED-24, No. 1, pp 36-44, Jan 1977.

COST EFFECTIVENESS AND EFFICIENCY OF CIRCULAR WAVEGUIDE FOR UHF-TV

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MANCHESTER, NH

INTRODUCTION

The efficiency of the RF System is as critical as the transmitter efficiency.

The RF System is a passive system and can only contribute a loss to the overall system. "Less loss", although sounding contradictory, is, in fact, a net gain.

"Less loss" will mean:

(a) More power delivered to the antenna

1. Lower antenna gain for a fixed ERP
2. Higher ERP for a fixed antenna gain

(b) Lower operating costs

1. Power dissipated in the transmission line represents dollars that were spent to generate the power initially.

Almost all of the new UHF systems are using tall towers and larger transmitter powers. Transmission line runs of 1,000, 1,500, and 2,000 ft. are common place. Transmitter power of 55 kW, 110 kW, and 220 kW are most often used.

The power dissipated in a coax transmission line can be as high as 50 - 70 kW. By using circular waveguide a new type of transmission line power savings of 40 kW can be realized.

CIRCULAR WAVEGUIDE

Coax line has been used exclusively for UHF TV until the mid seventies. With the increase in power from 55 kW to 110 kW and finally to 220 kW transmitters, it was necessary to increase the size of transmission line from 6-1/8" to 8-3/16" and 9-3/16" coax line. Because of higher order moding the higher UHF channels cannot use coax line. The EIA has recommended the upper frequency limits on coax of:

Coax Line Size	Limit Upper Freq.	Channel
6-1/8-75Ω	919	70
8-3/16-50Ω	613	38
8-3/16-75Ω	689	50
9-3/16-50Ω	543	26
9-3/16-75Ω	611	37

It is therefore necessary to use waveguide for the most high power installations.

EFFICIENCY

Rectangular waveguide has been used for some time, but it has wind load problems. Circular waveguide was developed to overcome this limitation and to take advantage of improved efficiency available with the circular cross section. The efficiency of circular waveguide is significantly better than coax or rectangular waveguide. This can be seen in Figure 1.

The efficiency and power dissipated in various transmission lines for a 1,000 and 2,000 ft. tower with transmitter power of 55 kW, 110 kW, and 220 kW are shown in Figures 2 and 3.

OPERATIONAL COST SAVINGS

The operational dollar saving of circular waveguide is very significant. The cost of generating RF power is dependent on the kilowatts used and the local cost per kilowatt hour. The kilowatts used are dependent on the efficiency of the transmitter.

Much has been said about improving the efficiency of the transmitter. A typical efficiency figure is 35%. With some of the newer techniques, such as, High Efficiency Tubes and Pulsers, efficiencies of 43% to 55% are obtainable.

The amount of line power used and the cost of that power based on 7¢ per kWH and an 18 hr. operating day are listed in Table I for a 55 kW, 110 kW, and 220 kW transmitter.

Based on these figures the power cost per kW/yr. varies from $1,313 to $835.

Using the 35% efficiency figure (most of the existing transmitters would be close to this) cost saving figures using circular guide were calculated:

Transmitter	Tower	
110 kW	1,000 ft.	Table II
110 kW	2,000 ft.	Table III
220 kW	2,000 ft.	Table IV

Using the efficiency tables the power dissipated in the transmission line was determined. The monies spent to generate that amount of lost power was calculated. The lower portion of the table shows the additional cost of using coaxial line as compared to circular guide. This is on a yearly basis. The dollar spent on the power that is lost continues with the life of the station.

INITIAL COST SAVINGS

In addition to the operating cost savings, there is also an initial cost savings using circular waveguide.

Using today's prices the cost of 1,000 ft. and 2,000 ft. runs were calculated for 6-1/8, 8-3/16, and 9-3/16 coax and compared with cost of a circular guide run. The cost per unit ft. shown below includes the line, hardware, and vertical spring hanger.

	Size	Cost Per Ft.	Channel
Coax Line	6-1/8	85.35	14-70
	8-3/16	153.00	14-50
	9-3/16	173.75	14-26
Circular Guide	WC-1150	67.00	55-70
	WC-1500	94.00	35-55
	WC-1800	102.90	14-35

As can be seen in Table V, the initial hardware costs of circular guide compares favorably with 6-1/8 coax but is much less expensive than 8-3/16 and 9-3/16 coax. Added to the initial hardware cost savings would be the operating cost savings.

ERP AND ANTENNA GAIN

The ultimate objective is to increase the ERP. By using circular waveguide less power is dissipated in the transmission line and more RF power is delivered to the antenna.

Assuming a fixed gain antenna, the ERP was determined for

Transmitter	Tower	
110 kW	1,000 ft.	Table VI
110 kW	2,000 ft.	Table VII
220 kW	2,000 ft.	Table VIII

As can be seen from the table, the ERP can be improved as much as 40%

Increasing the power delivered to the antenna permits the use of a lower gain antenna. Lower antenna gain will have a wide elevation beam width which will reduce the beam sensitive to wind variation and increase the near-in coverage. This would also reduce initial costs on not only the antenna but also the tower; since with a smaller antenna the overturning moment and tower loading are reduced.

PERFORMANCE

Circular guide systems installed to date (photo 1) have performed very well. The VSWR tends to be much lower than rectangular waveguide installation. This is so because one is able to hold much closer tolerance with the circular cross section. Because the impedance determining dimension is twice that of rectangular waveguide, the dimension to tolerance ratio is 4:1 in favor of circular guide.

The VSWR of the two transmission lines are shown in Figure 4.

Of considerable interest is the variation of VSWR with Time or Temperature. The VSWR was monitored continuously and plotted against the maximum and minimum temperature and the maximum wind as recorded by the local Weather Bureau. The results show (Figures 5 & 6) there are no measurable VSWR changes. The weather in Raleigh, NC (WLFL) during this period was the worst recorded by the Weather Bureau in 50 years.

CIRCULAR WAVEGUIDE THEORY

To propagate energy in circular waveguide it is necessary to be above the critical cut-off frequency of the Dominant or Lowest Mode (TE_{11}). To take full advantage of low attenuation, as compared with rectangular waveguide, it is necessary to operate considerably above this cut-off frequency.

The lower order modes and the electric and magnetic field configurations are plotted in Figure 7.

The attenuation values shown in Figure 1 are due to finite conductivity. It is necessary to insure that there are no polarization and mode conversion losses.

If a linear polarized wave is launched at the bottom of the vertical tower run when the signal arrives at the top of the tower, it will be slightly elliptically polarized due to the finite tolerance. Techniques have been developed whereby a complementary ellipse can be generated at the base of the tower so a linear polarized wave arrives at the top of the tower.

Mode conversion losses are minimized by operating below the TE_{21} mode. Close mechanical tolerances are necessary to limit polarization and mode conversion loss.

ATTENUATION CHARACTERISTICS

OF

UHF TRANSMISSION LINES

TRANSMISSION LINE	RESISTIVITY (μΩ-CM)
RECTANGULAR WAVEGUIDE	4.3
CIRCULAR WAVEGUIDE	3.4
COAX LINE	1.7

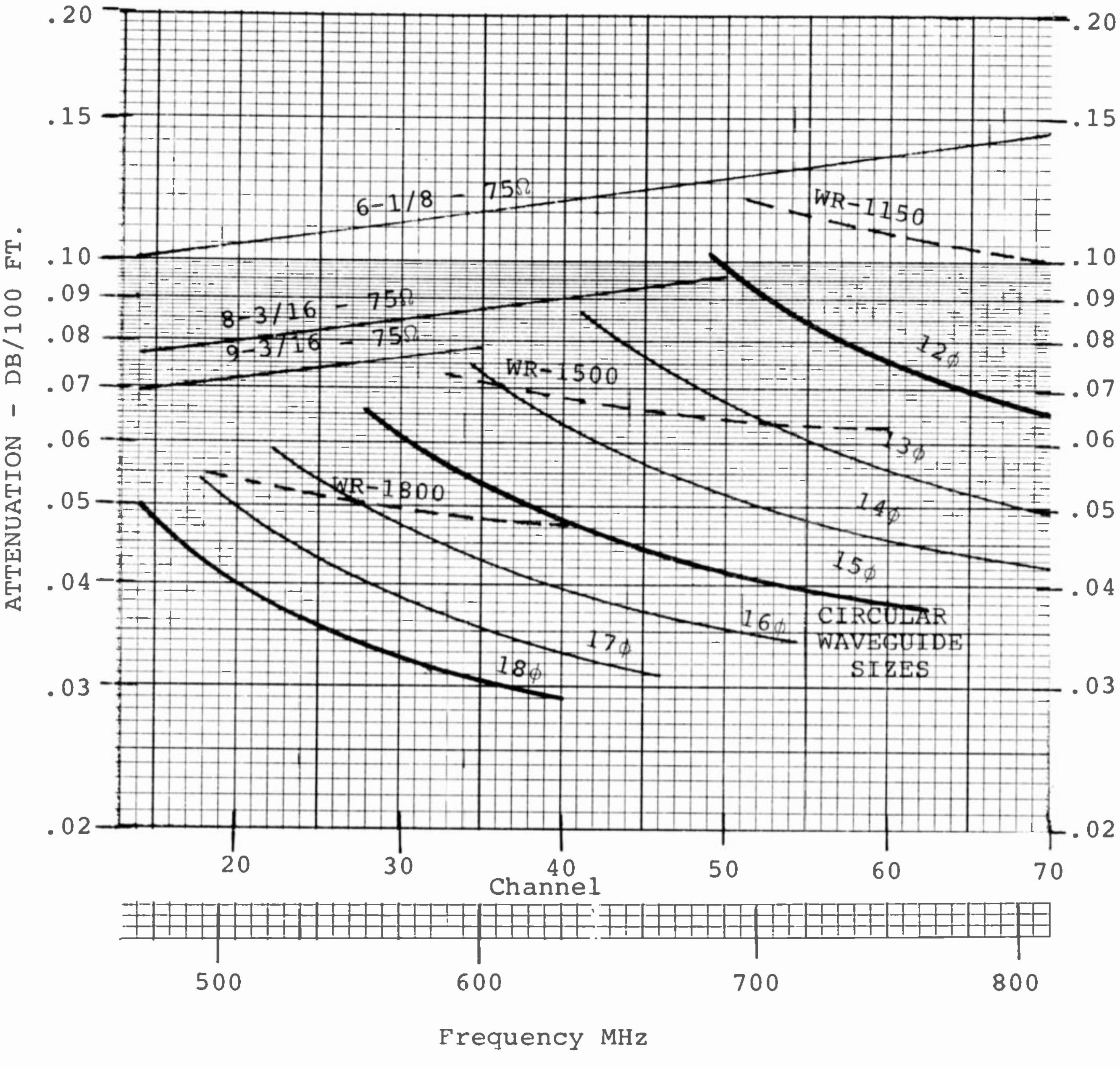
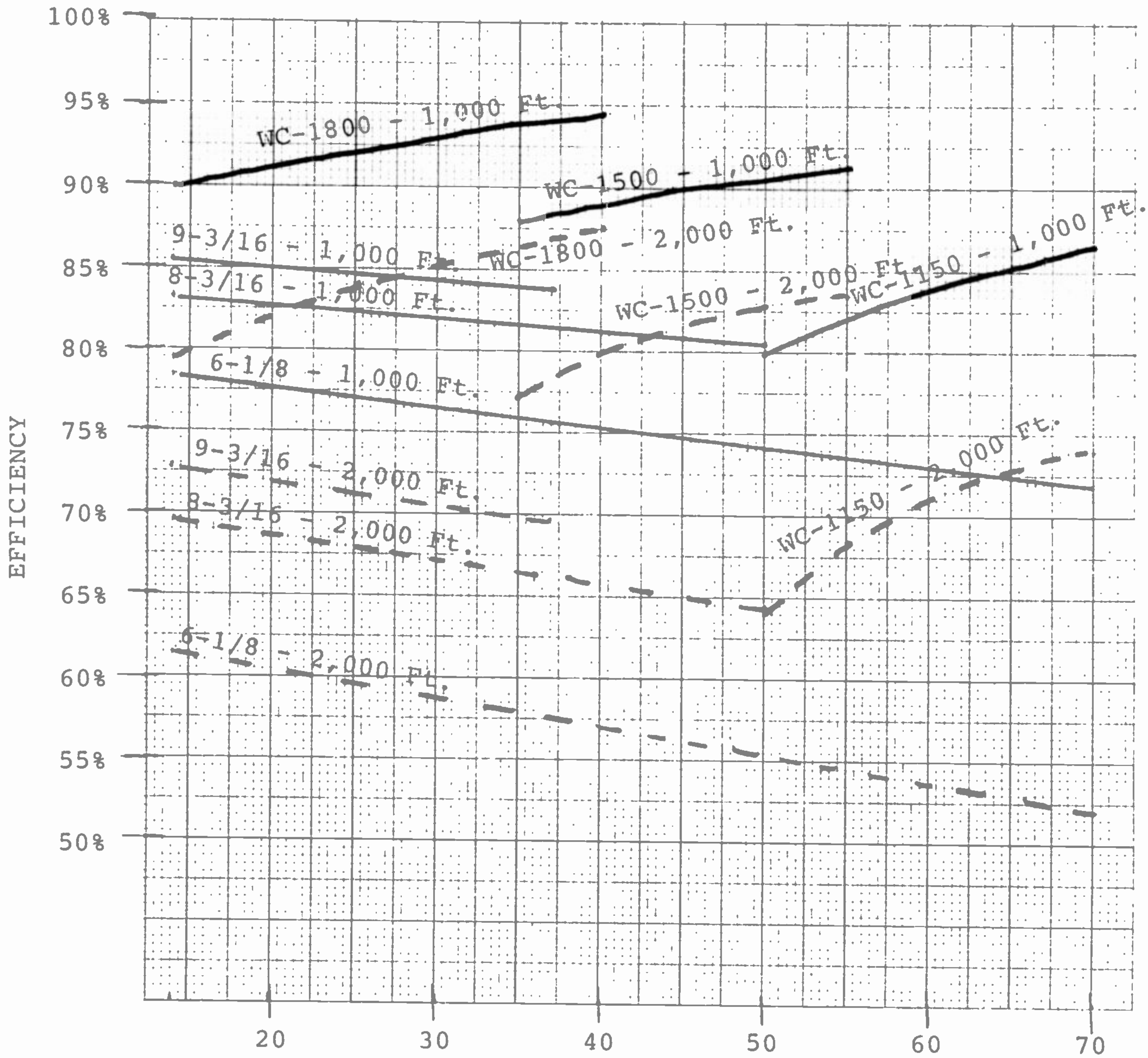


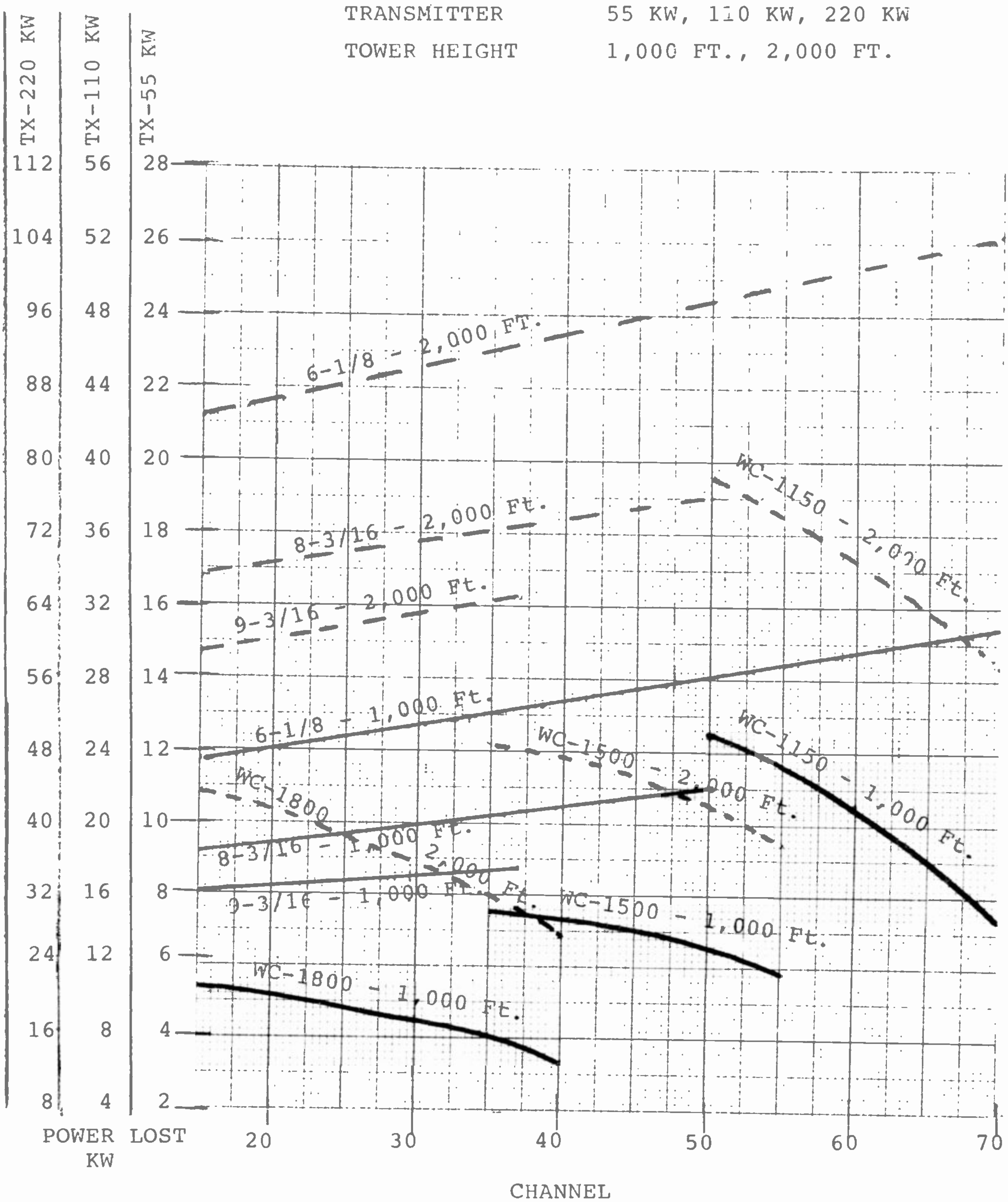
FIGURE 1

TOWER HEIGHT - 1,000 FT., 2,000 FT.



TRANSMISSION LINE EFFICIENCY

FIGURE 2



POWER LOST IN TRANSMISSION LINE

FIGURE 3

TRANSMITTER EFFICIENCY	35%	43%	55%
55 kW Transmitter Power Used/Yr. Cost Power/Yr.	1,032,428 kW $ 72,269	840,548 kW $ 58,824	656,624 kW $ 45,963
110 kW Transmitter Power Used/Yr. Cost Power/Yr.	2,064,856 kW $144,538	1,680,696 kW $117,648	1,313,248 kW $ 91,926
220 kW Transmitter Power Used/Yr. Cost Power/Yr.	4,129,712 kW $289,076	3,361,392 kW $235,296	2,626,496 kW $183,852
Power Cost kW/Yr.	$1,313/kW/Yr.	$1,069/kW/Yr.	$ 835/kW/Yr.

$$\text{COST OF POWER} = \frac{\text{kW/Yr. X Rate}}{\text{Efficiency}}$$

POWER COST BASED ON 7¢/kWH
18 Hr. Day

COST OF POWER

TABLE I

CHANNEL	20	50	70
EFFICIENCY OF LINE			
6-1/8 Coax	77.8%	74%	72%
8-3/16 Coax	83.2%	80%	-
9-3/16 Coax	84.8%	-	-
Circular Guide	91.0%	91%	86.5%
POWER LOST LINE			
6-1/8 Coax	24.5 kW	28.6 kW	30.8 kW
8-3/16 Coax	18.5 kW	22.0 kW	-
9-3/16 Coax	16.8 kW	-	-
Circular Guide	10.0 kW	19.9 kW	14.8 kW
COST OF LOST POWER*			
6-1/8 Coax	$32,000	$37,460	$40,348
8-3/16 Coax	$24,233	$28,820	-
9-3/16 Coax	$22,608	-	-
Circular Guide	$13,100	$12,969	$19,453
ADD'L OPERATING COSTS/YR.**			
6-1/8 Coax	$18,900	$24,491	$20,895
8-3/16 Coax	$11,133	$15,851	-
9-3/16 Coax	$ 9,508	-	-

*COST OF POWER - $1,313/kW/Yr.

**ADDITIONAL COSTS RELATIVE TO CIRCULAR GUIDE

OPERATING COST SAVINGS

TOWER 1,000 Ft.

TRANSMITTER POWER 110 kW

TABLE II

CHANNEL	20	50	70
EFFICIENCY OF LINE			
6-1/8 Coax	60.5%	55.0%	52.0%
8-3/16 Coax	69.0%	64.6%	-
9-3/16 Coax	71.8%	-	-
Circular Guide	82.0%	82.0%	73.5%
POWER LOST LINE			
6-1/8 Coax	43.5 kW	49.5 kW	52.8 kW
8-3/16 Coax	34.1 kW	38.9 kW	-
9-3/16 Coax	31.0 kW	-	-
Circular Guide	19.8 kW	19.8 kW	29.1 kW
COST OF LOST POWER*			
6-1/8 Coax	$56,900	$64,845	$69,168
8-3/16 Coax	$44,671	$51,011	-
9-3/16 Coax	$40,636	-	-
Circular Guide	$25,938	$25,938	$38,186
ADD'L COSTS/YR.**			
6-1/8 Coax	$30,962	$38,907	$30,982
8-3/16 Coax	$18,733	$25,073	-
9-3/16 Coax	$14,698	-	-

*COST OF POWER - $1,313/kW/Yr.

**ADDITIONAL COSTS RELATIVE TO CIRCULAR GUIDE

OPERATING COST SAVINGS
TOWER 2,000 Ft.
TRANSMITTER POWER 110 kW

TABLE III

CHANNEL	20	50	70
EFFICIENCY OF LINE			
8-3/16 Coax	69.0%	64.6%	-
9-3/16 Coax	71.8%	-	-
Circular Guide	82.0%	82.0%	73.5%
POWER LOST LINE			
8-3/16 Coax	68.2 kW	77.8 kW	-
9-3/16 Coax	62.0 kW	-	-
Circular Guide	39.8 kW	39.6 kW	58.2 kW
COST OF LOST POWER*			
8-3/16 Coax	$ 89,342	$102,022	-
9-3/16 Coax	$ 81,272	-	-
Circular Guide	$ 51,966	$ 51,876	$ 76,372
ADD'L COSTS/YR.**			
8-3/16 Coax	$ 37,376	$ 50,146	-
9-3/16 Coax	$ 29,306	-	-

*COST OF POWER - $1,313/kW/Yr.

**ADDITIONAL COSTS RELATIVE TO CIRCULAR GUIDE

OPERATING COST SAVINGS

TOWER 2,000 Ft.

TRANSMITTER POWER 220 kW

TABLE IV

CHANNEL	20	50	70
SIZE	WC-1800	WC-1500	WC-1150

TOWER - 1,000 FT.

6-1/8 Coax	-$ 29,496	-$ 18,342	+$ 10,028
8-3/16 Coax	+$ 51,913	+$ 62,067	
9-3/16 Coax	+$ 77,206		

TOWER - 2,000 FT.

6-1/8 Coax	-$ 45,821	-$ 26,313	+$ 29,917
8-3/16 Coax	+$112,533	+$132,041	
9-3/16 Coax	+$155,941		

THESE FIGURES REPRESENT THE ADDITIONAL COST
ABOVE THAT OF CIRCULAR GUIDE WHEN USING COAX LINE

COST FIGURES INCLUDE GAS BARRIER, HORIZONTAL
RUN, HANGERS, VERTICAL RUN, OUTPUT TRANSITION, AND FINE MATCHER

INITIAL HARDWARE

COST/SAVINGS
OF CIRCULAR GUIDE

TABLE V

CHANNEL	20	50	70

POWER INTO ANTENNA

6-1/8 Coax	85.6 kW	81.4 kW	79.2 kW
8-3/16 Coax	91.5 kW	88.0 kW	-
9-3/16 Coax	93.2 kW	-	-
Circular Guide	100.1 kW	100.1 kW	95.1 kW

ERP
ANTENNA GAIN 26

6-1/8 Coax	2,225 kW	2,110 kW	2,060 kW
8-3/16 Coax	2,380 kW	2,280 kW	-
9-3/16 Coax	2,420 kW	-	-
Circular Guide	2,600 kW	2,600	2,470 kW

ANTENNA GAIN
ERP = 2,500 KW

6-1/8 Coax	29.2	30.7	31.5
8-3/16 Coax	27.3	28.4	-
9-3/16 Coax	26.8	-	-
Circular Guide	24.9	24.9	26.2

TOWER 1,000 Ft.

TRANSMITTER POWER 110 kW

ERP - ANTENNA GAIN

TABLE VI

CHANNEL	20	50	70

POWER INTO ANTENNA

6-1/8 Coax	66.5 kW	60.5 kW	57.2 kW
8-3/16 Coax	75.9 kW	70.9 kW	-
9-3/16 Coax	78.9 kW	-	-
Circular Guide	90.2 kW	90.2 kW	80.8 kW

ERP

ANTENNA GAIN 26

6-1/8 Coax	1,729 kW	1,573 kW	1,487 kW
8-3/16 Coax	1,973 kW	1,840 kW	-
9-3/16 Coax	2,051 kW	-	-
Circular Guide	2,345 kW	2,345 kW	2,102 kW

ANTENNA GAIN

ERP = 2,500 KW

6-1/8 Coax	37.5	41.3	43.0
8-3/16 Coax	32.9	35.2	-
9-3/16 Coax	31.6	-	-
Circular Guide	27.7	27.7	30.9

TOWER 2,000 Ft.

TRANSMITTER POWER 110 kW

ERP - ANTENNA GAIN

TABLE VII

CHANNEL	20	50	70

POWER INTO ANTENNA

8-3/16 Coax	151 kW	142 kW	-
9-3/16 Coax	157 kW	-	-
Circular Guide	180 kW	180 kW	162 kW

ERP

ANTENNA GAIN 26

8-3/16 Coax	3,926 kW	3,692 kW	-
9-3/16 Coax	4,082 kW	-	-
Circular Guide	4,680 kW	4,680 kW	4,212 kW

ANTENNA GAIN

ERP = 5,000 KW

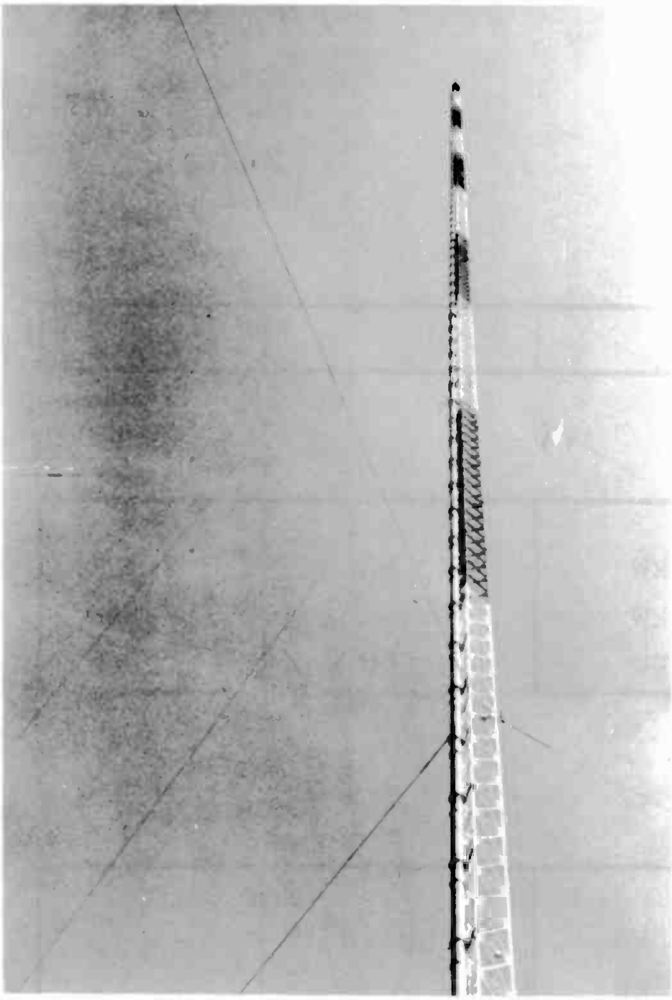
8-3/16 Coax	32.9	35.2	-
9-3/16 Coax	31.6	-	-
Circular Guide	27.7	27.7	30.9

TOWER 2,000 Ft.

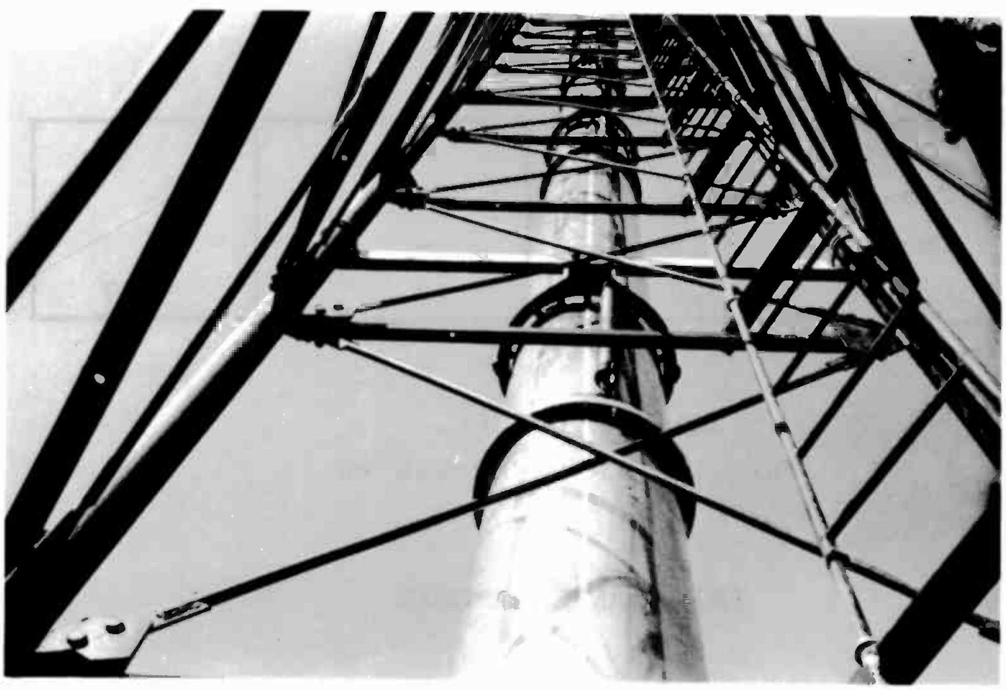
TRANSMITTER POWER 220 kW

ERP - ANTENNA GAIN

TABLE VIII



WLFL, CH 22
RALEIGH, NC



WFTS, CH 28
TAMPA, FL

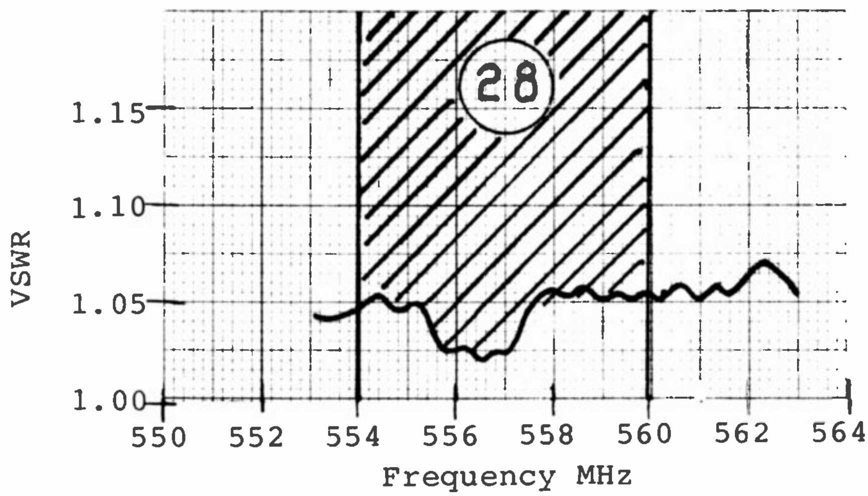
PHOTO 1

WFTS - CH 28

TAMPA, FL

TOWER - 1,000 FT.

WC-1800



WLFL - RALEIGH, NC

TOWER - 1,000 FT.

WC-1800

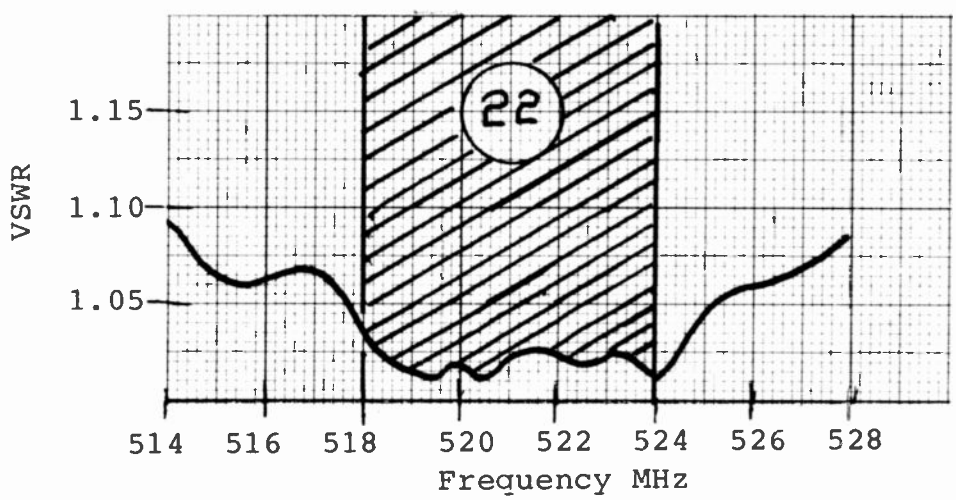
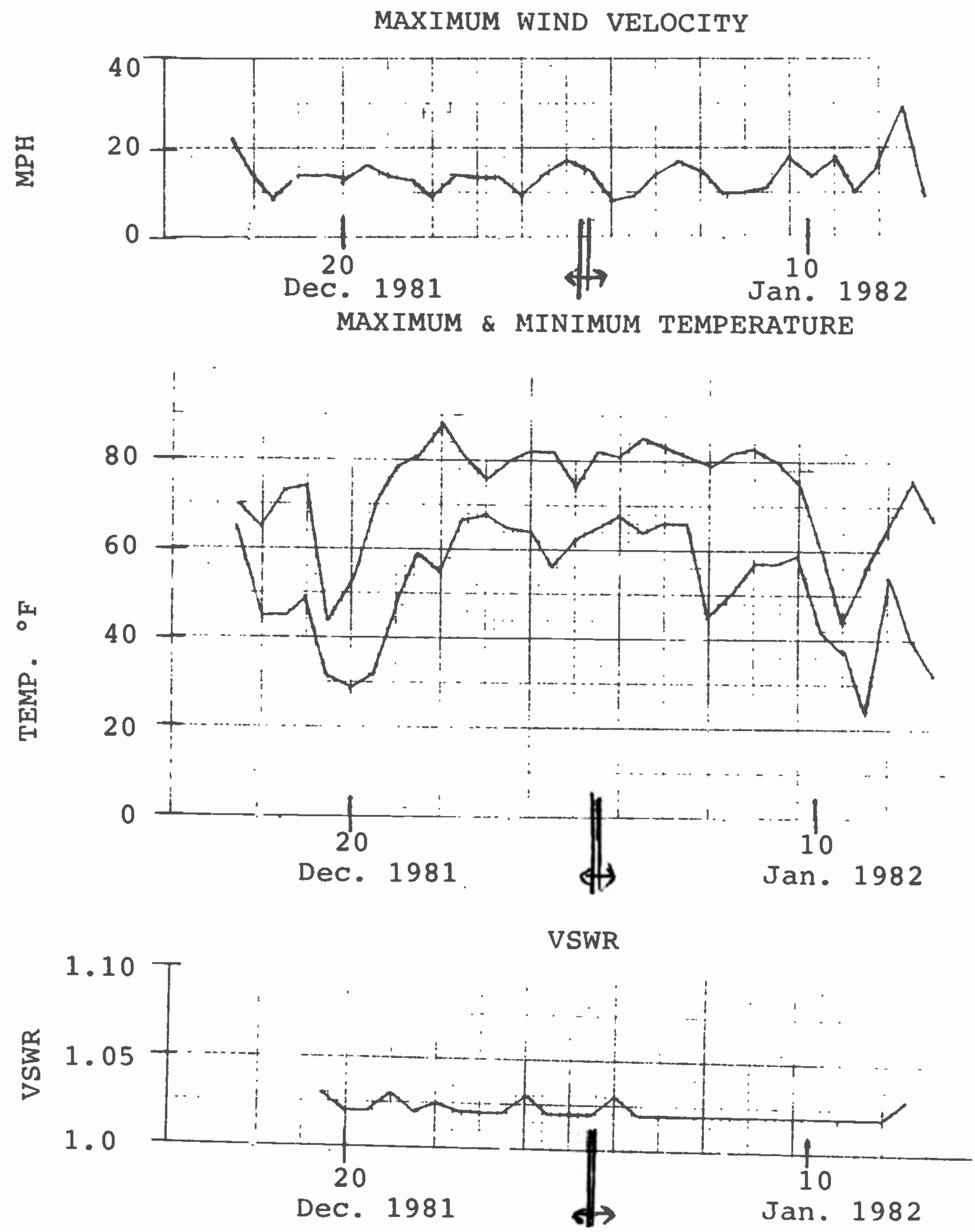


FIGURE 4

WFTS - CH 28

TAMPA, FL

TOWER - 1,000 FT.



VSWR

TEMPERATURE/WIND

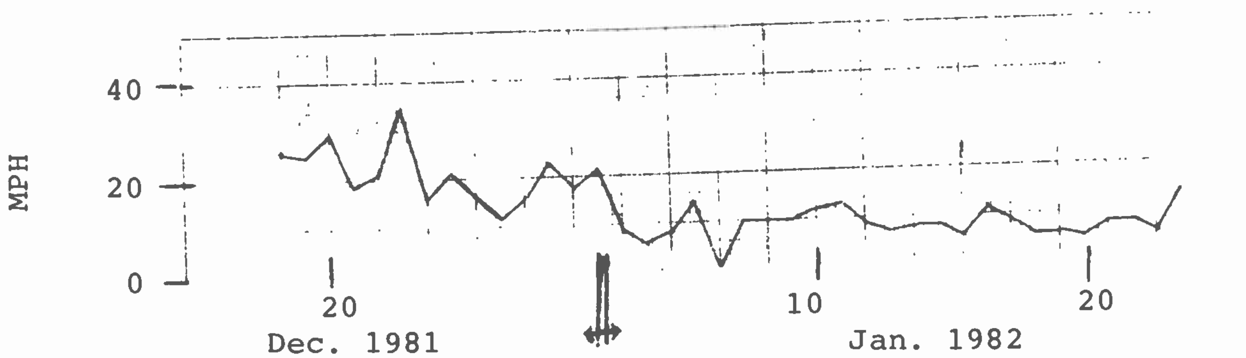
FIGURE 5

WLFL - CH 22

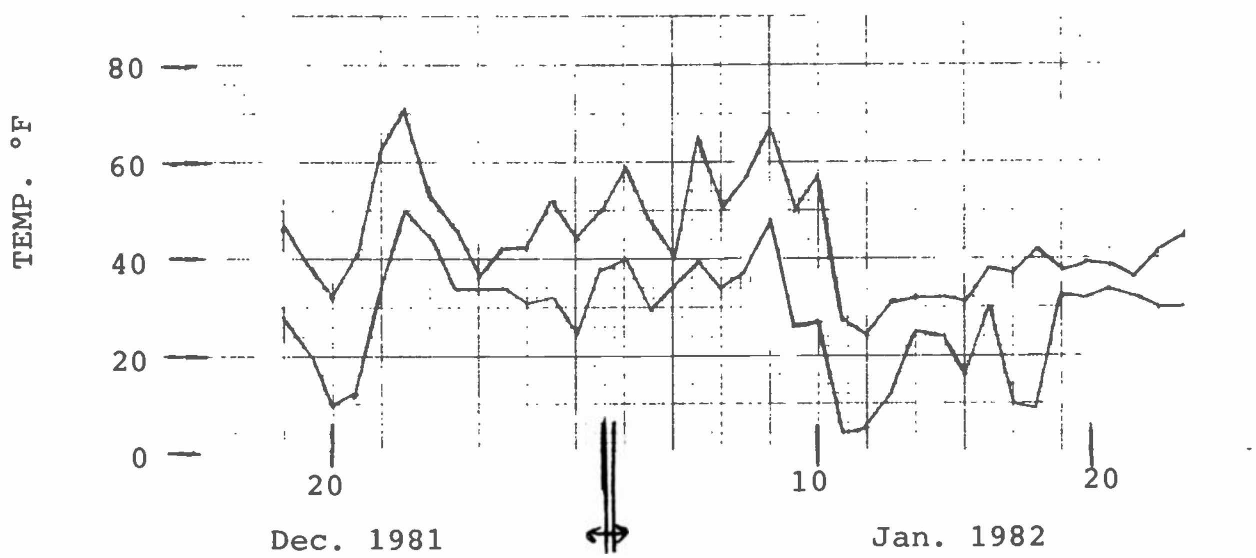
RALEIGH, NC

TOWER - 1,000 FT.

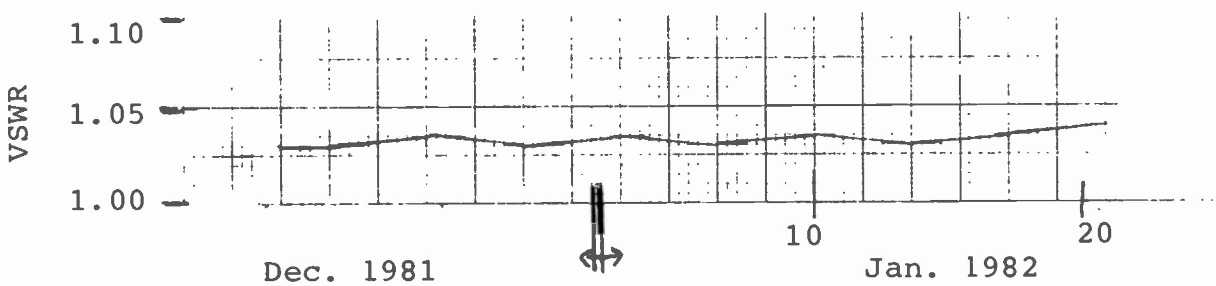
MAXIMUM WIND VELOCITY



MAXIMUM & MINIMUM TEMPERATURE



VSWR



VSWR

TEMPERATURE/WIND

FIGURE 6

CUT-OFF FREQUENCY VS DIAMETER

TE_{11}, TM_{01}, TE_{21} MODE

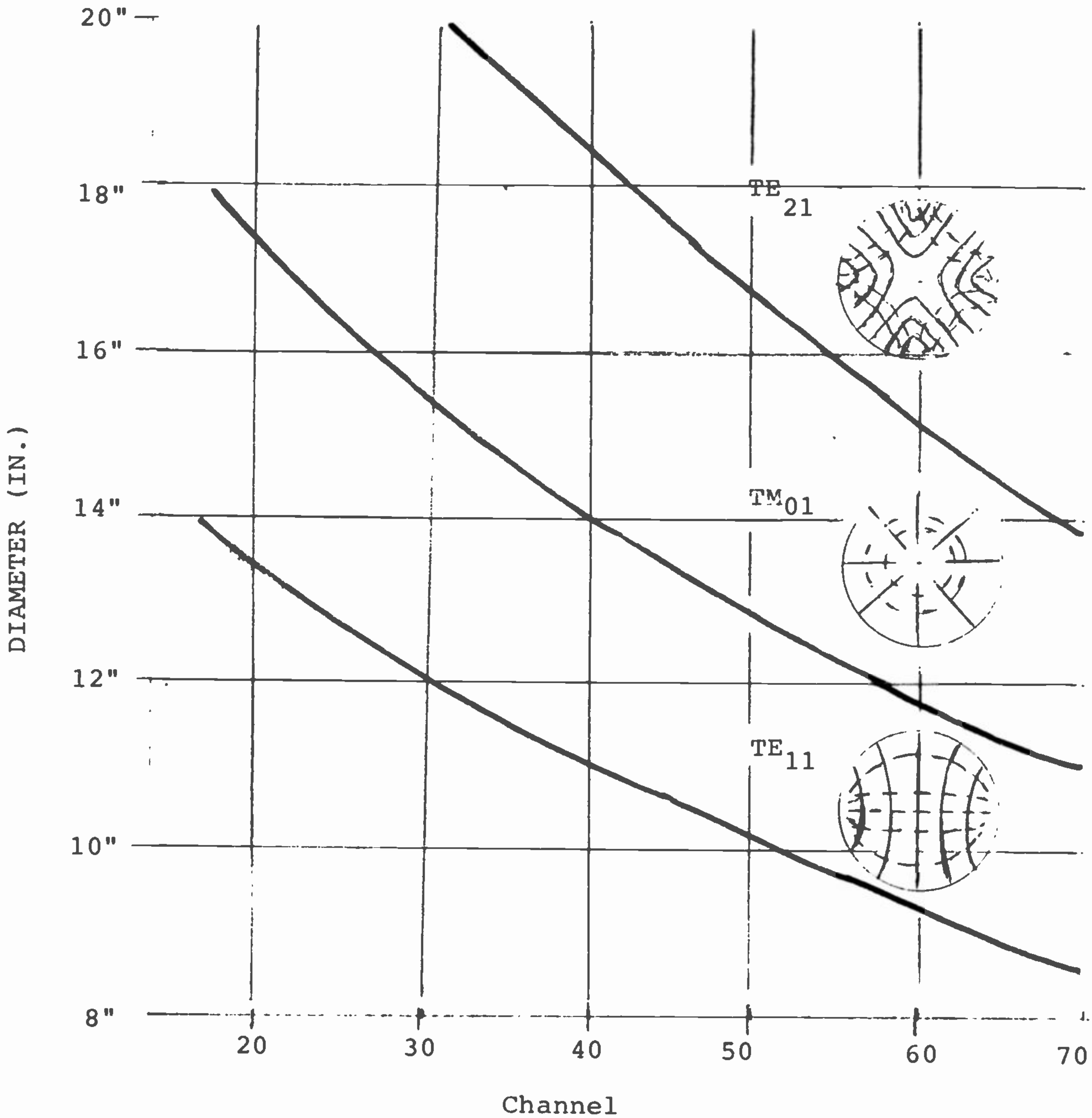


FIGURE 7

Progress in Low-Voltage Beam Modulation of TV-Klystrons

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Introduction: Beam modulation - but how?

Beam modulation of TV-klystrons by a grid offers the advantage of low modulator voltages contrary to the requirements of a modulation-anode. However, considerable objections have been raised, preferably by klystron transmitter users, against incorporating a "fragile" thing like a grid in such a rugged tube like a klystron.

One answer could be that a low frequency grid must not be of a fragile design and that there are means to protect it. Nevertheless, a grid is an intercepting electrode, its beam interception varies with the amplitude of the modulating signal, and it needs a stabilized offset-voltage.

A different approach to the problem is a low-voltage non-intercepting modulation electrode.

Features of a non-intercepting low-voltage modulation electrode

Realization of such an objective by introducing an electrode of annular shape (Annular Beam Control electrode) into the electron gun produces the following features, compared to using a grid:

- a design as rugged as any gun electrode in a conventional klystron, requiring no special protection means,
- no requirement for a DC offset-voltage, the peak sync level of the modulating signal may simply be clamped to the cathode potential,
- possibility of conventional operation of the klystron by simply applying a short-circuit between modulation electrode connector and cathode,
- in case of modulator break-down the klystron automatically switches to conventional operation,
- klystron types comprising modulation electrodes will be compatible with existing sockets.

Performance and characteristics of klystrons comprising ABC electrodes

ABC electrodes have been introduced into two families of existing TV-klystron types. It was demonstrated that the modification did not affect the conventional operation mode of the tubes: zero bias at the modulation electrode showed no difference to the previous performance, whilst application of a modulating voltage did not have negative influence on focusing etc.

Fig. 1 shows a typical characteristic of beam modulation depth versus beam current, using the type YK 1233 as an example. There is good reason for the fact that the characteristic ends with a modulation depth of almost 50 %. From the following fig. 2 (showing the output signal voltage of the klystron versus the input signal voltage, using the modulation depth as a parameter) we learn that a modulation depth of the beam current exceeding 40 or say 50 % is meaningless due to the reduction of the klystron gain.

Application of ABC modulation technique

The Annular Beam Control electrode technique offers different ways to reduce still further the power consumption of already highly efficient klystrons.

A relatively simple methode to reduce the average power consumption is applying the well-known sync pulsing technique. This can be done now with extremely low effort due to the low voltage requirements of the ABC electrode, compared to the modulating anode, while maintaining the low capacitance. As indicated in fig. 2 by the 75 % output signal level, an average power reduction of almost 20 % may be expected from this operation.

Increasing the effort by applying full time modulation, i.e. a modulation following the low frequency part of the luminance signal spectrum, special precorrection means for the signal will be required. This will, however, lead to another remarkable step in the reduction of power consumption. Using modulation characteristics like the one indicated by the diagonal dotted line in fig. 2 for instance, saving rates exceeding 30 % are realistic targets.

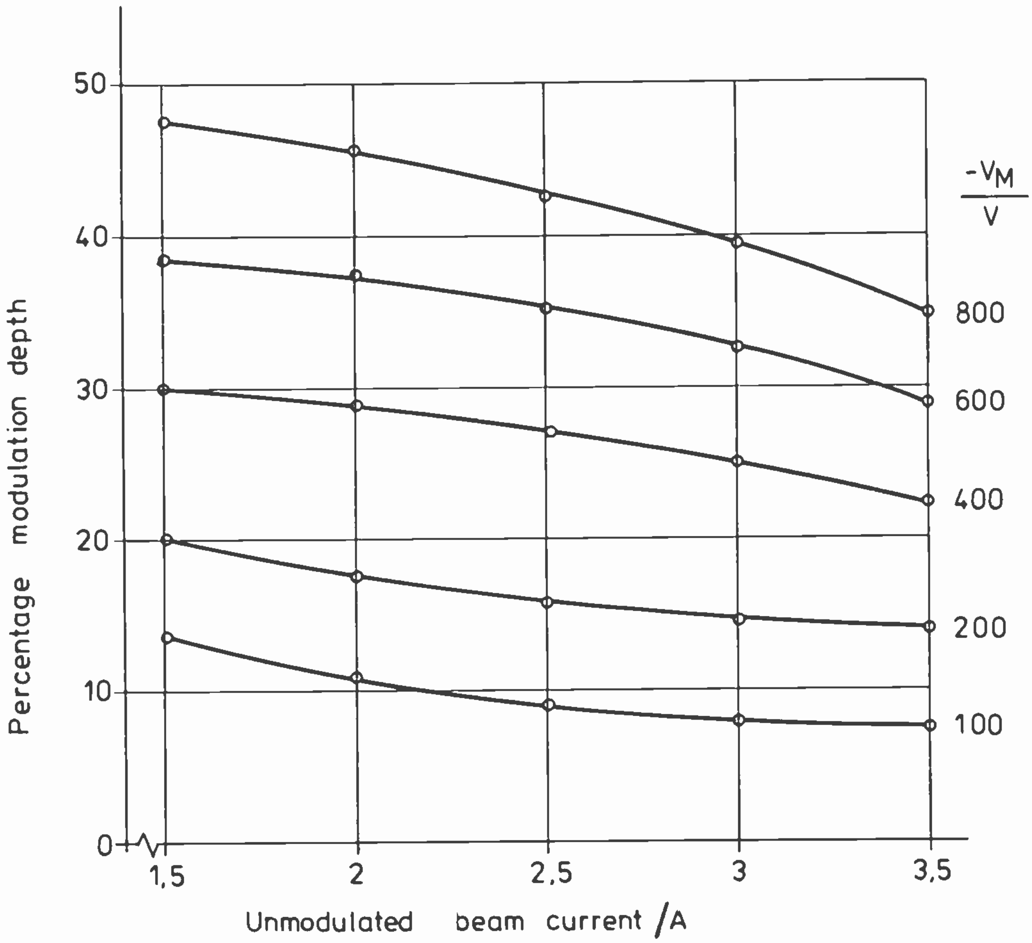

Fig. 1 25 kW-TV-Klystron YK 1233
Beam modulation depth versus unmodulated beam current. Parameter: Modulation electrode voltage V_M.

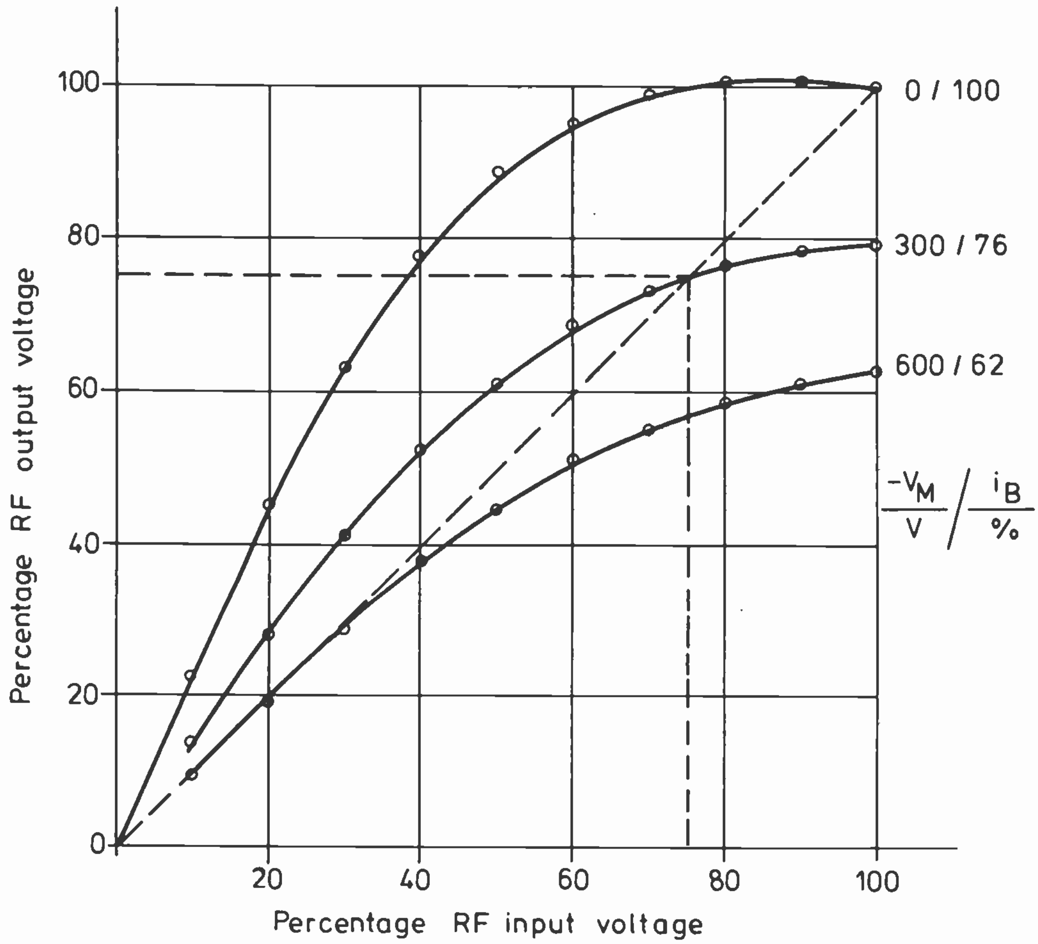

Fig. 2 25 kW-TV-Klystron YK 1233
RF output voltage versus RF input voltage
Parameters: Modulation electrode voltage V_M
and percentage of beam current i_B

DIRECT BROADCAST SATELLITE SYSTEMS

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1.0 INTRODUCTION

As far back as the early 1960's, Direct Broadcast Satellites were being considered and the basic technology for such systems was being investigated, however at that time the basic technology for such satellite systems was not yet available. Since that time, significant advances in both satellite and ground receiver technology has resulted in today having direct broadcast satellite systems feasible with no insurmountable technical barriers preventing implementation. The direct broadcast satellite service refers to a synchronous satellite relaying signals intended for reception by the general public either directly by individuals or by community receivers for distribution to local recipients. The major characteristic of such systems is that the receiver used for home reception should be small, with antennas on the order of 3 feet or less in diameter, with costs under $1,000. The major issues to be considered in providing a direct broadcast satellite service are as follows:

1. Service and Operational Requirements
2. Regulatory Issues
3. Technical Issues
4. Compatibility with Alternate Delivery Systems

In this paper, each of these issues will be examined in greater detail. The material for this paper comes from a number of sources such as the FCC report by B. Patton, WARC-77 and WARC-79 Final Acts, C.C.I.R. documents, FCC filings and a number of published articles. However, let us first examine previous experiments that have been carried out in direct broadcast satellites and the experience gained from such experiments.

In 1974, NASA launched the ATS 6 Satellite. While this satellite carried a large number of payload packages, direct television broadcast was provided for in both the UHF-band and S-band. In a joint program with India, television was broadcast to a large number of villages throughout India for a period of one year, starting in June of 1975 as part of a program called SITE. The receiving antenna used in the Indian village was a 3 meter UHF mesh parabaloid with a point focus feed. The nominal frequency in the downlink from the satellite was 860 MHz with a satellite EIRP of 48 dBw. The uplink broadcast frequency to the satellite was at 6 gHz and the transmit polarization from the satellite was right hand circular polarization. ATS 6 was also used in the U.S. for educational television experiments directly to educational institutions sponsored by HEW and the Public Broadcasting Corporation. This experiment was done at S-Band with a 3 1/2 meter diameter antenna used for receiving the signals from the satellite. In the case of the Indian experiment, a signal-to-noise ratio at the receiver was on the order of 46 dB and provided high fidelity video pictures, with the receiving earth station having a noise figure of about 4 dB. For the U.S. S-band experiment with a downlink transmission frequency of around 2.6 gHz, the video picture quality was also quite good. In this case, the receiving earth station converted the FM received signal to an AM-VSB format for use by conventional television receivers.

In January of 1976, the Communication Technology Satellite (CTS) was launched by NASA also to be used in experiments with direct broadcast and video. This satellite contained a 200 watt traveling wave tube and provided a EIRP from the satellite of 58 dbw. Transmit polarization for this satellite was linear, and the downlink frequencies were nominally around 12 gHz. A number of experiments were carried out both in the U.S. and Canada, and jointly between the U.S. and Canada, transmitting television signals for educational and teleconferencing purposes as well as experiments with program material.

In Japan, broadcast satellite experiments were conducted with the BSE satellite launched in April of 1978, which also used a downlink frequency of around 12 gHz and had a radiated power from the satellite of about 57 dBw. Experiments for this satellite also showed good quality television signals could be transmitted to reasonably small earth stations using FM transmission of the video signal.

ANIK-B, a Canadian satellite launched in December, 1978, was also used for direct broadcast satellite experiments, using a 14 gHz uplink to the satellite and 12 gHz downlink from the satellite having a radiated power of 48 dBw per video channel.

In addition, the Soviet Union has launched the EKRAN satellite, the first one being launched in 1976 with a downlink frequency in the UHF band at around 700 MHz, using frequency modulation with transmission of the video, with a radiated power from the satellite of about 49 dBw. The receiving system of this satellite used the Yagi-Uda array antennas, with antenna gains from 23-30 dB. EKRAN is today an operational system used in the Soviet Union. Table 1 summarizes the characteristics of some of these experimental satellite's receiving stations.

Table 1 - EXPERIMENTAL GROUND RECEIVING EQUIPMENT CHARACTERISTICS

SYSTEM	ATS-6	ATS-6	CTS	BSE-Japan
Frequency (gHz)	0.860	2.5	12	12
G/T (dB/K)	-6	+8	+15	+16
Antenna Dia. (m)	3	3	1.8	1.6
Material	Expanded Aluminum Mesh	Epoxy Fibreglass	Epoxy Fibreglass	Aluminum Press-stretch ([3])
Tracking	None	Limited Manual	Limited Step Track	None
Input Stage	Bipolar Silicon Transistor	Bipolar Silicon Transistor	Image Enhanced Diode Mixer	Image Enhanced Diode Mixer
Noise Figure (dB)	6.5	4	6	4.5
No. of Channels	1	1	1	2
No. of Frequency Changes	1	None	1	1
Intermediate Frequency (MHz)	70	None	1	1
Cost in Quantity	$800 (2400)	$4000 (130)	$10000 (1-10)	-
Est. Cost in Quantity	-	$3200 (1000)	$4900 (1000)	
Installation Cost	-	$1000	-	-

In addition to the above experimental satellites, a number of projected operational satellites are being considered for direct broadcast operation including NORDSAT (the Scandinavian satellite system), ARABSAT (the Arab satellite system), INSAT (the Indian domestic satellite system), a number of European broadcast satellites, and ANIK C (a new Canadian satellite system). These satellites vary considerably in the downlink frequency, varying from UHF frequencies up through 11-12 gHz for the downlink frequency, varying in radiated power from 42 dBw up through 65 dBw, and varying in terms of satellite usable bandwidth per video channel from 23 MHz through 72 MHz. In almost all cases the type of modulation is frequency modulation.

The experiments that have been conducted clearly demonstrate that earth stations with antenna sizes of 3 meters or less can be used for providing good quality video signals directly from a broadcast satellite. Let us go on to consider the above four issues relating to broadcast satellites.

2.0 SERVICE AND OPERATIONAL REQUIREMENTS

While the major service to be provided by direct broadcast satellite systems is the distribution of television signals directly to the home, the type of signal offered and how the service is to be provided operationally varies among potential operators. If we examine Table 2, put together by Dr. John Clark of RCA, we can see that in the case of the U.S. there are eight different applicants, and the applicants potentially are offering a variety of different services, or at least different operational modes for this broadcast service.

Table 2 - DBS U.S. FCC Applicants

Appli-cant	Prime Service	CONUS Zones	Number Chan	Orbital Loc.	TWTA (Watts)	EIRP (dBW)	Band. (MHz)
CBS	Sponsored HDTV Rebroadcast	4	3	RARC-83	400	60.4	27.
DBSC	Common Carrier	3	6 [+8][1]	103°, 123° 143°	200 [20][1]	56.	22.5
GSC	Subscription Leased	2	2	115°, 143° 110°, 125°	300	53.7 25	18.
RCA	Channels	4	6	140°, 155° 115°, 135°	230 185	58. 57.	[72][3] 16
STC	Subscription Sponsored	4	6	155°, 175° 115°, 135°	-	-	[28,100][3]
USSB	Rebroadcast Sponsored	4	6	[+2][2] 115°, 135°	230	57.	16

Table 2 - DBS U.S. FCC Applicants (continued)

Appli-cant	Prime Service	CONUS Zones	Number Chan	Orbital Loc.	TWTA (Watts)	EIRP (dBW)	Band. (MHz)
VSS	Rebroadcast Licensed	4	2	115°,175° 80°,100°	150	56.	18
WUTC	Channels	4	4	120°,140°	100	55.5	16

1 - spot beams, 2 - undesignated, 3 - HDTV

In the case of CBS, for example, we are talking about using a direct broadcast satellite for a new service which is high definition television, an 1,125 line television signal different from the current 525 line NTSC signal. In the cast of Direct Broadcast Satellite Corporation, they view the direct broadcast satellite system as a common carrier system which would be used for distributing a variety of different kinds of information depending on the leased time on the system. For some applicants, such as STC, a COMSAT subsidiary, the service would be totally provided by STC including both distribution of the video signals and the program material and would be a subscription service in which each subscriber paid to receive the video signal which would be normally transmitted in a scrambled fashion and would have to be unscrambled at the subscriber location. Other applicants view it as a rebroadcast service rather than as a subscription service. In U.S. Study Group document BC/835 submitted to Study Groups 10 and 11 of the C.C.I.R., ten major applications of these satellites were defined. These are as follows:

1. Education
2. Health and Medical
3. Electronic Mail
4. Law Enforcement
5. General Computer Networks
6. Emergency Communication and Disaster Warning Advisories
7. General Broadcasting
8. Business and Financial Information Networks
9. Electronic Publishing
10. Public Telephone and Telegraph

As one can see from this list, the possible applications are quite varied. Also, one can see from Table 2 that there still exists considerable differences in opinion as to some of the technical parameters to be used for broadcast satellites as well as the many possible services for which it can be used.

One of the important considerations in defining the broadcast satellite system is not only the service but also the performance requirements the system will have to meet. While there are no standards yet developed for Region 2, we can get some guidance on television performance standards from the Final Acts of the World Administrative Radio Conference which was held in 1977 on Direct Broadcast Satellites (WARC-77) and from the filings made to the F.C.C. for direct broadcast satellites. Based on FM transmission and television channel bandwidths on the order of 20 MHz, a specification from the Final Acts of WARC-77 call for a carrier-to-noise ratio of 14 dB for 99% of the worst month which would correspond to a video signal-to-noise ratio (peak signal to RMS weighted noise) of approximately 43 dB. And, since rain margins of from 2-6 dB have been projected for the direct broadcast satellite systems, this would correspond to a clear weather signal-to-noise ratio of possibly as high as 45-49 dB. In terms of protection ratio from interference, a specification of 35 dB carrier-to-interference ratio as a minimum was specified for a single interfering source and a cumulative carrier-to-interference ratio of at least 30 dB for all interferring sources. The overall subjective quality target is equivalent to a grade of 4.5 on the C.C.I.R 5-point scale.

One of the performance questions also refers to the percentage or number of hours of outage during the year when the signal quality is too poor to be useful. One of the specifications on this relates to a carrier-to-noise ratio of greater than or equal to 10 dB for 99.8% of the worst month, where 10 dB would correspond to the FM threshold point. This would correspond to outages from 2 to 12 hours per year for a 1-meter receiving antenna depending on the amount of rainfall and the location of the spacecraft or elevation angle to the spacecraft from a given site. Other performance requirements relate also to interference into other services such as terrestrial microwave or other broadcast satellite systems. Specifications which came out of the WARC meetings indicate a minimum power flux density at edge of coverage area for Region 2 to be -105 dBw per meter squared for individual reception at the edge of the coverage area for 99% of the worst month. Also, frequency guardbands of 9 MHz are specified for Region 2 at the upper edge of the 12.5-12.5 gHz band. This assumes a satellite EIRP of 63 dBw for Region 2.

We discussed a protection ratio of 35 dB carrier-to-interference for single entry when the interference is co-channel or at the same frequency. Figure 1 shows the protection ratios relative to the co-channel value as a function of the carrier frequency offset between the interfering carrier and the desired carrier.

3.0 REGULATORY CONSIDERATIONS

The 1977 WARC was the first conference to prepare a plan for the Broadcast Satellite Service (BSS). At this conference plans were developed for Regions 1 and 3 (see Figure 2) but not for the Americas in Region 2, where a postponement was requested. For Regions 1 and 3 the downlink frequency allocations were 11.7-12.5 Hz and 11.7-12.2, 12.5-12.75 Hz, respectively. The items specified in the plan for each region were:

1. Orbital Position
2. Polarization of the Transmission
3. Maximum Satellite EIRP
4. Frequency Plan for the Television Carriers
5. Satellite Transmitting Antenna Footprint per Service Area

The Plan for Regions 1 and 3 covers a total of 35 orbital positions. The polarization selected was circular in both directions. The satellite orbital positions are spaced every 6 degrees covering over 220 degrees of orbital arc. The minimum power flux density at the edge of the coverage area for 99% of the worst month is -103 dBw/m^2 for Regions 1 and 3 for individual reception and -111 dBw/m^2 for community reception. The frequency channels are uniformly spaced about every 20 MHz, providing 40 channels in Region 1 and 25 channels in Region 3. There are a total of 252 service areas for the two regions resulting in from 1 to 35 service areas per country, with most countries having a single service area. In Region 1, most service areas receive up to 5 television channels while in Region 3 most areas receive up to 4 television channels.

While a plan was not devised for Region 2 at WARC-77, a segmented orbital arc assignment was proposed for Region 2 as shown in Figure 3, with alternating segments of arc assigned to BSS and Fixed Satellite Service (FSS). At this meeting, the 11.7-12.2 gHz band was to be shared equally with BSS and FSS service, leaving 6 BSS positions and 17 FSS orbital positions for Region 2.

At the World Administrative Radio Conference in 1979 (WARC-79) this plan was changed with FSS and BSS utilizing the full orbital arc for Region 2 from 37^{o}W to 170^{o}W, with no segmentation. The downlink frequency allocations were changed so that BSS had first use of the 12.3-12.7 gHz band, FSS had first use of the 11.7-12.1 gHz band and 12.1-12.3 was shared equally between FSS and BSS. This results in 22 BSS positions available for service to North America with 6^{o} separation and 52 FSS position with 4^{o} spacing, 28 for North/Central America.

A Regional Administrative Radio Conference is being planned for 1983 (RARC-83) to come up with a BSS plan for Region 2. The FCC has taken the lead in planning for this conference within the U.S. and has formed advisory committees made up of industry and government representatives to help establish a U.S. position for such a regional Plan. There are 32 administrations in Region 2. It is likely in RARC-83 that all administrations will have at least 4-5 television channels with large nations having more. All countries covering several time zones are likely to have a minimum of one beam (service area) per time zone. It is likely that the currently shared 12.1-12.3 gHz band will be split with 12.1-12.2 given to FSS primarily and 12.2-12.3 given to BSS primarily, so that the BSS will have the 12.2-12.7 band for downlink service. It is likely that all TV channels transmitting to one service area will have the same polarization, and adjacent service areas using adjacent TV channels will use opposite polarization when served from the same satellite. Elevation angles to the satellite will be kept as high as possible by proper location of the satellite. The satellite should be west of the service area so that eclipses occur after midnight in the service area.

The possible uplink frequencies to the broadcast satellite (based on WARC-79) are 10.7-11.7 gHz, 14.5-14.8 gHz and 17.3-18.1 gHz, with the 14.5-14.8 gHz band being exclusive for BSS. There appears to be a preference for the 17.3-18.1 gHz for Region 2 and this may be the likely band selected in RARC-83.

Figures 4 and 5 show the satellite transmitting antenna pattern and earth station receiving patterns based on WARC-77. The shaped beam pattern shows what can be done with today's technology, which may permit satellite spacing as close as 2^{o}-3^{o} as compared to the proposed spacing of 6^{o}. Figure 6 also shows for individual receiver antenna patterns what improvements in sidelobe response can be expected for offset feed antenna designs.

4.0 TECHNOLOGY CONSIDERATIONS

At the WARC-77 meeting, a receive earth station G/T = 6 dB/^{o}K was specified for individual receiving stations with minimum power flex density of -105dBw/m^2 in Region 2 for individual reception. This was based on 1977 technology. Since that time technology improvements have been realized in both the satellite and receiving earth station areas. Let us first consider the receiving earth station.

4.1 Receiving Earth Station

There are three major receiver configurations that have been proposed for individual reception from the broadcast satellite. These are:

1. Low Noise GaAs Fet Preamplifier Front End followed by a single stage of down conversion.

2. Konishi Low Noise mixer front end with single i-f stage.

3. Low Noise mixer front end with double down conversion, having i-f frequencies of 1 gHz and 100 MHz.

At the time of WARC-77, the individual receiving earth station was assumed to have a noise figure of about 8.6 dB, and with an antenna gain of 38.6 dB (1.0 meter diameter) a figure of merit of 6 dB/^{o}K was selected. However, today GaAs Fet low noise amplifiers are capable of noise figures in the range of 2-5 dB and image enhanced low noise mixer designs are capable of providing noise figures in the 4-5 dB range. Thereford G/T values from 10-15 dB/^{o}K are possible with receiving antennas having diameters from 1-2 meters. This would permit a reduction in the minimum flux densities required from the satellite.

Figure 7 shows the three different receiver configurations. The depolarizer preceeding the receiver converts the circular polarized received signal to linear polarization. The low noise mixer front end avoids the requirement for a separate low noise amplifier at some penalty in increased noise temperature. The double conversion receiver while being more expensive than a single conversion system, permits easier rejection of spurious signals, a less stable first local oscillator and multichannel operation by switching the second local oscillator frequency.

Another technique associated with FM transmission is the use of feedback demodulators, which can extend the FM threshold by 2-5 dB depending on the modulation index and the receive filter bandwidths used.

Improvements in earth station antenna design are also being made with the objective of reducing the sidelobe response and lowering the cost. Offset feed antennas improve sidelobe response since feed blockage is reduced and provide improved VSWR (see Figure 6).

4.2 Spacecraft Technology

Since 1977, significant improvements have been realized in spaceceraft technology. For the DBS spacecraft some of the key technologies are:

1. High Power Amplifiers
2. Prime Power Generation
3. Antenna Designs

These technologies are in turn effected by:

a) Size of service areas
b) Number of TV channels to be provided
c) Receiving earth station G/T

To achieve high i-f power from the satellite TWT amplifiers would have to be used. Through the development of multiple collector tubes, prime power efficiencies in the range of 40-50% have been achieved with saturated power outputs from 100 watts to 400 watts with projections as high as 700 watts. Table 3 (from reference[1]) summarizes some of the high power tubes available today.

Table 3 - High Power Tubes Operating in the 12 gHz Band

SATURATED POWER OUTPUT	MANUFACTURER	APPLICATION
100/150 W	Thomson	H-SAT
200	AEG Telefunken	German Sat.
450	Siemens	German Sat.
450	AEG Telefunken	H-SAT or German Sat.
700	Siemens	German Sat.
100	Hughes	BSE (14.25-14.43 gHz)
200	Litton	CTS
20	Thomson/CSF	CTS/QTS/ECS
20	AEG Telefunken	SBS/OTS

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Solid state power amplifiers such as GaAs MESFETS, may be paralleled in the future to produce 10 or 20 watts if the satellite EIRP can be reduced to that level, however, in the immediate future TWTAs appear to be the primary high powered amplifier for the DBS application.

Prime power requirements can be quite significant. If we consider 200 watts of r-f power per channel and a TWTA efficiency of 40%, we would require on the order of 500 watts of prime power per channel and with five channels per time zone we would require 2500 watts of prime power per time zone. This would require sun oriented roll out arrays or roll out paddle arrays. Such large prime power requirements also makes it difficult to provide battery powered operation during eclipses and makes it important to choose orbital locations which have high elevation angles for the area being serviced and eclipse times which occur in the early hours of the morning (non-prime time).

Antenna techniques considered for producing shaped beams are:

- o multiple feed single or dual offset feed reflectors
- o shaped reflector
- o phased arrays
- o multiple feed lens antennas

Shaped beam antennas are attractive because they can produce lower side lobe responses, confine the power to a given service area, and more evenly distribute the power within the service area.

Shaped beam antennas result in a requirement for more accurate station-keeping and altitude control. A 0.1^{o} accuracy in east-west and north-south station-keeping appear to be practical today and altitude control of $\pm 0.1^{o}$ in pointing accuracy appears practical.

One of the transmission parameters to be selected is the use of circular versus linear polarization. In the Region 1 and 3 plans, circular polarization was selected. Reference [2] discusses some of the pros and cons of each polarization. These are summarized in Table 4 below.

Table 4 - Some Aspects of Linear Versus Circular Polarization

Factor
1. Alignment of Receiving Antenna

Remarks
Alignment of the polarizaion direction is not necessary for circular polarization

Advantage
C (Circular)

Factor
2. Effect of Misalignment on Cross-Polarization

Table 4 - Some Aspects of Linear Versus Circular Polarization (continued)

Remarks
Misalignment of polarization direction of both transmitting and receiving polarization antennae required with linear polarization, 2 to dB extra cross-polar protection margins in comparison with circular polarization.

Advantage
C

Factor
3. Orientation of Satellite Antenna

Remarks
With linear polarization the plane of polarization antenna will not in general correspond to the major or minor axes of a beam with elliptical cross-section; therefore:

a) It may be difficult to produce a good cross-polar response with linear polarization (in particular for elliptical beams).

Advantage
C

b) Transfer to a spare satellite at a different orbital position would probably be more difficult with linear polarization because of the need to realign the polarization plane.

Advantage
C

Factor
4. Sharing with Other Services

Remarks

a) If circular polarization is chosen for the broadcasting-satellite service and other services use linear polarization, 3 dB protection between these services and the broadcasting-satellite service is assured.

Advantage
C

b) If both the broadcasting satellite and other services, e.g., fixed satellite and terrestrial services, use linear polarization, then in isolated cases, where the dominat interference arrives near the main beam of a receiving antenna, it may bepossible to increase the isolation by the use of orthogonal polarization.

Advantage
L (Linear

Factor
5. Propagation Effects

Digital modulation is combined with converting the analog television into digital form. This is especially attractive with High Definition Television Signals (HDTV) where digital processing of this signal can significantly reduce the needed bandwidth and power for transmission. While this type of processing is also possible for the NTSC video signal, the cost of the digital receiver and digital D/A conversion equipment is still high compared to FM equipment. For HDTV, the power and bandwidth required for FM transmission are so large, that the efficiency of digital transmission may justify the cost of the digital receiving equipment.

One of the key operating features for the BSS is to position the satellite so that elevation angles in the service area are 20^o or greater, while at the same time having the satellite at least 15^o west from the edge of the service area so that eclipse will occur after 1:00 a.m.. In Region 2 for CONUS coverage, the most favorable orbital segment will be from 105^oW to 155^oW longitude, with optimum positions being 115^oW for the eastern satellite with East and Central time zone beams and 143^oW for the western satellite with Pacific and Mountain time zone beams.

By employing space and polarization diversity, so that for example, the East and Central time zone beams would have right and left hand polarization transmission respectively, it is possible to obtain some frequency reuse of the 500 MHz band and obtain on the order of 32 TV channels for CONUS coverage.

One of the important problems facing BSS in the U.S. is that the 12.2-12.7 gHz is also currently used for private terrestrial microwave radio with some 1400 licenses already issued in that band. These systems would have to be moved to 18 gHz or some other band to avoid causing excessive interference into the individual satellite receiving earth stations. This is one of the controversial problems still under consideration by the FCC.

5.0 ALTERNATE DELIVERY SYSTEMS

Other controversial issues relate to alternate delivery systems for entertainment television signals. Currently television signals are delivered to the home using fixed satellite services combined with local distribution via a CATV system or through community broadcast via a local TV broadcast station, or through an MDS station. Television signals are also distributed by the major broadcast networks to affiliate stations via an extensive terrestrial microwave and cable network. Some people have argued that the BSS should be used primarily for new services such as HDTV or electronic publishing. Others argue that it will permit service to areas not well served by existing delivery systems such as the Rocky Mountain or Alaskan areas, and this type of service alone is enough to justify a BSS. Still others feel it will complement existing services by offering specialized programming for subscribers who are willing to pay for this service.

The BSS is a one way information delivery system capable of reaching a very large receiving audience at reasonable cost. As indicated in the section on service requirements, this delivery system can be used for a large number of services in addition to general TV broadcasting, many of which are presently not available. What services this delivery system will be used for will depend on regulatory policies adopted by the FCC and internationally at

RARC-83, and by the marketplace. It is likely that BSS will be used for a variety of applications.

6.0 SUMMARY

It is evident that a BSS is technically feasible today and since WARC-77 (the first international conference to come up with a BSS plan), that considerable progress has been made in spacecraft technology in the areas of:

- High Powered more efficient TWT Amplifiers (100-700 watts)
- Shaped Beam Antenna Designs
- Improved Prime Power Solar Cell Arrays
- Improved Station-Keeping and Altitude Control
- Lightweight r-f Filters
- Microwave Integrated Circuits for LNA and HPA applications as well as in mixers, combiners, and other r-f subsystems.

In the individual receiving earth station significant improvements in performance and lower cost have been realized through the use of newer Microwave Integrated Circuits (MIC) and devices such as GaAs FET low noise amplifiers, and low noise solid state down-converters. Monolithic MICs offer the hope of even lower costs in the future. Advances have also been made in antenna designs both in lower production costs and installation costs as well as improving sidelobe response through offset feed arrangements.

While BSS plans have been discussed at WARC-77 and WARC-79, it will be up to the RARC-83 meeting scheduled for 1983 to come up with a BSS plan for Region 2 covering the Americas region. It is likely that the 12.2-12.7 gHz band will be provided for the satellite downlink transmissions, and the 17.3-18.1 gHz band will be used for the uplink transmission.

The orbital arc for Region 2 extends from 37^{o} to 170^{o}West longitude. This is to be shared among 32 countries who are in Region 2. This would permit 22 satellite locations with 6^{o} separation and for concurrent service to North and South America satellites could be separated by 1^{o}. However, requirements for minimum elevation angles greater than 20^{o} and eclipse outages occuring after midnight limit the CONUS coverage orbital arc to $105-155^{o}$ leaving some eight satellite positions.

It would appear that 4 or 5 standard TV channels could be provided per service area, with small countries having a single service area while larger countries such as the U.S. might have four or more service areas such as one corresponding to each time zone. Through the use of space and polarization diversity it will be possible to have some frequency reuse and obtain as many as 32-20 MHz television channels across the 500 MHz band, made available for BSS. Circular polarization is the likely transmission mode in the downlink to avoid having to require polarization orientation of the receiving antenna, however, linear polarization offers superior performance with respect to propagation behavior in rain.

FM modulation is the preferred modulation technique today, however, digital transmission may ultimately be the preferred transmission mode because of superior performance.

It is evident that BSS as an information delivery system, can be used for a large number of services and today there are eight applications pending at the FCC for such a variety of services. In the U.S. the pacing item for operational use of such systems, will be determined by regulatory and business environment, not the technical area.

References

[1] B. Patton, "Technical Aspects Related to Direct Broadcast Satellite Systems," FCC Report FCC/OST RA 80-1, Sept. 1980.

[2] USSG Document BC/838 for CCIR, Draft Report, "Factors to be Considered in the Choice of Polarization for Planning the BSS," May 1977.

[3] H. Akima, "Sharing of the Band 12.2-12.7 gHz Between the Broadcasting Satellite and Fixed Services," NTIA Report 80-32, Jan. 1980.

[4] WARC-77 Final Acts

[5] WARC-79 Final Acts

[6] P. H. Sawitz, "The Effects of Geography on Domestic Fixed and Broadcasting Satellite Systems in ITU, REGION 2," Proc. of AIAA 8th Comm. Satellite Systems Conf., April 1980.

Figure 1

REFERENCE CASE PROTECTION RATIOS RELATIVE TO CO-CHANNEL VALUES

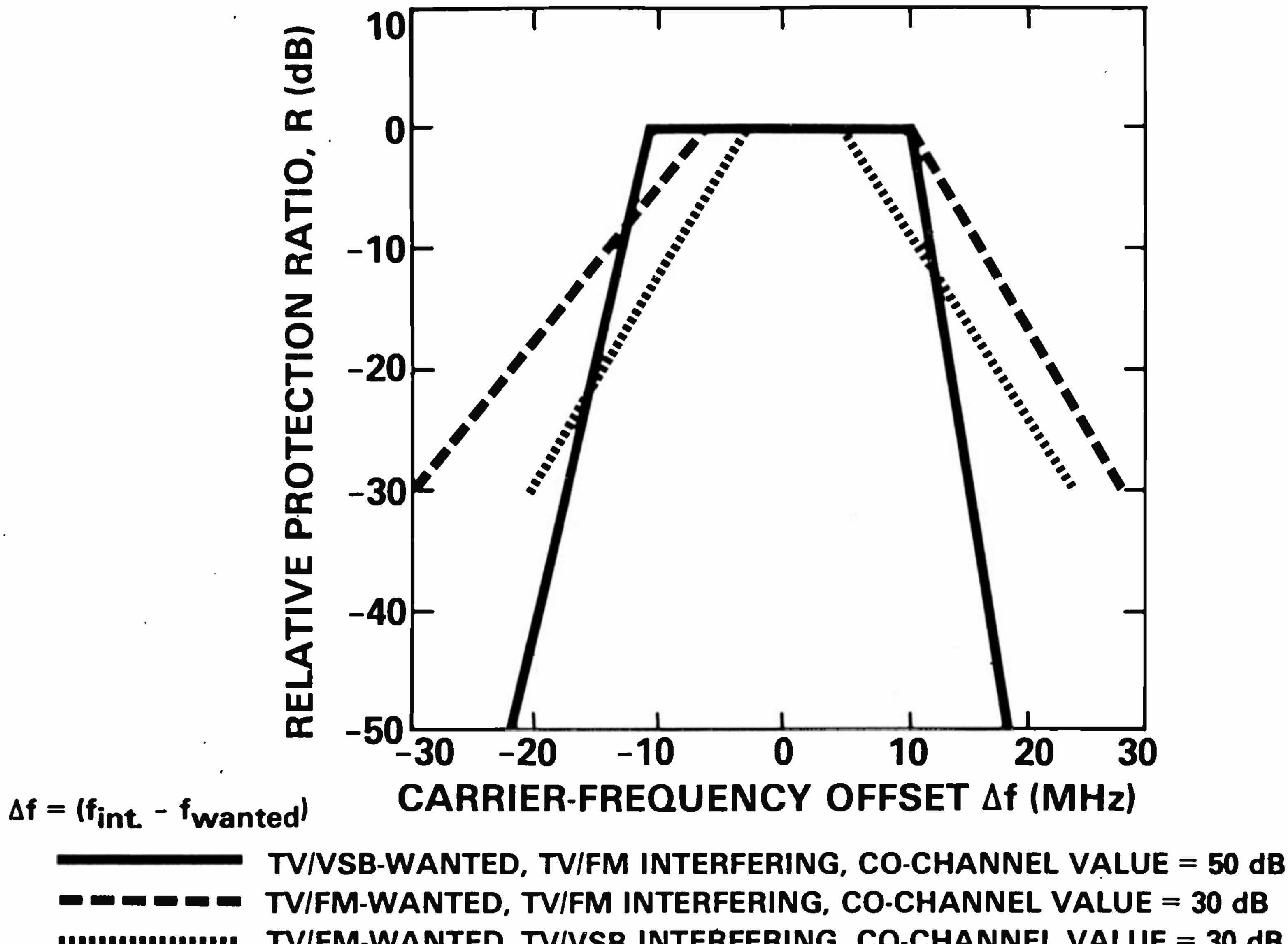


Figure 2

ITU REGIONS OF THE WORLD

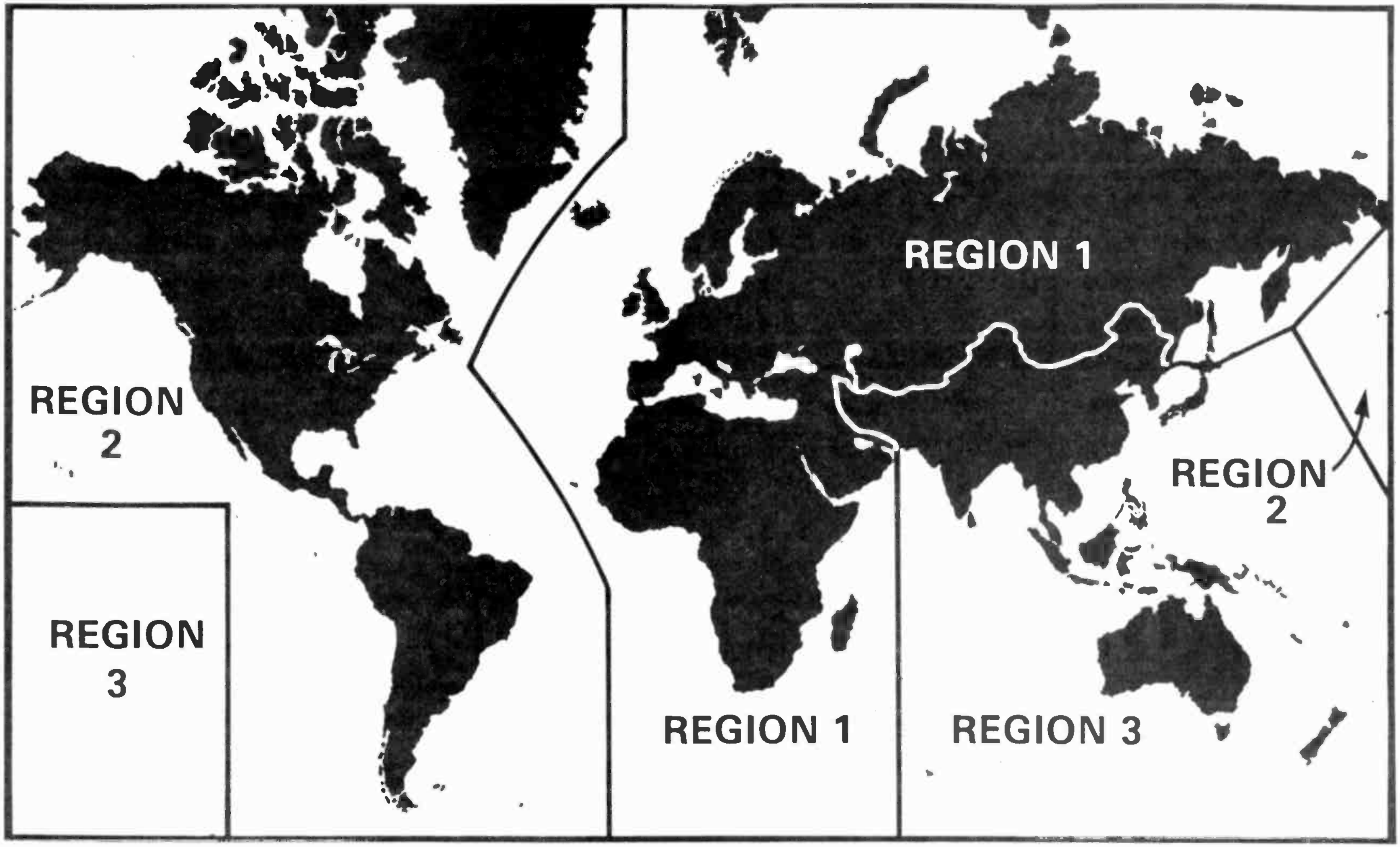


Figure 3

WARC-77 REGION 2 SEGMENTED ARC ASSIGNMENT

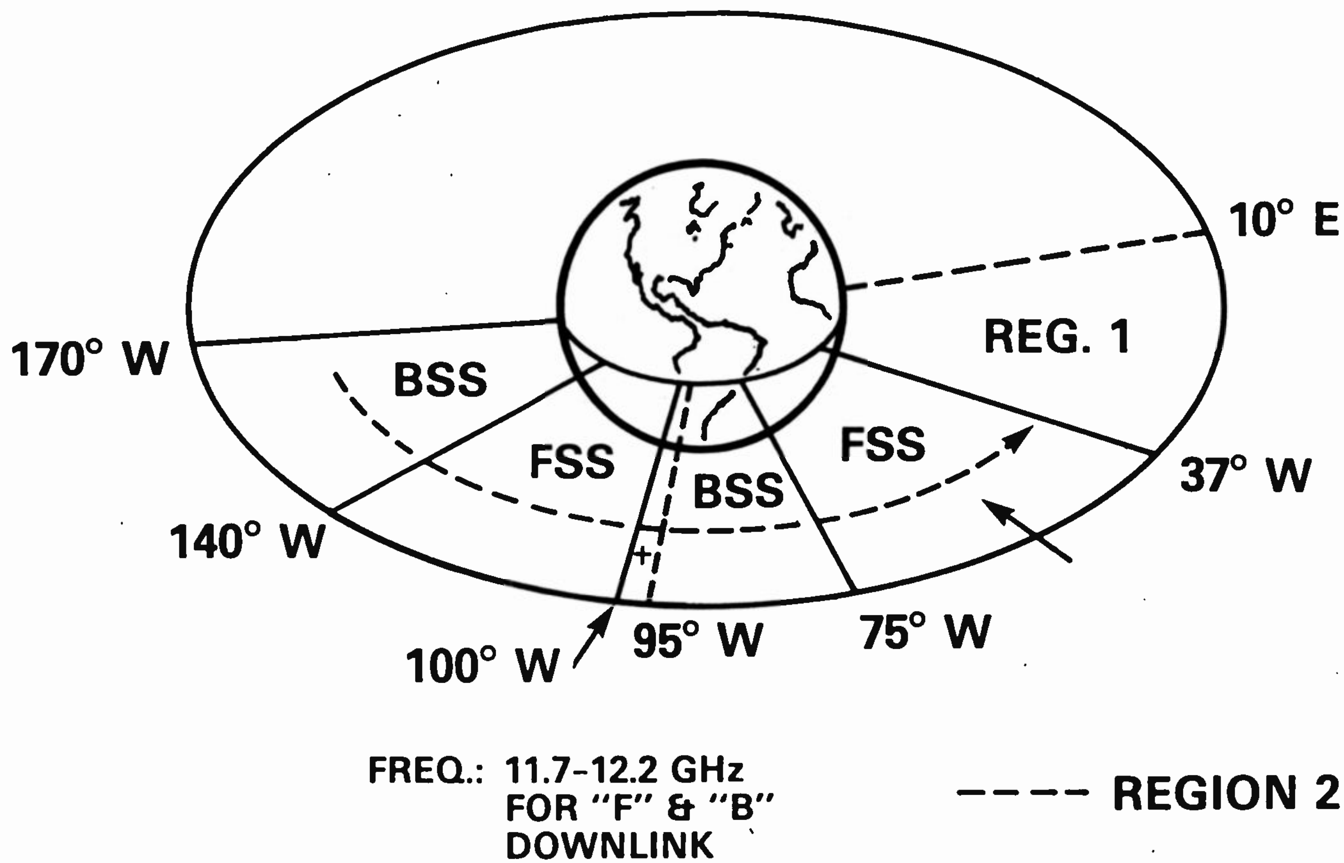
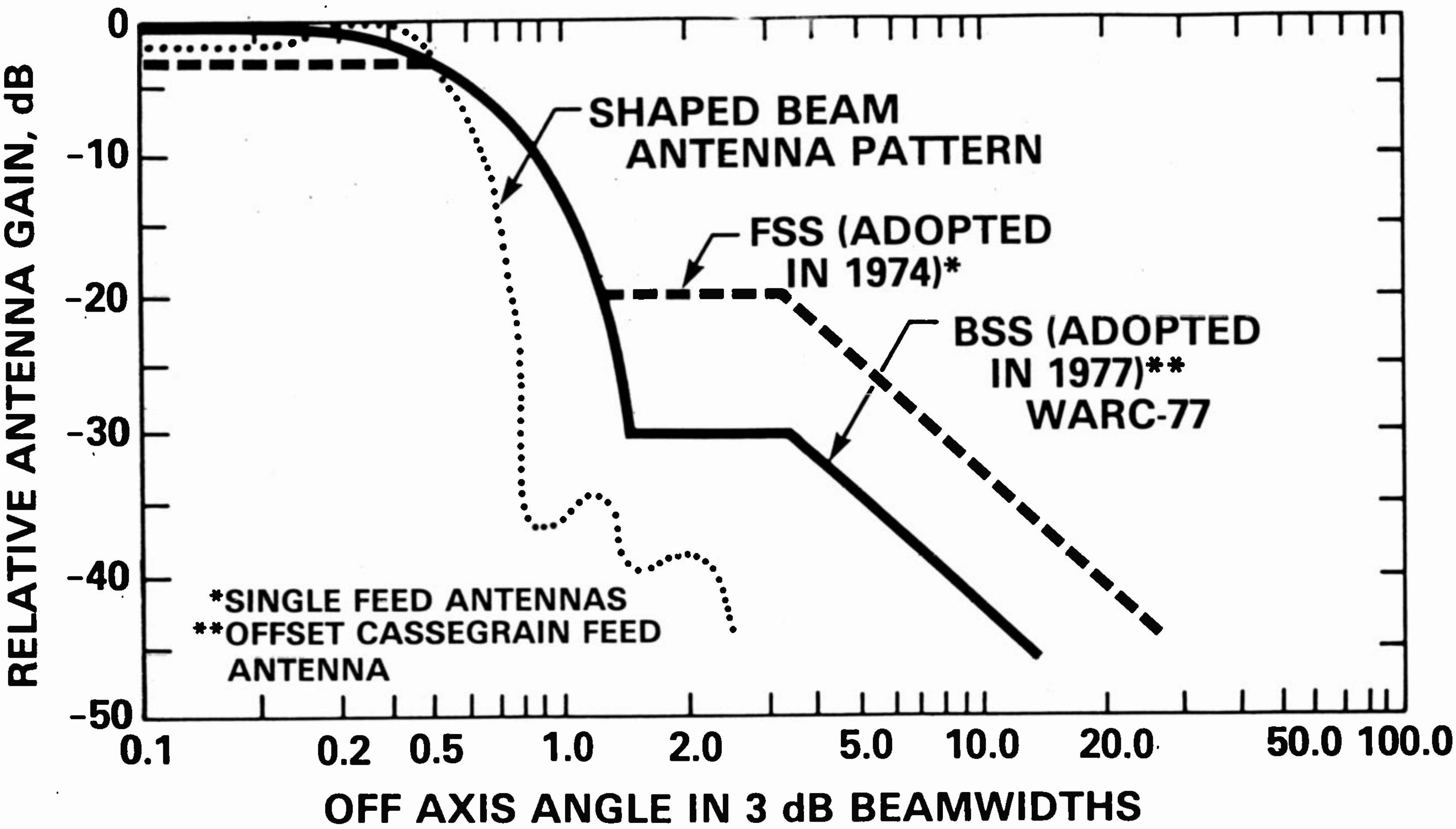


Figure 4

REFERENCE PATTERNS FOR SATELLITE TRANSMITTING ANTENNA



NOTE: THE VALUE OF 35 dB IS FREQUENTLY USED FOR THE NECESSARY PROTECTION RATIO BETWEEN TWO SATELLITE NETWORKS (31 dB TOTAL FOR CO-CH, 15 dB FOR ADJ.-CH INTERFERENCE)

Figure 5

REFERENCE PATTERNS FOR CO-POLAR AND CROSS-POLAR COMPONENTS FOR RECEIVING ANTENNAE FOR INDIVIDUAL RECEPTION IN REGION 2

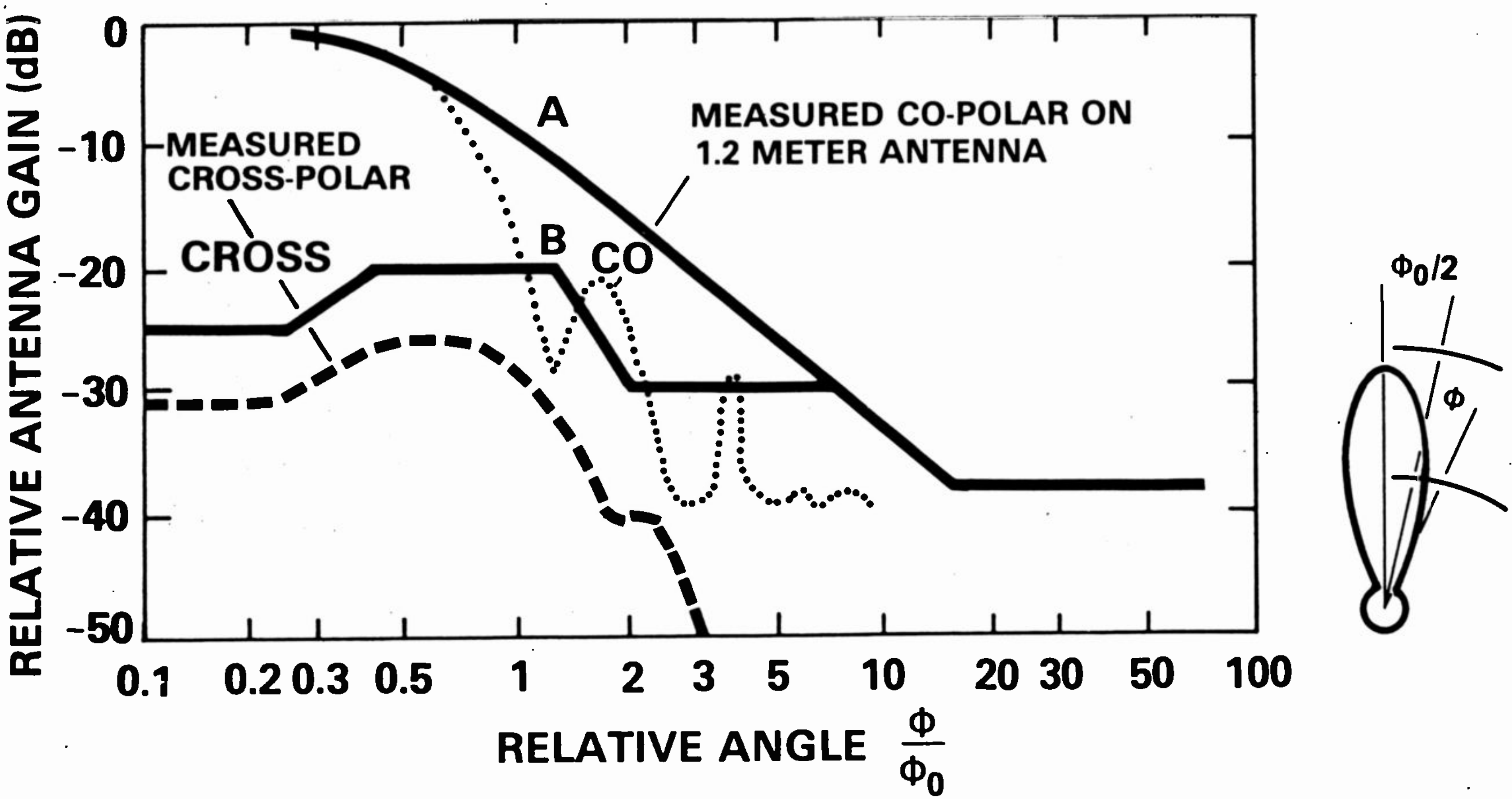


Figure 6

CO-POLAR RECEIVER PATTERNS FOR INDIVIDUAL RECEPTION

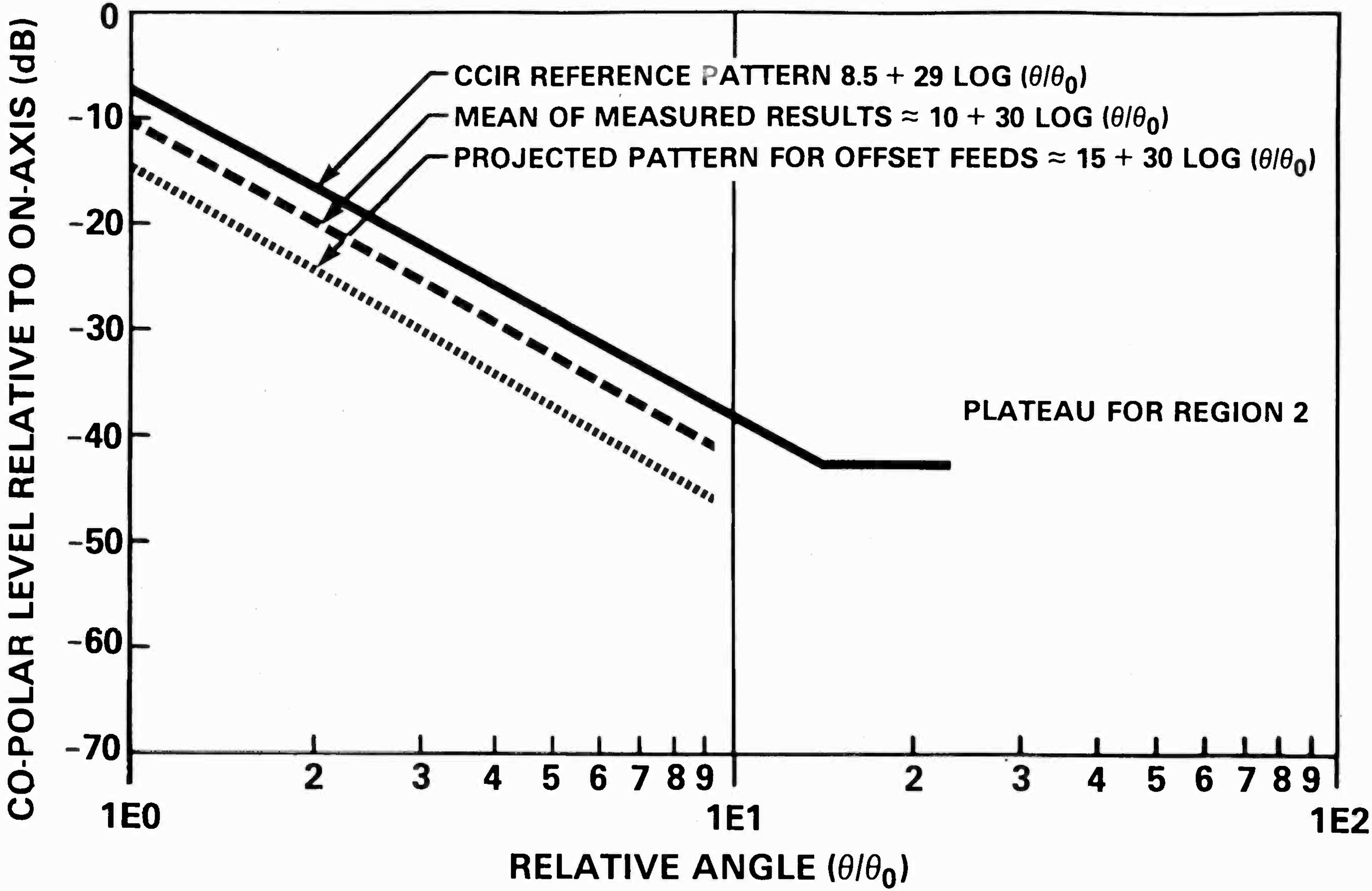
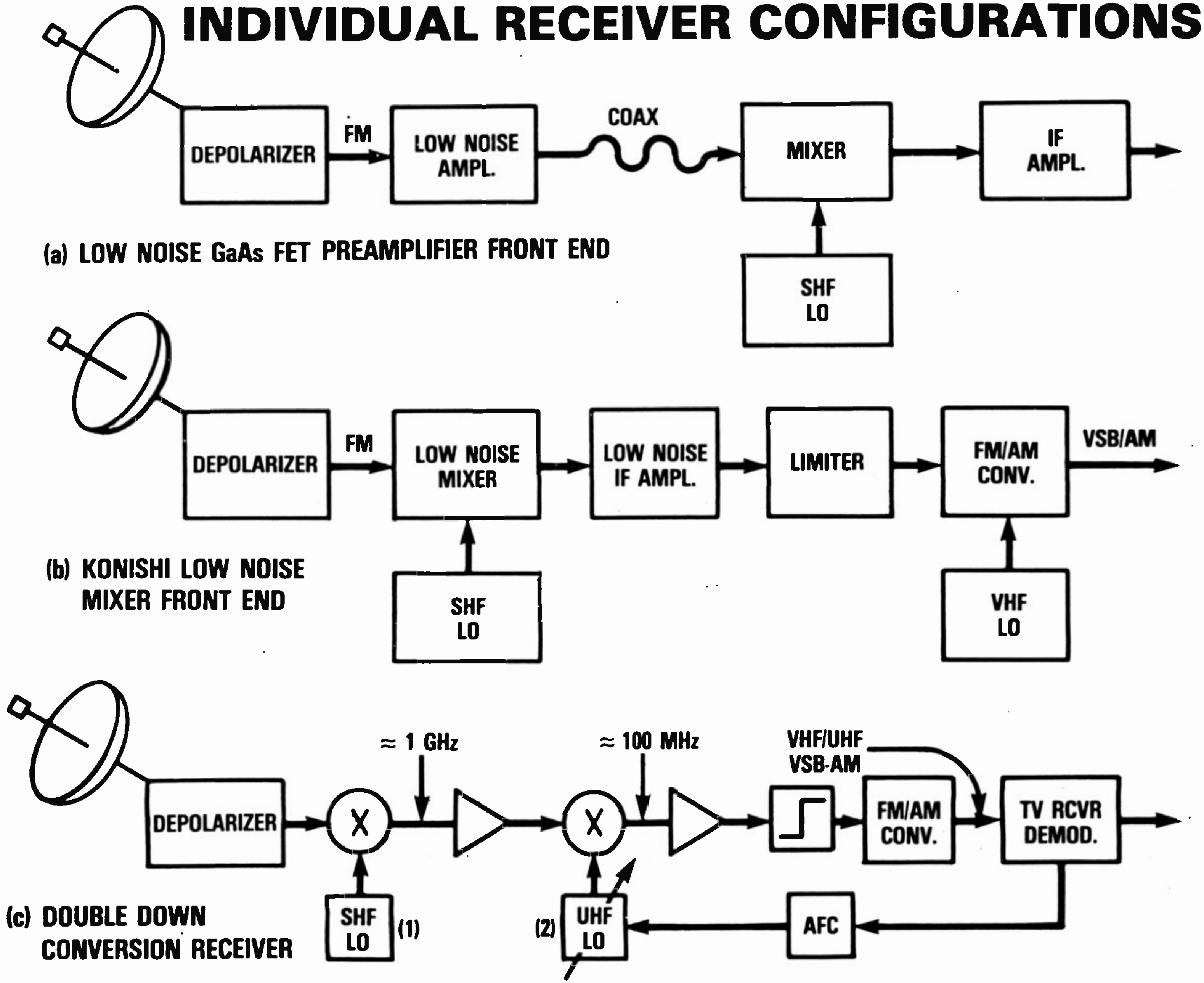


Figure 7

INDIVIDUAL RECEIVER CONFIGURATIONS



(a) LOW NOISE GaAs FET PREAMPLIFIER FRONT END

(b) KONISHI LOW NOISE MIXER FRONT END

(c) DOUBLE DOWN CONVERSION RECEIVER

Digital Audio: Where It's Been, Where It's Going

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Currently, the Broadcaster is using analog open reel-to-reel machines and endless loop cartridge machines for most locally recorded program material. He is also using the stereo LP disc which is used primarily for reproduction of music. Obviously, each of these two systems, both the analog tape recording and today's disc, contain certain deficiencies. In both systems we have reached a plateau in so far as further improving the performance.

Manufacturers and researchers are looking to alternatives for further improvement in recorded sound. These alternatives can be categorized by the word 'digital'. Presently there is nothing new in digital. The North American Indian smoke signals and the African tom-toms are digital in substance as was the Morse Telegraph. As far back as 1926, Nyquist showed mathematically how to process speech and music to transmit them digitally. Recently, technology has advanced sufficiently to make digital systems viable. Various digital stand-alone devices, such as delay units, pitch changers, and flangers, which are popular with 'Pop' groups, have appeared.

In the late 1960's the BBC initiated distributing its networks up and down the British Isles by a digital pulse code modulation system. Television sound has been carried along with the picture by the 'digital sound in syncs' method since the early 1970's in both the British Isles and the European Broadcasting Union's Continental Network. In 1972 the BBC developed a stereophonic digital audio recorder. In 1975 the BBC developed a multi-channel digital audio recorder. Both of these recorders were open reel-type recorders.

In 1982 we will see the introduction of the digital audio disc. This disc will be 12 centimeters in diameter, approximately 4.7 inches, and will overcome many of the shortcomings in today's LP records.

In order to provide a better feel for what digital audio sounds like, I will now play a demonstration digital recording. This recording is being played on a 3M digital audio mastering system which makes use of 1/2 inch tape. It is a four-channel machine which uses 50 kHz sampling rate and 16 bit quantization. The tape speed is 45 ips.

A digital audio system contains up to five distinct sections: input analog; analog-to-digital conversion; digital processing and storage; digital-to-analog conversion; and output analog. The two conversion sections can be designed using any number of conversion techniques. They can all be viewed as information transformations between the analog domain and the digital domain.

The analog domain is generally thought of as a voltage which can take on infinite resolution between some maximum and minimum level. To describe the digital domain we will only consider PCM or linear pulse code modulation since it generally is considered to provide the highest possible quality.

In the digital domain all information is represented by a bit value (1 and 0) in a word which generally consists of 16 bits. In order for each digital word to represent a signal which originated in the analog domain, each word is assigned a region of the analog signal range. This requires that the analog domain be divided (quantizied) into the same number of regions as there are digital words. In this fashion a single analog voltage is converted into a single digital word. However, the audio signal is time varying, thus requires that we partition the continuous time variable into a discrete series of time points referred to as 'sampling time'. Thus, a sequence of digital words is generated at the same rate as the sampling occurs. All changes in the analog signal between discrete sampling times are ignored. If the analog signal is band-limited relative to the sampling rate, the information in the sampled analog value is identical to that contained in the complete unsampled analog signal. Even though the sampling process ignores all signal changes between samples, no information is lost.

In a complete digital audio system, the incoming analog signal passes through a sharp, low-pass filter to restrict the bandwidth to a frequency below the Nyquist frequency. The signal is then sampled and each sample is held to allow the analog-to-digital converter time to convert the information into a digital word. Once in the digital domain, the signal can be either stored or transmitted, or any other number of functions can occur. At the output, the reverse process takes place. The sequence of digital words is converted into a discrete series of analog voltages by the digital-to-analog converter. An output, low-pass filter smoothes the discrete analog samples, thus the conversion back to the analog domain occurs. The only source of degredation in this system is the low-pass filtering and the quantization process at the input. The digitalization process does not create degredation. A band-limited, time-sampled, quantized analog signal has the identical information as the sequence of digital words. If we ignore technological imperfections, such as the imperfections in magnetic tape which result in drop-outs, it is possible to make several generations of dubs without any degredation of quality in digital recording.

To enable digital recordings to be exchanged between the various types of users, it can be shown that the parts of the analog-to-digital conversion previously discussed should be compatible. To enable this to be achieved, it will be necessary to standardize at least the basic fundamental parameters such as sampling frequency and bit rate.

There is already in existance, equipments in which the sampling frequency range varies from 32 kHz to 54 kHz and bit rates which vary from 10 bits per word to 16 bits per word. Various organizations, such as the European Broadcasting Union, the Audio Engineering Society, the Digital Audio Disc Committee of Japan, and the International Electrotechnical Commission, have held meetings to discuss the possibility of arriving at suitable standards and recently, as a result of these various discussions, it is becoming apparent that there will be at least two standard sampling frequencies - 48 kHz and 44.1 kHz.

To enable different manufacturer's equipment to be interconnected for dubbing, etc., it will be necessary to have an agreement on the digital-to-digital interface. A number of proposals for this interface have been put forward, but, at present, no consensus appears to be forthcoming in the near future.

Other components of the digital system will no doubt be standardized, if only to reduce manufacturing costs, but those previously mentioned are absolutely necessary before the industry will be in a position to go digital.

You may ask what all this has to do with me as a Broadcaster. During this year, perhaps late fall, we will probably see the introduction of the digital audio disc. Initially, the available library will be very limited. Depending upon the acceptance of the digital audio disc in the consumer market, we might see rapid growth of available music on digital discs. There is no question that the quality of audio from digital disc will far surpass that which is achieved on today's LP records. The recording industry is already making extensive use of open reel mastering-type digital reel-to-reel machines such as the one you just heard. I believe that it is inevitable that the broadcast industry will soon enter the digital domain. Digital audio discs will come in to common use. Digital open reel machines will be used for production work. Digital recording systems, similar to the cartridge machines, will be developed. Digital mixing consoles will become available, and the transmission of audio will be done digitally. The limiting factor to audio quality will eventually be restricted by microphones and loudspeakers.

The New Region 2 AM Broadcasting Agreement

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A new AM broadcasting agreement for the western hemisphere (Region 2) was concluded in Rio de Janeiro, Brazil November 9 - December 18, 1981. The United States is a party to the Agreement, subject to ratification by the Senate. The new Agreement will be the basis for development of AM broadcasting for at least the next ten years.

Up until this point AM broadcasting in the United States has developed under two long standing agreements - the North American Regional Broadcasting Agreement (NARBA) and the Agreement Between the United States of America and the United Mexican States Covering Radio Broadcasting in the Standard Broadcast Band (U.S./Mexican Agreement).

The NARBA countries include Canada, Cuba, Bahama Islands, and the Dominican Republic. Cuba ceased complying with NARBA when Castro came into power. The Bahama Islands and the Dominican Republic have had some difficulties complying with the Agreement and have caused some problems for the United States. The NARBA has become, in effect, a bi-lateral agreement with Canada. However, Canada has become dissatisfied with certain aspects of the Agreement, particularly with regard to clear channels. In December of 1981 Canada joined by the Bahama Islands stated that they were going to withdraw from the Agreement. This would leave us with an agreement with Mexico, but not with any of our other neighboring countries.

The U.S./Mexican Agreement has served both countries reasonably well and neither country has given any indication that the Agreement should be discontinued.

Since 1975 various steps were taken by the Inter-American Telecommunications Conference (CITEL), and countries in our Western Hemisphere looking towards an AM agreement for Region 2. Regions 1 and 3 (Africa, Asia, Australia, and Europe) concluded AM Broadcasting Agreements in 1975. The Region 2 steps included formation of a working group of broadcasting specialists which developed technical criteria and a draft agreement during nine meetings, with the United States active in all of the meetings. In March of 1980 the First Conference was held in Buenos Aires where the technical criteria was developed, which was needed for development of a regional plan. Agreement could not be reached on the issue regarding 9 or 10 kiloHertz channel spacing, so that decision was delayed until the Second Conference. A Panel of Experts, with representatives from eight countries, including the U.S.A., was assembled to help the International Frequency Registration Board (IFRB) to make a comparative study of 9 versus 10 kiloHertz channel spacing. The Panel of Experts met in Geneva for eight weeks in the spring of 1981 with very inconclusive results regarding channel spacing and this was reported to the Second Conference.

The United States was one of the main proponents of 9 kiloHertz channel spacing at the First Conference. Between the First and Second Conferences, studies which were made by the Federal Communications Commission and the industry convinced the Commission that the costs involved to the broadcasters and the industry, such as conversion costs, receiver obsolescence, and loss in service due to adjacent channel interference outweighed any potential benefits. Therefore the U.S. Delegation to Rio supported 10 kiloHertz channel spacing, which of course reversed the very active U.S. position for 9 kiloHertz at the First Conference.

The first major task of the Rio Conference was to decide on the channel spacing to be used in our hemisphere. Cuba was the main proponent of 9 kiloHertz channel spacing and they received support from Denmark and the United Kingdom. With the active support that 10 kiloHertz channel spacing had from the U.S.A. and thirteen other nations at the Conference 10 kiloHertz became a unanimous decision.

Following are some important time aspects of the new Agreement:

> The Final Acts of the Conference entered into force January 1, 1982. This resulted in all assignments in the Basic Inventory (existing stations and proposed stations which will be on the air by December 31, 1982) being listed in the Geneva Master Registry. This action was important in regard to achieving international protection for those assignments.
>
> The Agreement will enter into force July 11, 1983. This should permit sufficient time for the Senate's ratification process.
>
> The Agreement is intended to remain in force for about 10 years from date of entry into force. It will remain in force until it is revised by a competent Region 2 Conference.
>
> There are very specific procedures in the Agreement regarding modification of the Plan by new or modified assignments and regarding the status of assignments not brought into service within a 4 year period.

IFRB will record acceptable assignments in the Master Register and will play a role in resolving incompatibilities as required, or as requested by administrations. It is expected that the U.S. assignment process will continue much as it is now with the Commission only granting applications which are in accordance with the terms of the Agreement and entering into bi-lateral discussions with other countries, primarily Canada and Mexico, as needed.

The Agreement is, of course, binding between contracting parties, but not with non-contracting parties. However, non-signatory countries may accede to the Agreement upon deposit in Geneva of an instrument of accession.

Twenty Six of the thirty four countries in Region 2 participated in the Conference. Not participating in the Conference were two close neighbors of the U.S.A. - the Dominican Republic and Haiti. Non-participating countries will have the assignments contained in their inventories protected until August 1, 1982, which will be an incentive for them to become parties to the Agreement.

Cuba, which was a participant of the Conference, withdrew on December 14th. They cited two main reasons for withdrawing. The first was the Conference decision to not agree to a proposed shift of 48 assignments, involving 28 frequencies on an all or nothing basis.(A number of these assignments would cause intolerable interference to U.S. stations). The second was the U.S.A. announcement of its intention to implement a Radio Marti operation to beam radio programs to Cuba. Since Cuba was the sole participating, non-signatory country the Conference decided to give their assignments protected status only to January 1, 1982. After that date their List B assignments would be unprotected, would not be taken into account in calculating interference, and they could not prevent our assignments in Category B from going into Category A.

The Agreement contains all of the technical provisions needed to develop a plan of assignments for the hemisphere. Most of the provisions are very similar to the technical provisions used in the U.S. with some different descriptive terms and, of course, use of the metric system. One of the main differences relates to our present use of a 10% of the time interferring propagation curve. All other countries, except Canada, Mexico, Greenland, and the French Department of Saint Pierre and Miquelon will use a 50% of the time interference curve.

The Plan at the present time contains all assignments in two categories - List A or List B. List A contains assignments which do not cause or receive unaccepted interference. Most of our stations are in List A, some with their daytime operation in List A and nighttime operation in List B. But,there are numbers of stations, particularly our Class A's, and Puerto Rican stations, which are in List B because of unacceptable interference received, mainly from Cuba, or Venezuela. With Cuba not becoming a party to the Agreement and if modifications are made to various Venezuelan stations, many of the stations in List B will move into List A. Post-Conference procedures which were developed recognize the need to resolve remaining interference problems.

The Post-Conference period is a critical period devoted to verifying and making corrections in the Plan. The IFRB had computer support for the Conference, but for various reasons they had many problems which resulted in a considerable number of errors in the existing Plan which must be corrected.

January 1 - August 1, 1982 is a period devoted to verifying and making corrections in the Plan. By March 31, 1982 administrations are to correct all errors in their assignments listed in the Plan and notify the IFRB.

Annex 2 in the Final Protocol contains the procedure for resolution of incompatibilities and protection of assignments appearing in the Plan during the Post-Conference period. Under these procedures it is important that interference problems involving stations in List B be resolved by December 31, 1983, since protection levels will be established at that time.

Bi-lateral meetings will be held with other countries the next few months to accomplish this task. One such meeting has been held with Canada in an attempt to resolve problems - mainly on each others clear channels. Additional meetings are scheduled.

The U.S. Delegation to Rio faced a monumental task. 5,000 of the 15,000 assignments in the hemisphere were U.S. assignments which had been made under the terms of two sub-regional agreements. The results of these assignments provide a very good, unique, aural service to the U.S. public. The U.S. Delegation needed to preserve our existing American broadcasting service and system and seek to blend it with the needs and systems of the other countries in our hemisphere. It is believed that this was accomplished through the efforts of the members of the U.S. Delegation and its Chairman, Kalmann Schaefer.

It is anticipated that the errors which presently exist in the assignment plan will be corrected. It is also expected that most of the interference problems will be resolved through negotiations. However, the problem presented by Cuba is serious. With Cuba withdrawing from the Conference and not being a party to the Agreement there is no apparent means of resolving existing or future interference problems.

There are some 187 assignments in the Cuban inventory. None of them employ directional antennas to restrict radiation in the direction of other stations. Seven of their assignments propose operation with power in excess of 50 kilowatts (2-500 kW, 1-200 kW, 2-150 kW, 2-75 kW). This is far in excess of the power needed to provide a domestic service and will only serve to result in devastating interference to service presently provided by stations operating in the United States. Other stations operating with powers of 50 kilowatts or less will also result in very serious interference to U.S. stations since they will not operate with directional antennas designed to protect operating stations, or with power reduced to the level necessary to protect existing stations. The operations identified by Cuba in their assignment plan will not only cause interference within Cuba between their own stations operating on the same frequencies, but also to neighboring countries including, of course, the United States.

Cuban operations presently cause and will cause increased daytime interference to stations relatively close to Cuba, particularly in Florida.

During nighttime the Cuban operations cause, or will cause, extreme interference to areas over most of the United States. For instance:

(1) Ten clear channel stations will have their secondary service area destroyed and their primary service area severely limited.

(2) A great many stations which operate on regional channels will lose a considerable amount of their primary nighttime coverage.

(3) Over 200 stations in the United States will be affected in varying degrees by the Cuban assignments.

The National Association of Broadcasters (NAB) has a report "Cuban Interference to United States AM Broadcast Stations" which graphically illustrates the Cuban interference problem.

In an attempt to try and protect the service areas of stations in the United States from Cuban interference the United States took a reservation to the new Agreement which reserves the right to take such action as may prove necessary to assure the provisions of needed services to adversely affected areas. In accordance with this resolution the Federal Communications Commission has authorized several Florida stations to operate with increased power in an attempt to counter interference received from existing Cuban operations. However, U.S. stations are limited by the technical means they can employ to counter Cuban interference. For instance, increases in power by U.S. stations may cause serious new interference to other U.S. stations, or to stations in other countries, such as Canada and Mexico, contrary to existing Agreements, or the new Agreement. Also, there is no assurance that Cuba will operate with the facilities they have specified in their station inventory relative to power, operate on the frequencies specified, or change frequencies as they may. We are facing a situation regarding Cuban operations where attempts to combat interference received may not be effective and present a considerable cost to the licensees of stations affected.

The political aspects of the interference problem with Cuba are recognized. It is also recognized that this is only one aspect of our present relations with Cuba. In hearings held before the House of Representatives Committee on Foreign Affairs on March 3 and 4 and the International Operations Subcommittee on March 10 several witnesses urged Congress and the Executive Branch to explore every feasible means to reopen discussions with Cuba in an attempt to resolve the interference problems. Since Cuban interference also affects other countries such discussions could be on a bi-lateral, or multi-lateral basis. Every possible step must be taken by our government to resolve the very serious interference problem which has developed with Cuba.

CUBAN INTERFERENCE TO UNITED STATES BROADCASTING

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Abstract

AM radio stations in 34 states plus the District of Columbia will experience interference and reduced listening areas should Cuba proceed with its proposed inventory of broadcast facilities. Over 200 U.S. stations would be affected. Ten clear channel radio stations will lose their entire secondary service area and much of their primary groundwave service area. Thirty-seven clear channel stations would lose most of their skywave and some groundwave service area in particular directions. Technical modifications providing more than minor recovery of areas lost to interference are not feasible. Only limited power increases and antenna system design changes can be made within the constraints of domestic and/or international rules.

Introduction

At the November 1981 Region 2 (Western Hemisphere) Radio Conference in Rio de Janeiro, Cuba submitted an inventory of 187 AM radio stations to operate throughout their country. The inventory consisted of both existing and proposed stations including frequencies, power levels, and locations. Two of the proposed stations would operate at 500 kw. Other stations would operate at power levels that the U.S. would consider to be excessively large to serve a small country like Cuba: 10, 30, 50, 75, and 150 kw. Near the end of the conference, the Cubans attempted to amend their inventory in a manner that shifted interference away from other Region 2 countries and towards the U.S. In what is regarded as one of several major victories for the U.S. at Rio de Janeiro, the Cuban proposed amendments to their inventory were rejected by the Region 2 conference - even though the working group that included Cuba, the U.S., and other North American countries recommended that the Cuban proposals be accepted. As a result, the Cuban delegation withdrew from the conference and is now a non-signatory to the Final Acts of the Region 2 broadcasting conference. Cuba currently operates its radio facilities with no regard for adherence to international technical standards.

Review of Interference Calculations

To understand the complexity of Cuban interference calculations and predictions, we should first review the concept of interference and how the presence of interference is determined. Non-technical readers may wish to skip to the next section, although it is believed that the discussion presented while technical in nature is relatively easy to understand.

For single, co-channel interfering radio signals, interference at a particular location or within an area is said to occur when the ratio of the desired signal to the undesired signal is less than 20:1. Therefore, to avoid interference, the undesired signal must be at least 26 dB below the desired signal. The method used in predicting desired and undesired signal strengths depends upon the type of signal. In a groundwave calculation, the signal strengths of the desired and undesired signals are computed from knowledge of frequencies of the signals, the unattenuated radiation (inverse-distance field) of the radio facilities at the appropriate azimuth and the conductivity of the earth over the particular propagation path. For skywave calculations, the angle of departure from the transmitting station must also be known. Additionally, the skywave field is calculated using appropriate 10% of the time or 50 % of the time skywave curves. The 10% curves are used domestically for interfering signals.

The 50% curves are used for determining the extent of available secondary (sky-wave) service from class 1 "clear channel" stations.

The presence or absence of interference is more difficult to predict when multiple interfering signals are present. On U.S. regional frequencies, for example, as many as 30 radio stations or more will be operating at night. Since each facility radiates to some degree in all directions, each facility will receive an "interfering" signal from every other station on the channel.

To "organize" the interference and make its calculation technically meaningful, the so-called "50% exclusion principle" is utilized. This principle is needed to determine the actual interference-free contour when multiple interfering signals are being received. First, the individual interfering skywave signals are calculated. Each signal is multiplied by 20 to yield individual "night limits" which are in fact the groundwave signal strength values at which interference-free service would end if the single interfering skywave signal is present. Under the 50% principle, the individual night limits are calculated, sorted, and placed in decreasing order of signal strength. Starting with the strongest limit at the top of the list, the root-sum-square (RSS) of the first two limits is calculated. If the next smallest limit on the list is less than 50% of the calculated RSS of the first two limits, the calculation is completed. If the next smallest limit is greater than 50% of the RSS, the RSS is re-calculated to include this limit: i.e. the "new" RSS will have three contributors. 50% of the "new" RSS calculation is then compared with the subsequent limit appearing on the list. This process continues until a limit is found which is less than 50% of the RSS of all previous individual limits.

The 50% RSS, as calculated above, is known as the "night limit" of the radio station. New AM facilities are required to "protect" other co-channel full-time stations by not "raising their night limit-- e.g. a new facility must limit radiation in appropriate directions to avoid imposing a limit which is in excess of 50% of the existing RSS or in excess of the smallest individual limit contributing to the existing RSS. Should either of these conditions occur, the interference limit at the protected station would be considered to have been raised. Because the night limit defines the outer boundary of interference-free reception, higher night limits result in smaller interference-free service areas. As an example of the 50% exclusion principle, the following interfering limits are presented:

Station No. 1 - 1.0 mV/m
Station No. 2 - 0.60 mV/m
Station No. 3 - 0.59 mV/m
Station No. 4 - 0.58 mV/m

The RSS value from Nos. 1, 2, and 3 is 1.31 mV/m: therefore interference from No. 4 is excluded from the calculation because it is less than 50% of 1.31 mV/m.

The problem of directional antenna design is to serve the city of license while meeting co-channel protection requirements. However, radio stations cannot be on the same frequency at the same location: <u>therefore every AM facility has a unique set of protection requirements not duplicated anywhere in the world</u>. This is the reason why the allocation structure of the U.S. system of AM broadcasting is so complex and carefully engineered. The situation is not unlike a jigsaw puzzle, with every station inexorably intertwined with every other station.

There is an additional complication involved in predicting interference. Because new stations are required to protect all existing stations, the night limits of older stations are for the most part lower than the night limits of relatively new stations. This leads to the possibility that two AM stations can be equidistant from an "interfering" facility radiating the same amount of signal toward each station yet only one station is considered to be receiving interference by having its night limit raised.

Cuban Interference Calculations

Calculating interference requires knowledge of the interfering station's operating parameters. In the case of Cuba, of course, we are unable to "read the meters" and otherwise observe their operation. Comprehensive measurements must be performed in an effort to estimate the Cuban operating parameters on a particular frequency. Specifically, we need to know the approximate location of the Cuban facility and its unattenuated radiation in particular directions. Yet, even if this information can be determined, the prediction of skywave interference must be inexact because we would not know the electrical height of the Cuban transmitting towers. Since the Cubans are known to be changing operating frequencies, it is not likely that each tower would be operated at a consistent electrical height.

By using the Cuban inventory submitted at the Region 2 conference in December, 1981, we can predict how U.S. broadcasters would be affected if the Cubans build and operate any or all of these facilities. The inventory is specific with regard to locations, operating powers, tower heights, and frequency. When the interference has been predicted, field strength measurements can be studied and correlated to determine the extent to which Cuba is actually operating with the facilities proposed in the inventory. We are then able to better estimate the extent of real interference occurring in the U.S.

The complexity of calculating nationwide interference requires the use of a computer. First, we "propose" a radio station - a Cuban inventory entry. Next, in the case of regional channels, we need to determine which co-channel stations would have their night limits raised - a process which requires the calculation of each "50% exclusion" RSS at the transmitting site of each co-channel station. In turn, each RSS calculation requires the skywave field contributions to a particular night facility from every other night facility. Additionally, each skywave field contribution calculation requires 1) knowledge of the distance and bearing from the contributing station to the station whose RSS is being calculated, 2) the operating power and directional antenna information of the contributing station, and 3) use of the FCC curves regarding angle of departure and skywave field.

The computer programs developed at the FCC may be employed for the job of predicting interference. All of the above calculations can be performed at great speed, enabling the identification of regional stations receiving interference. Once the night limits are established, the service areas can be estimated by predicting the distances to the interference-free contours in a sufficient number of directions and by using the resulting root-mean-square (RMS) radius in the service area calculation. The difference between service areas with and without interference is the number of square miles where radio service to the public has been lost.

The calculation of interference to class 1 (class "A" under Rio technical standards) clear channel stations is somewhat more difficult. This is true because the presence or absence of interference to class 1 stations is not necessarily a yes/no calculation, but may depend on direction. The "regional" calculation is strictly a site-to-site calculation: if the night limit is raised, interference is occurring. For clear channel stations, the normally protected service area may be 1500 miles in diameter or more - consisting of a groundwave, or primary, service area "augumented" by a secondary, or skywave, service area. There can be a substantial difference in the predicted interfering skywave fields from one end of the class 1 service area to the other. Accordingly, it is possible that interference would be experienced in some portions of the class 1 service area, while others would remain interference free. It is thus necessary to treat interference to class 1 stations as a direction-dependent phenomenon. For each direction from the class 1 stations that is analyzed, a determination must be made as to the existence of a secondary service area. If a secondary skywave interference free service area exists, the individual limit (20 times the interfering skywave) must be found. The distance to this contour value can then be used in the RMS interference-free service area. On the other hand, if certain directions show that no interference-free secondary service exists, the calculated individual limit will result in an interference-free groundwave signal contour.

The class 1 interference situation is further complicated when there are multiple interfering skywave signals such as other class 2 nighttime facilities and perhaps another class 1B station on the channel. In these cases the interfering limits should be RSS'd to determine the resultant night limit in particular directions. If all secondary skywave service area is destroyed, the class 1 station can be effectively treated as a regional station in that a site-to-site RSS can be performed eliminating the need for direction-dependent calculations.

Results of the NAB Cuban Interference Study

Each station on the Cuban inventory was entered into the FCC interference computer program. When the dust (in this case the printer ribbon) had settled, it was found that AM radio stations in 34 states and the District of Columbia will experience interference and reduced listening areas should Cuba proceed with its proposed inventory of broadcast facilities. Altogether, over 200 U.S. stations are involved. Ten clear channel stations would lose their entire secondary service area and much of their primary service area. Thirty-seven clear channel stations would lose large portions of their nighttime coverage. Very few clear channel stations would remain interference-free.

Interference on clear channels will result in the greatest amount of service area lost per station. This is primarily because it takes relatively little interfering skywave signal to destroy vast areas of secondary skywave service area. The clear channels are thus most vulnerable to interference from Cuba.

Widespread interference can be caused by other "low power" Cuban stations as well as the Cuban 500 kw stations. The 30, 50, and 75 kw facilities on U.S. regional channels are of particular concern. For example, on 1040 kHz, the frequency tentatively selected for Radio Marti and a 500 kw Cuban proposal, only one U.S. nighttime station resides: WHO, Des Moines, Iowa. Consequently, only one U.S. station can experience co-channel interference. On 910 kHz, however, a 75 kw station would cause interference to 17 U.S. stations in 15 states from Florida to Iowa and Virginia. Florida would not be the only state that is affected: KAMC, Camden, Arkansas, would have its night limit raised from 8.15 to 1531 mV/m, a 56.1 percent reduction in interference-free service. Even low power stations have a tremendous potential to cause interference. A 5 kw Cuban facility on 950 kHz would cause interference to 5 stations in 5 states: WLSQ, Montgomery, Alabama (24.7% reduction); WLOF, Orlando, Florida (55.7%); WGOV, Valdosta, Georgia (56.7%); WSPA, Spartanburg, South Carolina (19.9%); and KPRC, Houston, Texas (1416%).

Solutions to Cuban Interference

Eight stations to date in Florida have applied and several have received Special Temporary Authority (STA) to change powers and/or antennas in the effort to recover service area lost to Cuban interference. For several reasons, the STA approach is unsatisfactory solution and should not be considered the whole answer. This is not to say that STA's serve no purpose; but for the majority of stations, the actual implementation of a STA is, in reality, a gamble on the intentions of Cuban broadcasting.

First, the affected station can never recover the entire service area lost to Cuban interference. It is a technical fact-of-life that less signal is required to create interference than to render satisfactory radio service. Accordingly, a doubling in power will not, in general, result in recovering the amount of service area one would expect. For regional stations, a simple power increase does not "remove" interference: it simply increases the distance to the interference-free contour. Doubling of station power will not double this distance. If a station increases power in an attempt to recover lost service area the interference has not been removed. The interference-free contour and the fact that the night limit has been raised remains unchanged.

Second, the design and construction of directional antennas for the purpose of recovering portions of service area lost to Cuban interference is a time-consuming, complex task. To meet FCC technical standards, the design engineer is faced with the formidable task of increasing radiation in certain directions but not others: the station must continue to protect other co-channel U.S. nighttime facilities while attempting to increase radiation in the directions where it is desired to recover service area. On regional channels, the situation is not unlike a "jigsaw puzzle" of radio stations where one piece near the edge is attempting to change its basic shape and still fit in the puzzle. Increased complexity of directional antenna design generally requires the addition of one or more radio towers, each of which needs more land for construction.

Third, if a suitable combination of directional antenna and transmitter power increase can be designed, the project may cost several hundred thousand dollars. This figure is substantially above the average radio stations's financing capability. Most radio stations have an average pre-tax profit margin of just $19,000.

Finally, there is no guarantee that the interference which governed the design of the "new" radio station will continue to exist or remain unchanged when the new facility is finally put on the air. In fact, a minor change in the Cuban operating parameters of location, frequency, or power will necessitate a re-design of the radio station to accomplish the original objectives of recovering service area lost to Cuban interference. While it may take six months to a year to design and build a directional antenna facility, it only takes a few minutes to change the operating frequency of a radio transmitter - and cause an entirely new set of interference problems for U.S. broadcasters.

Some people have maintained that many of the above technical problems can be solved if the FCC were to "relax" its technical standards in a way which allows the "interference burden" to be equitably shared by all nighttime AM broadcasters. This proposal is without merit. For the reasons discussed in an earlier section of this paper, each AM radio station is unique. Each AM facility has a unique set of protection requirements not duplicated anywhere in the world. Each station has a different night limit. Older stations have typically low nighttime limits while newer stations have higher limits.

The design of "interference-sharing" radio facilities is nearly impossible for the following reasons: a) a stronger signal is required to raise a high limit than to raise a low one; b) two stations in the same approximate direction or distance from a prospective "rebuilt" facility are unlikely to possess identical night limits; and c) radio stations are capable of one radiation value per azimuth and elevation angle. One could argue that if a high-powered Cuban station is involved, many U.S. stations will be interfered with by the Cuban with the result that any "additional" interference by "rebuilt" U.S. facilities would be "masked". Thus, the U.S. broadcast system conceivably could take advantage of the fact that interference imposed by an illegal Cuban operation could help some stations avoid overly complex directional antennas. However, it is important to note that the Cuban interference can not be "equitably distributed" - let alone eliminated - due to the inherent nature and structure of nighttime broadcast allocations. Some stations will be impacted to a greater degree than others - and this fact will not change regardless of any "relaxation" of FCC technical standards.

The clear channel stations are in a very difficult situation. While there are few co-channel U.S. night facilities the class 1 stations would have to worry about, raising power would be difficult because they are already at the FCC maximum 50 kw. Further, as it is easy to destroy secondary skywave service area, it is virtually impossible to recover it once it is gone.

If we in the U.S. cannot "repair" damage to our broadcast system caused by Cuba, what <u>can</u> we do? The technical answer to this question is: very little. Interference in AM broadcasting must be corrected at the <u>source</u> of the interference, not at the point of reception. In the U.S., we prevent interference by the widespread use of directional antennas. Nor, apparently, do they intend to. The political answer to the above question is: persuade Cuba to agree to employ the interference avoiding techniques long established for AM broadcasting. Such agreement is in Cuba's interest as well as our own. Both countries have legitimate needs for radio coverage. Our ignoring protection of Cuban stations because of the need to overcome interference from those stations hinders Cuba from providing the service they desire - as well as creating large areas where neither the Cuban station or the U.S. station can be received satisfactorily.

"DEVELOPMENT OF ANALOG SCPC AUDIO SYSTEMS"

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INTRODUCTION

The Mutual Broadcasting System, Inc. (MBS) is one of the pioneers in the use of domestic communications satellites to distribute programming to broadcast stations. MBS is the first commercial broadcast network to purchase and install its own uplink and downlinks and convert totally to satellite distribution. MBS' efforts to develop a satellite distribution system began in the mid-1970's, and in 1977 MBS contracted with the Western Union Telegraph Co. (WU) for capacity on the WESTAR satellite system for single channel per carrier (SCPC) service to be used in delivering programming to MBS-affiliated broadcast stations via MBS-owned earth terminals.

MBS was active, along with WU and others, in seeking a relaxation of the initial FCC regulations applicable to receiving earth terminals, as it was economically infeasible, for many potential applications, to install earth terminals which met the stringent requirements initially applicable. While awaiting the anticipated FCC deregulation of small aperture terminals, (SATs), MBS worked with equipment manufacturers and WU to specify appropriate hardware and operating parameters to perform the program distribution function in keeping with technical objectives. Immediately upon the FCC action deregulating SATS, MBS entered into a contract with California Microwave, Inc., (CMI) to provide hardware for its satellite distribution system, and to install earth terminals at MBS-affiliated stations, with 650 total systems acquired for installation in the first contract.

On March 1, 1982, MBS earth terminals installed and operating totalled 571, with 79 additional systems manufactured and awaiting, or in the process of, installation. Of these 650 terminals, approximately 50 were equipped with 15' (4.6M) antennas, approximately 260 with 10' (3.0M), and approximately 340 with 6' (1.8M), depending on the area of the country in which they were to be installed.

Many different factors must be considered in the design of an audio satellite distribution system. This paper describes these parameters and discusses the implications of various designs upon the finished operating system.

SIGNAL-TO-NOISE RATIO

There are many different standards which are used to determine the quality of a system. However, in all systems the final determination is how it sounds to the listener. Until system performance exceeds approximately 70 dB S/N, the signal-to-noise ratio (S/N) provides an excellent way to determine the quality of the transmission. It also provides a starting point in determining design criteria for the system. The signal-to-noise ratio for the system is dependent upon three items:

(1) Total carrier-to-noise ratio, $(C/N)_t$

(2) FM improvement factor, (FMI)
(3) Additional improvements, such as companding and emphasis

TOTAL CARRIER-TO-NOISE

For purposes of discussion in this document, the total carrier-to-noise ratio, $(C/N)_t$, is that received at the downlink on the output of the LNA. Total carrier-to-noise consists of:

(1) uplink carrier-to-noise ratio $(C/N)_{up}$
(2) downlink carrier-to-noise ratio $(C/N)_{dn}$
(3) total carrier-to-interference ratio $(C/I)_t$

$(C/N)_{up}$ is a function of the transmit power, location of the satellite, ground station and satellite parameters. $(C/N)_{dn}$ is a function of the downlink power, location of satellite, ground station and ground station parameters, (G/T).

IMPROVEMENTS IN THE SYSTEM DESIGN

Optimization of parameters can be achieved by the use of companding, emphasis and threshold extension demodulators. Audio distribution by satellite on a national level has been made possible by these concepts along with the use of small aperture antennas. For example, by using a compandor specifically designed for use in a satellite operating environment, small aperture antennas can be used with one-third the downlink power that would normally be required from the satellite. This results in a lower cost per terminal and more efficient use of the space segment, a consideration becoming more important everyday.

TOTAL CARRIER-TO-INTERFERENCE RATIO

The total carrier-to-interference ratio is comprised of several terms:

(1) Adjacent satellite interference
(2) Adjacent transponder interference
(3) Terrestrial interference
(4) Transponder intermodulation

Each of these forms of interference have a dramatic impact upon the frequency plan. The frequency plan utilized within the transponder must take all forms of interference into account when each of the individual operating frequencies are selected.

For the case of adjacent satellite interference, two common forms of interference to an SCPC system are FDM/FM and TV/FM. As a result of the spectral density of these types of signals, in the center region of the transponder, a total of 2 - 3.4 MHz is inappropriate for use in SCPC service.

In the case of terrestrial interference, a different portion of the transponder is affected. A typical terrestrial microwave system (using a standard 4 GHz spectrum plan) will have carriers located +/-10 MHz from the center of a transponder. SCPC carriers must also avoid this area of the transponder. The Mutual carriers are spaced not less than 2.1 MHz away from these carriers and are relatively unaffected by terrestrial interference.

Adjacent transponder interference provides yet another form of interference which must be avoided. The amount of interference, or IM noise, received from an adjacent transponder is limited by the spacecraft design parameters. The adjacent transponder interference will be a minimum at the center of the desired transponder, and increase as one approaches the transponder band edge. In the case of SCPC transmissions, acceptable performance is achieved by excluding the outer 6 MHz of the transponder.

Mutual has designed the bandpass filter in the downconverter to aid in the rejection of terrestrial and adjacent transponder interference. The filters pass all signals +/-8 MHz from the transponder center.

The final form of interference which affects SCPC performance is transponder intermodulation. This form of interference is a function of the transponder input/output power transfer curve, back-off from saturation, number of carriers, their power levels and occupied bandwidths. Methods to calculate and predict the intermodulation level in the transponder have been developed over the years, and have been the target of intensive computer analysis and modeling. Much of today's work on intermodulation analysis is based on a paper written by Wallace Babcock in 1952 in the Bell System Technical Journal [1]. Fang and Sandrin [2] did additional work on the intermodulation problem.

Today, using computer modeling programs, analysis can be done for any number of carriers under a variety of configurations. The particular transponder used by Mutual Broadcasting normally carries approximately 15-18 carriers serving such companies as Mutual Broadcasting, National Public Radio and MUZAK. By judicious selection of carrier frequencies, downlink power and earth station G/T, an optimum space segment design may be engineered.

FM IMPROVEMENT FACTOR

The FM improvement factor, a result of wideband frequency modulation, is a function of the highest modulating frequency peak frequency deviation and the noise bandwidth. The Mutual 15 kHz audio channel and the program channels used by National Public Radio are similar in terms of the FM improvement factor. In fact, the MBS system is similar to that of commercial FM broadcasting: 75 kHz peak frequency deviation and 15 kHz as the highest modulating frequency for the program channel.

ADDITIONAL SIGNAL-TO-NOISE IMPROVEMENTS

Other improvements in the signal-to-noise ratio are possible, such as companding and use of emphasis. In fact, both of these approaches are in use in many systems today. The MBS system uses both companding and emphasis.

The compandor used by Mutual was designed specifically for use in satellite transmission systems. It is also used by National Public Radio, who pioneered the design of it. The use of compandors in the audio chain is essential, as it is this device which allows a significant increase in the subjective S/N, 29 dB. The companding ratio used by Mutual is 3:1, and was selected only after extensive subjective listening tests. The use of companders in SCPC satellite system was discussed in a paper by Myron Ferguson [3].

Because such a large compression ratio is used, the performance of the compressor results in a better subjective improvement. However, it "does have greater sensitivity to gain variations in the transmission path" [Ibid]. For the Mutual system, the companding system is inherently part of the modulator/demodulator system. Therefore, gain variations in the transmission path are essentially eliminated.

Another area of improvement is in the FM demodulator itself. The use of threshold extension in demodulators is now a common practice. "The object of such a demodulator is to extend the linear relationship existing between the input carrier-to-noise ratio and the output signal-to-noise ratio, which in a normal wideband demodulator breaks down as the received carrier-to-noise ratio reaches small values" [4]. By the use of threshold extension, the demodulator threshold is lowered by 3 dB. Such an improvement has an impact upon the link budget, and thus upon the cost and parameters of the downlink.

Finally, the use of emphasis in the circuit must be evaluated. If any emphasis is used, pre-emphasis must be inserted after the compressor and the de-emphasis before the expander. Use of de-emphasis after the expander will result in an overload of the expander. Mutual uses 25 msec emphasis in its system. Because noise is related to the square of the voltage, noise power per unit bandwidth increases exponentially across the audio band at the rate of 6 dB/octave [3]. The use of emphasis must be analyzed with the use of the compandor system.

For Mutual, using the dBx compandor, compressor and expander feedback loops use emphasis in them to provide a "dynamic" type of companding. "The overall effect is to give masking [like that achieved in] multiband systems, but without the danger of high-frequency overload" [Ibid]. After extensive subjective listening tests, it was felt that the use of 25 msec emphasis is effective in low threshold environments. However, as the operating C/N increases above threshold, the effectiveness of emphasis becomes less pronounced. NPR, which also uses the dBx expander, does not use emphasis.

How all of these considerations add up to produce a final signal-to-noise ratio is of great importance. The Mutual system, using 6', 10' and 15' antennas have an operating signal-to-noise between 65 and 72 dB (PPL) depending upon antenna size. The frequency response is +/-0.5 dB from 50 Hz to 15,000 Hz. Distortion is less than 0.3%, typically 0.08% at 1 kHz. To interpret these figures, they have been compared to a well-engineered FM station in Washington, D.C., with a classical format. At that station, signal-to-noise, in mono, is 64 dB (PPL). For stereo, signal-to-noise is approximately 62.5 dB (PPL). It can be seen that the performance of an SCPC/FM transmission format such as that used by MBS will actually exceed that of a commercial FM broadcasting station.

SCPC/FM MODULATION

This leads to the question of analog SCPC compared to other types of modulation systems. The type of modulation desired is a function of the type of programming, total transmission path, end user and desired quality.

For example, using an SCPC transponder, satellite users, such as Mutual, have the ability to uplink from many different locations directly to the affiliates. Each SCPC signal can be uplinked from a different location. This results in two advantages: superior quality due to direct transmissions and, increased reliability, because the equipment normally required for backhauling is eliminated.

When all factors are considered, it is felt that analog SCPC/FM provides the best type of audio distribution available today. SCPC/FM provides a flexibility that will be needed as new programming and competion increases.

Because Mutual owns the downlinks, it is in a unique position to assure affiliates of the best service and quality possible. Mutual can provide an integrated downlink to the affiliates, at no cost. The affiliates are assured of a quality downlink system.

Although Mutual is using SCPC transmission at present, it is constantly on the lookout for new and improved methods in system design, such as different types of modulation, multiplexing, and improvements in companders. Mutual intends to provide the best possible service to its affiliates.

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SATELLITE DISTRIBUTED DIGITAL AUDIO

FOR NETWOPK RADIO

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INTRODUCTION

Scientific-Atlanta recently signed contracts with ABC, CBS, and NBC as the supplier for 3-meter digital earth terminals for their network radio stations. The resulting terminal is a time-division-multiplexed (TDM) digital system with a transmission rate of 8.78 Mb/s. This system has been tested and qualified with rooftop reception in downtown Manhattan. All digital audio testing was performed directly over the satellite in the presence of severe terrestrial interference. Excellent performance was recorded on all tests. It is forecast that by early 1984, over 3,000 of these 3-meter terminals will be installed throughout the nation, receiving high-quality program feeds from the major radio networks.

Scientific-Atlanta's 3-meter TDM digital audio terminal is designed to receive high-speed TDM digital data at 8.78 Mb/s, demodulate, decode, and then demultiplex the data into the desired audio and data channels. The terminals will support data rates equivalent to twenty 15-kHz audio channels (384 kb/s each) or an equivalent larger number of lower bandwidth channels. Each 384 kb/s channel can support either one 15-kHz program, two 7.5-kHz programs, twelve 32-kb/s auxiliary channels (voice cue or data) or one 7.5 kHz program channel and six 32-kb/s auxiliary channels. One 32-kb/s auxiliary channel slot is reserved for system synchronization.

Using this TDM terminal, all radio stations have access simultaneously to any of the twenty 15-kHz channels. This allows the station to receive their time-zone network feed, news services, data, voice cue, as well as other program material. This simultaneous access is obtained by adding plug-in cards to the mainframe equipment.

Digital transmission was selected instead of analog transmission for its data and channel use flexibility, the outstanding quality of the received

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program audio and the efficient usage of the satellite transponder. To obtain analog-transmitted high-quality audio into a 3-meter network requires the use of analog compression techniques. Unlike digital compression, a residue compression is retained in the expanded analog audio. Digital transmission provides expanded dynamic range (about 15 dB more than analog) and high signal-to-noise ratio of a very high level that has never been experienced in network reception.

Two types of analog radio transmission are presently used for network transmission - the single channel per carrier (SCPC) and the video subcarrier or diplex method. For quality audio reception into a 3-meter network, only the SCPC provides acceptable performance for a large number of audio channels.

Analog earth terminals are less expensive than digital, but their satellite transmission capacity is reduced. For high quality reception only about ten (10) channels of 15-kHz SCPC analog transmission can be used in the satellite 3-meter network reception. Sixteen is probably the maximum number which could be supported. TDM digital transmission, on the other hand, allows twenty high-quality 15-kHz channels in the satellite.

Other competing digital approaches such as the 5-carrier T1-rate transmission were usually more expensive for receiving full network service. For example, to simultaneously access digital audio or data channels in two T1-rate transmitted carriers require additional RF and demodulating equipment. To access data simultaneously in all five T1-rate carriers, the terminal cost will be prohibitive. Using the TDM approach allows access simultaneously to all audio and digital data on the full transponder. The TDM approach is also more immune to terrestrial interference than is the T1-rate approach.

Scientific-Atlanta's TDM digital earth terminal provides:

- efficient use of the satellite
- inexpensive full transponder service
- very high quality audio reception
- built-in expansion capability for future services
- one-way data service capability
- relative immunity to terrestrial interference

Scientific-Atlanta will manufacture and sell the terminals and will offer installation and 24-hour maintenance service.

EQUIPMENT DESCRIPTION

The 15-kHz program signal is sampled at 32 kilosamples per second and digitized to a 15-bit word. Digital commanding techniques are used to instantaneously compress the 15-bit word to an 11-bit word. A parity bit is added, resulting in a word length of 12 bits. The parity bit is used in error concealment encoding that allows bit error rate to degrade to 10^{-5} with "just perceptible" audio degradation in the 15-kHz audio channel unit. The noise performance of the terminal is enhanced by forward-error-correction (FEC) encoding. The combined data is bi-phase modulated (BPSK) at 70 MHz, then upconverted to the 6-GHz band for transmission to the satellite. Figure 1 is a block diagram of the 3-meter receive-only terminal.

The outside equipment for the basic terminal consists of Scientific-Atlanta's 3-meter antenna, a 120 Kelvin low-noise amplifier (LNA) and 200 feet of 1/2 inch foam-filled coaxial cable. This outside equipment configuration will be satisfactory for 96% of the radio stations located within the Continental USA. Larger antennas or lower temperature LNA's are recommended for radio stations in Florida, Southern Texas and some areas in the upper New England States.

The inside equipment is contained in two chassis that are mounted in the radio stations' equipment rack. The wideband receiver converts 4 GHz received transmission down to 70 MHz where it is bi-phase demodulated, the forward error correction (FEC) coding is decoded and the resulting 7.68 Mb/s data stream is supplied to the data processing unit (DPU) mainframe.

The 7.68 Mb/s data is demultiplexed in the DPU chassis where the full transponder service is available with plug-in cards. A summary specification of the terminal is provided in Table I.

The optional plug-in units are as follows:

- Dual 15-kHz channel unit -------- two independent 15-kHz channels on one card
- Dual 7.5-kHz channel unit ------- Two independent 7.5-kHz channels on one card
- Voice cue channel unit ---------- a 32 kb/s delta modulated (CVSD) 3-kHz voice unit on one card
- Data channel unit --------------- three-port addressable asynchronous for slow and medium rate terminal equipment

A three transponder select switch will also be offered as optional equipment. Scientific-Atlanta will also offer a full line of outside equipment including lower temperature LNA's, larger antennas, and a low sidelobe antenna for improved performance for small orbital satellite spacing now being considered.

RECEIVE-ONLY TERMINAL LINK PERFORMANCE

Performance Analysis - In this section, the RF performance of Scientific-Atlanta's 3-meter TDM digital earth terminal is reviewed. For all practical purposes, the effect of bit errors on the digital audio performance is eliminated completely whenever the bit error rate (BER) is equal to or better than 10^{-7}.

The next paragraph reviews the location of radio stations, illustrating that a majority of these stations are located on or above the 33-dBw power contour of the satellite. Link fade margin is addressed, followed by a tabulation of the link calculations. The last paragraph discusses the effects of bit errors on digital audio performance.

Satellite EIRP Contours vs. Terminal Location - Performance of the receive-only earth station depends on the earth station equipment configuration and the satellite power or EIRP contours. The majority of the end users for digital audio will be the existing radio stations, and fortunately most of these radio stations are located on the higher EIRP contours of the satellite. Table 2 shows a count of the top 300 Areas of Dominant Influence (ADI) for radio market as to their location in the various contours of a particular satellite.

TABLE 2. NUMBER OF TERMINALS LOCATED IN EIRP CONTOURS FOR A 600 TERMINAL NETWORK

EIRP (dBW):	36-35	35-34	34-33	33-32	32-31	TOTAL
No. of Terminals:	92	254	228	16	10	600

These estimates of terminal count versus satellite power contours are based on the EIRP contours of SATCOM F1.

Link Fade Margin - The margins required for a satellite link are substantially different than for a terrestrial microwave system. The margins required for a satellite link depend on system availability and must include allocations (margins) for:

a. Atmospheric absorption

b. Rain fades

c. Antenna pointing errors (including wind loading effects)

d. Transmitter power level variations.

One major advantage of a system which uses a saturating carrier in the transponder is that uplink fades can be neglected, since the transponder is operated beyond saturation by the expected amount of the uplink fade. Thus, this guarantees that during maximum uplink fades, the transponder does not come out of saturation. The downlink fade margins at 4 GHz required for a BER availability of 99.99% are shown in Table 3.

Interference - The interference into the receive-only digital audio terminal is a composite of interference from various sources. The interference model used in the calculations is shown in Table 4.

The top two C/I rations in Table 4 are typical values while the C/I ratios for adjacent satellite and terrestrial interference are obtained from analysis.

Terrestrial Interference - Scientific-Atlanta employs a notch filter designed as part of its bandpass-matched filter which reduces the effect of terrestrial interference while enhancing the desired signal detection.

The frequency response on an ideal matched filter for 9-Mb/s BPSK modulated signal obeys the $(\sin x/x)^2$ response with its first nulls at $\pm$ 9 MHz. This ideal filter is approximated by a filter with nulls at $\pm$ 10 MHz, which significantly reduces the effect of most terrestrial interference while enhancing the detection of the desired signal.

Scientific-Atlanta's 3-meter TDM digital terminal has been tested in severe terrestrial interference environment. Figures 2 and 3 are photos of frequency spectrum measured during the ABC-Radio test with a 3-meter rooftop antenna at 1926 Broadway, New York, N.Y. The top photo shows the wideband modulated BPSK signal between the two TD-2 terrestrial interfering signals (arrows). The interfering signal was about the same level as the unmodulated satellite carrier (Figure 3), or about -123 dBW. Excellent reception was obtained during the test in the presence of this interference.

The notched matched filter provides a protection of about 38 dB from two equal-level TD-2 levels of -116 dBW, which provides a terrestrial C/I ratio of 30 dB.

Link Calculations - The link calculations for Scientific-Atlanta's 3-meter digital audio terminal are illustrated in Table 5. Direct addition of the positive and negative numbers in the top part of the table establishes the ideal downlink E_b/N_o. Power adding the composite C/I ratio results in the effective E_b/N_o. Subtracting link implementation margin of 1.5 dB results in the realized E_b/N_o into the rate 7/8 threshold decoder. The last table entries are the margins above BER = 10^{-7} for the various satellite contours. In all cases, adequate margin is available for excellent digital audio reception.

If additional margin is required at some stations, it can be obtained by increasing the terminal G/T. Typical G/T increases obtained by modifying the outside equipment are listed in Table 6.

Interpretation of the effect of BER on Digital Audio Performance - The bit error rate (BER) is the average rate at which bit errors occur; it is the estimate of the bit error probability. The BER remains constant whether we consider the high-speed 7.68 Mb/s data stream or a single 15-kHz channel (384 kb/s) demultiplexed from this high-speed data stream. Scientific-Atlanta uses a single bit error concealment encoding to further enhance the performance of the digital audio terminal. The subjective performance of this terminal is illustrated in Figure 4 on a subjective impairment grade. Since the RF link is designed to operate at 10^{-7} error rate or better, the effects of bit errors will not degrade the performance.

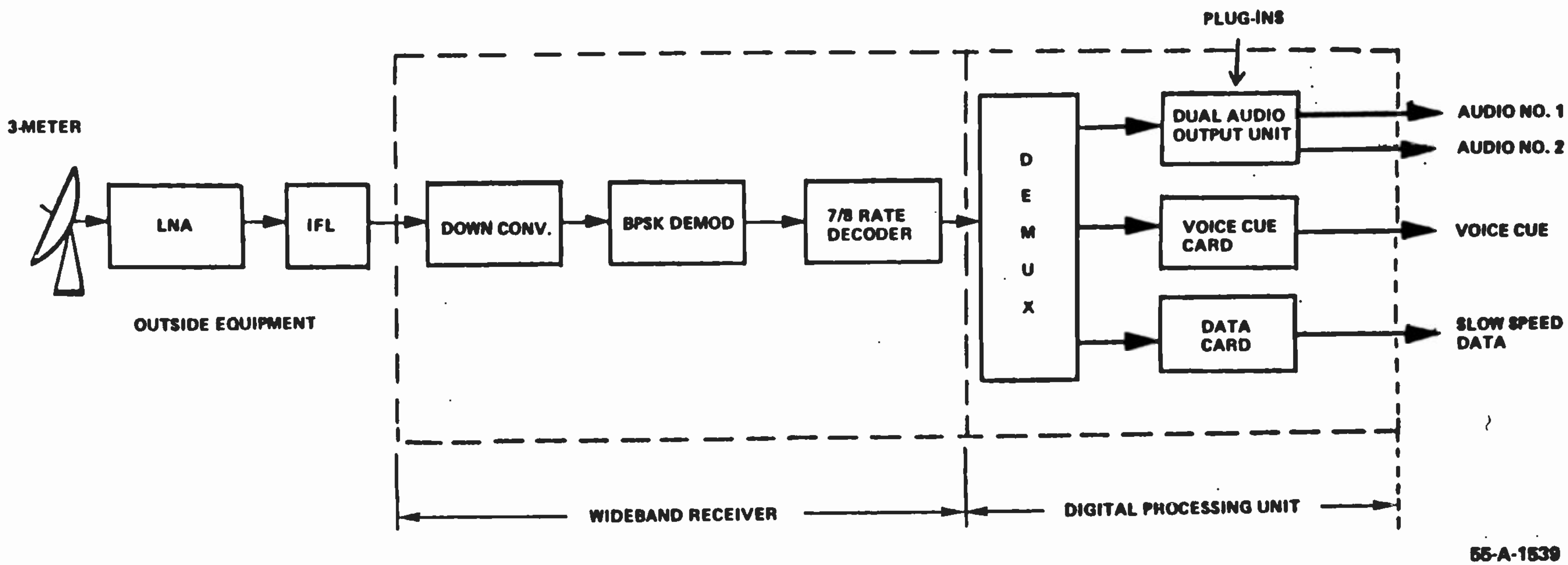


Figure 1. Three-Meter Digital Audio Terminal

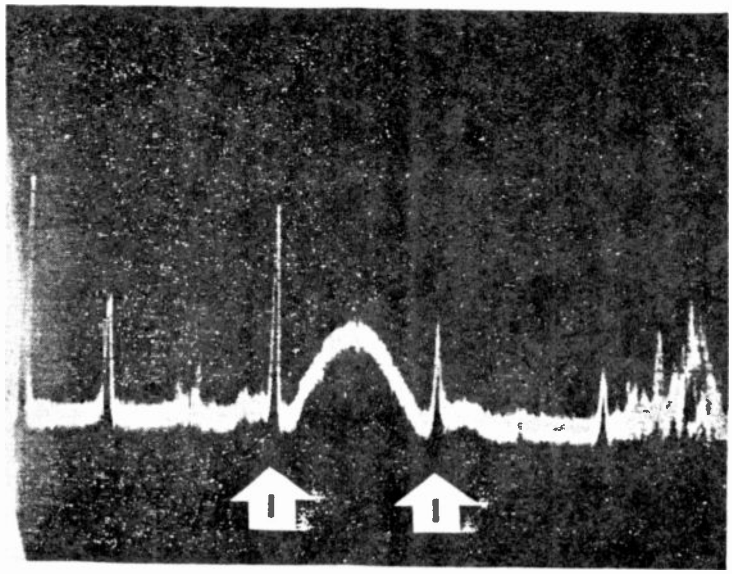


Figure 2. Wideband BPSK Signal Received by 3-Meter Rooftop Antenna at 1926 Broadway, New York, NY. Terrestrial Interfering Signal at ±10 MHz from Transponder Center

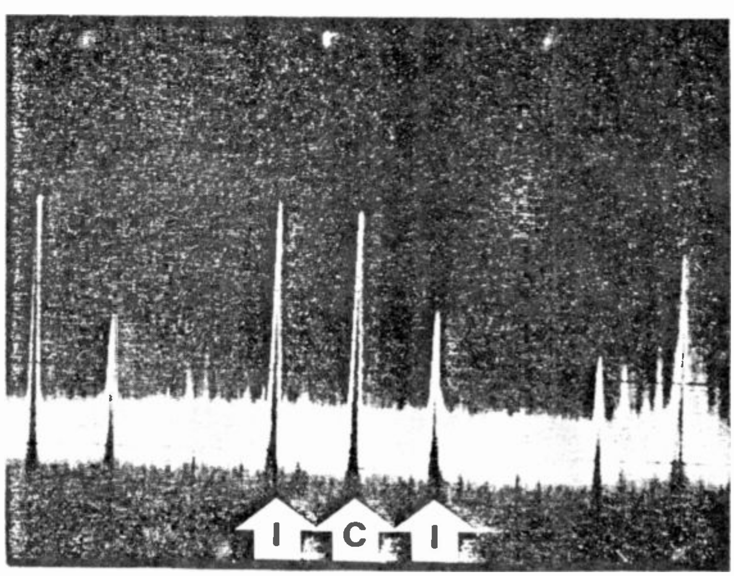
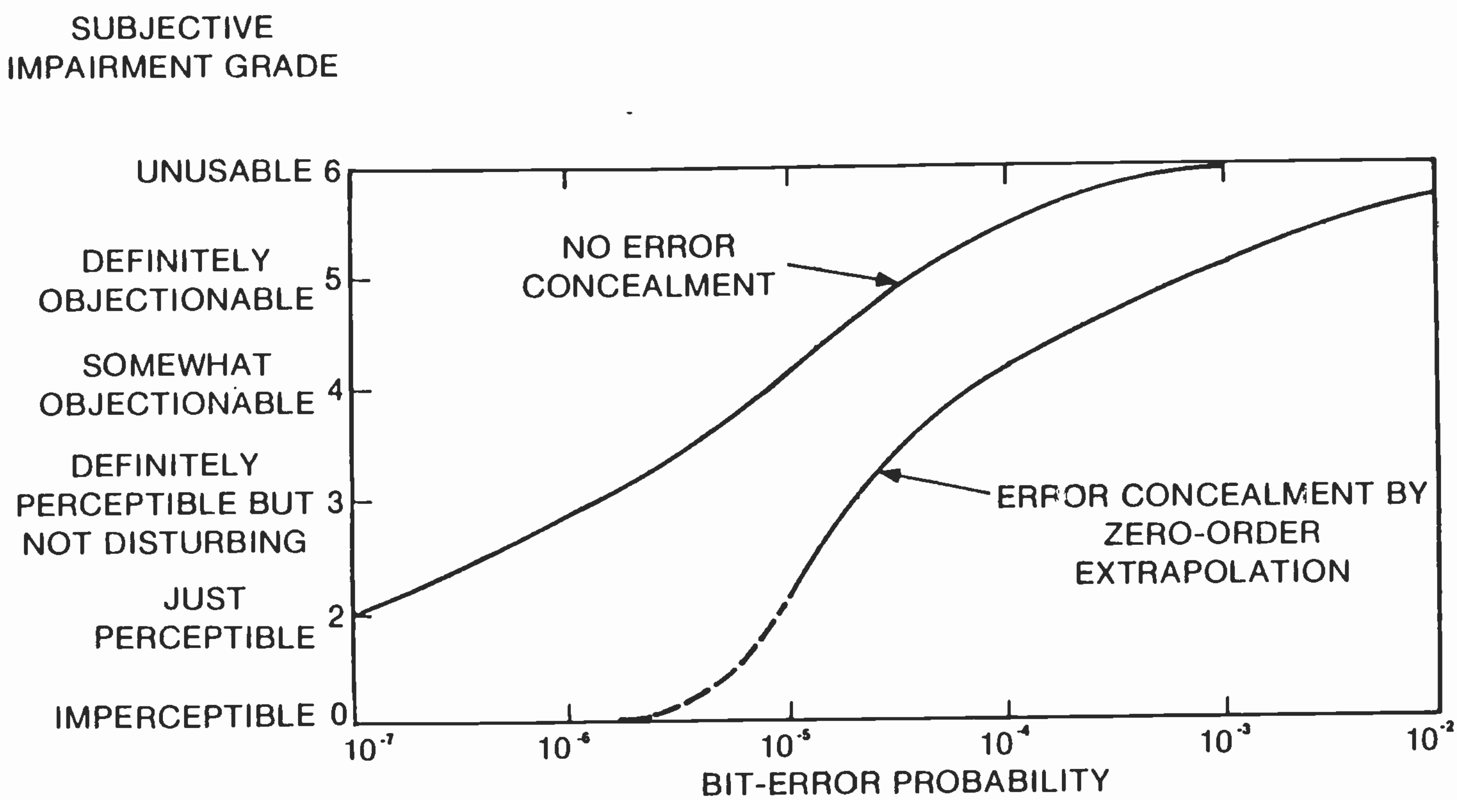


Figure 3. Center Spike is Unmodulated Carrier Received from Satellite into 3-Meter Rooftop Antenna

ERROR CONCEALMENT PERFORMANCE {1}



SUBJECTIVE EFFECT OF BIT ERRORS IN PRESENCE OF PROGRAM (SOLO TRUMPET)

{1} J. R. CHEW AND M.E.B. MOFFAT, "PULSE-CODE" MODULATION FOR HIGH-QUALITY SOUND-SIGNAL DISTRIBUTION: PROTECTION AGAINST DIGIT ERRORS," BRITISH BROADCASTING CORPORATION REPORT NO. 1972/18 UDC 534.86-626376.56

Figure 4

Table 1. 3-Meter Digital Audio Terminal Specification Summary

Item	Specifications
Antenna	
Diameter	3 meters
Gain	39.5 dB
Low-Noise Amplifier	
Noise Temperature	120K
Gain	50 dB
Wideband BPSK Receiver	
Downconversion	Dual
Received Data Rate	8.78 Mb/s
FEC Decoding	Rate 7/8 hard decision threshold decoding
Bit Error Rate	Better than 10^{-7} on or above 33 dBW satellite power contours
Digital Audio (15-kHz unit)	
Sample Frequency	32 kHz
Digital Compression	15 to 11
Sine wave Overload Level	+28 dBm0*
Peak Unaffected Signal	+24 dBm0
Signal-to-Quantizing Noise Ratio	≥56 dB (at +22 dBm)
Idle Channel Noise	≥81 dB (below +24 dBm)
Total Harmonic Distortion	0.3%

* 0 dBm0 = 0 dBm into 600 ohms

TABLE 3. LINK MARGIN FOR 99.99% BER AVAILABILITY

Parameter	Margin (dB)	Note
Atmospheric Absorption	0.1	a
Rain Fades	0.6	b
Antenna Pointing	0.1	c
Transmit Power Variation	0.0	d
	0.8 dB	

NOTES:

a. Atmospheric absorption: 0.1 dB.

b. Rain fades: This is a function of rain rate, earth station elevation angle, and vertical depth of rain cells. The baseline contour (32 dBW) for the RCA system (cutting through the Southern U.S.) is at higher elevation angles (resulting in lower fades) while the better EIRP contours are at lower elevation angles so that the increased rain fade is compensated by the higher available EIRP. For 99.99% of the time, the downlink rain fades (heavy rain over 60mm/hr) will be under 0.6 dB (note that a typical rain cell might extend 30 km or more in length, but only about 2 km in depth). A satellite link will experience a fade corresponding to the 2-km depth times the sin(x) of the elevation angle x, whereas a microwave system would experience a fade corresponding to the entire horizontal extent of the rain cell.

c. Antenna pointing errors: This is dependent on the antenna aperture and is well under 0.1 dB for 3-meter antennas.

d. Transmit power level variations: For saturated transponder, this is negligible since it only affects uplink losses.

TABLE 4. TDM DIGITAL INTERFERENCE MODEL

Source	C/I (dB)
Uplink	31
Cross Polarization	33
Adjacent Satellite	18.1
Terrestrial	30
Composite C/I*	17.5

*Composite C/I calculation: $10^{-1.75} = 10^{-3.0} + 10^{-3.3} + 10^{-1.81} + 10^{-3.0}$

Table 5 Three-Meter Network Link Calculation

Parameters	Values			Notes
Satellite Saturated EIRP (dBW)	32.0	33.0	34.0	
Multiple Carrier Loss (dBW)	0.0	0.0	0.0	
Space Loss (dB)	-196.8	-196.8	-196.8	
Boltzmann's Constant (dB)	+228.6	+228.6	+228.6	
Information Rate (7.68 Mb/s)	- 68.9	- 68.9	- 68.9	
3-Meter, 120° LNA G/T	+ 17.5	+ 17.5	+ 17.5	(1)
Ideal Downlink E_b/N_o (dB)	12.4	13.4	14.4	
Fade Margin (99.99%) (dB)	- 0.8	- 0.8	- 0.8	(2)
Downlink E_b/N_o (dB)	11.6	12.6	13.6	5)
Composite C/I (dB) (4.0° orbital spacing)	17.5	17.5	17.5	(3)
Effective E_b/N_o (dB)	10.6	11.4	12.1	
Link Implementation Loss (dB)	- 1.5	- 1.5	- 1.5	(4)
FEC Decoder E_b/N_o	9.1	9.9	10.6	
Decoder E_b/N_o for BER = 10^{-7}	8.1	8.1	8.1	(5)
Extra Margin above BER = 10^{-7}	1.0	1.8	2.5	

NOTES:

(1) The G/T varies with the elevation angle of the antenna: G/T = 17 dB/K at 5° elevation up to 17.9 dB/K at 40° elevation. An average value of 17.5 dB/K is used in the calculations.

(2) 99.99% fade margin discussed in link fade margin section above.

(3) Composite C/I discussed in interference section above. For 3° orbital spacing C/I reduces to 15 dB, which reduces the FEC decoder E_b/N_o by 0.6, 0.8, and 0.9 dB on the 32, 33, and 34 dBW contours, respectively.

(4) Typical measured performance over simulated satellite link using Scientific-Atlanta's BPSK modulator and demodulator is better than 1 dB from theory.

(5) The decoder is rate 7/8 1-bit hard decision threshold decoder. Value is measured value.

TABLE 6 . METHODS OF INCREASING G/T

Modification Antenna Dia.	LNA (K)	Typical G/T Increase (dB)
3.0 M	120	-0-
3.0 M	100	0.5
3.0 M	90	0.9
3.0 M	80	1.2
3.6 M	120	1.0
3.6 M	100	1.5
3.6 M	90	1.9
3.6 M	80	2.2
4.6 M	120	3.7

Local Satellite Radio Interconnects

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INTRODUCTION

By mid-decade, virtually every radio network affiliate will receive its feeds via satellite, either through a receive terminal on station premises, or over a local loop from a nearby receive terminal. While radio industry interest in satellite technology has been growing, most of that interest has revolved around the replacement of traditional network phone lines by satellite transmission, with consequent improvements in fidelity and reductions in cost.

This conversion from terrestrial to satellite distribution entails a substantial capital investment on the part of the radio industry, investment in transmission facilities, investment in receive terminals, and investment in personnel training. The second of these, investment in receive terminals, is by far the largest capital item in radio's transition into the satellite age.

In the discussion below, I shall argue that radio is about to spend too much money for many more receive terminals than the industry needs to achieve optimal satellite service. In fact, excessive capital expenditures associated with conversion to satellite may inhibit potential industry-wide benefits of satellite distribution.

The principal reasons for this conclusion are external to radio, and result from market forces in the satellite industry, specifically the radio industry's role as a consumer of satellite products, both space segment and equipment. A secondary reason can be found within the radio industry, in the changing relationship between networks and their affiliates.

OPTIMAL SATELLITE SERVICE

Improved fidelity and lower cost are cited as major reasons for network conversion to satelite distribution, and they are good reasons. However, reducing the operating costs of radio networks and giving affiliates better than a voice grade line are just the beginning. While new technologies are often first applied to old uses, it is often new uses that determine the ultimate significance of new technologies. Many observers believe that this will be the case in radio, that satellite distribution will evolve new applications that will provide radio much more than incremental improvements in technical quality and distribution cost.

With this in mind, it is important to think beyond, and to plan beyond, the mere substitution of one type of distribution network for another. The long-term interests of radio broadcasters will be served by a satellite system that can encourage and nurture experiments such as new programing forms or new combinations of audiences and advertisers. I would call such a system an optimal satellite system for radio, and suggest that it would have the following characteristics:

1) Universality: the ability to reach every radio station with uniform quality.

2) Flexibility: the ability to provide a variety of system configurations.

3) Simplicity: for both program supplier and affiliate station.

4) Economy: to exploit the traditional price-elasticity of telecommunications.

5) User Control: management by radio broadcasters.

This "optimal" system is clearly utopian, but one has only to look at the National Public Radio system, which is close to optimal within its own small universe, to see the potential benefits. For NPR affiliates, the satellite system has meant an unprecedented increase in programing alternatives, many of which come from independent producers, and many from other NPR affiliates themselves. In fact, with an optimal satellite system in commercial radio, the NPR experience suggests that several hundred commercial radio stations would become involved in program production and distribution.

OBSTACLES TO OPTIMIZATION

How utopian, how far from reality, is an optimal satellite system for the radio industry? The answer to this question lies in the economics of the satellite business itself, and in radio's role as a consumer. I have identified five obstacles to development of an optimal system. These problem areas will be the major determinants of the kind of satellite service the radio industry will have in this decade:

1) Capacity Shortage: Space segment demand will continue to exceed supply in the near term. This, coupled with regulatory questions that have yet to be settled, inhibits radio industry planning for its future space segment requirements.

2) Radio is a Minor User: All radio traffic on all satellites comprises less than two per cent of total available space segment. It is expected that by mid-decade, radio's share of space segment will actually decline as new capacity is brought on line. Because of radio's relative insignificance as a satellite customer, the industry's ability to force carriers to consider radio's priorities is negligible.

3) Incompatibility: Satellite programing intended for radio stations will be available from seven or more satellites within the next 18 months. Some of these satellites carry radio programing on more than one transponder. In addition to these different frequencies and different spots on the orbital arc, several different transmission methods are used for radio, two different digital methods (RCA ADDS and AT&T), single-channel-per-carrier, and multiplexed subcarriers. These incompatibilities join forces to virtually proscribe universal satellite service. The cost of a universal receive terminal capable of receiving all satellite radio transmissions is beyond the financial capabilities of all but the largest radio stations.

4) Supplier-Driven Market: While radio's space segment consumption is minor, the radio industry represents an important market to manufacturers of satellite receive hardware, simply because there are so many radio stations. Proliferation of satellites, transponders, and transmission modes only increases radio's attractiveness as an arena into which earth stations can be dumped. In the satellite hardware business, market share is the name of the game, and sup-

liers prefer to sell large lots to a few customers, like radio networks. As a result, individual broadcasters are being denied the opportunity to specify and negotiate the receive terminals that will best meet their individual needs.

5) Disproportionate Cost: Many radio broadcasters are astonished to learn that satellite receive terminals for radio can cost more than terminals for television. The usual cost advantages of audio vs. video are missing at this crucial junction. Ironically, satellite distribution is expected by many to improve radio's cost-effectiveness against television, yet the crucial link between satellite and station, an important capital expense, lacks this very cost-effectiveness.

HOW MANY TERMINALS?

The combination of factors summarized above leads to the question: how many satellite receive terminals does the radio industry need? And, equally important, how many will it wind up buying? I will consider six hypothetical cases:

1) If all radio traffic were on one satellite, and if all stations in a given market shared the use of one terminal, then the radio industry would need no more than 1,000 terminals to provide nearly universal service. (This case is not unlike a one-network plan, such as Mutual's.)

2) If all radio traffic were on one satellite, but each station were to have its own terminal, then the radio market would be 8,000 dishes.

3) Since all radio traffic is on seven satellites, if all stations in a given market shared the use of seven terminals, or a "universal" terminal, then radio's needs would be 7,000 dishes.

4) If this seven-satellite system were accessible to all stations with seven terminals per station, then the radio market would be 56,000 terminals.

5) If an intermediate position, such as a three-satellite system (RCA ADDS, Westar I and III), were considered and all stations in a given market used the same three terminals, then radio would need 3,000 dishes.

6) Finally, if each station had three terminals, the market total would be 18,000 dishes for radio.

The difference between 1,000 and 56,000 dishes is enormous. But setting aside the extreme cases, even the difference between 3,000 and 18,000 is huge (@ $10,000!), representing an industry-wide capital investment of over $100 million.

How many dishes does the radio industry need? To assure optimal service, it needs only enough to make certain that each satellite is accessible by each station. A "universal" terminal in 1,000 markets with all stations sharing would require 7,000 dishes. How many dishes will the radio industry get? Network plans indicate one dish per station per network, which yields a figure in the 6,000-8,000 range.

Consider however the physical distribution of the receive terminals in both of these cases. With roughly the same capital investment (7,000 dishes), the radio industry can either have one dish per station capable of receiving only one satellite, or a "universal" terminal in each market giving every station access to every transponder, every satellite, every transmission mode.

LOCAL INTERCONNECTS

The concept of local market radio broadcasters jointly owning and operating expensive technical facilities is well-known, particularly FM broadcasters sharing the cost of tall towers with candelabra. This concept can be extended to include satellite receive facilities.

A local satellite radio interconnect (LSRI) comprises these components:

1) A "universal" terminal capable of receiving all satellite transmissions intended for radio (or all transmissions local broadcasters wish to be able to access.)

2) Dedicated 15 kHz lines connecting the universal terminal to each participating station.

3) A switcher to switch satellite channels to station lines , using dial-up from the stations.

An LSRI uses off-the-shelf technology, and can be adapted to local market conditions and requirements. Typical capital costs in a medium-to-large market are on the order of $5,000 per station.

The benefits of LSRIs are consistent with the characterisitics of an optimal satellite system. Each LSRI member station has universal access to all radio programs on satellite. The LSRI is simple to operate, and sharing of capital costs yields the lowest possible per-station expenditure on the satellite age. The LSRI bypasses the notorious "last mile" problem in radio, and puts satellite planning and management under the control of radio broadcasters themselves.

Adapting AM Transmitters
For Stereo Transmission

David Hershberger

Harris Corporation

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INTRODUCTION

"Will my present AM transmitter work in stereo?" is a question being asked by many broadcasters now that the beginning of AM Stereo broadcasting seems near. This paper discusses the techniques for adapting AM transmitters for stereo, some of the problems which might arise, and their solutions.

All five proposed stereo systems modulate both the amplitude and phase of the carrier, whereas mono AM modulates only the amplitude of the carrier. With the phase modulated (PM) signal present, some areas of transmitter performance which were irrelevant for mono transmission will become crucial for stereo transmission.

Figure 1 shows the basic technique for converting an AM transmitter to stereo. The stereo generator takes the left and right channel audio signals and produces an audio frequency (AF) signal and a radio frequency (RF) signal which carries the PM part of the stereo signal. The AF signal is applied to the transmitter's audio input and the PM signal is applied to a low-level RF stage in place of the carrier signal generated by the crystal oscillator. The two signals are amplified separately and mixed in the modulated stage of the transmitter to produce the desired stereo signal which, in general, has very different characteristics than either of the two signals produced by the stereo generator. The complete stereo signal with its AM and PM components is then passed through the transmitter's output tuning network and the station's antenna system to the listener's receiver.

PERFORMANCE REQUIREMENTS FOR AM STEREO TRANSMITTERS

A good AM stereo transmitter must, of course, first be a good monophonic transmitter, but there are some additional requirements for making a good stereo transmitter. First, the AF and PM signals from the stereo generator must arrive at the modulated stage of the transmitter with the correct amplitude and phase for all modulating frequencies. This requires that:

1) The phase versus frequency characteristic of the transmitter's RF amplifier chain up to the PA input be linear; i.e., phase shift must be directly proportional to frequency over the bandwidth of the PM signal coming from the stereo generator. Another way to say this is that all frequency components in the PM signal must experience the same time delay.
2) The transmitter's AF amplifier chain must also have a linear phase characteristic (constant time delay for all frequencies) plus flat frequency response over the bandwidth of the AF signal coming from the stereo generator. The delay time through the two channels must be equal to ensure simultaneous arrival of the two signals at the modulated stage.
3) The transmitter's output network and the station's antenna system must be able to pass all the significant frequency components in the final stereo signal.

The five proposed stereo systems can be classified as additive systems or multiplicative systems, and there are some differences in the bandwidths required for the two types of systems. Figure 2 illustrates the differences by showing the spectral distributions of the two types of systems for a carrier frequency of 1000 kHz and 10 kHz modulation of either left or right channel.

The spectra for 10 kHz modulation of a monophonic transmitter is shown for comparison. In the mono case, the AF channel needs good response to 10 kHz; the RF channel passes only the unmodulated carrier. The output network and antenna system must have a 20 kHz bandwidth. The response need not be perfectly flat but should be symmetric; i.e., the attenuation and phase shift should be the same for the upper and lower sidebands.

Both types of stereo systems require 30 to 60 kHz bandwidth in the RF channel to pass the PM signal from the stereo generator. The additive system requires 20 to 30 kHz bandwidth in the AF channel but the harmonics in the AF spectrum cancel with the extra sidebands in the PM spectrum to leave only a single pair of sidebands in the output spectrum. Thus, the transmitter's output tuning network and the station's antenna system need only the same bandwidth as for monophonic broadcasting.

The multiplicative stereo systems need only 10 kHz bandwidth in the AF channel, but the extra sidebands in the PM spectrum are still present in the output spectrum. Thus, the output network and antenna system must have a 40 to 60 kHz bandwidth.

No transmitter will meet these requirements perfectly, so the question arises, how good must the transmitter and antenna system be to produce an acceptable stereo signal? The different stereo systems have different sensitivities to the various imperfections found in transmitters and antenna systems, so no one set of numbers can apply in all cases. Even so, some averages help to understand just what is required. To achieve 20 dB channel separation and 2% harmonic distortion for modulating frequencies of 30 Hz - 10 kHz, the following specifications might typically be required:

1) RF channel phase linearity: ± 2 deg. over a 40 kHz bandwidth: ± 4 deg. over a 60 kHz bandwidth.
2) AF channel frequency response and phase linearity:
   a) Additive system: ± .5 dB, ± 2 deg. 30 Hz to 10 kHz
      ± 2 dB, ± 10 deg. 30 Hz to 20 kHz
   b) Multiplicative system: ± .5 dB, ± 15 deg. 30 Hz to 10 kHz
3) Output network and antenna system:
   a) Additive system: Symmetric response ± .6 dB, ± 4 deg. over a 20 kHz bandwidth.
   b) Multiplicative system: flat response and linear phase:
      ± .5 dB, ± 2 deg. over a 40 kHz bandwidth
      ± 1 dB, ± 4 deg. over a 60 kHz bandwidth

SOME PROBLEMS WITH EXISTING TRANSMITTERS

A number of performance factors of existing AM transmitters may combine to degrade the quality of stereo transmissions:

<u>AF Channel</u>: The transmitter's RF amplifier chain may not have the required frequency response or phase linearity. Modulation transformers and coupling capacitors may cause high and low frequency rolloffs which will adversely affect response and phase linearity. If so, distortion and channel separation will be affected in the additive system and some of the multiplicative systems; separation only in other multiplicative systems.

RF Channel: High-Q interstage coupling networks and harmonic resonators may adversely affect the phase linearity of the transmitter's RF amplifier chain. Distortion and separation will be affected in all systems.

Time Delay Differences Between AF and RF Channels: In order for the stereo signal to be correctly generated in the PA stage, the RF and AF signals must arrive there at the same time. There are many contributing factors to time delays. In the RF chains, the number of stages and the "Q" of associated tuned circuits determine the delay. In the AF channel, time delay is determined by the modulation method, filters, and reactive components. If there is a time delay differential, separation and distortion will be affected in the additive systems and in some multiplicative systems. In other multiplicative systems, only separation is affected.

Output Network and Antenna System: For the additive system, deviations from symmetric response will affect channel separation but not distortion. For the multiplicative systems, deviations from the required bandwidth or phase linearity will affect separation and distortion.

Incidental Phase Modulation (IPM): IPM is undesired phase modulation of the carrier which results from the simultaneous amplitude modulation of the carrier. Two sources of IPM can be identified. The first is due to feedthrough of the PM signal from the RF driver stage to the transmitter's output. This results when the transmitter's neutralization is imperfect and affects primarily the multiplicative systems. Channel separation and distortion are affected.

The second type of IPM is due to nonlinear impedances in and around the transmitter's modulated stage. For example, as the final stage of the transmitter is amplitude modulated, it may present a varying load impedance to the driver stage and a varying source impedance to the output tuning network. These varying impedances will cause the phase shift through the tuning networks to vary with the amplitude modulation and the result is IPM. A similar effect can be seen in solid state transmitters where the storage times of the transistors vary with the amplitude modulation. This second type of IPM affects distortion and separation in all of the stereo systems.

Extraneous Phase Modulation (EPM): EPM appears as noise in the difference (L-R) channel of all stereo systems. The causes are several. PM hum may be present in RF amplifier stages which have inadequate power supply filtering. If the transmitter's original crystal oscillator is used to derive the PM signal, low frequency oscillator noise may appear in the L-R channel. PM "white" noise may also be present in low level RF stages of the transmitter.

SOLUTIONS FOR THE PROBLEMS

After learning of these potential problems, one might believe there is no way AM stereo can really be made to work, but fortunately there are some simple solutions to most of the problems which will arise in converting AM transmitters for stereo.

Frequency response and phase linearity of the transmitter's AF chain might in some cases be improved by a feedback system, but most transmitters will have too much phase shift in their audio sections to implement a stable feedback loop. In most cases, open loop type amplitude and phase correction will be incorporated in the stereo generator to correct the response of the transmitter's audio section. This is similar to the equalization system used in TV transmitters: The system is simply adjusted for best square wave response. Nearly all transmitters can be equalized for flat frequency response to 20 kHz for the additive system or 10 kHz for the multiplicative systems. The high frequency audio harmonics present in the additive system are quite low in amplitude and so will not cause an overload problem even in transmitters which rolloff sharply above 10 kHz. The characteristics of additive systems which achieve compatibility with mono radios also result in compatibility with modulators.

A stable feedback loop usually can be implemented around the transmitter's RF amplifier section. Such a feedback loop is self adjusting and will simultaneously improve phase linearity, IPM and EPM. The stereo generator will take an RF sample from the transmitter's modulated stage to be combined with the PM signal to implement the feedback loop. Figure 3 shows a complete AM transmitter conversion with AF correction and PM feedback. If necessary, phase linearity can be further improved by lowering the Q of one or more RF coupling networks. Time delay differential is most easily corrected through the use of an adjustable delay line in the stereo generator. To equalize delays, the delay line is placed in the channel which has the smaller amount of delay. IPM can be further improved by careful neutralization and tuning. EPM in the form of hum can be improved by additional power supply filtering. Oscillatory noise can be eliminated by including a suitable oscillator in the stereo generator.

Most transmitters' output tuning networks are broad enough to pass the stereo signal with no modifications required, but there may be a few cases where re-configuration of the output network will be necessary. For the additive system, the bandpass requirements for antenna coupling units (ACU's) and phasors are the same for stereo as for mono, so an antenna system which works well in mono will also work well in stereo. However, if there are some problems with the system, they will be magnified in stereo, and retuning for more nearly symmetric response may be required. For the multiplicative systems, phasors, and ACU's which cannot pass the 40 to 60 kHz bandwidth may need to be reconfigured for wider bandwidth.

MONO STATIONS

The mono broadcaster may occasionally need to improve the stereo parameters of his transmitter for the sake of those listening on stereo receivers. This may be necessary to:

1) Avoid stereo indicator "falsing"
2) To deliver an acceptable signal to any low cost receivers which do not employ automatic stereo/mono switching. It is expected that the vast majority of receivers will employ automatic switching. Therefore, the prime consideration for the mono broadcaster is to ensure that his transmitter will not trigger the stereo pilot tone detector circuits of the stereo receivers.

False stereo indications in the presence of modulation are most likely due to extreme amounts of IPM. False stereo indications independent of modulation are most likely due to PM hum or subsonic FM noise from the crystal oscillator. These undesired effects may be corrected by careful tuning procedures or factory-supplied transmitter modification kits.

STANDARDS

The FCC will be choosing an AM stereo system and setting standards for its performance. These actions will be the single most important factor in determining the difficulties and expense incurred by the broadcaster to convert his facility for stereo. The FCC has indicated that standards for frequency response, distortion, and noise will be similar to those now existing for AM broadcasting. However, one new standard, channel separation, will be seen in the stereo rules.

Experience has shown that about 10 to 15 dB channel separation is necessary to produce the complete stereo effect; that is, a left channel signal still appears to come from the left speaker even though the same signal is present in the right channel at a 15 dB lower level. In direct A-B listening tests it is virtually impossible to distinguish 15 dB separation from infinite separation.

Thus, it seems there would be little practical benefit to requiring more than 15 dB channel separation, but whichever stereo system is chosen, a tighter standard, such as 30 dB, might mean wholesale replacement of transmitters, ACU's and phasors for many stations having older equipment. If a tighter standard is desirable, it would seem desirable to phase it in gradually by requiring better performance of new type-accepted transmitters and new antenna system installations.

RESPONSIBILITY FOR TRANSMITTER CONVERSIONS

Harris Corporation hopes to manufacture an AM stereo generator package that can be added to most manufacturers' AM transmitters. The combination would then be type accepted as a composite system.

It will be the responsibility of all transmitter manufacturers to determine the adaptability of their transmitters to stereo broadcasting and supply modification kits if required to make them adaptable. Harris' approach to stereo generators and transmitter modifications will be to include transmitter correction circuits in the stereo generator wherever possible, e.g., phase correction and feedback circuits. This is a minimum cost approach because the corrections can be made at small signal levels rather than making modifications to the high power stages of transmitters.

The consulting engineers will have to determine the acceptability of phasors, ACU's, etc., for stereo transmission.

CONCLUSION

There has been a great deal of speculation within the broadcasting industry about the relative ease or difficulty of converting AM installations for stereo broadcasting. Some say there is nothing to do but add a stereo generator and start broadcasting. Others say there is no way to pass stereo signals through directional arrays or existing transmitters. Neither view is correct. Most AM transmitters can be successfully converted to stereo by adding a stereo generator and making some simple modifications. In some cases, particularly where the transmitting equipment is old or in poor condition, the conversion may be difficult or impossible.

The difficulty and cost of the conversion will depend on the type and condition of the transmitter and antenna system, which of the five competing systems is adopted and what performance standards the FCC requires.

The conversion process may sound like a big headache for the station's chief engineer, but all the details of transmitter conversion will be worked out by the transmitter manufacturers, and the consulting engineers will work out the antenna system problems. This should make for a smooth conversion process for the broadcaster.

ADAPTING AM TRANSMITTERS FOR STEREO TRANSMISSION

PART II

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Part I of this series, presented at the 1979 convention of the National Association of Broadcasters, discussed the basic concepts of AM stereo as they relate to generation of AM stereo signals by conventional transmitters. Part I discussed transmitter requirements, correction methods which can significantly improve transmitter performance for AM stereo, and how those requirements vary for different kinds of AM stereo signals. Part I also presented some average performance specifications required to ensure a minimal level of AM stereo performance.

For an AM transmitter to be a good stereo transmitter it must first be a good monaural transmitter. An AM transmitter should have low distortion, good transient response, and wide frequency response both in the modulator and in the output tuning and matching network.

For AM stereo, however, there are further requirements which have little to do with AM monaural. These parameters are generally:

1. Phase linearity and flat response in the modulator
2. Phase linearity and flat response in the RF driver stages
3. Low L+R to L-R crosstalk, which manifests itself as one or both of:
    a. incidental phase modulation (IPM) or
    b. incidental quadrature modulation (IQM).

Part I of this series discussed general techniques for improving these stereo parameters; in Part II specific methods will be presented.

AUDIO (MODULATOR) CORRECTION METHODS

AM transmitters envelope (audio) performance will be important for good AM stereo performance. Fortunately, most AM transmitters have reasonably good envelope channel performance in order to achieve good mono performance. However, some subtleties distinguish the requirements for good AM stereo. The most frequently occurring parameters requiring correction are:

1. High frequency rolloff
2. Low frequency rolloff
3. Phase linearity.

High frequency rolloff is obviously important. Good separation at the higher modulating frequencies requires that L+R amplitude be matched to L-R amplitude. If the L+R signal, essentially contained in the modulator signal, is attenuated, then separation will be reduced. High frequency rolloff is corrected in a simple way: equalization, or a boost complementary to the attenuation characteristic. The rolloff is usually gentle, which means that the equalizer circuit is simple and easy to use.

What is not so obvious is how low frequency rolloff can affect stereo performance. In addition to affecting low frequency separation, a low frequency rolloff can also affect intermodulation distortion characteristics when there is a high frequency tone in addition to a low frequency tone, as in the standard SMPTE test. In most AM stereo systems, the phase modulation component is correlated with the envelope modulation. If the low frequency component is phase shifted and/or rolled off by the modulator, then the phase modulation signal will not match the modulator signal. The result will be distortion and a loss of separation.

As in the high frequency case, the remedy is a low frequency equalizer. The low frequency rolloff of most transmitters is gentle, so again a simple equalizer will be all that is required.

Phase linearity is required for the same reason that low frequency response flatness is required: all frequency components of the envelope signal must arrive in synchronism with the phase modulation component, in order that the AM stereo signal is assembled correctly in the modulated power amplifier. Generally, correcting low frequency response problems will also correct low frequency group delay problems. But equalizing high frequency response does not generally simultaneously correct high frequency group delay variations. For this, a separate high frequency group delay equalizer can be provided. Again, the delay equalizer can be a simple low-order device. (The delay equalizer should not be confused with the AF/RF delay line, which has constant delay versus frequency and whose purpose is to eliminate the differential delay between modulator and RF channel. The delay equalizer corrects delay variations versus frequency in the modulator.)

Figure 1 shows the suggested form for the transmitter modulator correction circuitry, including low frequency equalizer, high frequency equalizer, and group delay equalizer.

RF CHANNEL CORRECTION METHODS

One of the more critical characteristics of AM transmitters is L+R to L-R crosstalk. Crosstalk is important in AM stereo because it affects stereo separation. Unfortunately, some of the literature incorrectly refers to crosstalk as incidental phase modulation, or IPM. For some systems, incidental phase modulation is exactly the same thing as L+R to L-R crosstalk. But for other systems, IPM may have little to do with L+R to L-R crosstalk, and a more appropriate parameter should be considered. Systems which transmit L-R by some kind of phase modulation have IPM as the most pertinent parameter. For systems using quadrature modulation or IQM is most relevant.

Incidental phase modulation may be expressed as:

$$v(t) = [1 + A(t)]\cos[wt + p(t)]$$

where

$v(t)$ = AM signal with IPM

t = time

$A(t)$ = audio modulating signal

w = carrier (radian) frequency

$p(t)$ = IPM signal (related to A)

while incidental quadrature modulation is given by:

$$s(t) = [1 + A(t)]\cos(wt) + q(t)\sin(wt)$$

where

$s(t)$ = AM signal with IQM

$q(t)$ = IQM component (related to a)

IPM is expressed as exponential, or angle modulation, while IQM is expressed as a linear, additive term. Again, the particular AM stereo system's L-R modulation method determines the most appropriate measurement parameter.

In many cases, measurement of IPM and IQM can give drastically different results. This is because IPM measurement requires a hard limiter while IQM measurement does not. Figures 2 and 3 show block diagrams for IPM and IQM measurement.

The limiter is a device which maintains a constant output level regardless of input level. Equivalently, its gain is inversely proportional to the envelope modulation. When the envelope is unmodulated, the limiter gain may be unity. When the envelope is positively modulated, the limiter gain decreases. When the envelope is negatively modulated, the limiter gain increases without bound. As the envelope approaches -100% modulation, gain approaches infinity. This characteristic of a hard limiter introduces a weighting factor. As envelope modulation -100%, phase modulation effects are amplified with increasing gain. Near -100% modulation, small nonsymmetrical sideband components or drive feedthrough components are grossly exaggerated by the hard limiter.

In contrast, the quadrature modulation detector shown does not have varying sensitivity as a function of envelope modulation. Because no limiter is present, its gain is constant.

IPM and IQM can result from RF drive feedthrough in the RF power amplifier. In an improperly neutralized modulated stage, an unmodulated drive component can appear as part of the output signal. For example, consider the case where a transmitter is modulated 100% but where a 2% drive feedthrough component appears (resulting in negative envelope modulation of 98%). Measurement of IPM and IQM will give drastically different results for this commonly occurring situation. Figure 4 shows calculated IPM for this case. The IPM characteristic shows large amplitude peaks and a distorted signal. Figure 5 shows the incidental quadrature component of the same signal. Notice that in this case the signal is small and undistorted.

The unmodulated feedthrough component merely shifts the static carrier phase, and does not add additional sideband components to the signal. The shifted carrier will move some of the AM sidebands to the quadrature ("Q") channel, which results in linear IQM crosstalk from L+R to L-R. The crosstalk signal is an undistorted version of the L+R (AM) signal.

The IPM signal is a rather "spikey" looking function, due to the fact that the hard limiter weights the negative modulation peaks more heavily. PM sidebands exist, but they are produced by the limiter and do not exist in the AM signal.

L+R to L-R crosstalk may be classified as follows:

1. Linear
    a. in-phase
    b. phase shifted
2. Nonlinear.

Linear crosstalk from L+R to L-R means that the signal appearing in L-R is an undistorted version of the L+R signal. This signal may be in-phase or phase shifted with respect to L+R. Nonlinear crosstalk produces in the L-R channel a distorted version of the L+R signal.

Different means of correcting these anomalies have been developed. A highly effective method for correcting linear crosstalk is open-loop correction. Because separation is the parameter affected by crosstalk, Harris has developed a "separation corrector" which, by improving stereo separation, reduces linear crosstalk.

The separation corrector produces small samples of left and right, and adds them to L-R with the proper sign and amplitude to cancel the crosstalk. Moreover, the separation corrector shifts the phases of these samples as required for optimum crosstalk cancellation. Cancellation of the crosstalk, whether it is IPM or IQM, results in improved stereo separation.

Figure 6 shows a simplified block diagram of the separation corrector. Not shown in the diagram are some frequency-dependent adjustments which compensate for narrow bandwidth RF circuits.

The separation corrector not only corrects for transmitter deficiencies, but it can also be used to moderate antenna system bandwidth problems, including sideband imbalance. The adverse effects of narrow antenna systems can be reduced by this simple audio circuit, which can result in significant savings to the radio station. In some cases, use of this audio separation corrector circuit will make costly antenna, phasor, and ACU improvements unnecessary.

However, the separation corrector can only correct for linear crosstalk terms. When the crosstalk is nonlinear, PM or QM (quadrature modulation) feedback is suggested. Although PM/QM feedback can also reduce linear crosstalk terms, it is best suited for reduction of nonlinear crosstalk components. The incidental phase or quadrature modulation is sensed, and fed back to the AM stereo exciter for closed loop control. The detection method (IPM or IQM) and hence the controlled parameter, will depend on the particular AM stereo system in use.

After optimizing transmitter tuning, audio level, etc., for best uncorrected performance, the setup procedure is as follows: first, apply PM or QM feedback. Then, the bandwidth and gain of the PM/QM feedback system is adjusted to best match the RF bandwidth of the transmitter in use. After optimizing closed loop performance, any remaining linear crosstalk terms can be further reduced through the use of the stereo separation corrector. Adjustments of the separation corrector are made at high and low modulating frequencies for both right and left channels. This complete adjustment of the RF correction circuits.

CONCLUSION

Recent developments in audio and RF transmitter correction techniques will allow improved AM stereo performance with existing transmitters. Moreover, these techniques will permit acceptable performance to be obtained from the few transmitters which would otherwise be marginally unacceptable. The techniques and hardware have been refined, and are easily used by broadcast technical personnel.

SOME TRANSMITTER PERFORMANCE CRITERIA

FOR AM STEREO

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AM stereo may become a widespread reality in the near future pending some important decisions by the FCC. When and if AM stereo does become a reality, every AM station chief engineer will be faced with the question, "Is my transmitter ready for stereo?"

Of the five systems under consideration for adoption by the FCC, all have a common criteria by which existing and new design transmitters must be judged, which heretofore has not been a primary concern for AM broadcasters. That criteria involves the term, "Incidental Phase Modulation" (IPM). IPM is defined as "Phase modulation produced concomitant with desired amplitude variations." The IEC has labeled the same criteria "Synchronous Frequency Modulation." IPM is important because all five systems of AM stereo under study by the FCC employ intentional phase modulation of the carrier to produce the L-R matrix signal.

To understand the basics of AM vs. PM modulation and the important effects of IPM on AM stereo, it is in order to briefly review simple modulation theory. The basic electrical signal shown in Figure 1 is mathematically described in Equation (1). Two elements of Equation (1) can be caused to vary in a defined manner to cause modulation of the signal: (1) Amplitude of A, producing Amplitude Modulation, and (2) Angle, producing either frequency or phase modulation depending upon whether the ωt or phase term is varied, or (3) a simultaneous variation of amplitude and an angle component.

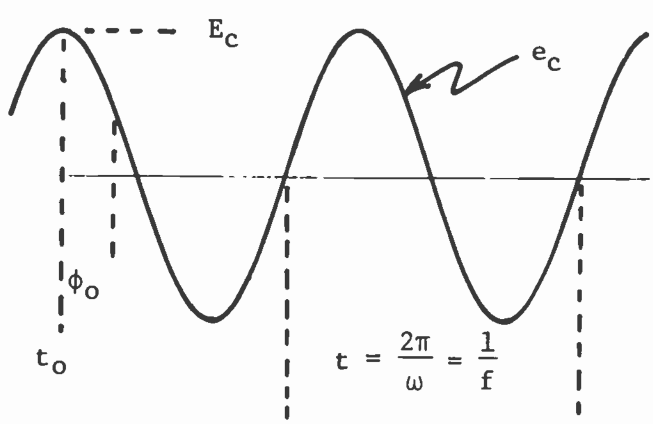


FIGURE 1

BASIC ELECTRIC SIGNAL

$$e_c = \underbrace{A}_{\text{Amplitude}} * [\text{Cos}(\underbrace{\omega_c t + \phi}_{\text{Angle}})] \qquad \text{EQ (1)}$$

If the modulating signal is assumed to be a similar signal but of much lower frequency, the expression for it can be as shown in Equation (2).

$$e_m = E_m \text{ Cos } \omega_m t \qquad \text{EQ (2)}$$

Since it is the amplitude A that is to vary at the modulating sinusoidal rate, we have the expression:

$$A = E_c + E_m \operatorname{Cos} \omega_m t \qquad \text{EQ (3)}$$

therefore, the complete amplitude modulated wave is expressed as:

$$e_c = E_c(1 + \frac{E_m}{E_c} \operatorname{Cos} \omega_m t) \operatorname{Cos} (\omega_c t + \phi_o) \qquad \text{EQ (4)}$$

where:

E_c = Peak carrier frequency voltage

E_m = Peak modulating frequency voltage

ω_c = Angular carrier frequency

ω_m = Angular modulating frequency

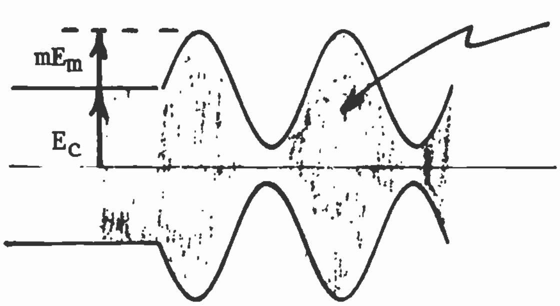
Equation (4) can be rewritten using trigonometric identities as:

$$e_c = E_c \operatorname{Cos}(\omega_c t) + \frac{mE_c}{2} \operatorname{Cos}(\omega_c + \omega_m)t + \frac{mE_c}{2} \operatorname{Cos} (\omega_c - \omega_m)t \qquad \text{EQ (5)}$$

where:

$$m = \frac{E_m}{E_c} = \text{"Modulation Factor"}$$

Now Equation (5) can be represented graphically in two ways, (1) in the time domain and (2) in the frequency domain. The well known time domain and frequency domain representation of Equation (5) is shown in Figure 2(a) and 2(b), respectively, for approximately 70% modulation depth.



Carrier frequency and phase inside envelope is constant at ω_c and independent of E_m.

FIGURE 2(a)

TIME DOMAIN REPRESENTATION OF AN AM SIGNAL

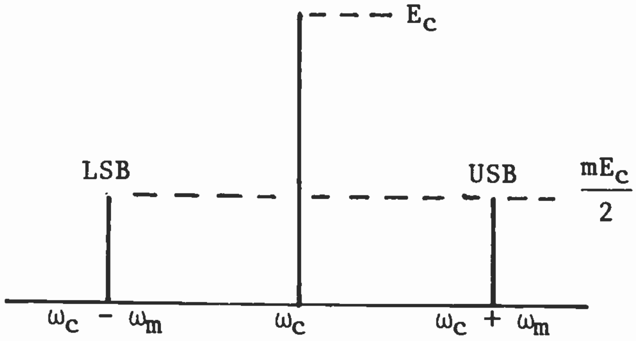


FIGURE 2(b)

FREQUENCY DOMAIN REPRESENTATION OF AM SIGNAL

Similarly, a relationship can be developed for an angular modulation signal by applying a modulating signal to the expression for either angular component of Equation (1). For phase modulation let:

$$\phi_m = m_p \mathrm{Cos}(\omega_m t) \qquad \text{EQ (6)}$$

where:

ω_c = Carrier angular frequency

m_p = Peak amplitude of modulation signal = peak phase deviation

Therefore, the complete expression for a phase modulated signal is expressed in Equation (7):

$$e_c = E_c \mathrm{Cos}(\omega_c t + m_p \mathrm{Cos}\, \omega_m t) \qquad \text{EQ (7)}$$

Equation (7) is evaluated by means of infinite series using Bessel functions to yield the frequency domain Equation (8).

$$\begin{aligned} e_c = E_c \{ & J_0 (m_p) \mathrm{Cos}\, \omega_c t \\ - & J_1 (m_p) [\mathrm{Cos}(\omega_c - \omega_m)t - \mathrm{Cos}(\omega_c + \omega_m)t] \\ + & J_2 (m_p) [\mathrm{Cos}(\omega_c - 2\omega_m)t + \mathrm{Cos}(\omega_c + 2\omega_m)t] \\ - & J_3 (m_p) [\mathrm{Cos}(\omega_c - 3\omega_m)t - \mathrm{Cos}(\omega_c + 3\omega_m)t] \\ + & J_4 (m_p) [\mathrm{Cos}(\omega_c - 4\omega_m)t + \mathrm{Cos}(\omega_c + 4\omega_m)t] \\ - & \text{- - - - - - - -}, \text{ etc.} \} \end{aligned} \qquad \text{EQ (8)}$$

where:

J_n = nth degree Bessel Function

The graphic representation of Equations (7) and (8) are shown in Figures 3(a) and (b), respectively.

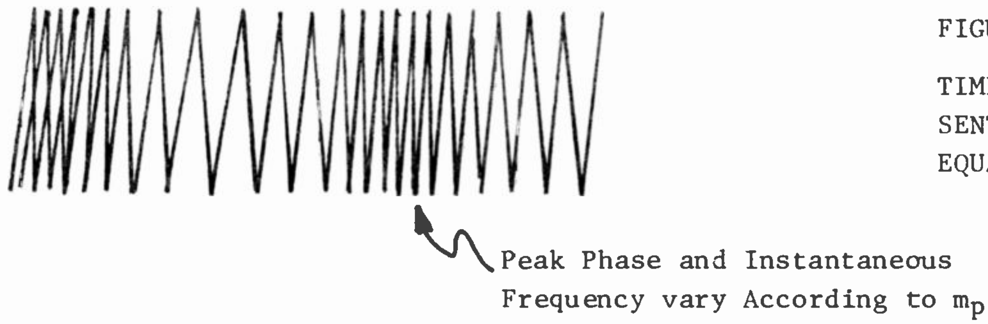


FIGURE 3(a)

TIME DOMAIN REPRESENTATION OF EQUATION (8)

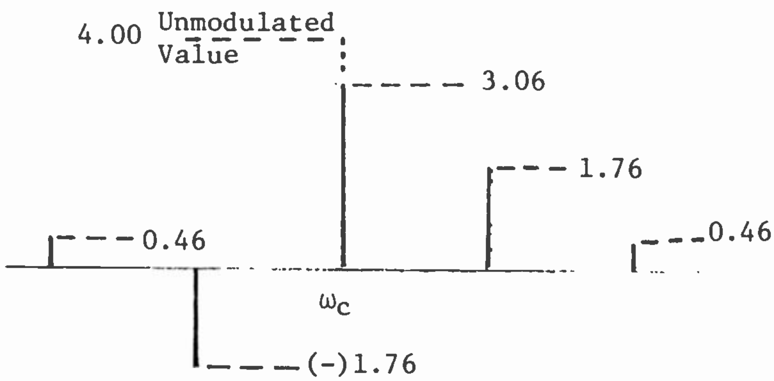


FIGURE 3(b)

FREQUENCY DOMAIN REPRESENTATION OF EQUATION (8)
(m_p = 1 RADIAN)

Thus far, the theoretical discussion of modulation systems has been based on the assumption that the system is linear, i.e., distortionless with regard to amplitude and/or phase linearity and has been presented preparatory to the topical subject of AM stereo. It is known from empirical observation that further mathematical development of simultaneous AM and PM will lead to erroneous conclusions due to the fact that for theoretical development, system linearity shall be assumed whereas in the real world system linearity is rare.

This writer believes it sufficient at this point to proceed with the observation that simultaneous AM and PM will produce frequency domain representations that are unlike either Figure 2(b) or Figure 3(b) but, rather, will be some combination of the two.

The interest in simultaneous AM and PM is due to the fact that some proposed systems of AM stereo utilize the separate and distinct types of amplitude and angular modulation for the (L-R) and (L+R) channels. Therefore, any amplitude modulation process which produces incidental phase modulation will have a deleterious effect on desired AM stereo characteristics.

Since it is believed that the AM stereo system to finally emerge will use concurrent phase modulation by some process, it is important for the prospective AM stereo broadcast engineer to know (1) if his equipment produces unwanted phase modulation concomitant with the desired AM, and (2) how to improve this incidental phase modulation characteristic.

QUALITATIVE ANALYSIS OF IPM

The first test that should be done to determine if a particular transmitter is afflicted with IPM is to make a spectrum analysis. This test will not yield how much IPM exists but only if it exists, and only to some degree how bad it is.

Figure 4 shows an RF spectral display of a hypothetical transmitter measuring 2.5% total harmonic distortion at 70% modulation level. The RF spectrum confirms the 2.5% THD figure and the balance of LSB and USB individual components indicates there is little (essentially none) IPM present. Now contrast the hypothetical transmitter shown in Figure 4 to that of an actual spectral analysis of a typical AM signal shown in Figure 5. The THD of the transmitter output of Figure 5 was measured at 1% by the standard method. The RF spectrum shown in Figure 5 also supports the 1% THD measurement figure, but a small amount of IPM is indicated by the unbalance of the LSB and USB components.

Figure 6 shows the RF spectrum of an AM signal which also measures 1% THD but which shows higher IPM than that of Figure 5. Figures 7 and 8 show examples of even larger amounts of IPM.

Therefore, a method of qualitative analysis exists through proper evaluation of an AM transmitter output spectrum. Once it is known that IPM exists in a specific transmitter, the next step is to quantify the measurement; i.e., determine exactly how much exists.

QUANTITATIVE ANALYSIS OF IPM

The simplest test to perform to quantify IPM is the Lissajous Pattern Test. Shown in Figure 11 is a block diagram of the Lissajous Pattern Test. Sinusoidal RF samples are taken of the exciter and transmitter outputs and applied to the X-Y inputs of a wideband oscilloscope. A phase shifter is inserted in either of the sampled signals and adjusted for a straight line representation on the scope. This straight line represents carrier zero reference phase. The transmitter is then modulated with a test tone and the peak trough and crest envelope phase deviations are displayed as Lissajous figures. The peak phase deviations are then calculated using Equation (9).

$$\phi_s = \mathrm{Sin}^{-1}(a/b) \qquad \text{EQ (9)}$$

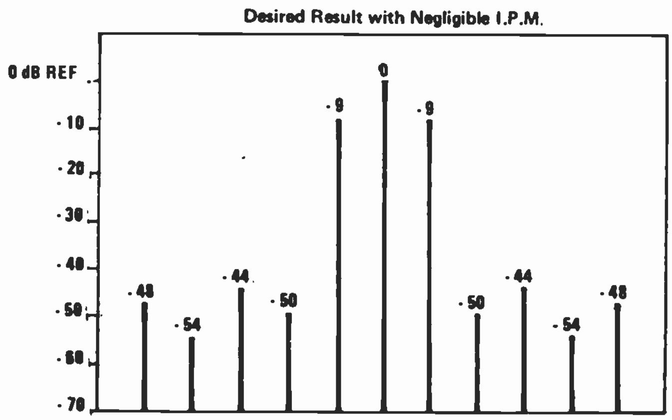


FIGURE 4. RF SPECTRUM OF HYPOTHETICAL TRANSMITTER HAVING PRACTICALLY NO INCIDENTAL PHASE MODULATION AND 2.5% TOTAL HARMONIC DISTORTION.

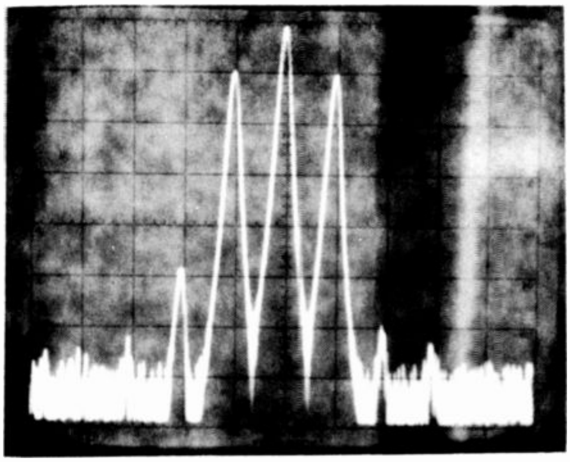


FIGURE 5. RF SPECTRUM OF ACTUAL AM SIGNAL INDICATING PRESENCE OF SMALL AMOUNT OF INCIDENTAL PHASE MODULATION. MEASURED AM DISTORTION OF 1%.

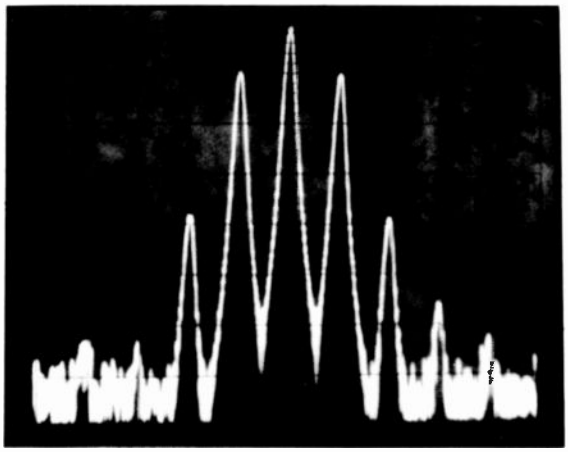


FIGURE 6. RF SPECTRUM OF AM SIGNAL INDICATING SIGNIFICANT I.P.M. NOTE: 2nd ORDER SIDEBANDS ONLY 30db BELOW 1st ORDER SIDEBANDS YET AM DISTORTIONS MEASURES 1%.

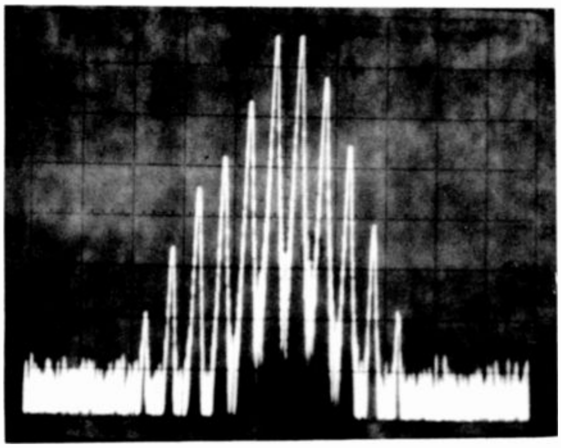


FIGURE 7. RF SPECTRUM OF AM SIGNAL INDICATING PRESENCE OF APPROX. 20° OF PEAK PHASE MODULATION. NOTE: THE 1dB REDUCTION IN APPARANT CARRIER AMPLITUDE AND 2nd ORDER SIDEBANDS ONLY 20dB BELOW DESIRED SIDEBANDS YET AM DISTORTION MEASURES 1%.

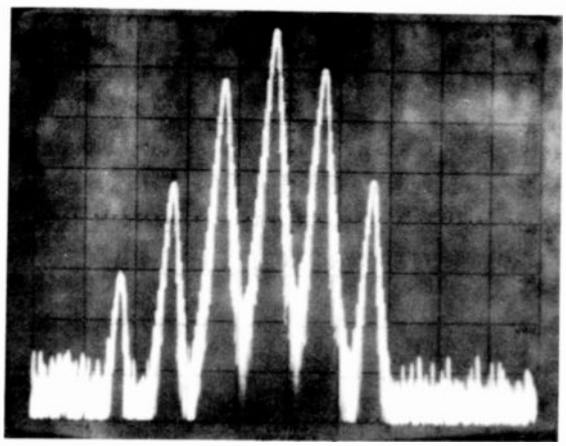


FIGURE 8. RF SPECTRUM OF AM SIGNAL INDICATING VERY EXCESSIVE I.P.M. OF APPROXIMATELY 1 RADIAN. NOTE: APPROX. 3dB REDUCTION IN APPARANT CARRIER AMPLITUDE (CARRIER SIGNAL IS HIGH SIGNAL ON THE LEFT) AND VERY HIGH ORDER AND UNBALANCED SIDEBAND PRODUCTS. YET AM DISTORTION MEASURES ONLY 1.5%.

where:

a = Ellipse width at zero axis

b = Peak ellipse excursion

Examples of this method for typical AM transmitters are given in Figures 12, 13, and 14.

The accuracy of the Lissajous pattern method is sufficiently accurate for peak phase deviations of one (1) degree or more. A method of measuring IPM below 1-degree values is shown in Figure 15. The same two RF samples as in the Lissajous Pattern Method are used and are fed at the proper level to the "RF" and "LO" inputs of a double balanced mixer. The "IF" output of the balanced mixer, once calibrated, responds to peak phase deviations which are measured on the oscilloscope. Other methods of measuring IPM also exist. One notable technique which is quite simple requires only one RF sample of the transmitter output which is connected to an HP-8901A Modulation Analyzer. This analyzer, and perhaps other similar analyzers, can measure IPM directly with better than 1 degree accuracy when operated and interpreted correctly.

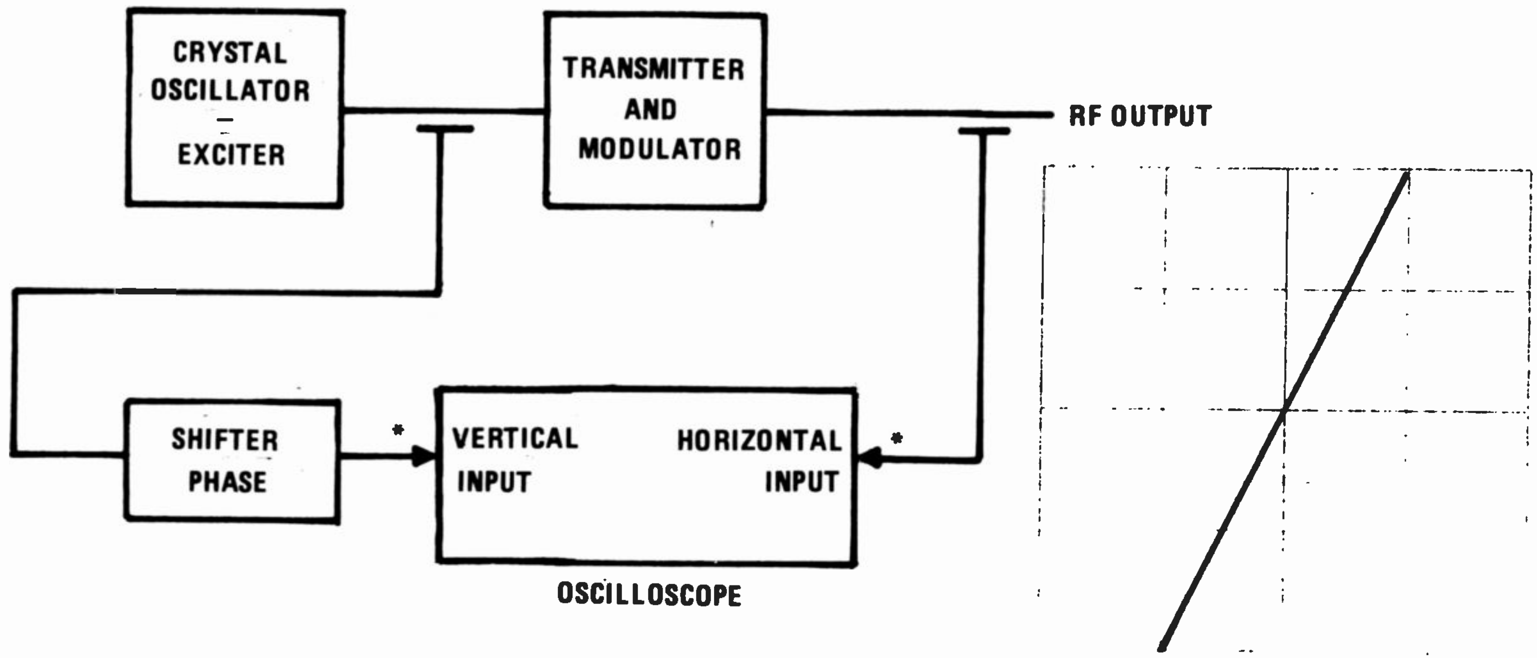
CORRECTING IPM

Once IPM has been measured, the next step is to make necessary transmitter modifications to reduce it to an acceptable level for good AM stereo performance. The major stereo performance indicator to be affected by IPM is stereo separation. The degree of IPM reduction required depends upon the type of system ultimately chosen for AM stereo and the specifications the FCC may impose regarding stereo separation.

It will be found, in general, that the older the transmitter, the worse it will be regarding IPM and also the more difficult it will be to correct.

Some transmitters designed since AM stereo became prominent have been designed to either minimize IPM or to make it easier to modify than some older transmitters.

It will be discovered that many older transmitters have as much as 20 degrees or more IPM and that reducing IPM on these older units to one (1) degree may be virtually impossible. Still, enough reduction may be quite easily possible to reach minimal acceptable stereo performance.



*SIGNALS AT BOTH INPUTS MUST BE SINUSOIDAL ADJUST PHASE SHIFTER AND VERTICAL AND HORIZONTAL CONTROLS FOR A STRAIGHT LINE DISPLAY WITH NO MODULATION.

Figure 11. Block Diagram – Lissajous Pattern

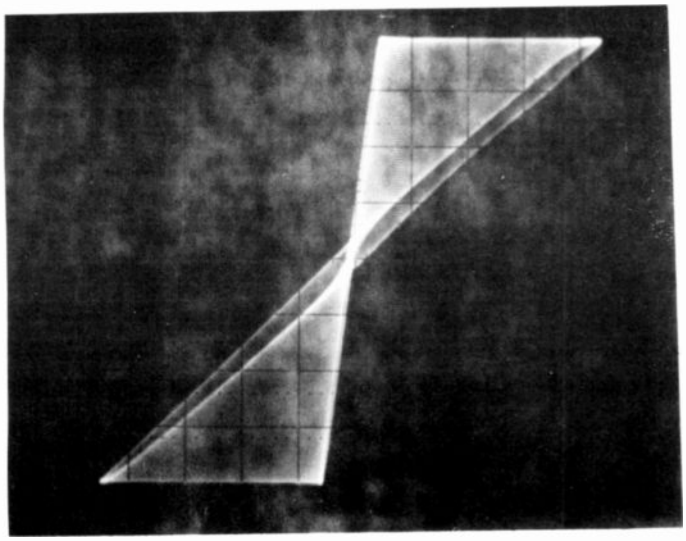


FIGURE 12. LISSAJOUS I.P.M. MEASUREMENT OF AM SIGNAL INDICATING APPROXIMATELY 3.8° PEAK PHASE DEVIATION AT MODULATION CREST. (80% MODULATION) AMPLITUDE OF TROUGH TOO SMALL TO MEASURE I.P.M.

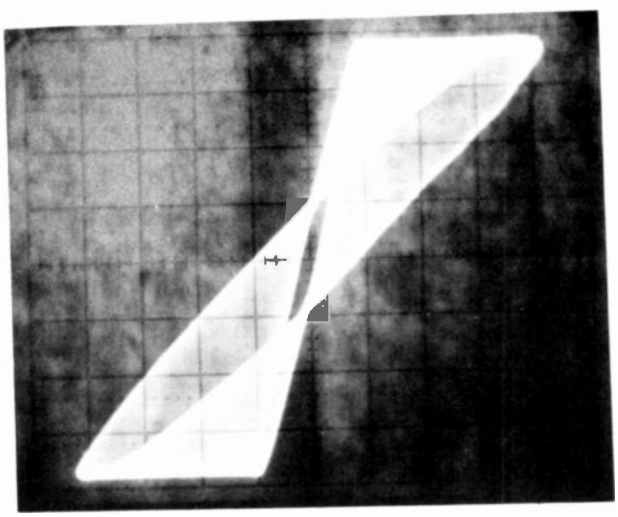


FIGURE 13. LISSAJOUS I.P.M. MEASUREMENT SHOWING 8.6° I.P.M. IN TROUGH AND 13.1° I.P.M. AT CREST OF MODULATION ENVELOPE (75% MODULATION DEPTH)

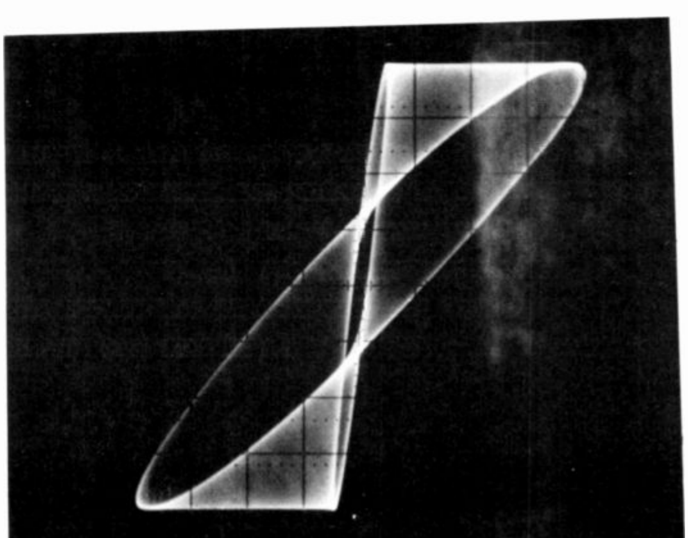
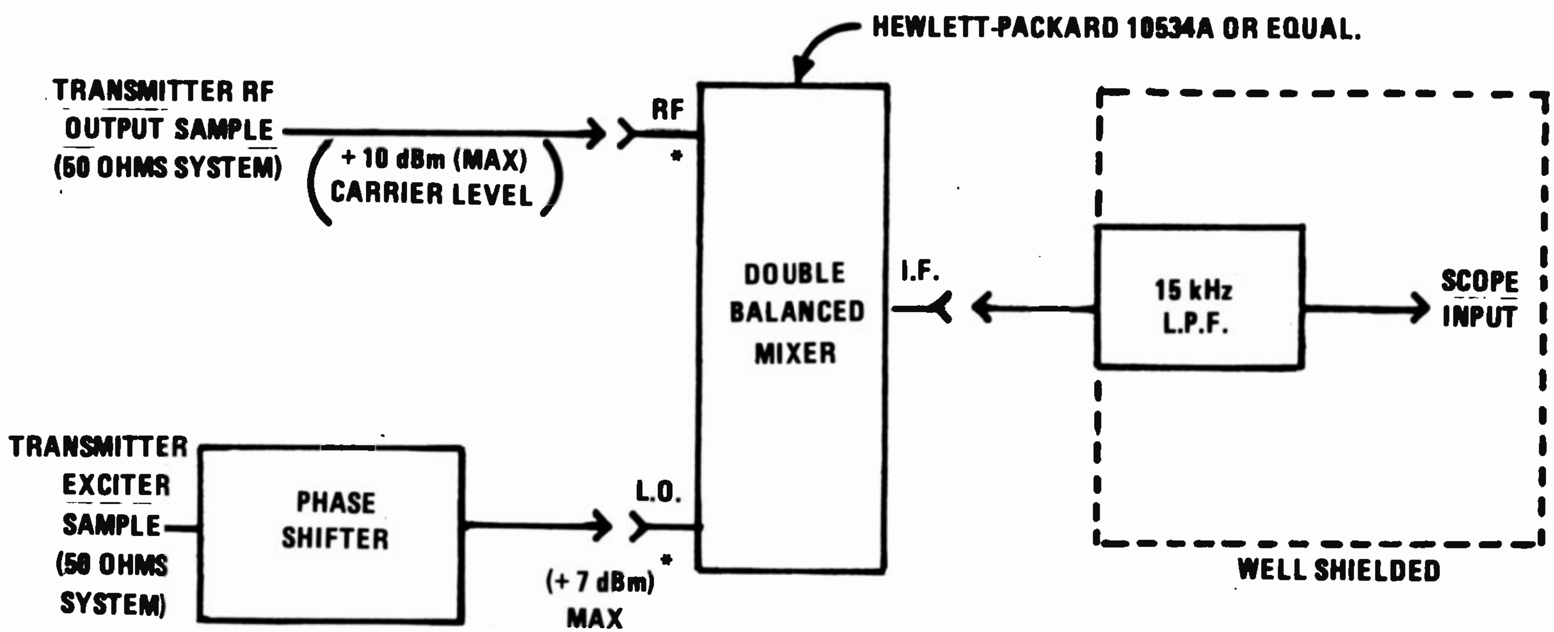


FIGURE 14. LISSAJOUS I.P.M. MEASUREMENT SHOWING 8.6° I.P.M. IN TROUGH AND 17.8° I.P.M. AT CREST OF MODULATION ENVELOPE (90% MODULATION DEPTH)



*SIGNALS SHOULD BE APPROXIMATELY SINUSOIDAL

METHOD:

STEP 1 ADJUST PHASE SHIFTER FOR MAXIMUM OUTPUT (EITHER POSITIVE OR NEGATIVE) ON SCOPE. THIS D.C. LEVEL REPRESENTS 57.3° of Dϕs

STEP 2 ADJUST PHASE SHIFTER FOR EXACTLY 0 VOLTS OUTPUT TO SCOPE. (90° PHASE SHIFT BETWEEN R.F. AND L.O.)

STEP 3 MODULATE TRANSMITTER WITH <10 kHz TONE AND NOTE PEAK PHASE SHIFT ON SCOPE.

QUADRATURE BALANCED MODULATOR METHOD TO MEASURE IPM

Figure 15. Typical Dϕs Error With This Technique is Approximately 0.1°

CAUSES OF TRANSMITTER IPM AND OTHER NEGATIVE SYSTEMS EFFECTS

The five major transmitter constraints to acceptable AM stereo operation are listed in Table 1 below. Items 1, 2, and 3 are directly related to basic causes of IPM and Items 4 and 5 can also have harmful effects on AM stereo performance.

TABLE 1
MAJOR CAUSES OF IPM AND OTHER
POOR AM STEREO PERFORMANCE CRITERIA

1. Improper or inadequate final PA neutralization.

2. Improper or inadequate RF driver neutralization.

3. Poor RF dynamic regulation of PA driver stage.

4. Insufficient bandwidth of RF chain (including antenna).

5. Assymetrical bandwidth of RF chain (including antenna).

CORRECTIVE PROCEDURES FOR DEFICIENCIES LISTED IN TABLE 1

The solution for Items 1 and 2 in Table 1 are simple to state though sometimes difficult to perform. Older transmitters which employ triode stages in either or both the driver and output stages are undoubtedly neutralized to the degree required for stable amplifier operation. The degree of neutralization required for acceptable IPM can be, and generally is, 10 to 20 dB better than for amplifier stability. The simple answer is to neutralize better.

Poor RF dynamic regulation of the PA driver stage has been found to be as equal a contributor to IPM as improper neutralization. The corrections to this possible cause of IPM, listed as Item 3 in Table 1, is not as obvious as Items 1 and 2 and can be more difficult to achieve. The four basic steps to correct RF dynamic regulation are:

A. Increase power capability of driver and swamp PA grid network with additional loss.

B. Determine the type of load change occurring on the PA driver tube. Possibly a phase shift network could be added between the driver and PA to present the correct shape of changing load at the driver output.

C. Combination of A and B.

D. Build or buy a solid-state broadband linear amplifier with sufficient output power to drive PA grid.

REFERENCES

The following list of references contain useful additional information regarding modulation theory, transmitter techniques and antenna system bandwidth and symmetry.

1. ELECTRONIC FUNDAMENTALS AND APPLICATIONS, J. D. Ryder, Prentice-Hall, 1959.

2. MODULATION, NOISE AND SPECTRAL ANALYSIS, P. F. Panter, McGraw-Hill, 1965.

3. MODULATION THEORY, H. S. Black, Van Nostrand, 1953.

4. RADIO ENGINEERS HANDBOOK, F. E. Terman, 1943.

5. "Sideband Energy of AM Transmissions," W. D. Mitchell, Continental Electronics Mfg. Co., paper presented at WABE, May, 1976.

6. "Operation of AM Broadcast Transmitters into Sharply Tuned Antenna Systems," W. H. Doherty, PROCEEDINGS OF IRE, July 1949.

7. IEC Recommendation 244-3B, 1972, "Unwanted Modulations."

ANOTHER LOOK AT AM DIRECTIONAL ANTENNAS AND PHASORS

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Consulting Engineers
San Francisco

ANOTHER LOOK AT AM DIRECTIONAL ANTENNAS AND PHASORS

The best AM broadcast directional antenna performance results from a team effort that pools the expertise of the consulting engineer, the phasing system designer, the phasor contractor, and the station personnel. Let's see what makes directional antenna systems good or bad by reviewing the following four stages in the life of a directional antenna. These stages are

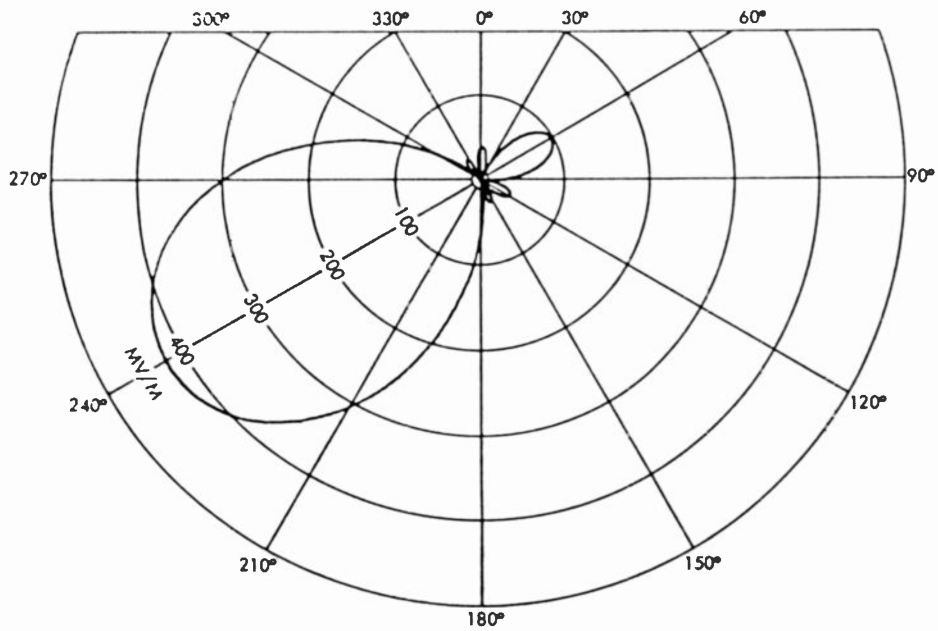
- Designing the directional pattern and obtaining an FCC construction permit.
- Designing, constructing, and installing the phasing system.
- Completing the field adjustment of the antenna system and the measurements necessary to prove that the authorized antenna pattern can be licensed.
- Maintaining the system once it has been licensed.

The first stage is almost exclusively the province of consulting engineers who specialize in the essential allocation studies and pattern designs necessary to obtain an FCC construction permit. Phasor design in the second stage is usually accomplished by the manufacturers of phasing equipment; they are often guided by the preferences of the consulting engineer who will eventually adjust the installed equipment. Phasor design was the subject of a paper I presented here three years ago at the 1979 NAB Engineering Conference; portions of that paper, which contained some rather radical ideas as to circuit design, have been integrated into this current presentation. The third stage, that of adjusting the directional antenna (DA) system, requires the efforts of both the consultant and station personnel to realize the actual directional antenna pattern that has up to this point existed only in the theory and the mathematics of the pattern design and the phasor design. Some of the brutish problems often confronted in the adjustment of a DA system will be discussed later. Finally, the licensed DA pattern must be maintained for many years within the authorized limits. This maintenance is usually the province of station personnel with occasional assistance from consultants. A directional antenna project is successful when the efforts of the consulting engineer, the phasor designer, the manufacturer, and the station engineers are effectively integrated.

DESIGNING THE DIRECTIONAL PATTERN

In considering the first stage, that of the DA pattern design, we find that decisions here can have a crucial effect on the ultimate performance of the array due to the operating impedance of each tower being defined almost entirely by the arrangement of the towers and the choices of height, field ratio, and phase. Historically, a great many arrays of three or more towers have been arranged with all of the towers in a single straight line, with a uniform spacing of approximately 90 electrical degrees between adjacent towers, and with tower heights at or near 90 degrees (one quarterwave length). For such in-line arrays that approximate 90 degree heights and spacings, the operating impedances are usually only marginally adequate for satisfactory bandwidth if the nulls or minima of the desired pattern are along the line of towers or at azimuths close to the line of towers. The pattern then approximates a cardioid. If, however, the desired pattern includes nulls that depart greatly from the line of towers and are substantially broadside of the line of towers, the resistive components of the individual tower operating impedances become very low or even negative. The Q of such towers becomes excessively high and the bandwidth is seriously impaired.

Figure 1 shows a real-life example of such a poor pattern design. The pattern has a pair of nulls that are even forward of broadside (less than 90 degrees from the axis of the main lobe). With such a pattern the radiation components from the individual towers do not add in-phase at the maximum of the main lobe. In other words the maximum radiation in the main lobe is far less than the arithmetic sum of the radiation from each of the four towers. A useful figure of merit when considering such systems is the ratio of the square root of the sum of the squares (RSS) of the individual tower radiation amplitudes to the root-mean-square (RMS) radiation of the horizontal plane pattern. This ratio of tower field RSS to horizontal plane RMS yields a useful figure of merit in evaluating alternative patterns. With the best patterns, the ratio of RSS to RMS is near unity and can even be considerably less than unity. A poorer array has a larger ratio of



Tower No.	Height	Bearing	Distance	Relative Field
1	90°	N 00° E	0°	1.00
2	90	60	90	2.26
3	90	60	180	2.26
4	90	60	270	1.00

Tower No.	Phase	Operating Impedance	Q
1	0°	-11.6 +j 85.5 ohms	7.37
2	161.9	19.2 +j 89.9	4.68
3	-37.6	21.8 +j 78.4	3.60
4	124.2	15.2 +j 58.1	3.82

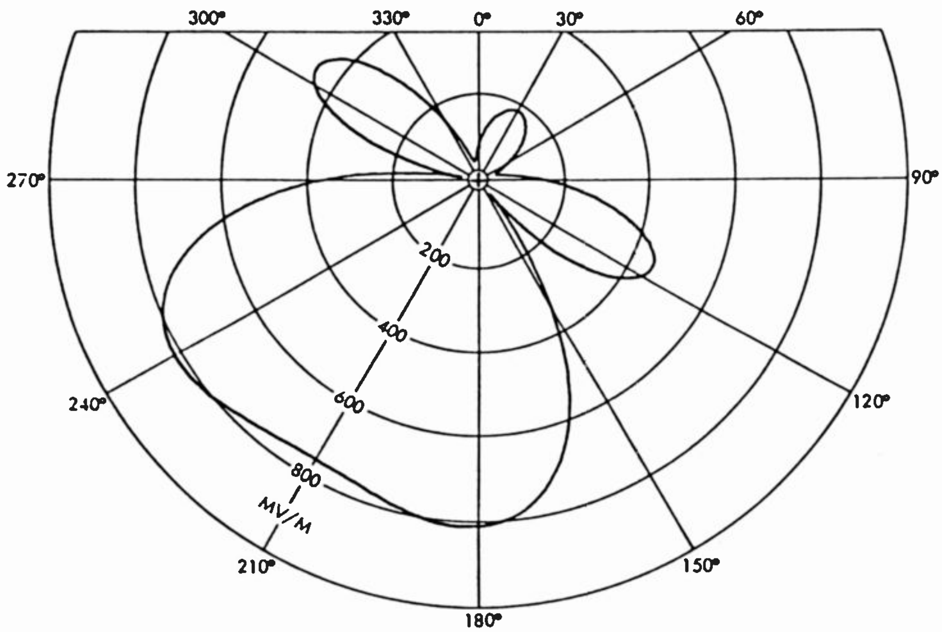
Theo. RMS=180 mV/m Theo. RSS=478 mV/m RSS/RMS=2.66

Measured 10 kHz sideband VSWR = 2.0:1

FIGURE 1. EXAMPLE OF POOR PATTERN DESIGN

RSS to RMS. This ratio for the poor example shown in Figure 1 is 2.66. I have discovered one old array in which this ratio exceeds 5. For the example in Figure 1, the measured worst-case common point VSWR at 10 kHz above or below the carrier was 2:1. We know of three old and poorly designed arrays in which the 10 kHz sideband VSWR performance was as bad as 3:1 even though perfectly matched at the carrier frequency.

The apparent Q of each tower (X ÷ R) is also a useful figure of merit. For a single-tower nondirectional operation, the tower Q thus indicated can provide a reasonably accurate indication of bandwidth by reference to the universal resonance curves that are familiar to all circuit designers. However, the tower Q's tabulated in Figure 1 bear no direct relationship to bandwith because the base impedance has been defined only for the



Tower No.	Height	Bearing	Distance	Relative Field
1	120°	N 00° E	0°	1.00
2	120	30.5	115	1.00
3	120	30.5	335	1.00
4	120	30.5	450	0.80

Tower No.	Phase	Operating Impedance	Q
1	66.8°	367 +j 320 ohms	0.872
2	162.0	277 +j 275	0.993
3	0.0	328 +j 307	0.936
4	96.3	204 +j 220	0.927

Theo. RMS=470 mV/m Theo. RSS=410 mV/m RSS/RMS=0.87

Measured 10 kHz sideband VSWR = 1.03:1

FIGURE 2. EXAMPLE OF GOOD PATTERN DESIGN

carrier frequency. Any departure from carrier frequency can produce a radical change in both the resistive and reactive components of base impedance due to the mutual coupling between the towers. The actual bandwidth of an element in a directional array can be far narrower than that implied by the tower Q's as derived and tabulated for the carrier frequency alone.

Figure 2 is an example of a rather good DA pattern design. Although the pattern shape in this example differs significantly from that of Figure 1, the important difference is not in the shape itself, but in the method by which the desired pattern is generated. Here again we have a pair of nulls forward of broadside to the line of towers, but the choices of tower spacing and heights have been adjusted so as to drastically improve the operating impedances of the individual towers and the corresponding indications of individual tower Q. In this particular array the ratio of the RSS of the tower fields to the RMS of the pattern is an impressively low 0.87 and the measured worst-case common point VSWR is only 1.03:1 at 10 kHz above or below carrier.

The essential point of this exposition of good and poor pattern designs is to stress that a poor design with poor operating impedances and a high RSS to RMS ratio yields an inherently poor directional antenna system. No amount of clever phasor design can ever remedy all of the deficiencies. The directional antennas with the best bandwidth and stability are those that give consideration to the individual tower operating impedances and the RSS to RMS ratio in the early stages of pattern design before a construction permit is sought.

DESIGNING THE PHASING SYSTEM

A good approach to designing a phasing system is to consider the entire collection of components between the transmitter and the towers as a whole, rather than by the building block approach implicit in Figure 3, which shows the classical block diagram of a directional antenna phasing system as it appears in the NAB Engineering Handbook. This diagram is useful for understanding the functions to be accomplished within a phasor. However, the electrical performance may be disappointing if the design process proceeds by a piecemeal block-by-block approach.

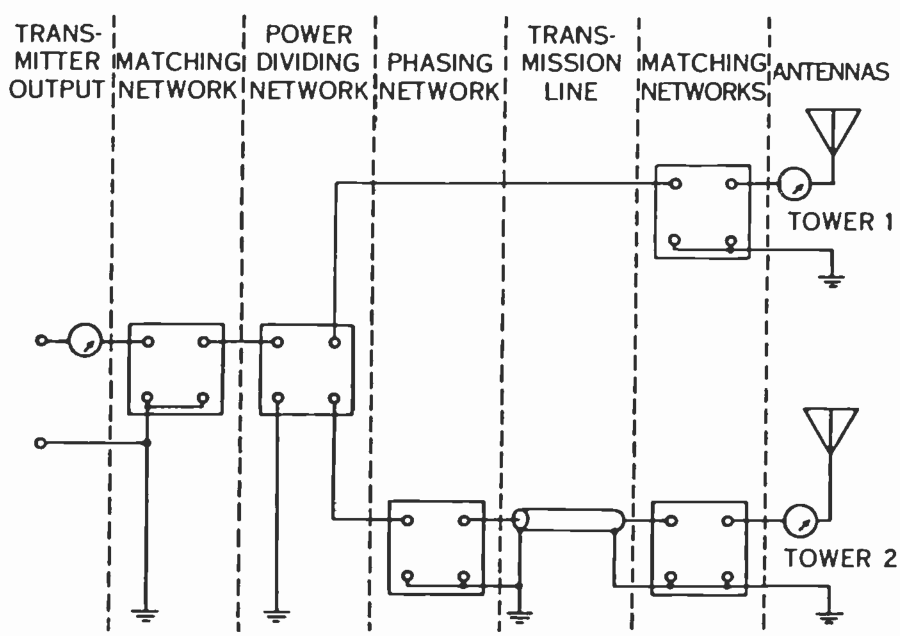


FIGURE 3. PHASOR BLOCK DIAGRAM

Phasor Performance Objectives

Two important performance objectives are to have both the phasor input impedance and the directional antenna radiation pattern remain essentially constant across the channel. To meet the first objective, an ideal phasor would present a load to the transmitter that is purely resistive and unchanging at all sideband frequencies of interest. Because phasors consist of coils and capacitors whose reactance varies with frequency and the antennas themselves represent impedances that also vary with frequency, this ideal can never be achieved. The term "impedance bandwidth" is used to describe the degree of constancy of the phasor input impedance across the entire range of sideband frequencies.

A useful figure of merit to express the impedance bandwidth as a single number is the worst-case voltage standing wave ratio (VSWR) at the common point for either of the two $\pm$ 10 kHz sideband frequencies when perfectly matched at carrier frequency. This approach yields a VSWR number that relates to antenna performance just as television and FM antenna performance is described by VSWR limits at various sideband frequencies.

Bandwidth requirements can vary greatly. At 1600 kHz, a total bandwidth of 20 kHz corresponds to only about 1% of the center frequency, while at 540 kHz such a bandwidth corresponds to nearly 4% of the center frequency. We can draw upon the experience of many antenna and phasor designs and make useful comparisons, even though they were devised for different frequencies, by considering the VSWR at frequencies which are 1% above and below the carrier frequency. These correspond to a bandwith of $\pm$ 10 kHz at 1000 kHz. The better directional antenna systems exhibit a measured common point VSWR of 1.1:1 or better at frequencies 1% above and below carrier.

The second objective is to have the antenna radiation pattern remain unchanged across the channel. Variations in the phase and ratio of tower currents at the sideband frequencies will result in changes in the location and depth of the intended pattern nulls. The frequencies for which the pattern remains useful describe the "pattern bandwidth". A real-life example of poor pattern bandwidth is shown in Figure 4. A spectrum analysis of the received signal in the null sector confirmed the extreme sideband dissymmetry. The cure was a redesigned phasing system. There is no simple method to define the pattern bandwidth by a single number, but it is not difficult to analyze alternative phasing systems to see which has the better pattern bandwidth.

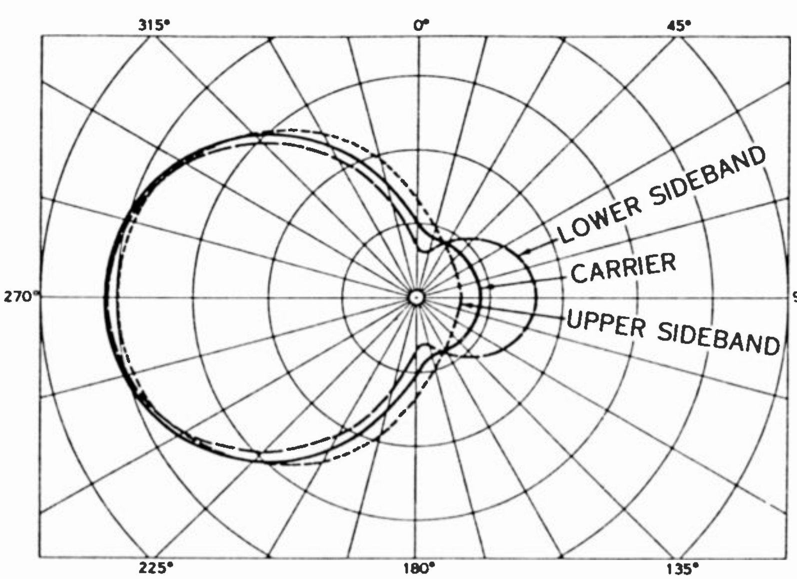


FIGURE 4. EXAMPLE OF POOR PATTERN BANDWIDTH

Considerations for Improved Phasor Designs

Many designs can be considered in arriving at optimum phasor performance. Let's look at five possible alternatives.

1. Consider alternative impedance matching network configurations. The variety of useful networks include leading and lagging "T" networks, "L" networks, and "phantom T" networks in which the output arm consists of the reactive component of a tower impedance. All of these are practical options that can offer improved performance in many cases.

2. Consider transmission lines as impedance transformers. The consequences of mismatched transmission lines are entirely different in an AM antenna system than in FM or television systems. Frequently, mismatched transmission lines can provide a desirable impedance transformation function.

3. Eliminate discrete power divider circuits. The power division function can be accomplished by adjustment of "T" networks or, in the case of two-tower arrays, by simple series elements arranged in a back-to-back configuration. Elimination of discrete power divider circuits reduces the number of components required.

4. Optimize the delay to the most narrow-band tower. Adjustments of the total phase shift from the point of common power division (which can be considered as a constant-voltage source) to the highest Q tower, so that this tower is an odd number of quarter-wavelengths away from the point of power division, can provide a useful improvement in bandwidth. Of course the phase delays to the other towers must also be adjusted so as to produce the proper pattern, but the procedure is effective in improving bandwidth of the tower that would otherwise exhibit the poorest bandwidth.

5. Reduce the total reactive power within the phasor. General experience indicates that minimizing the total reactive power stored in all the coils and capacitors of a phasing system improves the impedance bandwidth. This is especially true when frequency-selective traps are involved to minimize reradiation or cross-modulation with nearby stations. The reactive power in any circuit element is measured in volt-amperes reactive (termed "vars" or "kilovars" in the power industry) and is simply the product of the current squared times the reactance.

The example of poor pattern bandwidth cited in Figure 4 was cured by redesigning the phasor. The total reactive power in that phasing system was reduced to 25% of its original value. Reducing the total reactive power usually means reducing the number of components in a phasor; therefore, we strive for elegant simplicity in phasor design.

Examples of Alternative Phasor Designs

Following are examples of phasor designs that demonstrate one or more of the considerations outlined above. Figure 5 shows a 50-kilowatt array in which approximately 32 kilowatts are delivered from the output of the common point matching network directly to the center tower through a single vacuum capacitor with no network or transmission line losses. As a result, the losses in this phasor are exceptionally low. The measured worst-case 10 kHz sideband VSWR is 1.07:1 and $\pm$ 1% bandwidth is 1.1:1.

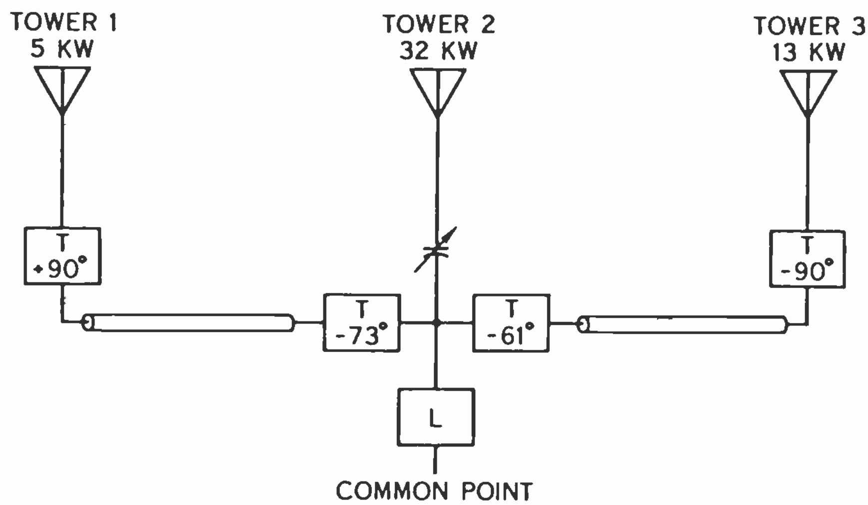


FIGURE 5. PHASOR WITH MINIMUM LOSSES

Figure 6 shows a two-tower, 1-kilowatt phasor of very simple design. This design not only locates the transmitter building and phasor at the base of one tower so as to eliminate a transmission line but also employs an unmatched transmission line to feed the second tower. This system has proved to exhibit an exceptionally flat common-point impedance characteristic. The measured worst-case 10 kHz sideband VSWR is 1.03:1 and the ± 1% bandwidth VSWR is 1.04:1. (Our firm designed this phasor in 1956.)

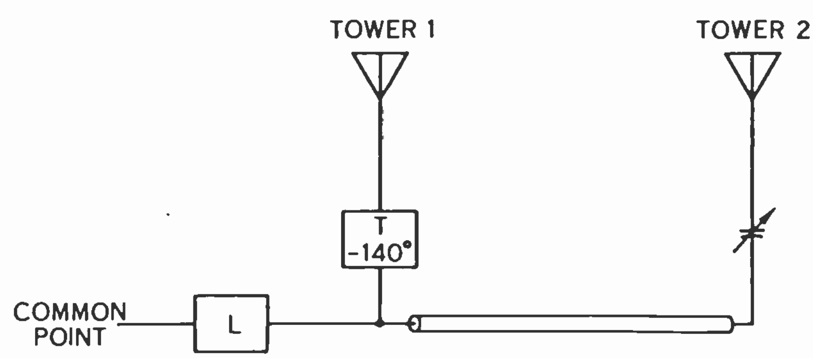


FIGURE 6. PHASOR WITH UNMATCHED TRANSMISSION LINE

Figure 7 is an example of a low-frequency (560 kHz) two-tower, 5-kilowatt system for electrically short (63 degree) towers. It incorporates leading L networks at both towers and a back-to-back arrangement consisting of L1 and C1 for adjusting phase and ratio. The measured worst-case 10 kHz sideband VSWR for this system is 1.09:1. The ± 1% bandwidth VSWR is 1.04:1.

Many factors affect the selection of an optimum phasor design, including phasor location, transmission line lengths, velocity, and tower operating impedances. Under favorable circumstances these factors can permit extreme simplicity in phasor design as shown in Figure 8. This 50-kilowatt two-tower system has a transmitter and phasor located at the base of one tower, utilizes a mismatched transmission line to the second tower, and employs a pair of vacuum variable capacitors (C1 and C3) in a back-to-back arrangement for phase and ratio control. Common-point impedance matching is accomplished with an L network consisting of L1 and C4. The sum total of essential parts in this entire directional antenna system is only four capacitors and one coil. The measured worst-case 10 kHz sideband VSWR is 1.04:1, and the VSWR 1% above and below carrier is also 1.04:1.

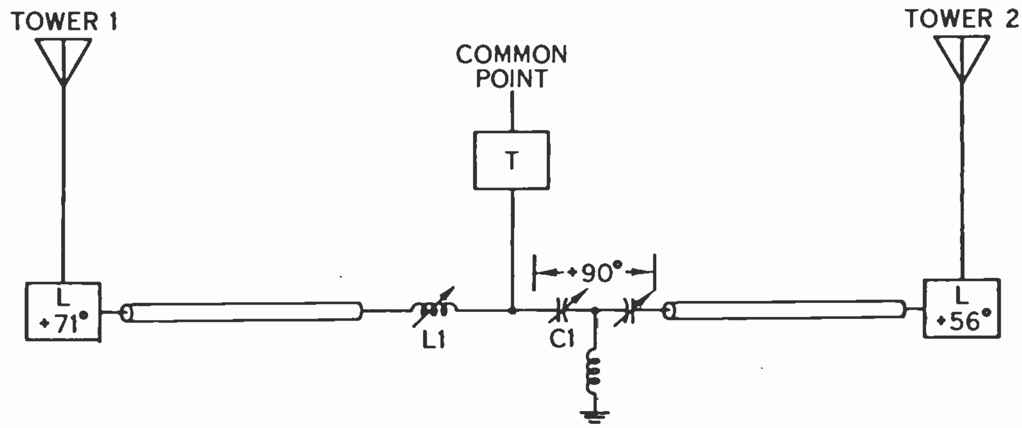


FIGURE 7. PHASOR WITH LEADING L NETWORKS AT TOWERS AND BACK-TO-BACK PHASE AND RATIO CONTROL

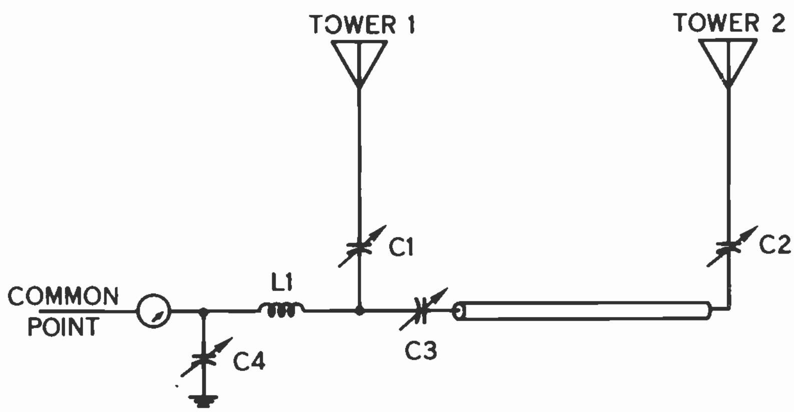


FIGURE 8. PHASOR WITH MINIMUM NUMBER OF COMPONENTS

ADJUSTING A DIRECTIONAL ANTENNA

The realities of inadequate pattern design or inadequate phasor design or an inappropriate selection of transmitter site may first become apparent when attempts are made to adjust an array to within its authorized pattern limits. Four frequently encountered problems and their solutions are discussed below.

1. The required tower base currents and phases may depart so far from the design values that much time is wasted in adjusting to the proper pattern. The tower base current ratios and phases bear a close relationship to the relative amplitudes and phases of the actual fields radiated by the individual towers only in the case of arrays with electrically short towers that utilize wide spacing between the elements so that mutual coupling effects are minimized. In most arrays, particularly those having electrically tall towers or different height towers within the array, extreme departures between the required base current ratios and phases and the desired individual tower radiation ratios and phases can result.

A simple example of the problem is illustrated in Figure 9. A tower that is electrically taller than 180 degrees must be driven with a base current 180 degrees out of phase with respect to the current that would have been required in a shorter tower to produce the same relative phase of tower radiation. Computer programs utilizing the method-of-moments technique now permit a complete determination of the current distribution throughout the height of each tower as to both amplitude and phase. With this information the amplitude and phase of the base currents necessary to generate the proper amplitude and phase of the individual tower radiation components can be calculated. This technique has proved to be a considerable timesaver by providing a much better starting point for field adjustments than simply setting the base current phases and ratios to the desired individual tower radiation field ratios and phases.

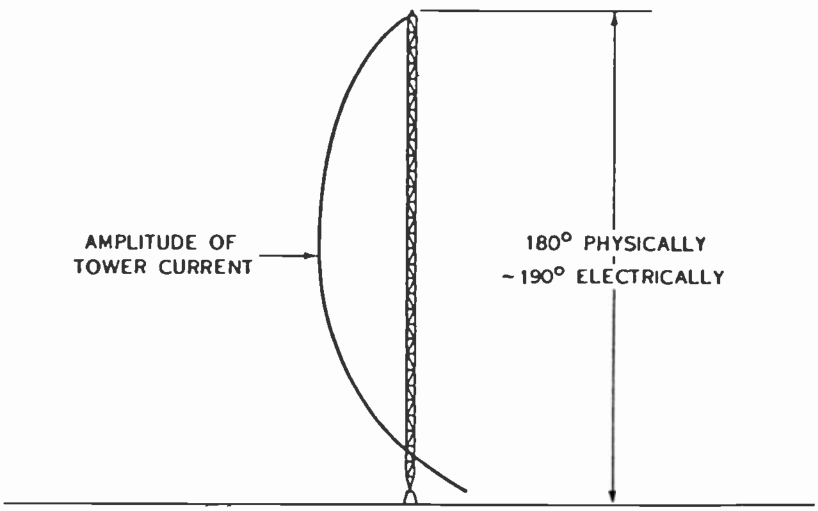


FIGURE 9. 180-DEGREE TOWER

2. Initial field strength measurements may be insufficient in number and location and thus be misleading as to the actual radiation in critical directions. Theoretically a directional antenna pattern is not fully formed except at an infinite distance where the separate towers can be considered as a point source. As a practical matter, near-field effects can often persist as far as 20 to 30 miles from an antenna before far-field conditions prevail. This is especially true in the deep nulls of wide-spaced arrays; however, misleading measurement results can often occur under apparently innocent circumstances. To guard against this eventuality, we routinely calculate the near-field results on all radials prior to all directional antenna proofs. This calculation, which is easily done by computer, involves consideration of the actual inverse distance attenuation and actual phase delay from each antenna element to a series of observation points along each radial.

   Figure 10 shows the results of such a calculation on a typical null radial and the resulting analysis of field strength measurements. Line A is the inverse distance line for the theoretical unattenuated radiation at one mile. Line B is the result of the near-field calculations assuming only inverse distance attenuation, that is, no soil losses. It converges with the inverse distance line with increasing distance. Line C represents a soil conductivity of 10 mmho/m as drawn in the conventional manner from analysis of the nondirectional measurements on the radial. Line D is a composite of Lines B and C. It includes the near-field calculations and is attenuated with distance in accordance with the soil conductivity previously established. This composite line converges with the near-field calculations at short distances where soil attenuation is negligible and converges with the soil line at great distances where the near-field effects disappear. Since Curve D accounts for both near-field effects and soil losses, it is the proper curve against which the field strength measurement data should be fitted. Note the excellent fit to the measurement data, both close to the array and at distant points even though the first 19 measurement points all fall considerably above the inverse distance Line A.

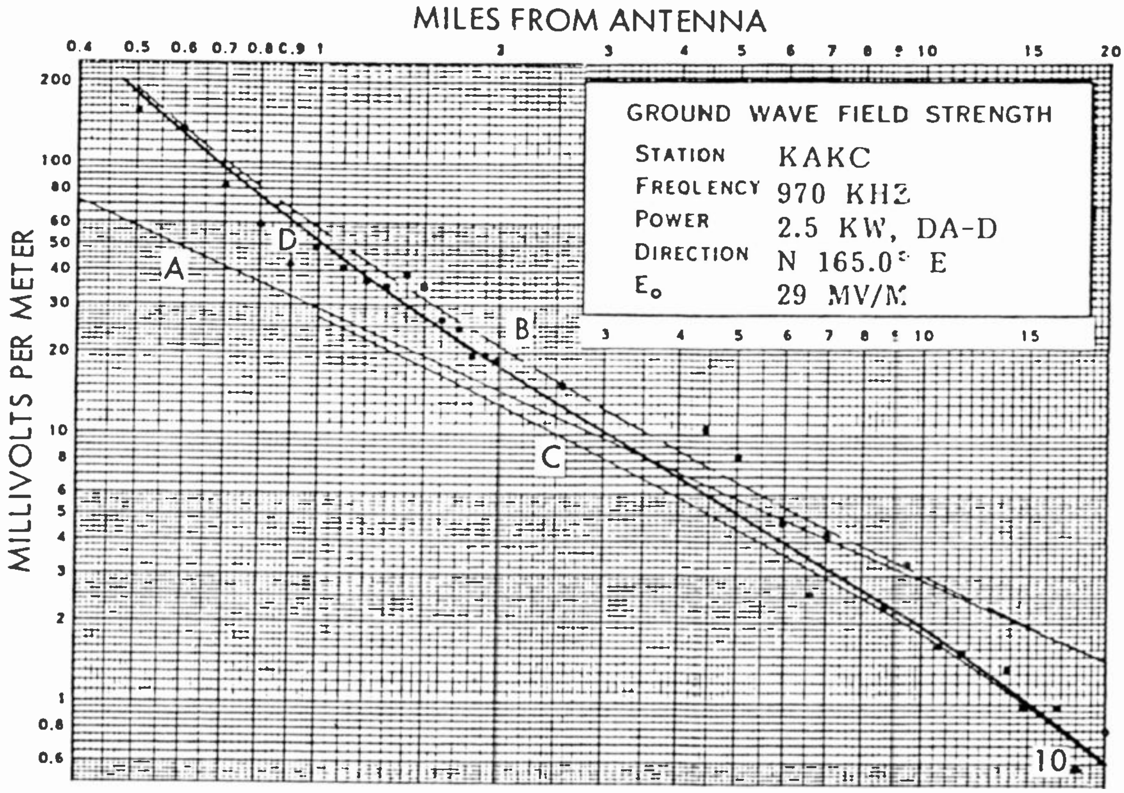


FIGURE 10. NEAR FIELD EFFECTS

3. Reradiation from objects external to the array may distort the pattern excessively. Of course the ultimate cures for excessive reradiation are to either detune every external structure having significant currents flowing in it or else to relocate the station to a better environment. However, there can be a practical alternative if this eventuality is considered in the pattern design stage. The effect of excessive reradiation on the patterns of arrays having all of the towers in a single line is to destroy the pattern symmetry that would otherwise exist. Reradiation will destroy the ability to simultaneously develop symmetrical deep nulls or to maintain suppression over a broad arc if that is what the allocation conditions and pattern design require. This problem with reradiation can be minimized by avoiding in-line arrays in situations where symmetrical nulls or suppression over broad sectors is required.

4. In multi-tower arrays, the combination of parameter changes necessary to bring the pattern simultaneously within all radiation limits may be too complex for simple analysis. The amount of cut-and-try necessary to progress from the initial pattern that results when an array is first turned on to the final adjusted

pattern that meets all radiation limits can be drastically reduced by the judicious use of computer optimization programs. We have developed three such programs, each of which is appropriate for specific circumstances. The first is for arrays that contain deep nulls in a sufficient number to uniquely define the one set of phase and ratio parameters necessary to develop the pattern. A simple "talkdown" procedure is utilized. Any arbitrary phasor adjustment is made that will (in turn) null the observed field strength at three or four points along each of the critical null radials (or small minor lobe radials). The resulting antenna monitor observations of tower current ratio and phase for each of the series of individual talkdowns are then fed to the computer along with the theoretical pattern parameters. The computer gives back the antenna monitor indications that will generate all of the necessary nulls and small minor lobes simultaneously. In this situation only one or two iterations are usually required to bring an array within all limits.

A second program works very well with patterns that have no deep nulls but have numerous radiation limits considerably below the pattern RMS. This program is based on the exceptional stability and resetability of the vacuum variable capacitors within the phasor and does not work as well with roller coils. First, the sensitivity of each phasor control is determined by observing the changes in dial reading necessary to produce a change no greater than approximately 2 degrees or 2% in any of the phase or ratio parameters. These changes in dial readings (both plus and minus from starting conditions) are marked near each dial, and the adjustment process begins. With two-way radio communication, the change in field strength at each of several measuring points along each of the measurement radials is recorded as each dial is varied in turn to the plus or minus dial limits previously established. The observed changes in field strength, together with dial variations, then form a computer input matrix along with information as to the increase or decrease in radiation desired on each of the radials. The computer then gives back a new set of dial readings which will bring the radiation pattern into closer agreement with the desired targets. Once the initial knob sensitivities are established, this program does not rely on any observations of the antenna monitor, only on the dial indications on the vacuum variable capacitors. This program works very well under heavy modulation conditions where antenna monitor indications of small changes would be difficult to read. Generally, several iterations are necessary because the program must assume that the effects of the knob changes on the resulting changes in field strength are linear and mutually independent.

A third optimization program has proved to be particularly successful with complex arrays and patterns. In this case very complete measurements are made to define the initial pattern immediately after the array has been turned on and adjusted to the best predicted phase and ratio parameters. Use is then made of the computer optimization program that was first developed years ago to optimize directional antenna patterns. In this application the program is used backwards. We start with the theoretical pattern as authorized and let the optimization program alter it to give the best possible fit to the actual pattern as measured. The inverse of the parameter changes found necessary in optimizing from the theoretical pattern to the measured pattern are then cranked into the array. The next set of complete field measurements shows that the pattern is at, or at least closer to, the desired theoretical pattern.

MAINTAINING A DIRECTIONAL ANTENNA

If a DA system has been properly designed and constructed, maintenance is minimized. Occasions for parameter readjustment can be extremely infrequent unless the directional antenna is classified by the FCC as a "critical array" or is unusual in some other respect. We are familiar with many arrays that have functioned for ten to fifteen years with no readjustment and virtually no maintenance attention. However, maintenance problems are occasionally encountered, so let's discuss a few of them.

1. RF connections within the phasor or antenna coupling units can become loose and overheat. Prevention consists of simply exploring with your fingers immediately after signoff for hot spots in the components and bolted connections in the phasing system. The rollers on roller coils are particularly vulnerable to increasing resistance with resulting heat build up.

2. Lightning strikes on the towers can occasionally destroy components in the phasing system. This hazard can be minimized by providing ball gaps across tower bases and adjustable horn gaps inside the cabinets at both ends of all transmission lines as well as on the antenna feed inside each antenna coupling unit. All gaps should be set to approximately twice the clearance at which they just break down on modulation peaks. This is a much closer gap spacing than many engineers expect. Inserting a small inductance in the RF feed from the antenna coupling unit to the tower is also common practice.

3. Excessive reradiation can develop as structures such as power line towers or buildings are erected in the vicinity of the array. Such reradiation is a function of the height and cross section of the structure, as well as the incident field it intercepts. Obviously tall structures close to the array and in the main lobe of radiation, where they are strongly illuminated, generate the most serious problems; however, reradiation from other radio towers as much as several miles away can cause observable and measurable effects. For reradiating towers less than a quarter-wavelength high, any technique that reduces the RF current flow to ground from the structure will reduce the reradiation. An effective toroidal current sampling transformer to measure such currents can be made from a length of 4-inch plastic clothes dryer vent hose. The spiral steel wire in such hose makes an excellent pickup device. A length of the hose can be wrapped around the base of the offending reradiator and connected to a field strength meter operating as a tunable voltmeter to measure the current flowing to ground. Calibration is accomplished by wrapping the hose around any conductor within the phasing system that is carrying a known current.

A SIMPLE BANDWIDTH TEST

If you want to evaluate the bandwidth of your DA system, let me suggest a simple test that can determine the amplitude of each high-frequency sideband component as radiated by each tower. This test is possible with any common field strength meter because the selectivity of such meters is just sufficient to resolve sideband components that are 10 kHz removed from the carrier. The procedure is as follows:

1. Keep the sampling lines terminated into the phase monitor, but add "T" connectors to bridge off samples into a field strength meter operating as a linear tuned voltmeter.

2. Modulate the transmitter 50% with 10 kHz sine waves. In a perfect system, this would result in sideband amplitudes equal to 25% of the carrier level.

3. For each tower sample first adjust the field strength meter gain so as to set the carrier level to full scale (100%). Then tune in each sideband in turn and log each sideband amplitude as a percent of full scale.

4. Repeat the process for each tower.

Figure 11 shows the 10 kHz sideband components in each tower of good and poor arrays as measured by this technique. I don't see an easy way to reduce these numbers to a meaningful single figure of merit but they certainly will pinpoint the bandwidth limitations within an array.

Poor Antenna Design				Good Antenna Design		
Lower S.B.	Carrier	Upper S.B.	Tower Number	Lower S.B.	Carrier	Upper S.B.
22.5%	100%	7.7%	1	25.5%	100%	26.2%
18.0	100	10.0	2	25.0	100	26.5
16.0	100	24.0	3	25.5	100	26.5
20.5	100	35.5	4	26.5	100	25.0

FIGURE 11. SIDEBAND MEASUREMENTS RESULTS

In the main lobe of a DA pattern, deficient 10 kHz sideband amplitudes do not usually result in measurable distortion because the radiation components from the individual towers add more-or-less in phase and the sideband deficiencies of any one tower are masked by the sum total of radiation from all of the other towers. However, in null sectors where radiation is suppressed because the vector sum of radiation from all the towers is at or near zero, inadequate sideband performance in only a single tower can result in serious audio distortion. This cause of audio distortion will still exist and probably become worse with any of the AM stereo systems proposed.

I will close with the wish that when you make this test you will find that both 10 kHz sidebands in all of your towers are alive and well and as close to 25% as in the example I cited. Then you'll know you have a DA system with the best possible bandwidth.

"Practical RFI Elimination Techniques

for the Broadcast Engineer"

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With the ever increasing emissions of RF, interference to broadcast station equipment is increasing. This paper will deal with three major areas of RFI (Radio Frequency Interference): sources and modes of interference, techniques used in newly designed equipment to prevent RFI, and a discussion of practical techniques a station engineer can use to minimize RFI in the station equipment.

The emphasis in this paper will be on the practical ideas and concepts that a station engineer can use to reduce interference to equipment. The first two sections of the paper will be brief overview of subject material.

A broadcast station's RFI problems are most often caused by its own transmissions, but in today's urban environment, interference from other sources can be just as troublesome. Potential sources of interference include, but are not limited to:

1. Other broadcast station operations (AM, FM, TV, STL, etc.,).

2. Citizens band transmissions

3. Public Service and Business Band transmissions.

4. Ham Radio

5. Digital Equipment

RF emissions, of course, vary widely in frequency and intensity from source to source. The list above is in no respect complete.

Interference to station equipment occurs primarily through two modes, Radiated and Conducted.

In the radiated mode interference is caused by components and circuits being directly affected by RF fields, a voltage is induced in the afflicted part and when rectified will cause interference. Interference due to this mode is usually worse in equipment with inferior shielding qualities.

RF entering the equipment via the various leads is called "Conducted RF". This form of RFI is usually the major cause of problems to broadcast equipment. RF enters the equipment indirectly thru the leads, is rectified by an active stage, and interference results.

These recent years station engineers have voiced increased concern over RFI problems to their station equipment.

Some equipment designs today are now taking into account the problems that the station engineer faces from RFI. The next area of discussion will focus in on some examples of RFI prevention techniques that have been incorporated into recently designed Broadcast Electronics equipment.

Figure 1 and Figure 2 show the RFI filter assembly from the Broadcast Electronics FX-30 FM exciter. All audio and control leads enter the exciter through this assembly. Each lead is filtered by a multi-section filter, with a feed-through capacitor as a shunt element. This assembly strips off RF energy from the leads before the RF reaches the main cavity of the exciter chassis.

The third area of discussion in this paper will deal with some practical applications, that a station engineer can use to minimize RFI to station equipment. We noted earlier that there were two principle forms of RFI to equipment (Radiated and Conducted). First, we will deal with some concepts and "Cures" for radiated RFI problems.

Radiated RFI is usually more pronounced in equipment that has poor shielding qualities. Loose fitting covers, doors or removeable panels may provide a single point low ohmic contact to the chassis of a given piece of equipment, but have little shielding ability at 100 megaHertz. A given piece of equipment may also work fine at AM broadcast frequencies but, in a strong RF field at 100 megaHertz, be susceptible to interference. A loose fitting or poorly bonded cover often does a beautiful job of reradiating RF into a chassis. At VHF frequencies fasteners on a cover or panel may be a quarter wave length apart, again seriously diminishing the shielding of the equipment. A practical demonstration of shielding effectiveness versus frequency is to take a battery powered AM-FM radio inside a car or van and compare AM reception versus FM reception. After noting the reception of the various stations on both bands, step outside the vehicle and compare reception on the two bands. Normally inside the vehicle you will note fairly poor reception AM band, and very little difference in FM band reception. The purpose of this exercise is to demonstrate that as frequency of a RF source rises (for this purpose comparing AM with FM) the shielding of an enclosure diminishes. The longer wavelength of AM can not penetrate the comparatively small openings in the vehicle but the short wavelengths of the FM broadcast can. This relation is also true for the various types of equipment found at a station, with respect to their shielding integrity at high (VHF-UHF) frequencies.

BASIC STEPS FOR CONTROLLING RADIATED RF

There are some steps that an engineer can take to minimize radiated interference by improving the RF integrity of a given piece of equipment. Listed below are some general concepts that an engineer can take to improve the shielding of the equipment.

1. Use internal tooth lock washers under screw heads and nuts that join chassis members together. This type of washer bites through paint and coatings to provide positive contact with the bare metal. Scraping paint off where fastners are used will also help.

2. Reduce the distance between the fasteners that bond chassis members together. Refer to the photograph of the Phono preamp (Fig.7). Note the short spacing between screws. This is done to ensure good cover bonding even at VHF frequencies.

3. Add shielding strips such as fingerstock and mesh gasket material to to removeable covers and panels.

4. Short, Wide, bonding straps between the chassis and swing out doors etc. can make a big difference in the RF integrity at VHF and UHF.

5. For small projects, a very RF tight box may be constructed out of blank double-sided PC board material. Solder all edges carefully, and the enclosure will be an effective RF shield.

Conducted RF enter equipment via various leads (power line, inputs, outputs, etc.), and causes interference by directly feeding RF to active circuits or reradiating RF inside the chassis. For example, at FM broadcast frequencies, the 36 inch leads from a turntable to a phono preamp often act as a quarter wave antenna, doing a fine job of delivering RF into the preamp.

A common symptom of conducted RFI is changing levels of interference to a piece of equipment as the leads are moved about.

Conducted RFI in existing equipment can often be minimized by the implementation of the following concepts. The most effective and practical method of minimizing conducted RFI, is the filtering of all leads entering the equipment. We looked earlier at some photographs on page 3 and 4, showing some filter networks in equipment. In many cases RF filtering may be added to existing equipment by an engineer.

Several basic filters make up the majority of decoupling networks used to curb conducted RFI.

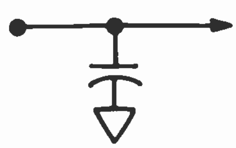


Fig. A

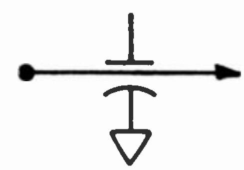


Fig. B

A single element shunt capacitor filter (Fig. A) will do a fair job of decoupling a line of RF if the reactance of the capacitor is low at the frequency of the interfering signal. The capacitor leads, as in all cases where the capacitor is used as a shunt element, must be kept short as possible. For effective operation into the UHF bands, a feedthrough capacitor should be used (Fig. B). A series inductor as a single element filter is not recommended as it may just reradiate RF into the chassis.

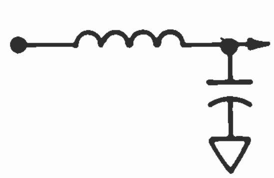


Fig. C

Fig. D

A two element filter network formed by a series inductor and a shunt capacitor (Fig. C) can exhibit good RF decoupling characteristics over a wide bandwidth. With the use of a 1000 pF feedthrough capacitor and ferrite choke, over 60dB of attenuation at FM broadcast frequencies can be obtained (Fig. D).

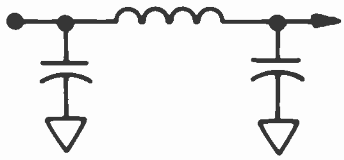


Fig. E

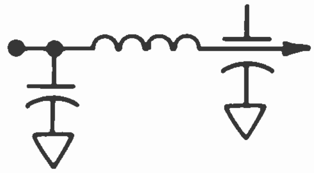


Fig. F

A three element "PI" filter, when components are properly selected, can provide 40 dB of RF attenuation over a wide bandwidth (Fig. E). As with the other two filters above, for good effectiveness at VHF and UHF frequencies, the third element should be a feedthrough capacitor (Fig. F).

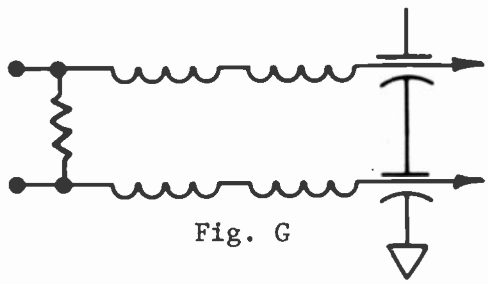


Fig. G

Figure G is an example of a wideband filter used on a balanced audio line. Two inductors are used in series to insure high series impedance over a wide bandwidth. Low impedance feedthrough capacitors insure effective operation into the UHF bands. The 620 ohm resistor was placed in front of the inductors to insure that they do not saturate under high frequency, high level audio signals.

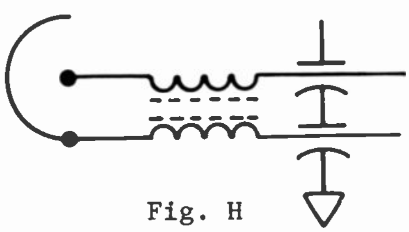


Fig. H

The filter network above (Fig. H) is a wideband network used on the input to a phono preamp. A shunt C shorts RF energy traveling on the lead outer conductor to the chassis. A bi-filar wound toroid choke effectively filters out differential mode RF. Because of the high impedance of the preamp, a 50 pF feedthrough was used to filter the signal lead.

The capacitors and inductors used to form filter networks, often behave in opposite ways from what we desire. For example, at 100 mHz some capacitors will act as inductors, and some inductors will act as capacitors. The impedance of seven inductors and five capacitors were measured from 500 kHz to 110 mHz on an RF vector impedance meter. The results were tabulated and are found in Table 1.

Table 1 reveals the impedance versus frequency of seven selected inductors/chokes. The impedance is given in ohms and phase angle for each frequency.

Note first the impedance of choke A versus frequency. This type of choke is often assumed to minimize RFI, however it has an impedance of only 14 ohms at

1 mHz and 185 ohms at 100 mHz. By taking one of the ferrite beads used to make up choke A, and winding 2 turns of #32 magnet wire through it (Refer to choke B), impedance over the measured range was nearly 3 times higher than choke A. This type of choke is useful to several hundred megahertz and is easily constructed.

Choke C is constructed in the same manner as choke B, but uses a larger ferrite bead with 7 turns of #32 magnet wire through it. Its impedance at 1 mHz is nearly 10 times that of choke B. Maximum impedance of this choke is at 25 mHz, at parallel resonance. This choke provides useful performance to several hundred magahertz.

Choke D is a commercially available ferrite choke, useful out to several hundred mHz. Its construction is similar to chokes A and B. Choke E is a tri-filar wound choke on a toroid ferrite core. This choke features good broadband performance to differential mode RF. A choke wound in this manner is much more immune to AC field pickup than the other chokes discussed so far.

A bi-filar wound pot core (choke F) exhibits a high impedance at AM broadcast frequencies (56k Ohms at 1.2 mHz). Being bi-filar wound makes this choke useful on balanced audio lines. This choke is mainly effective below 10 mHz and has several series resonant modes above 15 mHz.

An effective choke for FM broadcast frequencies is choke G, exhibiting an impedance of 18k Ohms at 110 mHz. However, at AM broadcast frequencies, this choke has the same impedance as choke A.

After studying the table it is apparent that some chokes are better than others for a given frequency. Table 2 gives the frequency versus impedance values for some small value capacitors frequently used for RF bypassing.

The first capacitor is a 390 pF mica, 100 volt (Capacitor A). One can see that it is not too effective for use at 1 mHz, but a series resonant frequency of 75 mHz makes it useful as a shunt element in VHF RF filter. Capacitor B is series resonant at 55 mHz and is a good choice for low VHF frequency bypassing. Capacitor C, although 10 times as large a value as B, is not as good a bypass capacitor at 100 mHz. Its impedance at that frequency is an inductive 8.4 ohms compared 4.6 ohms inductive for capacitor B. At 1 mHz capacitor C however exhibits an impedance more than 10 times lower than capacitor B.

Construction of a capacitor can also affect its impedance at high frequencies. Capacitor C and D both have a value of 0.01 uF and are both radial type constructions. Capacitor C is a polycarbonate and capacitor D is a ceramic disc. In this case the ceramic capacitor has slightly better characteristics at VHF.

Capacitor E is 10 times the value of capacitors C and D, and at 1 mHz exhibits roughly 10 times lower impedance than C or D. It has the lowest series resonant frequency of any of the capacitors tested and is an effective bypass up to about 50 mHz.

Earlier in the text it was recommended that feedthrough capacitors be used in applications that require effective RF filtering out to VHF. From the data presented in Table 2, it was shown that all of the capacitors tested lose effectiveness beyond about 80 mHz. A good 1000 pF feedthrough capacitor however, may measure 2 ohms -85° at 100 mHz and be an effective bypass to 1 gHz. Refer to some applications of feedthrough capacitors discussed earlier in the text.

In our discussion of controlling conducted RFI we have looked at some simple filter networks and performance of some capacitors and chokes at various frequencies.

By integrating the data from Tables 1 and 2 to the examples of filter networks presented, one can derive an effective RF filter for most frequencies.

For example, a piece of equipment is operated in a strong channel 2 (51.25 - 55.75 mHz) RF environment, and the station engineer wishes to reduce conducted RF into that equipment. A good choice of components for a two element filter (Fig. C), might be choke C listed in Table 1 (series element), and capacitor B listed in Table 2 (shunt element).

These components were chosen due to the choke (C), having a high impedance at channel 2, and the capacitor (B) being nearly series resonant (A short) at that frequency. Another example would be a component choice for RF filtering at FM frequencies. A good choice for a series element would be choke G (table 1), and a shunt element capacitor A (Table 2). In this case an even more effective shunt element would be the feedthrough capacitor described in Figure I.

It is beyond the scope of this paper to cover all aspects of controlling RFI, either conducted or radiated. This paper was presented to show some concepts of RFI elimination which have not often been discussed.

At this point a summary is in order.

1. There are two modes of RFI to equipment, conducted and radiated, and in most cases the interference caused by both can be minimized.

2. We looked at some pictorial examples of equipment which have integral RF shielding or filter networks. These pictures, it is hoped, will help and engineer implement ideas in the station equipment.

3. Conducted RF can be controlled with several basic filter networks.

4. Components do not behave at RF frequencies as reactance formulas would derive. At RF, components exhibit resonances and can act as capacitive or inductive elements.

5. Referring to data in Table 1, it was shown that some chokes are nearly transparent (a short) to RF at some frequencies.

6. The data presented in Table 2 shows that the most effective capacitor to bypass a given frequency is a capacitor which series resonates as a shunt element at or near that given frequency.

7. When constructing RF filter networks, the shunt element should exhibit a low impedance. It is very important that the leads of the capacitor be as short as possible for best performance.

It is the authors sincere hope, that some or all of the information presented here will help solve the problems you may encounter with RFI.

TABLE I

Choke Impedance versus Frequency (in Ohms) *

FREQUENCY IN MEGAHERTZ	A		B		C		D		E		F		G	
.5	7	+85°	23	+85°	210	+80°	12	+85°	15	+90°	9.6k	+85°	7	+85°
1	14	+85°	48	+85°	400	+78°	24	+87°	26	+90°	54k	+15°	14	+85°
1.5	22	+80°	74	+85°	600	+73°	36	+87°	38	+90°	16k	-70°	21	+85°
2	29	+72°	100	+72°	780	+70°	48	+87°	50	+90°	11k	-80°	28	+87°
5	56	+55°	180	+70°	1.60k	+55°	125	+87°	130	+90°	3.30k	-90°	68	+88°
10	80	+47°	250	+35°	2.60k	+40°	270	+74°	280	+90°	1.45k	-90°	135	+90°
20	115	+38°	350	+33°	3.60k	+42°	490	+53°	700	+70°	260	+0°	275	+90°
30	137	+33°	410	+30°	4.05k	-43°	640	+40°	1.0k	+40°	1.80k	-85°	425	+90°
40	152	+30°	460	+26°	3.80k	-54°	770	+28°	1.15k	+38°	840	-90°	600	+90°
50	162	+26°	500	+22°	3.20k	-62°	870	+16°	1.30k	+28°	600	-90°	800	+90°
60	170	+24°	510	+20°	2.80k	-67°	940	+6°	1.43k	+18°	480	-90°	1.10k	+90°
70	175	+22°	540	+18°	2.45k	-70°	980	-6°	1.55k	+10°	400	-90°	1.50k	+90°
80	180	+20°	555	+15°	2.15k	-72°	960	-15°	1.6k	0°	340	-85°	2.10k	+90°
90	185	+20°	560	+14°	1.95k	-75°	960	-25°	1.65k	-12°	300	-90°	3.40k	+85°
100	185	+20°	560	+13°	1.75k	-75°	880	-30°	1.55k	-28°	260	-85°	6.00k	+80°
110	190	+20°	580	+12°	1.65k	-75°	840	-38°	1.50k	-25°	240	-85°	18k	+40°

KEY TO CHOKE DATA:

A - 3 - .125 dia. Ferrite Beads on #22 gage wire, Fair-Rite Type #2643000301

B - .125 dia. Ferrite Bead w/2T #32 wire,Fair-Rite Type #2643000301

C - 7 Turns #32 wire on .291 dia. Ferrite Bead, Fair-Rite Type #2643000801

D - Choke, Ferroxcube UK200-20/4B Ferrite Choke

E - 6 Turns #32 Tri-Filar wound on .500 dia. Torrid Ferrite Core,
Fair-Rite Type #5961001103

F - Pot Core Ferroxcube, 3B7 Core, 30 Turns Bi-Filar Wound

G - 2.2 uh molded choke, J.W. Miller #9250-222

*Measured on Hewlett Packard 4815A RF Vector Impedance Meter

TABLE 2

Capacitor Impedance Versus Frequency (in Ohms)*

FREQUENCY IN MEGAHERTZ	A		B		C		D		E	
.5	840	-90^{0}	370	-90^{0}	32	-90^{0}	36	-90^{0}	3.0	-90^{0}
1.0	420	-90^{0}	185	-90^{0}	16	-90^{0}	18	-85^{0}	1.6	-70^{0}
1.5	280	-90^{0}	125	-90^{0}	11	-90^{0}	13	-85^{0}	1.2	-58^{0}
2.0	210	-90^{0}	91	-90^{0}	8	-90^{0}	10	-85^{0}	1.0	-45^{0}
5.0	81	-90^{0}	37	-90^{0}	3	-80^{0}	4.0	-75^{0}	1.0	$+0^{0}$
10.0	41	-90^{0}	18	-90^{0}	1.0	-50^{0}	2.0	-58^{0}	1.0	$+37^{0}$
20	19	-90^{0}	8.0	-90^{0}	1.0	$+54^{0}$	1.0	$+ 0^{0}$	1.0	$+60^{0}$
30	12	-90^{0}	4.4	-85^{0}	2.2	$+70^{0}$	1.4	$+45^{0}$	2.2	$+70^{0}$
40	7.5	-90^{0}	2.2	-75^{0}	3.0	$+78^{0}$	2.0	$+60^{0}$	3.0	$+72^{0}$
50	4.8	-80^{0}	1.0	-40^{0}	4.0	$+82^{0}$	2.8	$+70^{0}$	3.6	$+80^{0}$
60	2.6	-70^{0}	1.0	$+45^{0}$	5.0	$+82^{0}$	3.4	$+75^{0}$	4.2	$+80^{0}$
70	1.2	-35^{0}	2.0	$+70^{0}$	5.8	$+84^{0}$	4.0	$+80^{0}$	5.0	$+80^{0}$
80	1.0	$+37^{0}$	3.0	$+80^{0}$	6.7	$+86^{0}$	4.8	$+80^{0}$	5.6	$+84^{0}$
90	2.2	$+70^{0}$	4.0	$+80^{0}$	7.7	$+86^{0}$	5.5	$+80^{0}$	6.4	$+84^{0}$
100	3.2	$+78^{0}$	4.6	$+85^{0}$	8.4	$+90^{0}$	6.0	$+82^{0}$	7.0	$+84^{0}$
110	4.4	$+82^{0}$	5.4	$+85^{0}$	9.5	$+90^{0}$	6.7	$+85^{0}$	7.6	$+86^{0}$

KEY TO CAPACITOR DATA:

A - 390 pF Mica, 100 Vdc

B - .001 uF, Ceramic Disc, 1 kVdc

C - .01 uF polycarbonate, Radial, 50 Vdc

D - .01 uF Ceramic Disc, 25 Vdc

E - .1 Mylar, Radial , 50 Vdc

*Measured on Hewlett Packard 4815A RF Vector Impedance Meter

Figure 1

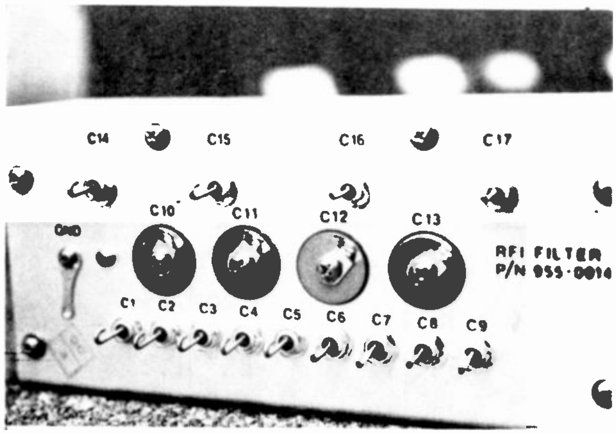


Figure 2

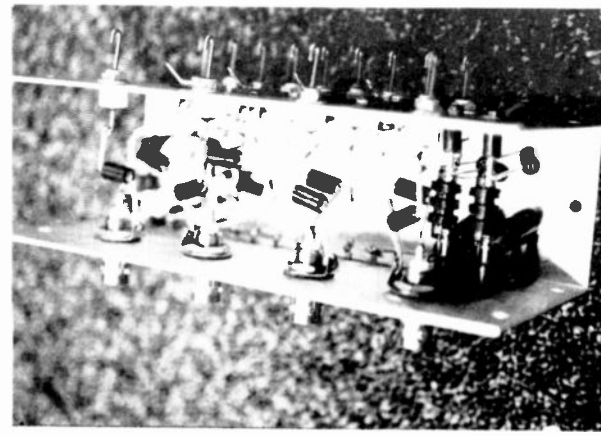


Figure 1 and 2 show the RFI filter assembly from the Broadcast Electronics FX-30 Exciter. All audio and control leads enter the exciter through this assembly. Each lead is filtered by a multi-section filter, with a feedthrough capacitor as a shunt element. This assembly strips off RF energy from the leads before the RF reaches the main cavity of the exciter chassis.

Figure 3

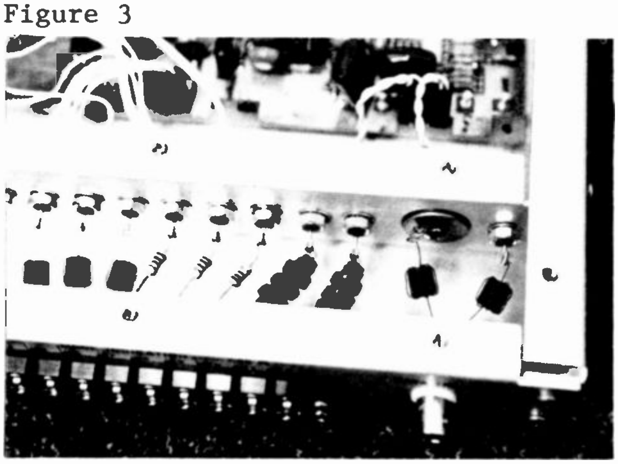
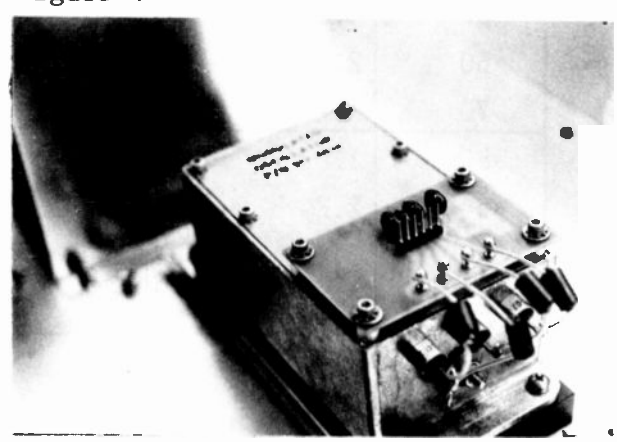


Figure 3 shows a similar filter arrangement on the Broadcast Electronics FC-30 SCA Generator.

Figure 4



The modulated oscillator assembly in the FX-30 exciter has a "PI" section filter for each lead going into the unit. All leads finally pass through the feedthrough capacitor before entering the chassis.

Figure 5

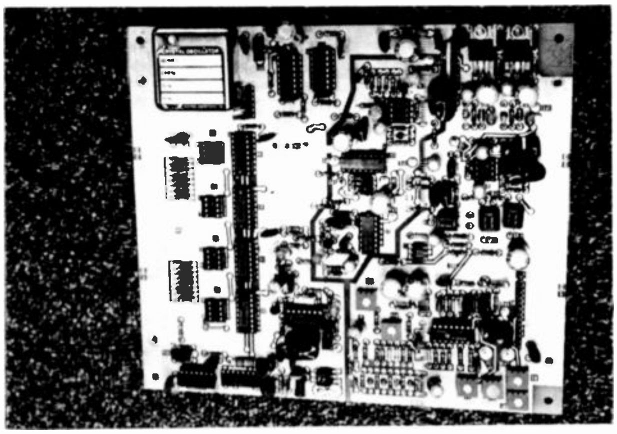


Figure 5 shows extensive ground plane used on this PC assembly from the FX-30. Ground plane on a PC board can add excellent shielding and grounding qualities to a design. Note the ground plane is broken in several places to eliminate ground loops.

Figure 6

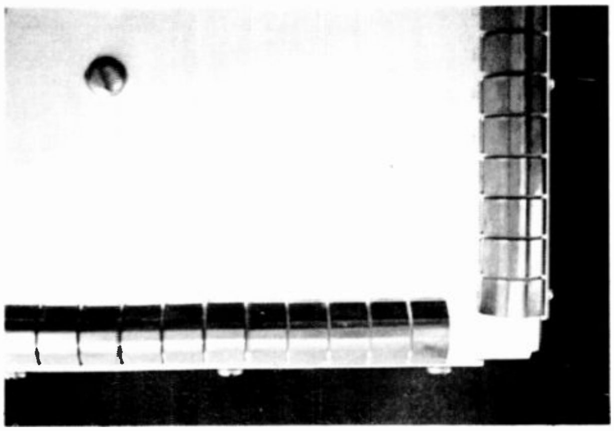
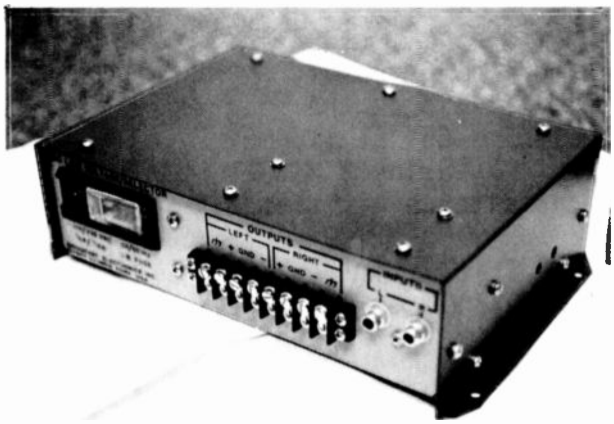


Figure 6 shows fingerstock used on the FM-30 transmitter to provide a positive RF connection betweek the door and the chassis.

Figure 7



Broadcast Electronics EP-1 Phono Preamp shows extensive RF shielding to minimize both conducted and radiated interference.

Figure 8

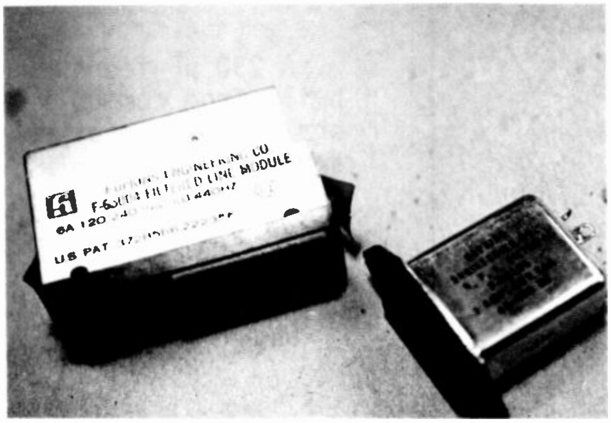
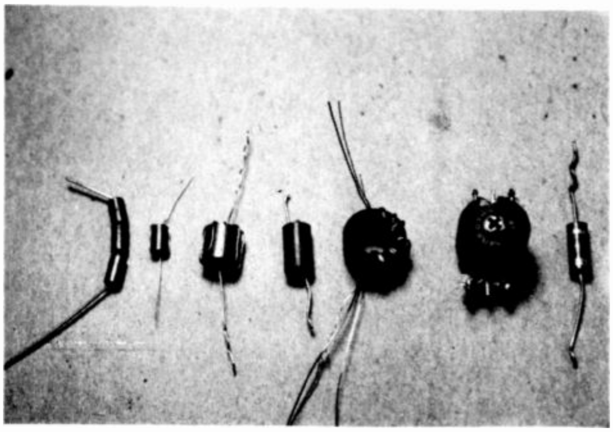


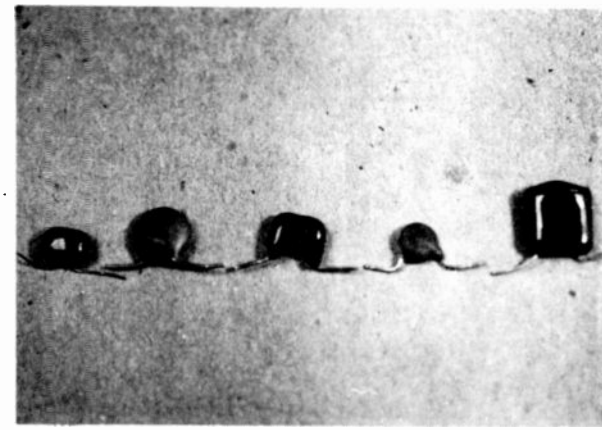
Figure 8 shows two commercial power line RF filters which provide good RF attenuation to VHF.

Figure 9



Seven chokes listed in Table 1.

Figure 10

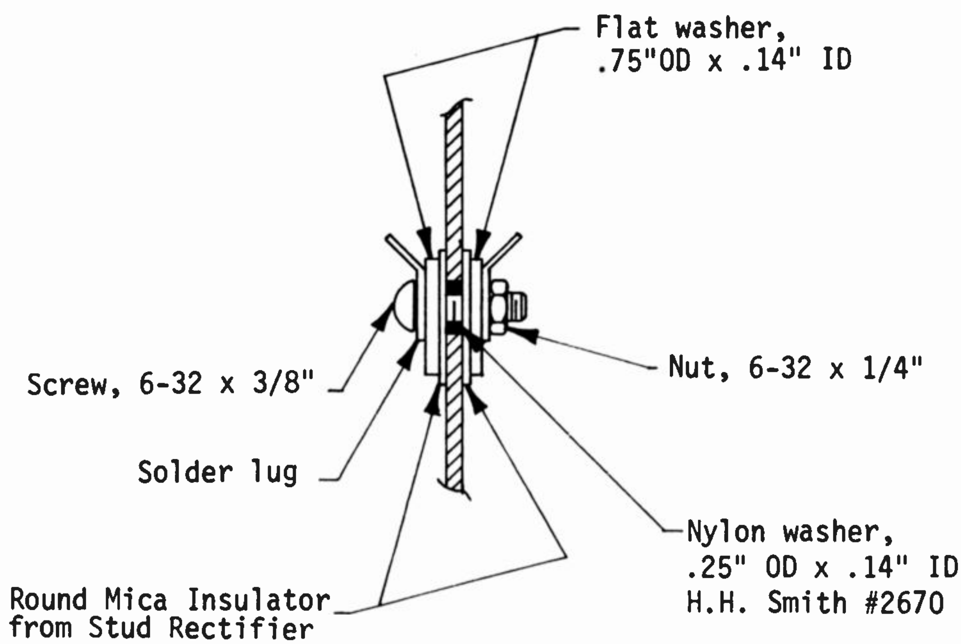


Five capacitors listed in Table 2.

Figure I

CONSTRUCTION OF AN EASY TO BUILD FEEDTHROUGH CAPACITOR

A 230 pF feedthrough capacitor may be easily built from commonly available materials as the diagram shows. The prototype measured 1 Ohm -65^{o} at 110 mHz and should be useful to 500 mHz. By using 2 Mica insulators on each side, the capacitance will roughly halve.



STATUS OF FREQUENCY COORDINATION IN BOSTON

ROSS B.KAUFFMAN

WCVB-TV

BOSTON,MASSACHUETTS

In 1976 the first attempt to establish a broadcast auxiliary coordinating procedure in the Boston area was undetaken.The project was initiated be the local Society of Broadcast Engineers chapter.The SBE was looking for a meaningful project that it could undertake to serve the industry.I volunteered at that time to act as coordinator.

This first attempt to create a coordinating procedure was made when Electronic News Gathering was just beginning in the Boston area.Use of the auxiliary frequencies was moderate and as such there was no pressing need to establish a coordinating procedure.It was never thought at the time that this service would eventually become as important as it is today.

The immediate Boston area has 9 TV's and over 30 AM/FM's.In the New England area,distances between adjacent markets is relatively short. There are about 16 TV's and over 50 AM/FM's within a 50 mile radius of Boston.Providence Rhode Island is approximately 40 miles from Boston while the New Hampshire border is about 30 miles away.

Our initial attempt to set up a frequency use data base was aided by the local FCC Engineer in Charge,Vincent Kajunski.With his cooperation the local FCC office,dated microfiche files were given to SBE each time the FCC recieved updated copies from Washington.After deciphering the data in the microfiche owr file data base was compiled.It included all assignments of auxiliary serice in Massachusetts,Rhode Island and New Hampshire.Unfortunately we listed only licensee,location and frequency.Later data was limited to the Boston area because of the volume of data to be processed.To supplement this data questionnaires were sent to all stations in Massachuetts,Rhode Island and southern New Hampshire.About 300 were sent with about 50% returned.

The following information was requested on the form.

Station Call letters
Address
Telephone Number
Chief Engineer

Do you use or are you assigned:
Remote pickup
Aural (STL)
TV (STL,TSL,IC,RPU)
Wireless Microphone(s)

List all Frequencies authorized or applications pending.If point to point service give azimuth.

The questionnaire was far from complete.

We added the pertinent data from the questionnaire to the initial data base.A list was compiled from the data base and distributed to each licensee that returned the questionnaire.

The local FCC office refers most licensees looking for auxiliary frequencies to me. I send a current copy to anyone that inquires with the admonition that I could not vouch for its completeness or accuracy but advised them they must make the decision on which frequencies to apply for.

Over the years additional data was added to the data base primarily in the 450-455 MHz and TV microwave frequency ranges because of increased usage and requests.

Currently the data base is computerized and we can now supply selected data to thoes who request it.

From the very beginning no specific frequencies were recommend to a licensee.The sevice provided is informational and intended as guideline only.The licensee has to make the final decision on their choice of frequencies.Once the licensee made his choice I urge him to advise me of his choice so the data base can be updated.

I would judge our attempts at frequency coordination in Boston as mediocre. It takes much time to establish and keep the data base current. The best way to do that is to encourage wide voluntary participation by all broadcasters who continually supply information on changes in their use of auxiliary frequencies.

The ideal data base in my opinion would contain the following information:

Licensee
Address
Phone
Chief Engineer

For each frequency used:
Frequency
Type of service

If point to point	Geographical coordinates of both points Brief description of both points
Control Point & Phone	Name of person who has authority to take action in emergencies
Times & length of time	facility in use

Antenna gain/size
Polarization
Azimuth
Auxiliary call

It is also helpful if the location of the ENG receive sites are included in the data along with the principal ENG receive frequency.

In the future it might be possible to establish a computer network which provides dial up acccess to the frequency coordination data base as some have suggested.

In summary frequency coordination data is difficult to establish and a monumental task to keep current.Only by the participation of every broadcaster and/or user in the program is it possible to make the undertaking meaningful. With the advent of more services vying for the same limited frequency sprectrum we will all need to communicate to avaert loss of programming due to interference in the auxiliary services

AM MODULATION MONITORS -

IMPACT OF IMPENDING FCC ACTIONS

Joseph Wu

TFT, Inc.

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In the FCC Docket No. 81-698, the Commission is proposing to delete requirements for type approval of modulation monitors for all broadcast stations. We, at TFT, believe that the adaptation of this proposal is desirable because it not only will give broadcasters new freedoms in the technical operations of their stations, but also gives manufacturers new opportunities to innovate and bring new cost effective products to the broadcasters by using state-of-the-art technology.

The development and documentation of a product with the submissions required for FCC type approval is very expensive for manufacturers. It can raise the selling price by 10 to 20 percent. If these extra costs do not produce benefits for the end users, then they should be removed and the savings passed on to the end users.

Different Modulation Monitors for Different Markets

We see that there will be a few changes in the AM monitors after the elimination of FCC type approval of audio modulation monitors. In general, there will be two major types of AM monitors: one to be designed for the small market stations and another to be designed for highly competitive major market stations. Let's examine the features and requirements of both types.

For small market operations, we believe cost is a major factor. An AM monitor using bargraph and peak lights to indicate average and peak modulation is adequate. These bargraphs should have user adjustable rise and decay time to suit the type of program signals broadcast by various station program formats. Stations in small markets usually employ consulting engineers who have test equipment to do proof-of-performance measurements and troubleshooting. A monitor just to show the modulation levels and alert the operator on duty when there is absence of modulation, over or under modulation and the loss of modulation, will be needed by these stations. Such a monitor can be manufactured at approximately 70% cost of the current FCC type approved AM monitor.

For the highly competitive market, a more sophisticated AM monitor will be needed. Audio processing can do strange things to the modulation. An instrument that can do routine monitoring, as well as performing analytical tasks, will be convenient for these stations. An oscilloscope presentation of the signals, both in time domain and frequency domain, will be extremely desirable. Storage capability is also desirable so that the engineers can do a closer analysis of the peaks and valleys of his modulation, even on the transient type of signals caused by interference, overshots, multipath and others. With a frequency domain presentation on the monitor, the engineers can locate noise sources in the audio paths where hum and intermodulation products are generated.

The modulation monitor with scope presentation appears to be expensive, but with today's technology, LSI and decreasing cost of CRTs, it can be manufactured at a cost within the reach of the major market AM broadcasters. We projected such a monitor will be in the four to six thousand dollar range.

On Peak Modulation Indicators

One of the most frequent doubts that an engineer has on his modulation monitor is "Is my peak modulation light telling the truth?" The answer for this question is "not always so" in a FCC typed approved monitor, especially when it is used in the off-air mode in conjunction with an RF amplifier. The peak lights meeting FCC type approval specifications (using tone burst testing) are too fast and non-discriminatory to noise peaks. Using a monitor with storage scope presentation will allow the engineers to analyze the modulation peaks and determine if the peaks are from his transmitter or from other sources.

An intelligent peak modulation light that only responds to the peaks of the program signals and ignors the transcient and noise is badly needed in an AM modulation monitor. TFT has developed a simple circuit that can detect the duration of modulation peaks. By calibrating the window width of this detector at the transmitter site, the peak light can discriminate 80 tc 90% of the externally generated peaks such as overshots, multipath, and ambient noises.

Off-Air Monitoring

AM stations operating by remote control have to use RF pre-selector to pick up their signal for the monitors. A single channel pre-selector is usually adequate for a small market station. In a highly competitive market, a frequency synthesized tunable pre-selector will allow stations to compare their program loudness against their competitors. A tunable monitor will also enable a consulting engineer to service several stations with one monitor.

The phase noise of the local oscillator in either the single channel or the tunable pre-selector will have to be kept low in order to meet the AM stereo performance requirements. The IF and RF filters must be of linear phase design to minimize overshots due to audio processing.

On AM Stereo

The obstacle of choosing a single standard for AM stereo is not a technical issue, but a financial one. Because of the "winner take all" contest for the millions of dollars in potential royalties from the AM stereo, it is too lucrative for the losers not to contest any decision made by anybody, including the FCC. One possible solution is that the five AM stereo proponents agree on a "profit-sharing" approach to adopting an AM stereo standard. After all, all five proponents have invested thousands of dollars in AM stereo development; even the losers should be compensated for their contributions. Today's high cost of money does not allow a manufacturer to invest in the development of products with a high degree of uncertainty. The market place decision allows the FCC to wash their hands off in making a decision, but it will certainly kill AM stereo in the long run. I hope the "profit-sharing" idea triggers some interest to the five proponents.

As far as the AM monitors are concerned, the monitor's manufacturers have done their homework, but no one can afford to introduce five different kinds of AM stereo monitors and hope one will be standardized later. Most of us will wait until the battles are over and the standard is adopted before introducing an AM Stereo Monitor.

The design approach of the AM monitor for future stereo applications is that the carrier and L+R signals should be brought to the rear panel. These signals can then be processed by an add-on AM stereo demodulator designed for one of the five proposed AM stereo standards. The add-on unit should perform the monitoring functions similar to the FM stereo monitors in the FM stereo broadcasting. Additionally, the AM Stereo Monitor can be part of the test equipment for making AM stereo proofs.

ATS Operations

Although ATS operation for non-directional AM stations has been authorized by the Commission since 1977, it has not become a popular mode of operation for AM broadcasters because the Rules requiring mandatory termination of the station's transmission if any one of the ATS Contro-lers, alarm circuits, input samples and mode switching time clocks fail and the problems are not corrected within 3 minutes. This requirement creates more problems than the ATS tries to solve. Unless the automatic shut-down requirement is removed from the ATS operation rules, it will remain unpopular.

Conclusion

The impending FCC rule makings on modulation monitors will bring about many changes in audio modulation monitoring and testing equipment. The de-regulation will bring cost effective products to the broadcasters due to more intensive competition among manufacturers. The features of the modulation monitors described in this paper can definitely be applied to FM and TV monitors too. The innovative manufacturers, who have done their homework, will see a higher return on their investment due to the fact that their products are designed to meet customers' exact needs: gain maximum coverage from his broadcasted signal, perform analytical and troubleshooting tasks, and operating within the FCC.

SBE CERTIFICATION - 1982

Jim Wulliman

WTMJ, Inc.

Milwaukee, Wisconsin

The SBE Certification Program has completed its first five year period of operation and is now processing renewals from the original group of Senior Broadcast Engineers who are certified.

Over 1750 broadcast engineers have been certified under the program.

The SBE Certification Program can trace its roots to informal discussions by some members of the NAB Engineering Advisory Committee during the late 1960's.

They suggested that the SBE propose to the FCC that they either upgrade the "First Class Operator" license or add an endorsement to the First Class license. The higher grade license would be required to be a Chief Engineer of a station. It would require experience at the operator level and passing a suitable examination.

The SBE Board recognized at that time that the First Class Operator license had become seriously weakened. A poll of the SBE membership in 1972 indicated that the majority was in favor of upgrading the First Class Operator license.

In April 1975, the SBE presented a plan for a voluntary Certification Program at the NAB Convention.

Consideration was given to three levels of certification - an associate level that was essentially an entry level - a broadcast engineer level for those who were actively employed as a broadcast engineer for at least five years - and a third level for the senior broadcast engineer who had been actively employed in a responsible position for at least ten years.

The Certification Committee prepared examinations representative of broadcast experience for the two latter categories. With the proper pre-requisite qualifications and a passing grade on the appropriate exam, the applicant would be awarded a certificate indicating his or her field of experience (AM/FM or TV). The associate level was dropped.

The certificate was to be valid for a period of five years and could be maintained by continuing education, employment, and active participation in the professional growth of broadcast engineering.

At least 25 professional credits were required for senior broadcast engineers and 20 professional credits for broadcast engineers to renew their certificates.

At the outset of the program, a limited "Grandfathering" provision was available for one six-month period. The senior broadcast engineering certificate was awarded to those applicants who met the guidelines governing the grandfathering clause. After the initial six month period only those having the required length of service and who passed the appropriate examination could be certified.

The recent FCC action, dropping the Radiotelephone First Class license, has prompted broadcasters and broadcast associations to suggest that the SBE reconsider the entry level certification program. This would provide a means of attracting technical students to train for entry into the broadcast industry. There would also be incentive for them to grow with broadcast technology by earning professional credits leading to the higher levels of certification.

The following is an outline of the entry level certification program as proposed by the SBE Certification Committee:

Certificate Name - Broadcast Technologist.

Examinations - Computer selected exams. Several hundred questions will be stored in a computer. The computer will randomly select 50 questions for each exam. Each exam will be different.

The entry level exam will be a General Exam consisting of questions on:

1. Electronic fundamentals.
2. FCC rules pertaining to operating tolerances.
3. Safety.

Exams would be given by:

1. SBE Chapters, proctored by certified SBE members.
2. Certified persons in areas where there is no SBE Chapter.
3. Professors of technical schools.
4. Chief Engineers at nearby stations where no other arrangements can be made.

How often will the exams be given?

Initially, bi-monthly. Later, quarterly.

Citizenship.

No citizenship requirement. Canadians are certified under our present program, as well as a few people overseas.

The exams will be given in English only.

Exams could be given overseas through technical schools or colleges. The applicant would have to provide a list of suitable schools for the Certification Committee to contact.

We will not publish a question and answer manual. However, sample questions will be included with the information packet.

The Broadcast Technologist certificate will be issued for a period of five years. It can be obtained in either of two ways:

1. By holders of a First Class license with two to three years of satisfactory service, OR
2. By passing the entry level examination. No experience requirement.

At the end of the five year period the Broadcast Technologist can renew his certificate by meeting a satisfactory service requirement, or the Broadcast Technologist can move up to the Broadcast Engineer level by passing a Broadcast Engineer examination in radio or television.

Fees.

It is the desire of the Certification Committee to hold the fees down to about the present $30.00 level if outside support can be obtained to meet the increased budget requirements of the expanded program.

Budget.

The certification Committee has developed a budget plan which will require support from NAB and/or other broadcast organizations to meet the start-up expenses of the expanded program. Our program will seek support for one-half of the first year expenses, decreasing each year until the program is self-supporting.

At the Annual Meeting, the SBE Board voted to absorb the added cost for the entry level program until December 31, 1982. The initial fee for Broadcast Technologist certification will be $10.00 for both SBE members and non members.

The Certification Committee is well aware of the shortage of qualified broadcast technical personnel, and intends to address this problem in two ways:

1. Prepare a list of colleges, Junior colleges, and technical schools which train people for broadcast related technical employment. We will need the help of the state broadcasters associations and SBE Chapters to assemble this information.
2. The Certification Committee will form the nucleus of an all industry committee to develop a suggested curriculum for training people to enter the broadcast technical field. This committee plans to utilize a program of instruction, which has been developed in Canada, as a starting point. The Canadian broadcasters have offered their cooperation.

To accomplish all of this, there will need to be a promotion campaign through NAB, state broadcast associations, and other industry associations, to reach high school and technical school students and guidance counselors. We must show them that broadcast technology offers not only an entry level job but also professional growth through an industry recognized Certification Program.

The SBE Certification Program offers the best method of identifying those broadcast engineers who are truly qualified, and also pointing the way for those who are entering the field, through the progressive levels of realistic certification from Broadcast Technologist to Senior Broadcast Engineer.

THE SOUTHERN CALIFORNIA FREQUENCY COORDINATING COMMITTEE

HISTORY AND PROGRESS

Richard A. Rudman

Group W KFWB

Los Angeles, California

EARLY HISTORY

The SCFCC was born in October, 1976 when CBS hosted a special meeting at Television City. CBS had hosted similar meetings in other markets where they operated. All stations were invited to send representatives to discuss recently announced revisions to Part 74, scheduled to take effect over the next two years.

The 450-451 and 455-456 mHz. remote pickup channels had been completely redesignated. Two channels were eliminated, many channels were split, and permissible uses for all channels were changed.

Before the FCC announced the 1976 plan, there were only ten channels in each 1 mHz. segment, for a total of twenty channels. With the 1976 Rulemaking, the FCC created five separate channel groups for specific uses, as follows:

Channel Designation	Total No. of Channels	Channel Bandwidth	Peak FM Deviation
N1	12	50 kHz.	10 kHz.
N2	24	25	5
R	10	50	10
S	2	100	35
P	8	10	1.5

N1 and N2 channels were to be used for broadcast program material, cues and orders necessary to that program material, and operational communications. Microwave path setup and dispatch functions fall under this last category. R and S channels were intended for program material, and cues and orders necessary to implement that programming. P channels were intended to be used for operational needs such as signaling, and telemetry.

If a station happened to be licensed for either 450.950 or 455.950, they would be allowed two years to vacate. This was necessary to accomodate the newly created S channels. They would be able to file for replacement channels in a simple manner. The SCFCC took root rapidly due to some simple realities that exist in our region.

These realities are terrain, user density and propagation anomalies. They explain why we felt a strong need to organize local frequency coordination at that time. The lack of some or all of these factors in other markets explain why our group acquired life, purpose, and momentum, while others did not.

TERRAIN

Many of the key mountain locations in Southern California are able to see one another, at least on clear days. There are more than 24 two-way sites scattered throughout the region. While many are shielded from each other, many are in line of sight relationship. The following is a partial list of key sites with this problem:

Name	Location	Height	Comments
Mt. Wilson	Los Angeles	5710 AMSL	One of the highest, most famous sites in the region.
Mt. Lukens	Los Angeles	5070 AMSL	While not as high as Mt. Wilson, is in line of sight relationship to several San Diego sites.
Oat Mountain	Los Angeles	3747 AMSL	North of San Fernando Valley. Can talk into and hear well south into Orange County and Northern San Diego County.
Mt. San Miguel	San Diego	2565 AMSL	Key San Diego site.
Otay Mountain	San Diego	3572 AMSL	Key San Diego site with good coverage in Southern Coastal Orange County.
Santiago Peak	Orange County	5687 AMSL	Orange County site that can talk and listen well in a four county region from Los Angeles south to San Diego, and east into Riverside County.

Terrain presents another problem to broadcast operations. Repeaters are necessary to carry out communications between studios and mobile units when line of sight conditions do not exist.

USER DENSITY

It is possible to tune in almost seventy AM and FM stations in the Southern California region. There is television activity on more than twenty VHF and UHF channels. A high percentage of these stations need frequencies for activities governed by Part 74. Many of these activities must be carried out in 450 mHz. channels. Other channels available to broadcasters were either being used to capacity already, or were unusable due to many types of interference.

PROPAGATION ANOMALIES

These include ducting, both air and over water; reflection due to inversion layers; and obstacle gain (knife edge refraction). The FCC has an ongoing study of these conditions in our region for both land mobile and broadcast portions of the spectrum.

All of these elements combined placed our region at least five years ahead of many others regarding frequency coordination. When we attended that first meeting, hosted by CBS, setting up a coordinating committee was viewed as a prudent, necessary self-defense measure. We felt the additional thirty six 450 mHz. channels added by the FCC would not be enough to fulfill our region's special needs without careful coordination.

THE WORK BEGINS

I was flattered when Bernie Koval, Chief Engineer for KFI at the time, suggested I become Chairman. I accepted the job with the provision when anyone else was willing to attempt it, I would step aside.

The story of what the SCFCC accomplished in its early days must include a listing of those who attended meetings, participated in discussions, and contributed their engineering experience.

Our organization and meetings soon acquired a tone of "structured informality". Our first task was the gathering of an accurate data base. Ed Edison and Ralph Grover, of Hammett and Edison, were invited to an early meeting held November 10, 1976. We heard confirmation that we could not rely on the FCC data base. The coordination efforts suggested for us by Ed and Ralph would require a lot of hard work.

We invited a firm that specialized in coordination to talk to us during another meeting held in November, 1976. We wanted to examine whether or not coordination could be computerized. We concluded it could not.

After considerable discussion, we decided to create our own data base. We conducted a thorough survey of existing users of 450 mHz. channels. We attempted also to ascertain what future usage patterns might be like.

We learned some valuable lessons at this point. Among these were some stations did not really know what frequencies they were using, some did not understand the FCC's plan for 450 mHz., and some were not planning for their long range needs.

The engineers who attended the early meetings realized the FCC's 1976 revisions, while helpful, were not adequate for our region. As licensees tried to adapt to the Part 74 changes, we realized the need for general guidance and organization. There was agreement that this should be done by forming a permanent committee with regular monthly meetings.

By mid-December, 1976 we had a preliminary data base showing current 450 mHz. activity. We compared it to the FCC data base and found many discrepancies.

During December we wrote to the NAB trying to establish a national liason. We believed frequency coordination was more than a regional problem. Many of our region's licensees were involved with operations in other cities. Their out of town crews often visited our area on short notice. CBS, NBC, and ABC felt a need to be licensed on common frequencies in all cities where they operated.

We found several conflicts in our region and were certain other regions must be experiencing similar problems. We soon discovered national support and interest was less than we had hoped for. Other regions were not experiencing the same level of problems in either day-to-day or special operations.

By the time of our December 16, 1976 meeting, a pattern had evolved for our meetings that has held to this day. We began by asking everyone present to introduce themselves. We exchanged information, comments, and changes to systems that were of general interest. All of this information went into our data base. We started our data base using forms designed by Bob O'Connor, of CBS. These forms showed existing channels and the ones added by the FCC. We revised our data base every month as we added new information. We wanted our listing to have a few basic items that would aid us in compiling a clear picture of actual and proposed activity, and hopefully resolve conflicts.

These items included the licensee's call letters, the two-way call, repeater locations, use of the frequency (s), a contact name and 24 hour telephone number, and mode of operation (simplex, repeater, mobile repeater, or mobile).

As each licensee described their system, questions were asked. By a process that might best be called "peer cooperation", compatible and sharing situations were outlined and agreed to in many cases. Some licensees relinquished channels in favor of others, ways were found to combine operations under the control of larger licensees, so these licensees required fewer channels.

CHANNEL SPLITTING

It was at the December 16th SCFCC meeting that the idea of splitting channels was first discussed. Discussion was underway regarding the newly created S chanhels, and how they were impractical in our region. Due largely to land mobile interference, high quality RPU stereo was quite risky. All of the "hi fi" remote pickup being done or contemplated at that time could be accomplished within an R or N1 channel. Bob Hess, then with KPBS, suggested we ask the FCC for permission to split the two S channels to create four more R channels for the region.

By the January 10, 1977 meeting the realization had dawned on everyone that there were more users than available channels, and some form of split frequency operation should be a high priority goal. It was at that meeting the idea to split the N1 channels was broached and several engineers present thought the concept was worth pursuing. We discussed many of the problems that would be associated with moving existing users if a waiver was granted. No one believed it would be an easy project if we decided to follow through on this concept.

By the February 2, 1977 meeting, we had compiled a first draft of the split channel plan. After talks with the FCC and the NAB, two issues emerged. The first was the suggestion that split channel operation would only be allowed by the FCC if both of the new split channels created from a single N1 channel be granted to the same licensee. The other issue centered on the fears that Land Mobile interests would use our request for splitting channels to counter broadcaster's statements that we must have wideband channels for RPU activities. Our discussions centered on how this last factor could be handled. By citing our plans for efficient spectrum use our plan would lessen the need to ask for more spectrum space. Discussion was also heard regarding the level of Land Mobile interference due to inadequate site management, poor or non-existent filtering of transmitters, and poor preventive maintenance on mobile units.

We decided the best approach because of valid concerns would be to seek a waiver for our region. Each station would apply for a waiver of Part 74.402 for their own needs. The first draft channel listing was the product of many hours of work by people who were familiar with the stations, the sites, and the various other factors making our region unique.

We concluded we might have to work around operations not yet on the air that could not be moved. We proposed that others who were on the air, or who were awaiting authority or equipment might also have to be worked around for many years, possibly on a permanent basis. Odd repeater splits had occured in the past. They would have to be reversed to put our complex puzzle together. New ones might have to be created out of necessity. As we proceeded, we heard more feedback from the NAB, our members, and other interests throughout the country. Our guidelines evolved from the following:

1. The NAB was very interested in our project, but would not endorse it at that time.

2. A basic philosophy embracing efficient spectrum use was thought to be highly desirable.

3. Any plan for split channel operation should reserve the right for full channel use on a coordinated basis.

4. Spectrum efficiency must be considered as essential to the plan.

5. There would have to be much more planning and discussion before we would be able to submit a plan that would have near unanimous backing.

6. We would have to satisfy the FCC our plan could be implemented with little economic impact on all those involved.

7. Our plan must never act to restrict any actual or potential licensee from access to spectrum space. There are legal penalties for any parties that act to restrain trade.

8. Our plan must have wide circulation to ensure all eligible licensees have a chance to actively participate.

9. Our plan would have to be regional in nature to succeed.

10. Under FCC Rules, exclusive RPU frequencies did not exist.

11. Rapid action was necessary to avoid problems created by licensees filing under the revised Rules due to immediate needs.

By our eighth meeting on March 2, 1977, we had established a clear direction for our plan. The SCFCC would make a written presentation to the FCC staff. The first part of that presentation would be a cover letter, supported by the proposed channel plan. It would present our case for a waiver of the Rules to allow channel splitting on a coordinated basis. The second part would be a collection of letters from as many actual and potential licensees as possible with their needs and comments on the plan. The last part of our presentation would be a paper defining the effect on lockout receivers on nearby adjacent and co-adjacent channel signals that might be encountered under split channel operation.[1]

The letter was received by the FCC on March 29, 1977. Over the next three years we had a dialog with the FCC staff regarding problems that had to be resolved before the FCC could reach a decision. Some of these related to the philosophy of split channel operation. Others related to how licensees would be able to work out conflicts.

The SCFCC received a letter from the FCC on January 8, 1979 from Jerold Jacobs, Acting Chief of the Broadcast Facilities Division. For the first time Mr. Jacobs told us that our request must be in the form of a Rules waiver. He outlined specifics on how we should request a waiver.

Each licensee was asked to submit a Form 313 with a request for waiver of Section 74.402 of the Rules. Referring to Section 0.281 of the Commission's Rules, Mr. Jacobs informed us these applications would have to be submitted "en banc" to the Commission for approval. He suggested each waiver make reference to the SCFCC's common justification for such a request. Mr. Jacobs asked for a detailed description of the operating practices of the SCFCC. This would assist the FCC Staff in their presentation of our case to the Commission.

Mr. Jacobs said in his letter Staff would recommend an experimental period of one year. He believed this experiment would help the FCC in any future considerations concerning split channel operation in other parts of the country, and the "appropriateness of initiating rulemaking looking toward a permanent split of certain remote pickup channels."

In February, 1979 we sent our response to that letter to the FCC. We gave an updated report on congestion problems in our region. We also described how our meetings worked to facilitate communications between licensees, mediate conflicts, and when needed constructively exert peer pressure.

We stated that since our first letter, we had tried to operate within the general framework of the allocation table we submitted with our 1977 proposal. The split frequency suggestions we had made were of course held in abeyance. Due to the worsening congestion licensees in our region were experiencing, we asked the Commission to take action as soon as possible.

During the 1979 NAB Convention, held in Dallas, the SCFCC participated in a meeting with several interests, including some Network engineering management people who had valid concerns about our plan. Resolution of these concerns, and FCC delays, took more than another year.

WAIVER TERMS

The SCFCC plan was approved by the FCC on June 25, 1980. In a letter signed by William Tricarico, Secretary of the FCC, the following points are of interest:

1. The FCC recognized the unique nature of our region and the problems it represented to broadcasters.

2. Acknowledgement was given to business conditions such as the rise in TV ENG activity that went beyond the provisions of the FCC's 1976 rewrite of Part 74 in 1976.

3. The SCFCC had been active since late 1976 to solve coordination problems, and our waiver suggestion was an important part of a carefully thought out plan.

4. Costs to any licensee involved in changes related to split channel operation would have been minimal.

5. The SCFCC plan was capable of yielding practical experience. This experience might help other parts of the country alleviate congestion problems.

6. The Commission, with that letter, granted our waiver request, contingent on a report on split channel operation we would submit to the FCC after one year of operation under the plan.

WAIVER IMPLEMENTATION - FIRST STEPS

The waiver implementation by the American Broadcasting Company was delayed due to the extent of their operations in our region. It was for that reason that ABC asked the Commission for an extention to the experimental period the FCC outlined in their June 25, 1980 letter. Budget considerations and extensive planning necessary also delayed National Public Radio's full use of a sophisticated repeater system to operate on a S channel split. This system is to involve a number of NPR stations who will use it to feed spot news into the NPR satellite distribution system, and also for the needs of numerous Southern California NPR outlets. The SCFCC's report on the waiver implementation must at this writing be termed "interim". No serious problems have been reported to the Committee by any licensee who has implemented split channel operation. Licensees operating on the center channel have not experienced difficulties of a serious nature to date. In January 1982, the SCFCC called for reports from the following licensees:

KABC-TV, Los Angeles	KCST, San Diego	KTLA, Los Angeles
KNBC, Los Angeles	KGB, San Diego	KRLA, Pasadena
KSDO, San Diego	KSBR, Mission Viejo	KUCI, Irvine
KHJ, Los Angeles	KGTV, San Diego	KFMB AM/TV, San Diego

Their comments are presented appended to this paper. Use of the center channels at some distance from the licensee ocupying the split channels has proven to be successful. This was the most critical application of the waiver and one that allowed us to accomodate several licensees who would have been forced into sharing situations which would have been detrimental to their operations. KUCI, in their report, mentions some interference. It should be noted they are experiencing less interference than before the waiver was implemented. At a recent meeting the SCFCC made several suggestions to KUCI that might lead to further reductions to interference they might experience. As time goes on and more licensees are accomodated, we expect new problems to develop for all licensees. We have given some thought to the solutions of these problems at many of our meetings since the waiver was approved. These problems involve interference, accomodating even more licensees, transient use by various entities, and changes in technology that will concern use of these frequencies.

INTERFERENCE

Types of interference causing problems today can be broken down into that from licensees on the same channel, interference from adjacent channel users, interference generated from improperly installed equipment, interference due to equipment malfunction, or any combination of these situations.

SAME CHANNEL INTERFERENCE

Prevention of this type of interference must rely on each licensee being aware of others sharing their channels, religious use of call letters by all licensees, and a willingness to check with other licensees before engaging in operations which might cause them interference. The SCFCC has been engaged in a continuing edcational campaign to ensure engineering and operations personnel are aware of all aspects of this problem. In the few instances where we have had problems, management personnel from stations involved have worked rapidly to correct these situations.

ADJACENT CHANNEL INTERFERENCE

The most common case of this type of interference arises when many users congregate at the same site, usually a news story. Actual program disruption is rare. Under certain circumstances, a transmitter operating at a news event can prevent consistent reception of repeater output by many licensees. Crews have learned to locate their trucks as far apart as possible under such circumstances. Most licensees have found this is a problem they must live with due to the nature of some story locations where trucks must all park in the same area. Many licensees have installed better filtering for their field equipment to minimize this problem.

Adjacent channel interference at fixed sites has been virtually eliminated except for cases where Subpart H operation places trucks using lower power IFB systems near fixed receivers. Cases of such interference have not been devastating according to reports received by the SCFCC. So far the parties involved have cooperated to limit such interference. Subpart H interference control does remain a high priority with the SCFCC.[2]

INTERFERENCE FROM IMPROPERLY INSTALLED EQUIPMENT

A few cases have come up in the past few years when either a licensee's transmitter was installed at a repeater site without using proper filtering methods, or a receiver was installed without using proper state-of-the-art protective measures. Broadcasters renting space from repeater site operators are occasionally subject to the same poor installation practices as we sometimes see in land mobile services which cause us out-of-band problems.

Education on proper installation techniques has helped. On-site help by brother engineers who attend committee meetings has in some cases been employed. It must be emphasized all parties involved in cases of this type of interference have been responsive, cooperative and patient.

INTERFERENCE DUE TO EQUIPMENT MALFUNCTION

This case is not limited to that caused by broadcast licensees. Many more problems come from outside the RPU band. Business radio, paging, government two-way services, and mobile telephone cause most of the interference problems the SCFCC deals with. We find such interference difficult to locate, terminate, and prevent. It is costly to licensees in terms in time involved in searching it out, and lost programming it causes. It is the greatest cause of interruptions to actual RPU programming.

Individual licensees can take steps to "harden" their receivers to exist in high levels of broadband energy that cause desensitization. They cannot guard against spurious products that actually appear within receiver passbands. The SCFCC and member licensees are still at the mercy of this type of interference and will remain so for some time.

TECHNOLOGICAL IMPROVEMENTS

By using modulation indexes approaching 1.0, it is possible to achieve audio bandwidth equal to or slightly greater than FM deviation in the 450 band for program delivery. Modifications to existing land mobile communications equipment are available that have enabled some licensees in our region to use N2 channels where they would have had to use N1 or R channels to achieve the same audio quality. Adding helical filters, high Q cavity filtering, and GasFET preamplifiers to stock receivers are proving to be workable solutions for some problems. Information on experiments in this direction is passed through the SCFCC so all can learn and hopefully benefit.

We await the introduction of a truly "bulletproof" receiver that uses wide dynamic range RF amplification, high level mixing, and other techniques to minimize susceptability to interference. We guarantee any manufacturer a proper test environment when such a receiver is available.

RECENT SCFCC RPU ACTIVITIES

The SCFCC continues to hear requests from new users for recommendations so they can begin operations. New networks, that go beyond users envisioned by our Committee, have been encouraged to join our membership. We have also encouraged cable system operators to attend our meetings.

SCFCC Bylaws have gone through two revisions and we believe they are near the stage where they can be adopted. These Bylaws are already being reviewed by other coordinating entities.

While we still think computerized coordination is not feasible for broadcast RPU bands, especially in our region, we are close to a computerized data base to make coordination "by hand" much easier. When this project is finished it will be possible to produce updated listings very rapidly. We are also looking into the possibility of making the data base available via a telephone Modem for any member who has a terminal.

We are working very closely with the San Francisco Bay coordination group, and have gathered data from Central California and Las Vegas licensees to assist in the compilation of what may become a Southwest regional coordination listing. These efforts should in turn mesh with other regions as frequency coordination comes of age within the broadcast community. Within a few years it should be possible to access a centralized source of RPU information for many parts of the country.

The SCFCC supports efforts by the Society of Broadcast Engineers and others who wish to establish coordination groups. We hope to be able to provide information like this report on subjects such as TV microwave coordination.

ORGANIZATIONAL CHANGES

During 1980 the SCFCC decided to create the office of Vice-Chairman to allow smooth flow of activities when the Chairman could not be present, and to assist in running the organization. The office of Communications Liason was devised to fill the need for an adminstrator. This position added much needed accountability to data gathering activities, mailing lists, and all communications to and from members.[3]

BYLAWS

We began to write a set of Bylaws in 1980. These Bylaws have since been revised twice and are now being reviewed by the legal departments of several member-licensees. We expect to adopt a set of Bylaws before the end of 1982. Many acting and forming coordinating groups have asked for our yet unapproved Bylaws so they can begin to organize their activities.

Several Bylaw issues are still unresolved. The first concerns voting. We have decided to base voting on licensees, rather than engineers who attend meetings, and to bar manufacturers and other associate members from voting priviledges. We have not yet decided how to resolve cases where an AM, FM, TV, and Network entity would all have collectively four votes to the one vote of a FM licensee.

There are very few issues that have required voting during our history, but we believe the issue of voting has to be addressed and answered. One proposal still being discussed at this writing would limit licensees to voting only on matters that concerned their section(s) of Part 74. Under this plan an AM or FM licensee would not vote on matters concerning TV-ENG.

Another issue concerns financing. We would like to avoid collecting dues and have expenses subsidized by continuing support by member licensees. We have discussed contracting publication of our monthly mailing list of the SCFCC Minutes and Newsletter to an outside firm who would sell subscriptions. The SCFCC data base might be published in the same manner. A subscription, including updates for a specific period of time would be sold to anyone who wanted access to this information. SCFCC members would indirectly subsidize most of the SCFCC's fixed expenses in this manner.

Our annual Christmas Party on Mt. Wilson in June would be a "pot luck" affair. To date it has been subsidized by KNXT-TV who has supplied both the meeting place and the food since our first such affair in 1980.

Another issue we are facing is how, or if, the SCFCC should make comments to the FCC. A formal procedure for making comments may be added to the proposed Bylaws.

OTHER SCFCC ACTIVITIES

The SCFCC was formed to coordinate 450 mHz. frequencies. This project began in November, 1976. We received FCC approval for the waiver for our region in June, 1980. While we were waiting for the FCC to rule on the waiver, we worked within our overall plan to be prepared if the ruling was favorable.

During this period we began to talk about coordination of VHF RPU activities, Aural STL, and TV microwave frequencies. Many of these activities are still in the planning stage at this writing. It seems to be axiomatic that frequency coordination efforts do not become concerted and serious until real problems manifest themselves.

AURAL STL AND TV MICROWAVE

Lack of 15 kHz. stereo telephone pairs to FM transmitter sites, the difficulty of achieving consistent phase relationships between those pairs that were available, and the tremendous inflation in cost of all leased telephone services increased demand for aural STL channels. AM stereo was looming on the horizon, leading many AM licensees to re-examine their STL needs.

Congestion in the 2 gHz. microwave band has also reached the critical point in Southern California. The SCFCC looks at both the Aural STL and the TV ENG microwave problems in a similar fashion. We can facilitate licensee-to-licensee communication that is essential to coordination. We can also make suggestions on equipment and operations standards that will help to solve many problems.

After the first landing of the Space Shuttle Columbia, the SCFCC became involved in coordination of broadcast frequencies at Edwards Air Force Base. We asked NASA and USAF frequency coordinators to attend our meetings, meet our members, and work with us to reduce interference to a minimum. We published NASA and USAF "ground rules" in our newsletter to inform both local broadcasters and network crews that would be coming into our region to cover this event.

The second shuttle landing had no instances of interference caused by broadcasters according to the USAF, NASA, and the FCC. We cooperated in a plan to check all equipment brought to Edwards for proper emissions. Both the FCC and the USAF had measurement trucks that checked all broadcast users. These precautions were necessary due to the extreme sensitivity of 2 gHz. microwave tracking receivers at Edwards, and the low power signals it must track from the spacecraft on approach for landing.

CONCLUSION

We foresee no "promised land" of coordination for any Part 74 section where every licensee in Southern California will have their private preserve of interference-free spectrum. We do see a time when we are able to achieve day-to-day operations with minimal interference.

FOOTNOTES

[1]General Electric, E.F. Johnson, Motorola, and RCA have all all supported our activities and participated in our meetings from the beginning. The paper referred to was made available by Motorola for our presentation to the FCC.

[2]Rental of wireless microphones by firms in our area has been and remains to be of concern to the SCFCC. We believe manufacturers of such equipment and rental agencies must be made more aware of local frequency coordination. Our efforts in this direction have not met with success to date.

IFB use under Subpart H is difficult to accomodate in our region. Licensees realize its use must be secondary, and on a non-interfering basis to actual program material.

[3]Donald Wilson was elected Vice-Chairman and still holds this position. He has been and continues to be one of the prime movers of the SCFCC. His analytical and thoughtful contributions are part of every important accomplishment of the SCFCC.

Howard Fine, now employed by KNXT, was elected Communications Laison in 1980 when the position was created. Howard has volunteered his time to the SCFCC to compile information. One of his many responsibilities is maintenance of our mailing list which has grown to over 450 names as of this writing.

WHAT DOES IT TAKE TO BUILD A WORKING FREQUENCY COORDINATING COMMITTEE?

Write a Letter to All Licensees in Region. Make every effort to get as many actual and potential licensees to your first meeting as possible. Do not overlook entities like CNN, cable systems, manufacturer's representatives, and managers of your local high buildings and hilltops that serve as common two-way locations.

Plan For Success. To assure success of your initial efforts, give thought to the politics of your locale. Should the SBE send the letter of invitation out on their letterhead, or should you ask a highly respected consultant?

Elements in Your Letter. Your letter should clearly define the mutual need for action and the immediate need to start a coordinating committee to deal with regional frequency problems.

Scope of Your Committee. Think and act on a regional level. Electromagnetic radiation does not respect political borders.

Key Benefit of Your Committee Activity. The most important activity your group can provide will be allowing the engineering personnel in your region to get to know one another. Everyone can then begin to build mutual respect. Start each meeting by having participants introduce themselves. Use name tags. Set a pattern for informal social interaction.

Sign Up List. Send one around early during each meeting so you will have a record of everyone who attended. Ask who will be willing to do Committee work.

Newsletter. Start a monthly newsletter immediately. Publish the sign up sheet, the minutes of your meetings, and new notes discussed during the meeting. Any information and correspondence that comes up between meetings can go in as well. Make sure you publish the agenda and date for the next meeting.

Conducting Meetings. Use Roberts Rules of Order as a guideline. Your meetings should be structured, but in an informal way.

Mailing List. Your mailing list is an important part of your data base. Include all actual or potential licensees, even if they do not attend meetings. Include cable operators, manufacturer's representatives, site managers, consultants, equipment dealers, Group and Headquarters personnel, and other frequency coordinating groups.

Alliances. If there are working coordinating groups for Amateur or Public Service Radio in your area, attempt to contact them and let them know you exist.

Leadership. Your group should decide on a Chairman and Vice-Chairman to run the meetings and provide liason and leadership for the mutual effort. Politics play a part here too. It is better to get everyone's concerns out in the open very early regarding your leadership.

Voting. There are few issues that require a formal vote. Your committee should really be a forum for information exchange. If voting is necessary, try to have the decision made on technical merits.

Conflict. For conflicts that cannot be resolved on technical merits, you will have to employ creativity. Split RPU frequency operation if approved for your region may be used to allow licensees to have more channels than otherwise possible. Do not overlook a lunch or dinner meeting to talk over concerns. Make sure neutral parties are present.

Costs. Try to get a few key local stations to finance your postage and duplication costs until you make a decision on financing. So far, the SCFCC has decided to avoid many problems associated with collecting dues. Some of your costs will be a meeting place if a free location cannot be obtained from a member, coffee and doughnuts for each meeting, and stationery supplies incidental to the newsletter and correspondence.

Minutes. Coerce or othewise enlist a responsible person to take careful notes at meetings, compile frequency information, and do other committee leg work. It seems to be a basic rule of committees that 10% of the members do 90% of the work. If you are one of this minority you should also be prepared to shoulder 90% of the blame when something goes wrong.

Goals. Pick one goal and see it through. Coordinating RPU, TV ENG, and Aural Stl activity in a few weeks is unlikely. If action is needed on all fronts, break the effort down into subcommittees.

Meeting Dates. Always set your next meeting date before the present meeting adjourns. Never skip months in the early years of formation. You will most likely find the need arises for meetings between the monthly regular meetings during the early months.

Perception of Committee Activities. Your committee will not be a police force. No one on it is a policeman. You will be most successful if you assume the role of facilitators. This just means you help communications between licensees to take place more easily.

Performance Measurement of UHF-TV

Broadcast Antenras by Helicopter

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Background

UHF TV broadcasters have been increasingly concerned about the performance of their transmitting antennas. This concern has been generated by a combination of energy costs and competitive pressures in the marketplace. With regard to the former, signal levels can be increased by increasing transmitter power but with high capital and operating costs. With regard to the latter, competitive pressures from other UHF stations--particularly newly constructed ones--and a desire to gain comparability with VHF stations, have caused stations to look at their transmission systems with a critical eye.

Some older transmitting antennas become suspect when new stations--having similar calculated effective radiated power--seem to do a better job of serving the market. Some stations have attempted to measure field strengths by conventional methods--such as the TASO prescribed approach--only to find that the resulting data is inconclusive with respect to making a determination regarding antenna performance. This is due primarily to terrain considerations (vegetation losses and reflections) which cause wide variations in received signal levels. The end result is that judgments are made by viewers and station staff on a relatively subjective basis by making "A/B" comparisons with other stations which are presumed to have a better signal.

This paper describes a method which can be used to assess the performance of medium and high gain UHF antennas, in situ, using a helicopter to obtain verifiable results.

Measurement of TV Broadcast Antennas

The manufacturers generally test completed broadcast antennas prior to shipment to a customer. We are all familiar with the generalities of range and turntable measurements. The antenna under test is usually mounted--with its axis horizontal--on an elevated turntable platform. A source radiator is located at some large distance (usually greater than one mile) and provides a uniform field in which the transmitting antenna under test can be used as a receiving antenna. This is a permissible procedure based on the reciprocity theorem.

The subject antenna can then be rotated about its longitudinal axis to get a horizontal pattern and can also be rotated in the horizontal plane to get vertical pattern "cuts." These cuts are usually made at equally spaced 30° azimuthal intervals and at specific points of interest (maxima and minima) in the case of directional antennas.

These data provide rather complete and accurate information regarding the shape of the horizontal and vertical patterns including such important characteristics as beam tilt, beam width, null locations and null-fill. But they do not provide for a direct measurement of antenna gain which, of course, is of primary concern to the broadcaster.

The gain of broadcast antennas is generally referenced to the gain of a dipole antenna; the dipole exhibits a maximum gain of 1.64 over an isotropic radiator. This gain can be calculated on a theoretical basis by considering such factors as effective aperture, null-filling, feed system losses, cross polarization losses, and pattern directivity or variations. These "losses" can then be deducted from the gain over a dipole by a -90° to +90° vertical pattern integration. Thus, it can be readily seen that the gain is not directly measured and its establishment is based on some assumptions.* One of the chief variables which make the calculation method superior to attempts at direct measurement is the variation in illuminating far-fields due to range reflections. (Direct measurement can also be made by near-field probing of elemental amplitude and phase distribution over the antenna aperture but this approach has its own set of error-generating problems.)

If one could eliminate reflections and achieve a near free-space propagation condition in the far-field, direct gain measurements could be made with a reasonable degree of accuracy.

Speaking to the subject of space measurements of ground based antennas, the Institute of Radio Engineers--in 1948--suggested in its publication Standards on Antennas: Methods of Testing that "... although it is possible that, in the future, the helicopter will be found more suitable, most measurements are now made with an airplane or blimp." This paper, written some 30 years later, discusses the practicalities--and problems--of making airborne measurements from a helicopter using techniques that minimize the potential sources of error.

*Assumptions include feed system losses, cross polarized components, radome losses, and scattering effects.

Airborne Measurements

The parameters of principal concern to the broadcaster include main lobe gain, horizontal pattern, beam tilt, beam width and null-fill. These antenna properties can be measured in situ by using helicopter-based instrumentation and procedures described herein.

As discussed earlier in this paper, one of the primary limitations on any measurement scheme is the presence of undesired reflections which cause errors in measured data. A TV antenna mounted on top of a tall tower with no nearby reflecting surfaces in its aperture would appear to present an excellent setting for direct measurement particularly if the measurements were made in close proximity of the antenna. Unfortunately, we have not as yet devised a method for placing a dimensionless isotropic antenna in the subject antenna's aperture.

A helicopter equipped with a receiving antenna makes a reasonable substitute provided that it is placed in the far-field of the antenna. The far-field measuring situation is required because the wave-front emerging from the transmitting antenna is approximately spherical and, therefore, the phase front across the receiving antenna will only be flat if the distance between the antennas is very large. However, at very large distances the problem of reflections from the ground will again be a problem. It is a generally accepted fact that the phase front in the so-called Fraunhofer region--or far-field--is reasonably flat to eliminate the measurement problems in the Fresnel region--near-field. It is also generally accepted that the far-field region begins at a distance defined by $2A^2/\lambda$ where A is the effective aperture of the antenna. For example, a calculation at Channel 14 for an antenna having a gain of approximately 15 dB would yield a far-field distance of 4,300 feet. Thus, one mile appears to be a reasonable distance for this particular measurement, and, in fact, is a good rule-of-thumb separation for most UHF TV situations. Caution must be exercised at shorter distances to consider the effects of radiation exposure to personnel. (At one mile from a one megawatt facility the power density in the main lobe is several times the "Russian" standard but below the proposed ANSI standard.)

The following figures depict the essential characteristics of the measurement set-up:

Figure 1 depicts a typical measurement condition

Figure 2 depicts the antenna installation on a helicopter

Figure 3 is a block diagram showing all equipment used and its interfacing

Figure 4 depicts the horizontal pattern of a typical receiving (measuring) antenna

Figure 5 depicts the vertical pattern (horizontally polarized) of the measuring antenna

In Figure 1, it should be noted that when measurements are made in the main beam area the reflection point is at a large depression angle (approximately 20° in this case). Generally, this should provide for reflected energy well below -20 dB relative to the main lobe (for medium or high gain antennas), assuming a reflection coefficient of 1.0 and far-field conditions. Since it is unlikely that either of these latter conditions can be met, the reflected energy can be assumed to be much less than -20 dB and, therefore, a negligible factor in the measurement accuracy. The geometry of the actual reflection situation must be studied on a case-by-case basis considering distance, helicopter elevation, intervening terrain and level of illumination of the reflection point by the transmitting antenna. This is particularly true when the lower regions of the vertical pattern are being probed with resultant low helicopter altitudes.

Figure 2 shows the boom used to support the measuring antenna. The length of the boom is determined by the vertical pattern of the receiving antenna and is calculated such that reflections from the helicopter main rotor will be about 20 dB below main lobe values. In this particular case the boom length is about ten feet.

Note in the block diagram of Figure 3 that altitude data is recorded simultaneously with the field strength data on a chart recorder. The altitude information can be merely pulse marks corresponding to major altitude divisions (such as 50 ft.) or can be continuous data generated from a radar altimeter. In the former case, the pulses are manually generated while observing a calibrated sensitive barometric altimeter. Altitude data is usually simultantously orally "called-off" and stored on an audio cassette.

Measurement Procedure

The station operator may wish to verify whether his horizontal pattern is omnidirectional (or has the desired directional characteristics) and may be tempted to measure this pattern in the horizontal plane of the main lobe at a relatively close distance. It seems simple enough to assume that flying a circular pattern with the helicopter at a known distance would permit this type of measurement. However, the experience with actual measurements indicate that this approach is impractical for the reasons listed below:

- The sector defining the maximum relative field (say $\pm 95\%$ relative field points) may be only 50 ft. deep at a distance of one mile from a high gain antenna. Maintenance of altitude with sufficient accuracy to assure that the helicopter is in the main beam is extremely difficult, if not impossible.

- Wind effects at altitude (rarely is a no-wind condition present) dictate that the helicopter tangential speed will not be constant which makes it difficult to determine azimuthal positions and correlate these data with field strength data.

- Wind effects and pilot techniques make the proper (and constant) orientation of the receiving antenna unlikely.

- Beam tilt may vary with azimuth and maintaining a constant altitude (if possible) would yield a result which indicates a horizontal pattern deficiency rather than the real nature of the problem.

The real-world case conditions would include sizeable wind components, turbulence and piloting errors which do not give this approach much chance of success. Thus, the procedure described below has been developed to eliminate or minimize the effects noted above.

The more practical approach is to make vertical cuts, similar to range measurements, which will reveal all essential characteristics of the vertical patterns and to make a sufficient number of these to have a good picture of the azimuth pattern. (A circular measurement can be made to spot any gross anomalies in the pattern between the points at which the cuts are made but, as noted, above this approach is not good enough for accurate measurements; azimuths needing further study can be identified.)

The procedure which has been successfully employed is described below:

1. Calculate the minimum distance to far-field.

2. Establish an appropriate radius for measurement after considering reflection conditions, nearby obstructions, and surface features which may affect the measurements or procedures. (For the purpose of this discussion the radius has been conveniently assumed to be one mile.)

3. Plot a circle--of the radius established above--around the subject tower/antenna on a topographic map. (Figure 6 shows a typical plot.)

4. Identify ground features (roads, buildings, rivers, etc.) which will be on or intersect the circle which can be used as visual reference points by the pilot. Select a sufficient number of these to achieve the desired azimuthal pattern data.

5. After equipping the helicopter as shown in Figures 2 and 3 and appropriate calibration tests and adjustments have been performed, the helicopter is flown to the measurement location. If possible, a landing is made adjacent to the tower to make final checks and to verify altimeter calibration.

6. Verify transmitter operating power. While slight power variations ($\pm 1\%$) will not affect the accuracy appreciably, all potential sources of error should be identified and parameters recorded.

7. Position helicopter over a measuring point at an altitude sufficient to allow probing of the pattern several degrees above horizontal plane. (Referring to the case depicted in Figure 1, this would probably be about 1,400 ft. AMSL.)

8. Start recorders and begin descent over point until low enough to obtain sufficient data to satisfy measurement objective. (400 ft.

AMSL will permit probing to approximately 8° below horizontal plane in this case.) Record altitude data.

9. Repeat for each pre-determined azimuthal direction.

10. Fly circular pattern in main beam to identify "holes" or other gross anomalies.

The resulting data from these measurements is contained on the strip charts. Figure 7 shows samples of the chart recordings from some actual field measurements. Data reduction is quite straightforward: calibration ticks (both amplitude and altitude) permit the development of vertical plane relative field pattern for each azimuth. Figure 8 shows typical helicopter data plotted over a vertical plane pattern made on the subject antenna at the manufacturer's test range.

Note that it is possible to determine beam tilt, beam width, null locations and null-fill. This sample pattern has been normalized, i.e., the maximum relative field has been assumed to be the same as the 100% relative field measured on the test range. Actual gain determination methodology is discussed in the next section.

Depending upon the purpose of the measurements and the type of antenna being measured, it may be necessary or desirable to take additional data to permit full antenna evaluation. This would include cross polarization (vertically polarized data) and high angle radiation (end fire) as some types of antennas are susceptible to these problems and, of course, these undesired radiation characteristics manifest themselves in the form of gain reductions.

Gain Determination

Theoretically, it should be possible to directly measure the radiated field in the main beam in absolute terms and, from this data, calculate the antenna gain. In practice, however, this is quite a difficult task as there are many potential sources of error including:

- transmitter operating power
- transmission line loss
- absolute gain of measuring antenna
- loss of interconnecting cables and connectors
- calibration of field intensity meter

- orientation of antenna during measurements
- altitude recordings
- direct pick-up (shielding of equipment)

The engineer must carefully plan the procedural aspects in order to minimize these errors. Antenna gain, measuring equipment calibration, and interconnecting line losses can be determined with a high degree of certainty (to within ± 1 dB of actual values) by sources traceable to NBS. The other error sources require carefuly attention to detail and, in the absence of good techniques, can easily result in errors of 2 dB and/or poor correlation of field strength-vs-elevation patterns.

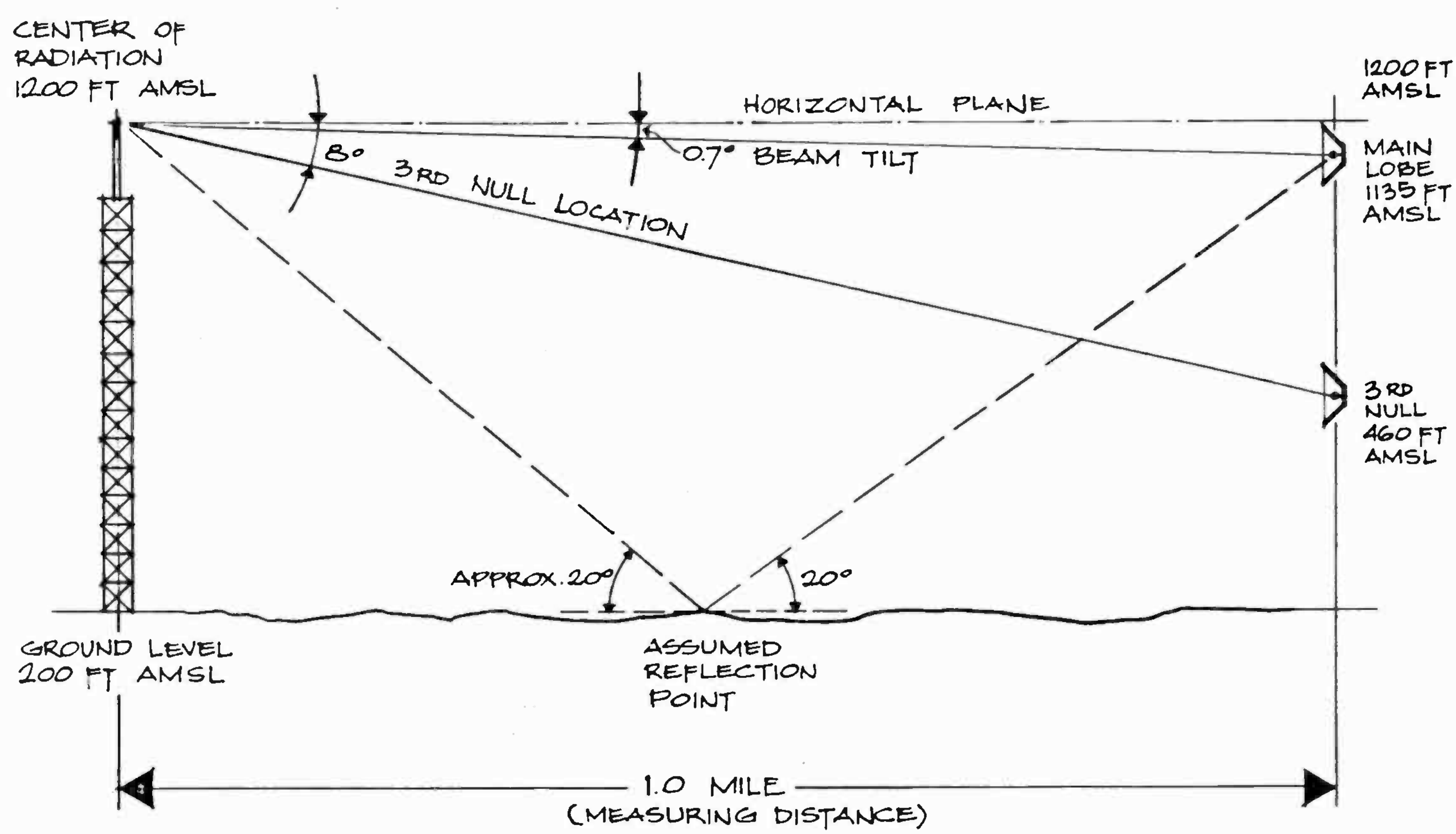
A hypothetical measurement case is given in Appendix 1. From these calculations, a good estimation of antenna performance can be made by plotting the resulting data.

Another approach which is recommended in cases where antenna performance is suspect because of "A/B" comparisons with other local TV station signals, is to make direct comparisons of antenna performance. Knowledge of the theoretical operating parameters of each facility will then permit assessments to be made on a relative basis.

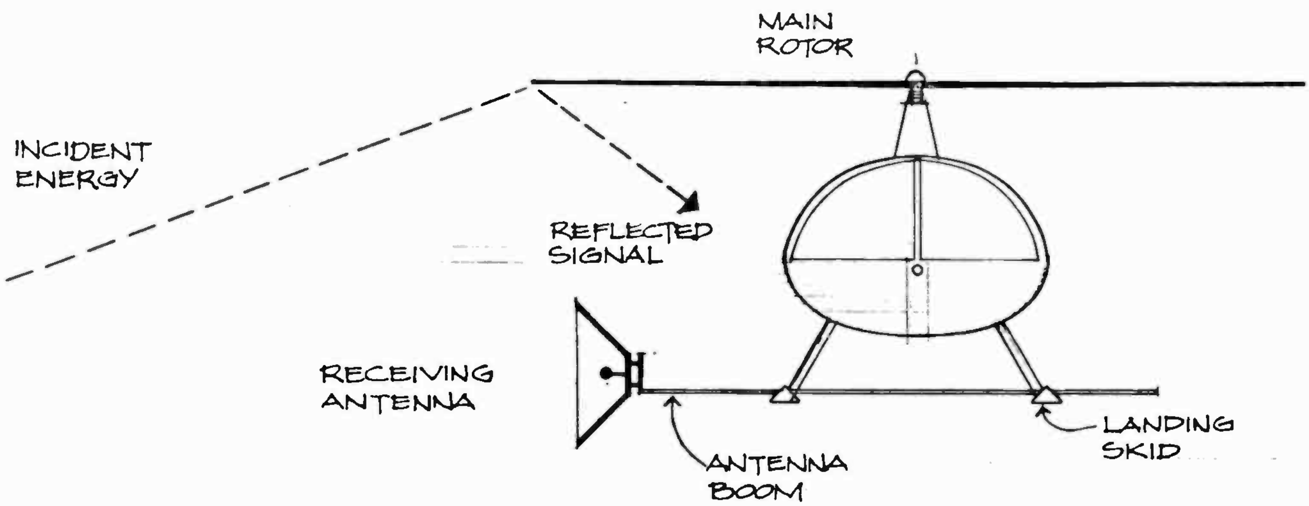
If it is assumed that virtually all of the power is contained in the region within $\pm 10°$ of the horizontal plane, it would be possible to integrate the measured data and arrive at a pattern gain factor. The actual gain of the antenna can then be determined by applying corrections for cross polarized radiation, feed system losses, and power contained in the sectors between $\pm 10°$ and $\pm 90°$. These latter corrections are usually based on assumed or calculated values, rather than empirical data, and is the approach normally used by the manufacturers in evaluating test range data.

Conclusions

The measurement techniques described in this paper can be used to determine the performance of medium and high gain UHF TV broadcast antennas, and considers measurements made at visual carrier. Similar measurements may be desired at aural carrier for beam tilt considerations. Application of the principles to VHF, low gain UHF antennas, antennas at low height above nearby terrain, and antennas located in congested areas must be done cautiously with due consideration given to ground reflection problems. Field test results in the measurement of five antennas including slot, pylon, helical, and "zig-zag" type radiators have yielded good results and have led to the identification of antenna problems as well as verification of original test range data.



	TYPICAL MEASUREMENT GEOMETRY PUBLIC BROADCASTING SYSTEM WASHINGTON, D.C.	FIGURE 1
	JOHN F.X. BROWNE & ASSOCIATES, INC.	June 1980

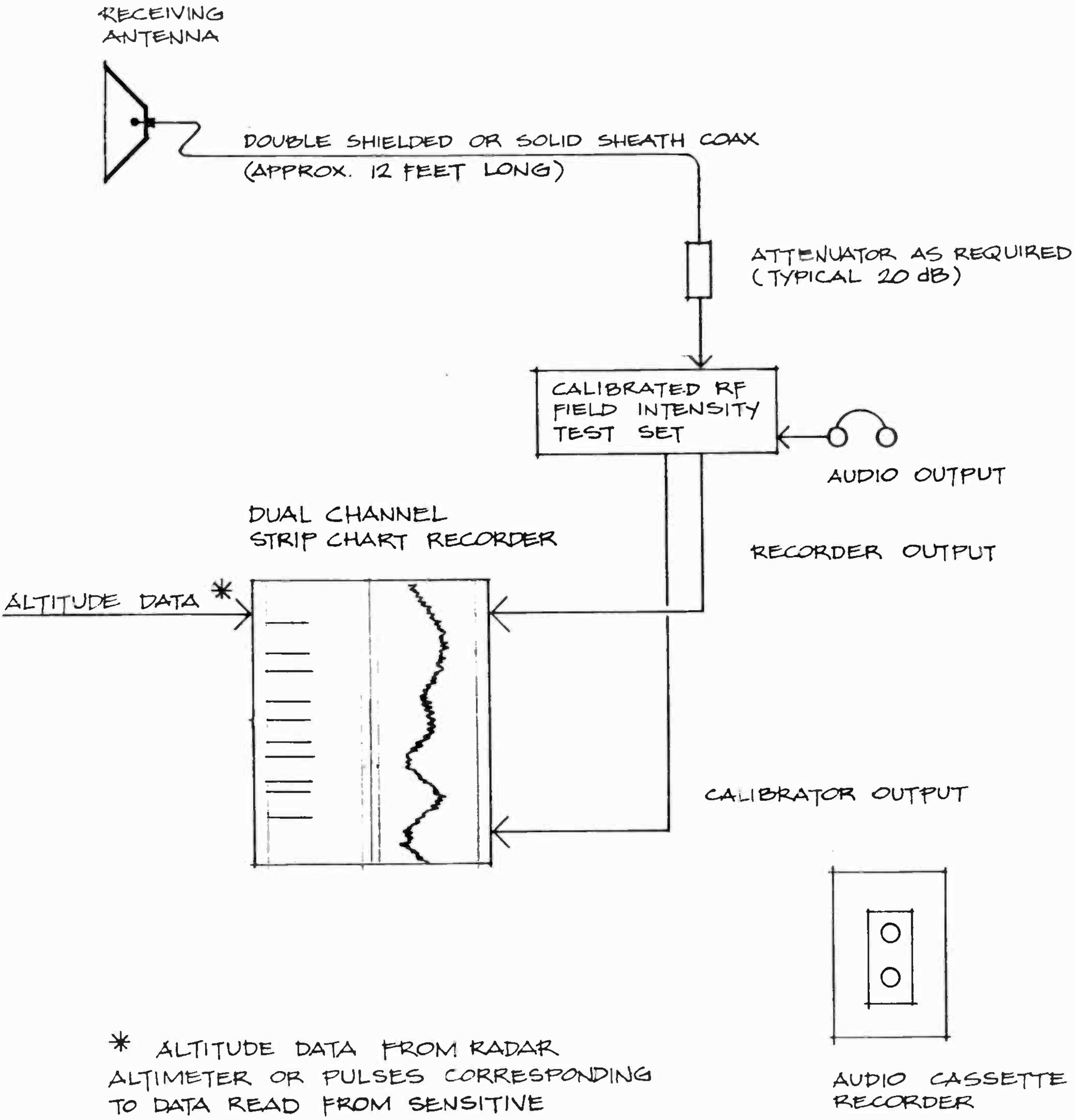


HELICOPTER INSTALLATION
PUBLIC BROADCASTING SYSTEM
WASHINGTON, D.C.

FIGURE 2

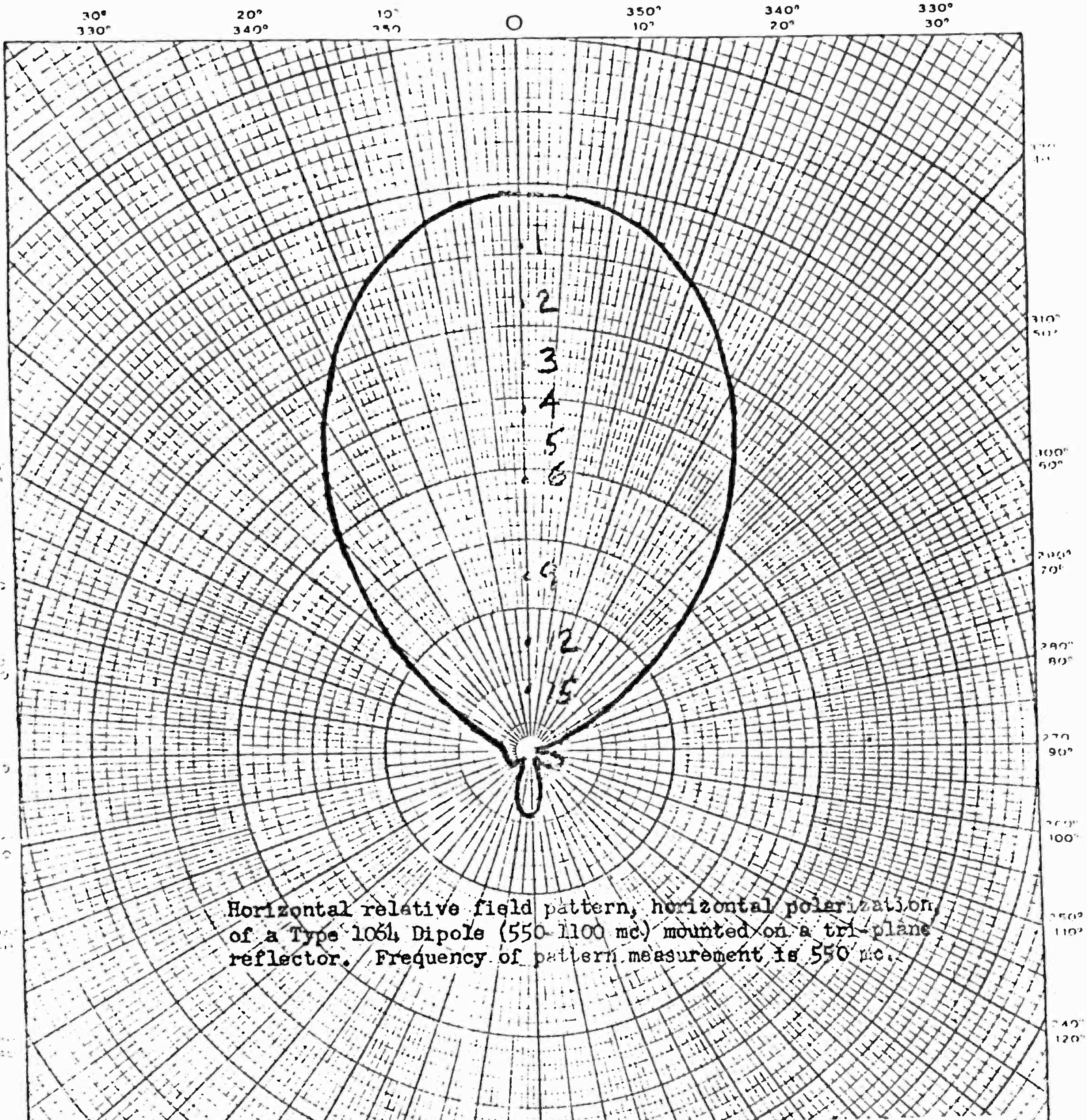
JOHN F.X. BROWNE & ASSOCIATES, INC.

June 1980



* ALTITUDE DATA FROM RADAR ALTIMETER OR PULSES CORRESPONDING TO DATA READ FROM SENSITIVE BAROMETRIC ALTIMETER

	BLOCK DIAGRAM PUBLIC BROADCASTING SYSTEM WASHINGTON, D.C.	FIGURE 3
	JOHN F.X. BROWNE & ASSOCIATES, INC.	June 1980



Horizontal relative field pattern, horizontal polarization, of a Type 1064 Dipole (550-1100 mc) mounted on a tri-plane reflector. Frequency of pattern measurement is 550 mc.

RECEIVING ANTENNA H PATTERN PUBLIC BROADCASTING SYSTEM WASHINGTON, D.C.	FIGURE 4
JOHN F.X. BROWNE & ASSOCIATES, INC.	June 1980

AMCI ALFORD MANUFACTURING COMPANY

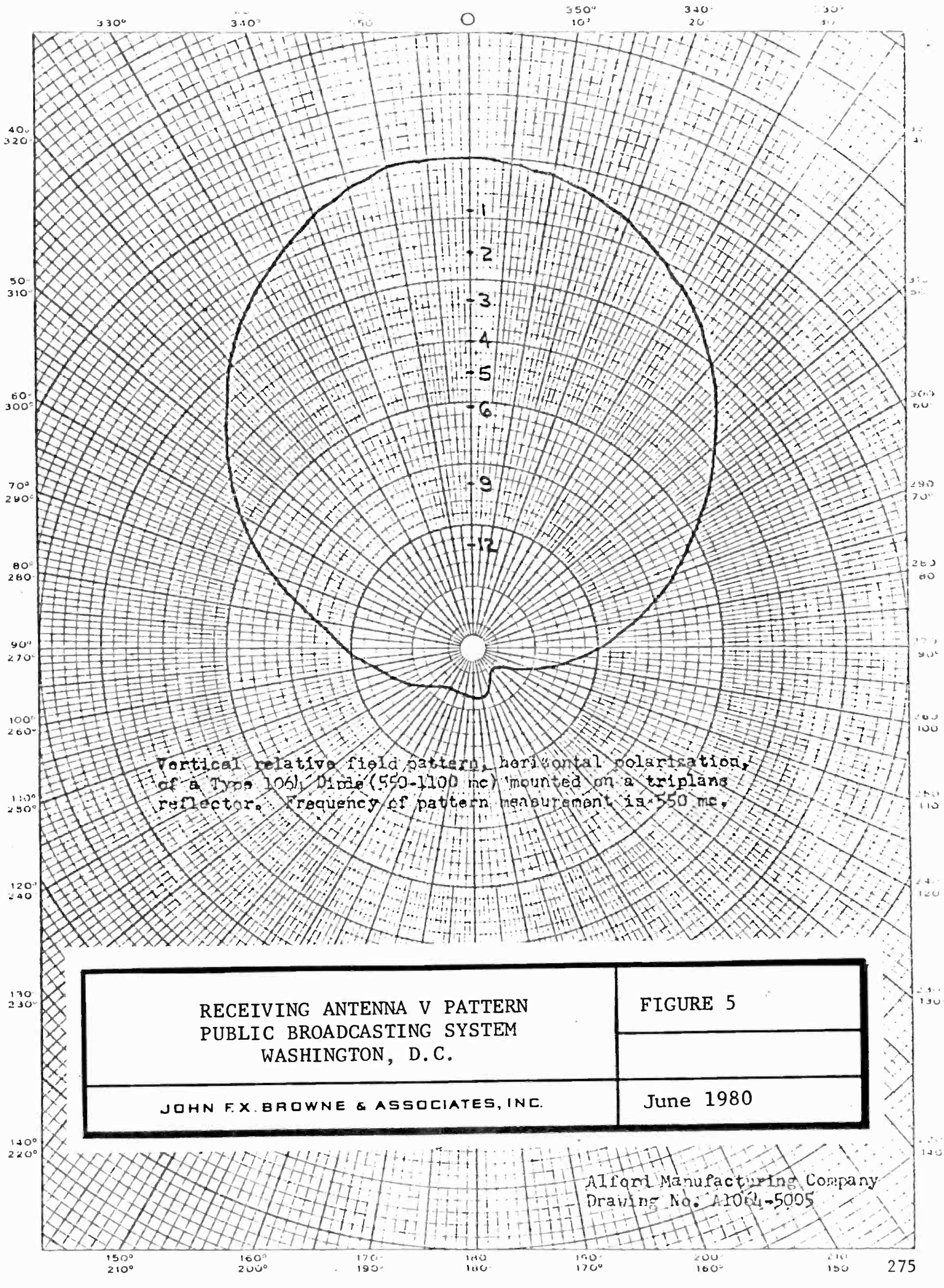
DR. BY TEM

DATE 4/26/61

TITLE 1064 Dipole Pattern

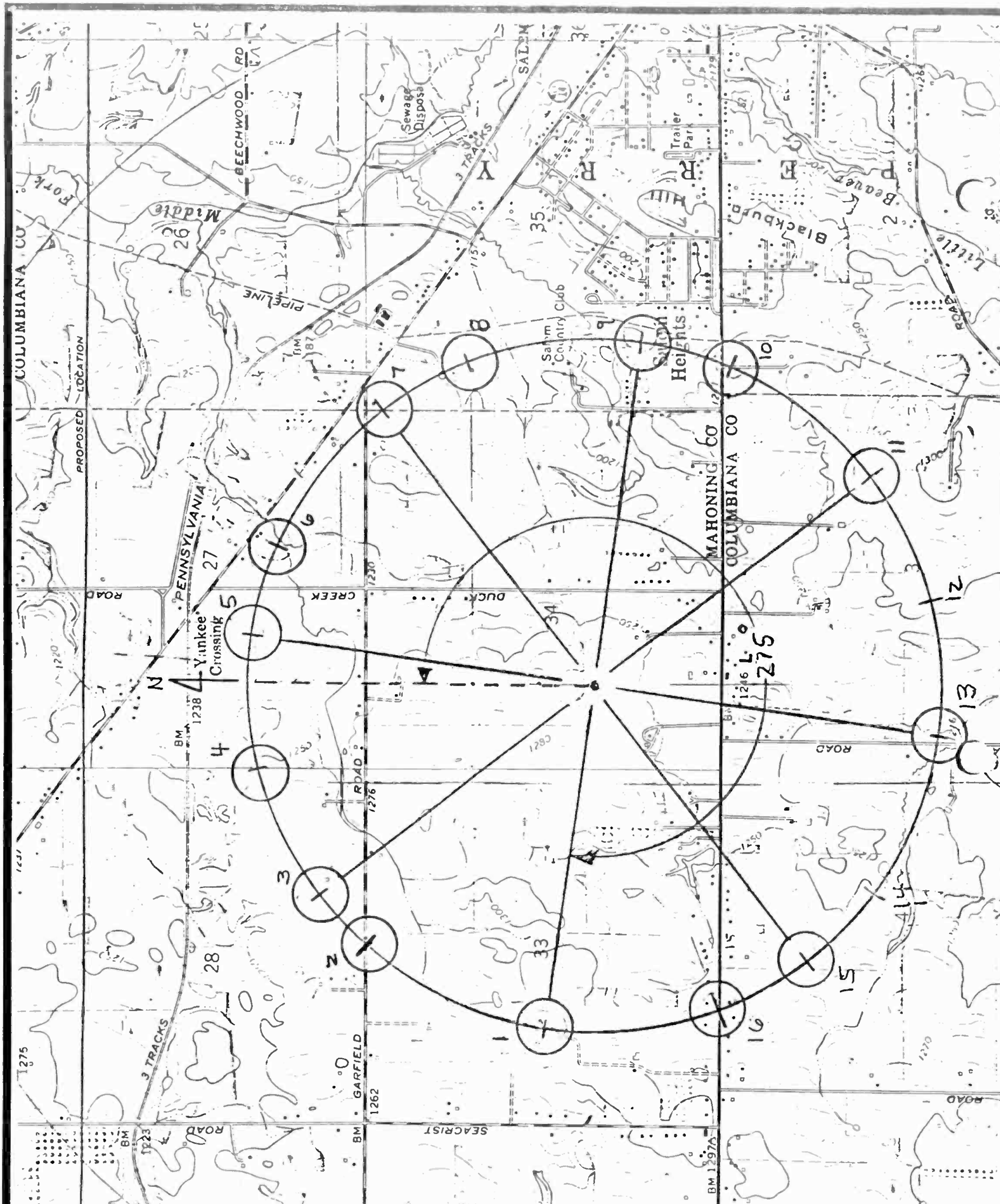
DWG. NO. A1064-5000A

AMCI 1000-60-033D

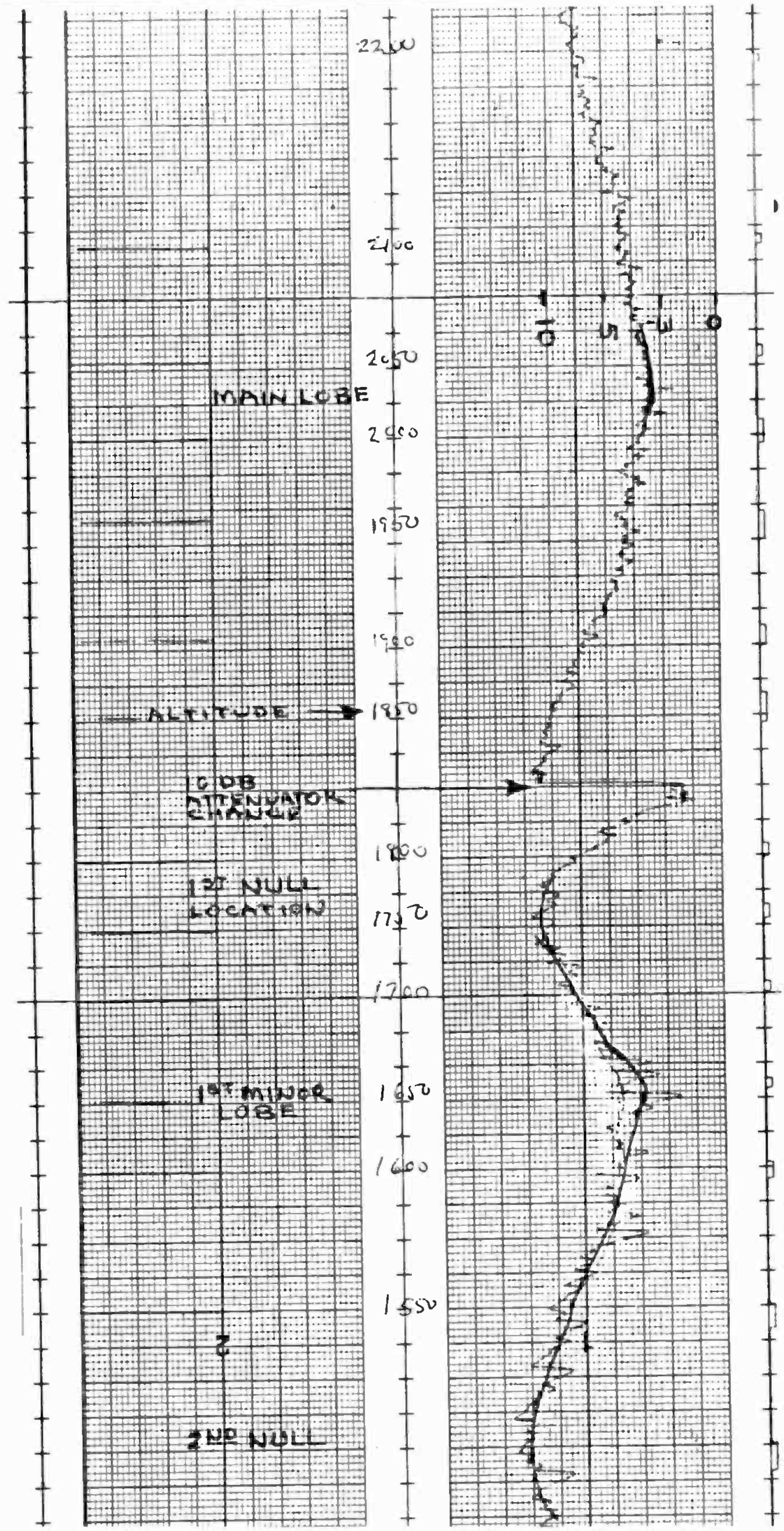


Vertical relative field pattern, horizontal polarization, of a Type 1064 Dipole (550-1100 mc) mounted on a triplane reflector. Frequency of pattern measurement is 550 mc.

RECEIVING ANTENNA V PATTERN PUBLIC BROADCASTING SYSTEM WASHINGTON, D.C.	FIGURE 5
JOHN F.X. BROWNE & ASSOCIATES, INC.	June 1980



	TYPICAL MEASUREMENT POINT PLOT PUBLIC BROADCASTING SYSTEM WASHINGTON, D.C.	FIGURE 6
	JOHN F.X. BROWNE & ASSOCIATES, INC.	June 1980



	SAMPLE CHART RECORDING PUBLIC BROADCASTING SYSTEM WASHINGTON, D.C.	FIGURE 7
	JOHN F.X. BROWNE & ASSOCIATES, INC.	June 1980

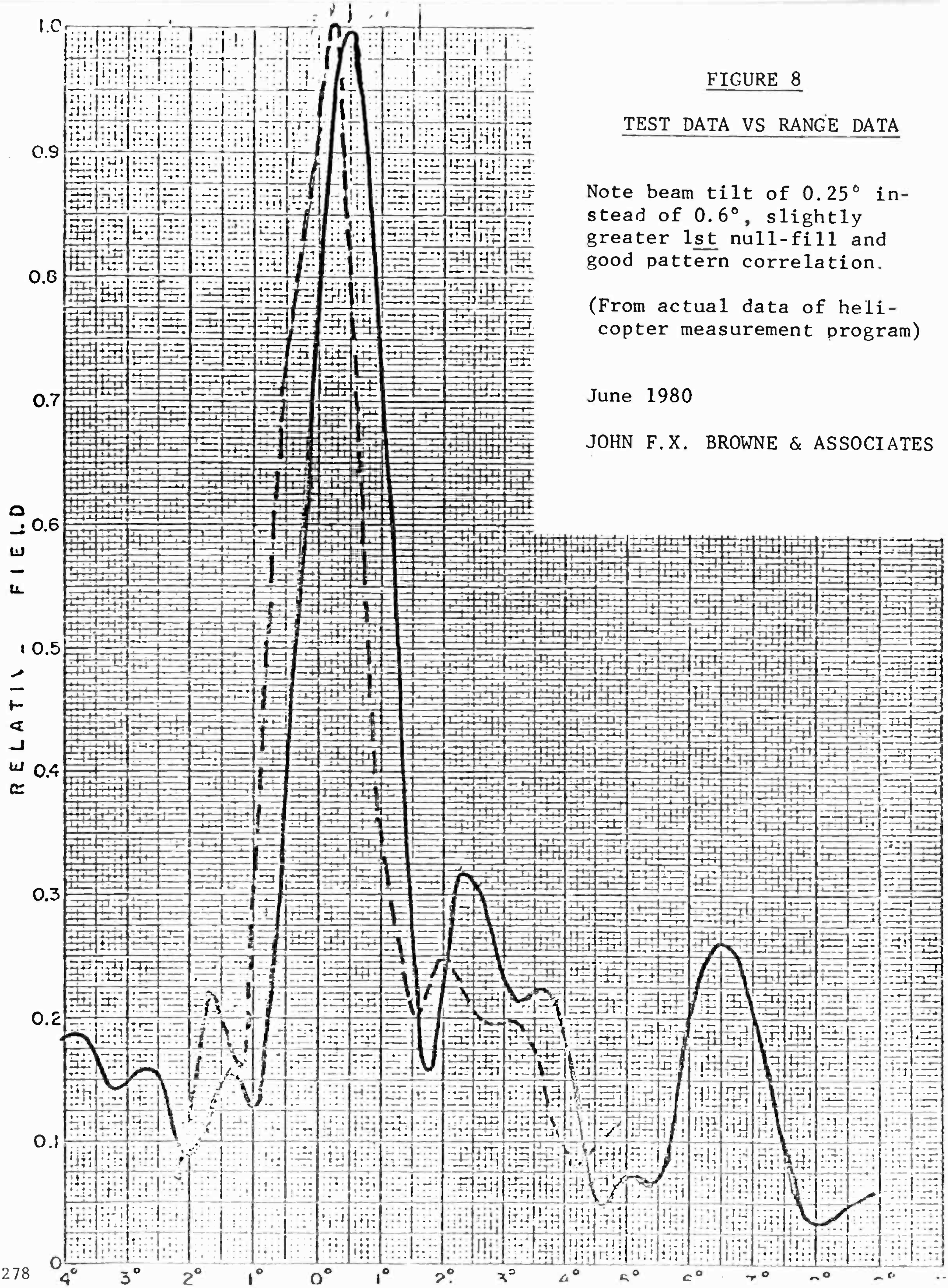


FIGURE 8

TEST DATA VS RANGE DATA

Note beam tilt of 0.25° instead of 0.6°, slightly greater 1st null-fill and good pattern correlation.

(From actual data of helicopter measurement program)

June 1980

JOHN F.X. BROWNE & ASSOCIATES

APPENDIX I

RECEIVED SIGNAL CALCULATION

The following is a sample calculation of the method of arriving at the expected received signal level (voltage) at the terminals of the measuring set shown in Figure 3. The example assumptions are as follows:

Station ERP (1.0 megawatt)	30.0 dBk
Operating Frequency (Channel 14)	471.25 MHz
Measurement Distance	1.0 miles
Path Loss	90.0 dB
Receiving Antenna Gain	6.3 dB
Cable and Pad Losses (Measured)	21.5 dB

ERP* - PL + GR* - Loss = Received Power

32.15 - 90.0 + 8.45 - 21.5 = -70.9 dBk = -10.9 dBm

(-10.9 + 106.99 = 96.01 dB above one microvolt @ 50 ohms)

*Corrected to isotropic reference.

AUDIO TIME BASE CORRECTION

John Pate

WSM, Inc.

Nashville, Tennessee

There are many devices used by radio broadcasters to maintain or enhance the technical quality of their air sound. Limiters, compressors and expanders are used to alter dynamic range; filters and equalizers are used to alter frequency response; and noise gates and noise reduction systems are used to lower system noise. All of these devices have one thing in common: they modify or control the amplitude of audio signals. The amplitude modification may be a function of level, frequency or some other characteristic. The problems and solutions involving amplitude processing are extensively documented and will not be treated in this paper.

There are a couple of significant technical problems however, which are of interest to broadcasters that have not received as much attention as those in the amplitude processing category. These problems fall into the category of time base errors involving a signal and its occurance in time. This is an area that has been dealt with for years in the field of video but has been neglected in the area of audio.

There are two specific time base errors of particular interest to radio broadcasters. These errors, which are associated with the playback of magnetic tape recordings, are mechanical in origin and are most commonly problems associated with the use of broadcast tape cartridges. In this particular case, much of the tape path is plastic with very loose mechanical tolerances and there is an almost total absence of tape dumping.

One of these specific time base problems is stereo delay error, sometimes called "stereo phasing". In this case, one channel leads or lags the other slightly in time. The time difference is seldom more than 200 microseconds and is of no consequence to the stereo listener. The mono listener, however, can hear a dramatically degraded frequency response of the center channel components. This effect is well known to broadcasters and has been an annoying problem to many since stereo broadcasting began.

Another specific time base problem is flutter, or cyclic speed error. This time base error is, again, the result of mechanical imperfections of tape machines and is worse with the use of the broadcast cartridge.

A time base correction device which deals with these problems has been developed by WSM and has been used successfully on the air for several months. We have found the device to be an effective and efficient means of controlling time base errors. The correction device operates by reading an inaudible pilot signal which has been recorded with the audio on magnetic tape. Upon playback, the pilot signal is decoded and an error signal is generated and fed to digital delays where the actual time base correction takes place.

A block diagram of the time base correction system in use at WSM is shown in Figure 1. It begins with an encoding device that generates a supersonic, and therefore inaudible, pilot signal which is recorded at a low level along with audio on the program tracks of the various cartridge, open reel and cassette recorders in use. The actual correction device is placed in the program line and becomes operative when the presence of the pilot signal is detected. In this "system" approach only one correction device is needed for an entire radio station regardless of how many tape machines are in use or even how many studios are in use. Whenever the pilot signal is detected by the correction system, it begins operating to minimize time base errors regardless of the source of the pilot. Correction at the system level, as opposed to correction at the machine level, does lead to one important limitation, however. No more than one pilot encoded source can be on the air at a time. While this initially may seem like a significant disadvantage, operating experience indicates that it is not. For example, on commercials the pilot signal is automatically made to end at the beginning of the secondary cue thereby allowing any number of commercials to be played sequentially without pilot overlap. On carted records, the announcer may talk over the beginning and end of the music as always. Momentary overlap of carted records and commercials is not harmful.

The actual time base correction is accomplished by two voltage controlled digital delays. This is illustrated in Figure 2. The delays we chose are Delta Lab DL-3's which were slightly modified. A decoding device samples the output of each delay, decodes the time base information from the pilot and then generates a feedback signal which is returned to the delay unit. This, therefore, is a closed loop correction system and continually acts to minimize time base error. Notch filters follow the correction system to remove the pilot before the signal is sent to the transmitter.

The pilot signal itself takes the form of a 19.000 kHz carrier which is 65% amplitude modulated by a 296.875 Hz sine wave. It is recorded on tape along with the left and right audio signals at a level of -25 dbVU. The carrier frequency was chosen so as to be supersonic but not so high in frequency that it would not be easily recordable on any tape recorder normally in use at a broadcast station. The specific frequency of 19.000 kHz was chosen to prevent any audible heterodyne from occurring in the event that slight leakage occurred through to the FM stereo generator which also uses a 19 kHz pilot. The modulating frequency was chosen because it is 1/64th of 19.000 kHz and therefore it is easily and precisely generated by a divider. This frequency is high enough to resolve accurate delay information yet low enough that the lower sideband produced by modulation, while lower in frequency than the carrier, is still supersonic. Modulation depth is not critical and reliable results are obtained anywhere from 50% to 75%. A modulation of higher than 75% is not used to simplify the process of regenerating the 19 kHz carrier. The pilot injection level of -25 dbVU was selected after tests were made to determine the minimum pilot signal needed for proper operation of the decoder, and the maximum level possible before modulation noise caused a measurable increase in audible noise floor. At -25 dbVU there is no increase in audible noise floor nor is any other audio performance criteria degraded in any way. Yet, this level is completely adequate to decode reliably at the output of the digital delays.

A block diagram of the pilot encoder is shown in Figure 3. The frequency standard consists of a 2.432 MHz crystal oscillator which is divided by 128 to obtain a 19.000 kHz (+/- 1 Hz) square wave. This is fed to switching transistor Q-1 and another divider to obtain a 296.875 Hz square wave which is then fed to a bandpass filter to obtain a 296.875 Hz sine wave. This signal drives modulating transistor, Q-2. The modulation is set to 65% with the potentiometer shown. The output of the modulated stage is then fed to a 19 kHz bandpass filter to remove 19 kHz harmonics and 296.875 Hz leakage. The bandpass filter output is then fed to two buffer amplifiers which drive the left and right program channels.

At WSM, the pilot encoder physically takes the form of a 1.5" by 6" module which plugs into a production console. The module has two momentary push buttons which turn the pilot ON and OFF. When "ON", a red LED illuminates and the pilot signal appears at the output of the console at the proper level of -25 dbVU. In addition to the OFF-ON push buttons on the module itself, the encoder is automatically turned ON when the stereo cart recorder's "record preset" button is pushed and automatically turned OFF when the "secondary cue" button is pushed. The pilot encoder module has no operational adjustments nor does it require any periodic checking or adjustment.

The time base corrector itself can be divided into two distinct error correcting sections which operate with shared digital delays. One section corrects for delay error between channels and the other corrects flutter.

Delay error is usually mechanical in origin and is normally the result of a record-to-playback gap azimuth error. The error voltage developed by the delay correction section is DC coupled and returns to only one of the two digital delays. In this way the delay of one channel is served in such a way as to minimize delay error at the outputs of the two delays. Figure 4 is a block diagram of the delay comparator and error voltage generator. Wideband samples of both delay outputs are first fed to 19 kHz bandpass filters which pass only the modulated 19 khz pilot. This is followed by an AGC stage and a detector to recover the 296.875 Hz modulation. Bandpass filters are then used to clean up the detected signal and to remove noise. The two 296.875 Hz signals are then fed to two comparators to accurately detect zero crossing and then on to a phase detector. The output of the phase detector is a voltage whose polarity is dependent on which channel leads in phase, and whose magnitude is proportional to the amount of phase error. This voltage is then amplified by the error amplifier and then integrated to average instantaneous error. A buffer amplifier follows and the output of this buffer is the control voltage which is fed back to one of the digital delays.

The second section of the time base corrector is the flutter reduction section. A block diagram of this section is shown in Figure 5. Since the nature of flutter is such that the same error occurs on all tracks of a multi-track machine, it is only necessary to sample one of the tracks. Also, for this reason, the same error correction signal can be returned to all the delays in a multichannel corrector. The 19 kHz bandpass filter and AGC are shared with the delay correction circuits. The modulated pilot is then fed to a limiter which removes the 296.875 Hz modulation and leaves only the 19.000 kHz carrier. This is filtered and applied to a frequency-to-voltage converter which produces an output voltage which is proportional to the carrier frequency. The AC component of this voltage, which correlates to flutter, is then amplified and comprises the flutter control voltage. This is fed to both delays and forms a feedback loop which acts to minimize flutter.

Operating tests of the time base corrector indicate very satisfactory results. The delay error corrector was designed to have a maximum correction range of plus or minus 350 microseconds and throughout this range the correction error is less than 5 microseconds. This equates to a delay error of less than 27 degrees of phase of a 15 kHz signal. The "lock-up" time for most signals is less than 2 seconds. In cases where extreme accuracy is required in delay error correction, a second feedback loop could be used which would compare the phase of the 19 kHz carriers. In this case, "coarse" delay correction would be done with the 296.875 Hz signals. When the delay error was brought to less than 26 microseconds (180 degrees of phase of 19 kHz), the second feedback loop would take over for "fine adjustment.

The flutter correction portion of the circuit was added after the mainframe was put on the air so only a small amount of time has been spent in "fine tuning" its operation. An ITC tri-deck was tested and found to have an average NAB weighted flutter of about 0.12% without correction and 0.034% with flutter correction. A Studer B-67 open reel machine was found to have 0.033% NAB weighted flutter without correction and 0.021% with correction. These results, while quite good, are those of a first generation device and are by no means the ultimate possible. Using this system, it is technically quite feasible to eliminate flutter for all practical purposes. In addition, for systems that require it, the feedback loop could easily be extended to include the capstan motor which would also eliminate speed error.

As we proceed up the learning curve by using this system, we found several characteristics that need to be taken into account in its operation, and we have found other characteristics which turned out to be unimportant. Some of these are as follows:

1) Since the correction system has a nominal delay of about 30 milliseconds when the pilot is present, "off-the-air" headphone monitoring by announcers is not practical. We have switched the headphones to monitor "off console" and have added equalization and compression to make it sound like it is "off air". The control room speaker monitoring remains "off air" as always.

2) The 19 kHz pilot reaches the transmitter suppressed to about 50 dB below program level and has not caused any detectable interference with the 19 kHz pilot put on by the stereo generator.

3) Tests have been made to determine DBX (a registered trademark) compatibility of the system using 303A cards. The companding of the pilot does not at all impair the operation of the time base corrector. There are, however, somewhat degraded noise reduction results because of the presence of the pilot. Our tests with broadcast cartridges indicated a 31 dB decrease in noise level using DBX without pilot encoding and a 21 dB decrease in noise level, over the audible range, using DBX with pilot encoding. This is the result of two factors. First, modulation noise is generated by the pilot in the audible range because it is recorded at a much higher level as a result of DBX encoding (compression) with no audio present. Second, the residual pilot level of -25 dB puts a limit on the degree of companding that can take place. Another complication is a slight tendency for audio-pilot intermodulation due to the higher record level of the low level passages. Neither of these obstacles is significant enough to outweigh the benefits of using noise reduction. A possible way to improve operation significantly using reduction would be to reduce pilot injection level to -27 dB or even -30 dB.

4) The 30 milliseconds of delay that is switched in and out as the pilot appears and disappears is not all noticeable. The red LED "Pilot" indicator in the corrector is the only means of telling that the unit has switched into operation.

5) In the first eight weeks of operation, of hundreds of cart and reel tapes that were played through the system, only two carts had enough program energy in the 19 kHz region to cause improper operation of the correction system. In both cases, they were agency supplied commercials and were severly overequalized. To guard against this, 19 kHz notch filters were installed in our

production consoles to limit program energy in the 19 kHz region. No problems have occurred since this was done.

6) There is plenty of margin in the pilot detector circuitry for low or high levels. If the pilot on the air console is anywhere within +/- 20 dB of normal operating level, the system will lock and operate normally. If two pilots inadvertently occur at the same time, there is a "capture" effect that tends to lock the system onto the stronger of the two pilots.

To summarize the advantages of the system:

1) Only one correction device is needed for an entire system, regardless of how many tape machines are used or even how many studios are used.

2) The pilot encoder is simple, inexpensive and has no operational adjustments.

3) The system corrects delay error and flutter from any tape machine without modification of the tape machine.

4) The installation of the system is simple.

5) The system is completely automatic in operation.

6) The system has no operational adjustments.

7) The system has long term stability.

8) Future upgrading would be easy and would not involve re-dubbing of tape already containing the pilot.

To summarize the disadvantages of the system:

1) The system is "double ended", i.e., program sources must be encoded for the system to work.

2) Only one encoded source can be on the air at a time, if "system level" correction is used.

3) If delay error exists on program material as it is being dubbed and encoded onto cart, the same delay error will exist as the cart is played on the air. (The "garbage-in-garbage-out" principle.)

There are two additional devices developed at WSM which although not directly related to time base correction, might be of interest to the stereo broadcaster and are worth mentioning briefly.

The first is a means for playing carts recorded in the newsroom on mono machines, on the stereo machines in the control room. Our system is a reflective optical sensor mounted in each playback deck that "looks" through a small (1/8" X 1/8") hole cut in the right hand cart guide. Carts recorded in the newsroom have a short piece of Scotch #850 reflective silver polyester tape, 1/2" wide, on the right side of the cart. When the optical sensor "sees" the reflective tape, a reed relay operates and the left playback amplifier feeds both the left and right outputs. Normal stereo operation occurs on any cart without the reflective tape. This system could also be used to enable a cart machine to differentiate between carts which are noise reduction encoded and those which are not. The system is inexpensive, simple and reliable. Also, the tape provides a means for the operator to know the status of the cart. A schematic diagram and parts list is available from the author on request.

The other device which we have developed and have been using for nearly three years is an unusual utility amplifier. It has a gain which is continuously variable from 8 dB to 68 dB, making it suitable for either a microphone or line amplifier. At input levels from -60 dBm to -40 dBm it has a residual noise level within 1 dB of thermal (for a 150 to 600 ohm source, 20 Hz to 20,000 Hz bandwidth), its frequency response is within +0 to -1 dB from 10 Hz to 43 kHz, it has very low distortion, a +25 dBm (into 600 ohms) output capability, and has a balanced and <u>floating</u> input and output. The unique thing about this amplifier is that it contains no audio transformers. A low noise Trans-Amp is used as a balanced front end and this is coupled to a balanced output stage with a linear optical isolator. The input and output stages are powered by their own separate floating power supplies. The input and output stages are therefore completely floating with respect to each other and to ground. A potential difference of 3000 volts can exist between input and output with no ill effect. With the transformers gone, there is no longer any problem with low frequency distortion, frequency response or group delay.

AM PREEMPHASIS AND TRANSMISSION BANDWIDTH

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SUMMARY

The success of FM broadcasting in recent years has caused AM broadcasters to become increasingly concerned with the technical quality of AM reception. Although many AM broadcasters transmit audio bandwidths to eight and even 15 kHz, their transmissions are seldom heard with such fidelity. The actual bandwidth of AM transmissions is discussed with a review of the typical day and night reception conditions. Existing receiver designs incorporate highly selective circuitry to minimize reception of interference which also severely limits the recovery of higher frequency audio. The use of a transmit preemphasis and the use of a 5 kHz audio cutoff filter for nighttime transmission is discussed. Tonal balance practices suggest that a 5 kHz limit in audio frequency response would complement a low frequency limit of 100 Hz which would be sufficient for the vast majority of automobile, table and portable radio receivers. The National Radio Systems Committee is studying the possible value of these practices especially as it relates to AM stereo.

INTRODUCTION

AM radio broadcasting in recent years has found itself a second class citizen to the increasingly popular FM service. FM, the "high fidelity" medium, had for many years been essentially dormant, with very few receivers in automobiles and only a small number in the homes of the audio enthusiast or the well-heeled. The FM programming was largely a duplication of the AM programming, with a few bold stations offering classical music...hardly a formula for rapid growth.

THE POPULARITY GROWTH OF FM

The change in the situation might be traced to the late fifties and early sixties when several things happened. First the stereo disc was invented in 1959 and the FCC established technical standards for broadcast FM stereo. They did not act at that time to provide similar standards for AM or television. Second, the transistor radio, or possibly more important, the low cost imported transistor radio which eventually had AM and FM became available to listeners of all ages. And thirdly, an FCC rule was established which prohibited broadcast stations in markets of over 100,000 people from simulcasting the same program on AM and FM for more than 25 percent of the programming week. Stereo, the low cost AM-FM radio and separate FM programming all contributed to the coming-of-age of FM broadcasting in the U.S.

Today, over 20 years later, more than 50 percent of radio listening is to FM radio. In fact the movement of audience has occurred at a more than three percent per year rate for the last seven years and we have every reason to believe that it will continue. (See Fig. 1.)

AM BROADCASTERS CONCERNED

This movement of audience away from AM to FM has caused a deep concern among AM broadcasters. This concern is reflected in the strong interest in AM stereo. AM stereo, it is hoped, will provide an attractive incentive for some of the migrated FM listeners to at least spend some of their time listening to AM radio.

AM RADIOS

Not only are the AM broadcasters concerned with the stereo feature for AM but they are vitally concerned with the sound of AM radios. They are convinced that the radio manufacturers couldn't care less about AM, and are concentrating all their design efforts on FM.....and maybe they are right, or partly so. It is with the arrival of AM stereo that it is hoped a new emphasis will be placed on the sound quality of AM radio.

AM radios may or may not have changed in sound quality over the years, with reasonably good and not so good ones appearing on the market every year. However there is a new standard for comparison....FM radio. When a person listens to his new AM-FM automobile radio and switches between FM to AM, there often is a considerable change in sound quality. This is very upsetting to the AM broadcaster because he knows that he is transmitting a high fidelity signal. Some AM stations transmit a signal just as high in fidelity as FM stations, and yet what comes out the other end is a poor second place.

Largely because of the built-in A/B test in every AM/FM radio, many AM broadcasters have deliberately changed the tone quality of their transmissions in an attempt to make AM sound like FM in a car radio. The result is a wide variation in sound transmission from one AM station to another, because each has adjusted its tone characteristics to the radios and the preference of station personnel involved. An NAB survey of AM stations showed that the tone characteristics varied from a flat response to more than 10 dB of treble boost to compensate for the AM radios. We have therefore developed an inconsistent standard for AM transmission of sound. How did we arrive at this situation and what can be done to improve the sound of AM radio?

AM TRANSMISSION CHARACTERISTICS

Although AM broadcast stations are frequently spaced every 10 kHz, the FCC actually permits the transmission of sidebands into the adjacent channels. AM stations are required to have a frequency response within 2 dB from 100 Hz to 5 kHz,[1] are required to measure the equipment performance from 50 Hz to 7.5 kHz,[2] and are permitted a frequency response out to 15 kHz and an occupied bandwidth of 30 kHz.[3] The FCC transmission limits and the spectrum of an AM station modulated with USASI noise is shown in Fig 2. USASI noise spectrum very closely matches the measured spectrum of modern music programming.[4]

AM RECEPTION CONDITIONS

Permitting AM stations to transmit sidebands well into the next channel appears incongruous...possibly that's why many have intuitively thought that AM stations are limited in fidelity to 5 kHz. However, the full bandwidth can be successfully transmitted and received during daytime hours because of the geographical and frequency spacing of local stations. Section 73.37 of the FCC rules describes the constraints on the overlap of the coverage contours daytime. For first adjacent channels (plus and minus 10 kHz) the 0.5 mV/M contours may not overlap. This means that at the outermost coverage of a local station, interference equal in amplitude can be experienced on the adjacent channels. However, traveling in toward the transmitter of the local station the desired to undesired ratio improves so that the local station becomes very strong and the adjacent channel interference becomes insignificant. Fig. 3 shows the actual measured spectrum daytime centering around a local Washington, D.C. station at 980 kHz. Notice that for more than plus and minus 20 kHz, there are neglible detected signals. In this case, the full bandwidth could be transmitted and successfully received.

Nighttime is a different matter. At night, the ionosphere reflects radio transmissions in the AM band over long distances, and adjacent channel protections which were enjoyed daytime largely disappear. See Fig. 4.

Figure 4. shows the spectrum around the same Washington, D.C. station a few hours later after sunset. New signals appear on every channel including the first adjacent channels. In this situation, the sidebands from adjacent channel stations cross over each other making it impossible for a receiver to successfully recover higher frequency sidebands of one station without receiving adjacent channel interference. It would appear that at night, the higher frequency sidebands could never be successfully recovered without interference except where the receiving location is very near the transmitter site of the desired station.

RECEIVER SELECTIVITY

The reflection of signals from the ionosphere at night contributes to the creation of strong adjacent channel interference. In order to successfully receive the desired station, the radio must be able to reject the unwanted adjacent channel interference. This requires the receiver selectivity to severely attenuate signals plus and minus 10 kHz and considerably attenuate signals even closer in frequency. The primary selectivity devices are the tuned circuits in the intermediate frequency amplifiers. These circuits when synchronously tuned typically produce the shape of the bandpass shown in Fig. 5. It is primarily the economics of circuit design that has produced this choice of circuitry.

The selectivity is peaked at the center frequency, and rapidly falls off either side in order to attenuate the adjacent channel carriers plus and minus 10 kHz and the sideband interference from the adjacent channel stations.

RECEIVER AUDIO RESPONSE

This type of selectivity shape successfully rejects the unwanted signals, but unfortunately also attenuates the higher frequency sidebands of the desired station producing a considerable rolloff in recovered audio frequency response. The measured audio response of a typical automobile AM radio station is shown in Fig. 6.

The measurement shown is of the electrical response only and does not show the effects of the acoustical environment, the loudspeaker or the enclosure. Note that the 3 dB response is out to only 1.7 kHz and that the response at 3 kHz is down more than 7 dB. The result is a sound that is quite soft, and because of speaker resonances on the low frequencies, a sound that can be muddy.

TONAL BALANCE

Tonal balance is an important factor of any sound reproduction device. Considerable research and testing is done by many manufacturers to determine what is best for their product. There are some general rules that have prevailed to this day as to a preferred response. The so-called 500,000 rule still is a useful design guide. This rule states that generally, the product of the low frequency response and the high frequency response should always equal 500,000. This rule is illustrated in Fig. 7. Each of the low and high frequency response combinations illustrated multiply out to 500,000.

In designing a sound reproducing system, consideration should be given to intrinsic limitations in high or low response so that the complimentary response can be derived from the rule. Generally, it is the low frequency acoustical response that is limited by either cost or physical size. A portable radio, table model, or automobile radio all have physical limits to the size of the loudspeaker and the enclosure which limits the acoustical low frequency response. The 500,000 rule tells us that for a certain low frequency response, there is an ideal complimentary high frequency response. Referring again to Fig. 7., a small portable radio could have a response of curve A, a table model radio curve B, an automobile radio curve C, and a home console unit curve D, all derived from the limits on the loudspeaker size.

Referring back to Fig. 6., the electrical response of the automobile radio has a 3 dB low frequency response lower than 100 Hz, while the high response goes only to 1.8 kHz. This is definitely not a balanced response and the product of the two is only 180,000 showing that it does not meet the 500,000 rule.

BROADCASTER EQUALIZATION

The broadcasters, after listening to their stations on automobile radios, conclude that the AM response is too bassy, especially in comparison with FM, and install various types of dynamic and fixed equalizing equipment to push up the high frequency response. There is no standard for this and the amount and frequency of the equalization is set up individually by station personnel.

The automobile receiver is very important to the broadcaster, and two conclusions can be derived from its characteristics. One is that a transmission preemphasis would help restore a tonal balance to a vast majority of receivers, and secondly, a 5 kHz high frequency response would be very adequate to complement the low frequency acoustical response available in the vast majority of automobile systems.

Fig. 8. shows the measured electrical audio response of a typical AM automobile receiver under several conditions: flat transmission to 20 kHz, preemphasis to 20 kHz, flat to 5 kHz and preemphasis to 5 kHz. Note that the 5 kHz limitation makes little difference to either the flat or preemphasized case and this is confirmed in listening tests. The preemphasis, however, makes a dramatic difference in the measured response and indeed in listening tests. When a loudspeaker system is used which provides a reasonable bass response, the preemphasis adds high frequency content which creates a very pleasing sound.

The concept of AM preemphasis and bandwidth limitations is not new. This area has been studied and reported in the December 1979 EBU Review.[5] The receiver limitations and the transmission bandwidth are similar to the U.S. but the interference environment is more difficult because of 9 kHz channel spacing and the high population and close spacing of radio stations.

If AM broadcast stations employed a uniform standard preemphasis, this would provide a consistent transmission characteristic for which both narrow and wide bandwidth, stereo and monaural receivers could be designed. It presumably would not be difficult to establish even a voluntary preemphasized transmission characteristic with the broadcasters, but the value of a limitation in bandwidth is more difficult to appreciate.

NIGHTTIME BANDWIDTH LIMITS

As shown in Figs. 3. and 4., the interference conditions are quite different in the AM band day and night. Although a wide bandwidth receiver can be expected to successfully recover 15 kHz of audio bandwidth daytime without substantial interference, for most listeners this would be difficult at night. The overlapping sidebands from adjacent channel stations make it impossible for any receiver to recover the full bandwidth of the desired station without also receiving adjacent channel sideband and carrier interference.

Because of the overlapping sidebands, interference from an adjacent channel actually becomes co-channel because higher frequency sideband components can appear in the center of the desired channel.

Therefore, in order to permit interference-free reception of the desired station, the higher frequency sidebands from the adjacent channel must be held back from appearing on the desired channel. This means that the audio modulating bandwidth must be reduced. Although it remains to be determined to what frequency, the obvious ultimate limit would be 5 kHz which prevents any overlapping of sidebands and theoretically makes possible reception of adjacent channels regardless of the amplitude.

CONCLUSION

This concept for nighttime operation is diagrammed in Fig. 10. The transmitted signal is preemphasized on the higher audio frequencies and cut off at 5 kHz. In a perfect embodiment, the transmitted signal has no sideband energy displaced more than 5 kHz from the carrier and the receiver has a perfectly square bandpass with an infinite slope skirt at plus and minus 5 kHz. There would never be any adjacent channel interference because there is no transmitted signal overlap and the receiver recovers signal from the desired channel only.

Although a 5 kHz bandwidth limitation appears very restrictive, based on the tonal balance rules, this is sufficient high frequency response to compliment a 100 Hz low frequency limit which covers all portable, table, and automobile receivers. It would only be the home audiophile who might have reproducing equipment capable of response lower than 100 Hz which might desire a wider complementary high frequency response.

There is no reason to limit the transmitted high frequency response during the daytime even with a preemphasis because the stations are geographically spaced to provide the adjacent channel protections. The same preemphasis curve can be used day and night except that a sharp cutoff filter at 5 kHz is to be used nighttime.

The transmission characteristic for this theoretical AM system is shown in Fig. 11.

Limiting the nighttime audio response of AM stations to 5 kHz may be a bitter pill for the AM broadcasters to swallow, but it is crystal clear that the vast majority of AM radio listeners will never be able to use wider bandwidth receivers nighttime if this is not done. It is to the mutual advantage of all AM broadcasters to limit their nighttime transmission response to some value to provide mutual adjacent channel protection and permit higher fidelity reception of their signals. A uniform fixed preemphasis for AM appears to be desirable to improve reception on existing receivers and to provide a consistent technical standard for the design of wider bandwidth monaural and stereo AM receivers.

The National Radio Systems Committee (NRSC) is co-sponsored by the Electronic Industries Association and the National Association of Broadcasters and presently has a subcommittee working on improved AM transmission and reception. The objective is to develop recommendations prior to the implementation of AM stereo so that the stereo transmitting and receiving equipment may incorporate the improved techniques.

The National Association of Broadcasters has not adopted a position on preemphasis or bandwidth limitations for AM broadcasting at this time. Further testing and analysis should be performed before the NAB Engineering Advisory Committee reviews the proposal. It is hoped that manufacturers of receiving and transmitting equipment will cooperate with the NRSC in evaluating the concepts discussed in this paper.

References

[1]FCC Rule 73.40 (a) (4).

[2]FCC Rule 73.47.

[3]FCC Rule 73.40 (a) (12).

[4]See Comments of Motorola, Inc. on the FCC Notice of Proposed Rulemaking in Docket 21313, AM Stereo.

[5]Processing Programme Signals for LF and MF Broadcasting, G. Petke, EBU Review, No. 178, December 1979.

HIGH POWER RF SYSTEMS

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INTRODUCTION

In most TV and FM installations, much time is spent in the selection of the transmitter to be used as well as the antenna. And after installation, much care is given in the care and feeding of the transmitter. However, there exists between the antenna and the transmitter a largely ignored area which, if not properly addressed in the beginning and on a regular basis during operation, can shut down a station as quickly as any other failure in the system. This area is the transmission line which carries the RF energy from the transmitter to the antenna.

This paper deals with practical approaches in the selection of the transmission line and practical experiences in the field, both on new installations and upgrading old installations.

Some of these experiences related here may seem comical but at the time they were more apt to bring on tears than laughter.

SYSTEM DESIGN FOR TRANSMISSION LINE SELECTION

For most installations, the selection of transmission line is a balance and a series of trade-offs between six major factors. They are;

1. The frequency at which you are operating
2. The loss of the transmission line desired
3. The wind loading of the line on the tower
4. The power required to be carried
5. The cost of the line
6. Whether you have a choice of using coax or waveguide

Decisions are also influenced by preferences of the people involved and their own personal experiences from the past.

In general, if you are in VHF or FM, you have little or no choice. You will have to use coax. In UHF however, a choice exixts between coax and waveguide.

Figure 1 is a flow chart for selecting a transmission line. The first decision point in the transmission line selection is the frequency. Practically speaking, this is already handed to you, being a gross generalization between UHF and VHF. With the VHF frequency to work with, you have little choice but to use coaxial line. The line size really is determined by the power, refer to Figure 2, and once this is determined, the wind loading should be checked, see Figure 3, to see that the tower can handle this load. If we have a 'go' in all these conditions, the problem has been reduced to selecting the source of transmission line. If, for some reason, you fail on the wind load test, you will have to move into an area I will call special tactics. This area will be discussed a little later.

In the UHF frequency range, the selection gets a bit more complicated, refer to Figure 2. Your first decision point after frequency is power level. Given the frequency and the power, your first question should be, "Can I even operate in coaxial line?" If the costing is favorable, then the coaxial line should be checked for the wind load on the tower. Once the wind load has been deemed acceptable, you have reduced the problem to the same point as you have in the VHF side. If you fail the wind load test, again you are in the same position as you are in the VHF side. You have to consider special tactics. However, if either power level vs. frequency or costs dictate to you a waveguide, the size should be selected and then go into the wind loading test with the same pass or fail criteria.

An interesting thing that should be noted is the fact that as you increase frequency the power handling reduces for coax while it increases for waveguide. Also, that the operating range of coax has an upper limit and is basically a broadband transmission system, where the waveguide runs have both a lower frequency limit and an upper frequency limit and tend to be quite a bit narrower in their fundamental operating range. If we now look at the attenuation vs. frequency, see Figure 4, we can see a very valid reason why the power handling on both coax and waveguide respond the way they do. You will note that as you go up in frequency, the attenuation of coax increases while the attenuation of waveguide decreases.

It is interesting to note that at some point the coax is actually lower loss than the waveguide. These are relative numbers and everybody has their favorite set of curves in calculating the efficiencies of their transmission line systems.

Figure 5 gives the relative cost of transmission line systems on a per foot basis. The interesting phenomenon here is that as your power goes up, your relative cost, along the power handling and attenuation in the UHF range seems to dictate going to a waveguide type system. The biggest drawback to this is the wind loading that the flat plate of the rectangular waveguide presents to the tower, as opposed to the smaller coaxial transmission line.

SYSTEM DESIGN - TOWER

The tower used must be capable of supporting the antenna selected and the largest sized transmission line expected to be used, (now or future). Microwave antenna equipment, two-way radio equipment, etc., must be considered also.

The responsibility for the tower, antenna and transmission line selection should rest with one individual or group in order to insure everything going together properly.

Transmission line layout, optimization, and installation can best be accomplished by those who will be responsible for the performance of the system. See Figure 7.

High power dictates that certain items, i.e.: patch panels, gas barriers, and elbows be located in such a manner that they receive sufficient air circulation to be able to operate at their rated power. Spacing between components should be such the VSWR's do not add up.

Optimization of the RF system insures that transmitters and diplexers "see" a good load; and when switching or patching the system, this load remains relatively constant. This will eliminate response changes in the transmitting system with switching or patching.

COAXIAL AND TRANSMISSION LINE DESIGN

The power RF system is often working at or near its maximum rating and to insure continued operation without burn-up, certain conditions must be met.

The inner conductor of any coax is going to expand and do so at a faster rate than the outer. To prevent problems such as over-run of the inner in relation to the outer, a captivated expansion joint is a must in the inner to accommodate this expansion. As this inner conductor will shrink when power is removed, the expansion joint must be "captivated" so that it will not come apart and open the line during this shrinking process.

The inner connector should also have sufficient "pull" so it cannot disengage. The inner connector should not have to move once it is made into the inner conductor. This "pull" or contact pressure will also insure the best possible RF contact with the inner conductor, eliminating the possibility of arcing the connector. The inner connector should have an "anti-split" protector to eliminate any possibility of improper installation in the inner conductor.

SPECIAL TACTICS

In some RF systems the system designer has the problem that his tower cannot accept the additional wind loading of waveguide, and yet he cannot because of frequency find a coax of suitable size and power handling capacity. The best alternative available at this point is to find some way of increasing the power handling of the coax line.

Dielectric has in the past worked with both Freon cooling as well as water cooling of transmission line which was to be operated severely above its maximum power for a given frequency. The Freon cooling does offer a reasonable alternative for the system designer to handle the special situations. However, since much greater care must be taken with the transmission line installation, it should only be used as a last measure in a system design. Freon is basically a gas which is probably better known for refrigerators rather than cooling of transmission line. However, certain grades of Freon are very suited to cooling transmission line without changing the other properties of the line.

The most common Freon used to date has been a Freon 116 or 116.5 which has been available from E. I. duPont and its distributors. The difference between the two 116's is a half a percent of some impurity which has little or no effect on the operation of the transmission line. Freon cools the line by transferring more heat from the inner conductor to the outer conductor than possible with either air or nitrogen. This means that the inner conductor will run cooler. However, the outer conductor, by virtue of carrying more power with the same attenuation, will run hotter under similar conditions and is directly related to the power in the transmission line and not the type of cooling involved.

The main parameters of the Freon which make it desirable for the cooling are a very heavy molecular weight giving very good heat transfer characteristics, a high voltage stand-off capability being, for all practical purposes, equivalent to that of SF_6, and the ability to operate at reasonable pressures in the transmission line, and the lower temperatures encountered in winter conditions without turning to liquid. This last property is very important since many of the other Freons that have been recommended for transmission lines at low temperatures in pressures of five to ten pounds per square inch could turn to a liquid in your transmission line. This would not seem to be your set of operating conditions for a high power transmission line system.

Experience has shown that approximately a 2 to 1 increase can be handled in a coaxial transmission line which has been purged of air and filled with Freon. This does not mean taking line and atmospheric pressure and adding Freon to bring the pressure up to five pounds gauge or even ten pounds gauge. This means evav-uating the line with a vacuum pump down to 20" or 30" of mercury of vacuum and refilling the evacuated transmission line with Freon. Care should be given to the contacts and the connectors within such a system since a substantial increase in current will be passing through these contacts. It has been our experience that line to be used at higher than rated power with Freon should be completely restored line that has existed on the tower and somebody just decides to put Freon in to it. However, with new or completely restored line, we have run transmission lines at up to twice their normally rated power in the high UHF range.

As stated earlier, there are many means of cooling coaxial transmission line such as forced air cooling, water cooling, heat sunk elbows, etc. However, for a general overall approach to coax line operating above its rated power, we have found that Freon offers the best solution.

WAVEGUIDE (WHAT TO LOOK FOR)

In a waveguide system we have gone from the familiar circular type of construction usually to a rectangular or square type of construction involving a hollow tube which is operating at the TE 01 mode and conducting power in the space between the walls. In general, due to the limitations of the material necessary to fabricate the waveguide, waveguide has been limited to approximately twelve foot lengths. This has usually been fabricated from four flat sheets welded at the corners. The flange passes over the rectangular tube and allows the tube to protrude through for a contact surface. After back-welding on the flange, the flange and waveguide tube are faced flush at the contact surface with a relief being built into the flange between the contact surface and the outer bearing surface. In general, we have found that such a flange with a built-in waveguide seal on one end allows for the most optimum type of installation from a cost and performance standpoint. The rubber seal in one face of each length of waveguide allows for pressure seal and does not require replacement for multiple use.

In addition, the contacts have proven to be reliable at levels three to four times what they see in a broadcast system, and the lack of knurled surfaces imbedding themselves in the contact surfaces of the waveguide negates many problems encountered when waveguide joints are made, opened, and then made again.

In general, the waveguide lengths are similar to coax lengths in the sense that two lengths are usually recommended depending on the frequency of the installation. This is to alleviate any possibility of flange build-up in a long waveguide run.

INSTALLATION

As mentioned in SYSTEM DESIGN, the responsibility for the installation should be with the designer of the system. In this regard, responsibility, design and supervision should be the same person or group. This will insure that the installation will be made as designed and responsibility for the system is established. This should obviously be done by those having experience in this area.

Installation of either coax or waveguide should be done by riggers experience in the field of transmission line work and not necessarily the same crew that erects the tower.

Where high powers are to be used, care must be exercised in such things as bolt tightening (torqued in all cases), "O" ring installation, hanger installation, etc.

In the case of some coax and all waveguide installation, optimizing is required and uses techniques not practiced by most riggers. This procedure requires the services of an engineer experienced in this work using such equipment as network analyzers, sometimes computer controlled, at the transmitter end of the system, working with a rigger at the tuners located on the sections of transmission line or waveguide.

A number of transmission lines have been measured and VSWR'S from good to bad have been recorded. The variation of VSWR between 1.1 and 1.2 is rather common in lines not optimized. See Figures 12 and 13.

Lines that have optimized elbows normally run under 1.1 and 1.08 is reasonable in runs to 1000 ft. See Figure 14.

An optimized waveguide run of 1000 ft. at 750 MHz was measured at a maximum VSWR of 1.06. This system uses older style tuners and could be 1.05 or less with the newer style tuners. VSWR's of 1.05 maximum should be possible in optimized transmission lines systems to 1000 ft. and 1.07 in runs to 1500 ft. Unoptimized transmission lines can be expected to be 1.1 to 1500 ft. See Figure 15.

A complete transmission line antenna system at Channel 41 shows a VSWR that is about normal for the systems measured and is used as an example. See Figure 16. The antenna used in this system was plotted after removing the transmission line. This plot shows the antenna has a good "shape" but is off center on the Smith Chart. See Figure 17. A fine matching unit at the input of the elbow complex results in "shifting" the antenna to the center of the Chart. See Figure 18. This optimizing eliminates a ghost and results in an overall lower VSWR of the system. This optimizing results in a system that is capable of operating at maximum power without reflections or failures.