PROCEEDINGS

38TH ANNUAL BROADCAST ENGINEERING CONFERENCE

NATIONAL ASSOCIATION OF BROADCASTERS
LAS VEGAS, NEVADA
1984

PROCEEDINGS

38TH ANNUAL BROADCAST ENGINEERING CONFERENCE

NATIONAL ASSOCIATION OF BROADCASTERS
LAS VEGAS, NEVADA
1984

R. 5-22-84

These proceedings contain technical papers presented at the NAB Engineering Conference April 28 - May 2, 1984, in Las Vegas, Nevada.

Published by the NAB Office of Science and Technology

Additional copies of the *1984 Proceedings* are available through NAB Services, 1771 N Street, N.W., Washington, D.C. 20036. For price information, please call (800) 368-5644.

1771 N Street, N.W.
Washington, D.C. 20036

Thomas B. Keller
Senior Vice President
Science and Technology
(202) 293-3557

National Association of Broadcasters

Dear Reader:

These Proceedings contain most of the papers presented at the 38th Annual Broadcast Engineering Conference held this year in Las Vegas, Nevada, April 28 through May 2, 1984.

Within this document are papers on important technical issues concerning all broadcasters (spectrum management and the broadcast auxiliary service), timely new papers on new uses of our broadcast spectrum (TV Multichannel Sound, AM and FM Radio Subcarriers, Radio Allocations), reference papers (Transmission Systems and High Definition Television) and state of the art papers on interesting and useful new technology in radio and television broadcasting.

Take the time to read and learn from the technical papers within this volume. To a large extent the future of our industry depends upon your desire and ability to understand and utilize the new ideas and technology presented here. In many respects, they are blueprints of our future; a how-to manual of success and a useful reference to compliment the NAB Engineering Handbook. Further, in the increasingly diverse and competitive broadcast marketplace, engineering becomes even more important to the maintenance of high quality signal transmission and the ability of a station to compete effectively. The size of these Proceedings reflect that concern and our efforts to provide broadcast engineers with important up-to-date technical information.

We at NAB are proud to publish these Proceedings. Your comments on any of the papers or any respect of the '84 Engineering Conference are always welcome.

Best personal regards,

Thomas B. Keller

Thomas B. Keller

1984 NAB BROADCAST ENGINEERING CONFERENCE PROCEEDINGS

CONTENTS

RADIO PRODUCTION TECHNIQUES

RADIO ENGINEERING

IMPLEMENTING TELEVISION MULTICHANNEL SOUND

TELEVISION PRODUCTION ENGINEERING

AM/FM ALLOCATIONS

RADIO SUBCARRIERS

TELEVISION NEW TECHNOLOGY

SATELLITE SYSTEMS

MULTICHANNEL TELEVISION SOUND: THE BTSC SYSTEM

UHF EFFICIENCY

NON-IONIZING RADIATION

BROADCAST AUXILIARY

RADIO - NEW TECHNOLOGY

SATELLITES AND DIGITAL TRANSMISSION

ADVANCED TELEVISION SYSTEMS

BROADCAST INTERFERENCE

SPECTRUM MANAGEMENT

Defining Signal-to-Noise Specifications

in Audio Tape Machines

Charlie Bates

International Tapetronics Corp./3M

Bloomington, Illinois

Comparing audio tape equipment performance on the basis of published specifications can be a difficult and misleading exercise because of the varying measurement methods that may be employed. Each manufacturer rates his machine by attaching a standard of performance derived from intensive measurement and study. These measurements may indeed be representative of that particular product's performance, but may not be comparable to competitive product specifications due to differences in test procedures or reference points. Signal to noise (S/N) figures can be particularly confusing due to the myriad of variables involved. A recent spec comparison of tape machines revealed a 20dB variation across published specifications in S/N for cartridge machines and an even wider spread in reel-to-reel S/N specs. In order to make an intelligent decision on equipment purchase, the buyer must be familiar with the test procedures used by equipment manufacturers to arrive at the published specs and how these different measurements relate to one another.

As an example, the NAB cartridge machine standard specifies the procedures for measuring signal to noise in a cartridge machine. It states, "The reproducer S/N shall be measured unweighted with a bandpass of 20Hz to 20kHz without tape running, but with an otherwise fully operating reproducer from 160nWb/m at 1kHz reference level. The minimum S/N shall be 50dB for mono and 47dB for stereo." The key words in this specification are "unweighted", "bandbass", and "reference level." Each of these factors affect the noise measurement in different ways depending on the type of noise source in the machine.

There are entire textbooks available on the subject of noise sources, but tape machines will be subject to four primary noise generators: thermal noise, shot noise, power supply noise and reproducer head induced electromagnetic interference (EMI). Lets take a brief look at each of these sources and see how they affect noise measurements.

Thermal noise is caused by thermal agitation of electrons in resistors. It is proportional to temperature and bandwidth, and is commonly called white noise (having constant energy per hertz bandwidth). The open circuit noise voltage generated by each source is given by:

$$V = \sqrt{4kTBR}$$

Where: k = Boltzmanns constant (1.38×10^{-23} Joules/°K)
T = Absolute temperature (°K)
B = Noise bandwidth (Hz)
R = Resistance (Ohms)

Shot noise originates in semiconductors. It is generated by the random diffusion of minority carriers and the random generation and recombination of electron-hole pairs. Shot noise is also a white noise and is expressed as a current:

$$I_{SH} = \sqrt{2qI_{dc}B}$$

Where: q = Electron charge (1.6×10^{-19} coulombs)
I = Average DC current (Amps)
B = Noise bandwidth (Hz)

The power supply related noise seen at the output of a tape machine will consist of both even and odd harmonics of the power line frequency. This noise can be induced in the reproducer electronics by stray magnetic fields, or generated by power supply ripple. Improper grounding techniques can also result in power supply related noise generation in a cartridge machine.

The fourth major source of noise is EMI coupling into the magnetic head. The head is painstakingly designed to be an effective magnetic field transducer and can only be protected from EMI by proper shielding.

When unrelated noise voltages or currents are added together the total power produced is equal to the sum of the individual powers from each source. As an example, two unrelated noise voltages would add as follows:

$$V_T^2 = V_1^2 + V_2^2$$

For more than two noise sources:

$$V_T^2 = V_1^2 + V_2^2 + ... + V_n^2$$

Fortunately, voltmeters are available to perform this mathematics for us. It is here that the NAB guidelines calling for an "unweighted bandpass" come into play. The noise voltages measured from thermal and shot sources are limited by the bandwidth being measured as is shown in their equations and also limited by any weighting curve that may be employed in the meter. The NAB recommended bandpass restricts measurements to frequencies that fall within the 20Hz to 20 kHz audio bandwidth.

Weighting Curves

Subjective testing over the years has resulted in a number of human auditory sensitivity curves which clearly show that our ears are not equally sensitive to all frequencies in the audio spectrum (see Figure 1). The Robinson and Dadson curves illustrate how our perception of loudness varies with frequency, and how this effect becomes more pronounced at lower audio levels. This type of data gave rise to the "A" and "B" weighting curves, which were developed to enable sound level measuring instruments more accurately depict what the human ear would actually hear (see Figure 2).

The "A" curve closely resembles the equal-loudness curve in the area of 40 phones (about the sound level of the average residence or business office) and is used in speech interference measurements. The "B" weighting curve typifies the equal-loudness curve of 70 phones (about the sound level of heavy traffic) and is generally employed in traffic noise studies. The "A" and "B" curves in Figure 2 illustrate the increased weighting effect of the "A" curve, employed for lower level measurements than the "B". The "A" weighting curve and the CCIR curve are employed in some audio analyzers for noise measurements. Their purpose is to tailor the meter's reading to the response of the human ear at these low levels.

Spectral Measurements

To graphically illustrate the effects of weighting lets look at a 20Hz to 20kHz spectral plot of the noise output of a cartridge machine (not pulling tape) measured unweighted according to the NAB S/N specification (see Figure 3). Note that the primary noise components are 60Hz and its harmonics. These are the result of power supply ripple and EMI induced into the tape head. The spike at 90Hz is the capstan motor commutation frequency radiating into the head.

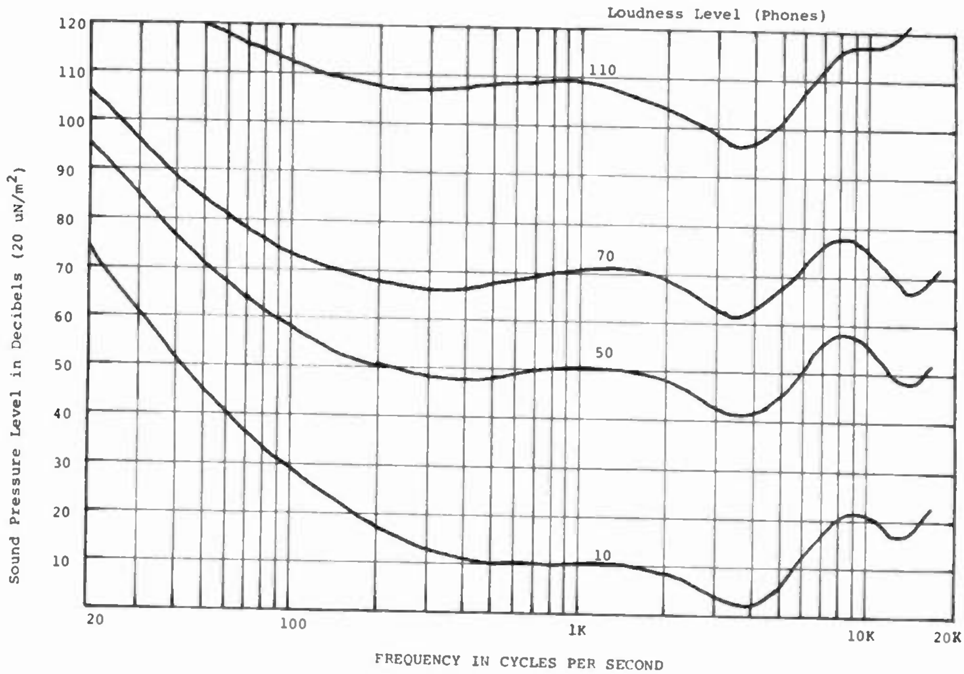


Figure 1. The equal loudness perception curves for pure tones determined by Robinson and Dadson in 1956 at the National Physical Laboratory, Teddington, England.

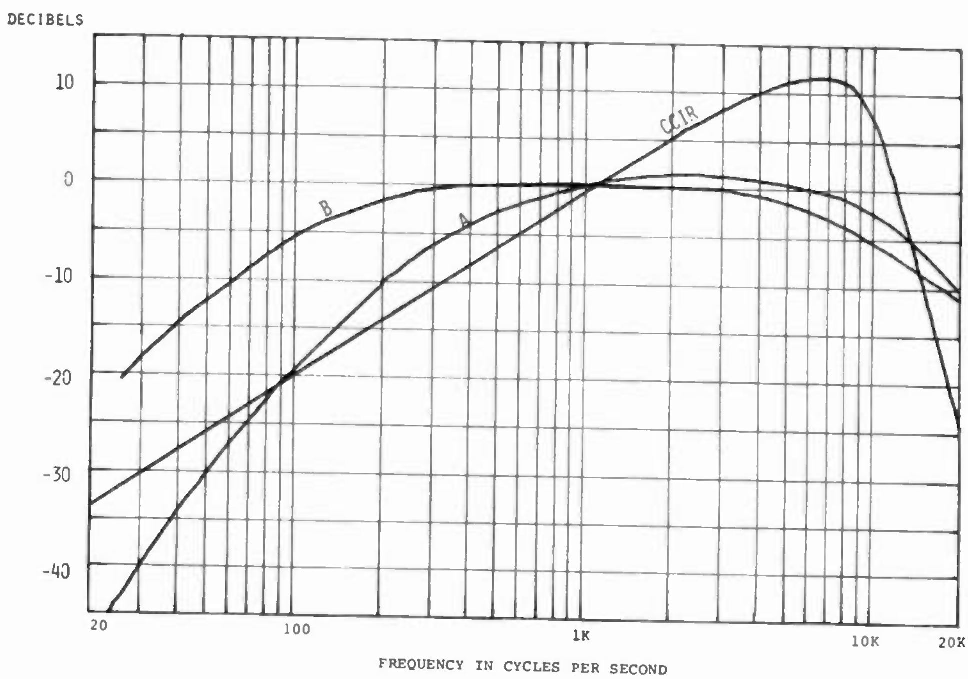


Figure 2. Frequency response or weighting curves used in measuring noise to discriminate against low and high frequencies in accordance with the equal loudness contours. "A" and "B" were developed for sound level meters. "A" and "CCIR" are used in audio analyzers.

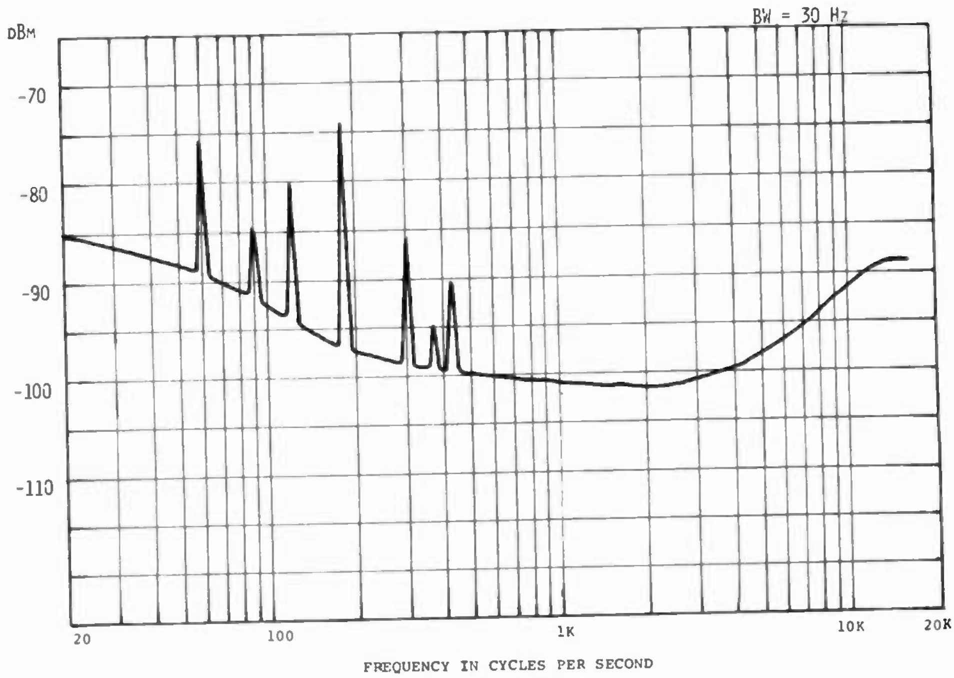


Figure 3. The spectrum of noise output from a cartridge machine measured unweighted, not pulling tape.

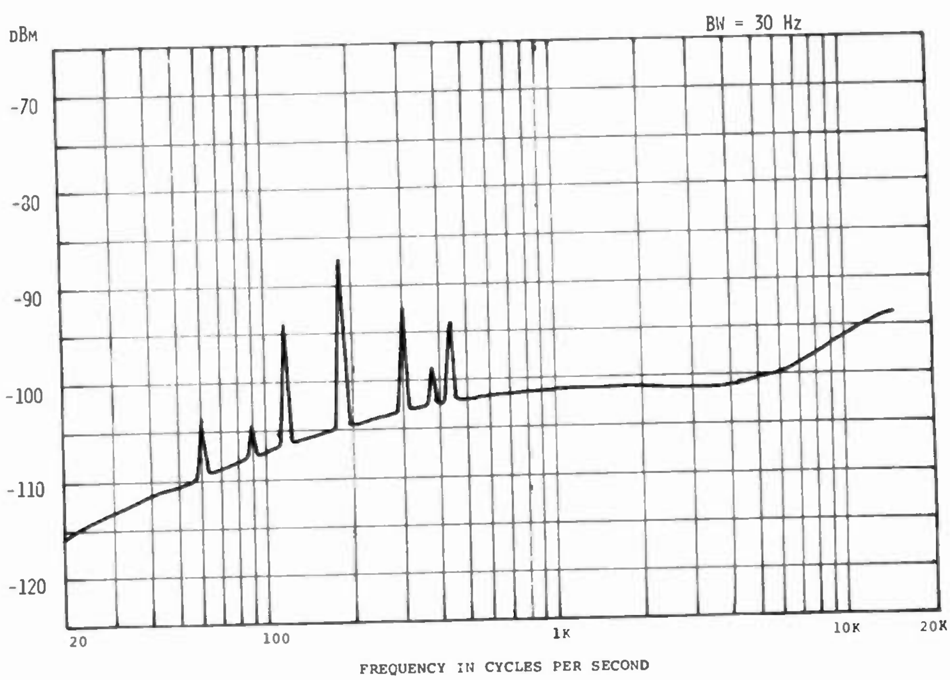


Figure 4. The spectrum of noise output from a cartridge machine measured with the "A" weighting curve employed, not pulling tape.

The same noise spectrum with an "A" weighting curve employed shows that the effects on frequencies below 500 Hz and on frequencies above 5kHz, are quite dramatic (see Figure 4). The meter readings associated with the unweighted NAB spectrum and the "A" weighted spectrum are -64dB and -70dB respectively. This 6 dB difference in noise level is achieved with the same cart machine, but different measuring techniques. While both of these noise measurements are quite legitimate, it is important that the potential buyer of the equipment understand how the numbers in the specifications were actually measured.

Today, we are recording at higher fluxivities thanks to improvements in magnetic tape. The tape machine which produced the noise spectra is calibrated for a fluxivity of 250nWb/m at the factory to accomodate this hotter tape and will therefore show a 4dB better noise figure than a machine set up at 160nWb/m as recommended by the NAB for noise measurements. The "A weighted" noise measurement of this tape machine is a full 10 dB lower than one would measure by using the NAB standard (60 dB per the NAB spec and 70 dB measured "A" weighted from 250 nWb/m reference.) Both of these measurements are useful for comparing machines and knowing both will tell us something about the frequency makeup of the noise. The published S/N specification on the test machine was 53dB, pointing up another area for consideration.

Tape machines coming off an assembly line can vary by as many as 6 dB in signal to noise with a Gausian type distribution. Each manufacturer elects to set his published specification at either a nominal level or at a worst case level below which no machine may pass. The buyer must know how each manufacturer has written his spec before comparing machines.

And now we are dawning on a new era of audio tape recording. Digital audio is certainly nothing new, for Nyquist demonstrated the mathematical possiblities of such a system in 1926. But digital recording is still in its infancy in practical applications as demonstrated by the ongoing struggle for standardization. However, bolstering specifications of better than 90 dB signal to noise radio and 90 dB dynamic range, digital audio recording will eventually find its way into our marketplace bringing with it yet another set of numbers to add to our confusion. The numbers specifying the signal to noise ratio of a digital audio system don't tell the whole story and cannot be directly compared with the signal to noise specifications of analog systems. This is an excellent example of comparing apples and oranges. The measurable noise generated by a digital system is so different from that of an analog machine that a new volcabulary of words have been used to describe it like quantization error, aliasing, granulation and bird singing. We will even add a little dither to turn granulation noise into white noise because white noise sounds better. And now we have clock jitter instead of flutter.

Measuring these performance parameters using conventional methods developed for analog recorders can produce some misleading results. The noise generated in the digital recording and reproducing process can consist of discrete frequencies which aren't harmonically related to the original signal. While these components may be quite small in terms of a measured voltage they can be much more offensive in terms of audio perception than a much higher harmonic distortion component or the more benign "white noise" to which our ears are much more tolerant. But, there are many very good papers already available on the topic of digital recording which cover in great detail the special problems involved and how they are being dealt with. The real test of signal to noise is in the listenting and the perception of the listener whether comparing analog or digital machines.

Specialised Trucks for Radio Remote Pick-Ups

Duncan MacEwan

British Broadcasting Corporation

London, England

Introduction

The BBC, now 62 years old, is properly regarded as the main national instrument of broadcasting in the United Kingdom. As such it operates four countrywide Radio Networks three of which are in stereo, National Regional Services in Scotland, Northern Ireland and Wales, a system of Local Radio Stations in England and the Channel Islands and a number of small Community operations scattered throughout its islands. All three wavebands are programmed viz. medium and long using AM and VHF Band II, FM. As such Radio has need of a comprehensive fleet of remote pick-up trucks (outside broadcast or O.B. vehicles in our parlance); O.B.s have always been considered to make a very important programme contribution to Radio's output and in recent years a considerable amount of effort has gone into perfecting the various techniques.

Since the Networks range in character from one which is predominantly news and public or current affairs oriented (Radio 4), through others of an M.O.R. (Radio 2) and serious orchestral, 'good music' character (Radio 3), to Radio 1, essentially a 'pop' music service, the trucks are of necessity very diverse in their type and style. In fact they extend from the converted London radio taxi and self-drive estate cars (complete with pneumatic mast and radio link) through to those of considerable sophistication costing $400,000 and even the latest one, a state-of-the-art digital control truck at a figure of approaching three quarters of a million dollars. The fleet which extends throughout the four countries which make up the U.K., now consists of some 60 fitted trucks supported by a further 35 auxiliary vehicles - caravans, aerial towers, trailer-mounted power supply generators, Land Rovers, etc. All have been custom-built for the various programme requirements which have either existed at the time or been foreseen on the horizon by creative engineers. Considerable trouble has been taken with their layout and equipment detail in order to meet the demands of the programme editors and producers most effectively. While many are specific to a particular need, economy of operation and flexibility in the use of each has also been part of the philosophy throughout the development.

Only by a careful study of the full range of remote pick-up programme situations has it been possible to design these various mobile facilities. That programme potential has been enhanced there can be little doubt with this range of fitted vehicles developed over the past 10 years. They are suitable for a very wide variety of outside broadcast work, matching the increased programme complexity and understandable desire on the part of the producers to take the programme to the public and to be seen to be out in the field transmitting 'live' or recording for later use. The PR factor is not to be under-rated.

1. Radio Cars (of the compact Station Wagon variety)

These are employed in Local Radio, one or at the most two per station and are also finding application in the National Regions notably in Belfast, Aberdeen and Edinburgh. The two channel mixer/transmitter is dashboard-mounted, programme cue and talkback facilities are included, as is a 6 metre extendable pneumatic mast; this and the ignition system are interlocked in order to avoid the mast getting progressively shortened! The cars are driven and operated by the Producers, Reporters and Station Assistants and the whole system made as fool-proof as possible. A reception point is chosen on the highest structure available and permanent PTT (AT & T) circuits ordered between it and the station's studios. The directional receiving aerial is remotely steered from the studio premises in order to maximise the signal, again a non-technical function involving only meter readings and listening to some test speech. Recently the basic system has been extended by the introduction of a second and even in some cases a third reception point in other towns within the editorial boundary of the local station, thus giving a more adequate County coverage.

2. Radio Taxis

For more comprehensive work in the Radio reporting field in the London area with the four Networks' heavy diet of news and current affairs output in mind, the use of a versatile and well equipped vehicle is called for. The London taxi with its durability, quite incredibly small turning circle of 7.62m (25 ft), load carrying capacity, generous headroom of 0.965m (3' 2") and a floor to headcloth height of 1.27m (4' 2") is an outstanding candidate for the role. The passenger compartment re-windowed with dark glass, is fitted out as a small interview studio and acts also as a base for location reports or recorded interviews. The programme material can be edited in the driver's compartment which is equipped with a 6 channel mini mixer, a Nagra IVs recorder and a Sony cassette machine. Cue and UHF talkback radio links are provided for both the Reporter and the one engineer who drives, operates, edits tape and carries out first line maintenance on the equipment. Again a pneumatic mast (9 metre) is included for the main programme radio link (VHF 40 Watts) and comprehensive battery arrangements provided along with mains charging facilities in the trunk at the rear of the vehicle. Facilities also exist for using the taxi's link equipment to relay commentary or reports taken on a hand-held walkie-talkie a hundred yards or so away.

In order to cover the London area satisfactorily a total of three permanent reception points is used in this large city conurbation case. These two taxis which give yeoman service also have cut into their roofs, an observation port to allow the Reporter to stand up and see out over the heads of others when the vehicle is parked e.g. in front of No. 10 Downing Street, the Prime Minister's residence, while awaiting an announcement from its steps;

A Radio Taxi

Radio Taxi - the engineer-operator's position

it proves advantageous to gain height in this way in any news situation where the media are present in abundance and might otherwise crowd Radio out.

3. The Mobile Studios

These two comparatively new major facilities were put on the road in May and June of 1982 specifically to meet the requirements of the bigger, longer-running current affairs and sports events. Week-long Political Party and Trades Union Conferences are examples of the former which like the British Open Golf Championship are peripatetic annual occasions. One was sent to Helsinki last year to cover the first World Athletic Championships, and there is an expectation that both will figure in the coverage of the Los Angeles Olympics. BBC Radio reflects these kinds of events throughout its output, but many other uses for these new studios not envisaged at an earlier stage are already becoming apparent. The 11 metre long (36 ft) 16 ton custom-built vehicles are equipped with a dual steerable front axle to allow manoeuvring in confined parking spaces, and have a width of 2.5m, the maximum permitted under the British Road Traffic Act. They have three main compartments - the studio, cubicle and radio link area and two subsidiary ones - the full length commentary platform on the roof protected by a guard rail, and the driving compartment which with its mic sockets and cue lights, can also be used for reporting or commentary work.

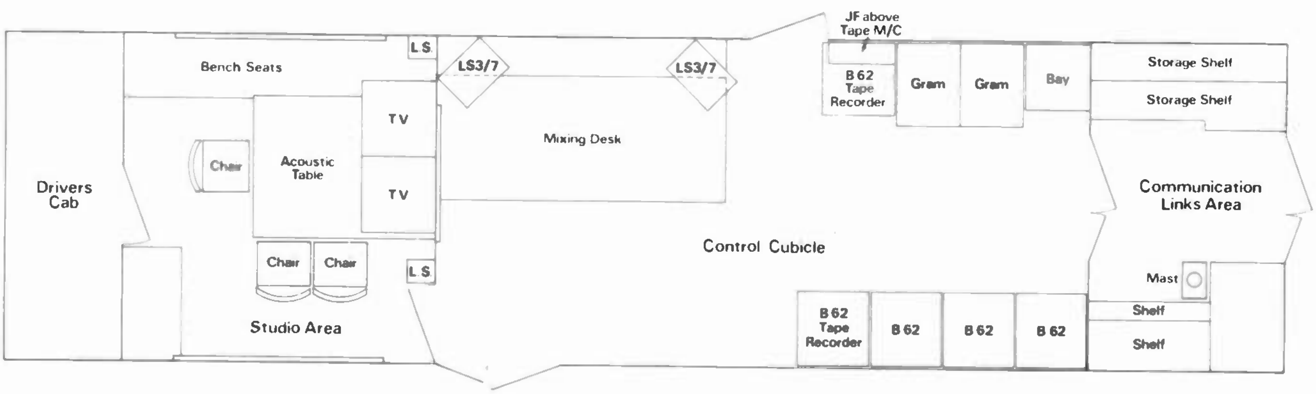


Fig. 1 Mobile Studio Layout

Access to the control room which is located in the centre of the vehicle can be had from the nearside or offside doors according to parking arrangements. Commentators and contributors gain access to the studio either via the vehicle cab or cubicle - the former causes minimum disturbance to the operation and is the normal way in. The studio is equipped with two TV monitor-receivers which allow for off-tube commentary when co-sited with BBC Television or one of the Independent Television Commercial Companies.

A specially designed 30-channel Glen Sound sound mixing desk which can handle 24 outside sources, each with individual talkback facilities, is the kernel of the unit. This large control desk is mounted longitudinally, and when the vehicle is operating in its alternative stereo music mode, it is withdrawn from the side wall on runners for improved monitoring conditions. The vehicle is acoustically treated and its dimensions allow for the accommodation of 2 disk reproducers and 4 professional tape recorders which can be remotely operated from the desk; these may also be used independently for editing, a useful facility for updating material when the programme is already 'on the air'.

The windows on either side of the studio are particularly large. This allows a 'fish bowl' view of the performers for PR purposes; however carpet-clad acoustic shutters make it possible to exclude all external light and public viewing when this is operationally desirable, or advisable from a security point of view.

Executive production control can be effected either from the studio or the cubicle in the conventional way, using a specially designed 24 channel communications facility. The radio link compartment, while fitted out and wired, is only equipped ad hoc to suit the particular operational requirement; the 10 metre pneumatic mast is however a fixed feature of the vehicle and can take a range of aerials. A small silent, well shrouded power generator (10 KVA) towed behind, ensures a degree of independence when faced with unexpected situations e.g. a long running hostage siege in a city centre or airport.

Mobile Studio - the Control Cubicle

A Type A Vehicle - the 40 channel SSL Control Desk

4. Music and Speech Programme Transmitting and Recording Vehicles

These come in three sizes viz.

The Type A: The latest versions of this vehicle have been equipped with

British Solid State Logic (SSL) computer assisted 40 channel mixers complete with total recall. (This control desk has also been selected for our many music studios, but the computer facility is only added where appropriate e.g. 'pop' and middle-of-the-road music). Three Studer A80 stereo reel-to-reel tape recorders and a 24 track A800 or Otari machine make up the rest of the major items. The computers and floppy disk drive unit are located at the front of the vehicle. On the outside are connections for up to 120 microphones. A colour monitor displays the SSL computer control settings while a closed-circuit television camera and monitor system is incorporated to allow the stage area to be viewed remotely.

The special body construction involves a triple 'skin' arrangement in which a jute based sound barrier mat is sandwiched between a glass-fibre filled outer cavity and an inner one containing fire-retardant absorbent foam. This latter layer is retained by perforated steel plate covered with an acoustically absorbent plastic material and carpet.

These facilities have been arranged to match the best of the equipment in our music studios, in our view a necessary requirement if the results are to be optimised under more difficult practical circumstances.

The Type B: For the majority of O.B. applications, the sophistication provided by the larger Type A control vehicles is not required. The Type B has a 40 channel Calrec manual mixing desk installed and no multitrack recorder.

The Type C: This, the smallest in the standard range of O.B. vehicles, is based on a Dodge 18 cwt van and has been designed to cover smaller, simpler broadcasts. It is equipped with two Glen Sound 6 channel mini mixers with simple response selection amplifiers and a pair of Nagra IVs stereo tape machines.

5. Foreign Commentators' Vehicle

In order to cope with those large show-piece events, ceremonial and sporting alike, which command such wide international interest, (this often increasing in an unpredicted way and at short notice as the date of the event approaches) BBC Radio has constructed a special vehicle which went into service the year before last. It is designed to cope with a multiplicity of

Foreign Commentators' Vehicle - Equipment Layout

Foreign Commentators' Vehicle - Commentator's Equipment commentators using the minimum number of technical staff in a flexible way. As examples of its use, large State Occasions, the Pope's visit, the Eurovision Song Contest, the World Cycling Championships, the Silverstone and Brands Hatch Motor Racing Grand Prix, Wimbledon, the F.A. Cup Final, European Football, and later the Commonwealth Games due to take place in Edinburgh in 1986 all figure in the list of such events.

The vehicle with its four specially designed Glen Sound engineering Control Desks and four sets of Producer's equipment, allows up to twenty commentators to be handled on the basis of one engineer per five within a fairly compact space, its overall length being only 8m (26' 3"). The actual units to be used by foreign broadcasters are portable and allow for a variety of different deployments to suit the various occasions and situations likely to be encountered in the field.

6. Multi-Track Recording Vehicle

For specialist recording applications, this vehicle can be equipped with two 24 track Studer A800 recording machines and may be used in conjunction with one of the Type A vehicles for the continuous recording of long 'pop' concerts, M.O.R. music and Country and Western festivals. To ease installation and temporary removal of one of the machines, the vehicle is fitted with double rear doors and a tail lift.

Multi-Track Recording Vehicle

Digital Recording Vehicle

7. Digital Recording Vehicle

While BBC Radio has been using digital recording techniques for $4\frac{1}{2}$ years now in a limited way, it has concentrated its activities in the area of experimental outside broadcasts. The small digital recording vehicle was first equipped in 1979 with the Sony PCM1600, 'U' Matic rotating head system and is now enjoying a period with second generation equipment - the $\frac{1}{4}$ inch tape, fixed head Mitsubishi Telefunken X80 range fed by one of the versatile 6 channel Glen Sound mixers and associated DK2/21 stereo monitoring facilities unit.

The truck used is a modified version of the small Type C vehicle, which with its extended roof space makes standing up possible.

The unit has been used extensively throughout the U.K. and has already journeyed across the English Channel to the Continent to record there as well.

More recently, a third generation of digital recording equipment, this time the portable Sony PCM F1 audio digital coder system using the Betamax VCR machine, has been in use for two main applications. Firstly as a straightforward, small lightweight digital stereo recording method in O.B. situations where e.g. air portability is a requirement for overseas music recording; a successful series of organ recitals was undertaken in Spain, and other programmes in France and New York. Secondly as a method of providing stereo contributions over a video circuit. The most notable examples of this have been the 'live' satellite transmissions to NPR in Washington of the 1982 and 1983 "Festival of Nine Lessons and Carols" from Kings College, Cambridge, England, on Christmas Eve, this being accomplished digitally end-to-end.

8. 'Radio 1 Roadshow'

A special articulated trailer has been converted to carry this disk jockey 'pop' programme-and-interview show quickly to more than thirty different locations during the short summer period between the end of July and September. At other times the unit features at public shows in selected towns or at sports events such as motor racing. It presents a young, modern image to our listening public and has proved very popular.

A side of the trailer opens out to form a 3 metre deep stage where the D.J. performs in full view of the public, simultaneously feeding the Network and the in-board P.A. system. The stage's operation is hydraulic, the floor and roof moving together.

Radio 1 'Roadshow' - Trailer

Radio 1 'Roadshow' - D.J. in action

The control cubicle is at one end and a small hospitality room at the other, these between them taking up the rest of the space. The brightly painted trailer is pulled by a Range Rover in the same livery while a second similarly be-decked combination converts into a Radio 1 sales promotion mobile shop. On stage the disk jockey has EMT 950 turntables, ITC/SP cartridge machines for 'jingles' and programme 'trails' and a self op mixer, while a radio microphone allows he or an assistant to move among the many people who watch the show, interviewing them and taking record 'dedications'. The stage can also be set for competitions, etc. involving members of the audience. The operation is engineer-assisted.

9. Radio Links Vehicles

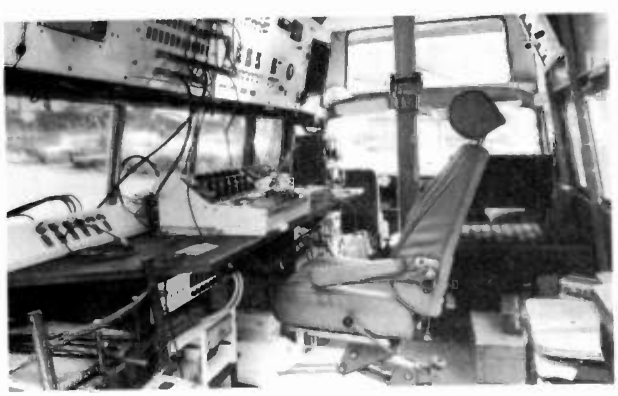
Certain programme situations are complicated enough in terms of radio link requirements to warrant a Base Station or communications centre being installed in a separate truck e.g. those calling for the use of 2 or 3 man-packs.

In others there is a need for an unattended radio link relay station to be set up in order to permit the range of the basic man-pack equipment to be extended e.g. away out at the far end of a hilly golf course. There are other occasions calling for the deployment of a second, distant commentator e.g. at the start of a horserace on the remote side of the course where the most economical and satisfactory way of establishing the necessary programme, cue and control circuits is by means of radio links. A fourth requirement involves 'live' transmission into one of the Networks while actually on the move, something that happens about half a dozen times a year and involves major, important strands of Radio's output.

Finally certain programmes, which are normally studio-based, are taken out-and-about a few times each year and may involve the personality Presenter being in five or six different locations within a city during his two hour segment e.g. in the market place, visiting a hospital ward, talking to fishermen at the docks, oil rig men at the helicopter base, etc. Good mono circuits can be provided by the employment of radio links for such 'live' broadcasting of this special and immediate kind. The production technique adopted is to link from the O.B. site first by radio and then via PTT circuits to London and have the disks or taped music played out in stereo hundreds of miles away from the venue over the permanent PCM distribution system. By this means, expensive stereo contribution circuit costs are saved while the quality of the composite output presented to the listening public remains very high.

For these five quite different tasks, a pair of custom built vehicles have been fitted out and equipped which are able to cope pretty well with any eventuality. The gear encompasses links operating in VHF Bands I, II and III, UHF Bands IV and V along with associated simple mixing equipment, but the communications control equipment has been specially designed.

The truck is a compact 15 cwt van complete with a host of aerials and a 9 metre pneumatic mast. Since the link equipment may be configured for each specific programme application, these vehicles represent a highly versatile, flexible pair of facilities.



Radio Link Vehicle

Digital Stereo Control Vehicle - Exterior

10. Digital Stereo Control Vehicle

The equipment in this truck when it comes into service during 1984 will represent the BBC's first attempt at experimenting with an all-digital sound control desk. The venture started in our Research Department a number of years ago when a novel digital computer processing system COPAS, was developed.

Subsequent to this, Neve have incorporated it in their new DSP 48 channel mixing desk. To be able to thoroughly test the equipment and associated new mixing techniques called for in a wide variety of operational conditions, it was decided to vehicle-mount the gear. This will allow its deployment anywhere throughout the U.K. or Western Europe where there is considerable interest in the project. It has the added advantage of being able to be used on a 'drive-in' basis with a wide range of studios and if required, operate in parallel with analogue equipment. The control console may be separated from the equipment bays by 1 kilometre of optical fibre cable.

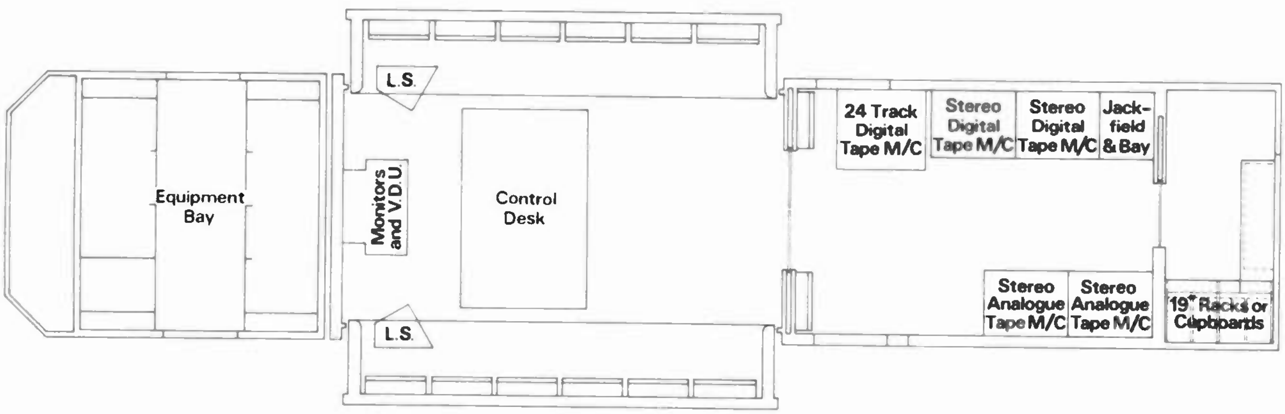


Fig. 2 Digital Sound Control Vehicle Layout

The 20 ton truck is itself of some interest; its length is 12.2 metres (40 ft) and width 2.5 metres, but the central section of the body expands out equally on both sides to 3.7 metres (12 ft) to afford better listening conditions for the sound balance. To the rear of this expanded control room is the equipment area.

The main features of the system are direct digital interfacing to digital recording and transmission systems, fully assignable controls, complete operational flexibility which allows the sound engineer to customise his own console (and before every session if he so wishes), 'snapshot' memory and time-code synchronised total automation available on all controls including faders, equalisers, limiter compressors, pan pots and routing.

The author wishes to thank the Managing Director of Radio, Richard Francis and the BBC's Director of Engineering, Bryce McCrirrick for permission to publish this Paper.

References: i) "BBC Radio Taxis and their Operational Role", J.D. MacEwan, Chief Engineer, Radio Broadcasting, BBC I.B.C. 1976
ii) BBC Eng. Inf. Quarterly No. 12 Spring 1983
iii) "The Digital Mixing Console", Dr. Martin Jones of Neve, Studio Sound October 1983

Spurious Emission Measurements Of Common Site FM Transmitter Installations Using Separate Co-located Antenna Systems

Warren G. Shulz

WFYR Division
RKO General Inc.
Chicago, Illinois

ABSTRACT- When two or more FM broadcast transmission systems share a common site with co-located antenna systems a degree of cross-coupling can occur between the systems. When this cross-coupling becomes excessive the transmitter final amplifiers can produce out-of-tolerance, non-harmonic related emissions. The frequencies and levels of these emissions are related to the amplitude and frequency of the cross-coupled transmission system(s).

This paper reviews spurious emission data collected from sites that have two or more transmitters co-located using separate but, adjacent antenna systems less than 200-feet apart. The data collected reviews performance using both bandpass and band reject (notch) filter methods. A discussion is presented showing a method to measure undesired products removed less than 1% from the carrier frequency.

The results show that the transmitter system can be successfully co-located using a separate antenna system(s) mounted near each other and be in compliance with FCC specifications when proper filter systems are applied.

The Problem

Transmission standards specify bandwidth occupancy limits for FM Broadcast systems. These are established by FCC rules and appear as:

Frequency				Suppression
F_c	to	+/-	120Khz	None
+/-120	to	+/-	240Khz	-25dB below carrier
+/-240	to	+/-	600Khz	-35dB
+/-600	to	+/-	infinity	-80dB *

*Minor relief will apply for carrier levels less than 5-kw as follows: In the event 43 + 10 Log_{10} (watts) equals a number less than 80dB it can be applied for suppression values for carrier frequencies greater than 600Khz removed from carrier. For example a 3 kw carrier requires a suppression of 77.8dB for products greater than 600Khz from carrier. The 80dB suppression value equals to a voltage distortion specification of 0.01%.

The manufacture insures, by proper design, that the transmitting equipment will meet these requirements as a functional system design. However, many of the transmitter equipment manufactures fail to specify the performance of the system in the presence of moderate levels (10 to 20 dBm) of in-band returning signal level back into the final amplifier's output stage. An excellent paper on this topic was prepared by Mendenhall (1). This paper clearly shows the effects of cross-coupled signals 40 and 60dB below the carrier level. The effects of cross-coupling are to produce new in-band intermodulation products of amplitudes in excess of 80dB below carrier reference. The levels and frequencies of the product(s) are dependent on the cross-coupled level and transmitter(s) carrier frequency.

Mechanics Of Intermodulation Generation

As in an audio system radio frequency intermodulation is the generation of new components with frequencies equal to the sums and differences of the multiples of the fundamental components. This theory can be reviewed in text books that discuss the mixer designs for receivers.[2] In the situation of FM broadcast tranmitters the fundamental signal from the loosely coupled external transmitter mixes with the carrier and harmonics in the final amplifier stage to create new in-band products. As Mendenhall reports the typical suppression techniques applied for harmonic content control offer little opposition to inband products from passing into or out of the final stage amplifier. A sweep study made of a Harris Corp. FM25K low-pass filter showed little attenuation below 150Mhz. The general equations for ascertaining possible intermodulation (IM) produts are listed in the Table 1.

The general expression is:

$$F_{im} = \pm n F_1 \pm m F_2$$

Where n,m = 1, 2, 3, ... with n + m = The Order

The lowest order will have the highest amplitude and the IM level will decrease in level as the order is increased. In particular, for an FM transmitter the level of the IM product would be at its worst nearest the operating frequency of the transmitter. This would appear as the sum and difference about the 2nd harmonic of the transmitter's carrier frequency. This product, 2A-B, is defined as a 3rd Order product.

Table-1 General Equations For Intermodulation Products

2nd order:	F_{im}	=	$\pm$	F_A	$\pm$	F_B	eq. 1
3rd order:	F_{im}	=	$\pm$	$2F_A$	$\pm$	F_B	eq. 2
	F_{im}	=	$\pm$	$2F_B$	$\pm$	F_A	eq. 3
4th order:	F_{im}	=	$\pm$	$3F_A$	$\pm$	F_B	eq. 4
	F_{im}	=	$\pm$	$3F_B$	$\pm$	F_A	eq. 5
5th order:	F_{im}	=	$\pm$	$3F_A$	$\pm$	$2F_B$	eq. 6
	F_{im}	=	$\pm$	$3F_B$	$\pm$	$2F_A$	eq. 7
	F_{im}	=	$\pm$	$4F_A$	$\pm$	F_B	eq. 8
	F_{im}	=	$\pm$	$4F_B$	$\pm$	F_A	eq. 9
6th order:	F_{im}	=	$\pm$	$4F_A$	$\pm$	$2F_B$	eq. 10
	F_{im}	=	$\pm$	$4F_B$	$\pm$	$2F_A$	eq. 11
	F_{im}	=	$\pm$	$5F_A$	$\pm$	F_B	eq. 12
	F_{im}	=	$\pm$	$5F_B$	$\pm$	F_A	eq. 13

Discussion Of Intermodulation Products

The general equations for determining intermodulation products through the 6th order can be reduced when the following items are taken into account: 1) Only the fundamental frequency of the interfering signal is going to make its way back to the transmitter's final stage, 2) Products removed by one octave above and below the carrier frequency can be discarded, and 3) Products appearing closest to the carrier frequency will have the highest amplitude.

After applying the 3 'rules' the lengthy list of equations 1 through 13 can be reduced to those shown in Table 2. While Table 2 gets to the point it is a good practice to run through the list generated from the 13 equations of Table 1. A numeric table of possible products can be prepared listing possible spurious emissions, carrier frequencies, and 2nd harmonic frequencies. This list can be used as a reference when making the measurements on site. In the practical world, the 3rd order product was always the most vigorous; in no case was 2nd order products found.

Table-2 Spurious Emission Products Of Major Concern

For Transmitter A (where A is less than B):

$F_{im} = 2A - B$ (3rd order in A from B) eq. 14

$F_{im} = 3A - 2B$ (5th order in A from B) eq. 15

$F_{im} = A + B$ (2nd order A/B) eq. 16

For Transmitter B (where B is greater than A):

$F_{im} = 2B - A$ (3rd order in B from A) eq. 17

$F_{im} = 3B - 2A$ (5th order in B from A) eq. 18

$F_{im} = B + A$ (2nd order A/B) eq. 19

Conversion Efficiency

The characteristic of a power amplifier to generate intermodulation products appears related to the level of the interfering signal, frequency spacing of the interfering signal, and bandwidth of the transmitter. It is this author's opinion that the bandwidth of the PA grid circuit has a direct bearing on the 'turn-around-loss' for the IM product amplitude level. If this hold true then the final stage tank circuit is a LC filter and the loading control varies the Q (bandwidth) of this 'filter'. The plate loading control has been observed to affect the turn-around-loss of the IM product level. The range is limited and as one approaches higer turn-around-loss the bandwidth of the tank circuit decreases and the efficiency of the amplifier falls. Thus, this adjustment provides a very limited range on the turn-around-loss of the amplifier. In some cases this slight adjustment can mean the difference between making the 80dB specification or not.

In general ground grid amplifiers, and recent tetrode design amplifiers (optimized for broad bandwidth) exhibit turn-around-losses on the order of 11dB for an interferring signal space 800Khz away. Whereas the old design tetrode amplifiers have been measured in the 30dB area for the 800Khz spacing. Do to the many variables it is unfair to establish specific levels without conducting a measurement of the actual transmitter. Mendenhall's paper provided specific information on the Broadcast Electronics model FM-30, 30 kw transmitter. This is the type of information needed to design, in advance, filtering systems that avoid any last minute surprises!

Steps To Document Spurious Emissions

After preparing a list of possible spuri products from equations 1 through 13 a method to measure these values is needed. A directional RF probe is used to measure the RF voltage levels within the transmission line to the antenna. (It is anticipated the spurious emission levels would be in near vicinity of the carrier and would be efficiently radiated by the station's antenna system.)

For the data presented a Bird Corp. model 552-75 RF probe was used in the Bird Corp., model 460 Thru-Line^R section. The performance of the probe was verified with a tracking generator and a spectrum analyzer. Over the range of 50 to 300 Mhz the probe's response variation was +/- 1dB and over the range of 75 to 150Mhz the probe's response was within a range of +/- 0.5dB. Figure 1 shows the response results.

With the probe's directivity it is possible to ascertain the source of the signal. IM products generated within the transmitter would have directivity towards the antenna termination. Products flowing from the antenna have no termination when looking towards the transmitter; thus these products exhibit the same level forward and reverse. An IM product will show up as an audio mix of both stations when monitored on a receiver connected to the RF probe. The 2nd harmonic will appear over deviated on the receiver.

Notch Filters

Suppression of the carrier is required to expand the useful range of the selected receivers to permit measurements in excess of 80dB below the carrier level. While a bandpass filter could accomplish this the filtering is difficult to tune and calibrate for each measured frequency. A carrier notch filter is most effective. A 3-cavity, ¼-wave, notch filter manufactured by DB Products, Dallas, Texas was selected. This filter set produces a notch, at carrier, in excess of 55dB. Calculated Q was 300 and the -3dB points are +/- 300Khz about the notch center. The filter set is tuned with threaded rod elements that run about 2Mhz per inch of adjustment. The coupling loops are fully rotable to control notch depth. The filter set loss is shown in Figure 2. This particular filter set has a usable range of 70 to 250Mhz with a response error of +/- 1dB which will require factoring into the final measurementresults.

Receivers

Two commonly found receivers were selected. 1) a tunable field strength receiver with a range of 50 to 225Mhz, mfg. by Potomac, model FIM-71, and 2) an RF spectrum analyzer mfg. by Tektronix, model 7L13. The FIM-71 has a usable dynamic range of 60dB, and the 7L13 has a dynamic range of about 75dB. In the case of the FIM-71 it is possible to read the level in volts and avoid any log amplifier generated errors as used in the 7L13. The field strength meter has a distinct advantage with the ability to make audio identification of the signal. With the application of the notch filter to provide 50dB of carrier suppression the dynamic range of the receiver is expanded to more than 100dB below carrier! It is important to verify the IM product being measured is external to the receiver. This is easily checked by introducing a specific amount of loss in the receiver input and watching for an identical change in the IM level. If this test fails then the IM product is being generated within the receiver.

Two Transmiters Sharing A Common Site

In this first example two transmitters spaced 800Khz apart are co-located at the same site using separate multi-element antenna systems. The facility list is shown in Table-3.

Table-3 Facility List; Prudential Site

Call Sign	Frequency	TPO	EPR	Antenna Type
WFYR	103.5Mhz	20 kw	17 kw	2-element, Harris,FMC-2
WJEZ	104.3	9.5	18.5	4-element, RCA, BFC-4

The antennas for these two stations area spaced by about 50-feet of vertical distance and mounted on opposing sides of the support structure. Cross-coupling between the two separate antenna systems was a nominal 45dB. In this case, notch filters were employed to reduce the level of the undesired signal appearing in the station's final amplifier stage. In working the frequencies of the two stations through the set of equations (1-13) for spurious emission products the possible mix products anticipated are:

For 103.5Mhz (Transmitter A)

2nd order:

A + B = 207.8Mhz

3rd order:

2A - B = 102.7

4th order:

3A - B = 206.7

5th order;

3A - 2B = 101.9

For 104.3Mhz (transmitter B)

2nd order:

A + B = 207.8Mhz

3rd order:

2B - A = 105.1

4th order:

3B - A = 209.4

5th order:

3B - 2A = 105.9

Re-stating these products, carrier frequencies, and harmonics in numeric order a table of anticipated products can be generated that will assist in identification of desired and spurious emissions from a site. This is shown in Table-4.

Table-4 WFYR/WJEZ Co-Located Emissions

101.9	5th order in WFYR from WJEZ
102.7	3rd order in WFYR from WJEZ
103.5	carrier WFYR
104.3	Carrier WJEZ
105.1	3rd order in WJEZ from WFYR
105.9	5th order in WJEZ from WFYR
206.2	4th order in WFYR from WJEZ
207.0	2nd harmonic WFYR
207.8	2nd harmonic WFYR/WJEZ
208.6	2nd harmonic of WJEZ
209.4	4th order on WJEZ from WFYR

This list assumes harmonics of the interferring signal do not appear within the power amplifier of the transmitter being measured. Based on the low-pass filter applied by the transmitter's manufacture products above the transmitter's 2nd harmonic carrier are ruled out. Furthermore, if the 3rd order product is within or near -80dB below carrier the amplitude of the higher order products will be lower and would have lesser amplitude than the 3rd order products nearest the carrier frequency.

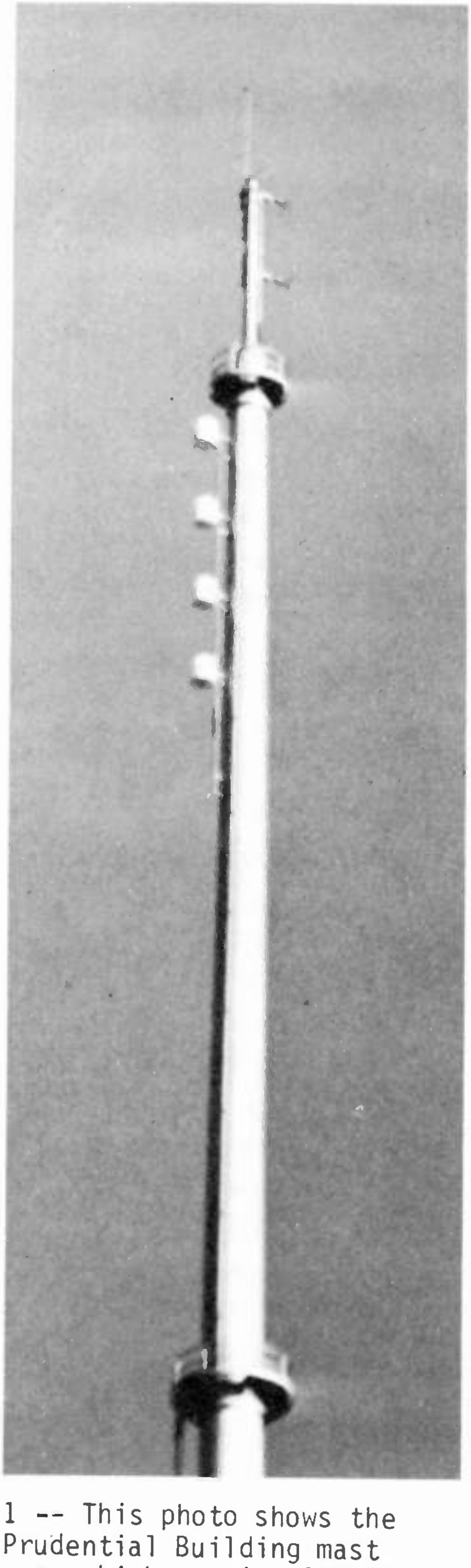
In the initial measurements the two transmitters were operating within compliance of spurious emission requirements. It was the desire of WFYR-103.5Mhz to increase its ERP from this auxiliary site to maintain its full class-B service area. Figure 3, a spectrograph, shows in an amplitude vs. frequency display the initial conditions. The 0dB reference is the carrier level of the WFYR transmitter. The spurious emission at 102.7Mhz was measured -85dB below the WFYR carrier reference. The turn-around-loss of the Harris FM20H2 transmitter was 39.5dB.

Figure 4, a spectrograph, shows the WFYR transmitter after the TPO was increased from 15 to 20 kw (indirect method). This 1.2dB increase was obtained by increasing the PA gain (higer screen voltage, and re-tuning). Under this condition the spurious emission at 102.7Mhz increased 13dB to a level of -73dB below the carrier reference. This clearly shows the need to apply filtering to lower the 3rd order product to a value less than -80dB.

An Electronics Research type 902 notch filter was installed on the transmission line of the WFYR transmitter. This filter was adjusted to pass 103.5Mhz (loss 0.06dB0 and notch 104.3Mhz (depth was -13dB) with the net result of improving the antenna isolation from 44 to 57dB. Figure-5, a spectrograph, shows the resulting performance. The spurious emission (3rd order product at 102.7Mhz) was lowered by 26dB and the turn-around-loss was improved by 10dB. The spurious emission was -97dB below the carrier reference. (Author-It would appear, in this case, the relationships between the interferring signal, turn-around-loss, and IM product level are a non-linear relation.)

Figure-6, a spectrograph of the opposing transmitter, WJEZ (104.3Mhz) operating at the same site. In this case the WJEZ transmitter had an ERI type 902 notch filter installed several years earlier when co-location of the two transmitters was started. The filter provided a notch depth of 16dB at the interferring carrier frequency of 103.5Mhz. In this case the turn-around-loss was 17.5dB and the resulting 3rd order product on 105.1Mhz was -80.5dB below the carrier reference.

Figure 7 shows the performance of the ERI 902 notch filter. Bandpass was measured to be 200Khz for a VSWR rating of 1.07. The insertion loss of the filter was verified to be 0.06dB. As a side note, in the case of the 102.7 and 105.1 carrier frequencies both were assigned to stations within 1 mile of the co-located site. In order to measurelevels of these products the stations on these channels must break carrier. The levels received from the transmitting antennas are about 80dB below the carrier reference.

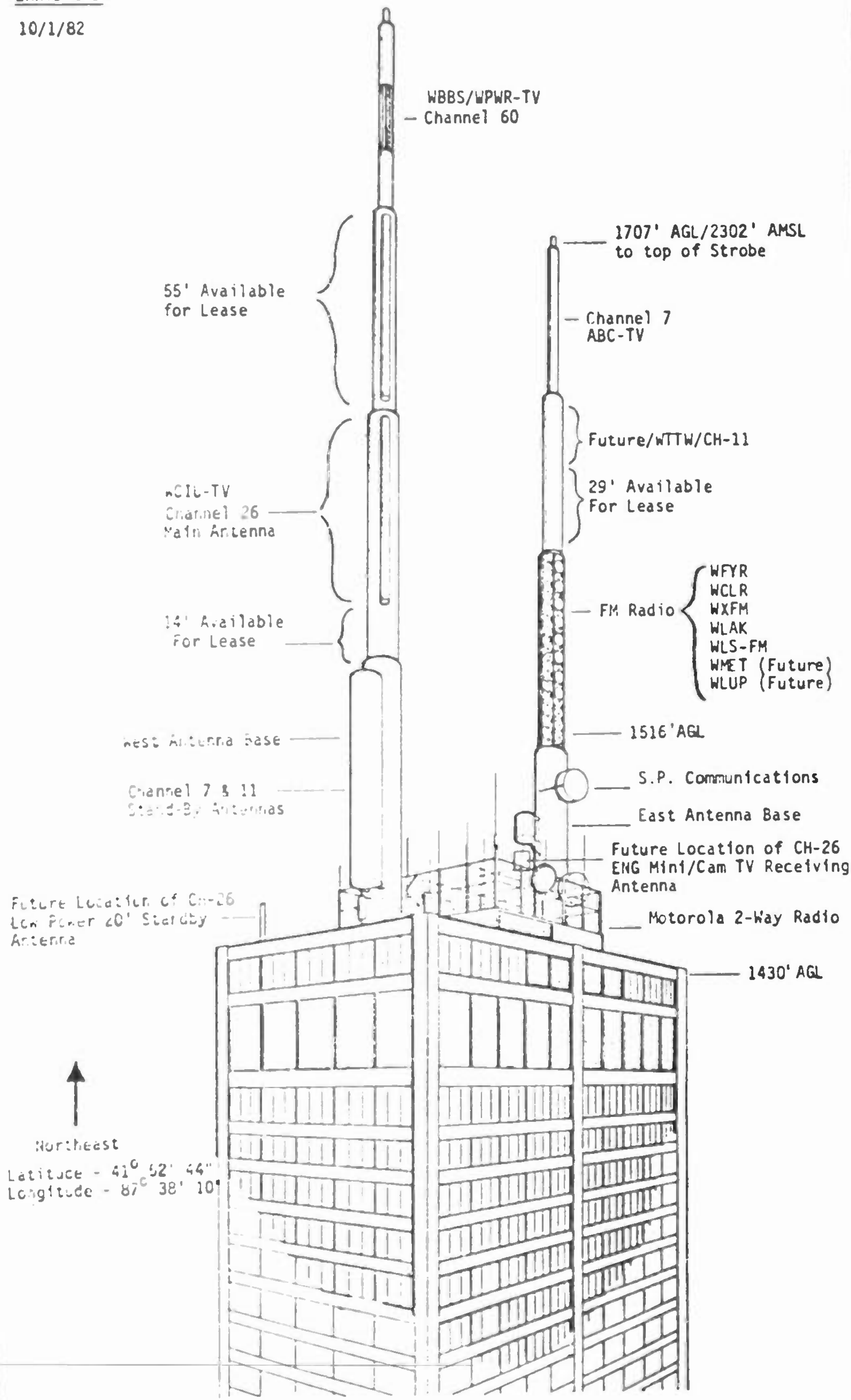


1 -- This photo shows the Prudential Building mast upon which stations' WFRY and WJEZ have co-located two separate antennas on a common structure.

1A -- This picture is the FM section of the east mast atop the Sears Tower Building. All five FM broadcast antennas are visible prior to radome installation.

Exhibit B

10/1/82



2 -- This photo shows the placement of various TV/FM and other related operations atop the Sears Tower Building, Chicago.

Five Transmitters Sharing A Common Site

In this example five stations share the same site with separate co-located antennas in a vertical space of 70-feet. Each of the five stations operate Class-B facilities with a typical antenna height of 1,560-feet, and a typical ERP of 4.3 kw. Photo-2 shows the five antennas in-place on the tower section prior to the installation of the radome covers. This tower section is located on the EAST mast atop the Sears Tower Building, Chicago. Each station's antenna consist of 3-pairs of dipoles arranged one pair to each side mounted to a triangular, 8-foot face tower section. The Cavity Backed Radiator (CBR)R antenna, manufactured by Harris Corp., are more commonly found in use for TV broadcasting antennas. Table-5 shows the facility list for FM broadcast operations from the Sears Tower.

Table-5 Sears Tower FM Broadcast Facility List

Station	Frequency	ERP	TPO	HAAT
WLAK	93.9Mhz	4.0 kw	9.9 kw	1,581 feet
WLS-FM	94.7	4.4	10.6	1,535
WCLR	101.9	4.2	10.5	1,561
WFYR	103.5	4.3	10.5	1,548
WXFM	105.9	*4.1	6.5	1,575

*Peak ERP (directional)

Sears Tower Spurious Emissions

Equations 1 through 13 listed in Table-1 are used to calculate the possible emission products relevant to the five transmitters sharing the Sears site. From these calculations a list of 73 IM products over a span of 70 to 223 Mhz is developed. Third order products number 22 and appear on 19 frequencies. The list of IM product is shown in Table-6.

Table-6 Sears Tower FM Broadcast Spurious Emission List

69.9	5th order WLAK/WXFM
72.3	5th order WLS-FM/WXFM
74.7	5th order WLAK/WFYR
77.1	5th order WLS-FM/WFYR
77.9	5th order WLAK/WCLR
80.3	5th order WLS-FM/WCLR
81.9	3rd order in WLAK from WXFM
83.5	3rd order in WLS-FM from WXFM
84.3	3rd order in WLAK from WFYR
85.5	3rd order in WLAK from WCLR
85.9	3rd order in WLAK from WCLR
	3rd order in WLS-FM from WFYR
87.5	3rd order in WLS-FM from WCLR
92.3	5th order WLAK/WLS-FM
93.1	3rd order in WLAK from WLS-FM
93.9	Carrier WLAK
	5th order WCLR/WXFM
94.7	Carrier WLS-FM
95.5	3rd order in WLS-FM from WLAK
96.3	5th order WLAK/WLS-FM
129.9	5th order WLAK/WXFM
175.8	4th order in WLAK from WXFM
178.2	4th order in WLAK from WFYR
	4th order in WLS from WXFM
179.8	4th order in WLAK from WCLR
180.6	4th order in WLS from WFYR
182.2	4th order in WLS from WCLR
187.0	4th order in WLAK from WLS
187.8	2nd harmonic WLAK
188.6	2nd order WLS-FM/WLAK
189.4	2nd harmonic WLS-FM
190.2	4th order in WLS from WLAK
195.8	2nd order WLAK/WCLR
196.6	2nd order WLS/WCLR
197.4	2nd order WLAK/WFYR
198.2	2nd order WLS-FM/WFYR
199.8	2nd order WLAK/WXFM
	4th order in WCLR from WXFM
200.6	2nd order WLS-FM/WXFM
202.2	4th order in WCYR from WFYR

97.9	3rd order in WCLR from WXFM
98.7	5th order WCLR/WFYR
100.3	3rd order in WCLR from WFYR
101.1	3rd order in WFYR from WXFM
101.9	Carrier WCLR
103.5	Carrier WFYR
105.1	3rd order in WFYR from WCLR
105.9	Carrier WXFM
106.7	5th order WCLR/WFYR
108.3	3rd order in WXFM from WFYR
109.1	3rd order in WLS-FM from WCLR
109.9	3rd order in WLS-FM from WLAK
	3rd order in WCLR from WLAK
	3rd order in WXFM from WCLR
110.7	5th order WFYR/WXFM
112.3	3rd order in WFYR from WLS-FM
113.1	3rd order in WFYR from WLAK
113.9	5th order in WCLR/WXFM
116.3	5th order WLS-FM/WCLR
117.1	3rd order in WXFM from WLS-FM
117.9	3rd order in WXFM from WLAK
121.1	5th order WLS-FM/WFYR
122.7	5th order WLAK/WCLR
	5th order WFYR/WLAK
128.3	5th order WLS-FM/WXFM
203.8	2nd harmonic WCLR
204.6	4th order in WCLR from WFYR
	4th order in WFYR from WXFM
205.4	2nd order WCLR/WFYR
207.0	2nd harmonic WFYR
207.8	2nd order WCLR/WXFM
208.6	4th order in WFYR from WCLR
209.4	2nd order WFYR/WXFM
211.0	4th order in WCLR from WLAK
	4th order in WCLR from WLS
211.8	2nd harmonic WXFM
	4th order in WCLR from WLAK
214.2	4th order in WXFM from WFYR
215.8	4th order in WFYR from WLS
	4th order in WXFM from WCLR
216.6	4th order in WFYR from WLAK
223.0	4th order in WLAK from WXFM
	4th order in WXFM from WLS

2nd order	- 10 products
3rd order	- 22 products
4th order	- 22 products
5th order	- 19 products
total	- 73 products

Stacking Plan And Filtering Equipment

Harris Corp. was selected as the supplier for the antenna equipment and all filtering equipment to control IM products at the Sears Tower site. Two items addressed were: 1) The stacking plan was optimized to reduce the filtering requirements, and 2) bandpass filters were employed on each transmitter system. The bandpass filter restriction provides losses to adjacent antenna signals and returning losses to the antenna fcr spurious products generated within the transmitter's output stage. At the same time the bandpass filter provides a suitable 'window' for the modulated signal of the transmitter's output. Figure-8 shows the measured VSWR and amplitude performance cf the WFYR filter. This filter, sub-contracted to and built by Dielectric Communications, is a 3-section unit with each section 28 X 60 inches in size.

The system specification for cross coupling of antenna systems, as measured from any antenna/filter to any other was 56dB loss minimum. (Isolation was measured as a terminated voltage into a 50-ohm load with all unused lines open.) The 56dB loss is made up of two components: 1) antenna isolation (typical end-to-end loss was in the area of 45dB), and 2) filter losses removed from carrier pass frequency (800Khz spacing loss is in the area of 11dB). Thus, for two stations spaced 800Khz, end-to-end on the structure, the 56dB isolation loss is achieved. If we add an estimated turn-around-loss of 18dB, 11dB of return loss through the filter to the antenna, the 3rd order product level on the antenna side of the filter is in the area of 85dB below carrier level for the 3rd order product.

Of course, as antenna spacing is reduced the cross coupling loss decreases. However, with increased frequency spacing higher losses occur when the filter losses are factored into the system. In addition, the turn-around-losses of the transmitter increase as the signals take on a larger spread.

It is this technique that makes the site usable for the five stations. Figure-9 shows the spectrograph for the WFYR transmitter located at the Sears site in the presences of the four adjacent interfering signals. The dashed line marked as, "filter loss" is the inverse response of the bandpass filter.

Measurements At Sears Tower

The previously discussed techniques can now be applied to document any spurious emissions found at the Sears site. Due to space limitations only the WFYR transmitter data will be reviewed. Table-7 shows the data collected from the FIM-71 field strength meter (with applicable notch filters).

Table-7 Data From FIM-71 With/Filters Prior to bandpass Filter

Field Strength Meter	dB Below Carrier*	Frequency	Remarks
23 uV	-95.0 dB	93.9 Mhz	WLAK carrier
90 uV	-83.1	94.7	WLS carrier
540 uV	-67.6	101.9	WCLR carrier
270 uV	-73.6	105.9	WXFM carrier
60 uV	-86.6 #	110.0	noise
16.5 uV	-97.9	207.0	WFYR 2nd Harm

*carrier reference no/filter = 1.30 V for 0 dB
#After filter level to antenna = -153.6dB below carrier

In making this data table-7, signals off-site (clearly identified), and those with a level lower than 100dB below carrier were not recorded. The spurious product at 110Mhz was noise. Upon review of Table-6 it was noted three 3rd order products appear at 109.9Mhz. While none of these are related to WFYR it appears as if a mix maybe be occurring in the WFYR transmitter. With the return loss through the filter the product at 110Mhz is suppressed beyond 150dB below the carrier reference on the antenna side of the filter. All the data collected was on the transmitter side of the filter.

To complete the measurements segemented sweeps were made with the spectrum analyzer. The 7L13 spectrum analyzer was substituted for the FIM-71. 50Mhz sweeps were made from 50 to 200Mhz and a narrow sweep 2Mhz about the carrier were documented. These four sweeps are shown in Photos-3,4,5, and 6. In all cases a notch filter was used to reduce carrier level. The top of screen reference is 40dB below carrier and -40dB on screen would indicate level -80dB below carrier. The signals appearing on Photo-6 +/- 800Khz from carrier were other transmitters from a site about 1 mile away. These signals could be clearly identified and heard on the FIM-71 receiver.

Field Measurements of Spruious Emissions

One may ask, is it possible to measure a spurious emission in the radiated field of the antenna? For all examples presented in this paper measurements were made using a directional RF probe in the station's transmission line. This was a matter of convenience and has been an accepted from of measurement for FCC submission of data collected. Recently, the author was invited to apply this paper's knowledge base at RKO General's station WAXY-105.9Mhz licensed to Fort Lauderdale, Florida. In this particular situation a question of 3rd order interference to an 800Khz (low) adjacent channel on the market was raised. RF voltage probe measurements showed this 3rd order product (caused by an

J. Bortowski	CE-WLAK/Viacom	D. Meyers	Harris Corp.
W. DiLucente	CE-WAXY/RKO	H. Priester	CE-WLS-FM/ABC
G. Capalbo	V-P Engr./RKO	N. Smith	Smith & Powstenko
G. Collins	Harris Corp.	R. Turner	CE-WCLR/Bonneville
B. Hoyt	Harris Corp.	T. Uyttendaele	ABC
J. Hurnie	CE-WJEZ/Plough	J. Valenta	CE-WXFM/Cox
G. Mendenhall	Broadcast Electronics		

References

(1) G. N. Mendenhall, "A Study of RF Intermodulation Between FM Broadcast Transmitters Sharing Filterplexed or Co-located Antenna Systems.", Broadcast Electronics Inc., Quincy, IL, (Presented at the IEEE Broadcast Symposium 1982)

(2) G. B. Gawler, "Frequency Converters and Detectors", ELECTRONICS ENGINEERING HANDBOOK, 1982, p. 14-57 to 14-69.

800Khz (high) adjacent on site transmitter) to be 94dB below the WAXY-105.9Mhz transmitter carrier level.

The field strength meter was used in conjunction with the filter set to measure the spurious emission in the field. For this situation the 3 notch filters were re-configured to a 2dB per filter insertion loss, bandpass filter tuned to pass 105.1Mhz. The station assigned to 105.1Mhz signed off and the level of the spurious emission at 105.1Mhz was determined. The 3rd order product appearing at 105.1Mhz from the WAXY-105.9Mhz transmitter was 90.5dB below the carrier level. The 3.5dB error, between the probe and field reading, appears due to a lack of sufficient filtering to protect the receiver in the field reading. However, the results in any case did show the emission was well below the 80dB suppression requirement. The field strength at this location was 118dBu for the WAXY signal. The receiving antenna used was that supplied with the FIM-71. This antenna was elevated about 10-feet above ground level. For anyone attempting to repeat this type of measurement I would suggest using 3 sections of bandpass and at least one section of notch filtering. This field reading was made in the near vicinity of five 100-kilowatt FM transmission systems!

Yes, field readings can be made if you are willing to tune filters. In the event you wish to measure a number of frequencies the filter tuning effort becomes a major problem not to mention the required shutdown of off-site broadcast operations to ascertain if the frequency is clear of any emissions. The directional characteristic of the voltage probe would yield the same confidence to ascertain the source of the spurious emission.

Conclusion

The procedures and examples presented show methods to measure and control spurious emission products when FM broadcast transmitters share a common site. All the information shown was collected from operating systems and in some cases the data was submitted as a part of the station's filing with the FCC licensing applications. The measuring procedures show a method to use the FIM-71 field strength meter to make measurements within 1% of the carrier frequency to levels 100dB below the carrier.

For the future, transmitter manufacturers need to provide data on the performance of their transmitter equipment for conditions when in-band signals are returning into the output circuit. It may be possible to configure a notch filter in the final stage input circuit to 'trap' the offending signal, thereby eliminating the need for high-power filtering circuits on the transmitter output.

As in any work of this type, and was pointed out earlier, each transmitter can perform in a nonpredicative manner in the presence of in-band returning signals. While general estimates of actual performance can be made it is difficult to predict exact numbers in advance. The phrase, "Give me your checkbook and I will return it when I am done" may well apply.

Acknowledgements

The author wishes to thank the following people for their assistance and support in gathering data for this paper:

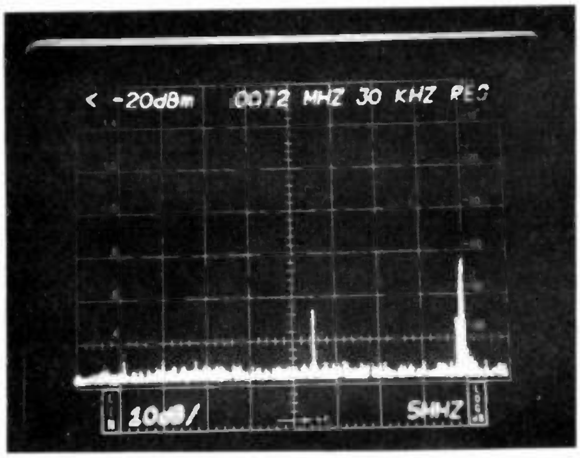


PHOTO - 3 Sweep 50 to 100 Mhz
Reference = - 40 dB
WFYR, Sears Site

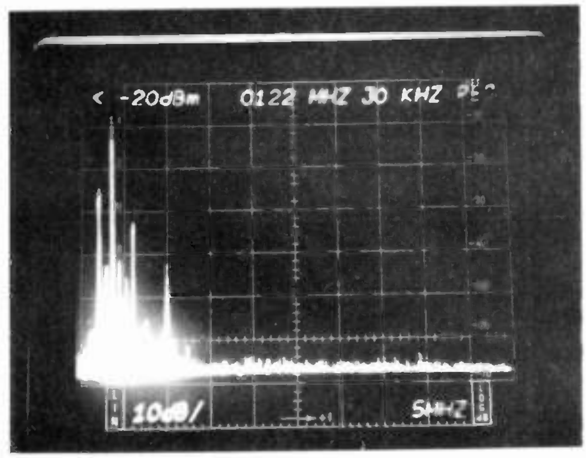


PHOTO - 4 Sweep 100 to 150 Mhz
Reference = - 40 dB
WFYR, Sears Site

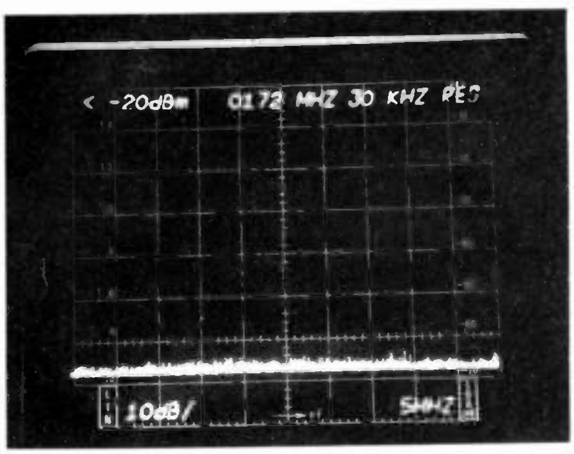


PHOTO - 5 Sweep 150 to 200 Mhz
Reference = -40 dB
WFYR, Sears Site

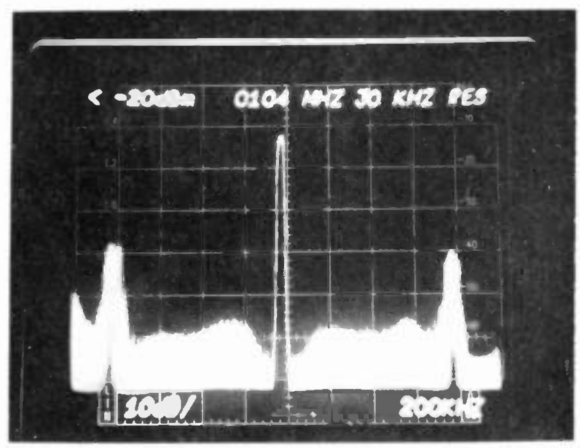


PHOTO - 6 Sweep from 103 to 104 Mhz
Reference = - 40 dB
WFYR, Sears Site

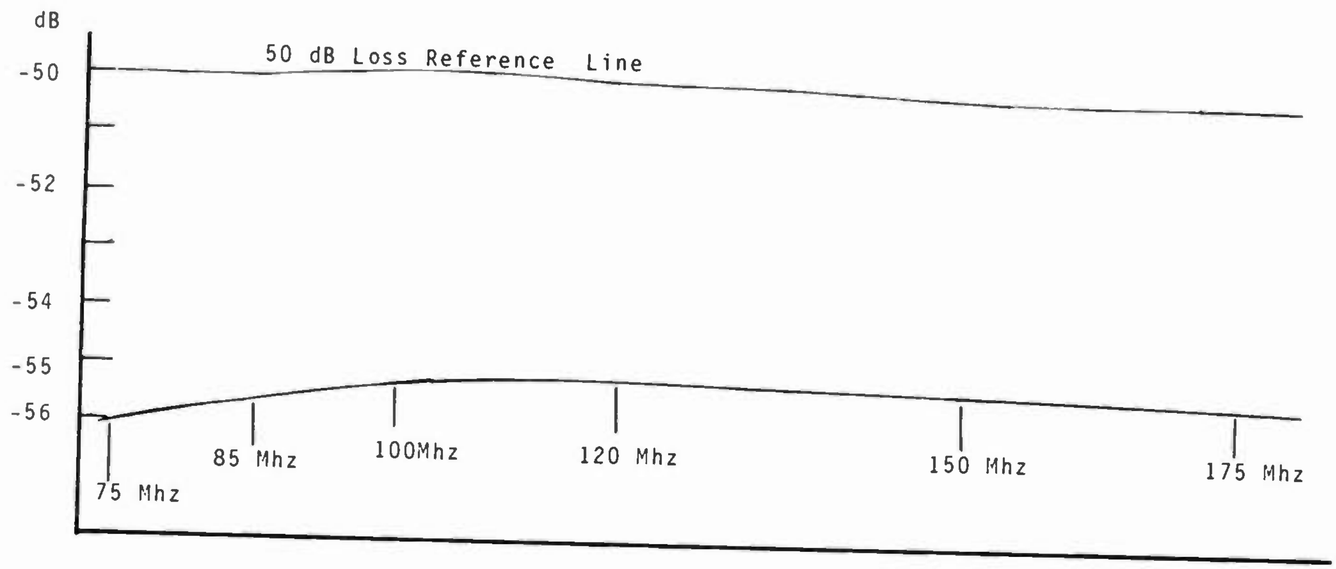


FIGURE - 1 Frequency response of Bird type 553-75 RF Sample probe in a Bird 460 Thru-Line[R] single element section.

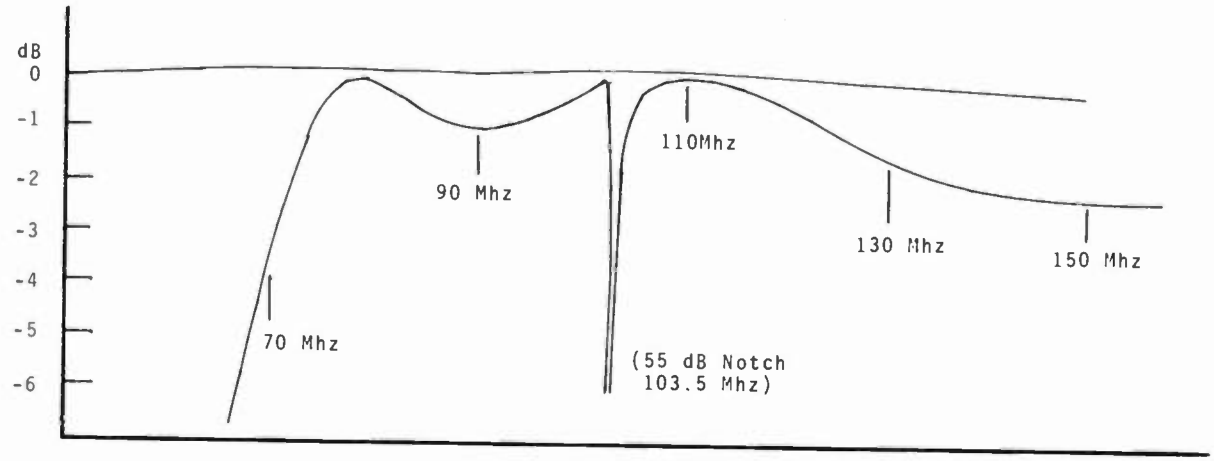


FIGURE - 2 Frequency response of 3-cavity filter set tuned to notch 103.5 Mhz.

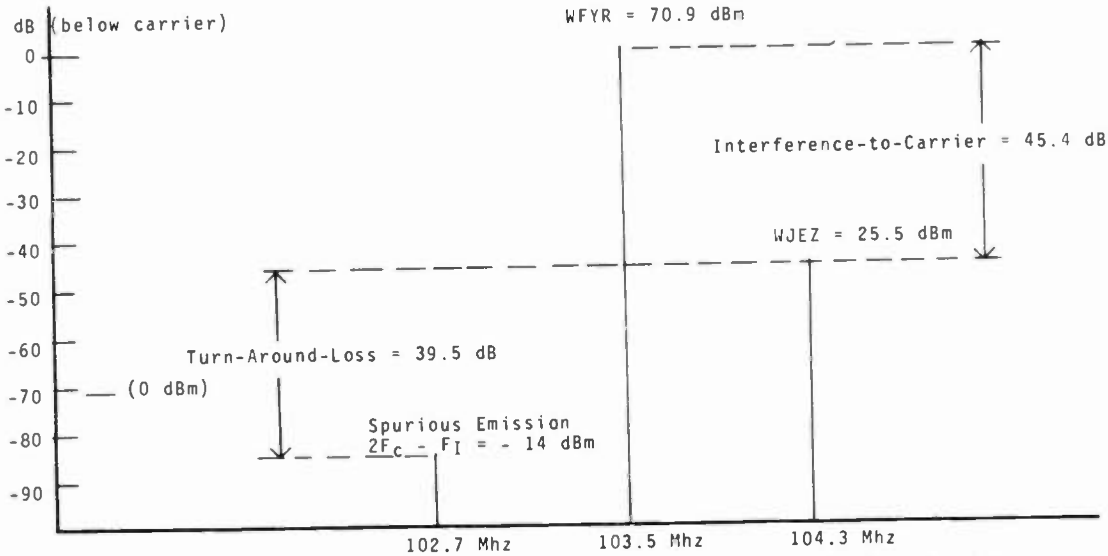


FIGURE - 3 Spectrograph, WFYR Auxiliary transmitter - Prudential Site operating with 15 Kw TPO, no filtering and in compliance.

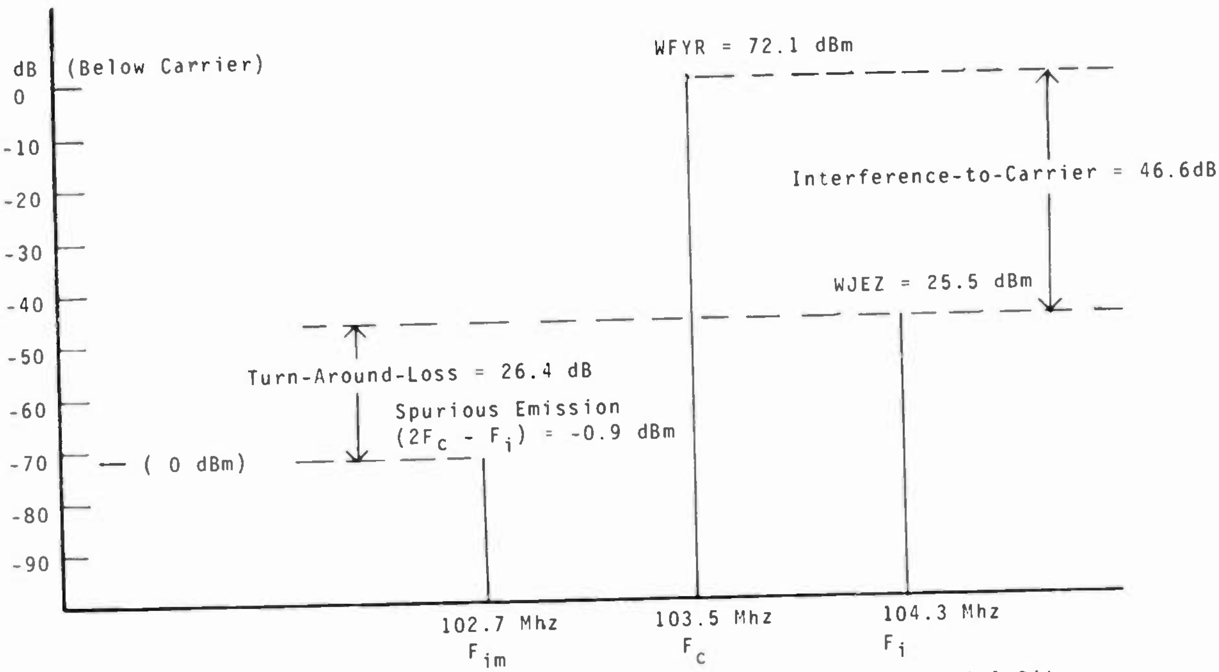


FIGURE - 4 Spectrograph, WFYR auxiliary transmitter - Prudential Site operating with 20 kw TPO, no filter, and out-of-Compliance

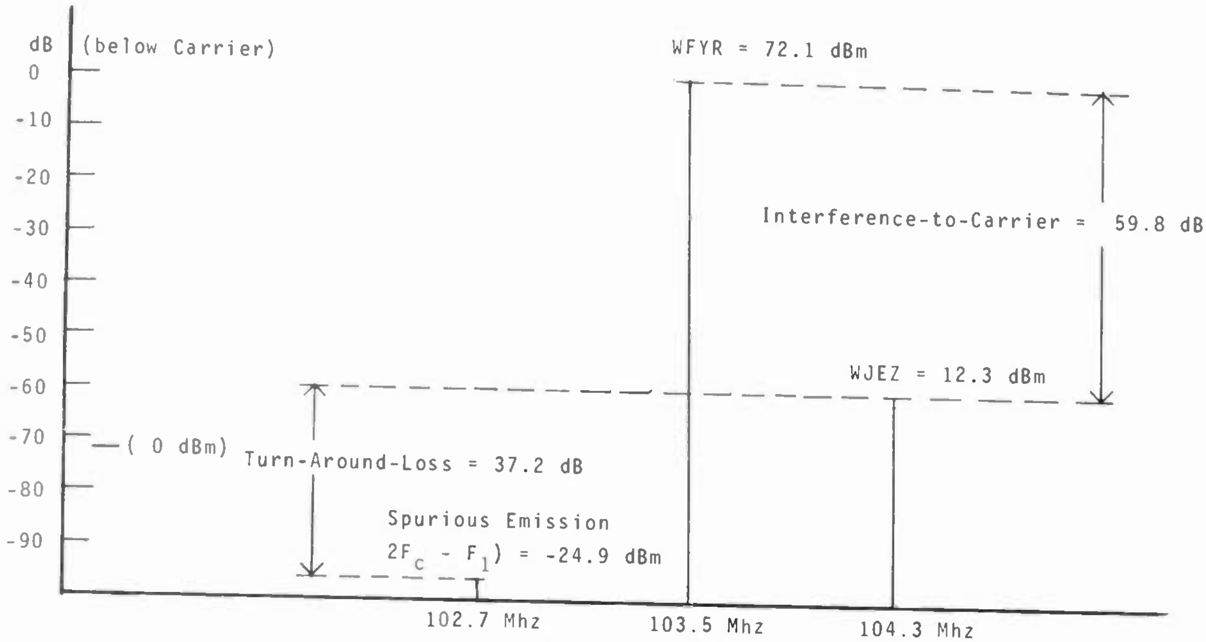


FIGURE - 5 Spectrograph, WFYR auxiliary transmitter - Prudential Site operating with 20 kw TPO, with filter, and in-compliance

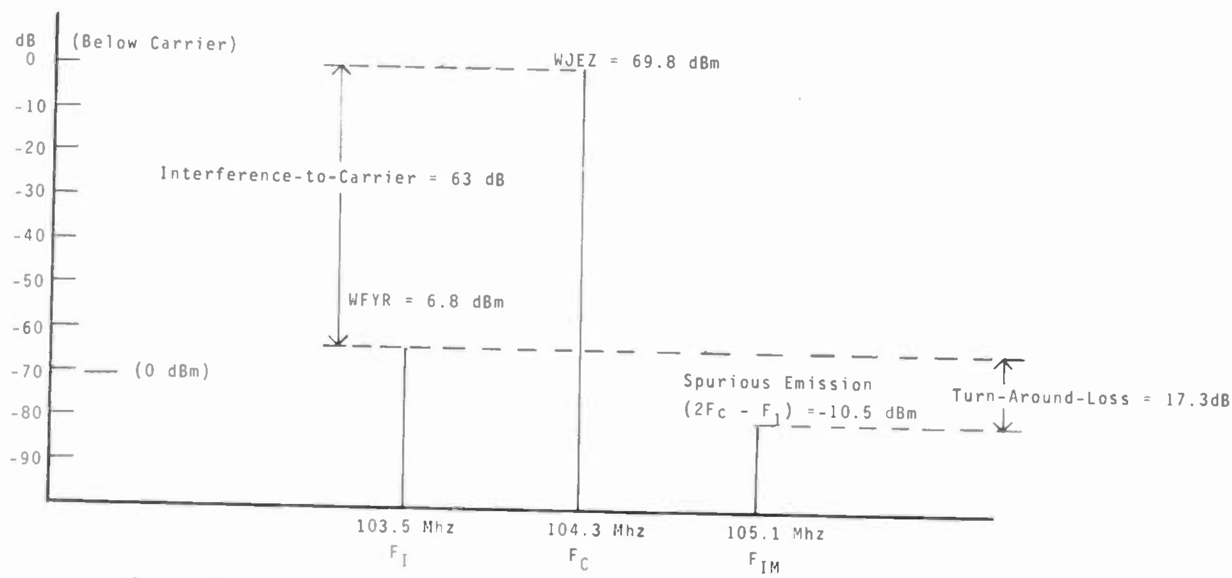


FIGURE - 6 Spectrograph, WJEZ main transmitter, Prudential Site operating with 9.5 kw TPO, and in-compliance

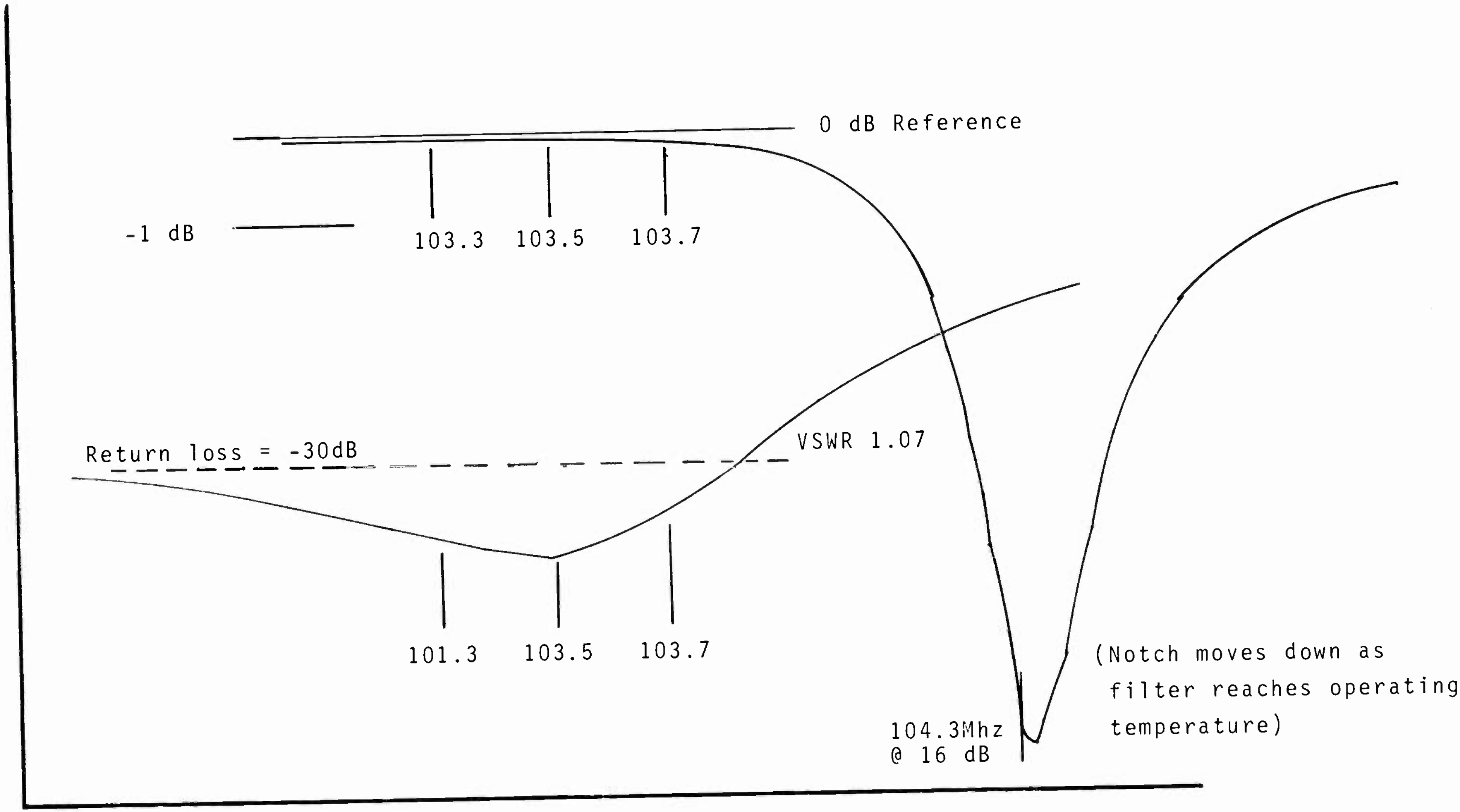


FIGURE - 7 Graph shows amplitude response and VSWR return loss for a ERI 902 notch filter tuned to pass 103.5MHz and reject 104.3 Mhz.

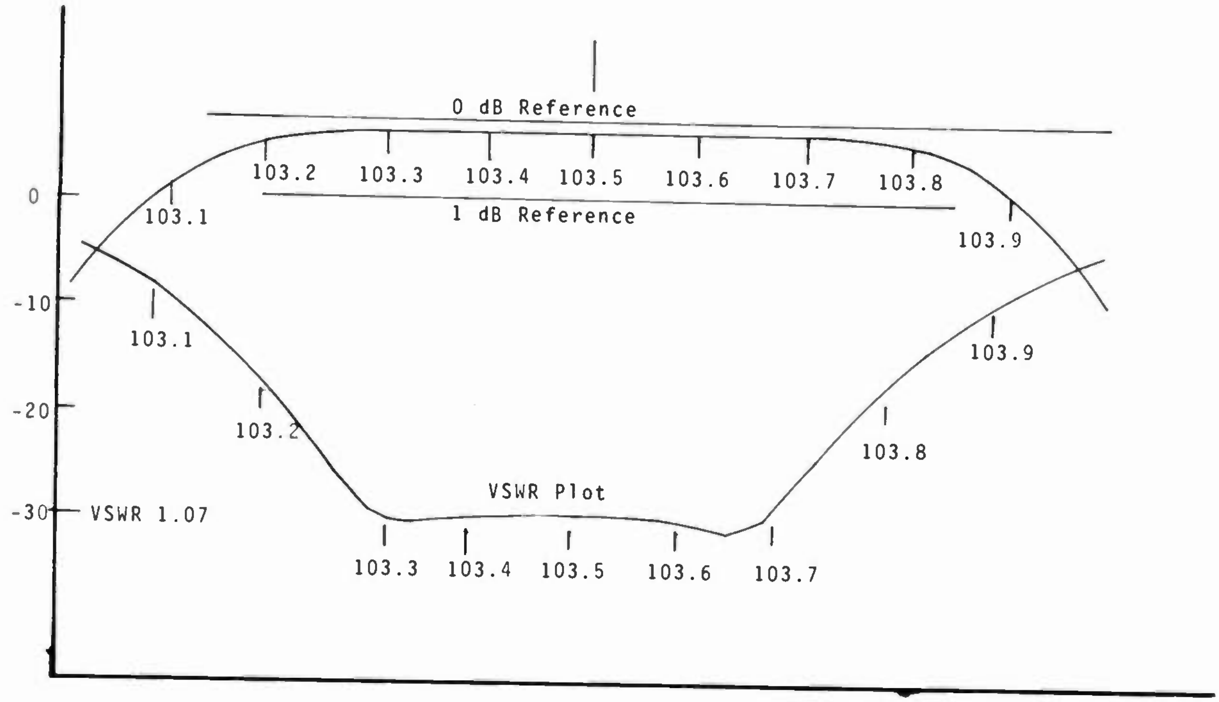


FIGURE - 8 Amplitude and VSWR performance of 3-section bandpass Filter Used by WFYR at the Sears Tower Transmitter Site.

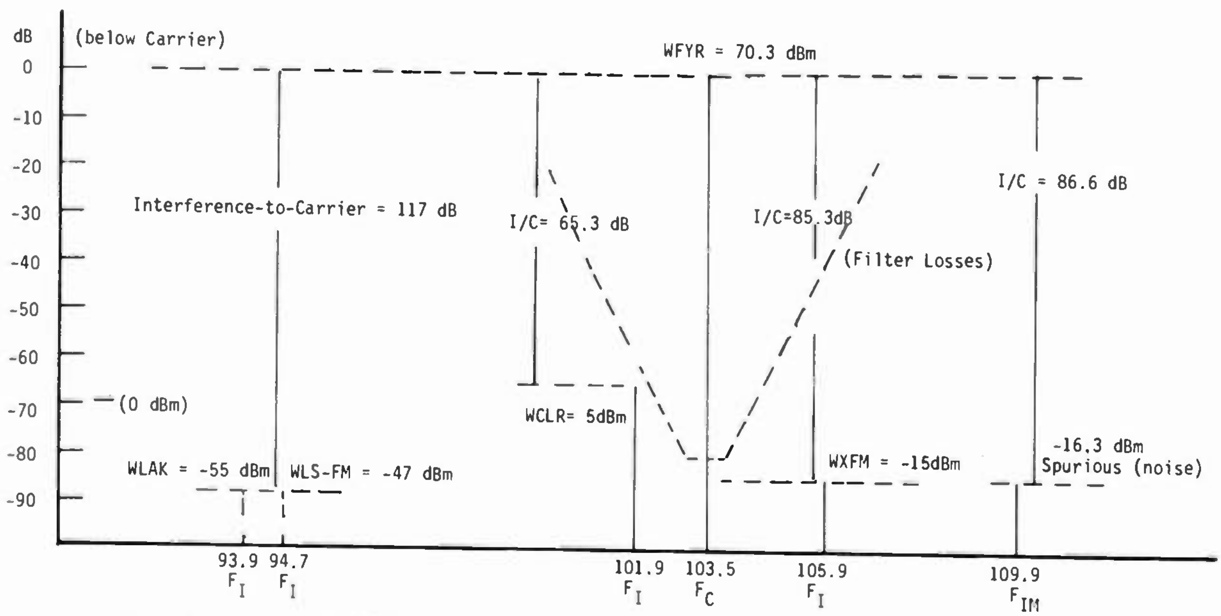


FIGURE - 9 Spectrograph, WFYR Transmitter - Sears Tower Transmitter site.

DESIGN AND APPLICATION OF A MULTIPLEXED

NINE STATION FM ANTENNA FOR THE SENIOR ROAD TOWER GROUP

R. E. FISK

HARRIS CORPORATION, BROADCAST TRANSMISSION DIVISION

QUINCY, ILLINOIS

ABSTRACT

This paper discusses the design of a broadband circularly polarized cavity backed radiator antenna for use by nine Class C FM broadcast stations.

By joining together and utilizing the same 2000 Ft. tower, broadband antenna system and transmitter building, construction costs for each station are reduced 50% to 75%. Combined average power antenna input at the antenna system input is in excess of 200 kW for these stations. The associated high peak power levels resulting from combining the nine FM channels requires careful consideration of feed system components and radiating element design. The design of these components is discussed.

The final design attains a 20% bandwidth with an input VSWR less than 1.1.

Test results including measured patterns, impedance data and power tests are discussed.

INTRODUCTION

On October 15, 1983, nine Class C FM Broadcasters in Houston, Texas signed on the air from a new 2000 ft. tower and multiplexed antenna facility. Known as the Senior Road Tower Group, each of these broadcasters while serving a major metropolitan area wanted to maximize their coverage and assure their future operation as Class C stations with maximum service facilities. By joining together and utilizing the same tower, antenna and transmitter building, construction and maintenance cost for each station was reduced 50% to 75%.

The Senior Road Tower Group was formed by a group of FM broadcasters located on the Shell Plaza Building in downtown Houston. Over a period of years many new tall buildings had been erected to heights that caused severe shadows in their coverage pattern. Since the Shell Plaza location and antenna facility was shared by eight stations, a logical plan was the construction of a new maximum service facility outside the congested growing metropolitan area.

A 2000 ft. tower supporting a common antenna for up to 10 stations, each having 100 kW effective radiated power was proposed in the initial planning.

Because of the magnitude of the project and the need for a high degree of cooperation and compromise among the stations, a committee was formed under the leadership of Mr. Bill Cordell, Chief Engineer of KIKK-FM as group administrator. Mr. Tony Uyttendaele, Director Allocations & RF Systems for American Broadcasting Companies, Inc., New York City, owner of KSSR-FM provided technical consultation to formulate the specifications for the antenna and multiplexer requirements for the project.

Harris Corporation was the successful bidder for the antenna, multiplexer and RF monitoring system portion of this project.

FCC Docket 80-90 was not in existence when the Senior Road Tower Group facility was planned. The farsightedness of this group was prophetic of the new FCC requirements for a maximum service Class C facility.

S.R.T.G. SPECIFICATIONS

The principal requirements for the S.R.T.G. antenna were:

FREQUENCY RANGE

Designed to operate with the combined output of nine FM transmitters on the following individual frequencies.

	STATION	FREQUENCY
1.	KYND - FM	(92.9 MHz)
2.	KLEF - FM	(94.5 MHz)
3.	KIKK - FM	(95.7 MHz)
4.	KSRR - FM	(96.5 MHz)
5.	KFMK - FM	(97.9 MHz)
6.	KODA - FM	(99.1 MHz)
7.	KILT - FM	(100.3 MHz)
8.	KLOL - FM	(101.1 MHz)
9.	KRBE - FM	(104.1 MHz)

POWER RATING:

Average Power: 290 kW

Peak Power: 5000 kW (Instantaneous) with 44% safety margin over published rating.

NUMBER OF BAYS

12 Bays with a 6-bay upper and a 6-bay lower section

POLARIZATION

Right-hand circular polarization. Axial ratio 2 dB or less.

POWER GAIN

Nominal power gain of 6 per polarization based on the RMS value of the azimuthal pattern.

PATTERN MEASUREMENTS

A. Azimuthal patterns H_{POL} and V_{POL} of each half of the antenna at all frequencies on the actual antenna. Continuously measured pattern to be within $\pm$ 2 dB of the calculated.

B. Elevation patterns. Each antenna half in 12 planes at each of the nine frequencies for H_{POL} and V_{POL} for an elevation range of +30° to -16°. Four azimuths to have measurements over a -90° to +90° range for gain calculations. (Total number of pattern measurements 468).

VSWR

Less than 1.1 at each operating carrier frequency $\pm$ 200 kHz.

TAC-12M CBR ANTENNA

The antenna proposed for the Senior Road Tower Group requirements by Harris was a type TAC-12M 12 bay cavity backed radiator design. The CBR design antenna had been used by Harris for a number of High Band TV stations over the past five years for circularly polarized TV applications. (1)

SCALE MODEL STUDY

The initial portion of the design program consisted of a single layer scale model study to determine pattern characteristics over the FM band frequency range and the required optimum cavity dimensions. Interface between the triangular mast section used to support the array and the cavities was developed through pattern work with the model.

From the study it was determined that a 72 inch diameter cavity mounted on the three faces of a 6' 6" triangular tower section would result in near optimum pattern characteristics.

ANTENNA DESCRIPTION

The TAC-12M antenna is an array of 36 cavity backed radiators (CBR) elements arranged 12-12-12 on the three faces of a triangular tower section to provide an omnidirectional radiation pattern. The elevation pattern of the antenna is tilted 0.75 degrees below a plane perpendicular to the axis of the antenna. The antenna is designed to radiate a circularly polarized signal over the frequency range of 88 to 108 MHz.

The nine FM stations required to be multiplexed are fed to the antenna through two 8-3/16 inch 75 ohm coaxial transmission lines. One line feeds the upper six bays and the other line the lower six bays of the antenna. See Figure 1, overall picture of antenna on the next page.

CAVITIES

For the FM frequency range, a circular cavity is used having a diameter of 0.6 wavelengths and a depth of about 0.2 wavelengths at mid-band. These

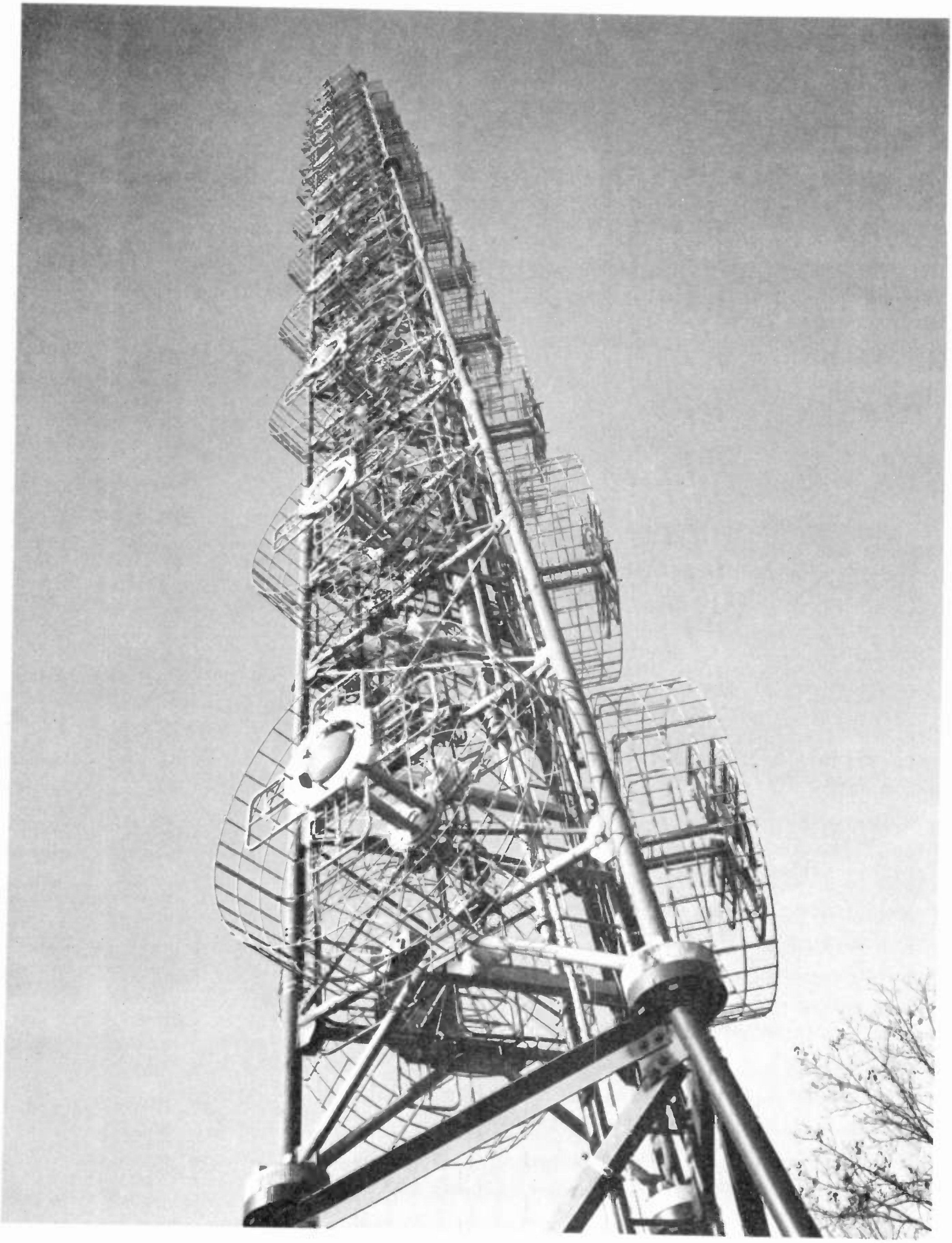


FIGURE 1. OVERALL VIEW OF TAC-12M ANTENNA FOR S.R.T.G.

dimensions result in a 120° half voltage beamwidth required for a three around array on a triangular supporting tower.

The cavities are of a galvanized welded steel grid construction to minimize wind loading. They are sturdy enough for a technician to recline inside the cavity while adjusting the dipole assembly.

THEORY OF OPERATION

The basic concept of the cavity backed radiator antenna for circular polarization is the use of crossed dipoles fed in quadrature phase located in the aperture of a backing cavity. The crossed dipoles excite the entire cavity with a rotating RF field in a plane parallel to the dipoles. The resultant RF field is thus circularly polarized and may be ideally represented by a rotating vector of constant magnitude revolving one revolution per wavelength of propagation distance. The field is considered to be right-hand polarized if the field rotation is clockwise as viewed from behind the radiators.

Radiation patterns, associated beam width and directivity are determined to a large extent by the size of the cavity aperture. The size and geometry of the dipole has less effect than the cavity to assure good pattern symmetry. The size and geometry of the dipole controls the antenna impedance and VSWR.

The backing cavity performs three important electrical functions:

First, it isolates the radiating elements from the tower or mounting structure, and, to a large degree, from adjacent elements.

Second, the backing cavity provides sharper beamwidth and more gain than achievable with the dipoles alone. The entire aperture of the cavity is excited to provide a larger effective area for each dipole.

Third, the backing cavity provides pattern control so that the antenna pattern beamwidth is nearly equal for both horizontal and vertical polarization.

PATTERN MEASUREMENT

Pattern measurements on the S.R.T.G. antenna were performed at the Harris Corporation Antenna Test Site located near Palmyra, Missouri. The antenna pattern measurement range at this facility is a ground-reflection range which utilizes the wide flat flood plain area along the Mississippi River. The transmitting antenna illumination source is located on a tower near the river and is directed over the flood plain toward the receiving site located on a steep bluff about 220 ft. above the flood plain. The path length is 15,830 ft. A remote controlled oscillator and two separate FM Yagi antennas (one for V POL and one for H POL transmission) were used at the transmitting antenna tower site.

WIDEBAND DIPOLES

Special Wideband flat dipoles are used for the excitation of the cavities. The dipole is designed to handle high peak power levels and for low Q to cover the desired bandwidth. The adjustable parasitic elements or tabs attached to the angular tuning ring shown in the photograph serve the function of an open sleeve parallel to the dipoles and extend the bandwidth. Similar designs are used for satellite antenna systems operating over a 1.8 to 1 frequency band. (See ref. 2)

A closeup of the radiating element is shown below.

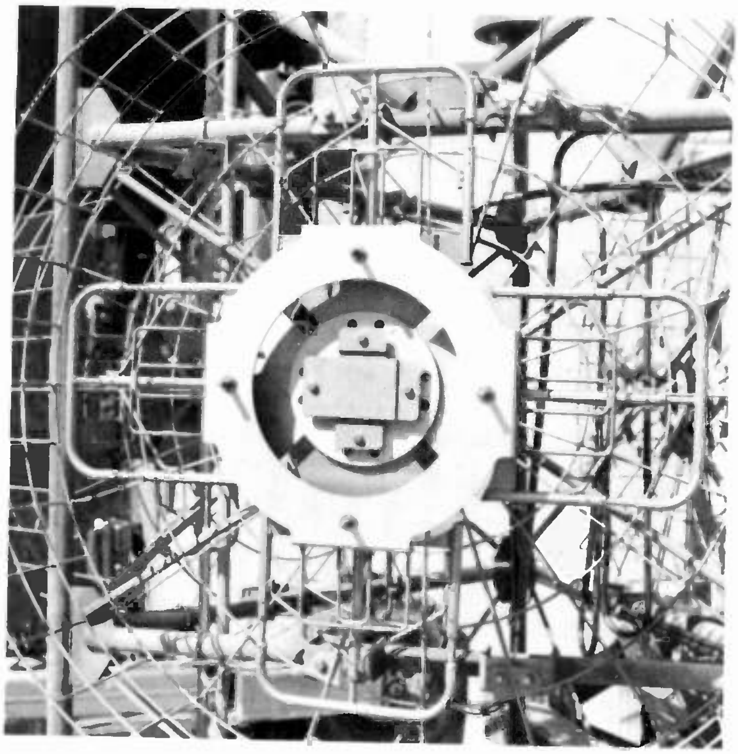


FIGURE 2
CROSS-DIPOLE
RADIATING ELEMENTS

FEED SYSTEM

The antenna is divided into two six-bay sections -- each fed with a 8-3/16 inch, 75 ohm coaxial line. The 8-3/16 inch lines each feed 3 dB coupler units to provide quadrature outputs for the horizontal and vertical dipole elements of the array. A schematic of the feed system is shown in Figure 3 on the following page.

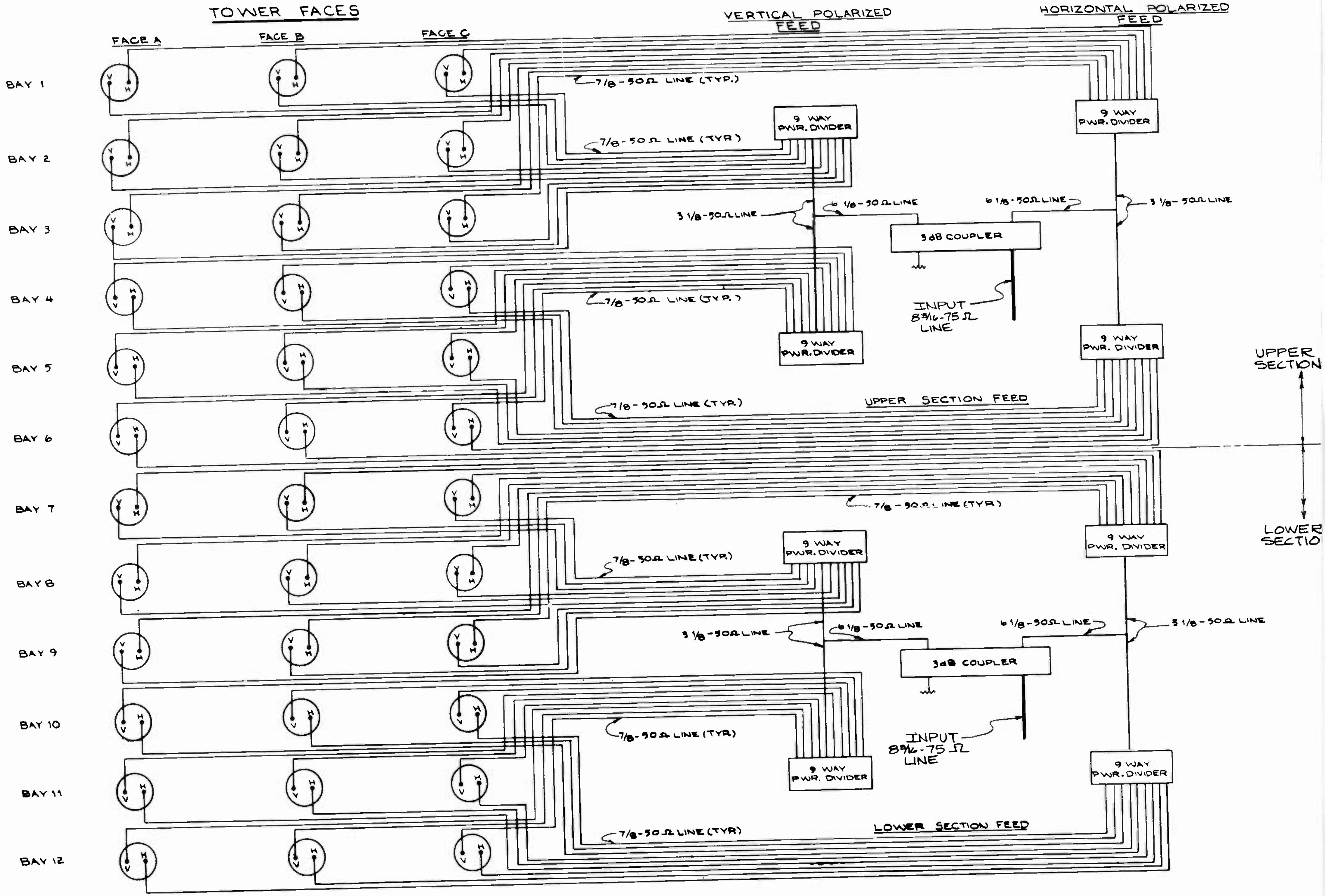
The power ratings of the various transmission lines used in the feed system are listed in the following table:

AVERAGE POWER RATINGS

50°C AMBIENT, 0 PSIG, 1.0 VSWR, 100 MHz

LINE SIZE	AVERAGE POWER RATING	NUMBER OF LINES	TOTAL RATING
7/8" 50 ohm	4.5 kW	72	324 kW
3-1/8" 50 ohm	45.5 kW	8	364 kW
6-1/8" 50 ohm	135.0 kW	4	540 kW
8-3/16" 75 ohm	213.0 kW	2	426 kW

Average antenna input power at tower top for 9 stations at 100 kW ERP is approximately 170 kW.



FEED SYSTEM S.R.T.G.
TAC-12M MULTIPLEXED FM ANTENNA

FIGURE 3

PEAK POWER RATINGS

0 PSIG

LINE SIZE	PEAK POWER RATING	NUMBER# OF LINES	TOTAL PEAK RATING
7-8" 50 ohm	58.8 kW	72	4.2×10^6
3-1/8" 50 ohm	920.0 kW	8	7.4×10^6
6-1/8" 50 ohm	3000.0 kW	2	12.0×10^6
8-3/16" 75 ohm	2500.0 kW	2	5×10^6

Peak power ratings will increase by a factor of 1.71 to a value of 7.2×10^6 ($5 \times 10^6 \times 144\%$) if lines are pressurized to 8.2 PSIG. Peak power operating value at antenna input for 9 stations is approximately 2.45×10^6 watts.

METHOD OF MEASUREMENT

Each half of the S.R.T.G. antenna was mounted on the test turntable with its axis horizontal for pattern tests as shown in Figure 4. The test stand is designed so that the antenna can be rotated to any desired angle in azimuth or elevation relative to the illumination source. Typical pattern plots as recorded are shown below.

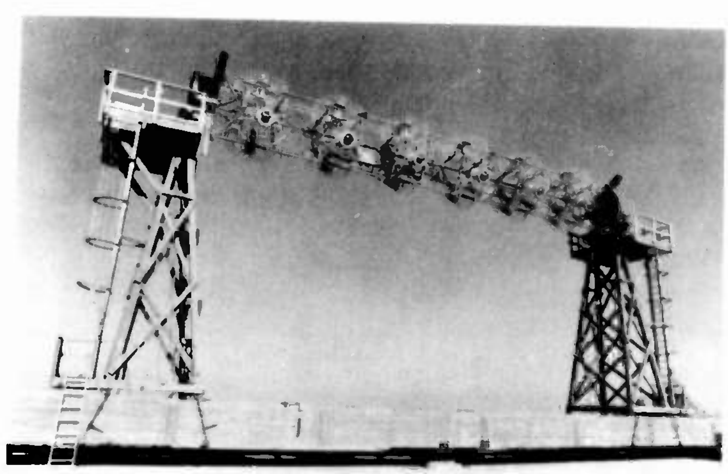


FIGURE 4. ANTENNA MOUNTED ON MAIN TURNTABLE 35 FT. ABOVE GROUND

MEASURED AZIMUTHAL PATTERN

TAC-12M MULTIPLEXED FM ANTENNA

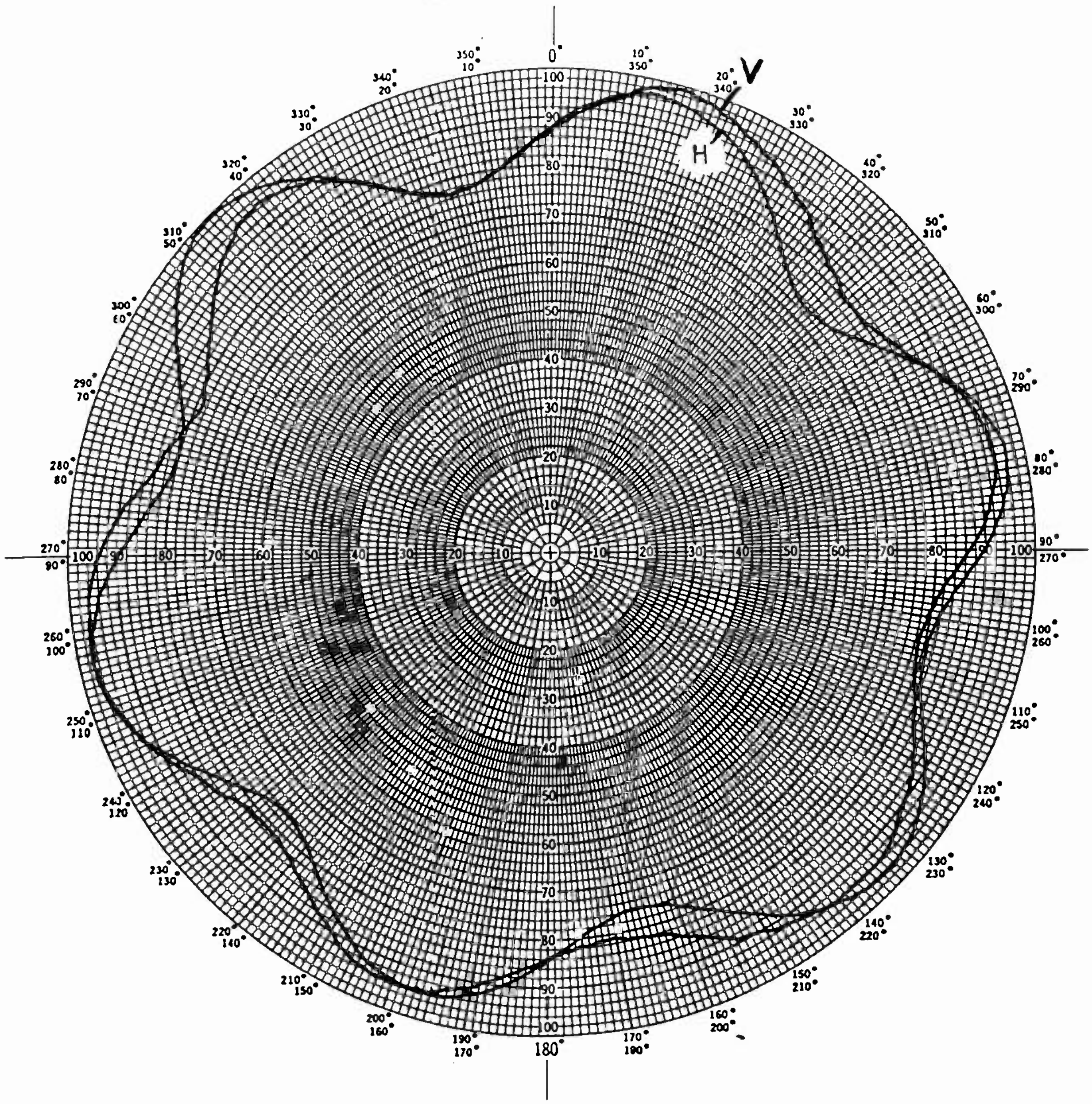


FIGURE 5. MEASURED HPOL AND V POL
RELATIVE FIELD PATTERNS

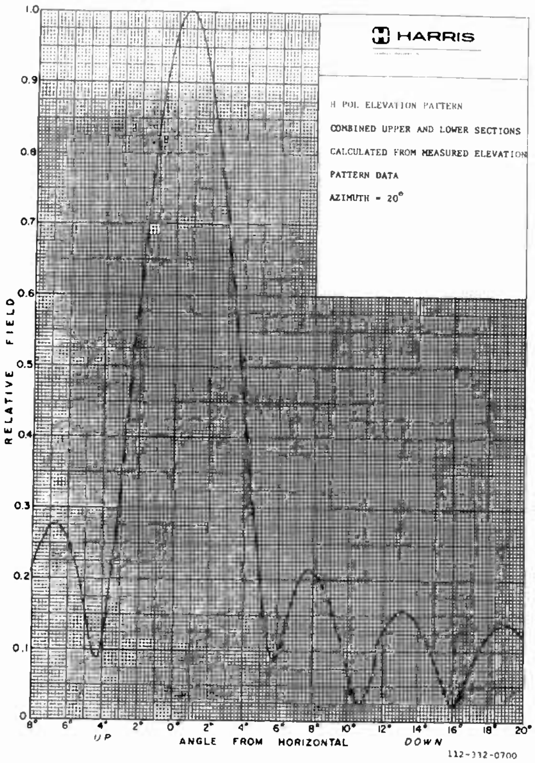


FIGURE 6

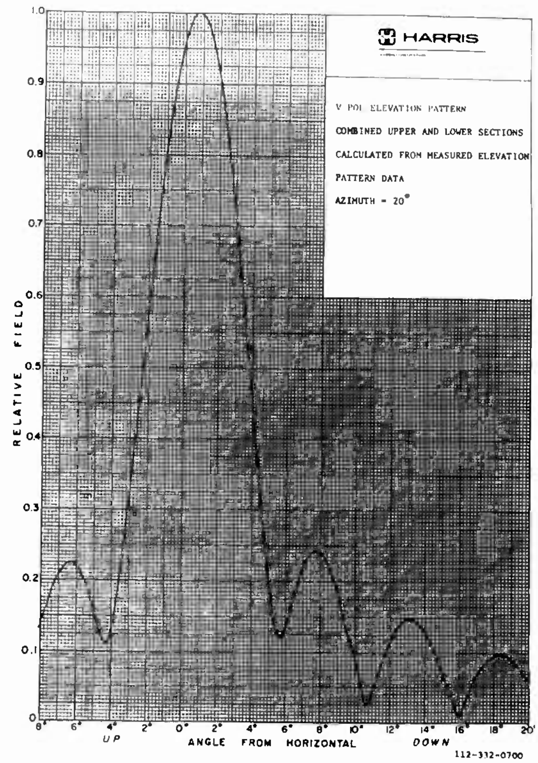


FIGURE 7

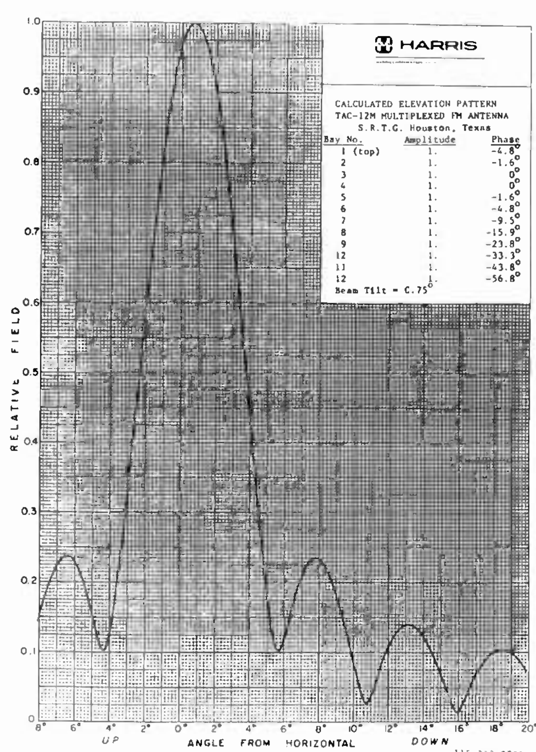


FIGURE 8

ELEVATION PATTERN MEASUREMENTS AND GAIN CALCULATIONS

Elevation pattern measurements were recorded by taking $\pm$ 90° cuts of the antenna pattern (relative to the horizontal pattern 90°) at azimuths of 20°, 110°, 200°, and 290°.

The relationship --

$$D = \int_{-90^\circ}^{90^\circ} \frac{69.9}{F^2(\theta)\cos\theta \, d\theta} \qquad (3) \text{ (Horizontal Plane = 0°)}$$

is used in a computer program to compute the antenna directivity or gain before derating.

The overall measured gain was determined to be 6.37 or 8.04 dB at 97.9 MHz from the measured pattern data compared to the calculated value of 6.0 or 7.78 dB for this frequency.

7/8 inch Feed Cables

The length of the 7/8 inch 50 ohm feed cables is varied for each bay to control the beam tilt and provide null-fill. The following table is a list of the length used for each bay and the resulting phase angle at the center frequency.

7/8 INCH FEED CABLE LENGTHS

BAY	PHYSICAL LENGTH	RELATIVE PHASE
1	19' - 1-1/2"	-4.8°
2	19' - 9-1/2"	-1.6°
3	19' - 0"	0.0
4	19' - 0"	0.0
5	19' - 0-1/2"	-1.6°
6	19' - 1-1/2"	-4.8°
7	19' - 3"	9.5°
8	19' - 5"	-15.9°
9	19' - 7.5"	-23.9°
10	19' - 10-7/16"	-33.3°
11	20' - 1-7/8"	-44.1°
12	20' - 5-7/8"	-57.0°

IMPEDANCE ADJUSTMENTS

Overall impedance measurements were made with the 120 ft. S.R.T.G. antenna mounted vertically on a 10 Ft. support tower near the edge of the bluff at the Harris Antenna Test Site as shown in Figure 9.

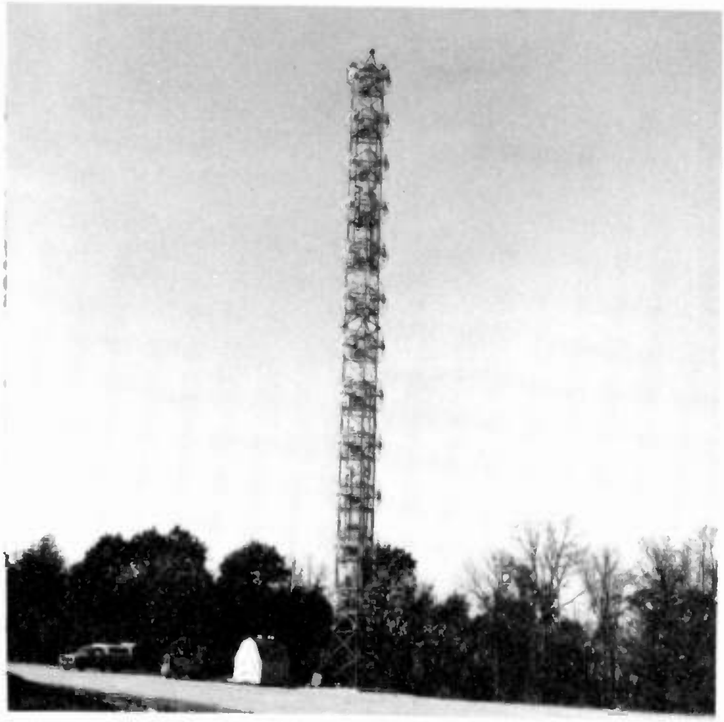


FIGURE 9.
ANTENNA MOUNTED
VERTICALLY FOR
IMPEDANCE MEASUREMENTS

Impedance tests require that each of the 72 dipoles be adjusted for an impedance match over the FM band with other cavities in the array excited to include mutual impedance effects on the match.

Adjustments include dipole length, parasite length and spacing from the dipole, and the strap width connecting the balun to the dipole arms.

By using a network analyzer with a X - Y plotter, the impedance function of each dipole was quickly plotted to facilitate adjustments.

The overall match of the array is shown in the expanded Smith Chart of the overall impedance. The frequency range of the nine stations' individual frequencies are identified by dots on the plot.

A VSWR tabulation from this plot follows:

SENIOR ROAD PROJECT

OVERALL VSWR MEASUREMENTS

	FREQUENCY	VSWR
KYND-FM	92.9	1.023
KLEF-FM	94.5	1.045
KIKK-FM	95.7	1.061
KSSR-FM	96.5	1.047
KFMK-FM	97.9	1.043
KODA-FM	99.1	1.061
KILT-FM	100.3	1.039
KLOL-FM	101.1	1.037
KRBE-FM	104.1	1.047

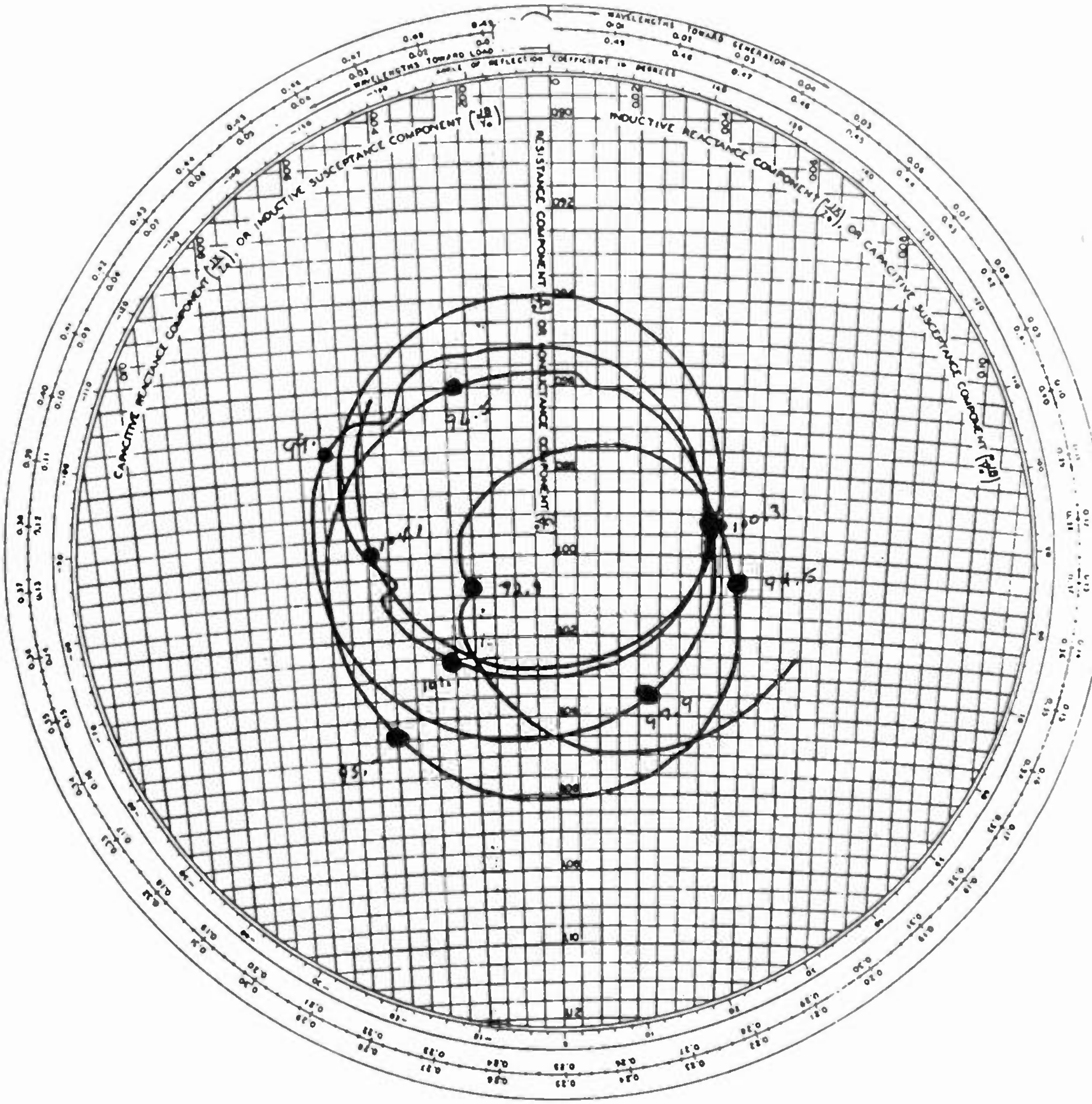


FIGURE 10. MEASURED OVERALL IMPEDANCE PLOT

The common point for the complete antenna system is located in the transmitter building. The output of the multiplexer is a 14 inch 50 ohm coaxial line to a 3 dB coupler which serves as a power divider to feed the two 8-3/16 inch 75 ohm lines to the antenna inputs.

An adjustable trombone section is used to equalize line lengths to the upper and lower antenna sections. The developed length of each line is approximately 2000 feet.

POWER TESTS

A single cavity was mounted near the building where Harris transmitters are manufactured and tested. A 50 kW FMK transmitter was pulsed at a 20% duty cycle and fed to one of the single dipoles of the cavity pointed vertically upward.

RF power at the 50 kW level was applied to the dipole while the element was kept wet by a fogger water spray. No evidence of corona or change in VSWR was noted during the test. Normal peak power operating level is about 2.5×10^6 watts or 35 kW per dipole.

At 5×10^6 peak instantaneous power the voltage level at the ends of the dipoles is about 5.3 kV or well below critical voltage levels for corona conditions. (See Reference 4)

CONCLUSIONS

FCC Docket 80-90 enacted on May 26, 1983 requires that Class B and Class C stations with less than maximum facilities must upgrade to a minimum facility or face being downgraded. This Docket has a widespread impact on existing FM Broadcast Stations. If a Class C station fails to upgrade to the minimum required antenna height of 300 meters or 984 ft. HAAT within three years of the effective date of the new rules, they will be downgraded to a lower classification.

The Cavity Backed Radiator antenna design has the high power broadband characteristics needed to combine a number of FM stations into a common antenna with exceptionally excellent pattern and impedance characteristics.

Its application to both multiplexed and single station facilities provide a versatile antenna system for many stringent FM Broadcast requirements.

ACKNOWLEDGEMENT

A project of this magnitude at Harris Corporation is controlled by a Project Management Group under the direction of Mr. Howard Young who worked closely with the Senior Road Tower Group to coordinate the antenna, multiplexer and RF monitoring facilities for the nine stations. His contributions were of great value to the project.

The author wishes to acknowledge the valuable suggestions and counselling of G. W. Collins and J. A. Donovan of the Harris Antenna Engineering Design Group.

The dedicated work of the antenna manufacturing and test team under Don Voros, Lee Smith, John Starr and Mark Riley contributed to the resultant high quality workmanship of the final design.

REFERENCES:

1. R. E. Fisk and J. A. Donovan, "A New CP Antenna for Television broadcast Service", IEEE Transactions on Broadcasting, VOL. BC-22, No 3. September, 1976. pp. 91 - 96

2. H. E. King and J. L. Wong, "An Experimental Study of a Balun-Fed Open-Sleeve Dipole in front of a Metallic Reflector", IEEE Transactions on Antennas and Propagation, March 1972 pp. 201-204.

3. Henry Jasik, "Antenna Engineering Handbook", McGraw Hill Book Company, Inc., 1961, pp 34-24 and 34-25.

4. Edmund a Laport, "Radio Antenna Engineering", McGraw-Hill Book Company, Inc., 1952, pp 239-242 and pp 444-448.

RECENT DEVELOPMENTS IN RADIAL GROUND SYSTEMS FOR AM BROADCASTING

ARCHIBALD C. DOTY, JR.

FLETCHER, NORTH CAROLINA

The vertical antenna used so extensively by AM Broadcast stations is extremely versatile. It can be constructed in a wide range of heights; it can consist of a thin wire or a thick tower; it can be top loaded, bottom loaded, capacitively loaded or inductively loaded; it's feed impedance can be varied to a convenient value by a number of methods.

The main problem with the vertical antenna is that to work efficiently it must be used with a good ground system.

This system must meet three basic requirements, briefly described as:

1) THE GROUND SYSTEM MUST EFFICIENTLY COLLECT "RETURN CURRENTS" FROM THE ANTENNA

When RF energy is fed to a vertical antenna, displacement currents flow from the antenna, through the surrounding space, and toward the earth below. To complete the circuit, and to permit these currents to flow, they must be collected from or near the earth, and returned to the RF feed point at the antenna base.

Considerable losses will be incurred if these currents must pass through the earth, as the earth presents a high resistance to RF current flow. Even sea water,which has quite high conductivity (low resistance) compared to earth does not compare with a so-called "Perfect" ground. In this connotation the ground might be thought of as having the electrical characteristics of a solid metal plate.

One special type of antenna, the so-called "Ground Plane"- often found in UHF service- operates independently from the earth's surface. This antenna incorporates two or more 1/4 wavelength radial wires or rods located at the base of a vertical radiator. However, in this configuration the radials are, in effect, a portion of the radiating element, rather than an artificial ground system.

Other antenna systems utilizing vertical radiators must include some method for collecting and returning the currents from the antenna. The method which has been most commonly used is to allow the return currents to enter the earth, and then collect them with an array of buried "radial" wires.

In past years there has also been occasional use of elevated arrays of radials under vertical antennas (counterpoises and ground screens) designed to intercept return currents before they enter the earth, or to collect them after they have entered the earth through the capacitive relationship between the radials and the ground (rather than by conduction through the earth).

2) THE ANTENNA'S RETURN CURRENTS MUST BE CONDUCTED THROUGH THE GROUND SYSTEM TO THE BASE OF THE VERTICAL RADIATOR WITH MINIMUM LOSSES

Once the return currents have been collected they must be returned to the base of the antenna. The complete circuit through which these currents flow can be considered as:

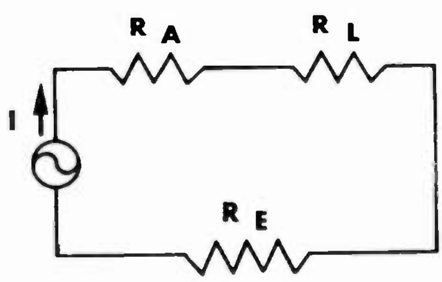


FIG.1

R_A= Radiation Resistance

R_L= Conductor Resistance

R_E= Ground Resistance

(R_E relates only to those return currents that enter the earth and are then collected by buried wires as conduction currents)

3) THE GROUND MUST PROVIDE A DISTINCT (ELECTRICALLY) REFLECTING SURFACE UNDER THE VERTICAL ANTENNA.

Some of the radiation from a vertical antenna is directed below the horizon, and will impinge on the surface of the earth. A portion of this radiation will be reflected from the earth's surface, and may combine with the direct radiation from the vertical to assist in establishing the radiation pattern of the antenna array, as shown in Figure 2.

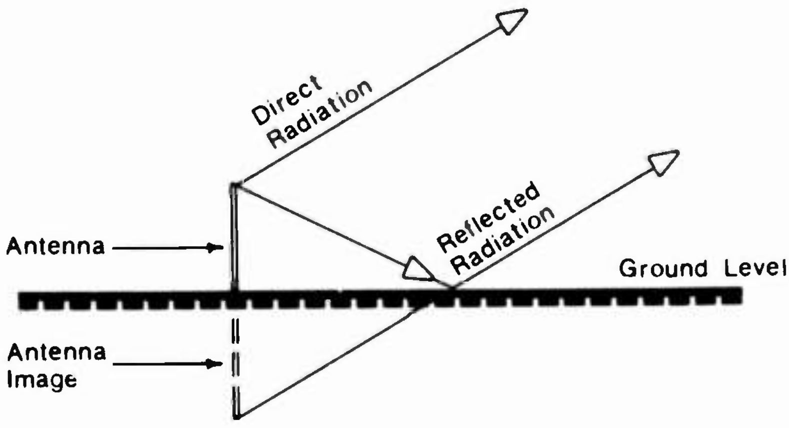


FIG. 2

Figure 3 shows that the image antenna is an exact reproduction of the vertical antenna as it would look if the earth's surface were a perfect "electrical mirror". Unfortunately, ground is far from being a good electrical conductor, as noted earlier.

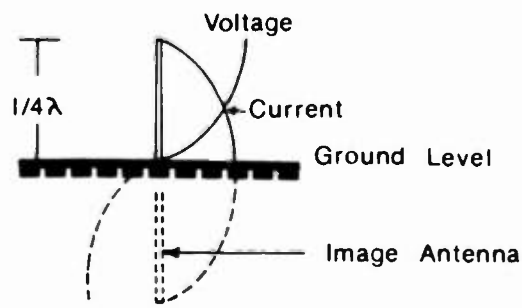


FIG. 3

A good artificial ground system should assist in correcting this situation by presenting a good reflecting surface.

PAST AND CURRENT PRACTICE

At the present time there are slightly under 5,000 AM Broadcasting stations in the United States. If each of these stations has followed FCC Rules[1], and has installed a minimum of 120 buried ground radial wires -each at least 1/4 wavelength long- under it's vertical transmitting antenna, there are more than 26,000 miles of these wires buried in this country.

The mole-like endeavors of the AM station owners have had good technical justification. They have been based on a landmark research effort undertaken almost fifty years ago. This research work, by Dr. George Brown and his associates[2], is acknowledged a true "classic" in it's field.

That work, however, has had two unfortunate, though probably unavoidable, results:

1) The overpowering completeness and excellence of the study had the effect of discouraging further research in the particular area that it covered.

2) The Federal Communications Commission was so impressed by the work that it became the basis of the current manditory requirements for the minimum number and length of radial wires required of each Standard Broadcast station.

The only problem with the FCC requirements is that the research on which they are based considered only one of the several possible configurations of artificial ground systems- the one using buried bare radial wires. No evidence has been found of any past research to learn the characteristics of insulated or elevated radials, for example.

RECENT RESEARCH

Dr. Brown was not alone in ignoring insulated or elevated radial "ground" systems. Although they were used to some degree in in early stations (including that used for the first transatlantic radio communications[3]), apparently no research was conducted on them after publication of the Brown paper-- until the extensive tests at Fletcher, North Carolina, in 1981 & 1982. The techniques and procedures used in the Fletcher tests have been extensively reported[4,5], and will not be repeated here. The following results of those tests, however, are pertinent to this discussion:

: As shown in Figure 4, overall ground conductivity at the test site varied by a factor of more than 2:1, with the variations in conductivity correlating well with variations in temperature and precipitation

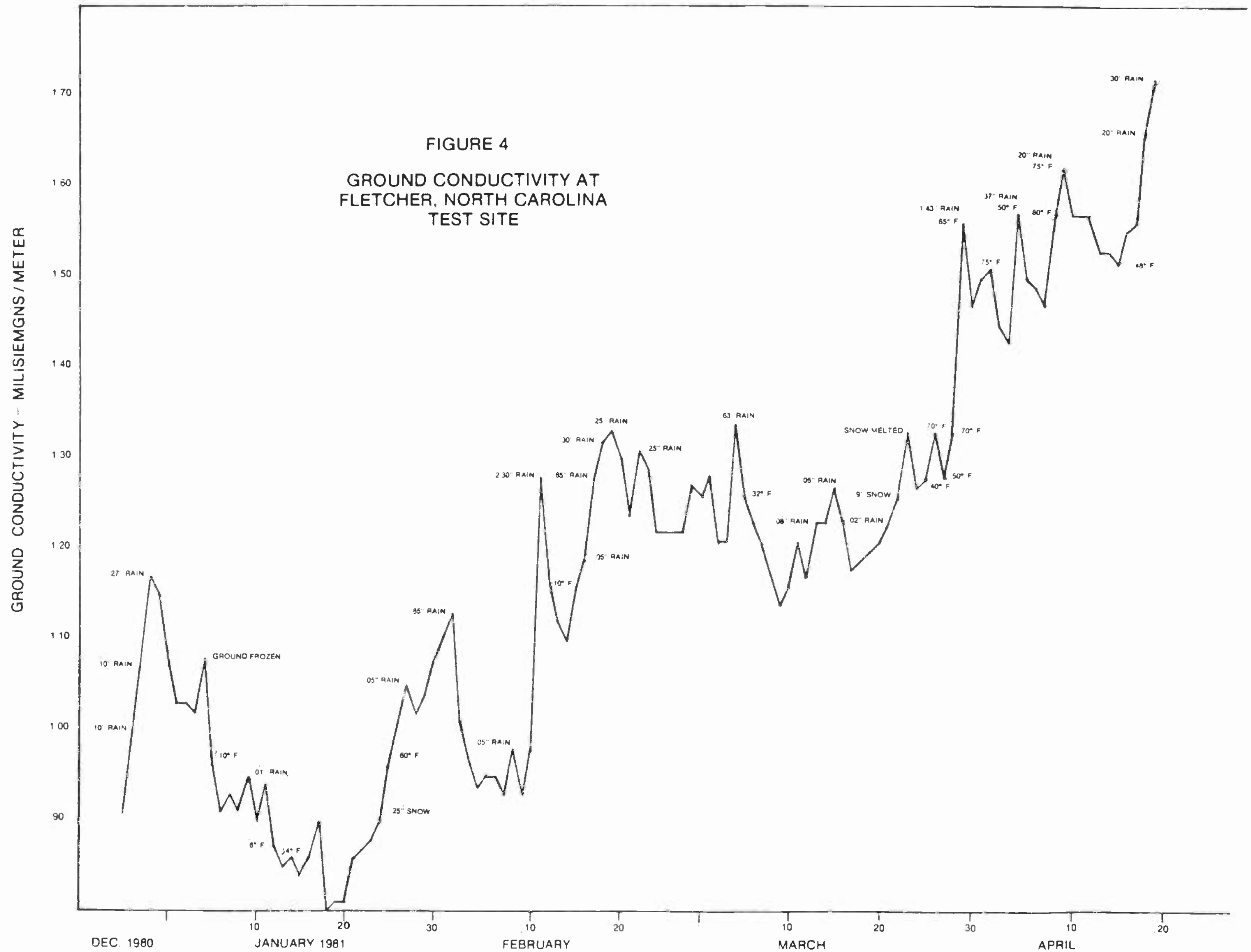


FIGURE 4
GROUND CONDUCTIVITY AT
FLETCHER, NORTH CAROLINA
TEST SITE
GROUND CONDUCTIVITY – MILISIEMGNS / METER
1 70
1 60
1 50
1 40
1 30
1 20
1 10
1 00
90
DEC. 1980
JANUARY 1981
FEBRUARY
MARCH
APRIL
10
20
30
GROUND FROZEN
SNOW MELTED
25" SNOW
.27" RAIN
.10" RAIN
.05" RAIN
.55" RAIN
2.30" RAIN
.25" RAIN
.30" RAIN
.63" RAIN
1.43" RAIN
.37" RAIN
.20" RAIN
.9" SNOW
60° F
75° F
50° F
70° F
40° F
48° F

: Ground conductivity found at the test points at the Fletcher site was **not** found to be as homoginous as expected through reference to the literature.
As seen in Figure 5, more than 100 measurements made at the test site showed variations greater than 6:1 in the ground conductivity of an area far smaller than that occupied by the ground radial system of an AM Broadcast station.
These values varied with temperature and precipitation, as did those at the basic reference point.

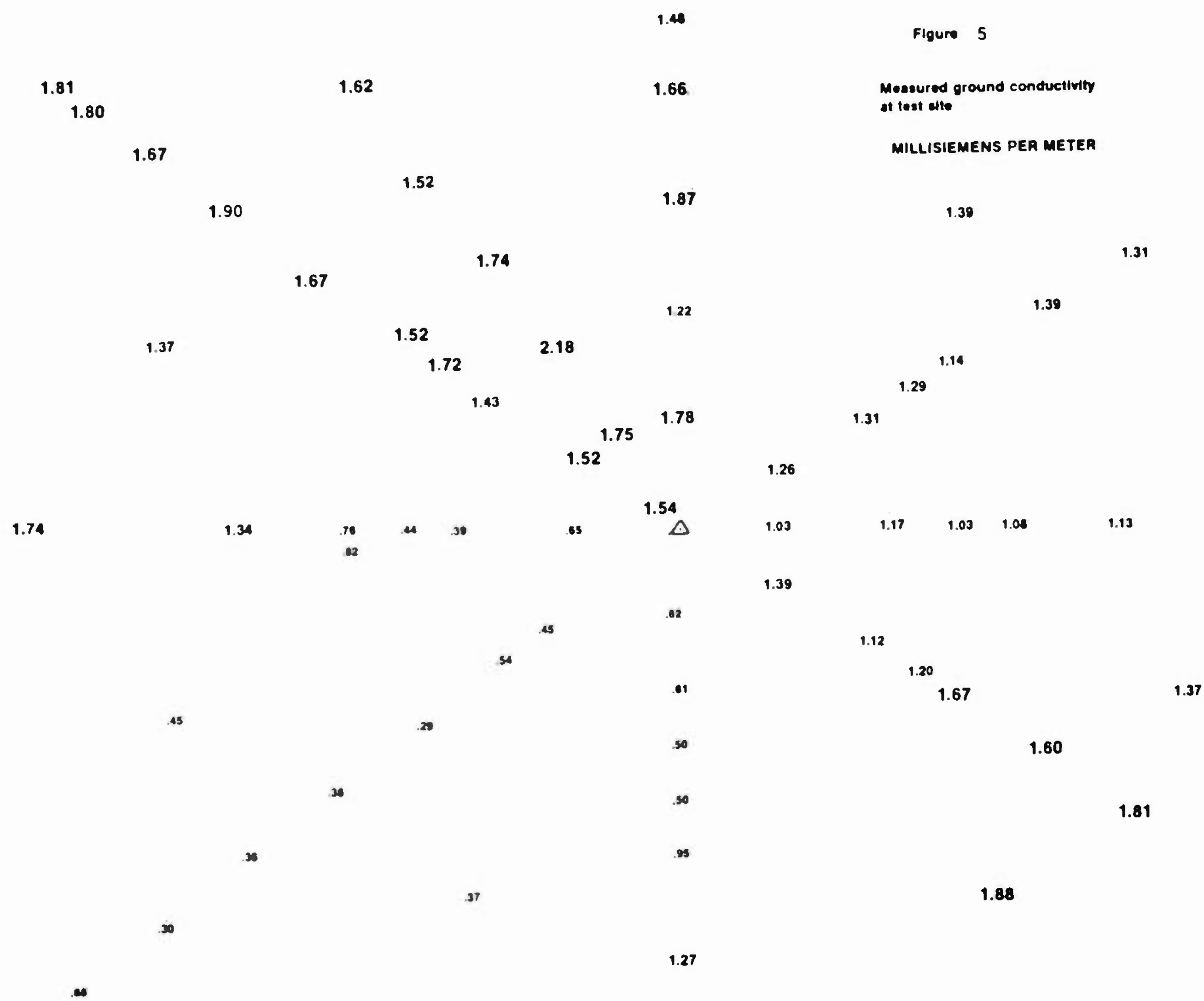


Figure 5

Measured ground conductivity at test site

MILLISIEMENS PER METER

: The return currents carried by the individual ground radials of the test counterpoise had very unusual and unexpected variations in distribution.

Also, the current distribution was not that antipicated from reference to the past 60 year's literature on the counterpoise. Figure 6 shows what was found.

: It was also found that when the radials were grounded so that the array consisted of a ground screen, the variations in current distribution persisted, as shown in Fig. 7.

: The variation in the distribution of current in the individual radial wires correlated closely with the conductivity of the ground under the radials- the higher the conductivity, the higher the current. (Compare Fig. 5 with Figs. 6 & 7)

The above findings may help to explain some of the unusual results that have been reported during Proof of Performance tests of AM Broadcast stations.

The initial research undertaken at Fletcher had the same failing as previous ground system test programs: it concentrated on one type of ground radial system, and it did not compare the relative performance of radial ground systems using buried bare wire with systems using insulated or elevated radials. An extensive review of the last 60 year's literature showed no record of such a comparison in the past.

Thus an additional test program was initiated, using the instrumentation and techniques developed in the earlier Fletcher counterpoise/ ground screen research. In this new program several thousand measurements were made of the magnitude and distribution of return currents carried by:

Elevated/ insulated radial wires.

Insulated radial wires lying on the ground.

Buried- bare wire- radials.

In these tests the radial wires terminated at the base of a 90 degree vertical antenna, and were the sole means for collecting the return currents.

So that tests could be conducted at a frequency adjacent to the AM Broadcast band- but without licensing difficulties- 1805 Khz. was used. (NB: The entire test program was conducted as a non-commercial, privately financed, project by engineers holding amateur radio licenses).

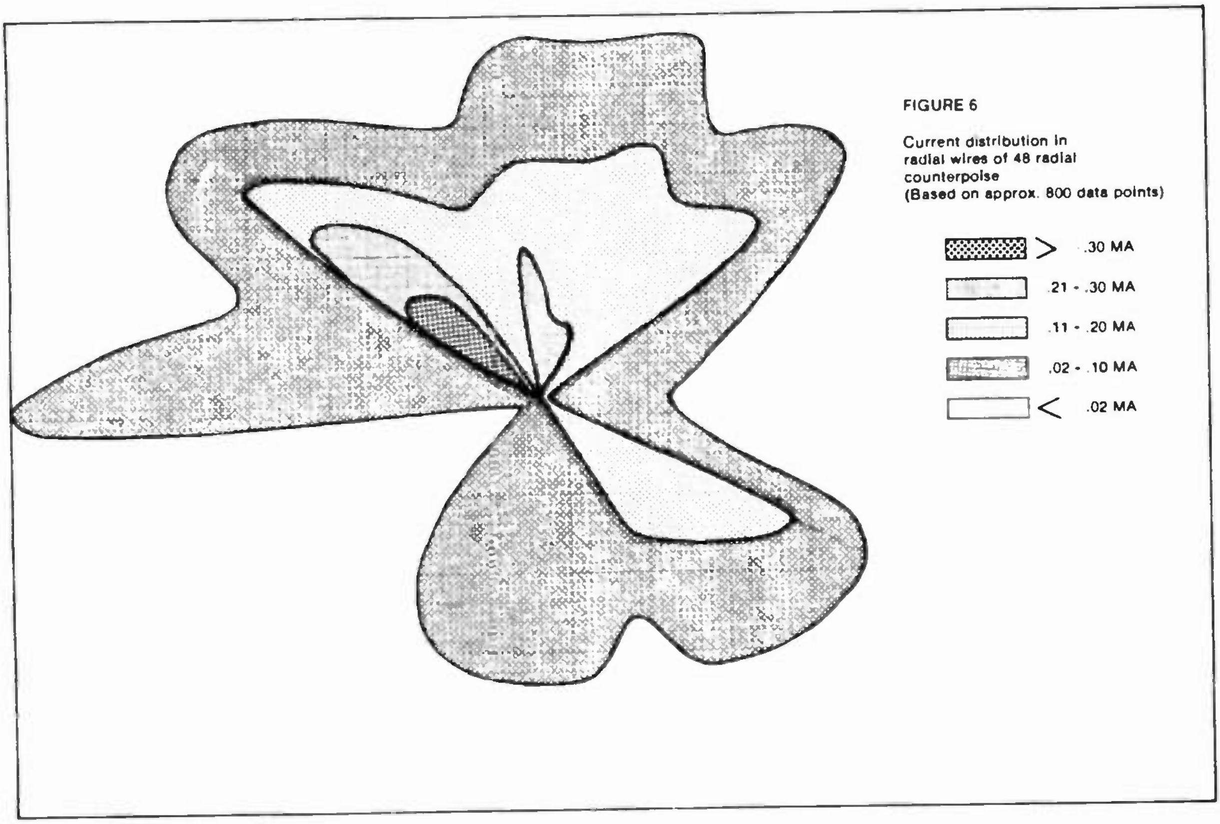


FIGURE 6

Current distribution in radial wires of 48 radial counterpoise (Based on approx. 800 data points)

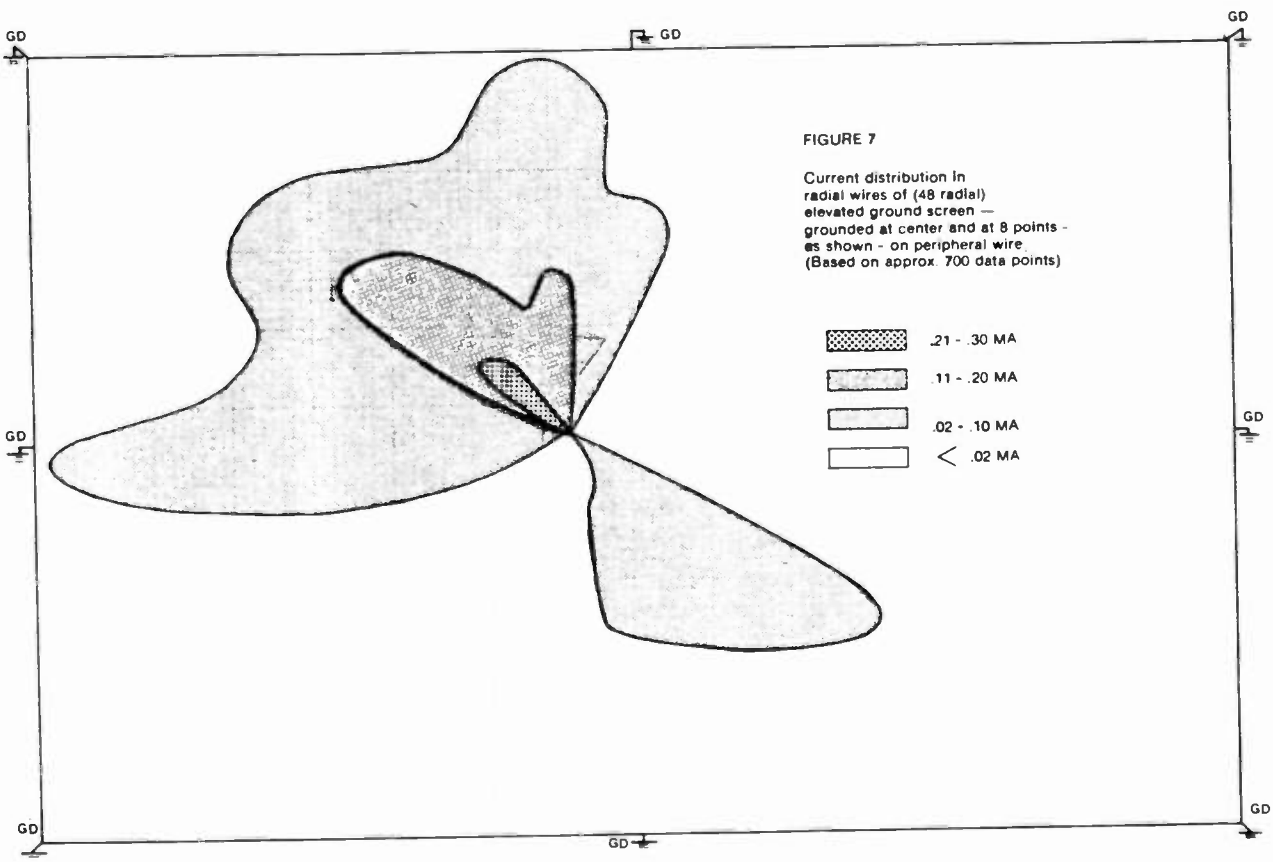


FIGURE 7

Current distribution in radial wires of (48 radial) elevated ground screen — grounded at center and at 8 points - as shown - on peripheral wire. (Based on approx. 700 data points)

The following combinations of single radial wires were tested:

SINGLE WIRES	LENGTHS OF WIRES TESTED
ELEVATED	45, 90, 135 & 180 feet
ON THE GROUND	45, 90, 135 & 180 feet
BURIED	45, 90, 135 & 180 feet

COMBINATIONS OF WIRES	LENGTHS OF WIRES TESTED
180 foot BURIED WIRE- under ELEVATED WIRES of	45, 90, 135 & 180 feet
180 foot ELEVATED WIRE- over BURIED WIRES of	45, 90, 135 & 180 feet
180 foot WIRE ON THE GROUND- over BURIED WIRES of	45, 90, 135 & 180 feet
180 foot BURIED WIRE- over WIRES ON THE GROUND of	45, 90, 135 & 180 feet

Radial current measurements were made every 10 feet on each of the various wires listed above.

In these tests the "Elevated" wires were 6' 6" above ground level, the "On the Ground" wires were lying loosely on the surface and the "Buried" radials were 8" below the surface, as shown in Figure 8.

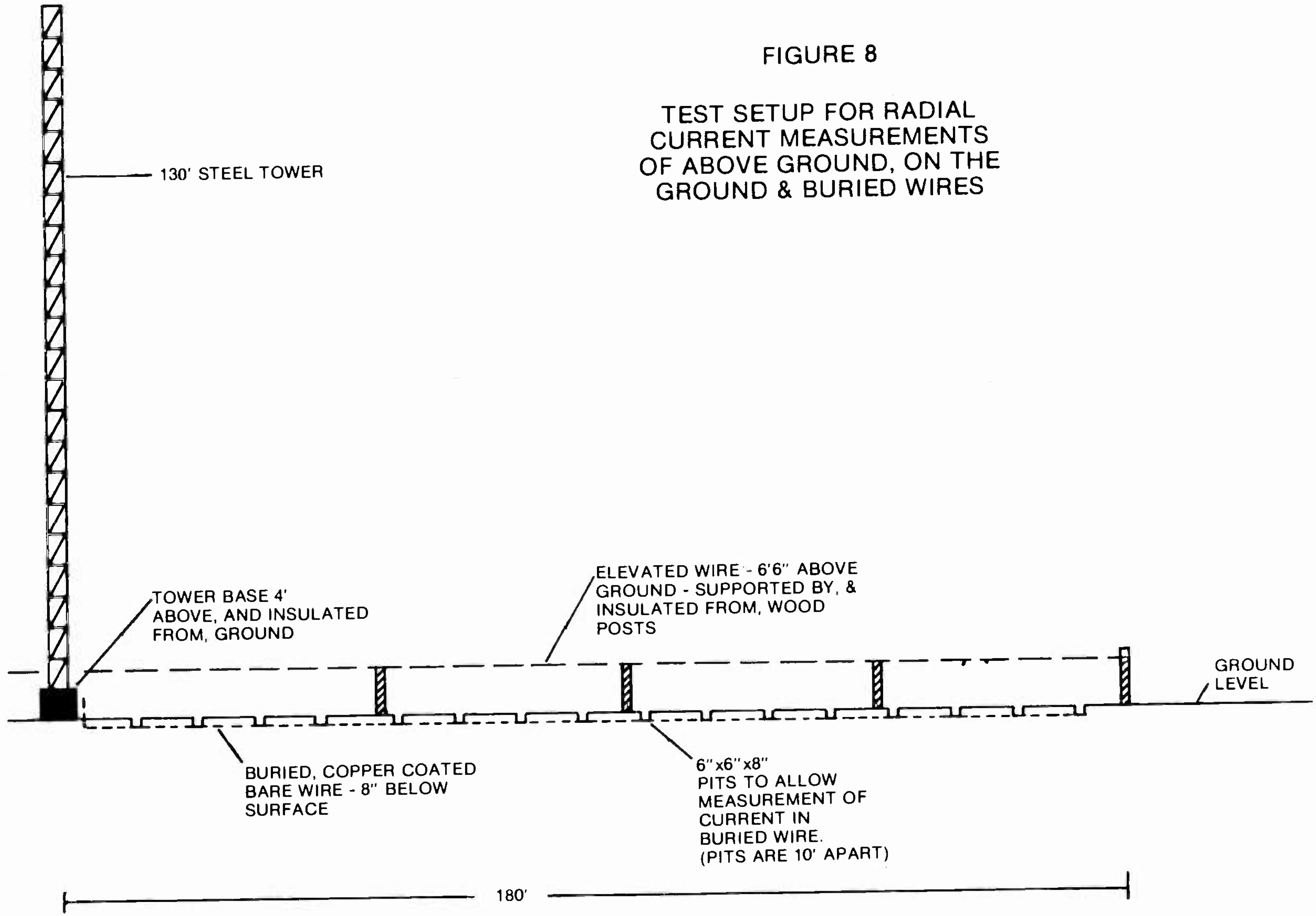
Location of the elevated and on the ground wires was directly above the buried wires.

The results of this series of tests were compared and supplemented by data from the 1981- 1982 tests. This was possible, since the earlier tests were conducted on the same site, and the same instrumentation and testing techniques were employed.

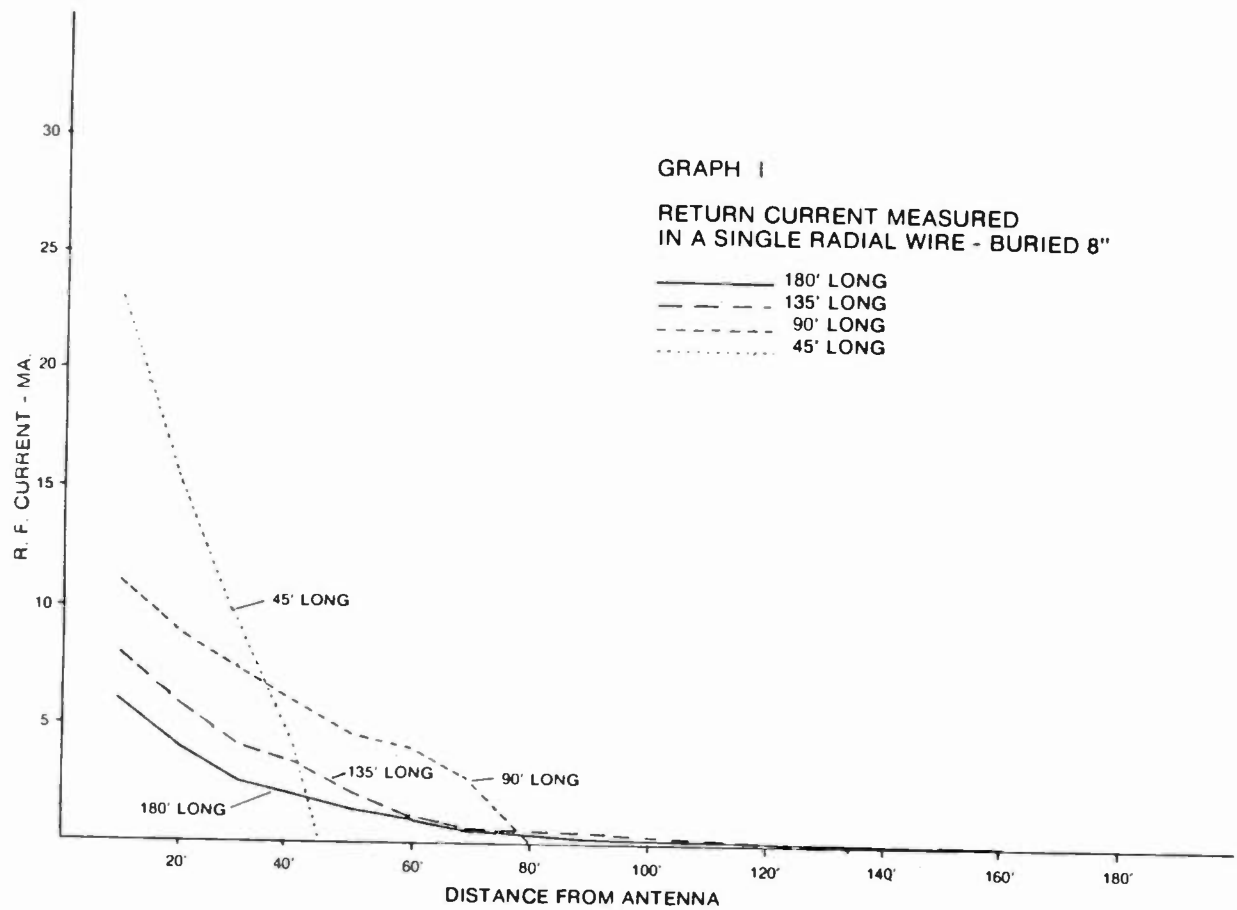
The following observations are based on these studies:

FIGURE 8

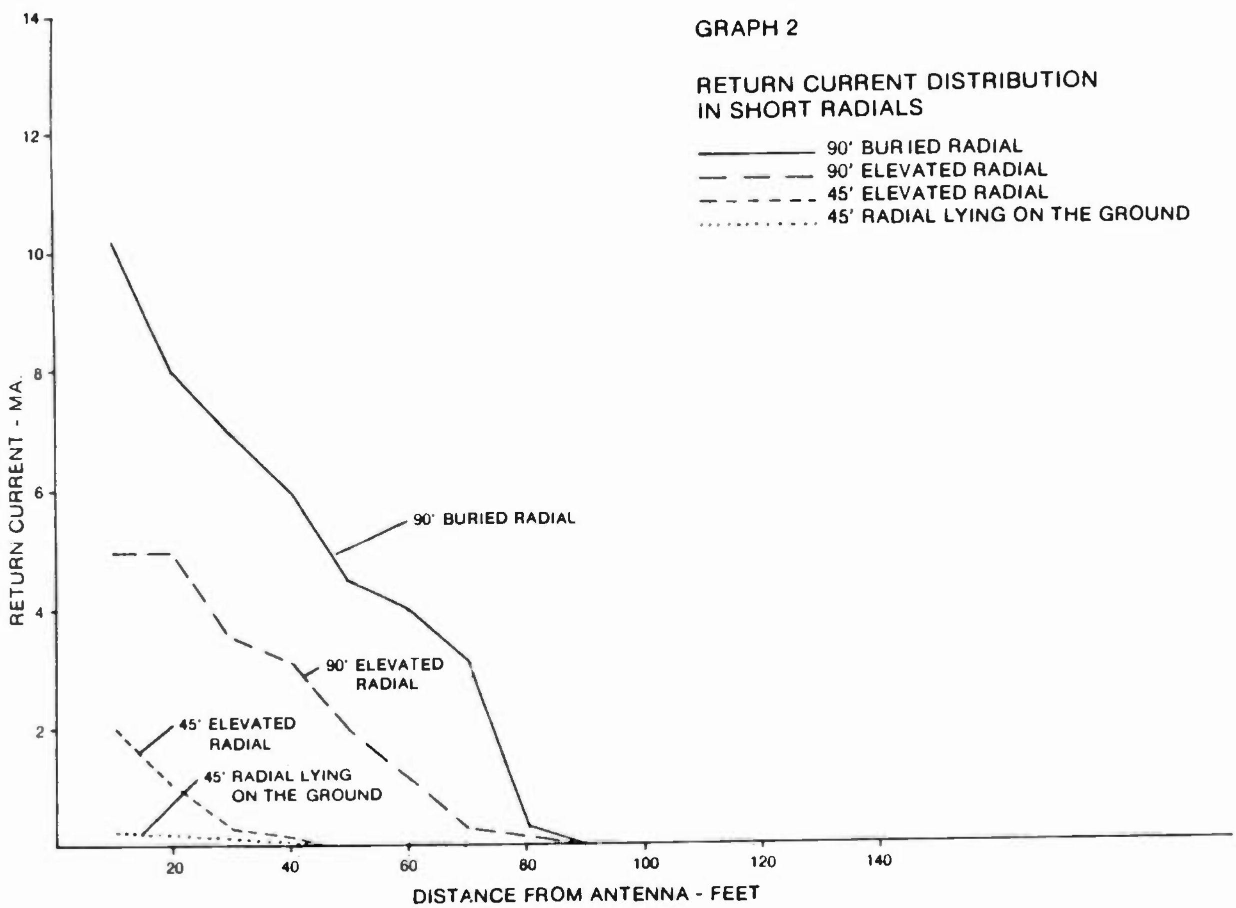
TEST SETUP FOR RADIAL CURRENT MEASUREMENTS OF ABOVE GROUND, ON THE GROUND & BURIED WIRES



I: The currents in buried, non insulated radial wires are concentrated near the base of the antenna, and decreases at a generally constant rate as the distance from the antenna increases. This is shown in Graph 1, and confirms what has been described in the literature for several decades.



II: As shown in Graph 2, the distribution of return currents in elevated or insulated radial wires having lengths less than approximately 0.20 wavelength is the same as that found in buried bare wires.

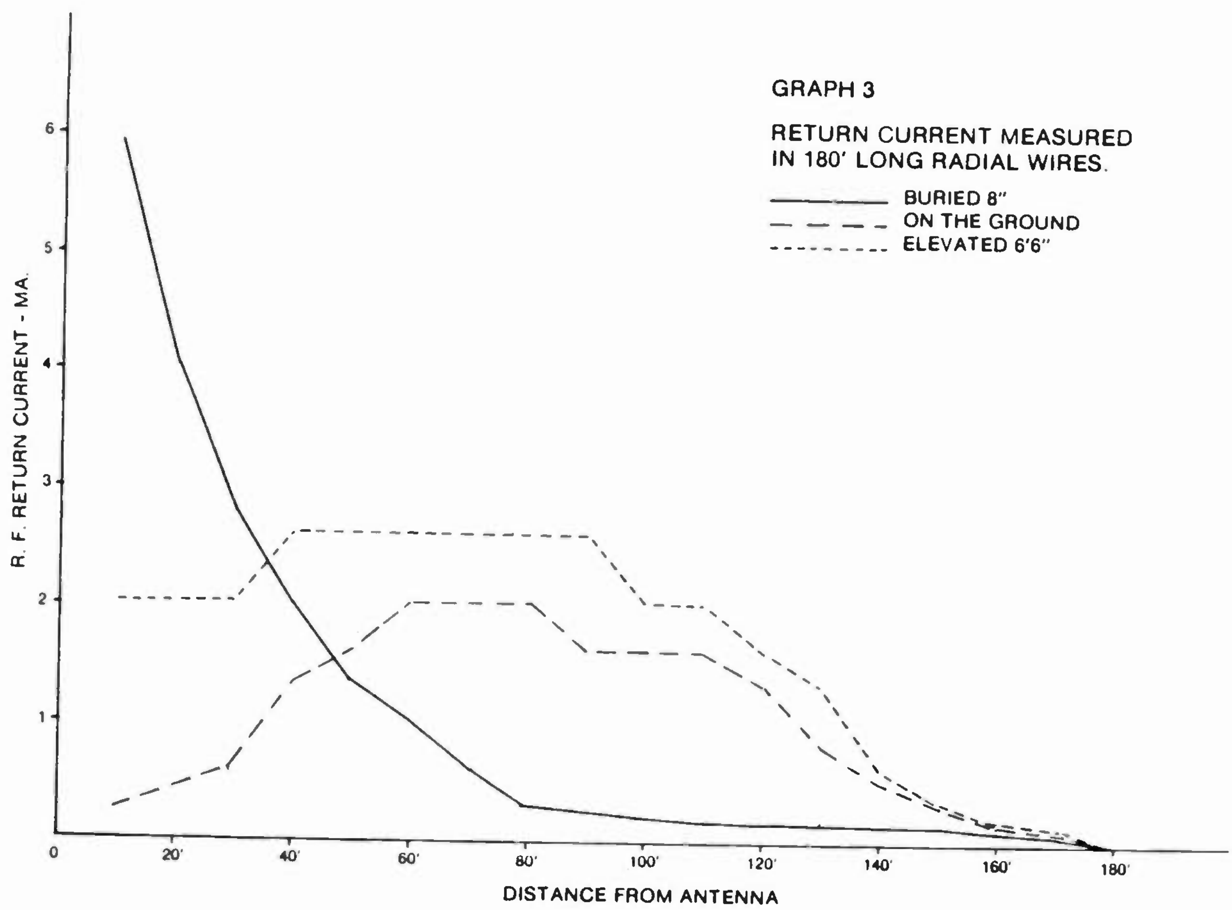


III: Graph 3 shows that the distribution of return currents collected by elevated wires can be distinctly different from that found in buried radial wires. Extensive testing showed that this phenomena exists so long as the elevated wire extends more than approximately 0.20 wavelength from the base of the antenna.

In these elevated wires the current level is moderate at the base of the antenna, rises very slightly as the distance from the antenna increases, remains at an almost constant level for a considerable distance and then graduately decreases to zero.

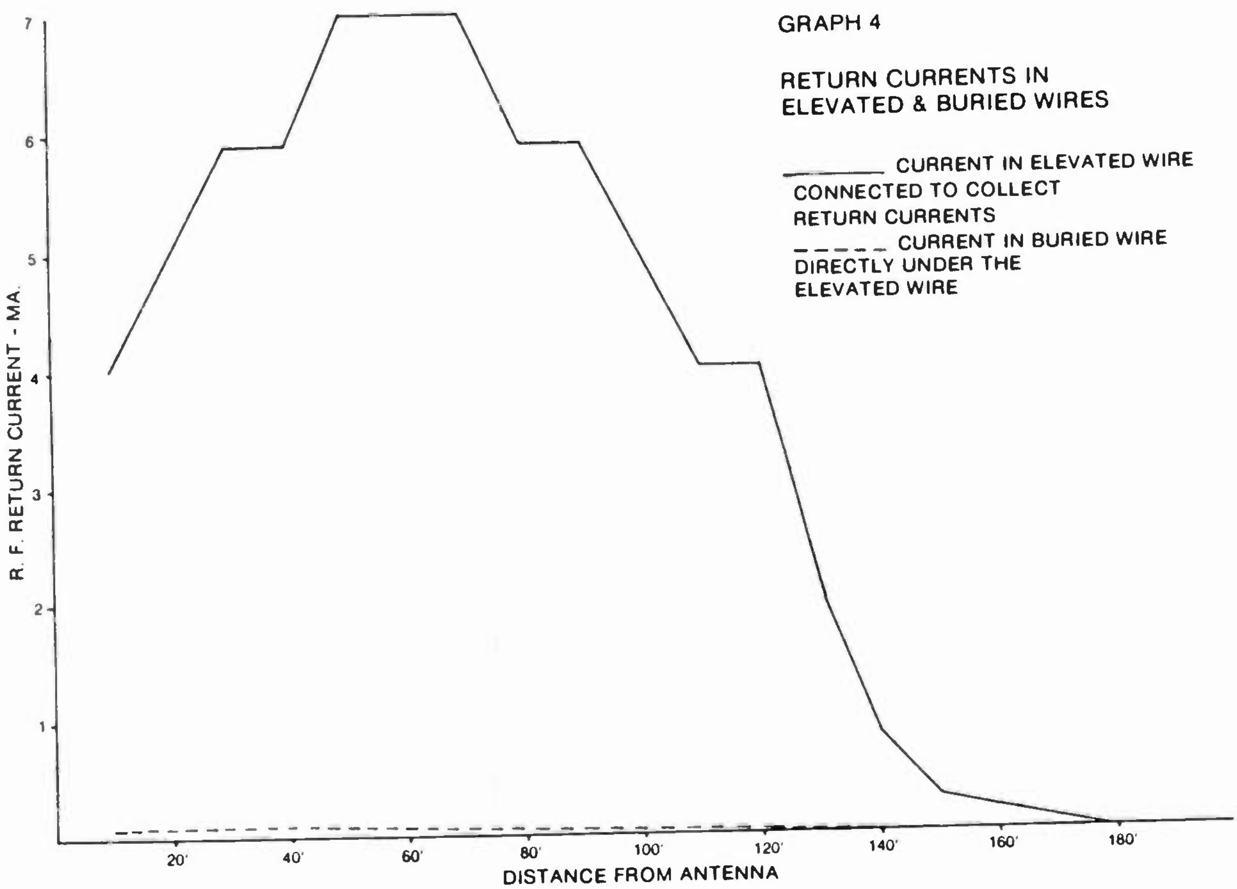
This same current distribution pattern was consistantly found in similar elevated radials tested in the very comprehensive (16,000 measurement) previous Fletcher test program. It has since been observed in similar counterpoise radials of another research antenna extensively tested at a different location.

It is of importance to note that the current distribution pattern of these elevated radial wires results in lower losses (R_1 in Figure 1) than those caused by the high level of currents adjacent to the antenna when buried radials are used. In practice the high level of current found near the base of the antenna with buried radials has resulted in such high heating (I^2R) losses as to set grass afire near the base of a low frequency vertical antenna!



IV: Graph 3 also shows that the current distribution found in insulated radial wires lying on the ground is similar to that found in elevated wires. Again, tests showed that this effect occurs so long as the radial is longer than approximately 0.20 wavelength. Thus radial wires on, but insulated from, the ground provide the much of the same advantage over buried bare radials as is discussed above for elevated radials.

V: Elevated ground systems- counterpoise or ground screen- are extremely efficient in intercepting return currents directly from the antenna- before they can reach the ground. As is shown in Graph 4, there are very small currents in the ground under elevated elevated radials connected to collect such currents. Those currents that do exist in the ground are conducted to the elevated wires as displacement currents.



The new test program has indicated that a properly dimensioned elevated ground system, or a radial system using insulated wire, operates more efficiently -with less losses- in collecting return currents from a vertical antenna than does the traditional buried, bare wire, ground system.

One of the reasons for the low efficiency of buried bare wires is that the return currents which they collect must pass through the ground for a greater or lesser distance before they reach the location of a radial wire. There are relatively high ohmic losses in this process, as the earth through which the currents must flow has finite resistance. This situation becomes serious near the antenna, where the return currents in buried wires are concentrated, and (I^2R) losses become high.

In contrast, elevated and/ or insulated radial wires collect the majority of the currents which they carry directly from the antenna -and the balance as displacement currents from the earth below the wires. The losses involved in the passage of these return (displacement) currents to the elevated wires in each case involves only the relatively low losses involved in their transference through the dielectric between the wire and the antenna or the ground--air.

The earlier Fletcher tests provided specific data that illustrated the degree of efficiency of elevated or insulated radials in collecting return currents and conducting them back to the feed point of the antenna: When at least 48 radials were used (in either counterpoise or ground screen configuration) losses became insignificant. No measurable improvement was found when additional radials were added to the ground system.

CONCLUSIONS

From analysis of the more than 20,000 measurements made in the Fletcher test programs we conclude that elevated, or buried but insulated, radial wire arrays fulfill the theoretical requirements for artificial ground systems at least as well as do the buried (bare) wire arrays commonly used in the Standard Broadcast Service.

The primary reason for this high efficiency can be best expressed[6]:

------IT IS MORE EFFICIENT TO COLLECT
RETURN CURRENTS FROM VERTICAL ANTENNAS
AS DISPLACEMENT CURRENTS RATHER THAN
AS CONDUCTION CURRENTS--------------

*   *   *   *   *

RECOMMENDATIONS

From the data presented above several specific recommendations are offered for consideration in the design of future artificial ground systems for use by AM Broadcast stations.

RADIAL WIRES SHOULD NOT BE BURIED IF THEY ARE TO COLLECT RETURN CURRENTS FROM A VERTICAL ANTENNA WITH MAXIMUM EFFICIENCY

IF POSSIBLE, ELEVATED RADIALS SHOULD BE USED. THESE CAN BE INSULATED (COUNTERPOISE) OR GROUNDED (GROUND SCREEN)

IF IT IS NOT POSSIBLE TO USE ELEVATED RADIALS, INSULATED WIRE LYING ON THE GROUND, OR BURIED AS CLOSE TO THE SURFACE AS POSSIBLE SHOULD BE USED.

ELEVATED RADIAL WIRES, OR INSULATED RADIALS AT GROUND LEVEL, SHOULD BE AT LEAST 0.20 WAVELENGTH LONG, IF PHYSICALLY POSSIBLE.

Implementation of these recommendations by AM Broadcast stations would, of course, require reevaluation of present manditory requirements. That such reevaluation might be in the public interest is illustrated by the following example:

Under present requirements a 1000 Khz AM Broadcast station must install a minimum of 120 buried copper wire radials- each at least 1/4 wavelength long. This totals 29,520 feet of wire.

If the above recommendations could be followed, and a counterpoise ground system, or insulated, rather than bare, radials were used, only 50 radials, each 1/5 wavelength long, should prove to be sufficient. This would require only 9,840 feet of radials- or 1/3 that presently used.

It is, of course, understood that esthetic requirements, economics, subsoil conditions and other factors must be considered in the determination of the proper ground system for an AM Broadcast station. As shown in this paper, however, there are alternatives to the traditional ground system that provide highly efficient electrical performance, require less space and, as an incidental bonus, provide a solution to problems of corrosion of buried wire systems.

ACKNOWLEDGEMENTS

This paper is dedicated to the late Edmund A. Laport. Without his encouragement, friendship- and an occasional prod the research reported here would not have been completed.

Appreciation is expressed to the Radio Club of America, for permission to include here portions of the paper first presented at the Radio Club's Diamond Jubilee meeting in New York, November, 1983.

The brilliant editing of this paper by John A. Frey will be as much appreciated by the reader as it was by the author.

REFERENCES

1. FCC Rules and Regulations, Part 73, Subpart A, 1972.

2. Brown, Lewis and Epstein, "Ground Systems As a Factor in Antenna Efficiency", Proceedings of the IRE, June, 1937.

3. Spencer, L., "John L. Reinartz", Ham Radio, Aug. 1981.

4. Doty, Frey and Mills, "Characteristics of the Counterpoise and Elevated Ground Screen", Professional Program, Session 9, Southcon '83 (IEEE), January, 1983.

5. Doty, Frey and Mills, "Efficient Ground Systems for Vertical Antennas", QST, February, 1983, pp. 20-25.

6. Laport, E. A., Private Communication, 1983.

7. Doty, A.C., Jr.,"Performance of Conventional Buried Wire Radials Versus Elevated/ Insulated Radial Wires Used As Ground Systems for Vertical Antennas", Diamond Jubilee Meeting, Radio Club of America, November, 1983.

Note: Copies of the full listing of the 73 references utilized in the design and implementation of the test programs referred to in this paper are available from the author.

AURAL BROADCAST STL SYSTEM DESIGN

John A. Moseley

Moseley Associates, Inc.

Goleta, California

The Aural Broadcast Studio-Transmitter Link (ABSTL) has become, in the past decade, a popular means to convey the aural intelligence to the modulation input terminals of a remotely-located broadcast transmitter. It has been proven to be an extremely reliable, cost-effective, high-quality audio link with considerable flexibility and versatility. It can accommodate all forms of broadcasting as it is practiced today. By no means has the STL development reached its zenith, for as technology advances, so will improvements in the equipment design. Thus, the ABSTL will continue to remain an important part in the future development of broadcasting.

One of the objectives of this paper is to assemble together the major design considerations associated with the installation of STL systems so that the reader will have an overview of a complete installation. Once these areas are recognized, more detailed information can be found in the published literature, from suppliers of hardware and from broadcast consulting engineers.

For those broadcast stations who are not presently using an ABSTL, the recent breakup of the Bell System, as mandated by the government, may motivate them to examine the use of a radio link from a purely economic point of view. Now, reports are being received about new rate increases for wire lines. Without implying any lack of interest, enthusiasm or service by telephone companies, it should be recognized that the growth of the data-related communications needs of our society has sharply risen over the past 10 to 15 years, while broadcast-quality circuits have declined in use. At least one of the new companies created by the break-up of AT & T is promoting data over voice for such tasks as reading utility meters, assisting in a medical diagnosis, shopping, home banking and computer conferencing. Some of you may recall the fate of the DC pairs that were widely used 20 to 30 years ago for remote control purposes. Experience now tells us that these circuits are essentially non-existent in this country today. And, so it is without any great stretch of the imagination for one to wonder about the future of hard-line program circuits. It is fair to say, however, that their demise is not imminent.

In considering the applications for which one can use an ABSTL, a brief survey of the history of the evolution of the STL might be helpful. In the late 1940's and into the 1950's, the STL was used primarily by a handful of FM broadcast stations. One may also recall that a number of TV stations were using the 950 MHz aural STL band for the conveyance of the TV sound information. In most cases, the motivating factor for using an STL for TV audio was the questionable performance of the diplex sound channel on certain video STLs. Besides the main channel modulation, some early FM stations used control tones in the 17-25 kHz spectrum in conjunction with the so-called simplex operation. In this mode, a high-frequency tone was superimposed above the audio information whenever the microphone circuits were activated. This allowed the broadcaster to sell to certain establishments a decoding box which would mute the announcer's voice. This practice was one of the motivating factors in the Commission's decision to establish standards and regulations for Subsidiary Communication Authorization (SCA) broadcast service. Once these rules were in place, simplex operations were prohibited.

With the establishment of a viable SCA service, a rebirth of the FM commercial broadcasting industry occurred. Also, the initial impact of television had worn off and broadcasters were looking for new avenues of expansion. For some FM stations, the aural STL provided the means to operate their transmitters without any program or control lines. The main audio channel, remote control tones and any SCA subcarriers were conveyed by the STL. Telemetry was returned using an SCA channel on the main FM carrier.

The adoption of FM stereo standards in 1961 presented new challenges in the design of STL systems. Up to this time, the major concern of the design engineer was for main channel frequency response, low distortion, high SNR, and in cases of SCA operation, low crosstalk. The addition of the second sound channel for stereo operation offered some difficult challenges with respect to matching the phase and amplitude responses of the two channels. While STL systems were significantly better in this respect than two Class AAA circuits, they still produced measurable vector or linear crosstalk levels when performance measurements were made. While not distracting to the listener, this distortion made proof measurements difficult for some stations. Bear in mind that the word 'crosstalk' has slightly different meanings for stereo and SCA operation.

At that point in time, virtually all ABSTL systems used phase-modulated transmitters. This precluded the transmission of composite stereo signals to the input of the FM transmitter. Also recall that most FM transmitters at that point used the serrasoid or phase-modulated exciter. Only one manufacturer offered a direct FM exciter for sale, it being retained from its earlier days. This was the birth of the dual ABSTLs for two of them were required to convey the left and right audio channels to the remote transmitter site. Because of their relatively narrow bandwidth requirements, it was possible to place them in one 500 kHz STL channel by offsetting the two carriers above and below the assigned channel center. SCA, remote control and cueing subcarriers were added with relative ease, with all modulating components still retained in the single 500 kHz channel. In some cases, separate antennas and transmission line systems were used for both STLs, while others combined the energy into one common antenna and transmission line system. Still others combined receiving antenna systems while maintaining separate transmitting antennas. When dual antenna systems were used, they were often cross-polarized to obtain additional isolation.

When the FM transmitter manufacturer converted to direct FM exciters, the

ABSTL design followed suit. Thus, the stereo generator could be located at the studio. Now, the audio signal was handled on a composite basis and was not demodulated until the signal arrived at the listener's receiver. As the design of the overall broadcast chain improved, the need for a "transparent" STL became more necessary. The noise, distortion, frequency and transient response became very important. Fortunately, the linearity required to achieve these characteristics also improved SCA performance. However, the interest in SCA operation declined as stereo grew. The design of the FM stereo home receiver was also undergoing change in that period as often the mere presence of an SCA signal caused birdies to appear in the stereo output channels. FM receiver design has made great strides in the past 20 years. So today, one may note that the design of the ABSTL has kept pace with the advances made in FM exciters and stereo receivers. Poor sound these days is usually the result of a poor audio source or, regrettably, sloppy engineering. Advances in the design of today's higher-quality STL systems include the use of low phase noise oscillators, frequency synthesizers, spectrally-pure broadband power amplifiers, low-noise, high dynamic range receiver front ends, and phase linear IF amplifiers and low-pass filters. These techniques allow transient reponse to be optimized. So, not only does one find superior audio performance in today's STL, but technology advances may be found in all areas of the STL design. Increased selectivity is needed to reduce adjacent channel interference resulting from spectrum congestion. Better dynamic range capability is required to minimize third order IMD products generated by simultaneous adjacent and alternate channel signals. These forms of interference become negligible when using new equipment, such as the Model PCL-606, designed for the high-density areas. A signal-to-noise ratio in excess of 70 dB can be obtained with this equipment in the presence of an adjacent and the second adjacent channels whose combined power exceeds the desired signal by 20 dB at the receiver input. STLs are more immune to high external RF fields, power line noise and variations, and with battery backup, power line failures. As an analog system, today's STL is state-of-the-art. Extremely flat and wideband frequency response, distortions in the 0.1 to 0.15% range, and signal-to-noise ratios in excess of 72 dB are common in new installations in the field. The breakthrough in these numbers, particularly the latter, will result with the development of digital STL systems.

The earlier dual STL systems provided a degree of redundancy for the less reliable vacuum tube systems. This was an important factor in many installations. Today, redundancy is still provided with automatic changeover systems at both ends of the link. However, the high reliability of today's solid state systems has decreased the redundancy desire in some small market stations. Redundancy should be considered for the ABSTL if the station has a policy of redundancy in general. If no other emergency back-up link is available, such as a telephone or RPU, a redundant STL can save valuable air time, especially during a rating period. Stations that lack adequate test equipment should also consider a standby link as more time than can be permitted may be required to locate and replace a defective component or module. While there have been significant changes in equipment and mode of operation in STL systems, the propagation considerations have remained essentially the same. That is, the average length of a typical path is still from 30 to 40 km in length, the physics of propagation remain intact, although we may know more about it, and antennas have changed very little. Congestion in the 950 MHz spectrum in some of the larger metropolitan areas has resulted in larger antennas to achieve greater directivity. Some new transmission lines offer reduced losses and provide improved performance in weather extremes. A proper path engineering study may involve assuring adequate Fresnel Zone clearance or, in some cases, diffraction effects with a knife-edge obstruction or possibly the effect of vegetation on propagation over grazing paths, especially during summer months or periods of heavy moisture.

In the past 10 to 15 years, the need for two-hop systems has increased due primarily to the cost and reliability of satisfactory wireline circuits. Also, as the economics of commercial broadcasting increases, the cost associated with delivering signals to the remote location has less impact on the viability of the broadcast station. The need for two-hop systems is relatively small, however. This limited market does not readily encourage the development of commercial heterodyne repeaters. The current practice is to use two separate STL systems for establishing a two-hop installation. This is referred to as a remodulating repeater and is considered by many to be superior to the heterodyne repeater. This may not appeal to the purist instinct in us, but it has its advantages. In some cases, it is necessary to have the baseband, or composite stereo signal, available at the repeater point. The need for a two-hop link, in some cases, has arisen from the presence of an interfering building in the radio path. Limited success has been obtained using back-to-back antenna configurations to "bend the beam" around or over the obstacle. Passive repeaters have not proven to be attractive at the 950 MHz spectrum because of size and cost. Akin to this concept, however, is one installation in a major metropolitan city that uses the side of a large nearby building to change the direction of the radio path some 150°. There have been isolated cases where an amplifier has been installed between the back-to-back antennas to offset losses. This practice should be carefully reviewed before implementation as the bandwidth of the amplifier is usually greater than 5 MHz, the width of the present STL band assignment at 950 MHz. The possibility of amplifying or creating unwanted signals, that is, IM products, is enhanced with its use to the possible detriment of other nearby users or to a frequency planning coordinating group. Further, if the gain becomes excessive in the amplifier, or the input/output isolation between the antennas is marginal, the system can become unstable and oscillate.

Assuming the proposed path has Fresnel Zone clearance, some simple but meaningful calculations can be made. One needs to know the power output of the transmitter, expressed in dBm, the sensitivity of the STL receiver, also expressed in dBm, which will provide an adequate signal-to-noise ratio plus fade margin for the system, the path attenuation in dB, and finally, the net dB gain of the antenna and transmission line systems. The manufacturers of the equipment can supply these figures (see Appendix A). Usually, shorter paths require less fade margin than the longer ones. For grazing paths, close attention should be paid to the intervening terrain and vegetation. Adequate Fresnel Zone clearance does not always necessarily ensure high reliability, as surprising effects can be experienced due to ducting atomospheric conditions. Several years ago, a path was established between two 2000 m peaks separated by slightly less than 200 km. When first installed, the received signal closely approximated the calculated value. However, within a month or so, the path suffered complete blackouts for several days. The designers apparently failed to consider the effects of a very hot valley at the middle of the path. On clear paths of approximately 40 km in length, unusual propagation disturbances are minimal. For a more rigorous path study, the system signal-to-noise ratio can be determined using RF bandwidth (B), absolute temperature (T) and Boltzman's constant (k) to determine the thermal noise, as modified by system losses and receiver noise figure. The noise figure of the receiver indicates its goodness compared to theoretical noise (kTB). This noise figure, kTB, path attenuation, antenna system gain and transmission line losses are all required in this more rigorous calculation.

As many see it today, the basic difficulty in placing new ABSTL systems in place is not equipment performance or path problems; rather, it is the lack of frequency spectrum. While the Commission has taken note of the problem, very little real progress has been made to relieve the congestion in the 950 MHz spectrum. In numerous areas, all of the STL channels have been assigned and are in use. To accommodate new applicants in these areas, channels must be shared if the geometry of the path permits. Consideration should be given to large objects in the path as reflections from structures, such as large storage tanks, may cause unwanted reflections to appear at the input of the undesired receiver. Polarization of the antennas has also aided in minimizing interference. For aural broadcast service, it was indeed a sad day when the Commission removed 5 MHz from the former 10 MHz STL band for services such as cellular radio, a plan which is yet to be fully implemented. In fact, the first system was inaugurated only last October in Chicago. New spectrum space is sorely needed. A number of foreign countries with less broadcast activity than the United States have recognized this problem by allocating other frequency bands for STL service. An expansion into the 1.7 GHz to 2.5 GHz spectrum will be most welcome. STL equipment for this frequency is now becoming available. The propagation characteristics for this band are not much different than those of the present 950 MHz band and should meet the requirements of broadcast ABSTL service. Allocation of channels in the 18 GHZ spectrum is not seen as a real solution to this problem. In this spectral region, propagation characteristics are seen to be more of a deterrent than on frequencies below 2.5 GHz, especially for all but the very short paths. Atmospheric gases and water vapor absorption become increasingly important. A 50 mm per hour rainfall can introduce some 5-6 dB/km of loss for a mean drop size distribution. In sophisticated millimeter systems, path diversity is an effective countermeasure against rainfall. However, this technique does not lend itself to the commercial restraints of the broadcasting industry. It appears at this time that the 18 GHz band has its greatest appeal to short hops, particularly in large cities. As a point of interest, a station facing a two-hop path created by a nearby building field tested a laser link for the first hop, a distance of some 200 m. The test was a success but never implemented as the entire studio complex was moved to a site with a clear radio path. It is extremely important that the broadcast industry keep pressure on the Federal Communications Commission for additional exclusive channels for aural broadcast STL service.

To sum up the foregoing thoughts, the ABSTL is an established means for interconnecting a broadcast studio with its remote transmitter. It can accommodate all modes of standard and FM broadcast operations, including remote control and data transmission. Today's equipment introduces virtually no signal degradation nor adds significantly to the system noise. The equipment is reliable and requires little or no maintenance, although a regular checkup is highly recommended to detect any deviations. It is cost-effective and it is under your control. All of these factors add up to making the ABSTL a wise investment for a broadcast station.

APPENDIX

STL PATH EVALUATION

INTRODUCTION

This appendix provides instructions for preparing the STL path profile and the path evaluation calculations. The STL path profile is needed to determine a clear unobstructed line-of-sight path with the main incident beam clearing all objects by a distance that is at least 0.6 of the radius of the first Fresnel zone.

The path should be plotted on a seven-and-one-half-minute topographical map, which is available from the United States Geological Survey. The maps can be ordered as follows:

East of the Mississippi:

Geological Survey
Distribution Section
Washington, DC 20242

West of the Mississippi:

Geological Survey
Distribution Section
Federal Center
Denver, CO 80225

STL PATH PROFILE

Using ground elevations obtained from the topographical map, prepare an STL path profile using true earth radius graph paper. Be sure to include all obstacles in the path, including buildings, towers, grain elevators, recently-erected structures, etc., that may not appear on the map.

The axis of the main incident beam should ideally clear all obstacles in the path by a distance that is at least 0.6 of the radius of the first Fresnel zone. This factor is used to define the transition between the obstruction and interface zones regardless of the reflection coefficient.

The following formula may be used to determine the first Fresnel zone clearance:

$$F_r = 2280 \sqrt{\frac{AB}{f_o P}}$$

where

F_r = Fresnel radius (in feet) perpendicular to the main beam path

A = distance (miles) from receiver to obstruction

P = total path length (miles)

B = P - A (miles)

f_0 = carrier frequency in MHz

Draw a straight line on the STL path profile graph which clears any obstacle in the path by the distance calculated using the above formula multiplied by 0.6 which yields the critical dimension. This line then determines the STL antenna and/or tower heights necessary at each end of the path.

PATH EVALUATION FOR AURAL STL SYSTEMS

Using the System and Path Information Sheet and the Path Evaluation Form, prepare path evaluation calculations for each path under consideration. Follow these instructions for completing the Path Evaluation Form:

Line 1. Transmitter Power Output: Enter the transmitter output power in dBm. Examples: 5 W = +37 dBm; 7 W = +38.5 dBm; 8 W = +39 dBm; 10 W = +40 dBm and 50 W = +47 dBm.

Line 2. Transmitter Antenna Gain: Enter the gain of the STL transmitting antenna over an isotropic antenna.

Line 3. Receiver Antenna Gain: Enter the gain of the STL receiving antenna over an isotropic antenna.

Line 4. Total Gain: Add the total of lines 1, 2 and 3 and enter here.

Line 5. Path Loss: Use a straight-edge ruler on the free-space attenuation nomograph provided on the System and Path Information Sheet. With the right side aligned on f_0 and the left side at the path distance, read the unobstructed free-space attenuation in dB from the center scale. Enter this value on Line 5.

Line 6. Transmission Line Loss: Determine the total length of transmission line required in the STL system (at both ends, the transmitter and receiver); then enter the line length and line loss in dB.

Line 7. Connector Loss: Enter total connector losses. (Example: A nominal figure is 0.5 dB based on 0.125 dB per mated connector pair at 950 MHz).

Line 8. Other Losses: Enter other miscellaneous system losses here. These may include power dividers, duplexers, diplexers, circulators, isocoupler, and the like.

Line 9. Total Loss: Enter the total of lines 5, 6, 7 and 8. This is the total STL system loss.

Line 10. Total Gain: Enter the total gain from line 4.

Line 11. Total Loss: Enter the total loss from line 9.

Line 12. Effective Received Signal: Enter the difference between line 10 and line 11. This is the signal level in dBm (without fading) that we expect at the RF input connector of the STL receiver.

Line 13. Minimum Signal for 60 dB SNR: Enter the model number of the STL receiver. Using the listing on the System and Path Information Sheet, enter the receiver signal input (in dBm) required to achieve a 60 dB SNR output.

Line 14. Fade Margin: Subtract line 13 from line 12 and enter here. At 950 MHz, typical minimum fade allowances are:

Monaural - 15 dB
Composite - 20 dB

PATH EVALUATION FORM

Reference Number ________ Prepared By ________ Date ________

CUSTOMER ____________________

SYSTEM GAINS

1.	Transmitter Power Output (Model # ________)	+ ________ dBm
2.	Transmitter Antenna Gain (Antenna Type ________)	+ ________ dBi
3.	Receiver Antenna Gain (Antenna Type ________)	+ ________ dBi
4.	Total Gain	+ ________ dB

SYSTEM LOSSES

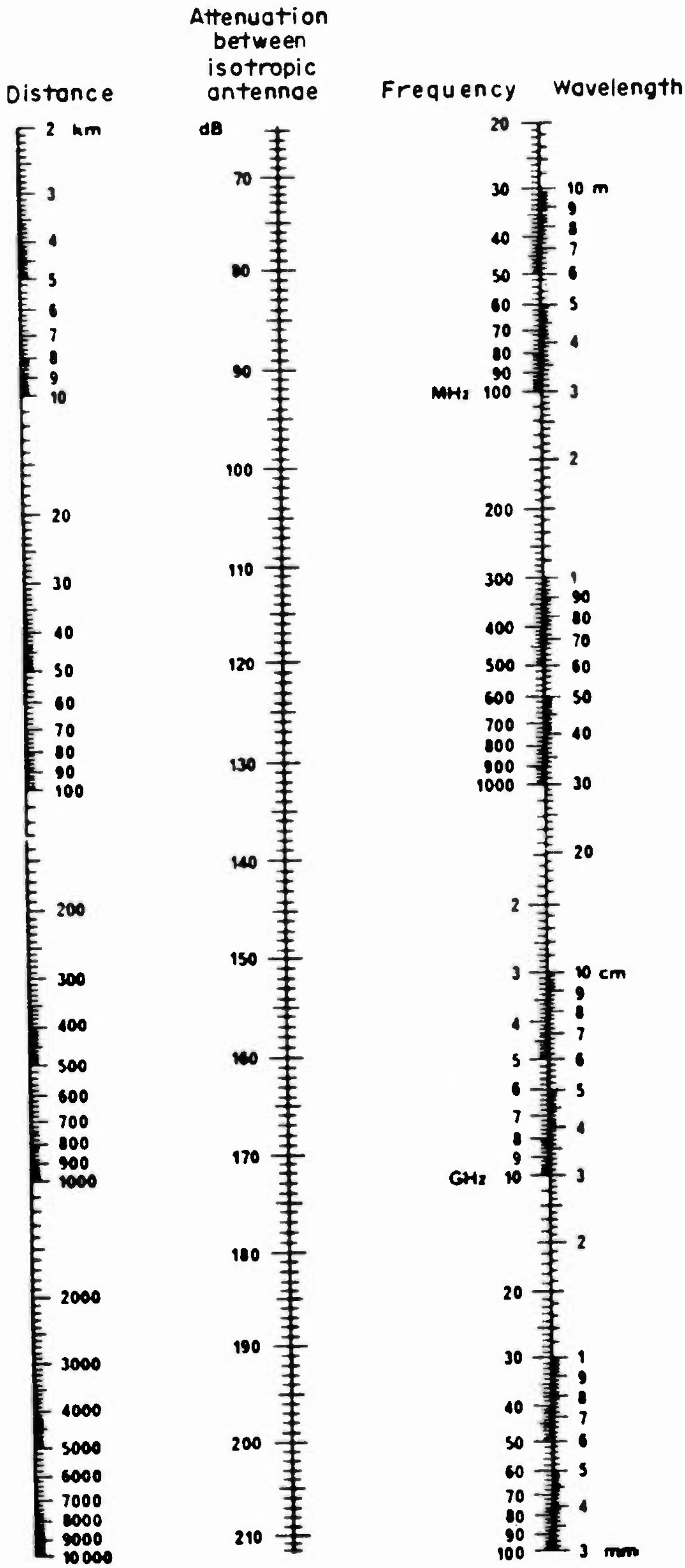
5.	Path Loss (________ miles/ ________ km)	- ________ dB
6.	Transmission Line Loss (Total ft ____/____ m) (Type ________)	- ________ dB
7.	Connector Loss (Total)	- ________ dB
8.	Other Losses ________	- ________ dB
9.	Total Loss	- ________ dB

SYSTEM CALCULATIONS

10.	Total Gain (Line 4)	+ ________ dB
11.	Total Loss (Line 9)	- ________ dB
12.	Effective Received Signal ________ µV =	________ dBm
13.	Minimum Signal Required for ________ dB SNR (Model ________)	- ________ dBm
14.	Fade Margin	________ dB

NOTES:

SYSTEM AND PATH INFORMATION SHEET
FOR TYPICAL 950 MHz STL SYSTEMS



Nomogram for Free Space
Attenuation Clear Path

TYPICAL RECEIVER SIGNAL
INPUT FOR 60 dB SNR DE-
MODULATED PROGRAM OUTPUT

PCL-101	-79.0 dBm
PCL-505	-83.0 dBm
PCL-505/C	-67.0 dBm
PCL-606	-84.0 dBm
PCL-606/C	-66.0 dBm

TYPICAL TRANSMISSION CABLES

Type	Attenuation (per 100 ft)
LDF4-50 (1/2 inch)	2.4 dB
LDF5-50 (7/8 inch)	1.4 dB

TYPICAL ANTENNAS

Type	Gain (dBi)
MF-960	16.1
PR-450U	20.1
4' P-948-GRN	18.9
6' P-972-GRN	22
8' P-996-GRN	25
10' P-9120-GRN	27
12' P-9140-GRN	28.5
15' P-9180-GRN	30

Audio Grounding for Hum and RFI Elimination

L. Scott Hochberg

Logitek Electronic Systems, Inc.

Houston, Texas

When we're designing our audio consoles and other audio products, we're confronted with many different audio signals confined within a small space. It's our responsibility to guarantee that the audio signals will not receive interference from each other, or from nearby transformers, power cords, transmitters or other devices. To do this, we've learned some audio "ground rules" which are equally applicable to larger audio systems, including radio or television studios. This paper will discuss those rules and some of their practical applications. We do not claim to have all the answers herein. But most stations we have visited limit their audio performance by failure to observe the teachings of the rules we are offering.

We've summarized our experience into five basic "laws", as follows:

1.) At least 90% of the troubles in the world are caused by bad grounds.

2.) The only good ground is the one below your feet.

3.) If one ground is good, two grounds are not better.

4.) If one ground in no good, two grounds may be better.

5.) Every amplifier is a differential amplifier.

We have proven the first rule time and time again. When we introduced a balanced-input phono preamp a few years ago, we guaranteed that it would solve turntable hum and RFI problems for the stations who used it. Our only condition on this guarantee was that before an engineer decided that the unit didn't work as we had promised, he call us and give us a chance to make it work right.

When an engineer would call to take us to task, we would discuss his audio grounding with him in detail, and recommeded specific ground connections for him to change. In virtually every case, hum and RFI pick-up were eliminated by making our recommended changes.

Of course, with the grounding system fixed, his old preamps which ours replaced might have worked fine, too. But we usually neglected to mention that.

The point is that many of the problems attributed to poor equipment are really ground installation errors. And that's the fault of manufacturers as well as engineers, since, in all the reems of paper that fill instruction manuals, I've never seen a good discussion of how to ground the equipment to get the specified performance away from the test bench.

We thought we had a pretty good handle on rule 1 when I worked at WTAQ in Chicago. I was a college student working summer vacation fill in, and our chief engineer had just come to WTAQ from one of the network shops downtown.

The station has an on-site 5 KW AM with four sticks, so there is enough of a field that sloppy grounding would cause trouble. And we indeed had trouble in our production room with a little bit of RF getting into recordings. It looked like a good project for us both to work on to impress the boss.

On first glance, we saw an obvious problem. All the shields were connected not only to the room's copper ground strap, but to every place else imaginable as well. We had more ground loops than you could shake a plier at. We got out our dikes and went to work. Strangely, though, as we cut more and more of the ground loops, the RFI got worse and worse. Nonetheless, knowing that ground loops work in strange ways, we pressed on, convinced that when we got rid of all the loops, we'd have a clean-sounding production room.

By the afternoon, all the loops were gone, and the production room sounded like an old Philco receiver. Not only were we getting great AM reception on our cart machines, but we picked up all kinds of pops, relays clicks, you name it. We wisely decided to take the rest of the day off, before the production manager came in to cut spots.

The next morning the chief was waiting for me, smiling. He had found the trouble. He led me to the mechanical room into which the production ground strap was fed. We followed it up the wall, over the furnace, across the dark corner where the emergency generator controls were hidden, past the "Air Alert" buzzer, and down to where it stopped -- three inches away from a pipe going into the ground.

Our "ground" strap was not a ground strap at all. It was a great antenna. Remember rule 2! The only good ground is the one below your feet!

By the way, there was quite a spark thrown when our chief brought the strap to the pipe. And, after connecting the ground strap properly, the production room sounded great.

Most examples of "grounds" which are not really grounded are not quite as obvious as this. For instance, one common practice is to assume that equipment chassis are good grounds. But think about this for a moment. As an equipment manufacturer, what are we concerned with when we design a chassis?

First, it should be rugged. The material must be sturdy enough to support the parts attached to it, even with mounting holes and often with vent holes punched into it. It should also be attractive. It also must conduct electricity, so that it can work as an electrostatic shield, but it does not have to be an extremely good conductor, since it is not designed to carry current.

The most common materials which meet all of the above requirements are aluminum and steel, so that is what most chassis are made from. There aren't many engineers who would use aluminum or steel ground straps, but many engineers connect shields and ground returns to "chassis ground."

By doing this, not only is the shield not attached to a good ground, but, by running ground currents across the chassis, you can reduce the effectiveness of the chassis in shielding the circuitry inside.

The real shame here is that most manufacturers go through great effort to design a central point on their equipment that grounds can properly be attached to. This point is usually either an isolated ground terminal or ground lug on a connector. If the manufacturer has done his job correctly, the designed ground point will easily accessible, and will be a far better place to connect ground wires than the chassis or the cabinet.

Another place where "grounds" are not what they seem to be is along so-called ground bus bars. Here I'm not talking about the two-inch solid copper strips that the networks install in their equipment racks. Rather, I'm refering to the copper ground wires mounted above the input terminal blocks on many older audio consoles for attaching shields and ground returns. The problem is that even copper has some resistance, and if you put enough current on a copper wire the voltage at one end will be higher than the voltage at the other end.

If one end of the bus is at ground, chances are the other end will be varying with the currents on the bus. If the bus is also a reference ground for an unbalanced input, all the unwanted signals on the bus will show up as noise and crosstalk when the unbalanced signal is amplified.

There are some places where this type of ground bus won't hurt you, but Murphy predicts that if you have a ground bus like this sooner or later the wrong signal will be hooked to it. The best plan is to avoid this type of structure.

Rule 3 refers, of course, to ground loops. We all know what ground loops are, so I won't dwell on them here. Sometimes, though, it's confusing trying to figure out whether or not we've created a ground loop. I always try to picture myself as an ant crawling along the ground wire. If I can get to true ground via more than one path, I have a ground loop.

One place to watch ground loops carefully is on unbalanced equipment. If you're feeding two signals to a stereo amp, for instance, make sure you don't ground the amp via the ground leads in both the left and right channel input cables.

Be careful on unbalanced phono preamp inputs, since the low level of phono signal makes it a good candidate for noise pickup. If the low side of the cartridge is grounded on both channels and your preamp is grounded via the power system or a ground strap, be sure not to carry your grounds through from the cartridge to the preamp. It's often difficult to visually tell how a phono cartridge is grounded, so it's a good idea to use an ohmmeter to check for continuity between the low side of each coil and ground so you can hook it up properly without guessing.

Now if rule 3 is for ground loops, then what is rule 4? If one ground is no good, how can two grounds be better?

We found out when we were designing our first audio console. Our original prototype popped loudly everytime we turned a channel on and off. We were using electronic switching, and solid state pushbuttons, so we were reasonably confident that it wasn't true "switch pop." Something else must have been happening.

A close look at voltages on the "ground" foils told the story. Whenever we turned on a channel, we also turned on an indicator light on the panel. We did this by saturating a transistor which had its collector connected to the light bulb and its emitter connected to ground. The other terminal of the bulb was connected to +12. When we turned on the transistor, it would pull current through the bulb to ground.

When you turn on a light bulb, the first thing that happens is that you get an in-rush of current, which quickly tapers off to the steady-state current consumption of the lamp. Thus, a current "spike" was created every time we turned on the bulb. The transistor shunted this spike to the ground bus on the circuit card, and, since the ground bus was not a perfect conductor, and voltage spike was created on the bus.

This ground bus also fed the "+" input of an op-amp which was being used "single-ended" to amplify the signal coming back from the channel gain fader. The problem was that rule 5 tells us that every amplifier is a differential amplifier, and this rule holds true even if the amplifier is in a "single-ended" circuit. Each op-amp still amplifies the difference between the signal on its "+" input and the signal on its "-" input.

If the "+" input is connected to a solid ground, the amplifier's output will be an inverted and amplified version of the signal on the "-" input. But if the ground is not constant, the variations on the ground bus will be amplified just as if there was a "real" signal there. That is what was happening in our console.

Whenever the light bulb turned on, there was a spike on the ground bus. This spike was amplified by the post-fader op-amp, creating a pop in th audio.

The solution was to separate the grounds, so that fluctuations on the light bulb ground couldn't affect the audio circuits. The same soultion was used to eliminate crosstalk between program and audition, between the monitor and the main channel, and everywhere else where signals could intermix improperly.

The same effects we found in our console show up repeatedly in broadcast stations we visit. Engineers often look for the most convenient way of grounding something, particularly when installing "non-critical" devices like control relays. But if you ground the coil of your turntable start relay through the shield of the phono preamp input wires, you may end up with relay pops in your phono audio. Similarly, if the ground from your monitor speakers joins up with the patch panel ground before it reaches your main ground point, you may hear your monitor feed all across your panel.

The key to good ground separation is to consider each ground wire and each shield as an integral part of the signal wire it works with. In other words, you wouldn't bus all your audio signals together, so don't bus your grounds. Use a separate ground wire for each piece of equipment you're grounding, and attach each ground wire to an substantial bus bar or to designated ground terminals on the central piece of equipment in each room, normally the audio console. Don't assume you can just "grab" a ground anywhere when you put in something new. Take the time to run proper ground wires back to the central ground point. The results will be well worth the effort.

There is obviously far more to proper grounding than we have covered in this quick overview. Nonetheless, if all broadcast engineers observed even these five simple rules, we believe "broadcast quality" would be substantially improved.

The Effects of Transient Overvoltages

On Broadcast Facilities

By Jerry Whitaker, Radio Editor

Broadcast Engineering Magazine

Overland Park, Kansas

The AC power line into a broadcast plant is the lifeblood of any operation. It is also, however, a frequent source of equipment malfunctions and component failures. The utility company AC feed contains not only the 60Hz power needed to run the facility, but also a variety of voltage sags, surges and transients. These abnormalities cause different problems for different types of equipment.

An AC voltage sag is generally defined as a decrease of 10-35% below the normal line voltage for a period of 16ms to 30 seconds. A surge, on the other hand, is a voltage increase of 10-35% above normal lasting 16ms to 30 seconds. Sags and surges may occasionally result in operational problems for the equipment on-line, but usually automatic protection (or correction) circuits will take appropriate action to ensure there is no equipment damage. Transients, however, are not so easily identified or eliminated. Many devices commonly used to correct for sag and surge conditions, such as ferrroresonant transformers or motor-driven autotransformers, are of limited value in protecting a load from high energy, fast rise-time spikes on the AC line.

The scope of the problem

Transient suppression is important to broadcasters because the sensitive, high-speed solid state equipment in use today can be disrupted, or even destroyed by random short-duration spikes riding on the AC waveform. If not attenuated, these brief pulses (sometimes only a few microseconds in duration) can destroy semiconductors, disturb logic operations or latch-up microcomputer routines.

Experience in the computer industry has shown that the vast majority of unexplained problems resulting in disallowed states of operation are actually caused by transient overvoltages on the utility feed. With the increased use of microcomputers in broadcasting, this warning cannot be ignored. The threat to broadcast facilities is compounded by the fact that microcomputers are being used at critical stages in the transmission chain, including program automation equipment and transmitter control systems.

The subject of transient overvoltages has been extensively studied in the computer industry. A pioneering effort by the IBM Systems Development division in 1974 (conducted by George Allen and Donald Segall), showed that voltage spikes lasting between 10 and 100ms in a frequency range of 10-100kHz occur on an average of better than 50 times per month in a typical commercial environment.

Other more recent studies have shown that power line transients caused by utility company switching, distribution system faults, large loads going on- and off-line, and lightning, can occur as often as 900 times per month. These spikes can reach 2kV (or more).

Assessing the threat

Someone once said that the best transient eliminator is a transient monitor. Anyone who has monitored primary power service lines with an oscilliscope for any length of time would surely agree with that statement. Recent developments in digital technology, though, have moved the business of assessing the threat posed by unprocessed AC from an educated guess to a fine science.

Sophisticated monitoring equipment can give the user a complete, detailed look at what is coming in from the power company. Such monitoring devices can provide a wealth of information on the problems that can be expected when operating data processing, transmitting or other sensitive electronic equipment from an unprotected AC line. Typically the power at a facility to be protected is monitored for a week or more, after which an assessment is made as to whether AC processing equipment is needed at the installation.

As a case in point, a recently-completed study for a San Francisco Bay area company planning to install a new data processing center graphically demonstrates the scope of the transient problem.

The firm wanted to determine the extent of transient activity that could be expected at the new site so that an informed decision could be made on the type of power conditioning needed. A Dranetz Engineering Laboratories model 606-3 AC line monitor was connected to the 480V dedicated drop at the new facility for a period of 6 days. During this time the monitor recorded <u>thousands</u> of spikes, many exceeding 2kV, on one or more of the three phase inputs. The transient recording threshold was 460V <u>above</u> the nominal AC voltage level of 480V, phase-to-phase.

An expert from the report summary states that on one particular day, the facility was plagued by many high-level transient periods, stretching from 8:30 a.m. until 3:00 p.m. In fact, the transient counters overflowed on the monitor's daily summary printout. The highest voltage recorded during this period was 4.08kV. (This transient activity occurred, by the way, during a period of good weather).

Figure 1 is part of the printout from this study. The data covers transients exceeding more than twice the normal line voltage that occurred

```
C  1792V  IMPULSE
A  0544V  IMPULSE
C  1984V  IMPULSE
A  0592V  IMPULSE
C  1856V  IMPULSE
A  0544V  IMPULSE
C  1824V  IMPULSE
A  0560V  IMPULSE
C  1488V  IMPULSE
A  0496V  IMPULSE
C  1664V  IMPULSE
A  0528V  IMPULSE
C  1600V  IMPULSE
A  0544V  IMPULSE
B  2480V  IMPULSE
11:19:11
```

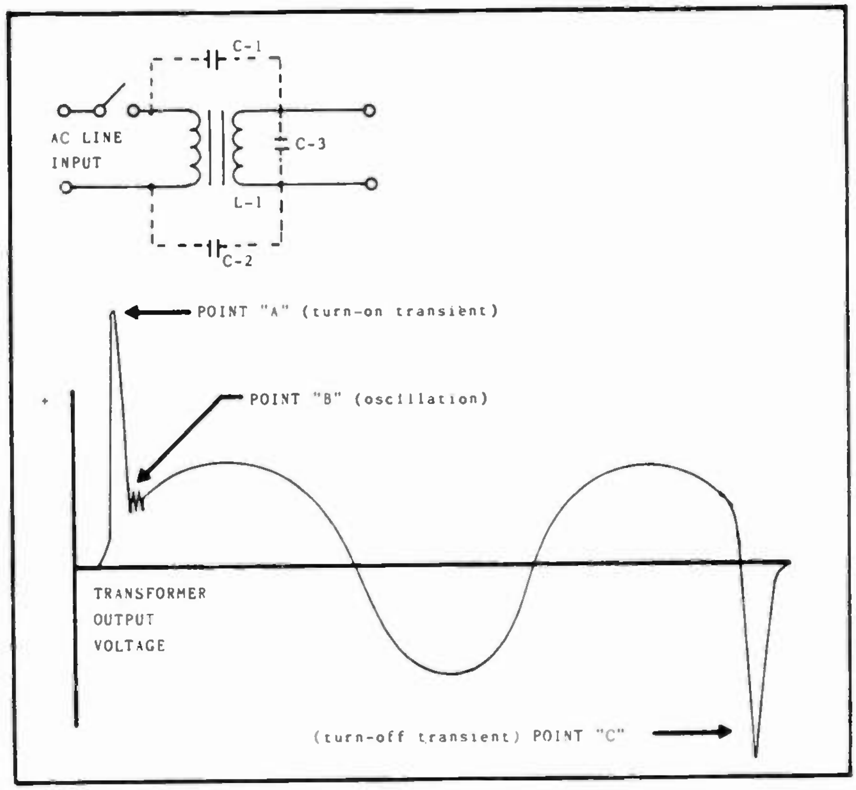


Figure 1 (on the left). A portion of the AC monitor readout from the San Francisco area power quality study. The first column indicates on which phase (A, B or C) the spike occurred. The second column is an actual readout of the transient (impulse) magnitude in volts.

Figure 2 (on the right). The causes of inductor turn-on and turn-off spikes. The waveforms are exaggerated to illustrate the transient effects.

C-1, C-2 and C-3 are stray capacitances which form a divider network between the primary and the secondary, causing the turn-on spike shown at point "A". The oscillation shown at point "B" is caused by the interaction of the inductance of the secondary (L-1) and C-3. The spike shown at point "C" is the result of power interruption to the transformer primary, which causes the collapsing lines of flux to couple a high voltage transient into the secondary circuit.

within a period of just 30 seconds. Even though these transients were very brief in duration, any sensitive equipment connected to the power line would suffer damage in a very short period of time.

While this is certainly a worst-case example of dirty AC, it points up the need for a minimal amount of spike protection on all incoming power lines. Studies such as the one detailed here should not be construed to be an indictment of utility company engineering standards. Very few power drops are as bad as the one analyzed in this study. Further, most transient activity on local AC lines is generated by power customers, not utility companies.

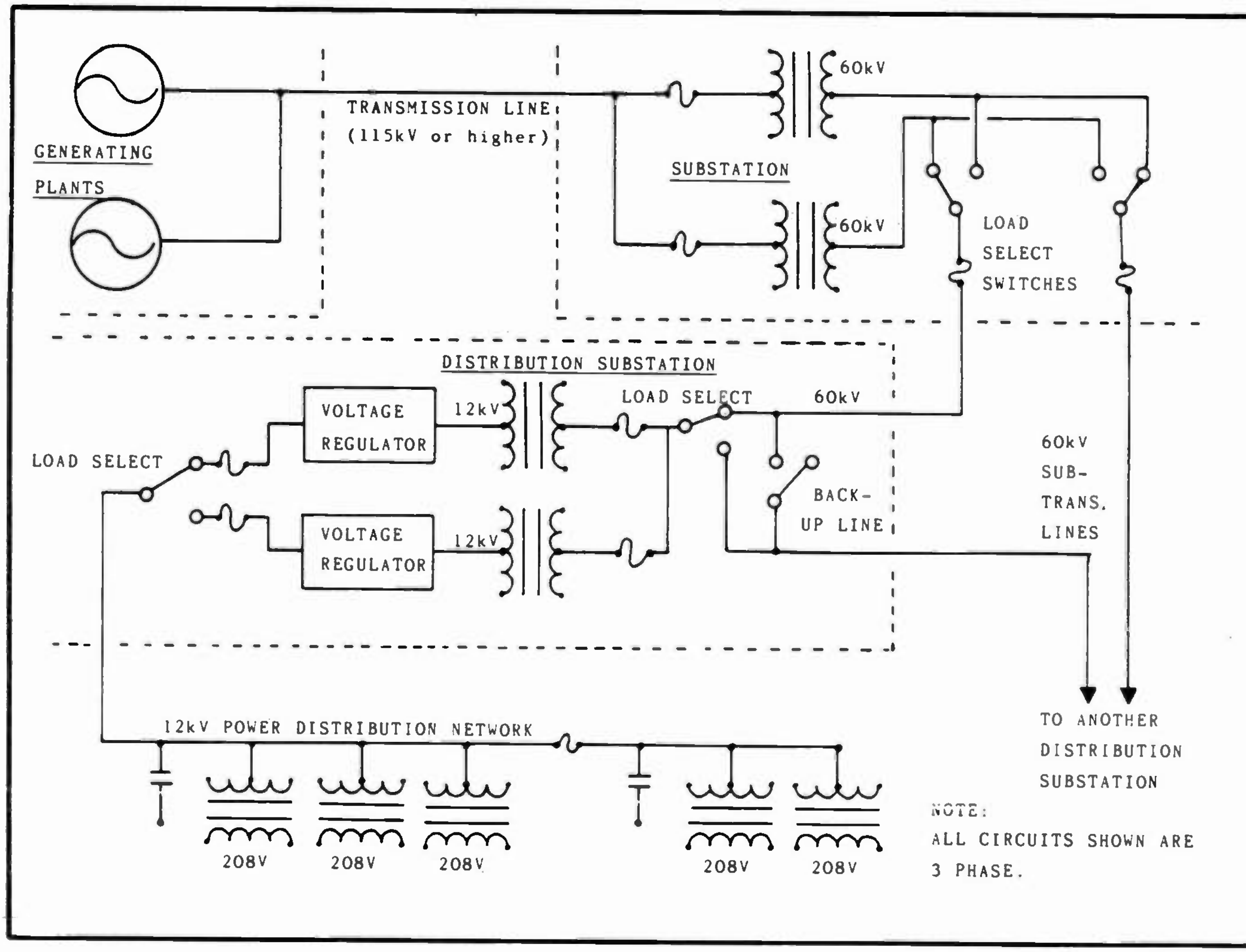


Figure 3. A simplified block diagram of a basic utility company power distribution system. The devices shown as fuses could be circuit breakers or reclosers, which function as automatic-resetting circuit breakers. A recloser will perform its interrupt-reset function a fixed number of times before being locked-out. The recloser must then be manually reset. All circuits shown are three phase. The capacitors perform power factor correction duty.

Unfortunately, the power quality problems affecting many areas of the nation are becoming worse, not better. Broadcasters cannot depend upon power suppliers to solve the transient problems that exist. Utility companies are rarely interested in discussing AC disturbances that are measured in the microseconds or nanoseconds. The problem must be solved instead, at the input point of sensitive loads.

Utilities have traditionally checked the quality of a customer's service drop by connecting a chart recorder to the line for a period of several days. The response time of such recorders, however, is far too slow to document any transient spike. Slow-speed analog recorders will only show long-term surge and sag conditions (as earlier defined), which can generally be dealt with by the regulated power supplies or high voltage protection systems normally used in broadcast equipment.

Transients on the utility power system are the result of basic laws of alternating current. A sudden change in an electrical circuit can cause a transient voltage to be generated due to the stored energy contained in the circuit inductances (L) and capacitances (C). The size and duration of the transient is dependent on the values of L and C and the waveform applied.

A large step-down transformer, the building block of a power system, can be an effective transient waveform generator when energized, or de-energized. As illustrated in Figure 2, the stray capacitances and inductances of the secondary can generate a brief oscillating transient of up to twice the peak secondary voltage when the transformer is energized. The length of this oscillation is determined by the values of L and C in the circuit. Another problem encountered when energizing a step-down transformer is that the load is looking into a capacitive divider from the primary. If the inter-winding capacitance is high and the load capacitance is low, a spike of as much as the full primary voltage can be induced onto the secondary, and thus the load. This spike does not carry much energy, because of its short duration, but sensitive equipment on the load side could be damaged upon re-application of power to a utility company pole transformer, for example, as would occur after a power outage.

De-energizing a large power transformer can also cause high voltage spikes to be generated. Unless switched-off at or near the zero-crossing, interrupting the current to the primary winding of a transformer can cause the collapsing magnetic lines of flux in the core to couple a high voltage transient into the secondary circuit. If a low impedance discharge path is not present, this spike will be impressed upon the load. Whether or not these turn-on, turn-off transients cause any damage is dependent on the size of the transformer involved and the sensitivity of the equipment connected to the transformer output.

These principles are the basis for several other commonly-experienced transients, such as switch arcing, fault spikes and relay transients.

The utility power system

The details of power distribution in the United States vary from one utility company to another, but the basics are the same. See figure 3.

Power from a generating station or distribution grid comes into an area substation at 115kV or higher. The substation consists of switching systems, step down transformers, fuses, circuit breakers, reclosers, monitoring and control equipment. The substation will deliver output voltages around 60kV to subtransmission circuits which feed distribution substations. These substations convert the energy to approximately 12kV and provide voltage regulation and switching arrangements that permit "patching around" a problem. The 12kV lines power the pole- and surface-mounted transformers, which supply various voltages (generally 208-240V 3 phase) to the individual loads.

Depending on the geographic location, varying levels of lightning protection is included as part of the AC power system design. Most service drop transformers (12kV to 208V) have integral lightning arresters. In areas of severe lightning activity, a ground wire will be strung between the top insulators of each pole, thus attracting lightning to the grounding wire, and away from the hot leads.

Capacitor banks are placed at various locations in the 12kV distribution system. The number of banks and where they are located in the system is determined by the load distribution and power factor of the circuit. The capacitors will improve the short-term line voltage regulation (in the millisecond range) and reduce transient activity on the line. Spikes are reduced by virtue of the fact that the capacitor will present a high impedance to the 60Hz line voltage frequency, but a low impedance to high frequency transients. The capacitors are placed on the line in order to keep the power factor as close to unity as possible. Transient suppression is simply a by-product.

Transients are generated in the utility company system in ways similiar to those outlined previously for inductors in general. Energizing or de-energizing a transformer can create a substantial spike on the AC line. Various utility fault conditions can also result in the generation of potentially damaging overvoltage transients.

For example, the occurrence of a fault somewhere in the utility company 12kV distribution system will cause a substantial increase in current in the 60 to 12kV step-down transformer at the local area distribution substation. When a fuse located near the fault opens the circuit, the excess stored energy in the magnetic lines of flux of the transformer will cause a large oscillating spike to be injected into the system. Routine load switching by the utility will have a similiar effect. These transient voltages can be quite frequent, and in some instances very damaging to equipment rectifier stacks, capacitors and transformers.

Effects of spikes on components

Transient protection is important in a modern broadcast facility because of the widespread use of high speed logic systems, sensitive analog integrated circuits and low voltage discrete semiconductors. These devices require a clean supply of power in order to perform correctly. The first line of defense in the protection of broadcast equipment from damaging transient over-voltages is the AC-to-DC power supply.

The power supply components most vulnerable to damage from an AC line spike are the rectifier diodes and filter capacitors. Diodes will occasionally fail from one large transient, but many more fail because of smaller, more frequent spikes that bit-by-bit punch through the semiconductor junction. Such occurrences explain why otherwise reliable systems fail "without apparent reason." Capacitors are vulnerable to damage due to transients because the working voltage of the device may be exceeded during the occurrence, punching a hole in the dielectric and leaving the capacitor useless at its normal operating value. The most damaging conditions result from the "right" timing of an operational change with the "right" amount of residual magnetism in the power supply transformer or DC reactor, or energy in the filter capacitor(s). These situations may well be rare in normal operation, however the possibility of such "worst case" conditions cannot be disregarded.

Thyristors (SCRs), like diodes, are subject to damage from transient overvoltages because of the possibility that the device's peak inverse voltage or instantaneous forward voltage (or current) ratings may be exceeded. Thryistors face an added problem because of transient occurrences due to the possibility of device mis-firing. A thyristor can break-over into a conduction state regardless of gate drive if either (1) too-high a positive voltage is applied between the anode and cathode, or (2) a positive anode-to-cathode voltage is applied too quickly. If the leading edge is sufficiently steep, even a small voltage pulse can turn a thyristor on. This represents a threat not only to the device, but to the load that it controlls as well.

Problems can be caused in broadcast facilities by transient overvoltages not only through device failure, but due to logic state upsets as well. Studies have shown[1] that an upset in the logic of typical digital circuitry can occur with transient energy levels as low as 1×10^{-9} Joules. Such logic state upsets can result in microcomputer latch-up, lost or incorrect data, program errors and control system shutdown.

In addition to the single-occurance logic upset, exposure of semiconductors to a high transient environment can cause a degrading of the devices, which can eventually result in total failure. Figure 4 shows the energy-versus-survival scale for several types of semiconductors. This chart clearly shows the importance of effective transient suppression.

Suppression methods

The decision on how to proceed with a transient protection program is not an easy one, but it is made somewhat less complicated by the economics involved. Most commonly-available discrete protection devices can be purchased for $5.00 to $20.00 each and installed by station personnel at critical points in the transmission plant. The alternate method is to purchase a transient suppressor unit designed for connection to the utility company primary input lines at the protected facility. The "system" approach is certainly the most effective way to prevent damaging transient voltages from entering a broadcast plant. It is also however, the most expensive way. A commercially available quality transient suppressor system will cost from $1,500 up. The price tag for a 200 Amp, three-phase service protector can easily run $7,500 or more.

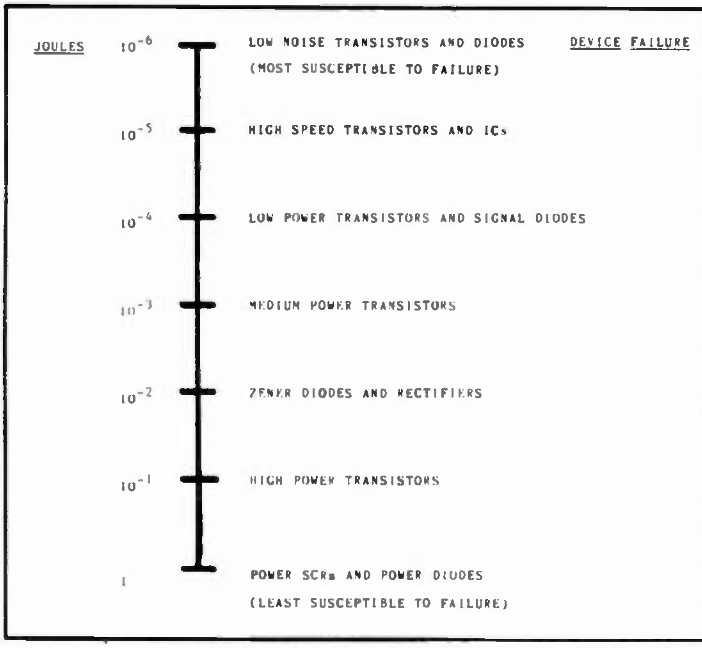
The amount of money a broadcaster is willing to spend on transient protection is generally a function of how much money is available in the engineering budget and how much the station has to lose. An expenditure of $10,000 for transient protection for a major-market station, where spot rates can run into the hundreds or thousands of dollars, is easily justifiable. At small or medium market stations, however, justification is not so easy to come by.

The systems approach

The high-performance end of transient suppression equipment available to broadcasters is occupied by motor-generated units, uninterruptable power systems and various types of power line filters.

Motor-Generator Units (MGU):

As the name implies, a motor-generator unit consists of a motor powered by the AC utility supply that is mechanically tied to a generator, which then feeds the load. Transients on the utility line will have no effect on the load



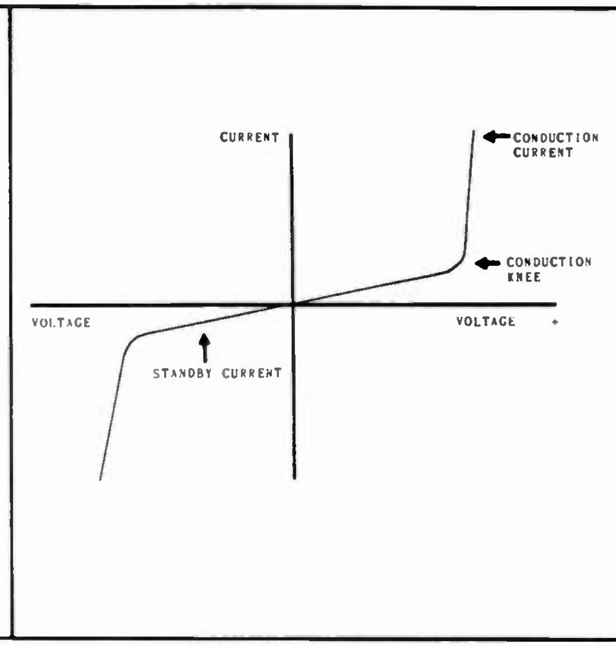


Figure 4 (on the left). An estimate of the susceptibility of semiconductor devices to failure due to transient energy. A transient duration of several microseconds is assumed.

Figure 5 (on the right). The voltage-versus-current curve for a typical varistor. The device is a non-polarized protection component that is essentially invisible to the circuit until the positive or negative voltage applied reaches or exceeds the "conduction knee." The varistor then effectively clamps the voltage excursion at a safe level.

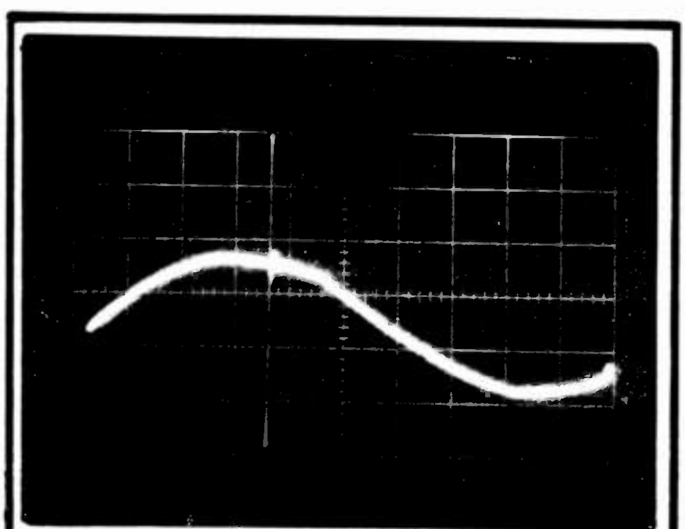


Figure 6. The transient waveform used to check the performance of AC power isolation and transient suppression methods. The spike is 800V P-P and is positioned just past the 90 degree point on the AC wave. The scope photo scale = 135V/division.

when an arrangement such as this is used. Further, the addition of a flywheel to the motor-to-generator shaft will provide protection against brief power dips (up to 1/2 second on many models).

Other features available from a motor-generator unit include output voltage and frequency regulation, ideal sine wave output, elimination of common-mode and transverse-mode noise, elimination of utility company power factor correction problems and true 120 degree phase shift for 3 phase models. The efficiency of a typical MGU ranges from 65% to 89%, depending on the size of the unit and the load.

Uninterruptable Power Systems (UPS):

Another method guaranteed to eliminate spikes on utility company power lines is the rectifier-inverter combination, used in many Uninterruptable Power Systems. AC from the utility is rectified to a given voltage, say 120 volts DC, across which sits a bank of batteries connected in series to yield slightly less than 120 volts. This DC power drives a closed-loop inverter, which provides voltage and frequency regulation. The output of the inverter is generally a sine wave, or pseudo sine wave (really a stepped square wave.) If the utility voltage should drop or disappear, current is drawn from the batteries. When AC power is restored, the batteries are re-charged. Obviously a load of any size will require a large number of very expensive batteries.

Many UPS systems incorporate a standby diesel generator that is started as soon as the utility company feed is interrupted. With this arrangement, the batteries are called upon to supply the operating current for only 10 seconds or so, until the generator gets up to speed.

Power Line Filters:

AC power processing systems are available from a number of manufacturers in a variety of configurations and designs. An example of one such processing unit is the Islatron system*, which is an active-filtering device that "tracks" the input power line voltage and then "triggers" itself into action when a deviation of more than ±2 volts from the ideal sine wave is detected.This response time is essentially instantaneous, less than 10ns. The device, because of its tracking feature, prevents ringing at the output regardless of the transient present at the input to the unit. The reference signal for this function is obtained through the use of a low pass filter that analyzes the incoming sine wave to produce a high quality 60Hz sinusoid "standard."

Discrete suppression devices

The number and sophistication of discrete spike suppression devices available to the broadcast engineer has greatly expanded within the last 10 years. Transient suppression technology has come a long way from the days of spark gaps and resistor-capacitor (RC) snubbers. The many new devices available at reasonable prices make tight control over unwanted voltage excursions possible, and allow the complicated electronic equipment being manufactured today to work as intended. Much of the credit for transient suppression work must go to the computer industry, which has been dealing with the problem for more than two decades.

*Manufactured by the Control Concepts Corporation.

A complete examination of the many discrete protection devices now on the market is beyond the scope of this paper. What follows, however, is a discussion of some of the more common types that have particular application to the broadcast industry.

The Varistor:

The MOV Varistor, made by the General Electric Company (and by other firms under different designations, but performing similar functions), is a transient suppression device available in a number of different voltage and power ratings. The component is a zinc oxide, voltage-dependent, symmetrical resistorthat performs in a manner similiar to back-to-back zener diodes. When the varistor is exposed to a high-energy voltage transient, the device impedance changes from a high standby value to a low value, thus clamping the voltage at a safe level. See Figure 5. The MOV Varistor has a fast response time, less than 50ns, and a low standby power dissipation. The device is designed to be immune to the power-follow problem common in some gas-filled spark-gap suppression devices (Gas-gaps).

The TranZorb:

The TranZorb series of spike suppression devices is manufactured by General Semiconductor Industries. The product line covers a wide range of voltage and current ratings. Available for unipolar or bipolar use, the TranZorb is a zener-type device with a fast reaction time and good clamping ratio, making it useful in protecting integrated circuits and other voltage-sensitive devices. Included in the General Semiconductor TranZorb line is a series of very low voltage units (available in standoff values of .8 to 6.4V) and a group of transient suppressors packaged in DIP enclosures for easy mounting on printed circuit boards.

The Klipvolt:

The Klipvolt transient suppression device, manufactured by ST Semicon, Inc., has (like the other protection components already discussed) two basic modes of operation: a high-resistance standby value and a bipolar (or unipolar, if required) low-resistance conduction value. The Klipvolt is manufactured by a pressed-powder selenium process that produces a selenium crystalline structure in a random arrangement on the cell plane. This construction allows the cell to dissipate considerable power under active operation.

Performance testing

The development of discrete protection devices such as those previously mentioned has made it possible to design transient suppression capabilities into virtually any piece of broadcast equipment. Methods used in years past, including RC snubbers, common isolation transformers and RFI filter assemblies, did little to eliminate the threat posed by AC line spikes. Figures 7(a)-(d) show graphically the improvement that the new generation of transient suppressors has made over previous methods.

In preparation for this report, four devices were checked for performance using a TS-III transient waveform generator test set, provided courtesy of the Control Concepts Corporation. The TS-III generates a waveform shown in Figure 6. It is approximately 800V P-P and 30us wide. The spike is synchronized to the AC line, allowing easy observation on an oscilloscope. The TS-III pulse is designed to be similar in nature to the transient typically generated by SCR controllers, contractor load switching and large motor mode changes.

Figure 7(a) shows the 800V P-P test spike applied to a common isolation transformer. As can be seen, the output transient is greater than the input transient! One of the new high-performance isolation transformers specifically designed to remove noise and spikes from the AC line would certainly perform better than the unit used in this test (a general purpose device not intended for transient or noise suppression applications). On the basis of this measurement, however, engineers should be cautioned against using just any transformer for isolation if transient suppression of some sort is not provided on the primary and secondary of the device.

Figure 7(b) shows the TS-III spike applied to a general-purpose RFI filter assembly. As the scope photo shows, there is almost no attentuation of the pulse. The RFI filter may have performed better if the input spike rise time had been different, but since the transient generator test set gives only a single output rise time, this was not examined.

Low-value capacitors are sometimes used for transient protection in broadcast equipment, generally across the AC input terminals. Figure 7(c) shows the attentuation gained when a .25uF capacitor is placed across the 800V P-P spike. The cap reduces the transient by half, centered on the sine wave. This is contrasted with the MOV varistor, which clips on maximum amplitude, positive and negative.

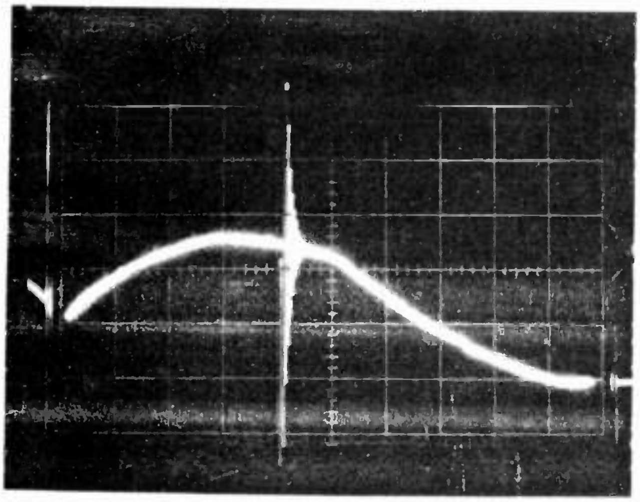
An examination of figure 7(c) also shows a brief period of oscillation following the transient. This is caused by the interaction of the capacitor and the inductance in the circuit. Placing a series resistance of 100 ohms reduces the oscillation to a negligible amount, however it also significantly reduces the effectiveness of the capacitor in snubbing the transient.

The most promising of the four discrete suppression devices used in this series of tests was a General Electric MOV varistor. Figure 7(d) shows the performance of the varistor under the application of the 800V P-P transient. The device clips the level at about 200% of the applied steady-state voltage. While this level of protection may not be sufficient to prevent disallowed logic states in microcomputer equipment or overloads in sensitive systems, it would probably prevent damage from occurring to the equipment on-line. This conclusion assumes that most power supplies used in broadcast equipment are designed for long term operation at about 50% of their maximum capability.

While the MOV varistor performs well by many standards, note the undershoot that occurs in the waveform. Varistors (and the Transzorb and Klipvolt devices already discussed) clip on maximum amplitude, without regard to the position of the spike on the sinusoid.

Circuit-level applications

Many low-voltage power supplies used in broadcast equipment are, at best, adequate. All too often power an expensive piece of equipment is derived from



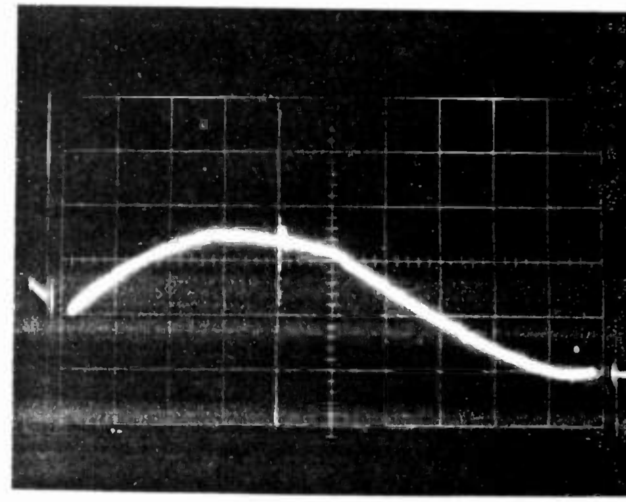
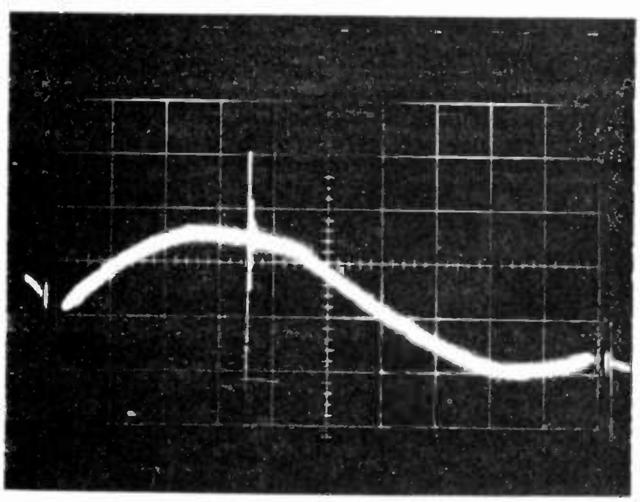


Figure 7 (a) (on the left). The effect of the test transient on a common isolation transformer. Note that the ringing caused by the spike exceeds the original disturbance.

Figure 7 (b) (on the right). The performance of a commonly-used RFI filter assembly when subjected to the 800V P-P test spike. Note there is very little attenuation of the transient.



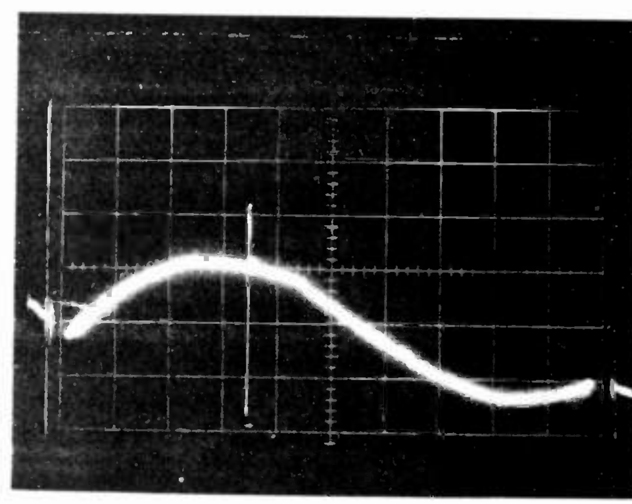


Figure 7 (c) (on the left). The transient suppression performance of a .25uF capacitor. Although the spike is substantially reduced, a small amount of oscillation can be seen in the display. The addition of a 100-ohm series resistor will eliminate the oscillation, however it also reduces the transient suppression effectiveness of the capacitor.

Figure 7 (d) (on the right). The transient suppression performance of a GE MOV Varistor device. Note the good clamping action and freedom from oscillation or ringing provided by the varistor. Of the discrete devices checked with the 800V P-P test signal, the varistor showed the greatest promise.

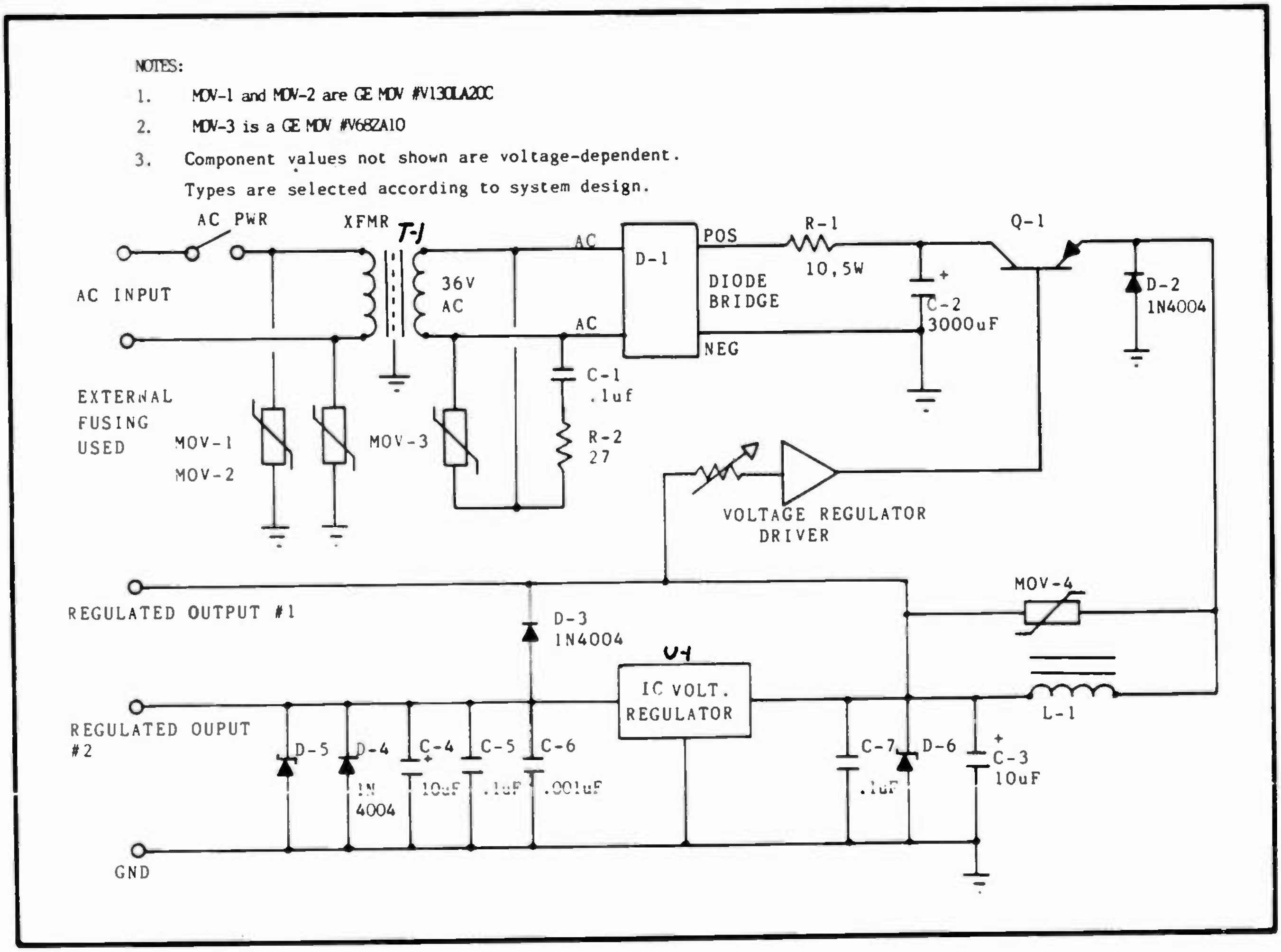


Figure 8. The recommended transient overvoltage protection for a low voltage power supply. A circuit such as this will have good survivability despite frequent transient disturbances. The protection methods shown will also reduce the possibility of transients appearing on the DC power supply rails that might cause logic state upsets.

a circuit that has virtually no transient overvoltage protection. While this type of supply will certainly work, it falls far short of the "state of the art" and is much less than we should be able to expect from professional equipment.

Figure 8 shows transient protection methods for a typical low-voltage series-regulated power supply. MOV-1 and 2 will clip spikes on the incoming AC line, and the combination of MOV-3 and C-1 will clip turn-on, turn-off, and fault spikes on the secondary of T-1. Resistor R-1 protects diode bridge D-1 by limiting the amount of current through D-1 during turn-on, when capacitor C-2 (the main filter) is fully discharged. MOV-4 protects series regulator Q-1 and the load from damage due to transients generated by fault conditions and load switching. The varistor is chosen so that it will conduct current when the voltage across L-1 is greater than would be encountered during normal operation. Diode D-2 protects Q-1 from back EMF kicks from L-1.

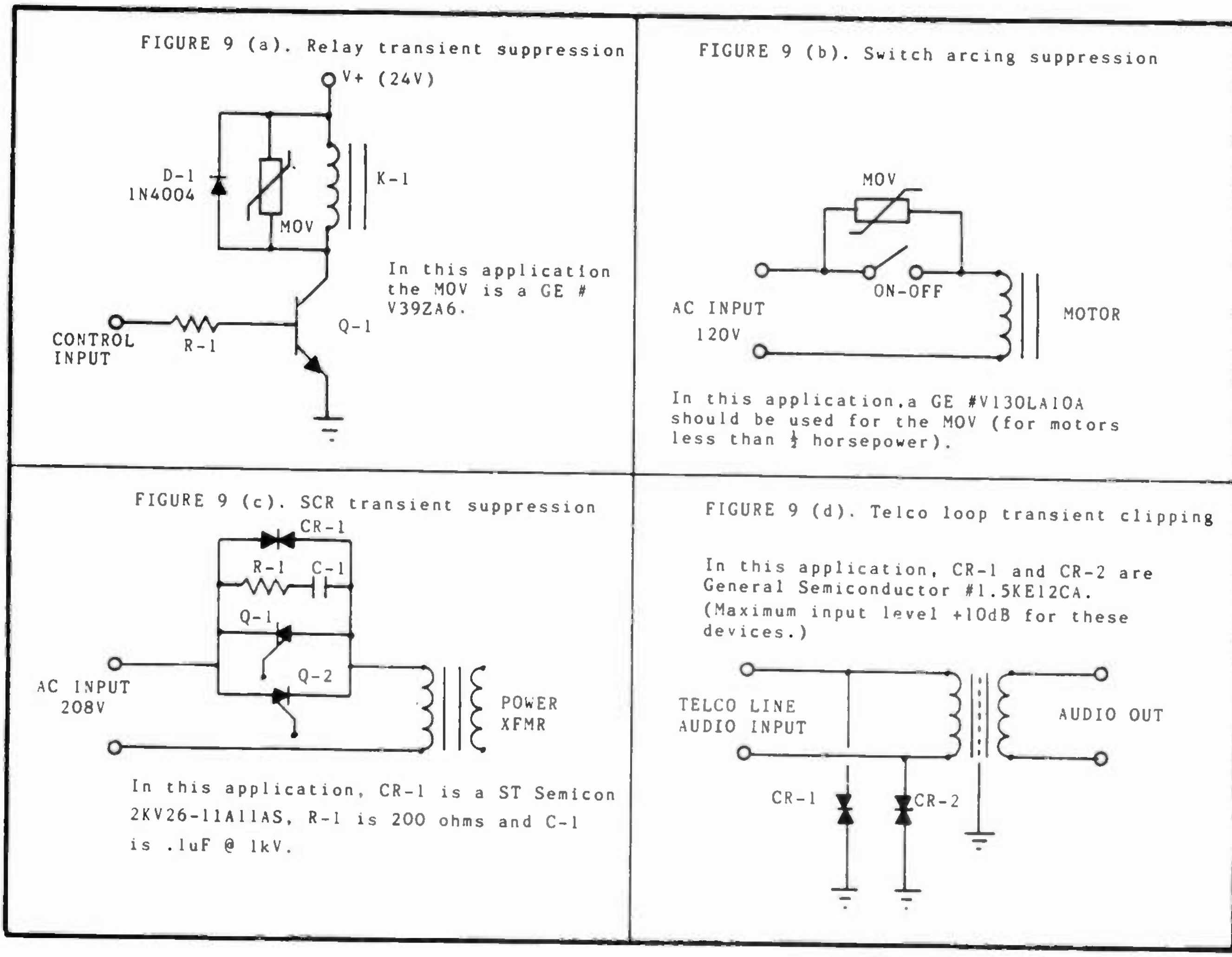


Figure 9. Transient suppressor applications for several common circuits. Diagram (a) shows back-EMF transient suppression for a relay switched by a transistor. Diagram (b) shows an application of a varistor for suppression of switch contact arcing. Diagram (c) illustrates a common application of a transient suppressor to protect a pair of SCRs. Diagram (d) shows a way to protect a telephone company audio or data loop from transients that can often appear on local- or long-lines.

Three-terminal integrated circuit voltage regulator U-1 is protected against excessive back-current due to a short circuit on its input side by diode D-3. Capacitors C-3, 4, 5, 6 and 7 provide filtering and protect against RF pickup on the supply lines. Diode D-4 protects U-1 against back-EMF kicks from an inductive load, and zener diode D-5 protects the load from excessive voltage should U-1 fail (possibly impressing the full input voltage onto the load.) D-5 also protects U-1 from overvoltages due to spikes generated by inductive load switching or fault conditions. D-6 performs a similar function for the input side of U-1.

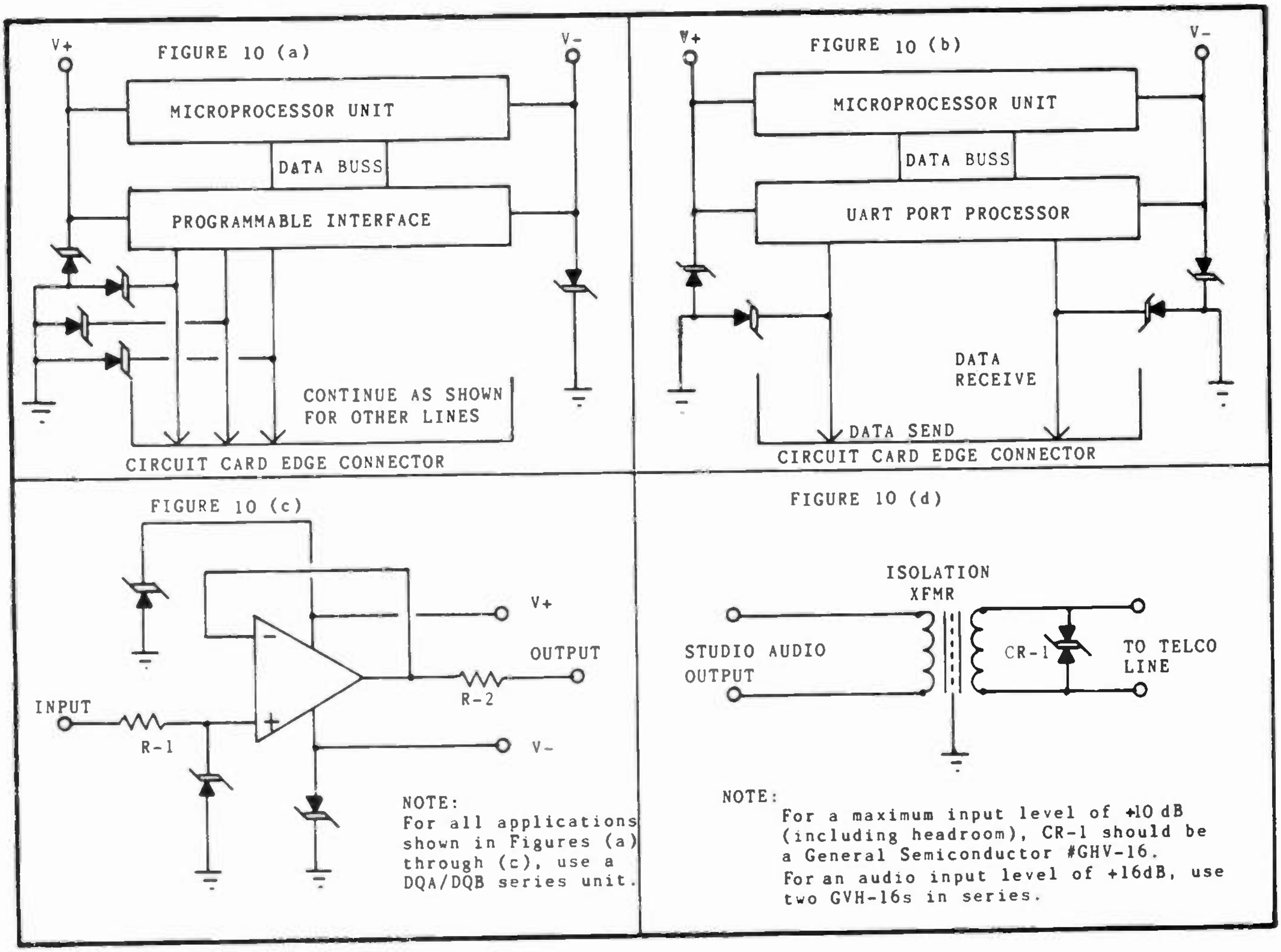


Figure 10. Applications of transient suppression devices to microcomputer, analog voltage sampling and audio circuits. Diagrams (a) and (b) show protection methods for common computer circuits. Diagram (c) illustrates protection methods that should be used on analog voltage sampling circuits. The devices used in these applications are General Semiconductor #DQA/DQB series DIP units. The packaging arrangement used in the DQA/DQB line makes these components ideal for circuit-board-level transient protection. Diagram (d) shows an application of transient suppression on a high-quality audio telephone company loop. The device specified is a low capacitance unit that will not degrade the high frequency audio performance of the Telco line.

Some additional applications for spike suppressors are shown in Figure 9. Any transistor that switches an inductive load should have transient protection, as should switches controlling any appreciable amount of power. The use of a spike suppressor across a switch or relay contact will greatly extend the life of the switching element. SCR control of the power input to a large transformer is common in transmitter equipment today, and some form of protection is vital to long-term reliability. The surge protector shown will clip spikes generated by the transformer during retarded-phase operation. Transient suppression is often desirable on telephone company audio or data loops. Spike clipping devices are selected based upon the normal audio voltage levels (including headroom) used on the line.

For maximum protection of microcomputer equipment, transient suppression should be designed into individual circuit boards. Figures 10 (a) and (b) show typical applications of on-the-board spike suppression. The devices used are General Semiconductor DQA/DQB series DIP TranZorbs. Four individual TranZorb devices are included in each DIP package, making convenient placement on crowded printed circuit boards possible. Figure 10 (c) shows an application of transient suppression in a voltage-follower circuit, common in many analog data acquisition systems. Note the use of suppression devices at the power supply pins of the circuits shown in (a) through (c). Figure 10 (d) illustrates an alternative to the transient protection arrangement shown in Figure 9 (d). The Figure 10 circuit will avoid the possibility of introducing noise into high-quality audio program lines due to common-mode imbalances that may result from transient suppressors being tied to ground. Further, the use of a low capacitance suppressor will ensure minimum capacitive loading on the telco circuit.

In conclusion

The modification of a piece of equipment to prevent the generation of transients, or to prevent damage from externally-generated spikes, should be weighed carefully. Transient suppression devices must be selected to allow for operation under all anticipated conditions. Further, the manufacturer of the equipment should be consulted before changes of any type are considered. An incorrectly-placed transient suppressor can end-up causing many more problems than it solves.

The sophisticated equipment used today at radio and television stations must have adequate protection against transient disturbances if long-term reliablity is to be achieved. The days when the AC power could be assumed to be clean are gone, and broadcast engineers must be prepared for damaging spikes that will occur. In view of the tremendous investments that broadcasters have made in high technology equipment, the cost of transient protection is very small. It is always cheaper to prevent a problem than it is to solve one.

References:

1. "Voltage Transients and the Semiconductor" by H. Niles Gloer, The Electronic Field Engineer, Volume 2, Number 2.

2. Noise Suppression Reference Manual, Topaz Electronics Company.

3. Computer Power Products engineering data, Computer Power Products, Inc.

4. General Electric MOV Varistor Data and Applications Manual.

5. Lightning Elimination Associates Technical Reports #LEA-9-1, LEA-0-10, LEA-0-8.

6. EMP Protection for AM Radio Stations, Defense Civil Preparedness Agency, TR-61-C, May, 1972.

7. EMP and Electric Power Systems, Defense Civil Preparedness Agency, TR-61-D, July, 1973.

8. EMP Protection for Emergency Operating Centers, Defense Civil Preparedness Agency.

Stereo Audio Production for TV

James F. Gibbings

National Broadcasting Company

New York, New York

At long last it really looks as if television broadcasting in stereo or alternately the provision of a second audio program for example, shows released in English and Spanish versions is almost here. The unanimous vote by the EIA-MTS to adopt a proposed standard for transmission in the United States, using the Zenith transmission system and the dbx companding or noise reduction system, clears the way for the FCC to finalize its Report and Order under Docket Number 21323.

In preparation for this event NBC along with the other networks has been working diligently to get ready for the advent of Multi-Channel Sound (MCS). At NBC we have set up a Multi- Channel Sound Listening Room in New York to enable us to experiment with different sound formats and to demonstrate the potential of MCS to our management, producers, and to outside interested partners, such as Advertising Agencies.

It should be realized that there is a significant difference between FM Stereo listening and the Stereo TV experience. In the case of FM listening, music is usually dominant. However, in the case of TV, dialogue is usually dominant. Because of this, coupled with the usual small size of the TV display it is important that dialogue appears to come from the center of the TV stage. In general Stereo TV will use monophonic dialogue with music and effects in stereo for the foreseeable future.

Television production using the proposed compatible stereo standard presents the producer, director and sound mixer with several interesting challenges. First, there is the fact that the sound stage is naturally wider than the picture stage for the majority of viewers. This not only deprives the artist from effectively exploiting on-camera sound panning but complicates the mix. This is not a great loss as proven by the motion picture industry. It seldoms relies on sound panning on stage to any great extent. More on mixing guidelines later.

Secondly, the production staff must be ever conscious of delivering two very important sound products simultaneously: that of discrete stereo (left and right channels) and the compatible monophonic (left plus equal right), the latter being that experienced most by viewers for some time to come.

A third challenge is that a considerable amount of mixing for production will be undertaken only one time with little or no opportunity to remix and process or "sweeten" before airing. This will require special techniques such as a menus of presets and possibly audio follow video switching. Certainly television mixing will come to be a unique art and a special satisfaction to those who master it.

Let's take a look at some of the troublesome parts of the first challenge and suggest some ways of coping.

Because of the NTSC Standards were built on the premise that the viewer would be at least the distance of 5 picture heights from the screen, the visual field of view is rather narrow (less that 16 degrees of angle). Recognized guidelines for listening to stereophonic sound suggest that for best results the optimum speaker placement is 8 feet at a listening distance of 8 feet or about 60 degrees of angle. Of course, the viewing environment limits the number of viewers that can be centrally located for stereo listening and suggest a narrower angle to the speakers to reduce the effect of off-picture dialogue for the viewer who is seated well off center. Truly, the reduction of off-picture dialogue and the optimum speaaker separation for the center listener is a compromise in a simple left/right delivery; that compromise dictated by the layout of viewers. TV Receiver Manufacturers who build in stereo speakers must make several assumptions about the compromise required. Those who supply detached speakers or otherwise adjustable channel separation should supply a diagram or written guidance to their customers to lead them to the optimum configuration in the viewing setting.

One very potent mixing technique can be employed to minimize off screen dialogue.

It is well known in professional audio circles in a two channel stereo format that the width of the stereo stage can be adjusted within limits by the technique of mixing a small amount of "L" with "R" and an equal amount of "R" with "L".

Mixing minus L with R and minus R with L expands the stage. With care in quantity and phasing the material, an effect of as much as 25%

increase in stage width can be achieved without creating a major void at center stage.

Mixing plus L with R and plus R with L narrows the stage. When the quantities become equal the obvious result in monophonic, placing the stereo focus as close to center as practical.

Choosing the negative values for the cross feed, for ambience material and off screen effects, and positive values for all on-screen effects and dialogue, helps to minimize the dicotomy and reduces the distraction created by the difference in stages. Continuously varying the mix allows continuity of the sound transitions on to off stage.

In the area of Television Production Multi-Channel sound there are many options to be considered.

Shows that are produced in motion picture format offer a number of different quality options for consideration. In order to evaluate quality options and to evaluate the ultimate benefits to be gained from the use of a third channel or center speaker, NBC commissioned Universal to prepare some excerpts from NBC shows with sound tracks of different quality and in both two and three track formats.

We have selected a number of sequences with different levels of quality and both two and three track versions, where appropriate, and have transferred this material to 1" type C video tape using channel one for the left speaker feed, channel two for the right speaker feed and channel three, which as you know has a slightly lower signal to noise ratio, for both the monophonic material and the center speaker for three track excerpts.

We have been conducting listening tests of all this material. Before we go further let's define the different quality levels.

Monophonic, of course, is what we've got now. It is usually limited in frequency range to approximately 200Hz to 8Khz because of the use of the academy roll-off to limit noise on optical sound tracks.

Stereo album format uses elaborate multi-mixing and mixdown of the music, and true stereo ambient sound. Sound effects which correlate with visual action (impacts, moving cars, etc.) would be monophonic originals spread or panned across the audio field. Dialogue would be monophonic and almost always located center stage. Frequency range is approximately 200 hz to 12Khz.

Stereo motion picture format differs from album format in using a considerably less elaborate stereo music recording and with ambient sound effects derived from a monophonic source and spread over the stereo stage. Action sound effects and dialogue are treated similarly to those in the album format. Frequency range is approximately the same as "album" quality.

Synthesized stereo creates an illusory stereo spread for music or ambient sound effects by passing a monophonic source through a phasing device. Again, action effects and dialogue treatment could be similar to album and motion picture formats. However, if synthesized stereo is used simply to inexpensively enhance older shows where separate sound effects are no longer available, then all sound effects would pass through the synthesizer and panning and precise location of action effects would not be possible.

There are two ways to achieve the center dialogue sound. One is an appropriate in phase mix of dialogue in the left and right channels to achieve a phantom center dialogue effect as discussed above.

For a listener on axis between the two speakers, the sound appears in the center as if coming from a phantom speaker located in the center. As the listener moves off axis, however, there is an erratic shifting of the localization of the apparent source due to phasing effects between the left and right speakers. Ideally, the dialog and other center sound would be recorded, transmitted, and reproduced with a discrete third track and center speaker. The center speaker can also be used for monophonic sound.

In dialogue, dominant scenes, we have found the third center channel to be very effective. Some sound effects are mixed into the center channel but most effects and music are assigned to the left and right channels.

The Zenith transmission system will permit the use of a third audio channel in quadrature with the difference signal (L-R).

Now let's look at shows that are originally taped in video format. As an example the Johnny Carson "Tonight" show has been recorded in stereo format since October 1982. Of course to date it has only been broadcast in monophonic sound using the L+R signal recorded on Channel 1 of our 1" tape equipment. This is almost a live show in that recording and broadcast follows within about two hours time. This means there is no time for post production audio sweetening. Since the show is virtually live "bundled" sound techniques must be used.

By this we mean that there is no complete separation between cast, effects and music microphones so there will always be some "leakage" from one mike to another. This is in contrast to motion picture sound where the audio mixing is all done in post production, with generally speaking, isolated dialogue/lyrics, sound effects and music.

At this time all the audio boards at NBC Burbank were designed and built 15 years ago. While that was a long time ago, a lot of thinking went into the boards. There is also a large custom-made jack panel accompanying each board that has over a thousand holes. You can swap modules, mult signals, jump modules, move modules, or do whatever you need to do with this jack panel. The flexibility is great for TV production, where you have to send six or seven different kinds of feeds to the floor or to the PA system, to the foldback system, people monitoring headphones and so forth. It's quite a lot of things to get around.

Each of the NBC-built consoles at the Burbank complex has 53 simultaneous inputs. These inputs are assignable, through a system of submixers, to 20 vertical faders. Ten of the vertical faders are fed directly from the inputs. Each of the remaining 10 faders has four submix inputs associated with it. Volume level on each submix input is controllable by a rotary pot. The four rotary submix pots for each of the 10 verticle faders are located directly above that fader.

The console has one equalizer per vertical fader, which of course means there are no separate EQ facilities for any of the individual submix channels. Equalization, is one of the present console's weak areas.

At the time the board was built, it was state-of-the-art. That's all there was for equalization in 1964. The EQ facilities that do exist, however, allow the user to select crossover points. Each equalizer boosts 12 dB and cuts 16 dB at 40 and 100 Hz on the low end, and at 3,5,10 and 15 kHz.

The 20 vertical faders on the console are assignable to six rotary submasters, which, in turn, are assignable to two masters: the Cast and Music masters. The two masters are assignable to the board master, which is the final output for everything on the board except for two auxiliary faders, having one input each, and the Nemo fader.

The Nemo fader, which takes its name from an old telephone company term, and stands for "Not Emanating from Main Office" - is for controlling the volume levels of all sends that originate outside of the studio, which in this case means film, videotape, and remote lines. There are six inputs to the Nemo fader, which can be controlled manually or remotely by the video switcher.

Summing up the consoles signal flows, it comes down like a giant tree and is essentially a 53 x 20 x 6 x 2 x 1 board.

To operate the board in stereo, three of the board's six submsters are used to group all of the Cast inputs into a left, center and right configuration; i.e. one submaster for left, another for center, and a third for right. The remaining three submasters are used to group all of the music inputs into a similar left, center and right configuration. All of the left channel information - both dialog and music - is then sent to the left output channel (Cast master in conventional operation), and all right information is sent to the right output channel (Music master in conventional operation). All center information is sent to both output channels.

Since the prsent NBC console is monaural, there are no pan pots, the system essentially creates 'hard panning' stereo. Several methods have been devised for reducing the "hard left/hard right" sound characteristics inherent in the system. For example, you can flip the reverb channels, sending the reverb for left-channel instruments to the right channel, and vice versa.

Something like panpots can be derived by sending a signal to two channels, and playing with the relative volume levels of each channel. Also, on occasions where the "Tonight" show hosts a large musical group, an auxiliary mixer is used to accommodate the resulting extra inputs to the board. Because the auxiliary mixer - generally a Yamaha 8x2 or 12x2 - is a stereo board, use can be made of the panpots it has.

From the main audio board, the left and right signals are sent into a custom, NBC-built matrix unit, which converts them into two new signals: a sum (L+R) channel and a difference (L-R) channel. The mono sum channel is recorded on track of the videotape. This is the signal that is currently transmitted over the network to the affiliates and, in turn, to the viewers' homes.

The difference (L-R) signal is recorded on track #2 of the videotape. This signal, is anything that is strictly left or strictly right. The center has disappeared because it has been cancelled. The difference signal presently remains in-house for future use. With the advent of Stereo TV, NBC will be able to transmit both the sum and difference channels.

NBC is presently in the process of installing new state-of-the-art consoles in our Burbank studios.

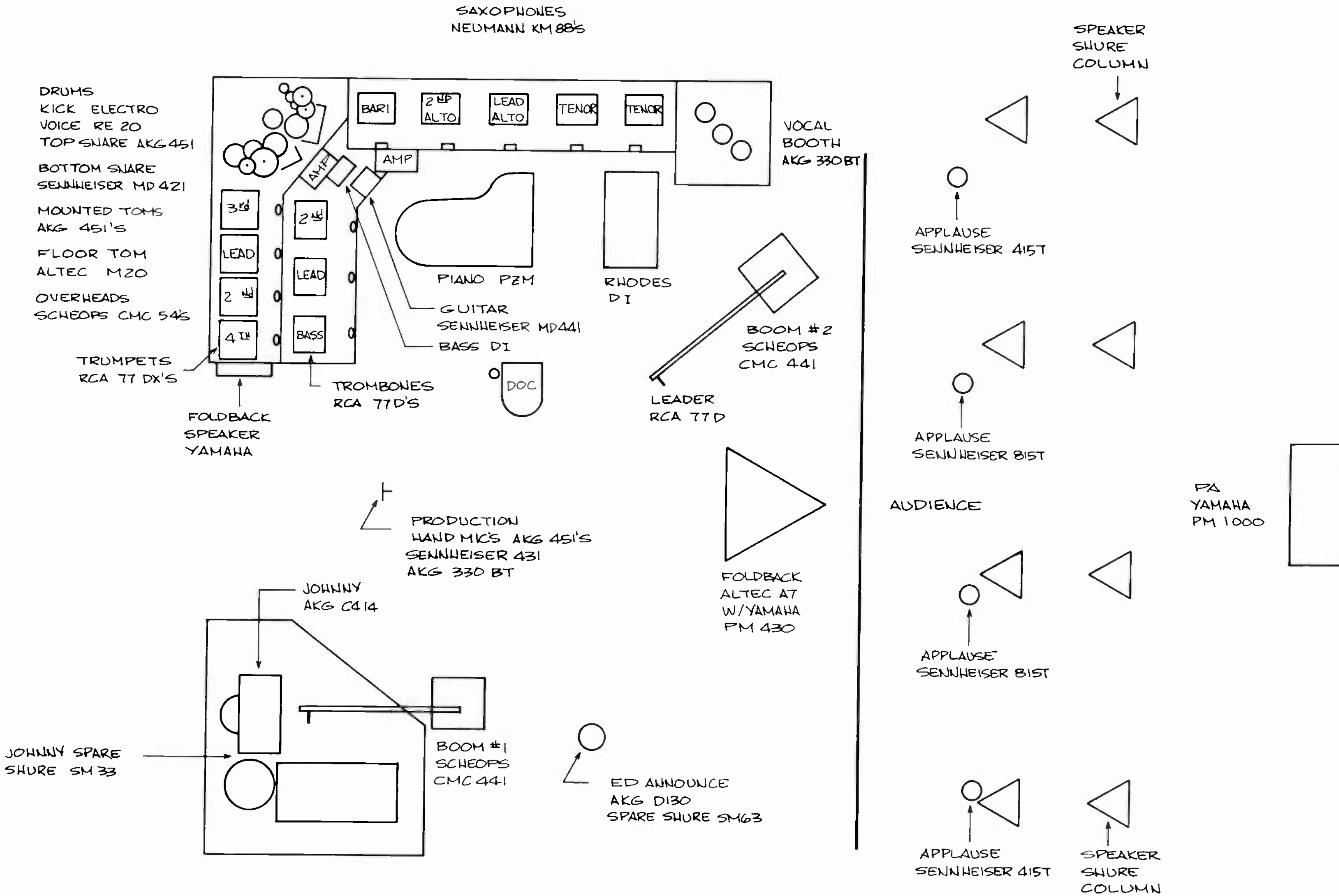
This is a diagram of a typical audio set up on the "Tonight" show. Microphones used on the "Tonight" show currently include RCA 77 DX's on the brass, Neumann KM 88's on the reeds, and Schoeps AKG and Electro-Voice mikes on the drums. Johnny Carson uses an AKG-C451 and Sennheiser mikes for audience applause. In addition a Lexicon 224 is used for reverberation on the band. Additional equipment includes an EXR Exciter, UREI LA-3 limiters and Valley People Gain Brain II Compressor-limiters.

It is our belief that NBC's production of Sports Events, Musical Shows, such as the "Tonight" show, will only add a very small cost increment to present production costs. Depending of course on the Producer's requirements.

In conclusion, it appears that NBC or in- house production and feature movies produced in stereo format will be introduced first. These will be closely followed by packaged sitcoms and then series shows produced in motion picture format.

We would like to thank MCA/Universal and the Johnny Carson show for permission to use our demonstration excerpts and Dick Stumpf, Universal, Director Sound and Electronics Department.

I would also like to thank Terry Byrne of NBC Corporate Planning who is spearheading our Multi-Channel sound program, my associate Don Musson, Director of Technical Development for his guidance and support and last but not least, Ron Estes, the sound mixer on the Tonight show whose original and continuing stereo accomplishments did much to motivate NBC's Multi-Channel sound program.


SAXOPHONES
NEUMANN KM 88'S
BARI
2 ND ALTO
LEAD ALTO
TENOR
TENOR
AMP
AMP
VOCAL
BOOTH
AKG 330 BT
SPEAKER
SHURE
COLUMN
DRUMS
KICK ELECTRO
VOICE RE 20
TOP SNARE AKG 451
BOTTOM SNARE
SENNHEISER MD 421
MOUNTED TOMS
AKG 451'S
FLOOR TOM
ALTEC M20
OVERHEADS
SCHEOPS CMC 54'S
3rd
LEAD
2 nd
4 TH
2 nd
LEAD
BASS
PIANO PZM
RHODES
DI
GUITAR
SENNHEISER MD441
BASS DI
BOOM #2
SCHEOPS
CMC 441
APPLAUSE
SENNHEISER 415T
TRUMPETS
RCA 77 DX'S
TROMBONES
RCA 77D'S
DOC
LEADER
RCA 77D
APPLAUSE
SENNHEISER 815T
FOLDBACK
SPEAKER
YAMAHA
PRODUCTION
HAND MIC'S AKG 451'S
SENNHEISER 431
AKG 330 BT
AUDIENCE
PA
YAMAHA
PM 1000
FOLDBACK
ALTEC A7
W/YAMAHA
PM 430
JOHNNY
AKG C414
APPLAUSE
SENNHEISER 815T
JOHNNY SPARE
SHURE SM 33
BOOM #1
SCHEOPS
CMC 441
ED ANNOUNCE
AKG D130
SPARE SHURE SM63
APPLAUSE
SENNHEISER 415T
SPEAKER
SHURE
COLUMN

Phase Considerations In Stereo TV Production, Post Production And Transmission

Mark Schubin

Technological Consultant

New York, New York

There are presently well over 200 million television sets in the United States. Nearly 20 million were sold in 1983 alone. Even if it is assumed that more than half of all television sets sold in the United States will be equipped for stereo sound once stereo TV broadcasting begins (a questionable assumption), it could take at least 10 years before the majority of the television audience listens in stereo. Therefore, it is crucial that monaural television sound retain its present quality during the transition to an all-stereo listening audience.

The problem of maintaining monaural sound quality is concentrated largely around the issue of relative phase between stereo channels. The Electronic Industries Association recently completed extensive tests of a wide variety of technical parameters in the evaluation of different stereo coding and noise reduction schemes to be used in television transmitters and receivers. However, little attention has been paid to other aspects of the television sound chain, including microphones, mixers, processing equipment, distribution equipment, monitoring equipment, and even cables and connectors. Not only technology, but also technique has been ignored in these areas.

Many forms of stereo production techniques are quite incompatible with monaural sound. One of the simplest stereo microphone techniques, for example, is the use of two omnidirectional or slightly directional microphones, spaced widely apart, over, or in front of, a group of performers. Such a technique creates a tremendous diversity of sound emanating from the listener's speakers, an effect often useful for selling stereo equipment, however enjoyable it may prove to be for long term listening. Unfortunately, it is virtually impossible to create a natural monaural sound by combining the signals from those microphones. One problem is acoustic phase cancellation at different frequencies (leading to a very uneven effective monaural frequency response, varying with performer placement).

Close microphone techniques are less susceptible to the problem of acoustic phase cancellation, since sounds not intended to be picked up by a particular microhone will tend to be tremendously lower in level than intended sounds. With such microphone techniques stereo imaging is almost entirely at the disposal of the audio mixer, who can electronically position any microphone's signal at any point between the stereo speakers.

Ideally, it should be possible simply to add the signals of the left and right channels to create a monaural channel matching the intent of the audio mixer. Unfortunately, in the real world, such recombination generally leads to the phenomenon known as "center channel buildup," where sounds positioned between the speakers will appear to be significantly louder to monaural listeners than they will to stereo listeners.

A number of techniques have been developed to reduce or eliminate center channel buildup, when stereo and monaural signals can be dealt with separately. These range from elaborate electronic circuits designed to shift the phase of the left and right channels by 90 degrees in opposite directions, to mixing consoles designed to create left, center, and right mixing channels that can be recombined into left, right, and monaural output channels in such a way as to reduce the center channel buildup.

Some audio mixers prefer to assign different proportions of various microphones to the monaural output channel through the use of an auxiliary, post-fader mixing channel, while still others prefer to use two completely separate mixing consoles (with different people in charge) to create separate stereo and monaural mixes, either by splitting microphone signals or even by using two sets of microphones. Different equalization and level compression techniques can be applied to these different mixes on the assumption that the stereo one will be heard through a high fidelity sound system, while the monaural one will be heard through a television set speaker.

Unfortunately, multichannel television sound offers none of the luxuries of separate stereo and monaural mixes. The left and right channels will always be recombined in monaural TV sets, and sound will be heard sometimes from high quality amplifiers and speakers and sometimes from the classic television receiver sound system that's been in use for years.

While there is no optimum solution to the problems of dynamic range compression and equalization for these two sound systems, there are excellent microphone techniques that have been developed for the absolute elimination of phase problems. Perhaps the best of these is the M-S or mid-side system (sometimes called the mono-stereo system), utilizing a single microphone positioned for optimum monaural sound, in conjunction with a figure-eight pattern microphone, coincidently placed, to pick up stereo difference information. Traditionally used for an ambient pickup of a musical group, there is no reason that the M-S technique cannot be applied to multiple microphones, or even to an announcer's microphone.

Whatever microphones are used, it is critical that they be wired in a common phase configuration. Unfortunately, not all microphones are wired in the same way. In fact, some 12-volt condenser microphones require different polarities of powering so that, even if the microphones are wired to a common standard in terms of signal phase, it is likely that an operator may use a phase reversing cable or adaptor to get appropriate power polarity to the microphone.

It is a simple matter to test microphone phase through the use of an acoustic impulse generator/detector set. In addition to making a popping sound, such generators are usually equipped with electrical signal generators, and the detectors, in addition to housing microphones, are usually equipped with electrical detectors. Thus, through the use of the four possible connections (acoustic-acoustic, acoustic-electric, electric-acoustic, and electric-electric), the phase of everything from microphones to speakers may be checked, omitting nothing in between.

The importance of this testing for (and correction of) phase in the connection chain cannot be overstressed. It is common practice in the United States to wire pin 1 of the three pin audio connector, and the sleeve of a tip-ring-sleeve connector as shield or ground. It is much less common that the remaining two conductors are wired equivalently throughout this country, or even in a single facility.

Audio amplifiers, mixing consoles, and processing equipment often introduce a signal phase reversal, sometimes offering differing phase at different outputs. For example, one popular mixing console offers identical phase at all inputs and outputs, with the exception of a single group of unbalanced outputs. Since other unbalanced outputs on the same console offer correct phase, the use of an identical balancing system on these two outputs will result in opposite phasing, even though the balancing system is correctly wired.

Out-of-phase conditions in a monaural broadcasting facility are generally innocuous. If a total phase cancellation exists, it will be immediately detected, since no audio will be heard. A less than total cancellation that goes undetected will be compensated for with automatic or manual level controls. An out-of-phase condition that does not cause cancellation will pass by totally undetected, though some listeners claim to be able to detect improper acoustic phase even when only one channel is heard. Thus, a television station about to transmit stereo might be riddled with out-of-phase conditions and never know it.

The only way to ensure that a facility is correctly phased is to test each and every input, output, and connection with something like impulse generator/detector pairs. A television station planning to transmit stereo without such equipment is simply asking for trouble, especially if monitoring practices call for listening to stereo rather than monaural sound. In the worst case, of course, listeners to the monaural signal could end up hearing absolutely nothing. A much more insidious problem could be the complete loss of the announcer in the monaural channel, while music, filled with phase differences, passes through apparently unharmed.

Even when all microphones, connections, cables, and amplifiers are correctly phased, it is possible to create an out of phase condition through accidental or intentional adjustments of the mixing console, or by playing a tape or receiving a transmission created incorrectly outside the facility. For that reason, it is important to continuously monitor the phase of the console output and/or transmitter input. The simplest phase indicator is the X-Y oscilloscope display, which may be immediately created with a maintenance oscilloscope, or even by adding some connections to the common vector display driven from the R-Y and B-Y outputs of an NTSC decoder. There are also commercial versions of this display available, specifically designed for stereo phase monitoring.

On such a display, an oblique line from bottom left to top right indicates correct monaural phase and an oblique line from bottom right to top left indicates incorrect monaural phase. The deviation of the trace from the line indicates stereo separation. A display filled with traces indicates vast separation; an almost linear display indicates low separation. A tendency towards the horizontal or vertical indicates a higher level left or right channel, depending on how the display is wired.

While a great deal of information (including an indication of clipping distortion) can be gleaned from such a display, it may be too complex for continuous viewing. Other phase indicators are available using meter movements or colored lamps. One such lamp display is small enough to fit into the space occupied by a single fader module on a mixing console.

Signal processing in stereo is trickier than for monaural sound. The simple act of peak limiting, for example, can suddenly shift a stereo image from the center to one side or the other, unless the limiters used on both channels are tied together in such a way that gain reduction in one channel will cause simultaneous and identical gain reduction in the opposite channel.

The sudden shifting of an acoustic image to match camera cuts has generally been discouraged in stereo simulcasting of television programming so as not to upset FM listeners who are not also viewers. It is probably impossible for a human mixer to follow camera cutting adequately in a live show or continuous recording. Therefore, as a good rule of thumb, dialogue and solo performer microphones should be kept in the center of a stereo image, while music and effects can be widely separated. Nevertheless, some stereo motion pictures have made extensive use of the positioning of dialogue to match visual cuts. Naturally, such positioning requires post production time.

The recording of a stereo television show calls into question more phase considerations. It is common recording studio practice to adjust the positioning of the recording and playback heads of audio tape recorders prior to each recording or mixing session to ensure that phase cancellation will not occur between different tracks. The same phase indicators used to monitor console output may be used in the head alignment process, though greater phase resolution is desirable.

Unfortunately, though Type C videotape recorders have offered multiple audio tracks since their inception in 1977, few such machines offer the possibility of such head adjustment. For that and other reasons, stereo television shows have often been recorded on separate audio and video tape recorders, the two being locked together for synchronous playback by the SMPTE time code.

Developed specifically for video tape editing, SMPTE time code has spread throughout the audio production field. One of the key specifications of the time code is that it use 80 digital bits per frame. However, in a typical audio-only facility, there is no such thing as a frame. Therefore, when using an audio-only facility for television audio post production, it is important to be sure that any SMPTE time code generated at the facility be locked to some source of video frame information, such as a video sync generator. Power line lock is not sufficient to prevent extensive speed variations when the video and audio tapes are synchronized.

Whether television sound recordings are made on one or two recorders, a delay is likely to accumulate between audio and video in the post production process as a result of the use of time base correctors, video frame synchronizers, and digital video effects units, all of which delay video (and, sometimes, vertical interval time code) without affecting either audio or longitudinal time code. Correcting for this delay differential is a simple matter with digital audio delay devices.

Unfortunately, just as limiters and compressors must be locked together to prevent stereo imaging shifts, digital audio delays must be locked to the same clock to prevent phase distortion. There are very few true stereo audio delays available; the use of two monaural delay units, even of the same model, is not an acceptable alternative unless these devices are locked together.

Phase distortion is also a problem in network transmission circuits. When wire circuits are used, a difference in path length will create a frequency dependent phase error. Generally, higher frequencies will be farther out of phase than lower ones, but sufficient delay may actually advance higher frequencies into the next cycle, giving the appearance of worst phase error at middle frequencies. Amplifiers, subcarrier modulators and demodulators, and intercity multiplexers and demultiplexers can also offer differing path delays between channels.

The fact that the subcarrier modulators and demodulators presently used for terrestrial network television transmission in the United States exhibit excellent phase response should not be taken as an indication that all subcarrier modulators and demodulators will offer good phase response. In fact, those used for international satellite television transmission offer poor phase response, with errors varying not only with respect to frequency, but also with respect to time.

Intelsat offers no stereo audio transmission service. AT&T offers stereo audio between New York and several European cities, but provides no specification for phase response. Even the "stereo conditioning" offered by many telephone companies for wire circuits simply ensures that the frequency response between the channels will not deviate by more than one dB; no consideration is given to phase response.

It is, of course, possible to purchase a second video transmission channel and to feed digitally encoded stereo audio through it using one of the many encoders and decoders presently available. However, Kokusai Denshin Denwa Company of Japan (the only authorized Intelsat carrier in that country) has specifically stated that they would disconnect any such transmissions. Although neither Comsat nor AT&T has made similar claims (and both have carried digital audio on video circuits in the past), there is no guarantee that such transmissions will not be considered violations of tariffs in the future.

One way to improve monaural compatibility regardless of the alignment of recording heads, the path length of transmission circuits, or any other miscellaneous delays, is to deal with sum (L+R) and difference (L-R) channels, rather than left and right channels, in all recording, signal processing, and transmission. The sum channel, by the definition of every stereo television system recently proposed, is the monaural signal. Regardless of what happens to its phase relative to the difference channel, audible problems will not be heard by monaural listeners. In this case, phase problems will affect stereo listeners, causing separation errors, image shifts, and possibly even frequency response problems, though these are generally less noticeable than phase errors in a combined monaural channel.

To ensure that phase errors do not creep into the sum channel, however, it must be carried as such all the way from the mixing console to the transmitter. If separate left and right channels pass through processing equipment and recorders and pick up phase errors, these errors will affect the sum channel. If sum and difference signals pass through the same devices and pick up the same phase errors, the errors will not affect the sum channel.

Therefore, all stereo television facilities should be wired for sum and difference channels rather than for left and right channels, with the single exception of loudspeaker monitoring facilities. Even these would benefit from a center sum speaker in addition to stereo speakers at the sides (ideally fed from a sum/difference decoder), and from the practice of periodically switching from stereo to sum monitoring.

At the present time, there is virtually no sum/difference stereo audio equipment available, just as transmission carriers offer virtually no phase coherent stereo audio transmission paths. It is a simple matter to convert left/right equipment to sum/difference equipment. In fact, once the signals emerging from a mixing console have been converted into sum and difference, all other parts of a facility other than loudspeaker monitoring systems and transmitter inputs need only have their labels changed.

Handled correctly, stereo television could be very successful. It is important to realize that stereo television begins at the microphone, however, despite the fact that most present concern seems to be concentrated on the transmitter.

STEREO AUDIO PRODUCTION AND POST-PRODUCTION FOR TELEVISION

Douglas F. Dickey

Solid State Logic Limited

Oxford, England

With the completion of the technical and administrative preliminaries, American broadcasters are now faced with the practical implementation of Multichannel Television Sound. Not surprisingly, everyone is discovering that there are more technical and creative issues involved than initially meet the eye (and/or the ear).

The first area which broadcasters are scrutinising is the programme distribution and transmission chain. With the acceptance of a single industry-wide standard, the changes required in this area are readily defined, and implementation should be a largely straightforward engineering matter. A lot has already been written about this aspect of MTS, and I will not add to it here.

The second requirement for implementing MTS is to incorporate stereo and multichannel audio capability into planned and existing production and post-production facilities. Unfortunately, the changes necessary to properly accomplish this are substantially more diverse and subjective than those required to upgrade the transmission chain, and substantially less well documented. 1:00

It is hoped that this paper will be useful to the broadcaster who is seeking to organise his or her thoughts on the requirements of stereo teleproduction, and to point them towards the right questions to ask during their own planning process. With any luck, you might even find a few useful answers to some of the questions you are already asking or being asked.

A Question of Formats

Because stereo teleproduction and post-production are among the newest of the electronic arts, technical and creative production standards are just beginning to evolve. Let me give you an illustration of where we are today.

The proposed MTS transmission standard calls for three audio channels. Two will be full bandwidth, and we'll call these the Stereo Channels. The third channel will have a limited bandwidth, and it has been given the unfortunate acronym "SAP", which stands for "Secondary Audio Programme".

The use to which these three channels are put has a major bearing on the design of all audio production and post-production areas - yet a consensus on their best application is only now being reached. 2:00

One view is that the MTS system should emulate the discrete left, centre and right format employed by the motion picture industry. The Stereo Channels would be used to carry the "hard" left and right channel information, and the SAP Channel would carry the centre channel information, which would always be dialogue.

Film makers use this standard because it provides a stable centre image from any seat in a large theatre. Experimental broadcasts and experience with home playback of stereo video discs and cassettes (which use an ordinary "phantom centre" stereo format) have shown that a discrete centre channel is less important in the home environment, not to mention its incompatibility with millions of home stereo components. It is increasingly unlikely that the 3 channel film-style format will find much use in television.

A more progressive view is that the three channels should be used to carry an encoded mono/stereo/surround sound format such as the very effective Ambisonics system. While this format offers some elegant possibilities, it also multiplies production difficulties significantly. Furthermore, it requires a more complicated playback system, for which only a tiny fraction of American homes are equipped. Though Ambisonics cannot be dismissed as a possibility in second or third generation MTS systems, it does not look to be a practical concern for American television broadcasters for at least the next five years.

The production standard which has emerged from all of the debate calls for the use of the Stereo Channels to carry ordinary stereo left and right information, with the same degree of mono compatibility as records and tapes. The SAP channel will be used to carry a seperate mono programme. One major use of this facility will be to provide second language soundtracks, which could increase audience significantly in the many areas of the country with large foreign-speaking populatations. Other good uses for the secondary audio programme suggest themselves with a bit of creative thought.

Consumer Acceptance

A fundamental point is that the transistion to stereo television must be accompanied by an overall improvement in audio quality.

The average consumer has a much better natural awareness of good sound today than was the case when stereo records were first introduced. Even modestly priced home stereo systems are capable of significant fidelity, and the mass audience will be comparing your sound quality with that of FM broadcasters and the pioneering AM Stereo outlets.

A more critical marketing consideration is that the first group of people to equip themselves with stereo television receivers are likely to be the same people who presently own Beta Hi-Fi's, Stereo VHS, Compact Disc players, and above-average stereo components. It is against their standards that stereo television will be initially judged. If the trend setting critics and consumers find the medium lacking in quality, the mass audience will almost certainly be slow to follow.

Audio Monitoring

Quality control in audio production begins with the listening environment. Having defined the main production format as standard stereo, the starting point is to replace the old mono speakers with a good stereo left/right pair. Which brings us very quickly to problem Number 1.

Having been a poor stepchild in television production for many years, audio frequently finds itself tucked into the odd spaces that were left after the video, lighting and machine rooms were accomodated. Most of these spaces are not suitable for proper stereo monitoring. Problems include the inability to achieve a solid centre image at the mixer's position, unacceptable frequency response due to room geometry and reflective surfaces - in fact a whole multitude of sins.

The film and recording industries have made significant advances in control room and dubbing theatre acoustics, but these do not apply directly to stereo television control room design. The television mixer must balance the sound to achieve a natural perspective in relation to the television picture as it will be perceived in the home. And that is a different function than mixing records or theatrical film soundtracks.

The major independent video sound houses have done a lot of work in this field, but they are generally starting with a purpose-built acoustically designed space. In the more typically awkward existing television sound environment, the optimum positioning of monitors is a more difficult matter.

This need not mean rebuilding your control rooms. In many cases, nearfield monitoring (to minimise the contribution of the room itself), plus judicious equalisation and some minor acoustic improvements will provide a workable solution. Be prepared for some experimentation. The point is not perfection, the point is that you must be certain that your listening areas provide a reference standard which results in excellent stereo quality at the receiver.

You'll also need to decide who needs to listen to what. Aside from the mixer's position, stereo monitoring will be required by the sound effects area, master control and probably by the technical director. Other areas can probably get by with good old mono.

It is important that the frequency response and perceived center image match closely in these different areas, so that everyone is able to reference their judgments to what the mixer is hearing.

In post-production areas, there will be no need for anyone to monitor the Secondary Audio Programme simultaneously with the stereo programme, but in live situations this will not be the case. On-Air control rooms must take this into consideration. It is also important to remember that the SAP transmission will have a limited frequency response. In mixing the secondary programme, it will be desirable to switch a matching filter into the monitors, similar to inserting the Academy curve into the centre channel monitor on a dubbing stage.

It is also important to consider the various feeds that each position might need to hear. The basic combinations are stereo channels in stereo, stereo channels summed to mono, and the SAP channel. To minimise communication, each position should be able to select the feed it wants independently.

Live Stereo Production

Turning for a moment to production itself, let's divide our programmes into two categories. The first category is live programming, or anything which goes to tape without post-production. Multi-channel sound for this type of programming need not be terribly difficult.

As an example, consider a stereo telecast of a football game. A mix with only a few stereo feeds plus the usual mono sources can yield effective and exciting results. One stereo feed would be the crowd reaction mikes. The second might be playing field ambience. Two properly placed coincident microphone pairs could satisfy the requirements. Opening and closing theme music could come from a stereo cart player.

The play-by-play announcer and colour commentatator would have their usual mikes panned slightly left and right of center, corresponding with their on-camera positions. Parabolic and shotgun mikes, field field reporters' mikes and the referree's wireless would all be panned to the center.

While this is a relatively simple set-up, the enhanced sense of openness it affords provides a dramatically improved sense of "being there" for the home viewer. And that excitement is a large part of what sells sports.

The same basic principles can be applied to live news and interview shows, and live-to-tape daytime dramas, sitcoms and variety shows. While each of these programmes entails varying degrees of complexity, producing them in stereo should not be an intimidating experience. Some thought must be given to the rules for matching camera angles and stereo perspective, some changes in microphone technique will be required, and this may reveal the need for some acoustic changes in the studio. Once again, the watchword is to be prepared for some experimentation.

Live SAP

Creating a live Secondary Audio Programme, particularly a foreign language soundtrack, complicates things a bit. Generally, it will require some mix minus feeds and at least one additional IFB channel. Going back to our football game example, let's assume that the SAP is carrying the programme in Spanish.

First, we need to obtain a mix minus feed of all microphones and tape sources other than announcers' and reporters' mikes. Stereo pairs will need to be summed to mono.

Then, we need to add the Spanish announcer's mikes to this mix minus feed, and to provide them with appropriate IFB. If the English announcers are to conduct live interviews during the game, we will need access to their mikes with independent level control to the SAP mix bus, so that we can add their voices underneath a simultaneous translation. The translator will also need appropriate foldback.

A second mixing engineer will be required to monitor and control the elements of the SAP mix. If the audio is originating from a truck, the SAP mix may well be performed on the same console and monitored on headphones, but a seperate mixing area would be preferable.

Storage of the Audio Programmes

Up to this point, the production requirements of Stereo Television Audio and the Secondary Audio Programme present no major challenge to the state of the art. Things get more interesting when we stop thinking live and start thinking tape.

The first problem is that the present generation of one inch VTRs provide three audio channels. This means that we can accomodate Stereo plus timecode, or Stereo plus SAP. Unfortunately, what we want to do is record Stereo, SAP and Timecode.

In the past, stereo simulcasts have simply locked the video master to a four track ATR which carried the required programme audio. This is adequate for occassional special programming, but when MTS becomes an everyday occurrence, it would be a horrible nightmare.

The costs and problems involved in changing over to a new one inch standard with four audio channels do not seem worth it at this time. There seem to be fewer drawbacks to printing Vertical Interval Timecode on the video master, thus freeing the track normally used for longitudinal SMPTE timecode to store the SAP signal.

This approach will probably require improved audio electronics for the third channel, and the addition of VITC to SMPTE translators on the ports of a lot of synchronisers, but it seems the lesser of two evils. The final word is not in on this, as many people are just beginning to think about the problem. I'm sure that anyone who comes up with a better solution will have no problem finding buyers.

Audio Post-Production for Television

By far the most interesting and creatively powerful aspect of stereo television is the audio post-production process. Borrowing freely from the technology of motion picture mixing, video editing and multi-track recording, audio-for-video post-production engineers are developing tremendously flexible and efficient methods for creating soundtracks.

Ten years ago, video audio post-production was in its infancy. Today, it has become capable of producing soundtracks which rival those of the most elaborate feature films, at a fraction of the cost in time, labour and materials.

Allow me to vastly oversimplify: In film sound post-production, each soundtrack element is first transferred to an individual sprocketed magnetic tape, which is conformed to the fine cut of the picture by splicing in the required lengths of leader. A half hour programme can require dozens of these reels to be assembled.

In mixdown, the reels are loaded on sprocketed playback machines called dummies. The sprockets are used to maintain lock between the many dummies, the projector, and the master dubbers. The original elements are first mixed to create seperate Dialogue, Music and Effects mixes. The final soundtrack is constructed from these.

The good news for the broadcast community is that you will not need a lot of machines to do very advanced stereo post-production. You will not need editing tables, you will not need dozens of sprocketed dummies and dubbers, you will not need a projector, projectionist, loader, editor and three mixing engineers.

You will need a 16, 24 or 32 track ATR, a 3/4" or 1" video machine, a 4 or 8 track ATR, a couple of quarter tracks and cart machines, and a good synchroniser capable of locking at least three machines. If you do a lot of audio post-production and any of it is intricate, a a good computer-assisted mixing and editing system will pay for itself in short order.

The basic concepts of audio post are the same whether the medium is film or video, but the tools and procedures used in video are electronic rather than mechanical. Each soundtrack element is transferred to a single track of a 16, 24 or 32 track ATR, one track of which has already been striped with SMPTE timecode from the edited video master.

In conforming sound to picture, the starting timecode of each cue to be synchronised is determined and compared with the timecode of the matching frame on a dub of the video master. The exact offset between the two timecodes is calculated and maintained as the cue is re-recorded in sync onto the multi-track master. Elements which are required only to start at a precise time may be manually transferred to the multi-track simply by rolling the cue when the video master reaches the corresponding start point. In a manner similar to video editing, this entire process may also be controlled by computer.

When this "pre-lay" process is complete, a single reel of two inch audio tape will contain all of the soundtrack elements. For the final mix, this is played back in sync with the video master. The third machine in the lock is traditionally a four track ATR, although eight track recorders are gaining ground with the advent of stereo television.

Video Mix Formats

In audio post for mono television, a four track ATR is usually employed as the mix master. Track 4 carries the SMPTE code. Tracks 1 through 3 may be divided into Dialogue, Effects and Music as in film, or they may be assigned as Cast, Music and Audience, or whatever the producer finds particularly convenient. Whatever the format, the various elements on the mix master are combined to form the final soundtrack, which is then layed back to the video master.

The mix requirements for stereo television with a secondary audio programme are a bit more elaborate. Again, this is an area in which production standards have not been universally agreed, but it may be helpful to look at a format which is gaining a lot of ground because it so neatly fills the bill.

Instead of a 4 track ATR, the mix master is an eight track machine. Timecode is carried on Track 8. In setting up the stereo programme mix, the elements are first grouped into stereo pairs, with one pair for dialogue, one pair for music, and one pair for effects. The left/right dialogue mix is recorded on tracks 1 and 2 of the eight track, left/right music is recorded on tracks 3 and 4, and left/right effects on tracks 5 and 6.

Depending on the sophistication of the console employed, these three stereo pairs may be either simultaneously laid back to the one inch master, providing first generation stereo DME and composite mixes, or the stereo DME mix may be subsequently combined and layed back to the video master.

Once the stereo mix has been accomplished, the secondary audio programme mix is set-up. Having divided our soundtrack as outlined, this becomes a fairly quick and simple process.

The most common usage for the SAP will be foreign language soundtracks, so all that will typically differ in this mix is the dialogue tracks. Assuming that the foreign dialogue elements have been pre-layed to the multi-track reel, all that is required is to perform a mono foreign dialogue only mix, which is recorded on track 7 of the 8 track machine.

Tracks 3 and 4 are then summed to create a mono music mix, and tracks 5 and 6 are summed to create a mono effects mix. These are then remixed with the foreign dialogue mix from track seven, and the composite is laid back to the SAP track on the video master.

Computers for Audio Post

For relatively simple projects, all of the pre-lay and mixing procedures I have described can be performed manually. As is true of complex video post sessions, complicated audio post sessions benefit greatly from computer assistance.

Most synchronisers have some degree of memory which can be applied to repetitive machine control functions. Above this basic level of automation, mixing systems are available which allow all fader and cut button moves to be programmed into memory for easy revision, and this can be very helpful in an involved mix.

At a higher level still, equalisation and panning changes can be programmed, complex signal paths can be stored in memory for subsequent recall, and the mixing automation and machine control can be thoroughly integrated.

Audio post-production computers can be real time savers for even small operations. Once again, the point is to anticipate what your operational requirements will be over the next several years. Providing for that growth as part of your initial changeover to stereo will save a lot of grief and money in the not to distant future.

Summary

In summary, the implementation of multichannel television sound will require a careful re-working of the entire audio chain. In production and post-production areas, much more care will need to be taken with control room acoustics and the quality of audio monitoring.

Simple stereo-capable audio consoles will not be sufficient to meet the needs of the stereo channels and the SAP. More sophisticated busing structures are called for, and provision must be made for more elaborate IFB along with this. Given the need to produce audio programming with quality comparble to that found in other consumer media, the sonic performance and creative flexibility of consoles assumes new importnace to the broadcaster.

Multi-track tape machines will become a common tool of the television industry over the next few years. Broadcasters must consider not only the track capacity they will require, but whether or not they should invest in digital ATR's as part of their initial upgrade.

In audio post-production, the requirements for synchronised machine control will become increasingly complex, as will the requirements for computerised mixing systems.

The initial implementation of stereo and SAP capability will require a lot of thought, work and capital investment. The re-training of operators for live teleproduction will not pose any great problems, but the post-production arena will introduce a large number of new requirements, machinery and techniques.

The subject is vast and interdisciplinary, and I have only had time to just outline the aspects requiring consideration. I hope this provides a useful framework to assist you in the planning process. The rewards of multichannel television will be manifold, provided that the potential is understood and thorougly provided for.

MULTICHANNEL TELEVISION SOUND

Rationale for Implementation of the Second Audio Program Channel

Kevin Dauphinee
Equipment Planning Engineering
American Broadcasting Companies, Inc.
New York, NY

Multichannel television sound is rapidly approaching the implementation stage of its development, following the EIA Steering Committee selection of a single transmission and noise reduction system. The television receiver industry has placed an inordinate amount of emphasis on the newly selected stereo audio capability, but broadcast authorities have determined that the second audio program channel is the multichannel television sound service capable of generating additional return on what will certainly become a substantial investment. Attracting viewers by competitive programming is fundamental to the success of the broadcast television industry. The second audio program channel removes the language barrier in minority communities, providing greater accessibility to popular, competitive television programming.

In service in Japan since 1979, one of the two stereo channels of a stereo television sound system has been used successfully for the transmission of second language translations of television programs. Subjective evaluation of the Japanese system has demonstrated satisfactory performance for second language applications, provided that the translation and the original dialog are reasonably well synchronized and the levels are relatively equivalent. Under these circumstances, the masking effect of the preferred channel is suitable for the prevention of crosstalk perception. Acceptable crosstalk performance is possible from a system utilizing both frequency modulated main channel and second channel subcarrier (FM/FM), under ideal conditions. The multichannel television sound system currently employed in Japan is capable of separation approaching 60 dB at all audio frequencies. However, in the presence of multipath or ICPM distortion mechanisms, crosstalk may be reduced to a level below 45 dB. The perception of crosstalk between two uncorrelated channelsis minimized when channel separation exceeds 60 dB.

In light of the potential crosstalk problem presented by multichannel audio multiplex techniques and the required subcarriers, the Broadcast Television Systems (BTS) Committee of the E.I.A. determined the desirability for a separate subcarrier for simultaneous second language transmissions. Crosstalk from the main program into the translation could be reduced to inaudibility at the expense of the spectrum currently used as a guard band. Further incentive for the use of an additional subcarrier is the ability to provide stereo audio programming to any viewer wishing to upgrade their receiver to a stereo capability without imposing the limitation of a program dialog translation on one of the stereo channels. The Second Audio Program (SAP) channel subcarrier offers a simultaneous stereo and second language capability, without penalizing either viewer preference population. Not only are viewers offered the benefits of stereo audio or dialog translations, but the broadcaster has the potential to generate greater revenue from advertisers who wish to provide both stereo and second language audio. A greater audience can be attracted by appealing to the native language of certain minority communities and the enhanced impact of a stereo sound field is capable of providing a more lasting impression on the viewing audience listening in stereo. These reasons, among others, motivate the broadcast industry toward the implementation of multichannel television sound.

From the inception of the E.I.A. multichannel television sound subcommittee, ABC Broadcast Operations and Engineering has supported the incorporation of a separate subcarrier for second language translations of network programs. The efficient utilization of the television aural spectrum is readily achieved by the system proposed along with the acknowledgement that the potential for audience expansion is clearly within reach. Large segments of various communities speak languages other than English and those communities are served by broadcasters specializing in the minority languages in question. As a service to the non-English speaking community, the television broadcast industry is anxious to offer popular programming with the appropriate dialog translation. Implicit in the program offerings is a competitive environment with the capability to achieve increased viewership for broadcasters inclined to participate. The percentage of television broadcasters serving the non-English speaking viewing public is miniscule (less than 3%) compared to the potential participation made possible by the SAP channel. Likewise, advertising directed toward the non-English speaking community will stand to increase significantly, once the system has been placed into service on a meaningful scale.

By far, the Hispanic population in the United States comprises the largest non-English speaking community. 1983 estimates of Hispanic residents approach 24 million and that number is likely to increase beyond 30 million by the end of the decade. The ability to reach a percentage of this population is extremely attractive to both broadcasters and advertisers. The second audio program channel provides such an ability and offers an immediate return on a capital investment.

A most attractive feature for the use of the second audio program for Spanish language translations of network programs is the concentration of the Spanish speaking population. Approximately 60% of the Spanish speaking population in the United States are located in six metropolitan areas: New York, Los Angeles, San Antonio, Miami, San Francisco and Chicago. Over 86% of the total Hispanic population are located in thirty metropolitan areas, mostly in the southwestern region of the United States. From a network point of view, the opportunity to develop effective translations for prime time programs may be tested on a limited scale and rolled out to the remaining markets once techniques have been refined. Advertisers find these circumstances attractive since measurement of effectiveness is simplified and reliability determined through community homogeneity.

Recent ABC focus group tests of Spanish translations of the network program "The Fall Guy", were conducted by simulcasting over AM and FM radio stations in select markets. 60% of the Spanish speaking individuals contacted who were aware of the simulcast, in fact, listened to the simulcast while viewing the program. 10% of the respondents said they would not have watched the program if the translation of the dialog had not been provided by the simulcast. Since the experiment ran for three weeks, further data provided greater insight into the promise of the second language service. Of the 40% of the respondents who were aware of the simulcasts, 50% learned of the fact from radio, 40% from television, and 10% by word of mouth. Of the viewers tuning into the simulcast, 70% prefer advertisements in Spanish and 25% held favorable opinions of the companies that provided translations for their ads.

There are currently 18 Spanish language television stations in the United States and 5 Mexican stations near the southwestern border. The total advertising expenditures in the top 30 U.S. Hispanic markets reached $224 million with $93 million spent on television advertising. Approximately 65% of that total was spent in the top ten Spanish speaking markets. Budweiser alone spent over $3 million on Spanish language advertising in the United States. Clearly, a market exists that attracts programming competition, anticipating growth at a rate that should exceed currently expanding broadcast advertising. The Hispanic population is growing at a rate 6.5% greater than the population of the United States as a whole. It is this potential market that motivates the implementation of multichannel television sound, and in particular, the second audio program channel.

The breadth of the equipment requiring replacement or modification for the implementation of multichannel television sound is extensive. What might be misconstrued as simply adding two audio channels to a television system includes replacement or modification of control consoles, routing switchers, videotape systems, FM, microwave and optical fiber STL's, common carrier systems and transmitter exciters. Components will be configured quite differently from currently familiar facilities designs. The scope of the equipment requirements for a complete, network-wide multichannel television sound capability obviously prohibits short term implementation. However, a limited capability for network distribution of stereo programs and second language translations is both achievable and practical. Upon availability of a commercial exciter and a monitor receiver, a local station transmitter can be modified to inspect the system over the air. The viewers' response to the new technology will be closely followed in the weekly reports of multichannel sound receiver sales.

A limitless variety of implementation methods for multichannel television sound can be conceptualized. A basic approach to network and local distribution is provided to demonstrate the simultaneous capability of stereo program audio and an independent second language channel.

Figure 1 illustrates the production of stereo and second language channels from the multiple microphone inputs, as well as tape machine and auxiliary line feeds. Mixing consoles, routing and channel switchers will require at least three discrete audio channels to provide the multichannel sound capability. Consideration is given to a fourth audio level in switching equipment for time code distribution within the plant. In Figure 2, the local station receives the multichannel signal from the network via common carrier from a channel switcher. The transmitter environment is shown in Figure 3. Modifications to the RF equipment will be determined by the age of the transmitter and the level of performance with respect to intercarrier phase modulation.

Production of multichannel sound programs will develop slowly at first, beginning with live event commentary, as well as dialog translations of prime time programs. The complexity of such productions can range from a single additonal microphone and an isolation booth for live commentary to elaborate post production facilities, requiring accurate synchronization and appropriate dramatic interpretation to achieve the desired effect. Obvious comparisons can be made between second language production and closed captioning with respect to the near term availability and penetration of the new program service.

Multichannel television sound is a technology that has been studied extensively and is now on the verge of implementation. What distinguishes the system recommended by the E.I.A. from techniques used elsewhere is the ability to provide an enhanced service to two distinct audiences. The second audio program channel will provide dialog and commentary translations simultaneously with stereo program audio. The motivation for this service is derived from the additional audience potential that may be attracted by second language translations of popular television programs. This additional audience should result in the attraction of incremental advertising revenues for the broadcaster. The investment required to modify existing equipment for the provision of the multichannel sound service may be offset by the increased revenues derived from additional sales opportunities created by second language translations of programs and advertisements. Research has indicated that the television audience, broadcasters and the advertising community welcome the additional service of multichannel television sound, and in particular, the second audio program channel.

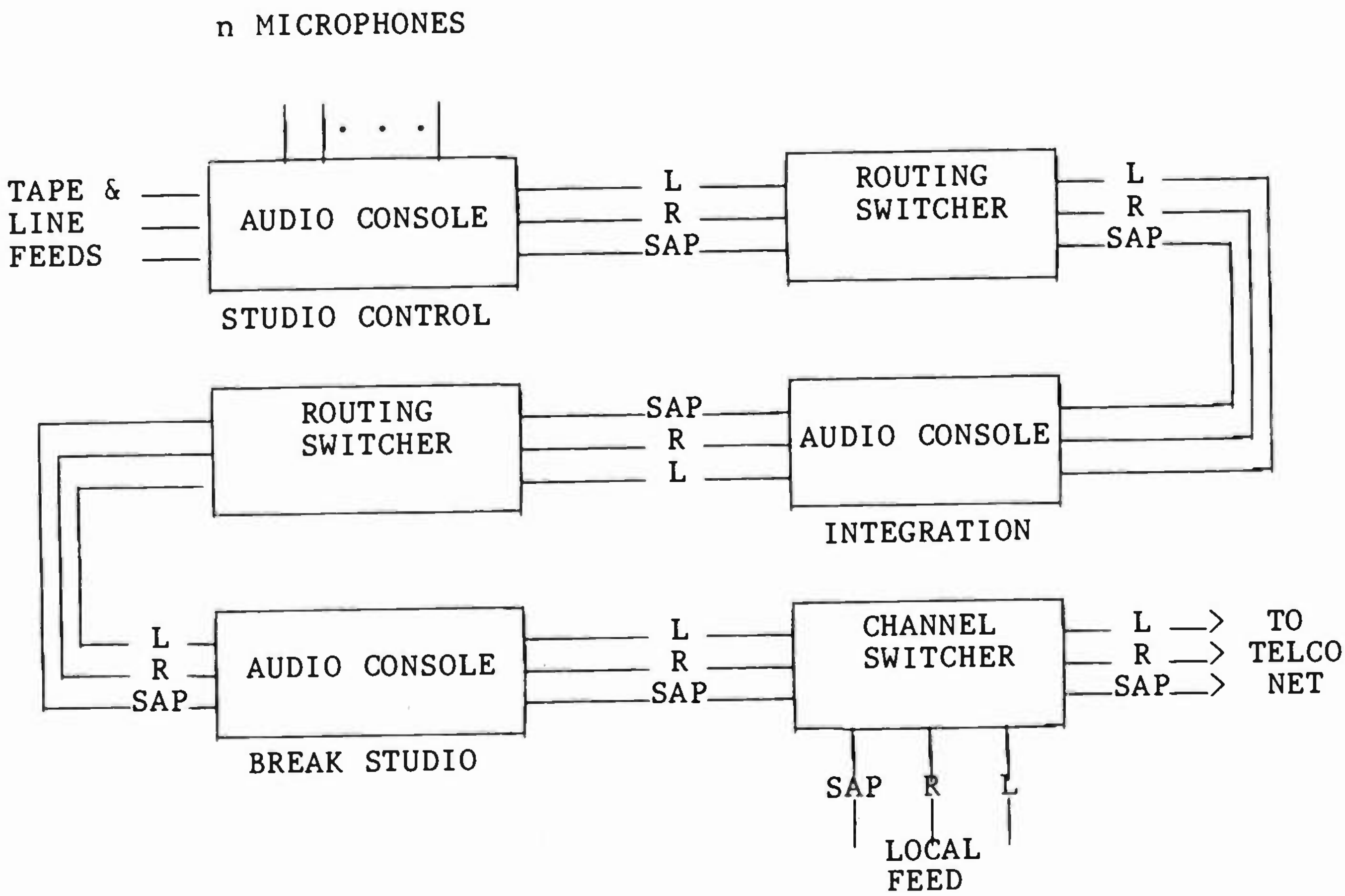


FIGURE 1. NETWORK STAGE ORIGINATION

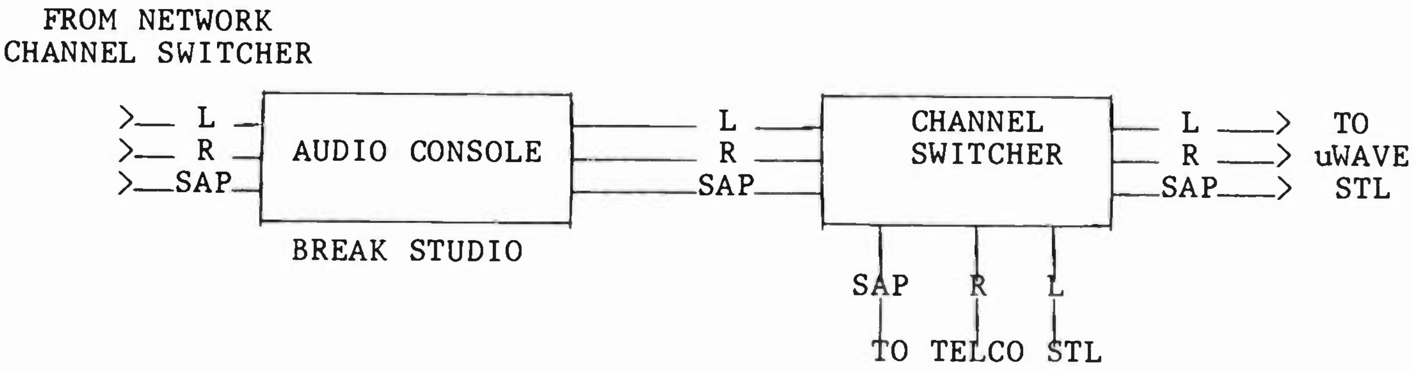


FIGURE 2. LOCAL STATION DISTRIBUTION

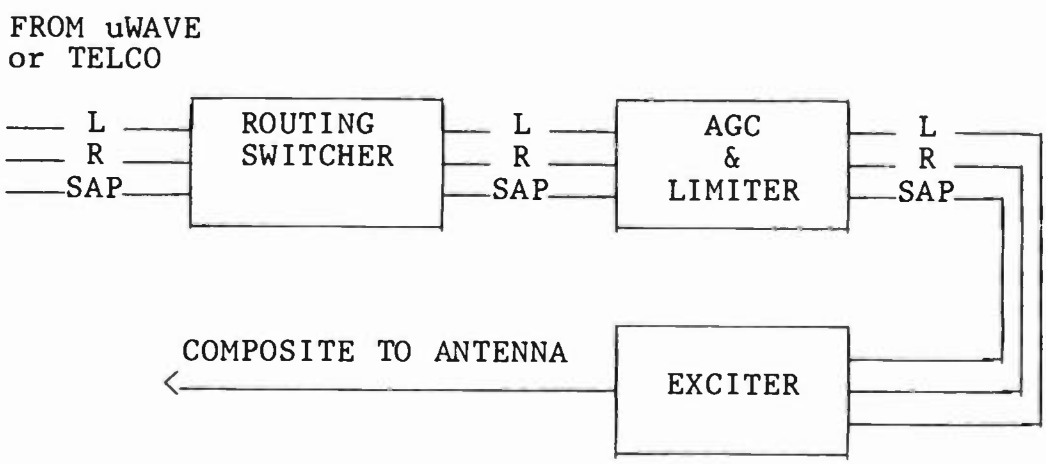


FIGURE 3. TRANSMITTER ENVIRONMENT

References

1. Alan L. Cohen, "ABC UNO: an Experiment in Dual Language Programming", Project Evaluation, July 25, 1983. ABC internal document.

2. Strategy Research Corporation, "A Comprehensive Study of U.S. Hispanics - A Market Profile, 1980.

3. Broadcasting/Cablecasting Yearbook, 1983.

4. Marketing & Media Decisions - December 1983.

5. E.I.A., B.T.S. Committee, Report on Multichannel Television Sound, The Basis for a Single Standard, Vol. 1-A, Nov. 9, 1983.

TRANSMITTER REQUIREMENTS AND CONVERSIONS

FOR TV MULTI-CHANNEL SOUND

G. L. Best

HARRIS CORPORATION

Broadcast Transmission Division
Quincy, Illinois

INTRODUCTION

With the adoption of multichannel sound by the FCC, the entire TV transmitter from the audio and video input terminals to the combined RF output terminals will need to be re-examined. To insure that transmitters are not the limiting factor in stereo reproduction, it is desired that the transmitters be as transparent to the incoming signal as possible.

In this paper we will examine the visual and aural requirements for TV multichannel sound and look at implementing changes that will be required from a transmitter system point of view. Let us first examine the aural portion of the transmitter.

AUDIO PERFORMANCE

Baseband audio of all 3 system proponents include components out to 105 kHz. Therefore, existing audio circuitry will need replacement with wideband audio circuitry. Emphasis will need to be placed on phase linearity and minimization of any amplitude ripples or roll-off over the audio passband to achieve good stereo separation and no cross talk between the stereo channel and the second audio program (SAP).

All errors in phase linearity or amplitude response within the aural transmitter tend to be cumulative. This is mainly due to the fact that roll-off normally occurs as the modulating frequency increases both in the audio and RF chains. Separation degradation due to amplitude roll-off will be more prominent than will departure from phase linearity. Table 1 gives separation degradation for various amplitude roll-offs.

Table 1. Stereo Separation vs Amplitude Roll-Off

Amplitude Roll-off	Fractional Bandwidth (fm/fc)	Stereo Separation
.04 dB	.1	52.0 dB
.16 dB	.2	40.7 dB
.36 dB	.3	33.6 dB

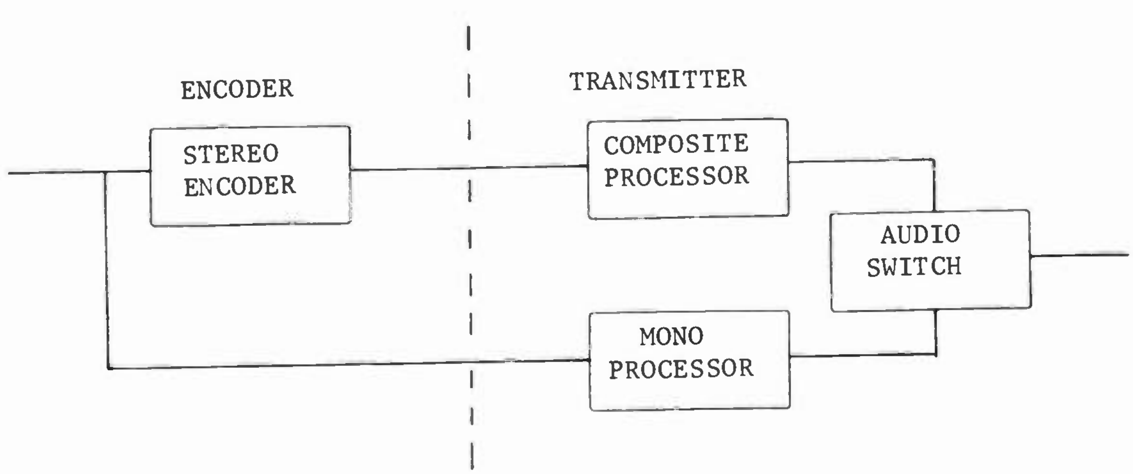
The data above assume a single pole roll-off and assumes no phase distortion. The phase distortion terms are roughly 7-8 dB below the amplitude components.

To achieve minimum distortion in the audio baseband signal, a minimum of circuit components should be used to avoid extra poles introducing extraneous phase shifts. To approach FM radio quality the FM signal to noise ratio needs to be significantly improved in transmitters. It is desired that the modulated oscillator be the limiting factor with respect to FM noise within the broadcast transmitter excluding the encoding scheme. Any noise contributed by the baseband audio circuitry should be kept to a minimum. Present day technology of low-noise large-bandwidth inexpensive OP amps can fulfill the requirements of flat amplitude response, linear phase response and signal to noise ratios of 80-90 dB. Figure 1 supplies measured data of the new Harris wideband audio card.

Since stereo program material is not yet in abundant supply, much of the material may still be monophonic. Rather than transmit the entire composite signal with no L - R information and thus increasing the amount of noise present at the receiving end, the broadcaster may choose to broadcast the mono signal by defeating the pilot at the encoder. If 75 kHz deviation is allowed the audio input level, if not automatically adjusted, must be readjusted so as to produce 100% modulation again. This may be provided in two ways:

a. The encoder can do all of this automatically.

b. The encoder can be bypassed and fed directly into a separate audio feed that also provides the pre-emphasis curve. An example of this is shown below.



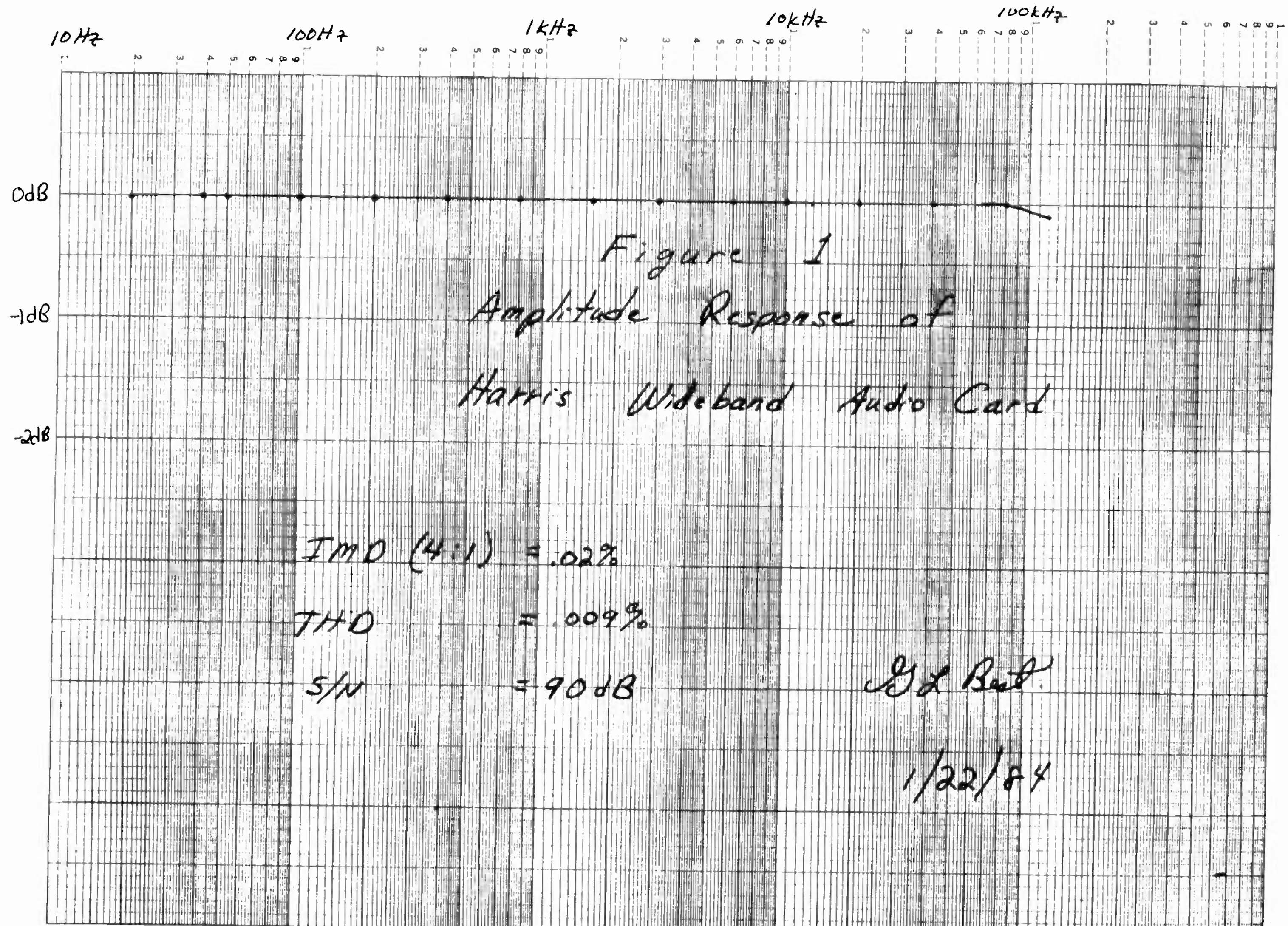


Figure 1
Amplitude Response of
Harris Wideband Audio Card
IMD (4:1) = .02%
THD = .009%
S/N = 90 dB
1/22/84
10Hz
100Hz
1kHz
10kHz
100kHz
0dB
-1dB
-2dB

All Harris aural exciters will include separate feeds for mono (with 2 SCA channels) as well as the composite feed, and provide the capability to remotely switch between the two feeds.

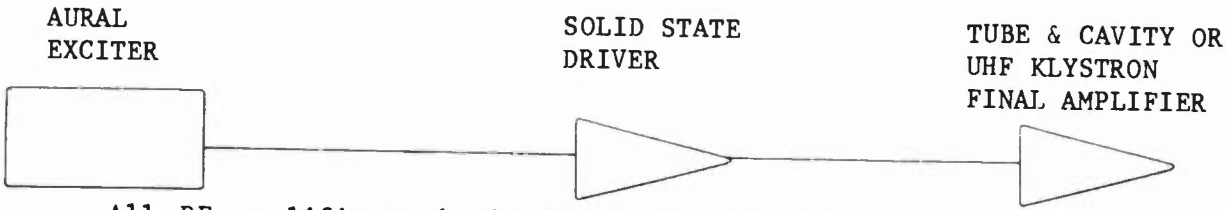
Modulated oscillator performance requirements will also increase. Instead of flat modulation sensitivity vs frequency characteristics up to 15 kHz, now the modulated oscillator will be required to have a flat response out to 47 kHz and it is desirable that it be flat out to 120 kHz. Because deviation will be increased by a factor of 3, FM S/N should improve at the transmitter by 20 log 3 = 10 dB. This will be true if a single tone test is conducted at 100% modulation. This is not an accurate representation of stereo FM S/N. A competent transmitter manufacturer will specify what the stereo S/N is for either channel (L or R) with de-emphasis and a given bandwidth. The true stereo SNR degradation is 23 dB if the deviation is held constant. But since we are increasing the deviation by 10 dB the net loss is 13 dB. This will be at the expense of higher distortion products generated. This is due to the fact that there is now more voltage swing on varactor capacitors causing any IM or harmonic nonlinearity also to be increased. Also the increased deviation must not impair the phase lock loops ability to maintain the average center frequency at its desired state.

A survey of modern TV transmitters indicate an average Intermodulation and harmonic distortion performance level of .5% to 1% as compared with that of .25% for FM broadcast equipment. Intermodulation and harmonic distortion products which, in monaural operation lie above 15 kHz, will now lie in the stereo channel or the SAP channel and degrade stereo separation or S/N in the SAP channel. In addition, IM and harmonic products generated in the stereo channel will now lie in main channel. Therefore it is almost certain that the modulated oscillator will need to be redesigned to improve A) IMD, B) THD, C) S/N, and D) frequency response.

The PLL must still function properly and any reference frequency sidebands (which show up in FM S/N) must not land in any of the designated channels. Those transmitters that do not employ PLL techniques may need to adopt them due to the following. By applying more voltage on the varactor diodes of the modulated oscillator, the varactors voltage to capacitance curve has become more non-linear. Thus the average frequency will tend to move from center frequency unless the varactor has perfect odd symmetry and such is not the case without custom varactors.

All modern transmitters are IF modulated. To insure these low level stages do not contribute any group delay or amplitude roll-off, wideband amplifiers should be used. As the signal is up converted to the RF aural carrier frequency, the FM of the local oscillator signal should be checked. The level of FM produced by LO should be 10 dB less than the modulated oscillator. This will result in a total FM noise degradation of only 2.3 dB. In many cases ovenized crystals are used as the source for the local oscillators. Synthesized LO sources should be avoided unless their contribution to the total noise is negligible. Synthesizer reference frequency spurs will show up as FM noise.

RF amplifiers in the aural chain that bring the stereo signal up to its desired levels must now be viewed from another angle. A typical RF chain is shown.



All RF amplifiers should be viewed from the point of obtaining a flat symmetric amplitude response and a minimized group delay across the passband. Since FM modulation and demodulation is a non-linear process there is not a one-to-one correspondence between RF amplitude/phase response and baseband stereo separation and crosstalk. Measurements indicate that a 2.5 MHz, 3 dB bandwidth is sufficient to achieve excellent separation. No longer can amplifiers be tuned just by maximizing efficiency without compromising stereo performance or crosstalk unless, the stage is designed to operate that way. The technology now exists for broadband solid state driver modules that do not require tuning over the whole band. These types of driver modules do not present any problems to multichannel sound. However, if these stages are of narrow band design, the driver stage must be scrutinized more carefully. With older transmitters a tradeoff will probably need to be made between achieving rated driver power and optimum stereo performance. The input circuits of these stages may need some redesign. However, in most cases where grounded grid circuits are used, the input Q is sufficiently low that no modifications are necessary. Now these amplifiers must be swept frequency analyzed to insure the passband is sufficiently wide to pass the stereo signal and that its passband shape is symmetric about the carrier frequency.

Since most aural PA (and tube driver stages) are single tuned circuits, the symmetry about the aural carrier is usually not a significant problem unless tuning is misadjusted. Weirather and Hershberger [1] suggest a method of tuning if swept response equipment is not available.

To get a rough idea of the input and output circuit bandwidth required, Carson's rule can be applied to the stereo portion of the baseband signal.

$BW = 2\ (\Delta F + fm)$ where F = peak frequency deviation
fm = the modulating frequency

For the stereo portion of the signal $F = 75$ kHz and $fm = 47$ kHz max, $BW = 2\ (75+47) = 244$ kHz.

If we desire that no group delay and amplitude distortion should exist over this portion of the band, the properties of a single tuned circuit require the 3 dB bandwidth be about 10 times the distortion-less bandwidth. This calculation returns us the approximately 2.5 MHz 3 dB bandwidth requirement of the input and output circuits on all amplifier stages following the modulated oscillator.

Klystron amplifiers have previously been adjusted with all cavities excluding the penultimate synchronously tuned on the aural carrier. On the lower channels this will result in an aural passband of approximately 1 mHz. Measurements will be conducted to if this bandwidth is sufficient for minimization of crosstalk and to achieve excellent stereo separation. If a 2.5 MHz bandwidth is required for optimum performance, the cavities must be stagger tuned in order to achieve the desired passband as shown in Figure 2. This will result in a decrease in the gain of the klystron. In tests conducted at Harris on the Varian S klystron, findings indicated that this loss in gain is roughly 6 dB when the klystron was tuned for saturation at rated power out (i.e. the most efficient operation condition. Thus in some cases an additional preamp may be required before the klystron depending on the drive power available from the exciter. This tuning is accomplished by moving the 3rd cavity approximately 1 MHz above the aural carrier and adjusting the 1st cavity approximately 1 MHz below the aural carrier frequency.

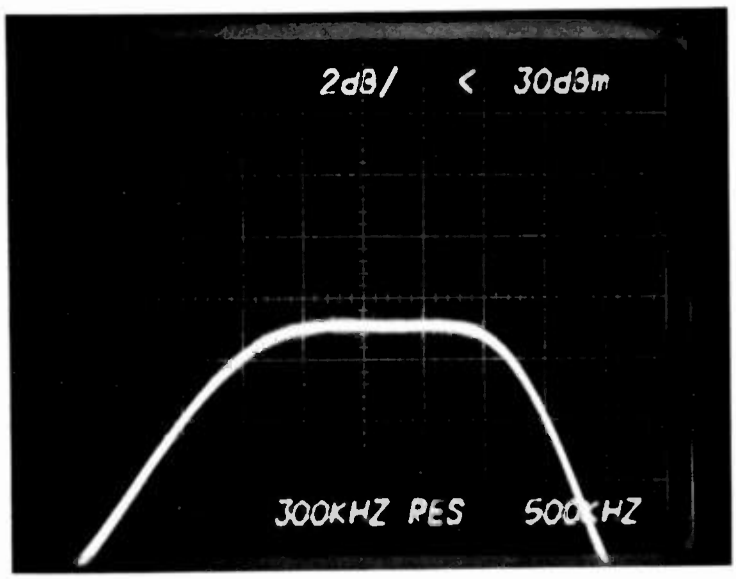
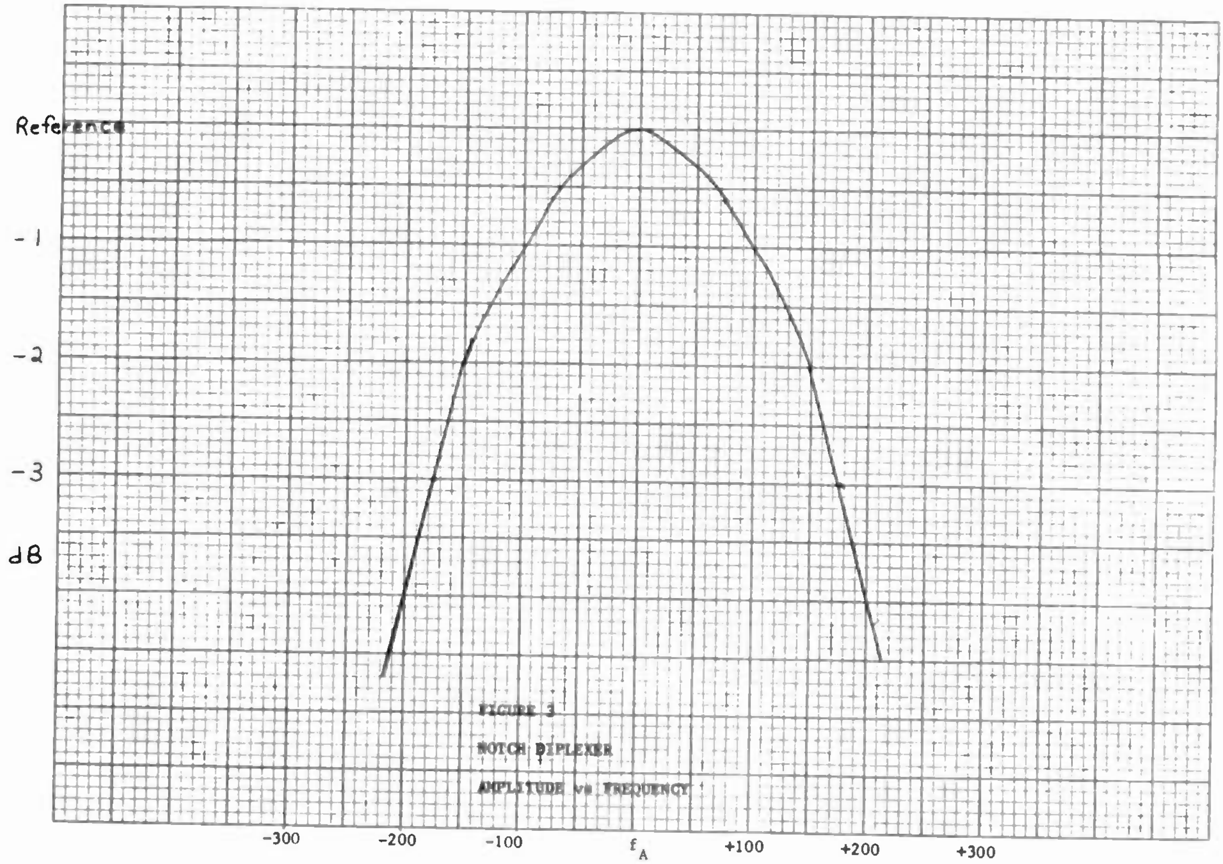
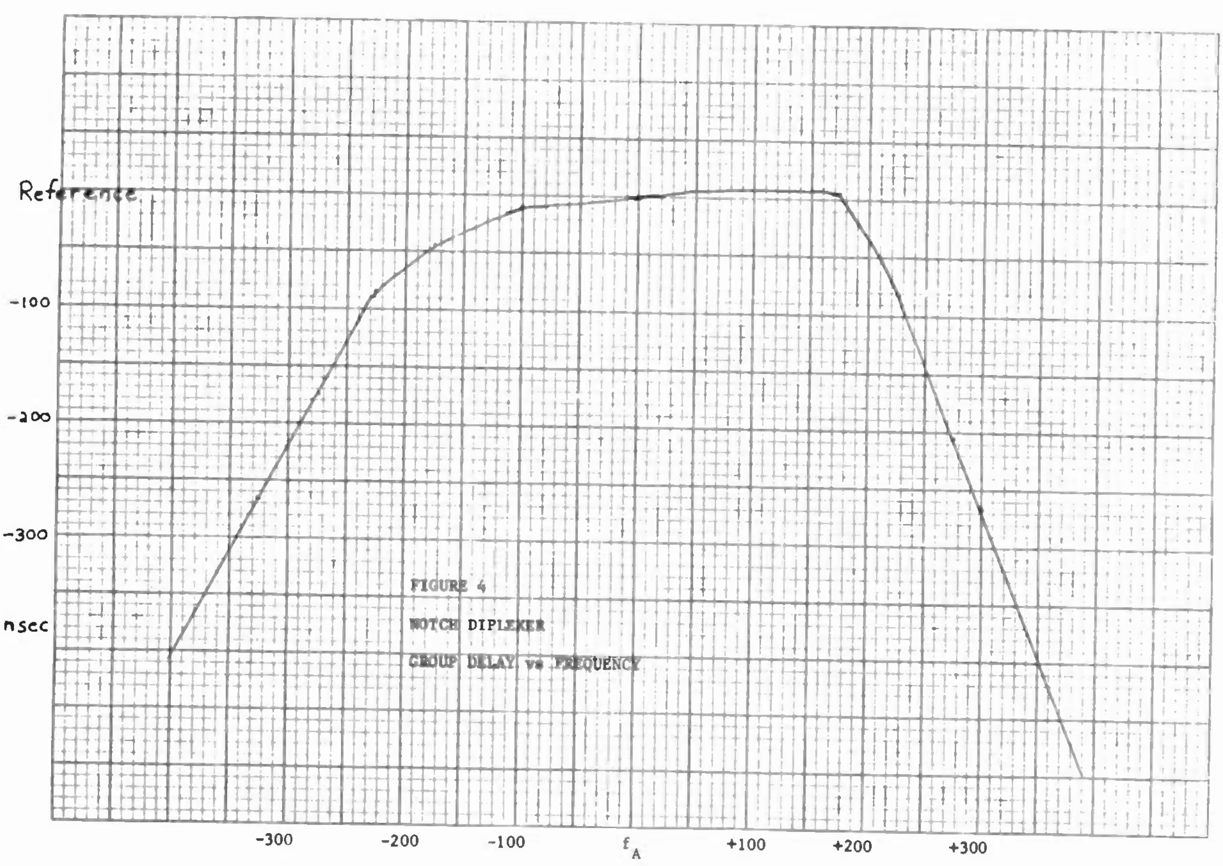


Figure 2. Wideband Klystron Tuning for Optimum Stereo Performance

The notch diplexer is the last and possibly the most critical element in the aural chain with regard to stereo separation. Hybrid diplexers do not present any degradations to multichannel sound. Standard notch diplexers are not necessarily tuned symmetric with respect to the aural carrier, so as to pass visual 4.18 MHz with minimum attenuation. Thus the group delay across the passband is both large and nonsymmetrical. A typical notch diplexer amplitude and group delay responses are given in figures 3 and 4. All pass group delay compensation located in the IF or RF chain can restore the group delay to a more desirable response. An amplitude corrector may be required to restore the aural passband to its desired shape. However, accompanying an amplitude corrector is an associated group delay characteristic. Thus another all-pass compensator is required to compensate for the amplitude corrector. Obviously this is a complex system and the tendency to go around in circles certainly exists.


Reference
-1
-2
-3
dB
FIGURE 3
NOTCH DIPLEXER
AMPLITUDE vs FREQUENCY
-300
-200
-100
f_A
+100
+200
+300


Reference
-100
-200
-300
nsec
FIGURE 4
NOTCH DIPLEXER
GROUP DELAY vs FREQUENCY
-300
-200
-100
f_A
+100
+200
+300
Khz

Another approach is to use 2 cavities in a double tuned arrangement in place of the standard cavity in the notch diplexer. This will yield a larger bandpass and smaller group delay over the bandpass. Tests have indicated that a stereo separation of 46 dB and crosstalk into the second audio program at less than 40 dB can be achieved without the use of special compensation or equalizer circuits.

In summary, these major changes must occur in the aural portion of the transmitter:

1. Wideband audio stages with high S/N should be employed to insure negligible distortion in the baseband signal.

2. Low distortion modulated oscillators with improved noise performance and PLL techniques should be used so that the quality of TV multichannel sound will approach that of the FM broadcast service.

3. The bandwidth of IF and RF stages should be widened. Power - bandwidth tradeoffs may need to be made to achieve optimum stereo performance. Where feasible, the TV station might consider 20% aural power to increase its stereo coverage area. This may compensate partially for the degradation in S/N due to the increased bandwidth of the stereo signal.

VISUAL TRANSMITTER PERFORMANCE REQUIREMENTS

ICPM

There is no defined level of ICPM for a given stereo performance level since the signal to buzz ratio is highly dependent on the picture spectral components. However, results of the EIA committee on multichannel sound provided important data with respect to buzz to signal ratio and signal to buzz beat ratio.

The video-test pattern selected represents a severe test. Significant levels of $nf_H \pm mf_V$ (f_H = horizontal scanning frequency and f_V = vertical scanning frequency) were present which are the largest contributors to buzz when ICPM is present. Analysis of the data in Table 2 reveals that for a signal to buzz ratio (SBR) of 45 dB, (weighted quasi-peak) a maximum of 2° ICPM can be tolerated. This data represents measurements made with a Nyquist receiver which is the dominant type in the field. A signal to buzz beat ratio of 30 dB in the SAP channel was obtained with 2° of ICPM. Further decreases in ICPM below 2° yielded no further increases in signal to buzz beat ratio.

Table 2. SBR vs ICPM[2]

Channel	ICPM	Worst Case SBR (weighted quasi-peak)
	$\pm 5.0°$	37 dB
STEREO	$\pm 2.5°$	45 dB
	$\leq \pm 1.0°$	50 dB
	$\pm 5.0°$	33 dB
SAP	$\pm 2.5°$	32 dB
	$\leq \pm 1.0°$	32 dB

The main limiting factors in SBR in the SAP channel will most likely be the receiver. The equivalent amount of ICPM generated by a transmitter to equal the buzz produced by the Nyquist slope of the TV receiver can be given by:

$d = 1.2\ (F/f_H)$ where F is the audio channel frequency and f_H = 15.734 kHz.

Therefore, at $F = 5\ f_H$, the SAP channel carrier frequency, the transmitter must produce 6.0 degrees of ICPM.

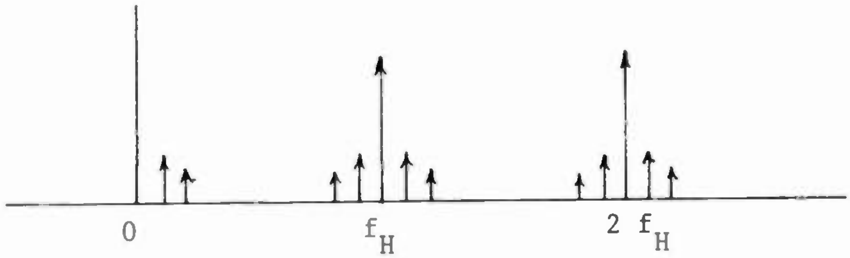
MISCELLANEOUS TOPICS

Internally Diplexed Transmitters

Internally diplexed transmitters offer about 20 dB less protection than externally diplexed transmitters using notch diplexers to IM products generated in circuitry following the IF SAW filter. A video notch at 4.5 MHz will reduce direct modulation components causing interference in the audio channel. This sharp notch must be both stable in frequency with all temperature variation and a phase equalizer must follow the notch filter to remove the phase perturbations in the video passband at 4.18 and below. A typical IF SAW filter like the one used in the MCP-2 visual exciter will provide typically 20 dB isolation to those 4.5 MHz components as well. Measured isolation of a class AB grounded-grid visual PA cavity indicate that the IM product at 4.5 MHz generated by 3rd order distortion between the visual carrier and a 100 IRE 2.25 MHz sine wave superimposed on a 50 IRE pedestal was 64 dB down with respect to the aural carrier. These IM products are not anticipated to be a significant problem.

Importance of Maintaining Precision 4.5 MHz Intercarrier

Video sidebands translated to the intercarrier will appear at nf_H in the audio baseband as follows.



If the 4.5 MHz intercarrier is maintained precisely, the buzz caused by ICPM transferred to the intercarrier will be minimized. If the 4.5 MHz intercarrier was to drift $\pm$ 400Hz the $nf_H \pm mf_V$ spectrum will be displaced and thus 400 Hz beat tones would be audible and possibly objectionable. One method of maintaining the 4.5MHz intercarrier is to lock it to a stable source.

Example of System Performance

Let us assume we want our overall transmitter system to achieve a 40dB separation and also 40dB crosstalk into the SAP from the main and stereo channels. Subjectively, 40dB separation is considered good quality. Working backwards from the notch diplexer we can arrive at a goal for each stage in the aural chain. Measurements previously mentioned indicate 46dB separation and 40dB SAP crosstalk are achievable with a 2 cavity double tuned multiplexer design similar to the proposed 2 cavity double-tuned notch diplexer. Power amplifier cavities with 2.5MHz 3dB bandwidths will typically yield 55dB separation. Furthermore, let us assume our drive stages are 4MHz wide at the 3dB bandwidth points so that any contributions in group delay or amplitude distortions from them are considered to be negligible. The low power RF and IF stages following the modulated oscillator are typically 10MHz at the 3dB points so their contributions may be considered negligible also. Our audio circuits have wideband cards so that a 50dB separation is achievable at baseband. Knowing these figures, we can calculate the resultant stereo separation assuming that all errors are cumulative (Murphy's Law).

$$\begin{aligned}\text{Separation} &= 20 \log \left(10^{\frac{-46}{20}} + 10^{\frac{-55}{20}} + 10^{\frac{-50}{20}}\right) \\ &= 40 \text{ dB}\end{aligned}$$

By using wideband PA, driver and low power RF and IF stages the stereo to SAP crosstalk should be 60dB down based on measured data from the FM broadcast service.

Thus stereo to SAP crosstalk

$$\begin{aligned}&= 20 \log \left(10^{\frac{-40}{20}} + 10^{\frac{-60}{20}}\right) \\ &= 39 \text{ dB, which also is quite acceptable.}\end{aligned}$$

SUMMARY

1. General design guidelines and performance requirements regarding aural TV transmitters for achieving FM stereo quality comparable to the FM radio broadcast service have been presented.

2. The influence of ICPM upon signal to Buzz ratio has been discussed and goals of acceptable ICPM outlined.

3. Transmitter system configurations have carefully examined to determine where possible tradeoffs must be decided upon and what modifications may have to be made.

4. A brief look was taken at modifying existing transmitters to yield better stereo performance.

FUTURE DEVELOPMENTS

1. Notch diplexers will be better characterized for the performance required to yield optimum stereo performance.

2. Many stations will be requesting retrofit kits to up-grade their existing system.

3. There will be an increase in the use of white clippers and ICPM correctors to minimize buzz components and the adjustments to control these parameters will be more closely monitored.

4. Eventually FM stereo generators and mod monitors may become a part of the standard TV multichannel sound transmitter.

REFERENCES

1. Weirather & Hershberger, "Amplitude Bandwidth, Incidental AM, and Saturation Characteristics of Power Tube Amplifiers for FM", IEEE Transactions on Broadcasting March 1983, Vol. BC-29, No. 1, p 14.

2. EIA Report on Multichannel Television Sound, Volume 2A, October 31, 1983.

MONITORING TV MULTICHANNEL SOUND

Joseph C. Wu

TFT, Inc.

Santa Clara, California

INTRODUCTION

This paper is based on the preliminary engineering work for the purpose of determing the design requirements of an aural modulation monitor for TV Multi-Channel Sound (TV/MCS) which is suitable for off-air monitoring of the aural transmitter signal as well as for proof of performance use. It is assumed that the Zenith Corporation's TV/MCS System will become the industrial standard.

Major design objectives for the TV/MCS Monitor are stated in this paper. Three major signal processing steps are necessary in demodulation a TV/MCS signal for measurement purposes. These are baseband demodulation, stereo demodulation and subchannel demodulation. In addition to the description of the three major block diagrams in the demodulating process, the choice of circuit design for optimizing the monitor's performance are discussed.

The modifications required to extend the service of a monaural modulation monitor for TV/MCS are also discussed.

WHY MONITOR MODULATION

The recent changes in the FCC Rules have eliminated the type approval requirements of the aural modulation monitors. The Commission has given the broadcasters freedom to choose their method of monitoring modulation.

Assurance of having a high quality aural signal is perhaps the most important reason for having a monitor. An aural monitor provides means of maintaining constant loudness from various program sources. To be able to do proof of performance is another important reason to have a monitor that demodulates the RF signal and separates the components in the composite signal for analysis.

An RF carrier without modulation does not provide coverage. Therefore, keeping a proper level of modulation assumes adequate signal coverage for the communities the station serves. Preventing over modulation in order to meet the FCC legal modulation limits is another good reason to have a monitor.

MONITOR VS SPECTRUM ANALYZER

Spectrum analysis is commonly used for analyzing modulation. A table of comparison (Table 1) is given to compare the features and capabilities of a modulation monitor and against a spectrum analyzer. Disregarding the cost factor, one can conclude that the spectrum analyzer is more suitable as an analytical tool for trouble shooting and a modulation monitor is a tool for day-to-day on-air operation monitoring.

MONITOR VS. SPECTRUM ANALYZER

FUNCTIONS	SPECTRUM ANALYZER	MODULATION MONITOR
PEAK MODULATION MEASUREMENT	Approximate	Absolute
RMS MODULATION (LOUDNESS)	Depending on User's Interpretation	Depending on Meter Circuit and Meter Ballistics
MODULATION CALIBRATION	Bessel Null (Additional Test Equipment needed)	Frequency Synthesis (Built-in)
AS ANALYTICAL TOOL	Excellent	Good
EASE OF OPERATION	Trained Personnel Required	Little Training Required
EQUIPMENT COSTS	$25K - $35K	$6K - $10K

TABLE 1

AN IDEAL MONITOR

An ideal modulation monitor should consist of, but not be limited to, the following functions:

a. Demodulates the composite signal from the aural carrier.
b. Separates the components in the composite signal for measurements.
c. Capable of off-air monitoring.
d. Covers all VHF and UHF channels.
e. Suitable for proof of performance use.

BASEBAND MONITOR

The very first step in monitoring the aural modulation of an aural transmitter is to demodulate the composite baseband signal from the aural carrier. Majority of the TV stations nowadays are operating their transmitter facilities by remote control, monitors having off-air monitoring capability is required. Converting the RF carrier to I.F. frequency, the monitor can cover all the TV channels by changing the local oscillator (L.O.) frequency and the channel filter frequency. The incidental phase noise in the L.O. must be kept below 75 dB in order to achieve a better than 70 dB signalto-noise ratio (SNR) in the output of the monitor. FM noise in the L.O. is directly translated into the demodulated composite signal.

Two methods of demodulating are desirable in a TV/MCS monitor. Being able to demodulate the composite signal directly from the aural carrier as well as from the inter-carrier allows the operator to compare SNR due to intercarrier phase modulation (ICPM) to the independent carrier demodulation method. A pulse counting FM discriminator circuit is chosen to demodulate the composite signal due to its high degree of frequency-to-voltage conversion linearity in this type of circuit. Better than 0.1% of harmonic distortion can be achieved from a properly designed pulse counting discriminator.

The phase linearity of the low pass filter immediately following the FM discriminator is extremely important for achieving high stereo separation. One-half of one degree phase linearity is required in the low pass filter to yield a 50 dB separation measurement requirement.

The sensitivity of the peak modulation indicators (peak flashers) has been a controversy of the industry for a long time due to the lack of definition in the FCC Rules for type approval as to what constitutes an incidental over modulation. A new type of peak modulation indicator has been developed at TFT which can differentiate the duration of the peaks modulation as well as measure the amplitude of the peaks, it allows the user to adjust the sensitivity of the peak flashers and set up a standard for measurement of pulse width and amplitude. Having this user adjustment capability, the monitor can differentiate short duration pulses generated by multi-path or overshots from over modulations. Figure 1A shows the circuit that allows varying the sensitivity of the peak flasher to the pulse width as well as amplitude.

An easy to use FM Calibrator is essential in a monitor designed for day- to-day use. Although Bessel-Function Null technique for calibrating frequency modulation has been known to many engineers, it is not the most convenient method nor least expensive method for calibration of FM. An FM Calibrator using frequency synthesis method is shown in Figure 1B, which illustrates the simplicity and economy of using frequency synthesis method to calibrate FM Modulation. The accuracy of this type of calibrator is better than 1% because its long term stability is only a function of the crystal aging.

When the RF signal to the monitor is picked up from an antenna, intermodulation products can be generated due to multi-path propogation and could degrade the performance of the monitor. Having a multi-path detector built in the monitor and the use of a directional antenna can help minimize the multi-path distortion. The multi-path detector is designed to detect the envelope (AM) in the I.F. amplifier before the signal goes into limiting. Rotating the antenna for minimum amount of AM on the envelope helps reduce distortion in the monitor due to multipath propogation.

STEREO MONITOR

The stereo monitor for the TV/MCS can be an integrated part of the Baseband Monitor or can be a separate external unit. It processes the composite signal through filters and stereo decoder for measurement purposes.

A 15 kHz low-pass filter and a bandpass filter having a bandpass from 16.5 kHz to 46.5 kHz are used to separate the (L + R) and (L - R) channels from the composite signal. Two narrow band filters are tuned to 31.5 kHz and 15.75 kHz to extract the pilot carrier and the subcarrier respectively. The output of these four filters are calibrated so that the relative levels of the four components, L + R, L - R, subcarrier level and pilot level, can be measured. The pilot carrier frequency also can be measured by using an external frequency counter.

The stereo decoder is the phase-lock-loop (P.L.L.) type. The subcarrier (31.468 kHz) is generated from a P.L.L. which is phase coherant to the pilot carrier. Phase adjustment of the subcarrier is not required for making stereo separation measurements as long as the P.L.L. is in "lock" condition. Over 55 dB of stereo separation can be achieved in this type of decoder.

FM SUBCARRIER MONITOR

Figure 3 shows a simplified block diagram for a TV/MCS FM modulated subcarrier demodulator using frequency-converting design. The composite signal of the TV/MCS is up-converted to an I.F. frequency and the subcarrier to be measured is filtered at the I.F. frequency using a properly designed I.F. filter. This allows the subcarrier channel monitor to operate at any subcarrier frequencies without having to change the filter design for each different subcarrier. Furthermore, the filter's characteristics and its design can be optimized at one common I.F. frequency for all subcarriers. Multiple subcarriers can be measured by changing the local oscillator frequency.

By converting the I.F. back down to the subcarrier again, using the same local oscillator signal, the subcarrier frequency error is maintaned at the output of the second mixer. By using a frequency counter, the subcarrier frequency accuracy can be measured by removing the FM modulation to the subcarrier.

UP GRADING MONO-MONITOR FOR TV/MCS

Up grading a aural modulation monitor for TV/MCS use requires circuit modification in two major areas; extending the linearity of the FM discriminator from 25 kHz to accomodate 75 kHz frequency deviation and extending the low-pass filter's cut off point to 110 kHz in order to allow the passage of the entire composite signal. The complexity of these modifications depend on the type of circuits originally used in the monitor design. The pulse-counting type FM discriminator can be up graded to 75 kHz operation by changing the time constant that affects the pulse width. It is necessary to replace the 15 kHz filter now used in the monaural monitor with a new filter having 110 kHz roll off.

Attention is required to make certain that the L.O. birdies (inter modulation products) are at least 65 dB below the desired signal within a 100 kHz bandwidth, so that the spurious signals will not be generated in the first mixer and I.F. amplifier. I.F. bandwidth also requires attention to assure adequate stereo and subcarrier response. The rule of thumb for bandwidth requirement is two times the sum of frequency deviation and the highest signal frequency. For TV/MCS the I.F. bandwidth requirement is approximately 370 kHz using the above rule of thumb.

After the baseband monitor has been up-graded, it is necessary to add a stereo monitor or subcarrier monitor or both to monitor the performance of the entire TV/MCS programs. The stereo and subcarrier channel monitor will require the composite baseband signal fed from the output of the modified mono monitor.

CONCLUSION

TV/MCS monitoring requirements are more complex than the monaural sound transmission. The technology for monitoring TV/MCS transmission has already been developed due to the similarity between the TV/MCS and the FM stereo and SCA broadcast. Furthermore, existing aural modulation monitor for monaural application can be up-graded for MCS use.

REFERENCES

1. Report on Multichannel TV Sound. Electronic Industries Association Broadcast Television System Committee, Volume 1-A, Page 14-36, Page 49-55, Page 235-238.

2. Intercarrier Buzz in TV Receivers, Report on Multichannel TV Sound, Appendix E. Carl G. Eilers, Zenith Radio Corp.

3. FM Modulation Calibration Using Frequency Synthesis Method, NRBA Papers, Oct. 1983 by Joseph Wu, TFT, Inc.

4. Model 701 and 702 TV Aural Modulation Monitor Operation Manual, by TFT, Inc., 1975.

5. Model 724B FM Stero Monitor Operation Manual, by TFT, Inc., 1976.

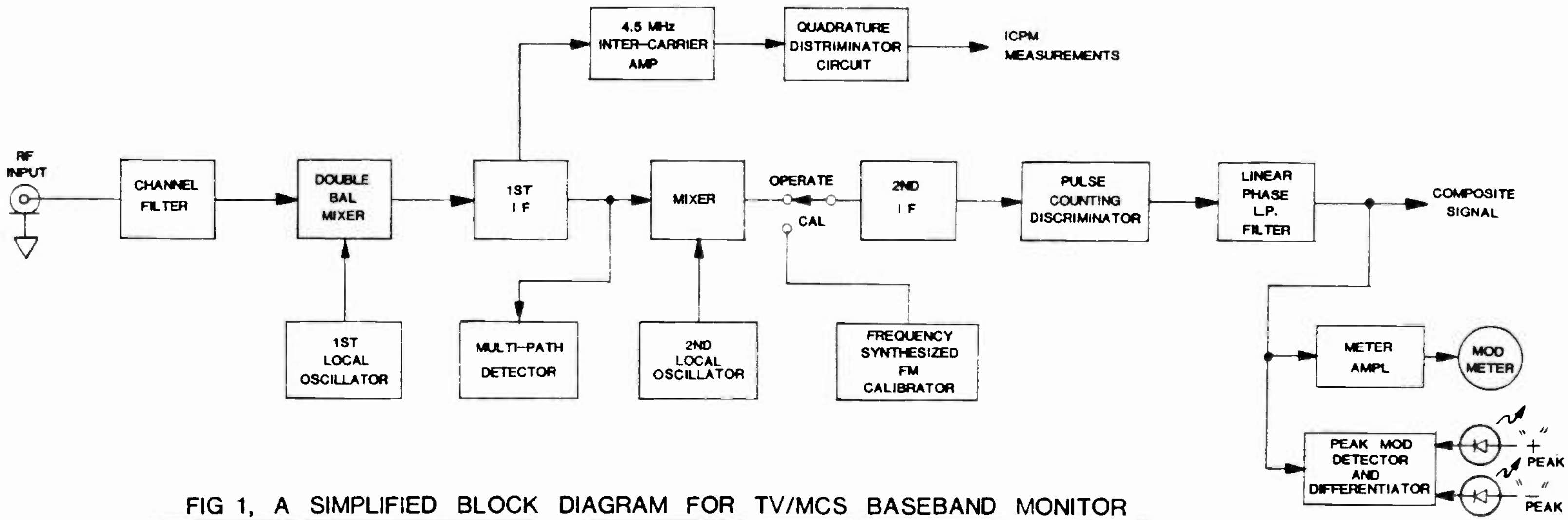


FIG 1, A SIMPLIFIED BLOCK DIAGRAM FOR TV/MCS BASEBAND MONITOR

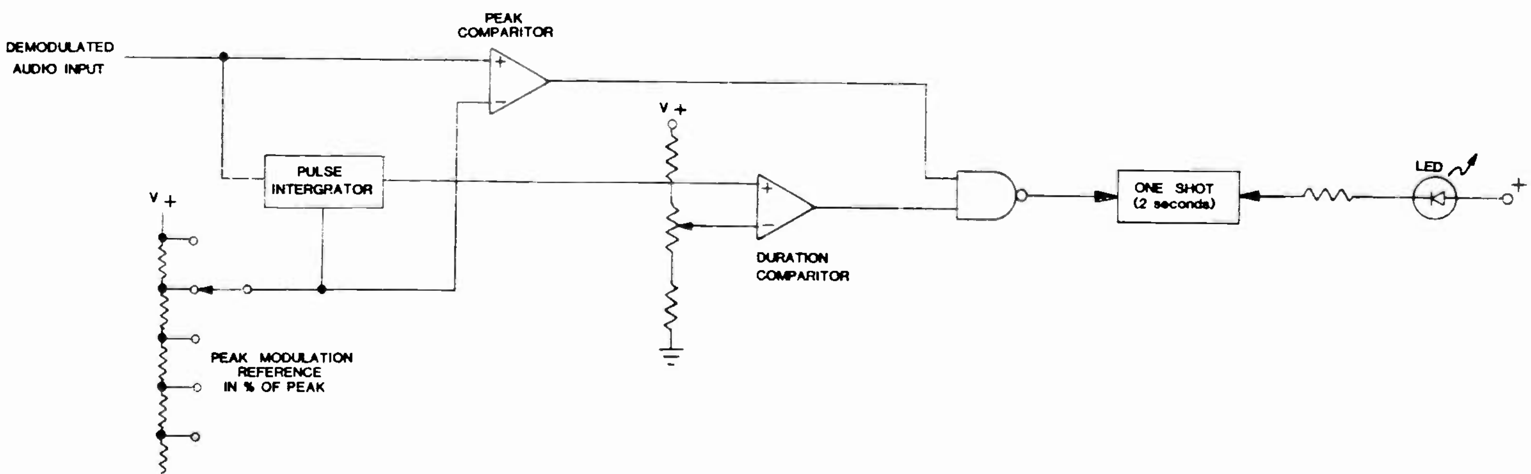


FIG 1A, PEAK MODULATION DURATION DIFFERENTIATOR

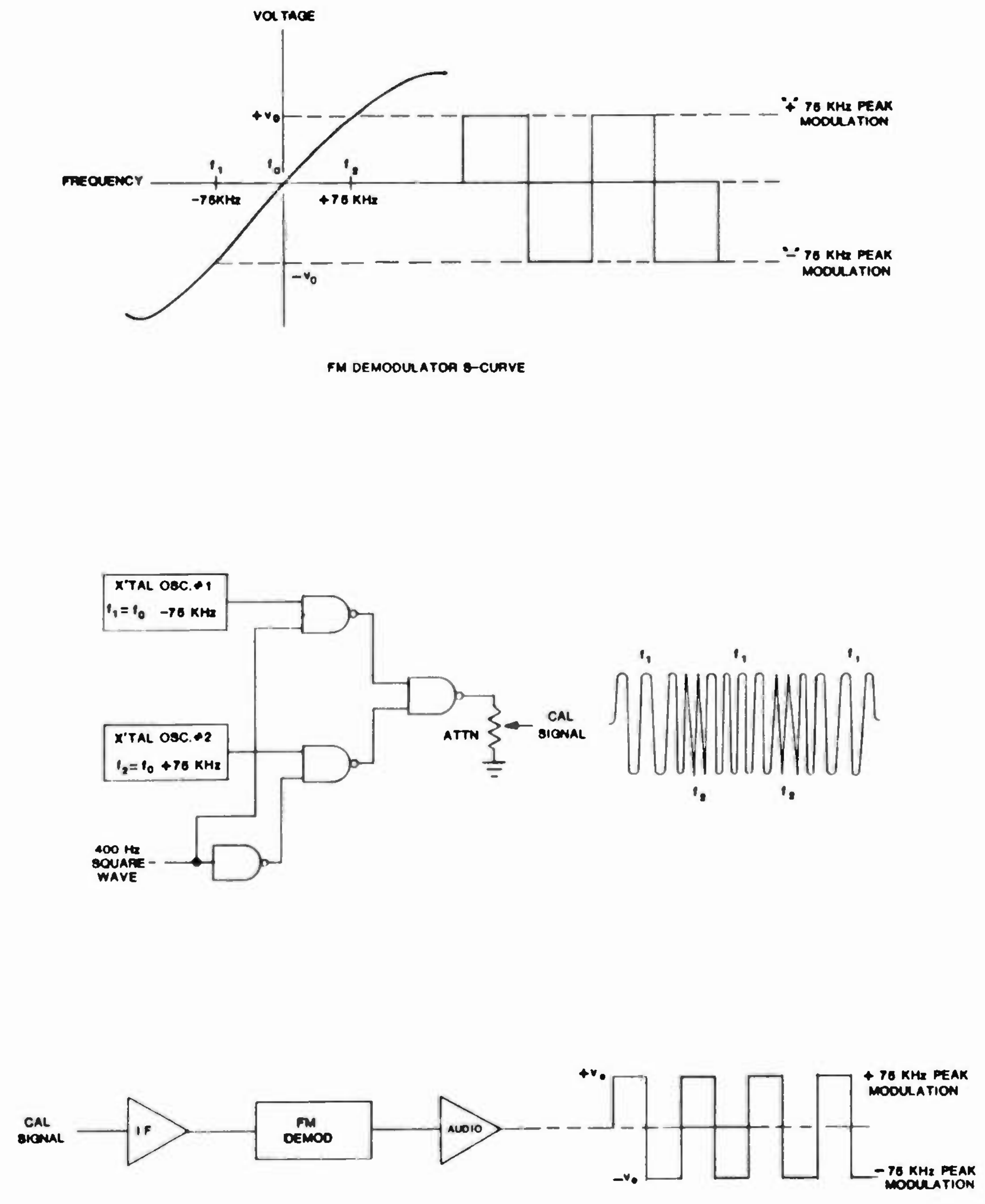


FIG 1B, FM CALIBRATION USING FREQUENCY SYNTHESIS METHOD

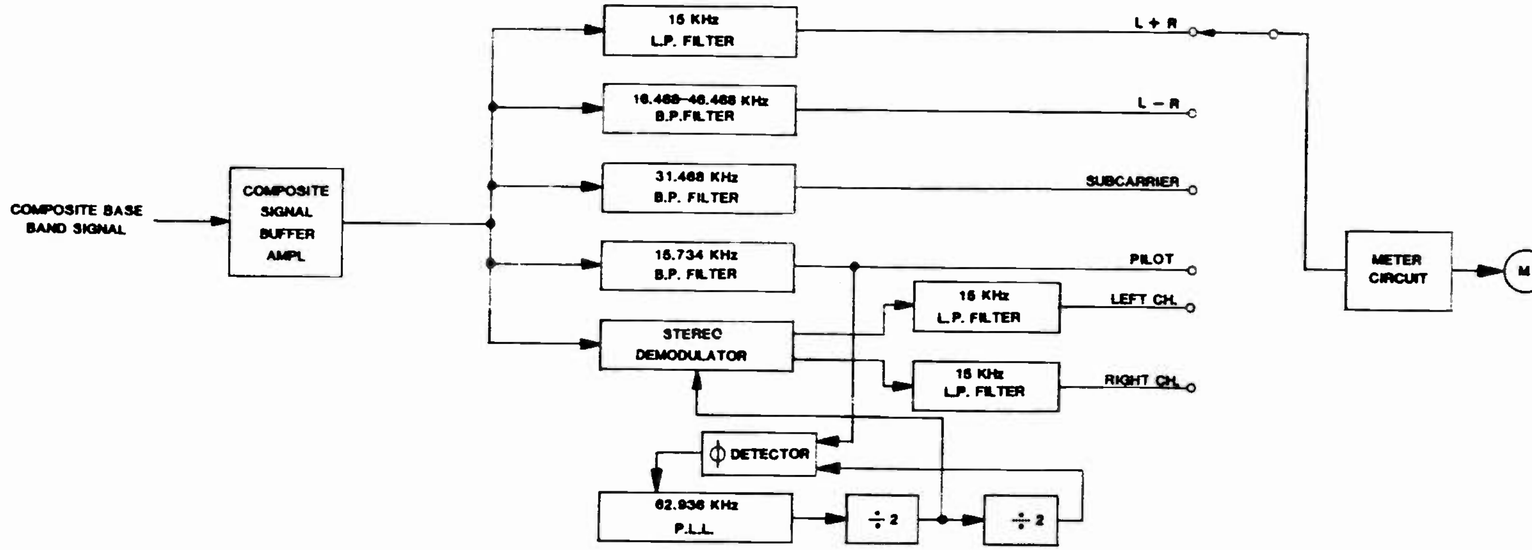


FIG 2, A SIMPLIFIED BLOCK DIAGRAM FOR TV/MCS STEREO DEMODULATOR

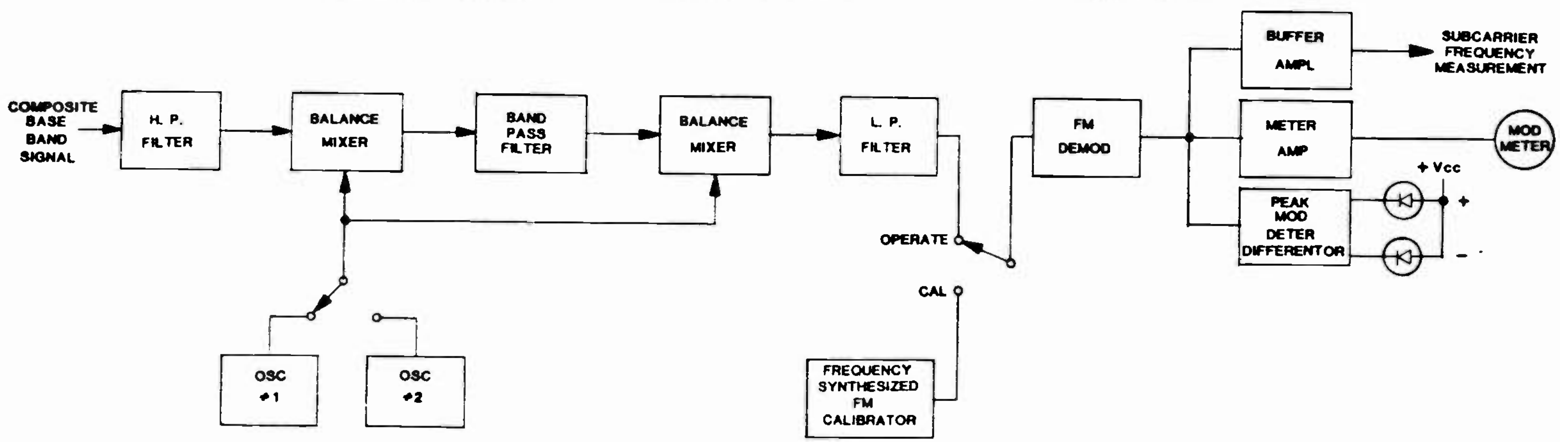


FIG 3, A SIMPLIFIED BLOCK DIAGRAM FOR TV/MCS SUBCARRIER DEMODULATION

Montage Picture Processing: An Interactive Production Tool for Producing Television and Motion Picture Programming

Ron Barker

Montage Computer Corporation

West Concord, Massachusetts

Both film and tape are accepted means of creative temporal communication. They are in fact neither.

When a sequential series of static frames are presented on a single screen, the temporal message exists only in its substantially completed form. During editing therefore, each picture clip is perceived by the editor only in its own non-temporal context. The ultimate impact in the eye of the viewing audience has of necessity to exist only in the mind's eye of the editor or director and may not be experienced by either until the editing process is completed. Consider a painter confined to a single color at a time. Like today's editor, he would specify his need alphanumerically, go to his storage room, locate say 'Cyan 12', return to his palette and only then be able to observe the actual color in the context of his work so far. Not only is serial access via alphanumeric tags cumbersome, it is entirely contrary to the instinctive reactions so necessary for creative expression. Most communication means, including the writing of this transcript for example, suffer from equally complex serial non-temporal limitations. The crucial ability to browse and iteratively rearrange in real time, however, saves the day. Most of us "know it's right when we see it's right", so we just keep at it until the desired impact is achieved.

The crux of the film and tape editing bottleneck is a complete inability to browse or to iterate. Montage Computer Corporation launched a development program early in 1982 to resolve these traditional editorial limitations which culminated this past February with production delivery of its MONTAGE PICTURE PROCESSOR (tm).

The ability to browse is achieved by instantly exposing all incoming video into more than 6000 Picture Labels, having a 120 by 100 pixel resolution and 16 levels of grey. Representing interactively the beginning and end of any picture or sound clip or, when required, the sequential contiguous series of frames around the head and tail of such clips, these labels are the primary communicating means for Montage editors. They are presented on 14 black and white monitors arranged in a horizontal row of vertical pairs.

The Montage editor may smoothly pan all Picture labeled clips through these displays using two control wheels and may point to any such clip and execute any of 20 Picture Processing commands including: TRIM, PULL, INSERT, DISCARD, COPY, MARK, RETURN, HOME, PLAY, SPLICE, DISSOLVE, WIPE, CUT, JOIN and SPLIT. Any command may be executed at any time, on any clip and in any sequence.

The Montage provides a total of seven operating bins arranged to appear in vertical smooth scroll sequence including dedicated DISCARD, SOURCE, and PULL bins plus four WORK BINS. Material may be moved throughout these bins at any time and in any sequence with the bin selection, like clip pointing, being achieved with dual seven position levers. Each clip may have its sound volume and routing controlled and edits can be layered to include three independent tracks, one picture and two sound.

The overall effect of these proprietary interactive display and control means is to provide a spatial representation of temporal content using both peripheral vision and the human ability to assimilate information from multiple images to enhance the perceived information in any one image.

The resulting ability to browse and act is essentially both complete and immediate.

The ability to iterate is achieved by simultaneously storing all incoming video (up to five hours) on multiple tape drives (7 to 14 Beta Hi Fi Format). Conventional information management techniques are then applied via a multitasking computer system (Motorola 68000 and up to 22 Z80's), to play back any source clip in any sequence at any time and with any requested transition effect.

The Montage editor is thus automatically and continuously presented with a full temporal presentation of picture and sound as represented by his arrangement of Picture Labels. This continuous random access, full look ahead, auto-assembly playback feature provides for essentially immediate and unlimited editorial iterations.

When combined, these two proprietary developments allow literally hundreds of editorial decisions to be made and seen per hour on a console that is learned much like a video game and which provides ergonomically satisfying tactile and accoustical feedback. The finished work may be recorded, distributed, removed for later use or a standard Edit Decision Disk for ON-LINE conformation may be generated.

All Montage EDL's are by definition clean as they are only computed after the finished program has been completed. Multiple vertical interval time code data is carried with all Montage Pictures for both system management and source time code preservation purposes.

Montage editors may continue their editorial decision making after leaving the system by using a third proprietary development which allows all Picture Labels to be printed with data as a hard copy Montage Storyboard. 600 picture clips representing approximately one hour of programming may be printed each 20 minutes for paper editing, library purposes or, in combination with a built-in electronic greasepencil, to plan and manage the pre-production and production process.

The MONTAGE PICTURE PROCESSOR (tm) thus allows a director to become involved in post-production as soon as he has his script, thus substantially increasing his ability to produce a creative temporal communication.

Although the first manifestation of these developments are now being delivered as completely integrated turn-key packages, the techniques described apply equally well to all video storage media from consumer 8 mm to Broadcast 'C' format, CAV laserdisc playback to DRAW disc and digital or high definition technology.

MIRAGE - THE PRODUCTION OF AN ILLUSION

Richard J. Taylor O.B.E., M.Sc., C.Eng., F.I.E.E.

Quantel Limited

Newbury, Berkshire, England

Introduction

From the early days of their involvement with video effects, Quantel has wished to design a machine which would free the creative user from the traditional constraints of rectilinear manipulations of the image.

The dream of giving the creative designer the freedom to choose whatever shape and form suited him, rather than accept the fixed parameters chiselled in silicon by the engineers, had for so long seemed so elusive that when the required result was finally achieved the equipment acquired the affectionate title "Mirage".

This paper gives the background to the development of Mirage, describes the control mechanism, and introduces a younger sister of Mirage - the Mirage Macro.

The Design Aim

The design aim was to make a machine in which each picture point was individually addressable and whose circuitry could cope with multiple address folds. These two criteria would in principle allow classic effects like a page turn, shown as the genuine article, to be undertaken. The individual addressability creates the curve as the page turns and the overlap circuitry copes with the fold back of the turn.

Similar arguments apply to shapes such as globes, cylinders, cubes or really whatever the creative designer wishes to do with the machine, and Figure 1 shows examples of what is achievable with Mirage. On the printed page only static shots are possible but remember that not only are the effects dynamic and, therefore, move in their own right but also real moving video can be on all of the surfaces.

In terms of the video processing circuits themselves it will be clear that

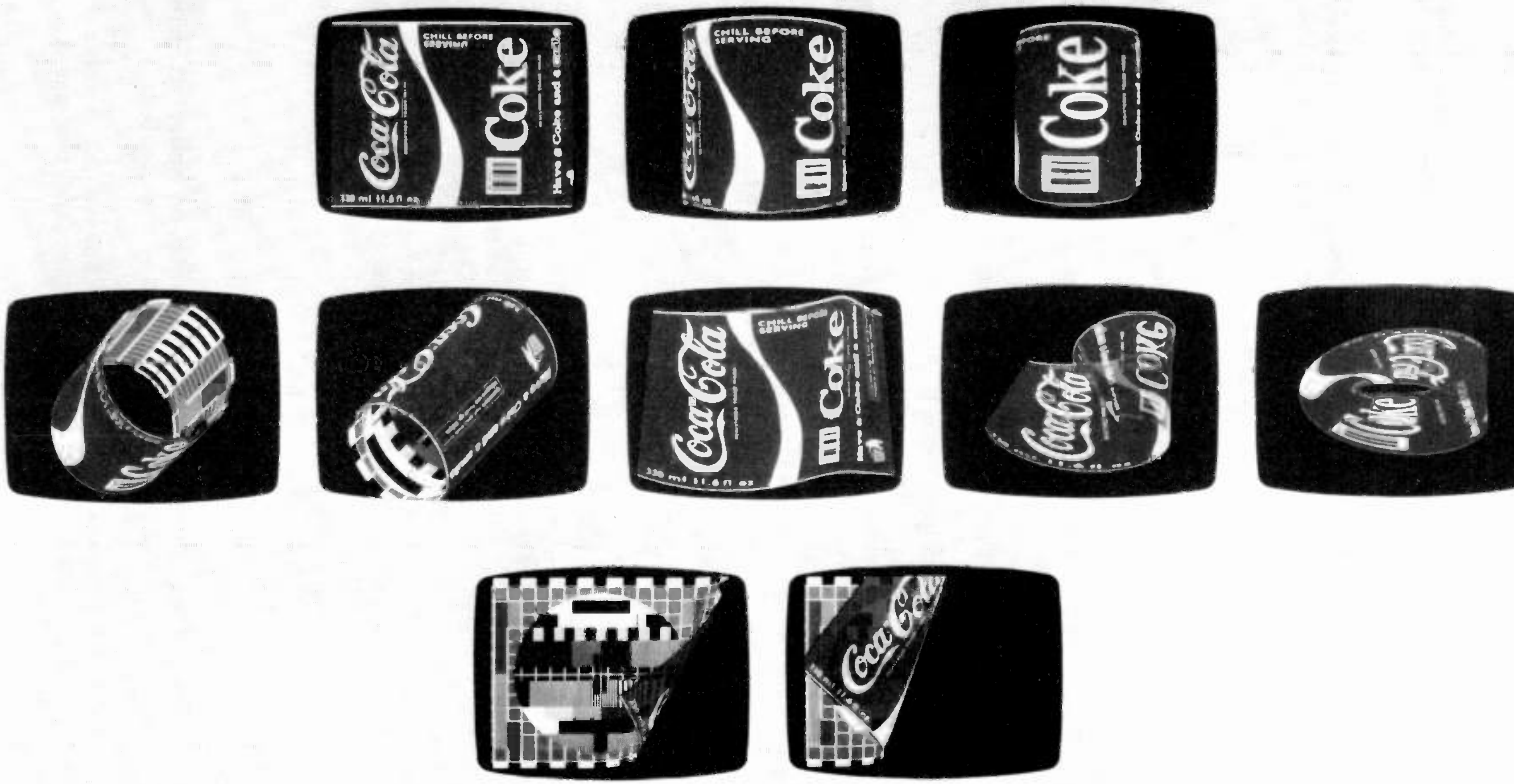


Figure 1. Examples of Mirage manipulation.

the free-form requirement of the original design aim can be met. Each point in the picture can be manipulated freely without any constraint on the adjacent points. Interpolation circuitry tracks the addressing of the points to ensure at all times the quality of the output picture is equal to that achievable with a top quality camera lens - except of course you can't do that with a lens anyway!

Automatic circuits permit the picture to be changed at the exact fold back point giving the realism of a front and a back or an inside and outside to the picture. Further, the machine can mix the front with the back giving the illusion of transparency.

Clearly, the address mechanism has to be special for no longer do the horizontal and vertical addresses simply increment. In conventional effects equipment the address mechanism merely comprises counters advanced by the appropriate picture point or line clocks. This is even true of rotation or perspective equipment where the relationship between the various addresses remains consistent and predetermined.

In Mirage, of course, there is no predetermined relationship between addresses. Indeed, in an explosion effect the addresses take up the relationship of a random number generator!

In order to overcome this apparent difficulty, the address circuits actually become an address store - equal in size to the amount of data required to store the picture itself. The Mirage computer deriving the shapes then loads this store at its leisure (by Quantel standards) and while the shape is created on the screen the picture point and line counters, instead of addressing the picture store directly, drive the separate address store which in turns accesses the picture memory.

The Hardware

A picture of the Mirage hardware is shown in Figure 2.

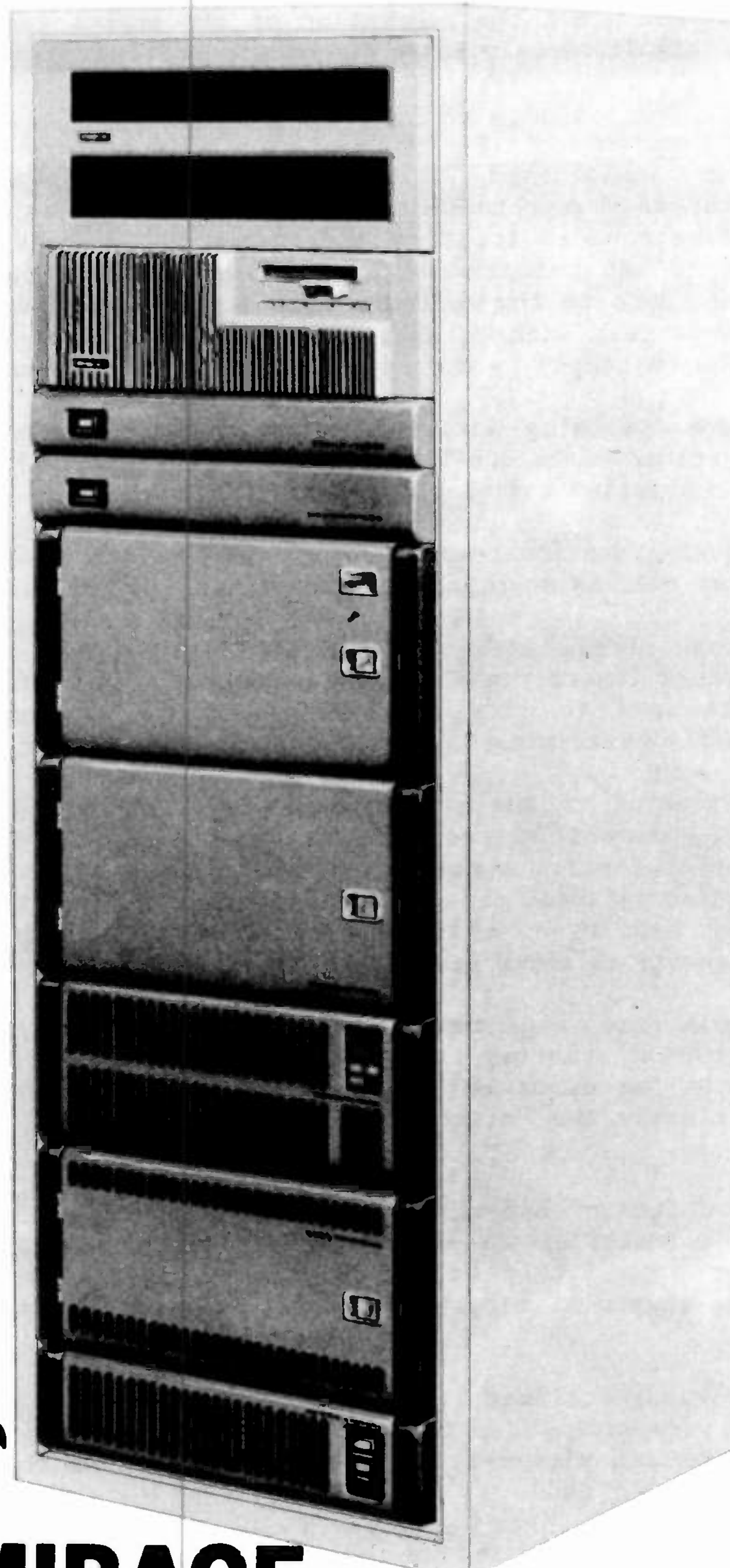
The system breaks into convenient sections. The top two boxes comprise the computer that drives Mirage together with the streaming tape cartridge for transporting effects. Below that, in the double fronted 2U high units, are the digital decoders and encoders. These are state of the art units leaving the minimum possible footprint on the video. The two main video processing racks are in the middle of the system immediately above the grille unit that houses the main power supply as well as a cooling air plenum for the video rack. The address generation circuits are below the main PSU with a disc for storing effects at the very bottom of the unit.

The Mirage Computer

The start of the shape generation process is a powerful mini computer optimised for arithmetic computation at high speed - the Hewlett Packard A700.

The A700 is able to prepare the necessary shapes at a remarkable speed providing good interaction between the shape designer and the machine.

An additional advantage of the A700 is that its time-shared operating system allows the computer to handle the control of the main Mirage as well as



MIRAGE

FIGURE 2 MIRAGE HARDWARE

simultaneously permitting the creation of new shapes and effects at a second composing station. Thus the generation of new shapes can be undertaken as an off-line task simultaneously with the on-air control system driving the Mirage on-line.

The Composing Station

One of the many aspects that makes Mirage unique is that it can have new shapes and effects added to it by the customer. In previous effects devices, major changes to add rotation or perspective or the ripple effects etc., all required alterations to the software source program as well as, in many cases, additional hardware. With Mirage, the users themselves can add entirely new effects and shapes simply by expressing the correct mathematical model.

This shape composing activity is not intended as an on-air routine but rather the operator works off-line returning to the main Mirage only when he is happy with his creation and wishes to try it for real.

The composing station comprises just a video display unit connected to the Mirage computer and, as such, can be some distance from the main Mirage machine.

The purpose of the screen on the VDU is two-fold. First, it is used to allow the various instructions to be entered in conjunction with the keyboard; second, it is used to view the results of the instructions as wire-frame drawings actually performing the effects, fast, but not in real time.

The application of the composing station can be split into two distinct areas using pre-prepared shapes or effects, or designing new ones from scratch. The first type of control and decision-making process for the composing station is when Quantel-originated effects are being used. Here the routines have been written so that even an unfamiliar operator can easily modify the effects simply by answering questions asked by the machine.

For example, the "page turn" routines ask questions like "start angle?" (a page can be turned starting at any corner or side); "finishing angle?" (the turned page can end up at any edge of the original picture); and "radius of curvature?" (clearly the "stiffness" of the page effects the tightness of the roll).

The second form of entering shapes or effects into the machine is to work out the solid geometry of the shape or effect required using standard geometric formulae. For this little or no programming experience is required but an ability to see shapes as a combination of classical solid geometric objects is helpful.

One of the aspects that leads to tremendous realism in the Mirage effects is that true perspective can be added, not only for the simple case of the spinning or tumbling picture but for all objects like the case of a cylinder being tipped end over end.

When looking exactly at one side of an upright cylinder little or no perspective is evident. As the cylinder tips the "straight" sides start to taper so that the end coming towards the viewer is bigger than the end going away. When the cylinder is exactly on end, surprisingly there should still be a picture visible since the near end will be a circle of larger diameter than the far end, and thus the inside of the cylinder should be shown heavily perspectived in between.

The mathematics of such perspective transformations can be cumbersome and time consuming to work out. Since providing the "correct" perspective adds so much to the illusion being created, special routines are included in the composing station software that allows the user to add perspective automatically as a second stage after the simple three dimensional object has been created. The routine allows full control of the viewing and vanishing points so giving the appearance of the eye moving all around the object in question. As is described later, Mirage now has an option of real time floating viewpoint control allowing the movement of this viewpoint at will by the on-air operator.

Once generated, effects and shapes may be stored on the micro floppy discs for transfer either to another machine or for re-use later in the same one.

Remember, at the composing station the operator can see all the effects he creates working in non-real time as wire-frame drawings but nevertheless fully performing the effects. Of course, if the effect is to be tried on a remote Mirage, then the tape cartridge merely has to be plugged into the Mirage to see the full real effect with real video.

The composing station also allows control of the front-back or inside-outside switch for Mirage. As was described earlier, the machine is able to handle two video inputs: one to be placed on the front or outside surface of the picture and one to be placed on the back or inside surface. Normally, control is automatic since the hardware is aware when a "fold back" has taken place but if for additional special effects the switch is required elsewhere, then this parameter may be added to the routine at this stage.

Not only can shapes be entered at the composing station, for example cylinders, globes, boxes or more complex objects, but also transitions can be entered. That is to say, the way the picture changes from a flat image into a cylinder can also be entered as part of the routine. In the absence of this information being programmed the hardware will automatically cope with a transition but if a more spectacular result is required it must be entered at this time. An example is again the cylinder, if left to it the hardware will fold back the two sides of the picture until a cylinder is achieved. The creative designer may choose to do something different and, in fact, if the left-hand side of the picture is made to set off twisting clockwise and the right-hand side anti-clockwise, a quite delightful transition results.

It may seem irreverent to say but nevertheless, it is true that some of the most interesting shapes and transitions come quite by mistake. The odd sign wrong in a routine can in fact turn out far better than the actual original thought, so the moral at the composing station is always observe one's mistakes before discarding them!

By now it will be quite clear why the composing station control mechanism is not intended as an on-air device but rather more for an activity to be undertaken in the peace and quiet away from the studio or edit room. After each alteration to a routine the programme has to be re-compiled by the computer operating system and this can take anything from a few tens of seconds for simple shapes, to a few minutes for very complex effects and shapes.

The On-Air Mechanism

If the composing station is an off-line mechanism then the on-air control mechanism is, exactly as its name implies, most definitely an on-line mechanism.

The on-air control system was designed for the edit suite or studio and can fully realise the facilities of Mirage in the on-air environment.

The system can be fully integrated with any switcher and editor, but also has mixing and editing functions of its own.

The secret of the on-air mechanism is that it allows a totally transparent co-existence of a lever arm for control of effects transitions with automated numeric control of transitions.

The standard control mechanism is shown in Figure 3 and it will be seen that it is dominated by the lever arm and two joysticks.

The power of the A700 computer is brought to bear on the on-air control as well as the composing station for it does allow this unique facility of having manual and automatic control intermixed at will.

Each shape or effect has a number assigned to it when entered via the composing station. If a shape has been called up by its number on the on-air keypad, then movement of the lever arm will cause the picture to ooze to that shape, in proportion to the movement of the lever. Alternatively, if an automatic transition is selected then the picture will ooze linearly or along a profiled acceleration-deacceleration track according to the time and style set in the automation mode. The interesting feature comes not so much that one or the other is available but that it is possible to have both. If the shape is appearing under automation and say it is going too slow then the lever can be moved, the effect caught up with and the manual control override the automation and complete the effect. The converse case of automation taking over from manual control is also possible. It should be emphasised that the overriding takes place without any glitches.

If an effect rather than a shape was called up then the movement of the arm will cause the effect to take place, again in direct proportion. So in the case of the page turn, moving the arm will cause the page to turn, waggling the arm will cause the page to waggle. Returning to the example of the spectacular cylinder roll-up, the movement of the lever will cause the effect as created at the composing station to take place. Selection of the buttons near the lever arm can either reverse the effect or go on to the next with subsequent movements of the lever. In fact whole strings of effects or shapes can be assembled and replayed with successive movements of the lever arm. Again intermixing of manual and automatic control is allowed.

In the classic effects equipment style, live control of the size and position of an effect or shape is provided by means of the size and position joysticks and, of course, control of the aspect ratio is also available.

It should be noted that at the composing station the mathematics allows the picture to exist in virtual space, i.e that off the edge of the screen. Using the page turn example again, if the effect is done at full size as the page turns beyond 90°, that portion of the picture will disappear off the edge.

However, because of virtual space, if that effect is reduced in size by means of the joystick then of course the entire page will be seen as it moves beyond 90°.

In addition to the automatic or pre-set control of the front-back or inside-outside switch, the on-air control mechanism offers real time control of the mix fractions during front-back overlap. This gives the appearance of controllable transparency of video.

An extension of this facility is to allow Mirage to accept an input "soft" or "linear" key and then use it to isolate the keyed subject prior to processing. In this way objects can be made to appear on the surface of globes, cylinders, boxes etc. without the rest of the rectangular picture being present or visible.

Being able to handle a soft key will, of course, provide the highest quality at the edge of the keyed subject. Control of the front-back mix effect busses as well as input key facilities are all achieved from the "state" portion of the control panel.

Floating Viewpoint Control

Flexibility is the raison d'etre of Mirage, so it was ironic that until NAB 1984 the one thing that the creative designer wished to be able to change in real time could only be done off-line at the composing station. The mathematics of changing the viewpoint of the viewer when looking at a three dimensional object are complex as was explained in an earlier section. Mirage did have routines for solving the necessary equations but only slowly off-line at the composing station and yet the real requirement was to be able to do it on-line in the control room when with the client.

Floating Viewpoint Control does allow this ultimate step in flexibility. The complex routines previously nestled in the HP have now been translated into hardware which allows the operator to wander around the object (or shape) at will by means of a tracker-ball. The technique has been seen before with computer aided design systems but, of course, with Mirage it is real moving video that is being manipulated!

An example of the control mechanism would be the tumbling Coke can. Previously, once the axis of the tumble had been chosen at the composing station it could not be changed without recourse to the composing station. Now, Floating Viewpoint Control allows the tumble axis to be at any orientation giving the illusion of the can floating in space.

Of course, the Mirage automation system works as before with Floating Viewpoint allowing sequences of effects and moves to be put together in strings.

Mirage Macro

The advent of Floating Viewpoint Control has moved the "run time" operation of Mirage from the HP computer to the new control mechanism. The computer is still needed to compute the shapes in the first phase but once the shape in question has been entered onto the Mirage Library Disc it takes no further part in the proceedings.

The Mirage Macro is a new machine in the Mirage range that has a removable disc in the library but no HP computer. Thus, new shapes can be entered by means of plugging in a new disc cartridge that has been created by Quantel or on a fully optional Mirage with a composing station. Once the shapes have been "plugged in" the on-air manipulative power of the Mirage Macro is just as powerful as the fully equipped Floating Viewpoint machine. The new machine is designed to be used by operators not interested in the computation of new shapes but rather the on-air manipulative power of a portfolio of shapes stored in the library. Thus, the Mirage Macro becomes the classical digital effects machine par excellence with all the range of effects at the operator's disposal, without the extra burden of the computer keyboard.

Conclusion

Mirage has proven in use to be a dramatic introduction to a new generation of image manipulation machines that are changing the look on the air.

It is hoped that the reader will feel that the original design criteria for the machine have been met and indeed, the freedom has been given to fully realise Mirage as the Production of an Illusion!

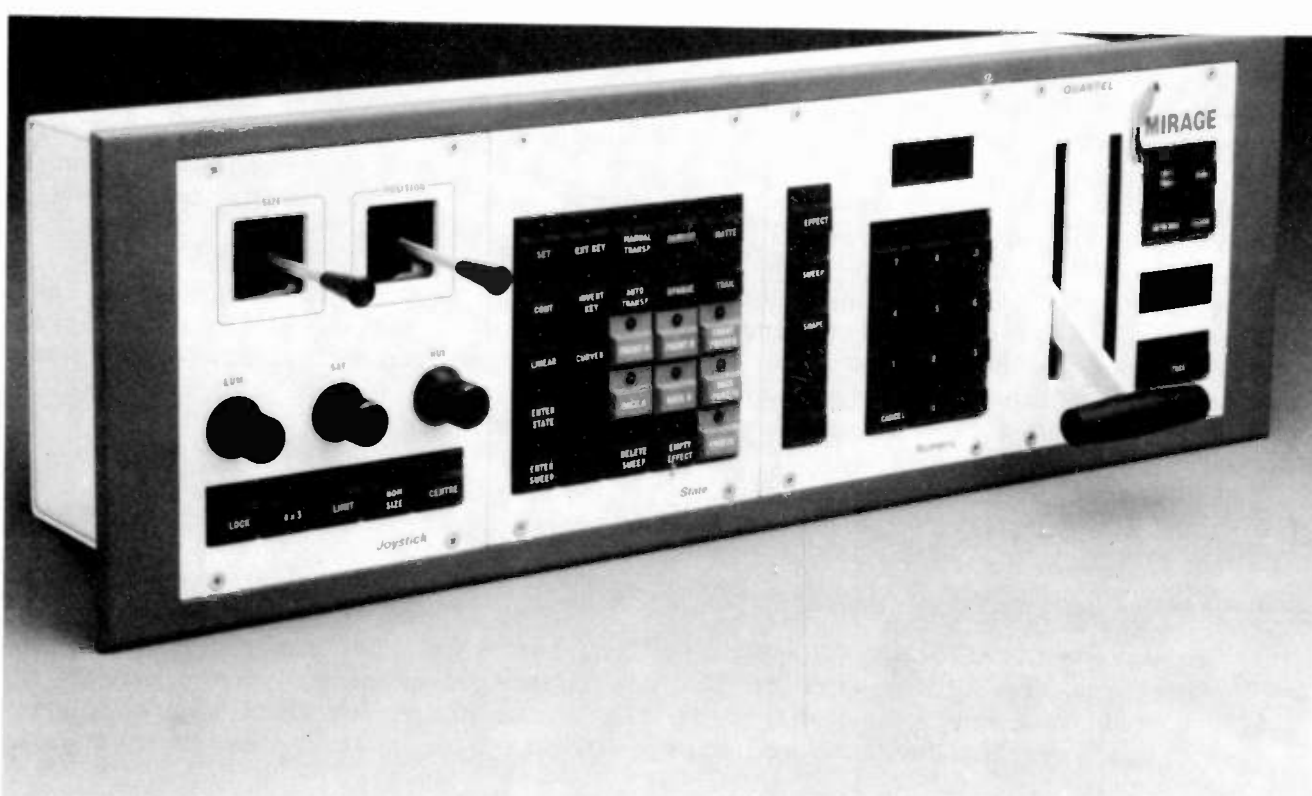


FIGURE 3 STANDARD CONTROL SYSTEM

ELECTRONIC GRAPHICS PRODUCTION - STILL STORE USAGE IN TELEVISION

FRANK ALIOTO

DIRECTOR OF BROADCAST SALES

ADDA CORPORATION
LOS GATOS, CALIFORNIA

When considering still storage in television it is useful to divide the discussions into two distinct parts:

1. What still store devices are, how they were viewed when first introduced to television and what they have become.

2. The technical side of a still store and how it is interfaced into a television station's overall facility.

---

1. Very simply, a still store takes a frame of video from any source, freezes it and stores it on a hard computer disk for instant retrieval. The idea of storing frames of video in computer disks has been around now for about ten years. It is an idea that grew out of the engineering revolution that began sweeping the world in the early 70's.

Since video tape was replacing 16mm film during newscasts, wouldn't it be great, electronics manufacturers reasoned, to further eliminate film by doing away with slides and thereby eliminate all the processing costs and processing delays associated with slides.

Not only would costs be cut and time saved, but there would be other benefits to the television station as well. For example, the television station's on-air look would be improved.

No longer would one need to worry about an important news or commercial slide being aired dirty, cracked, or upside down in the gate. Since storage was digital, colors would play back true and there would never be much concern about a blue color appearing on air as a red or green. Since storage was digital and the playback heads would never touch the recording medium, stills could be played back an infinite number of times without fear of the picture quality degrading.

Now that the still stores have become widely used, we take those assets mentioned above for granted, but just seven to ten years ago, the thought that a machine existed which could do those things was a rather revolutionary thought indeed.

After much consideration and research, the first electronic still store was brought to market by Ampex in 1976-1977. The ESS as it was called could not only freeze frames of video but could also do real time animation. ADDA Corporation introduced their own ESP to the market shortly thereafter. It also was capable of storing freeze frames and completed art on a computer disc drive but could not perform real time animation.

Both devices, although possessing production potential, were primarily viewed as optical slide chain replacements. As a result, when selling a still store to a TV station, the salesperson would simply seek out the station's chief engineer to see if he had an optical slide chain he wished to obsolete and replace it with a still store.

In addition to the rationale for still store usage discussed earlier, we at ADDA could also delineate quite clearly how a $65,000 ESP could be used as an optical slide chain replacement and pay for itself as a capital expenditure in two and one-half years or less. We would do studies of the number of stills used by the potential customer each day, the number of art cards used during the newscast and commercial productions, even the cost of maintenance on the studio camera used to shoot the art card during a commercial session was tossed in for good measure. In a nutshell, the argument boiled down to "We can clean up your on-air look and save you money. What else can you hope for in a great engineering tool?"

The idea of still store as optical slide chain replacement and money saving engineering tool was a good one from the start and we sold a large number of still stores that way.

About 3½ years ago however, the perception of still store as optical slide chain replacement began to change rapidly. It began to change when ADDA introduced its ESP-C, a still store system that has two key advantages over previous generation still stores.

First of all the system is virtually transparent. The system translates regular video to digital signals with no apparent signal to noise or bandwidth degradation.

Secondly, the machine possesses the ability to refreeze images in its buffer memories. Consequently, elements can be added to a picture, deleted or changed many times in a buffer memory before any image is ever committed to the computer disk for storage. This ability to add elements to a frame is called a multi-generation re-record ability. This multi-generation re-record is in effect a digital sandwich or multi-layered digital canvas, upon which the artist can place layer upon layer of digital paint. When using an ESP, an artist can go down twenty generations or layers in creating his images for new and commercial production.

As soon as multi-generation machines began delivery, artists, production and news types all found that they could create better graphics on the still store than they could have ever created with slide and art cards. Not only that, but art cards which previously took two to three hours to create could now be created in five minutes or less. As a result, the still store quickly grew into a broad based production tool ... so much that at least 90% of the still stores sold today are sold as graphics and production systems. Some are still sold as optical slide chain replacements, but seldom is this the primary reason for purchase.

2. When it was an optical slide chain replacement, typical configuration of the system was to have a mainframe (analog/digital processor) in the equipment rack with a couple of disk drives nearby. A record/recall panel was generally located in the master control area and a second record/recall panel was located in a production booth.

Today, the configurations vary as widely as the number of users. Figures a,b and c illustrate in block diagram from three of the more typical installation schemes with a and b most commonly used when building graphics.

When building graphics with a still store, the artist is not limited in his choice of video sources. The Series C Electronic Still Processor (ESP) can capture and store stills from any stable direct color (not interlaced) or monochrome video source. It stores stills on computer type magnetic discs that provide immediate on-line access, as well as various amounts of off-line storage. During acquisition, storage, and retrieval operations there are two independent video channels in the ESP system. The ESP system can receive only one video input (into either channel), but at all times outputs two independent video signals. A still stored on disc can be recalled into either channel. The two channels present various options to the operator. For example, one still can be placed on the operator designated on-air channel, while the other channel is used to preview upcoming stills from the disc, or the alternate channel can simultaneously receive video input to the system for still storage.

A standard ESP system consists of an analog/digital processor chassis, a master control panel, and two disc drives. The system can be expanded to include up to four disc drives and up to fifteen slave control panels (a system may also have only one disc drive). All video signals input and output from the ESP system connect to a connector panel on back of the analog/digital processor. Likewise, the disc drives and the master control panel also connect to the back of the analog/digital processor.

The analog/digital processor, commonly referred to as the mainframe, contains all video input and output processing circuits, two separate semiconductor memories (the frame memories), each of which can buffer store one video still, the system microprocessor and its associated operating program, as well as a separate microprocessor based disc drive controller. The mainframe is an enclosed rack mountable chassis, that contains a 41 slot circuit card cage, power supplies, and two cooling fans.

The master and slave control panels, referred to as the remote control units, are physically identical. Each is a rack-mountable unit, consisting of a front panel mounted onto a chassis. All operating controls (pushbuttons and indicator lights) and a message display panel (the prompter panel), and two LED displays are on the front panel. Mounted to the back of the panel are several circuit boards that interact with the switches and drive the displays. Near the back of the chassis is a microprocessor control circuit board (the micro remote board), complete with a program memory, and power supply. There is no cooling fan within the chassis - the remote control unit depends on the cooling provisions of the rack in which it is mounted.

The order in which the ESP system performs its basic function during a typical operating situation is as follows:

1. Incoming video is previewed on the preview channel of the ESP.

2. A frame of video is frozen in the preview channel frame memory (it becomes a still).

3. The still is viewed by the operator, evaluated, and stored on a disc.

4. Later the still is recalled from the disc and output from the ESP as video on the operator designated channel.

Once stills are stored on a disc, it is also possible to edit the disc (add or delete stills), define sequences of stills, and then recall them as a series. If an ESP system is equipped with the multipix option stills can be recalled in a compressed form and selectively positioned on the screen. Futhermore, the ESP system can perform various auxiliary functions, such as scanning through the contents of an entire disc pack onto another, format disc packs, and so on.

The incoming video is input into the mainframe, converted into digital form and stored into one of the two frame memories a frame at a time (the operator selects one memory, that designates it as the preview memory). The incoming video is written into the preview memory continuously, but the writing can be stopped on operator command, whereupon the last frame is retained (frozen). The operator can also at any time select the alternate memory as the preview memory.

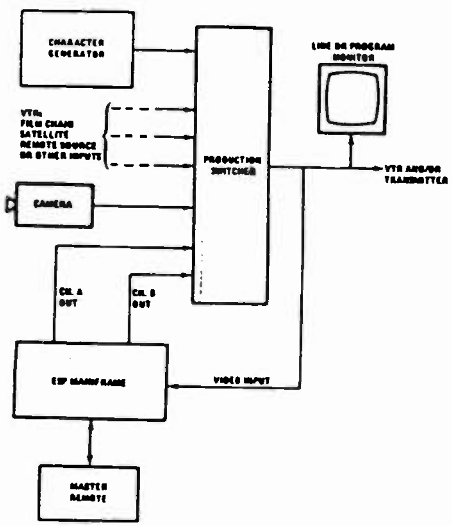
At all times the contents of both frame memories are continuously converted back into analog video, integrated with an internally generated composite sync signal, and output to the back panel output connectors of the ESP system, synchronous with the local sync reference signal. The contents of the frame memories can also be stored onto the magnetic disc surface in the disc drives. When a still is recalled from the disc, it again is written into the preview memory, converted to analog video, and output from the ESP.

The remote control units are the operator interface to the ESP system. The operator enters commands on the remote control front panel push-buttons, the commands are received by the remote control interface program, and interfaced to the mainframe operating program. The remote control interface program also receives from the mainframe messages for the operator, that indicate the status of the system or request the operator input data. If there are multiple remote control units, only one can be in control of the system at one time. System control may be assumed by any remote unit, although the master remote can always retrieve the control.

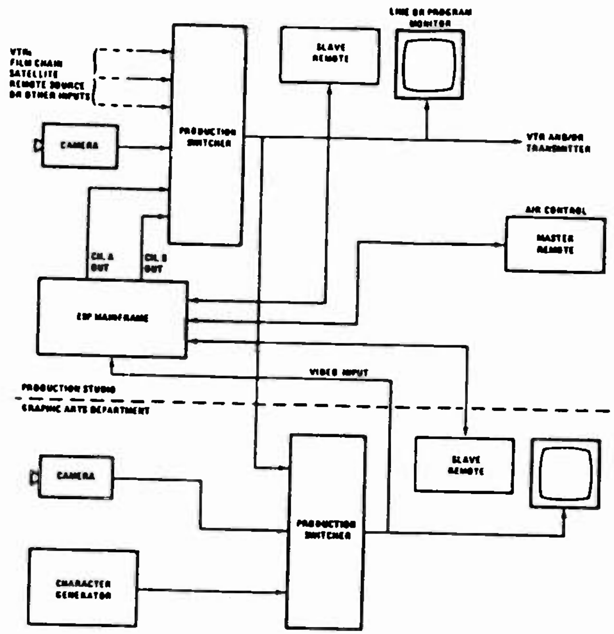
The Series C Electronic Still Processor is designed for broadcast, industrial, and institutional applications. For broadcasting it is both a rapid access live program material source as well as a quick, flexible and innovative production tool. It serves as a still library from which program material can be variously processed by other equipment and assembled for on-air usage. It replaces 35 mm film/slide chains and expands the creative capabilities of graphic arts departments and program production studios.

As an example, for newscasts, it can be used in conjunction with a character generator to quickly create stills with tilting and graphics superimposed, and arrange them in sequences for recall at air time. Because sequences can be edited rapidly, last minute changes are possible. During coverage of sports events, the ESP can be a source library of portraits of athletes, or be used to capture, save and immediately recall stills of key play action for on-the-spot analysis. Likewise, for other programs the ESP can be used in conjunction with special effects equipment to assemble series of stills into dynamic visual displays, create animation, colors that change and highlight different areas of stills (for maps, charts, etc.), montages, and other similar effects. In each of the above examples stills can be recalled from the ESP individually or in sequences - in full screen size, or various compressed and selectively positioned formats.

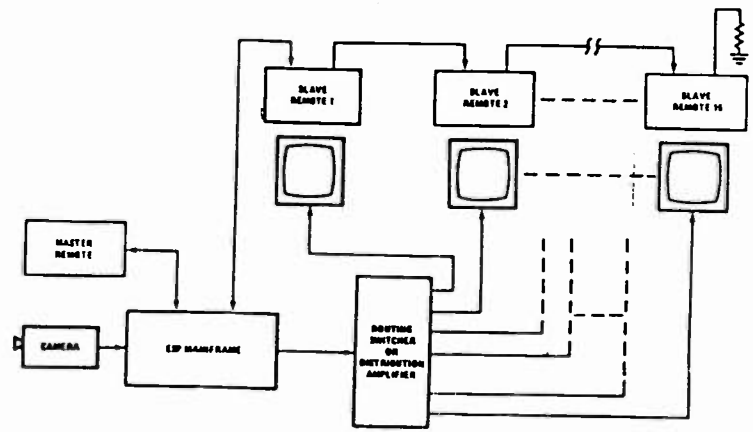
In summary, the ESP is a versatile creative working tool that is continuously finding new applications in the broadcast industry. In industrial and institutional applications the ESP can be used in closed circuit television systems for pictorial document storage, surveilance, training and other applications where a large volume of visual storage and fast access are required.



a. ESP System Dedicated to Production Studio.



b. ESP System Shared By Production Studio and Graphics Department.



c. Multiple Station Closed Circuit Installation.

Figure 1–Block Diagrams of Typical ESP Applications

A Digital Line Array Telecine

Ray Matchell

Marconi Communication Systems Limited

Chelmsford, Essex, England

When Marconi decided to proceed with a new telecine, they established certain basic design criteria. These were:-

1. It should provide the highest quality picture from all film gauges and on all TV Systems.
2. Its performance should be predictable and drift-free.
3. It should not require any of the routine line-up procedures demanded by other currently available telecines.
4. The film transport mechanism should be simple and require no maintenance other than cleaning.
5. Change of film gauge should be achievable in a few seconds.

In recent years there have been two main types of systems, one with camera tubes and another with the scanning performed by a CRT - the flying spot type. The camera tube type is very easily understood in that the camera "looks" at the film which could be in almost standard film projector. One point to note is that the film pull-down must be very nearly synchronized to the TV system. Failure to do this would give rise to application bars on the picture. This synchronization is quite easy for a 625/50 Hz system, but for 525/60 Hz we require a special projector which displays frames alternately for 2/60 sec and 3/60 seconds, i.e., the so called 2-3 action. The projector presents no real technical problems, but the camera system will always have the defects present in the tube/scan area, i.e., mis-registration, geometric errors and worst of all, lag. The high contrast ratios encountered in film make lag a more serious problem than in a live camera.

Turning now to the flying spot, the film motion is continuous and vertical scanning is provided both by the film motion and the raster on the CRT. In the latest systems a digital store is employed to enable the scan patch to be kept in a constant position and the film is read in one pass with the information being presented to the store in a sequential form. This signal is then converted to a conventional interlace form by manipulating the stored data.

This system eliminates the problems connected with variations of focus and shading in the field direction but it does not remove these problems in the line direction.

There is still a possible problem with phosphor grain and both camera tubes and CRTs are subject to degradation and are expendable, with re-registration and re-alignment being necessary on replacement.

Thus Marconi looked at and eventually decided upon a telecine employing line array CCDs. The line array requires none of the scanning and EHT paraphernalia associated with a high power CRT system and a number of drift sources and adjustments are therefore immediately eliminated. Further, linear sensor CCDs have now been developed to the point where they can produce pictures of the very highest quality. Their lives are virtually infinite and they require no electrical registration. In addition, Marconi decided to introduce a totally digital signal processing system to give predictable drift-free performance and to remove all pre-set and alignment controls from the signal processing circuits. Finally, the use of a capstan-driven continuous motion film transport provides the necessary mechanical simplicity for freedom from routine maintenance. The combination of these features has enabled the Marconi telecine to achieve all original design objectives.

THE PICTURE SENSING SYSTEM

The picture sensing devices used are Fairchild 133 second-generation line sensors. They have 1024 sensing elements plus eight other elements for black and white references. The photo-elements are each 13 micro inches wide. The second generation devices have higher sensitivity, better blue response, and lower dark current than the earlier version. The chip includes its own clock driver circuitry and is capable of operation at up to a 20 MHz data rate. The devices have a typical dynamic range of 2500:1.

On both sides of the line of sensors there are shift registers. The charge produced by light incident on the photo sensors is allowed to accumulate over the period of one television line. It is then transferred via the shift registers, to one of two output ports. Each output port receives the signals from alternate sensing elements and the two outputs are combined to produce a single TV line of information.

The film is imaged onto three sensors for red, green and blue via a conventional light splitter block. If the film is stationary, the output from any array will be the same on all lines and will correspond with that on the one line of film image which falls on the array. In order to obtain a picture from moving film, it is necessary to generate different lines of information to make up a complete television frame (two fields), during the time that one film frame takes to move through the film gate.

In the 625-line, 50 field/sec television system, the film is run at 25 frames/sec. At normal speed, one film frame is equal in time to two television fields. In 525 line, 60 field systems, the situation is slightly more complicated. The film is run at 24 frames/sec. and the passage of two film frames occupies the same time as five television fields.

The read-out of the arrays occupies the normal active TV line length, but to match the active number of lines per field to the active portion of the film frame, the "wait" period between lines is adjusted. The correct number of TV lines for each frame is accumulated, but has been obtained in sequential rather than interlaced form. In order to produce a correctly interlaced frame, the signals are first written into digital storage. All odd numbered lines are then read out in sequence during one television field, and then all even numbered lines are read out during the next television field. At this stage, the total line length is also corrected and a correctly interlaced television frame is produced. On the 525 line system, a 3,2,3 utilization of stored fields takes place in order to match two film frames to five television fields.

Sequential to interlace conversion could theoretically be performed before or after signal processing and on either R,G,B, or luminance and chroma signals.

In order to reduce storage requirements, the conversion is, in fact, performed after the main signal processing and on luminance and colour difference signals. 16K RAMs are used for storage, 240 being used for luminance and a further 240 for chroma. The chroma signals are stored at only half the bandwidth of the luminance.

THE FILM TRANSPORT SYSTEM (See Fig. 1)

The line array scanning system demands the use of a continuous motion film transport which must be locked to TV syncs when running at standard film speed. The design of the film transport was influenced by operational as well as technical considerations. The following are some of these:-

1. The spools should be at a comfortable height for loading.

2. The film gate should be near eye level and be vertical.

3. The hotter parts of the equipment, i.e., the video lamphouse should be at the top where hot air can be directly evacuated.

4. The mechanism should need no maintenance other than cleaning.

Film is taken from the feed spool, past the feed-spool reversal switch, which is designed to foul the film path if put into the wrong position, to the first compliance arm system. The compliance arms are connected to optical sensors behind the film deck, and control the film tension via the spooling servos.

The only sprocket in the system is placed close to the film gate and is driven by the film. An optical pulse generator, which produces framing pulses for each of the three film gauges, is attached to this sprocket. The film gate is curved to maintain lateral rigidity, and because sensing is along only a single line, there is no loss of peripheral resolution due to this curvature.

The capstan diameter is approximately 50 mm (2 ins) and there is no pinch roller. The film wrap is approximately 150 degrees for 16 mm and Super 8 mm and 190 degrees for 35 mm film. The capstan carries a 5000-line tachometer disc at the rear to generate pulses for the capstan servo. Optical and magnetic sound is picked off at the capstan, because at this point the film is under the greatest velocity control. Spool diameters up to 21 ins (approx 5000 ft.) can be used.

THE OPTICAL SYSTEM

Direct imaging of the film onto the arrays is obtained from the main projector lens. The lamphouse includes a field lens relay system to provide room for the necessary filters. These include infra-red filters, a variable neutral density disc for light control and a cyan filter for use with negative film. The main projector lens and the gate lens, which is the last component of the illumination system, are different for the three film gauges, and are mounted in the interchangeable film gate assembly.

The projection lenses for the three film gauges have been designed as a matched set to remove any need to adjust the registration after changing from one film gauge to another. Video and sound lamphouses both incorporate automatic lamp changeover in the event of failure, although in practice, we tend to disable this for post-production work. A 250 W Q.I. lamp is used for video and a 20 W Q.I. lamp for sound. Neutral density filters to balance the red, green and blue sensitivities are used at the exit ports of the colour-splitting block.

The arrays are rigidly mounted in mechanical assemblies that provide four degrees of freedom for initial registration, which is done when fitting the arrays. The adjustments are then locked and do not need further attention unless the arrays are removed. Thus we have eliminated the requirements of routine registration.

The three printed circuit boards associated with the arrays plug onto the arrays and can be removed without disturbing the registration.

VIDEO PROCESSSING

The line array presents an analogue output of a few hundred milli-volts and this must be converted to a digital signal of 11-bit accuracy at about 13 MHz clock rate. Once past this hurdle, the signal is in the safety of the precise digital system.

Full logarithmic masking and true power-law gamma correction are provided, being achieved mainly by the use of look-up tables. Needless to say, the resultant performance is completely predictable and absolutely stable. Thus, we have achieved another of our performance criteria.

Following gamma correction, the red, green and blue signals are formed into luminance and colour difference signals, the latter being filtered and then reduced to half the clock rate. The signal from the line array corresponds to a sequential scan of the film frame and it must be converted to interlaced form by use of a storage system. The store may be operated in various modes to allow for slow motion, still frame, alternative standards (625/525-line), etc.

After retrieval from the store, the luminance signal is passed through the horizontal and vertical aperture correction stages, which are a great improvement upon the analogue counterparts. Finally, the signals are converted back to analogue form for use in the internal colour encoder.

It is obvious that the system must be controlled digitally and this is done by the use of a microprocessor, which is also responsible for communciating with the projector system to obtain film gauge, speed details and other control information.

CHOICE OF CLOCK FREQUENCY

The choice of clock frequency was somewhat complicated by the fact that at the time of planning the design of the B3410 processing system there was no agreed standard sampling frequency for digital television studios. It was therefore, decided to choose a frequency convenient to the telecine system. However, there is now an approved standard for digital TV signals (CCIR recommendation 601) and future machines for use in all-digital studios will operate with the approved 13.5 MHz clock rate.

It is obvious that a line locked system is preferable as this enables the scanning clocks for the line array system to be directly locked to the main digital system clocks. Also the store design is simplified if the same pulses are used for the read and write clocks. The store system utilizes fairly slow 16K by 1 dynamic RAMs and these are multiplexed 15 ways. This led to the use of a clock frequency which was a multiple of 15 times line frequency. The final figure chosen was 825 times line frequency for 625 line systems and 810 times line frequency for 525 line systems. This gives a main clock frequency of approximately 12.9 MHz and with the line array clocks related by a factor of 1.5, 19.35 MHz in this case. A further variable relationship is applied to allow for the digitizing clocks used in the Cinemascope system.

ANALOGUE-TO-DIGITAL CONVERSION

For analogue-to-digital conversion the accuracy for digitizing coded signals is almost always 8 bits. However, when digitizing unprocessed, and in particular pre-gamma correction signals, a greater degree of accuracy is required. Calculation of gamma laws and the various contrast ratios and mask factors involved, indicated that 11-bit accuracy would be ideal for pre-gamma digitization. Unfortunately, there was no standard 11-bit flash converter of sufficient speed on the market and it was therefore necessary to adapt an 8-bit device to give the required performance.

Before analogue to digital conversion is performed it is important to pass the signal through an anti-alias filter, which should substantially attenuate frequency components higher than half the sampling frequency. The design of this filter demands a fairly critical design compromise. Too sharp a cut off results in ringing on sharp edges and too little attenuation of frequencies exceeding half the sampling rate results in aliasing or the production of spurious beat components.

Figures 2 and 3 show the block diagrams of the video system. There are three separate full bandwidth channels for red, green and blue. The A/D converter includes sin x/x correction to compensate for the 1024-element sampling of the line array devices. A similar form of correction for the digital sampling is included in the D/A converter filters. As is well known, the elements of the line array sensors can have variations in sensitivity giving rise to a form of line shading pattern. By storing data on an open gate this patterning and any lamphouse shading can be removed by the depattern amplifier.

Thus the system provides an excellent line shading performance and as the sensors are looking at only one line, there can be no vertical defects. Following depatterning the colour balance is adjusted in a multiplying D/A converter which is controlled from the microprocessor bus. Two 8-bit A/D converter ICs are employed to give digitization to an accuracy of 11 bits at low levels and 8 bits at higher levels. The changeover between the two data streams is controlled by monitoring the data from the low level digitizer and detecting the full range condition. Just prior to the actual A/D ICs there is an offset system to allow low contrast films to be digitized more accurately. The rate at which the signal is digitized depends upon the horizontal display mode. With normal non-widescreen prints the digitizer will be run at the same rate as the main system clocks which is 12.9 MHz but for full cinemascope expansion the rate will be about 22 MHz. The actual mode of operation is determined by the cinemascope control panel.

The line arrays used have 1024 elements which makes them capable of resolving 1024 lines per picture width before aliasing occurs. This is equivalent to 768 lines per picture height, which is the more conventional expression. The maximum expansion for cinemascope use requires that approximately 4/7 of the film frame is expanded to fill the full width of the TV screen. It can be seen that this still permits resolution of greater than 400 lines per picture height to be obtained without aliasing. However in order to achieve this it is essential that the bandwidth of the circuits preceding the expansion system is adequate to handle over 700 lines per picture height or effectively about 9 MHz.

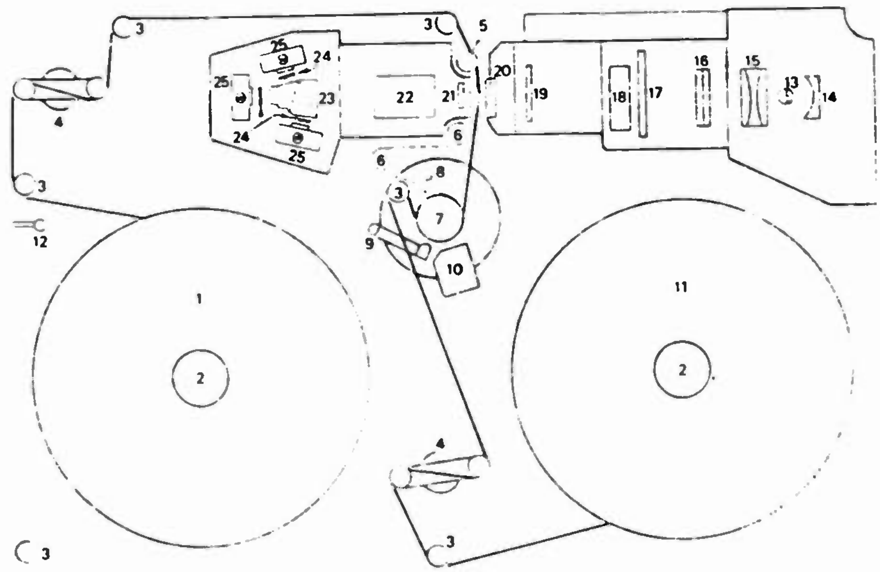
Figure 4 shows this system employed.

The image of the whole of the picture-bearing width of the film is arranged to occupy the full width of the line array. The filters between the line arrays and the A/D converters are switched to permit transmission of the required higher bandwidth. The A/D boards then digitize the signal at a frequency determined by the amount of expansion required, which may be anything from 1:1 to approximately 1.75 to 1.

The digitizing frequency is determined by the relationship:-

$$\text{expansion ratio} = \frac{\text{Digitizing frequency}}{\text{System clock ratio}}$$

The digitized lines are written into stores A and B alternately. The output is read from one store whilst writing into the other store. The output always contains 680 samples read out over the normal active line period. The input may have up to 1.75 times this number of samples, depending on the expansion ratio used.



Key
1. Feed spool
2. Spool retainer
3. Guidance rollers
4. Compliance arm
5. Frame pulse generator sprocket
6. Auxiliary 35mm rollers
7. Capstan
8. Optical sound sensor support
9. Magnetic head support
10. Optical sound illumination
11. Take-up spool
12. Feed spool reversal switch
13. Lamp
14. Reflector
15. Condenser lens
16. Infra-red filters
17. Variable n.d filter
18. Field lens
19. Cyan filter
20. Gate lens
21. Emulsion position corrector
22. Main projection lens
23. Optical beam splitter
24. N.D filters
25. Line arrays and mounting blocks

Figure 1

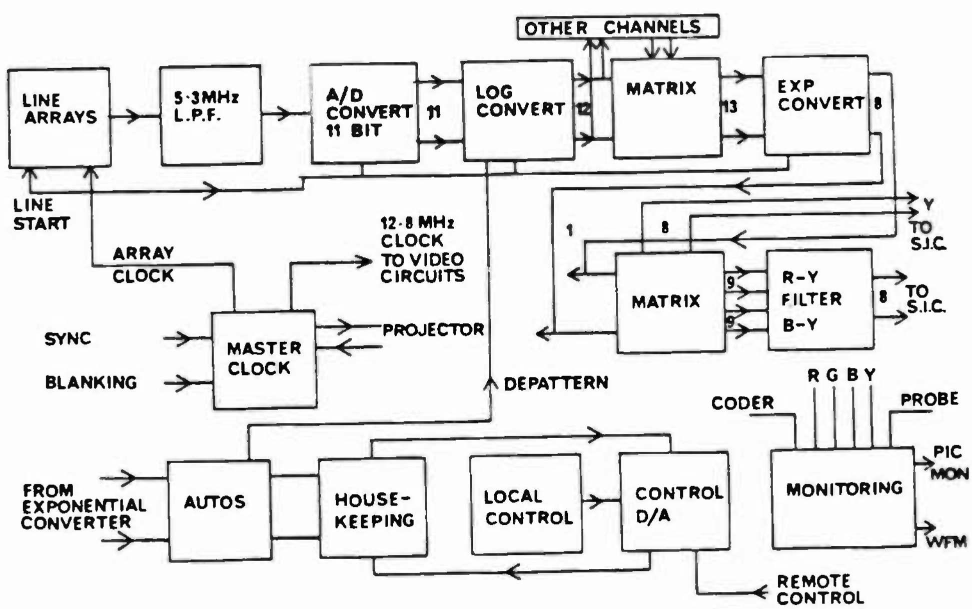


Figure 2

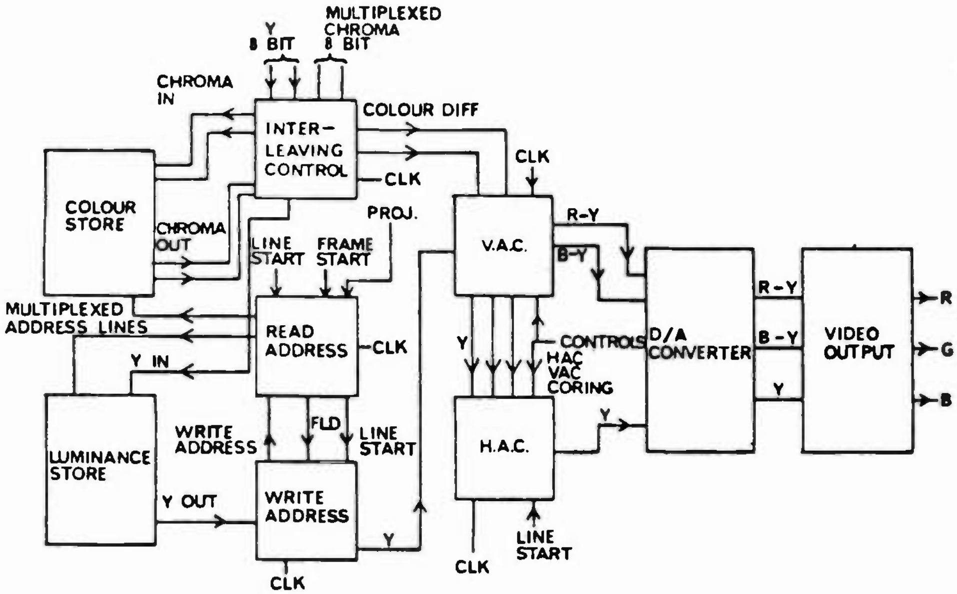
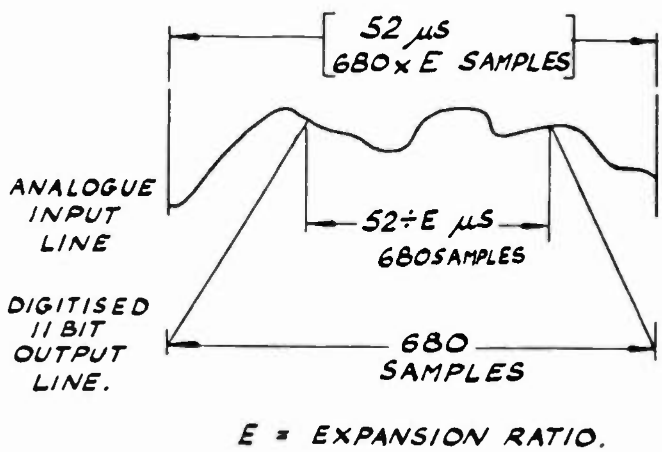


Figure 3



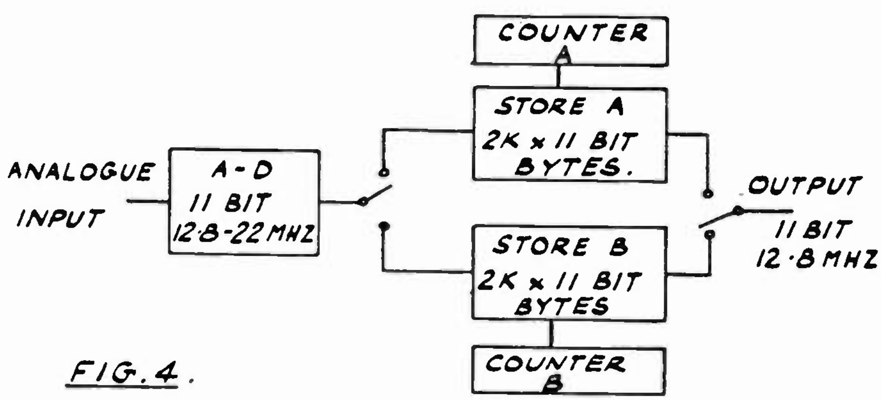


FIG. 4.

When writing into a store the corresponding counter has been cleared to zero. For output from a store the counter is loaded with the "Pan Word" which defines the position of the output line with respect to the film frame.

Each store has a capacity of two thousand eleven-bit bytes, and the output signal from the store is processed by the following circuits in exactly the same way as film of normal aspect ratio.

LOG CONVERTER

Once the signal has been digitized it is converted into a logarithmic form. Not only does this allow for a simple method of achieving a continuously variable true power law gamma correction but it also allows the multiplicative process required for the colour correcting matrix to be performed. This colour correction is the process used to offset the effects of reading the film dyes where they overlap. The defect is a form of cross-modulation between the three colour channels. The actual base of the logarithm is chosen to suitably fill the available number range and the result of this is a base of 1.0024857. The log function is implemented by the use of a bipolar PROM look-up table, the data for which was generated by a computer program. This technique allows direct computer-to-PROM working and removes the need for any additional handling, somewhat different from the old days of a handful of diodes and resistors! A test signal, which consists of a sawtooth and pulse and bar, may be selected at the front of the converter.

After log conversion a black reference signal is inserted during the line blanking interval. This data is later extracted and is the value that will be finally set to black clipping level. The logarithmic output signal is of 12-bit accuracy.

MATRIX

As previously mentioned a masking matrix is used to correct for errors in reading the film dyes. As the error is a multiplicative one, the correction can be achieved by subtracting signals in the log form, (equivalent to division). Additionally, we may see that when we multiply a log by a constant we are actually raising the log term to a power. Hence, we may use this as a method of performing gamma correction at this stage. In the B3410 all the required gamma and masking functions are provided by a single nine-term matrix. Obviously considerable 'number crunching' must take place in the control processor. The mask factors are obtained from fixed data in PROM (nominally for Technicolor printing or Eastmancolor dyes) or from six preset controls. These values are multiplexed by the master masking demand control which has effectively 32 steps. The resultant data is further modified to include the settings of gamma trim and master gamma controls before it is fed, once a film frame, to the masking printed boards. Look-up tables provide partial products which are then suitably combined. Once again, bipolar PROMS are used for the look-up tables. The multiplier in the main channel is of 7-bit accuracy whilst the other channels, where the multipliers are smaller, are performed at 5-bit accuracy. Finally all signals are summed to an accuracy of 13-bits. Throughout the digital system, the signal is latched after each processing operation. As a point of interest, the digital system takes around 90 amps at 5 volts, all clocked synchronously.

EXPONENTIAL CONVERSION

Once the signal has been masked it is then required to convert it back to a more normal form by the use of an exponential converter. At this stage it is necessary to perform a number of compensations to allow for the effects of limited system contrast ratios and hence avoid transmitting unwanted pedestals etc.

The first compensation is to prevent a variation of gamma causing substantial movements of black level. A data word is latched from the microprocessor and this is added to the incoming video data. This causes the signal to be moved along the exponential curve and hence the output is varied. Next the black data inserted in the logarithmic converter is extracted and fed to a look-up table. The data from the table is used for providing gain compensation for black level variation, however, before this correction is added in, the inversion for negative operation is performed. The signal can then be passed to the exponential converter table. Finally the black level data is latched once more and this is subtracted from the main video data, thus setting the final black levels. The signal is now in an 8-bit form but is still sequential rather than interlaced.

LUMINANCE AND CHROMINANCE MATRIX AND FILTERS

The signals from the exponential amplifier are matrixed together to produce digital Y, R-Y and B-Y signals. The matrix uses bipolar PROMS preprogrammed so that for any eight bit input address code the output will be multiplied by a specific factor, e.g., a luminosity coefficient. A combination of this technique and binary addition enables the required luminance and chrominance signals to be obtained.

The colour difference signals are filtered to half the bandwidth of the chroma signals and they are sampled at half the normal clock rate so that both chroma signals are effectively multiplexed into the same bandwidth as the luminance signal. The colour difference filter is a digital linear phase response low pass filter which is implemented using a finite impulse response technique with an impulse duration of 21 samples. In practice the choice of a cut off frequency equal to $\frac{1}{4}$ of the sampling frequency and an odd number of samples makes all the even coefficients zero. As a result only eleven "past" and "future" samples are required. The samples are obtained by delaying the required signals by integral numbers of clock periods.

SEQUENTIAL TO INTERLACE CONVERSION

Sequential to interlace conversion is now a well known art. Figure 3 shows the sequential to interlace converter and the vertical and horizontal aperture correction. The actual conversion system uses approximately two fields of storage for luminance and two for chrominance. Although we have already noted that the data from the line array is sequential, it is also non-standard in another way. This is due to the variation of the film frame bar width for different gauges and formats. For example, with 35 mm non-anamorphic film the frame bar is about 20% of the total frame where as with cinemascope format the bar is very narrow. It follows that, as the whole frame always passes the array in a fixed time, the active part of the frame must pass more quickly for standard format than for cinemascope. This variation is accounted for by adjusting the line blanking interval, while leaving the active part of the line fixed at the normal 52 micro-seconds.

Before the data can be passed to the store proper, it must undergo an interleaving process to allow for coincident reading and writing cycles from the same RAM. This is done by a suitably controlled system of shift registers and data selectors. Following this the data feeds to a 15-way demultiplexing system to reduce the data rate to that which is suitable for the rather slow RAMs (relatively slow dynamic RAMs are used for reasons of cost and power). For normal speed 625-line operation the frame write cycle must start sufficiently ahead so that the complete frame has been scanned before the TV system comes to the end of its first field. At this point in time there will be one complete field of stored information which is valid and is required to make up the next TV field; the other data which is the field just output may now be overwritten.

We could see then that only a little over 1.5 fields of store is required, however, for 525 lines where the 2-3 system is needed and for speeds other than normal, additional storage is needed. In fact, a little over two complete fields of 625-line store are provided, 7.86M bits altogether. When data is output from the store, it is passed through a corresponding 15-way multiplexer to provide a normal TV system interlaced output.

APERTURE CORRECTION

The signal from the sequential to interlace converter is passed to the aperture correction circuit which includes horizontal and vertical correctors plus a coring circuit. The data is first fed to the vertical corrector line delays which are constructed from fast static RAM. Signals from the delays are then added in the usual proportions to form a vertical-correction signal. This signal is passed on to a comparator circuit which performs a coring operation to remove unwanted low level signals to the desired degree. The vertical corrector has a conventional response with the first maximum at ½ line frequency, and following peaks spaced at line frequency intervals thereafter. The response shape is the usual cosine characteristic. The video signal for the main path and the horizontal corrector is extracted after the first line delay and a correction signal is formed with two one clock delays. This gives the correction a theoretical first peak at 6.4 MHz. The horizontal correction signal is also passed through a coring circuit.

The correction signals then undergo a gain adjustment in accord with the setting of the correction controls.

The coring and correction levels are commanded from the microprocessor data bus. The balance of horizontal to vertical correction may be controlled by two separate controls with overall level being set by the master aperture correction control.

DIGITAL TO ANALOGUE CONVERSION

The digital to analogue conversion follows immediately after the aperture corrector. Once the signal is back in analogue form it must be filtered to reject the digitization products outside the video band. A matrix is then used to remake the separate red, green and blue signals which may then be passed to the internal colour encoder.

MONITORING AND TESTING

The telecine has built-in monitoring for system checking. This may be used in conjunction with either the analogue test signal injected into the line array board or with the digital test signal of the logarithmic converter.

It is also possible to insert external test signals. To enable the video signals throughout the digital system to be examined for maintenance purposes, a test board with a D/A converter may be plugged into any of the digital video board positions. This provides a normal analogue video signal and thus enables checking with a standard oscilloscope. In addition to this the video control processor can be switched to provide fixed stimuli for the video system and thus enable signature analysis techniques to be used.

VIDEO CONTROL SYSTEM

The whole of the video system is controlled by the 'housekeeping' microprocessor. The device employed is an Intel 8085A. All analogue panel control signals are digitized after having been selected by an analogue multiplexer. The digitization is performed in one sequence and the resultant values are then held in RAM. Additionally the conditional controls, masking, C'scope, picture/bars etc. are read and the system then computes all the necessary control function.

System information is communicated to the various boards to give pulse rates and speeds etc.

Data to the video circuitry is only output during frame blanking to avoid a sharp change part way through the picture. Boards prior to the sequential to interlace converter are serviced during film frame blanking and those after during TV system blanking.

A test unit is provided to allow any of the microprocessor systems within the telecine to be stopped or single-stepped. Any memory location can be examined or changed if it is RAM.

Also, simple test routines may also be directly typed in.

AUTOMATION

The standard telecine is equipped with an automatic control system operating on the integrate-to-grey principle for white balance. Black levels and overall white level may also be controlled. This system acts on the digital video and is intended for on-air operational use where the film is of unknown quality.

SERVO SYSTEMS

Turning now to the servo systems employed, three systems are used, these being:-

1. Capstan servo

2. Spooling servos

3. Light control servo

The capstan servo is responsible for moving the film at the correct velocity and phasing to provide the whole of the vertical scanning of the picture. This obviously requires a high degree of both mechanical and electronic precision to avoid picture unsteadiness and vertical non-linearities.

The capstan revolves at approximately one rev/sec on 16 mm film and 2.5 rev/sec on 35 mm film. It is directly driven, and the system is especially designed to have adequate inertia controllable at very slow speeds, yet it is capable of being started and locked up in approximately 100 msec. The motor has multiple poles to prevent cogging and carries a 5000-line optical tachometer at the rear. Velocity lock is achieved by comparing the number of master clock (12.9 MHz) pulses occuring between successive tachometer pulses against a pre-set number. Positional information for framing and phase lock are obtained by comparing film frame information from the sprocket pulse generator to pulses generated from television frame pulses and the film "racking" or "framing" circuits. The phasing error signal is fed into the velocity servo as an offset. A microprocessor assesses the effect this offset has on the velocity drive and adjusts the preset count requirement to take care of film stretch or shrinkage. This enables tight velocity lock to be maintained. Changes of standard film gauge or speed are performed by a change of preset count and by division of either tachometer or clock pulses. "Shuttle" which is continuously variable between 1/5 speed and the fast rewind speed, is achieved by the use of a variable preset number obtained by A to D conversion from a variable dc.

The spooling servos control the film tension and remove spooling drag from the capstan. Direct-drive, high torque, printed-circuit motors are used. The shafts of the compliance arms carry slitted discs which are part of a spiral curve. These slits control the position of the light falling on optically variable resistors, and cause progressive forward or reverse drive to be applied to the spooling motors as the arms move away from their correct tension position.

The light control servo is a positional servo that controls the variable N.D. disc. It receives control via the main microprocessor system from either the light control knob or the automatic system.

PREFIX

This is a fully programmable system for colour correction. It has access to all the telecine controls including Pan Scan and Sound. It is capable of changing control settings on a frame-by-frame basis or dynamically over a number of frames. It provides either linear or "S" pan modes, and outputs programmable cues for pre-roll etc. It will accept a cue input from another device such as a shot change detector.

It has editing facilities to enable any control settings to be changed and in the edit mode the actual settings of all controls can be read for any individual "event".

The normal display shows events against frame numbers.

LATEST DEVELOPMENTS

As a result of operational experience and the demand for additional facilities, a number of items are in the course of development. These include the following:-

Black Stretch

The combination of black crushing which frequently occurs in high contrast films and an observed non-linearity in the line arrays, which occurs at very low light levels, has made it desirable to employ additional "black stretch". This black stretch is only used in the bottom few percent of the video signal and is manually variable. It is implemented by obtaining digital luminance information at the aperture corrector and using this to derive a correction signal which is first fed through a D/A conversion circuit and added back into the analogue system following the main D/A conversion.

Variable Speed Option

A variable speed facility is in course of development which will enable time compression or expansion of feature films or other non-standard forms of operation. This facility requires more storage and different store organisation. The storage will be in the form of 64K RAM which will permit the system to be incorporated within the basic telecine.

Selective Colour Correction

A post store corrector which permits independent control of primary and complementary colours for scene by scene correction on still frame is also included in the development programme.

ACKNOWLEDGEMENT

The author wishes to thank the Directors of Marconi Communication Systems Limited for permission to publish this paper and also his many colleagues who brought the design concepts to fruition.

Using Small Diameter Pickup Tube for High Quality Television Production

N. V. Rao and A. Franken

Amperex Electronic Corp & Philips

Slatersville, RI

High quality television production has to be defined before we discuss the components that go to make the system. Literature search lead me to the conclusion that there is no precise definition of "high quality TV production". My discussions with specialists on TV production resulted in a broad definition of what makes a high quality TV production. In these discussions it was clear that one inch tape was a definite requirement; using this tape one should be able to make a few generations - say five or six copies - without noticeable deterioration in the quality of the picture.

In order to get that kind of performance the camera should be capable of performing with a 58db to 60db S/N. Well, this is a severe demand on the camera and camera tube. Other qualifying parameters in our high quality TV production are listed below:

TABLE I

Registration	-	0.05% of picture height and stable.
Geometry	-	Error less than 1%.
Center Resolution	-	50% for green corner resolution within 10 to 20% of center resolution.
Colorimetry	-	Good.
Dynamic Range	-	3 to 4F stops.
Flare	-	5% or less.
Lag	-	Good.
Burn-in	-	Not visible with additional 18db gain.

The above requirements are more or less minimum.

Four or five years ago such a demand on 18mm tube would have been considered an impossible task. During the last few years the technology has improved and many new developments have been made in the 18mm tubes. The major improvements are:

Low capacitance output of target.
Introduction of diode gun. [1,2]
New concept of electrostatic focus - accelarating lens.
Electrostatic deflection - mixed field tube construction.

Improvement in signal to noise ratio can be achieved by increasing the signal or reducing the noise current. Increasing the signal current under given lighting conditions by increasing the sensitivity of the photo conductor is difficult to achieve because the photoconductors used in camera tubes for TV production purposes - lead oxide and selinum layers - are already working under high quantum efficiency. Other possibility of improving the signal is using more efficient lenses. It is much easier and less expensive to get lenses with lower F numbers for smaller format tubes.

The other possibility of improving S/N is to reduce the noise current.

The expression for the noise current is given by:

$$i_n^2 = \frac{4kTB}{R} + \frac{16}{3} kTR_{eq} C^2 \pi^2 B^3. \qquad (1)$$

Where k is Boltzmann's constant

B	-	Bandwidth
R	-	Signal resistance
R_{eq}	-	The equivalent noise resistance
C	-	Stray capacitance (including target capacitance)

The noise current can be reduced by reducing the target capacitance. The smaller format tube has smaller target capacitance and contributes less noise due to stray capacitance. In addition the recent developments in tube technology has resulted in tubes with low target capacitance tubes (LOC Tubes). As a result of this improvement it is possible to achieve 58 to 60db signal to noise in 18mm tubes.

The diode gun introduced in the 18mm tubes, for example in XQ2427 and XQ3427 and other tube types has the advantage of high beam capability without increasing the energy spread of the electron beam. The high beam availability of the diode gun helps to handle the highlights and the low energy spread of electron velocity results in excellent lag conditions. The decay lag and rise lag have been analyzed in detail by L. Heyne.[3] This analysis shows the signal current as a function of time is given by the equation:

$$i_s(t) = i_2 \cdot \left\{ 1 + \left(\frac{i_2}{i_1} - 1 \right) \cdot e^{-\frac{i_2 t}{c V_0}} \right\}^{-1}$$

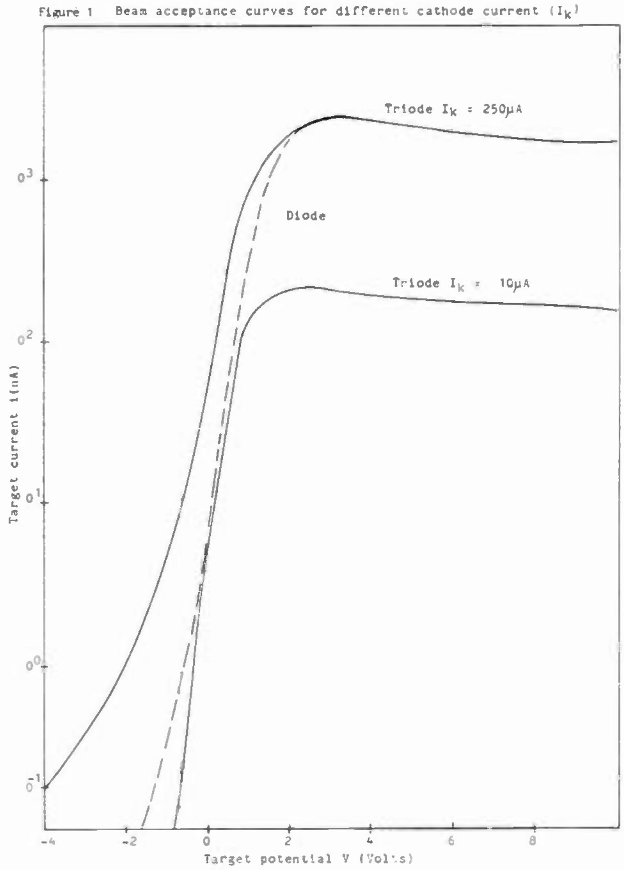
i_s = signal current
i_1 = initial current
i_2 = final current
c = capacity of the layer
t = time
V_0 = potential difference within which the current accepted by the layer increases by a factor e.

If V_0 is stated in volts the beam temperature is 1600 V_0K.

The figure (1) shows the beam acceptance curves for the conventional gun and the diode gun. The diode gun has small energy spread even at high beam currents. This results in much better beam acceptance and better lag of tube.

Figure 1

Figure 1 Beam acceptance curves for different cathode current (I_k)



From equation (2) it is clear that the low capacity of the photoconductive layer improves the beam acceptance and consequently the lag improves. Again 18mm tube with smaller area target has only half as much layer capacity as a 30mm tube - hence better lag characteristics. Also, Plumbicon® with thicker lead oxide layer (≈ 8μm) has better lag characteristics than thin layer (~ 4μm) selenium photo conductive Saticon ® tubes.

Another important requirement as shown in Table I is the modulation depth of the tube. The response required is 45 to 50% at 400 TV lines in the center. The modulation depth of the 30mm tube and 25mm tube is well in excess of 50%. With the improvement in the electron gun, electron optics and the photoconductive layer the modulation depth of 18mm tube has reached the 50% level. The Saticon tube with 4μm thick photoconductive layer has better modulation depth than the Plumbicon with 8μm thick photoconductive layer.

In addition to the center resolution, the corner resolution should be within 10 to 20% of the center resolution. Two new developments - 18mm tube with electrostatic focus/magnetic deflection and 18mm tubes with magnetic focus/electrostatic deflection are of great interest. I am showing the construction of XQ4187.

Figure 2

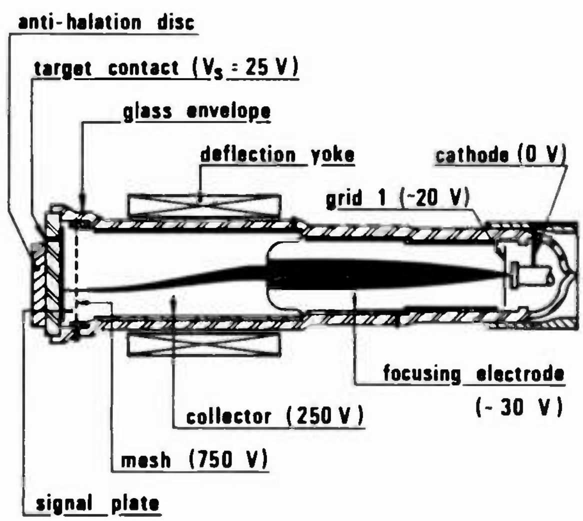


Figure 2

Arrangement of the electrodes, deflection coil, and base connector of the XQ4187.

® Plumbicon - Registered Trade Mark for television camera tube of N. V. Philips.

® Saticon - Registered Trade Mark for television camera tube of Hitachi.

This tube has an evaporated wall electrode for the focus. The other electrodes, for example the mesh, diaphram etc. are rigidly mounted as shown. This provides the required stability of registration. Also the mechanical centering is done in the yoke before the whole assembly is mounted in the camera. Figure (3) shows the center and corner resolution of the tube and highlight handling capability.

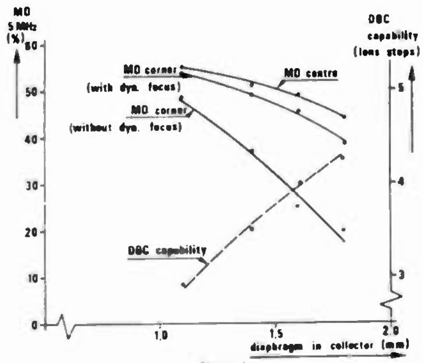
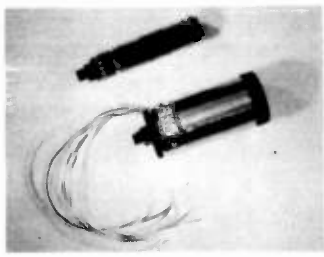


Figure 3
Modulation depth and dynamic beam control capability of XQ4187.

The next figure (4) shows the photograph of XQ3457.



Again the metal coating is evaporated on the walls of the glass bulb and the deflectron elements are formed using laser beam technique. The stability of this type of construction is also good.

An important data missing in Table II of items is the geometry and registration. From all the measurements we have it is possible to meet the geometry and registration error of $<0.2\%$ without any correction. With the necessary corrections, it is possible to achieve the required registration of better than 0.05%. However, one major problem used to be stability of registration. In this respect the XQ4187 and XQ3457 show better stability due to the evaporated wall electrode.

The image burn-in due to high intensity light should be negligible. This factor is more important in a camera with automatic set up using diascope. The type of photoconductor you choose may play an important role here.

The camera tube flare should under all circumstances be much less than the flare due to lens. The camera tubes described above meet these requirement without any problem.

TABLE II

18mm Camera Tubes for TV Production

	P8462 XQ3427 BC4937 Lead Oxide	85XQ XQ4187 Lead Oxide	87XQ XQ3457 Lead Oxide	H9386 Selenium	CT-2322 Selenium
Length/Diameter (mm)	105/38	97/22	85/35	105/38	85/33
Mass (grams)	230	110	160	230	190
Focus/Deflection	Mag/Mag	Elec/Mag	Mag/Elec	Mag/Mag	Mag/Elec
Gun Type	Diode	Diode	Diode	Diode	Diode
Mod. Depth 400TVL	50%	50%	50%	60%	55%
Sensitivity μa/lm	320	320	320	350	320
Power consumption (W)					
Filament (tube/circ)	0.6/0.5(6.3V)	0.5/0.15(9V)	0.6/0.5(6.3V)	.6/.5(6.3)	0.6/0.5(6.3V)
Focus (tube/circ)	0.9/∼0.5	-	1.0/0.5	0.9/∼0.5	1.0/0.5
Deflection (yoke/circ)	0.1/∼0.7	0.1/∼0.7	∼ .4	0.1/∼.7	∼.4
TOTAL	∼ 3.3	∼1.45	∼ 3.0	∼3.3	∼3.0

From all the information that we have here, we can conclude that the 18mm pickup tubes are capable of meeting all the requirements of the high quality television production. In addition, these tubes have the added advantage of low power and small size. Also, there is a wider choice of lenses with various F stops and zoom ratios when small diameter camera tubes are used in TV production.

References

(1) Van Roosmalen - US Patent No. 3,831,058.

(2) N. V. Rao - US Patent No. 4,388,556.

(3) L. Heijne - Acta Electronica 2, 124-131, 1957 -1958.

International and Domestic Aspects of

Recent AM Band Revisions

Wallace E. Johnson

Moffet, Larson & Johnson, P.C.

Arlington, Virginia

It is not red hot news that a new AM Broadcasting Agreement was concluded for our hemisphere in December of 1981. But, that event is very important because it will greatly impact on future development of AM broadcasting in the United States.

In order to understand the importance of the new Rio Agreement it is well to review a little bit of the AM history.

AM Broadcasting in the United States developed under two long standing agreements - NARBA (North American Regional Broadcasting Agreement) and the U.S./Mexican Agreement.

NARBA countries included Canada, Cuba, Bahama Islands and the Dominican Republic. Cuba ceased complying with NARBA when Castro came into power. Bahama Islands and the Dominican Republic had some problems complying with the terms of NARBA which resulted in some difficulties for the U.S. NARBA became, in effect, a bi-lateral agreement with Canada. However, Canada had become dissatisfied with the Agreement, particularly with regard to clear channel priorities and protections, so in December of 1981 they stated that they were going to withdraw from the Agreement. This would have left us with no effective agreement with any of our neighboring countries except Mexico - and of course more distant countries in Central and South America, which were becoming increasingly important because of growing interference problems. Therefore a new, effective, agreement with our neighbors and more distant countries in our hemisphere became extremely important for protection of our present broadcast system and to permit an effective, continuing expansion of AM broadcasting.

Since 1973 various steps had been taken by the International Telecommunications Union (ITU), the Inter-American Telecommunications Conference (CITEL), and countries in the Western Hemisphere looking towards an AM Agreement for Region 2. Regions 1 and 3 (Africa, Asia, Australia and Europe) concluded AM

Broadcast Agreements in 1975. In March, 1980 the First Conference was held in Buenos Aires where the technical criteria needed for a regional plan was developed. Agreement could not be reached on the issue regarding 9 or 10 kiloHertz spacing, so that decision was delayed until the Second Conference. A Panel of Experts, with representatives from eight countries, including the United States was assembled to help the International Frequency Registration Board (IFRB) to make a comparative study of 9 versus 10 kiloHertz spacing. The Panel of Experts met 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 were made which convinced the FCC that the costs involved to the broadcast industry and the public, 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 reversed the U.S. position for 9 kiloHertz at the First Conference. With the active support that 10 kiloHertz spacing had from the U.S. and thirteen other nations at the Conference 10 kiloHertz became the decision.

Following are some major aspects of the Rio Agreement:

1. 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) being listed in the Geneva Master Registry. This action was important in regard to achieving international protection for those assignments.

2. The Agreement was scheduled to enter into force July 11, 1983. Even though the Agreement is considered to be in effect it has not been submitted to the U.S. Senate for ratification. It is anticipated that this process will be initiated this summer.

3. 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.

4. There are very specific procedures in the Agreement regarding modification of the Assignment Plan by new or modified assignments. IFRB will record acceptable assignments in the Master Register and will play a role in resolving incompatibilities as required, or as requested by administrations. The U.S. assignment process will continue much as it is now with the FCC only granting applications which are in accordance with the terms of the Agreement.

5. The Agreement is binding between the contracting

parties, but not with non-contracting parties. However, there is a process whereby non-signatory countries may accede to the Agreement. The Dominican Republic was not a contracting party, but have stated their intentions to become a party. Twenty six of the thirty two countries in the Hemisphere participated in the Conference.

6. Cuba, which was a participant in the Conference, withdrew on December 14th, 1981. 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. The second was the U.S. announcement of its intention to implement a Radio Marti operation to beam radio programs to Cuba.

7. 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 the 10% of the time interferring propagation curve. All other countries, except Canada, Mexico, Greenland, and the French Department of St. Pierre and Miquelon will use a 50% of the time interference curve which will tend to increase interference to some U.S. stations.

8. In the Final Protocol 18 reservations and 2 counter-reservations were taken. Reservation #14 was by the U.S. which reserves the right to take such action as necessary to assure provision of needed service to adversely affected areas if efforts to eliminate such interference fails to lead to satisfactory solutions.

The major problem regarding the Rio Agreement is with reference to Cuba. If Cuba actually implements their proposed assignments there will be some very serious interference to a number of U.S. stations. Most of this interference will not be recoverable to any significant degree. The Cuban problem is primarily a political matter with broadcasting only one part of the total picture of our relations with them. There does not appear to be any possibility of a negotiated agreement at the present time on technical standards between the U.S. and Cuba. With no door presently open with Cuba to attempt resolution of interference problems the best we can hope for is that we may possibly benefit from negotiations between Cuba and other countries which are a part of the Region 2 Agreement.

The Broadcasting to Cuba Bill (Radio Marti) was passed by Congress in September, 1983. While authorizing broadcasts to Cuba, the legislation states that those broadcasts must be under the jurisdiction of the Voice of America on the VOA 1180 kiloHertz frequency. The Bill also authorizes the appropriation of five million dollars to the United States Information Agency (USIA) to help broadcasters suffering losses due to Cuban interference. The FCC is now in the process of developing a method for determining how this money is to be allocated to broadcasters impacted by Cuban interference.

A new bi-lateral agreement has been negotiated with Canada and was signed and entered into force January 17, 1984. It reflects the provisions of the Region 2 Agreement, but also has some provisions of specific interest to broadcasters in the U.S., such as:

1. An extended hours provision which permits operation during nighttime hours which occur between 6 a.m. and two hours past local sunset time, with protection requirements based upon use of diurnal curves.

2. Class IV stations presently operating on local channels will be able to increase nighttime operating power by a factor of four, such as 250 watts to one kiloWatt. The effective date of such power increase will be established by an exchange of letters between the administrations and also coordinated with Mexico.

3. Nighttime operations will be permitted on the clear channels, as specified in NARBA, on the basis of protection of the Class A station in the other country within their 0.5 mV/m 50% skywave contours. Protection will not be required outside of the country where the station is located.

The FCC will be issuing a Notice of Proposed Rule Making which will specify the rule changes necessary to implement the new U.S./Canadian Agreement.

Negotiations are also taking place with Mexico regarding a new bi-lateral agreement which will reflect the Region 2 Agreement together with specific provisions such as in the U.S./Canadian Agreement, including extended hours of operation and nighttime increase in power by Class IV stations. The new agreement will also permit operation on the clear channels of each country on the basis of protection to the 0.5 mV/m 50% skywave contours. It is anticipated that this agreement will be completed before the end of 1984.

The only remaining international matter pertaining to AM broadcasting relates to the World Administrative Radio Conference of 1979 which provided for an extension of the AM band to include 525-535 kiloHertz and 1605-1705 kiloHertz. Originally, 525-535 kiloHertz was to be used on an equal primary basis by broadcasting and aeronautical navigation; 1605-1625 kiloHertz exclusively broadcasting; 1625-1705 kiloHertz where broadcasting would be allowed on a primary basis, fixed and mobile on a permitted basis and radio location on a secondary basis. Broadcasting in 1625-1665 kiloHertz was not to commence before July 1, 1987 and operation in 1665-1705 kiloHertz not to commence before July 1, 1990.

Part 2 of the FCC Rules as revised and adopted by the Commission following conclusion of the 1979 WARC proceedings has listed 525-535 kiloHertz for use by mobile travellers information service on a primary basis, 1605-1615 kiloHertz is listed for use by mobile travellers information service and 1625-1705 kiloHertz is for radio location use. These allocations will be listed until implementation proceedings are concluded by the FCC, which process will be initiated by a Notice of Inquiry scheduled to be released in April of 1984. Upon conclusion of the implementation proceedings it is anticipated that 1615-1625 kiloHertz will be exclusively broadcast, while 1625-1705 kiloHertz will be primary broadcast with radio location secondary.

The extended band usage and implementation dates are also dependent upon two Western Hemisphere conferences to be held under the auspices of the ITU in 1986 and 1988. The 1986 Conference will develop the technical standards for the expanded band while the 1988 Conference will develop the assignment plan. As a result of these Conferences and finalization of FCC rule making proceedings to implement the extended bands it is anticipated that it will be 1990 before use of these expanded bands can take place.

From this rather brief, summary report it is hoped that there will develop an appreciation of the impact which international activities during the last four years will have on the future of AM broadcasting in the United States.

THE EFFECTS OF INCREASED DEVIATION ON ADJACENT FM CHANNEL PROTECTION

Michael C. Rau
National Association of Broadcasters
Washington, D.C.

Harrison J. Klein
Westinghouse Broadcasting and Cable, Inc.
New York, N.Y.

John Kean
National Public Radio
Washington, D.C.

Introduction

When the Federal Communications Commission issued its First Report and Order in the Notice of Proposed Rule Making concerning use of the Subsidiary Communications Authorization (SCA), 1/ it deferred decision on whether to increase the peak deviation allowed an FM station during SCA use. In a subsequent Order, 2/ the Commission requested comments on two issues: "(1) the degree of reception degradation, if any, caused by adjacent channel stations using peak modulation exceeding 100%; and (2) whether short-spaced stations may suffer adjacent channel interference to any greater extent than normally-spaced stations under the condition of typical FM program transmission practices." The National Association of Broadcasters (NAB), Westinghouse Broadcasting and Cable, Inc. (Group W), and National Public Radio (NPR), jointly sponsored a technical study to comment on these issues. The resulting Technical Report was submitted to the FCC, 3/ and this paper describes the study and its findings.

1/ Notice of Proposed Rule Making, BC Docket 82-536, (FCC 82-368), 47 Fed. Reg. 36235 (August 19, 1982). First Report and Order, FCC 83-154, adopted April 7, 1983, 48 Fed. Reg. 28445 (June 22, 1983). Corrected by 48 Fed. Reg. 37216 (August 17, 1983).

2/ Order Extending Time for Filing Comments to Notice of Proposed Rule Making, 48 Fed. Reg. 24945 (June 3, 1983).

3/ Increased FM Deviation, Additional Subcarriers and FM Broadcasting: A Technical Report, August 30, 1983, filed in BC Docket 82-536 by NAB, Group W and NPR.

Conclusions and Recommendations

Increasing Peak FM Modulation

Analysis of the test data reveals that peak modulation can be increased to as much as 110% under certain modulation conditions with a negligible change in first and second adjacent channel radio-frequency (RF) protection ratios. 4/ Accordingly, an appropriate FCC requirement for modulation during subcarrier use is:

> When one or more subcarriers are used in addition to the main channel program, the main channel modulation shall be reduced by one-half the total deviation of all subcarriers. The maximum peak modulation would thus be 100% plus one-half the total subcarrier modulation. In addition, peak modulation under any modulation condition is not to exceed 110%.

For example, a stereo station operating with one 10% subcarrier would be allowed 100% + 1/2 (10%) = 105% peak modulation. A station operating with two 10% subcarriers would be allowed 100% + 1/2 (20%) = 110% modulation. A monaural station operating with three 10% subcarriers would be allowed the maximum permissible modulation of 110%.

It should be noted that the reduction in main channel modulation (the "backoff") required when subcarriers are used is reduced by half from former requirements. Under this proposal, main channel modulation would be reduced less than 0.5 dB when one 10% subcarrier is added. This should help to increase the incentive for stations to begin taking advantage of their subcarrier capability.

Increased Peak FM Modulation and Short-Spaced FM Stations

Received interference is not directly determined by the actual distances between stations, but by the relative distances between desired and interfering stations at a particular receiver location. While it is true that a short-spaced station will have a smaller geographic circumference where a given undesired-to-desired RF protection ratio exists (since the undesired station is closer and the undesired signal is stronger), additional interference due to subcarrier operation and increased peak modulation should not occur because the proposed modulation levels are not expected to change existing U/D ratios. In other words, a receiver that has a particular RF protection ratio at its input will be affected equally by the proposed modulation increase irrespective of the spacing between the desired and undesired stations.

Thus, if peak FM modulation is increased to 110%, short-spaced stations are not affected to any greater extent than normally-spaced stations.

Test Description

Technical Background

The research summarized in this paper explores the effects of increasing

---

4/ 100% modulation is defined as 75 kHz peak FM deviation (47 CFR 73.310(a)). 110% modulation would be defined as a peak deviation of 82.5 kHz.

permissible FM modulation from 100% to 120% in 5% increments with various combinations of subcarriers. Because an increase in FM modulation may bring additional energy into adjacent channels, there is a potential for added interference. The extent of this interference can be assessed by observing how demodulated interference in a typical FM receiver changes as the modulation characteristics of an undesired signal on an adjacent channel are varied.

To measure relative levels of interference in an FM receiver, these tests compared RF undesired-to-desired protection ratios necessary to provide a 50 dB audio signal-to-noise ratio (S/N) in the receiver under test. For example, assume that to obtain a 50 dB audio S/N in a receiver, an undesired second adjacent channel station could be as much as 30 dB stronger than the desired station at the receiver's antenna terminals. 5/ The undesired-to-desired (U/D) RF protection ratio is then said to be 30 dB.

If the modulation characteristics of the undesired carrier are changed, and the new U/D RF protection ratio is found to be, say, 35 dB, interference has decreased. It now takes a stronger adjacent channel signal to cause the same amount of interference. Therefore, a change in modulation that causes an increased (more positive) U/D RF protection ratio results in less interference, and a change which causes a decreased (more negative) protection ratio results in more interference.

The tests measured the RF levels of undesired signals of varying transmission formats necessary to produce an audio S/N of 50 dB in the receiver under test. From this data, the U/D ratios were calculated and the increase in interference, if any, is apparent.

The Receivers and Signals

Five receivers were tested. The results of each receiver for each test were averaged together in order to provide a "composite" result table indicative of a "typical" FM broadcast receiver. The receivers were carefully chosen to be representative of the FM receiver industry in 1980 and 1981 as part of an industry-wide effort to gain insight into the relationship of RF protection ratios and recovered S/N. 6/

The desired signal was unmodulated in order to measure changes in noise floor as the undesired signal was varied. A 19 kHz pilot tone was injected at 10%. The desired signal at the input of the receiver was maintained at a level of 500 uV across 75 ohms (-54.8 dBm).

In all cases, the undesired signal was modulated with noise into the left and right program channels. A 19 kHz pilot tone was injected at 10%. The subcarriers, if present, were centered on either 67 kHz or 92 kHz, or both, and injected at a level of 10% modulation for each subcarrier.

The subcarriers themselves were unmodulated. After some testing in the

---

5/ A signal of higher than 30 dB above the desired station would degrade the audio S/N below 50 dB.

6/ See Report to the National Radio Systems Committee, July 24, 1981, filed in BC Docket No. 80-90 by the Subgroup on Technical Matters of the Advisory Committee on Radio Broadcasting, July 12, 1982, Wallace E. Johnson, Chairman.

NPR laboratory it was determined that modulation of the subcarriers had no noticeable effect on the measured RF protection ratios. Because subcarrier modulation would have increased the overall complexity of the tests, it was decided not to modulate the subcarriers with noise or any other test signals.

Two separate test runs were made, one employing a noise signal developed within the International Radio Consultative Committee (CCIR), and the other using "Synthetic Program Noise," a noise signal custom-developed for these tests at National Public Radio.

CCIR Noise Modulation. This noise signal is explained and presented in CCIR Recommendation 559. 7/ Ordinarily, its employment as a test signal is sufficient for the purposes of RF protection ratio measurements. In this instance the FCC, however, specifically requested information based on "typical FM program transmission practices." While CCIR noise was intended to represent "typical" music programming, its spectral characteristics and peak-to-average ratio are significantly different from program material commonly broadcast by FM stations in the U.S. This difference occurs principally because most FM stations "process" their audio prior to transmission, a factor not considered in the CCIR deliberations. 8/ Other technical analyses of the spectral characteristics of program material imply limits to the usefulness of the CCIR noise model. 9/

Figures 1 and 2 in Appendix A show the readily apparent differences between the FM spectra of CCIR noise modulation and typical FM radio stations. Compare the smooth nature of the CCIR noise spectrum with the sharp, abrupt spectrum of typical FM stations.

It was therefore necessary to develop a new type of noise that frequency-modulates an RF carrier in a way that occupies 200 kHz FM channels just like actual processed FM stations. After observing the characteristic spectrum "signatures" of several Washington, D.C. area FM stations with a spectrum analyzer, a noise was developed that met the noise-design objectives.

Synthetic Program Noise. Since the usual objective of audio processing is to assure that all frequencies at all times have the potential of fully modulating the FM transmitter, the new noise signal originated with so-called "pink" noise -- noise with equal energy per octave bandwidth. The pink noise was modified with a high-pass filter to attenuate low frequency energy below 50 Hz not normally present in actual FM programming, processed or otherwise. A low-pass filter then reduced energy above 15 kHz at a rate of 60 dB per octave. Finally, the noise was clipped in order to simulate the peak-to-average, or density, characteristics of typical audio processors. The resulting signal has

---

7/ Recommendations and Reports of the CCIR, Volume X, Rec. 559, XIVth Plenary Assembly, Kyoto, 1978, Curve B at Figure 2.

8/ Actually, the CCIR noise was originally developed for use in AM transmission systems. The corresponding Report for FM systems, Report 796, says that the "standardized coloured noise. . .described in detail in Recommendation 559" is to be used.

9/ See, for example, McKnight, J.G., Signal-to-Noise Problems and a New Equalization for Magnetic Recording of Music, Journal of the Audio Engineering Society, 7,5 (1959).

a synthesized spectrum typical of FM program transmission practices. This signal was named "Synthetic Program Noise," or "SPN."

The spectrum of an FM carrier modulated with SPN closely matches spectra of typical FM stations. See Figure 2 in Appendix A. Note the similar spectral distributions, particularly the steep skirts and uniform amplitude in the vicinity of the center frequency.

Measurement Procedures

Data was taken for both upper and lower first and second adjacent channels. The left and right channels of the undesired signal were modulated with either CCIR noise or SPN. Subcarriers were added, each at 10% injection, and various maximum peak modulation levels were used. The subcarriers were unmodulated. The RF level of the undesired signal was adjusted until the receiver under test provided a 50 dB audio S/N, and the level of the undesired signal in dBm was noted. 10/ The 19 kHz pilot tone modulation allowed the receivers to be measured in the stereo mode. This was considered to be a more stringent test than monaural measurements because stereo S/N is more sensitive to adjacent channel interference than monaural S/N.

Reduction of the data entailed the following tasks: first, the undesired signal levels for upper and lower adjacent channels were averaged together to provide a "net" response for each receiver at +/- 400 and +/- 200 kHz. Second, the undesired-to-desired protection ratios were calculated. Finally, for each noise type (CCIR and SPN), the protection ratios of each receiver employed in corresponding test situations were averaged together. The appropriate standard deviations also were calculated. The result is a "composite" protection ratio that reflects the mean effect on adjacent channel interference for various combinations of modulation percentages and subcarriers, calculated for each noise type.

Equipment Used and Block Diagrams Showing Interconnection

A list of equipment and block diagrams used in these tests are attached as Appendix B. Figures A, B, C and D are block diagrams detailing the equipment set-up actually used for data collection. Figures E, F and G are diagrams of detailed circuits of critical test points. Table 1 in Appendix B is a list of test equipment.

Interpretation of Results

The composite protection ratios are shown in Table 1 (Synthetic Program Noise) and Table 2 (CCIR Noise) of Appendix A. Recall that a change in modulation which causes an increased (more positive) protection ratio results in less interference. A decreased (more negative) protection ratio means more interference.

Analysis of SPN Data

The data shown in Table 1 for both second and first adjacent channels

---

10/ Audio frequency noise voltage was measured in accordance with CCIR Recommendation 468-2 Mod. F.

correlate well with the previous investigations of subcarrier modulation and occupied bandwidth conducted by NPR and Group W. 11/ Examining the table starting with stereo-only 100% modulation, observe that the protection ratios increase when a single subcarrier is added with no increase in modulation. This corroborates the conclusions reached previously by NPR and Group W that a full 10% backoff is not needed to maintain existing occupied bandwidth and protection criteria when a single subcarrier is in use.

The reduction in interference occurs because the sideband energy created on each side of the carrier by a low modulation index 12/ subcarrier has less potential for causing adjacent channel interference than the energy removed by the reduction in deviation of the high modulation index stereo program. Adding a second subcarrier while maintaining 100% modulation (for a main channel backoff of 20%) results again in a protection ratio higher than stereo-only 100% modulation.

The 105% column shows the protection ratio effects of the proposal to increase modulation to 105% when one subcarrier is in use. The second adjacent channel figures are nearly identical to the stereo-only 100% value. Although the first adjacent channel figure for a 92 kHz subcarrier is slightly more negative than stereo-only, the 67 kHz subcarrier figure is more positive by about the same amount, making the average almost exactly equal to stereo-only. On balance, it is evident that 105% is an appropriate modulation level when one 10% subcarrier is used.

When a second 10% subcarrier is added, the table shows that 110% is the most logical choice for a peak modulation limit. At this level the second adjacent channel 110% figure is within 0.1 dB of the stereo-only 100% peak modulation value. The first adjacent channel protection ratio at 110% peak modulation is actually better than stereo-only at 100% peak modulation.

Using a backoff of half the total subcarrier injection would allow a monophonic station to have one 30% subcarrier and 115% peak modulation. While no measurements were made for this particular condition, they would be expected to show a significant decrease in protection ratios. Thus peak modulation should be restricted to a maximum of 110% regardless of the number and injection levels of subcarriers used.

The mean figures in Table 1 of Appendix A correlate well with both theory and past experience. It should be noted, however, that the standard deviations shown in the table for each measurement condition are significantly greater than the variation in the mean calculated from all measurements. Any effect on FM service caused by higher FM deviations is eclipsed by differences in receiver design and performance that, among the receivers tested, varies by as much as 20 dB. In other words, from a listener's viewpoint, the quality of his receiver is much more determinative of audio quality than the particular transmission characteristics of the interfering RF signal.

11/ See Reply Comments of Westinghouse Broadcasting and Cable, Inc., filed November 17, 1982, in BC Docket No. 82-536, Engineering Report at 1. See also Comments of National Public Radio, filed October 18, 1982, in BC Docket No. 82-536.

12/ Modulation index is defined as the instantaneous deviation divided by the modulating frequency.

In addition, although the actual protection ratios differed widely from receiver to receiver, the changes in protection ratios as the modulation characteristics were varied remained relatively consistent from receiver to receiver. Therefore, decisions as to appropriate modulation limits can be made on the basis of the mean protection ratios without fear that many receivers will be affected differently.

Analysis of CCIR Noise Data

The data in Table 2 correlate with what would be expected of a low density modulating signal causing a more narrow occupied bandwidth than typical FM transmission practices. Second adjacent channel protection ratios deteriorate somewhat as modulation is increased above 100%. The average protection ratio at 105% modulation is 2.3 dB lower than stereo-only at 100%. At 110% the average ratio is 3.2 dB lower, at 115% it is 4.4 dB lower, and at 120% it is 5.0 dB lower.

The first adjacent channel protection ratios do not deteriorate to the same degree as the second adjacent ratios. Even at 120% modulation, the protection ratio is only 1 dB worse than stereo-only at 100%. This occurs because the distribution of sideband energy caused by the subcarrier has less potential for first adjacent channel interference than the distribution of sideband energy from the main channel. It should be remembered that the reduced second adjacent channel protection ratios would not be expected in practice, because the occupied bandwidths are not representative of typical FM transmissions.

Analysis of Short-Spacing Between FM Stations

A satisfactory signal-to-noise ratio in an FM receiver depends, among other things, upon the ratio of undesired-to-desired RF signal levels at the receiver's antenna terminals. The absolute levels of the desired and undesired signals are not significant, only their relative amplitudes. 13/ In the example described above (see "Technical Background"), a 30 dB U/D ratio provided a 50 dB audio S/N; it was unnecessary to specify actual desired or undesired signal strengths.

Because the signal strength of an FM station depends on its distance from a receiver, interference is not directly determined by the distance between stations; only by the relative distances from the receiver to both the desired and undesired stations. A receiver that has a particular U/D ratio at its input will be affected equally by increased peak modulation irrespective of the spacing between the desired and undesired stations.

This paper concludes that virtually no change in undesired-to-desired RF protection ratios -- no change in relative RF signal strengths -- is expected if the FCC increases permissible peak FM modulation to 110% with the use of two subcarriers. Accordingly, if no change occurs in relative signal strengths, short-spaced stations will not be affected by an increase in peak FM modulation to any greater or lesser extent than normally-spaced stations.

13/ This assumes that the desired signal is above the minimum 50 dB quieting threshold of the receiver, and that neither signal is above the receiver's overload point.

Appendix A

Table 1

Composite Undesired-to-Desired RF Protection Ratios Necessary to Provide 50 dB Audio S/N

Evaluated for Certain Modulation Percentages and Subcarriers

AVERAGE OF RESULTS

Noise: SPN

dB/Standard Deviation

Receivers: Sony STR-V45
Pioneer KP-3500
Technics SA-505
GE 3-5256
Magnavox PA1839

Second Adjacent Channel

	100%/S.D.	105%/S.D.	110%/S.D.	115%/S.D.	120%/S.D.
Stereo	+36.8/12.8	--	--	--	--
Stereo + 67 kHz	+38.0/12.8	+37.0/12.6	+36.4/12.6	--	--
Stereo + 92 kHz	+37.2/12.5	+36.7/12.5	+36.3/12.6	--	--
St. + 67 + 92 kHz	+37.9/12.5	--	+36.9/12.4	+36.4/12.4	+35.7/12.3

First Adjacent Channel

	100%/S.D.	105%/S.D.	110%/S.D.	115%/S.D.	120%/S.D.
Stereo	-5.3/10.3	--	--	--	--
Stereo + 67 kHz	-3.4/10.0	-4.6/10.7	-5.6/10.0	--	--
Stereo + 92 kHz	-4.4/8.6	-6.1/9.1	-6.9/8.7	--	--
St. + 67 + 92 kHz	-4.4/9.4	--	-4.6/8.9	-6.2/9.2	-7.1/8.9

Appendix A

Table 2

Composite Undesired-to-Desired RF Protection Ratios Necessary to Provide a 50 dB Audio S/N

Evaluated for Certain Modulation Percentages and Subcarriers

AVERAGE OF RESULTS

Noise: CCIR

Receivers: Magnavox PA1839
Technics SA-505
GE 3-5256
STR-V45
Pioneer KP3500

dB/Standard Deviation

Second Adjacent Channel

	100%/S.D.	105%/S.D.	110%/S.D.	115%/S.D.	120%/S.D.
Stereo	+38.4/10.8	--	--	--	--
Stereo + 67 kHz	+38.5/9.5	+37.6/9.6	+36.7/9.8	--	--
Stereo + 92 kHz	+35.3/9.6	+34.6/9.5	+34.0/9.6	--	--
St. + 67 + 92 kHz	+33.6/8.4	--	+34.8/9.6	+34.0/9.6	+33.4/9.9

First Adjacent Channel

	100%/S.D.	105%/S.D.	110%/S.D.	115%/S.D.	120%/S.D.
Stereo	-10.0/10.4	--	--	--	--
Stereo + 67 kHz	-9.4/12.8	-10.1/11.8	-10.9/11.1	--	--
Stereo + 92 kHz	-9.3/12.5	-9.7/11.5	-10.5/10.6	--	--
St. + 67 + 92 kHz	-9.2/13.2	--	-9.6/12.6	-10.2/11.8	-11.0/10.9

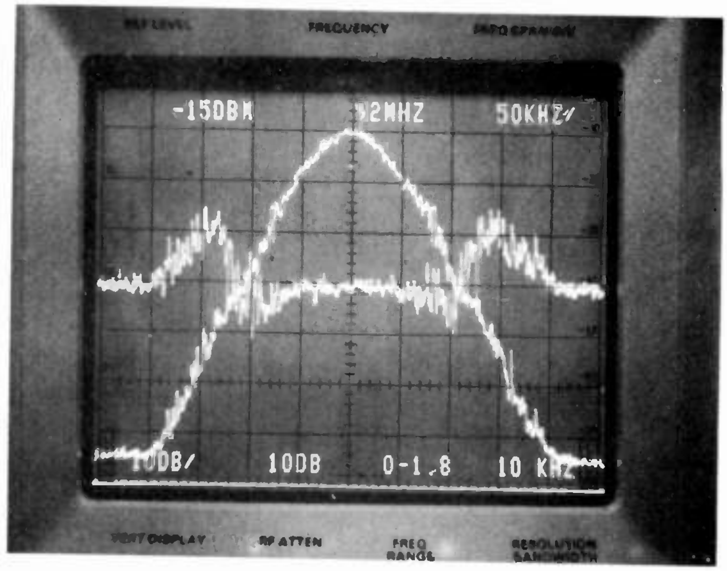
CCIR STANDARD NOISE SPECTRUM

VIEW A:

RF SPECTRUM OF MODULATED FM CARRIER PER CCIR RECOMMENDATION 796, REDUCED 20% FOR 10% PILOT AND 10% FOR 67 KHZ SCA. COMPOSITE DEVIATION IS 32 KHZ QUASI-PEAK AS SPECIFIED IN STANDARD.

VIEW B-A:

DIFFERENCE IN RF SPECTRUM BETWEEN STEREO (L+R) CCIR NOISE (REDUCED 10% FOR 10% PILOT) AND VIEW A (DESCRIBED ABOVE).



VIEW A:

RF SPECTRUM OF MODULATED FM CARRIER PER CCIR RECOMMENDATION 796, REDUCED 20% FOR 10% PILOT AND 10% FOR 92 KHZ SCA. COMPOSITE DEVIATION IS 32 KHZ QUASI-PEAK AS SPECIFIED IN STANDARD.

VIEW B-A:

DIFFERENCE IN RF SPECTRUM BETWEEN STEREO(L+R) CCIR NOISE (REDUCED 10% FOR 10% PILOT) AND VIEW A (DESCRIBED ABOVE).

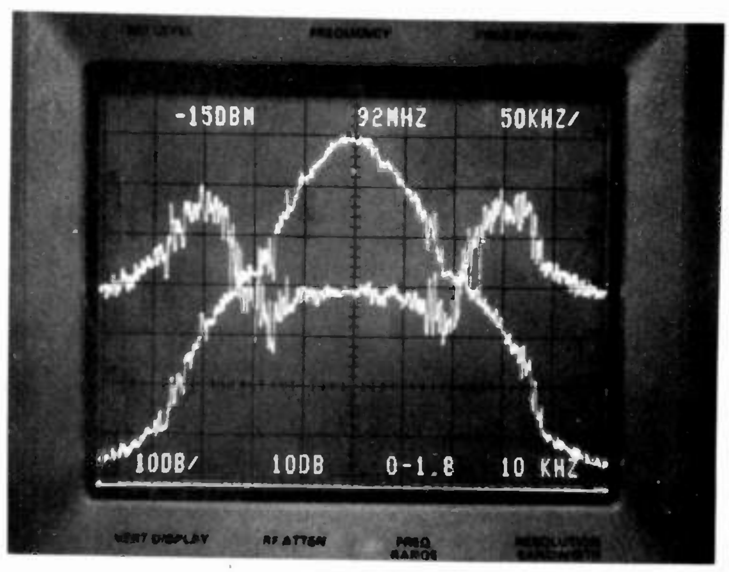
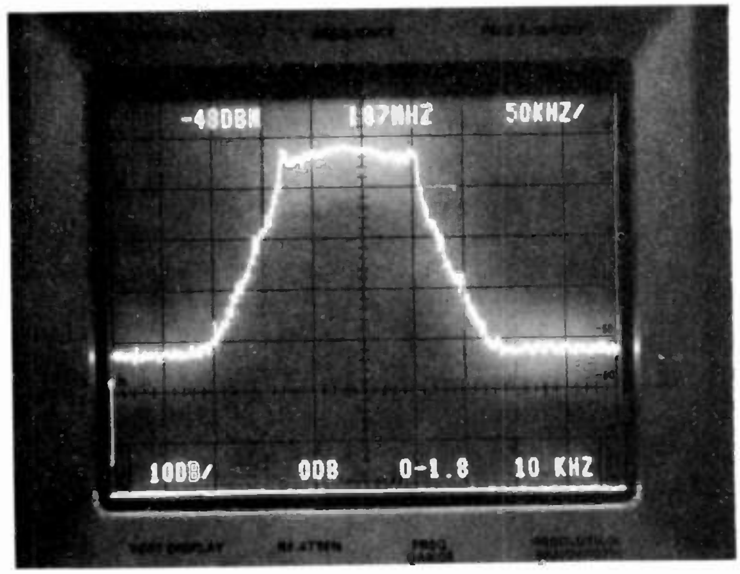


Figure 1

SPN-DEVELOPED RF NOISE SPECTRUM
AND RF SPECTRUM OF TWO WASHINGTON, D.C. AREA FM STATIONS

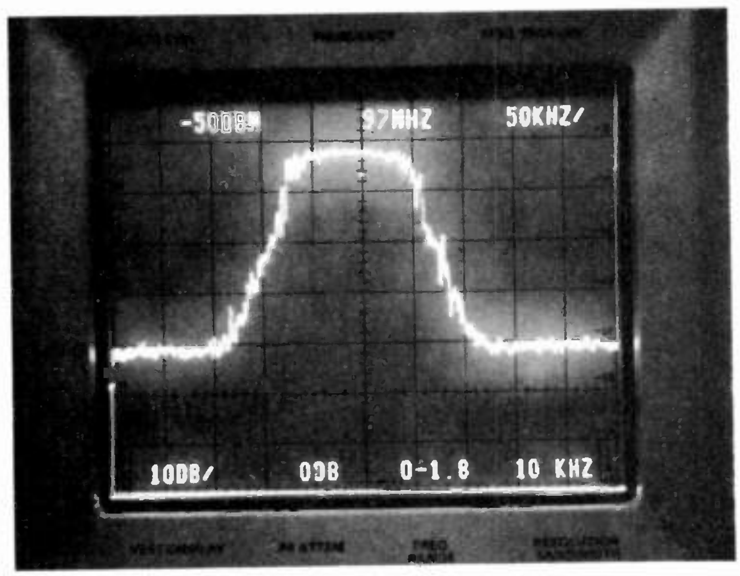
WRQX-FM

MEASURED OFF NPR'S MATV SYSTEM IN DOWNTOWN WASHINGTON USING ROOFTOP DIRECTIONAL ANTENNA. NOISE LEVEL AT -52 DB FROM CABLE BOOSTER AMPS. SAMPLE PERIOD WAS ONE MINUTE AND WAS DIGITALLY-STORED WITH PEAK RESPONSE DETECTOR IN "MAX HOLD" MODE.



WASH-FM

SAME MEASUREMENT CHARACTERISTICS AS ABOVE.



NPR-DEVELOPED NOISE

SAME MEASUREMENT CHARACTERISTICS AS ABOVE, HOWEVER, GAUSSIAN NOISE WAS ADDED TO THE RF PRESENTATION TO APPROXIMATE THE EFFECT OF AMPLIFIER NOISE ABOVE. THE NOISE WAS NOT PRESENT DURING PROTECTION RATIO MEASUREMENTS.

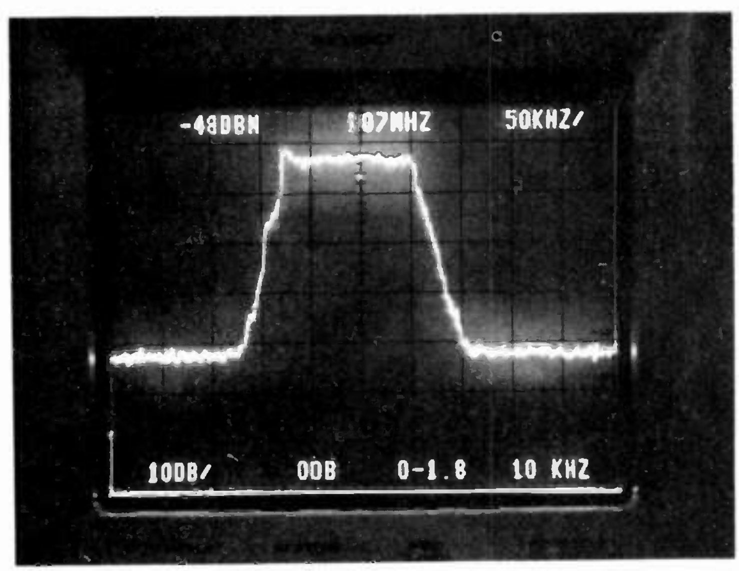


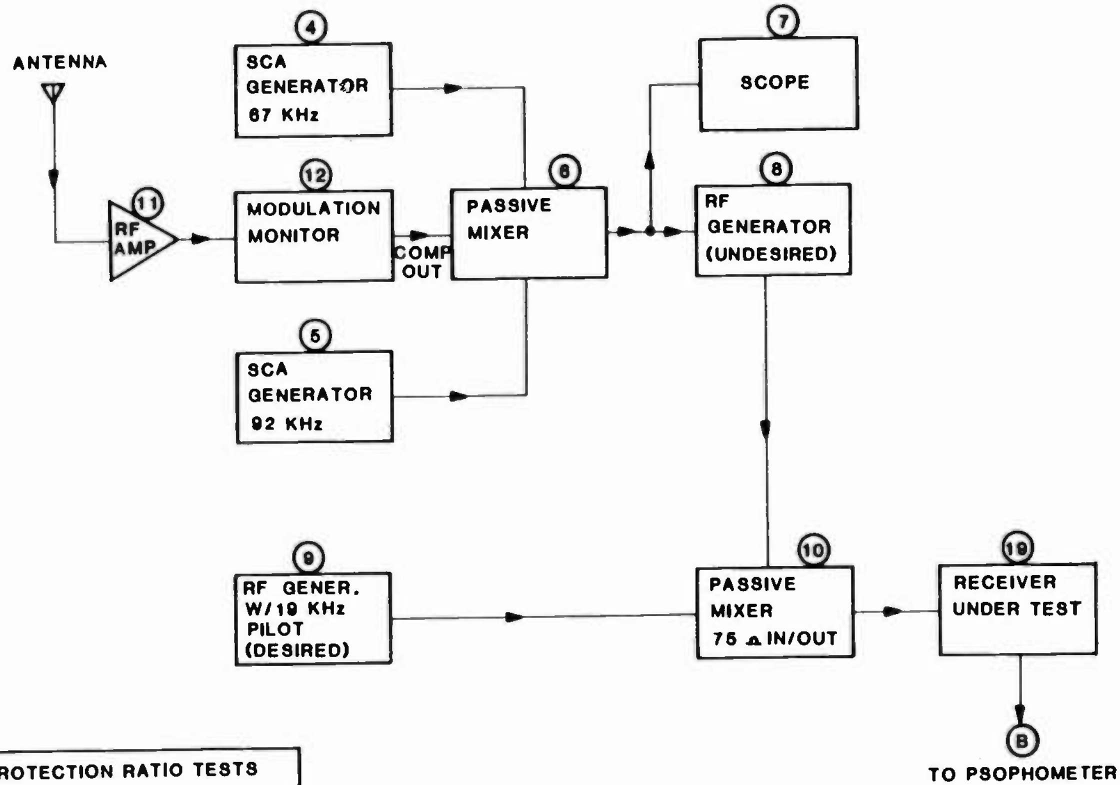
Figure 2

Appendix B

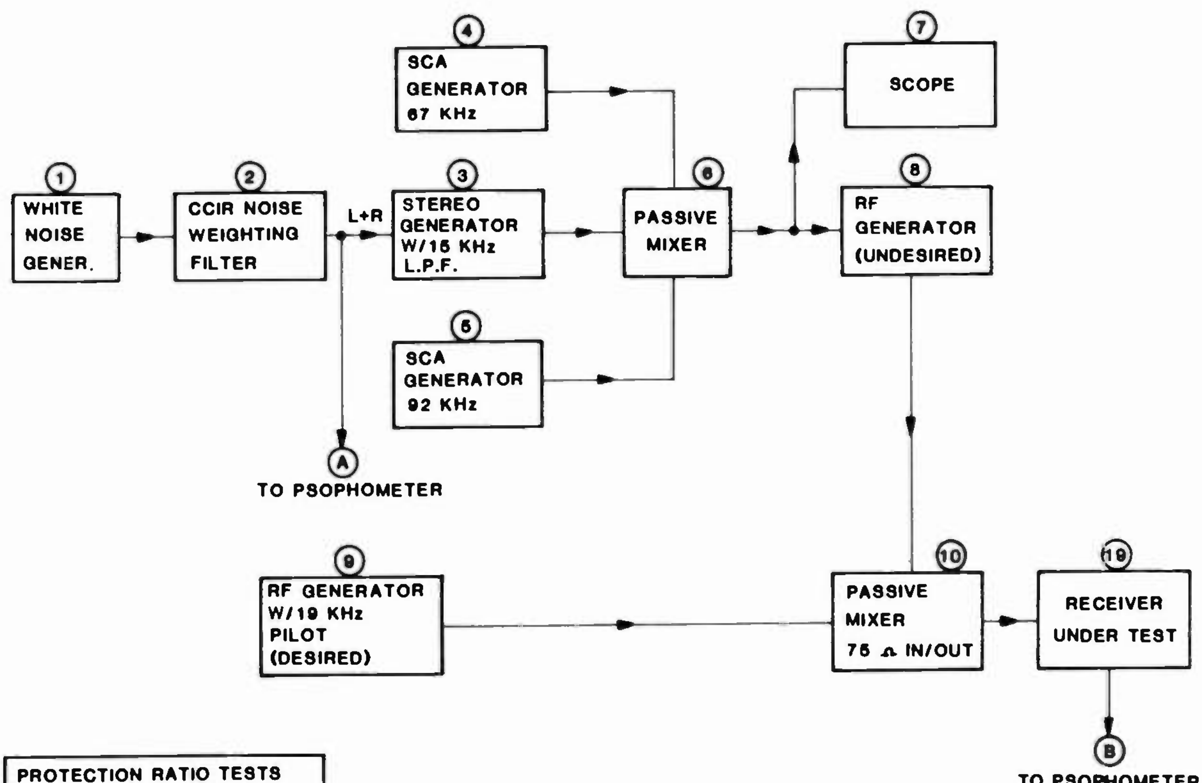
Table 1

List of Test Equipment

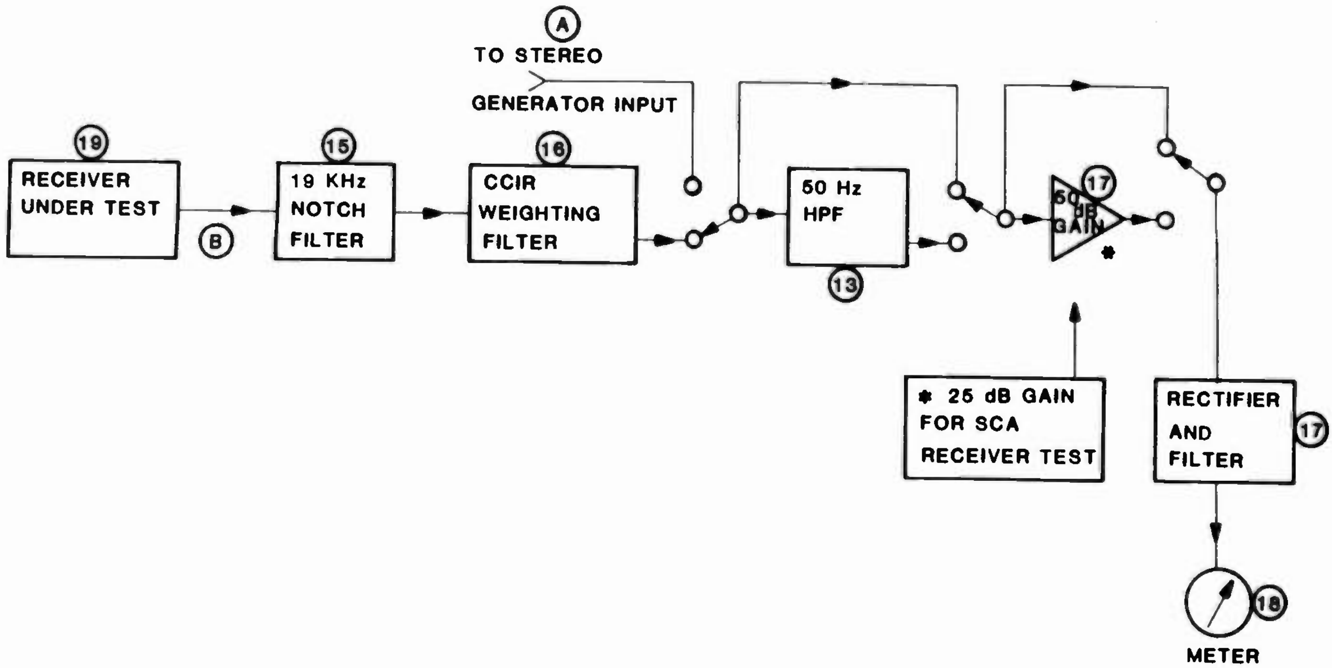
Number	Description	Make/Model
1	White/Pink Noise Generator	Ivie IE-20B
2	CCIR Noise Weighting Filter	Orban Parametric Eq.622B
3	Stereo Generator	Rood SC203
4	67 kHz Subcarrier Generator	Moseley SCG-8
5	92 kHz Subcarrier Generator	Moseley SCG-8
6	Passive Mixer (Built at NPR)	see Figure E
7	Oscilloscope	Tektronix SC-502
8	"Undesired" RF Generator	Boonton 102D
9	"Desired" RF Generator	Boonton 103D
10	75 ohm Passive Mixer	Mini-Circuits Lab ZSC-2375B
11	RF Amplifier	Blonder-Tongue MVB-25
12	Modulation Monitor	QEI-691
13	50 Hz HPF 1 Pole RC (NPR)	Figure E
14	Clipper - (NPR)	Figure F
15	19 kHz Notch Filter (NPR)	Figure G
16	CCIR 468 Noise Weighting Filter	Dolby Labs 98A
17	50 dB Gain Block/Rectifier and Filter	Selco 34A
18	Meter	Selco (BBC Stand. 4297)
19	Receivers Tested (FM)	GE 3-5256 Technics SA-505 Pioneer KP-3500 Sony STR-V45 Magnavox PA1839



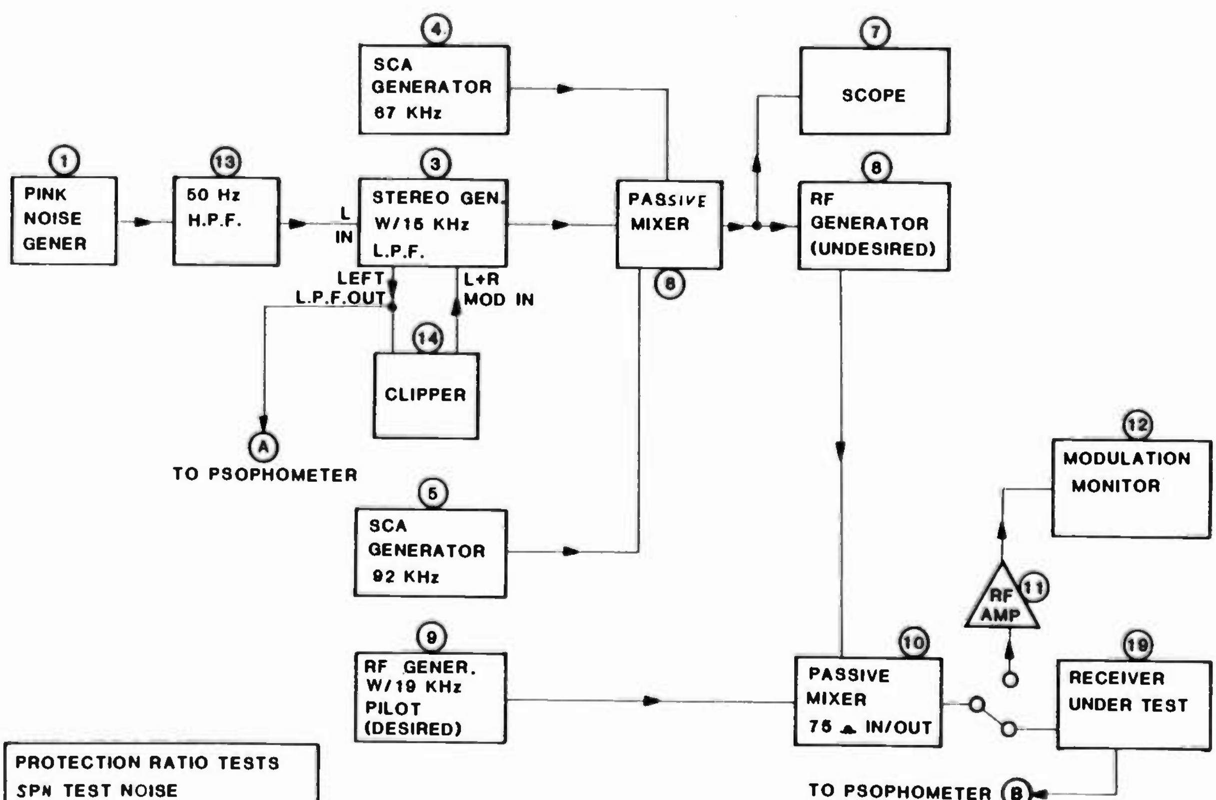
PROTECTION RATIO TESTS
REAL-WORD NOISE
FIGURE B



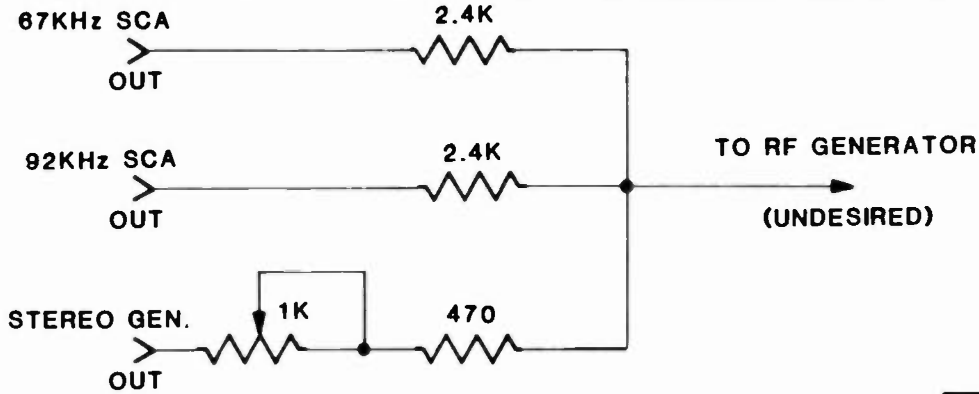
PROTECTION RATIO TESTS
CCIR STANDARD NOISE
FIGURE A



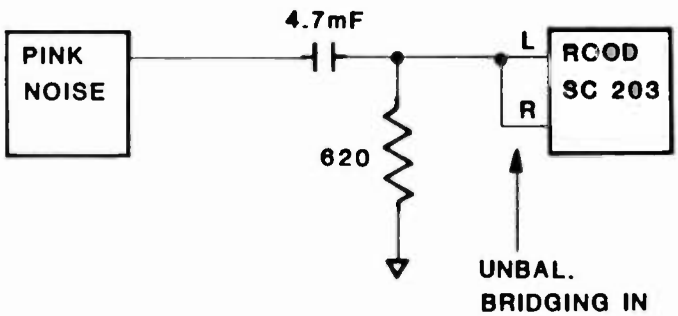
PROTECTION RATIO TEST
NPR–PSOPHOMETER
FIGURE D



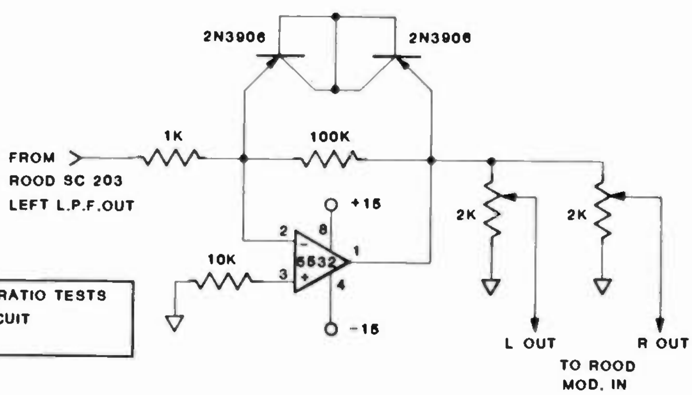
PROTECTION RATIO TESTS
SPN TEST NOISE
FIGURE C



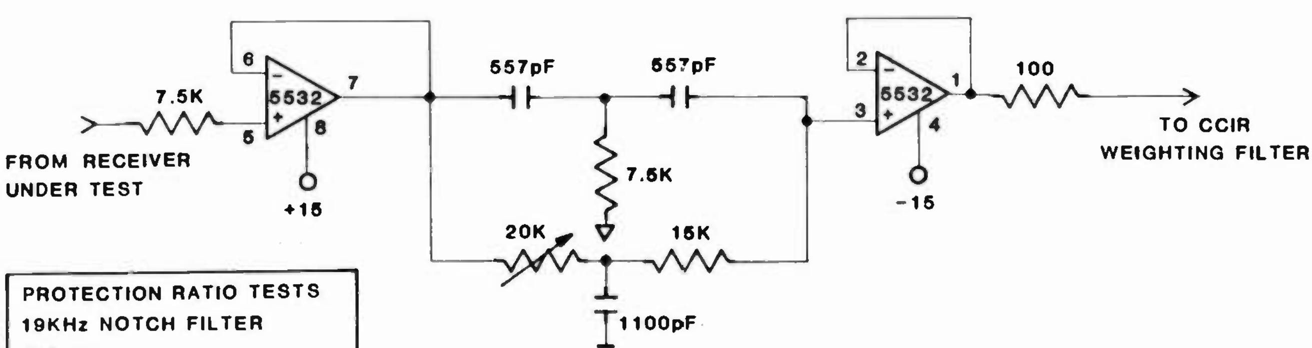
PROTECTION RATIO TESTS
PASSIVE MIXER
FIGURE E



PROTECTION RATIO TESTS
50 Hz HPF
FIGURE E



PROTECTION RATIO TESTS
CLIPPER CURCUIT
FIGURE F



PROTECTION RATIO TESTS
19KHz NOTCH FILTER
FIGURE G

TRANSMITTER PERFORMANCE REQUIREMENTS

FOR SUBCARRIER OPERATION

John T.M. Lyles - Mukunda B. Shrestha

Broadcast Electronics, Inc.

Quincy, Illinois

I. INTRODUCTION.

Stereophonic subcarrier operation in FM broadcasting has made it the profitable radio medium of today. SCA subcarriers, on the other hand, have not attained this universal acceptance, partly because of crosstalk with the main programming channels, and the desire to maintain maximum modulation levels. The key to optimal subcarrier performance is to examine the entire transmitting system and correct the bandwidth and distortion limitations of all stages through the chain. This discussion will cover design techniques and insights realized during the development of a new line of FM transmitters. Specific improvements in FM exciter linearity, stereo and SCA generator spectral purity, and amplifier bandwidth and stability have allowed new levels of performance with simplified field adjustment.

Subcarriers commonly use either AM-on-FM modulation (as in stereo) or FM-on-FM modulation (as in traditional SCAs). Both processes are complicated by the principle that frequency modulation requires transmission of an infinite number of sidebands for perfect demodulation of information. In practice, however, the information of FM can be carried in a broadcast channel with acceptably low distortion. There are two different frequency bands involved: 1) The composite baseband, which contains the modulating audio plus one or more amplitude or frequency modulated subcarriers, and 2) the FM carrier frequency band of the transmitter. Waveform linearity, amplitude bandwidth, and phase linearity must be maintained at acceptable limits in the baseband chain from the audio inputs through subcarrier generators to the FM exciter modulated oscillator. From here, the FM carrier is usually amplified in a series of class C nonlinear power amplifiers, where most amplitude variation is removed. However, the amplitude and phase responses of all the networks which follow must also be controlled to minimize degradation. Figure 1b shows the effects of a narrowband

RF filter on the RF spectrum of a composite signal consisting of a stereophonic subcarrier modulated only on the left channel with 4.5 kHz and with a 67 kHz unmodulated SCA subcarrier. The only distortion evident on the spectrogram is the loss of some side-bands greater than 150 kHz from the center frequency and amplitude differences between the lower and upper sideband pairs. Figure 1d shows the corresponding effects observed on the demodulated baseband spectrum for the same signal. Note the creation of many undesired intermodulation terms which could cause crosstalk into both the stereophonic and SCA subcarrier bands.

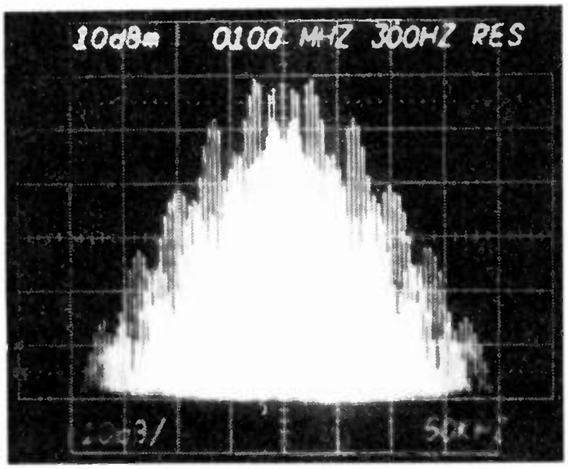


FIGURE 1a.
WIDEBAND RF SPECTRUM TO DEMODULATOR

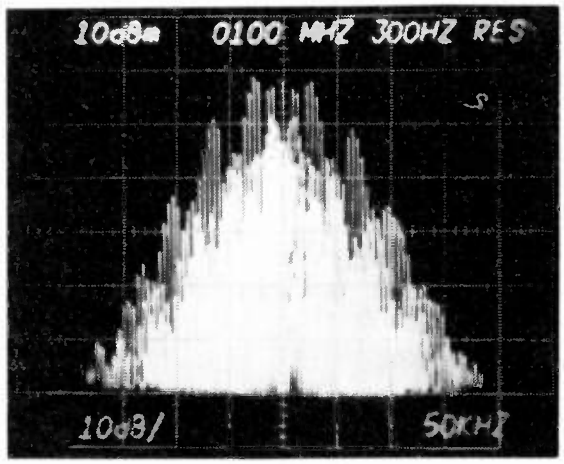


FIGURE 1b.
BANDWIDTH LIMITED RF SPECTRUM TO DEMODULATOR

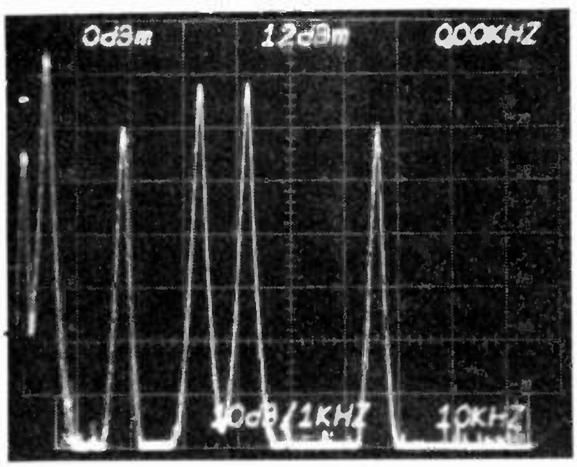


FIGURE 1c.
DEMODULATED BASEBAND SPECTRUM FOR WIDEBAND RF SPECTRUM

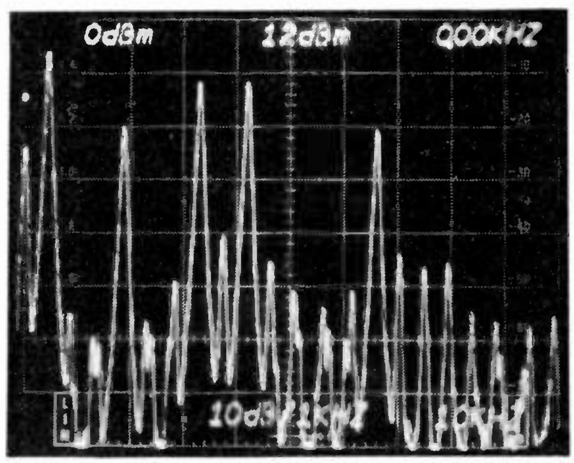


FIGURE 1d.
DEMODULATED BASEBAND SPECTRUM FOR BANDWIDTH LIMITED RF SPECTRUM SHOWING DISTORTION PRODUCTS

II. ELEMENTS OF FM BROADCASTING SYSTEM WHICH AFFECT SUBCARRIERS.

The following components affect subcarrier performance through the system:

1. SCA generator
2. Stereo generator
3. FM Exciter

4. Composite STL, when used
5. All transmitter RF amplifiers
6. Antenna system, including diplexers and combiners
7. Multipath and other propagation phenomena
8. Receiver antenna, IF passband, and demodulators

In this paper we will concentrate on those components which are part of the transmitting equipment. Information pertaining to receivers, multipath, antennas, and combiner effects can be found in other articles and reports mentioned at the end.

2.1 SCA Generator.

The SCA generator frequency modulates a subcarrier with bandlimited audio information or data. Audio frequency shift keying and direct FSK for data channels can be used with some new generators. Multiple narrowband or a single wideband SCA can be used in the baseband. The audio frequency response of the standard narrowband SCA must be tailored to prevent the FM sidebands of the SCA from overlapping the stereophonic sidebands by greater than -60 dB. With a 67 kHz SCA, to minimize crosstalk from the SCA into the stereophonic subchannel, the audio should be bandlimited to 4.3 kHz and the peak deviation of the subcarrier limited to 6 kHz or less. Various Bessel function tabulations have been prepared for use with 67 kHz FM SCAs. They can be found in the NAB handbook. Any system utilizing new center frequencies, multiple subcarriers, or different modulation forms will require careful spectral analysis of the baseband to assure minimum interference and maximum compatability with stereo.

The SCA generator should generate a subcarrier sinewave with low harmonic distortion, requiring minimal bandpass filtering; bandpass filtering of FM can generate additional unwanted intermodulation products in the demodulated SCA information. The audio input should be conditioned by a lowpass filter, as mentioned above.

A modern high-performance SCA generator, shown simplified in Figure 2, uses a linear VCO IC. It produces a sinewave at any frequency from 39 to 95 kHz with less than 0.5 percent distortion. A 100 kHz lowpass filter is used on the output. The audio input is conditioned with a 6th order lowpass filter which is 3 dB down at 4.3 kHz. This filter may be bypassed for wider bandwidth SCA's or different preemphasis. A DC coupled data input is included for direct frequency shift keying. Note that the subcarrier output is faded on and off at a controlled decay rate rather than switched to prevent squelch problems at the SCA receiver.

2.2 Stereo Generator.

The stereo generator must have good 38 kHz subcarrier suppression with modulation applied. 38 kHz leakage may cause additional 76 kHz regeneration in the system. It must also have good 76 kHz (second harmonic) suppression. The second harmonic modulation sidebands should be attenuated as well, because they add crosstalk into the SCA subchannel. For an SCA signal to noise ratio of -60 dB the stereophonic harmonics should be suppressed by -70 dB. This number allows for some degradation through the entire system. Manufacturers of stereo generators have traditionally chosen either linear or switching modulators.

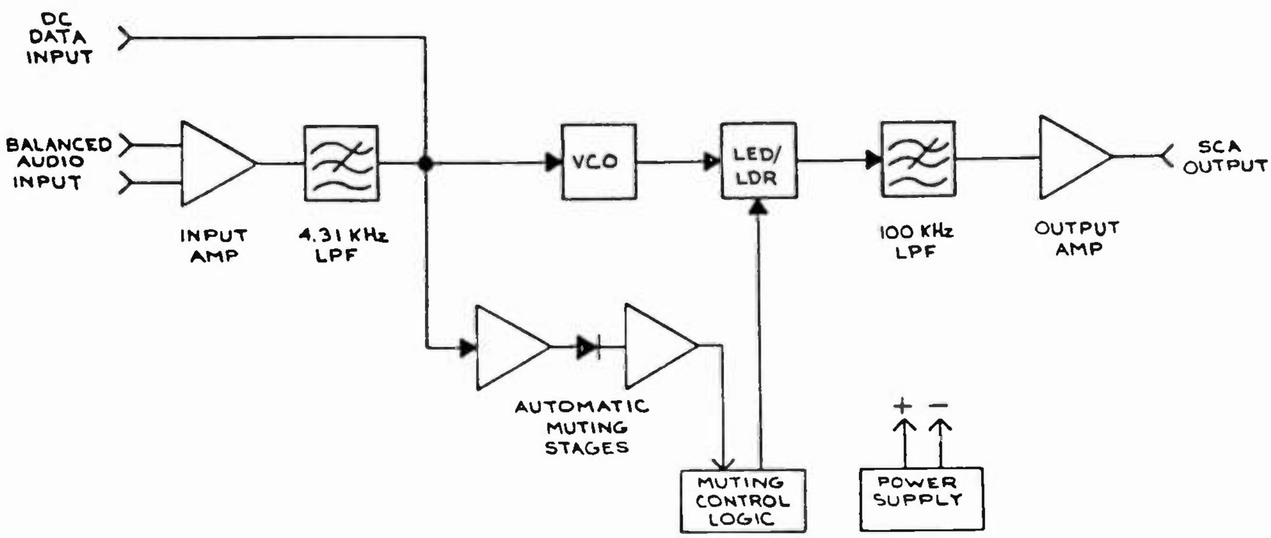


FIGURE 2. SIMPLIFIED BLOCK DIAGRAM OF FC-30 SCA GENERATOR

The linear modulators may use balanced analog circuits with trimmer controls for 38 and 76 kHz nulls. These adjustments should be maintained if excessive 76 kHz becomes evident using spectrum analysis of the baseband. The switching modulator is popular because of less critically toleranced components and good long term stability. However, the modulator waveform usually requires filtering with a steep cutoff 53 kHz lowpass filter. This filter is not a trivial design, because passband amplitude ripple and phase non-linearities cause degraded high frequency separation by adversely affecting the upper sideband of the L-R subcarrier.

The stereo generator shown in Figure 3 uses a digital stairstep generator to synthesize the subcarrier and pilot simultaneously, eliminating any pilot phase adjustment. Appropriate components are added in the synthesis to approximate a sinewave with lower harmonic content. The composite lowpass filter then has a gradual rolloff, with the -3 dB point beyond 100 kHz. Separation is better than 50 dB at 15 kHz. In this design, subcarrier suppression is specified at -75 dB and the 76 kHz sidebands are -80 dB below 100% modulation. 57 and 95 kHz, the third and fifth pilot harmonics, are suppressed -80 dB or more.

Audio input lowpass filters are necessary in all stereo generators. These "brickwall" filters protect the pilot and subcarrier by greatly attenuating audio components above 15 kHz. Some designs have filters which, due to poor passband group delay, ring or overshoot on transient program waveforms. This overshoot can be measured as overmodulation. Without audio lowpass filtering, increased spectrum occupation and spillover into the SCA band occurs. If the pilot level is observed fluctuating during modulation, defective filtering may be suspected. The FS-30 generator uses carefully aligned 5-pole active lowpass filters with controlled delay equalization to keep overshoot below 2 dB, while providing adequate protection to the pilot, stereophonic subcarrier, and SCA subcarrier.

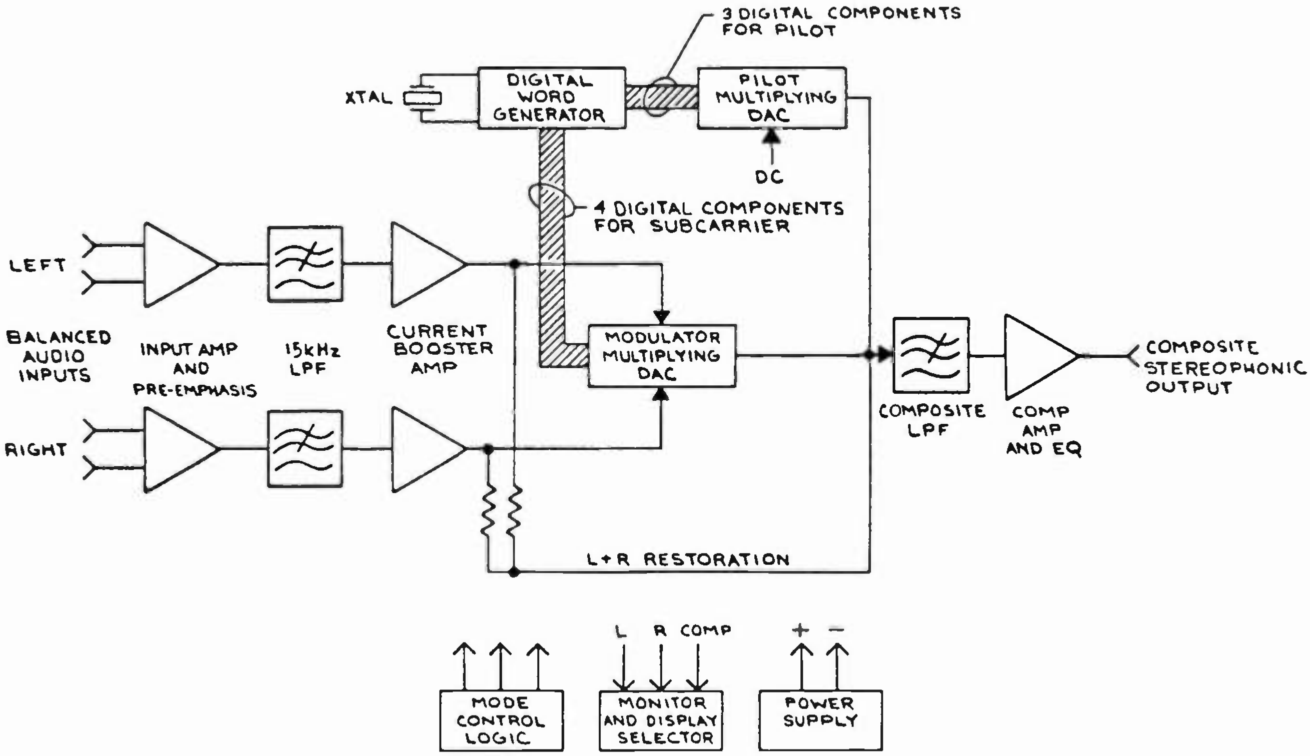


FIGURE 3. SIMPLIFIED BLOCK DIAGRAM OF FS-30 STEREO GENERATOR

2.3 FM Exciter.

The exciter characteristics are foremost in importance for good subcarrier performance. The frequency modulated oscillator in most current units is a direct varicap diode VCO. The voltage-to-capacitance transfer function of these devices is not linear over the wide range used, so linearization may be necessary. Non-linearities in the FM oscillator can, by altering the waveform of the baseband signal, create distortion in the demodulated output at the receiver. Secondary effects of this distortion may include stereo crosstalk into SCA. Modulator linearization using a piecewise approximation pre-distortion network has reduced harmonic and intermodulation distortion to less than 0.05% in the FX-30 exciter (Figure 4). All exciter stages after the oscillator operate as broadband amplifiers with no bandwidth limitations.

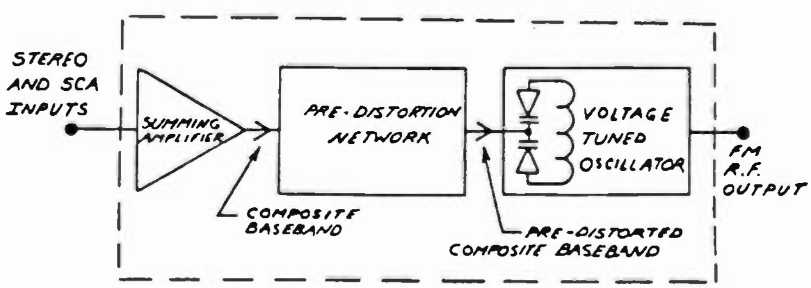


FIGURE 4. SIMPLIFIED BLOCK DIAGRAM OF LINEARIZED FM MODULATOR

2.4 Composite STL.

The composite STL is really a transmission subsystem within a system. Its transmitter requirements are identical to those of the main FM exciter and power amplifiers detailed in this report. For minimum degradation of the composite signal it is recommended that all SCA channel information be fed into the exciter at the transmitter. Telco landlines or another narrowband link can usually handle the program bandwidths of SCAs. This reduces the technical burden of maintaining very low intermodulation performance through the entire STL modulation/demodulation process. However, the STL should have a flat bandwidth through 53 kHz for minimum stereophonic subcarrier degradation. For stereophonic separation of 50 dB, it is necessary to maintain a composite amplitude flatness of ±0.04 dB and phase linearity within ±0.2 degrees through 53 kHz in the baseband. Many exciters and STLs cannot meet this requirement. The stereo generator should be engineered to compensate for this deficiency through the use of a built-in composite equalizer with low and high frequency adjustments. The range of amplitude correction of the FS-30 stereo generator is shown in Figures 5a and 5b. These figures show the maximum boost and cut with the low (Figure 5a) and high (Figure 5b) frequency controls at both ends of their ranges.

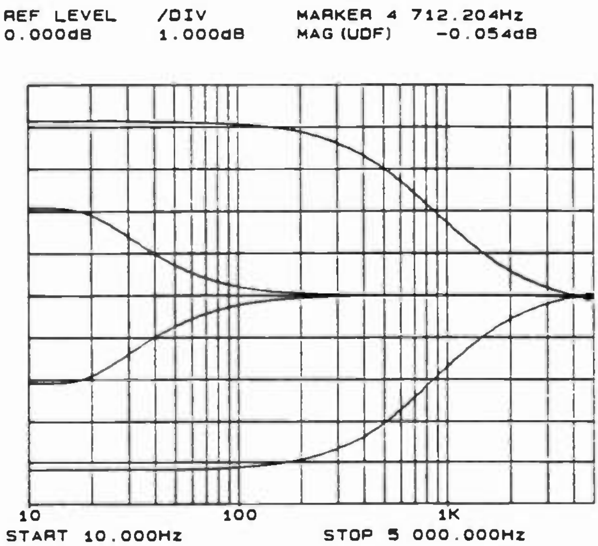


FIGURE 5a.
RANGE OF FS-30 LOW FREQUENCY EQ

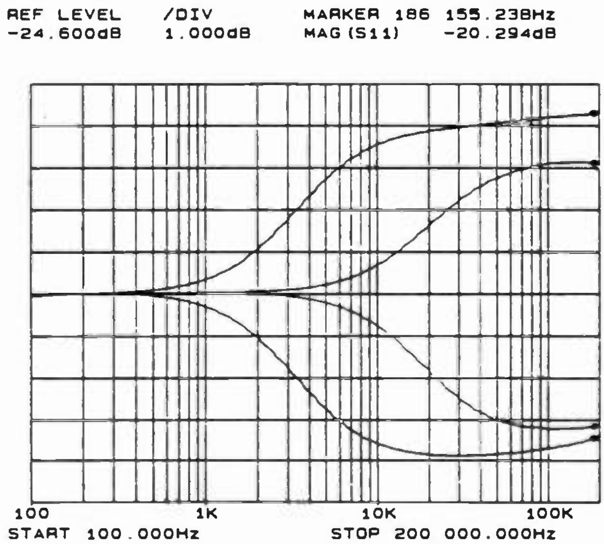


FIGURE 5b.
RANGE OF FS-30 HIGH FREQUENCY EQ

2.5 RF Power Amplifiers.

The remainder of the FM transmitter consists of a chain of power amplifiers, each having from 6 to 20 dB of power gain. Ideally, the transmitter should have as wide a bandwidth as practical with a minimum of tuned stages. Broadband solid-state amplifiers are preferred to eliminate tuned networks in the RF path. A new generation of class C bipolar and MOSFET broadband amplifier stages exhibit both high efficiency and greater than 20 percent bandwidth to cover the FM broadcast band. These solid-state amplifiers may be combined for higher power. Tuned output bandpass filters may still be necessary when operated in a dense RF environment to prevent intermodulation from being generated in the PA modules.

Higher powered transmitters in the multi-kilowatt range may use a single tube PA stage with high efficiency. The dollars/watt economics of single-tube transmitters outweigh the bandwidth benefits of solid-state transmitters at the higher power levels. Design improvements in tube-type power amplifiers have concentrated on improving bandwidth, reliability, and cost effectiveness.

III. POWER AMPLIFIER CIRCUIT DESIGN.

3.1 Bandwidth Considerations.

As mentioned earlier, the FM signal theoretically occupies infinite bandwidth. In practice, however, truncation of the insignificant sidebands (typically less than 1 percent of the carrier) makes the system practical by accepting a certain degree of signal degradation. The input and output tuned circuits of the PA limit the bandwidth of the FM signal. The degree of bandwidth reduction is a design constraint which affects the gain and efficiency in all tuned PA stages.

The bandwidth of an amplifier is determined by the load resistance across the tuned circuit and the output or input capacitance of the amplifier. Thus, for a single-tuned circuit, the bandwidth is proportional to the ratio of reactance to resistance:

$$BW \cong \frac{1}{2\pi R_L C} \cong \frac{X_C}{R_L} \qquad \text{(eq. 1)}$$

where BW = bandwidth between half-power points
R_L = load resistance (appearing across tuned circuit)
C = total capacitance of tuned circuit (includes stray capacitances and output or input capacitances of the tube)
X_C = capacitive reactance

The load resistance is directly related to the RF voltage swing on the tube element. For the same power and efficiency, the bandwidth can be increased if the capacitance is reduced.

3.2 Grounded-Grid Versus Grid-Driven Operation.

Since the input capacitance of tube amplifiers in a grounded-grid configuration is smaller than that of a grid-driven configuration by as much as 50 percent, an investigation was carried out in 1982 to determine the advantages of using a grounded-grid circuit for a tetrode tube amplifier. Input capacitances of typical tubes are shown in Table 1.

TABLE 1

TUBE TYPE	Cin (pF) Grounded Grid	Cin (pF) Grounded Cathode
4CX3000A	67	140
4CX3500A	58.5	111
4CX5000A	53	115
8990/4CX20,000A	83	190

Prototype input circuits were developed for grounded-grid and grid-driven operation of a 5 kilowatt PA using the Eimac 4CX3500A tetrode. A series of measurements were made to evaluate the performance of grounded-grid versus grid-driven operation of the tetrode PA with respect to gain, efficiency, amplitude bandwidth, phase bandwidth, and synchronous AM under equivalent operating conditions. Measurements were made at normal and reduced plate voltage for both saturated and unsaturated PA operation. Saturation is noted when little change in output power occurs with increasing drive power. Maximum efficiency occurs at this point. The PA gain and efficiencies are tabulated in Table 2. Swept amplitude and phase responses of the different PA configurations are shown in Figures 6a thru 6d.

The significant findings of the tests and measurements are as follows:

1. When driving the PA into saturation, the bandwidth of the PA is limited by the output cavity bandwidth in the grounded-grid amplifier. The PA bandwidth in the grid-driven amplifier is limited by the input circuit Q, which is basically determined by the extent of swamping resistance used. PA bandwidth under saturation can be improved in either configuration by reducing the plate voltage as evident from equation (1). However, this involves a trade-off in efficiency with a smaller voltage swing. For example, in the grid-driven saturated configuration a 25 percent bandwidth improvement was observed with 1.4 dB loss of PA gain and 2.3 percent efficiency loss with reduced plate voltage.

2. When the PA is not driven into saturation, the grounded-grid amplifier does not appear to give any bandwidth improvement over the grid-driven amplifier at the 0.25 dB points (see Figures 6c and 6d). At the 3 dB points however, there is a slight ($\cong$ 15%) improvement in bandwidth when using the grounded-grid unsaturated PA.

3. A grounded-grid saturated PA improves bandwidth over a grid-driven saturated PA at the expense of amplifier gain. A 15 percent improvement in the PA bandwidth was observed while losing 3 dB of the amplifier gain. For a grid-driven amplifier, a 25 percent reduction of input circuit resistive swamping results in the same 15 percent bandwidth improvement at the expense of only 0.5 dB in gain.

4. The phase linearity in the 0.5 dB bandwidth appears to be better using the grid-driven PA. The grounded-grid PA exhibits a more nonlinear phase slope within the passband, yet has a wider amplitude bandwidth. This phenomenon is due to interaction of the input and output circuits because they are effectively connected in series in the grounded-grid configuration. The neutralized grid-driven PA provides more isolation of these networks, so they behave like independent filters.

TABLE 2

MEASUREMENTS FROM 5KW PA (GD) GRID-DRIVEN (GG) GROUNDED-GRID CONFIGURATION

CONFIGURATION	GD	GG	GD	GG	GD	GG	GD	GG
PA CONDITION	---- SATURATED ----				---- UNSATURATED ----			
RF POWER OUTPUT (W)	4900	5000	4800	4900	3225	3350	3325	3200
PLATE VOLTAGE (V)	5220	5200	4500	4480	5320	5315	4550	4600
PLATE CURRENT (A)	1.27	1.26	1.49	1.4	0.81	0.9	1.08	1.0
DRIVE POWER (W)	140	280	190	340	70	170	70	175
EFFICIENCY (%)	73.9	72	71.6	72.7	74.8	66.5	67.7	65.8
GAIN (dB)	15.4	12.5	14.0	11.6	16.6	13.0	16.8	12.6
SYNCH. AM (dB)	-54	-56	-56	-58	-46	-48	-51	-52

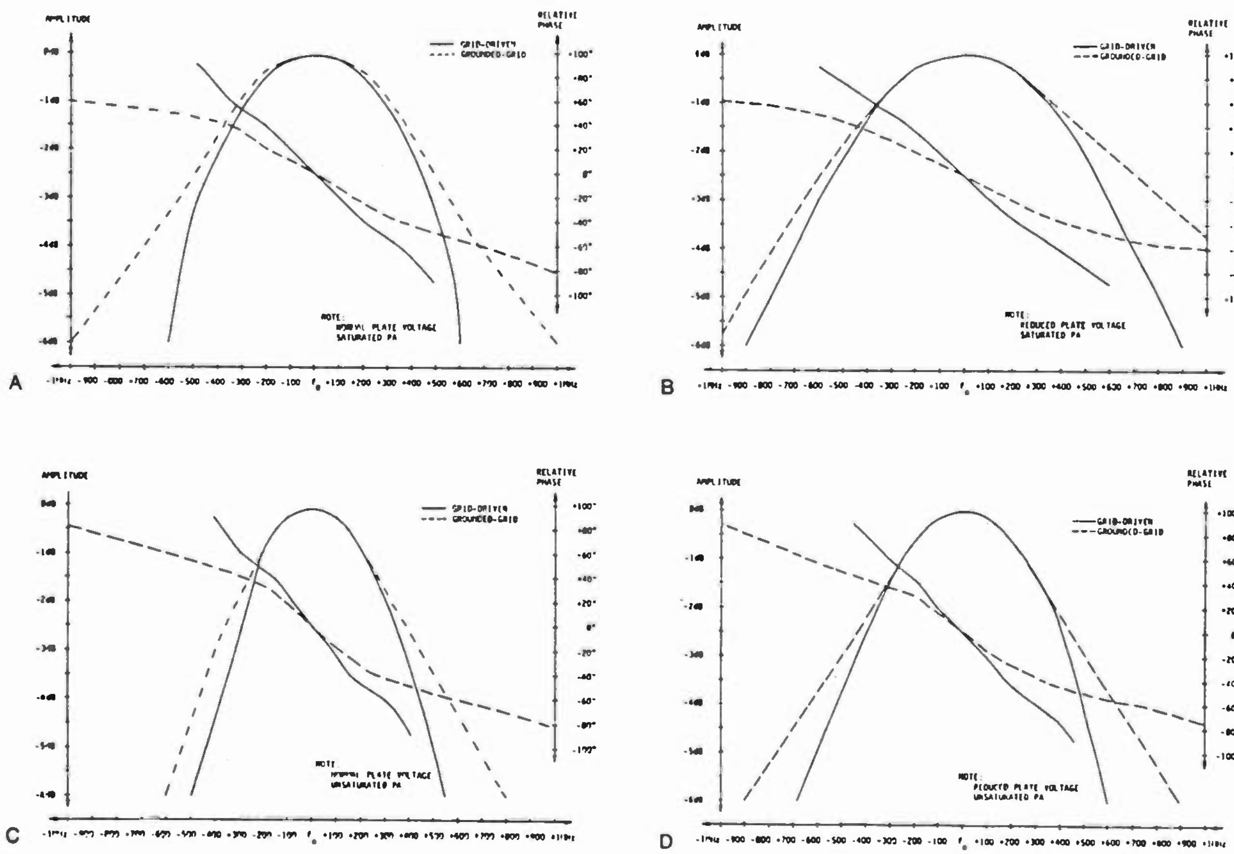


FIGURE 6. MEASURED AMPLITUDE AND PHASE RESPONSES OF GRID-DRIVEN AND GROUNDED-GRID TETRODE POWER AMPLIFIERS

In view of the findings mentioned above (in particular item No. 3) the use of a tetrode in a grounded-grid configuration did not appear to be feasible economically. An additional intermediate power amplifier would have been required to fulfill the higher drive power requirements, thereby affecting the overall cost and reliability of the transmitter. The decision was made to use a grid-driven PA for our FM-3.5A and FM-5A transmitters. Bandwidth limitations of the grid-driven PA were overcome by swamping the input circuit and by developing a novel impedance-matching device to achieve optimum transfer of power from the driver stage into the PA. The loss of PA gain due to swamping was limited to 0.5 dB, while achieving bandwidth nearly equivalent to a grounded-grid amplifier, yet providing a more linear phase response.

3.3 Broadband Impedance-Matching.

A broadband impedance matching circuit was developed to match the high grid input impedance of a tetrode RF power amplifier to the 50 Ohm impedance of a solid-state driver. The conventional matching circuits used in transmitter applications are generally of the type known as L, π , or T networks. All of these circuits require interactive adjustment of one or more circuit elements to provide a satisfactory impedance match for each frequency and RF power level.

The new impedance-matching circuit developed for the FM-3.5A and FM-5A transmitters consists of a combination of series inductor (L) and shunt capacitor (C) circuit elements, implemented as a printed circuit with inductors and capacitors etched on a copper-clad laminate. Multiple LC sections are used to match the 50 Ohm source impedance to the high input impedance of the grid-driven RF power amplifier. The impedance-matching device is shown in Figures 7 and 8.

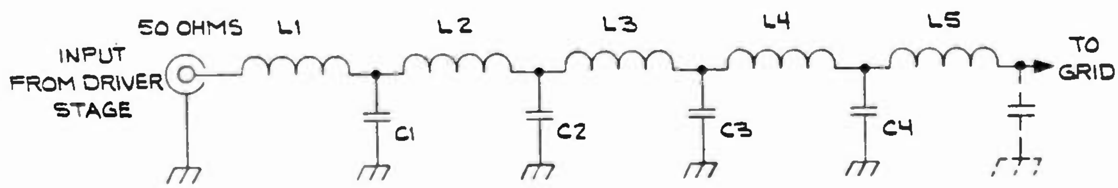


FIGURE 7. CIRCUIT DIAGRAM OF PA INPUT IMPEDANCE MATCHING DEVICE

This impedance-matching circuit improves the transmitter's operation and maintainability, compared to previous methods. A single tuning control in the input circuit is sufficient to tune and match the 50 Ohm driver impedance to the high input impedance of the grid over the 88-108 MHz FM broadcast band with a 4:1 range of RF power levels. The input-matching circuit eliminates separately mounted components which can be microphonic (sensitive to vibration) due to mechanical instability. By incorporating this new impedance-matching device, we have been able to improve the bandwidth, reliability and stability of the transmitter.

The typical performance figures of the FM-3.5A and FM-5A transmitters with regard to PA bandwidth, efficiency, gain, and synchronous AM (measured for 3500W and 5000W power outputs, respectively) are presented in Table 3.

FIGURE 8. PHOTO OF INPUT MATCHING DEVICE. TUBE SOCKET GRID RING CONNECTION AT RIGHT OF CENTER, 50 OHM INPUT AT LEFT

TABLE 3

TYPICAL PERFORMANCE OF TRANSMITTER PA's

BE Model	3dB Bandwidth	Efficiency	Gain	Sync. AM
FM-3.5A	1.2 MHz	75%	14.5 dB	-47 dB
FM-5A	1.3 MHz	75%	15.0 dB	-51 dB

3.4 Power Amplifier Cavity.

The vacuum-tube power amplifier is constructed in an enclosure containing distributed tank circuit elements designed for minimum loss. The efficiency of the PA depends on the RF plate voltage swing, the plate current conduction angle, and the cavity efficiency. The cavity efficiency is related to the ratio of loaded and unloaded Q as follows:

$$N = \left(1 - \frac{Q_L}{Q_u}\right) \times 100 \qquad \text{(eq. 2)}$$

where N = efficiency in percent
Q_L = loaded Q of cavity
Q_u = unloaded Q of cavity

Loaded Q depends on the plate load impedance and output circuit capacitance. Unloaded Q depends on the cavity volume and the RF resistivity of the conductors due to skin effects. A high unloaded Q is desirable, as is a low loaded Q, for best efficiency. As Q goes up, the bandwidth decreases. For a given tube output capacitance and power level, loaded Q decreases with plate voltage or with increasing plate current. This explains the improved bandwidth for the reduced high voltage measurements in Table 2 and Figures 6b and 6d.

Other methods popular in improving the bandwidth of PA output circuits include minimizing added capacitance, as manufacturers of quarter-wave cavities have attempted. The ideal case would be to resonate the plate capacitance alone with a "perfect" inductor, but practical quarter-wave cavities require either the addition of a variable capacitor or a variable inductor using sliding contacts for tuning. An inherent mechanical and electrical compromise in these designs has always been the requirement for a plate blocking capacitor and the presence of maximum RF current at the grounded end of the line where the conductor is nonhomogeneous. A new approach to VHF power amplification uses a folded half-wave cavity design. This is shown compared to conventional designs in Figure 9. The half-wave line is tuned without the use of variable capacitors or sliding contacts. The blocking capacitor is unnecessary and the high current point is located in the central area of the tank line where no joints, fasteners, or obstructions occur. This design is inherently more reliable and, due to the folded nature, requires only slightly more physical height than the quarter-wave design.

The bandwidth of the PA cavity is optimized by a choice of the highest characteristic impedance mechanically allowable. The center conductor is sized for minimum impedance discontinuity and is directly clamped to the outer surface of the anode fins for best heat transfer. The secondary tuning line (with adjustable bellows) is sized to maintain a similar characteristic impedance without appreciable end-loading distributed capacitance. An inductive loop couples the strong fundamental magnetic field near the center of the cavity. The loaded Q of the cavity varies as the square of the effective loop area and inversely as the square of the distance of the loop center from the cavity center axis. This loop is positioned so that it links more or less magnetic field and determines the output loading of the transmitter. This unique approach yielded the bandwidths in Table 3 which provide excellent subcarrier operation.

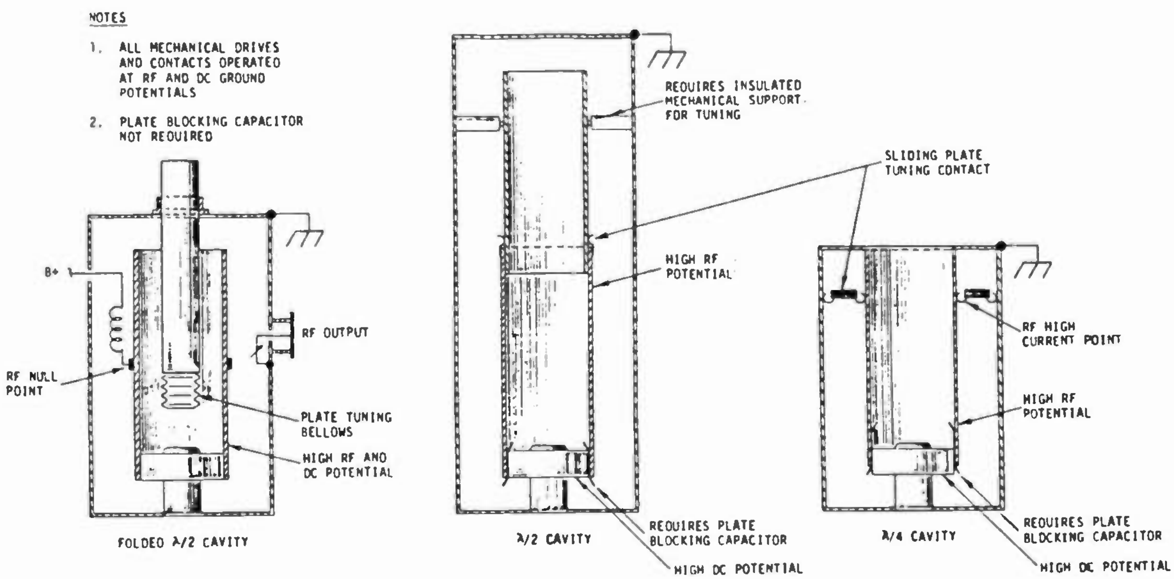


FIGURE 9. COMPARISON OF PA CAVITIES

3.5 PA Adjustments For Subcarrier Optimization.

The power amplifiers which have been discussed operate with improved reliability and power efficiency without compromising subcarrier performance. By providing a broadband input matching circuit with a single control, adjustment of these transmitters for optimal subcarrier performance (minimum crosstalk, maximum separation, etc.) is very repeatable. A typical adjustment procedure involves tuning the transmitter for minimum audio output from an envelope detector while FM modulating the transmitter to 100% with a single 400 Hz tone. When the minimum is reached, the audio output from the envelope detector will double in frequency to 800 Hz. This indicates correct tuning at the center of the passband. Tuning for best synchronous AM will simultaneously result in high efficiency. This also coincides with minimum stereo-to-SCA crosstalk. The rigid mechanical construction of both the input matching circuit and the folded half-wave cavity contributes to the overall electrical stability of the tuned circuits, a benefit for long-term SCA operation where constant "tweaking" is undesirable.

IV. CONCLUSION.

The development of new FM transmitting equipment requires attention to design details in bandwidth and linearity of all sub-systems, including the stereo and SCA generators, the FM exciter, and all RF amplifier stages. New techniques have been developed which reduce the number of controls throughout the transmitting system, minimizing field adjustment. The key design criterion for new transmitters is to optimize SCA and stereophonic subcarrier performance while retaining high reliability.

ACKNOWLEDGEMENTS.

The authors are grateful to Mr. Geoff Mendenhall for his assistance in editing this paper. The authors also wish to thank Charlotte Steffen for preparing this manuscript and Mike Hayden along with Jeff Houghton for providing illustrations.

For Further Information On The Technical Aspects Of Stereo/SCA Operation.

Bott,H., "Analysis of Certain System Characteristics of Stereo and SCA Operation", IEEE Transactions on Broadcasting, Vol. BC-13, January 1967.

Bartlett,G., ed., NAB Engineering Handbook, National Association of Broadcasters, 1975.

Clark,G., "Let's Minimize Those FM Transmitter Problems", Broadcast Engineering, February, 1974.

Clark,K., and Hess,D., Communication Circuits: Analysis and Design, Addison-Wesley Publishing Company, Second Edition, 1978.

Denny,R.,Jr. "Report on SCA Operation", Proceedings of 37th NAB Engineering Conference, 1983.

Federal Communications Commission Rules and Regulations, Vol. III, Part 73, Government Printing Office, October, 1982.

Gray,L., and Graham,R., Radio Transmitters, McGraw-Hill Book Company, 1961.

Federal Communications Commission, "FM Subcarriers", Docket 82-536.

Hedlund,L., "Stereo and SCA Are Totally Compatible In FM Broadcasting", Communications News, April, 1974.

Hidle,J., Priester,H., and Resnick,A., "Modulation Levels During SCA Transmission", Broadcast Engineering, February, 1984.

Kean,J., "The Effect of Additional SCA Subcarriers on FM Stereo Performance and RF Protection Ratios", Proceedings of 37th NAB Engineering Conference, 1983.

Kean,J., "What is SCA Interference and "Birdies"?", SCA Techtalk from NRBA bulletin.

Klein,H., "Issues Concerning Modulation Levels During FM SCA Operation", Proceedings of 37th NAB Engineering Conference, 1983.

Lanier,R., "FM SCA: An Engineering Perspective", Broadcast Management/Engineering, September, 1983.

Mendenhall,G., "The Composite Signal-Key To Quality FM Broadcasting", Broadcast Electronics, Inc.

Mendenhall,G., "SCA Basics", Telocator, September/October, 1983.

Middlekamp,L., "Stereophonic Seperation in Transmission", IEEE Transactions on Broadcasting, Vol. BC-14, September, 1968.

Motorola, Inc., "Subsidiary Communications Authorizations, and Their Application To Radio Paging", Motorola, Inc., Communication Sector Paging Division, September, 1983.

National Association of Broadcasters, Westinghouse Broadcasting and Cable, Inc., and National Public Radio, "Increased FM Deviation, Additional Subcarriers and FM Broadcasting: A Technical Report", Comments on BC Docket No. 82-536, August 30, 1983.

Onnigan,P., "Transmitting Antenna VSWR Effects on FM Stereo", Jampro Antenna Company.

Schober,E., "FM Multipath and Distortion Reduction Through RF Amplifier Optimization", Broadcast Engineering, May, 1983.

Schrock,C., "FM Broadcast Measurements Using the Spectrum Analyzer", Tektronix Application Note 26AX-3582-3.

Waldee,S., "SCA and Stereo FM", Audio, January, 1979.

Weirather,R., and Hershberger,D., "Amplitude Bandwidth, Phase Bandwidth Incidental AM, and Saturation Characteristics of Power Tube Cavity Amplifiers for FM", IEEE Transactions on Broadcasting, Vol. BC-29, March, 1983.

Weirather,R., and Smith,S., "Design Criteria for Multi-station Combining Systems", Proceedings of 37th NAB Engineering Conference, 1983.

INNOVATIVE USE OF AM SCA'S

Bill Holbrow and Rex Trobridge

McGraw-Edison Company

Carson, California

Our objective here is to describe an AM broadcast based communications and control system that is so modularized and flexible that it can serve a variety of uses. The system operator is able to expand, modify, and restructure its system to extend the capability to such control functions as load management, beeper systems, emergency warning, etc.

The primary elements of the system, which are being supplied currently by McGraw-Edison Company, are: a computer based control center, a special modulator installed at a local AM broadcast station, and remotely located radio-controlled switches or data terminals (Figure 1).

The control center, with a properly sized computer, provides considerable flexibility through properly designed software with a wide variety of commands (Figure 2). In one-way applications, remote switches are controlled in groups, a small microcomputer with no external disc capability is sufficient. A Z-80 based microprocessor, for example, can handle 32,000 controlled groups with a basic 32 command repertoire. For expanded operations involving individual customer addressing, status reporting of all group and individual control actions, multiple control devices, etc., a larger microcomputer (on the order of the DEC PDP-11/23) is needed, primarily for its larger software processing capability and available memory. Such a system can address and control well over a million individual customers. For two-way applications (which will be described later), a larger micro- or small mini-computer (such as the DEC VAX series) may be needed. Modular flexibility is provided by low cost software changes. Major expansions are accomodated by replacing the smaller computer with larger, upward compatible computers that can augment existing software, data base and command formats.

The second primary element of the system is the computer/modulator at an AM broadcast station. Since service territory may cover regions as

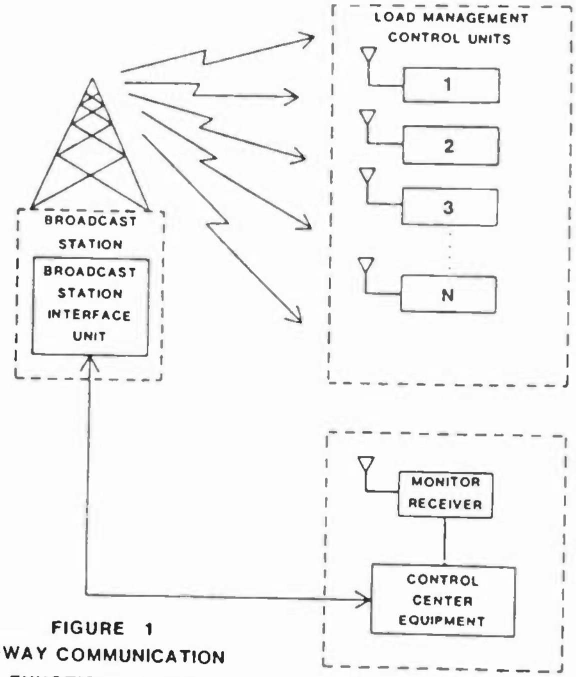


FIGURE 1
ONE-WAY COMMUNICATION
SYSTEM FUNCTIONAL SCHEMATIC

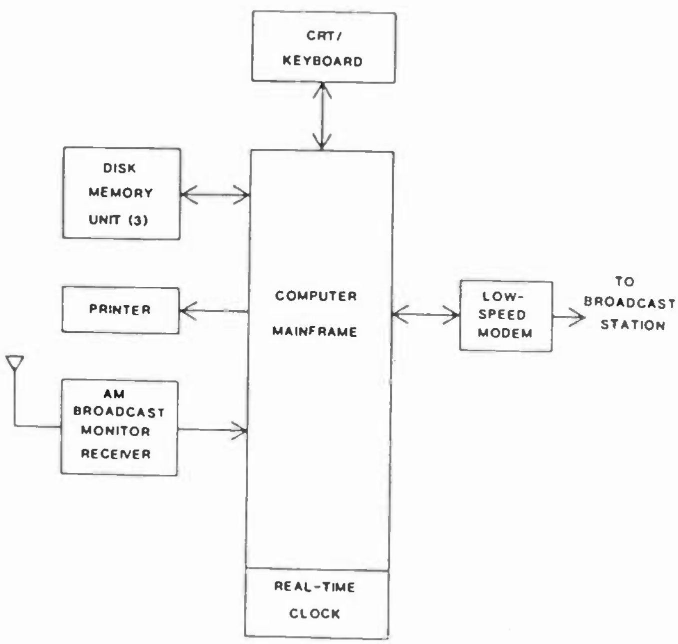


FIGURE 2
UTILITY CONTROL CENTER
COMPUTER BLOCK DIAGRAM
ONE-WAY SYSTEM

geographically disparate as hills and canyons and include heavily populated urban as well as sparsely populated rural areas, a reliable communications system is essential. The signal, from the central control to a receiver 50 miles away, must be as reliably detected as the signal to a receiver across the street from the central control.

Using an established AM broadcast radio station as the communications medium provides a reliable, low-cost foundation for an effective, flexible digital communication control system. An AM broadcast station operates in the 535-1605 kHz, medium frequency, protected band. Man-made interference within a given service territory is non-existent. The propagation characteristics of the AM broadcast band make it possible for a single radio station to cover a large service territory, with coverage on the order of an achievable 30,000 square miles.

The principle on which the AM radio communications system is based is relatively simple. For the forward link, subaudible (less than 80 Hz) digital signals are superimposed on an existing AM radio station's broadcast signal using small-angle, synchronous, quadrature (phase) modulation to transmit commands to receivers installed on customers' premises. The subaudible signals do not interfere with the AM station's programs nor do they in any way deplete the carrier power and coverage area of the AM station. This forward-link concept offers important benefits:

- No new radio spectrum is required.
- Powerful AM radio stations -- the reliability of which is probably unmatched by any other known communications medium -- already exist.
- The coverage area is very large.

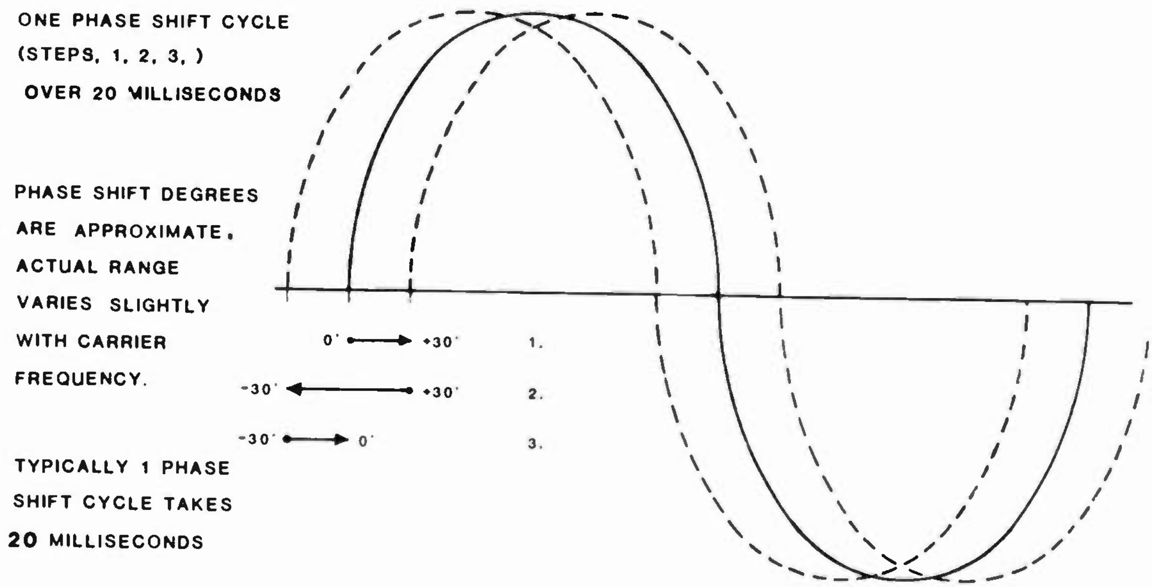
The effective use of phase-modulated signals superimposed upon the AM station's carrier signal is the unique feature of the McGraw-Edison system. Phase modulation is illustrated in Figure 3. The carrier (Part A of the figure), showing only one cycle, is advanced and delayed by +/-30 degrees to produce a digital "1" in the transmitted message. The absence of a modulation in the appropriate time slot indicates a "0".

As stated earlier, phase modulation does not interfere with the regular broadcast program material. McGraw-Edison is currently developing the technology required to assure non-interference with AM stereo transmission with promising results.

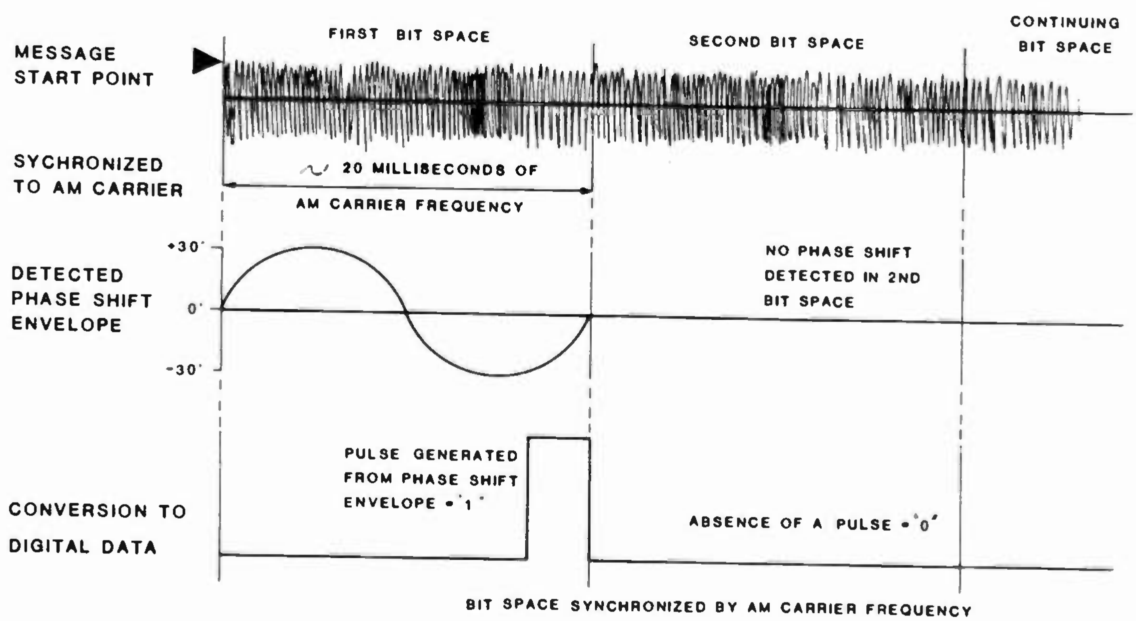
The phase-modulated signal of the McGraw-Edison system is transmitted at 80 bits per second (bps). Figure 4 is a reproduction from the AM band centered about KNX 1070 kHz (KNX is the station in the Los Angeles area currently being used for transmitting energy management signals). The large spike in the center represents the AM carrier and the small ripples on either side represent the program modulation.

The equipment at the broadcast station provides an interface between messages sent from the control center (via a data link -- Telco or equivalent) to the broadcast transmitter site. The interface equipment is shown in a block diagram in Figure 5. The subsystem has two identical units

A. ONE PHASE SHIFT CYCLE



B. CONVERSION OF PHASE SHIFT TO DIGITAL DATA



GRAPHIC DISPLAY OF SMALL-ANGLE PHASE MODULATION

FIGURE 3

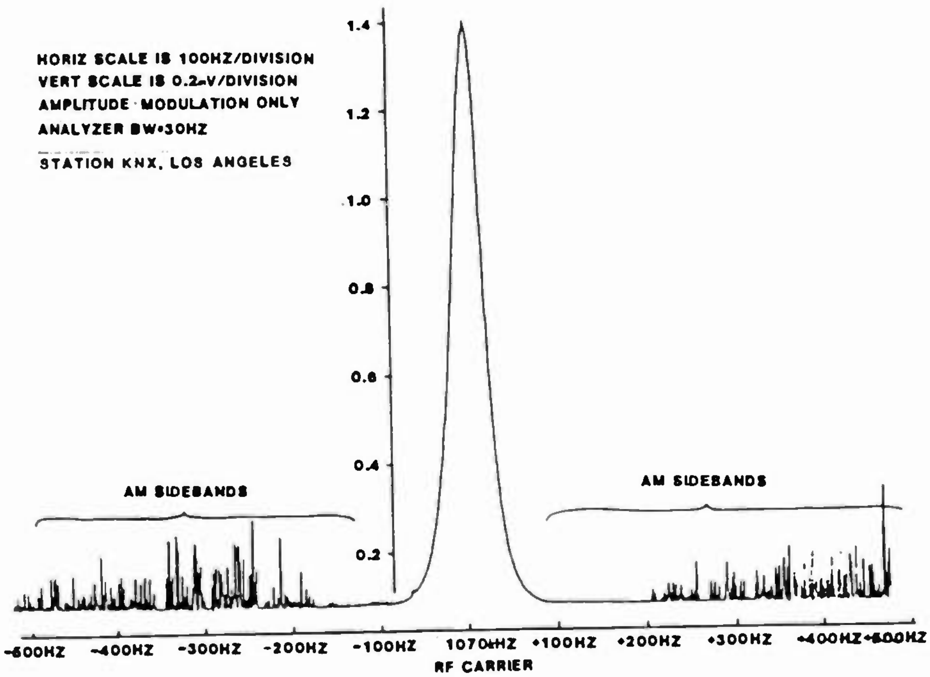


FIGURE 4 RF SPECTRUM OF STATION KNX LINEAR PLOT, ANALYZER BANDWIDTH IS 30 HZ--WITHOUT PHASE MODULATION

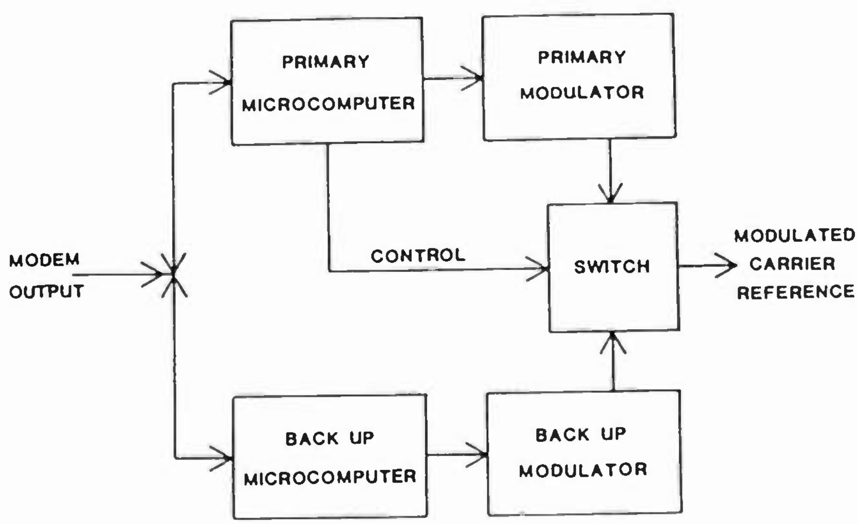


FIGURE 5
BROADCAST STATION INTERFACE
FUNCTIONAL BLOCK DIAGRAM

providing full redundancy, each unit consisting of a microcomputer and a modulator. One unit is designed as the primary; the other as the backup. Both microcomputers will receive messages from the control center via the data line modem.

A switch (Figure 5) enables the primary system, as long as the primary microcomputer is operating properly and the primary modulator is producing a correct signal. Messages received by the primary microcomputer are error checked and, if correct, are put in a queue. When a message is ready to be transmitted, it is transferred to the modulator.

The modulator generates a signal at the broadcast station carrier frequency and applies quadrature modulation, in accordance with the transmitted message. The modulated carrier is then sent to the broadcast station transmitter.

If the primary computer or modulator experiences a malfunction, the switch operates and the backup system goes on-line. The backup unit will have valid copies of all current messages to reduce message loss.

Frequently, full system coverage can be obtained using one station. When several stations are required to cover a larger territory, one of them is designated as the primary station (a function of distance from the control center and from the other stations). The others are designated as relay stations. Each relay station operates a special receiver tuned to the primary station's carrier and is capable of reading the phase-modulated signal. Signals transmitted by the primary station are retransmitted by the relay stations so full territory coverage is obtained almost simultaneously.

An important by-product in using the AM carrier is the availability of the radio carrier frequency for time synchronization of the entire system. This provides very precise control of customer remote devices over the entire large coverage area of the AM station. In addition, the same digital synchronization circuits used by the forward link receiver are shared by the transmitter of the return link. This vastly simplifies the problem of digital circuit design and guarantees system-wide precision timing.

The third basic element of the McGraw-Edison system are the radio-controlled receivers, located throughout the service area. As with the control center, there is an austere or basic radio switch that receives, decodes and controls a single relay on command. McGraw-Edison also provide advanced levels of radio switches using microcomputers and non-volatile memory (NOVRAM).

Activate commands can be sent to separate customer or equipment groups (up to 32,000 groups as indicated previously). Each radio switch can, simultaneously, be a member of several groups, typically four. The actual number will depend directly upon available addresses in the NOVRAM memory chip, storing various reference quantities (such as time-out or minimum reactivate times). Activate commands can be addressed to all members of all groups for simultaneous control (the SCRAM command). At the other extreme, activate commands can be sent to individual equipments or sites.

The change value message can be applied in the same manner as activate commands, except that the result is a revision of the code in the NOVRAM memory chip. A SCRAM-type message can be applied here as well as individual value changes.

One standard microprocessor-based control unit will serve most users, without hardware modification; the software (firmware) is tailored to meet various application requirements.

The two-way system features a "return" communication link in either the VHF or UHF bands, using the broadcast station signal as a frequency reference for the transmitter and as a coherent source for a synchronous detector in the receiver.

In the two-way communication system (Figure 6), messages are generated by the control center and sent to the customer locations via the broadcast station AM signal. Bidirectional devices at each of the locations monitor key activity intrusion (temperatures, smoke, etc.). The bidirectional devices may be polled at any time using a command over the forward link. When polled, the devices transmit data over the return radio link to the central receiver units. These units, placed at prominent geographic locations, send the return link messages back to the control center. The data is received, formatted, and output for operator review at the control center.

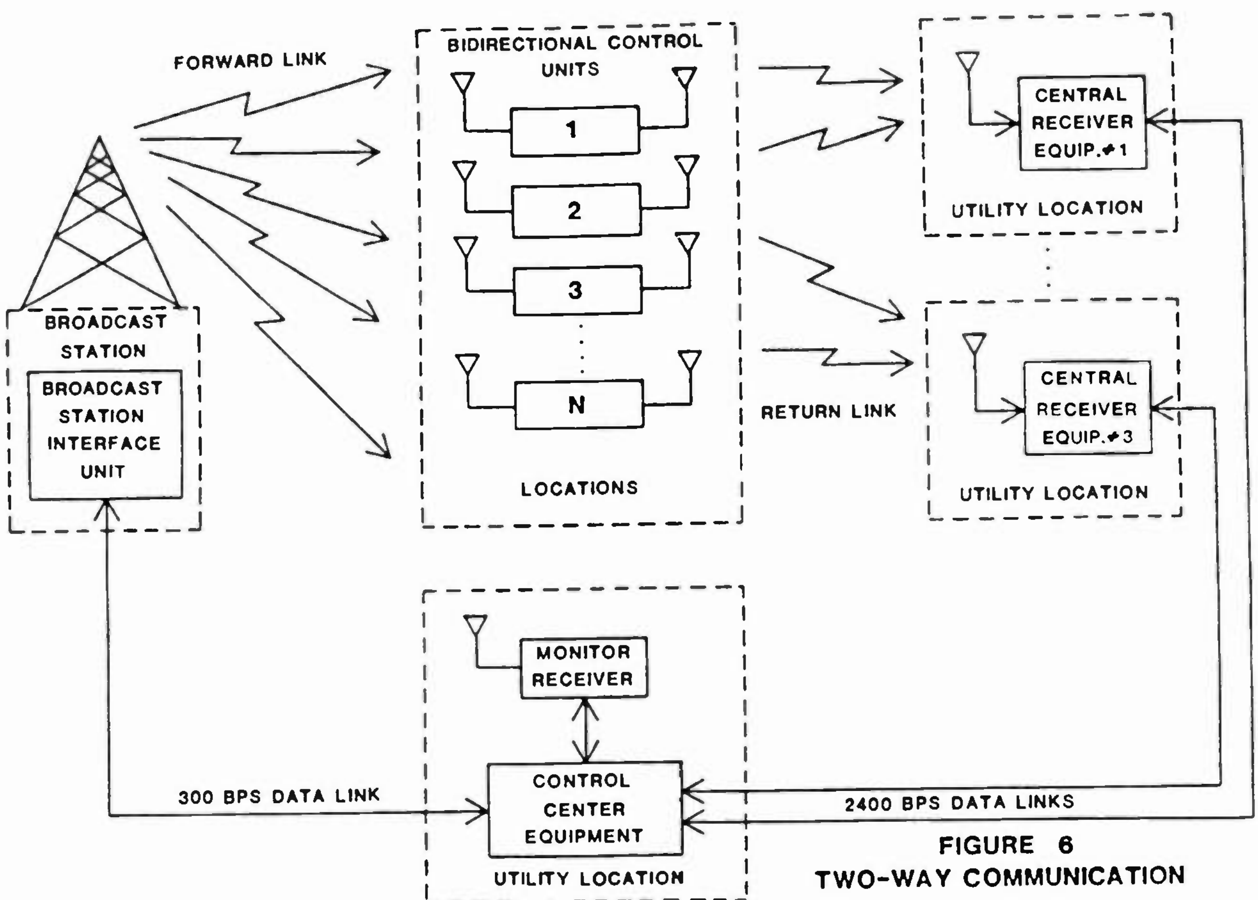


FIGURE 6
TWO-WAY COMMUNICATION
SYSTEM FUNCTIONAL SCHEMATIC

The one-way control receivers can be expanded as desired. The one-way control receivers, with their direct links from the broadcast station, can be replaced with two-way units. This paves the way for a comprehensive capability, centered about the local control center. A great variety of application options become available through the two-way capability, with the use of increasingly more powerful microcomputers and NOVRAM chips at each locations. A well-defined growth plan is highly recommended so that the system can achieve maximum benefits.

The transition from one-way to two-way operation is primarily one of adding capability and capacity. The broadcast station interface unit remains unchanged. At the control center, the microcomputer should be replaced by a minicomputer and additional storage devices. Additional CRT's and printers may be desired.

The original software can be either supplemented by specific routines for two-way operations or replaced by a more powerful integrated package for extended capabilities, in anticipation of system application growth. Central receivers are added to receive signals from the two-way control devices and to relay them to the control center. The number of central receivers needed depends upon customer density and local geography. As a general rule, each central receiver can serve a ten-mile radius.

Applications

Electronic technology and FCC deregultion open up a wide range of other uses. Digital data transmission with expanded addressing techniques allows several groups to share a single carrier. In order to remain sub-audible, channel capacity today on AM broadcast is limited (80 baud); however, the number of potential control points can be very high (more than 500,000).

Opportunities exists for paging, traffic light control, highway sign display control, timing signals for coordinating vehicle location and two-way communication systems. The synchronization provided by the stable AM carrier controlling a low cost phase-locked-loop receiver provides a highly efficient technique for time-division multiplexing as well as precise frequency control. This represents a key to orchestrating communication networks or cellular systems involving multi-point communications. Some of these potential applications may not be too far off given the rapid advances in electronics.

Other applications, where selected groups require notification, include warning at and around nuclear power plants, priority messages to the deaf via visual displays or teleprinters and recalling fireman, policemen and other select groups in emergencies.

Potential Revenue

A question commonly asked is, "What should the broadcaster expect in revenue from this new business area?" In terms of return on investment, the profit is extremely high. The broadcast station interface equipment may be paid entirely by the system operator or user. If this is the case, the user would require long-term business arrangements. Although no industry standard yet exists for determining a fair market price, an annual fee for use of the sub-carrier is a starting point for negotiations. Fees are somewhat less than the FM-SCA because of the reduced bandwidth available at AM.

To make this system even more attractive cost-wise and increase the overall revenue to the station, a multi-user approach is an option. In this approach, a cost for usage time becomes a prime factor. For utility load management, only a small fraction of the available air time is used by the utility. Two-way communications such as alarms and status reporting may require more air time than one-way systems (Figure 7). Several users could share this data communication channel by frequency separation and/or the addressing capability made possibe by digital technology. More than 32,000 groups can be individually addressed on the present micro-processor based system. A message unit of five seconds would allow over 10,000 messages per day, which, even at a low cost per message, represents a significant business opportunity to the broadcast station. Innovation in fee structure is without limit including joint ventures and profit sharing.

A M EMERGING OPPORTUNTIES

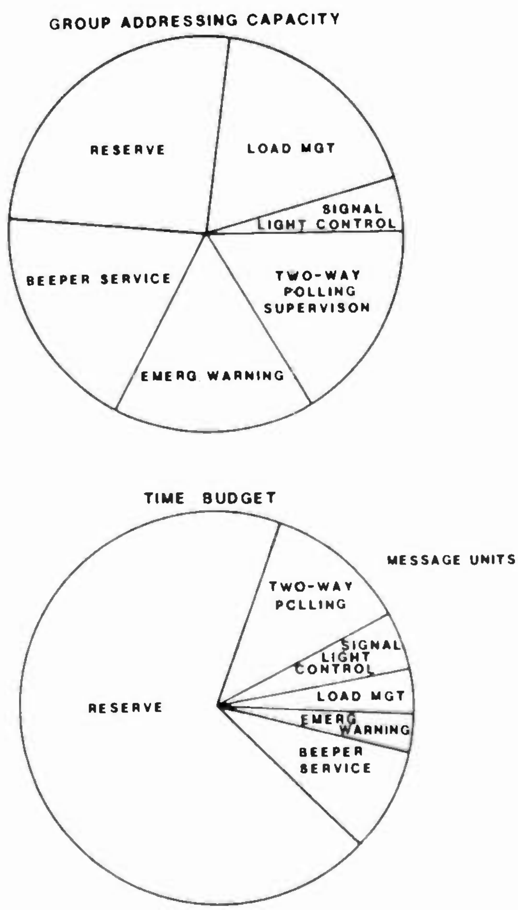


FIGURE 7

Conclusions

AM Radio auxillary data communications systems have unique advantages for digital communication and the control of remote devices. The high effective radiated power of AM broadcast radio -- along with small-angle, quadrature (phase) modulation and stable carrier frequency for system synchronization -- offers a powerful, reliable, expandable, wide-range cost-effective communications system. With the basic communications system in place, advances in technology will permit the implementation of additional applications.

PAGING VIA FM SUBCARRIERS

Motorola, Inc.*

Communications Sector Paging Division

On April 7, 1983, the FCC amended Part 2 and 73 of the Commission's rules to allow FM stations to use their Subsidiary Communications Authorizations (SCA's) for one-way communications of a non-broadcast nature. Prior to this rulemaking, FM SCA's were only authorized for services which were classified as being of a broadcasting nature (an example of this is reading services for the blind).

On May 19, 1983, the FCC released the text of its "First Report and Order" in BC Docket No. 82-536 which allows FM SCA's to be used for services such as (but not limited to): paging, inventory, price and delivery information, dispatching, and other one-way communications services. The rule amendments effective date was June 27, 1983.

In addition to allowing SCA's to be used for non-broadcast services, the rulemaking also:

- Permits the operation of the subchannels whether or not the main channel is in operation.

- Eliminated the need to retain program logs for SCA operations.

- Allows subcarrier modulation techniques to be at the discretion of the FM station licensee as long as there is no main channel interference.

- Expanded the FM baseband from 75 to 99 KHz. (This does not apply to stations within 200 miles of the Mexican border).

If an FM station licensee chooses to offer a paging service, they may do so as either a private carrier or a common carrier. A distinction between these two types of carriers is explained in the following excerpts from the Commission's text.

"Therefore, the determination as to whether a particular service offered on an FM subchannel is private or common carriage will be made in accordance with the NARUC I test for all non-broadcast related services except mobile radio. For mobile radio services, the new Section 331(c) standard will govern."

"Once a licensee has determined that the proposed service is common carriage under the appropriate standard, it must seek authorization to provide that service from the Common Carrier Bureau

*This paper was presented by Mr. Steven Shapiro and is reproduced with permission of Motorola, Inc. Communications Sector, Paging Division.

(and state commissions, as appropriate). Because existing broadcast licensing procedures may not afford the needed mechanism by which necessary Commission determinations related to common carrier service offerings can be made, we will require any licensee intending to provide such services with subcarriers first to seek authorization by filing a suitable request under Parts 21 or 22, as appropriate. Public notice will be given of each such request received, and a 30-day period will be afforded to all parties wishing to file comments in connection therewith. After considering any comments submitted and the substance of the underlying request, the Commission will issue a decision disposing of the matter. It is our intention that, in seeking such authorizations, the FM subchannel operator will be in the same position, entitled to the same privileges and subject to the same obligations and regulations as a traditional offerer of such services. For example, the Commission has established as a general matter that competition in the provision of certain common carrier services is in the public interest. Therefore, our policy is that applications to provide common carrier services from qualified applicants will be granted unless there is some basis to believe that such grant "is likely to produce results that conflict with the goals of the Communications Act." This policy applies equally to FM subchannel operators proposing to provide these common carrier services."

"FM broadcast licensees seeking to provide private carrier service on subcarrier facilities must notify the Licensing Division of the Private Radio Bureau at Gettysburg, Pennsylvania, 17325, by letter, prior to initiating service. In the letter, they must certify that their facilities will be used in this regard only for permissible purposes. When providing land mobile service, they must also certify that service will be offered only to users eligible under Part 90 of the Commission's Rules, and that any interconnection of the station with a telephone exchange or interexchange service or facility will be obtained in accordance with new Section 331 of the Communications Act. Such notifications will not give rise to a comment period, and no separate authorization will be issued by the Commission. As in the case of common carrier services, the FM subchannel operator offering a private service will be in the same position, entitled to the same privileges and subject to the same obligations and regulations as a traditional offerer of such services."

"In all cases, involving either private or common carrier services, the applicant will not be seeking approval for the technical facilities of the FM station or the subchannel. The Commission regards FM subcarrier use as a secondary privilege that runs with the primary FM station license. That right is conferred on the primary station licensee only. In this regard, it should be noted that an FM broadcaster that elects to use a subchannel for private or common carriage remains a broadcaster for all other purposes. Only the use of the subchannel for non-broadcast related purposes would be regulated in accordance with private radio or common carrier regulations."

In summary, if an FM broadcaster chooses to offer service as a common carrier, the FM licensee must obtain authorization from the FCC and state commissions (in most states). However, as of this writing, the Commission has placed a moratorium on SCA common carrier filings (Form 214) until the Commission examines what procedures and forms they will require.

If an FM broadcaster chooses to offer service as a private carrier, the FM licensee must notify the Licensing Division of the Private Radio Bureau at Gettysburg, Pennsylvania, by letter before initiating the service. Note that no state approval is required to offer service as a private carrier since the broadcaster cannot interconnect with the public switched telephone service as a common carrier can. The interconnect rules are more strict for private carriers, and any interconnection must be in accordance with the new Section 331 of the Communications Act. The Communications Act as amended can be obtained from:

Supt. of Documents
U.S. Government Printing Office
Washington, D.C. 20402

FM STATION CLASSIFICATIONS, ERP AND RANGE CHARACTERISTICS

There are three classes of FM stations (in addition to the 10-watt educational classifications) which determine the station's RF power limits and maximum allowable antenna heights. They are:

STATION CLASS	MAXIMUM POWER (ERP)	MAXIMUM ANTENNA HEIGHT (AAT)
Class A	3 kw	300 feet
Class B	50 kw	500 feet
Class C	100 kw	2000 feet

According to Section 73.206 of the Commission's Rules, "a Class A station is a station which operates on a Class A channel and is designated to render service to a relatively small community, city, or town, and the surrounding area."

Twenty channels (of the total 300) are designated as Class A channels. On these channels, there are approximately 1,803 Class A stations which is more than one-third of all licensed FM stations.

The Commission's Rules also define Class B and C stations. According to Section 73.206(b), "A Class B station is a station which operates on a Class B-C channel in Zone 1 or Zone 1-A and is designed to render service to a sizeable community, city, or town, or to the principal city or cities of an urbanized area, and to the surrounding area"

"A Class C station is a station which operates on a Class B-C channel in Zone II and is designed to render service to a community, city, or town, and large surrounding area."

FM stations have various contours for measuring their area of signal coverage. The FM station's "principal city" contour is that contour which has a signal strength of 3.16 millivolts per meter. In addition to the "principal city" contour, FM stations also measure their "primary" coverage area with a signal strength of 1 millivolt per meter. There's also the 50 microvolt contour, but this "secondary" is usually not referenced by many stations since other stations on adjacent channels cause reception problems beyond the "primary" contour.

The most common reference for an FM station's coverage area is that which is covered by the "primary" 1 millivolt per meter contour. According to the FM Atlas and Station Directory, Bruce F. Elving, FM Atlas Publishing Co., 1982, we can see that stations which fall into the various classifications have typical "primary" ranges as follows:

CLASSIFICATION	1 MV/METER CONTOUR RANGES*
Class A	12-15 miles
Class B	26-38 miles
Class C	40-50 miles

*This is what one might experience with an automobile FM radio and a 39 inch whip antenna.

For comparison purposes, the 3.16 mV/meter "principal city" contour is approximately 10 dB stronger than the "primary" contour as shown in the following equation. Thus, the "principal city" contour is considerably less. The 10 dB figure is arrived at by calculating:

$$20 \log 3.16/1.00 = 9.99 \text{ dB}$$

According to the FM Atlas and Station Directory (p. 12), "secondary coverages may be estimated as from about three times the primary distance for 3,000 watt stations to about twice the primary distance for 100,000 watt stations."

NEW FM SPECTRUM

In May 1983, the FCC created three new classes of stations and changed station separation requirements which should result in 600-1000 new FM stations.

According to "Broadcasting" May 30, 1983, "the Commission amended those rules to permit Class A stations to operate on Class B and C channels; to permit a new intermediate class of station, B1 (maximum power of 25 kw and maximum antenna height of 100 meters (329 feet)), to operate in Zones I and IA; to permit two new classes of stations, Classes C1 (maximum power of 100 kw and maximum antenna height of 300 meters (984 feet)) and C2 (a maximum power of 50 kw and a maximum antenna height of 150 meters (492 feet)) to operate in Zone II."

With this rulemaking, the FCC also reduced the stations separation requirements. This means that selectivity and IM problems for an SCA receiver may become greater in the future.

SPECULATION ON SCA PAGING

Since many FM stations have antenna heights that are between 500' and 1500', and since most FM stations have power limits that exceed those of land mobile or common carrier transmitters, there has been considerable speculation with respect to radio paging on FM SCA's. Some of these speculations include:

MISCONCEPTION #1

- FM radio stations can offer paging over a wider geographic area which may be as much as 60 to 80 miles in diameter.

MISCONCEPTION #2

- An FM subcarrier could transmit data as fast as 5000 characters/second which is equivalent to sending a single spaced typed page in one second.

MISCONCEPTION #3

- FM subcarriers are better suited for nationwide paging because there are only three nationwide frequencies, and over 2800 applications have been filed with the FCC.

MISCONCEPTION #4

- There will be 20-25 million pagers in service by 1990 to be utilized by FM radio stations.

Note that these points are not necessarily true but are believed to be true by some people. Here's why they are not necessarily true.

FACT #1

-FM radio stations can cover 60 to 80 miles in diameter, but a radio with an external antenna is required. For example, an automobile may receive FM stations at those distances, but the radio has either an external whip or windshield antenna. A pager's internal antenna is at a 20-30 dB disadvantage which means that the distances for paging will be less than half of that which a car radio experiences on the main channel. A general rule of thumb is that 12 dB is required to double range. Therefore an SCA pager's range must be less than half that which is experienced in most automobiles.

FACT #2

-Given that an alphabetic character requires 8 binary bits and 5000 characters multiplied by 8 bits equals 40,000 bits, a data rate of 40,000 bits/second would be required to send 5000 alphabetic characters in one second. Forty thousand BPS is not practical due to greater bandwidths required and loss of range. However up to 9600 BPS may be practical.

FACT #3

-Two thousand eight hundred (2800) applications were filed with the FCC for 37 local paging channels, not the three nationwide paging channels. These 37 channels can be used over and over again in each of 200 plus cities. Applications for the three nationwide channels were accepted until August 11, 1983. NOTE: 16 such applications for network organizers for nationwide paging were received.

FACT#4

-While some analysts believe that there could be 20-25 million pagers in service by 1990, that estimate is for the total market, including radio common carriers, telephone common carriers, and the government and private sectors. At this time, no one can be sure what percentage of that number the FM stations will obtain.

To determine the share of market a broadcaster's SCA paging system may obtain will be determined by:

- -What geographic area is to be covered by the service?
- -Will subscribers want service in other cities?
- -Where are the bulk of the subscribers located?
- -What types of paging services are now offered?
- -What will be different about this service?

-How is the paging service to be sold?
-At what expected placement rate?
-What promotional advertising program is planned?
-What types of users need paging?
-What type of service is desired by the users?
-How does your state regulate common carriage and what effect does that have on you offering a paging service?

In other words, every market is unique and your specific situation will determine your potential share of market. The total available market for the United States does not provide enough information for you to determine whether paging will be profitable for you.

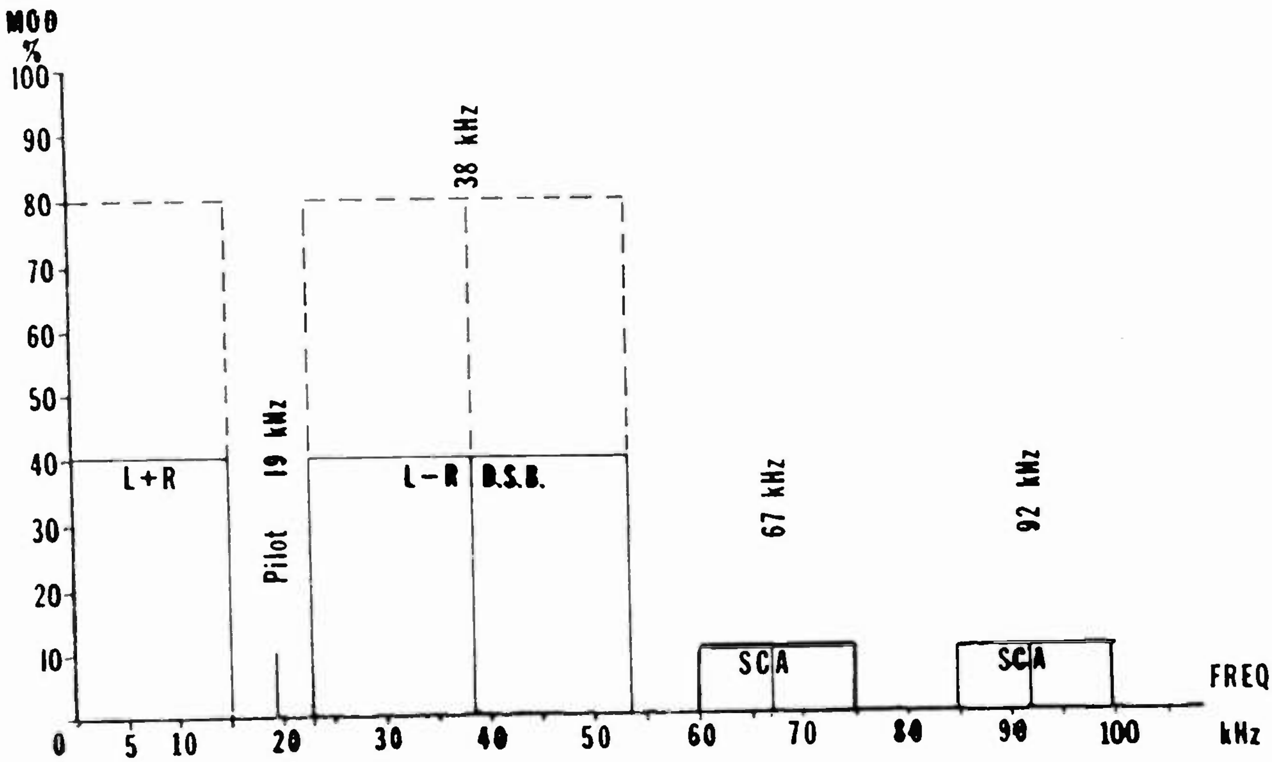


FIGURE 1

II. SCA TRANSMISSION AND DEMODULATION PROCESS

A. NUMBER OF SCA'S

With the FM subcarrier baseband covering 53 KHz to 99 KHz (total of 46 KHz) during stereophonic transmissions (20 KHz to 99 KHz during monophonic transmission or when the main channel is not modulated), more than one subcarrier can be transmitted. This will depend upon the method of modulation used in modulating the subcarriers. For example, if an analog signalling scheme (such as voice paging) using 10% modulation is used, only two subcarriers could be transmitted (see Figure 1). Each subcarrier would occupy 15 KHz, so therefore, these two subcarriers would occupy 30 KHz of the available 46 KHz. The remaining 16 KHz would be used in separating the subcarriers as well as separating the first subcarrier from the main channel programming. Therefore, voice transmissions or any analog signalling (with 4-5 Khz deviation) limits the total available subcarriers to two.

If low rate digital information such as binary paging is transmitted, multiple subcarriers can be used, provided that the modulation level for each subcarrier is reduced to prevent overmodulation of the transmitter by the combined subcarriers. Binary tone alert paging requires very little spectrum (200 Hz) in the FM baseband to signal effectively and since there could be up to 100,000 binary tone alert pagers on a single channel, there is no apparent advantage in using more than one subcarrier for tone alert paging. (Note that this does not hold true for two-tone signalling. See Appendix   for further information).

With binary digital tone alert paging occupying so little spectrum, the possibility exists for one or more additional subcarrier channels to be used for wide band digital or analog transmissions providing point-to-point services. Point-to-point means that fixed receivers with directional gain antennas are used to receive information from the transmitted SCA. In that way, the paging reliability would be optimized, particularly if the paging subcarrier is put on 67 KHz and the other services from 75 to 99 KHz.

B. SIGNAL TO NOISE RATIO IN THE SUBCARRIER CHANNEL

The signal to noise ratio (S/N) in the subcarrier channel is determined by the following factors:

1. The S/N in the main channel receiver IF.
2. The deviation of the main carrier by the subcarrier.
3. The frequency of the subcarrier.
4. The width of the subcarrier channel.
5. The response shape of the main IF filter and discriminator in the receiver.
6. The amplitude and frequency content of the main channel modulation (including stereo).
7. Multipath distortion of the transmitted RF signal.

The first of these factors is the most critical because of the FM threshold effect which causes the discriminator output S/N to change much faster than the IF S/N when the input is below a certain threshold. In a well-defined SCA system, reliable tone alert paging can be achieved with the S/N in the main channel IF slightly below the FM threshold. In this region of operation, the subcarrier S/N varies 2 to 3 dB for every dB change in the main channel S/N.

The other main factors affecting the subcarrier channel S/N are the deviation of the main carrier by the subcarrier, the subcarrier frequency and the subcarrier channel width. The ratio of the deviation to the subcarrier frequency is the subcarrier modulation index. The remaining four factors listed above are secondary; they can degrade the S/N if they are inappropriate or excessive.

The relationship among the various factors can be expressed approximately by the following equation for above the threshold operation:

$$(S/N)_{SC} = 1/2\ (S/N)_{IF}\ M^2\ (W_{IF}/W_{SC})$$

where $(S/N)_{SC}$ and $(S/N)_{IF}$ are the signal to noise power ratios in the subcarrier channel and the main IF channels respectively, M is the subcarrier modulation index, and W_{IF} and W_{SC} are the main IF and subcarrier channel widths. This equation is reasonably accurate for $(S/N)_{IF}$ equal to or greater than 12 dB. For $(S/N)_{IF}$ = 10 dB, the equation is about 2 dB optimistic.

The undesirable effects of inadequate IF and discriminator response can be avoided by proper design of the receiver. The main channel modulation should cause very little trouble provided that the broadcast transmitter is operating with a normal modulating signal and is meeting FCC requirements on peak deviation and distortion. Severe multipath distortion caused by reflections of the transmitted signal by natural or man-made terrain features can have the same effect as poor frequency response in the receiver IF. This condition would be noticed in FM broadcast reception as poor performance in stereo. Poor performance is evidenced by loss of stereo information. In SCA paging, multipath would cause distortion products of the main modulation to fall in the subcarrier signal amplitude at the receiver. The use of directional receiving antennas to minimize multipath is possible in fixed receiver installations but obviously not in portable or mobile paging systems. There is no obvious solution to the multipath problem which is very important to remember in a paging system. However, it is expected to be serious only in a few locations. Field tests may be necessary to determine the feasibility of SCA paging in some specific cases.

Assuming that there are no problems with IF response, abnormal main channel modulation or multipath, we can use the S/N equation to calculate the subcarrier channel signal to noise ratio for some representative examples. For instance, if the IF bandwidth is 200 KHz, the subcarrier channel is 1 KHz wide (binary signalling), the subcarrier frequency is 67 KHz, the subcarrier deviation of the main carrier is 7.5 (producing a modulation index of 0.112) and if the (S/N)IF is 16 (12 dB), then (S/N)SC = 1/2 x 16 x (0.112)2 x 200 = 20 or 13 dB. If (S/N)IF drops to 10 (10 dB) with other conditions unchanged, then (S/N)SC would be about 8 (9 dB).

C. SUBCARRIER DEMODULATION AND SENSITIVITY

The signal to noise ratio in the subcarrier channel required for reliable paging depends on the code, the bit rate, the methods of modulation, detection, and decoding, and on the subcarrier bandwidth. Assuming that the code structure and decoding scheme are the same as in Motorola's present binary pagers and that frequency modulation and noncoherent detection of the subcarrier is used, the required signal energy per bit divided by noise density varies from about 8 dB to 14 dB for 90 percent reliability, depending on the subcarrier deviation level and channel width. These results are obtained from computer simulation and bench measurements. Optimum sensitivity is obtained with deviation approximately plus and minus 0.4 of the bit rate and with the filter bandwidth ahead of the subcarrier discriminator equal to the peak to peak deviation. However, circuit stability falsing and battery saver considerations may require the use of somewhat greater deviation and filter bandwidth.

Other modulation and detection methods such as PSK or coherent FSK could be used on the subcarrier to improve sensitivity by 2 to 4 dB, but at the expense of a substantial increase in complexity. A further improvement of several dB could be obtained by using a synchronous long block coding scheme and a somewhat higher bit rate, but again at the cost of greater complexity and system changes affecting the paging terminals as well as the receivers. These potential improvements do not appear to be feasible at this time.

Taking 12 dB signal energy per bit divided by noise density as a reasonable subcarrier channel sensitivity, and using the same assumptions for the main channel as in II B, we find that a signal to noise ratio of 13 dB in the main IF channel is required. Further assuming a 4 dB noise figure for the receiver, we calculate an overall receiver sensitivity of -104 dBm at the antenna terminals.

In the 88 to 108 MHz FM broadcast band, an antenna efficiency of approximately -24 dBi (dB with respect to an isotropic radiator) or -26 dBd (dB with respect to a dipole antenna) is expected for a small magnetic dipole contained in the pager. Considering that information with the other information just presented, it is easy to see why an SCA paging receiver will require a very strong signal to provide reliable paging. Therefore, while VHF land mobile paging receivers have sensitivities of approximately 5 uV/M, SCA paging receivers will have typical sensitivities of approximately 75 uV/M.

III. RANGE ESTIMATES

A. FIXED POINT INSTALLATIONS

Fixed point installations such as Muzak or reading services for the blind provide excellent range for subcarrier reception. According to the FM Atlas and Station Directory (p. 12) as mentioned earlier, "secondary coverages may be estimated as from about three times the primary distances for 3000 watt stations to about twice the primary distance for a 100,000 watt station."

Therefore, a fixed point installation using a high gain, directional antenna (such as a Yagi) could experience ranges as follows:

CLASSIFICATION	SECONDARY CONTOUR RANGES*
Class A	45 miles
Class B	75 miles
Class C	100 miles

*NOTE: This applies to high gain fixed antennas, not to a pager with an internal antenna.

B. BUILDING PENETRATION

A pager uses an internal antenna which is 20-30 dB below a dipole. Then to make matters worse, the pager normally must operate in steel or concrete buildings which can add approximately 25-30 dB more to signal degradation.

These problems are not unique to SCA paging, but in fact, are problems which paging services have to overcome regardless of the frequency they operate on. However, a land mobile paging system can overcome these problems fairly easily.

If a land mobile paging system has to cover a warehouse district, remote corner of town, or perhaps even the central downtown area of an urban city, additional transmitters can be added. An FM broadcast station does not have that opportunity.

For example, the antenna site for FM station KPLU in Tacoma, Washington, is 15 miles from Seattle. We know from our own testing with National Public Radio that first floor building penetration in Seattle was only approximately 50% for tone alert paging. If a land mobile system had that problem, a transmitter in Seattle could be added (for further information on this test, see Section III.D.).

C. FM TRANSLATORS AND BOOSTERS

In trying to overcome these problems, a common question is, "Why can't I add a booster or translator to solve these problems?"

A translator does broadcast everything the main channel does, including its SCA, but it operates on a different frequency. A multifrequency pager would be needed which would add size, weight, and cost to it. A further problem with translators is that they vary in power between one and ten watts. Even at ten watts, they won't transmit very far.

Boosters overcome the multiple frequency problem because they operate on the same frequency as the main channel. However, boosters don't lend themselves to SCA paging either because their use is only permitted in areas where sharp terrain differences (such as a mountain or valley) shield the main transmitter's signal. Therefore, a hole in paging reception will exist.

D. ACTUAL TEST RESULTS

With a better understanding of some of the technical issues, we can now discuss what is likely to happen in a typical SCA system.

In field testing, we tested tone alert paging using a binary digital decoder in a receiver with a 90 uV/M sensitivity. This is typical of what can be expected in an actual SCA paging system.

Tone alert paging was tested because it will have the greatest range and is the predominant method of paging in use today. Also, 90% reliability is considered to be the minimum acceptable reliability for paging. Ninety percent reliability means that the user will receive nine out of ten pages, or miss one in ten pages. We also tested for 50% reliability (receiving only half the pages transmitted), although few pager users would tolerate such a low reliability rate. Fifty percent reliability for paging does not constitute a useable system.

Field tests were conducted by National Public Radio and Motorola in the Seattle, Washington, area on June 12 and June 13, 1983. A binary paging encoder and subcarrier modulator were installed at FM station KPLU, on 88.5 MHz in Tacoma, Washington. This 100 KW ERP station has its 600 foot antenna approximately 15 miles from downtown Seattle. A 67 KHz subcarrier was used to modulate the transmitter with 7.5 KHz deviation. The subcarrier deviation was plus/minus 1000 Hz. During most of the test, tone alert pages were sent automatically every 15 seconds.

The tests covered most of Seattle and some of the surrounding area. In the downtown area of the city, coverage on the street outside buildings was nearly 100%. Inside buildings on the first floor, it dropped to less than 50% (remember that 50% is unusable in practice). Reliability improved progressively as the pagers were taken to higher levels in the building. In most other parts of Seattle, coverage on the street was better than 90%. However, there were a number of weak signal areas where paging was unreliable or impossible. Similar results were obtained along highways outside the city with good and bad locations and a gradual decrease of paging probability with increasing distance from the transmitting antenna. At the KPLU station site in Tacoma, approximately 22 miles from the antenna, approximately 50% paging (again unusable in practice) was obtained inside the building with up to 90% outside.

In looking at Figure 2, the chart shows various coverages for a Class C station with a 1000' antenna and 100 kw ERP. The chart shows that in an open area such as a street with no buildings or other obstructions, a distance of approximately 33 miles is the maximum distance for 90% reliability. However, in many cities, a pager user is likely to be in a building, and therefore, if they were on the fourth floor, 90% reliability doesn't exceed 12 miles. Furthermore, if the user was on the first floor, 90% reliability occurs at approximately 7-8 miles.

Since many FM stations do not have 100 kw ERP with antennas at 1000' AAT, we ran some further tests using the signal of a 100 kw ERP station with its antenna at 350 feet. We then tested tone alert paging at two Motorola plants which were within the primary coverage area of that station. Both plants are concrete structures, one and two story buildings in areas where there are no tall structures.

In looking at Figure 3, you can see the results achieved at each plant. Note that before we tested, we made predictions on the paging reliability and found that the measured reliability was very close to our predictions. At a distance of 28 miles from the transmitter, paging would not satisfy any user.

In predicting the reliability of SCA paging vs. a two-transmitter land mobile paging system, Figure 4 shows that the area covered by the SCA system exceeds the two-transmitter system by 21%. This assumes the land mobile transmitters are spaced at five miles from the same origin point as the SCA transmitter.

If we spaced the transmitters at ten miles from the same origin point (Figure 5), the total area covered by the SCA exceeds the land mobile system by 16%. However, note that the land mobile system shows directivity which may be important if the SCA's antenna isn't in the center of a town.

It is worth noting that in the top 30 Standard Metropolitan Statistical Areas (SMSA) as defined by the FCC, the typical number of

COVERAGE VERSUS RELIABILITY FOR A 1000′ FM STATION ANTENNA 100 KILOWATTS

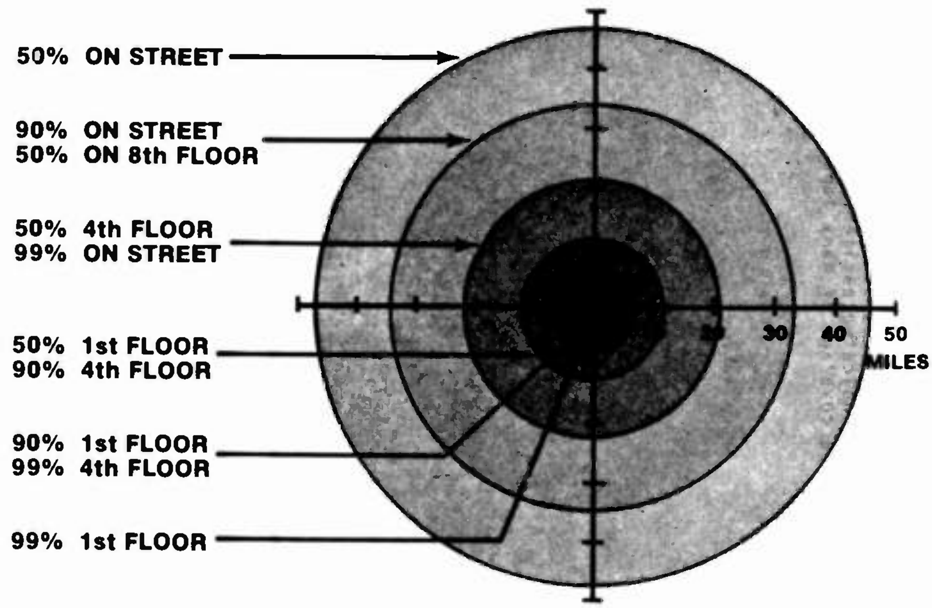


Figure 2

PAGING TEST RESULTS

	DISTANCE FROM XMTR	PREDICTED RELIABILITY OF PAGING	MEASURED RELIABILITY OF PAGING	STREET SIGNAL LEVEL dB/PAGING THRESHOLD	MEAN BLDG PENETRATION LOSS
MOTOROLA PLANTATION PLANT	3.4 MI.	95%	98.7%	+ 46 dB*	25 dB*
MOTOROLA BOYNTON PLANT	28 MI.	<1%	3.6%	+ 15 dB	27 dB

* ESTIMATED STREET LEVEL SIGNAL

Figure 3

COMPARISON OF SCA VS. TWO LM XMTRS
99% RELIABILITY ON STREET

SCA AT ORIGIN
LM XMTRS AT X = ± 5 MI.

AREA COVERED
BY SCA = + 21%

ANTENNA HEIGHTS:
SCA = 1000′
LM = 500′

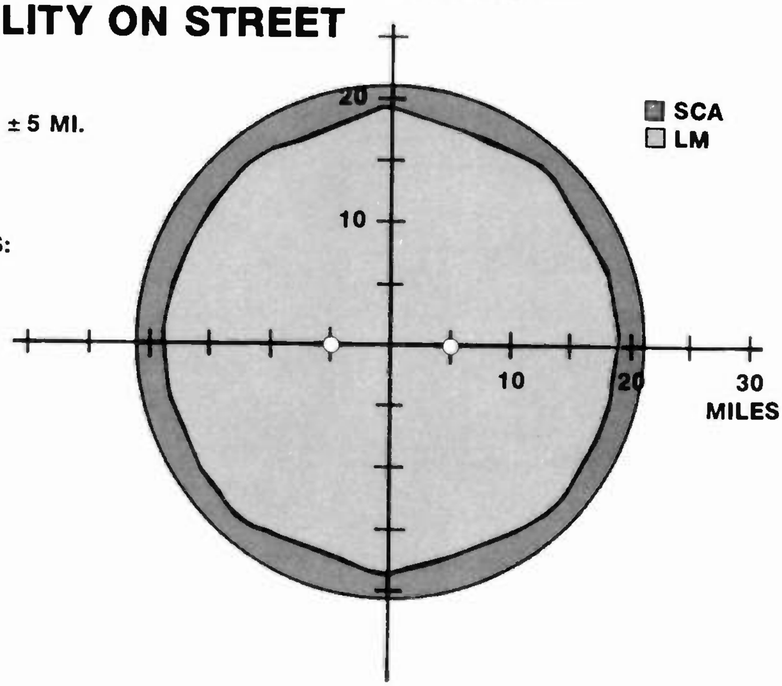


Figure 4

COMPARISON OF SCA VS. TWO LM XMTRS
99% RELIABILITY ON STREET

SCA AT ORIGIN
LM XMTRS AT X = ± 10 MI.

AREA COVERED
BY SCA = + 16%

ANTENNA HEIGHTS:
SCA = 1000′
LM = 500′

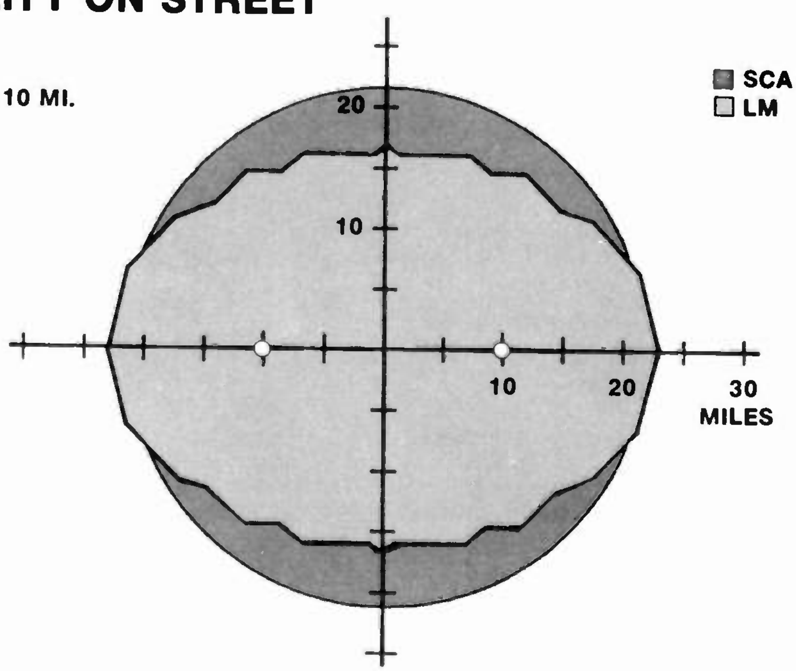


Figure 5

transmitters on a land mobile paging system varies between 3-5. Also, a few very large common carriers have 30 or 40 transmitters! With SCA coverage approximating the coverage of a two-transmitter land mobile system as Figures 4 and 5 show, SCA paging will probably have less range than most land mobile systems in the top 30 SMSA's. This also assumes that the SCA's antenna is situated in the optimal place to serve the desired geographic area.

While all testing and predictions were for tone alert paging, we know that data paging results (assumes GSC coding as discussed in Appendix) will be similar due to the similarity in signalling methods. However, for voice paging, 9 dB more signal is required for a user to clearly understand a voice message so the useful range for SCA voice paging will be approximately 40% less.

E. NATIONWIDE PAGING

In FCC Docket 80-183, the Commission reserved three channels at 929 MHz for nationwide or network paging. By having specific channels allocated for nationwide service, a pager wearer could subscribe to a network and move anywhere within the coverage of the system in the U.S. and be paged. However, a person could not be in their home city and be paged on the network frequencies since local paging must occur on one of 37 different local channels. Therefore, it appears now that a person who requires local and nationwide paging will need a two-frequency pager.

With 900 MHz transmitters and receivers costing more to manufacture and purchase than 150 MHz equipment, some believed that nationwide paging on SCA's would be lower in cost. This is not correct for the following reasons.

SCA nationwide paging has been in service since 1978 in Sweden on the 87-104 MHz band. It covers 99% of the nation's geography, but there are "holes" in some areas due to building penetration problems. For example, in downtown Stockholm, signal reception is difficult when the wearer of the pager isn't near a window. The SCA paging receivers used for this purpose scan the 87-104 MHz band once every ten seconds until the system identification code is found and then stays on that frequency. Therefore, the receiver must have a wide front-end to cover the entire FM band which has a tendency to accept more extraneous signal interference. Other requirements are a frequency synthesizer to cover 100 (U.S.) FM channels and the SCA's on each one, and a means of scanning across the band. These devices are not required in a 900 MHz pager. So, any cost savings in using a lower frequency range are more than offset by adding circuitry to make the system function.

Also worth considering is that the United States has many more cities and many more tall buildings which are built with concrete and steel. Therefore, the building penetration problems will exist in at least the top 50 major U.S. cities, and these problems cannot be overcome by adding transmitters on SCA's as can be done at 900 MHz. As indicated earlier, the FM SCA in large cities will not achieve the coverage of many land-mobile common carriers.

F. Calculating Range

With all this technical information and field test information, the obvious question of many FM station licensees is, "What can I expect from my station?" To answer that question the following formula is used.

$$R=D*10^{-2}* H^{1/2}* P^{1/4}* F$$

Where:

R=Desired range of the station in question.
H=Antenna height (AAT) in feet of the station in question.
P=ERP in KW of station in question.
F=Factor of 1 for tone-alert paging.
.83 for GSC data paging (see Appendix II).
.60 for tone and voice paging.
D=Distance of a 1000', 100 KW SCA in miles under the desired reliability conditions. This is obtained from Figure 6.
NOTE: If in building coverage range is desired, be certain to use the figures (i.e. 90% first floor or 90% fourth floor depending upon desired conditions) that includes the building loss factors.

This relationship assumes plane earth propagation (ignores earth curvature) and directly relates range of the reference SCA station with 1000', 100 KW antenna height and ERP respectively to the new station given the same reliability conditions. As long as D is less than 30 miles, the relationship will be quite accurate. AS D approaches 50 miles then some earth curvature is in effect and as much as a 10% error will result. The calculated range for R will in this case be less than the true range.

To clarify use of this formula, three examples are given.

EXAMPLE 1

An FM station with its antenna at 600' AAT and 33 KW ERP wishes to calculate the estimated range for tone-alert paging for 90% street reliability. The calculation is as follows:

D= 32.5 (from Figure 6)
H= 600
P= 33
F= 1

Therefore R=19.1 miles for 90% street reliability.

If tone and voice paging was used, F=.6 and then R=11.5 miles.

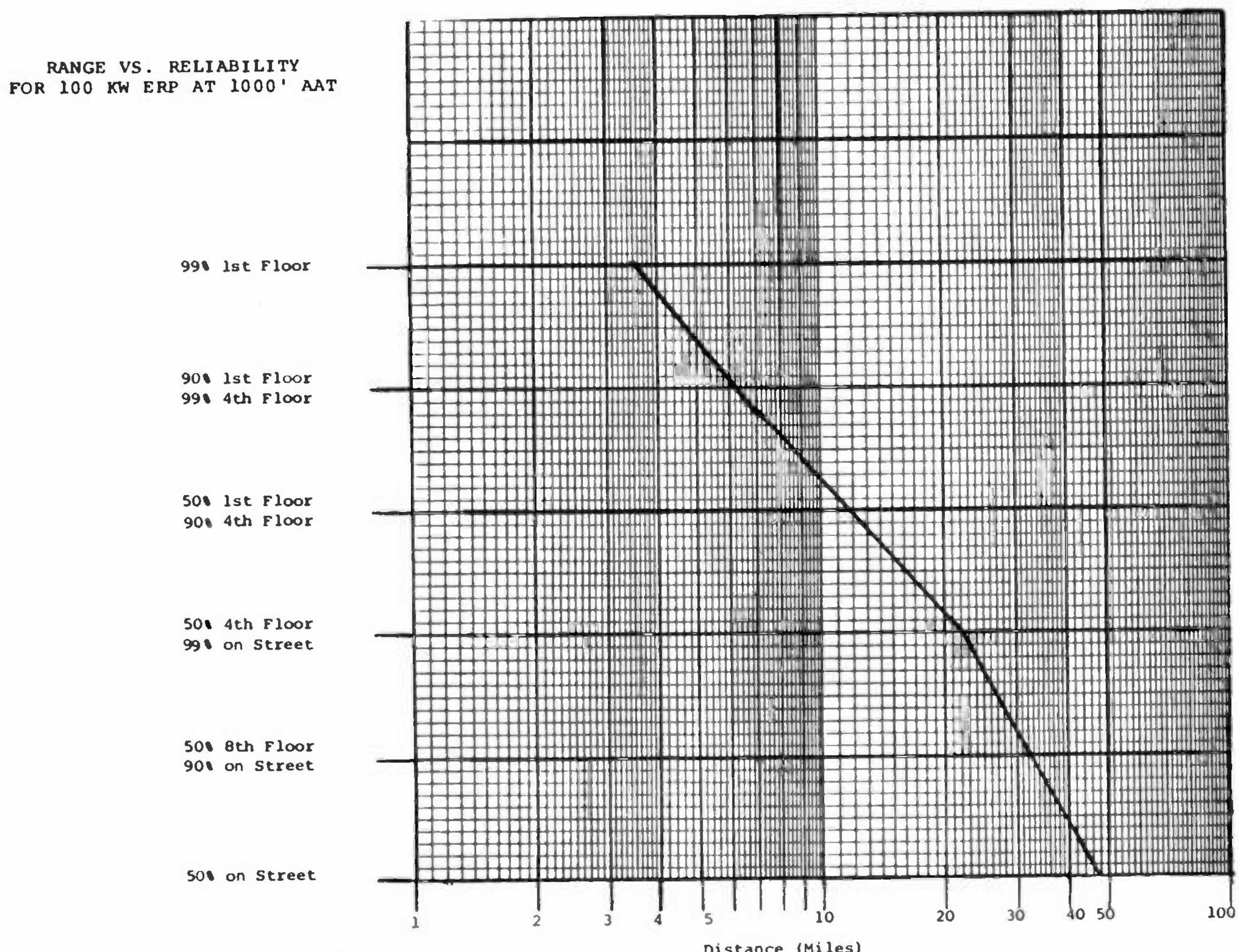


Figure 6

EXAMPLE 2

An FM station with its antenna at 240' AAT and it's power of 3KW ERP wishes to calculate the estimated range for tone alert and tone and voice paging with 90% reliability on the first floor of an urban building. The calculation is as follows:

D= 6.2 miles (from Figure 6)
H= 240
P= 3
F= 1 for tone alert paging
  .6 for tone and voice paging

Therefore R= 1.3 miles for tone alert paging and
R= .76 miles for tone and voice paging.

EXAMPLE 3

An FM station with its antenna at 1548 feet and 4.6 KW ERP wishes to calculate range for tone alert and data paging at 90% reliability in the first floor of an urban building. The calculation is as follows:

D= 6.2 miles (from Figure 6)
H= 1548
P= 4.6
F= 1 for tone alert and
  .83 for data paging

Therefore R= 3.6 miles for tone-alert paging and
3.0 miles for data paging.

Note that for both tone alert and data paging calculations, binary digital coding was assumed. Should a slower form of coding be utilized for tone-alert paging (at the expense of speed and maximum channel capacity), the range calculations could improve. This is highly dependent upon the coding scheme chosen.

For tone and voice paging, the calculations are for the signal strength required for a user to clearly understand the voice message so the calculations hold true for whatever coding scheme is chosen.

Also, if the geographic area in question is rural with very few tall buildings, calculations should be made referencing "90% on street" in Figure 6.

IV. SYSTEM NEEDS

A. GENERAL DESCRIPTION

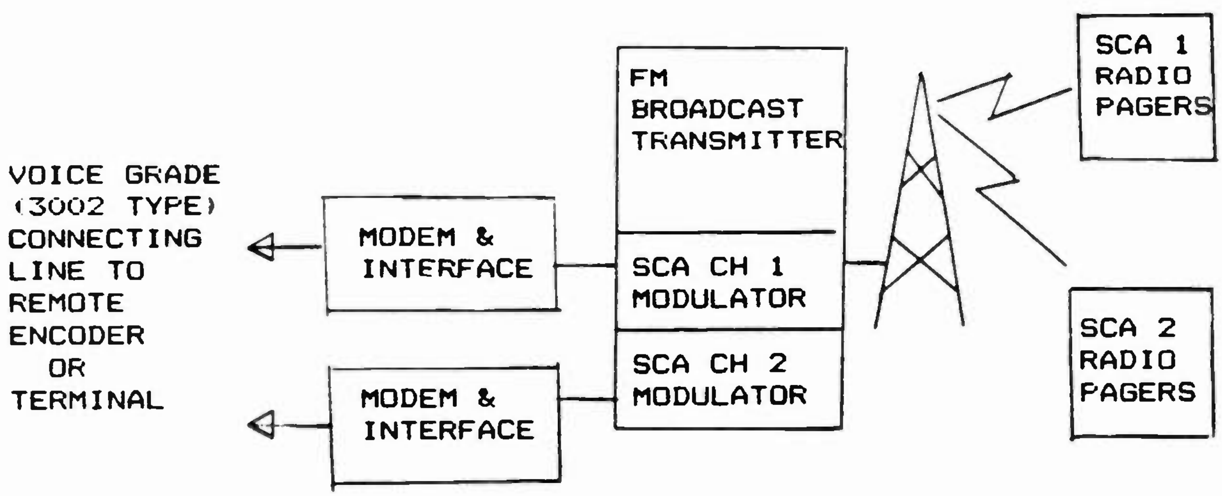
In a land-mobile paging system, four pieces of hardware are required. They are:

- Transmitter
- Antenna (Supporting Structure and Cable)
- Automatic Paging Terminal or Manual Encoder
- Pagers

With an FM broadcast station, the transmitter and antenna already exist (unless the station is under construction), but one more piece of hardware is required--an SCA modulator (and its associated SCA monitor to insure proper operation).

SCA modulators currently on the market are capable of modulating analog signals. However, with binary signalling, an SCA modulator will have to modulate signals much lower in frequency than analog or voice transmissions. Analog or voice transmissions are typically in the 300 Hz to 3500 Hz audio ranges, but digital signals require the SCA modulator to signal with frequencies as low as D.C. for binary FSK. Therefore, most modulators may require some modifications at the audio input points.

For illustrative purposes, if we assume a station transmitting two subcarriers, we can see what the system looks like.



B. SYSTEM OPERATION

SCA radio paging will function like a conventional radio paging system. Calls are entered into an encoding device. This may be a manual encoder with operator dispatch or a fully automatic dial terminal interconnected to the Public Switched Telephone Network. In the automatic system, calls are dialed directly by the person wanting to make the page. The encoding device translates the information received into the signals sent to the transmitter. An interface at the transmitter modulates the subcarrier with the code of a particular pager or group of pagers. Radio signals are then sent to the pagers. A pager with a code corresponding to the one transmitted will respond with an alerting signal.

The encoding device uses an internal modem (modulator-demodulator) to permit sending the binary information via tone carrier in the audio frequency range. (It should be noted that the modem is not required if the encoding device is located within 50 feet of the SCA modulator). This enables use of inexpensive voice grade leased telephone facilities, such as a 3002 type line. A modem is added at the SCA modulator. The modem and interface circuit at the transmitter demodulates the signal from the encoding device and connecting telephone line. It then applies the binary signal to the SCA modulator.

C. ENCODING EQUIPMENT

Motorola provides a line of terminals and encoders which will equip systems ranging from the smallest start-up manual entry paging encoder up to the largest automatic dial terminal systems. The following Motorola products are available now to handle SCA paging requirements.

A MODEN PLUS manual encoder for the start-up/small system of up to 2000 pagers

A MODAX 500A automatic dial interconnect terminal for a small system of up to 1000 pagers

A METRO-PAGE automatic dial terminal for the mid to large size system of up to 10,000 pagers

A METRO-PAGE 200 automatic Radio Paging Exchange for the very large city/metropolitan area, multichannel systems up to 200,000 pagers

The encoding device generates Bell type 202 modem signalling tones to a standard voice grade two wire telephone line. The telephone line links the encoder site to the SCA modulator site. At the FM station, a type 202 modem decodes the signalling tones from the encoder, detects the presence of tones to activate the SCA modulator, and transfers the decoded binary signal to the SCA modulator.

For operation with a binary pager, only three connecting wires are required from the transmitter site modem to the SCA modulator:

Carrier (Data) Detect
Data
Ground

All signals are standard RS232C EIA compatible signals with +/- 12 VDC transitions.

D. PAGER

The final element required in an SCA paging system is, of course, the pager. Motorola has introduced the SCA 1000 FM RADIO PAGER which will meet the needs of any SCA paging system. The SCA 1000 FM Radio Pager includes the following features:

- Tone Alert Signalling
- Binary Digital Decoding
- Battery Saver Circuitry
- Small, Attractive Belt-Worn Styling
- Loud, Clear Alert Tone
- Battery Monitor
- Use of either an AA Alkaline battery or an AA Rechargeable Nickel-Cadmium battery

More detailed information is included in the back of this document in Appendix IV.

With the use of a binary digital tone alert pager, an FM station will:

- Be able to put more pagers on a single SCA, thereby maximizing SCA revenues.

- Utilize a paging method which has the greatest range capabilities.

- Utilize a paging signalling method which allows more SCA's to be used within the FM subcarrier baseband.

V. CONCLUSION

The most important word to keep in mind with respect to SCA paging is: caution. Yes, SCA's may provide a viable means of offering paging, but one must forget the easily made assumptions and remember what the real facts are:

- Class C FM stations comprise only 24.2% of approximately 5000 existing FM broadcast stations.

- A typical radio common carrier in the top 30 SMSA markets usually has 3-5 transmitters in a given geographical area and will provide greater range and building penetration than even a Class C station.

- Class B and C stations do generate a strong RF signal, but they will also generate IM and desense problems that have not been as great a factor in traditional land mobile systems.

- FM broadcast antennas are set in place and fixed. The antenna must be located in the proper place to cover the targeted geographic area.

- A pager cannot have the range of the 1 mV/M primary contour area since its antenna is internal and is 20-30 dB below a dipole.

- In a viable subscriber system, pagers must work in steel and concrete buildings in major cities, and building losses are typically 25 dB.

- Class A stations may be practical for SCA paging when the required area of coverage is small and the FM station's antenna is located properly since fill-in transmitters can't be added.

Paging will work on an SCA, but the issues must be understood. An SCA paging system will have limited range, and building penetration problems.

With all factors considered, many suburban and rural areas are extremely viable candidates for practical systems as are limited areas in larger metropolitan areas.

APPENDIX

CODING SCHEMES

A. EXPLANATION OF VARIOUS CODING SCHEMES

While pagers are available in three basic types (tone-alert, voice, and display), there are various coding schemes which can be employed for each type of paging. Basically, there are two classifications of paging signalling, analog and binary. Analog signalling includes two-tone sequential signalling and 5/6 tone sequential signalling. Binary signalling, the transmission of a series of 0's and 1's, includes Binary Digital and Golay Sequential Coding (GSC).

1. Analog Signalling

a. Two-Tone

Two-tone signalling has been in use for over twenty years. The name two-tone describes exactly how the signalling works. A paging encoder or terminal sends out a sequence of two tones within the frequency range of 300 Hz to 3100 Hz. For voice paging, the first tone (A) is sent for one second, and the second (B) tone is sent for three seconds. The pager alerts during the three-second B tone, and the voice message follows the B tone.

Therefore, four seconds of signalling plus the length of the voice message (typically ten seconds) of airtime is used for one page. For tone-alert signalling, the tones are sent at a faster rate of .4 seconds (A) and .8 seconds (B). Therefore, only 1.2 seconds of airtime is used since there is no voice message. (Note that some two-tone systems utilize 2 seconds (A) and .1 second (B) for tone alert paging and 2 seconds (A) and .5 second (B) for tone and voice signalling).

Although two-tone signalling is considered to be slow, it has proved to be an excellent method of offering tone alert and tone and voice paging in weak signal areas. This is due to the longer signalling time which allows the pager decoder more time to decode the tones in a fading environment. However, a voice pager may alert in a weak signal environment, but occasionally the voice message may not be understood if it is too weak, since a voice message requires an additional 9 dB more signal to be easily understood. There are approximately 6500 two-tone codes available for a single channel which seriously limits the number of pagers that can be placed on a channel since individual codes may not be reused.

b. 5/6 Tone Signalling

To overcome the slow speed of two-tone signalling and increase the number of available codes, five-tone signalling was developed.

Five-tone signalling consists of a series of five tones sent in a rapid sequence. The total time required for five-tone signalling is approximately .22 seconds. For voice models, there is a one-second gap for the pager's alert tone which is then followed by the voice message.

To add a battery saver, a sixth tone, the preamble is used which increases the signalling time from .22 seconds to approximately .92 seconds for voice paging. For tone alert paging, the additional tone adds little extra time if a "batch" transmission mode is used. This is because the preamble needs only to be sent once, followed by numerous five-tone codes (each having the same preamble) which negates the need of sending the preamble for each page.

Battery saver enables the receiver to practically turn off, to save battery drain. The pager "strobes" on periodically, and if it sees its preamble, it stays on "looking to see" if its code is sent. If the proper preamble is not detected or its code isn't sent, the pager goes back into its battery saver mode. Battery saver can increase the pager's battery life by up to a 3:1 ratio.

With a six-tone system, nearly one million codes are available.

2. Binary Signalling

a. Binary Digital Signalling

Binary Digital signalling was introduced in 1973, and since then, hundreds of thousands of pagers have been put into use with this coding. Binary Digital coding is used for tone alert signalling and offers very fast signalling speed for both battery saver and non-battery saver applications. Non-battery saver signalling takes approximately .2 seconds while battery saver signalling takes approximately one second more (1.20 seconds total). However, as with 5/6 Tone Signalling, "batch" transmissions can be sent which means that this additional one second is practically insignificant. Binary Digital signalling has proven to be an excellent method of signalling pagers, especially in weak signal environments.

b. Golay Sequential Coding (GSC)

Binary Digital coding proved itself as an excellent method for tone only paging, however, it did not lend itself (in its original form) to voice and data paging applications. Therefore, in order to make it possible to perform those coding functions, it was upgraded and called Golay Sequential Coding (GSC, named after the inventor of the binary word structure, Marcel Golay).

GSC enables tone alert, voice, and data paging to be mixed in a single system allowing the utmost flexibility in system design. With GSC, there are four million addresses available and battery saver as well. It is an asynchronous code (as is Binary Digital Signalling) with excellent fade protection and bit correction which is very important for a single transmitter system such as SCA's.

B. WHICH IS BEST FOR AN SCA?

Since an SCA is a single transmitter paging system, the important considerations are; what type of paging will be offered, how many users are expected to be subscribing to the system, and which signalling scheme provides the greatest range in a typical urban, fading environment?

1. Paging Service Capacities

Before deciding on what type of paging you will ultimately offer, you must consider the capabilities of the channel with respect to various forms of paging. If a call rate in a busy hour of .2 is assumed (i.e., two calls per pager in a ten-hour period), we can estimate the maximum capacity of a channel.

PAGING SERVICE	MAX. NUMBER OF USERS (APROX.)
Tone-Alert	100,000
Tone and Voice	1,500
Numeric Data Paging (12 digits average)	25,000*
Alphanumeric Data Paging	
16-character average message	15,000*
40-character average message	8,000*

*Note: The numbers shown above assume GSC coding in all cases.

PAGING SERVICE	TYPICAL SERVICE CHARGES PER MONTH	TOTAL POSSIBLE USERS	MAXIMUM REVENUE
Tone Alert	$ 8.00	100,000	$800,000
Tone and Voice	$15.00	1,500	$ 22,500
Numeric Display	$13.00	25,000	$325,000

Therefore, the greatest potential for profit lies with tone alert paging, and the least potential for profit lies in voice paging. Note that the service charges do not include the monthly lease rates of the pager and that the maximum revenue figures are hypothetical since 100,000 tone alert or 25,000 numerical display pagers on a single system would be very unlikely. Also, alphanumeric paging is not shown since it is too early to target a typical price for service charges. Alphanumeric paging is new to the marketplace. In addition, with any type of paging, tariffing could include a per message charge over and above a set monthly charge.

Therefore, in a small, rural community where the potential for pagers in use is less than 1500 units (no matter what service is offered), voice models will yield the system operator with the highest revenues. However, in large cities where more than 1500 pagers will be in service, tone alert or numerical display paging will provide the best revenues since the system will probably reach maximum capacity quickly if voice paging is used.

2. Coding Scheme Performance

If voice paging is chosen, the coding scheme used to alert the pager becomes a very minor consideration. The limiting factor in voice paging is intelligibility of the voice message received, not in the pager's sensitivity to picking out a code in a weak signal environment. In other words, if the wearer of the pager cannot understand the voice message, it may be unimportant whether or not the pager even alerted in the first place.

However, for tone alert and display paging, there are some important differences in coding schemes performance.

For tone alert paging, the different coding schemes yield:

CODING TYPE	CALLS/SECOND	TOTAL PAGES PER MINUTE
Two-Tone	.83	50
Five-Tone	5	300
Binary Digital	5	300
GSC	5	300

Note that for two-tone signalling, .4 second (A)/.8 second (B) timing was assumed. If 2 seconds (A)/.1 second (B) timing is used, the total pages per minute DECREASES from 50 to 28.6 pages per minute.

Therefore, for most paging systems, the optimal coding to use is either Five-Tone, Binary Digital, or GSC coding schemes. Three hundred pagers per minute will satisfy the needs of practically all paging systems. Two-tone signalling is too slow for tone alert paging and is limited in capacity by the low number of individual codes available.

NOTE: The information contained in this paper represents the best information available at the time of printing. This document is not intended to be a financial planner or system design guide, but is intended as a reference for those who are considering the use of FM Subsidiary Communications Authorizations for radio paging.

The calculations included in this document to calculate radio paging distances may not apply to all geographic areas since many environmental factors can affect radio paging. However, every attempt has been made to develop general guidelines that will accommodate most situations. This information will be updated as the experience with SCA paging increases.

ARI—Automatic Radio Information

Gerald M. LeBow
Technical Marketing Consultants, Inc.
Valhalla, New York
And
Hans G. Duckeck
Blaupunkt Werke GmbH
Hildesheim, West Germany

HISTORY

Road traffic problems in major cities in the United States continue to plague city planners and motorists alike. As the economy continues to improve we can expect to see ever increasing numbers of individuals driving for pleasure and business. With that will come increased traffic jams, air pollution from idling vehicles and frustrated motorists.

This same scenario took place in Europe more than 10 years ago and because of its negative effects on the economy and ecology, various governments decided to find a way of improving traffic flow in Europe. Under the auspices of the European Broadcasting Union (EBU), these governments wanted to first determine if timely, relevant information regarding traffic situations, alternate routing, severe weather and accidents was getting to drivers via radio.

In Europe, broadcasting facilities are often operated by governmental agencies. Similar governmental agencies control the traffic and highway departments as well as the local and state police authorities. Because of this close relationship between broadcasters and traffic authorities, traffic information was and is readily available to European radio stations. The question then was why drivers didn't receive this information and/or take advantage of it? Research found 4 major reasons why available traffic information on radio was not effective in diverting or controlling traffic.

1) Drivers were in their car and had the radio turned off, hence no traffic information was reaching them.

2) Drivers did not know which station(s) were providing traffic information for their area. They might be listening to a station giving traffic information for another area or might dial across the AM or FM band for hours trying to find any traffic information.

3) The most common problem was that the driver had the radio turned on, tuned to a station delivering traffic information but he simply didn't "hear" it through the clutter of music, news, sports, weather or because he was thinking about the days events.

4) The driver was listening to a cassette and did not hear the traffic or severe weather messages.

Various European governments determined that if a system could be developed to automatically deliver traffic information in a high profile manner, it would go a long way to solving traffic problems. In 1970 investigations were started as to how traffic and severe weather information could be delivered. Proposals ranged from the creation of a network of new radio stations on long wave, medium wave and VHF frequencies to using existing AM and FM broadcast stations. After detailed studies, it was decided that FM radio stations were the most practical because FM broadcast stations already existed throughout Western Europe and covered the entire land area. Second, FM radio stations have a defined coverage area, unlike AM stations which, at night, can cover areas far removed from their primary service area.

In order to solve the abovementioned 4 problems using existing FM broadcast stations, the **ARI** system was developed in Germany in the early 1970s.

STATION IDENTIFICATION

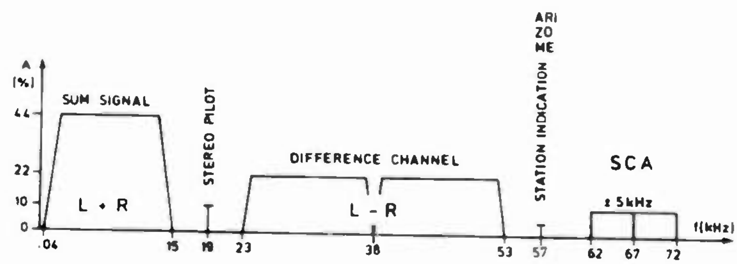
First of all, there must be a means to distinguish those radio stations that transmit traffic messages from other stations. This can be accomplished with the aid of a pilot signal, but this pilot signal must fulfill three important conditions:

1) The pilot signal must be inaudible and shall not cause any interference to main channel and sub-channels, even under difficult reception conditions.

2) It must be compatible for mono and stereo transmitters and other subcarrier systems.

3) It should contribute as little as possible to the frequency deviation (modulation) of the transmitter in order to avoid reducing the loudness of the station.

Because this pilot signal has to be inaudible, the frequency must be above 16 kHz. If we look at the modulation spectrum of an FM stereo transmitter, we see that the stereo pilot is at 19 kHz. The left-minus-right channel is between 23 and 53 kHz. Therefore the inaudible signal must be above 53 kHz. After careful investigations it was proven that the frequency of 57 kHz can fulfill all three conditions if it is properly used.

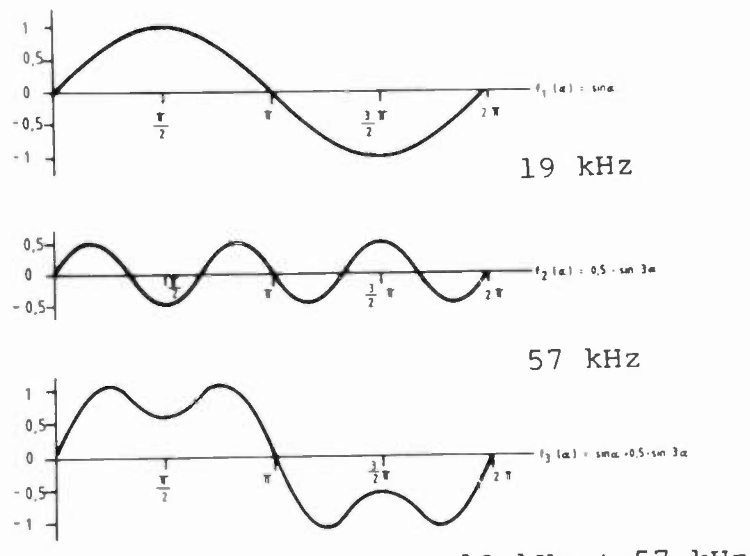


MODULATION-SPECTRUM OF FM-STEREO-STATIONS

A. 57 kHz is far above the audible range, and cannot interfere with mono transmitters. It is also above the band which is used for stereo transmission, and cannot interfere with such transmissions either And it is sufficiently below the 67 kHz SCA channel. Thus, condition No. 2 is fulfilled.

B. This frequency being derived from the 19 kHz stereo pilot (3 X 19 = 57), no beat frequencies can occur if, for example, in multipath reception areas harmonics of the 19 kHz frequency are generated which may happen to be 57 kHz. Thus, condition No. 1 is fulfilled, too.

C. If the phasing of the frequencies 19 kHz and 57 kHz is adjusted so that the zero crossings occur at the same time and in the same direction and an amplitude of the 57 kHz is 50% of the 19 kHz, the addition of these two frequencies yields a saddle curve that increases the total frequency deviation of the transmitter by approximately .75% which can be neglected. Thus, condition No. 3 is also fulfilled.



STEREO PILOT + ARI SIGNAL

Radio stations using the **ARI** system constantly transmit this Station Indication Frequency of 57 kHz. **ARI** car radios are equipped with a decoder for detecting the 57 kHz frequency. If the motorist wants to tune to an **ARI** station, he depresses the **ARI** button on his car radio.

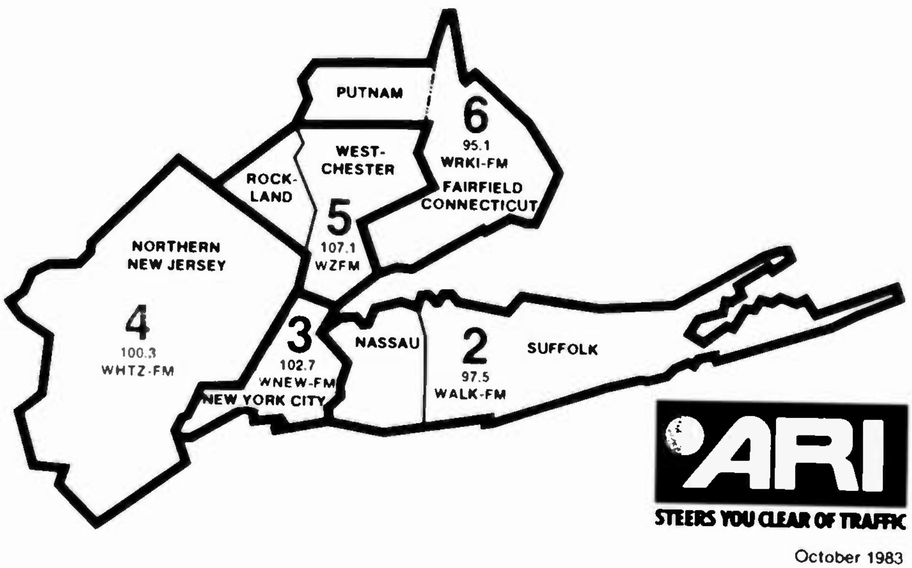
If he is tuned to a non-**ARI** station, the radio is muted. Now the motorist turns the tuning knob without looking at his radio. As soon as he has found an **ARI** station, the 57 kHz signal is detected by the decoder which, in turn, activates the receiver. The program of the station then becomes audible. Thus, the motorist can tune to a traffic information station while paying full attention to the traffic. There is no need to know the frequency of the station or to look at the dial scale.

With an Electronically Tuned Radio (ETR) model, the search tuner starts to run when the **ARI** button is depressed and will stop only on stations that transmit the **ARI** signal.

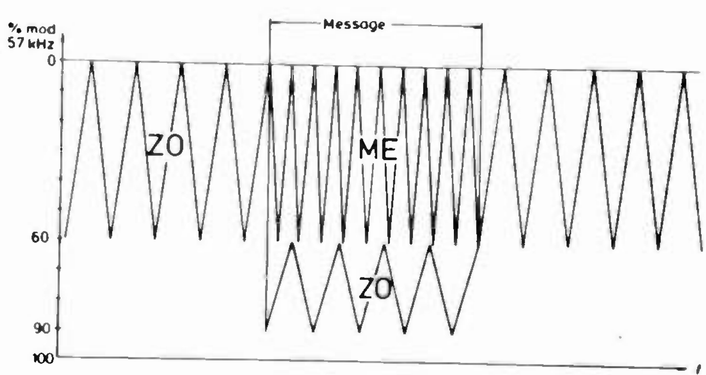
ZONE INDICATION (ZO)

In areas of high traffic density, one radio station cannot efficiency transmit all information on traffic incidents that occur in its coverage area, e.g., motorists being south of the radio station may not be interested in incidents that are happening in the northern area. For that reason such areas are sub-divided into traffic information zones and the incidents of the different zones are transmitted by different radio stations, although their coverage areas might be nearly identical. In order to distinguish these areas, the zone indication (ZO) was introduced. For this purpose, ten very low frequencies between 20 and 122 Hz were selected.

In order to simplify the operation for the motorist, these frequencies are designated by numericals from 0 through 9. Stations serving different areas transmit different zone indication frequencies. The indication numerical is shown on the **ARI** maps that accompany the **ARI** car radio. When an **ARI** station is tuned in, the zone number appears in a display on the dial scale or on the **ARI** adapter.

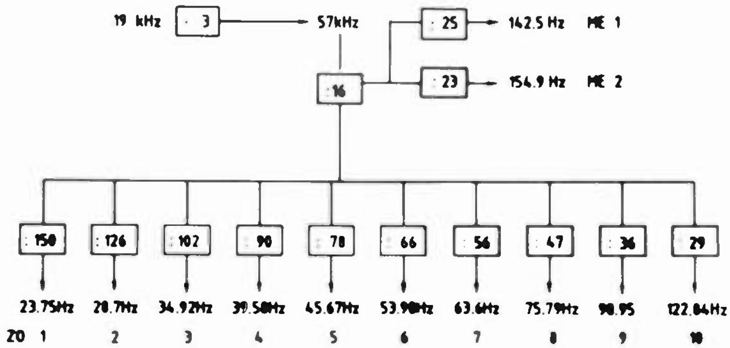


The ZO signal, which is constantly transmitted, is amplitude-modulated to the 57 kHz **ARI** pilot. The modulation degree of this signal on the 57 kHz pilot is normally 60%. During the traffic message, when the message signal is modulated at 60%, the modulation degree of the ZO frequency is reduced to 30% to ensure that the 57 kHz **ARI** frequency is not overmodulated.



ME/ZO MODULATION-SCHEME

The zone frequencies are created by repeatedly dividing 57 kHz. All zone and message frequencies are correlated directly to the 57 kHz and 19 kHz pilots eliminating the possiblity of interference or noise during severe multipath.



ME AND ZO FREQUENCIES

In the case of a more sophistiated car radio with active **ARI** decoder, the zone indicating number can be pre-selected. In this case the electronic tuning of the radio will automatically search for a station that transmits the pre-selected zone signal and will only stop on such a station. This is not only useful as long as the motorist is driving in the given zone, but it will also help him to identify the station, in a zone where he is going so as to get advanced information about road conditions ahead.

TRAFFIC MESSAGE (ME 1)

This message signal has a frequency of 142.5 Hz and is amplitude-modulated to the 57 kHz **ARI** pilot with a modulation degree of 60%. Being modulated on the 57 kHz pilot, it remains inaudible on the main channel. This frequency is transmitted throughout a traffic message and is detected in the car radio by the ME 1 decoder. The volume of the car radio can be turned down to an entertainment level during music and normal program. As soon as the radio station transmits the Message Indication signal, this will be detected by the decoder, which in turn raises the volume of the radio to a pre-set level which ensures that the motorist hears the traffic message.

If the motorist is listening to a cassette, the cassette motor is stopped by the ME decoder as soon as the message signal appears. The message is then switched to the audio amplifier and the motorist can hear it. At the end of the traffic message, when the ME signal is switched off at the station, the decoder will switch back to cassette.

EMERGENCY MESSAGE (ME 2)

If a motorist does not want to be interrupted by traffic messages, he can release the **ARI** button.

For purposes of emergency warning, we introduced a second message signal (ME 2) the frequency of which is 154.9 Hz and which is also amplitude-modulated to the 57 kHz pilot with 60% modulation degree during important emergency messages. The decoder for this signal is always alive in the **ARI** decoder, which ensures that important emergency messages will be heard by the motorist even with the traffic facility switched off.

This emergency alerting system is now being tested over radio station WSTW-FM, Wilmington, Delaware in cooperation with the states of New Jersey, Delaware and Public Service Electric and Gas Company at their Salem Nuclear Power Station.

DIGITAL DATA (ME 3)

Using a specially designed FSK/PSK technique, it is possible to transmit slow speed data on the 57 kHz pilot. This data could be used as a utility load management system to turn on and off addressable loads. It can also be used as a point to point data signaling system.

ARI WARNING TONE

A motorist driving his car while listening to a cassette, or having turned the volume down, may leave the coverage area of his **ARI** station without noticing it. For that reason a device is incorporated in the **ARI** decoder that, approximately 30 seconds after the station has disappered, generates an audible warning signal. The motorist will then release the cassette or turn the volume up, turn the tuning knob until he can hear the program of an **ARI** station, and then return to the original setting.

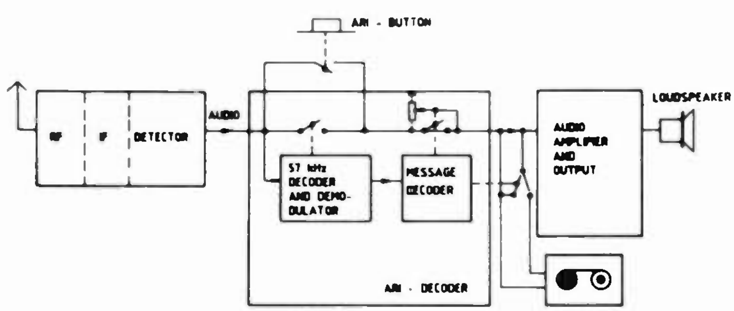
If he has an ETR model, it is sufficient to press the search button. The radio will automatically tune to another **ARI** station. If within 30 seconds no **ARI** station is found, the warning tone will come back, telling the driver that there is no **ARI** station in the area. He will then release the **ARI** button and the warning tone will stop.

DIFFERENT TYPES OF ARI CAR RADIO MODELS

With the **ARI** system, the motorist has the possibility to purchase the level of sophistication he needs.

BASIC ARI

Low cost car radios are equipped with a decoder for the **ARI** pilot signal and for the message indication. With such a radio the motorist can distinguish **ARI** stations from non-**ARI** stations, he can easily tune in the **ARI** stations. He can turn the volume to normal listening level or he can listen to a cassette, but will always get the traffic messages.



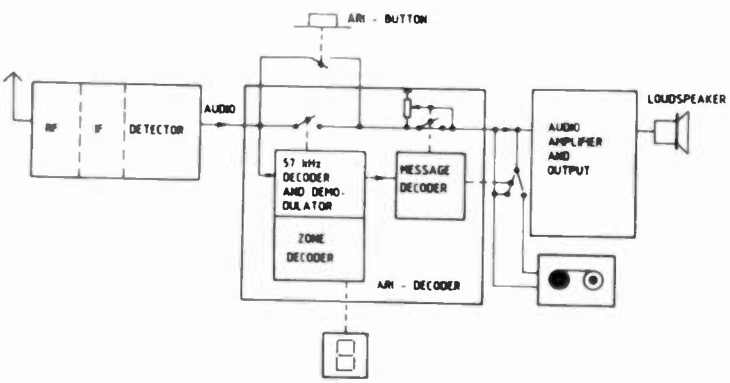
BASIC ARI MODEL

PASSIVE **ARI** DECODERS

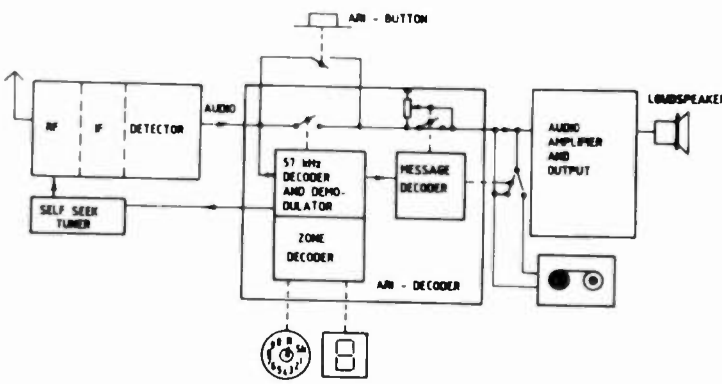
In this case, the **ARI** decoder does not only detects the **ARI** signal and the ME signals, but also the ZO signal and shows it on a display. With this decoder, the motorist can see to which of the **ARI** stations in the **ARI** traffic area he is tuned.

ACTIVE **ARI** DECODERS

An active **ARI** decoder has an additional selector, push-button, sequential switch or rotating switch which is used to pre-select the desired zone, the messages of which the motorist wants to hear. After selecting the zone number, the search tuner starts, and stops only at a station transmitting the selected zone number. A display shows the zone number of the station that has been found.



PASSIVE ARI MODEL

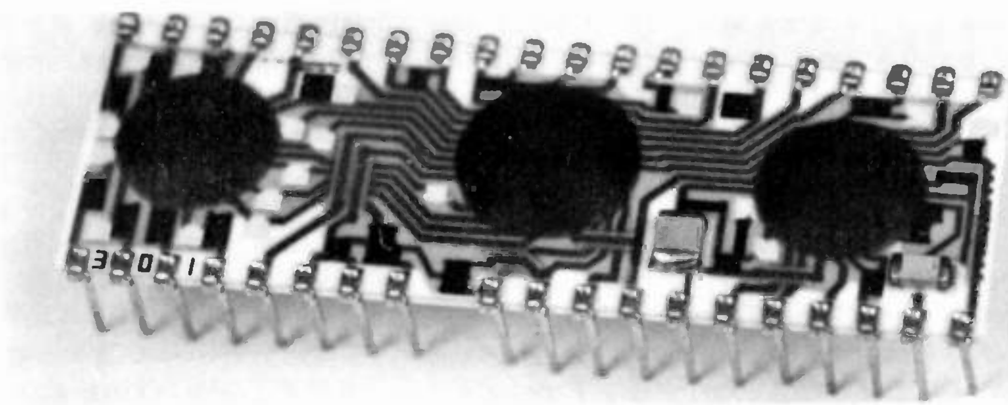


ADD-ON DECODERS

During the introductory period of the **ARI** system in the United States, it would be premature to sell all car radios with integrated **ARI** in areas where the system is not yet operational. For that reason, recent car radio models are equipped with a connector on the rear for subsequent connection of an **ARI** decoder. As soon as the **ARI** system is operational in the area, the motorist can buy an add-on decoder, which is either of the active or of the passive type, plug it into the car radio and fasten it under the dashboard.

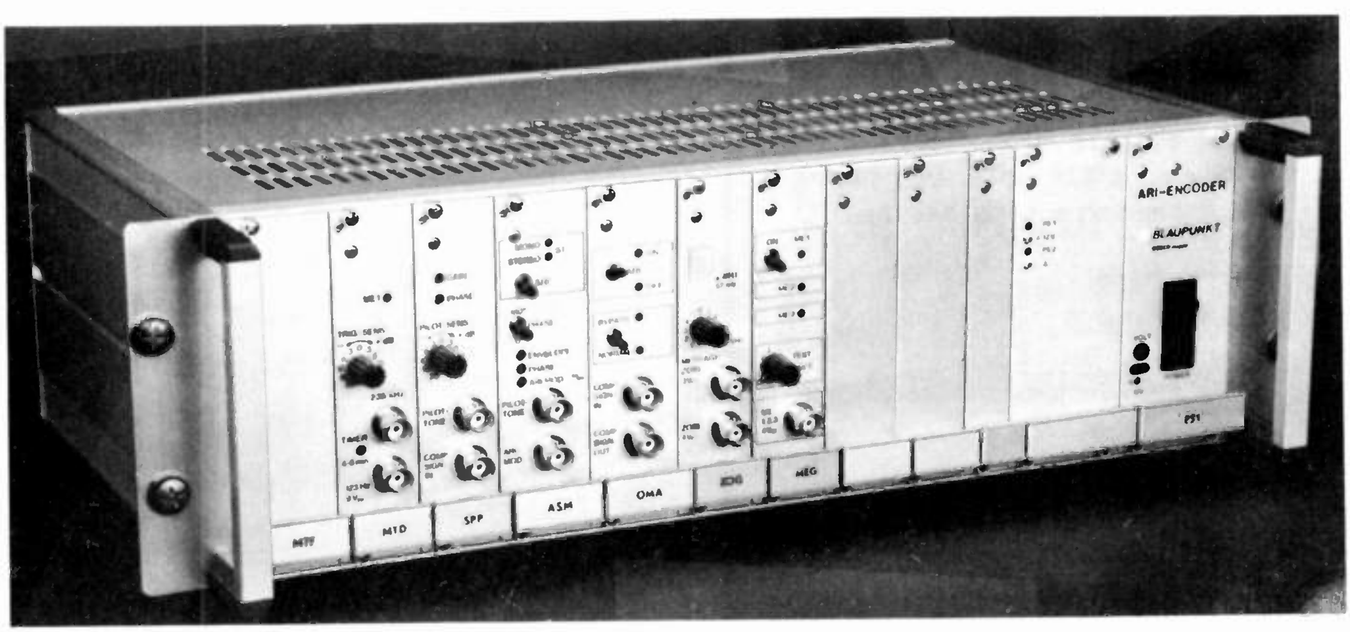
ARI DECODER CIRCUITS

Besides thick-film hybrid circuits for Basic **ARI** receivers, Blaupunkt developed a microprocessor decoder that can be used for active and passive decoders. This device is located on a thick film substrate, the size of which is only 18 mm by 50 mm.



ARI **ENCODERS**

For generation and injection of the **ARI** signals into the modulation of an FM station, an **ARI** encoder is needed. This device was developed especially for use in North America. Built into a 19" standard cabinet, it is connected into the MPX path between stereo generator and exciter of the transmitter. It is designed to interface with practically all types of transmitting equipment used in North America. In case of a problem, the encoder switches automatically or manually to a hard-wired bypass mode. Once installed and adjusted, the encoder is very stable and requires no special maintenance or testing.



ARI IN THE UNITED STATES

Prior to FCC deregulation of subcarriers, specialized subcarrier use required extensive testing and reports to the Federal Communications Commission. In 1981 the **ARI** technology was operated on a test basis at WVNJ-FM licensed to Newark, New Jersey but serving the New York metro area. Silent tests were conducted for over 8 months and numerous measurements and reports were filed with the Commission. On the basis of those reports and the fact that not a single listener had ever noticed the presence of the **ARI** system on WVNJ-FM, the Commission authorized commercial use of the **ARI** system in the United States in September 1981.

In April of 1983, the **ARI** system was officially launched in the New York metropolitan area on 4 radio stations WNEW-FM, New York City, WALK-FM Patchogue, Long Island, WZFM White Plains, New York and WVNJ-FM (now WHTZ-FM) Newark, New Jersey. In September of 1983 WPST-FM Trenton, New Jersey, WIOQ-FM Philadelphia, Pennsylvania and WSTW-FM Wilmington, Delaware were added as well as WRKI-FM Brookfield, Connecticut. This created an **ARI** traffic corridor from New Haven, Conecticut to the Maryland border. On March 13, 1983, 2 more **ARI** radio stations, WOMC-FM and WHYT-FM were operational in the Detroit Metro area.

The **ARI** traffic system is now in the process of being implemented in the top 20 markets of the United States with the next major area being Southern California from San Diego to north of Los Angeles in time for the 1984 Summer Olympics.

CONCLUSION

The **ARI** system has proven itself to be useful and effective in the dissemination of traffic, severe weather, mass transit and emergency information. Due to the low cost of receiver implementation, virtually every car radio sold in West Germany and many other European countries now contain **ARI**. If the same experience follows in the United States, we can anticipate improvement in traffic flow over the next few years and can expect to see **ARI** as a standard feature in car and home radios.

REFERENCES

Paul Ad. Becke
Das **ARI**-Verkehrswarnfunksystem, ADAC 1971

P. Bragas
Verkehrsrundfunk; Bosch Techn. Berichte
1973/5

H. Duckeck, Blaupunkt Werke GmbH
ARI-Automatic Radio Information;
SAE Detroit 1984

DATA TRANSMISSION VIA FM SUBCARRIER

Eric Small

Modulation Sciences, Inc.

Brooklyn, NY

Transmitting data on the SCA of an FM station places unique demands on the broadcast system. Quantifying the needs of data transmission and relating how one company designed hardware to cope with those requirements will, hopefully, provide a general insight into the nature of data communications via an FM subcarrier.

Traditionally, broadcast engineers have been able to characterize transmission channels in terms of frequency response, distortion, and signal-to-noise ratio. Whether AM or FM, the performance of the channel could be described in a meaningful way by these analog parameters. While there is a relationship between these well understood analog parameters and digital performance, it is indirect and complex. Data communications managers need the channels they use characterized in a way that relates directly to data transmission. The parameters of most interest are <u>speed</u> and <u>accuracy</u>.

Speed

Speed refers to how much information can be transmitted per unit time. It is measured in bits per second (bits/sec) not to be confused with baud. The frequently misused term "baud" measures something entirely different from information rate. More on the correct definition of baud later in this paper. The speed parameters of a system are fixed, deterministic values. They are most often designed into a network and not subject to change.

Accuracy

All "raw" transmission channels have some finite error rate, regardless of how well designed. Any system that has been properly tested has a number attached to its error rate. That number may be very small, error rates of one bit for every one hundred million transmitted are not unheard of, but it still is an error rate. Beware of any claims that a system has such a low error rate munications attorney should be sought before putting any encrypted SCA on the air.

Conclusion

Data transmission via SCA is a viable technology, especially for point to point applications. Because it promises to be the least expensive technology to deliver one way data to many locations over a wide area, it is not easy to access what its final impact will be.

Data on subcarriers are a new technology with vast potential for the entrepreneur. I hope this paper has shed some light on their technical potential.

REFERENCES

The following books will give a good overview of data communications:

1. "Modern Data Communications, Concepts, Language, and Media" by William P. Davenport, Hayden Book Company, Inc.

   The best introductory level data communications book around. Very light on the math and theory, but full of good examples. Mostly telephone line oriented, but still a good start.

2. "Technical Aspects of Data Communication", second ed. by John E. McNamara, Digital Press.

   An excellent engineering text. It does not assume any background in computers or communications and develops all its concepts very clearly. It is also practical, containing the text and full explanations of many popular data communication standards. Casual reference to various standards is often one of the most confusing aspects of trying to understand data communications.

3. "Telecommunications and the Computer", second ed. by James Martin, Prentice-Hall.

   The standard reference on data communications from the point of view of the computer user. Essential reading for anyone wanting to talk to computer people about data transmission.

4. "Coding and Information Theory", by Richard W. Hamming, Prentice-Hall.

   The definitive book on error detecting and correcting codes by the man who started it all. Hamming is a gifted teacher and keeps the math understandable. This is probably the only readable book on the subject.

5. "The Network Revolution", by Jacque Vallee, AND/OR Press.

as to be insignificant. Such a statement shows a lack of understanding of data transmission or insufficient testing. Bear in mind that even an error rate as low as one part in one hundred million (almost never achieved in the real world) through a system carrying ten thousand bits/second (a typical speed) means an average of one error every three hours.

Error rates are determined statistically. It takes many hours of testing under various conditions to determine error rate. As a rule of thumb, testing should continue until at least ten errors are accumulated. For a low error rate system this may take days or even weeks of continuous testing. Error rates are generally expressed as errors in so many bits. Numbers like 1 in 10^5 or 1 in 10^7 are typical. Accuracy can also be expressed as an error probability such as 1 in 10^{-7} chance of an error. The numbers are reciprocals of one another and which expression is used depends only on taste.

Synchronous vs Asynchronous

Another important characteristic of a data circuit is whether it is synchronous or asynchronous. Asynchronous is the more common technique. It has been employed with electromechanical teleprinters for more than fifty years and is universal with personal computers today. Typically, eight bits of information (seven data bits plus parity) occur between a start and stop bit. The presence of the start and stop bits allows the information to be transmitted randomly in time. A string of characters may be transmitted one right after the other, or seconds (even hours) may elapse between characters. Either way, the data will be received correctly. In synchronous transmission no start or stop bits are transmitted. The data must be transmitted in large groups, from several hundred to many thousands of characters long. Synchronization is established at the beginning of the block and then the data must be transmitted in a continuous stream.

Synchronous transmission is obviously more efficient because each character takes only seven bits to transmit versus ten for asynchronous, thus for the same channel, synchronous will carry 30 percent more data. The disadvantages are that the data must be grouped in blocks prior to transmission and a transmission error during a block will invalidate the whole block.

Direct and Indirect Modulation

There are many methods of putting data on an SCA channel, especially since deregulation has removed all restrictions on allowable modulation. Broadly viewed, there are two methods of data modulating on SCA—indirect and direct. Indirect has been in use for some time and many engineers are familiar with it, so we'll explore it first.

Indirect data modulation of an SCA could also be called "compatible modulation" because the thing that you plug into the SCA generator is audio--good old garden variety stick-your-headphones-across-it-and-listen audio. It's nice because a standard SCA generator is used unmodified--just like background music. In its simplest form, called AFSK for audio frequency shift keying, an audio oscillator is switched between two frequencies. One frequency, like 1070 Hz, represents a zero and another, like 1270 Hz, is a one. This method is cheap to generate and easy to handle, but it is slow. The maximum speed possible on a standard SCA channel is 1200 bits/second. The receivers, although conventional audio SCA units, need a relatively expensive demodulator to turn the audio back into data. By switching the phase of an audio tone instead of its frequency, higher data rates can be achieved, but the cost of the receiver demodulator rises very rapidly with speed, and conversely, the tolerance of the system to interference (crosstalk) drops quickly.

One reason why indirect data modulation remains popular is that it employs the same technology used to send data via the dial telephone network. Thus there is a lot of hardware around and many people understand it. Direct data modulation however, takes advantage of something available to the SCA user--which is not available to the dial telephone network user--the subcarrier itself. Subcarriers exist in telephone transmission, but only at the central office. The SCA subcarrier, typically at 67 or 92 kHz, is available for direct manipulation by the broadcaster. By doing so, higher data speeds at greater accuracy can be achieved with less bandwidth--and the receiver demodulators are cheap too.

Direct modulation, when compared with indirect, is definitely a free lunch. The technique is very simple. Instead of varying the frequency of an audio tone that in turn modulates the SCA subcarrier, the frequency of the subcarrier is itself varied. For example, a "zero" could be 66 kHz and a one 68 kHz. There are some niceties to be observed, (like controlling the spectral garbage created by the sudden switch in frequency and having the kind of coding scheme so the average frequency of the signal remains stable regardless of the data pattern); but these problems were solved long ago for other applications and the solutions are in the text books.

<u>One Company's Solution</u>

When Modulation Sciences approached the problem of designing a data SCA system we had the goals to produce a system that:

1. Was fast

2. Occupied no more baseband spectrum than a music SCA

3. Had an error rate equal to or better than that provided by telephone data service

4. Used the cheapest possible data decoders in receivers

The system that resulted, called DATA-SIDEKICK, has the following specifications:

1. 4800 bits/second data rate

2. A minimum bandwidth of only 16 kHz

3. A <u>measured</u> bit error rate of 1 part in 10^7 (the telephone company struggles to deliver 1 in 10^5)

4. A data decoder that uses about eight dollars worth of parts

There is nothing sacred about these performance figures. The transmission speed could easily be doubled or even tripled, but the error rate would degrade somewhat (probably still better than Telco) and mostly the receiver demodulator cost would skyrocket.

Performance Limits

What limits the performance of a data SCA? First, foremost and hardest to quantify is multipath. Under the right conditions a second RF signal 20 dB below the desired can completely wipe out an SCA-- any type SCA, if its time delay is of the critical value. In assessing how resistant to multipath a data SCA system will be, the first question must be "What is the lowest signal to noise ratio at which it will operate at for a specified error rate?" For the MSI system a 10 to 15 dB SNR will produce a minimum error rate of 1 in 10^5. It should be noted for indirect modulation an SNR of 20 to 25 dB is needed. Because no "standard multipath" test exists, any statements beyond the preceeding must be viewed as subjective, and since they are being made by a less-than-disinterested party (the author is Vice President of Engineering of Modulation Sciences), they may be viewed with some suspicion. However they are useful to the extent that others are encouraged to duplicate these tests as a basis for comparison between systems.

Modulation Sciences is located in downtown Brooklyn, New York. The Empire State Building and the World Trade Center from which the test FM stations broadcast are only a few miles away, but due to local buildings, neither are line-of-sight. In addition,several large local buildings are on more or less reciprocal bearings to transmitter sites. For the purpose of controlled tests, a SCALA FM log periodic was erected with an antenna rotor. It may seem odd that what is without doubt the finest available FM receiving antenna was used in these tests, but after attempting to use other more "typical" antennas it was felt no option existed. The VSWR of consumer FM antennas varied wildly over the FM band, some as bad as 4:1. Their patterns appeared to vary from day to day. When the SCALA was installed, with its maximum VSWR of 1.4:1 across the band a front to back ratio of 25 dB, the data became consistent over time. Only heavy rain, which changes the coefficient of reflection of brick at 100 mHz, altered the results significantly.

Many weeks of testing took place. Most involved WBAI transmitting data at 92 kHz from the Empire State Building. Using the outdoor antenna the claimed error rate is 1 in 10^7. However, after weeks went by with no errors, and a case could be made for a rate of 1 in 10^9. Such figures seemed too good to be true, so the more conservative claim is made.

Using a simple whip antenna the error rate degraded to 1 in 10^5. That is still much better than the telephone can normally provide. With a whip antenna most errors could be traced to electrical noise in the building or the cancellation effects from people moving about the room. Thus it should be expected that the performance of a whip antenna will vary tremendously with the exact circumstances of its location. Our tests also convinced us that any simple whip antenna will not perform acceptably in a high rise office building in a densely populated urban area. The same appears to be true for body worn or hand held receivers except in wide open spaces. However, the performance of data SCA with even a simple outdoor antenna is nothing short of spectacular. Error rates of a hundred or even a thousand times better than the best the telephone company can deliver are the rule rather than the exception. And the cost of terminal equipment is about 25 percent that of a modem. Most importantly, there is no monthly line charge.

Multipath Tests

Testing for multipath susceptibility was accomplished by two techniques. The first by noting how much the beam could be turned from pointing directly at the station before the error rate climbed dramatically. Although this method lacks any quantitative standard, it does allow for comparison of different systems in a dramatic way. This method makes it easy to see if a circuit change has improved or degraded performance.

A second method simulates the effect of multipath. Ths is most useful in closed circuit laboratory tests. A notch filter with a Q of about 3000 is inserted in the FM baseband. A fine tuning capacitor controls the position (frequency) of the notch, and a bypass attenuator controls the depth of the notch. While this is not true multipath, the effect is similar. And a filter is cheaper and much more versatile than a fixed length coaxial delay line.

Other Types of Modulation

The direct frequency shift approach to data SCA taken by Modulation Sciences is hardly the only practical technique available. Rather, we felt it was the best suited to the goals outlined earlier. For alternate objectives, such as increased speed, other types of modulation should be considered. The obvious approach to speeding up the system--increasing the frequency shift and moving the center frequency to the middle of SCA portion of the baseband works poorly. That's because the bandwidth of the demodulator must be increased to a point that allows more additional noise than additional signal. Also, direct FSK does not use the baseband spectrum very efficiently.

One way to gain increased data rate without greatly increasing bandwidth is, instead of merely switching between two frequencies for 0 and 1, to use four frequencies to represent, 0, 1, 2, 3. This is called a four level or "dibit" code. It is more expensive to decode, but effectively can double the data rate in a given bandwidth. We need not stop at four levels. Eight levels will still further increase the transmission speed.

It is in these systems of signalling that the difference between the terms "bit per second" and "baud" becomes apparent. For example, if a 2400 bit/second data stream is transmitted on a subcarrier that has four possible states, then the baud rate of the transmission system is 1200 baud. A baud refers to the number of times the line condition changes per second. If the transmission system were two level instead of four, then the baud rate would be 2400. So baud rate, when used correctly (a rare occurrence), is a good indication of the bandwidth needed to transmit data. Of course there is a price for the reduced bandwidth of multilevel coding--the signal-to-noise ratio must be good.

Phase shift keying (PSK) is another technique for higher speed data transmission. With this technique, the phase of signal is switched between two or more states to represent digital information. To decode such modulation a phase reference is needed. It may be internal--the average phase of the signal itself--or external. Deriving the phase reference or clock from the signal itself is the most difficult part of receiving PSK. However in stereo FM broadcasting an ideal external phase reference already exists--the 19 kHz pilot. A pilot tone locked PSK system seems to hold the greatest potential for high speed data transmission on SCA. It has the potential for as much as 56,000 bits/sec. This of course could take the entire spectrum from 53 to 99 kHz and the receivers would be quite expensive, but it certainly can be done.

Error Prevention

Error coding is an essential part of any data transmission system, although this function is generally handled in the software of the sending and receiving computers. Error coding can be used to absolutely guaranty that the data received is correct. There is error detection coding which does exactly what the name implies. Error correction coding allows errors to be fixed.

In error detection, some extra data is transmitted every few hundred characters. This extra data, representing only a few percent of the total data transmitted, carries a code that when processed with the data just received will indicate with incredible reliability if the data was correctly transmitted. It is not unusual for error detecting schemes to be able to insure that information can be transmitted continuously for years before an error will slip through.

Modern error detecting codes should not be confused with character parity as implemented by the ASCII standard. Most experts agree that parity is of little value in detecting transmission errors. The Cyclic Redundancy Checks (CRC) are the class of codes commonly used for reliable error detection.

Error correcting codes are even fancier. As their name implies, they allow for the repair of corrupted data. Error correcting codes work similarly to error detecting codes, but require the dedication of a much greater percentage of the total data transmitted. Depending on the exact type code used, they allow for the correction of errors of one, two or even more bits in each block. They also function as error detectors, so even if there are too many errors to correct, they will indicate that the data is wrong.

Data error rate is measured by transmitting a known data pattern. The recovered data is compared at the receiver with what should have been transmitted. Any discrepancies are counted. In practice a linear sequence generator (LSG) creates a pseudo random bit pattern for transmission. At the receiver, an identical LSG is forced into synchronization with the transmitted data. An exclusive or (XOR) circuit then compares the two data streams.

More Than One Data Stream

Multiplexing means several signals sharing the same carrier. Stereo FM is multiplexing. SCA is multiplexing. In the analog world multiplexing is very imperfect. A 15 kHz channel can not be easily split into three 5 kHz channels. Guard bands are needed between channels and the filtering becomes complex. Multiplexing in the data domain, however, can be nearly ideal.

A 4800 bit/second channel can be split into sixteen 300 bit/sec. channels. Only a few percent of the data channel is required for additional housekeeping. In practical terms this means that sixteen different wire services all running flat out at 30 characters per second can be transmitted at the same time over one SCA. The technique to accomplish this is called time division multiplexing (TDM). Each of the individual channels is allocated a slot in a fixed data frame and that slot is wasted if there is no data to transmit when its turn comes up. That is why TDM is really only efficient when all the channels are used most of the time.

If the demands of each individual channel are not constant, then other forms of multiplexing become practical. Depending on what the total need of all the individual channels are, more than one hundred can be multiplexed together. This is called statistical multiplexing or packet multiplexing. In this scheme, the data for each channel is broken into groups or packets and assigned a header designating its channel number. When a packet is assembled, it is placed in a queue for transmission. At the receive end, the packets are identified and sent to the appropriate destination. The advantage of such a system is that the transmission system is used at its maximum efficiency much of the time. If the usage is light, then each user gets nearly the full capacity of the system. When the demand exceeds the total capacity of the transmission system all that happens is a delay is created. So long as the average demand on the system is not greater than the total capacity, then the delays will be only temporary.

Multiplexing offers the capability of serving a large number of diverse users of data SCA in an inexpensive and efficient fashion.

Security

For many applications the need to provide security for transmitted data is essential. This is particularly important in the case of SCA, since the data is being broadcast on an essentially open channel.

Security can be thought of as existing in layers. The first layer is provided by being an SCA. That discourages at least casual listeners, especially at 92 kHz where decoders are less common. The use of other nonstandard SCA frequencies provides somewhat greater protection. The nature of the data modulation provides additional protection to eavesdropping. Most of the data transmission techniques are proprietary and observation on standard test equipment does not provide too much insight into the coding techniques. Multiplexing and synchronous transmission makes decoding without prior knowledge of the transmission parameters very difficult.

All of the above will provide good protection against interception by an outsider, but if any users' messages must be protected from interception by another user, then some form of encryption will be needed. The simplest way to accomplish encryption is to use the National Bureau of Standards Data Encryption Standard (DES). The details of the standard are beyond the scope of this paper. The DES is effective and universally accepted by data processing people. It is however, expensive to implement because the standard demands that the encoding and decoding be done in hardware and the special IC's are costly. Other, less expensive coding schemes exist that can be efficiently done in software, although not as well respected as the DES, they will provide a good deal of security.

Any station licensee planing a data SCA, especially an encrypted one, should bear in mind the position of the FCC regarding the responsibility for information broadcast on the SCA. See Section 73.295(e). The advice of a com-

A nontechnical paperback about the impact of data communications on everyday life. Well wriiten and entertaining.

6. "Digital Communications: Microwave Applications", by Kamilo Feher, Prentice-Hall.

A heavy-duty engineering text. Lots of math. Everything you ever wanted to know about putting data on the radio.

MERPS - A New Generation of Multicassette Machines

Rupert L. Stow

CBS Broadcast Group

New York, New York

INTRODUCTION

Nine months ago, the NAB Engineering Advisory Committee elected to form a subcommittee to define the user requirements for a standardized video cartridge for the next generation of multi-cart machines.

As Chairman of this subcommittee, I would like to report today on its work, and discuss the genesis of the requirements that have been defined.

The members of the subcommittee were chosen to represent all types of user, and included advertisers, advertising agencies, distributors, TV networks, group broadcasters, independent stations and cable MSOS.

The acronym for the subcommittee is MERPS, or Multi-Event Record/Playback System.

This rather arcane title is more general than "multicassette machine", because it was foreseen that some medium other than a tape cassette could meet the requirements, such as a videodisc or some other as yet undeveloped medium.

However for convenience and illustrative purposes, this report makes reference to "cassettes", meaning a physical medium for the recording, storage, and playback of video and audio signals.

Firstly, I would emphasize that in this first phase of our work, the MERPS Subcommittee was concerned only with the functions and features which users require. We leave to the manufacturers to determine how the requirements are to be met.

The specification of users' performance objectives has been sent to over 120 manufacturers, and we have asked each of them to comment without delay on the requirements sent out, and to indicate the contribution they can make.

BACKGROUND AND NEED

For the last decade, broadcasters have employed multicartridge machines for the playback of commercial and promotional messages. These machines have performed yeoman service, and some 900 of them are currently in place.

But they are old and tired, maintenance costs are rising, and they all employ the now obsolescent two-inch quadruplex format.

In fact, two manufacturers, Ampex and RCA, developed excellent but mechanically incompatible systems. Because these two systems used mechanically different cartridges, and because many stations used no multicart machines at all, the distributors of commercial messages were inhibited from shipping cartridges. The cartridges, moreover, were relatively expensive; they were bulky and heavy, and were costly to ship.

As a result, distributors continued to ship commercials in a single common format, originally two-inch quad tape reels, and more recently, one-inch "C" format tape reels.

While this was an economically realistic solution for the distributor, the broadcaster was left with the task of transferring the tapes received to the two-inch cartridge format. This format transfer operation has been onerous for broadcasters, and the annual costs to a typical television station in labor, expense and the use of capital facilities, is over $250,000.

Moreover, the transfer operation necessarily led to a degradation of video quality.

Thus it is that a need - and an urgent need, now exists for a new generation of multicassette machine systems, employing current technology and providing a high quality and cost effective operation for the playback of prerecorded material.

While a machine system meeting the MERPS requirements may use many different designs, it is especially important for all sides of the industry that a single uniform recording format be employed by producer, distributor, and broadcaster.

THE PURPOSE OF THE MERPS SUBCOMMITTEE

With this need in mind, it was the primary purpose of the MERPS Subcommittee to develop user performance requirements for a machine system to be used in the playback of commercial and promotional messages for broadcast or cable distribution, all on a single industry-wide recording format.

A machine system meeting such requirements would save the time, labor, and expense now borne by broadcasters, who today must transfer the messages as received from one format to another format required by their particular multicartridge machine.

It was a further purpose of the MERPS Subcommittee to require that the cassette or other message container, be small and light to reduce shipping costs, and that the format chosen should provide a high audio and video signal quality, at least equal to that provided by one-inch "C" format videotape. If therefore these purposes could be effected, three main objectives could be achieved.

Labor, expense and time now expended in format transfer would be obviated.

Distribution and shipping costs of cassettes would be reduced.

The highest quality of technical performance would be obtained by the choice of recording format and by reducing the number of dubbing generations required in the distribution and broadcasting operations.

THE SCOPE OF MERPS SUBCOMMITTEE

The scope of the subcommittee's work was designed to develop the following sets of objective requirements: -

1. Functional performance requirements that describe what the machine will do for the user.
2. Mechanical performance requirements of the system that indicate how well and how quickly and reliably the system will perform.
3. Technical requirements of the electronic recording format chosen that will assure the quality of the broadcast message.

The scope included the consideration and definition of the requirements for three associated but different applications.

1. A machine system for the largely automated playback of commercial messages and promos.
2. A machine system for the playback of television news programs, providing random access to any story held in store.
3. A machine system for the largely automated playback of full-length prerecorded television programs.

While we would not expect that one machine system configuration can be developed which meets the particular requirements of all these applications, we do expect that one recording format will serve all applications, if only because all applications will benefit from the highest audio and video quality.

The specification of requirements developed for MERPS recognizes all three applications, and I would now like to discuss the particular needs of each.

OBJECTIVE REQUIREMENTS

The MERPS whose functional requirements we have defined, is commonly called a multicassette machine, but it should be noted, and the exhibits at this convention demonstrate, that there are in fact two basic design approaches to a multi-event playback system.

The first is the multicassette machine in which a large number of cassettes are held in a magazine, and a mechanism is commanded to select a particular cassette and place it in a playback transport or VCR. Such a machine has many cassettes and only one or a small number of transports depending upon the application. A machine control system orders the loading and playback operation.

The second design approach is a multi-transport machine in which no magazine is provided for the storage or selection and loading of cassettes. Rather, a large number of cassette transports are employed, and these must be loaded and unloaded manually with cassettes as the need arises. This system is integrated with, and its actual playback operation is controlled by, an automated machine control system. This design approach has of course evolved from the banks of VCR's often used in the broadcast of local news programs at many stations. It has the merit of simplicity, but may be more costly to procure than a multi-cassette machine system with fewer transports.

Commercial Messages and Promos

One application for a MERPS is commercial and promotional message playback. Here the need is greatest, and the requirements most clearly defined. Since a commercial is rarely more than two minutes in duration, a short cassette length only is required, provided that a single message is recorded on each cassette.

For economy in distribution, the cassette should be as small and light as possible, and it should be low in cost.

The signal quality should be at least equal to one-inch "C" format.

For the economical mass distribution of commercials, and above all to avoid the need of format transfer, a single industry-wide electronic recording format is desired, and would be used by advertiser, advertising agency, distributor, and broadcaster alike.

A network or independent television station may broadcast as many as 5,000 commercial messages or promos a week, and thus identification of each cassette is a vital consideration for a MERPS. One solution to this problem is to print a label and affix it to each cassette. The label would carry the title and some identifying number, together with the same information in bar code format.

Bar code readers might be provided at the MERPS machine, such that cassettes may be loaded randomly in vacant bins and the identification of each read out automatically. The location of each cassette in the magazine is stored in the machine control system.

It is envisioned that the television station's host computer which stores the day's playback schedule, will steer the machine control system and thus direct the selection, loading and playback of the required cassettes on schedule.

A special requirement of commercial playback machines is that a sequence of short duration messages may be played continuously and under computer control. This requires that more than one playback transport be provided, in order that one transport may be rewinding, ejecting and loading a new cassette ready to play immediately after the other transport has completed its playback.

However, because the cycle time for the rewinding, ejecting, selecting, threading and cueing, takes, let us say, 20 seconds, this transport will not be ready to play on schedule at the end of a 10-second message played on the first transport.

As a practical matter, a two-minute commercial break may include as shown in this diagram two 5-second "bumpers", followed by three 10-second promos, and followed by a 20-second commercial.

Now if the cycle time, that is, the total time required for the rewinding of a cassette that has played, its ejection from the transport and replacement in the magazine, the selection, loading, threading and cueing of a new cassette in the transport occupies 20 seconds, we see at once that three transports are insufficient. When the third message, C, has played on transport 3, transport 1 has not yet completed its turn-around cycle, and thus a fourth transport, transport 4, must be used to play message D. When message D is completed, transport 1 is now cued up with message E and is ready to play.

It is thus clear that the cycle time of the MERPS will dictate the number of transports required for the sequential playback of a number of short messages.

For the commercial message application, we have considered a MERPS with playback only capability, but occasions may arise when one or more of the transports should be able to record an incoming signal.

In this connection, Group W and Tape-Film Industries (TFI), a major distributor of television commercials, will launch SPOTNET, a new satellite distribution system for television commercials. In this system, master copies of commercial messages received by the distributor are played back and the signal transmitted by satellite to the client stations equipped with ground receiving stations.

Each message received thus may be directly recorded on a blank cassette loaded in the MERPS machine at the client station.

The transmission of the commercial message might be preceded by a data stream which commands the MERPS to prepare for a record session, and delivers also to the machine control system of the MERPS, the identification of the message in a form compatible with the bar code identification system referred to earlier.

News Programming

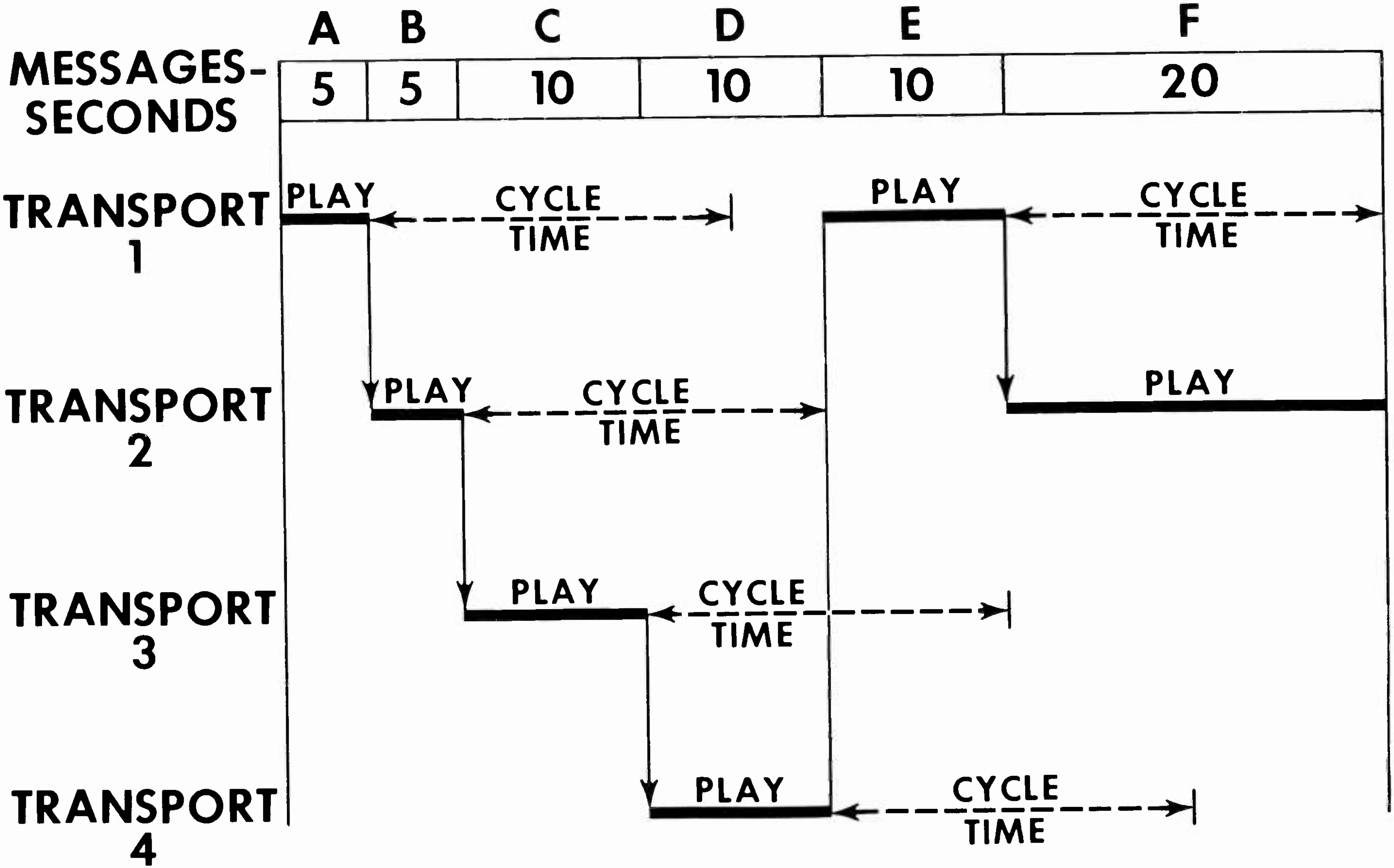
The steady increase in the volume of local news programming in the United States suggests an important application for a MERPS at many stations, and involves some special requirements.

Firstly, the cassette length for a news story is seldom more than five minutes, though a 20-minute capacity is valuable for many documentaries and other special circumstances.

While news stories are edited off-line, the completed cassette with a suitable bar code label identifying the title and number of the story may be inserted in a MERPS magazine, ready for broadcast. Thus a bar code or equivalent identification system is a requirement for this application also. Because a news program consists of a number of stories each of which is generally introduced by the anchor-person whose length of live introduction is not precisely predetermined, the actual playback command for the story must be given manually by a technical director in a control room.

It is also characteristic of a news program that the order in which stories are

MERPS-SEQUENTIAL PLAYBACK



played back may be changed up to and even during a broadcast. Thus it is a requirement that the MERPS should be provided with a video display unit on which the current playlist or lineup of stories may be displayed. Such a display might usefully include the title of each story, its identifying number and other data of importance to the TD, such as the duration of each story, and the transport on which it is being played back.

Further, it is valuable for the technical director to observe on monitors, not only the story being aired, but a preview of the next two stories scheduled to follow. It is therefore a special requirement for the news application that the MERPS should provide for three output channels to the control room, and means for the technical director to select any one channel and play it back under manual control.

Clearly then, the MERPS should have provision for the remote control of the system, starting, stopping, freezing or rejecting at will any of the cassettes playing or waiting to play in the three output channels.

Easy random access to any story at any time is an important requirement for news applications, and a keyboard associated with the playlist display, is one means of achieving this.

While three transports may be sufficient for news, if a MERPS with, say, four transports is used, a simple 4 X 3 output switcher may be required to route the output of any transport to any of the three system output channels.

Finally, it is important to have a magazine with a capacity for all the cassettes required on a news program. For a one-hour local news program this could pose a requirement for up to 60 cassettes.

Full-Length Programs

The third application foreseen for MERPS is the playback of full-length television programs. Steered by the station's host computer which stores the day's broadcast schedule, we can conceive, and indeed it is an objective to have a system for the computer-controlled playback to air of an entire day's programming on a single MERPS machine.

The requirements for this application may be rigorous, but they are clear.

Firstly, the cassette should have a minimum of 60 minutes play time.

While the cassette itself may well be larger than that used for commercials and news applications, ideally the bins or other receiving devices and the transports themselves, should be able to accept either size cassette.

The same uniform electronic recording format is of course required for this application.

Obviously the highest technical quality for program playback is required, and indeed the technical quality objectives presented in the report bespeak a performance superior to one-inch "C" format videotape. For the program application, a MERPS with only two playback transports would be sufficient, but there is no fundamental reason why a single MERPS machine cannot be used for both the playback of full-length programs and the commercials with which they are interspersed.

As with other applications, a cassette identification system associated with a bar code reader is a desirable feature.

If a MERPS is dedicated to the application of program length playback only, a magazine containing 40 cassettes would normally be adequate for a full day's programming.

CONCLUSIONS

I have described in illustrative terms the results of the MERPS Committee's work in developing user objectives for the next generation of multi-event record and playback systems.

In summary, users need a MERPS which will play back many short messages or full-length programs for broadcast or cable distribution.

Users require high technical quality of signal output employing a single industry-wide recording format.

Users require machine systems whose operation can be largely automated, or manually controlled.

With these objectives met, we expect MERPS to save time, labor and expense in the production, distribution and playback for broadcast of all prerecorded material.

Finally, I would like to express my appreciation - and I hope yours - to the members of the MERPS Committee for their enthusiasm and technical contributions; to NAB Counsel Valerie Schulte, for her rewording of the report so that it now enjoys a mantle of legal grandeur of which it was previously innocent; and to the NAB ex officio member Ralph Justus, who provided the minutes of our deliberations, and edited the present report.

To all manufacturers I would urge that you study our needs as users. Tell us what you can do and what you cannot do. Above all, and while there is yet time, I pray you find common cause with your competition to reap the benefits of a larger unified market by using a single common recording format. By all means design a better mousetrap, but use a common cheese.

THE RCA CCD-I BROADCAST COLOR CAMERA

Dennis J. Woywood

Vice President, Marketing, Broadcast Systems Division

RCA Corporation

INTRODUCTION

Television cameras have been improving steadily over the last 20 years. These improvements have taken the form of increased sensitivity, better signal-to-noise ratio, improved registration, greater ease in use and improved optics.

Now an advance in solid state technology has provided the basis for still further improvements in camera performance. The camera described in this paper provides a substantial additional improvement in sensitivity and signal-to-noise ratio and, for the first time in modern cameras, total relief from the disturbing effects due to lag. This camera will deliver a full-level signal with only 20 footlamberts of available light to the camera in normal gain setting, and 2.5 footlamberts in the high sensitivity mode. Thus we achieve an improvement of close to 5 dB in effective sensitivity and 6dB in signal-to-noise ratio compared to tube pickup devices.

Improvements due to the absence of lag, image burn and highlight smear are dramatic. In some ways such as the total elimination of the tails behind headlights, this is obvious. In addition, it has the attribute of providing significant resolution in motion. When a sensor exhibits lag, then all images have smear when movement is present; this immediately wipes out detail in the image.

in the presence of substantial motion, lag produces broad-based blurring. For example, when looking at a moving car, people inside cannot be seen because both they and the pillars of the car windows blur together. In the CCD-I camera all lag related effects are eliminated.

THE NEW CCD

The sensor used in this camera, is a new frame-transfer-type CCD image sensor. It is a fully manufacturable, high sensitivity device developed for high performance television applications. A novel feature of these CCD imagers is a thinned configuration. This allows illumination from the back side of the device and provides quantum efficiency in excess of 60% over the entire visible spectrum for the whole imager area. Noise in the CCD read-out is less than 35 electrons rms per pixel. The high quantum efficiency, together with the very low noise output permits outstanding low-light-level camera performance.

The CCD sensor in this camera is a three-phase 403 (H) x 512 (V) pixel frame-transfer imager, having a chip area of 319 x 416 mil^2 and 16μm (H) x 10μm (V) pixels. Photo generated electrons are integrated in the illuminated A-register as shown in Figure 1. During the vertical interval the entire charge field is transferred to the B-register. One line at a time is then transferred to the C-register during the horizontal interval; this line is then clocked out as the video part of the TV line.

The imager is constructed with three levels of polysilicon, buried channel registers, anti-blooming drains in the photosensitive area, and operates with 2:1 vertical interlace. The device is thinned to about 10μm total thickness to allow back side illumination, thus providing contiguous pixels with 100% utilization of the incident light.

The resulting high quantum efficiency achieved is illustrated by the typical behavior shown in Figure 2, in which a Q.E. exceeding 60% over the entire visible spectrum is demonstrated. Each device in a color camera is selected for optimum performance in its portion of the spectrum.

Excellent blooming control is provided by the anti-blooming drains. Overloads greater than 10,000X can be handled, and transfer smear can be totally eliminated by the use of a shutter. The complete lack of lag results in dynamic resolution that is greater than tubes.

This excellent performance is achieved through the combination of low noise read-out of the CCD, high quantum efficiency of the thinned devices, low dark current, and excellent uniformity of dark current and photo response.

THE COLOR SEPARATION ASSEMBLY

The optics in this camera uses a color separation prism similar to assemblies suitable for use with one-half inch pick-up tube devices. Three CCD's are positioned facing the exit ports of the prism. The precision of the position must be such that the electronic signals correspond to proper registration of the three separated colors.

When using tube-type sensors, the raster position, size and geometry are individually adjustable. Thus, while basic mechanical alignment is important, there is some room for electronic adjustment of registration

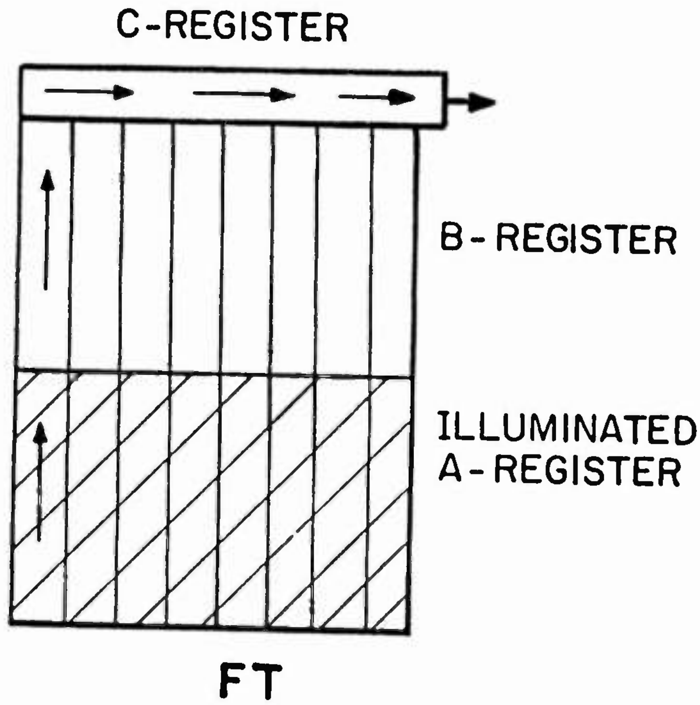
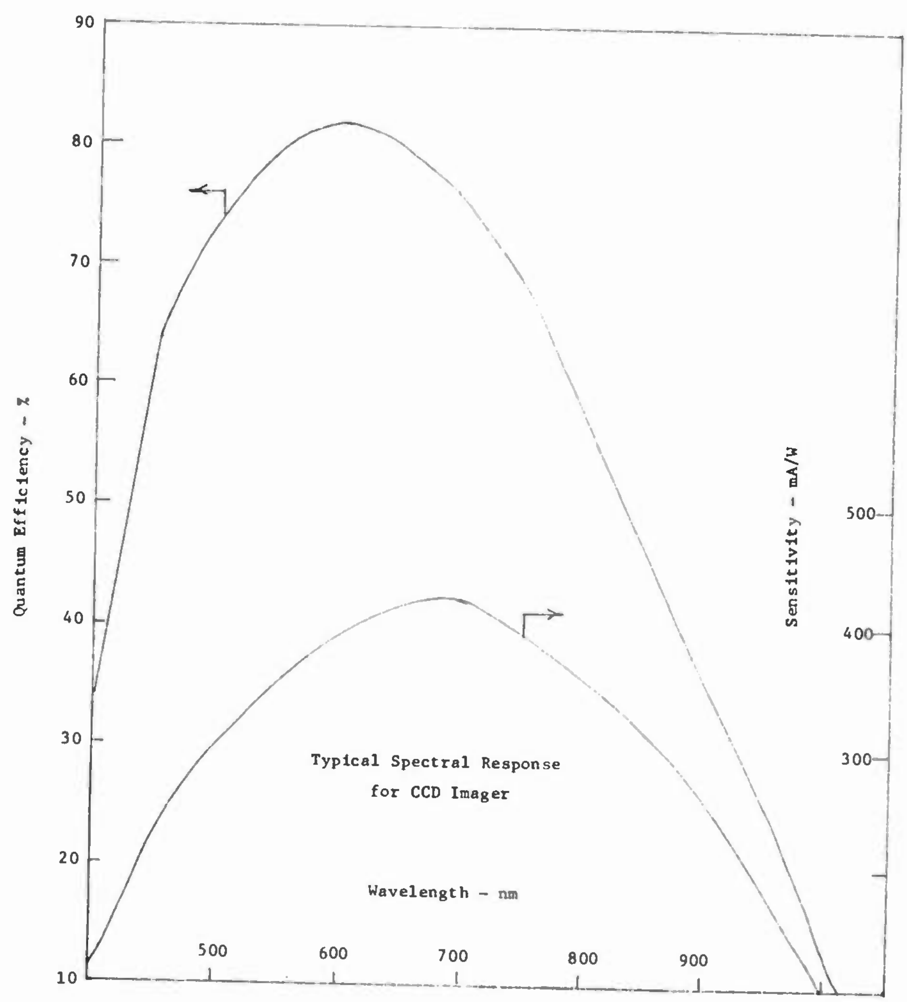

C-REGISTER
B-REGISTER
ILLUMINATED
A-REGISTER
FT

Figure 1


Quantum Efficiency - %
Sensitivity - mA/W
Typical Spectral Response
for CCD Imager
Wavelength - nm
10
20
30
40
50
60
70
80
90
500
600
700
800
900
300
400
500

after assembly. The CCD sensor, however, has a fixed raster. The patterns used to define the CCD elements in the sensor also define the position of the raster. So, except for small timing variations, there is very little one can do electronically to correct misregistration. All registration must be set mechanically during assembly. In this camera, basic alignment is done once and is permanent.

The assembly process is performed with manipulators that can resolve a movement of less than one micron and are stable to better than that for periods of days. Once the manipulators have placed the CCD's in their correct position, highly stable cements are used to permanently secure them. The manipulators are then removed as no further adjustments are necessary or possible.

Measurement techniques were developed to provide both alignment and validation of position and registration. Building on the knowledge gained in developing the Tk-47 automatic registration system, special circuits were combined with standard instrumentation to permit resolving registration errors as small as 0.01 micron. These tools were used during the development of the camera to determine small degrees of relative movement during various types of environmental stresses. Movements with time, temperature, vibration and shock were evaluated.

As a consequence of the precision adjustment techniques, the stability of the adhesives used, and the ability to measure with high precision, this camera has achieved intercolor registration accuracy of better than 0.05 per cent at the center of the raster. This value is stable over the entire temperature range of operation and for the life of the camera.

Chromatic aberration caused by the external lens system results in image sizes being different for the three colors. The inability to adjust the raster size means that there will be registration errors at the edges of the picture due to the different image sizes. These errors are less than 0.05 percent of the vertical height of the image; however, the cumulative error can result in errors at the sides of the picture being as great as 0.10 per cent.

SYSTEMS CONSIDERATIONS

There are three areas of camera system change related to the use of CCDs:

- Device drive circuits and imager set-up requirements replace the deflection drivers, registration correction circuits and tube control circuits.

- Video signal recovery from the transducer is completely different from the familiar transresistance preamplifier.

- Pickup tube power requirements (high voltage filament current and precision focus coil current) are absent which eases the power supply design.

The last item will not be further discussed.

The CCD Imager Drive Electronics

As described earlier, the CCD imagers are divided into two field store sections, the A and B registers, and a line structured output section, the C register. The picture information integrated in the A register during the field is transferred into the B register during the vertical blanking interval, and, line-by-line, from the B to the C register during the horizontal blanking interval.

The timing of the A and B register transfer clocks is controlled by a custom LSI circuit. This device contains all of the counters required for the field transfer. In addition, it contains two externally tunable oscillators, which supply the clocks for the A to B transfer counter function and the B to C transfer functions. The output from this LSI provides the timing relationships for these transfers and are buffered by hybrid amplifiers capable of driving the three imagers in parallel. The chip is synchronized to the camera system by the application of horizontal and vertical drive pulses, as well as system sync and blanking.

The line output, or C register, timing is controlled by a separate crystal oscillator, which is phase-locked to the camera's horizontal drive. It produces the three-phase C register transfer-gate drives by division in a high speed shift register. Each imager is driven separately by a hybrid amplifier similar to the one used to drive the A and B gates. All 403 illuminated pixels are clocked out during the unblanked portion of the TV line.

Picture Signal Recovery

The signal output from the CCD imager is a series of amplitude-modulated pulses, supplied by a MOS source follower. This signal contains information at the base band frequencies, as well as on modulated carriers which are the harmonics of the clock drivers.

The baseband signal is not used since it contains the 1/f noise contribution of the MOS source followers which provide the CCD outputs. A synchronous detector provides the lowest noise video signal. The three phase transfer clock circuitry supplies the pulse train for use as a carrier. This additional pulse train must be time referenced to the transfer clocks to assure reliable signal recovery. The syncronously detected baseband video is filtered to remove any residual clock signal.

The remaining processing step is the detection and cancellation of the effects of the finite dark current inherent in the silicon device. The two major components of this current are:

- a black shift, or bias, caused by dark current integrated in the A register at the time the picture information is being stored;

- a second sloped black shift added as the signal is transferred through the B register during the readout process.

These two are affected by both the original CCD processing, and the temperature of the device in use.

After the picture information is read out, another component of dark current is accumulated in the B register. It is this last component, which is directly related to the first two, which is measured and used for all required correction.

CAMERA PERFORMANCE

Only key performance features of the CCD-I Broadcast Camera will be listed.

1. Registration of the three color images is set to the order of 0.02 percent of picture height at the center. Elsewhere in the picture registration is limited by the chromatic aberration in the lens.

2. This registration is set at manufacture and is permanent. Thus problems are avoided from inadvertent misadjustment.

3. Shock and vibration tests have shown the optical structure to be strong and stable. No changes in registration were caused by 10G forces. Tests of the physical structure showed no damage at levels exceeding 50G.

4. Registration loss in motion is caused solely by field rate considerations: image integration during each field will cause blurring during motion just as in a film camera. However, there is no blurring from tube lag since this effect is not present.

In the CCD camera signal-to-noise and sensitivity require new ways to be described. In tube cameras, performance was limited by scene-dependant low-light-level effects. The lack of these effects in the CCD provides the ability to trade signal-to-noise and sensitivity to better advantage. The best approach to this has not yet been determined.

However, under comparable conditions, whatever the choices made, the CCD-I provides a 15dB improvement over current tube cameras.

WHAT LIES AHEAD?

The RCA CCD-I Color Camera provides greatly improved sensitivity, elimination of lag and, for the first time, dynamic resolution matching that of motion picture film cameras.

Work is continuing to develop still better sensors for the television camera market. While much may be off in the distance, we can look forward to sensors with the same sensitivity as the present CCDs, but with greatly improved horizontal and vertical resolution.

As skill and knowledge improve, we envision devices with up to two or more times the present number of horizontal elements. A similar improvement in the vertical direction will occur. The major factor in developing such improved devices relates to device manufacturing yield. This is no small task, but is being diligently addressed.

CCD's are inherently smaller and lighter than the tube-type sensors that they replace. Therefore we envision cameras that will be much lighter and smaller, aimed at taking full advantage of the inherent capabilities in this sphere.

We see RCA's new CCD-I Broadcast Color Camera as a major step in color television camera technology.

ACKNOWEDGEMENTS

The CCD-I camera was a cooperative effort between the Broadcast System Division, New Products Division and the David Sarnoff Research Center. Many people contributed to the success of this development. The author thanks these engineers and scientists; they will each report the details of their work in appropriate forums.

Reduced Bandwidth Requirements for Compatible High Definition Television Transmission

W.E. Glenn, K.G. Glenn, J. Marcinka, R. Dhein, I.C. Abrahams

New York Institute of Technology, Science & Technology Research Center

Dania, Florida

The objective of HDTV has been to produce images of significantly higher quality than standard 525 line transmissions. It has been assumed that a quality equal or better than 35mm projected film should be a reasonable target. L. L. Pourciau[1] presented a comparison of the static resolution of 35mm film to what could be achieved with television. According to his analysis, a scanning system of about 1000 lines resolution with equivalent horizontal resolution and with proper compensation should match 35mm film performance very closely. In our paper last year[2] we presented the relative bandwidths of luminance, R-Y and B-Y. Figure 1 shows the measured relative response of the visual system to luminance and isoluminance chromaticity gratings over the range of spatial frequencies perceived when viewing television. The spatial frequency dimension corresponds directly to resolution. The high frequency cut-off of R-Y is about a factor of two lower than Y, and B-Y about a factor of four lower. These values suggest that color information have a much higher resolution relative to luminance information than was used in the NTSC system.

If a system were to be designed based on these values using a 3 to 5 aspect ratio, the bandwidth requirements would be:

Y	-	21.00 MHz
R-Y	-	5.25 MHz
B-Y	-	1.31 MHz
Total:		27.56 MHz

This is very close to the system design recommended by NHK.[3]

Let us first consider how we might reduce bandwidth by taking advantage of the "oblique effect" in vision. Typically, the human visual system has poorer resolution on the diagonal than it does in the cardinal (horizontal and vertical) directions. This phenomenon has been often reported in the vision literature[4,5] and we have also made an independent measurement of it. Figure 2 shows the

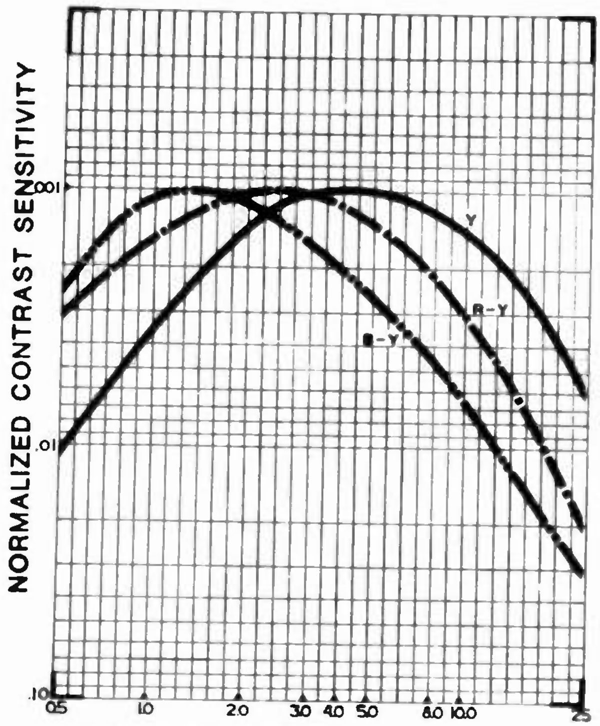


FIG.1 STEADY STATE CONTRAST SENSITIVITY

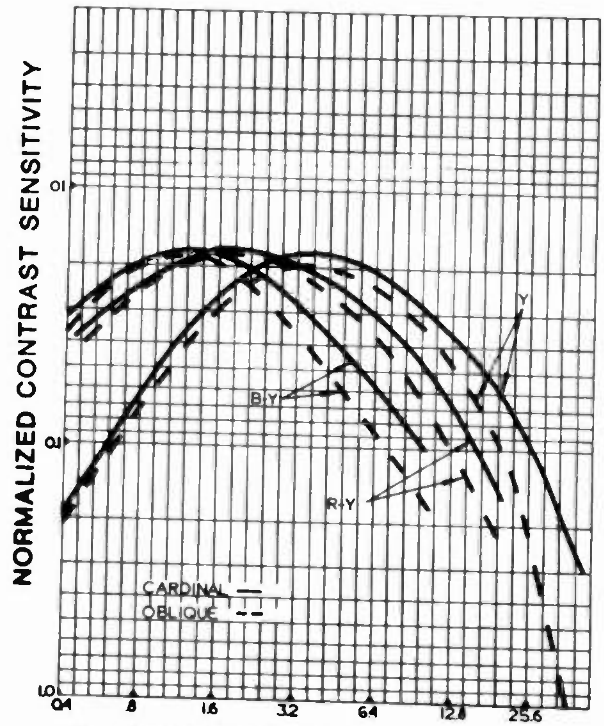


FIG.2 CARDINAL AND OBLIQUE CONTRAST SENSITIVITY

average MTF of eight subjects for Y, R-Y and B-Y gratings oriented in oblique and cardinal directions. As you can see, at high spatial frequencies the visual system has lower resolution in the oblique orientation for all three signals by about $\sqrt{2}$. If we plot a polar distribution of limiting resolution response at 30% of the peak sensitivity of the visual system, we get a plot as shown in Figure 3A.

Also shown in this figure is the distribution for film (3B) and for the HDTV system described above (3C). Obviously, the standard sampling method for television does not match the distribution of the visual system. The techniques described by Wendland[6] using diagonal sampling, oversampling the camera and display, and pre and post-filtering, will produce a distribution as shown in Figure 3D with half the bandwidth. Diagonal sampling has also been used by NHK[7] in their recently proposed system. The slight loss in perceived resolution on the diagonal is compensated for by the slightly better "Kell factor" obtained by better processing of the image. If we employ these techniques for Y, R-Y and B-Y, we now

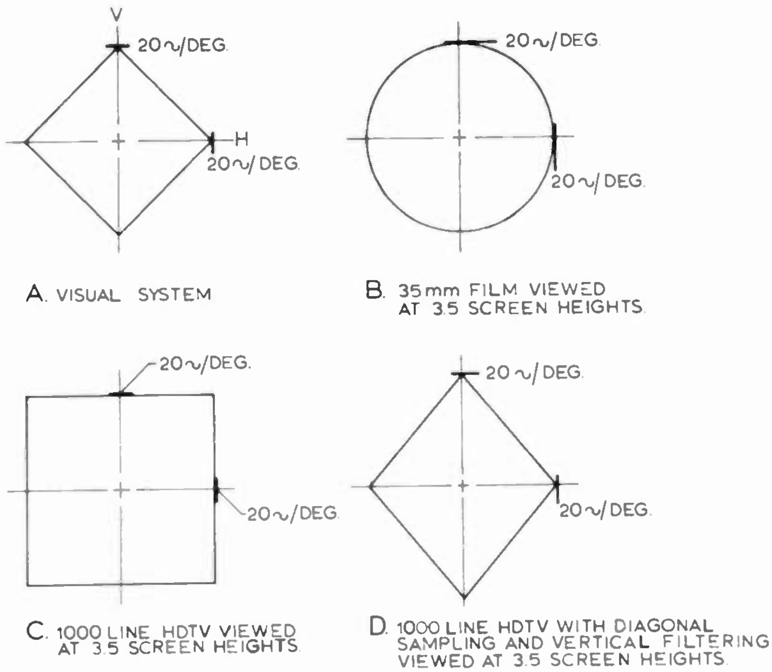


FIG. 3 POLAR SPATIAL FREQUENCY PLOT AT 30% OF MAX. MTF IN CYCLES/DEGREE (~/DEG.)

have a static image of very close to the same perceived resolution with the following bandwidths:

Y	-	10.50 MHz
R-Y	-	2.63 MHz
B-Y	-	.66 MHz
Total:		13.79 MHz

Next we must consider the requirements for depicting moving images. In our previous paper[2] we presented information regarding the dynamic properties of the human visual system. Basically, this system is divided into two parallel channels. One "channel," termed the "transient neuron" system, has poor spatial resolution and fast temporal resolution. This channel is most effective for detecting motion. The other "channel," the "sustained neuron" system, has high spatial resolution and poor temporal resolution. This channel functions primarily as a high resolution pattern detector. Figure 4 shows a plot of the approximate spatial and temporal resolutions of these two pathways.

In addition to the poor temporal response of the high resolution system, it is also masked by transients caused by motion for a period of about 1/3 sec. To quantify the response sensitivity to transients and the amount of masking, we measured these effects using luminance and isoluminance chromaticity gratings.

Figure 5 shows the response of the visual system to Y, R-Y and B-Y for gratings of three spatial frequencies as a function of the duration of the presentation.

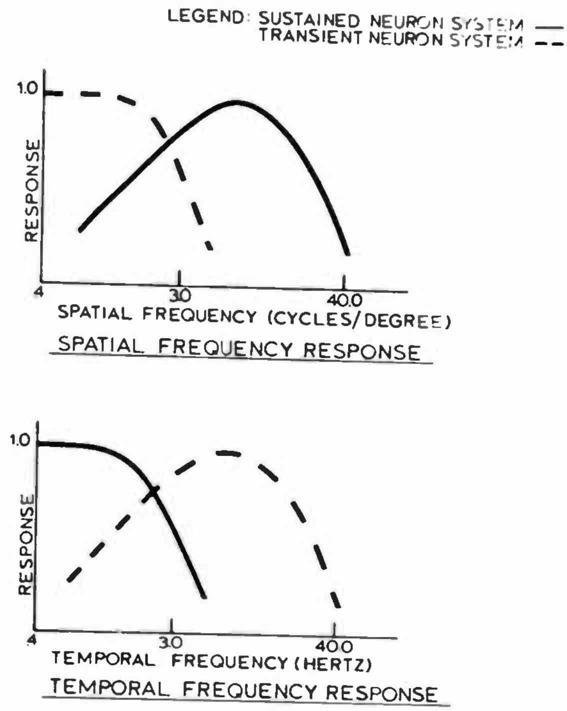


FIG. 4 SPATIAL AND TEMPORAL RESPONSE OF TRANSIENT SUSTAINED NEURON SYSTEMS.

Figure 6 shows the perception of these gratings when they are presented adjacent in time to a masking signal composed of .1 sec. off-the-air television. (The details of this experiment are described in Reference 2.) We may consider these short duration effects demonstrated in the laboratory as analogous to transient scene changes produced by motion. From these data we infer that when a scene changes rapidly, the visual system has an enhanced sensitivity to spatial frequencies below about three cycles per degree and a surpressed sensitivity for spatial frequencies higher than this. It is, therefore, very important to protect the dynamic resolution of spatial frequencies below 3 cycles per degree. This is the domain of the "transient" neuron system which is especially adapted for motion detection. The dynamic resolution above this spatial frequency can degrade with motion, without being perceived.

As a result of this division of performance of the visual system, the "detail" (high spatial frequency) information in both luminance and color can be updated at much lower frame rates than the low resolution information. Last year at the NAB Convention we demonstrated a simulation of a transmission system in which the low resolution information was transmitted at 30 frames per second, and the detail was transmitted at 5 frames per second. Images transmitted in this way and reconstructed using a frame store in the receiver depict motion and detail quite faithfully.

A conventional interlace system does not electronically separate spatial or temporal frequency componants. However, the process does inherently display low

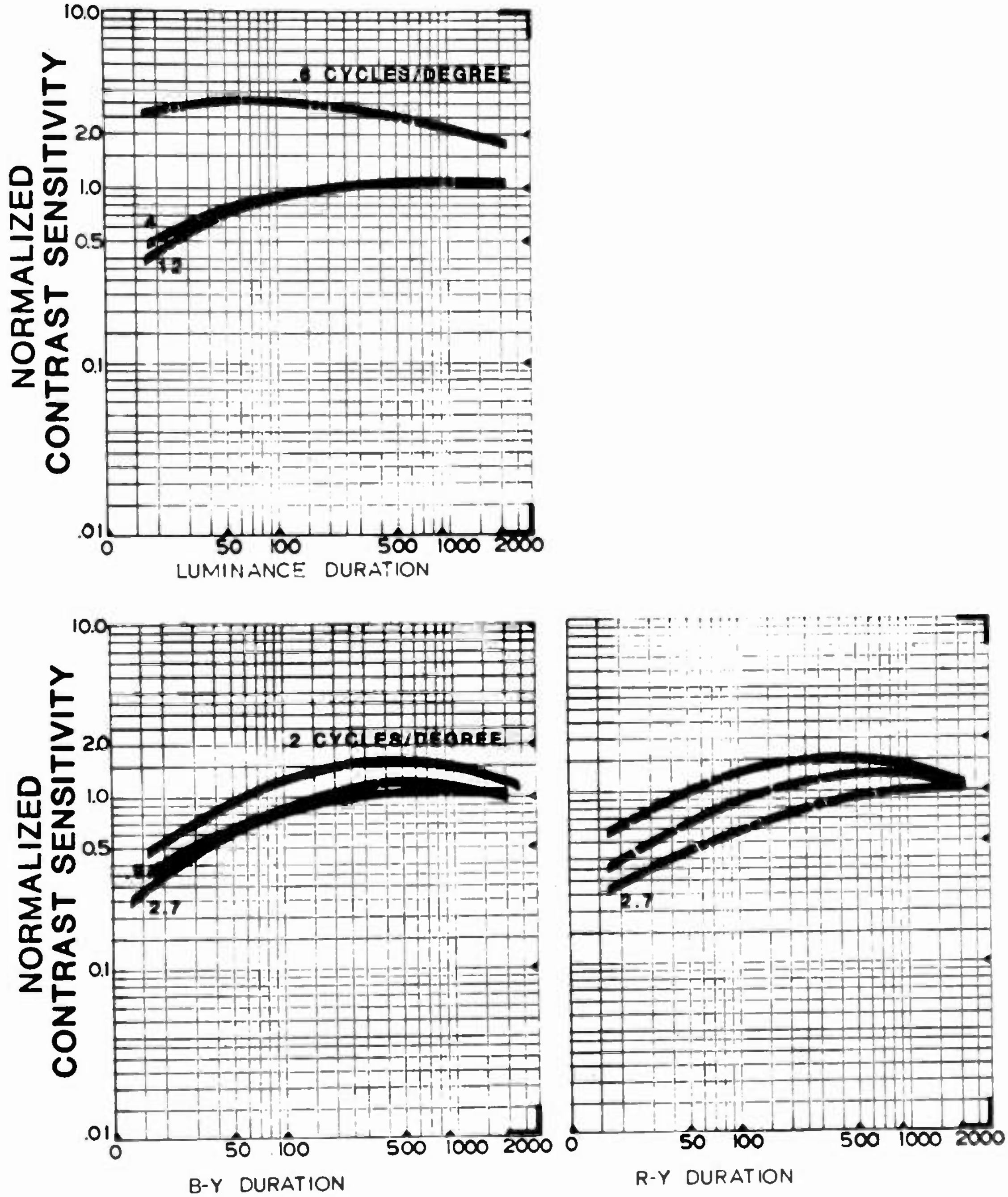


FIG. 5 CONTRAST SENSITIVITY AS A FUNCTION OF DURATION (MILLISECONDS)

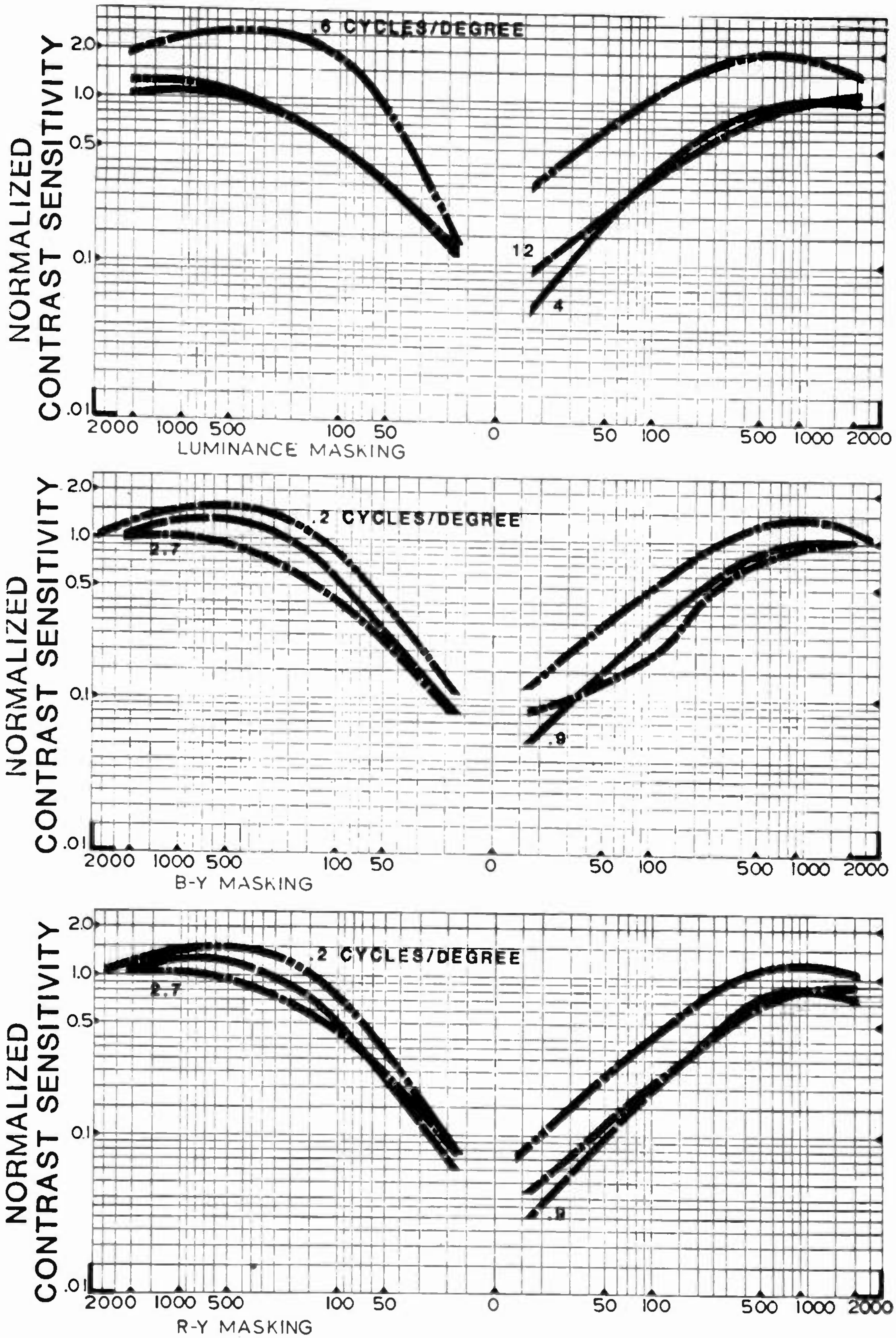


FIG. 6 CONTRAST SENSITIVITY WITH MASKING BY A .1 SECOND MASK

spatial frequencies at a higher frame rate. Since a field has half as many scan lines as a frame, the limiting vertical resolution for 1/60 second (one field) is, therefore, half that of the vertical resolution for 1/30 second (one frame). The 1/60 second field rate produces good dynamic resolution at low spatial frequencies, and thus depicts motion much more precisely than the 1/24 second frame rate used in motion picture film. The more "jerky" motion of film is primarily due to the effect of this lower frame rate which is longer than the time constant needed for perception of low spatial frequencies by the transient neuron (motion detecting) system. With interlacing, the detail information does not appear until after 1/30 of a second, which is about a factor of ten shorter duration than the time needed for perception by the sustained neuron (pattern detecting) system.

This procedure of sending low resolution information rapidly and detail information slowly can be used to salvage the portion of the present video spectrum above 2.5 megahertz where color and luminance information cannot be separated adequately by processing. To use this part of the spectrum, we can first make "frame combing" really work. In the normal NTSC transmission we can add together the R-Y, B-Y and Y signals above 2.5 megahertz for two successive frames. This information is simply repeated for two successive frames with the color carrier reversed in phase as it normally does. A frame store can then be used to add these same two frames or subtract them. The sum gives us the Y signal above 2.5 megahertz without any color carrier. (The color signal was made identical for those two frames but the carrier is reversed in phase.) The difference of the two frames gives us color without cross color. Since the luminance is identical for the two frames, the difference is zero in this frequency range. In both the standard receiver and enhanced receiver (with frame comb), and the HDTV receiver, the frame rate for color and for detail above 2.5 MHz would be 15 frames per second. From the above psychophysical data and our simulation with detail at 5 frames per second, this should be adequate.

The above modification to the standard NTSC transmission would allow us to transmit Y up to 4.2 megahertz and I and Q up to their normal bandwidths without cross-color or cross-luminance. The remaining luminance and chrominance detail information can be transmitted at an even lower frame rate. We could use 7.5 frames per second and diagonal sampling (to take advantage of the oblique effect). The 4.2 megahertz of the almost standard NTSC transmission carries the rapidly changing low resolution information. Another low bandwidth channel is used to carry the remaining detail and color information at 7.5 frames per second (probably in analog component format). Figure 7 shows the video spectrum of the proposed system. The normal channel transmits luminance up to 2.5 megahertz unchanged at 30 frames per second interlaced. From 2.5 megahertz to 4.2 megahertz the detail luminance and chrominance are repeated to give 15 frame per second presentation and are separated with a frame store. Another frequency spectrum is used to transmit the higher detail luminance and detail chrominance at 7.5 frames per second. This additional channel needs the following bandwidth to transmit these detail signals:

Y	-	1.30 MHz
R-Y	-	.60 MHz
B-Y	-	.15 MHz
Total:		2.05 MHz

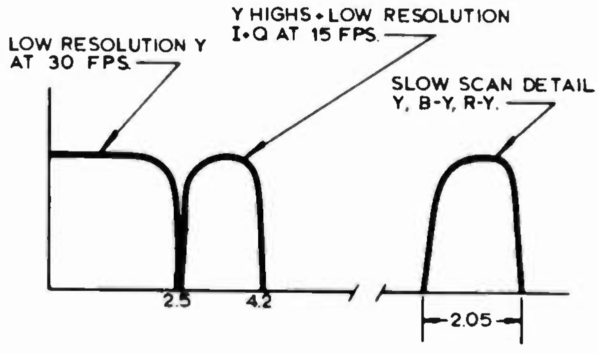


FIG. 7 VIDEO SPECTRUM

In this presentation we have shown a system concept that is designed to take advantage of several known limitations of the visual system. This should make it possible to transmit HDTV compatibly with the addition of approximately two megahertz to one standard bandwidth NTSC channel, and provide an image quality very close to that of 35mm films.

REFERENCES

[1]Pourciau, L. L. "High Resolution TV for Motion Picture Production." Paper presented at SMPTE Conference, Los Angeles, California. October, 1983.

[2]Glenn, W.E., Glenn, K., Dhein, R.L., Abrahams, I.C. "Compatible Transmission of High Definition Television Using Bandwidth Reduction." Proceedings of 37th Annual Broadcast Engineering Conference, National Association of Broadcasters, Las Vegas, Nevada. 1983, p 341-349.

[3]Fujio, Takashi, Ishida, J., Komoto, T., Nishizawa, T. "High Definition Television System - Signal Standard and Transmission", NHK Laboratories Note #239. August, 1979.

[4]Appelle, S. "Perception and Discrimination as a Function of Stimulus Orientation: The Oblique Effect in Man and Animals", Psychol. Bull., 78, 266-278, 1972.

[5]Timney, B.N., Muir, D.W. "Orientation Anisotropy, Incidence and Magnitude in Caucasian and Chinese Subjects", Science, 193, 699-701, 1976.

[6]Wendland, B. "Extended Definition Television with High Picture Quality", SMPTE Journal, 92, 1028-1035. October, 1983.

[7]Sugimoto, Maseo. "A Single Channel HDTV Broadcast System." NHK Technical Report presented to HDTV Subcommittee, SMPTE Conference, Montreal, Quebec, February, 1984.

A COMPATIBLE HDTV BROADCAST SYSTEM

John P. Rossi, A. A. Goldberg, R. H. McMann

CBS Technology Center
Stamford, Connecticut

A television system for broadcasting HDTV on two channels of a direct broadcast satellite is described. Channel 1 carries a compatible 525-line picture. Channel 2 carries an augmentation signal, that in conjunction with the signal from Channel 1, produces a 1050-line HDTV picture with a 5:3 aspect ratio. Four high quality sound and/or data channels are provided via digital modulation. The HDTV receiver does not require field or frame memories that will be expensive during the near future.

The suggested HDTV system is for broadcast in Region 2 and evolves naturally from existing 525-line television standards. It should not be confused with a still to be defined HDTV production standard, which is hoped to be used throughout the world. Major programs will be produced in the HDTV production format and converted to the proposed 1050-line HDTV standard as well as to the 525-line standard prior to broadcasting.

1. INTRODUCTION

CBS has urged that the 12.2-12.7 GHz DBS spectrum in Region 2 should be used to provide new and innovative services such as high definition television (HDTV). Technical developments provide confidence that commercial implementation of an HDTV service within this decade is a reasonably realizable goal. The 525-line NTSC service broadcast from Terrestrial transmitters and effusively via cable systems is serving the American public well. However, with the continuing trend towards larger screens, the many weaknesses of the 525-line NTSC signal are becoming obvious, in spite of significant improvements that have been made in the NTSC picture quality. HDTV broadcasts must serve the large population of 525-line television viewers as well as the new generation of HDTV viewers, which implies a need for a dual standard system.

CBS proposes to broadcast compatible 525-line pictures with an augmentation signal on a second channel to produce 1050-line, wide screen HDTV pictures, i.e., a dual standard system. Two 24 MHz wide channels, not necessarily contiguous, will be used. Channel 1 will carry 525-line, 60 field, 4:3 aspect ratio, time multiplexed component (TMC) color TV video plus three or more audio signals. Channel 2 will carry an additional 525-lines of video but with a 5:3 aspect ratio, that when matrixed with the signal from Channel 1 will constitute an HDTV picture. As shown in Figure 1, an HDTV receiver will tune to channels 1 and 2, whereas a 525-line receiver can receive a standard quality picture from Channel 1.

The term compatibility is somewhat ambiguous because DBS can never be fully compatible with existing broadcasting means. A simple definition of questionable utility is that the degree of compatibility varies inversely with the cost of conversion circuits. DBS will require a 12 GHz FM receiver. With TMC transmission, the receiver will also employ video processing circuits to reconstruct the 525-line signals. We think that for DBS applications TMC has the potential for producing superior picture and sound quality. Therefore, the Channel 1 will be compatible with TMC broadcasts.

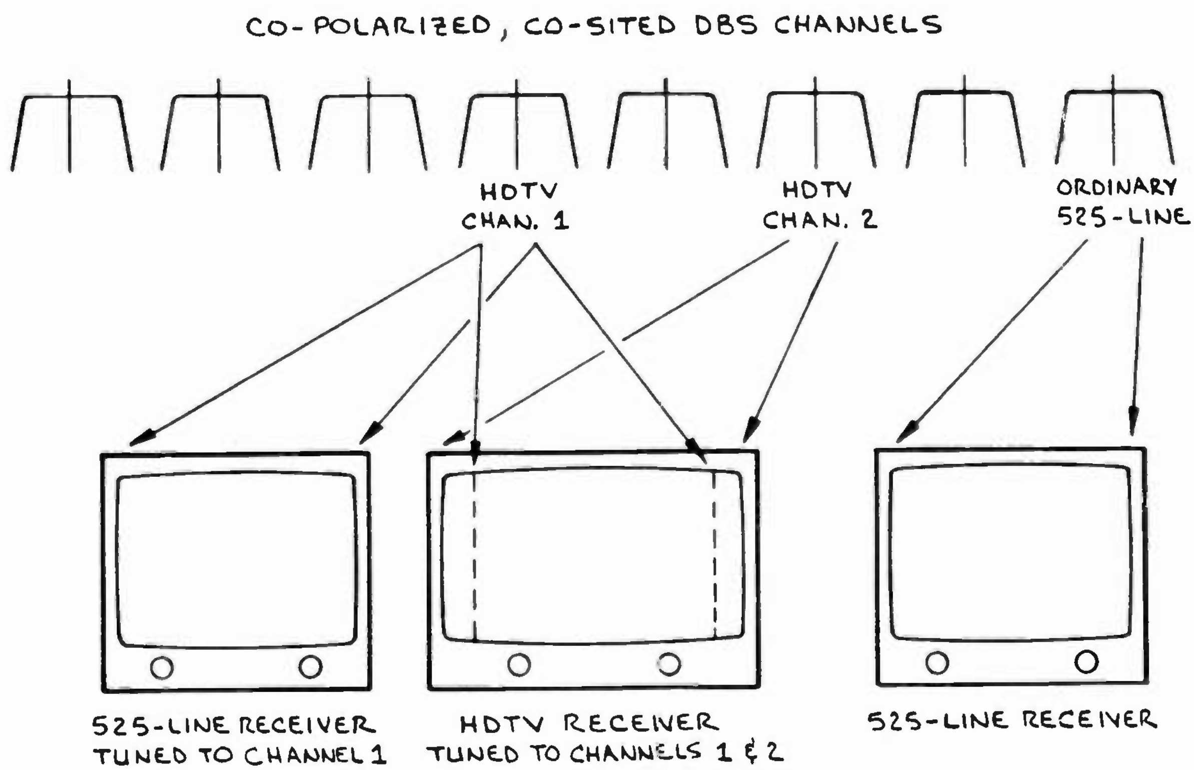


FIGURE 1. DBS CHANNEL USAGE

Various TMC formats are being examined with an intent to optimize the 525-line and 1050-line systems. Tentative specifications for Channel 1 are as follows:

o Color information will be sent via time compressed line-sequential R-Y, B-Y signals.

- o Time compression of the color-difference signals will be three times greater than that of the luminance time compression.
- o Digital audio will be time-multiplexed with the video.
- o Scrambling methods are still under study and will not be addressed in this paper.

The HDTV system to be described here is for broadcast in Region 2 and evolves naturally from existing 525-line television standards. It should not be confused with a still to be determined HDTV production standard, which it is hoped will be used throughout the world. Major programs will be produced in the HDTV production format and converted to the proposed 1050-line HDTV standard as well as the 525-line standard prior to broadcasting.

HDTV requires high horizontal and vertical resolutions, a wide image aspect ratio, and high fidelity stereophonic sound. A primary design objective is an optimization of these factors within the constraints of channel bandwidth and noise, compatibility, receiver cost, and psychovisual needs.

Superior vertical resolution calls for more than the 525 scan lines presently in use. In the interest of compatibility, we have elected to use twice 525, or 1050 lines for the HDTV picture. Increased horizontal resolution is obtained with higher video signal bandwidths.

Dual-aspect ratio pictures is one of the more interesting features of this compatible 525, 1050-line television system. At the input to the DBS uplink there will exist a 1050-line, 5:3 aspect ratio TMC video signal. Its source may be a production standard HDTV camera and a standards converter. Alternatively, a telecine chain can directly generate the 1050-line HDTV signal. There is, of course, also the possibility of employing 1050-line live cameras for local originations.

Sending a 5:3 image in Channel 1 was considered and rejected on the basis of compatibility. Those using 525-line, 4:3 receivers would truncate 20% of the picture edge information thus lowering the transmission efficiency. Alternatively, a 4:3 kinescope would display dark bands top and bottom. Therefore, only the center 4:3 area of the picture is transmitted with 525 lines through Channel 1.

The balance of the picture consists of another 525 lines, 5:3 aspect ratio, which is transmitted via Channel 2. A two-channel HDTV receiver reconstitutes the 5:3 picture by combining the two video signals into a 1050-line, 2:1 interlace, 60-field raster. This concept is illustrated in Figure 2.

High fidelity sound is provided by digital audio signals transmitted in time sequence with the video. The number of sound channels may be as high as four.

2. IMAGE SCANNING

System M/NTSC television employs 525 lines, 60 fields, 2 to 1 interlace scanning standards. The aspect ratio of the active picture area is four units wide to three units high (4:3). Channel 1 broadcasts, therefore, must conform to these standards if they are to be compatible with other 525-line broadcasts.

Channel 2 augments Channel 1 signals in order to produce HDTV pictures and

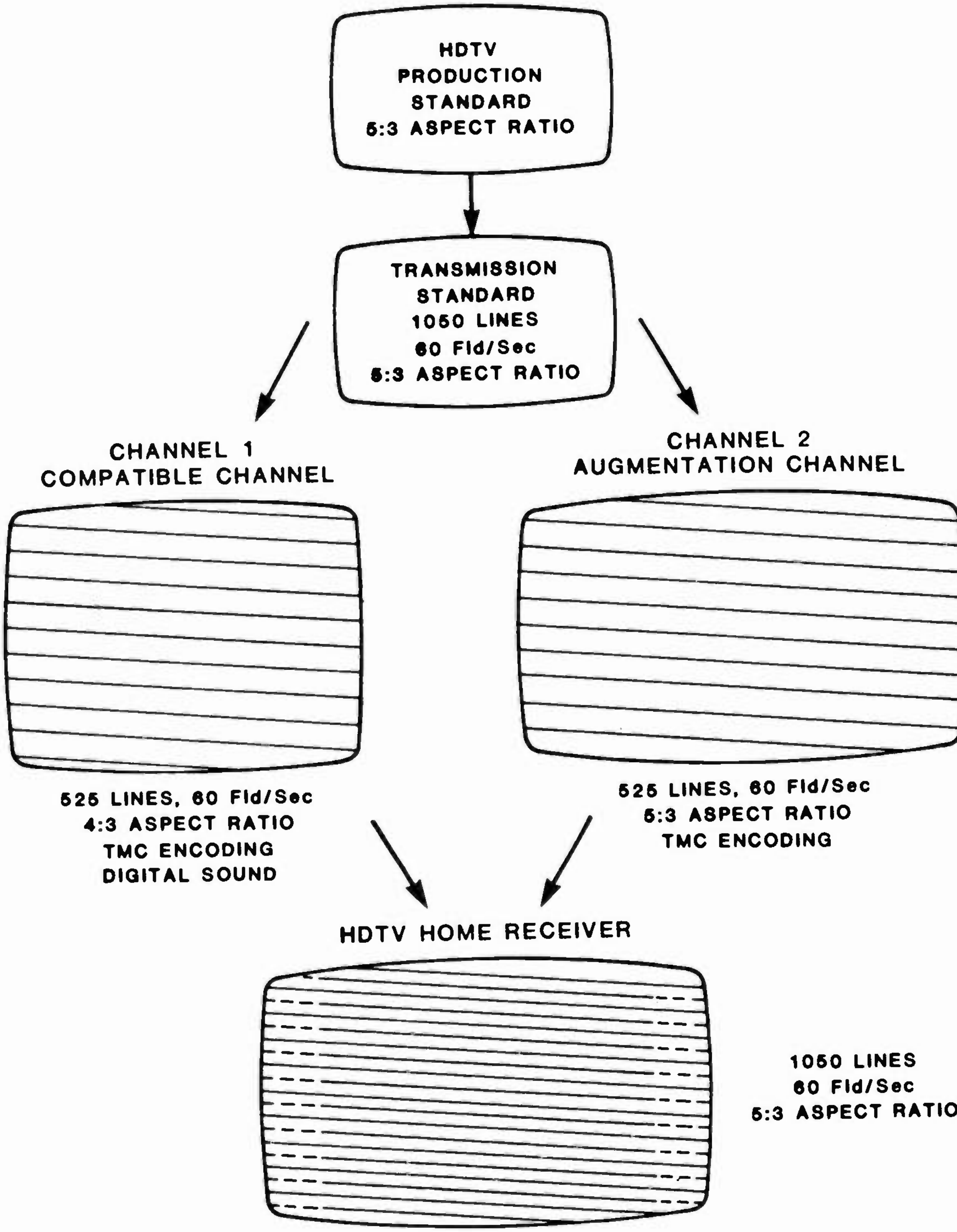


FIGURE 2. DBS HDTV 2-CHANNEL SYSTEM

intuitively should have an integer scanning relationship to 525 lines. Twice 525 is 1050 lines, which can provide the desired vertical resolution for HDTV. Fujio of NHK has reported that the optimum viewing distance for 940 lines, 2:1 interlace, 5:3 aspect ratio pictures is four times picture height (4H) and for 1125 lines is 3.3H. By interpolation, 1050 lines may be viewed at 3.6H.

Figure 3 shows how the 1050-line HDTV scanned luminance image can be carried in Channels 1 and 2. For simplicity, assume a TV camera is scanned in the 1050-line format, one field of which is shown in Figure 3a. The odd-numbered lines are labeled A and the even-numbered lines B. Note the horizontal bounds of the active picture are designated for 5:3 and 4:3 aspect ratios. Channel 1 carries 262½ lines per field, each line being the summation of lines A and B, in the 4:3 area only, as shown in Figure 3b. Channel 2 carries only line B of the 5:3 area and A+B of the two side portions of the 5:3 area, again 262½ lines per field, as shown in Figure 3c. The reconstructed HDTV field in the receiver is shown in Figure 3d. Odd-numbered lines A in the 4:3 area consist of A+B of Channel 1 less B from Channel 2. Even-numbered lines B are lines B from Channel 2. The side portions of lines A are obtained from Channel 2.

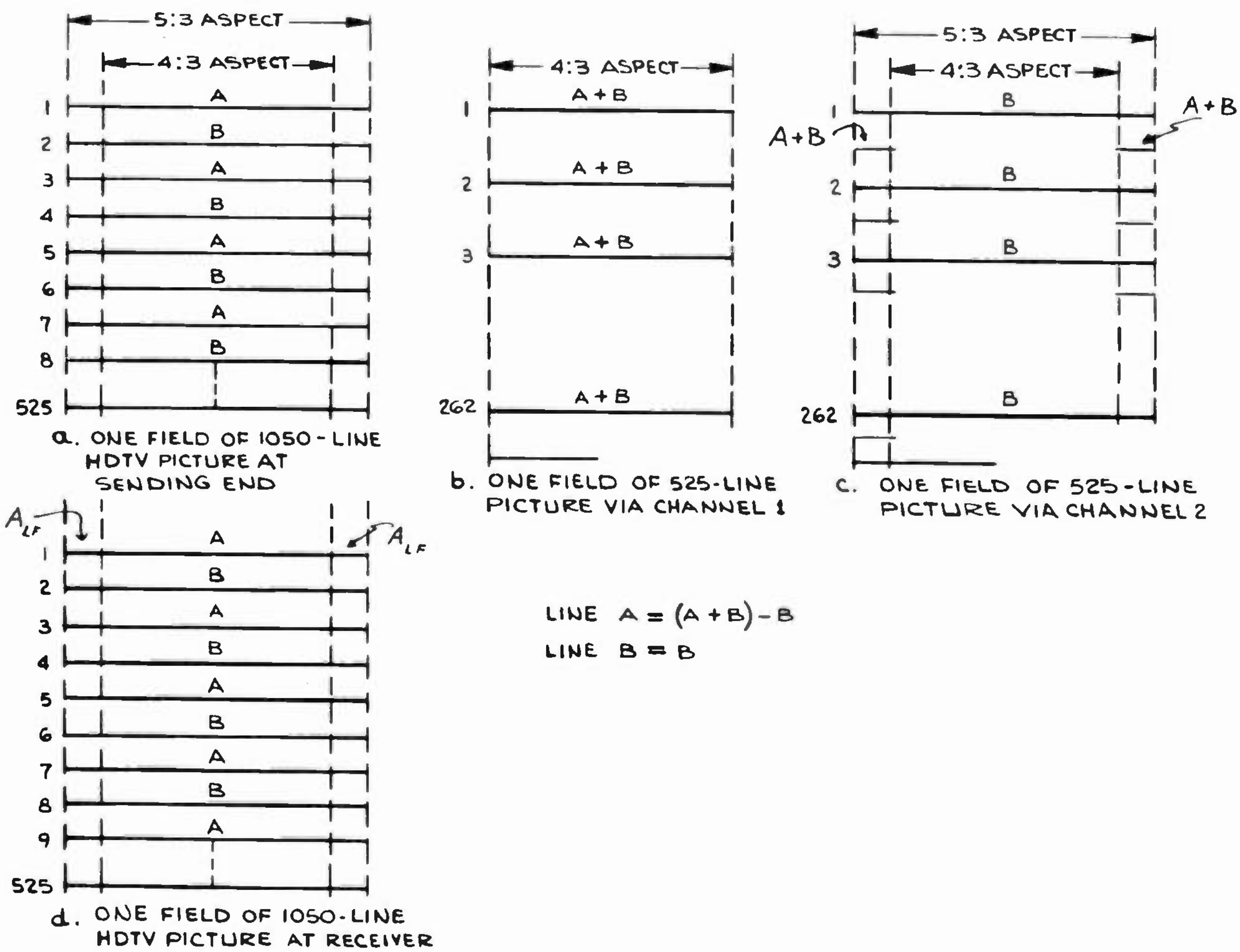


FIGURE 3. SIMPLIFIED SCANNING GEOMETRY

The line structure of Figure 3 has some spatial sampling shortcomings, therefore, the actual matrixing coefficients of lines A and B are different in order to produce a more pleasing picture in the compatible channel. Three line matrixes are also being considered for channel 1 to reduce possible aliasing.

3. TRANSMISSION SIGNAL FORMAT

As previously mentioned, standards for an HDTV production system are still to be determined. One likely prospect is 1125-line, 60 field, 2:1 interlace, 5:3 aspect ratio with 20 MHz luminance bandwidth and 6 MHz color difference signal bandwidths. However, these parameters are subject to change as worldwide deliberations take place. For example, 75 to 100 fields per second have been mentioned. The still to be determined standards converter to interface production and broadcast HDTV signals, therefore, will not be addressed in this paper.

Interlaced television rasters typically contain an odd number of lines but even numbered line rasters can also be interlaced. One method is to add a frame rate square wave to the vertical deflection of the display CRT to displace the scanning beam half a line pitch on alternate fields. Another approach is to modify the vertical scanning period to assign 525½ lines to the odd field and 524½ lines to the even field.

The 1050-line, 2:1 interlace, 5:3 aspect ratio picture must be reformatted to be carried 525-lines in each of the two 24 MHz DBS channels. Channel 1 carries a 525-line 4:3 aspect color picture plus sound while Channel 2 carries the additional picture information to reconstitute the HDTV signal.

DBS Channels 1 and 2 carry time multiplex component analog television signals frequency modulated on each RF carrier. Audio, data, picture sync and sundry control signals are time multiplexed with the video. Consequently, only one component signal exists at a time on each carrier, thereby avoiding intermodulation distortions. Signals on Channels 1 and 2 are time related; that is, data headers and sync signals are time coincident in order to synchronize the video line periods in the HDTV receiver. The two channels traverse identical up and down link paths and time differences are negligible.

Figure 4 illustrates the TMC formats in the transmission channels. Different video components with different amounts of time compression are transmitted during each TV line period of 63.55 μs. In channel 1, the luminance ($Y_{4:3}$) video occupies 42 μs and the color difference signal ($C_{4:3}$) occupies 14 μs. Audio, data, signal partitioning and clamping occupies the remainder of the line time. It should be noted that since the active time of the 4:3 video for the 1050-line picture is 21 μs, the luminance in channel 1 has undergone a 2 to 1 expansion and the color difference signal has been compressed by a 3/2 factor. With respect to a 525-line, 4:3 picture having a standard active line time of 52.5 μs, the luminance signal in channel 1 exhibits a compression of 5 to 4 and the color difference signal has three times the compression of the luminance signal.

In the augmentation channel 2, luminance ($Y_{5:3}$) occupies 33 μs and the color difference signal ($C_{5:3}$) occupies 17.5 μs. In addition to these signals the left and right signal sides C_ℓ, C_r, Y_ℓ, Y_r of the 5:3 picture not transmitted in channel 1 are transmitted in channel 2. The color sides C_ℓ and C_r occupy 2 μs each. The luminance sides Y_ℓ and Y_r occupy 3.5 μs each. 2 μs are allocated for clamping, sync recovery and signal transitions.

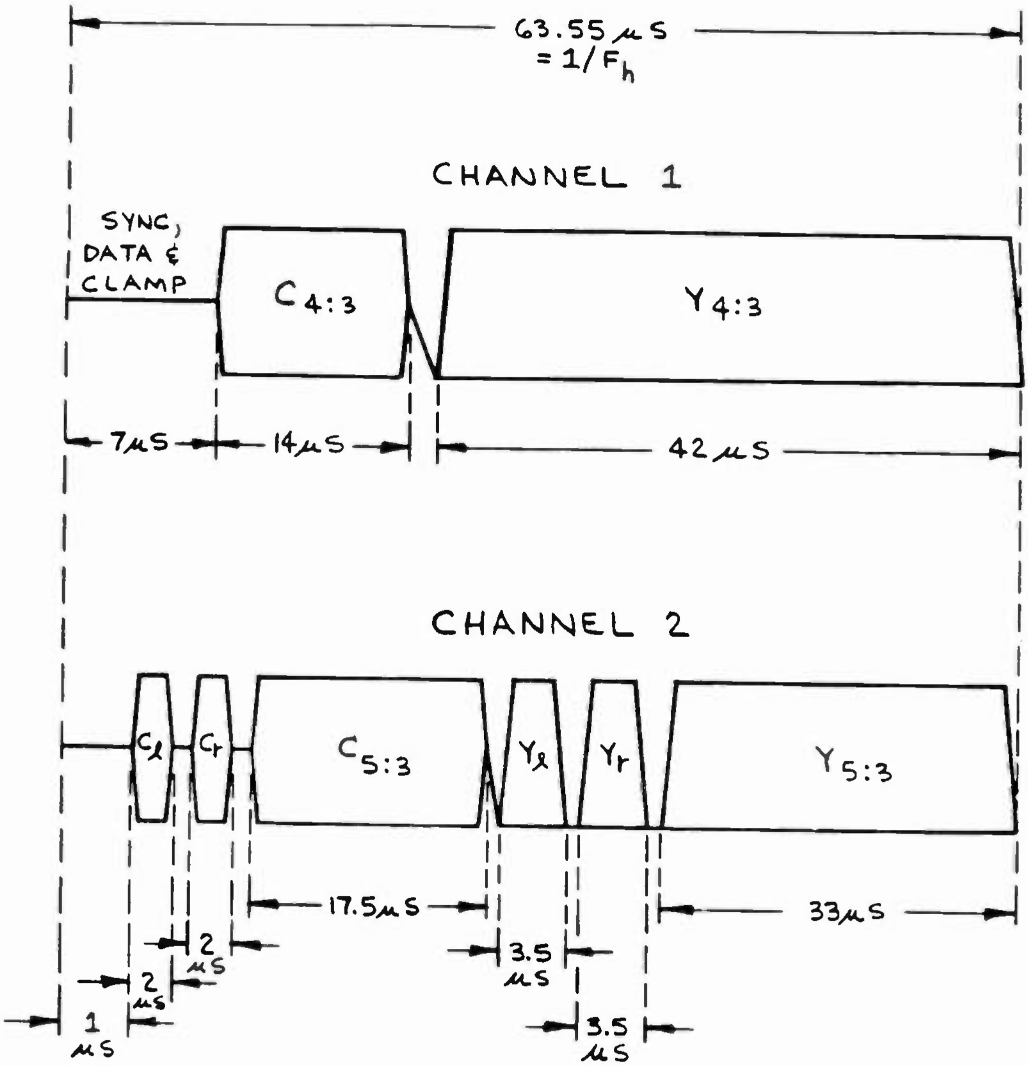


FIGURE 4. TMC FORMATS

From the stated time allocations for the different signals, it can be seen that the color difference signals compression ratio in channel 2 is the same as in channel 1. The luminance signals, however, are compressed about 37% more than in channel 1. This implies that for equal modulating frequency bandwidth in the two channels, the luminance signals bandwidth available from channel 2 will be 37% lower than the luminance bandwidth from channel 1. Therefore, after dematrixing of the 1050 TV lines in the HDTV receiver, the high frequency luminance information not carried in channel 2 can only be obtained from channel 1 and thus it will be common for each line pairs. In effect, this means that the recovered HDTV luminance signal will exhibit a comb filter characteristic for the upper 37% of the frequency band. It also implies that the luminance signal horizontal resolution at the two picture sides will be 37% lower than at the picture center. Tests have shown this to be unnoticeable. The vertical resolution will be constant throughout the 5:3 picture area. The color difference signal will exhibit homogeneous horizontal and vertical resolution throughout the 5:3 received picture.

4. SIGNAL BANDWIDTHS

The maximum luminance and color difference signal bandwidths available at the HDTV receiver are, of course, determined by the practical signal bandwidth that each 24 MHz DBS channel can carry. Given the power restrictions for satellite transponders, the coverage area, the size of the receiving antenna, interference requirements, etc., it was determined that each channel could carry at least 8 MHz of baseband signal.

In channel 1, the HDTV signal undergoes an expansion of 2 to 1, therefore, at the receiver, after recompression, the 8 MHz baseband signal provides 16 MHz of luminance bandwidth. The color difference signal, of course, will have one third the luminance bandwidth, or 5.3 MHz.

In channel 2, 37% higher luminance compression ratio is used, and that means that the luminance signal from channel 2 will have 37% lower bandwidth, or 10 MHz. The color difference signals from channel 2 will have the same bandwidth as the channel 1 color signal, i.e., 5.3 MHz.

It is also of interest to calculate the signal bandwidths available from channel 1 for the compatible 525-line receiver. The luminance compression, related to the 525-line 4:3 active picture time, is 5/4. Therefore, the 8 MHz transmitted luminance signal bandwidth decompresses to 6.4 MHz in the 525-line receiver. The color difference signals will have one third the luminance bandwidth, or 2.13 MHz. It is expected, therefore, that the compatible 525-line picture will be of excellent quality, i.e., it will be an enhanced 525 line picture having far better quality than standard NTSC pictures.

5. SOUND TRANSMISSION

Digital sound and data are time multiplexed with video and carried via frequency shift keying of the RF carrier. The multiplexed signal format allocates in channel 1 about 7 μs for audio or data, sync, clamping and video transitions as shown in Figure 4.

Digital sound coding using adaptive delta modulation (ADM) will be used. A data rate of approximately 330 K bits/sec per audio channel will be used. Such data rate has been shown to provide the high fidelity quality desired for HDTV. Delta modulation also fails very gracefully in the presence of high bit error rates and it can, therefore, provide acceptable performance in marginal signal areas.

Three audio channels will be time multiplexed with video on a line by line basis. An additional audio or data channel may also be carried in the signal's vertical interval.

6. PERFORMANCE

Baseband video concepts of the Two Channel Compatible HDTV Broadcast System have been implemented in hardware and are being evaluated in the laboratory. Following this, a simulated DBS system complete with RF links will be tested. In the meanwhile, attention is being paid to the evolving technology of satellite television broadcasting and, where necessary, appropriate changes are being made to the system.

The 1050-line compatible HDTV system can provide optimum performance within the bandwidth, noise constraints of two DBS channels. The picture quality closely approaches the 1125 line proposed production standard. Receiver cost is reasonable due to the absence of compulsory frame stores.

Received video signal to noise ratio will be influenced by power flux density PFD, and the performance of the receiving antenna plus the low noise amplifier. PFD depends upon the satellite's EIRP, which in turn is a function of transmitter power and the area of coverage. PFD will also depend on rain attenuation and the definition of reliability. Ideally, the DBS system should provide pictures with a weighted S/N of at least 49 dB. Viewers in extremely rainy locations may have to use larger antennas to obtain this performance. RARC-83 specified a PFD of -107 dBW/m^2 but the U.S. reserved the right to use -105 dBW/m^2. The performance of the HDTV system should be satisfactory within this range, but the latter value of PFD is preferred because it permits the use of smaller receiving antennas.

7. CONCLUSION

We have described a DBS-HDTV system featuring:

- Dual 525 and 1050-line scanning standards.
- Dual 4:3 and 5:3 aspect ratios.
- Channel 1 compatible with 525-line broadcasts of others.
- Time multiplexed component video signals.
- Three or more digital sound/data channels.

The performance of this system should give a significant and very noticeable improvement over present day NTSC television pictures.

The system is currently being tested at the CBS Technology Center and it is hoped that demonstrations can be given during the Fall of 1984, using a 1125 Line System for comparison purposes.

Terrestrial Interference Suppression

Using Filters and Phase Cancellation

William P. Johnson, Chief Engineer

Microwave Filter Company, Inc.

East Syracuse, New York

Terrestrial Interference (TI) is quickly becoming a major concern during the planning stage of proposed satellite earth stations. It has always been considered during the earth station licensing procedure and often the solution was a philosophically simple if not usually costly approach - move the site to a "clean" location. But practical and economic limitations frequently exclude such a move as a first choice option, and other less expensive alternatives for suppressing TI are being considered at the same time as specifications for antennas, LNA's and receivers.

The TI Problem

The majority of TI is the result of common carrier microwave links operating within the 3.7-4.2 GHz band -- the earth station receive band. Such links transmit at center frequencies 10 MHz removed from satellite transponder center frequencies. Since a domestic video transponder is 36 MHz wide, the TI center frequency will be 8 MHz inside the transponder bandwidth. The big question for the earth station receiver then is, "What is the desired signal, and what is the interference?" Demodulators being the faithful circuits they are interpret the entire input as signal. Therefore, with TI present, the resulting output may only slightly resembles the original signal, if at all.

The microwave links carry voice, data and video, and the significant bandwidth can vary from a few megahertz (narrow-band) to up to 30 MHz (wide-band). Wide-band TI requires site shielding or microwave phase cancellation if adjacent transponders are to be protected since any filter which removes the TI bandwidth would clearly remove much of the desired signal. Narrow-band TI can be removed by application of notch filters in addition to wide-band suppression methods.

When notch filters are used, a small amount of the desired signal spectrum will be removed along with the narrow band TI. This implies the introduction of some distortion. Each application must determine whether such small distortion is really significant or not. If it is deemed too severe, then microwave phase cancellation and finally the ultimate site-relocation option must be exercised.

Filtering Narrow Band TI in the I.F. and Microwave Paths

Whether to place notch filters in the earth station I.F. or Microwave path depends upon two considerations:

(1) Is the TI level low enough that it does not saturate the LNA or downconverter circuitry and does not detuned AFC controlled local oscillators?

(2) Is the TI multiple frequency?

As long as TI levels are low such that non-linear operation of mixer and amplifier stages does not occur and as long as the AFC functions properly, then one is free to filter at either location. When TI levels are high, intermodulation products and detuned AFC operation is typical. In such cases, some suppression before these stages is required - usually between the LNA and downconverter. Where only one or two TI frequencies are involved and levels are low, it may be economically advantageous to price out both possibilities -- individual notch filters at microwave, or a pair of I.F. Traps.

Since the downconverter will retain the relative frequency spacing of the transponder and the TI, the TI will always center up at plus or minus 10 MHz from the I.F. center frequency -- providing the AFC is tuned and functioning normally. Therefore, for a 70 MHz I.F., a notch filter would be placed at 60 MHz or 80 MHz. To prevent unnecessary removal of desired signal energy, a bypass switch can select one or both traps as required, or remove both in the absence of TI. In cases of high level TI, notch filters at each TI frequency would be inserted into the microwave path, usually after the LNA and before the downconverter.

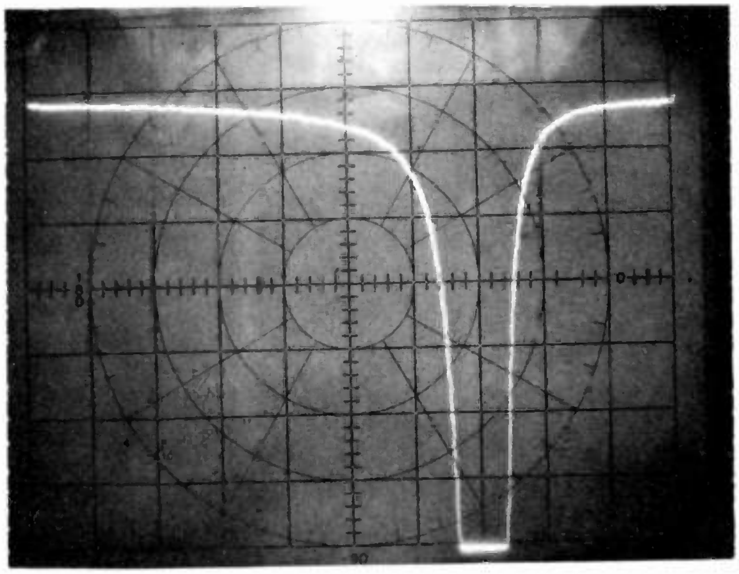
Notch Filters -- Response and System Performance

As an example, Figure 1a and b show the insertion loss and insertion phase response of an 80 MHz I.F. notch filter. The 40 MHz wide transponder allocation containing the 36 MHz wide video spectrum would extend +/- 4 division. It is clear that the 1 dB points at 80 +/- 2.5 MHz and the corresponding differential group delay of 48nS to those points will introduce some amplitude and phase distortion of the FM transponder signal. Calculation of such distortion as it disrupts the video signal is extremely difficult and impractical. However, measurement of the Verticle Interval Test Signals (VITS) will quickly shed some light on the subject.

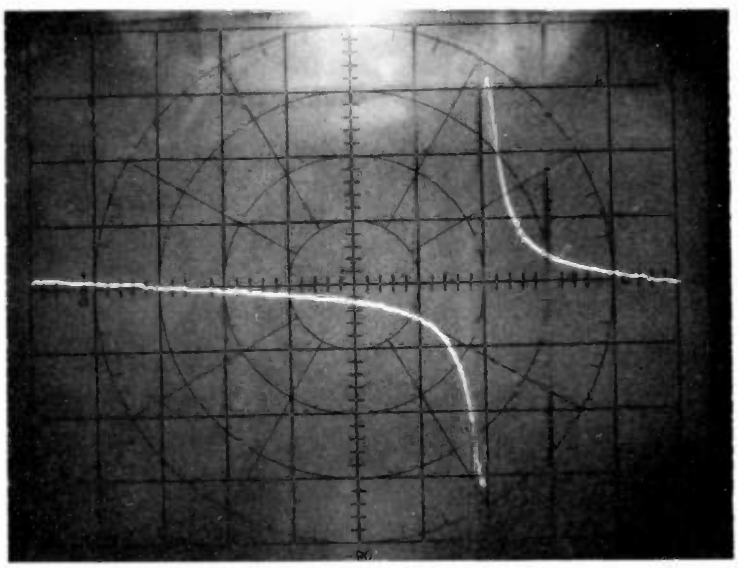
VITS Measurements on a Filtered Receiver

To illustrate the affects of filters on the received signal, a 30 MHz I.F. bandwidth commercial receiver was used -- an Agile 24/M manufactured by Standard. Its output VITS response was measured using a Tektronix model 380 NTSC test monitor. Measurements were made using no filters, an 80 MHz I.F. trap, and a 4170 MHz microwave trap while viewing transponder 23 on SATCOM F3. Measurement and interpretation is as described in NTC report No. 7.[1] Test performed were Chrominance-Luminance Gain and Delay Inequality, Gain/Frequency

Distortion, and Differential Gain.



(a)



(b)

Figure 1: 80 MHz 45 dB I.F. trap response. Satellite signal bandwidth would extend about +/- 4 divisions.

(a) Insertion Loss:	0.25 dB/Div Vert.	5 MHz/Div Horiz.
	0 dB=+3 Div from ctr.	70 MHz=Center
(b) Insertion Phase:	45°/Div Vert.	5 MHz/Div Horiz.
	0°= Center	70 MHz=Center

Chrominance-Luminance Gain and Delay Inequality[2]

Figure 2a shows the 2T pulse and chrominance pulse of the vertical interval test signal as processed through the satellite receiver. As measured, the

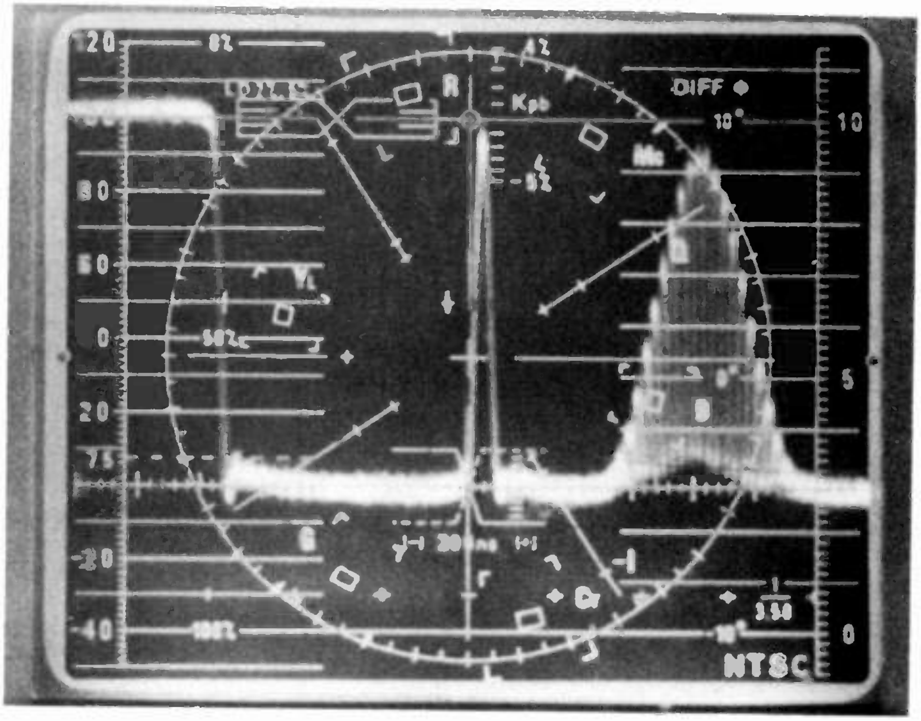


Figure 2a: Chrominance-Luminance Gain and Delay Inequality for Unfiltered Receiver

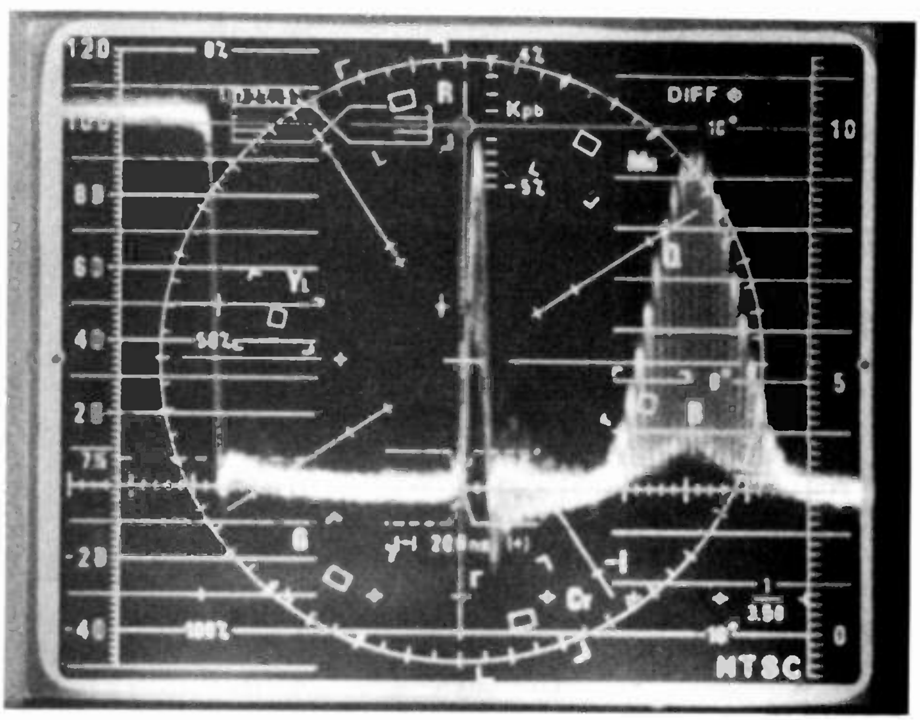


Figure 2b: Chrominance-Luminance Gain and Delay Inequality for Receiver with I.F. Filter at 80 MHz.

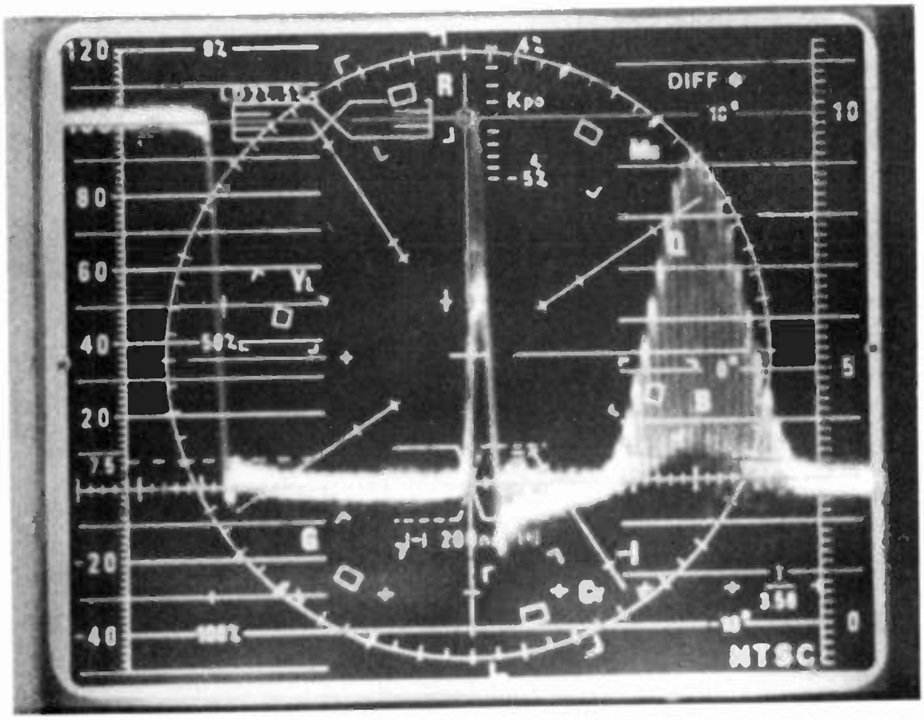


Figure 2c: Chrominance-Luminance Gain and Delay Inequality with Microwave Filter at 4170 MHz.

gain inequality is 6 IRE units with respect to 100 IRE units. When the 80 MHz I.F. trap is inserted, Figure 2b shows the gain inequality remains the same indicating no C-L gain distortion. Note that significant "ringing" occurs at falling edges of sharp amplitude changes. This is characteristic of amplifier response in the presence of non-optimum loading. When the microwave trap is installed, Figure 2c shows the C-L gain distortion rises to 7 IRE units. Worst case C-L gain distortion above the un-filtered receiver: 1 IRE unit. The NTC limit is 100 +/- 3 IRE units.

All three measurements indicate only slight C-L delay distortion.

VITS Gain/Frequency Distortion[3]

Figure 3a shows the multiburst portion of the verticle interval combination test signal as processed by the satellite receiver. Each of the six frequencies are within the 45-53 IRE unit window allowed by the NTC. When the 80 MHz I.F. trap is inserted, Figure 3b shows a slight rolloff in high frequency response. The insertion of the 4170 MHz microwave trap and removal of the I.F. trap results in the display shown in Figure 3c which is not measurably different than that of the I.F. trap. Table I summarizes the information obtained in these measurements.

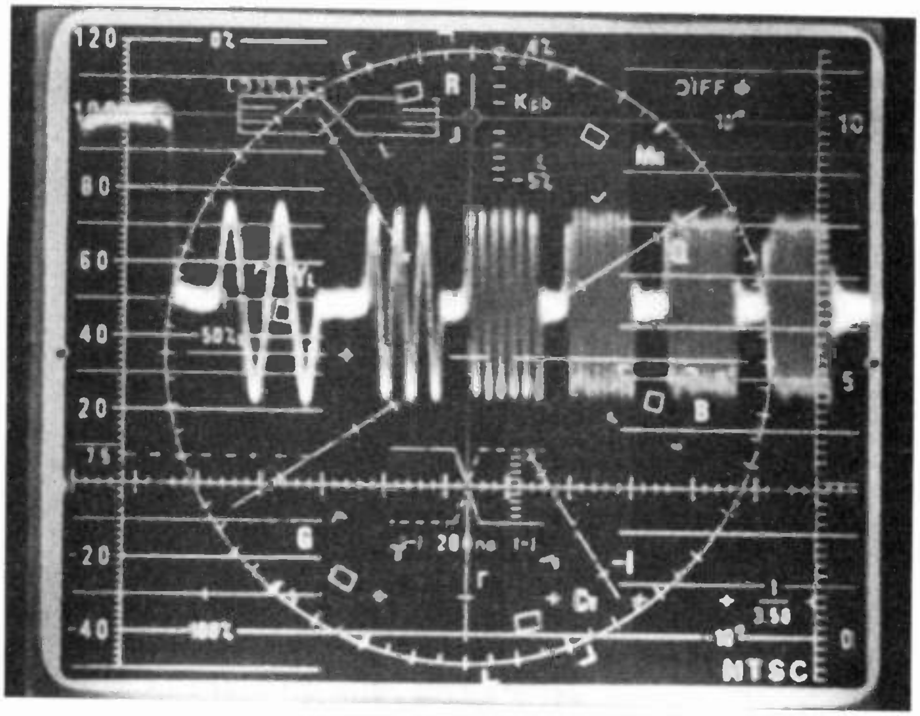


Figure 3a: Gain-Frequency Distortion for Unfiltered Receiver

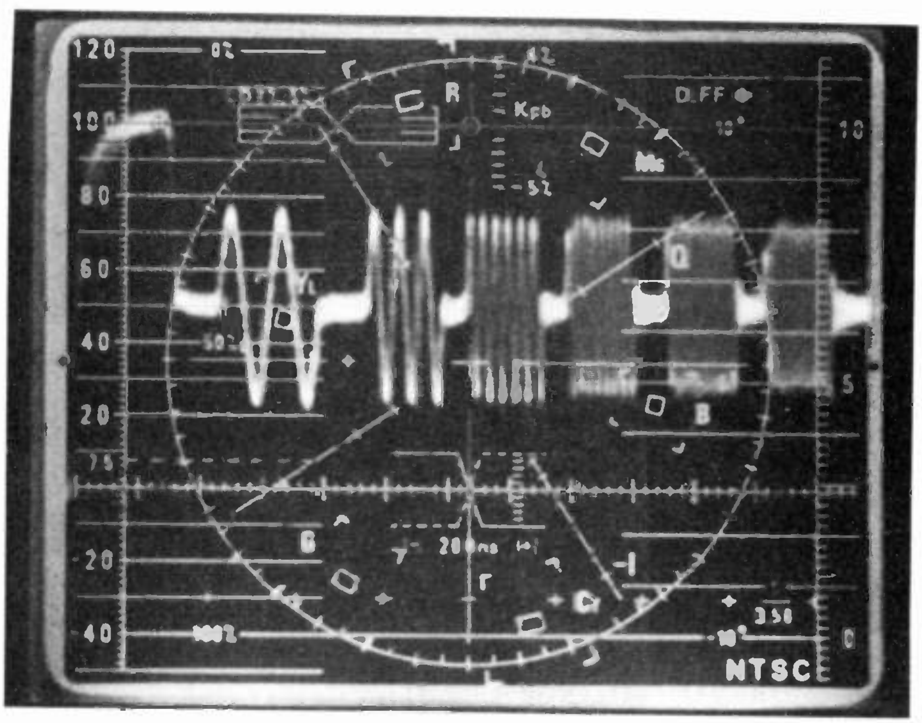


Figure 3b: Gain-Frequency Distortion with I.F. Filter at 80 MHz.

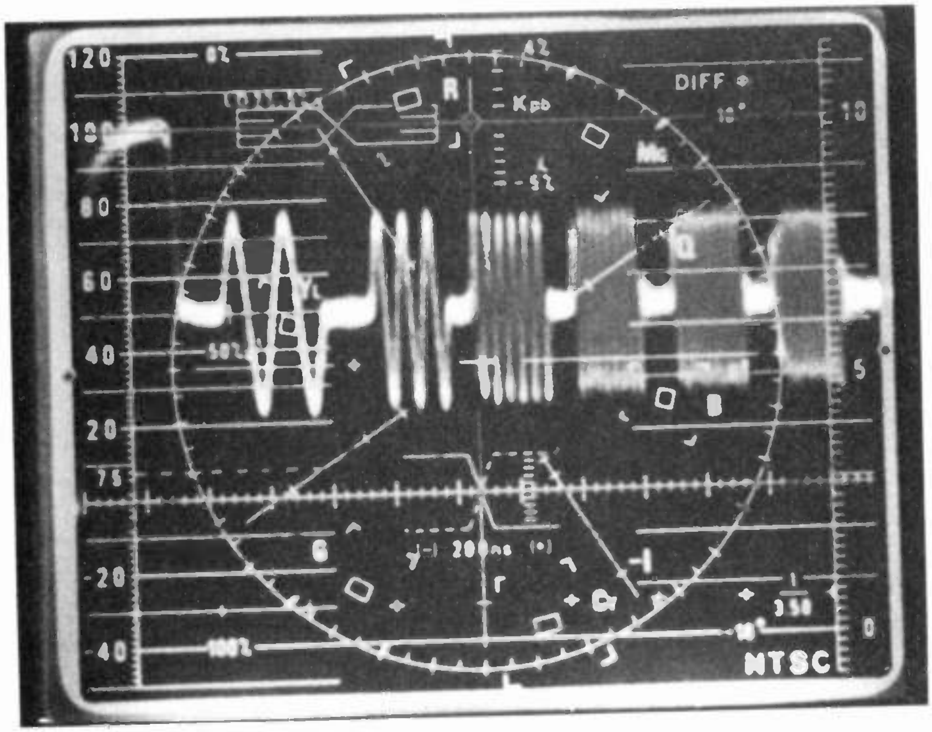


Figure 3c: Gain-Frequency Distortion with Microwave Filter at 4170 MHz

Frequency (MHz)	---- Amplitude (IRE Units) ----		
	No Filters	80 MHz I.F. Traps	4170 MHz Trap
0.5	55	55	55
1.0	54	54	54
2.0	51	51	51
3.0	50	50	50
3.58	47	47	47
4.2	47	45	45

Table I: Summary of Multiburst measurements for Gain/Frequency Distortion. The recommended NTC limit is 45-53 IRE units.

VITS Differential Gain[4]

Figures 4a, b, and c show the change in amplitude of the chrominance sub-carrier superimposed on the six step luminance signal varying from blanking level to white level. Figure 4a is with no filtering, Figure 4b is with the 80 MHz I.F. trap, and Figure 4c is with the 4170 MHz trap in place. The differential gain is 8% without filters and 9% worst case with either the I.F. trap or the microwave notch filter. The NTC maximum is 15%.

While numerous other tests could be performed, these indicate that filtering does not introduce a large amount of distortion into a video receive system.

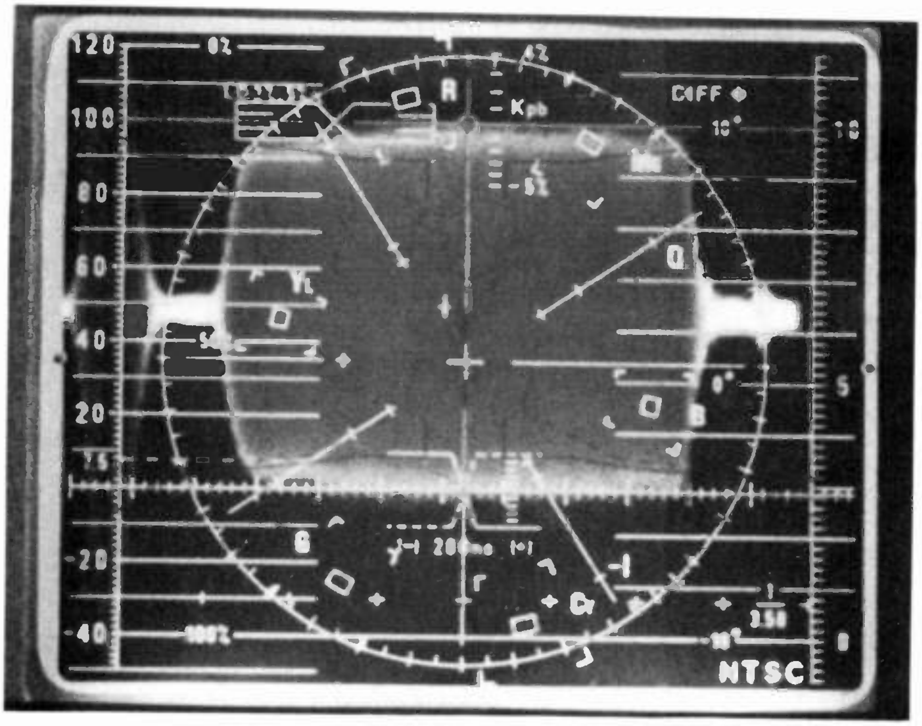


Figure 4a: Differential Gain Measurement for Unfiltered Receiver

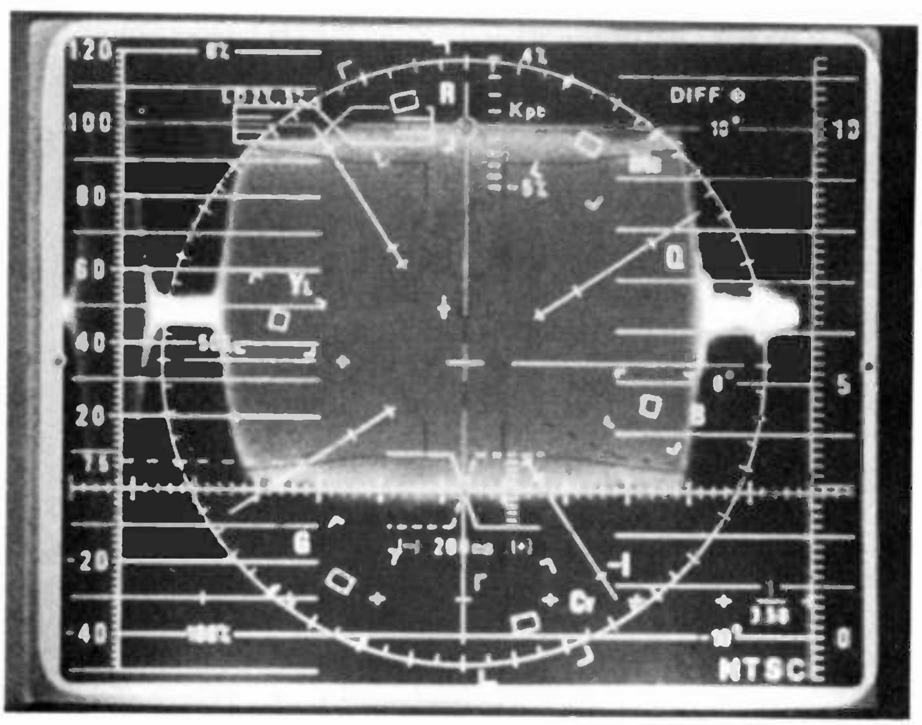


Figure 4b: Differential Gain Measurement with I.F. Filter at 80 MHz.

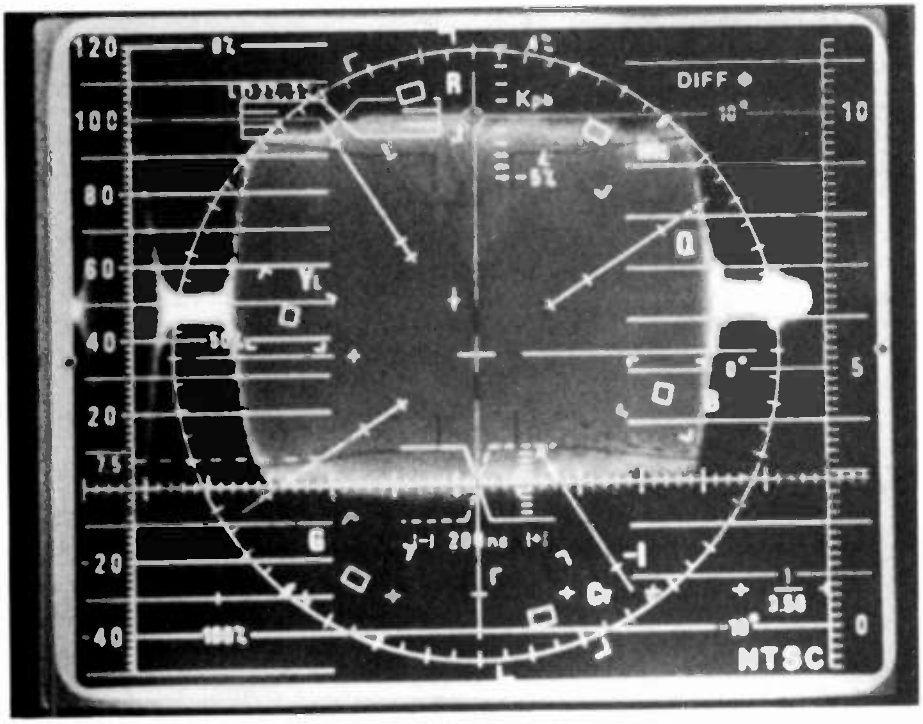


Figure 4c: Differential Gain Measurement with Microwave Filter at 4170 MHz.

However, pulse rise and fall times can become somewhat distorted as shown in some of the measurements. This is due less to the action of the filtering process and more to the introduction of filters after active stages which may respond irradically to such a varying load impedance. This might be overcome by such a simple "fix" as padding the input and outputs of the filters where absolute signal level can be reduced, such as in an I.F. stage, or where it can be easily restored, such as with a line driver ahead of a microwave filter. At any rate, the data demonstrates that filtering of TI should not be viewed as savage butchery of video signals and that it may instead provide cost effective options to the overall reception problem in the presence of narrow band TI.

But, one thing that filtering is not -- a general solution to wide-band TI, where significant portions of the desired signal are overlapped by interference.

The reason wide-band TI requires special attention is that, by the time it and the desired signal are mixed, there is no receiver which can distinguish between them. The TI is too wide to be filtered, which is a frequency differential technique anyway. However, a big difference does exist which will allow removal of the TI -- the desired signal and the TI are approaching the earth station from different directions. Enter the Microwave Phase Canceller.

Microwave Phase Cancellation

The technique of Microwave Phase Cancellation has been used extensively by the military for years to prevent "jamming" of radar and communication equipment. It's application to earth stations is a natural.

Microwave Phase Cancellation consists of using a "test" antenna to pick up a sample of the TI. By adjusting its level and phase properly it can be mixed with the original interference causing a vector addition which leaves only the desired signal. This is illustrated in Figure 5.

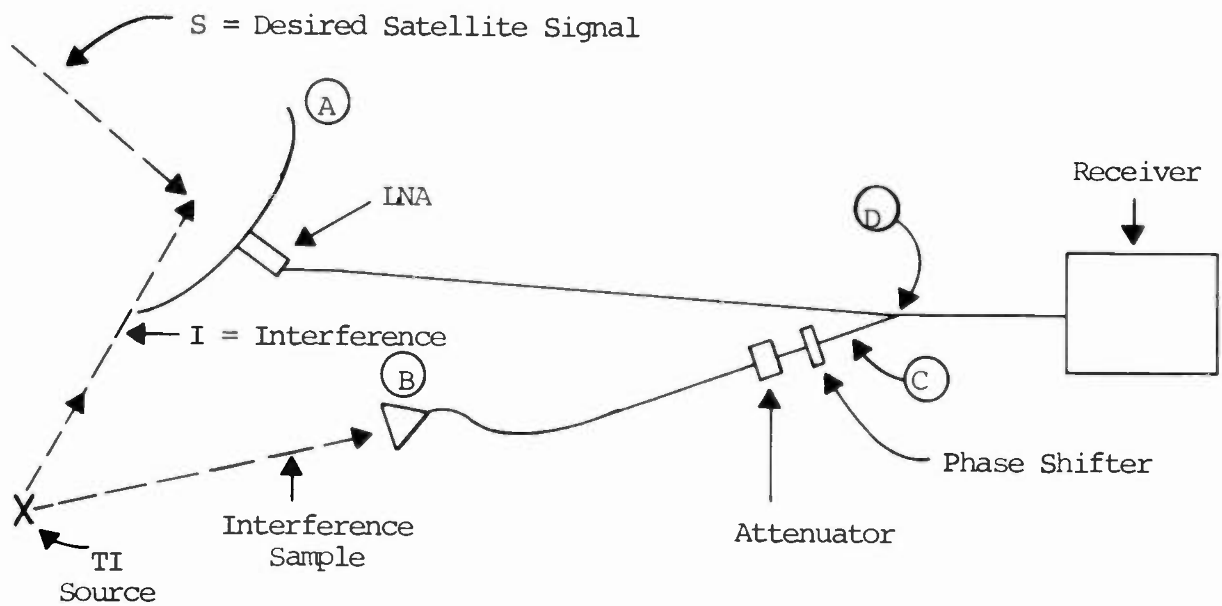


Figure 5: Illustration of the Microwave Phase Canceller. A sample of TI is picked up at B, modified in phase and amplitude then mixed with the main path signal and interference at D. The vector term at C performs the cancellation of the main path TI.

The Theory Behind Microwave Phase Cancellation

Refer to Figure 6. Assume the following signal voltages.

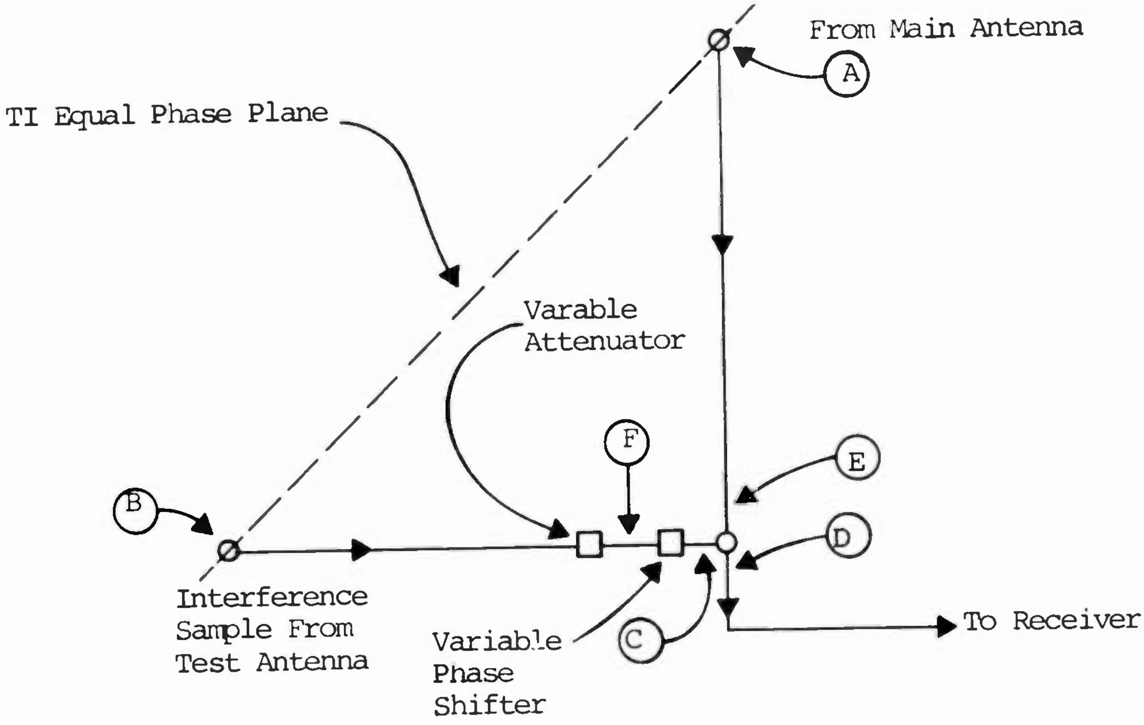
At A: $S + Ie^{j0}$ (1)

At B: kIe^{j0} $k > 1$ (2)

Where S = Complex (magnitude and phase) of the desired satellite signal

I = Interference magnitude

This condition is easy to accomplish in the field by moving the test antenna, and it simplifies the analysis. The factor k represents the desired condition that the interference is stronger in the test antenna path -- remember that it is pointed directly at the TI source. An amplifier can be added if required.



Main Path Length AE = L_s inches

Sample Path Length BF = L_i inches

Figure 6: Schematic of basic Microwave Phase Cancellation System

---

Moving from point A to point E, the input signal at point A (Equation 1) experiences a phase shift such that it becomes:

$$\text{At point E:} \quad S + Ie^{-j\theta_s} \qquad (3)$$

$$\text{Where:} \quad \theta_s = 2\pi \frac{L_s}{\lambda_o} \frac{F}{F_o}$$

L_s = Path Length AE
λ_o = Wavelength at Center Frequency of Interference
F_o = Center Frequency of Interference
F = Frequency Variable

The sample of interference at point B (equation 2) experiences a similar phase shift as it progresses toward point C. Note that up to point F, both phase and magnitude are changed. The phase changes due to path length, and the magnitude is adjusted by the variable attenuator set to A dB.

$$\text{At point F:} \quad I_e{}^{-j\theta_I} \qquad (4)$$

$$\text{Where:} \quad \theta_I = 2\pi \frac{L_I}{\lambda_o} \frac{F}{F_o}$$

L_I = Path Length BF (see Figure 6)
λ_o, F_o and F as in equation 3
A dB = 20 Log k

And,

$$\text{At point C:}\quad S + I_e^{-j(\theta I+\phi)} \qquad (5)$$

Where: $(\theta_I + \phi)$ = Total phase shift along path BC (see Figure 6)

The additional Phase Shift (ϕ) is purposely introduced such that a vector term is created which will cancel the interference in the main path AE (see Figure 6).

The value of ϕ_o, the phase shift at the interference center frequency, can be calculated as follows.

There exists, in general, a path difference of:

$$\Delta\theta_o = \frac{2\pi}{\lambda_o}(| L_s - L_I |) \text{ (radians)} \qquad (6)$$

The difference between the paths, excluding multiples of a full wavelength is:

$$\Delta P_\lambda = [\frac{\Delta\theta_o}{2\pi} - \text{INT } (\frac{\Delta\theta_o}{2\pi})] \text{ (wavelengths)} \qquad (7)$$

If the satellite signal path length, L_s, is greater than the sample path length, L_I, then the positive phase shift at the interference center frequency which needs to be introducted in the sample path is:

$$\phi_o = 2\pi(\Delta P_\lambda - \tfrac{1}{2}) \quad : \quad \Delta P_\lambda \geq .5 \qquad (8a)$$

$$\phi_o = 2\pi(\Delta P_\lambda + \tfrac{1}{2}) \quad : \quad \Delta P_\lambda < .5 \qquad (8b)$$

And as a result,

$$\phi = \phi_o\frac{F}{F_o} \qquad (9)$$

The phase length of path AE and BC are equal, except for multiples of a full wavelength at F_o, with the addition of a ϕ_o Phase Shift (at F_o) on path BC.

The net voltage, V, at point D therefore, is:

$$V = S + [I_e^{-j\theta_s} + I_e^{-j(\theta_I + \phi_o \frac{F}{F_o}}] \qquad (10)$$

Since it is desired to have only S, the satellite signal remain at point D, then it is necessary to minimize the term in parenthesis over the band of interest -- that of the interference. It will be zero at F_o. This expression assumes that the attenuator and phase shifter are set once and not adjusted further for a particular F_o. The parenthesis term in Eq (10) will be small for frequencies near F_o, and it will be optimumly small when $\theta_s = \theta_I$.

The attenuation of the interference is a function of the match between θ_s and θ_I, and of the frequency of interest. Figure 7 shows a graph of attenuation vs frequency for F_o = 3950 MHz. Note the degradation of bandwidth due to path length differences (i.e., θ_s and θ_I). It is important to match lengths and take care in placement of the test antenna to achieve best results. Any lengths will achieve good rejection at a single frequency, but only matched path lengths will allow the best bandwidth characteristics, and hence the best suppression of the TI.

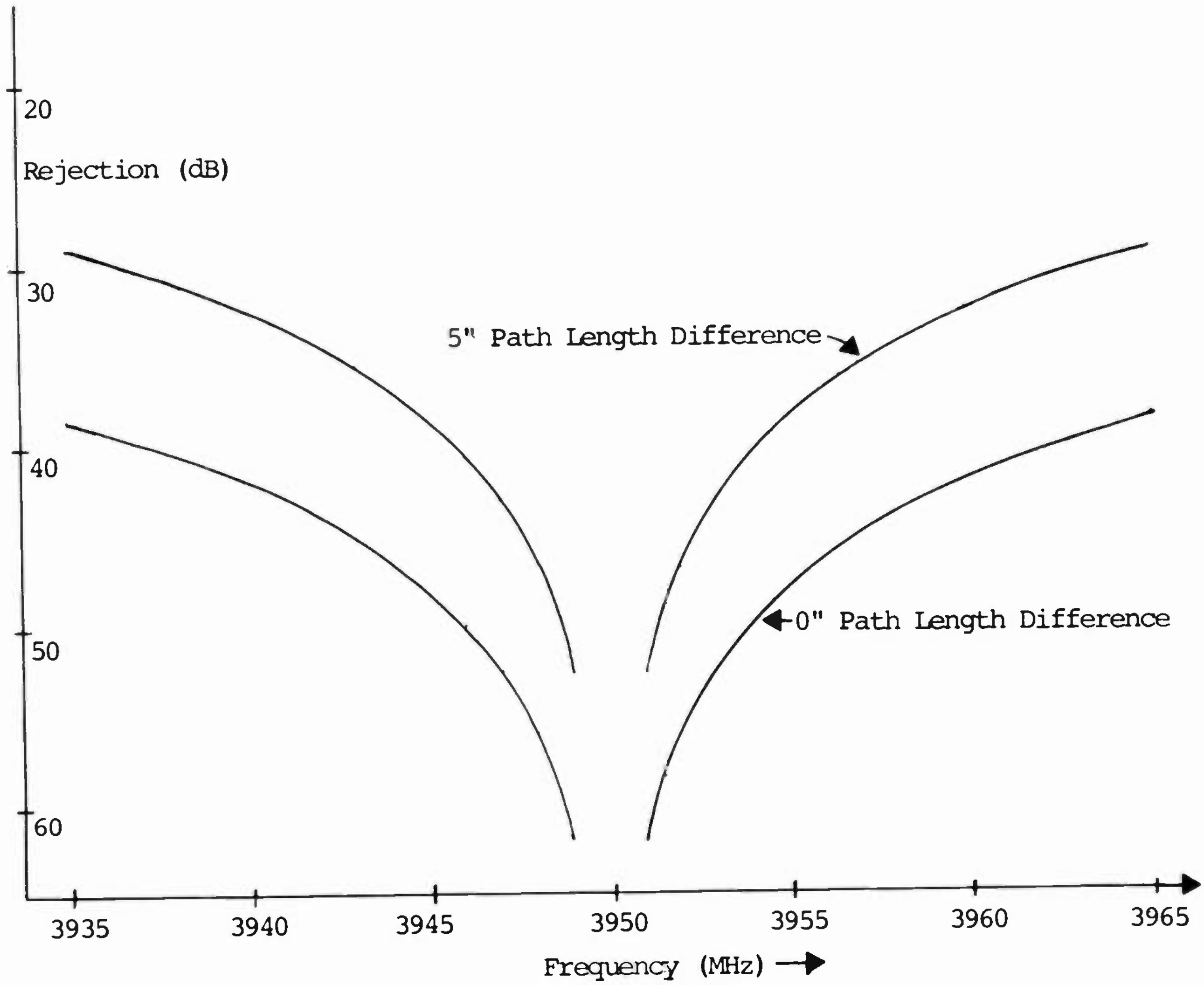


Figure 7: TI Rejection vs. Frequency For Microwave Phase Canceller

Practical Application of Microwave Phase Cancellation

The concept of microwave phase cancellation can be expanded upon to overcome a variety of particular difficulties, or it can be applied in a far less costly manner as a basic system.

The basic system uses an manually adjusted attenuator and phase shifter to set up the proper magnitude and phase relationships. The test antenna is pointed in the direction of the interfering source. It is positioned such that its phase center and that of the main dish are on equal phase planes. This is not necessary, but any deviation from that location will require correction of cable lengths to assure equal phase paths from the TI sources. Once the basic system is set up for a given F_o, it will require readjustment to move the optimum cancellation to a new center frequency. Additional test antennas, attenuators and phase shifters will be required to handle multiple azimuth and multiple center frequency TI.

A modification of the basic system utilizes control system technology and real-time computer surveillance of the system output to optimize the microwave phase cancellation. While such a system will modify adjustments in response to changing interference magnitude and phase parameters, like the basic system, it will only handle one frequency at a time. The advantage is the ability to optimize itself in response to dynamic inputs.

The functional difference between a computer controlled system and a basic manual setup is in the automatic adjustment features. The economics, which may be a real concern, are vastly different. The basic system including the specialized installation and hardware conceivably could run between $5,000 - $7,000 as an estimate -- probably conservative. The computer controlled sytem will be about 5 times the cost. As usual, one must carefully weigh the features gained as compared to the cost.

Conclusion

The vague generality that filters are not useful to the broadcaster in combating TI has been questioned by comparing an unfiltered and filtered receiver in the absence of TI and comparing some VITS responses. It is clear that the rumored severe distortion simply is not the case, and such techniques should be considered as one of the many options available to allow earth station operation in the presence of narrow-band TI.

Where wide-band TI is involved, another method involving microwave phase cancellation was presented, and some discussion of its theory and application was highlighted.

These are a small but powerful set of tools to the operator or installer of earth stations. And in the face of increasing TI as communications links expand, they will show themselves as reliable approaches to the problem of getting the signal without giving away the checkbook. And that's not static.

Footnotes

1. "NTC Report No. 7," The Public Broadcasting Service for the Network Transmission Committee of the Video transmission Engineering Advisory Committee, June 1975 Revised January 1976.
2. IBID, pp 28-31
3. IBID, pp 32-33
4. IBID, pp 44-45

The author wishes to express gratitude to the management of Microwave Filter Company, Inc. for their encouragement and support of this paper, as well as providing the resources to allow its preparation.

HIGH PERFORMANCE MULTIBEAM TORUS ANTENNA

FOR ACCESSING C AND KU BAND SATELLITES

Marvin D. Shoemake

SatCom Technologies, Inc.

Norcross, Georgia

The Torus antenna is a multiple-beam offset fed reflector capable of viewing broad segments of the satellite arc. Additionally, the antenna operates across the C- and Ku-Band frequencies simultaneously and meets the stringent sidelobe pattern requirements for 2-degree satellite spacing.

The multiple beam, multiple frequency Torus antenna has a geometry quite different from conventional parabolic earth station antennas. The reflector surface is formed by rotating an offset parabolic section about a generating axis. The resulting shape is parabolic in one dimension and circular in the other. Figure 1 illustrates the geometry of the Torus. This geometry generates a focal arc through the point "F" which is produced by rotating about the "U" axis. Independent multiple feeds placed along the focal arc generate beams which can view any satellite located in the 50^{o} field-of-view of the antenna.

Phase Aberrations

The Torus is not a surface of revolution, therefore some phase aberrations are generated across the aperture section of a given feed. The aberrations cause the resultant pattern and gain of the antenna to degrade. The amount of degradation is directly related to the amount of phase error across the aperture section being used. These aberrations are of course frequency dependent and become worse as the frequency is increased. This degrading effect can be minimized by proper feed and resultant pattern illumination design. The Torus geometry directly affects the amount of phase aberration and therefore considerable effort is expended in choosing the proper design to minimize this phenomena. Typical patterns of 4.5 and 7 meter Torus antennas at C- and Ku Bands are shown in figures 2, 3, 4 and 5. The aberration can be noted in these patterns at Ku-Band frequencies. There is some slight amount of first sidelobe asymmetry and the lack of a first null on one side of the pattern. This results in a five to ten percent reduction in efficiency versus an equivalent parabolic antenna at the Ku-Band frequencies. While this performance is slightly less than optimum, it is felt that the advantages of simultaneously viewing any satellite, whether C- or Ku-Band within a 50^{o} orbital arc segment outweigh the small degradations in performance.

Offset Geometry

Conventional axis symmetric antennas have an inherent problem of supporting the feed or subreflector at the focal point of the reflector. This support structure blocks a part of the aperture and scatters the energy impinging on this blocked reflector section. The blockage and resultant scattering results in increased antenna sidelobe levels which lead to potential adjacent satellite interference and characteristics that do not meet the FCC sidelobe criteria for two-degree satellite spacing.

The Torus antenna design geometry provides a focal arc located outside the reflector aperture. This offset reflector design eliminates the blockage and scattering support mechanisms present in axis symmetric antennas, resulting in much lower sidelobe characteristics, which in fact do meet the sidelobe criteria defined by the FCC for two degree satellite spacing.

Physical Description and Specifications

The Torus antennas are available in 4.5 and 7 meter equivalent sizes. The specifications for these Torus antennas are given in Table 1. The major difference between the TC and TCK versions of the two antenna sizes is the surface accuracy required at Ku-Band. The TCK Torus will provide optimum performance at both C- and Ku-Band frequencies which meets the FCC requirements for two degree satellite spacing.

The Torus antenna is much larger than a single beam equivalent gain parabolic antenna. Figures 6 and 7 illustrate typical outline dimensions for a 7 meter Torus antenna. Although the size is significantly greater than a single beam antenna it requires much less installation real estate than two or three antennas required to view two or three satellites. When one considers that as many as 25 satellites within the antenna's field-of-view can be viewed simultaneously the size and real estate required for the Torus is very small comparatively.

The Torus antenna is aligned with the orbital arc along the scan plane ("L" refer to figure 1) of the aperture. Therefore, depending on the earth station location, the reflector will be oriented in various positions ranging from the scan plane perpendicular to local vertical to a position of about 45 degrees to local vertical. The mounting structure of the Torus is designed to accommodate the different mounting configurations. A typical installation is shown in figure 8.

The specially shaped aluminum panels of the Torus are attached to a spine truss back structure designed to translate loads during high wind conditions into the mount tripod and bipod leg assemblies, which in turn transfer loads into the concrete foundation. This structure is clearly illustrated in figure 9. The structural design of this very stiff antenna structure is consistent with surviving 125 mph winds and operating without degraded performance up to 45 mph. The reflector panels and support structure provides a reflector surface accuracy of .025 inches rms, and pointing accuracies of .028° in 45 mph winds on the TCK models.

Feeds

Feeds for the Torus antenna are designed to optimize performance consistent with the geometry of the antenna at C- and Ku-Band frequencies; receive and transmit, linear and circular polarizations with dual polarized and full frequency reuse capabilities available. The feed horn designs are corrugated or multimode horns to achieve circular symmetric pattern characteristics and very low cross polarization isolation. Figure 10 illustrates a typical feed arrangement. The feeds shown are three C-Band feeds and a single Ku-Band feed mounted on the feed tray of a Torus antenna.

Applications

Today, approximately 23 satellites occupy the 70-143° west longitude geostationary arc: 19 C-Band and 4 Ku-Band. Presently there are approximately 25 additional satellites being planned for launch through 1985: 14 C-Band, 5 Ku-Band, 6 C/Ku-Band. There are about 6 satellites, all C-Band, whose expected life will exceed ten years through 1985, leaving a total potential of 42 satellites in orbit: 27 C-Band, 9 Ku-Band and 6 hybrid C/Ku-Band.

The Torus antenna is capable of viewing any 50° segment of the orbital arc providing access to any one or all satellites within this field-of-view. This provides the earth station owner with a great deal of flexibility since to view additional satellites it is only necessary to add a feed and associated electronic equipment. No additional real estate, construction permits or zoning hearings are required. The Torus antenna should be seriously considered for earth stations requiring access to multiple satellites. It offers a viable economic alternative which will meet stringent technical requirements of future two degree satellite environments.

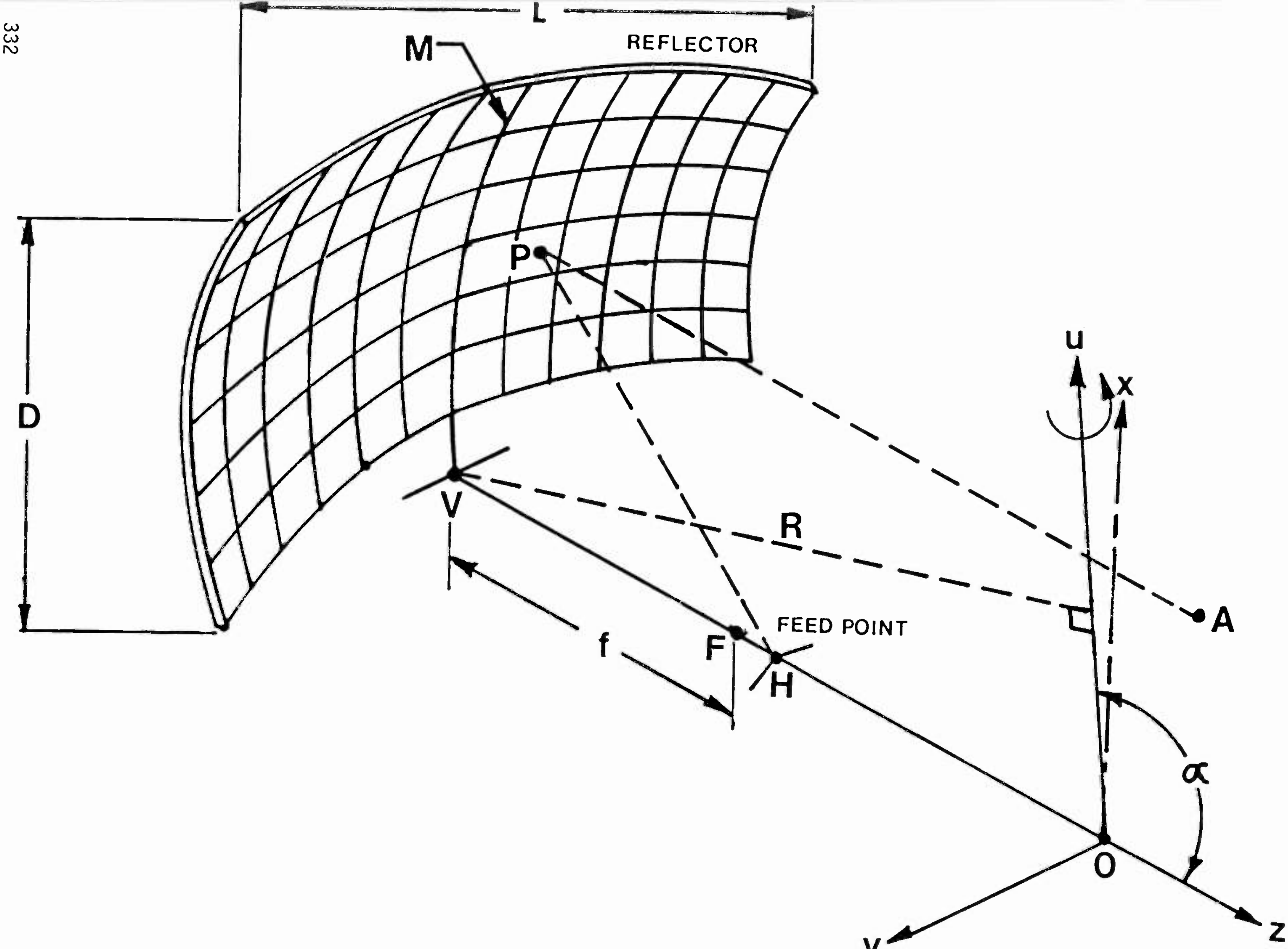


Figure 1. Torus Antenna Geometry

TABLE 1

ELECTRICAL	MODEL 450TC	MODEL 450TCK	MODEL 700TC	MODEL 700TCK
Frequency (GHz)				
Receive	3.7-4.2	11.7-12.2	3.7-4.2	11.7-12.2
Transmit	5.925-6.425	14.0-14.5	5.925-6.425	14.0-14.5
Gain at Midband (dB)				
Receive	43.5	50.0	47.3	54.1
Transmit	46.1	51.3	49.3	54.8
VSWR	1.3:1	1.3:1	1.3:1	1.3:1
Beamwidth at Midband				
Receive -3 dB	1.15°	.43°	.75°	.28°
-15 dB	2.25°	.88°	1.50°	.57°
Transmit -3 dB	.75°	.38°	.50°	.23°
-15 dB	1.45°	.78°	1.05°	.46°
First Sidelobe Level	-15 dB	-15 dB	-15 dB	-15 dB
Radiation Pattern	Meets current FCC specifications for 2° Satellite Spacing			
Antenna Noise Temperature (ref. omt port), Typical 10° Elevation	36°K	46°K	35°K	44°K
Cross-polarization Discrimination 1 dB Beamwidth	-30 dB	-30 dB	-30 dB	-30 dB
Power Handing Capability	5 kW CW	2 KW CW	5 kW CW	2 KW CW
Isolation Between Ports	35 dB min	35 dB min	35 dB min	35 dB min
MECHANICAL				
Reflector Size	36'W x 15'H	36'W x 15'H	W x 23'H	W x 23'H
Equivalent Diameter	4.5-meter	4.5-meter	7-meter	7-meter
Reflector Construction-AccuShape	12 panels	12 panels	27 panels	27 panels
Field of View	50°	50°	50°	50°
Net Weight	5,500 lbs	5,500 lbs	10,000 lbs	10,000 lbs
Shipping Weight	7,500 lbs	7,500 lbs	13,000 lbs	13,000 lbs
Shipping Volume	1,500 cu ft	1,500 cu ft	3,000 cu ft	3,000 cu ft
ENVIRONMENTAL				
Wind Loading at 32°F				
Operational	30 mph gusting to 60 mph			
Survival	125 mph, no ice-87 mph with 2" radial ice			
Pointing Accuracy				
Wind 30 mph gusting to 45 mph	0.035° rms	0.028° rms	0.035° rms	0.028° rms
Wind 45 mph gusting to 60 mph	0.070° rms	0.056° rms	0.070° rms	0.056° rms
Temperature Range				
Operational or Survival	-40°F to 140°F	-40°F to 140°F	-40°F to 140°F	-40°F to 140F
Atmospheric Conditions	Salt, pollutants and corrosive contaminants as found in coastal and industrial areas			

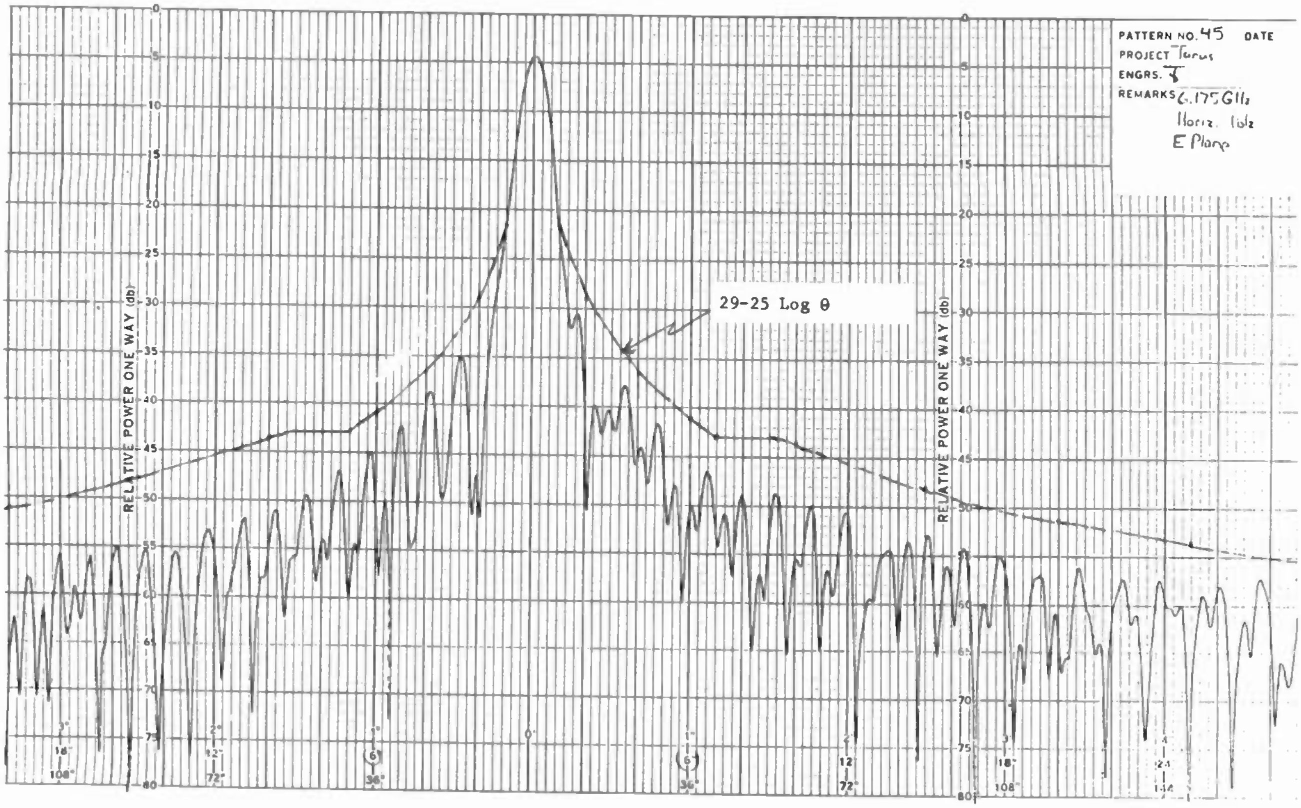


Figure 2. TYPICAL 4.5 METER TORUS 6 GHZ PATTERN

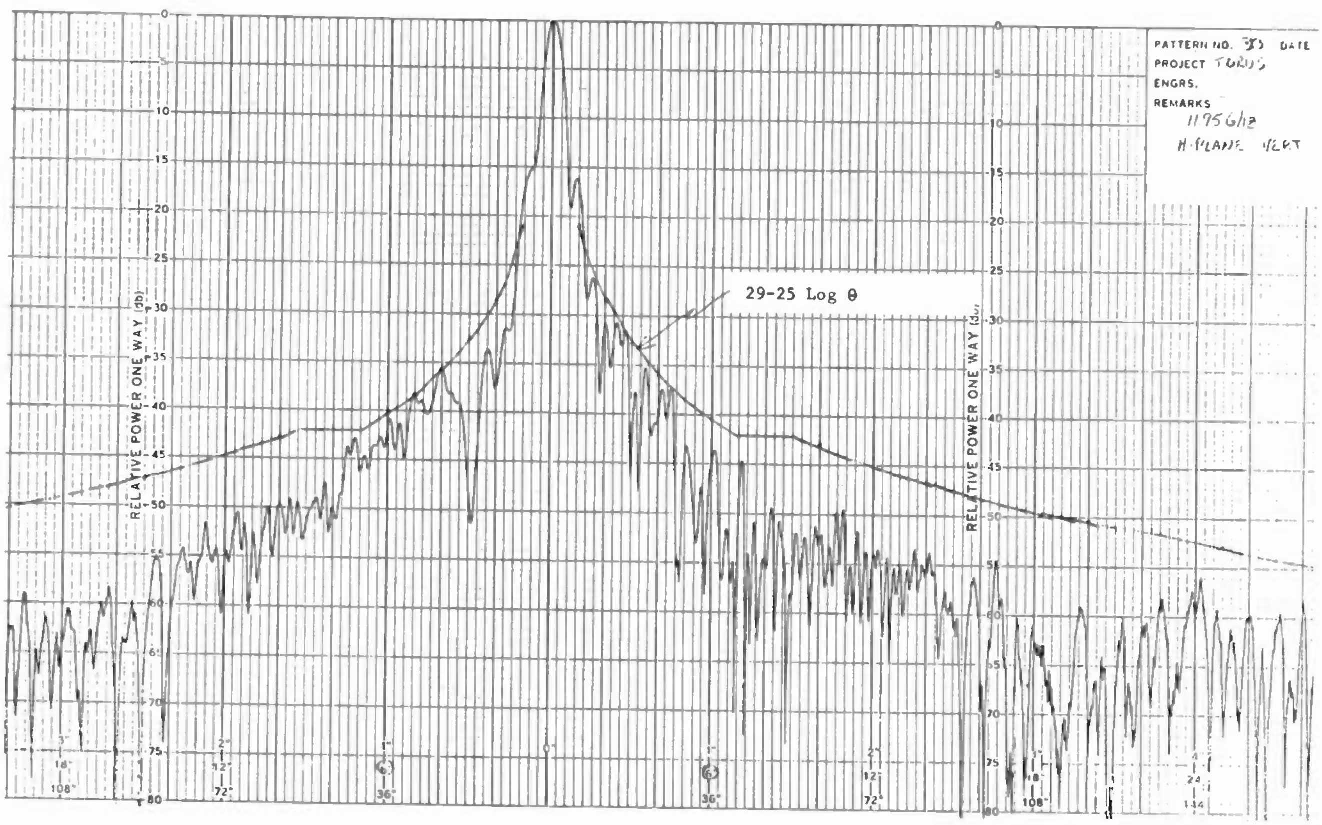


FIGURE 3. TYPICAL 4.5 METER 12 GHZ PATTERN

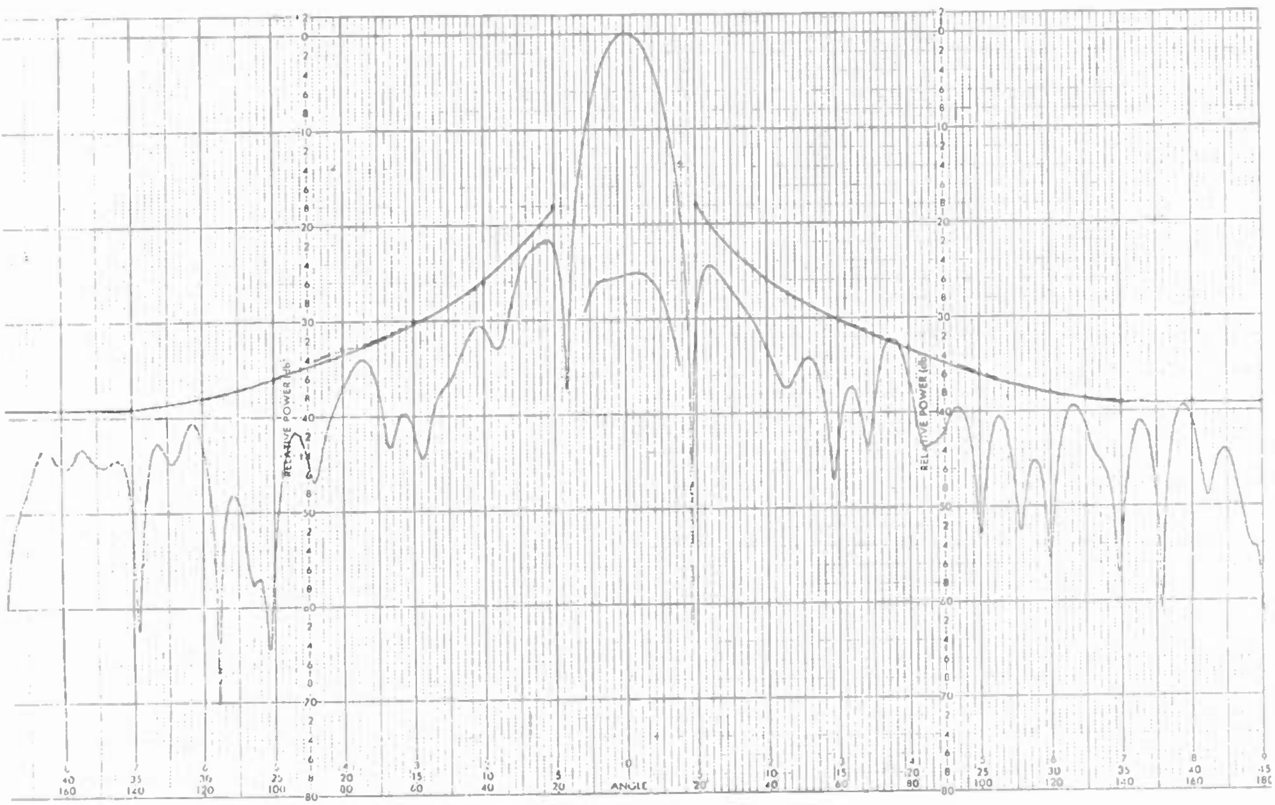


FIGURE 4. TYPICAL 7 METER TORUS 4 GHZ PATTERN

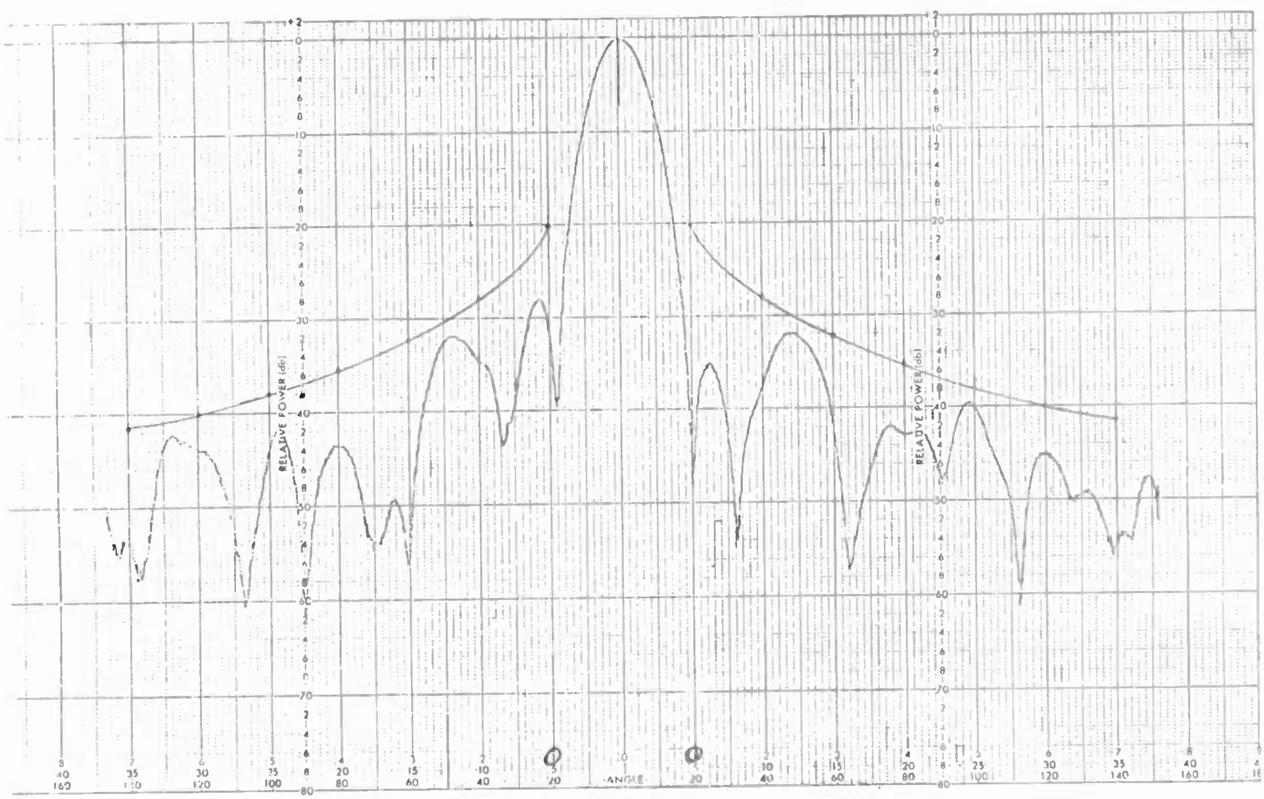


FIGURE 5. TYPICAL 7 METER TORUS 6 GHZ PATTERN

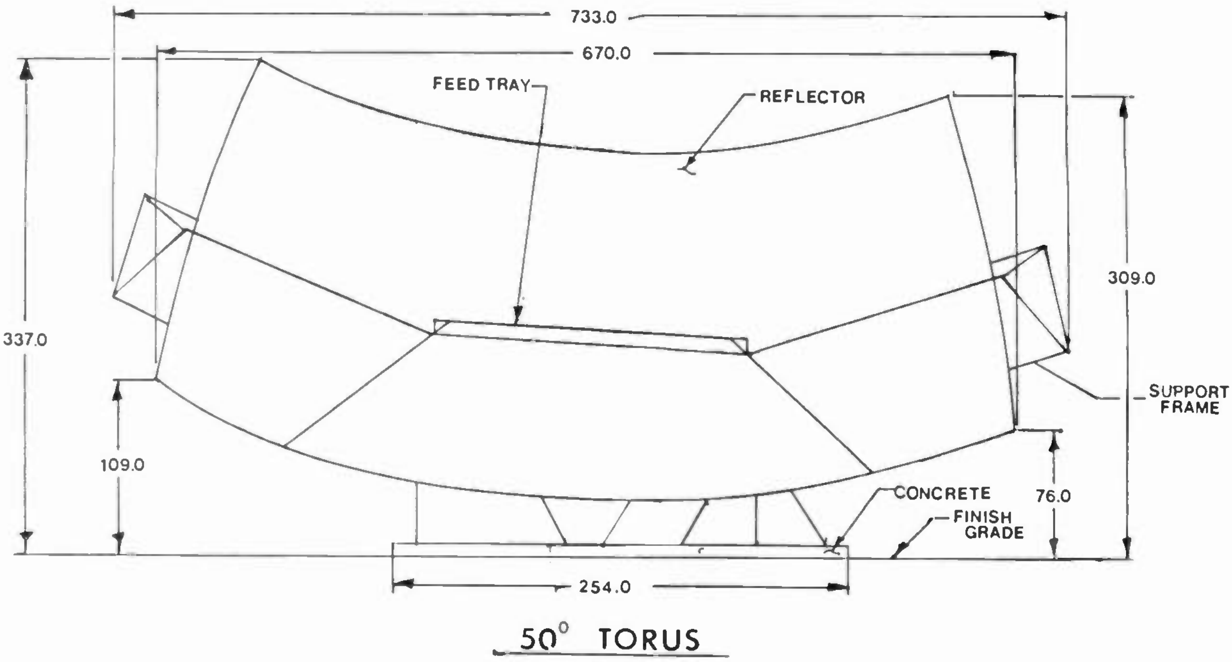


FIGURE 6. 7 METER 50° TORUS

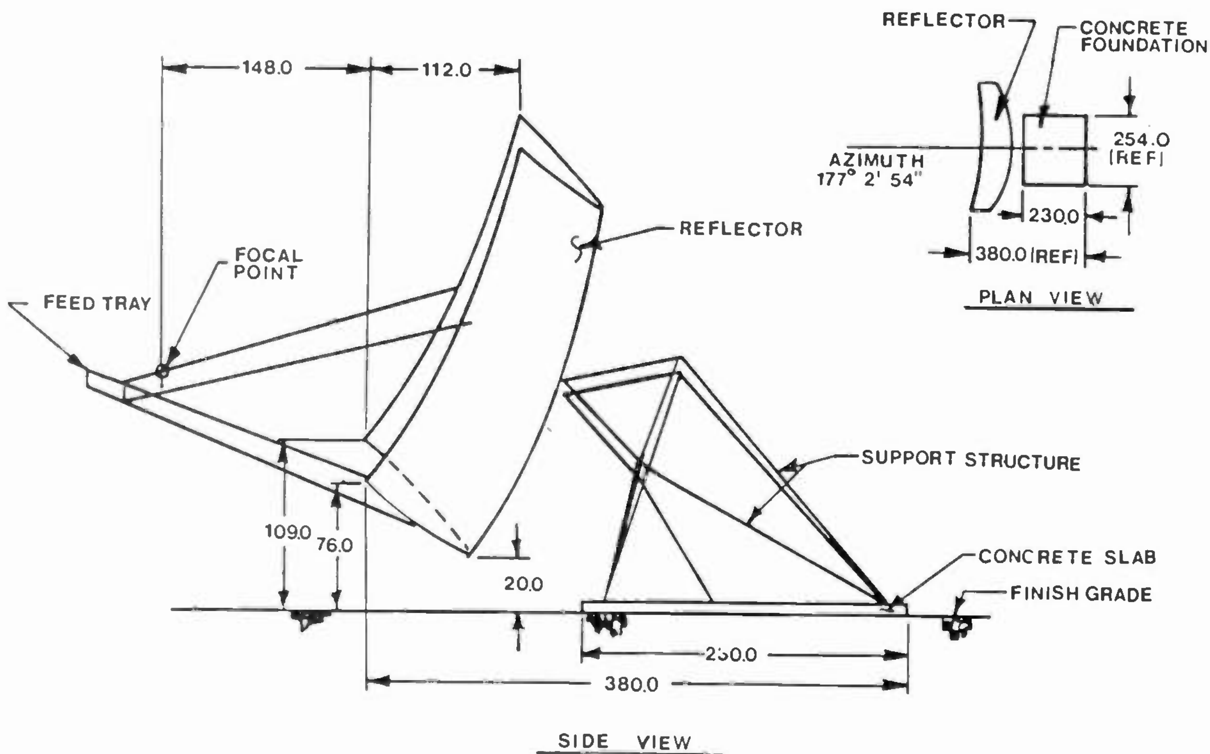


FIGURE 7. 7 METER 50° TORUS ANTENNA

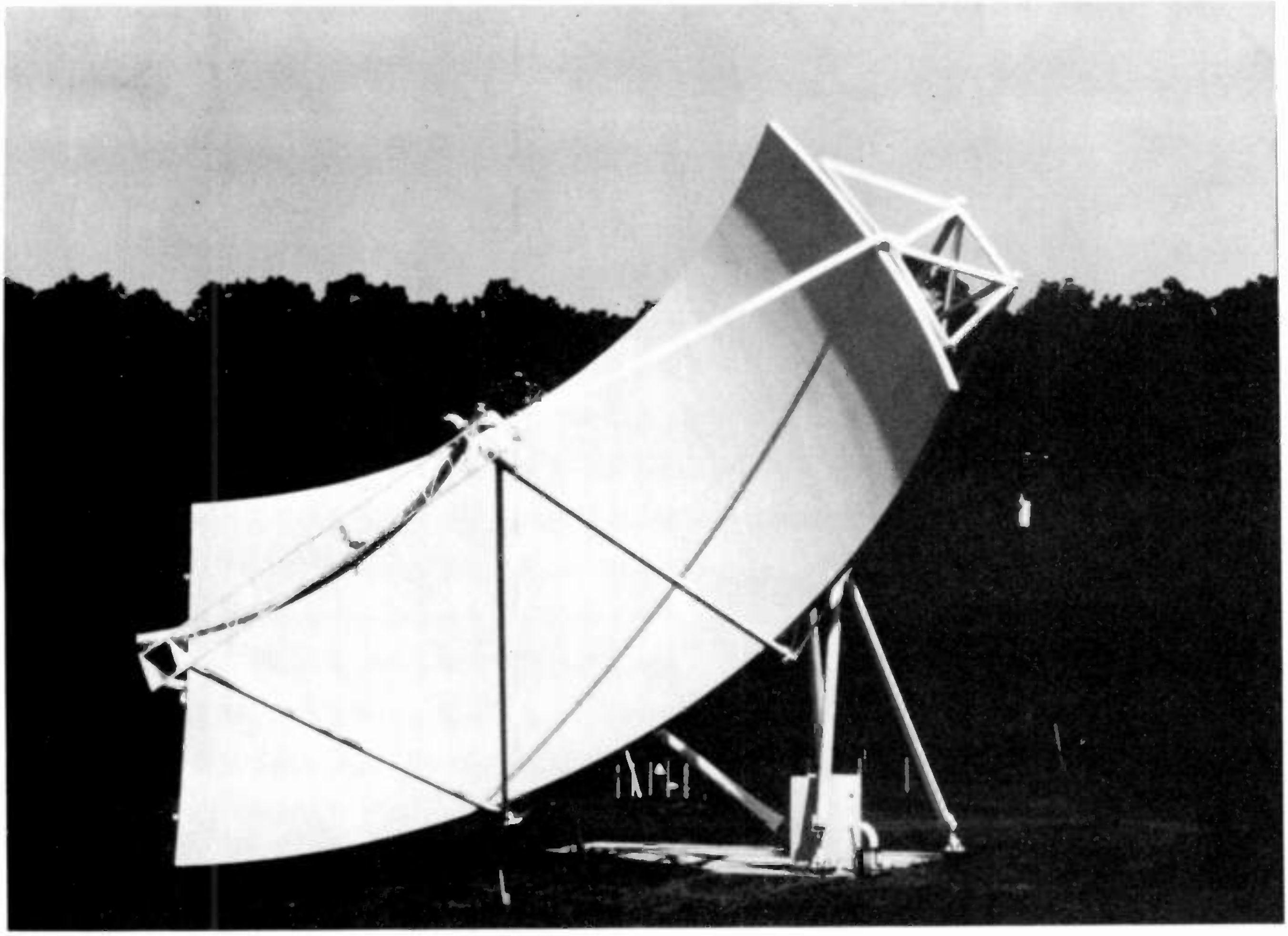


FIGURE 8. TYPICAL 4.5 METER TORUS INSTALLATION

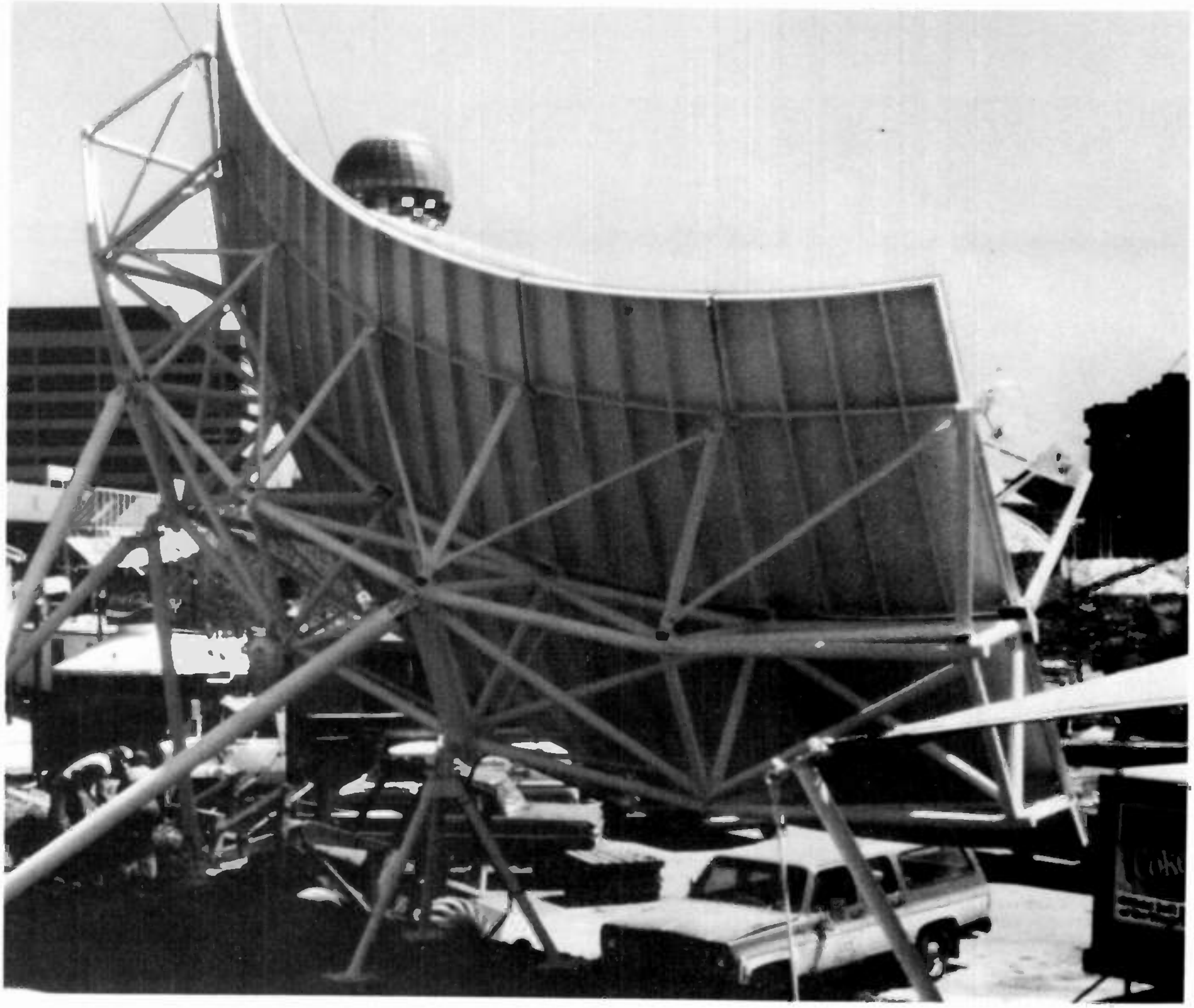


FIGURE 9. TORUS BACK STRUCTURE AND MOUNT

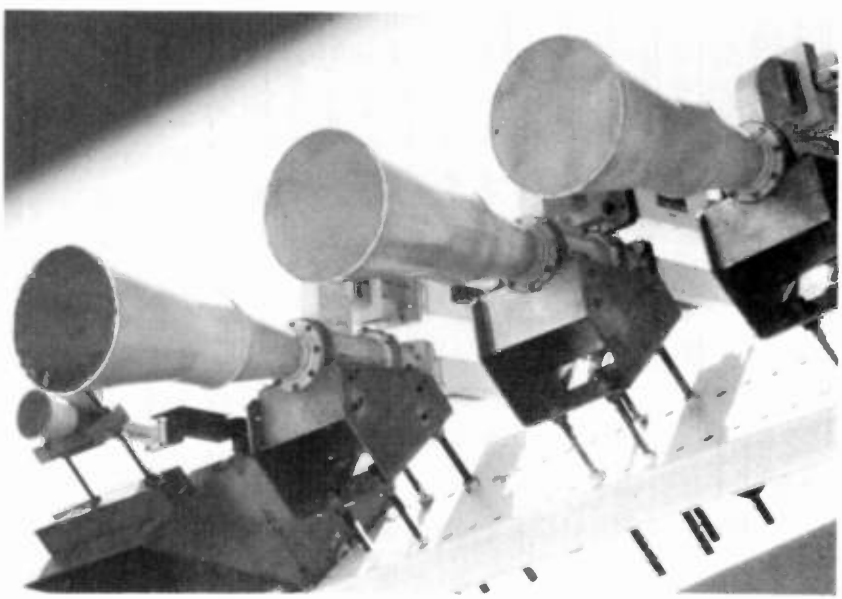


FIGURE 10. TORUS FEED MOUNTING ARRANGEMENT

The EIA Television Multichannel Sound Laboratory

Carl D. Olson

EIA Consultant

Chicago, Illinois

In 1980 the Laboratory and Field Tests (Task Force D) proposed by the Multichannel Television Sound Subcommittee (MCS) of the Broadcast Television Systems Committee (BTS) began to take form at the Matsushita Industrial Company (MIC) laboratory site (offered to the Subcommittee for their use) in Franklin Park, Illinois. The task assigned was to test the performance of the three competing proposed Television Multichannel Sound Systems and report the results for industry evaluation, which would be expected to lead eventually to a decision to recommend a single system.

The Laboratory and Field Tests took place in two separate time periods. The first testing period ran from October 1980 through October 1981, and the second from March 1983 through November 1983. The second testing period was undertaken after completion of the first test for the many reasons which are cited in the Electronic Industries Association (EIA) and National Association of Broadcasters (NAB) two volume publication (1). Certain aspects of this second test period are addressed in this paper.

Planning the Test Program

In this second phase of the testing program it was determined that a common, high quality television receiver would be incorporated into the test-bed to avoid hardware-limited test results, and to provide a receiver of sufficient versatility to explore all of the proponent systems with three modes of operation. This would allow for a complete study and fill in the gaps left in the first test phase. This was accomplished by the use of a modified commercial television demodulator (2).

It was also determined from the first test phase that the noise floor of the baseband modulated aural carrier could be improved by the use of a separate commercial FM signal generator (3).

In addition to the MCS specified tests, provisions were made so that the

National Cable Television Association (NCTA) could perform testing on cable systems, and especially on cable scrambling systems, in the presence of MCS signals.

The test program objectives were established as, a) assessment of the compatibility of the proposed system with the existing television service, and b) to determine the quality of the proposed service. The tests consisted of both objective as well as subjective tests.

Objective data was gathered by use of an automatic distortion analyzer (4), a gain-phase meter (5), a true rms meter (6) and a weighted quasi-peak meter (7). Photographs were taken with a camera (8) from an oscilloscope (9) and from spectrum analyzer displays (10).

Subjective data was gathered by playing tape recordings into the desired aural channel and observing the results of both the desired and adjacent channels of a population of television receivers, both properly tuned as well as detuned, and scoring the results compared to a normal monophonic broadcast of the same material. In the final test that was conducted the companding systems were combined with the transmission systems and select program materials were played through all system combinations. The combined systems outputs of both stereo and SAP program channels were simultaneously PCM (11) recorded for later audition, duplication, distribution and evaluation.

<u>Implementing the Test Plan</u>

It was decided by the committee that each test was to be performed on each proponent's system as close to simultaneously as possible, so that the procedures and test results would be subjected to a similar set of test conditions, which would aid in maintaining uniform test results. This required that all of the test equipment and proponents equipment must be in place at one time, and that it must be possible to switch from one transmission system to another at will. Also, that switching from test to test would require a minimum of changes in equipment configuration. While this is good in principle, the implementation of such a laboratory testing arrangement called for the construction of many switching systems to accomodate this requirement.

The basic plan to accomodate this was to isolate all proponents from each other for both high and low sides of all signal inputs and outputs. Further, all ac supply voltage lines were distributed from a common source and individually branched off to each proponent or test-bed equipment group and were individually switchable.

Inputs and outputs to the switch boxes were accomplished through isolated BNC connectors. The signal paths were switched using two-deck, ten position rotary wafer switches. Possible crosstalk between switch positions was controlled by using alternate switch positions as active positions, and using the adjacent switch positions as a grounded shield. To simplify the wiring and avoid the possibility of input/output crosstalk separate switching boxes were used for the proponents encoder input signal and for the baseband output signal. With this arrangement it was possible to achieve excellent noise floors, through this independent handling of each proponent.

Matching of the proponent's encoder inputs and output was constrained by the following signal parameters:

(a) The maximum output of the automatic distortion analyzer was 1.0v rms into a 600 ohm load.

(b) The maximum output of the automatic distortion analyzer must be reserved for the 50kHz "L only" deviation, which is 2.5dB greater than normal drive level to obtain 25kHz sum channel deviation when L = R.

(c) The signal drive level was adjusted to produce the correct drive on the lowest sensitivity encoder, and the other encoders inputs were attenuated to match this.

(d) Since the FM signal generator accomodates a wide range of input signal level, there was no difficulty in driving it from the proponent's encoder baseband output. This required that the modulation sensitivity be set for proper modulation with the lowest encoder output, and that the others be attenuated to match it.

Matching the decoders was an easy task:

(a) The output of the television demodulator was 10mv/kHz and all decoders were able to function properly with this input level.

(b) The output impedance of all of the decoder outputs was 600 ohms, and properly matched the 600 ohm passive 15kHz low pass filter, which was terminated in 600 ohms.

(c) The output signal level to properly drive the test equipment was 1.0v rms when the system was operating at rated deviation. Only one proponent's decoder outputs had to be supplied with additional attenuators to match this requirement.

Establishing Procedural Disciplines

Assurance that consistent results would be obtained necessitated the establishment of certain disciplines in the laboratory. This was particularly true because of the number of engineering personnel directly or indirectly involved in the test program, as well as the complexity of the test-bed and associated equipment, both commercial and laboratory made, which amounted to more than 100 units with different ac line voltage requirements as well as differing input/output signal levels, impedances and connector types. Associated with these complexities were a myriad of knobs, switches and controls. Often it would be necessary to conduct special investigatory tests at the end of a day, and therefore it was necessary to reestablish precisely the same base of operation for each test day.

A strict procedure was established which was followed each day. First, a warm-up time of two hours was found to be necessary to fully stabilize the operation of all of the equipment. The next step in the procedure was to test the test-bed as a monophonic system without any proponent equipment connected. This step was all accomplished by switching, and required no physical disconnection of any equipment. The test involved injection of a calibrated level 400Hz tone directly into the aural exciter input and operating it as a monophonic system. The television demodulator with its de-emphasized monophonic output was examined for amplitude, distortion and noise floors. When this spot check yielded performance that was no more than 0.5dB from the established norms, the test-bed was pronounced acceptable (actual variation from day to day was more like 0.25dB). This exceptional stability was made possible as a result of the selection of the equipment, test instruments and careful design and assembly of the test-bed. After the test-bed system check out was satisfactorily completed, then individually each proponent's encoder and decoder were tested for amplitude, distortion and noise floors for stereo operation and for SAP operation. Proponents were then required to verify that their equipment was operating up to its expected performance criteria before proceeding with the tests.

Plotting Amplitude and Distortion vs. Frequency

The automatic distortion analyzer, which normally sweeps from 20kHz to 20Hz in 60 seconds, was not immediately adaptable to the task requested by the committee of swept response curves due to the 15kHz low pass filter employed at the input to the test instrumentation. To achieve satisfactory automatic tracking it was necessary to provide an external positive voltage ramp to drive the automatic distortion analyzer from 30Hz up to 15.7kHz in 99 seconds by utilizing the ramp output of a synthesizer/function generator (12), which would initiate sweep on command.

The analog outputs of the swept frequency and distortion measurement from the automatic distortion analyzer were fed to the x and y axes of an x-y recorder (13).

Amplitude plotting was achieved by feeding the decoder signal output through the 15kHz low pass filter to a gain-phase meter and utilizing the analog output to drive the x-y recorder.

Unfortunately, the analog outputs of the automatic distortion analyzer and the gain-phase meter were not matched, so an attenuator was placed into the path of the larger (distortion) output, and adjusted to match the output of the smaller (amplitude) output, which assured that graphs were not spoiled by an inadvertant error of failing to readjust the y-axis sensitivity when plotting both amplitude and distortion on the same piece of graph paper.

Challenge of Test #7

At the time when the committee presented the idea of a Test #7, preparation of the test results for publication was occupying the major emphasis. To add this task at this time presented the most challenging aspect of the entire test program.

In this test the companding and transmission systems were to be married, and the product of this union was to provide documentary evidence of both objective as well as subjective test results for evaluation. On the surface the merging of the two systems appeared to be faced with difficulties in several ways:

(a) All of the companding system inputs were wired in-parallel.
(b) Phono connectors were used for all inputs and outputs.
(c) An additional piece of companding equipment had to be supplied by each proponent for simultaneous operation of both stereo and SAP program channels.

The timing of the test program was such that it must be completed so that audition tapes and the published report would be available in sufficient time to allow for evaluation before the December 1983 industry meeting.

The test was to include all of the transmission impairments and companding/transmission systems combinations and all three receiving modes, so that all of these could be evaluated. The total combinations exceeded 600 tests, and even a five-minute tape of each would have required several weeks to complete. The audition and comparative evaluation of the tapes and test results would have added additional time, making the task quite impossible.

The solution to this dilemma was to create a program sequence of all of the test conditions which could, in turn, be live-switched through a programmed

Photograph #1

Test #7 recording session "in-progress". Two technicians shown at their stations. One technician in screen room, not visible. One recording technician at recorders on left side, not visible.

Photograph #2

Back view showing additional wiring added to transmission system test-bed to incorporate companding system test-bed for Test #7 recordings.

sequence without interruption of the taping session This must include:

(a) Selection of program material which would exercise the combined systems through a wide variety of program materials. This had to include alternate silent periods in the program material to be matched against heavy modulation on the other program channel to evaluate crosstalk.

(b) Selection of a sequence of transmission system impairments related to the program material which would best exercise that particular transmission system impairment.

(c) Provision for a director's cueing tape which could be used to instruct the technicians to switch transmission system impairments on cue.

Two additional transmission impairments were added to the test that had not been part of the second test phase, but that had been part of the first test phase. They were:

(a) Addition of random noise to create a grade B signal reception condition.

(b) Addition of impulse noise.

In addition, a large number of peripheral equipments were added to achieve the required amplitude levels and controls as well as switching. This is partially depicted in the photographs. Photograph #1 depicts Test #7 recording "in-progress" with technicians at their stations. Photograph #2 gives some idea of what it means to be "wired for sound".

Additional tasks included removal of pre-emphasis and de-emphasis components from the transmission systems encoders and decoders, and accomplishing the live-switching task required that several new switching boxes be constructed. Live-switching of RF type transmission impairments was accomplished by means of coax relay switches and A/B switches. Video patterns were switched by a production switcher (14), and ICPM was switched by switching the visual exciter (15) quadrature corrector "on" to an adjusted five degrees, and "off" to an essentially zero degree condition.

The final master tape materials were placed on two 14 minute PCM tapes (one for stereo and the other for SAP). To provide a permanent program director's cueing tape, a cue track was recorded on the audio track of the master stereo PCM tape. This cueing track was played into a public address system, which instructed the technicians to switch transmission system impairments on cue.

The final taping amounted to forty 14 minute taping sessions, which generated 80 tapes. This included all three receiver modes, four noise reduction systems (standard pre-emphasis/de-emphasis and three companding systems), and three transmission systems and their variations.

Marrying of the companding and transmission systems test-beds into a single unit appeared to be anything but integrated when completed due to the large number of individual pieces and tended to remind one of the proverbial "house of cards". And however much we wished to improve the marriage by reconfiguring the companding system test-bed to have the advantage of isolated inputs and outputs, and to replace the overworked phono connectors, which were now intermittent, the time constraint dictated otherwise. Therefore, our only direction was to forge ahead and exercise all due care and diligence to hold things together long enough to obtain good representative tapes of all of the system combinations.

When the taping task was completed, the duplication of tapes for distribution began. Eight taping machines were employed in the duplication process, but the large quantity of tapes ordered extended the duplication process for several weeks.

Multipath Oversight

In spite of our diligent efforts to run a very tight ship, we were in for a surprise when we reviewed the effects of multipath crosstalk on the test tapes.

When the multipath (bipath) signal was set-up, great care was taken to insure that the line was equalized and that the levels and line length were carefully controlled. A strict procedure was followed, by which the bipath signal was precisely nulled at the aural carrier through adjustment of gain and line length, and then exactly 1/4 wavelength was removed from the delay path. This assured an exact 90 degree phase relationship between the desired aural carrier and the undesired delayed aural carrier. The undesired carrier was then reduced by 20dB and the test proceeded. There was substantial crosstalk generated in the bipath mode, so no particular note was made of the relationship between the desired and undesired visual carrier.

It was noted; however, when analyzing the Test #7 multipath performance as recorded on the tape that it was more severe than had been anticipated based on the results of Test #1 and #4B. The taping; however, was finished, so all that could be done was to investigate the reason.

Investigation revealed that the total line length had been increased when the RF distribution had been revised to add the impulse noise and grade B noise impairments. This change in length resulted in an increased amplitude slope around the visual carrier. This; in turn, caused more AM-to-PM conversion of the video sidebands, resulting in increased buzz. It was determined that the conditions for Test #1 and #4B were average; whereas, Test #7 conditions approached worst case. This is discussed in the supplement dated December 19, 1983.

Conclusion

There is a constant learning process involved in a program of this type, and I believe that what was learned could provide the basis for an even more refined testing process. While engineers are never satisfied, and would continue to labor ad infinitum, there is considerable satisfaction in having participated in these tests, and hope that they will serve as a valuable reference for future work.

Acknowledgment

The author wishes to thank John Landeck who functioned as co-consultant, and whose special skills and abilities contributed immeasurably to the successful completion of this project.

Appreciation is also expressed to James Gibson of the David Sarnoff Research Center, RCA, and Edmund Williams of the Science and Technology Laboratory, NAB, and to all of the proponent engineers who were so cooperative and helpful in assisting with the tests.

Special acknowledgment must be given to the MIC Staff, and particularly Robert Wolff, without whose assistance this project could not have been completed

so well.

Further acknowledgment must be expressed to all of the organizations who loaned their equipment for use in this test program.

Footnotes:

(1) "Multichannel Television Sound: The Basis for Selection of a Single Standard", by Electronic Industries Association's BTS Committee, Vol. 1-A published by the National Association of Broadcasters, November 9, 1983, Vol. 2-A published bu the Electronic Industries Association, November 5, 1983.
(2) Tektronix 1450-1, modified
(3) Boonton 103D, FM Signal Generator
(4) RE Electronics BKF 10, Automatic Distortion Analyzer
(5) Hewlett Packard 3575A, Gain-Phase Meter
(6) Hewlett Packard 3400A, RMS Meter
(7) Bruel & Kjaer 2429, Quasi-Peak Voltmeter/Psophometer
(8) Tektronix C5C, Oscilloscope Camera
(9) Tektronix 465, Oscilloscope
(10) Tektronix 7L5 and 7L14, Spectrum Analyzers
(11) Sony PCM-701ES, Digital Audio Processor, and Sony SL-2500 Betamax Videocassette Recorder
(12) Hewlett Packard 3325A, Synthesizer/Function Generator
(13) Hewlett Packard 7045B, X-Y Recorder
(14) 3-M Company 812, Production Switcher
(15) Harris Corporation, 994-8713-002, Channel 3 Visual Exciter

TV RECEIVER DESIGN OPTIONS FOR MULTICHANNEL SOUND

L. H. Hoke, Jr.

NAP Consumer Electronics Co.
A North American Philips Company
Knoxville, TN 37914

INTRODUCTION

The selection and recommendation of a standard by the EIA/BTS Multichannel Sound (MCS) Subcommittee will revitalize the marketing of television receivers. Sets now in the home will begin to become obsolete in function, thereby leading to a replacement of a complete generation of home equipment in a much shorter time than the normal wear-out. Such was the case with FM broadcast receivers in the mid-1960s. Television receiver manufacturers view this as an opportunity to sell a large quantity of units over the next few years. The rush to bring products to market is on and the timing is critical. The public will become sensitized to the need (want) for a new receiver, and with that, the proliferation of configurations and features, spurred by the fierce competition within our industry, will explode.

NEW PRODUCTS

Three distinct classes of television receiver will evolve. First, the standard monophonic, single-speaker, table model/portable will continue to dominate the 19 inch and smaller portions of the product line. These sets are predominately the second or third set in the home and, as such, are found in the kitchen and bedroom. Audio output power of 1 to 2 watts will continue to be used. Selectable second audio program (SAP) option may find its way into some areas of this product.

The second major class will be the table model with twin extension speakers. Products for home use having the outward appearance of monitors have been available for the past one or two years. Sound power is in the 1 to 5 watt area and screen sizes are typically 13 and 19 inch, with a few 25 inch models available. These models will be equipped eventually with full multichannel sound decoder to provide the customer with the choice of monophonic, stereo or SAP audio programming. This section of the market appears to be growing.

The newest class will be the balanced or symmetrical cabinet with identical speakers located to the left and right of the picture. This design concept will be applied to both table model receivers in the 19 inch range as well as the larger floor model consoles. Multichannel sound decoders will provide both stereo and SAP audio reproduction. Audio output systems in this class will range from 1 to 2 watts per channel to 10 to 12 watts,and will feature full range bass and treble controls as well as the contoured loudness control feature. Stereo separation enhancement circuits, now popular in portable AM/FM audio units called "boom boxes," may find application in table model television receivers. Enhanced stereo created by adding to each channel a negative fraction of the opposite channel, for example L = (L-0.2R) and R = (R-0.2L), can give the illusion of far greater separation and depth than that directly attributable to the distance between the speakers, which for a 19 inch table model is only 21 inches. These circuits can also provide a pseudo-stereo effect on a monophonic transmission, thereby providing a more lively program sound even when stereo is not broadcast. These products will undoubtedly be hot market items during the next few years.

In addition to new receivers equipped to receive multichannel sound, an after-market adapter business will develop. This will take two major forms. First, several manufacturers will supply "set-top" or "set-back" decoders which will convert their 1982-1984 built stereo-ready or stereo-adaptable models to full stereo performance. These manufacturers have been anticipating the stereo revolution, but could not include a decoder in the unit until the MCS subcommittee made their final system choice. In some cases, the function switching will be on the adapter box. For others, the switching has been already included in the TV receiver control panel.

A second group of after-market products will provide stereo sound or SAP with older standard monophonic TV receivers. These products will contain a television band tuner, sound detector and MCS decoder, as well as stereo amplifiers and separate twin speakers. Again, the tuner unit may occupy a set-top location, but will have no connection to the TV set except for an antenna splitter. Undoubtedly, this group will contain a multitude of configurations offering various audio power levels as well as a selection of features and services. The SAP-only, single speaker unit will likely become a specialty product in major metropolitan areas having a significant non-english speaking population.

RECEIVER SOUND SYSTEMS

Two types of sound system will be found in the broad group of MCS television receivers, ie. Nyquist intercarrier and quasi-parallel. Under normal conditions, as demonstrated by the results of the MCS Subcommittee Test 7 tapes[1], both systems can provide adequate service to the viewing and listening public. The relatively low-cost conventional intercarrier system, Figure 1, will be the most prevalent system for the table model units which have small self-contained speakers. The poorer buzz performance resulting from the Nyquist slope can be somewhat alleviated by the lack of low frequency response available from the low-cost speaker system.

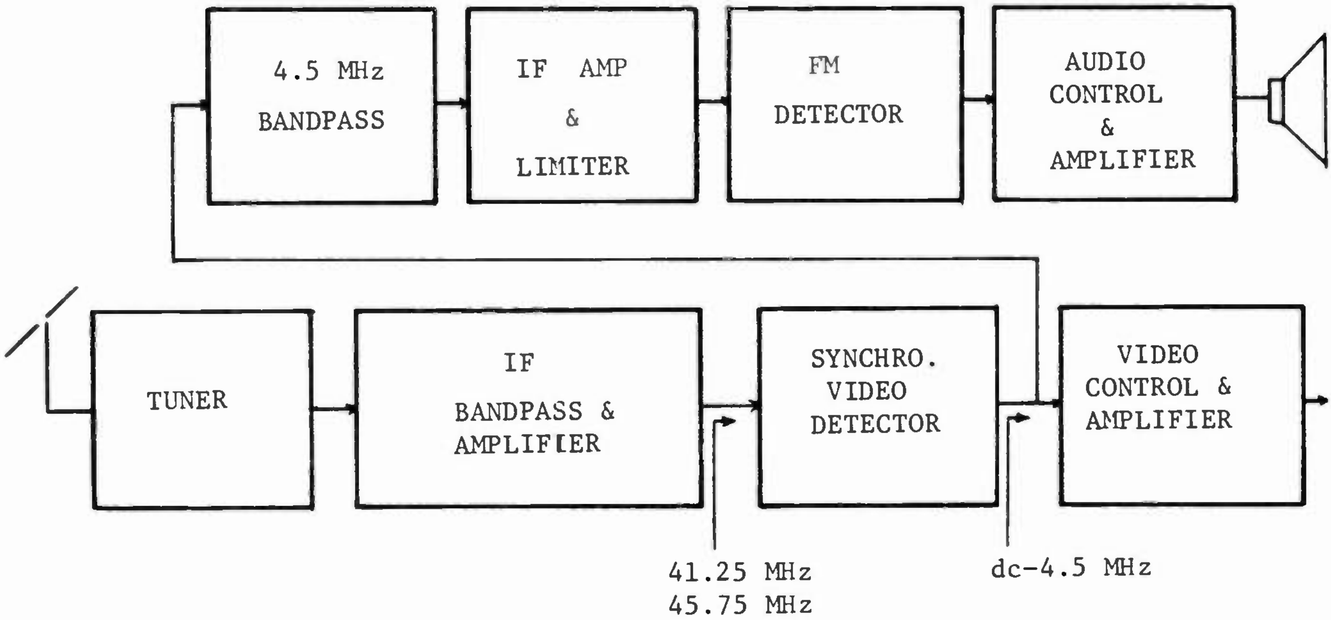


Figure 1. Conventional Intercarrier Sound System

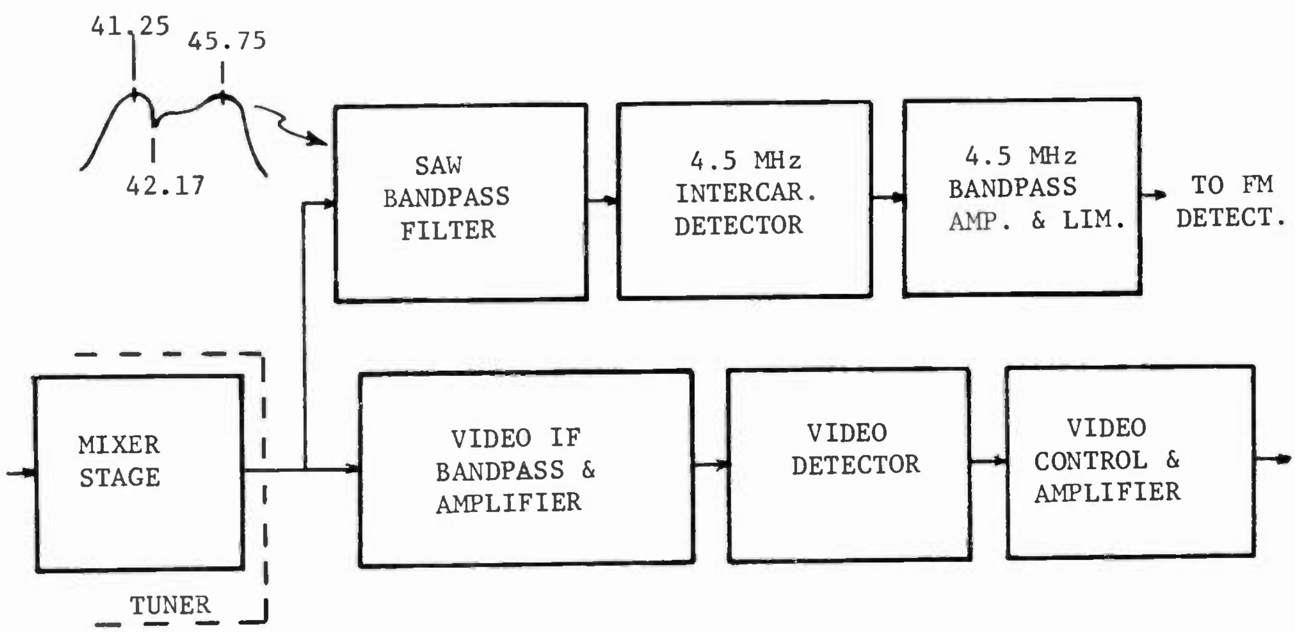


Figure 2. Quasi-Parallel Sound System

The second and newer system, the quasi-parallel, is actually a hybrid of the split-sound used in the earliest televison receivers and the conventional intercarrier system, Figure 2. Parallel paths for sound and picture carriers are created at the output of the tuner. For the sound channel, a portion of the video and sound carriers are passed through a double-peaked bandpass which has a rejection trap at the color subcarrier frequency. The object of this design approach is to ensure that sound and picture carriers are both on flat response curves so that little or no Nyquist incidental carrier phase modulation (ICPM) can occur. The design of a surface acoustic wave (SAW) IF filter for this unique application has been described by Yamada and Uematsu[2]. Following the bandpass, either a simple envelope or more complex quadrature detector creates the 4.5 MHz intercarrier signal which can then be processed in the conventional manner. Although the block diagram appears to be on the order of twice the complexity of the conventional intercarrier system, the cost in the future will decrease from its present level of 2 to 1 compared to the intercarrier. In Europe, this system has been used in their high-fidelity grade of television receivers. At least one manufacturer is marketing an integrated circuit which handles all the sound channel functions following the SAW IF filter.[3]

A third type, split-sound, possesses neither an intercarrier mixer nor a Nyquist slope, and, therefore, produces no buzz from either of these mechanisms. Data given in MCS Test 8 shows the superiority of the ideal split-sound over the standard intercarrier (Nyquist) and quasi-parallel systems when receiving signals in which the video carrier contains a few degrees of ICPM.[4] Eilers[5], however, describes several mechanisms by which this system can be degraded by buzz. In the current TV receiver environment, for example, low-cost RF modulators used with video cassette recorders, games, CATV set-top converters, accessories and internally diplexed transmitters will cause the ICPM to be transferred from the visual carrier to the aural carrier. In these cases, the performance of the split system is noticeably poorer than that of either of the intercarrier types. Unfortunately, these drawbacks are sufficiently prevalent to discourage the use of split-sound in high volume consumer receivers.

CONCLUSIONS

Multichannel sound will offer the viewer a new dimension of realism in entertainment as well as the possibility of a second language presentation. To accomplish these features, the television receiver will change character, not only in its internal systems and circuits, but also in the styling and shape of its cabinetry. Manufacturers will offer a broader line of product concepts and shapes, thus giving opportunity to the consumer to pick and choose to fill his specific need. MCS will certainly introduce a revolution in our industry.

REFERENCES

1. Multichannel Television Sound, Vol.2-A, pp. 113-124 and Supplement, pp. 48-55, Electronic Industries Association, 1983.

2. J. Yamada and M. Uematsu, "New Color TV Receiver with Composite SAW IF Filter Separating the Sound and Picture Signals," IEEE Trans. CE, Vol. CE-28, No. 3, August 1982, pp. 192-195.

3. "TDA2546 Quasi-Split-Sound Circuit with 5.5 MHz Demodulation," Handbook of Philips ICs for Video Equipment, Signetics Corporation, 1983.

4. Multichannel Television Sound, Vol. 2-A, pp. 125-145, Electronic Industries Association, 1983.

5. C. G. Eilers, "Intercarrier Buzz in Television Receivers," Multichannel Television Sound, Vol. 1-A, Electronic Industries Association, Nov. 1983, Appendix E.

The Zenith Multichannel TV Sound System

Carl G. Eilers

Zenith Radio Corporation

Glenview, Illinois

Introduction

In late 1978, the Broadcast Television System Committee of the Electronic Industries Association, on behalf of the television industry, formed a subcommittee on multichannel television sound for the purpose of formulating standards for the broadcasting and reception of multichannel television sound which was to include stereophonic as well as second program (second language, for example) enhancements of the main audio program. On December 22, 1983, the industry chose the Zenith transmission system coupled with the dbx noise reduction system and submitted the combined system, the BTSC system, to the Federal Communications Commission on January 30, 1984.

A Brief Description of the Zenith MTS System

The transmission standards are illustrated in Figure 1 and summarized in Table I.

The main channel modulation consists of an (L+R) audio signal. The pre-emphasis is 75 microseconds. The L-R audio signal is subjected to level encoding according to Figure 7 which is part of the dbx Companding System that includes complementary decoding (expansion) in the receiver. The encoded L-R signal causes double sideband, suppressed carrier amplitude modulation of a subcarrier at $2f_H$. The audio bandwidth of pre-emphasized L+R and of encoded L-R is 15 kHz.

The main channel peak deviation is 25kHz. With level encoding temporarily replaced by 75 microseconds pre-emphasis the subchannel peak deviation is 50 kHz. When L and R are statistically independent, the peak deviation of the main channel and the stereophonic subchannel combined is also 50 kHz due to the interleaving property. When L and R signals are not statistically independent or when (L+R) and (L-R) signals do not have matching pre-emphasis characteristics (as is the case when (L-R) is encoded), the combined deviation of main channel and stereo-

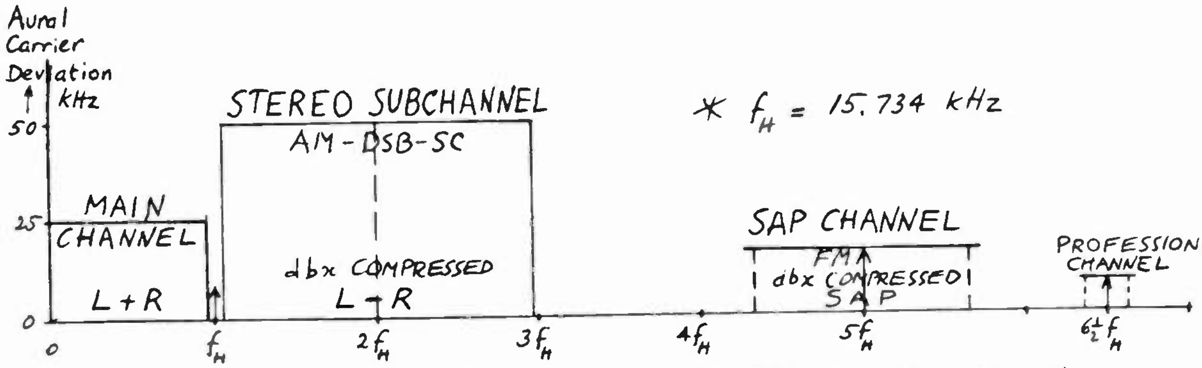


Figure 1 Stereo + SAP Service

TABLE I

AURAL CARRIER MODULATION STANDARDS							
Service or Signal	Modulating Signal	Modulating Frequency Range kHz	Audio Processing or Pre-Emphasis	Subcarrier Frequency *	Subcarrier Modulation Type	Subcarrier Deviation kHz	Aural Carrier Peak Deviation kHz
Monophonic	L + R	.05-15	75 μsec				25 †
Pilot				f_H			5
Stereophonic	L - R	.05-15	dbx Compression	$2f_H$	AM-DSB SC		50 †
Second Program		.05-10	dbx Compression	$5f_H$	FM	10	15
Professional Channel	Voice or Data	.3-3.4 0-1.5	150 μsec 0	$6\frac{1}{2}f_H$	FM FSK	3	3
						Total	73

* f_H = 15.734 kHz

† Sum does not exceed 50 kHz

phonic subchannel is constrained to 50kHz and the separate components assume their respective natural levels dictated by the acoustic scene.

Note in Table I that the listed 25 kHz main channel (monophonic) peak deviation is not part of the 73kHz total deviation due to the above mentioned constraint.

A CW pilot subcarrier signal of frequency f_H is transmitted with a main carrier deviation of 5 kHz.

The subcarrier for the SAP channel has a frequency of $5f_H$ (78.670 kHz) and is frequency locked to $5f_H$ in the absence of modulation. The SAP audio signal is subjected to level encoding identical to that of the L-R signal. The resulting SAP modulating signal is bandlimited to 10 kHz and frequency modulates the SAP subcarrier to a peak deviation of 10kHz. The main carrier deviation by this subcarrier is 15kHz.

The professional subchannel has a subcarrier located at approximately 6.5 f_H and modulates the main carrier by 3 kHz peak deviation.

Other Baseband Configurations

The foregoing description illustrated the case for a fully loaded multichannel sound baseband.

Of course, some transmissions will consist of monophonic audio with second audio program (SAP) with or without non-public subcarrier(s). This is illustrated in Figure 3.

Another base band configuration might consist of main (L+R) channel, pilot and stereophonic (L-R) subchannel with or without non-public subcarrier(s). This is illustrated in Figure 2.

The last baseband configuration, consisting of monophonic audio with or without non-public subcarrier(s) is shown in Figure 4.

Broadcast Standards

The radiated signal parameters and accompanying performance standards are summarized in Chart I. The similarity and differences between these multichannel television sound standards and the FM Broadcast Stereophonic standards may be noted.

Audio Signal Processing and dbx Companding

Audio signal processing is not new to the broadcaster and can be used optionally in conjunction with television multichannel sound source material. The dbx Companding System is, however, the mandatory noise and interference reduction companion to the Zenith transmission system and is based on complementary audio processing at the transmitter and at the receiver.

In order to obtain respectable stereo signal-to- thermal noise ratios at the Grade B contour, it was clear at an early stage in the system development that noise reduction would be a desirable feature. In order to maintain monophonic

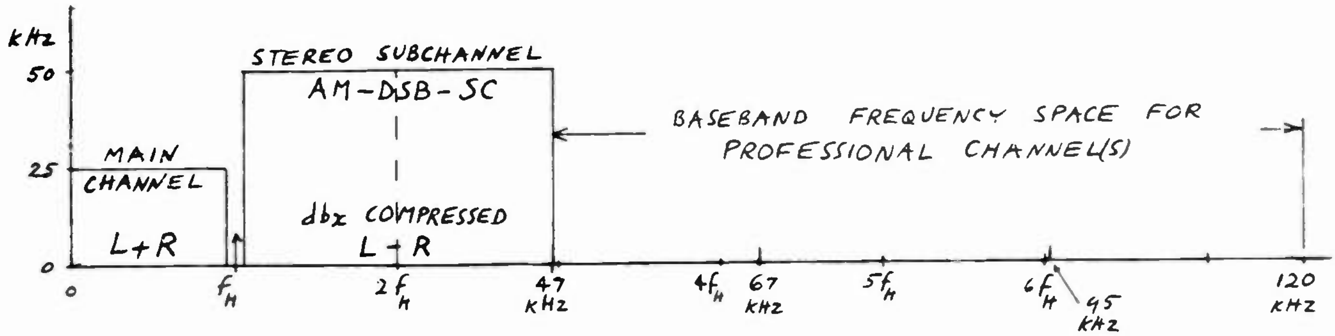


Figure 2 Stereo Service

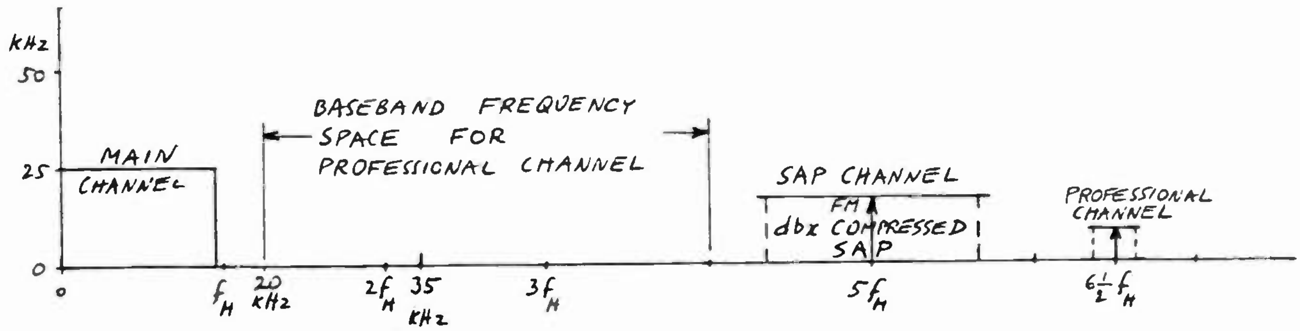


Figure 3 Mono + SAP Service

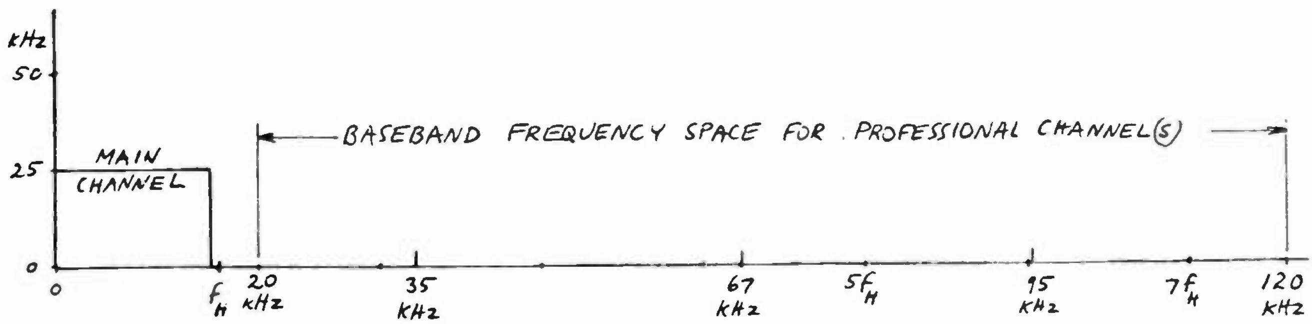


Figure 4 Mono Service

(L+R) compatibility, it was decided not to compand the main channel but to compand only (L-R). Noise reduction was accomplished because most of the noise is introduced in the subchannel. The dbx System includes a Stereo difference compressor and a separate SAP compressor on the transmitting end and a single expander for both Stereo and SAP in the receiver. Stereo separation will be influenced by the degree to which compressor and expander processing are complementary.

TV Multichannel Sound Signal Generation

A block diagram illustrating MTS composite base band signal generation is shown in Figure 5.

Stereo Generator. The Stereo Generator shown is one typical of that used in FM stereophonic broadcasting. The Stereo Generator is usually one of two types. One operates on the time-division multiplex (TDM) principle and the other on the frequency-division multiplex (FDM) principle. In the TDM type Stereo Encoder, left and right inputs are commutated to the output at the subcarrier rate. The usual subsequent bandlimiting raises the subcarrier's output compared to the main channel by a factor $4/\pi$. This increase is compensated for in a variety of ways. In one method, the left and right audio inputs are bypassed around the commutator and are added to the output via an attenuator of $(2/\pi - 1/2)$ times. One way to achieve the increased subcarrier level required for the television stereo sound standard is by reducing the main channel contribution at the output. This can be done in similar bypasses, attenuating first by a factor $(1/2 - 1/\pi)$ and then inverting before adding to the output. Another way is by controlled cross-coupling of the inputs before commutating. Note that the commutating action combines matrixing and subcarrier generation.

In a stereo Encoder, according to the FDM principle, matrixing and subcarrier generation are completely separate operations. Thus, changing levels of either main channel signal or of subcarrier or of both is a straightforward process.

The MTS Stereo Generator uses an oscillator circuit locked to horizontal pulses derived from the video modulating the television transmitter. This circuit could take on the form of a VCO (Voltage-Controlled Oscillator) in a phase-locked loop.

It is noted that the FM Broadcast Stereo Generator usually includes audio lowpass filters and pre-emphasis networks in the inputs. These units are moved forward in the processing chain to the section which includes the dbx compressor. The functions of protecting the pilot and preventing crosstalk from (L+R) to (L-R) and vice-versa are still required, however, and are performed by two Cauer filters.

The Block Diagram shows 11th order Cauer filters (for the filter circuit and characteristics see Figure 6) which have a loss pole at f_H assuring clear spectrum space for the pilot. This lowpass filter is in the dbx compressor when the compressor is active. The switching (for measurement purposes) is so arranged that when the compressor is off the Cauer filter remains in the circuit and a 75 microseconds pre-emphasis network is substituted for the compressor in the circuit. The identical filter in the L+R path assures equalized response in both paths, and thus good separation, as well as protecting the pilot and stereo subschannel from main channel crosstalk.

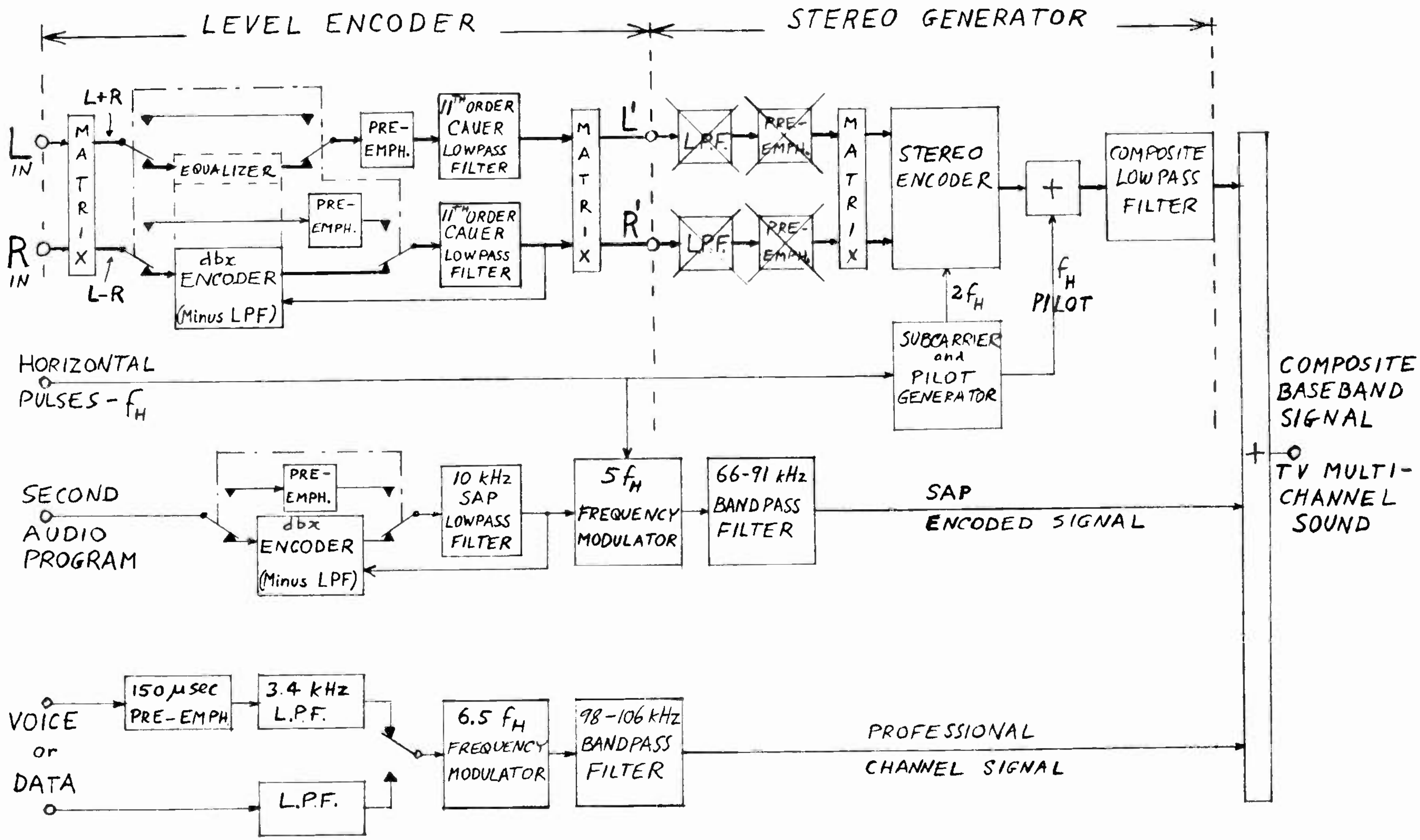
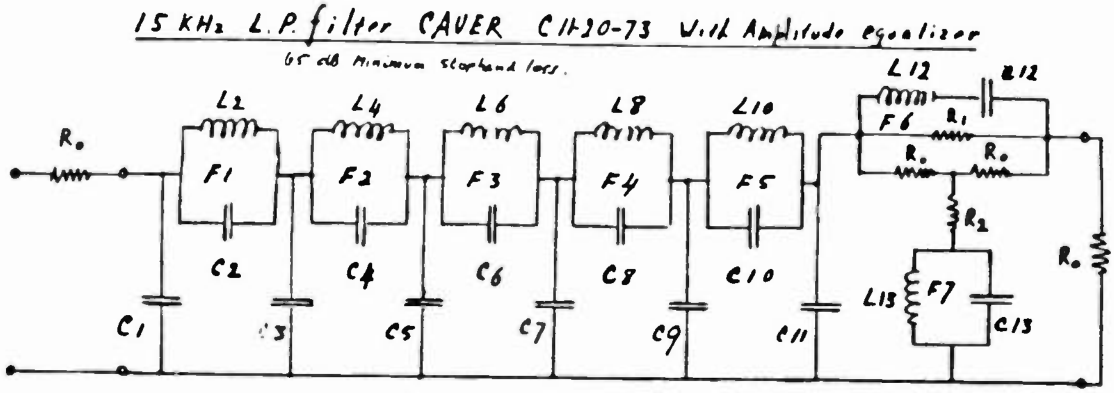
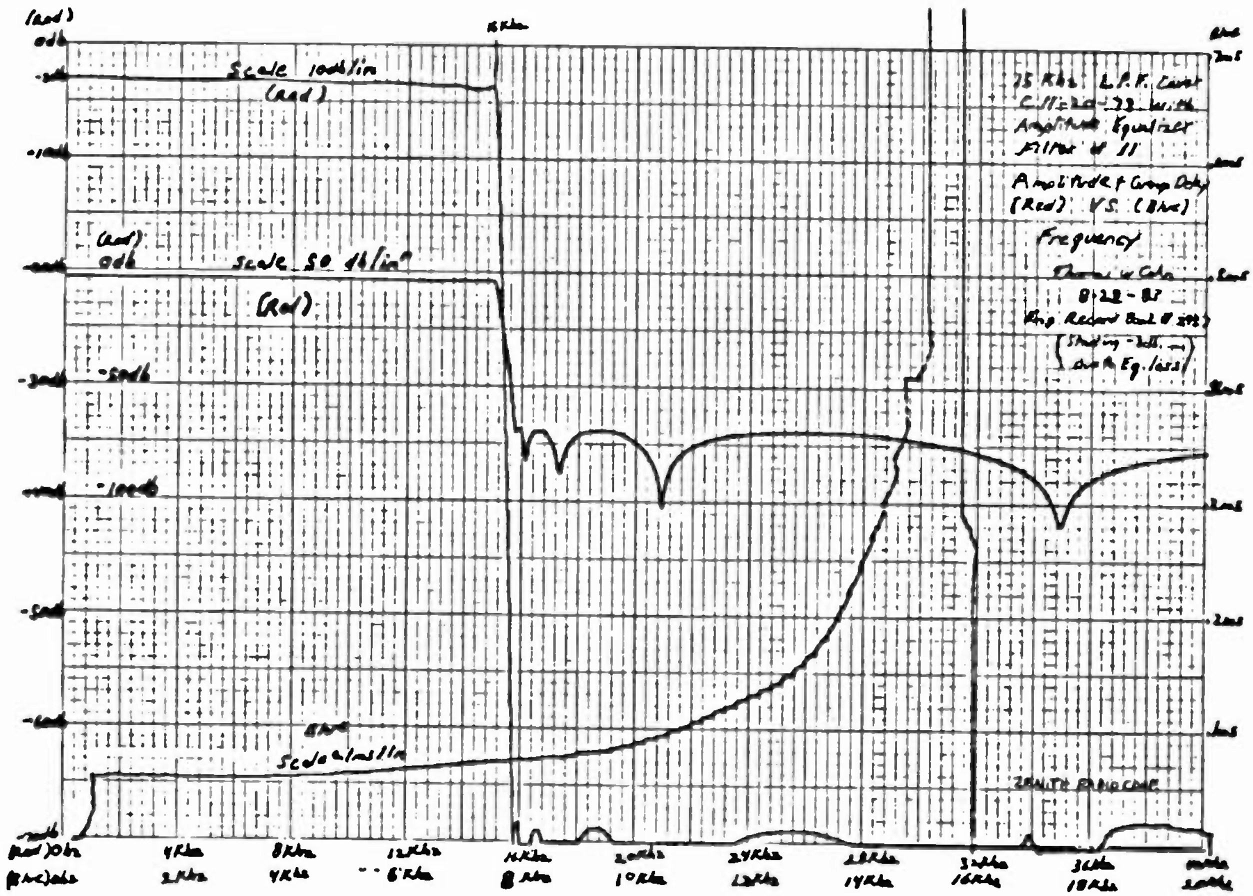


Figure 5 TELEVISION MULTICHANNEL SOUND - COMPOSITE BASEBAND SIGNAL





11th Order Cauer Low-Pass Filter

Figure 6

Level Encoder (dbx Compressor)

The Level Encoder's main component is the dbx compressor in the (L-R) path. Figure 7 shows a more detailed Block Diagram with the frequency independent compressor, the spectral compressor, and the lowpass filter arrangement (Reference [1]). That part of the Encoder response which is due to parasitic effects or to deliberate bandlimiting (mainly phase shift) is duplicated as an Equalizer in the (L+R) path, which also contains normal 75 microseconds pre-emphasis. The compressor can be switched off in which case it is replaced by another 75 microseconds pre-emphasis network. The output matrix without the compressor then delivers perfectly separated, pre-emphasized left and right signals. When the compressor is in the circuit, the dynamics of (L-R) have changed and the Level Encoder outputs, L′ and R′, are contaminated by R and L, respectively. After the matrix in the Stereo Generator, however, the required signal relationships are restored.

Some Comparisons with FM Stereophonic Broadcasting

In this section, some comparisons will be made between the Zenith television multichannel sound transmission system which includes the dbx encoding process and the familar Broadcast FM stereo system.

Both Broadcast FM Stereo and Television Stereo Sound use the (L+R) signal on the main channel; they also both use an AM, double-sideband, suppressed carrier modulated subcarrier and a pilot at half the subcarrier frequency.

The new AM subcarrier has a frequency of $2f_H$ (f_H = 15.734 kHz, the picture horizontal scanning frequency). Another difference is the modified stereophonic subcarrier injection. The L-R signal is encoded (compressed) while the L+R signal is not. This requires a different arrangement of pre-emphasis networks, lowpass filters and matrix circuits. The pilot subcarrier has increased injection. Broadcast FM stereo generator lowpass filters typically have loss poles at 19kHz. For television stereo, these filters have a loss pole at f_H = 15.734 kHz while maintaining 15kHz audio bandwidth. Second Audio Program (SAP) generators are similar to SCA generators.

Composite Filtering

The Composite Lowpass Filter at the usual Broadcast FM Stereo Generator output has a passband width of 53kHz while the bandwidth of the television composite stereo signal is 46.5 kHz. When qualitatively comparing the FM and TV services, there are some factors that tend to increase the required stopband loss of this filter, if used, while others lead to a decrease. The factors leading to an increase are: 1) the increased stereo subchannel injection; 2) the fact that SAP overlaps a larger part of the 3rd harmonic frequency band of the stereo subchannel than does the SCA; 3) the fact that SAP is protected according to the proposed rules while SCA, under the current broadcast FM rules, is not. Among the factors that tend to decrease the required stopband loss are: 1) increased SAP injection; 2) SAP with compression as compared to SCA with 150 microseconds pre-emphasis.

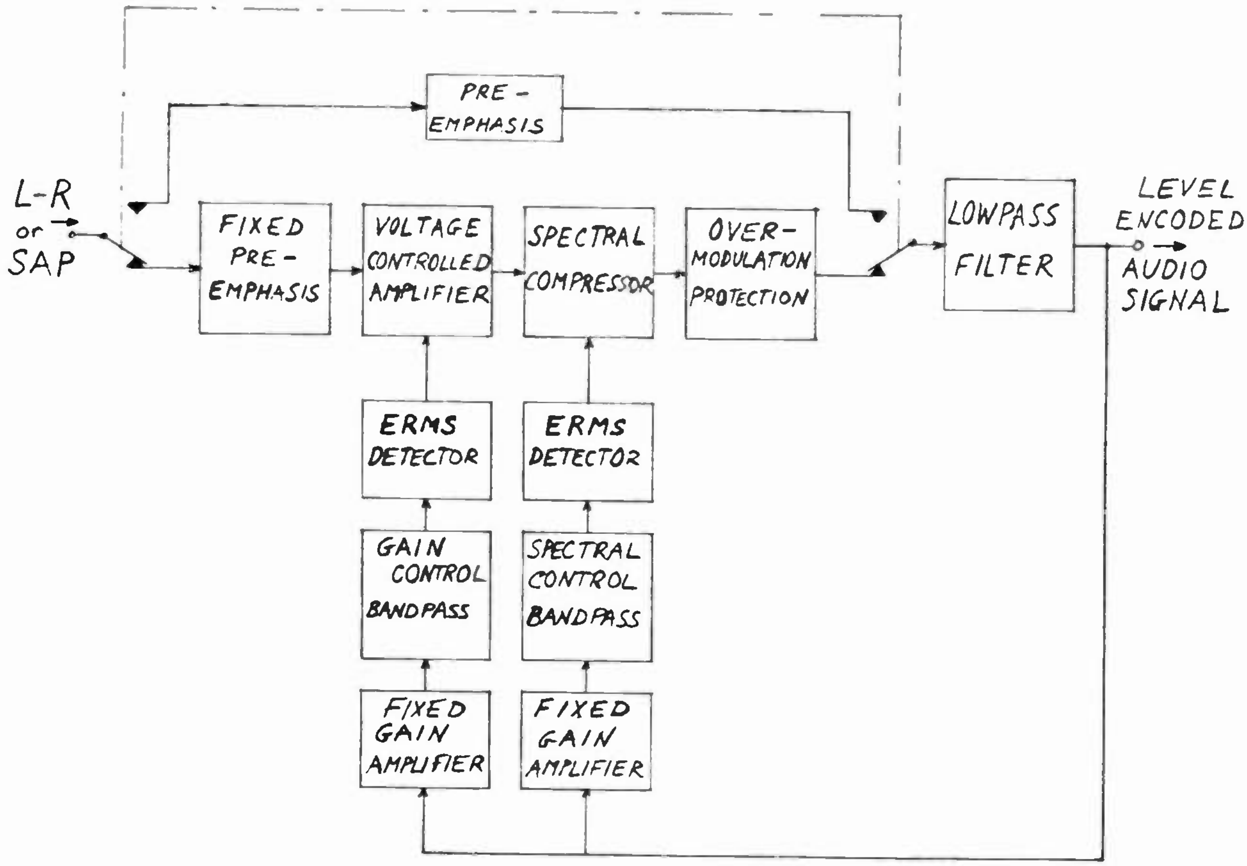


Figure 7 dbx ENCODER BLOCK DIAGRAM

Second Audio Program (SAP) Generator

Continuing the Block Diagram description with the SAP channel, it is noted that this channel also starts with a dbx Encoder. The control circuits in this unit should not be affected by spectral components above 10 kHz as defined in the standards so that appropriate filtering is required. This filtering is provided by the 10 kHz lowpass filter shown in Figure 8. The filter is the only difference between SAP and L-R Encoders. The frequency modulator has to be locked to the fifth harmonic of the horizontal deflection frequency.

While SCA Generators almost universally use 67 kHz as the subcarrier frequency, the SAP Generator must be centered at $5f_H$ = 78.67 kHz. The proposed rules include a clause "that, when modulated, the center frequency deviate maximally 500 Hz from $5f_H$." However, the same proposed rules require the SAP carrier to be locked to the 5th harmonic of horizontal scanning rate in the absence of modulation, however short that time may be, and to shut off the SAP carrier when the program ends. This necessitates a locking circuit that is active at least during pauses of modulation. The 10 kHz subcarrier deviation for SAP is higher than for SCA which is 7.5 kHz or less.

The subsequent bandpass filter's lower skirt has to protect the stereo signal which reaches up to 46.5 kHz. The upper skirt can protect the Professional channel (although the proposed rules do not require such protection).

Professional Channel

The Block Diagram includes Professional channel blocks for either Voice or Data.

Final Summation

All of these signals are added together in the final summation to form the composite signal which modulates the aural FM transmitter.

Television Multichannel Sound Decoding

Included in this section is the Decoder Block Diagram (Figure 9) based on the unit used by Zenith during the EIA Field Tests. Trap and filter details are found on the same page. Since this is a fast evolving technology, the decoder and receiver designer will have available a multiplicity of tools and circuits at his disposal, including analog and digital, and thus the diagram is shown only to illustrate the receiving process.

The FM Detector in the TV receiver or TV station monitor provides an output which is the composite baseband signal. The 4500-type Stereo Decoder does not have inherent $5f_H$ rejection in the pilot path or in the stereo subchannel path. For this reason a $5f_H$ Trap circuit precedes the Stereo Decoder. The 4500 peripheral circuits are as found in the manufacturers application note except for the outputs. The IC unit is designed for L and R outputs but L+R and L-R are needed for the companding process. In addition, no de-emphasis is wanted at the 4500 outputs. The modified output circuit is detailed on the Diagram.

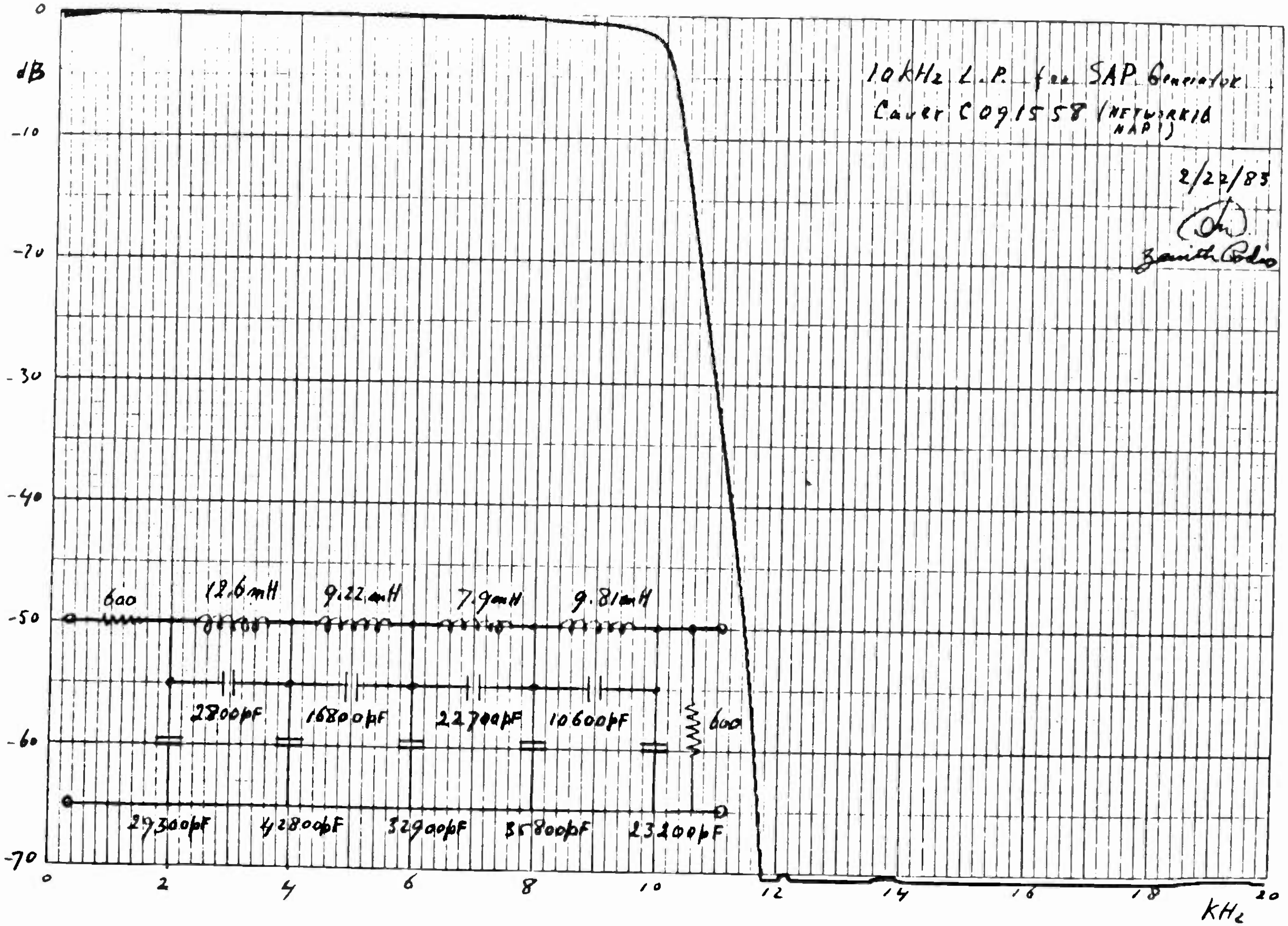
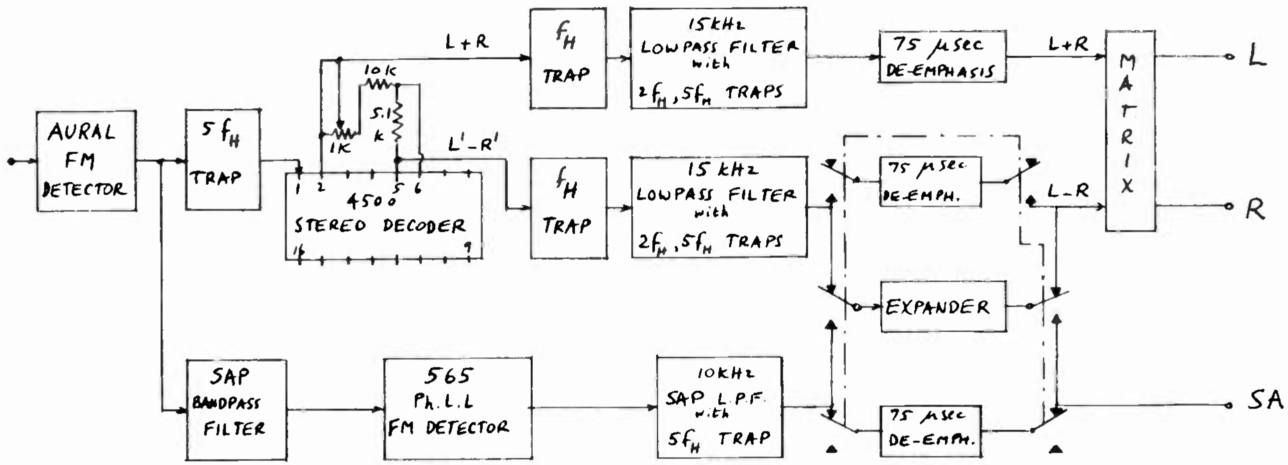
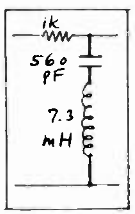


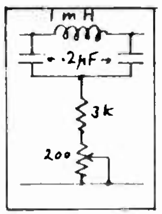
Figure 8 10 kHz SAP Low-Pass Filter

TELEVISION MULTICHANNEL SOUND — DECODER

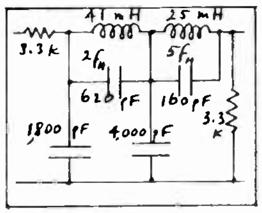




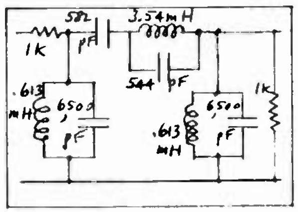
5f_H
TRAP



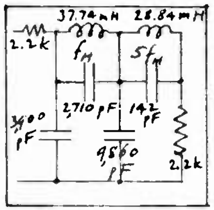
f_H
TRAP



15 kHz
LOWPASS FILTER



SAP
BANDPASS FILTER



SAP
LOWPASS FILTER

Figure 9

ZENITH RADIO CORP

The encoded audio differencesignal has to be bandlimited in order not to mislead the Expander Control circuits. The Stereo Decoder has inherent unwanted subcarrier outputs at fH, 2fH, 4fH and at the sidebands of the latter two. For this reason, the lowpass filter (Figure 10) has traps at these frequencies. If high precision, high stereo separation expansion is desired, a compensating circuit should be included in the L+R path (cascaded with the 75 microsecond de-emphasis circuit) to duplicate that part of the Expander amplitude and phase response that is due to parasitics and to deliberate bandlimiting just as is done in the Level Encoder of Figure 5. The L+R path includes a lowpass filter identical to the one in the L-R path, assuring identical response in both paths and thus good separation.

The SAP path begins with a bandpass filter. This is followed by a 565-type integrated circuit phase-locked loop FM detector. Next follows a 10 kHz lowpass filter intended to eliminate spurious signals above 10 kHz that might influence expansion control in a non-complementary manner to that of the compressor. The fixed de-emphasis constituting the last stage in the Expander keeps the spurious signals out of the audio.

The Block Diagram shows the use of a single switchable Expander for L-R and SAP. Two expanders are desirable for station monitoring - one for stereo and one for SAP.

In the case of the TV receiver, design techniques should be used to minimize intercarrier buzz since intercarrier sound is most likely going to be the choice for the consumer market. See References [2] and [3].

References

[1] "Compandor Complexity Analysis," EIA, December 12, 1983, pp 22-48.

[2] "Intercarrier buzz phenomena Analysis and cures," IEEE Transactions on Consumer Electronics, Vol. CE-27, No. 3, August, 1981, pp 398-409.

[3] "Intercarrier Buzz in Television Receivers," Proceedings, 37th NAB Engineering Conference, pp 251-266.

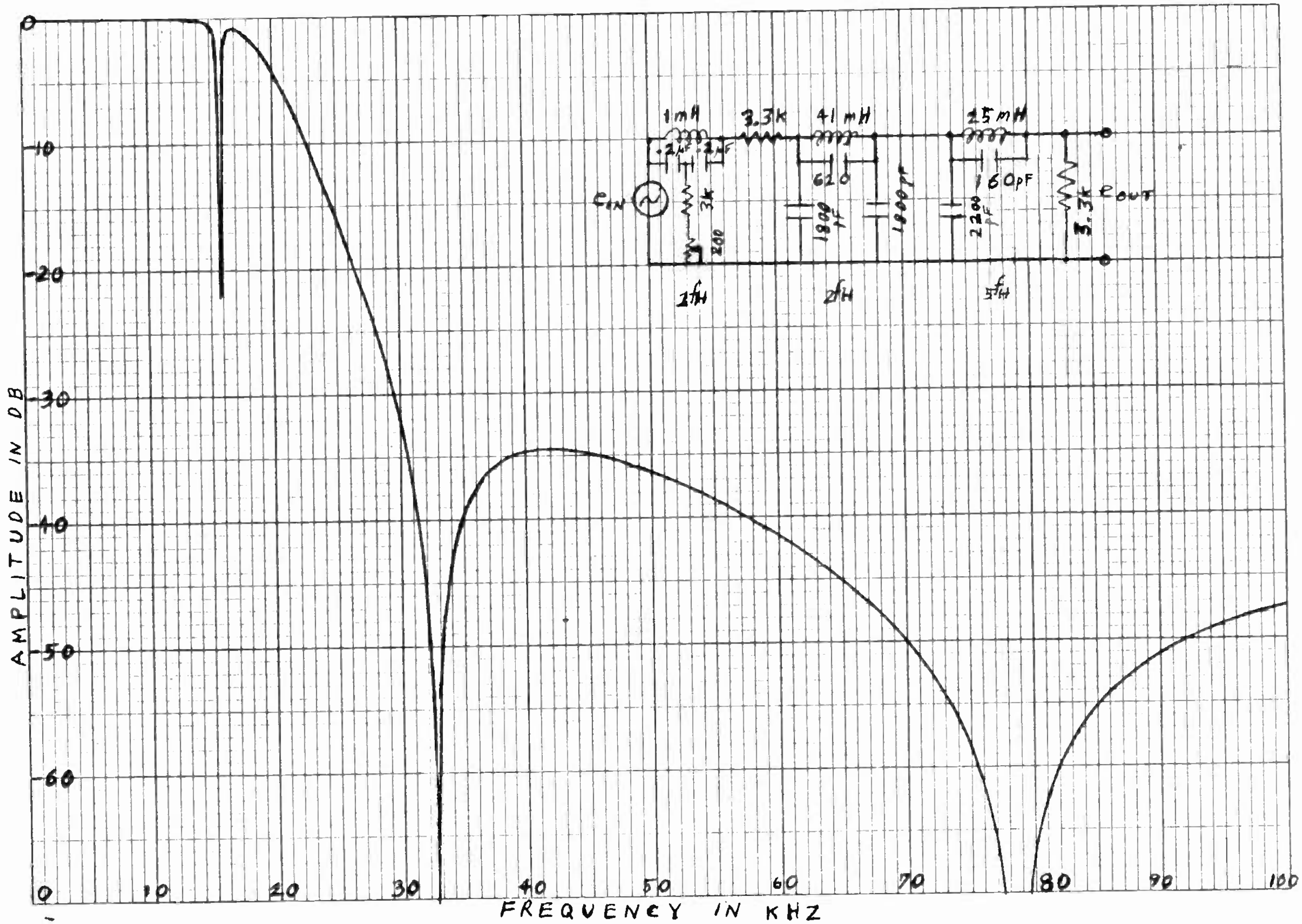


Figure 10 Stereo Decoder Low-Pass Filter

CHART I

Radiated Signal Parameters and Tolerances

Main Channel

1.	Modulating Signal		L+R
2.	Frequency Range		50 Hz - 15 kHz
3.	Pre-emphasis		75 sec
4.	Aural Carrier Deviation by Main Channel	max.	25 kHz

Pilot Subcarrier

1.	Frequency (color program)		f_H = 15.734 kHz
2.	Frequency (black-white program)(no burst)		15,734 ± 2 Hz
3.	Aural Carrier Deviation by pilot subcarrier		5 kHz
4.	Pilot-to-Interference Ratio (1,000 Hz band) (Reference 5 kHz deviation)	min.	40 dB

Stereophonic Subchannel

1.	Modulating Signal		L-R
2.	Frequency Range	(dbx encoded)	50 Hz - 15 kHz
3.	Subcarrier Frequency		2 f_H = 31.468 kHz
4.	Subcarrier Modulation Method		AM-DSB-SC
5.	Aural Carrier Deviation by modulated stereophonic subcarrier	max.	50 kHz
6.	Aural Carrier Deviation by main channel signal plus modulated stereophonic subcarrier	max.	50 kHz
7.	Stereo Subcarrier suppressed to Aural Carrier Deviation of	max.	0.25 kHz
8.	The Stereophonic subcarrier shall cross the time axis with a positive slope simultaneously with each time axis crossing of the pilot subcarrier		
9.	The difference between time axis crossings of the pilot subcarrier and of the stereophonic subcarrier (in pilot frequency degrees) is	max.	3 degrees

Composite Stereophonic Modulation

1.	Stereophonic Separation (50 Hz - 15 kHz) (no dbx Encoding)			40 dB
2.	Equivalent Input Separation (@ 10%, 75 microsec equivalent modulation):	50 - 100 Hz	min.	26 dB
		100 - 8,000 Hz	min.	30 dB
		8,000 - 15,000 Hz	min.	20 dB
3.	Total Distortion (30 kHz band; includes dbx Encoding):	50 - 100 Hz	max.	3.5%
		100 - 7,500 Hz	max.	2.5%
		7,500 - 15,000 Hz	max.	3%

4. Crosstalk of stereo subchannel signal into main channel (reference 25 kHz deviation) — max. -40 dB
   Crosstalk into main channel (reference 25 kHz deviation) — max. -60 dB
   Crosstalk of main channel signal into stereophonic subchannel (reference 50 kHz deviation) — max. -40 dB
   Crosstalk of all multiplex signals into the stereophonic subchannel (reference 50 kHz deviation) — max. -60 dB
5. FM Noise on aural carrier in main channel range (reference 25 kHz deviation) — max. -58 dB
   FM Noise on aural carrier in stereophonic subchannel range (reference 50 kHz deviation) — max. -55 dB
   AM Noise (50 - 47,000 Hz band; (reference 100% AM) — max. -50 dB

Second Audio Program Subchannel

1. Modulating Signal		(dbx Encoded)	SAP Signal
2. Frequency Range			50 Hz - 10 kHz
3. Subcarrier Frequency			$5f_H$ = 78.67 kHz
4. Subcarrier Frequency Tolerance		max.	$\pm$500 Hz
5. Subcarrier Modulation Method			FM
6. Subcarrier Deviation		max.	10 kHz
7. Aural Carrier Deviation by SAP Subcarrier		max.	15 kHz
8. Total Distortion (20 kHz band; (includes dbx Encoding)	50 - 100 Hz	max.	3.5%
	100 - 7,500 Hz	max.	7%
	7,500 - 10,000 Hz	max.	3%
9. Crosstalk from stereo into SAP (reference 10 kHz deviation)		max.	-50 dB
10. FM Noise on aural carrier in SAP Channel Range (reference 10 kHz deviation)		max.	-50 dB

dbx Encoding

1. The gain through the dbx Encoder for a: 300 Hz tone, causing 14.1% modulation, equals		0 dB
8,000 Hz tone, causing 32% modulation, equals		18.4 dB
2. The equivalent input tracking (50 Hz - 10 kHz or 15 kHz band; including equivalent input noise) is	max.	0.3 dB
3. The equivalent Encoder input noise level (10 or 15 kHz band; reference 100 Hz, 100% 75 microsec. equivalent modulation) is	max.	-70 dB

Aural Transmitter

1. Deviation range	min.	73 kHz
	recommended	100 kHz
2. Modulation bandwidth	recommended	120 kHz

Visual Transmitter

1. The amplitude of a 4.5 MHz sine wave video signal in the radiated signal referenced to the amplitude of a 200 kHz sine wave video signal in the radiated signal shall be	max.	-30 dB
2. Incidental Phase Modulation of the visual carrier (1 - 94 kHz band) for a:		
carrier amplitude between white and blanking level is	max.	3 degrees
carrier amplitude between blanking level and sync tip is	max.	5 degrees

The dbx-TV Companding System for Multichannel TV Sound

Leslie B. Tyler, Mark F. Davis, William A. Allen

dbx Inc

Newton, Massachusetts

Introduction

Without companding, the Zenith multichannel-TV-sound transmission system is capable of very high-quality transmission and reception of stereo audio. However, its full potential will never be realized without a good companding system. The reasons for this lie in the nature of the stereo television signal itself and in the practical limits to the quality (signal level, interference level, etc.) of the received signal. In order to understand these limits and how the dbx-TV system can circumvent them, this paper will first discuss the basic characteristics of the Zenith system and then explain how dbx TV works to improve the system.

The millions of mono television sets now in use make compatibility of the stereo signal with them of prime importance. Accordingly, the Zenith system transmits the sum of the left and right stereo audio signals (L+R) in the spectrum space now occupied by the mono audio signal. The stereo information is encoded by subtracting the left audio signal from the right (this is L-R) and transmitting it over a subcarrier, which will be received only by new, stereo TV sets. While an existing, mono set will ignore this subcarrier, a stereo set will use the L-R signal to reconstruct the original left and right audio signals by adding and subtracting the L+R and L-R signals.

Stereo-Noise Levels

The L-R signal is transmitted via an AM subcarrier located at 31.468 kHz (twice the video horizontal-scanning frequency of 15.734 kHz), superimposed on the present FM audio carrier. Except for the subcarrier frequency, the system is essentially the same as that used for transmitting stereo FM in the United States. FM transmission systems like these have a parabolic noise characteristic, which means that they have relatively more noise at higher frequencies.

Because the AM subcarrier is at a higher frequency than the L+R signal (the L+R signal contains only frequencies up to 15 kHz, while the modulated sub-carrier bandwidth extends from about 17 kHz to 46 kHz), the L-R signal contains much more noise when demodulated, assuming equal modulation levels.

One way to improve this situation is to increase the modulation level of the L-R subcarrier, but this technique is limited by the introduction of interference with too much modulation. In the Zenith system, 6 dB more L-R modulation is allowed than for L+R. The result is that the subcarrier adds approximately 15 dB of noise to stereo reception as compared with mono, even under ideal reception conditions.

To make matters worse, when transmission and/or reception conditions are impaired (transmitter ICPM, multipath, etc.), buzz or hum can be introduced into the audio. This would further degrade the stereo signal-to-noise ratio as compared with mono, if no companding system were used.

The significance of this added noise can be appreciated by calculating the expected S/N ratio under typical, less-than-ideal reception conditions. Grade-B reception describes a condition wherein the picture is somewhat snowy but still acceptable to most viewers. The mono S/N ratio in this case is about 65 dB, which means that most mono listeners are not bothered by the noise. In stereo, however, the situation is 15 dB worse, for an S/N ratio of 50dB. These figures refer to the difference between peak-sinewave-signal levels and unweighted noise floors after 75-microsecond deemphasis. (Without 75-microsecond deemphasis, the S/N ratio is about 43 dB.) To put this in perspective, a Philips compact cassette without noise reduction provides between 55 and 60 dB from the peak signal to the unweighted noise floor. Grade-B audio reception without companding is not even this good -- certainly not acceptable for high fidelity, and quite noisy for general music listening.

Perhaps the most important consequence of this noise is in its effect on coverage area. A TV viewer who becomes excited about stereo reception and runs out to purchase a stereo-equipped set will be very disappointed if he lives in a grade-B reception area. Imagine bringing the new set home, hooking everything up, turning on a favorite channel (one which has been touting the new stereo service) and hearing hiss that was never before audible! What's worse, the hiss disappears whan the set is switched into mono! Will the customer blame the set manufacturer, the TV station, or both? Or will he just tune to a local station which doesn't suffer from this noise?

SAP-Noise Levels

The situation is worse for the Second Audio Program (SAP) channel. Because of the location of the SAP subcarrier, at a frequency of 78.67 kHz, even more noise is introduced because of the FM parabolic noise curve. Owing to the potential for interference at this frequency, the SAP signal modulation level may not be allowed to be as large as that of the L+R signal. Furthermore, this is an FM subcarrier, which makes it additionally subject to buzz beat, an inter-modulation of the picture with the audio that causes a particularly obnoxious type of non-harmonically related distortion.

The SAP S/N ratio with 75-microsecond deemphasis is about 33 dB in grade-B reception and reaches only about 43 dB in grade A. (Without 75-microsecond deemphasis, these S/N ratios are 26 and 36 dB.) Again, these figures compare poorly even with a cassette without noise reduction.

AM-Interleave Effects

Another factor which must be considered in understanding the limits to performance of the Zenith system is that the modulation level of the composite signal (L+R audio plus L-R modulated subcarrier) is dependent on the sum of the L+R and L-R signals. Not only must the individual signal levels be limited in order to prevent overmodulation, but the sum of the two must also be limited. If full modulation of the L-R subcarrier is achieved, the L+R signal cannot simultaneously reach its full modulation level. If the noise-reduction system used with the Zenith transmission system reduces the possibility of the L-R subcarrier's reaching full modulation, the L+R signal need not be constrained as much.

Fortunately, early on in the multichannel-TV-sound evaluation program, the EIA recognized the likely limits to performance of all the proposed transmission systems (Zenith, Telesonics, and EIA-J), and sought out companding proposals that would reduce or eliminate these problems.

Design Goals

The dbx TV-noise-reduction system was designed to aid the Zenith multichannel-sound TV-transmission system in delivering a clean, noise-free audio signal into the home. dbx TV is capable of providing consistently high-quality audio in the face of the likely variety of possible channel degradations. Specifically, the system was designed to:

1) Provide significant amounts of noise reduction even in poor reception areas (grade B or worse).
2) Preserve input-signal dynamic range without headroom loss or other anomalies.
3) Prevent the stereo subcarrier from interfering with overall transmitted power levels (AM-interleave effects).
4) Ensure reliable and effective performance even in the face of severe manmade noise and transmission/reception-system impairments.
5) Provide this noise reduction at reasonable cost and simplicity.

Where To Use Noise Reduction

It was clear that the noise in the Zenith transmission system comes from the subcarriers and not from the main (L+R) audio channel. Therefore, the system was designed to work only on the L-R and SAP channels and to transmit the L+R signal without companding. This decision also answered the question of the compander's compatibility with the TV sets now in use. If no changes were made to the mono signal before transmission and the new subcarriers would not interfere with present receivers, then compatibility would be guaranteed.

Another decision that was made early on in the dbx-TV design program was to use the same compander for both the L-R and SAP channels if at all possible. This would allow TV-receiver manufacturers to produce sets at the lowest possible cost, because a single noise-reduction circuit could be switched between the L-R and SAP channel. If only one decoder must be included in a set, the implementation cost would be reduced as well.

A block diagram of a complete system embodying this philosophy is shown as Fig. 1.

In view of the limited dynamic range in the impaired channels and the ambitious goals listed above, it was necessary to depart from previous design approaches used by dbx. To understand how dbx accomplished this, it is necessary to discuss the psychoacoustic phenomenon known as masking.

Masking

All audio-noise-reduction systems work on the principle of masking. This principle can be simply stated:

> If a desired program signal (music or speech) is loud enough and broad enough in its spectral content, then the ear's attention will be captured by this signal rather than by the noise of the transmission medium.

For example, if the program consists of a single low-frequency sinewave, it must be transmitted at a very high level relative to the background noise of the stereo-subcarrier channel in order for the ear's attention to be completely captured by the low-frequency note and for the listener to be unaware of the background noise (Fig. 2). On the other hand, if the music is a raucous electric guitar, its spectrum is so broad that it does not need to be very much higher in level than the background noise for the noise to fade below the listener's perception threshold (Fig. 3). (See I. M. Young and C. H. Wenner, "Masking of White Noise by Pure Tone, Frequency-Modulated Tone, and Narrow-Band Noise," J. Acoust. Soc. Am. 41, 700-705, 1967.)

The noise-reduction system must encode (compress) the audio signal in such a way that it will consistently mask the noise of the channel during transmission and then decode (expand) the transmitted signal to recover the original audio. Of course, there should be no distortion or other degradation of the audio itself in passing through the encode/decode (companding) cycle. And in the decoding process, all the audible noise should be eliminated. As noted above, this requires:

1) The level of the transmitted audio is high relative to the background noise.
2) The spectrum must be conducive to masking.

The background-noise spectrum of the Zenith stereo subcarrier is white, having a 3-dB/octave rising characteristic, while the SAP subcarrier has a 9-dB/octave rising characteristic. Proper masking of the noise in the presence of signal will therefore take place only if the transmitted-signal spectrum contains substantial high-frequency content, especially in the case of the SAP channel. If the program material itself could be relied upon to

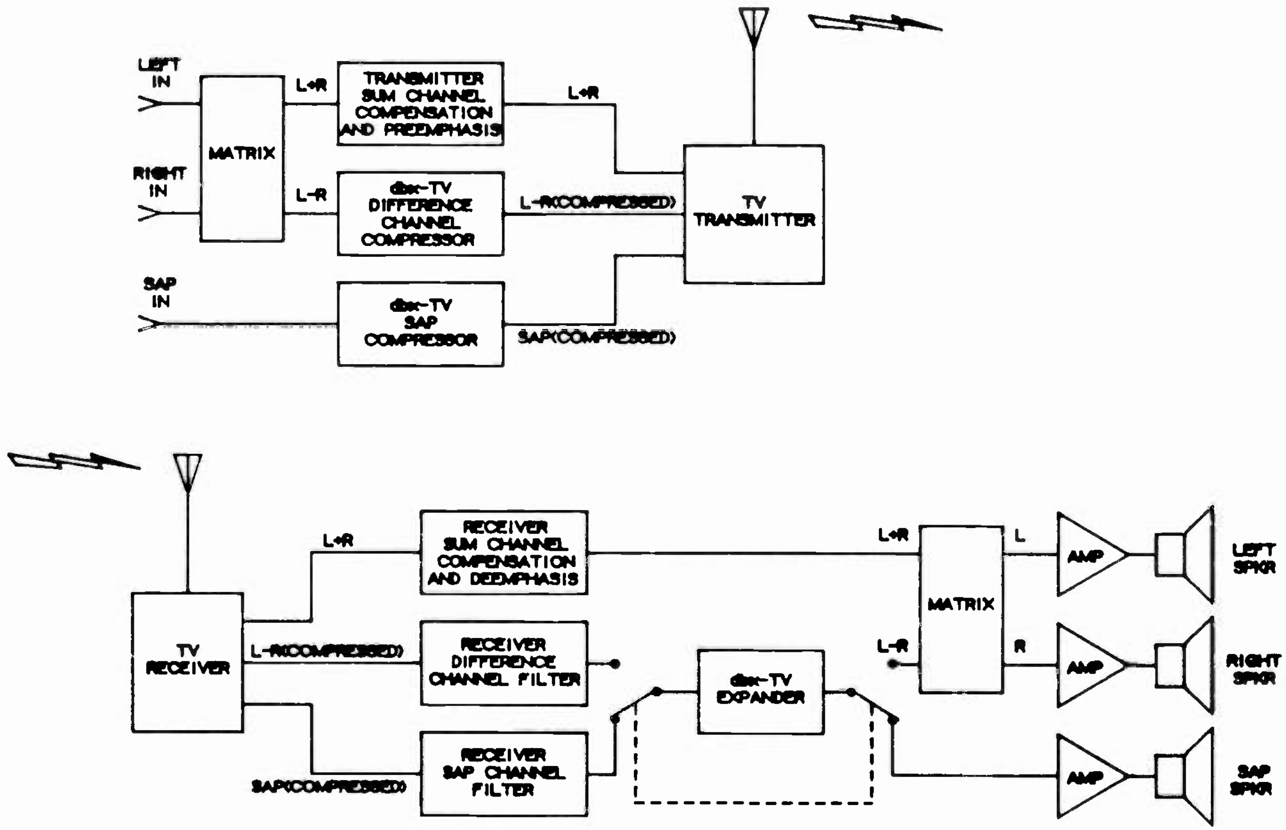


FIGURE 1
BLOCK DIAGRAM OF THE COMPLETE AUDIO SYSTEM

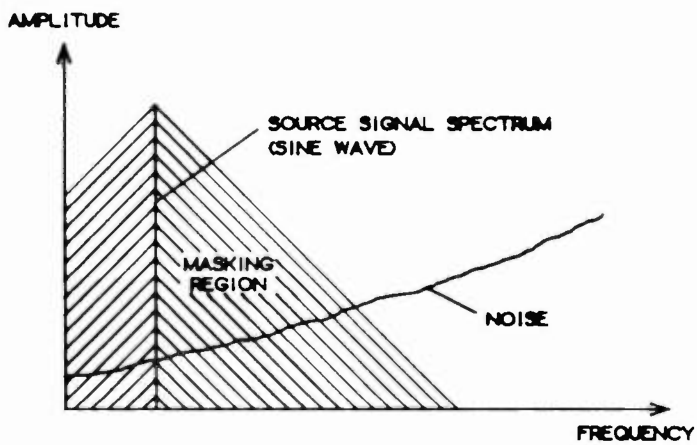


FIGURE 2
MASKING OF NOISE BY A LOW FREQUENCY TONE

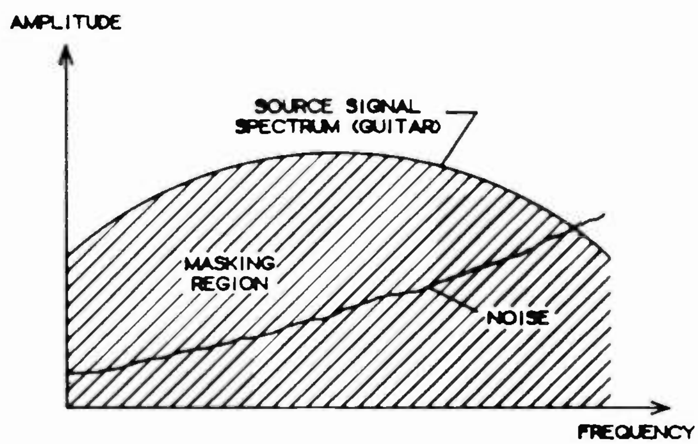


FIGURE 3
MASKING OF NOISE BY A BROAD SPECTRUM SIGNAL

have sufficient high frequencies, then the compander would only have to keep amplitude levels high through the transmission channel. However, most program material has its dominant energy at low frequencies.

Preemphasis and Deemphasis

The simplest way to transform such program material into a signal conducive to masking is to apply fixed preemphasis to the signal before transmission. This is done in present mono TV audio and in FM broadcasting by the use of 75-microsecond preemphasis. This preemphasis changes the spectrum of the average program material (containing mostly low frequencies) to be more evenly balanced between highs and lows. In the TV or FM receiver, the program signal is deemphasized, which restores correct tonal balance and reduces the audibility of the high-frequency hiss at the same time.

Fixed pre- and deemphasis is used in the dbx-TV system, too, for the same purpose. In dbx TV, however, two preemphasis networks are used. One is essentially the same as 75 microseconds (actually 72.7 microseconds). The other is 390 microseconds, but the rising frequency response created by this preemphasis is curtailed at 30 microseconds. The frequency-response curve of the complete preemphasis (Fig. 4) has a very steep section between about 2 kHz and 5.5 kHz, which helps dbx TV overcome the large amounts of noise present in grade-B reception. In the TV receiver, a corresponding deemphasis restores the correct tonal balance to the program material and reduces the audibility of the hiss picked up in transmission.

Unfortunately, audio program material is inconsistent in its spectral balance and its level. With fixed preemphasis alone, two problems would remain. First, some audio signals contain predominantly high frequencies and the preemphasis would boost them too much, causing overmodulation. This is called lack of headroom -- insufficient room for the peaks of the program to be transmitted cleanly. Second, some audio signals would be too low in level and too lacking in high frequencies to properly mask the channel noise, even with the strong preemphasis used in dbx TV.

Spectral Companding

In order to help make the audio more consistent in its spectral balance before transmission, dbx TV uses a second stage, in which the preemphasis adapts its characteristics to suit the signal. We call this spectral companding. The spectral compressor (in the encoder) monitors the spectral balance of the input signal ("How much high-frequency material is there?") and varies the high-frequency preemphasis accordingly. When very little high-frequency information is present (this is when masking is least likely), the spectral compressor provides large high-frequency preemphasis. When strong high frequencies are present (this is when overload is most likely), the spectral compressor actually provides deemphasis, thereby reducing the potential for high-frequency overload. The resulting encoded signal is therefore dynamically adjusted to consistently contain a substantial proportion of high frequencies before transmission, providing masking of the channel noise.

How Spectral Compression Works

The range of variation in the frequency-response curves that the spectral compressor can produce is substantial: from +27 dB to -27 dB at 15 kHz (Fig. 5). Because the spectral compressor follows the fixed preemphasis, the range of possible responses from the system as a whole (Fig. 6) varies from nearly flat (spectral compressor at maximum cut) to drastic high-frequency boost (+55 dB at 15 kHz).

The variable-preemphasis/deemphasis stage works by varying the gain of a voltage-controlled amplifier (VCA) embedded in a frequency-selective network (Fig. 7). When the VCA gain is low, no signal appears at point B. The only transfer from the imput to the output must go from point C to the output, which passes through the deemphasis network. This attenuates high frequencies.

On the other hand, when the VCA gain is high, the response between point A and the output is essentially flat (this is analogous to an op amp with a closed feedback loop). The signal at point C is small compared to the signal at point B (due to the high VCA gain), so the output signal is essentially the same as that at point A, which is the input signal after preemphasis. This provides gain to high frequencies.

At intermediate VCA-gain settings, the response varies smoothly between these two extremes, with flat occurring at unity (0-dB) gain.

The VCA gain is controlled by an rms-level detector sensing the amount of high-frequency energy at the output of the compressor (Fig. 8). The bandpass filter center frequency is about 10 kHz (Fig. 9). An rms detector is used because of its unusual combination of properties, such as relative insensitivity to phase shifts in transmission and a unique blend of signal-dependent time constants governing its acquisition and release behavior. A complete description of the rms-level detector is beyond the scope of this paper.

When high levels of high-frequency energy are detected, the VCA gain is high, and deemphasis results, reducing high frequencies. When low levels of high-frequency energy are detected, the VCA gain is low, and preemphasis results, boosting high frequencies. In effect, the range of spectral variation at the output is reduced, "compressing" the spectral dynamic range.

Spectral Expansion

During reception, the spectral expander (in the decoder) will restore the high frequencies to their proper amplitude. If the original input signal contained predominantly low frequencies, the decoder will attenuate the high-frequency background noise, leaving the low-frequency signal and only the low-frequency background noise, which will be masked by the signal. If the original input signal contains predominantly high frequencies, the decoder will not need to attenuate the high frequencies to restore correct frequency response; in this case, the signal itself masks the noise.

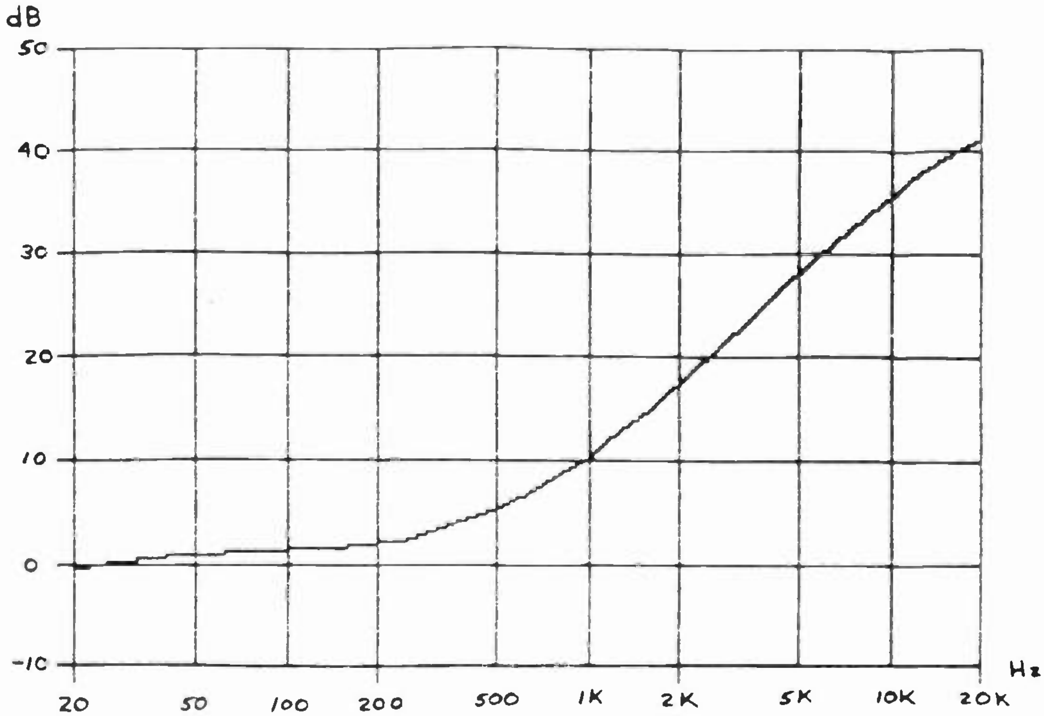


FIGURE 4
FIXED PRE-EMPHASIS FREQUENCY RESPONSE

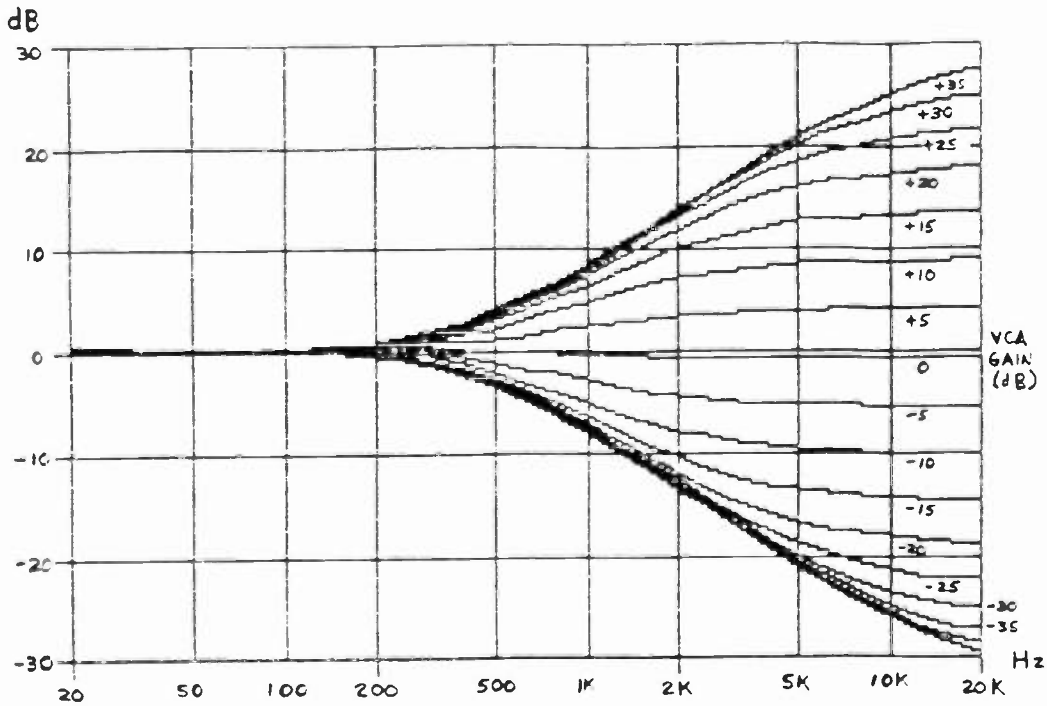


FIGURE 5
SPECTRAL COMPRESSOR FREQUENCY RESPONSE RANGE

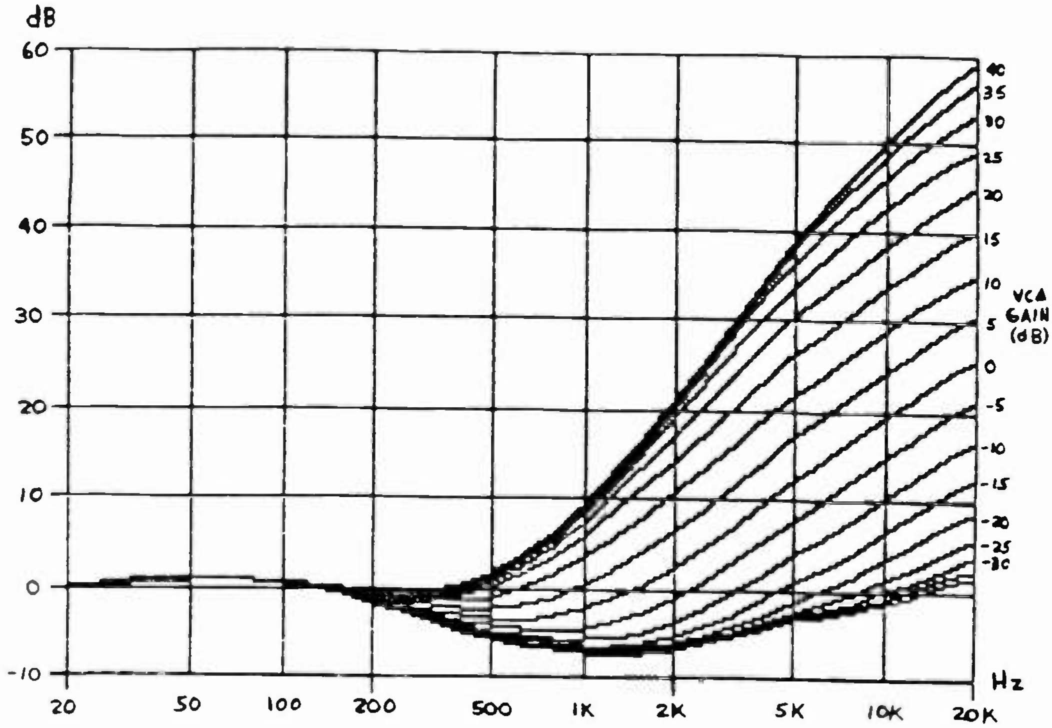


FIGURE 6
FREQUENCY RESPONSE RANGE OF SPECTRAL COMPRESSOR AND FIXED PREEMPHASIS

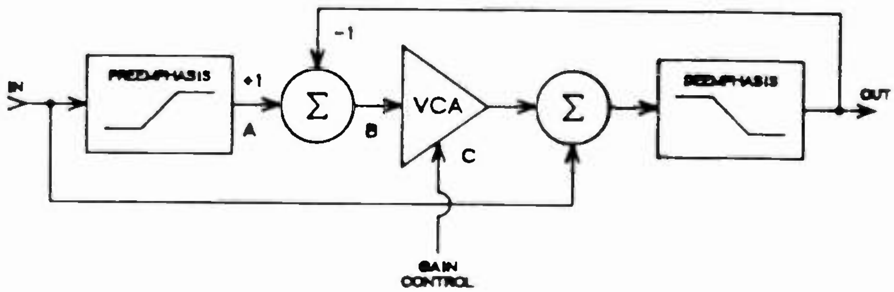
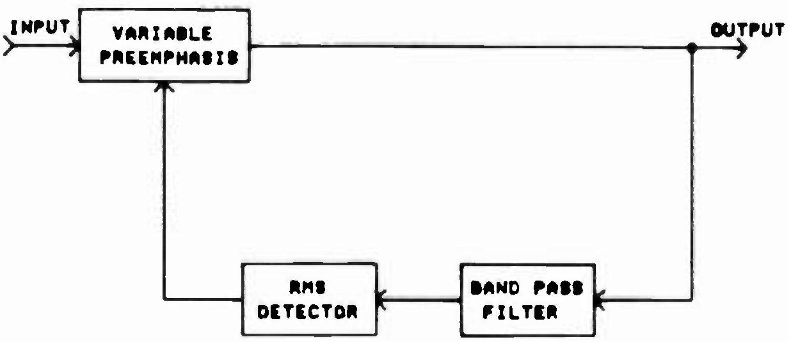


FIGURE 7
VARIABLE PREEMPHASIS/DEEMPHASIS NETWORK



SPECTRAL COMPRESSOR
FIGURE 8

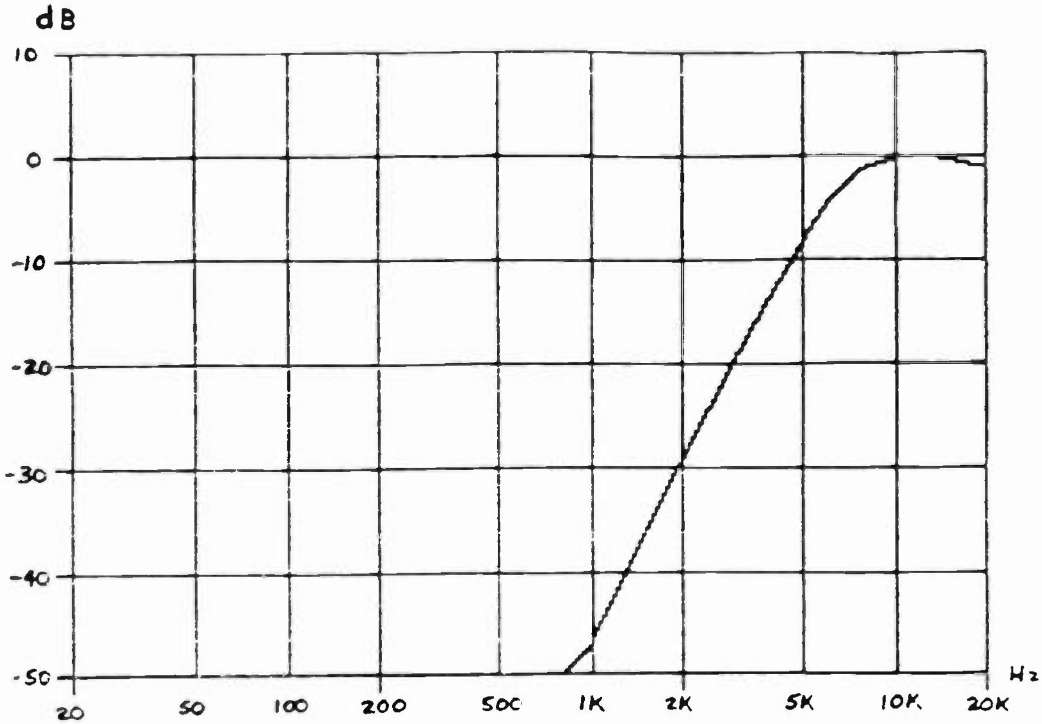
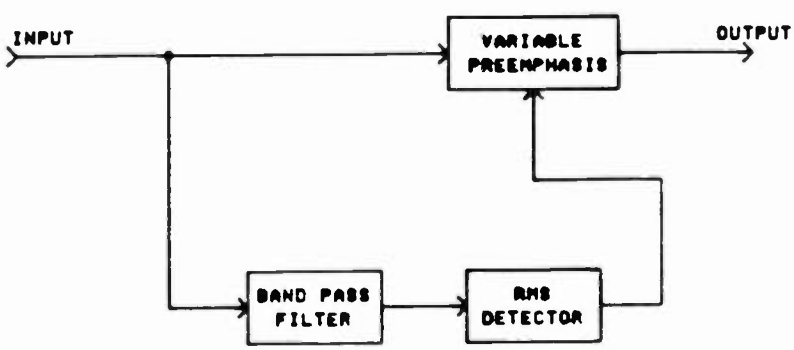


FIGURE 9
SPECTRAL COMPANDER RMS DETECTOR FILTER
FREQUENCY RESPONSE



SPECTRAL EXPANDER

FIGURE 10

The spectral expander (Fig. 10) is a mirror image of the spectral compressor. The same variable preemphasis/deemphasis network is used, controlled by an identical rms-level detector sensing energy in the same passband. Only the control polarity is changed. In the expander, high levels of high frequencies cause preemphasis, while low levels cause deemphasis. Since the input signal to the expander is identical to the output signal from the compressor (except for noise), the rms-level detector monitors the same signal in each case. This ensures that the expander exactly mirrors the action of the compressor, thereby maintaining audio transparency.

By using the spectral compressor, two simultaneous requirements are met:

1) The system is extremely forgiving of high-background-noise environments because the spectral shaping of the input signal is adjusted according to the needs of the input signal to provide high masking at all times.
2) Headroom is maintained throughout the frequency range because extreme preemphasis is used only when it's really needed. (The dbx-TV high-frequency-overload characteristic is quite close to the 75-microsecond preemphasis characteristic used in the mono -- L+R -- channel.)

Wideband Amplitude Companding

Neither the spectral compander nor the fixed pre-/deemphasis will help reduce noise when the signal is very low in level, especially if the signal has little high-frequency content. This is where the third stage of dbx TV becomes important: the wideband compander. This element is responsible for adjusting the level of all frequencies simultaneously to keep the signal level in the transmission channel high at all times.

The wideband compressor reduces the dynamic range of input signals by a factor of 2:1 in dB. Not only are small signals raised in level but large signals are reduced (Fig. 11). The output level tends toward about 14% modulation (an "unaffected level" point of -17 dB), which has three benefits:

1) The signal is consistently above the noise floor.
2) The signal it is consistently below 100% modulation, reducing AM-interleaving effects (see page 8).
3) Headroom is maintained for transient peaks to overshoot the nominal level at the compressor output without causing overmodulation.

How the Wideband Amplitude Compressor Works

The wideband amplitude compressor works by controlling the gain of a VCA in response to an rms-level detector, which senses the low- and mid-frequency energy level at the output of the compressor (Fig. 12). The filter restricting the rms-level sensor rolls off at about 35 Hz and 2.1 kHz, thereby limiting the sensed energy to the dominant energy in most program material. An rms-level detector is used here for the same reasons as in the spectral compressor, although the time constants are slower in the wideband amplitude compressor. Because both the rms-detector output and the VCA gain-control input are linear in dB, the combination provides precisely "decilinear" response (Fig. 13).

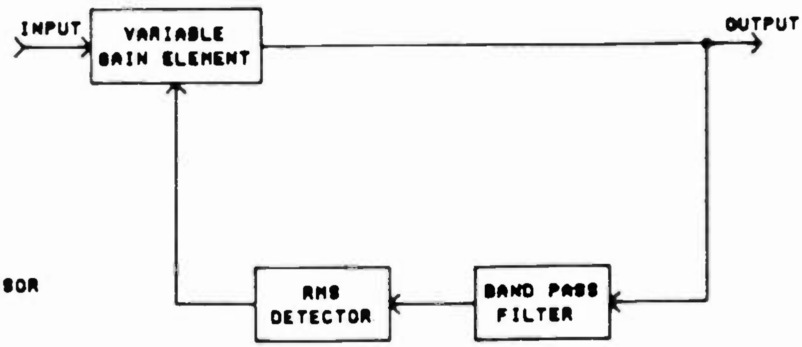


FIGURE 12
WIDEBAND AMPLITUDE COMPRESSOR

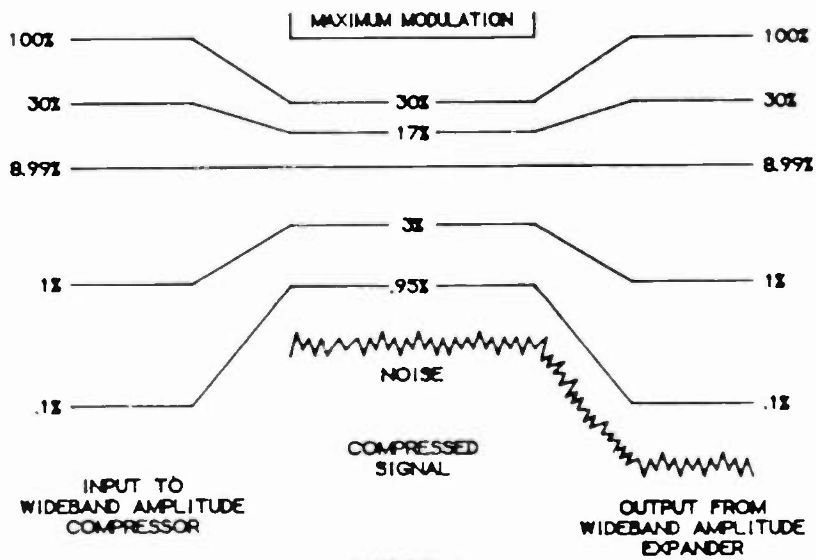


FIGURE 11
WIDEBAND AMPLITUDE COMPRESSOR/EXPANDER RESPONSE

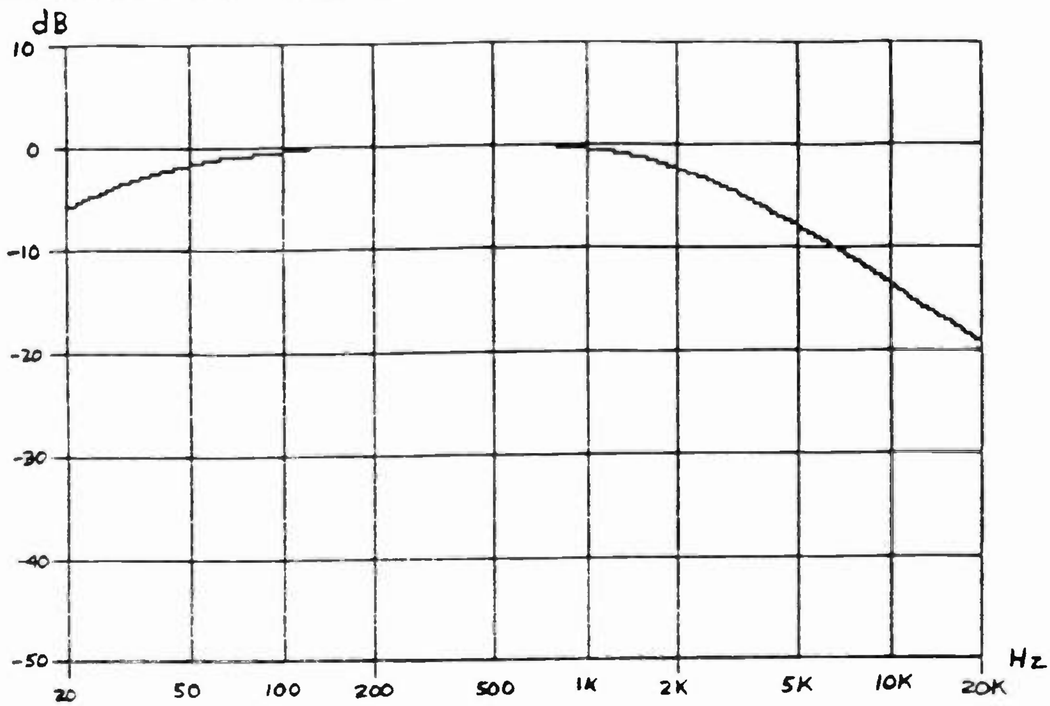


FIGURE 13 WIDEBAND COMPANDER RMS DETECTOR FILTER FREQUENCY RESPONSE

The Wideband Amplitude Expander

During reception, the wideband amplitude expander (in the decoder) will restore the signal levels to their proper amplitude. If the signal in the transmission channel is high in level, the decoder will increase the level further, restoring the signal to its original amplitude. If the signal in the transmission channel is low in level, the decoder will attenuate it further, restoring it to its original amplitude and pushing channel noise down toward inaudibility. When there is only noise in the transmission channel, it is sensed as a small signal, which is strongly attenuated by the decoder, making it altogether inaudible.

The wideband amplitude expander (Fig. 14) is a mirror image of the wideband amplitude compressor. The same VCA is used, controlled by an identical rms-level detector sensing energy in the same passband. As in the spectral compander, only the control polarity is changed. In the wideband amplitude expander, high signal levels cause gain while low levels cause attenuation. Since the input signal to the expander is identical to the output signal from the compressor (except for noise), the rms-level detector monitors the same signal in each case. This ensures that the expander exactly mirrors the action of the compressor, thereby maintaining audio transparency.

Transient Protection

Another important aspect of the dbx design was to protect against very large transients causing excessive modulation of the transmission system. In a peak-limited medium, such as the Zenith television-transmission system, the peak excursion of the compressor output is an important parameter to control. This requirement can most easily be met by a clipper or limiter, incorporated into the noise-reduction compressor and set to operate at 100% modulation.

In the system submitted to the EIA for testing, a clipper was used to control peak modulation. In actual practice, this clipper may be very sophisticated (such as the "smart clipper" designs used in the broadcast industry). The discussion which follows applies to any of the many methods of peak limiting the audio-channel signal excursion.

Clippers are relatively inaudible in operation if the amount of clipping that takes place is small. That is, the clipper must clip only transients that last under a few milliseconds. In order to accomplish this:

1) The unaffected level point of the wideband compressor is set to be substantially below (approximately 17 dB) the point that would cause 100% modulation in the transmisson channel.
2) The wideband and spectral compressors, which precede the clipper, are allowed to operate quickly enough to let only brief overloads reach the clipper.
3) The static preemphasis precedes the clipper, further reducing the transient-overload duration.

By including a clipper of known characteristics within the noise-reduction circuitry, dbx-TV ensures compliance with FCC rules while maintaining transparent compander operation.

AM-Interleave Effects

Another side benefit of this unaffected level alignment, but a most important one, is that the signal level broadcast over the stereo subcarrier tends to have its amplitude distribution average between 10% and 30% modulation. For the Zenith transmission system, this is extremely important in providing for large amplitudes in the mono carrier without exceeding the allowable modulation limits of mono-plus-stereo carriers. In this case, since the stereo subcarrier tends to average below 30% modulation, the mono carrier will be allowed to stay around 70% modulation, which is not too different from present-day broadcast practice.

Difference-Channel Filtering

In order to protect the 15.734-kHz pilot, it is necessary to prevent the audio channel from containing any information above about 15 kHz. This is accomplished with a lowpass filter in the compressor. This lowpass filter is included within the feedback loop of the compressor (Fig. 15), so that the rms-level detectors in both the compressor (encoder) and expander (decoder) sense the same bandlimited signal. As of this writing, dbx is working with Zenith and the EIA to determine a recommended practice for the type, cutoff frequency, and order of this filter. Note that the phase shift introduced by this filtering must be compensated by an identical phase shift in the sum (L+R) channel, or stereo separation will suffer. In practice, this is easily accomplished by inserting an identical filter in the sum channel, since both channels must be bandlimited for proper operation.

In order to provide for future improvements to the system as a whole, any filtering required at the transmission end must be compensated fully at the transmission end of the system only. Similarly, any filtering required at the receiver end will be compensated only in the receiver. In this way, hardware limitations to the phase linearity of filters, ac coupling required to eliminate dc offsets, and other such limitations of the present state of the art will not be perpetuated by the system implementation.

The received, demodulated difference (L-R) signal will have high-frequency components (caused by the other audio channels in the system) that could interfere with proper decoding if they are not attenuated before reaching the expander detectors (particularly the spectral-expander rms detector). Filtering is therefore necessary to eliminate mistracking. Again, as of this writing, dbx is working with Zenith and the EIA to determine the type, order, and cutoff frequency of this filter. Additionally, it may be placed either in the L-R signal path or in the expander control path. If it is placed in the signal path, a compensating filter with the same phase and amplitude characteristics must be placed in the sum-channel path, or separation will suffer. If it is in the control path, only one filter is needed (Fig. 16).

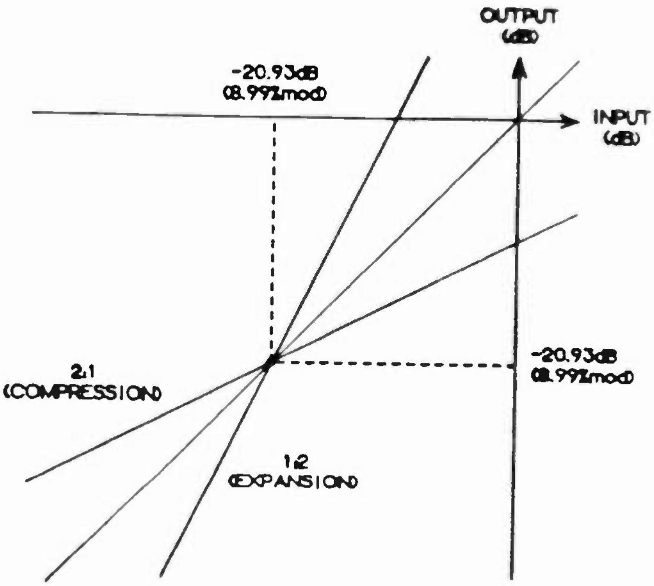


FIGURE 14
WIDEBAND AMPLITUDE COMPANDER
RESPONSE CHARACTERISTICS

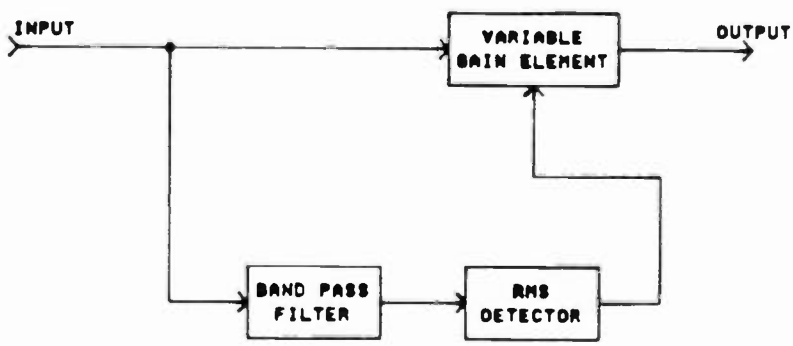


FIGURE 15
WIDEBAND AMPLITUDE EXPANDER

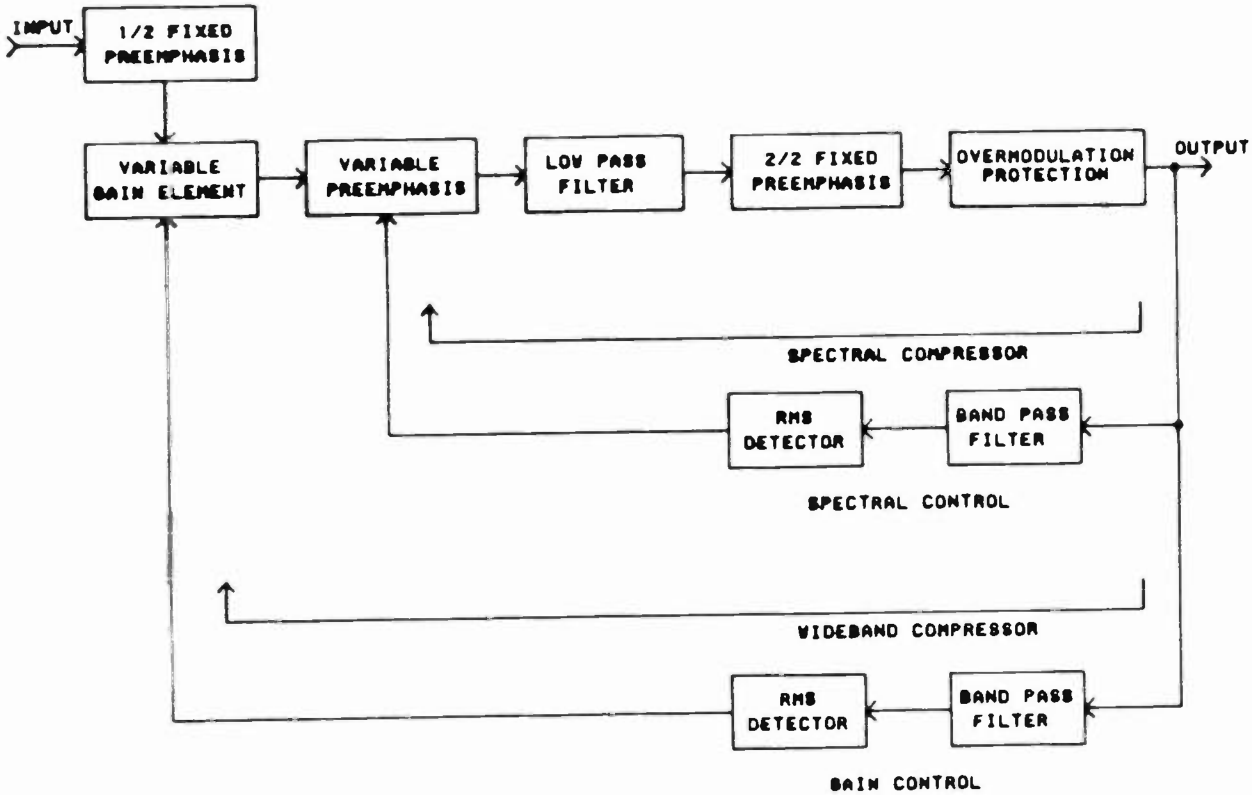


FIGURE 16
COMPRESSOR (ENCODER)

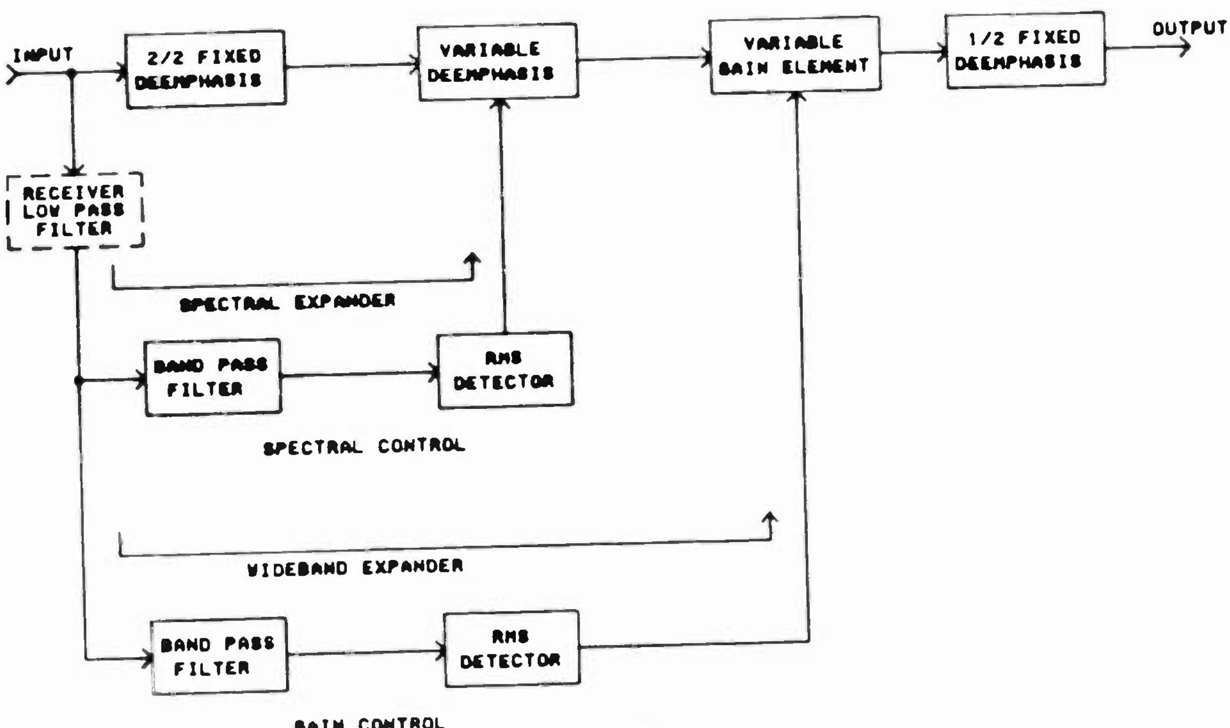


FIGURE 17
EXPANDER (DECODER)

Summary

The combination of dbx TV and the Zenith Multichannel Television Sound Transmission System (refered to as the BTSC -- Broadcast Television Systems Committee -- System) provides true high-fidelity stereo performance for all viewers within the present range of TV signals. When switching into stereo, there will be none of the added hiss that accompanies the switch into stereo with FM radio. Neither the fidelity nor the coverage area need be compromised.

For the SAP channel, the combination of dbx TV and Zenith allows adequate performance to be achieved out to the grade-B contour. City-grade (grade-A) performance will be quite good, consistent with a 10-kHz-bandwidth channel. The uncompanded channel S/N ratio is so low that some compander artifacts are inevitable in grade-B reception, but this will not interfere with intelligibility and the service quality is still acceptable.

Full compatibility with the TV sets now in use is maintained, while new receivers will be able to obtain the benefit of stereo and SAP reception at reasonable cost. It is also possible to produce add-on receivers that will allow stereo or SAP reception with a present-day TV set.

Acknowledgement

The system presented here represents the work of many individuals both inside and outside dbx. Particular note is taken of the contributions to the EIA committee effort, which crossed company lines. While we each brought our own philosophies to the committee table and our opinions were often divergent, there was a tremendous sharing of ideas; the final result reflects contributions from all. dbx is proud to have been a part of this industry-wide effort to devise and select the BTSC system, and we are of course especially pleased to have been selected to play such a vital role in the compander design.

The authors would like to express their appreciation for the aid and support of the EIA and of all the individuals and their respective companies making up the EIA Multichannel Television Sound Subcommittee, particularly those individuals making up the Companding Working Group. It has been most rewarding to participate in a effort characterized by such cooperation among competing companies.

RECENT DEVELOPMENTS IN HIGH EFFICIENCY KLYSTRONS

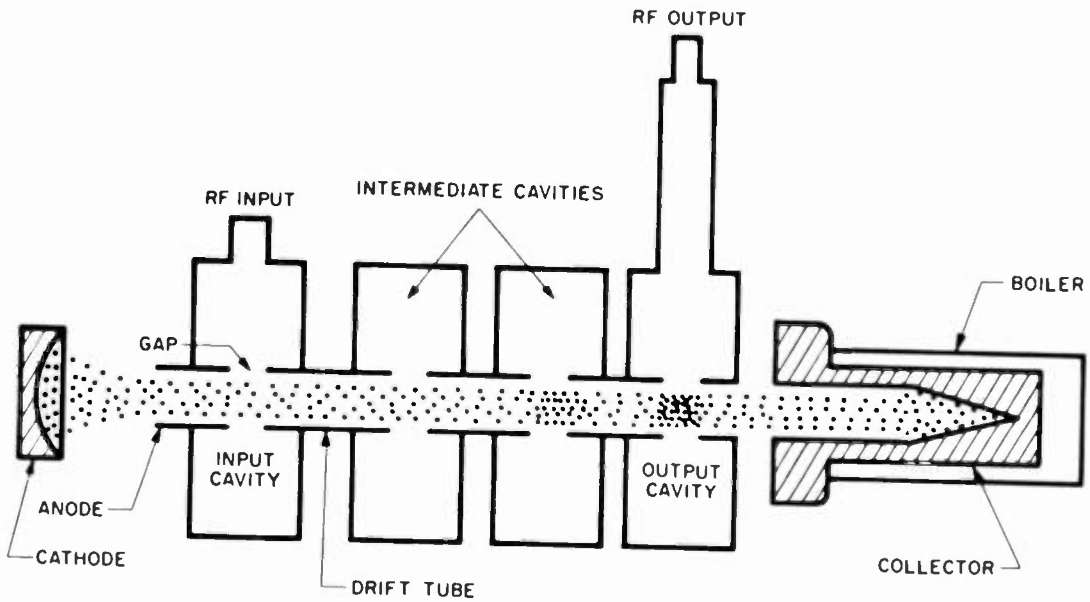
HOWARD FOSTER
VARIAN ASSOCIATES
PALO ALTO, CA 94303

Last year at this time we (1) reported on some initial results involving a new technique of tuning five-cavity klystrons. During the intervening 12 months we have expanded and completed that work resulting in the introduction of a new series of efficiency-tuned five-cavity tubes.

The original investigation into this technique at Varian was performed by Lien (2) and made use of second harmonic pre-bunching. The electron dynamics developed in the Lien model have been applied to production type five-cavity klystron tubes having adequate bandwidth for television service and which employ fundamental pre-bunching. The result has been a new series of tubes which are both cost effective and provide an increase in efficiency of ten percentage points. To differentiate these from earlier types, we have referred to these tubes as being tuned in the S-configuration, or more simply, S-tubes.

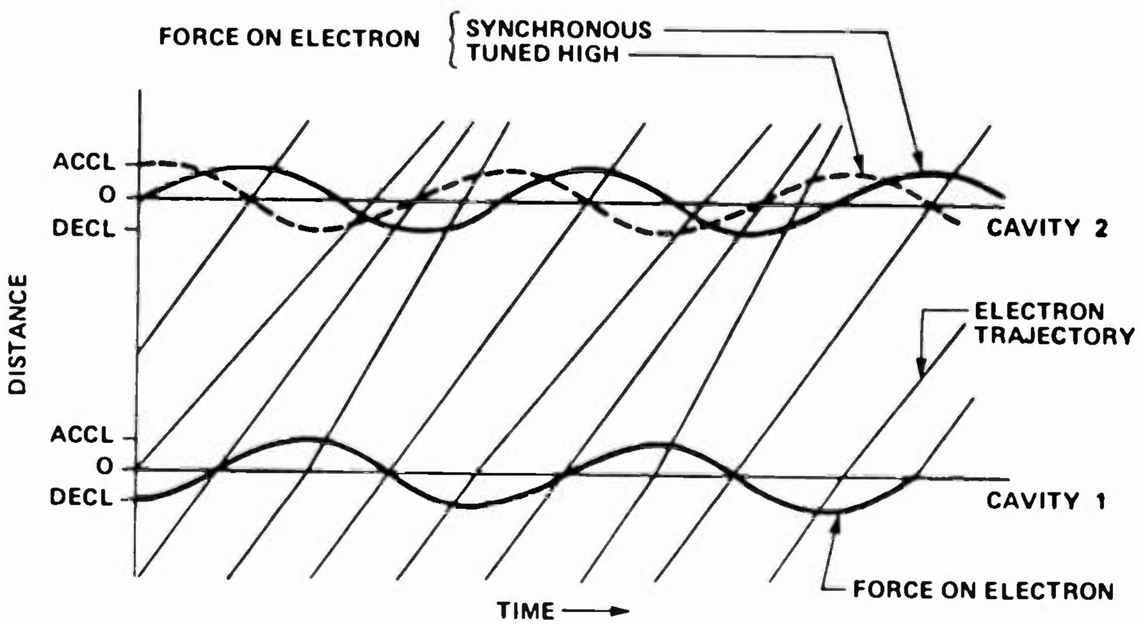
Figure 1 is a simplified diagram of a typical integral-cavity klystron used as the final high power amplifier in a UHF-TV transmitter. The RF input signal is coupled into the input cavity by an inductive loop. The RF voltage impressed across the capacitive gap of this cavity applies an RF component to the beam current. The intermediate cavities enhance this process until at the gap of the output cavity, the electron plasma is formed into more or less tight bunches. By judicious design of the output cavity, gap and loading varian

FIGURE 1
SCHEMATIC DRAWING OF KLYSTRON



varian

FIGURE 2
APPLEGATE DIAGRAM



loop, (3) the energy in the bunched beam can be efficiently extracted, and delivered to the RF transmission line load. The conditions for high conversion efficiency are the formation of bunches which occupy a small region in velocity space and the formation of interbunch regions with low electron density. The latter is particularly important since these electrons are phased to be accelerated into the collector at the expense of the RF field. Feenberg (4) has analyzed the energy exchange between the RF fields across the gap and an electron beam of given initial velocity and plasma wavelength. It can be shown that the energy loss due to an electron accelerated into the collector can exceed the energy delivered to the field by an equally and properly phased electron.

Figure 2 is an Applegate diagram showing the effect of cavity tuning on electron bunch formation. Each diagonal line represents an electron having velocity proportional to the slope of the line. All electrons enter the bottom of the diagram from the gun region with equal velocity. The pre-bunching of the electrons due to the RF field in cavity 1 is reinforced in cavity 2 when this cavity is asynchronously tuned to the high end of the pass band. Moreover, in order to obtain the proper phase relationship between the plasma wave and the RF fields in intermediate cavities, it is necessary to tune these cavities well above the pass band. Figure 3 is a pictogram showing where these cavities are tuned with respect to the passband of the VKP-7553S klystron. In such a tuning arrangement, the gain is quite low, typically 36 db as against normal gain of about 50 db for a five-cavity klystron. When the tube is "efficiency tuned" in this manner, it is necessary to substantially lower the Q of the second cavity. This is accomplished in the S-Klystron by the addition of a coupling loop and external load on this cavity. The power delivered to this load may be as much as 500 watts on some channels.

Figure 4 is a phase-space diagram comparing a conventional five-cavity klystron with the new S-tube. This diagram is a computer prediction of the phase sorting of all electrons entering the interaction region. The model predicts that fewer electrons will appear in the interbunch region of the S-tube than for the conventional tube.

FIGURE 3
TUNING PICTOGRAM

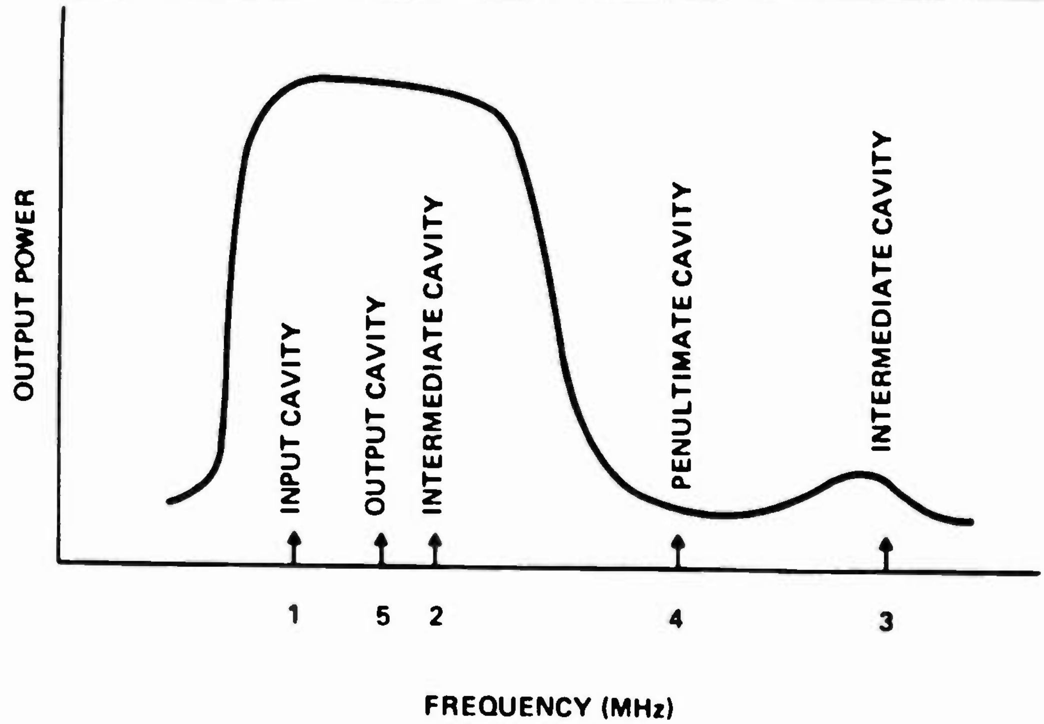


FIGURE 4
EFFECTS OF KLYSTRON TUNING ON BUNCHING

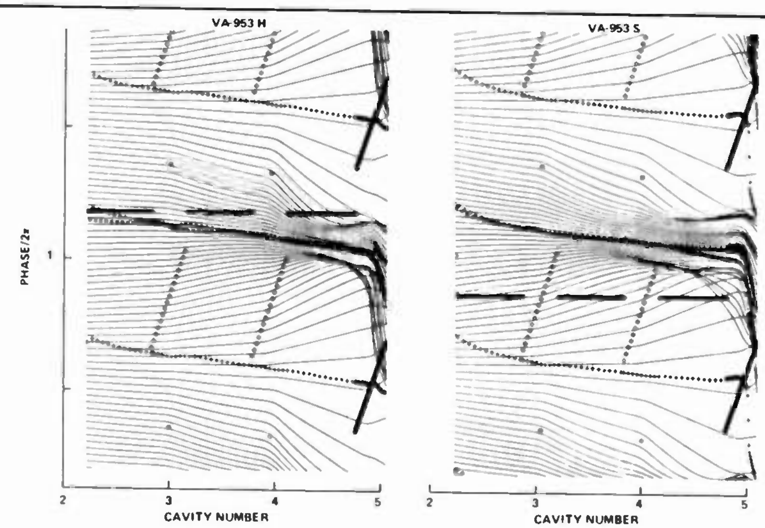


Figure 5 shows the growth of the bunching process by plotting the normalized instantaneous RF component of beam current for the fundamental, second and third harmonics of the plasma wave. Note that the fundamental (i_1) shows a continuously increasing positive slope just before entering the output gap while the (i_2) and (i_3) components peak right at the gap (5).

These computer simulation programs also predict the conversion efficiency for a given output cavity design. As of the time of this writing, several production prototype tubes have been built and tested in the low, mid and high band channels. The performance of these tubes has matched the predictions with a good degree of accuracy.

Figures 6, 7, and 8 are a compilation of the major performance characteristics of the new S-tubes compared to standard tubes, together with transfer characteristics and efficiency response of the new tube. An improvement in the saturated efficiency of ten percentage points has been realized. This may yield a peak-of-sync (6.) efficiency in excess of 63% to 65%. The relatively low gain of the S-tube requires RF drive powers up to 25W for the worst case channel. With the exception of this and the external load requirement for the second cavity, these tubes are equal to the standard tube in form, fit and function. In fact, when these tubes are retuned to the standard configuration, performance is essentially equal in all respects to the standard tube.

Since the new S-tube makes maximum use of common parts and is essentially equal to current production types having well known characteristics, we believe it offers a very cost effective method by which broadcasters can realize improved UHF transmitter efficiency.

FIGURE 5
PLASMA WAVE INSTANTANEOUS CURRENTS

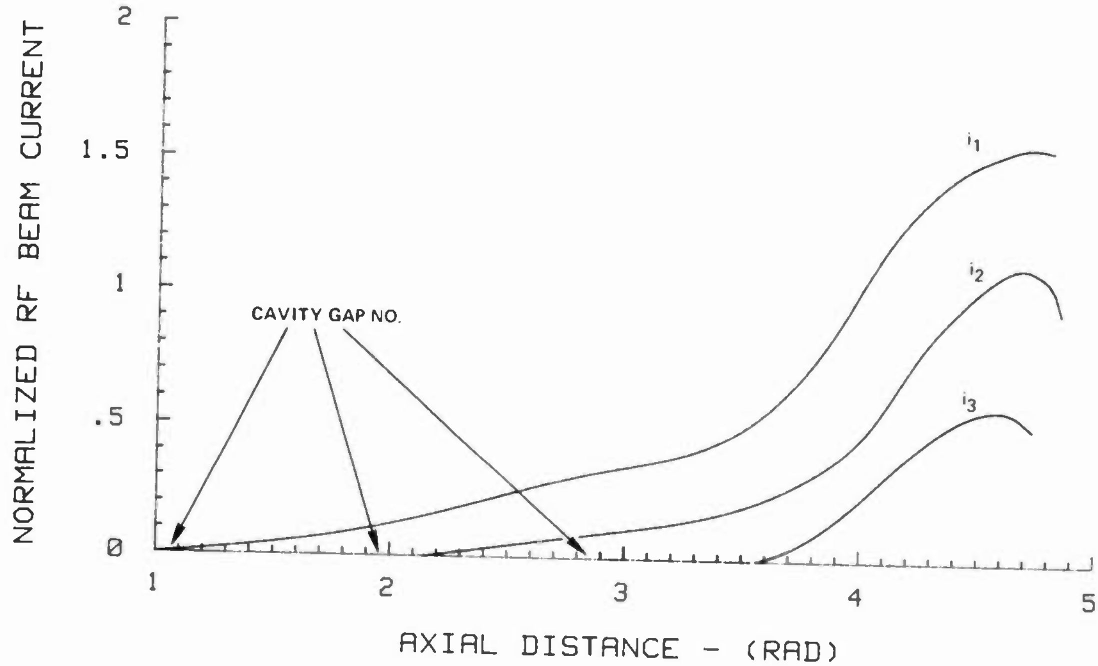
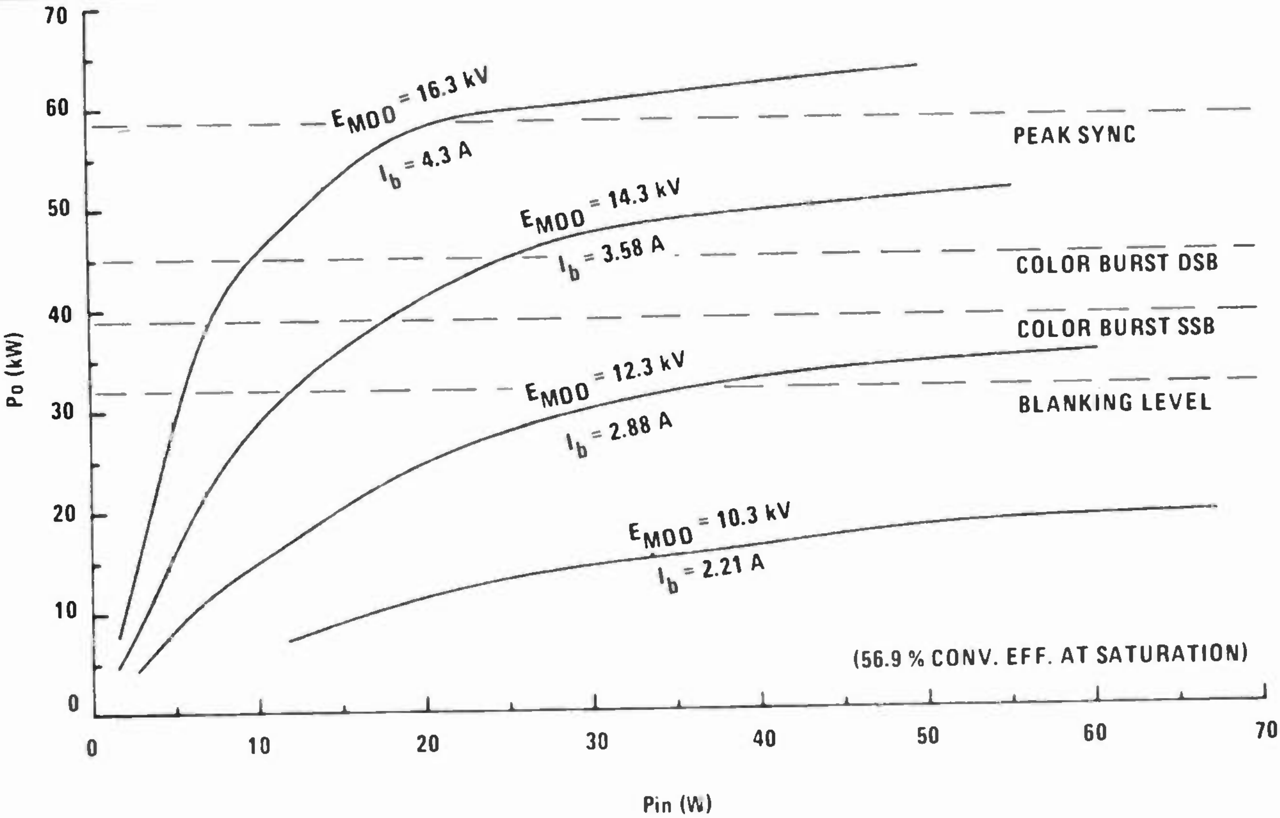


FIGURE 6
PERFORMANCE COMPARISON

CHARACTERISTIC	STD. TUBE	VKP-7553S	UNITS
OUTPUT POWER (SATURATED)	65	62	(kW)
BANDWIDTH (-1 dB POINTS)	8	6	(MHz)
BEAM VOLTAGE	24.5	24.5	(kV)
BEAM CURRENT	6.3	4.8	(A)
HEATER VOLTAGE	7.0	7.0	(V)
HEATER CURRENT	16.5	16.5	(A)
MOD ANODE VOLTAGE (REF. TO K)	19.5	16.8	(kV)
TRUE EFFICIENCY	42	52	(%)
GAIN	50	34	(dB)
DRIVE POWER REQUIRED	0.5	25	(W)

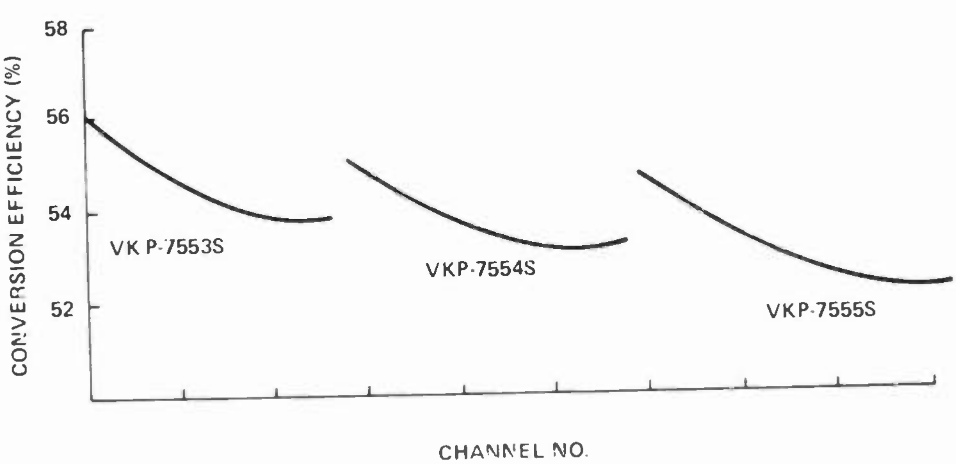
varian

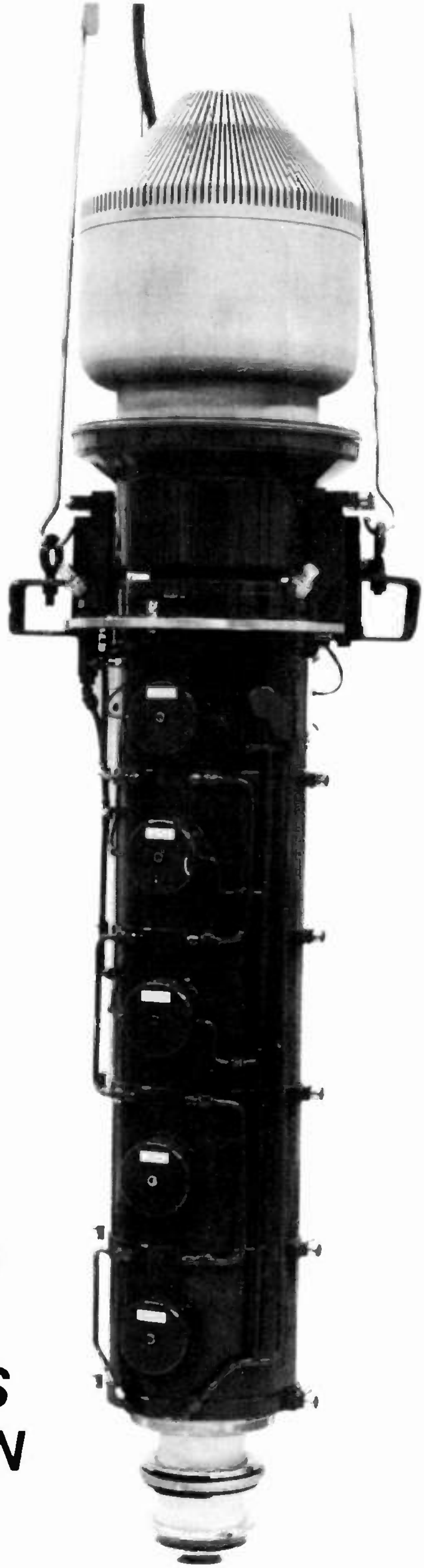
FIGURE 7
TRANSFER CHARACTERISTICS, VKP-7553S



varian

FIGURE 8
SATURATED CONVERSION EFFICIENCY, VKP-7553S SERIES KLYSTRON





varian

FIGURE 9

VKP-7553S KLYSTRON

REFERENCES

1. R. S. Symons, "Increased Efficiency from Five-Cavity Klystrons", Proceedings, 37th Annual NAB Broadcast Engineering Conference.

2. E. L. Lien, "High Efficiency Klystron Amplifiers" presented at the Conference on Microwave and Optical Generation and Amplification, Amsterdam, The Netherlands, 1970.

3. H. G. Kosmahl & L. V. Alhers, "3-Dimensional Evaluation of Energy Extraction in Output Cavities of Klystron Amplifiers", IEEE Trans. Electron Devices, Vol. ED-20 pp 883-890, October 1973.

4. E. Feenberg, "Notes on Velocity Modulation", Sperry Gyroscope Co. Report No. 5221-0143 September 1953.

5. T. Kageyama, Y. Morizumi, & E. Watanabe, "Large Signal Analysis of Broad-Band Klystrons with Design Applications. IEEE Transactions on Electron Devices, January 1977.

6. C. Cutler, "New Opoortunities for UHF-TV Transmitters", UHF Compatibility Task Force, Office of Plans and Policy, F.C.C. Washington, D.C., February 1980.

7. R. Symons, "Klystrons of UHF Television", IEEE Proceedings Vol. 70 No. 11, pp. 1304 - 1312, November 1982.

IMPROVED EFFICIENCY KLYSTRONS IN UHF TV TRANSMITTERS

Charles F. Smiley and Greg L. Best

Harris Corporation, Broadcast Transmission Division

Quincy, Illinois

ABSTRACT

TV transmitters using 60 W solid state drivers and improved efficiency 60 kW 5-cavity integral klystrons exhibit excellent picture quality while achieving 63-69% visual efficiency under pulsed conditions. The integration of these improved klystrons into existing UHF station installations will be discussed considering the tradeoffs that are possible in drive power, linearity and efficiency.

Introduction

UHF broadcasters, searching for methods to become more competitive with VHF stations, have demanded higher efficiency and secondarily more effective radiated power[1]. The latest UHF transmitters have achieved both of these goals. At 60 kW transmitter power output, transmitters with single visual klystrons have 5-10 kW less power consumption than 55 kW transmitters which were considered highly efficient just 2 years ago. This means power cost savings will be $2,600 to $6,000 per year. The extra 5 kW can be used in the antenna's main beam for higher ERP or can be used for null fill with the same ERP on the horizon.

In comparison with other 60 kW dual visual klystron transmitters, the new 60 kW single visual klystron transmitter consumes from 16 to 30 kW less power which can mean as much as $15,800 in power savings yearly.

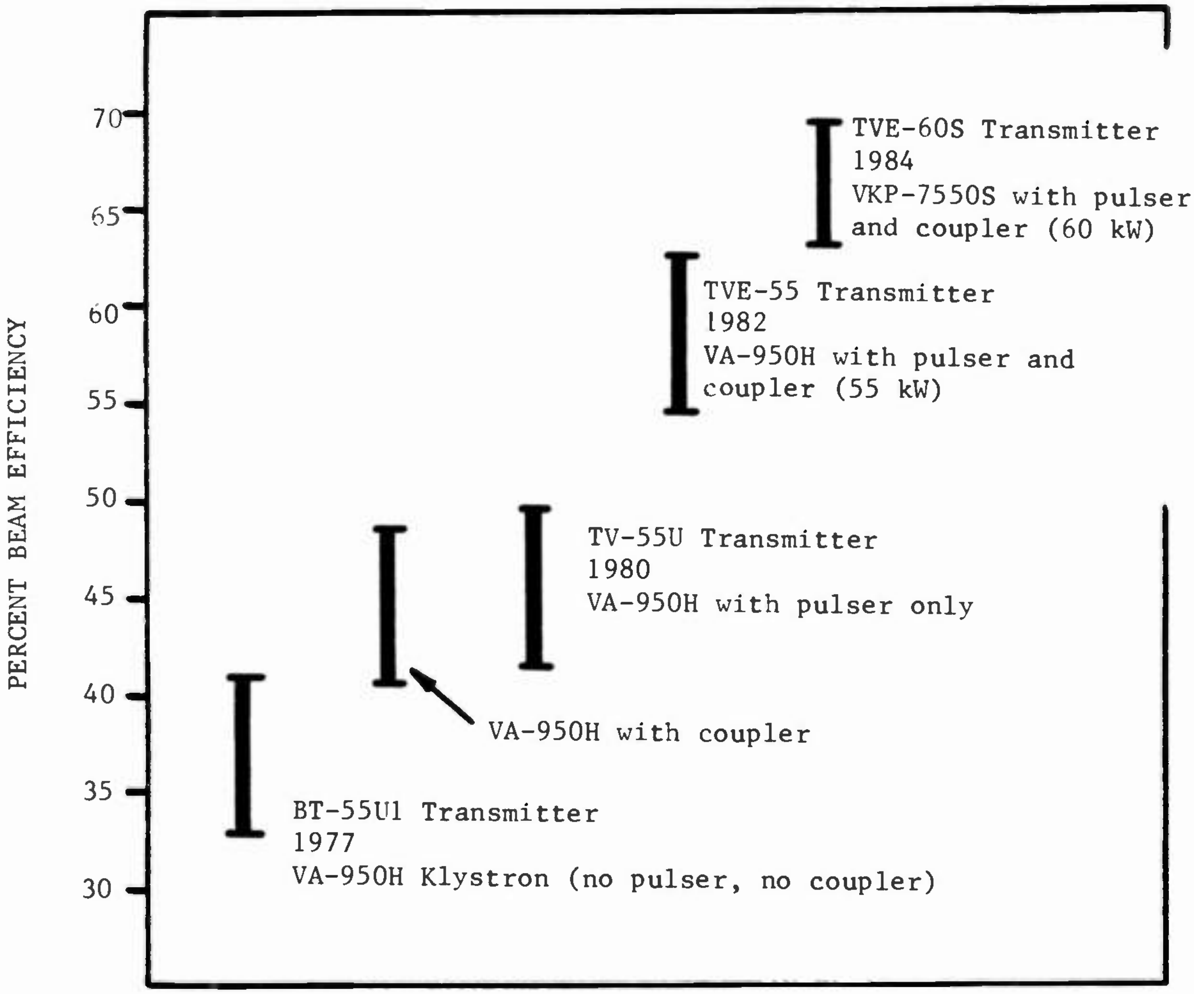


FIGURE 1

FIGURE 1. Improvements of efficiency in Harris UHF TV transmitters. Klystron type and accessories are given.

The S Klystron in Visual

The new 60 kW transmitter employs a new Varian VKP-7550S series[2] of integral 5-cavity klystrons. The VKP-7550S "S" klystrons attain 52 to 54% beam efficiency under CW or unpulsed TV conditions. The S klystron can be pulsed to produce 63 to 69% beam efficiency in a TV transmitter.

Achieving high efficiency with the VKP-7550S series klystrons puts some new requirements on a TV transmitter system - namely increased drive power, more amplitude linearity correction and increased levels of modulating anode pulsing.

The efficiency of the S klystron is made possible by tuning the 3rd and 4th klystron cavities +12 and +20 MHz above visual carrier respectively, but this also reduces the gain of the klystron by about 10 dB. This tuning is outlined in Table 1. A pulsed klystron can require as much as 50W of peak drive from a linear amplifier to pass a pre-corrected TV RF waveform. A solid state Intermediate Power Amplifier in a class A linear operation can readily produce 50 W output power by combining 4 recently developed RF power transistors.

A variable visual load coupler is used to increase the load to the output klystron cavity to use more of the beam voltage for the AC swing in the output cavity gap and increase efficiency[3,4]. The coupler is tuned to increase load impedance on the output cavity within the manufacturer's recommended tuning range or until sync tip oscillations are observed. Sync tip oscillations appear as ringing or high frequency sine wave signals on the sync pulse and are believed to be caused by secondary electron feedback enhanced by the reverse gain of the klystron cavities. When sync tip oscillations are observed, they can be eliminated by tuning the coupler for a lower load impedance on the output klystron cavity. In order to maintain the visual passband flatness of the S klystron, the Q of the second cavity has been reduced with heavier loading of the cavity by coupling in an external load. A 300 W 50 ohm load is located in the transmitter for this purpose.

Another consideration in operation of this klystron is a significant increase in amplitude non-linearity. The S tube exhibits 30-40% differential gain in typical operation. To compensate this non-linearity, a new linearity corrector with a maximum range of 50% differential gain pre-correction has been added to the exciter.

The efficient S klystron requires higher voltage levels for pulsing the modulating anode in order to decrease the beam current. In the older VA-953H klystron where 4.8A beam current was necessary to amplify 55 kW sync peak, 6 - 8 kV on the mod anode was sufficient to attain this beam current. The S klystron typically requires only 4.6 A to amplify 60 kW sync peak, and thus 7 - 9 kV is required on the modulating anode. For the video portion of the TV

TABLE 1

Comparison of tuning for Integral 5-Cavity Klystrons

KLYSTRON SERIES	VA-950H	VA-950H	VKP-7550S	VKP-7550S
CONDITION		W/COUPLER	W/COUPLER TUNED AS VA950H	EFFICIENCY TUNED
Power Output	55 kW	55 kW	55 kW	60 kW
1st Cavity	+2 MHz	+2 MHz	+2 MHz	-2 MHz
2nd Cavity	+1.5 MHz	-.75 MHz	-.75 MHz	+1 MHz
3rd Cavity	+4 MHz	+10 MHz	+10 MHz	+12 MHz
4th Cavity	+8 MHz	+12 MHz	+12 MHz	+20 MHz
5th Cavity	+0 MHz	+0 MHz	+0 MHz	+0 MHz
Typical Efficiency (Unpulsed)	37%	48%	45-49%	52-54%
Typical Efficiency (Pulsed)	45%	62%	56%	63-69%

signal 10-11 kV will be required on the modulating anode to produce the 3.5 A beam current. A new higher voltage pulser provides these voltage levels and in addition, provides filtering of the pulser waveform to decrease out-of-band spurious radiation caused by high level modulation of the output TV signal.

The operating lifetime of S klystrons is expected to parallel the field experience of the VA-950 H and G series klystrons which are demonstrating average life of 22,000 hours considering failures for all reasons[5]. Even better tube lifetimes have been demonstrated by 5-cavity integral klystrons in Harris transmitters where a customer survey of 174 klystrons in 1983 indicated an MTBF of 40,789 hours. These figures are a large improvement from a 1981 Harris survey of 179 5-cavity integrated klystrons which indicated an MTBF of 16,810 hours. Tube replacement costs for the VKP-7550S series will be the same as the older VA-950H series of 5-cavity integral klystrons.

The TVE-60S Transmitter

The block diagram for the 60 kW transmitter using the VKP-7550S klystrons in visual and aural is shown in Figure 2. Both klystrons operate at 24 kV from an external beam supply.

In the visual RF chain the MCP-2U Visual Exciter houses the modulation and pre-correction circuitry. The VIDEO[6] Saw Filter in the exciter for vestigial sideband shaping gives the transmitter its ± 0.5 dB low ripple passband, 2% K factor 2T response and more than 15 dB of visual rejection at the aural carrier for aural signal protection. The visual exciter linearity corrector has been improved to pre-correct for the increased amplitude non-linearity of the S klystron. A quadrature corrector has more than 10° of incidental phase correction capability which is sufficient for the typical 6° of incidental phase exhibited by the klystron in the video portions of the TV waveform. A separate corrector over the sync portion of the video waveform, compensates the typical 20° of incidental phase which is caused by the change in the klystron beam current by the modulating anode pulser.

To provide the 50 W of drive for the visual klystron and to overcome other RF system losses, a 60 W linear IPA has been developed. A block diagram showing typical power levels, stage gains and transistor types is shown in figure 3. Four Acrian UTV-150 power RF transistors are combined for the highest power stage of the IPA. The RF transistors in the Power Amp and Pre-Amp stages are operated in a class A push-pull pair configuration for stability into large load VSWR variations. All stages have two RF transistors operating in parallel and combined in a hybrid. This provides redundancy and greatly improves the input VSWR for each stage. A total of 4 stages of amplification provides more than 33 dB of gain so that less than 50 mW of drive is required from the exciter. An AGC is provided to stabilize the gain over temperature and time. The AGC uses the input RF signal as a reference so that IPA output power is controlled from the exciter. All pre-amplifier stages are broadbanded and the power amplifier stage is factory tuned so that no field adjustment of the amplifiers is required.

An often overlooked aspect of solid state RF amplifiers, the air cooling system, has been carefully designed. An extremely efficient aluminum heat sink with a thermal resistance to the air stream of 0.12°C/W directly cools the RF transistors avoiding the multiple thermal interfaces so common in other

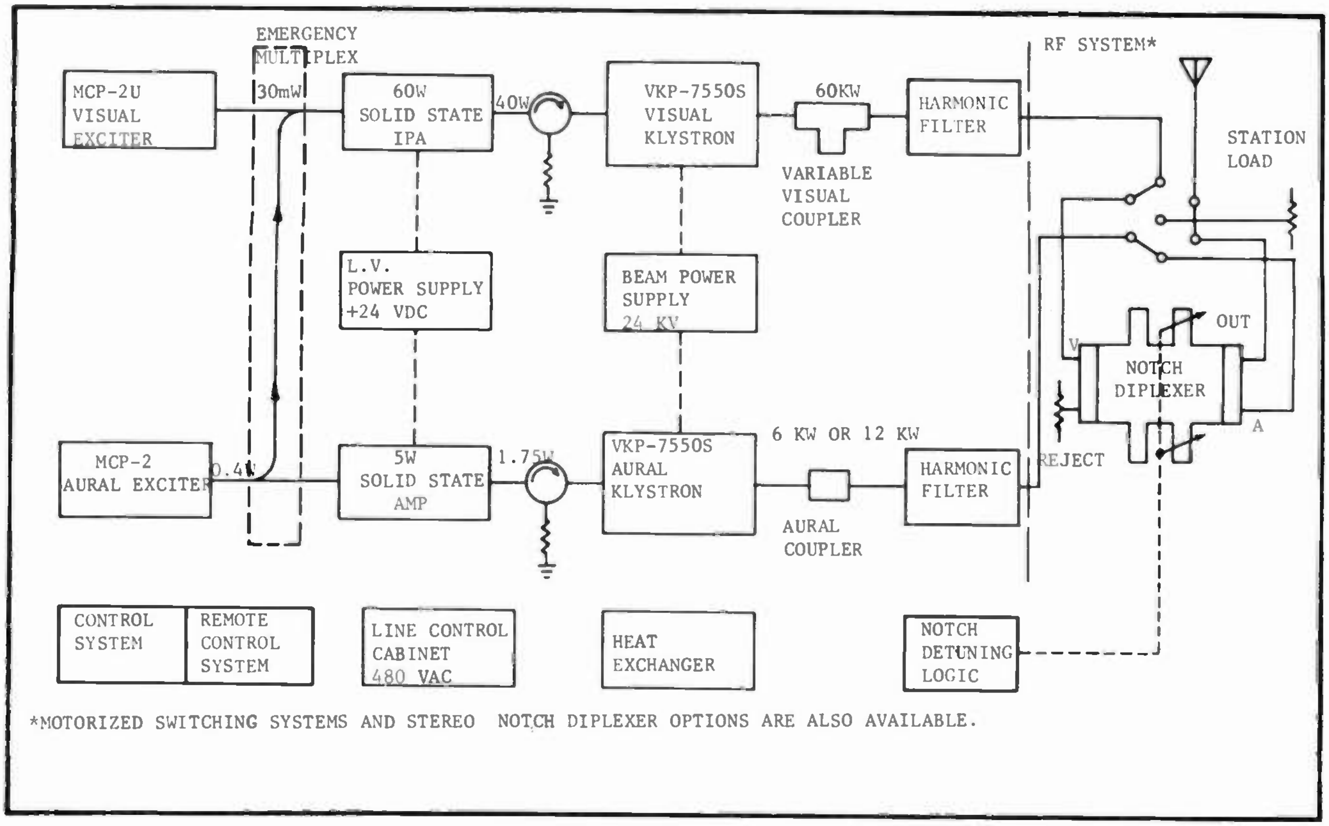
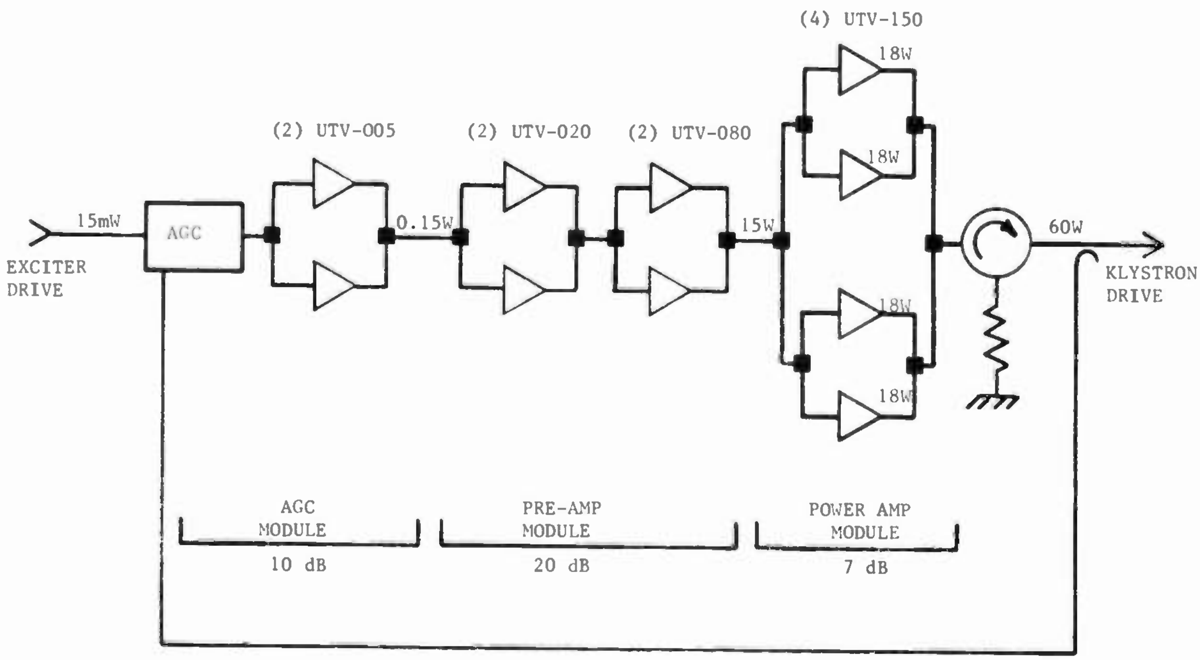


FIGURE 2. HARRIS TVE-60S BLOCK DIAGRAM

FIGURE 3

60 W SOLID STATE INTERMEDIATE POWER AMPLIFIER



commercially available amplifier modules. This inherently better cooling system has a direct effect on RF transistor lifetime reducing transistor die temperature by more than 10°C and doubling lifetime.

The linear IPA amplifier has very low intermodulation distortion for a multiplexed TV signal and makes possible emergency multiplex operation where the visual klystron is used to transmit a 20 kW multiplexed TV signal with 10% aural on an emergency basis. The klystron is the only stage with appreciable intermodulation products and typical values of the +920 kHz beat frequency are -46 dB below peak of sync under a standard 3-tone test. The emergency multiplex system reconfigures the aural RF path and resets visual and aural levels by operator initiation from either a front panel push button or by remote control.

The RF output system shown in the block diagram of Figure 2 has a notch diplexer with detunable aural notches. During emergency multiplex operation, the notches are tuned higher in frequency so that the multiplex TV signal can be passed through the diplexer to the antenna line. This type of diplexer is available with either monophonic or stereophonic aural specifications.

The S Klystron in Aural

Efficient tuning of the S klystron in aural service is obtained by slightly stagger tuning all cavities except the 4th around aural carrier. The 4th cavity is always tuned higher than aural carrier for maximum saturated power. An aural coupler or a variable visual coupler, depending on power level, is used to correctly load the S klystron output cavity. The 4th cavity is tuned typically +2 MHz higher than aural carrier when the aural coupler is used and + 4 MHz higher than aural carrier when the variable visual coupler is used.

The aural tuning method described above can provide a widened bandwidth and an even symmetric frequency response around aural carrier for good stereo performance. When stereo tuned, the 3 dB bandwidth is typically 2.3 to 3 MHz as shown in Figure 4. Drive power requirements can be as high as 1.75 W for 6 kW aural power where the gain of the klystron is reduced because of the low value of beam current (typically 0.9A) and because of stereo tuning. For this reason, a solid state amplifier with 6 dB gain has been added following the aural exciter to allow the efficiency of the aural klystron to be optimized.

Visual Transmitter Performance

The VKP-7553S klystron was tested in the Harris TVE-60S transmitter on channel 26. Performance is outlined in Table 3 and shown in figures 5 - 10. Efficiency of the transmitter was 69.4% and peak drive power to the S klystron was 53 W under pulsed conditions.

The response of the klystron bandpass was corrected to $\pm$ 0.3 dB in the exciter. Out-of-band spurious at +7.75 MHz was more than -73 dB RMS (-63 dB peak) from peak of sync level. The pulser filter proved to be more than sufficient to prevent out-of-band spurious signals from pulser sidebands.

The differential gain corrector in the exciter had sufficient range to correct 36% differential gain of the klystron to within 3%. Differential phase was corrected from 13.5° to 1.5°. Incidental phase was corrected from +6° over video to 2°. The sync tip suffers 20° incidental phase and this is separately corrected to 1° by a sync keyed corrector in the exciter.

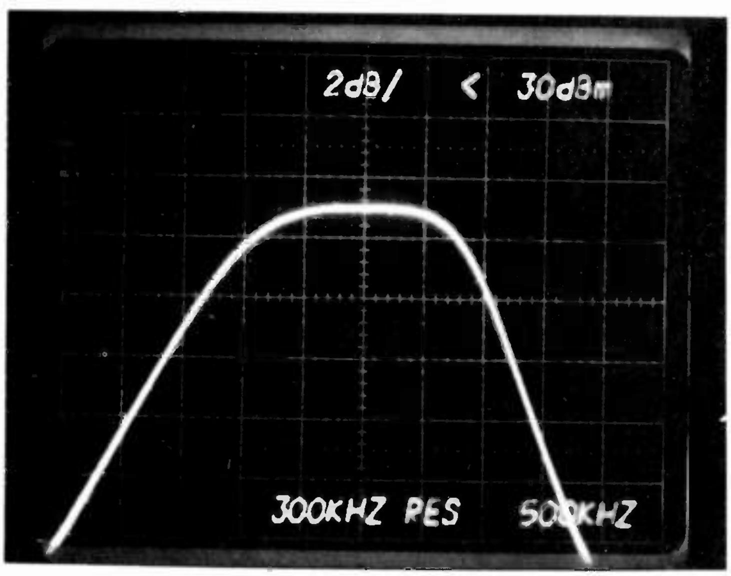
Table 2 delineates the use of the VKP-7553S klystron in aural service at channel 26. Experiments were run to compare the use of the aural coupler with

TABLE 2

The VKP-7553S Klystron in Aural Service with Stereo Tuning

Output Power	6 kW	12 kW	24 kW
Beam Voltage	24 kV	24 kV	24 kV
Beam Current	.9 A	1.3 A	2.6 A
Efficiency	27.8%	38.5%	38.5%
Drive Power	1.75W	.5 W	.25 W
Gain	35.4 dB	43.8 dB	49.8 dB
Condition	Aural Coupler	Aural Coupler	Visual coupler at maximum impedance
Tuning of 4th Cavity	+2 MHz	+2 MHz	+6 MHz

Figure 4. Tuning of the VKP-7553H klystron with aural coupler for stereo performance at 12 kW. The 3 dB bandwidth is 2.3 MHz.



Vertical scale 2 dB/div.

Horizontal scale 0.5 MHz/div.

the visual coupler for aural service. The efficiency of the aural coupler was markedly superior to the visual coupler at power levels below 12 kW which is the maximum power rating of the aural coupler. Therefore, the aural coupler will be used in the 60 kW transmitter for both 10% and 20% aural. At 24 kW aural power, the visual coupler made 38.5% efficiency for stereo tuning.

TABLE 3

TVE-60S Performance and Specifications

	SPECIFICATION	TVE-60S VISUAL PERFORMANCE	UNCORRECTED
Beam Efficiency		69.4%	
Power Output	60 kW	60 kW	
Beam Voltage		24 kV	
Beam Current		3.6 A	
Diff. Gain	5%	3%	36%
Diff. Phase	$\pm$ 3%	1.5%	13.5%
Incid. Phase	$\pm$ 2°	2°	6°
Low Frequency Linearity	11%	10%	
Sync Overshoot	5%	3%	
2T K Factor	2%	1%	
20T Baseline Disturbance	5%	1%	
+7.75 Spurious	-60 dB	-73 dB RMS (-63 dB Peak)	

Retrofit Kits

UHF broadcasters are interested in using the S klystron in existing station installations to increase efficiency. To fully take advantage of the efficiency offered by the S klystron would require the transmitter changes which have been listed in this paper. These changes include 50W of RF drive, sufficient linearity correction to pre-correct 30-40% of differential gain, and a pulser with sufficient voltage output and filtering to drive the modulating anode.

A second possibility concerns the use of an S klystron in a transmitter equipped to use a VA-953H or G series klystron. To clarify this situation, Table 4 shows experimental data on S klystrons under different tuning conditions at channel 26.

TABLE 4
Comparison of Operating Parameters and Performance
for Integral 5-cavity Klystrons

	VA-953H	VKP-7553S TUNED AS H	VKP-7553S TUNED AS H	VKP-7553S EFFICIENCY TUNED	VKP-7553S EFFICIENCY TUNED
Coupler	Yes	Yes	Yes	Yes	Yes
Pulser	Yes		Yes		Yes
Power Output	55 kW	58.8 kW	55 kW	60 kW	60 kW
Beam Current	3.65 A	5.0A	4.1A	4.8A	3.6A
Beam Voltage	24.1 kV	23.7 kV	24.0 kV	24.0 kV	24.0 kV
Efficiency	62.5%	49.6%	55.8%	53.0%	69.4%
Peak Drive Power		5.46 W	3.38 W	20 W	53 W
Uncorrected Diff. Gain	~25%	25%	28%	20%	36%
Corrected Diff. Gain	2%	5%	3%	2%	3%
Uncorrected Diff. Phase	8°	4.5°	14°	6°	13.5°
Corrected Diff. Phase	1°	1°	2°	1°	1.5°
Uncorrected Incidental Phase					
over video	+2°	0°	+6°	4°	+6°
at sync	-2°*	-2°	~-15°	-2°	-20°
Corrected Incidental Phase					
over video	+0.5°	< 1°	+1°	0.5°	+ 2°
at sync	-0.5°	< 1°	-0.5°	0.5°	0°
Out-of-band spurious at +7.75 MHz					
RMS			-72 dB		-73 dB
Peak			-62 dB		-63 dB

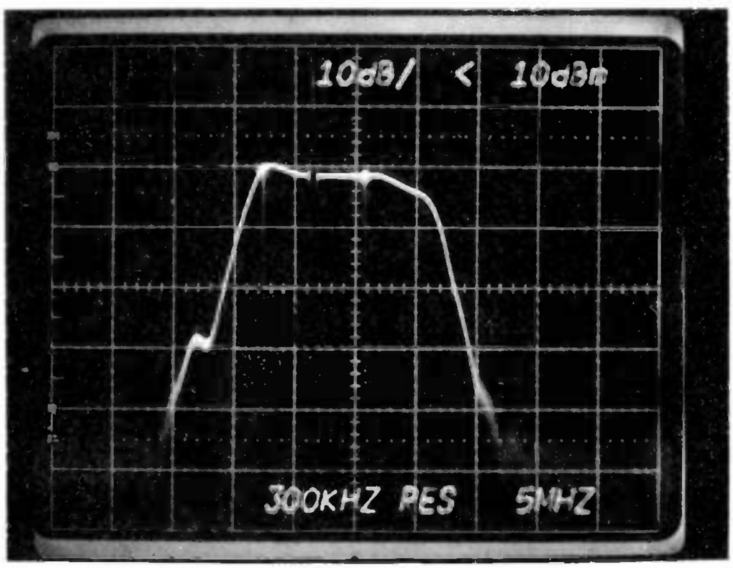
*Pulser Induced Phase Corrected

From analysis of this data it appears that a properly tuned S klystron will be a replacement for the H klystron. The first column of Table 4 shows a VA-953H with a pulser and coupler. The data in the third column, an S klystron tuned as an H with coupler and pulser, shows a similar tuning condition to the H klystron. Efficiency is lower and signal distortions are higher, but the drive levels and distortions should be correctable in a transmitter equipped with H klystrons. The efficiency would be considered as a normal variation between klystrons or slightly different tuning conditions. The data in the second column of Table 4 is a tuning condition which is typical of a problem which can arise using an S klystron for an H with coupler where the S klystron has been tuned too efficiently. In the case of the second column, the transmitter would probably not have enough drive since more than 5 W is required. Many transmitters have only 0.75 W of drive for unpulsed operation or 3 W of drive for pulsed operation. Similarly, tuning too efficiently can increase amplitude distortion, and the exciter may not be able to sufficiently pre-correct the TV signal.

At Varian, data on VKP-7553S klystrons tuned as H klystrons for different channels throughout the band have shown lower gain at lower channels (9 dB variation across low band). Thus, drive power problems also appear to have a channel dependence.

When the S is used as an H, a 50 W 50 ohm load must be connected to the second cavity loading port located on the pole piece of the klystron. Under H tuning conditions, the power in this load has been found to be less than 5 W.

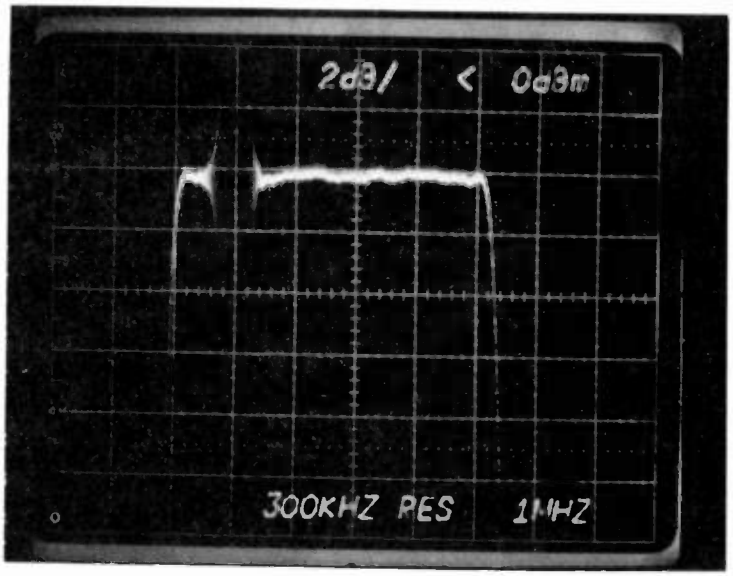
FIGURE 5. Bandpass response of the TVE-60S transmitter.



Vertical scale
10 dB/div.

Horizontal scale
5 MHz/div.

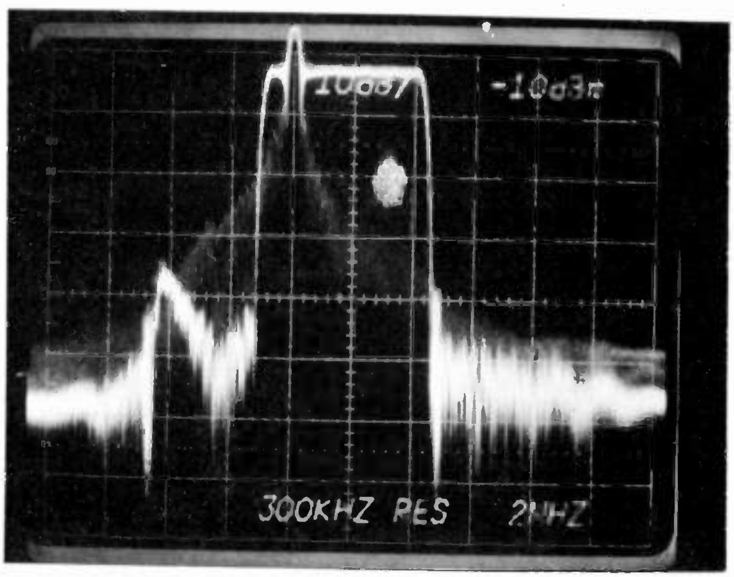
FIGURE 6. TVE-60S in-band frequency response with the VIDEO SAW Filter is $\pm$ 0.3 dB from -0.75 MHz to +4.1 MHz.



Vertical scale
2 dB/div.

Horizontal scale
1 MHz/div.

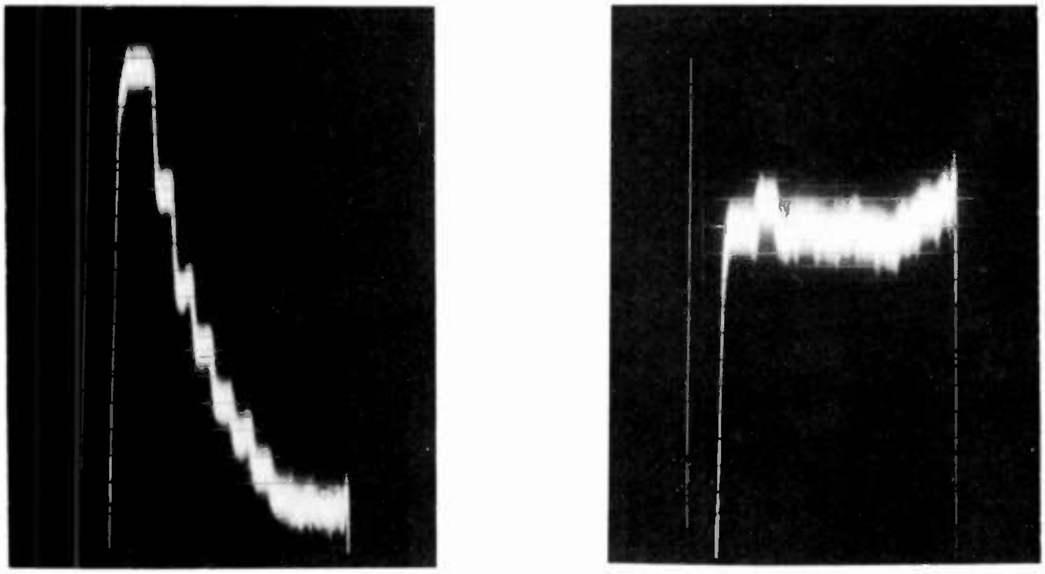
FIGURE 7. TVE-60S out-of-band frequency response is -65 dB peak at +7.75 MHz with reference to peak of sync.



Vertical scale
10 dB/div.
Center line
-60 dB from
sync peak.

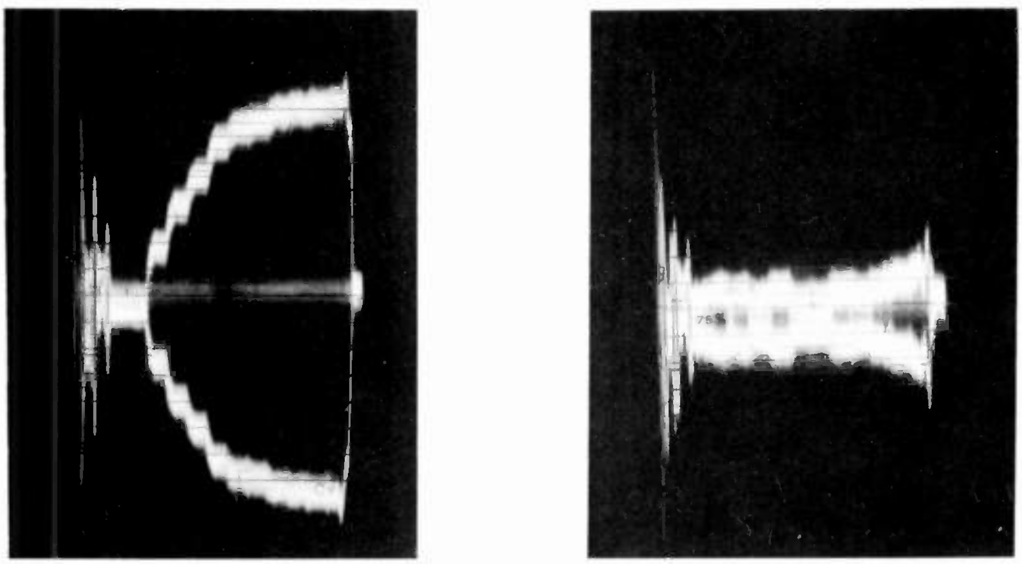
Horizontal scale
2MHz/div.

FIGURE 8. Differential gain at 69% efficiency, 60 kW. a) Pre-corrected signal driving klystron shows 36% differential gain. b) Output signal shows 3% differential gain. Vertical scale is 2% div.



a) Pre-corrected. b) Output signal.

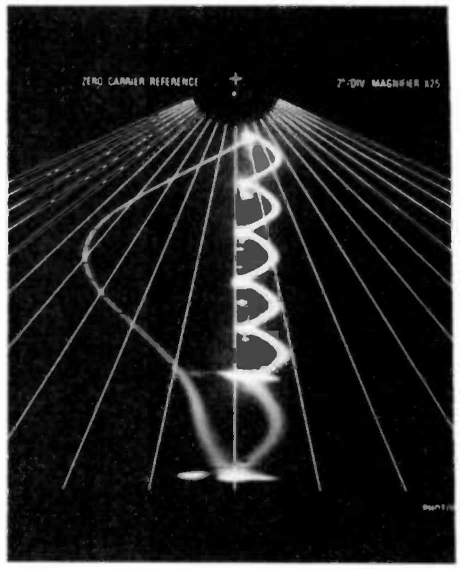
FIGURE 9. Differential phase at 69% efficiency, 60 kW. a) Pre-corrected signal driving klystron shows 13.5° differential phase. b) Output signal shows 1.5° differential phase. Vertical scale is 1°/div.



a) Pre-corrected. b) Output signal.

FIGURE 10. Incidental phase at 69% efficiency, 60 kW. a) Pre-corrected signal driving klystron shows -6° over video and +20° at sync. b) Output signal shows +2° incidental phase.

a) Pre-corrected.



b) Output signal.

Conclusion

The TVE-60S transmitter uses the VKP-7550S klystron to achieve 60 kW of output power with a beam efficiency of typically 63-69% in visual, 28% in aural service at 6 kW, and 38% in aural service at 12 kW. The transmitter has been shown to meet excellent performance specifications while meeting these efficiencies.

If a broadcaster wishes to put an S klystron in an existing H klystron transmitter, careful tuning can result in performance and efficiency similar to the H klystron although slightly increased distortion and slightly lower efficiency were observed during testing.

Footnotes

1. Gieseler, P.B., et.al., Comparability for UHF Television, UHF Comparability Task Force, Federal Communications Commission, September, 1980, pp. 125-206.

2. VKP-7553S for Channels 14-29, VKP-7554S for channels 30-51 and VKP-7555S for channels 52-69.

3. Unetich, R.M., et.al., "Perveance Reduction and Efficiency Enhancement of Integral Cavity Klystrons", Varian Special Report, Varian Associates, Inc., April, 1981.

4. Smiley, C.F., et.al., "Efficiency and Performance Improvements in UHF Television Transmitters", Harris Corporation, Broadcast Division, April 2, 1982.

5. Data from Howard Foster, Varian Associates, Inc., November, 1983.

6. Visual IF Delay Equalized Output

Acknowledgements

The author wishes to thank Jim Keller, Dave Danielsons, Jim Ruxlow and John Deemer for their assistance in obtaining data for this paper. Varian Associates, Inc. compiled the data for the 1983 Harris Customer Survey for klystron life.

UHF Transmitter Retrofit for Efficiency and Reliability

R. Wayne Estabrooks

Connecticut Public Broadcasting

Hartford, Connecticut

WEDW Channel 49 in Bridgeport, Connecticut is one of four television stations owned and operated by the Connecticut Public Television network. WEDW first began operating in November 1967, utilizing a General Electric thirty kilowatt transmitter and a General Electric TY-24 omnidirectional helical antenna side mounted on a tower at 370 ft. above the ground. The station was underpowered by modern day standards, specifications were difficult to meet, stability was poor and of course remote control was not practical.

We began planning to replace this facility in 1979. Initially plans called for replacing the old transmitter with a new one. We decided not to buy a new transmitter, but to completely rebuild the existing one in an effort to take advantage of the UHF Efficiency work done by the Public Broadcasting Service using as many of the PBS recommendations as possible.[1]

Replacement of the 14 year old GE helical antenna was deemed a priority. In addition to suspicions that much of the radiated power was going into outer space, the side mounting of the antenna caused serious problems. The signal relecting off the six foot face of the tower interfered with the circularity of the pattern and caused nulls as deep as 12 to 15 db to occur in the signal.

Failing in attempts to lease space at the top of the tower, we chose an antenna with a peanut pattern and mounted it in such a way as to minimize tower bounce by placing a null toward the tower face.

1 UHF-TV Transmitter Improvements Progress Report Compiled By John T. Wilner and Thomas B. Keller, Jr., Public Broadcasting Service July 30, 1980.

The antenna was constructed with an input harness capable of 90 kilowatts visual input power to accommodate a possible future power increase.

In the Fall of 1980, we finalized plans, equipment was ordered, and construction began in June 1981.

DESIGN OBJECTIVES

Our design objectives for the transmitter included;

1. The use of two 30 kilowatt parallel visual output amplifiers in order to increase power, improve redundancy, and provide for possible modification to 55 kilowatt units at a later date.
2. To achieve the maximum efficiency possible utilizing available technology in 1980.
3. To eliminate all coaxial elbows in the visual signal path, between the output of the transmitter cabinets and the tower elbow complex.
4. To achieve a reliable, stable, transmitting operation capable of state of the art quality, remote control capability, and improved klystron protection.
5. To attempt to accomplish the retrofit with no loss of air time.

In order to accomplish the rebuild, most of the major components of the transmitter had to be replaced. These items included the beam contactor, beam supply, exciter, diplexer, klystrons, harmonic filter, and the entire cooling system. As each item was chosen, careful consideration was given to efficiency. We would retain the original visual and aural amplifier cabinets and the high voltage rectifier and control cabinets. The original exciter/driver would be retired and a second visual amplifier installed in its place.

We decided to use the original transmitter cabinets to help avoid licensing problems. We found a used General Electric visual amplifier cabinet in Louisiana and purchased it. The F.C.C. allowed us to license this new installation as a General Electric T T-61 modified because G.E. had manufactured a 60 kilowatt parallel transmitter with that model number. The modification consisted of a Harris exciter retrofit package, a Comark waveguide diplexer, and a Dielectric Communications waveguide harmonic filter.

In order to operate the second visual amplifier, much modification had to be done. New control and protection circuits, arc detector chassis, and a complete magnet supply were constructed. A rack was installed next to the rectifier and control cubicle to house the new visual amplifier circuitry. (Fig. 1) Numerous parts in the transmitter were upgraded or

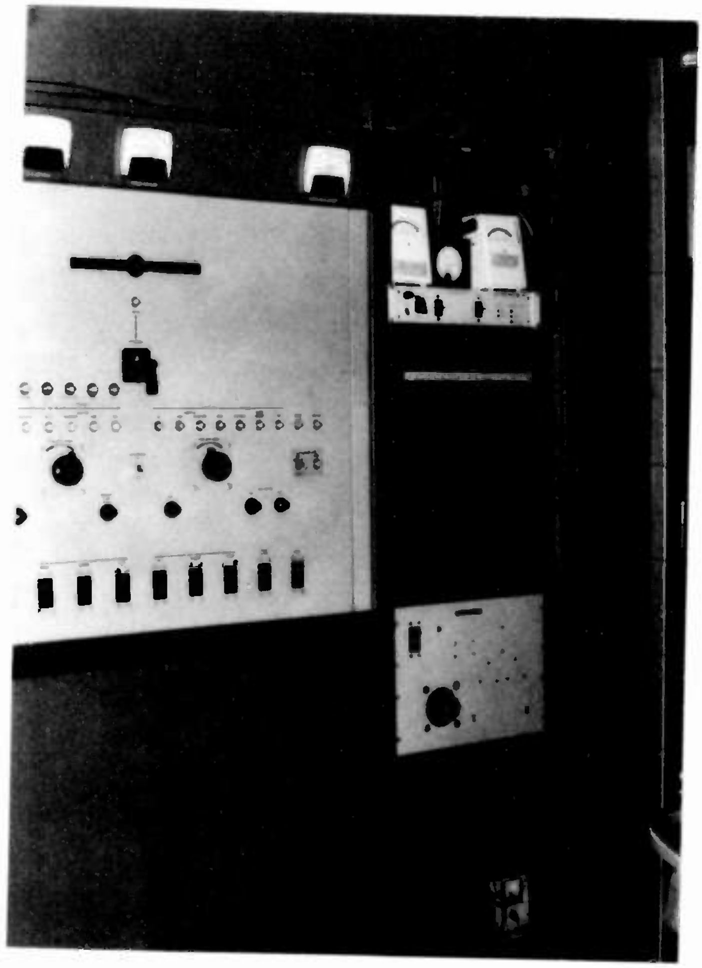


Fig 1

RACK ADDED TO TRANSMITTER TO HOUSE VISUAL #2 MAGNET SUPPLY, CONTROL CIRCUITS AND CONTROL PANELS FOR WAVEGUIDE SWITCHING AND PULSER replaced. New high voltage resistor divider assemblies were built and installed in each visual amplifier to provide bias taps for the pulser installation.

The transmitter cooling system was completely rebuilt to accommodate the additional water flow required by the second visual amplifier and to make use of a high efficiency heat exchanger. (Fig. 2) The original heat exchanger in the 30 kilowatt GE used a 15 horsepower motor. The new heat exchanger uses four 1 horsepower fan motors which turn on in sequence with rising temperature. This unit uses only one fan motor to cool the transmitter in cold weather and we have never needed more than three fan motors running to cool the transmitter even in the hottest weather. This results in a very significant energy saving.

The new heat exchanger is mounted outdoors necessitating the use of a 50/50 mixture of ethylene glycol and water to prevent freezing.

A temperature controlled motorized mix valve was incorporated into the cooling system design to prevent very cold coolant from circulating through the klystrons at startup. (Fig. 3)

The water pumps for the system were designed to provide adequate flow using only one 5 horsepower pump motor. A 60 kilowatt external cavity klystron transmitter normally uses a 7.5 to 10 horsepower motor. Three inch pipe was used on all the main lines to minimize restriction.

We decided to use waveguide as extensively as possible to take advantage of its power handling capability, reliability, and efficiency. The waveguide system starts directly at the output of the visual amplifier cabinets. (Fig. 4) The output of the amplifiers connect directly to the visual hybrid combiner. We decided not to use switching at the output of the visual amplifiers due to space and cost factors. We learned from our experience with parallel visual amplifiers at another site that switching around the visual combiner to permit operation at 50% power is seldom used. We did purchase waveguide sections to bypass the visual combiner and allow the connection of either visual amplifier directly to the diplexer in the event it is ever necessary to have one amplifier out of service for an extended period. This arrangement also permitted the operation of the single 30 kw transmitter into the waveguide system until the installation of the second visual amplifier cabinet was complete.

At the output of the visual combiner is a waveguide harmonic filter. (Fig. 5) The original harmonic filter was not capable of high power operation. We were able to use the original harmonic filter on the aural transmitter.

At the output of the visual harmonic filter, the waveguide size transitions from WR1150 to WR1500. A bidirectional coupler was provided at this point to allow combined visual power metering and VSWR measurement and

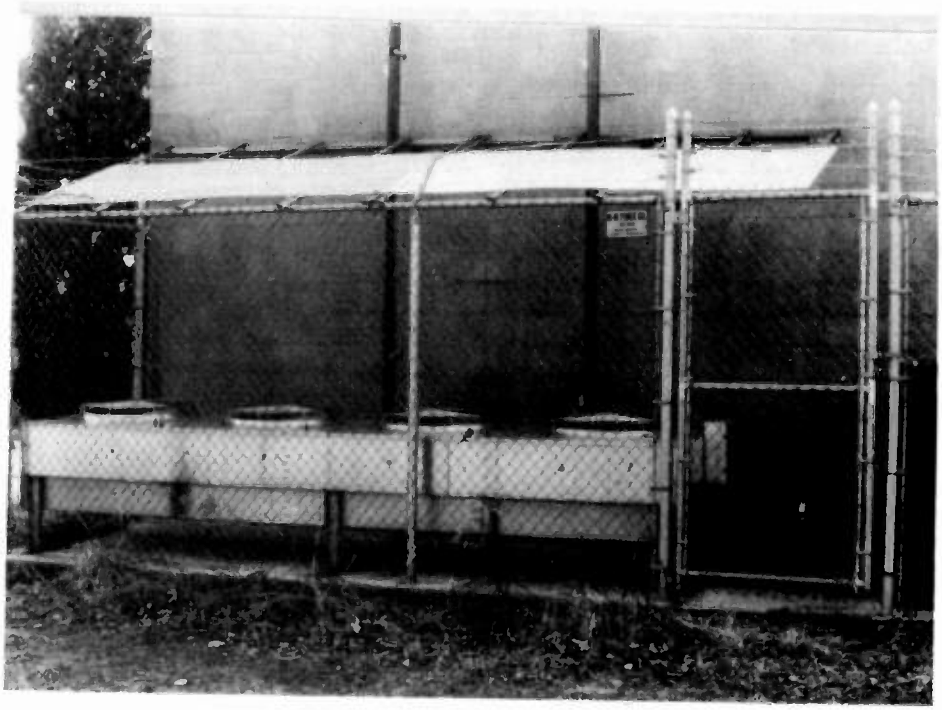


Fig. 2

HALSTEAD & MITCHELL 550,000 BTU HEAT EXCHANGER MODEL GLY-38

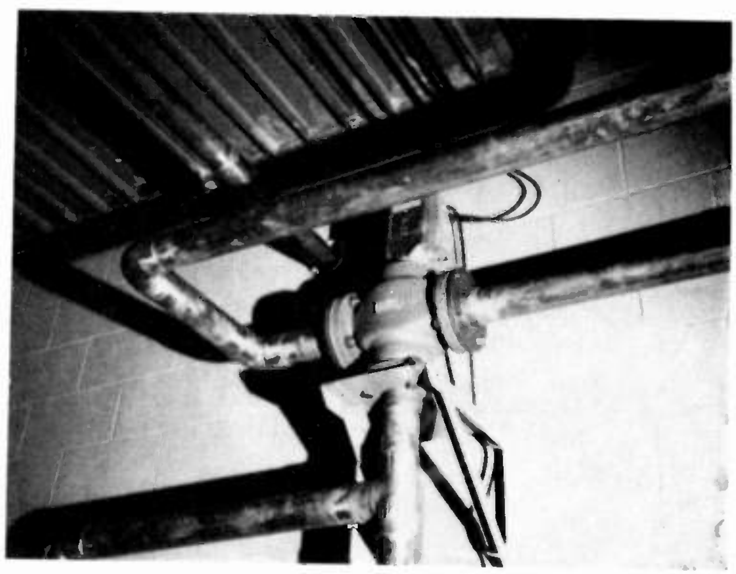


Fig. 3

TEMPERATURE CONTROLLED MOTORIZED MIX VALVE

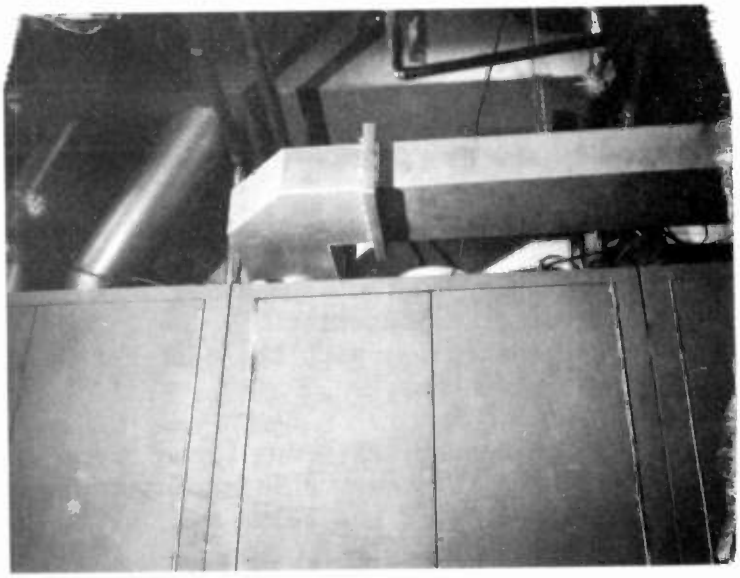


Fig. 4

WAVEGUIDE CONNECTED DIRECTLY TO OUTPUT OF VISUAL AMPLIFIER

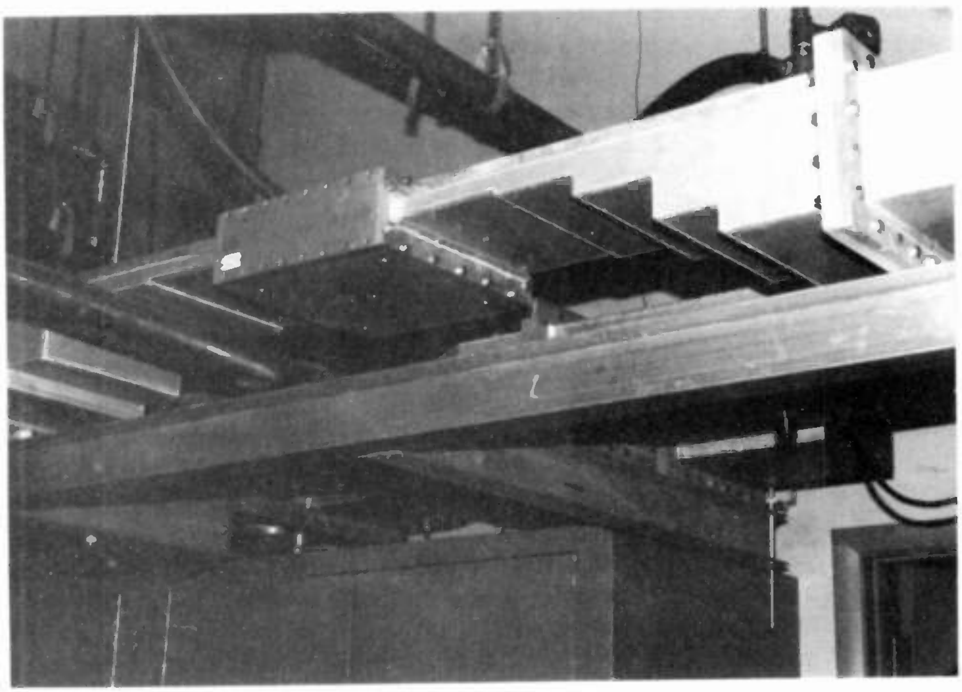


Fig. 5

WAVEGUIDE HARMONIC FILTER

protection. The combined visual signal then enters a three port wave-guide switch at the input of the diplexer which in conjunction with a four port waveguide switch on the output of the diplexer provides for four modes of operation.

1. Bypass of the diplexer for emergency multiplex operation.
2. Visual amplifiers direct into the dummy load.
3. Diplexer output into the dummy load for power meter calibration.
4. Normal operation, with the diplexer output into the antenna.

A horizontal waveguide run to the tower transitions directly into the 6 1/8 vertical run of coax up the tower eliminating the coaxial elbow at the base. A waveguide run up the tower was ruled out due to windload considerations.

A high speed vacuum beam contactor rated at 600 A. was installed to improve klystron protection. A high speed solid state AC overcurrent protection circuit designed by Comark is used in this contactor. (Fig. 6)

An integrated beam supply with a -60 db hum and noise specification was chosen. The supply is rated at 92% efficiency. The original beam supply was thought to be only about 88% efficient.

The original beam supply was left intact and could be used in an emergency in the event of a catastrophic failure of the integrated beam supply.

The line voltage regulator was completely eliminated for a power savings of at least 3 kilowatts and no adverse effects were realized. Since then we have disconnected the line regulators on our Hartford station. Both locations have had good power line regulation.

The final phase of the project was to install a pulser. After we had the transmitter up and running as a parallel transmitter we proceeded with the pulser installation. The pulser we bought was designed to install in a Harris integral cavity klystron transmitter and we were installing it in a GE external cavity klystron transmitter. We had to fabricate or modify many parts. High voltage resistor divider assemblies had already been fabricated previously. The zener diode board, mod anode resistor assembly and the vacuum spark gaps were installed inside the visual amplifier cabinets. The pulser electronics (Fig. 7 & 8) were installed on mounting brackets fastened to the wall directly over the front of the visual amplifier cabinets as close to the mod anode of the klystron as possible. Location of the pulser chassis on top of the transmitter cabinets was not possible due to space limitations. A benefit of mounting them on the front of the transmitter is easy access to the timing adjustments in view of the monitoring equipment. The motorized potentiometer for mod anode bias raise/lower was installed on the top of the visual cabinets and enclosed in a perforated metal housing for safety.

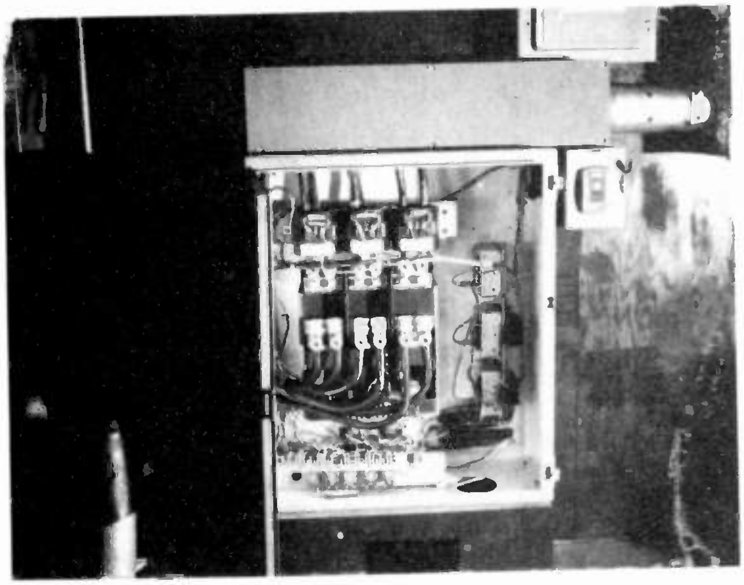


Fig. 6

VACUUM CONTACTOR ABOVE

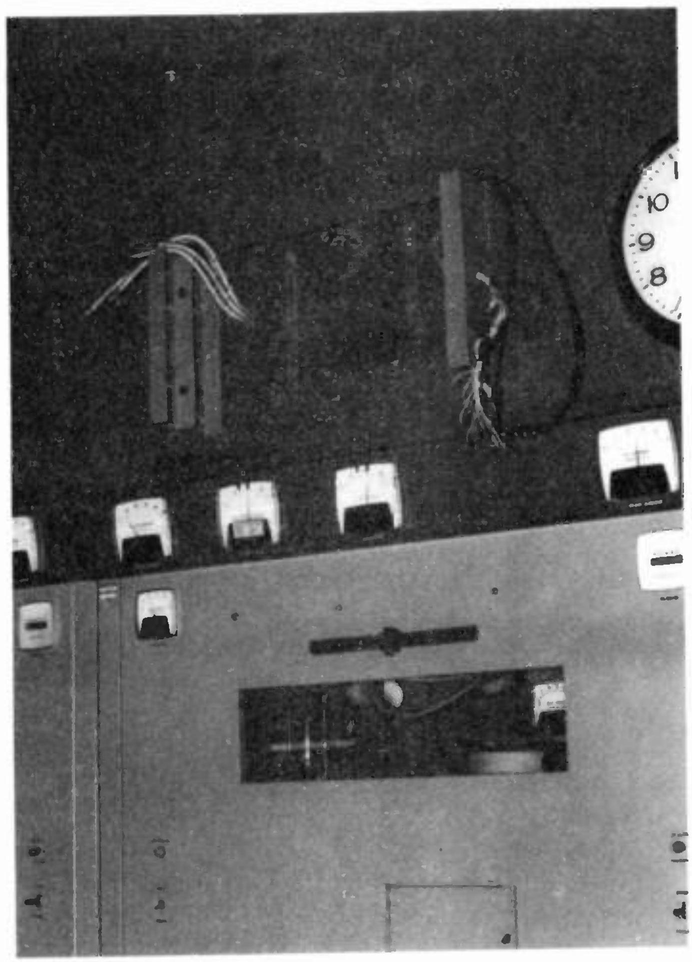


Fig. 8

PULSER MOUNTING BRACKET

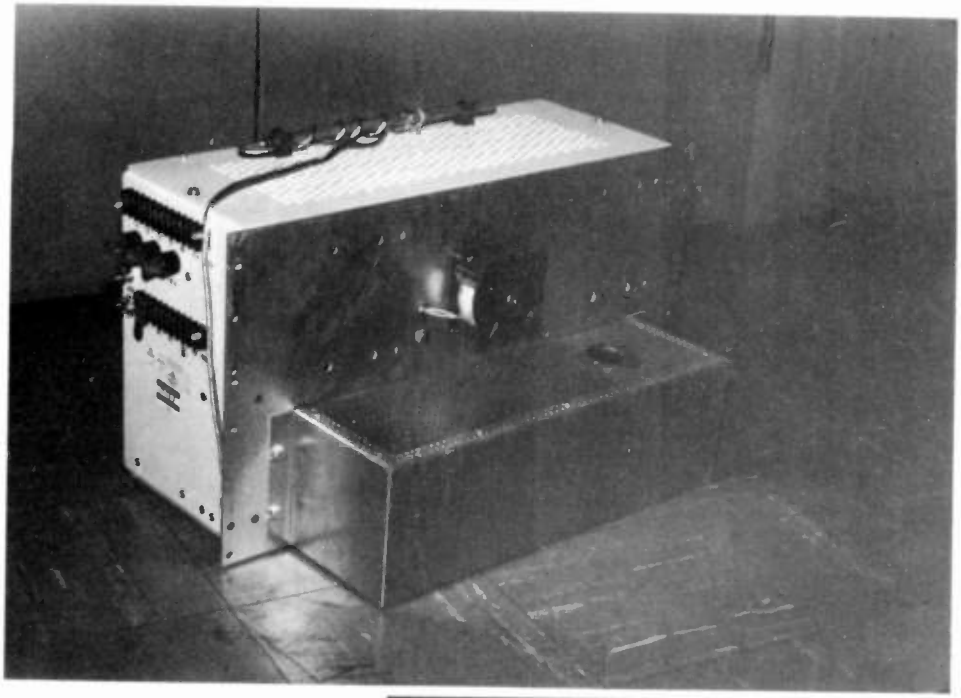


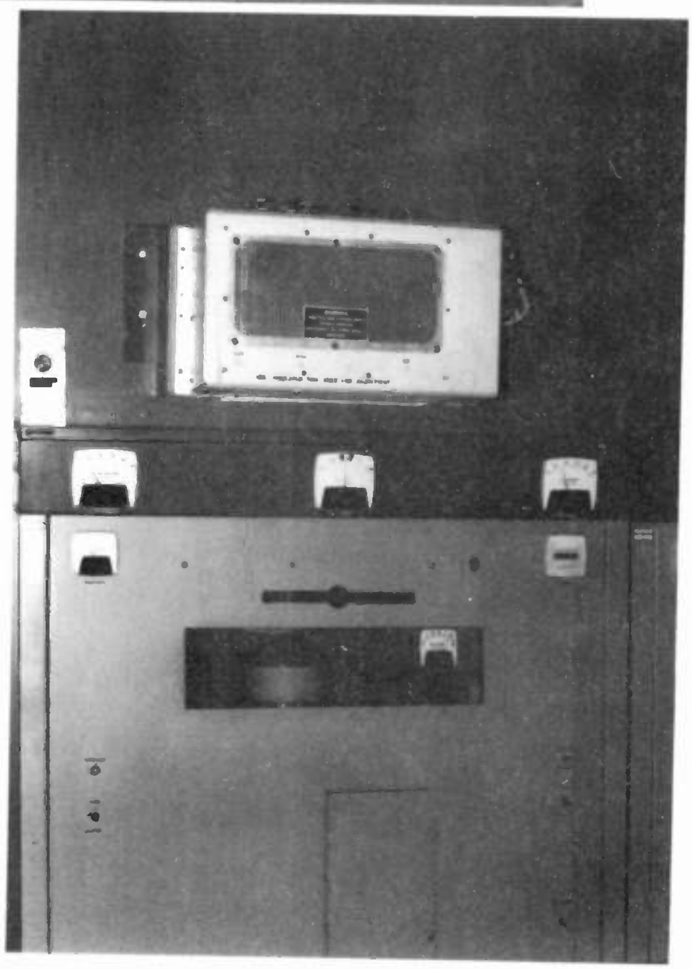
Fig 7

PULSER CHASSIS REMOVED

FROM TRANSMITTER ABOVE

PULSER CHASSIS MOUNTED

ON TRANSMITTER RIGHT



A chassis pan was added to the rear of the pulsers to enclose the rear high voltage connections for safety and shielding. High voltage wiring was routed from the rear of the pulser chassis through a short length of 1 1/2" conduit to the top of the associated cubicle. The pulser control panel was modified for rack mounting. A 24v power supply was added for pulser control and A.C. power was wired and fused. The pulser control panel was interfaced to the stations remote control so that the pulser could be turned on or off, or mod anode voltage adjusted from the studio.

The pulser can be turned on or off with no interruption in service. In the off mode, a vacuum relay selects a bias tap for normal non pulsed operation and the sync ICPM corrector is switched off.

With a slight wiring change, either pulser can be connected to pulse both visual klystrons in the event of a failure of either pulser chassis.

Initially the Harris exciter only provided for ICPM correction at sync. In the spring of 1982,Harris Corp. introduced its quadrature corrector. We subsequently purchased a retrofit kit and installed it. ICPM can now be corrected to 1 degree.

CONCLUSIONS

The complete transmitter retrofit was accomplished for under $300,000 including labor. This did not include the new antenna.

The plant has been operating for the past 2 years in its completed form. The reliability has been much better than expected.

We are very pleased with the overall efficiency we achieved. We believe the efficiency is equivalent to, or better than, any manufactured transmitter of the period. (1981) Our power bills show that the unmodified General Electric transmitter operating at 30 Kw peak visual and 3 Kw aural, consumed 127 Kw of power as indicated in the peak demand figures in the power bills (Fig. 9). The retrofitted transmitter operating at 55Kw peak visual and 5.5 Kw aural, consumes 140 Kw of power. (Fig. 10)

The technical performance exceeds manufacturers published specfications. (Fig. 11)

Based on the success of this project, we plan to accomplish a similar retrofit at WEDN Channel 53 in Norwich, Connecticut in the near future.

UI United Illuminating

statement

account number

11502-2673095

CONN EDUCATIONAL T V
VIDEO LANE
SHELTON CT 06484

demand KW

126.0

service from	to	rate	reading	constant	kilowatt hours
9/19	10/18	P2	22181	120	55200

FUEL COST ADJUSTMENT $.012850 PER KWH

	PB	2949.07
	BF	2949.07
	EL	2606.02
		730.60

Any charge or portion thereof not paid within 28 days after mailing of the first bill containing such charge is subject to interest from such date to the date of payment.

RATE OF INTEREST

1 PERCENT/MO.

OCT 1979 $6285.69

Mailing date of this bill OCTOBER 23, 1979

please pay this amount

PB—previous bill BF—balance forward CR—credit OP—off peak ES—estimate MI—minimum CO—combined EL—electric PY—payment

UI United Illuminating

statement

account number

11502-26~~73095~~

CONN EDUCATIONAL T V
VIDEO LANE
SHELTON CT 06484

demand KW

127.0

service from	to	rate	reading	constant	kilowatt hours
10/18	11/19	P2	22696	120	61800

FUEL COST ADJUSTMENT $.013110 PER KWH

LATE PAYMENT CHARGE

	PB	6285.69
10/22	PY	2949.07CR
10/30	PY	3336.62CR
	EL	2858.50
		834.50
		30.05

Any charge or portion thereof not paid within 28 days after mailing of the first bill containing such charge is subject to interest from such date to the date of payment.

RATE OF INTEREST

1 PERCENT/MO.

NOV 1979 $3723.05

Mailing date of this bill NOVEMBER 21, 1979

please pay this amount

PB—previous bill BF—balance forward CR—credit OP—off peak ES—estimate MI—minimum CO—combined EL—electric PY—payment

Fig. 9

POWER BILLS FOR UNMODIFIED 30 KW GE TRANSMITTER

SHOWING 127 KW PEAK DEMAND

UI United Illuminating

statement

account number	demand KW			
11536-2350007	140.0	2/11	PB	6371.80
CONN EDUCATIONAL T V			PY	6371.80CR
6 VIDEO LANE				
SHELTON CT 06484				

service from	to	rate	reading	constant	kilowatt hours			
1/18	2/16	P2	39717	120	57840		EL	5551.03
FUEL COST ADJUSTMENT $.007090CR PER KWH								422.39CR

Any charge or portion thereof not paid within 28 days after mailing of the first bill containing such charge is subject to interest from such date to the date of payment

RATE OF INTEREST 1.25 PERCENT/MO.

FEB 1983 $5128.64

please pay this amount

Mailing date of this bill FEBRUARY 18, 1983

PB-previous bill BF-balance forward CR-credit OP-off peak ES-estimate MI-minimum CO-combined EL-electric PY-payment

UI United Illuminating

statement

account number	demand KW			
11536-2350007	140.0	3/14	PB	5128.64
CONN EDUCATIONAL T V			PY	5128.64CR
6 VIDEO LANE				
SHELTON CT 06484				

service from	to	rate	reading	constant	kilowatt hours			
2/16	3/17	P2	40202	120	58200		EL	5580.31
FUEL COST ADJUSTMENT $.008660CR PER KWH								519.13CR

Any charge or portion thereof not paid within 28 days after mailing of the first bill containing such charge is subject to interest from such date to the date of payment

RATE OF INTEREST 1.25 PERCENT/MO.

MAR 1983 $5061.18

please pay this amount

Mailing date of this bill MARCH 21, 1983

PB-previous bill BF-balance forward CR-credit OP-off peak ES-estimate MI-minimum CO-combined EL-electric PY-payment

Fig. 10

POWER BILLS FOR MODIFIED TRANSMITTER OPERATING AT 55 KW

SHOWING 140 KW PEAK DEMAND

TRANSMITTER PERFORMANCE

Overall Plant Efficiency		43%
Plant Power Consumption		140 Kw
Visual Beam Efficiency		58%
Power Output		55 Kw
Beam Voltage		19.7 Kv
Visual Beam Amps	2.4 Amps Ea. Tube	
Aural Beam Efficiency	(Non H Tube)	37%
Aural Beam Amps		.76A
Aural Power Output	At Diplexer Output	5.5 Kw
Differential Gain		2%
Differential Ø		2°
Incidental Carrier Phase Modulation (ICPM)		1°
Low Frequency Linearity		7%
Sync Overshoot		3%
2T K Factor		2%

Fig. 11

ACHIEVABLE PARAMETERS FOR

UHF TRAMSMISSION SYSTEMS

Spencer J. Smith

Vice Pres. Engineering

Dielectric Communications

INTRODUCTION

In the past several years I have read, listened to and discussed many articles concerning waveguide transmission line, especially circular waveguide for UHF TV application. All of these articles seem to compare apples against oranges, oranges against pears, and pears against apples, but never all three at one time.

In preparing this paper, my goal was to compare all three at one time, both in a theoretical sense and a practical sense, actual measurements of the three systems, (coaxial, rectangular and circular). The data shown is actual measured data taken at the same time with three identical 500' runs of transmission line, one of each type. The formulae used in doing the theoretical calculations are given in Appendix A of the paper, however, since they are well known, they will not be a subject for this paper itself.

Many of us have gambled since coming here to Las Vegas this week, however, in deciding on a transmission line to use in your installation, I feel that the best way not to gamble is to have all the information available prior to making your decision. It is not my intent to recommend one transmission system over another only to compare the measurements vs. the theoretical calculations and do a relative cost overlay of the three systems.

The electrical measurements made on the transmission systems were done with an HP8505 Network Analyzer assembled on our test rack in Raymond, Maine. These measurements were run on non-optimized transmission lines to give the best possible point of comparison. The results are given and compared to the theoretical calculated values. Through optimization signifigant improvements can be made to some parameters, however, parameters such as insertion loss, group delay, wind load, etc., are inherent properties of a given style and size of transmission line.

THEORETICAL

In examining the three styles of transmission line which we will be discussing today, I tend to view these as having various degrees of freedom. Starting with coaxial line, we literally operate the transmission line from dc up to a magical fraction, (typically .95), of cut off for the TEM mode,one degree of freedom. In rectangular waveguide we have a low frequency restraint due to cut off of the dominant mode as well as a high frequency restraint with regard to the generation of the next higher order mode in the waveguide, two degrees of freedom. With circular guide, which is fairly new to the broadcast scene, the tendency has been to be a little on the greedy side and push into the twilight zone of higher order modes, namely the high end of the TE11 mode, and yet try to restrain the generation of the TM01 mode, three degrees of freedom.

The mechanical parameters of the transmission systems are pretty well defined through calculations. Such things as weight, wind load and size are usually given and not subject to measurements.

The calculated values cover a wide range of transmission lines, however, the actual measurements were made as follows:

The coaxial line used was 6-1/8" 75 ohm high conductivity copper transmission line. The waveguide used was WR1150 rectangular manufactured from 1100 aluminum. The circular waveguide was WC1050 manufactured from 1100 aluminum. The mechanical lengths of the sections in coax were 20 ft., the rectangular were 12 ft. and the circular were 12 ft. The mechanical parameters such as weight and wind load for these sizes and others are given in Table 1. No further comparisons other than those given in Table 1 will be made.

TRANSMISSION LINE	SIZE	LENGTH	WT/100' (lbs)	WINDLOAD/100' @ 50/33 (lbf)
COAXIAL	6-1/8	20	680	1700
WAVEGUIDE	8-3/16	20	925	2255
RECTANGULAR WAVEGUIDE	WR1150	12	575	4908
	WR1500	12	750	6366
	WR1800	12	1312	7670
CIRCULAR WAVEGUIDE	WC1050	12	540	2965
	WC1200	12	615	3377
	WC1400	12	717	3927
	WC1600	12	817	4477

Table 1. Mechanical Parameters

The electrical parameters to be considered are insertion loss, VSWR, and group delay. The parameters of insertion loss is a function of transmission line type, transmission line size and material. The main reason for its inclusion is to show a corelation between both calculated and measured values for various transmission line so that the end user will have a better understanding of his actual transmission line loss when he installs his system. VSWR in general is highly dependent on mechanical tolerances and mechanical authenticity of the transmission line system. It has been used as almost a sole measure of the quality of transmission line for many years. (This is not necessarily a good measure). Group delay is a parameter which will become more significant and more tightly controlled as broadcast systems progress into more elaborate modulation schemes and into transmission systems which do not have a flat group delay curve. Group delay is an important characteristic of any transmission system and is critical to its ability to transmit a signal with minimum distortion. Distortion results when the phase shift through a device is a nonlinear function of frequency. Group delay is a convenient indication of this nonlinear phase shift. If the phase shift through a device is a linear function of frequency, the group delay, ($-d\theta/d\omega$- see Figure 1), will remain constant and a signal can be transmitted without distortion.

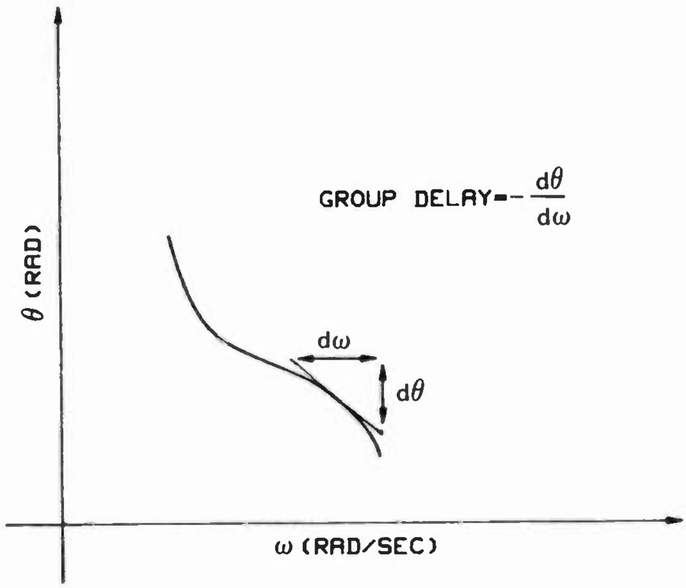


Figure 1. Phase vs. Frequency through a Transmission Device

INSERTION LOSS

A key parameter of any transmission system is the amount of power dissipation in the transmission line for a given input power to the system. This is normally expressed in dB levels. This reduces the problem to one of ratio of input power to power dissipated in the line. In the theoretical calculation only dissipative losses are calculated. These measure from the relative losses of the transmission system and manifest themselves in the form of heat generated in the transmission line. Figure 2 shows insertion losses for various types of transmission line. It should be noted that for the circular waveguide, the insertion loss curves become dashed at frequencies which support the TM01 mode and these particular waveguide sizes. Later these values will be compared to actual measured points of insertion loss for various cases of transmission line. In addition, Figure 2A shows the insertion loss of pure TEM mode circular waveguide for comparison sake.

GROUP DELAY

Figure 3 shows curves of group delay for various sizes and types of transmission line. It is quite obvious that for coaxial line the group delay is constant over the frequency range, while for the waveguide, both rectangular and circular, the group delay is a function of guide wave length vs. free space wave length. These variations in group delay will be compared to actual measured values later in the report. In addition, actual measurements of group delay will be shown for waveguide in which operation above the low frequency cut off of the next higher order mode is accomplished.

V S W R

Table 2 displays the VSWR capable for the various systems over their normal operating band of frequency. As in the other two cases, these values will be compared against actual measured values later in the report. It should be noted that these VSWR's are a product of manufacturing tolerances which are on the order of ±.005" or less on critical dimensions. Departure from the tolerance range will result in serious degradation of VSWR's for the transmission system. The second order effect of a VSWR degradation will be in higher than normal insertion loss for a transmission system even though the input VSWR looks to be within normal specifications. This is caused by a multiple of internal reflections which can be canceled out at the input. However, the loss generated by these multiple reflections will show up as insertion loss.

LINE LENGTH and FREQUENCY BAND	COAXIAL WAVEGUIDE	RECTANGULAR WAVEGUIDE	CIRCULAR WAVEGUIDE
0 - 1000'			
LOW BAND VHF	1.035:1		
HIGH BAND VHF	1.035:1		
UHF	1.035:1	1.05	1.05
1000' - 2000'			
LOW BAND VHF	1.04 :1		
HIGH BAND VHF	1.04 :1		
UHF	1.05 :1	1.06	1.06

Table 2. Typical VSWR Values

REFERENCES

1. "A Dual Channel Symmetrical Waveguide System", R. Fiore and M. Aitken, NAB; 1983
2. "Analytical and Experimental Data on Circular Waveguide for UHF TV", A Holtum, T. Greenway, T. Vaughan, NAB; 1983
3. "Fight Waveguide Losses 5 Ways", R. Cox and W. Rupp, MICROWAVES, August 1966
4. "High Power RF Systems", D. Sargent and S. Smith, NAB; 1981
5. "Waveguide or Coax for Five Megawatts", T. Vaughan, IEEE Transactions on Broadcasting, Vol. B.C.-16, Number 1, March 1970

Complete paper with measured data available from Dielectric Communications, Raymond, Maine 04071

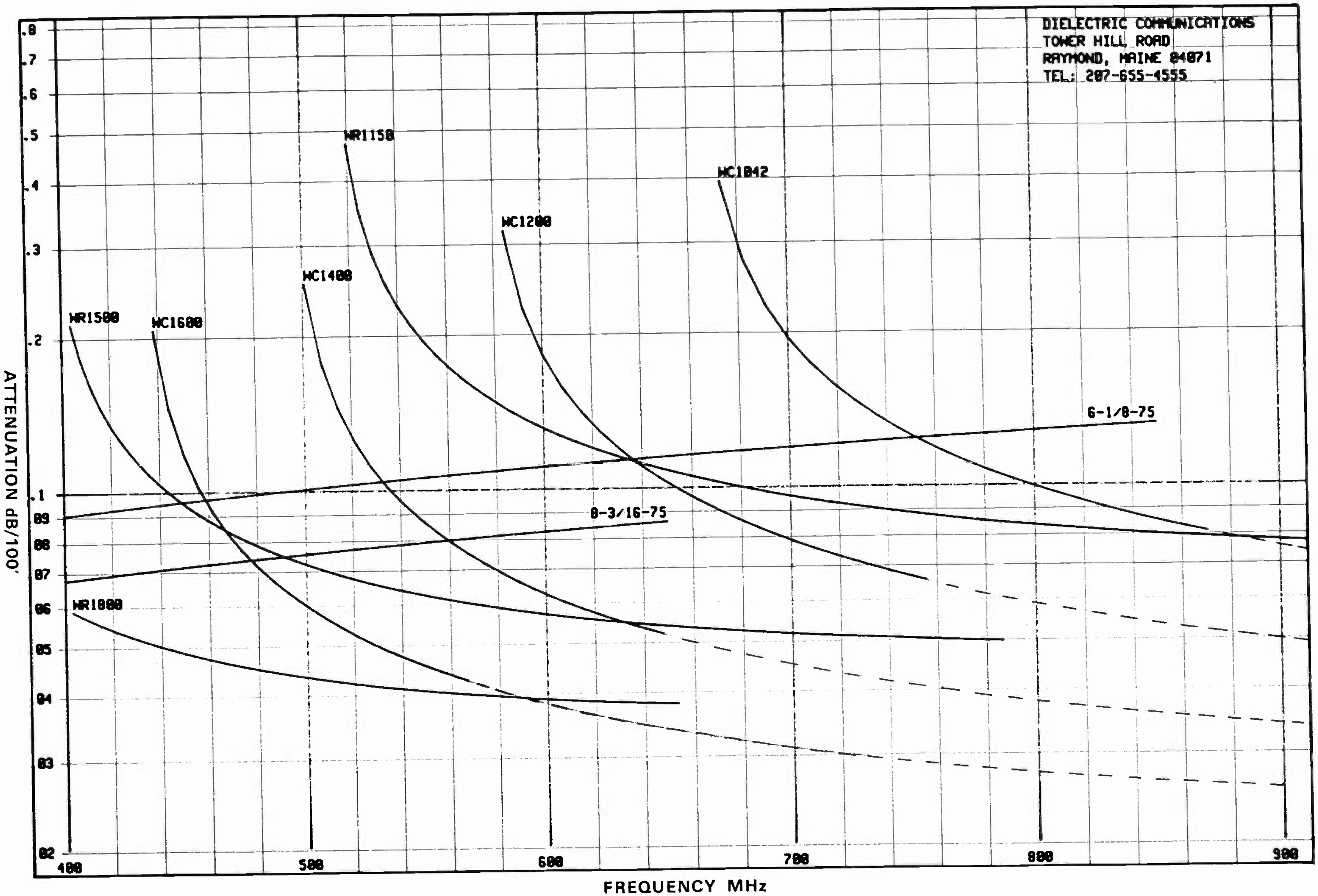


Figure 2. Normal Mode Attenuation for Various Transmission Lines

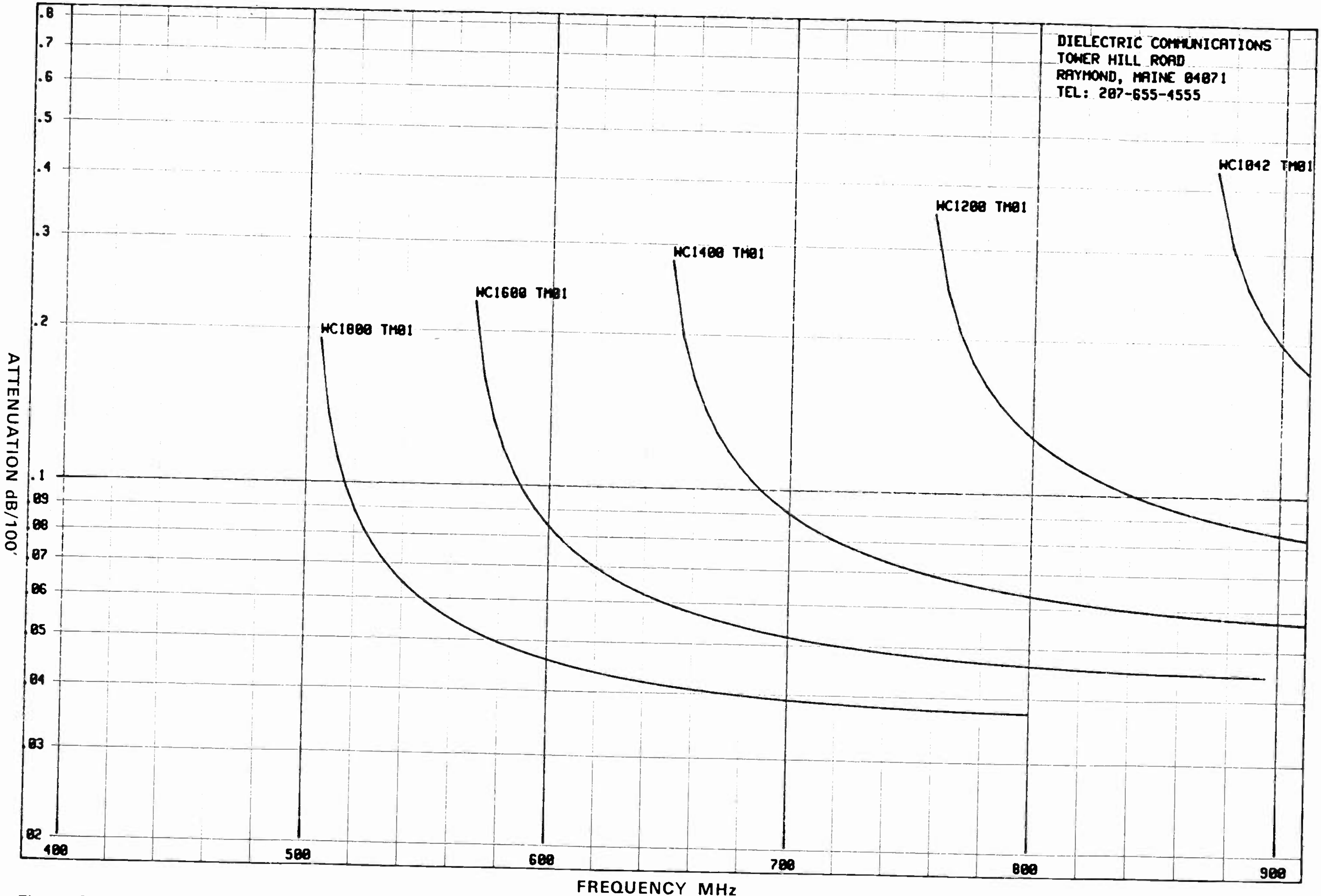


Figure 2A. TM01 Mode Attenuation for Circular Waveguide

UHF TABOO PERFORMANCE OF AN ADVANCED TECHNOLOGY TELEVISION RECEIVER

Hector Davis

Federal Communications Commission

Washington. D.C.

INTRODUCTION

The FCC's Sixth Report and Order (1). 1952, gives assignment principles for the Table of Assignments for the television service. It was apparently intended that the broadcast television service be subject to acceptable interference. For example, cochannel station assignments have to be separated by a minimum distance consistent with acceptable cochannel interference. Similar considerations of acceptable interference were used in establishing the minimum distance separations of adjacent channel assignments. Minimum distance separations for cochannel and adjacent channel station assignments apply to both VHF and UHF television. Minimum distance separations for other television channel combinations which apply only to UHF assignments are known as "UHF taboos". The UHF taboos would be expected to result in levels of television interference which are acceptable, like the cochannel and adjacent channel distance separations.

Certain performance characteristics of UHF television receivers were taken into account in establishing the UHF taboo distance separations. Since the taboos do limit UHF television channel usage, there could be considerable improvements in spectrum utilization if taboo-related receiver performance characteristics were improved.

In 1976. the Federal Communications Commission sponsored a contract for the development of an advanced technology television receiver. This receiver was to have improved performance relative to the UHF taboo performance of a group of television receivers tested in 1974 (2). The contract was awarded to Texas Instruments. Incorporated. (TI). The FCC (TI) receiver utilized new technology, including a new type of intermediate frequency amplifier. The improvements. if widely implemented. could lead to reevaluations of the UHF taboos. Although the FCC (TI) receiver was superior to off-the-shelf receivers in taboo-related performance characteristics. there were areas of its performance that could be further improved. particularly noise figure and internally generated spurious products (3,4).

In early 1979, the FCC awarded another contract for development of an advanced technology television receiver. The contract was awarded to RF Monolithics, Incorporated (RFM). In April 1982, the receiver was delivered to the FCC's Office of Science and Technology for evaluation (5).

RECEIVER DESCRIPTION

The FCC (RFM) receiver as tested was modified from a consumer-type receiver by the contractor. The consumer-type receiver was one of three color television receivers purchased by the FCC using the following criteria:

1) Contemporary (1979). solid state design
2) Electronic tuning
3) From the same production run. preferably with sequential serial numbers
4) Comb filter video for improved picture detail

Each of the receivers. Magnavox Model RK4245, was tested by the FCC Laboratory at the outset. Two of the three receivers were sent to the contractor. The third was kept at the laboratory as a control receiver. The contractor thus had one receiver to modify and one loaned for comparison. All three receivers are now at the FCC Laboratory.

A block diagram of the advanced technology receiver is given in Figure 1. Basically, the contractor modified the consumer-type receiver from its antenna terminals to its detector output. This output was interfaced to the receiver's existing circuitry for luminance, chrominance, synchronization, and sound signals.

In order to remove the UHF taboos associated with the intermediate frequency (IF) band of conventional receivers (41 to 47 MHz). the FCC (RFM) receiver employs a first IF of 444 to 450 MHz. A second IF of 41 to 47 MHz is used for low cost IF gain and pseudo-synchronous demodulator circuitry. Surface acoustic wave (SAW) filter technology for the first IF provides IF selectivity with low loss. Additional performance improvements result from the use of gallium arsenide varactor filters in the UHF tuner and from the use of a high performance double-balanced mixer.

BACKGROUND FOR THE DATA PRESENTATION

Television receivers perform well over a wide range of desired signal levels. The data reported here was obtained over the range of -65 dBm to -5 dBm. The value, -65 dBm, is taken to be representative of a lower end of acceptable service with respect to receiver noise.

The strongest desired signal level likely to be encountered by a television receiver is represented by -5 dBm.

For television to television signal interference, it is reasonable to consider -5 dBm to be representative also of the strongest undesired signal level likely to be encountered by a television receiver. Levels up to 0 dBm are given in the data presentation.

The test procedure has been previously described in some detail (2, 5). Typically, a desired television signal level is established at -65, -55, -45, -35, -25, -15, or -5 dBm. Then "just perceptible" interference is established with one or two undesired television signals, depending on the particular taboo being tested. Tests with two undesired signals were made with the signals at equal level. The undesired signals were translated off-the-air television signals. This maintained effects observable because of such characteristics as lack of frame synchronization and saturation changes in the programming of the undesired channel(s). The desired signal was video modulated with a 50% average picture level full-screen pedestal with color burst. Its aural carrier was unmodulated. This modulation yielded data agreeing within plus or minus four decibels with data obtained with picture modulation, according to previous tests.

UHF TABOO PERFORMANCE

IF Beat:

Quoting the FCC's Sixth Report and Order (1 at Paragraph 174),

> "It is recognized that when two stations in a city are separated by (a receiver intermediate frequency - Ed.) it is possible that the two signals will combine to provide a beat signal which will be picked up by the I.F. Amplifier. Where a 41.25 mc I.F. is in use, such signals may exist in channels which are separated by - - - eight channels from the desired station. The effect is similar to that of intermodulation. - - - "

Table 1 shows that the FCC (RFM) advanced technology receiver does not display interference for undesired television signal levels less than 0 dBm. This is because of the intermediate frequency system which the receiver employs. Television signals separated by eight channels from the desired channel are well attenuated by the FCC (RFM) receiver's UHF tuner and first IF, and apparently do not affect the receiver's second IF.

TABLE 1

UHF IF Beat*
Desired Ch. n. Undesired Ch. n+8 or n-8
Undesired Signal Levels in dBm for "Just Perceptible Interference"

	Desired Signal Levels in dBm						
	-65	-55	-45	-35	-25	-15	-5
RFM Rcvr	>0	>0	>0	>0	>0	>0	>0

*: Name applicable to conventional television receivers, but not the FCC (RFM) receiver for the channels used.

Intermodulation:

The UHF television intermodulation taboo is to account for the fact that undesired television signals can mix to produce a spurious interfering signal within a desired television channel. In the FCC's Sixth Report and Order (1 at paragraph 175), the Commission concluded that the best method of avoiding problems of UHF television channel intermodulation was to use a normal minimum separation of six channels in a city. A separation of two channels is allowed at VHF. UHF tuners are assumed to be more vulnerable than VHF tuners to intermodulation channel combinations.

Table 2 shows two channel separation UHF intermodulation data for the advanced technology receiver.

TABLE 2

UHF Intermodulation
Desired Ch. n. Undesired Chs. n+2, n+4
Undesired Signal Levels in dBm for "Just Perceptible Interference"

	Desired Signal Levels in dBm						
	-65	-55	-45	-35	-25	-15	-5
RFM Rcvr	-14	-8	-2	-3	>0	>0	>0

TABLE 2 continued

Desired Ch. n. Undesired Chs. n-2, n-4

Undesired Signal Levels in dBm for "Just Perceptible Interference"

	Desired Signal Levels in dBm						
	-65	-55	-45	-35	-25	-15	-5
RFM Rcvr	-12	-10	-6	-4	>0	>0	>0

The FCC (RFM) receiver is not totally free of interference for the intermodulation test condition. However. Table 2 and Figure 2 do indicate relative immunity to strong interfering signals for moderate and strong desired signal levels. The differences between the n+2, n+4 and the n-2, n-4 data are within the allowances for repeatibility of subjective observations.

Oscillator:

The UHF television oscillator taboo is intended to lessen the probability that the local oscillator of one television receiver will interfere with the reception of a desired channel by another television receiver. A receiver's local oscillator is used to convert a desired channel to the receiver's intermediate frequency (IF). The oscillator interference problem cannot occur at VHF because of the frequency separation between channel 6 and channel 7. VHF local oscillator frequencies do not fall wihin television channels. However, at UHF such interference can occur if one receiver is tuned to a channel seven channels below that to which a victim receiver is tuned. The UHF assignment table takes the standard IF into account to avoid interference which could result from assignments separated by seven channels.

Since the FCC (RFM) advanced technology receiver does not employ the standard intermediate frequency system, for UHF reception its local oscillator frequencies do not fall within television channels. However, it should be noted that there are other issues related to its local oscillator frequencies and intermediate frequency system (5).

Table 3 gives data for television signal to television signal interference for the FCC (RFM) receiver. The factors considered for the oscillator taboo for conventional receivers do not apply to the advanced technology receiver.

TABLE 3

"Oscillator"*
Desired Ch. n. Undesired Ch. n+7 or n-7
Undesired Signal Levels in dBm for "Just Perceptible Interference"

	Desired Signal Levels in dBm						
	-65	-55	-45	-35	-25	-15	-5
RFM Rcvr	>0	>0	>0	>0	>0	>0	>0

*: Name applicable to conventional television receivers. but not the FCC (RFM) receiver. The UHF taboo for n+7 and n-7 is based on local oscillator radiation and not the TV signal to TV signal channel combinations tested.

Image:

A desired channel, below a receiver's local oscillator frequency, is converted to the receiver's intermediate frequency (IF) band because the desired channel differs from the local oscillator frequency by the IF. Undesired signals as much above the local oscillator frequency as the desired channel is below will also be converted to IF and can cause interference if sufficiently strong. Such signals are called images.

For conventional television receivers, broadcast television signals do not cause image interference in the VHF band. VHF television channel images do not fall on television channels. For UHF, those channels which could cause such interference are fourteen and fifteen channels above the desired channel. The likelihood of interference has justified the UHF image taboos.

The undesired channel fifteen channels above the desired channel is called the picture image channel. The name is derived from the fact that its visual (picture) carrier, and not its aural (sound) carrier is in the image channel of a conventional television receiver. Similarly, the sound image channel, fourteen channels above the desired, is so-called because its aural carrier and not its visual carrier is in a receiver's image channel.

Tables 4 and 5 show that the FCC (RFM) receiver is, for all practical purposes, unaffected by the UHF image taboo channel combinations tested. This is due to the receiver's unconventional IF system. Its UHF television images do not fall on UHF television channels.

TABLE 4

UHF Picture Image*
Desired Ch. n. Undesired Ch. n+15
Undesired Signal Levels in dBm for "Just Perceptible Interference"

	Desired Signal Levels in dBm						
	-65	-55	-45	-35	-25	-15	-5
RFM Rcvr	>0	>0	>0	>0	>0	>0	>0

*: Name applicable to conventional television receivers.
but not the FCC (RFM) receiver for the channels used.

TABLE 5

UHF Sound Image*
Desired Ch. n. Undesired Ch. n+14
Undesired Signal Levels in dBm for "Just Perceptible Interference"

	Desired Signal Levels in dBm						
	-65	-55	-45	-35	-25	-15	-5
RFM Rcvr	>0	>0	>0	>0	>0	>0	>0

*: Name applicable to conventional television receivers.
but not the FCC (RFM) receiver for the channels used.

CONCLUSION

It is obvious from this presentation that advanced receiver technology such as used in the FCC (RFM) receiver could have significant impacts on the UHF taboos and could offer new opportunities for improved UHF spectrum utilization. The FCC report (5) from which much of the text and the data describing the FCC (RFM) receiver in this presentation was extracted states that the receiver is generally better than 90% of the conventional receiver data referenced for the following cases:

UHF television adjacent channel
Conventional receiver UHF picture image
Conventional receiver UHF sound image
UHF television cross modulation: n, n-2
UHF televiwion cross modulation: n, n+4
UHF television intermodulation: n, n+2, n+4
n, n-2, n-4

Also, compared to the same data base, the receiver was generally better than mean receiver performance for the following channel combinations:

UHF television cross modulation: n, n+2
UHF television cross modulation: n, n-4
Conventional receiver UHF IF beat: n, n-8
Conventional receiver UHF IF beat: n, n+7
Conventional receiver UHF IF beat: n, n-7

With regard to the conventional receiver IF beat combination, channel n with channel n+8, the FCC (RFM) receiver is generally better than only 10% of the data base. However, the mean receiver data exceeds 0 dBm, a signal level which is probably higher than those representative of broadcast television.

The FCC (RFM) receiver has other advantages, particularly with regard to upper and lower adjacent channel performance. Some potential problems have also been documented (5). The receiver is an important implement in analyzing the costs and benefits of various approaches to more efficient spectrum utilization.

The views expressed are those of the author and do not necessarily reflect the views of the Commission. Documents prepared in the course of, and publications relating to, an employee's official duties shall not be used for his private gain.

REFERENCES

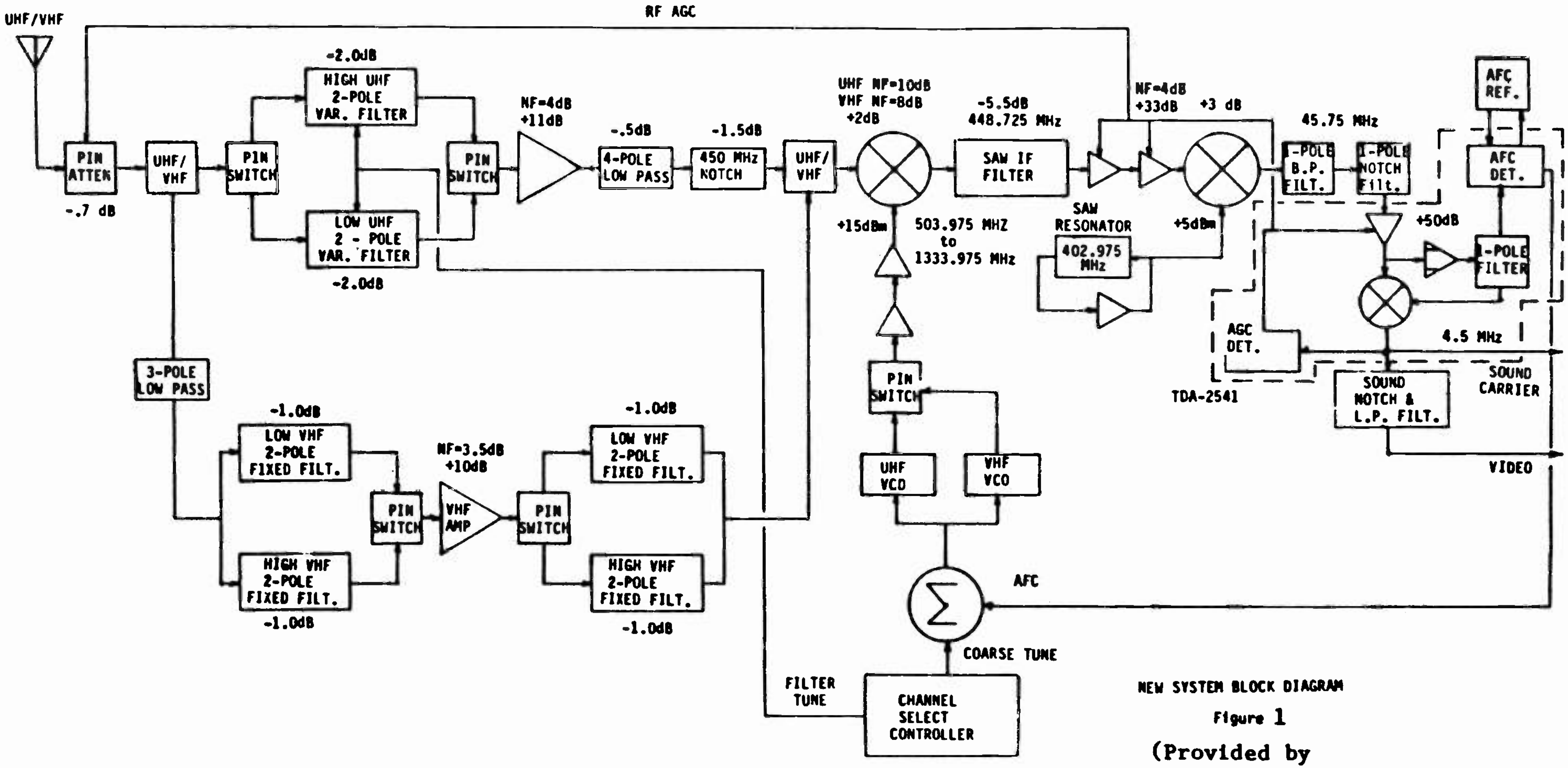
1. Sixth Report and Order. Docket Nos. 8736, 8975, 8976, 9175, Federal Communications Commission. April 11, 1952.

2. A Study of the Characteristics of Typical Television Receivers Relative to the UHF Taboos, W.K. Roberts. Laboratory Project Number 2229-63, Federal Communications Commission, June 1974, National Technical Information Service Number PB-235057.

3. A Study of the Characteristics of the FCC Prototype Receiver Relative to Conventional Receiver UHF Taboos, L.C. Middlekamp. Laboratory Project Number 2229-72, FCC/OCE LAB 78-01, Federal Communications Commission. February 1978.

4. High Performance TV Receiver, D.L. Ash and C.S. Hartmann, FCC/OCE CE78-01, Texas Instruments Inc., prepared for the Federal Communications Commission. March 1978, National Technical Information Service Number PB-277186.

5. Advanced Technology UHF Receiver Study, Part I. Receiver Performance Measurements, H. Davis, FCC/OST R-83-1, Federal Communications Commission, February 1983.

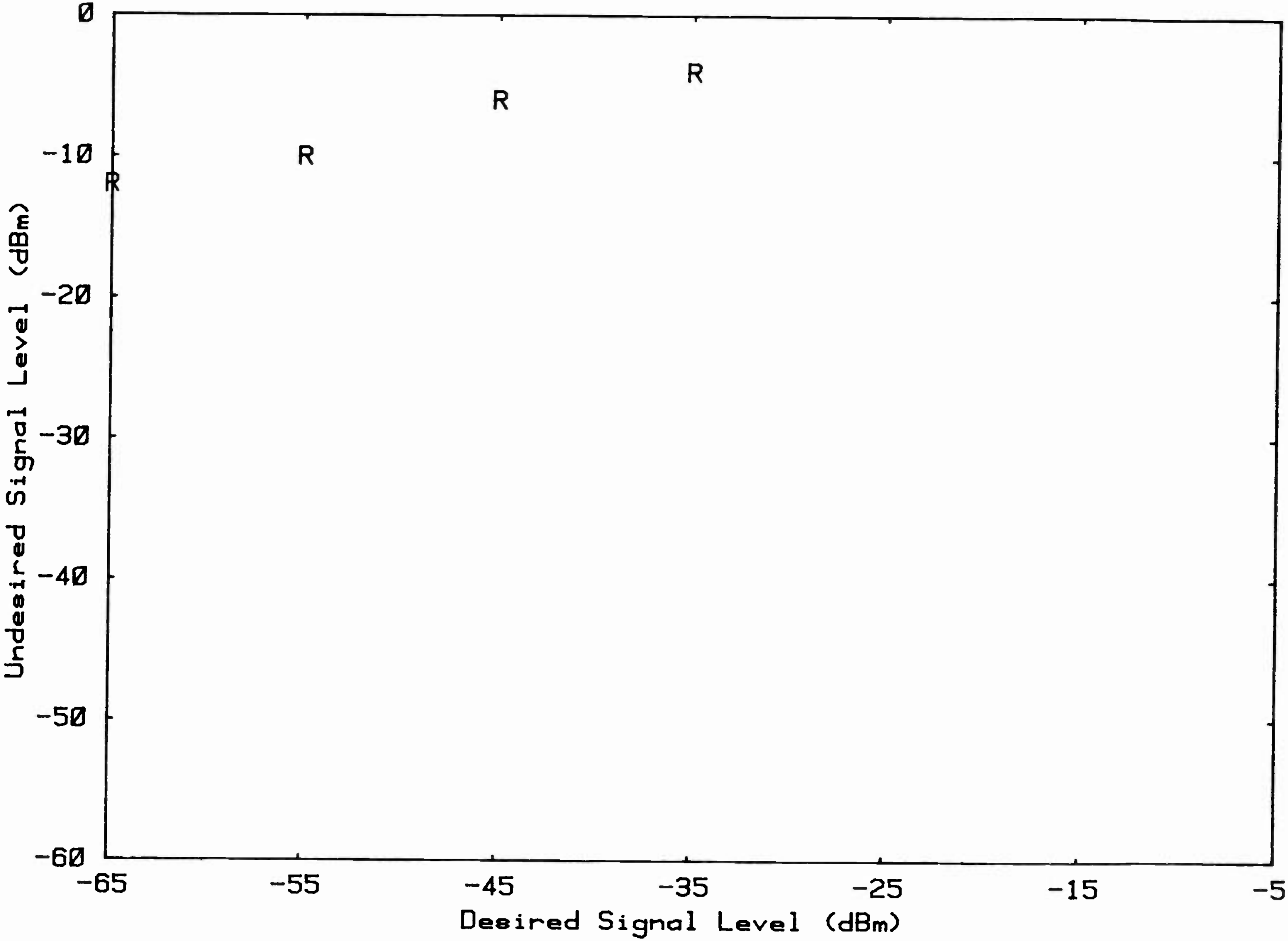


NEW SYSTEM BLOCK DIAGRAM

Figure 1

(Provided by
RF Monolithics, Inc.)

Figure 2: UHF Intermodulation, Ch n, n-2, n-4, R = FCC(RFM) Rcvr



Electromagnetic Energy Policy Alliance -

A Source of Facts

John M. Osepchuk, Ph.D.

Raytheon Research Division
Lexington, MA 02173

Abstract

A new organization has been formed to promote rational viewpoints and policies with respect to potential hazards of the electromagnetic fields associated with modern electrical/electronics technology. The historical background of this development shows that "soft science" and other factors had fed the spread of misinformation by the media about electromagnetic systems of all types and at all parts of the non-ionizing portion of the electromagnetic spectrum. Thus the new organization is an alliance of manufacturers and users ranging from utilities, VDT users, broadcasters, radar manufacturers, microwave-oven manufacturers, heat-sealer operators, amateur radio operators, etc. - the Electromagnetic Energy Policy Alliance (EEPA). This paper reviews the background which led to the formation of this Alliance and the plans of the Alliance in standards development, public education, research and above all, the development of a reliable source of facts in the murky field of biological hazards of electromagnetic fields.

Background and Need

After the passage of the Radiation Control for Health and Safety Act of 1968 (P.L. 90-602), there has been a steady buildup of concern about the hazards of exposure to electromagnetic fields - or non-ionizing radiation (D.C. to infrared) along with a buildup of bioeffect research in the United States (∿ \$10 - \$20 M/year). In addition, there has been a concomitant buildup of regulatory activity and media coverage. The latter has been mostly in the category of misinformation and in some cases of such mischievous nature as to deserve the description as "propaganda".

The first wave[1] of media misinformation was directed against "microwaves" based mostly on the ideas of Dr. Milton M. Zaret. Radar systems and microwave ovens were the particular targets of the time. This culminated in the simultaneous condemnation[2] of microwave ovens by Consumers Union and Senate Hearings[3] under Senator Tunney in 1973. At that time Dr. Zaret testified:

"There is a clear present and ever-increasing danger to the entire population of our country from exposure to the entire non-ionizing portion of the electromagnetic spectrum." Since then Dr. Zaret has attacked the whole spectrum from 60 Hz to infrared with emphasis on the "Hertzian" portion recently defined[4] by Dr. Zaret as roughly the range of 1 kHz to 100 GHz and with an emphasis on allegations of cataracts resulting from low-level exposure to Hertzian radiation. On the other hand, Consumers' Union has reversed itself and now promotes[5] the buying of microwave ovens.

The second wave[6] of media misinformation was stimulated by the book The Zapping of America[7] and the microwave irradiation[8] of the U.S. Embassy in Moscow. The author, Paul Brodeur, promoted his book by many appearances on radio and TV as well as in newspaper and magazine articles. The resulting spread of irrational fear of "microwave/RF" radiation no doubt led to lawsuits like the no-merit decision[9] that very-low level (unmeasurable by hazard meters) microwaves caused brain disease or the many environmentalist protests of systems like the large Air-Force radar,[10] PAVE PAWS in Cape Cod, Massachusetts or even innocuous low-power, microwave relay towers. In one case, the City of New York delayed for 2 years a part of a Coast-Guard system - an action which was strongly condemned[11] by Dr. Merrill Eisenbud because there was no real microwave hazard. Much literature and Congressional hearings resulted in the several years after Brodeur's book.

In the last five years a third wave of media misinformation has joined the entire non-ionizing range of the electromagnetic spectrum as if in response to Dr. Zaret.[3] Thus the fears associated with 60 Hz power lines and microwaves have been joined. Numerous articles[12-15] have increasingly been directed toward the whole non-ionizing radiation (NIR) spectrum or electromagnetic[14] or electrical[15] radiation. A small minority of research scientists supply these articles with questions on safety of 60 Hz or microwave systems. Speculations[16] about "non-thermal" effects are spread in the technical literature as well. Suspicions[17] have been raised about a link between leukemia and occupational exposure to electromagnetic fields. A related phenomenon is the spreading fear[18] that some type of electromagnetic radiation from video-display terminals causes health hazards.

Various lawsuits in the areas of product liability and siting disputes have resulted. At least two review articles[19,20] have been prepared to aid plaintiff's lawyers. Rational exposure standards[21] have been continually developed by voluntary bodies. Although these represent the consensus of the U.S. scientific community, they have not led to uniform preemptive rational standards, either for the environment or in occupational areas. Because federal standards appear[22] somewhat distant some states[23] have developed environmental regulations.

Impact on Broadcasting

During the first wave of propaganda, broadcasters felt little impact since the attack was directed towards "microwave" sources which broadcasters felt[1] were disassociated from broadcasting. It soon became clear, however, that there was little distinction between radio frequency or microwave sources in the bio-effect literature. In fact, it had become accepted that for humans the most penetrating radiation would be in the resonance regime[24] of man, i.e., roughly 30 to 300 MHz. This is, of course, the range of FM radio and VHF TV broadcasting.

Thus broadcasting sources have been included as targets in the second and third waves of media misinformation. FM, TV and microwave relay towers were implicated[25] in speculative association with increased cancer rates near Syracuse, New York. This has led to delays in new transmitter tower construction.

A proposed environmental standard[26] of 50 $\mu W/cm^2$, independent of frequency, was interpreted[27] as a serious potential disruption of broadcasting in New York City. Although this proposal was rejected[28] by an advisory committee for the New York City Commissioner of Health, it probably encouraged later controversy[29] over the safety of the broadcasting antennae at the World Trade Center.

Later the safety of broadcasting towers was attacked[30] in Oregon. Stringent environmental standards[30] equivalent to or tighter than that[23] of Massachusetts were set by local authorities near Portland. This precipitated a pilot epidemiological study (under EPA contract) by a local university professor that has been publicized[31] as linking TV towers with cancer.

Soon there were scattered protests[32] of microwave satellite transmitting dishes used by broadcasters and even receiving dishes[33] in relay links for cable TV operators. These protests, significantly, occur predominantly in sophisticated suburbs in the Northeast.

It is clear that any potential source of broadcast radiation could be attacked by local groups even though the reasons may be truly other than concern for health. The latter issue is a convenient one for local citizens to use in opposing construction or even present uses of broadcast technology. It is not inconceivable that protests could be extended to ENG (electronic newsgathering) systems, mobile radio and even amateur radio.

In recent years various broadcasting groups have become aware of the general problem of RF radiation in the environment, public perception and appropriate mitigation and management policies for broadcasters. An extensive study by Clemmensen[34] was funded by the National Association of Broadcasters. Her conclusions are that although uniform preemptive standards may be an ideal solution, this should not be assumed until an adequate cost-benefit study is made by industry. In the interim, she recommends that users such as broadcasters give serious attention to techniques of "issue management."

Need for Public Education

Although all parties concerned should devote serious attention to bioeffect research, it can be shown that existing environmental fields, including those from broadcasting, are far below safety standards of any official entity in the world. Broadcasters should be especially aware of this fact since broadcasting is the principal source[35] of environmental RF radiation.

Why then does public controversy exist over microwave/RF sources? The answer is in the main that the general public is confused and even frightened by the misinformation or even mischievous "propaganda" that has characterized the coverage of "microwave/RF" radiation in the media. The ideal solution is public education, but is it practical?

The public often confuses "microwave," "RF" or "electrical" radiation with ionizing radiation. This is a crucial problem because the latter subject is fraught with philosophical difficulties[36,37] that lead from the so-called linear dose-response hypothesis, i.e., that there is no threshold for effects of ionizing radiation. In fact, for non-ionizing radiation there is universal agreement that thresholds for effects and hazards do exist and therefore safety thresholds can be set below which people are safe!

Thus, Dr. Czerski, a leading researcher in Eastern Europe, describes[38] the basis for the most conservative safety exposure limits as "the principle of 'zero' interaction: This level is safe; no effects are demonstrable."

In a key Soviet book[39] on Microwaves and Human Safety, it is stated: "The data from biophysical and medical investigations which are now available make it possible to conclude that the principal pathological changes arise during continuous prolonged irradiation by a field with an intensity 1.... 10 mW/cm^2 or more.. Accordingly, exposure to a field with an intensity of about a few or even tens of microwatts per square centimeter cannot be regarded as in any way dangerous."

There are safe levels for microwave radiation. When the radiation levels in the U.S. environment are compared with ideas on safety limits, without exception potential exposures from existing environmental fields, accessible to the general public, are well below even limits set by Eastern Europe, especially if exposure duration is taken into account.

In addition to the scientific issue which leads to safety levels, and the issue of propaganda (or misinformation) there is also the philosophical issue, i.e., can one prove absolute safety? Even extensive long-term exposure studies will not produce absolute proof. The latter is impossible but it also applies to most things in life, e.g., printer's ink on newspapers, but it should not be a cause for alarm or even concern if this subject is viewed rationally.

The misperception[40] of risks of electromagnetic radiation is real and causes wasteful scenarios like the litigation[40] over the use of a 35 mW microwave source as a tower for antenna tests. Some information about these misperceptions were gained by the author in a pilot survey. Table I is one result of the survey. Note that the word "radiation" itself serves as a red flag in forming perceptions of risk while "broadcast" is still relatively benign.

Although education seems to be a solution for public misperceptions there are those who disagree - claiming that it is impossible, hopeless or even counter-productive in that it raises the awareness level of the public. They counsel that left unanswered, such propaganda will eventually die out. One wonders, however, if this was not the attitude adopted by utilities in dealing with their radiation issue until recently, concluding[41] that a $30M program of public education is desirable.

The practical importance of the public misperception problem cannot be neglected. Recently a plaintiff's lawyer pointed out[42] that if there is little danger from RF radiation sources, then industry has done a poor job of public education because as he perceives it, there is a considerable reservoir of fear and distrust of this technology among the general public - from which juries are chosen.

For simple logical reasons professional engineers within IEEE formed[43] the Committee on Man and Radiation (COMAR) in order to deal with misperception and to conduct public education projects. Thus COMAR rebutted[44] Consumers'

TABLE 1. Results of Survey of Risk Assessment for Environmental Radiation

Rank	Physical Entity	Risk Index
1	Radiation source	3.09
2	Microwave-radiation source	3.09
3	Uses ionizing radiation	3.00
4	RF radiation source	3.80
5*	ELF source	2.77
6	Source of non-ionizing radiation	2.72
7	Broadcast tower	2.68
8	Microwave relay tower	2.62
9	Uses microwave energy internally	2.57
10	High-voltage line	2.24
11	Generates much heat energy	2.14
12	Power line	2.00
13	Broadcast studio	1.50
14	Uses arc welding	1.18

Risk index - Average score of those surveyed (20) per assignment of scores to answers

Score	Answer
5	Oppose its presence in my town.
4	Request town boards to study risk before acceptance
3	Request town to institute controls as conditions of acceptance.
2	Try to find out more before deciding.
1	Support its acceptance.

*Risk assessment not completed by 30% of those surveyed.

Union when it attacked safety of microwave ovens, successfully processed an IEEE position paper[45] on the subject, published a reprint volume[46] and is working on an educational film for television.

Still there are skeptics[47] and it must be admitted that COMAR has been of limited effectiveness, both in timing and in strength of its actions, e.g., it has been even ineffective in preventing misinformation from being published in the general-interest magazine Spectrum which is subsidized by IEEE. Clemmensen claims[47] few people are interested in the subject - yet real problems of litigation for alleged radiation injury and opposition to transmitting towers continue.

History of the Founding of the Electromagnetic Energy Policy Alliance (EEPA)

For some time manufacturers and users have perceived that problems of siting or litigation related to alleged hazards of low-level electromagnetic radiation are in large part the result of an "undirected media campaign"[48] which spreads misinformation and/or outright propaganda. The media is fed by the results of "soft" science, i.e., work which ultimately cannot be replicated or suffers from ambiguity, e.g., a common feature of a small pilot epidemiological study. One example is the failure[49] to replicate earlier reports[50] that low-level pulsed microwaves can stop the heart of a frog. Another example is the reported[51] association of paternal radar exposure with the parenting of children with Down's syndrome. Less well known is the fact that this association was not found in a much larger (and more expensive) follow-up study.[52]

A tiny minority of the scientific community is willing to support lawsuits alleging potential or actual hazard from low-level electromagnetic radiation, principally on the thesis (potentially self-serving) that until much more research is done, one cannot be certain about the safety of any level - particularly if there are as yet undiscovered new mechanisms of interaction as well as windows[53] in frequency, power level, time and even modulation index. Armed with such pernicious concepts intervenors have been successful in halting earth satellite terminal construction[54] and even an important ELF defense communication system.[55]

As early as 1974 microwave-oven manufacturers through the Association of Home Appliance Manufacturers (AHAM) found it desirable to have a special educational seminar at the University of Washington in order to allow internal and external legal counsel to become informed. The truism that legal counsel, as well as the general public, can be miseducated by the media was increasingly recognized so that by 1980 internal seminars were held for lawyers and other non-specialists. In 1981 personnel from RCA, Bell Labs and Raytheon organized a seminar at the Homestead, Virginia for the purpose of such education with the help of leading scientific experts such as Professors Guy, Michaelson and Justesen; Drs. Pollack, Appleton, Hirsch and others. In 1982 this seminar was repeated under the cooperative sponsorship of three trade associations - the National Association of Broadcasters (NAB), Electronic Industries Association (EIA) and AHAM and the key efforts of Messrs Umansky, Bennett and Weizeorick respectively. By then the scope was broadened to appeal to those involved with the VDT "radiation" problem as well as 60 Hz. The organizers of the 1982 seminar reviewed the concept of an "Alliance" as a more effective means of fighting misperception and the other negative aspects of the radiation problem. This was unanimously accepted by those attending the 1982 Homestead Seminar and there was created an Organizing Committee of lawyers, engineers and executives. (The membership is listed in Appendix A.) This committee met seven times during 1983 in Washington, DC and prepared the various documents (e.g., by-laws), office and facilities plans and membership plans required for founding the organization (as well as the choosing of a name). (See. Fig. 1.)

On February 2, 1984, executive representatives of eight founders met with the Organizing Committee in Washington and officially launched the Electromagnetic Energy Policy Alliance. The founders are AT&T, GTE, MCI, Motorola, NAB, RCA, Raytheon and Rockwell/Collins. The members of the initial Board of Directors are listed in Appendix B. (See Fig. 2.)

Plans for EEPA

The Electromagnetic Energy Policy Alliance was organized in response to the controversy over the public and occupational health implications of exposure to non-ionizing electromagnetic energy. Alliance organizers "believe that fear of electromagnetic energy is based primarily on misinformation, misunderstanding and misconceptions" and therefore that "manufacturers and users must act positively in the areas of public policy, regulation, research and education to preserve the health of the U.S. electromagnetic technology and its beneficial uses for society."

According to its articles of incorporation, the Alliance will serve the following specific purposes:

1. To promote the establishment of responsible uniform national standards related to the use of non-ionizing electromagnetic energy;

2. To conduct and/or sponsor research to develop a more thorough understanding of the effects of non-ionizing electromagnetic energy on humans and other living organisms;

3. To formulate policies and to develop educational programs designed to promote an informed public understanding of the benefits and risks associated with the use of non-ionizing electromagnetic energy;

4. To develop and maintain an information resource system with respect to the uses, effects and governmental regulation of non-ionizing electromagnetic energy; and

5. To act in an advisory capacity to government and the public on the technical, scientific and other relevant aspects of the production, use and effects of non-ionizing electromagnetic energy.

Members are expected to include electric utilities, broadcasters, appliance and electronic equipment manufacturers, radar and mobile radio communications companies, and manufacturers and users of video-display terminals and other equipment utilizing electromagnetic energy across the frequency spectrum from 50/60 Hz to microwave.

As EEPA is formed and grows, technical committees as well as standing committees will be formed. The technical and legal staffs of the founders and initial members, in fact, represent a prime asset of the Alliance, i.e., a source of facts based on their expertise and involvement with government agencies and the scientific community over many years. In the past, when manufacturing or user organizations encountered a "radiation" (non-ionizing) problem, they have been genuinely confused by the media or even trade publications (which are also a version of the media). Successful approaches to their problems often have begun only after having made contact with other organizations experienced in these problems, i.e., "issue management" as denoted by Clemmenson.[34] The Alliance will be a more effective way of sharing their expertise for the ultimate benefit of society.

In 1985 the first annual meeting of EEPA will take place. It is hoped that a broad array of members will then exist. Certainly all broadcasting organizations are urged to seriously consider membership as a means of staying currently informed and supporting activities that will underlie positive steps to a rational policy by society toward electrical/electronic technology.

APPENDIX A

Organizing Committee

ELECTROMAGNETIC ENERGY POLICY ALLIANCE

Name	Affiliation
John M. Osepchuk, Ph.D., Chairman	Raytheon Co.
Kent Anderson	AHAM
Peter Bennett	EIA
Charles Buffler, Ph.D.	Litton Microwave
Jules Cohen	Jules Cohen & Assoc., P.E.
Shirley Fujimoto, Esq.	Keller & Heckman (UTC)
Robert Harvey, Esq.	Nixon, Hargrave, Devans & Doyle
Howard Johnson	Consultant
Tom Keller	NAB
George Kiessling	RCA
John Lyons, Esq.	Motorola
John Mitchell	USAF School of A.S. Med.
Ron Petersen	Bell Labs
Barry Umansky, Esq.	NAB
William Van Gemert, Esq.	Raytheon Co.
John Wald, Esq.	Litton Microwave
Don Walker	Motorola
Max Weiss, Ph.D.	Bell Labs
John Weizeorick	AHAM
Michael H. Ladd	GTE
David D. Davidson, Ph.D.	GTE
James T. Carter	Rockwell/Collins

APPENDIX B

INITIAL BOARD OF DIRECTORS (EEPA)

(February 1984)

Organization	Director
AT&T	To be named
GTE	John A. Whittaker, Vice President, Govt Affairs
Motorola, Inc.	Mort Topfer, Vice President
MCI	Dr. Daniel Walters, Vice President and Chairman Pro-Tem, EEPA Board
NAB	Edward O. Fritts, President
RCA	Howard Rosenthal, Staff Vice President, Eng.
Raytheon	Dr. Joseph F. Shea, Senior Vice President, Eng.
Rockwell/Collins	Dr. Harold Sobol, Director R&D

Members of the EEPA Organizing Committee at the February 2, 1984 Founding meeting. Left to right, Thomas B. Keller, Michael Ladd, James T. Carter, Jr., John Weizeorick, George Kiessling, Kent Anderson, Shirley Fujimoto, William Van Gemert, Jules Cohen, John M. Osepchuk, Max Weiss, Ron Petersen, Barry Umansky, Don Walker, John Lyons, Robert Harvey.

Members of the initial Board of Directors at the February 2, 1984 founding of EEPA; left to right, Dr. Harold Sobol, Howard Rosenthal, Dr. Daniel Walters, Mort Topfer, Edward O. Fritts.

References

1. John M. Osepchuk, "Microwaves in the Media: Truth or Consequences," Digest IEEE Int. Microwave Symposium, p. 11 (1979).
2. Consumers' Union, "Microwave Ovens; Not Recommended," Consumer Reports Magazine (April 1973).
3. U.S. Senate Record of Hearings on "Radiation Control for Health and Safety Act of 1968," (March 8, 9, and 12, 1973). U. S. Govt Printing Office Document 96-087.
4. Deposition of Dr. Milton M. Zaret, p. 58, relative to Baugh et al. vs Raytheon Co. (Dec. 6, 1983).
5. Consumer Reports, p. 584 (November 1983).
6. Joseph F. White, "Were You Zapped?" Microwave J., p. 16 (January 1981).
7. Paul Brodeur, The Zapping of America, W. W. Norton Co., New York (1978).
8. Herbert Pollack, "The Microwave Syndrome," Bull. N.Y. Acad. of Medicine, Vol. 55, p. 1240 (Dec. 1979).
9. Yannon vs N.Y. Telephone Co., N.Y. W.C.B. Nos. 07142308, 07523602 (1980).
10. UPI, "Suit over PAVE PAWS Rejected," Boston Globe (Dec. 7, 1979).
11. Merrill Eisenbud, "An Overview," Bull. N.Y. Acad. Medicine 55, p. 1274 (Dec. 1979).
12. Susan Schiefelbein, "The Invisible Threat," Saturday Review (Sept. 15, 1979).
13. Lowell Ponte, "The Menace of Electric Smog," Readers Digest, p. 65 (Jan. 1980).
14. Robert Gannon, "Electromagnetic Pollution - Are They Zapping You?" Popular Science, p. 98 (Dec. 1983).
15. H. McCann, "Is Electricity a Health Risk?" The Detroit News, p. 1D (Nov. 17, 1983).
16. Eric Lerner, "RF Radiation: Biological Effects," IEEE Spectrum, p. 51 (Dec. 1980).
17. Samuel Milham, "Mortality From Leukemia in Workers Exposed to Electrical and Magnetic Fields," New England J. of Medicine, Vol. 307, p. 249 (July 22, 1982).
18. Jeff Sorensen and Jon Swan, "VDTs: The Overlooked Story Right in the Newsroom," Columbia Journalism Review, p. 32 (Jan./Feb. 1981).
19. Karen A. Massey, "The Challenge of Non-Ionizing Radiation: A Proposal for Legislation," Duke Law J., Vol. 1979:105, pp. 105-189 (1979).
20. Pamela J. Laquidara, "Litigating Non-Ionizing Radiation Injury Claims: Traditional Approaches to a Contemporary Problem," Boston College Environmental Affairs Law Review, Vol. 10:963, pp. 965-1018 (1982/1983).
21. ANSI C95.1 - 1982, Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields, 300 kHz to 100 GHz, American National Standards Institute, New York, N.Y. (1982).
22. Microwave News, Vol. III, No. 8 (Oct. 1983).

23. 105 CMR 122:000, Regulations Governing Fixed Facilities which Generate Electromagnetic Fields in the Frequency Range of 300 kHz to 100 GHz and Microwave Ovens, Massachusetts Register Issue No. 379, p. 1 (Sept. 1, 1983).

24. Om P. Gandhi, "Dosimetry - The Absorption Properties of Man and Experimental Animals," Bull. N.Y. Acad. of Medicine, Vol. 55, p. 999 (Dec. 1979).

25. R. O. Becker, "Microwave Radiation," New York State J. of Medicine, Vol. 73, p. 2176 (Nov. 1977).

26. Leonard K. Solon, "A Local Health Agency Approach to a Permissible Environmental Level for Microwave and Radio Frequency Radiation," Bull. N.Y. Acad. Medicine 55, p. 1251 (Dec. 1979).

27. E. Holsendoph, "Proposed New York 'Health' Rule May Force Radio, TV Out of City," Washington Star (Sept. 26, 1978).

28. Merrill Eisenbud, Chairman, Radiation Advisory Committee, Report to the Commissioner of Health, City of New York, "Consideration of the Proposed New York Microwave Standards," (April 12, 1979).

29. "At the World Trade Center - Confusion and Controversy in New York City," Microwave News, Vol. I, No. 2, p. 6 (March 1981).

30. "Curbs on Radiation from Broadcasting in Oregon," Wall St. Journal, p. 30 (July 7, 1982).

31. Jim Dawson, "Study Alleges TV Tower Beam Programs...and Cancer," Minneapolis Star, p. 1C (Jan. 26, 1982).

32. "Danger from Microwaves," Stamford (CT) Advocate, p. A10 (May 13, 1982).

33. Matt Kramer, "Making Waves," TAB, Newton, Mass. (Dec. 22, 1982).

34. Jane M. Clemmensen, "Non-Ionizing Radiation: A Case for Federal Standards," Univ. of California/Berkeley (Sept. 1983).

35. David E. Janes, Jr. "Radiofrequency Environments in the United States," Proc. IEEE Int'l Communications Conference, p. 31.4.1 (1979).

36. Helen M. Caldicott, Nuclear Madness, Autumn Press, Brookline, Mass. (1978).

37. John W. Gofman, An Irreverent Illustrated View of Nuclear Power, particularly p. 30; Committee for Nuclear Responsibility, San Francisco, Calif. (1979).

38. S. Baranski and P. Czerski, Biological Effects of Microwaves, p. 183, Dowden, Hutchinson, and Ross, Inc., Stroudsburg, Pa. (1977).

39. B. A. Minin, Microwaves and Human Safety, Moscow, JPRS-65506-2, p. 327 (1974).

40. John M. Osepchuk, "Dealing with Misperception: On the Costs of Perceived Versus Real Risks of Microwave Radiation," p. 97 in Nicholas Steneck, Ed. Risk/Benefit Analysis: The Microwave Case, San Francisco Press (1982).

41. Ian Fells, Book Review (Before It's Too Late by Bernard Cohen, Plenum Press, N.Y. (1983);) New Scientist, p. 752 (Dec. 1983).

42. Marc Moller, "Litigation Implications of Microwave Exposure and Personal Injury Claims," 18th Annual Microwave Power Symposium (IMPI), Philadelphia, Pa. (July 20, 1983).

43. Don R. Justesen, "The Committee on Man and Radiation: A commitment to the Public Interest," Bull. N.Y. Acad. Medicine, Vol. 55, p. 1267 (Dec. 1979).

44. COMAR Statement on Consumer Reports (1973), Article on Microwave Ovens, IEEE Spectrum, Vol. 11, p. 24 (April 1974).

45. IEEE Position Paper, "Human Exposure to Microwaves and Other Radiofrequency Electromagnetic Fields," (Jan. 16, 1982).

46. John M. Osepchuk, Editor, Biological Effects of Electromagnetic Radiation, IEEE Press, N.Y. (1983).

47. Reference 34, pp. 3-16.

48. John M. Osepchuk, "The Conning of America," Microwave J., p. 18 (Oct., 1978).

49. R. M. Clapman and C. A. Cain, "Absence of Heart-Rate Effects in Isolated Frog Heart Irradiated with Pulse-Modulated Microwave Energy," J. Microwave Power, Vol. 10, p. 411 (Dec. 1975).

50. R. L. Adams and R. A. Williams, Biological Effect of Electromagnetic Radiation, Report DST-18103-074-76, Defense Intelligence Agency (March 1976).

51. A. T. Sigler et al., "Radiation Exposure in Parents of Children with Mongolism (Down's Syndrome)," Bull. Johns Hopkins Hosp. Vol. 117, p. 374 (1965).

52. B. H. Cohen et al., "Parental Factors in Down's Syndrome - Results of the Second Baltimore Case-Control Study," in Population Cytogenetics, Studies in Humans, E. B. Houk and I. H. Porter, Editors, p. 301, Academic Press, N.Y. (1977).

53. "Time Windows in Bioeffects Research," Microwave News, Vol. III, No. 6, p. 1 (July/Aug. 1983).

54. "RCA's Satcom Station Blocked Again in Seattle Area," Microwave News, Vol. II, No. 9, p. 2 (Nov. 1982).

55. "Federal Judge Halts Construction of Naval Communications Project," Washington Post (Feb. 1, 1984).

A Review of Auxiliary Broadcasting

Michael C. Rau

National Association of Broadcasters

Washington, D.C.

Broadcasters use auxiliary broadcast services for a multitude of essential purposes. With recent deregulation, the number of potential uses is expanding rapidly. Yet, while there has been inexorable growth in the number of AM, FM and TV stations, there has not been a corresponding growth in the allocated frequencies for auxiliary services; in many broadcast markets, these frequencies are congested and further growth is questionable, if not impossible. Pressure from competing radio services, such as private, land mobile, government, cable and other services, make it unlikely that, in the future, additional spectrum for auxiliary services will be allocated. It is therefore imperative that broadcasters make efficient use of the spectrum they have now. To do this, it is first necessary to understand what spectrum is allocated and available and how it is used; what portion of available spectrum is at present most controversial and why; the extent that cost efficient technology can alleviate existing spectrum shortages and assist broadcasters; and recent FCC rule changes of a procedural or policy nature that impact on auxiliary spectrum use.

The purpose of this paper is to concisely address these issues.

I. Auxiliary Broadcast Allocations; Radio Broadcasting

Radio stations use auxiliary broadcast facilities for studio to transmitter links (STL's), remote pickup units (RPU's) and on occasion, intercity relay links (ICR's). STL's carry the live on-the-air programming, remote metering, control telemetry, and any subsidiary communications from the studio location to the actual broadcast transmitting facility. STL stations are operational whenever the broadcast station is on the air, often a full 24 hours. RPU's are mobile facilities that are used to also transmit live on-the-air programming from a temporary remote location such as a shopping center or football game to the station's studio facilities for taping, for later rebroadcast, or for on-the-scene-reports incorporated into the actual on-going live broadcast. Radio stations typically have two, three or more RPU's that may be licensed to one or more frequencies. ICR stations are used to circumvent obstacles in an otherwise clear path from studio to transmitter link. ICR's generally aid the broadcaster's capability to relay programming from studio location to transmitter site, even though the terrain and buildings in between do not allow a clear path. In these cases, the ICR generally is located on the highest point on the path between studio and transmitter. Typically, ICR's transmit on different frequencies than they receive; however, recently the Commission has authorized use of "microwave boosters" which retransmit on the same frequency as they receive. Because STL's, RPU's and ICR stations frequently, if not always, carry live on-the-air programming, they must be protected from undesired interference from other telecommunications allocations which may be sharing their frequencies or operating in adjacent spectrum. Absent such protection, interference received by operating STL stations can destroy programming intended for the general public's reception. Moreover, any interference or interruption to the STL transmission, or RPU or ICR transmission, is immediately present on the actual broadcast and is present everywhere within the broadcast station's service area. In the case of STL's, proper operation is so critical to the technical integrity of the broadcast, that many stations purchase and erect "hot stand-by" STL stations which immediately begin transmitting (on the same frequency) if the "main" STL station fails.

Available Frequency Allocations

1. <u>1605-1705 kHz</u>: These frequencies are available for mobile remote pickup broadcast stations. There are three channels available: 1606, 1622 and 1646 kHz. At 1606 kHz, use is subject to the condition that no harmful interference is caused to the reception of AM stations operating on 1600 kHz. The entire band, however, is subject to modification at a 1985 Regional Administrative Radio Conference (RARC), which will consider the establishment of technical standards for AM broadcast service from 1605-1705 kHz. Also, at 1610 kHz, many local government entities operate "Travelers' Information Stations" or "TIS".

2. <u>25.67-26.48 MHz</u>: There are a total of 26 frequencies available here for use by remote pickup broadcast stations. All are subject to the condition that no interference is caused to high frequency broadcasting, primarily at night. Some of these frequencies are shared with the Maritime Mobile Services; but, as yet, this service has not been implemented. These frequencies are also subject to other fixed station uses which may be established at future RARC's.

3. <u>152.855-156.2475 MHz</u>: There are nine channels available here, between 152.84 and 153.38 MHz provided no interference is caused to other services. Also, these frequencies will not be licensed to "network entities," and "will not be authorized to new stations for use onboard aircraft."

4. <u>161.625-173.2 MHz</u>: There are seven channels here that are available only in certain geographic areas. Five of these channels, between 161.64 and 161.76, are not available to network entities and are also not available in Puerto Rico or the U.S. Virgin Islands.

5. <u>942-952 MHz</u>: This spectrum is without question the most controversial spectrum at present for radio broadcasters. Only half -- 947-952 MHz -- is actually allocated for radio STL's on a primary basis. The lower half, 942-947 MHz, belongs to "land mobile reserve" and has yet to be implemented by that service.

In the early 1970's as part of a massive restructuring of radio spectrum to accommodate the long term needs of the land mobile services, available STL spectrum was halved. The approximately 400 STL that operate there at that time continue to do so today; they are permitted continued renewal of their licenses pending a decision as to their disposition through a future FCC rule making proceeding. That rule making proceeding began on June 23, 1982 when the FCC adopted <u>Notice of Proposed Rule Making</u> in General Docket No. 82-335. Here, the Commission brought together a number of different proposals for the disposition of STL's operating below 947 MHz that had accumulated during the ten years or so since the band had been reallocated. In 1976, NAB petitioned the FCC to reallocate 942-947 MHz for all STL stations on a primary basis. A similar petition, filed by Moseley Associates, a manufacturer of STL's, proposed permitting broadcast STL's to operate on a secondary non-interfering basis within unassigned UHF television channels. Docket 82-335 also included a proposed decision in two previous FCC proposals (1) to allow STL stations to operate in the band 2150-2160 MHz; and (2) to allow STL stations to operate at 2110-2113 MHz. Both of these proposals envisioned "sharing" spectrum with other fixed users; the Commission suggested that these proposals be abandoned, chiefly because of spectrum congestion.

But the essence of Docket 82-335, however, proposed to permit STL's access to spectrum at 2130-2150 MHz and 2180-2200 MHz. Broadcasters now operating STL's below 947 MHz would be required to move to the new frequencies within five years of the adoption of rules in this proceeding.

At present, the outcome of Docket 82-335 is not known. According to the Commission's unified agenda of federal proceedings, a decision is expected sometime this spring and may be available by the time this paper is printed. In any event, it is NAB's view that the best way for the Commission to accommodate existing and future growth of aural STL's is to reallocate 942-947 MHz for the exclusive use of STL stations. Our comments in General Docket 82-335 advocated this reallocation; took no position with respect to accommodation of aural STL's in the UHF band; agreed with the FCC in that STL stations sharing of 2110-2113 MHz and 2150-2160 MHz should be abandoned; and, based on several technical studies, asserted that the Commission's proposed shared use of 2130-2150 MHz and 2180-2200 MHz is inadequate to meet the present and projected needs of STL's stations. It turned out that the 2100 MHz bands the Commission had proposed are at present very crowded and are growing ever more so through private operational fixed service users as well as common carrier users. In fact, a study in the Los Angeles area showed that it was unlikely that existing STL occupants at 942-947 MHz could be successfully accommodated at 2130-2150 and 2180-2200 MHz. These bands are just too crowded. Unfortunately, the Commission's Notice had provided no occupancy studies of the proposed frequency bands.

Pending Regulatory Proceedings

Besides Docket 82-335, there is one additional proceeding that promises to significantly affect broadcaster's use of auxiliary spectrum. That proceeding is MM Docket No. 84-280, a Notice of Proposed Rule Making adopted March 16, 1984, that proposes to amend current frequency assignment procedures in the "Broadcast Remote Pickup Service to facilitate more efficient use of available spectrum. Noting that demand for RPU frequencies at the 153, 161, and 450 MHz bands has increased, the Commission proposed "splitting" existing channels in order to encourage use of more spectrum efficient technologies, such as Amplitude Compandored Single Sideband ("ACSB"). Additionally, for the first time, "repeaters" using ACSB would be permitted wherever feasible.

II. Auxiliary Broadcast Allocations: Television Broadcasting

Television broadcasters use auxiliary broadcast spectrum for many of the same reasons that regular broadcasters do. The most popular uses include television STL's and Electronic News Gathering ("ENG"). Unlike radio broadcasters, however, television stations frequently employ intercity relay links when news events require mobile coverage. The broadcaster will erect an intercity relay link located at a strategic place within his city of license. Then, the mobile ENG equipment will direct their transmissions to the relay point instead of the station itself. The relay station improves the station's capability to gather live news by extending the area within which ENG transmissions can be received.

On balance, the spectrum allocated for television auxiliary services is significantly more competitive than spectrum allocated for radio broadcasting services. This has to do with the fact that network entities often travel extensively and can compete with local broadcasters for available frequencies. This results in an enhanced potential for spectrum congestion. Also, the variety of different activities that television broadcasters undertake usually requires more complex auxiliary systems than radio broadcasters. Also, most of the auxiliary spectrum allocated for television broadcasters is located above 1 GHz, popular spectrum used by private and common-carrier communications services.

There are three principal bands used for television broadcast auxiliary services: 1990-2500 MHz ("the 2 GHz band"), 6425-7125 MHz ("the 7 GHz band"), and 12.7-13.25 GHz ("the 13 GHz band").

1. The 2 GHz Band. While the 2 GHz band extends from 1990-2500 MHz, only 1990-2110 and 2400-2500 MHz is available for television broadcasters who wish to own and operate their own auxiliary news gathering services. There are seven channels at 1990-2110 MHz. There are few potential interfering services that also operate at these frequencies, although they can be used for certain earth-to-space and space-to-earth transmissions at 2025-2110 MHz and at certain locations at 1990-2120 MHz. Also, there may be government operated earth exploration satellite services at 2106.4 MHz. Fortunately, all of these services operate on a non-interfering basis with television auxiliary broadcasters. At certain geographical sites in the United States, space research radio services are authorized on a co-equal basis with auxiliary broadcasters at 2025-2035 MHz. There are three channels available at 2400-2500 MHz.

The 2 GHz band is used for both fixed STL uses and mobile ENG uses by television broadcasters. To date, local frequency coordinating committees have done an admirable job of ensuring that both these fixed and mobile uses at 2 GHz coordinate the mutual use of these frequencies.

2. The 7 GHz band. Here, there are four channels available between 6425 and 6525 MHz and ten channels available between 6875 and 7125 MHz. A standard satellite service time signal operates at 6427 MHz, and occasionally passive microwave sensor measurements are taken at 6425-7025 MHz. At 6425-6525 MHz, ENG stations operate on a secondary basis. Between 6875 and 7125, TV relay stations can operate on a secondary basis. These frequencies are primarily available for television broadcasters who wish to own and operate their own auxiliary services. There are other frequencies available for common carrier entities that wish to sell or lease their services to broadcasters. Although the communications provided by a broadcaster who owns his auxiliary service and the common carrier entity who leases or sells the broadcaster the same service are identical, they currently occupy different spectrum.

3. The 13 GHz Band. The 13 GHz band is without question becoming increasingly congested. In addition to television broadcasters, cable television systems ("CARS") use these frequencies as do private operational fixed service users ("POFS"). Thirteen GHz can support both mobile and fixed operations; however, mobile ENG is nearly always operated on a secondary basis to fixed service users. This spectrum is also highly controversial -- as explained in detail below.

Pending Proceedings

The foremost pending proceeding is General Docket No. 82-334. This rulemaking begun in early 1982, seeks to establish "a spectrum utilization policy for the fixed and mobile services, use of certain bands between 947 MHz and 40 GHz". The proposals espoused by the FCC in the most recent Notice of Proposed Rule Making could have a disastrous effect on television broadcaster auxiliary frequency use. In brief, the FCC proposed to displace present occupants of what is now the direct broadcast satellite ("DBS") downlink band, 12.2-12.7 GHz, to bands that included those allotted for television broadcast auxiliary services (TV ENG and STL at 2, 7 and 13 GHz). Not only would sharing be permitted but, for the first time, the FCC would impose technical standards it believes would improve spectrum efficiency. These proposed standards include required minimum path lengths for transmission on each microwave band, minimum antenna suppression standards, a new channeling plan for 2 GHz and revised frequency coordination procedures.

Fortunately, thanks to widespread participation by NAB and the broadcast industry, in September, 1983, the FCC adopted a favorable First Report and Order in Docket 82-334. The FCC preserved the integrity of 2 and 7 GHz, and allowed only limited sharing at 13 GHz and opened up the 18 GHz to all. For the time being, no action was taken on other technical proposals such as minimum path length requirements. The most important victory for broadcasters here was that no sharing will occur at 2 and 7 GHz.

The Commission did, however, alter some frequency coordination procedures, especially at 13 GHz. In this respect, on December 3, 1983, NAB filed a Petition for Partial Reconsideration and Clarification in General Docket No. 82-334. The petition examined the status of the frequencies 12.7-13.25 GHz in detail and determined:

1) private operational fixed service users displaced from 12 GHz may operate on the frequencies 12.7-13.20 GHz, on a co-primary basis with existing fixed users of that band;

2) mobile ENG units may operate on the secondary basis to fixed users on the frequencies 12.7-13.20 GHz, except that, on the frequencies 13.15-13.20 GHz, within 50 kilometers of the top 100 television markets, they are co-primary with community antenna relay stations ("CARS") ENG; and

3) mobile ENG units may operate on a co-primary basis with existing fixed users on the frequencies 13.20-13.25 GHz.

NAB's petition attempted to ensure that mobile ENG users at 13 GHz, now secondary will always remain at 13 GHz despite the expected preponderance of new primary fixed links arising out of Docket 82-334.

Docket 82-334 also opened the 18 GHz band to any and all wideband users, including broadcast auxiliary services. Since this proceeding remains under FCC reconsideration, the 18 GHz band channeling plans, bandwidths, and frequency coordination procedures may be revised and, accordingly, not presented in this paper.

III. Practice and Procedure

There have been many changes in the last year regarding the practice and procedure of broadcasters now using auxiliary frequency spectrum or commencing expanded auxiliary operations. Here are several that you may not be aware of:

1) if you are using an aural STL operating at 947-952 MHz for the transmission of television sound, you may continue to do so provided that a demand for that channel is not expressed by an AM or FM station in your area;

2) the FCC now permits system licensing of TV auxiliary mobile stations. However, the maximum area that may be requested in conjunction with a system license is that of a standard metropolitan statistical area, or SMSA. Applicants for a system license will have to specify the type of transmitter to be used at each station, but it will not be necessary that the equipment, to be operated under the terms of the system license, be manufactured before April 1981 (as used to be the case). Further, this equipment does not necessarily have to be type accepted by the Commission. Applications for consolidation into a system license will be accepted only at the time application is made for renewal of the main Part 73 broadcast station license; and

3) if you have an "excess capacity" of TV ENG or subcarriers on TV STL's or ICR's, you may use them for non-broadcast and potentially profitable new communications services.

4) The FCC considers broadcasters to be included within the category of entities eligible for Private Radio Service frequencies ("POFS") administered under Section 90.75(a)(1) of the Commission's Rules. The FCC will accept applications for the use of frequencies (4 channels maximum) in the 21.8-22.4 and 23.0-23.6 GHz bands.

IV. How to Use the Appendix

The Appendix attached to this paper presents the broadcast auxiliary spectrum as presented by the FCC's Second Report and Order in General Docket No. 80-739: domestic implementation of the Final Acts of the 1979 World Administrative Radio Conference ("1979 WARC"). This "grandaddy" of allocations proceedings sets the stage for the next 20 years of frequency assignment decisions.

Available spectrum is listed in order of increasing frequency. If a given "slice of spectrum is divided horizontally, the "upper" service is primary and the "lower" service is secondary. If, on the other hand, a slice of spectrum is divided diagonally, both services enjoy "co-equal" status.

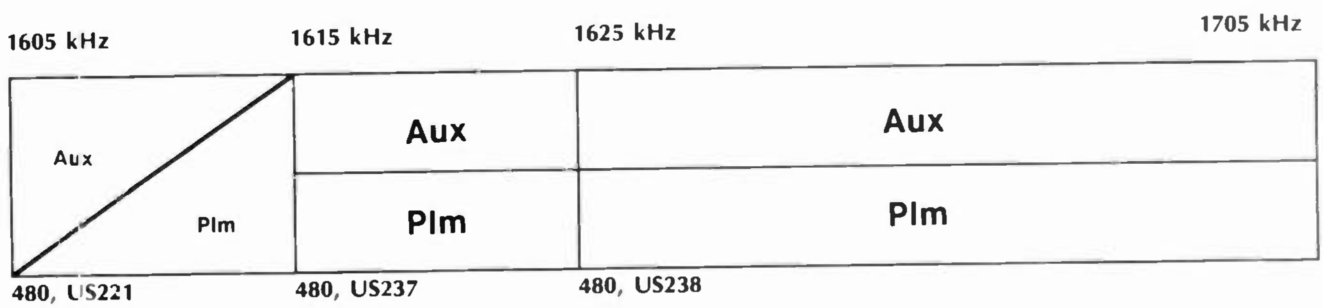
Note the allocation footnotes for each spectrum slice. These are located in Section 2.100 to 2.108 of the Commission's Rules. Here are the relevant terms:

POFS = Private Operational Fixed Service
Aux = Auxiliary Broadcast Service
Mar = Maritime Mobile
Plm = Private Land Mobile
DPlm = Private Land Mobile (common-carrier)
Hf = High frequency broadcasting
Ham = Amateur Radio

V. Conclusion

This paper began as an ambitious attempt to consolidate present knowledge of (1) allocated frequency spectrum, (2) pending FCC spectrum allocation proceedings, (3) spectrum congestion, and (4) prospects for the future into one technical presentation. It is not complete. To do so would have required a great deal more time and effort than has been expended. But a start has been made; a "snapshot" taken of auxiliary spectrum use. It is my hope that all broadcasters can use this information to first make maximum use of the auxiliary systems they employ, and second, to benefit our industry by understanding and participating in the major spectrum allocation questions of auxiliary broadcasting that remain to resolved.

Radio Broadcast Auxiliary Services
1605 — 1705 kHz



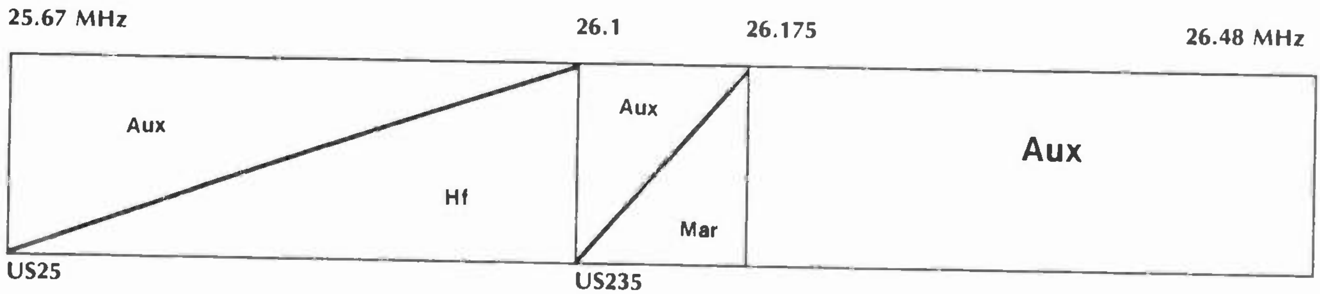
480: Use of 1605-1705 kHz is subject to 1985 RARC (AM band expansion).
US221: Use of 1605-1615 kHz limited to distribution of Traveler's Information Stations ("TIS"), operating at 1610 kHz.
US237, US238: At 1615-1625 kHz, Radiolocation services are primary until 1985 RARC.

Group A channels: 1606, 1622, and 1646.

May 1984

1

Radio Broadcast Auxiliary Services
25.67 — 26.48 MHz



US25: No interference to high-frequency broadcasting.

US235: These frequencies have yet to be implemented by the Maritime mobile services; also, they may be used as an alternative fixed allocation subject to future RARC's.

Group D: 25.87, 26.15, 26.25, 26.35;

Group E: 25.91, 26.17, 26.27, 26.37;

Group F: 25.95, 26.19, 26.29, 26.39;

Group G: 25.99, 26.21, 26.31, 26.41;

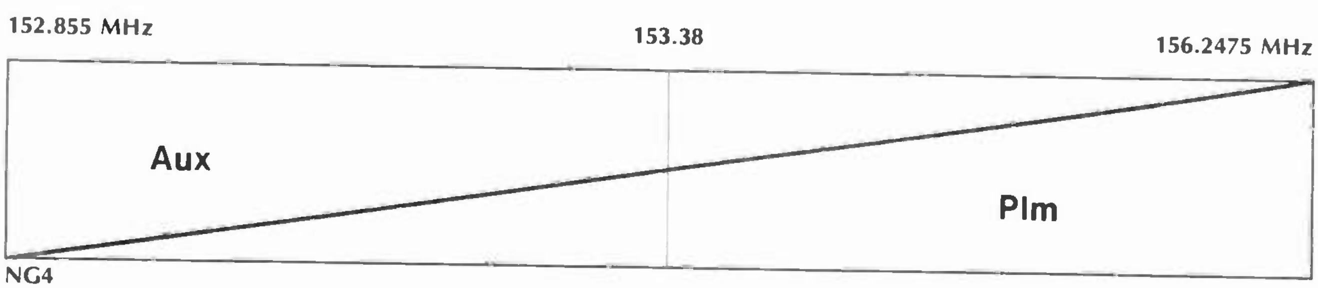
Group H: 26.03, 26.23, 26.33, 26.43;

Group I: 26.07, 26.45; and 26.11;

Group J: 26.09, 26.47; and 26.13.

May 1984

Radio Broadcast Auxiliary Services
152.855 — 156.2475 MHz



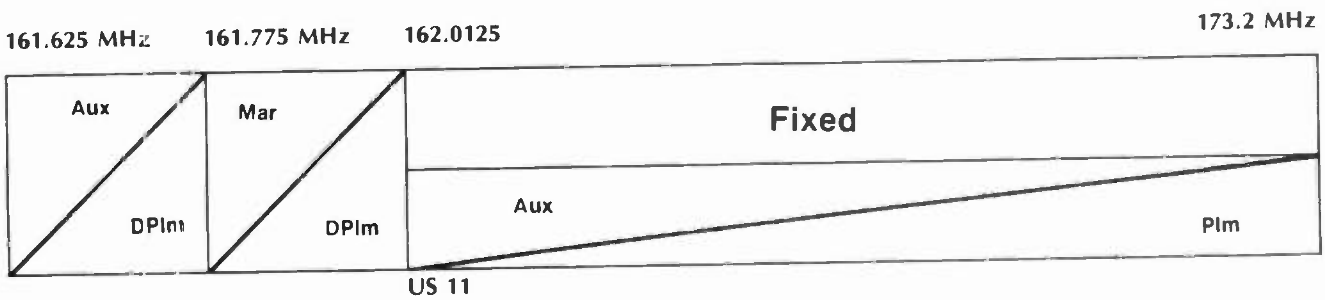
NG4: Limits available frequencies to 152.84-153.38 MHz, provided no interference is caused to other services.*

Group K1: 152.87, 152.93, 152.99, 153.05, 153.11, 153.17, 153.23, 153.29, 153.35.

* These frequencies will not be licensed to "network entities," and "will not be authorized to new stations for use on board aircraft."

May 1984

Radio Broadcast Auxiliary Services
161.625 — 173.2 MHz

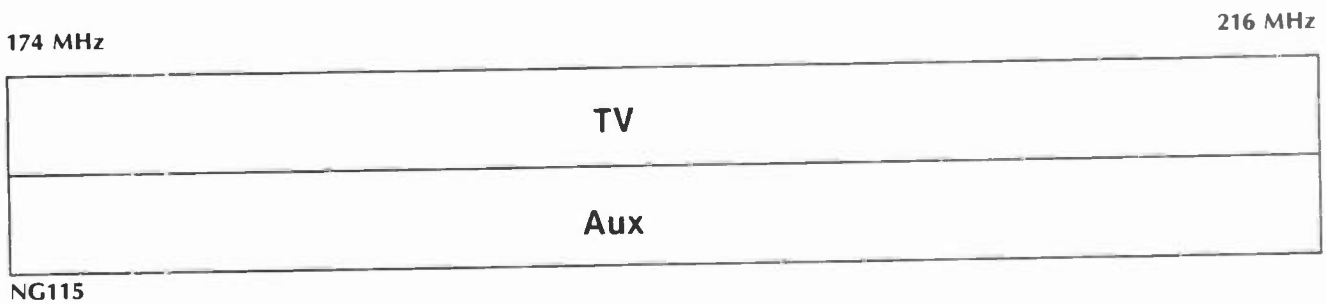


US11: Limits available frequencies to 166.25 and 170.15 MHz only in certain geographic areas.
Group K2: 161.64, 161.67, 161.70, 161.73, 161.76 *
Group L: 166.25 MHz
Group M: 170.15 MHz
* Not available to "network entities" or in Puerto Rico and the U.S. Virgin Islands.

May 1984

4

Radio Broadcast Auxiliary Services
174 — 216 MHz

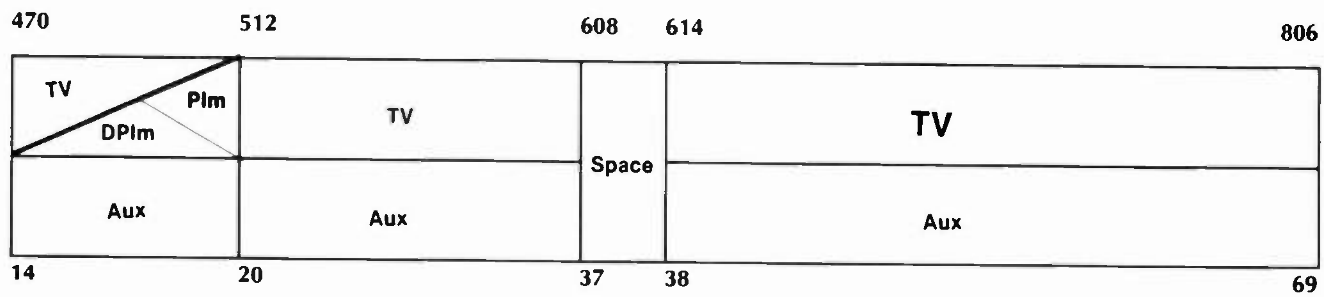


NG115: "Wireless microphones" may operate here but strictly on a secondary non-interfering basis.

May 1984

5

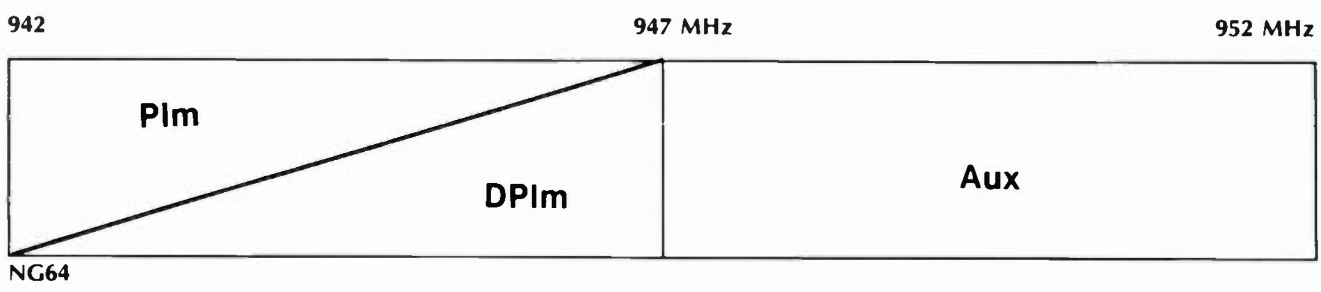
Television Broadcast Auxiliary Services
470 — 806 MHz



Auxiliary operation at 470-806 MHz is limited to low power television stations and television translator relay stations.

May 1984

Radio Broadcast Auxiliary Services
942 — 952 MHz

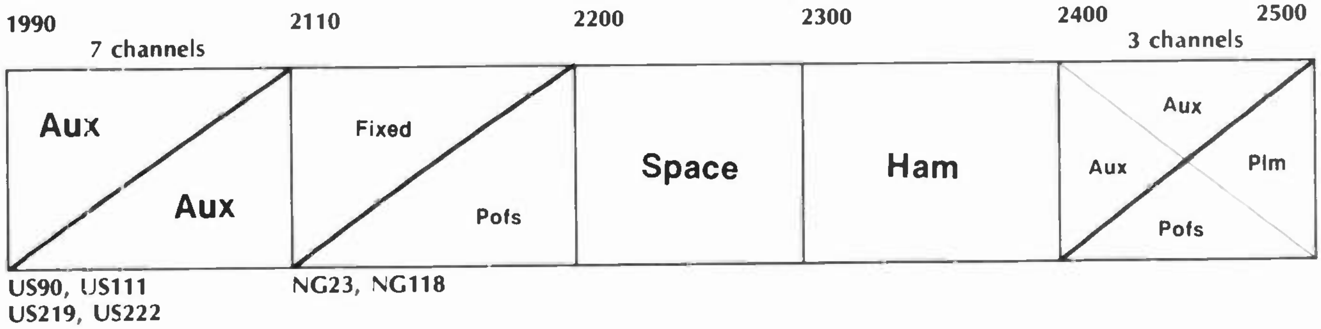


NG64: Permits Aural STL's operating at 942-947 MHz continued license renewal "pending a decision as to their disposition through a future rule making proceeding." Also, there are several STL stations that continue to operate below 942 MHz (licensed prior to April 16, 1958).

Channels: Available frequencies are every 0.5 MHz. "Split channel" operation may also be authorized.

May 1984

Television Broadcast Auxiliary Services
1990 — 2500 MHz



US90: There may be earth-to-space and space-to-earth transmissions authorized (on a non-interfering basis) at 2025-2110 MHz.

US111: There may be government space research earth stations authorized (on a non-interfering basis) at certain locations at 1990-2120 MHz.

US219: There may be government earth exploration-satellite services at 2106.4 MHz (on a non-interfering basis).

US222: At certain U.S. sites, certain space research radio services are authorized (on a co-equal basis) at 2025-2035 MHz.

NG23: Frequencies at 2100-2200 MHz may be assigned to stations in the international fixed public radio service (but only in Florida and U.S. Carribean).

NG118: TV translator relay stations may operate on a secondary basis.

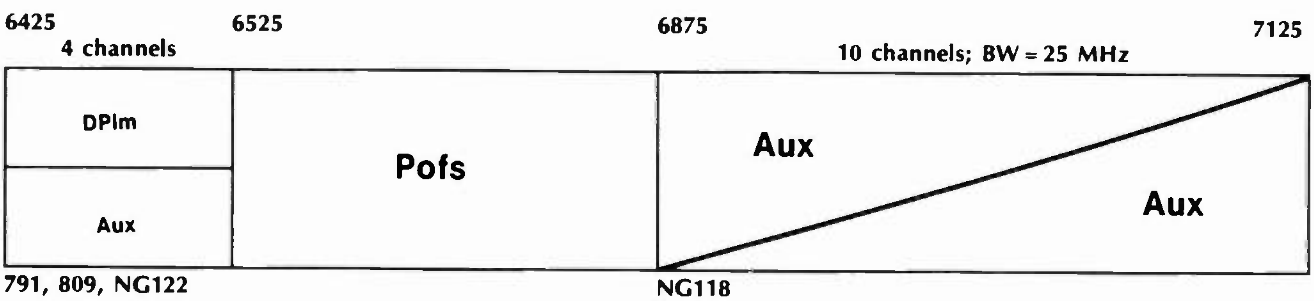
US41: Government radio location service at 2450-2500 MHz can operate on a non-interference basis with non-government services.

752: All radio services must accept ISM interference from equipment operating at 2450 MHz.

Band A: 1990-2008; 2008-2025; 2025-2042; 2042-2059; 2059-2076; 2076-2093; 2093-2110.
2450-2467; 2467-2484; 2484-2500.

May 1984

Television Broadcast Auxiliary Services
6425 — 7125 MHz



791: 6427 MHz is a "standard" satellite service time signal.

809: "Passive" microwave sensor measurements at 6425—7025 MHz.

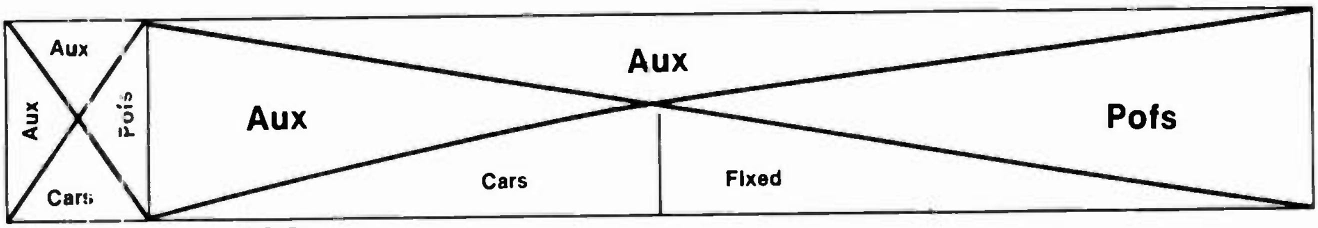
NG122: TV pickup stations may operate on a secondary basis.

NG118: TV translator relay stations may operate on a secondary basis.

Band 8: 6425-6450; 6450-6475; 6475-6500; 6500-6525 (NG122).
and 6875-6900; 6900-6925; 6925-6950; 6950-6975; 6975-7000; 7000-7025; 7025-7050; 7050-7075; 7075-7100; 7100-7125.

May 1984

Television Broadcast Auxiliary Services
12.7 — 13.25 GHz



NG53: TV pickup and CARS pickup are co-equal at 12.7-13.15 GHz. At 13.15-13.20 GHz, they are co-equal and exclusive within 50 km. of the top 100 markets.

NG118: TV translator relay stations may operate on a secondary basis.

US251: Space research at 12.75-13.25 GHz, but only at Goldstone, CA.

Band D:

Group A:

12700-12725; 12725-12750; 12750-12775; 12775-12800;
12800-12825; 12825-12850; 12850-12875; 12875-12900;
12900-12925; 12925-12950; 12950-12975; 12975-13000;
13000-13025; 13025-13050; 13050-13075; 13075-13100;
13100-13125; 13125-13150; 13150-13175; 13175-13200;*
13200-13225; 13225-13250.

Group B:

12.7125-12.7375; 12.7375-12.7625; 12.7625-12.7875; 12.7875-12.8125;
12.8125-12.8375; 12.8375-12.8625; 12.8625-12.8875; 12.8875-12.9125;
12.9125-12.9375; 12.9375-12.9625; 12.9625-12.9875; 12.9875-13.0125;
13.0125-13.0375; 13.0375-13.0625; 13.0625-13.0875; 13.0875-13.1125;
13.1125-13.1375; 13.1375-13.1625; 13.1625-13.1875; 13.1875-13.2125;*
13.2125-13.2375.

* note NG53.

May 1984

Spectrum Conservation With High Performance SSB Microwave Carriage of Multiple Television Signals

T. M. Straus, R. T. Hsu, and W. C. Margiotta

Hughes Aircraft Company, Microwave Communications Products

Torrance, California

Frequency congestion is increasingly a problem in the 11 GHz Common Carrier and 13 GHz Broadcast Auxiliary bands. In accordance with normal practice a 40 MHz wide segment is provided for the carriage of a single FM channel in the 11 GHz Common Carrier band while 25 MHz wide channels are specified in Section 74.602 of the FCC Rules and Regulations for each television signal transmitted by FM in the 13 GHz Broadcast Auxiliary band. By comparison, up to four TV signals could fit into the 25 MHz allocation if the standard 6 MHz VSB bandwidth were used. Figure 1 shows an arrangement in which the residual 1 MHz is utilized for audio transmission. Similarly, up to six television signals could be accomodated within the 40 MHz wide channel at 11 GHz. Clearly, the scheme provides a highly spectrum efficient means of signal transport. However, spectrum efficiency cannot be the only measure of a communication system. RS 250B continues to be the standard for high quality microwave transmission of television signals. Note, however, that "This standard deals primarily with analog microwave systems using angle modulation".[1] Further, "Standards for a short haul system such as an STL are as stringent as the practical state-of-the-art permits". Both the challenge and the question was then the extent to which high quality spectrum efficient VSB techniques could match the RS 250B criteria applicable to FM transmission.

High quality VHF modulators and demodulators exist to serve the needs of the broadcast industry and as test instrumentation. Several of these were investigated from the point of view of the present application. It soon became evident that standard high quality VSB modulators employed IF filters which were too wide for adjacent VSB channel implementation. At best the filters provided only 20 dB rejection of signal in the adjacent channel and this led to excessive interference. The same was true of the VSB filter used in the demodulator. Measurement showed an unacceptable degradation in S/N when the adjacent channels were turned on. This was particularly evident when the sound trap in the demodulator was switched out. Yet with the sound trap switched in the system, frequency response, as evidenced by the multiburst test shown in Figure 2, was severely curtailed. Since separation of the sound from the video was in any case also beneficial to the reduction of intermodulation distortion in the modulator and the microwave transmitter, it soon became evident that a frequency plan separating audio from video, as shown in Figure 1, was required to achieve the highest quality transmission.

Another factor in the back to back testing of VHF modulators and demodulators which soon became evident was that the highest S/N obtainable was less than the 67 dB criterion stated in RS 250B for short haul systems. However, the highest quality demodulators offered two alternative detection methods; synchronous and envelope detection. Utilization of the former permitted the introduction of partial carrier suppression in the modulator. This in turn enhanced the maximum obtainable S/N and also led to further reduction in intermodulation distortion. When coupled with the introduction of IF-band limiting SAW filters into both the modulator and demodulator, the back-to-back performance characteristics delineated in Table I were obtained. The amplitude frequency response is limited by ripple in the SAW filter. Further improvement in this parameter is anticipated by the use of SAW filters having complementary ripple patterns in the modulator and demodulator. With this modification it is expected that the RS 250B criterion of ± 0.2 dB for short haul systems may be met. Short-time distortion is another matter. Table II shows the variation of short-time distortion with bandwidth. The table is based upon a theoretical transform calculation of the response of a T rise-time bar signal through an ideal filter having the indicated bandwidth. For adjacent channel carriage of VSB television signals the SAW filter must start to roll off at 4.2 MHz above the vestigial sideband carrier in order to provide sufficient isolation at the band edge which is only 4.75 MHz above the carrier. The theoretical calculations are supported by the measurements; when a SAW filter with cut-off 5 MHz above the carrier was used the RS 250B short and medium-haul criterion of 4 IRE was obtained. However, as previously stated such a filter leads to intolerable interference between the adjacent channels.

TABLE I - Key VHF Modulator/Demodulator Performance Parameters

Amplitude Frequency Response	±0.3 DB
S/N	69 dB
Differential Gain	0.5%
Differential Phase	0.2°
Short-Time Distortion	6 IRE

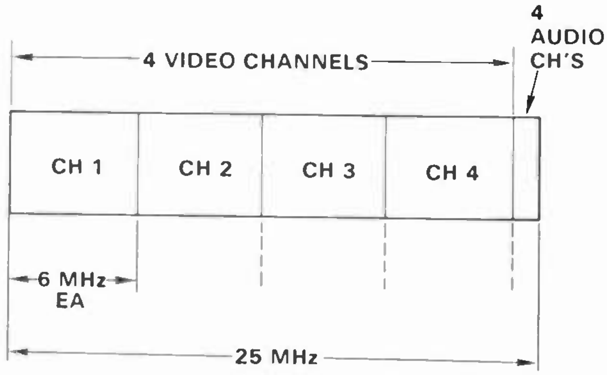


Figure 1 — High Quality Multichannel VSB Transmission Frequency Plan

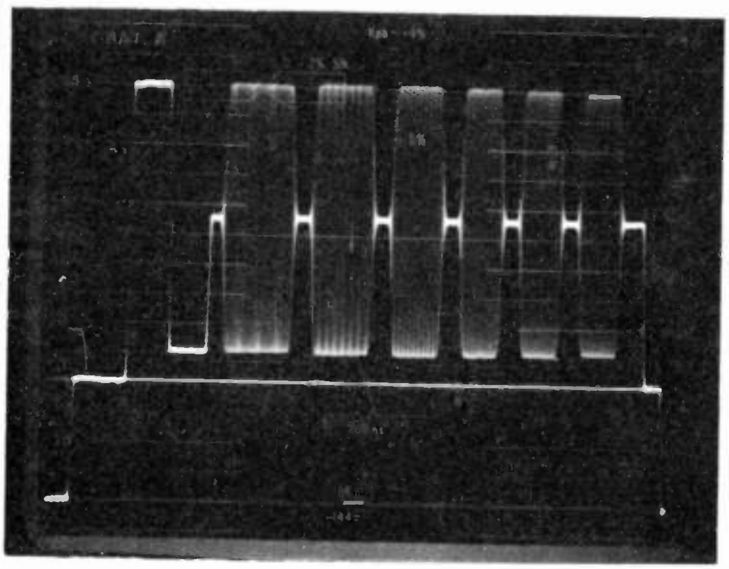


Figure 2 — Effect of the Demodulator Sound Trap on Multiburst Response
a) Sound Trap Out

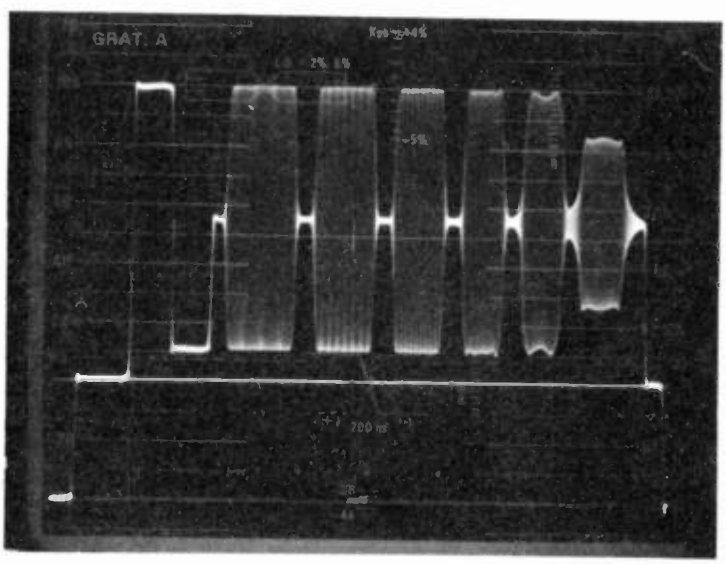


Figure 2 — Effect of the Demodulator Sound Trap on Multiburst Response
b) Sound Trap In

TABLE II - Short-Time Distortions vs System Bandwidth

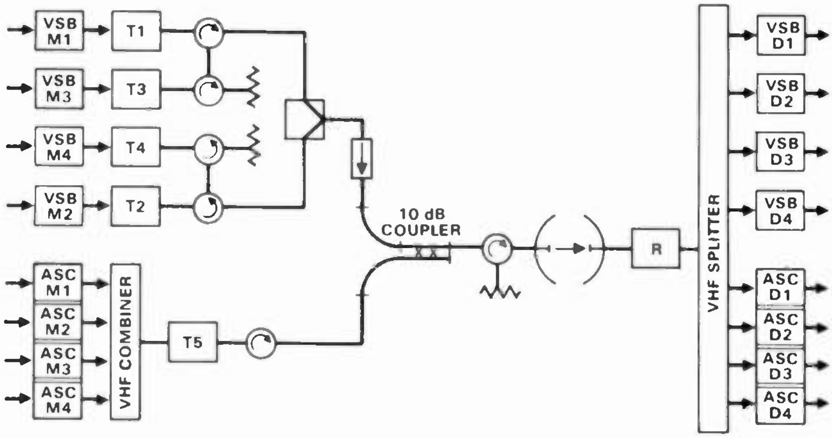
Baseband Width (MHz)	Short-Time Distortion (P-P IRE units)
4.0	6.2
4.4	5.0
4.8	3.9
5.2	3.0

Before addressing the overall link performance it may be of interest to briefly address the quality of lower cost VHF modulators and demodulators widely used by the CATV industry. Here adjacent channel carriage is routine but the other specifications are not as good. Selected optimized units can approach the short-haul RS 250B differential gain and phase criteria (2% and 0.5° respectively) but the normal specification may not even comply with the medium-haul RS 250B. S/N is also typically limited to 60 dB and amplitude frequency response is at best ±0.5 dB and specified at ±1 dB. Short time distortion is more typically stated in terms of a 2T rise waveform to allow for the 4.2 MHz bandwidth limitation. Thus, although these modulators and demodulators can provide a reasonably high quality signal to the CATV subscriber, they cannot be considered for utilization in a system where state-of-the-art type performance is desired.

Figure 3 shows the configuration of a high quality system carrying four VSB TV signals. At the transmitter each video channel is separately processed first through a IF/VHF modulator and then through a microwave upconverter/amplifier. Next adjacent channels are combined through a circulator-filter multiplexing network and then further combined with the adjacent channels by means of a "magic tee". The four audio signals are either FDM or TDM combined and then converted up to microwave in a single unit. A 10 dB coupler provides for combination of the audio and video signals at the output of the transmitter.

The receive site utilizes a single broadband microwave downconverter to convert the signals to VHF. At this point a power splitter leads to the separate video and audio demodulators.

Table III provides a composite summary of measured performance of the overall experimental 4-channel system. Aside from short time distortion, all parameters met or exceeded the RS 250B medium haul criteria and with the further exception of amplitude frequency response and S/N, even the short haul RS 250B criteria were matched or bettered. Probable improvement in amplitude response has already been alluded to and the overall S/N performance is discussed further in a subsequent paragraph. Figures 4-7 show photographs of the system performance relating to a number of the key parameters. Variation in performance among the various high quality VHF modulators and demodulators evaluated during the course of the study was generally minor when the aforementioned modifications are taken into account. The microwave portions of the system utilize standard "AML" equipment which is widely used for



VSB M1-4 : VESTIGIAL-SIDEBAND MODULATOR
T1-5 : AML TRANSMITTER
R : AML WIDE-BAND RECEIVER
VSB D1-4 : VSB DEMODULATOR
ASC M1-4 : AUDIO SUBCARRIER MODULATOR
ASC D1-4 : AUDIO SUBCARRIER DEMODULATOR

Figure 3 — 4-Channel System Configuration

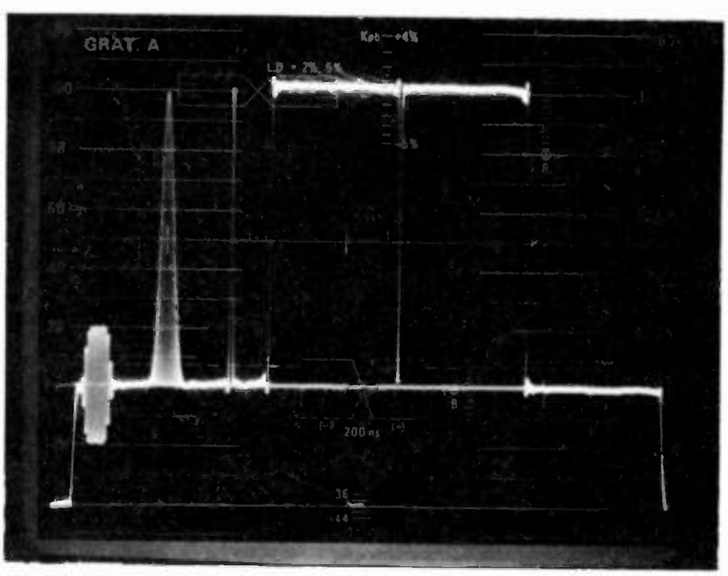


Figure 4 — System 12.5T Pulse and T Rise Bar Signal Outputs

TABLE III - Overall System Performance Summary

Parameter	Test Result
Amplitude Frequency Response	± 0.35 dB
Field Time Distortion	2 IRE
Line Time Distortion	1 IRE
Short Time Distortion	6 IRE
Chrom. to Lum. Gain	1 IRE
Chrom. to Lum. Delay	20 ns
Differential Gain	1%
Differential Phase	0.5°
Signal to Noise Ratio	63 dB
Luminance Nonlinearity	1%
Chrom-to-Lum Intermodulation	0.5%
Chrom. non linear gain	0.7%
Chrom. non linear phase	1°
Dynamic Gain of Picture Signal	0.6%
Dynamic Gain of Synch Signal	1.3%
Transient Synch Signal Nonlinearity	0.6%
Long-Time Waveform Distortion	2 IRE
Signal-to-low freq noise Ratio	55 dB $\frac{\text{p-p}}{\text{p-p}}$
	73 dB $\frac{\text{p-p}}{\text{rms}}$

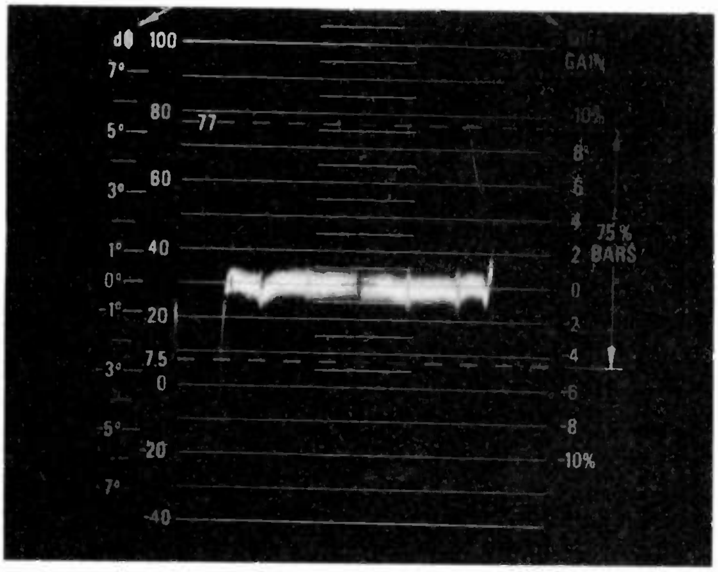


Figure 5 — System Differential Gain Measurement

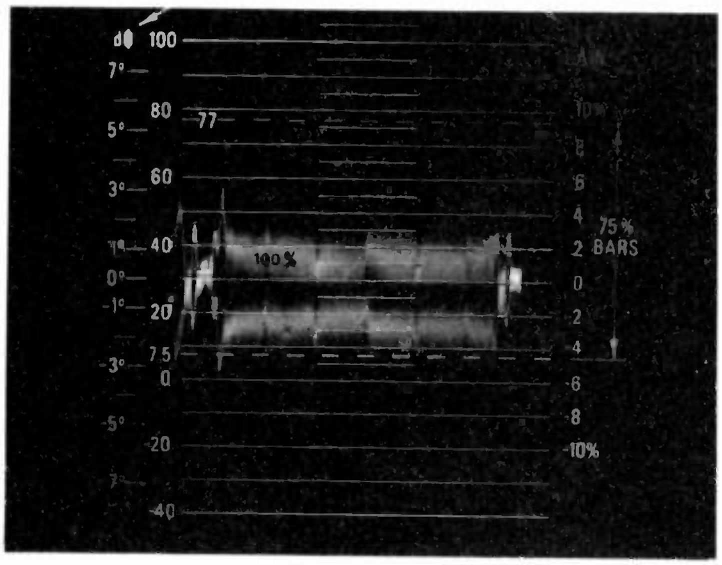


Figure 6 — System Differential Phase Measurement

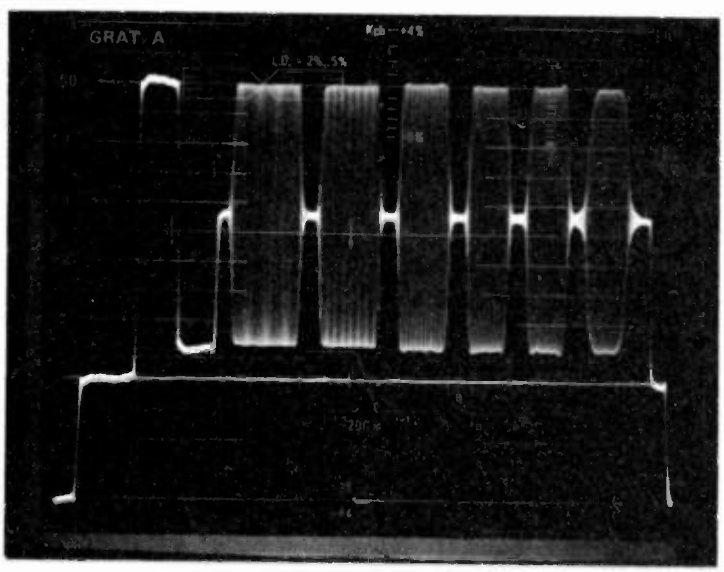


Figure 7 — System Multiburst Response

local distribution of CATV signals. However, in order to achieve the best possible performance, both with respect to noise and nonlinear distortion, several modifications were also made to this equipment. In the transmitter the principal change was the insertion of a FET amplifier between the up-converter output and the klystron amplifier input. This change allowed the upconverter signal level to be reduced sufficiently to make its contribution to non-linear distortion negligible. Further, with the inclusion of the FET amplifier, the klystron gain could be reduced, thus lowering the overall noise output while at the same time permitting optimization of klystron tuning for linearization performance. Nevertheless, it was necessary to utilize precorrection in the VHF modulator in some of the channels in order to achieve the differential gain and phase performance given in Table III. With no precorrection, differential gain and phase performance in one channel degraded to 2.5% and 0.8° respectively.

Figure 8 shows the spectrum analyzer presentation of next adjacent channels at the VHF output of the AML receiver. Note that, with the exception of a small component of intermodulation distortion due to the upper adjacent channel, there is no spillover of the adjacent channel test signals into the (temporarily) unoccupied central channel. This is due to the aforementioned SAW filter. Even the intermodulation product is here 70 dB down relative to the unsuppressed carrier.

Table IV returns to the question of S/N in the overall system by setting forth the contributions of the various elements for adjacent channel performance. The single channel performance has only three contributors and comes within 1 dB of matching the 67 dB RS 250B short haul criterion. The gap could in fact be narrowed to ½ dB by allowing the receiver to operate at a higher S/N. This is only limited by the AGC setting which utilizes an input ferrite attenuator to maintain constant S/N and multichannel intermodulation performance throughout the (35 dB-wide) AGC range. For single channel applications there is no drawback to raising the AGC set point. However, for multichannel application the receiver intermodulation products become significant at high signal levels. The table assumes the worst case six-channel loading applicable to a 11 GHz common carrier band transmission. As seen, the S/N is further degraded by adjacent channel signal and transmitter noise so that an overall 63 dB S/N is obtained. Thus the performance falls between the 67 dB and 60 dB RS 250B short and medium haul criteria. Further improvement would be extremely difficult since there are so many contributors to the overall system S/N. In any case, the performance applies only to the clear weather situation. Figure 9 shows a typical plot of S/N versus signal level at the microwave receiver input. The example shows the receiver AGC set at the 70 dB level and an overall system S/N of 63 dB in the microwave receiver AGC range. As the signal fades below the AGC threshold, the microwave receiver S/N becomes dominant. The example illustrates the case of a 10 mile path using 6' antennas at either end. Total waveguide loss was assumed to be 5 dB. With these assumptions a 42.5 dB fade margin to 35 dB S/N is indicated. Additional fade margin could be obtained by tower mounting of the microwave receiver (cutting waveguide loss by 2 dB) and the use of larger antennas. However, throughout much of the United States a 40 dB fade margin on a 10 mile path should result in less than an 1 hour of path "outage" per year. Note in this regard that an AM system, unlike FM transmission, has no "threshold" below which the S/N degrades much faster than 1 dB for each additional dB of path fade.

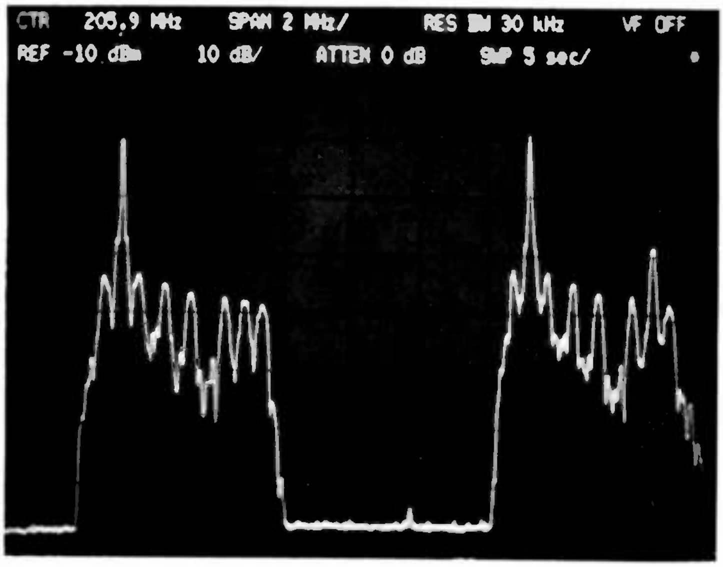


Figure 8 – Spectral Isolation Performance of Adjacent Channels

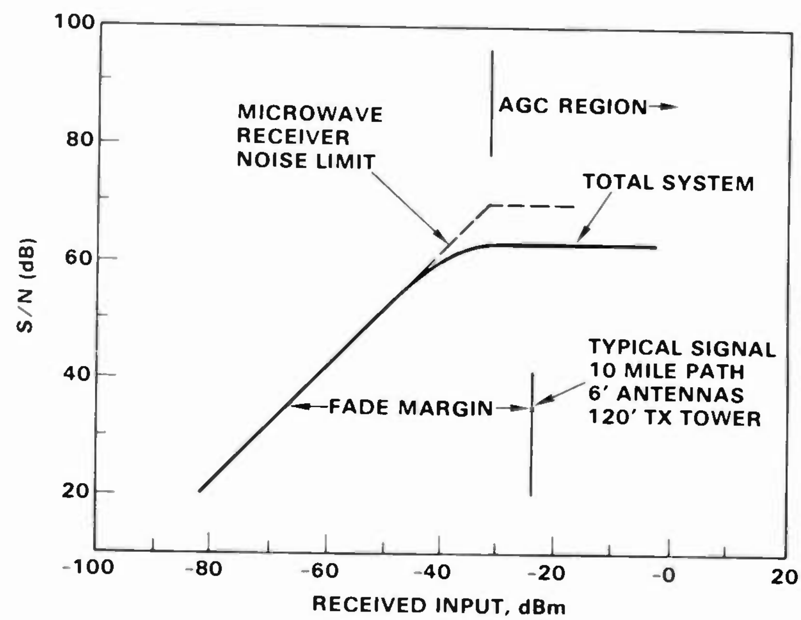


Figure 9 – Typical System S/N Fade Performance

TABLE IV - System Noise Contributions

Contributor	Equivalent S/N (dB)	
Microwave Receiver Thermal Noise	70	
VHF Modulator/Demodulator (1 channel)	69	
Transmitter (klystron) noise (1 channel)	76	
Total System (single channel)		66
Receiver Intermodulation (6 adjacent channels)	72	
Adjacent transmitters (noise)	76	
Adjacent channel signals	69	
Total System (six channels)		63

In conclusion, high quality microwave transmission of adjacent VSB television channels is possible using standard equipments with certain modifications. The quality of the transmissions is generally equivalent to the rigid standards of RS 250B short-haul with a major exception of S/N in which case the performance falls roughly half way between the RS 250B short and medium haul criteria. Carriage of television signals in this manner may thus provide an attractive alternative in situations where available spectrum is insufficient to accomodate standard FM type transmissions.

References

1) Electrical Performance Standards for Television Relay Facilities, EIA RS 250B, page 1.

PROPAGATION CHARACTERISTICS FOR THE

NEW 18 GHZ AURAL BROADCAST STL BAND

STEVEN R. SMITH

GENERAL TELEPHONE COMPANY OF THE NORTHEST, INC.

EVERETT, WASHINGTON

The Federal Communications Commission, through General Docket No. 79-188 Second Report and order issued September 9, 1983, allocated a portion of the radio spectrum in the 18 GHz Band for Broadcast Auxiliary Services. This allocation consisted of 12 two-way paired frequencies with band width of 5 MHz.

The following is an account of the effects of atmospheric conditions on 18 GHz propagation in the Puget Sound Basin. General Telephone Company of the Northwest, Inc., established a nine-mile 18 GHz test link in August 1981 using a Farinon DM-18 microwave radio. The system operated through the winter of 1981/82 between the Bothell Radio Remote Station and the Everett Primary Center. The receive signal level and rain rate were recorded on a strip chart recorder for the nine-month period.

This paper deals with the effect of rain on 18 GHz signal propagation. Other propagation anomalies, such as the observed effect of K-factor layering with respect to up fades and the attenuation effects of snow, will also be examined.

INTRODUCTION

General Telephone Company of the Northwest, Inc., using equipment on loan from Farinon, Inc., established an experimental nine-mile 18 GHz radio link in the Puget Sound Basin. The objective was to determine whether 18 GHz propagation in the Northwest would provide adequate service for telecommunications toll traffic. The objective of toll grade service is 99.999 percent reliability. This corresponds to approximately five minutes per year of outage time for each complete system.

The experimental path was a 14.4 km hop between the Everett Primary Center (EPC) at Everett, Washington and the Bothell Radio Remote site (Bothell) at Bothell, Washington and was licensed under GTNW's experimental license. The system was designed with a 35.2 db fade margin and a calculated availability of 99.99975 percent. (See data sheet.) This provides for 1.3 minutes of outage time per year. The propagation outage of the experimental system exceeded this value but was within the 99.999 percent availability. The total outage time due to propagation anomalies was found to be approximately two minutes. However, rain did not cause the system to drop below the fade margin at any time. Two atmospheric conditions caused the system to approach the fade margin. The first rain following installation of the system caused extreme, abrupt attenuation of the signal. (See Figure 1.) Secondly, K-factor layering caused beam bending and again, attenuation of the system approached the fade margin. (See Figure 2.)

The system was operational between August 1981 and May 1982. It was felt that rain would be the major cause of attenuation resulting in poor reliability. The wavelength in free space of this frequency is 1.67 cm (16.7 mm). It was felt that the approximate size of raindrops typical in this locality is much smaller than would be required to cause quarter wavelength attenuation of the signal. Even though the Pacific Northwest region receives abundant rainfall, rain showers are usually not as heavy as in many other parts of the country. For example, the rain rate in Miami, Florida has been recorded at more than 20 times that of the Pacific Northwest. (Freeman, 1981)

METHODS & MATERIALS

As can be seen in Figure 3, there was no chance of fresnel zone interference due to path obstructions. The transmit frequencies were checked with a frequency counter manufactured by E.I.P. (Model 548) and was recorded as 18585 MHz at EPC and 18805 109 MHz at Bothell. After checking and recording the transmit power as 22.9 dBm at EPC and 21.0 dBm at Bothell with a Hewlett Packard 432A equipped with the K-486A thermistor mount, the system was turned on and path alignment was performed. Due to an oversight during installation, the E planes of the WR42 connector going into the radio cabinet were incorrectly oriented at 90 to each other resulting in 20 dB low RSL. After the oversight was corrected, the system came up to predicted RSL per the AGC reading on the meter panel. The predicted RSL was -42.8 and the actual RSL at EPC was -40. (See line 30 of Figure 4.)

The equipment used in the test consisted of DM-18 radios set up for one T-loading. Bandwidth on this radio is 10 MHz. Andrew HPM-180C antennas were used while the waveguide consisted of WR42 flex twist. To monitor activity on the path, the AGC was hooked up to one of the two channels of a slow running (1 cm/hr) strip chart recorder. During the first month of the study, there were chart recorders on both ends of the path, but subsequently one was removed for use elsewhere. The recorder at the Everett Primary Center remained in use for the duration of the test. Upon comparison of the two records, it was found that due to this small value of Delta f/f, identical results were obtained at the two sites.

To measure rain rate, an event counter was connected to the B Channel of the recorder at the EPC. Upon receipt of 0.01 inches of rain, the event would be recorded. Calibration of the system was done using a 50 dB vane attenuator inserted into the waveguide. 5 dB increments were noted on the AGC meter and the strip chart recorder. A weather log was maintained throughout the study to correlate with the Channel B data of the chart recorder.

FINDINGS

Four conditions were found to cause the R.F. signal to be attenuated. Continuous rain, wet and heavy snow, beam decoupling, and the first rain on the system will be discussed separately.

Rain, with the exception of the first rain, never caused the system to drop below its fade margin of 35 dB. The deepest fade found to be caused by rain was 18 dB which occurred when the rain rate was 15.24 mm/hr. (See Figure 5.) The predicted fade for this rain rate (using $A = a R$) is 15.22 dB with a = .05 and b = 1.12. Values of "a" and "b" are according to O. j. Laws and S. A. Parsons, 1943. This particular rain cell was unusual in that it covered the entire path. Most rain cells with a 15 mm/hr rate would cover a smaller area (see CCIR No. 563-7). Findings agree with the CCIR report that the heavier the rain, the smaller the cell size.

A maximum rain rate of 20 mm/hr was recorded which should correspond to rain cell size of 4 km in length. This, of course, would vary greatly from region to region. To cause the path to fade 35 dB, the rain rate would have to be 25 mm/hr along the entire path. This, in fact, never happened during the ten-month period when the system was functional. The phenomena correlating rain rate to rain cell size explained the results of the following millimeter wave experiment. Rep. 338-2 issued by the CCIR is partially quoted below:

> "Measurements in the United Kingdom over a period of two years ... at 11, 20, and 37 GHz on links of 4-22 km in length show the attenuation due to rain and multipath, which is exceeded for 0.01% of the time and less, increased rapidly with path length up to 10 km, but a further increase up to 22 km produced only a small additional effect ..."

Snow attenuation occurred when snow flakes approached a quarter wavelength and contained appreciable moisture; dry snow had very little effect on the signal strength. The most severe fade observed was caused by a wet, heavy snowfall. (See Figure 6.) This deep fade was 29 dB for two minutes; the signal remaining below 25 dB for a total of six minutes. This occurred on December 28, 1981.

Wet snow falling in large flakes occurs only infrequently in the Puget Sound Basin. Its duration is usually short since the snow is generally in a transitory state between drier snow and rain.

During an episode of snow as described above, the surface of the EPC antenna was wiped clean. The EPC antenna is oriented toward the south and a south wind had deposited approximately 3 mm of snow on the face of the antenna. The signal increased only 1.8 dB after the face of the dish was cleaned off. Therefore, most of the attenuation seems to have been due to absorption or scattering and not to a physical blockage on the face of the antenna.

At times during the month of September 1981, atmospheric conditions resulted in ducting. The results of this phenomena were two outages (see Figure 2.), although these outages were almost instantaneous. Conditions which caused this effect are quite rare, with extremely hot autumn days and cool nights. Daytime September temperatures in this area are usually much cooler.

This phenomena was found not to be frequency selective. The transmit frequencies for this system were separated by 20 MHz and both ends behaved identically. (See Figure 2.) Hence, a frequency diversity system would not have been effective in alleviating this outage.

A unique anomaly was recorded during the first rain following turnup of the experimental system in September. (See Figure 1.) This fade was a single event and occurred on both ends of the path. Members of the Radio Engineering group at GTNW have been aware of this phenomena on the 11 GHz system for many years, but did not have experimental results from which to draw conclusions. It is suspected that the antenna cover is the cause of the attenuation. Upon getting wet for the first time, the cover may attenuate the R. F. signal. One possible explanation for this is that maximum water shedding ability may not be evident until the cover is fully saturated one time.

CONCLUSION

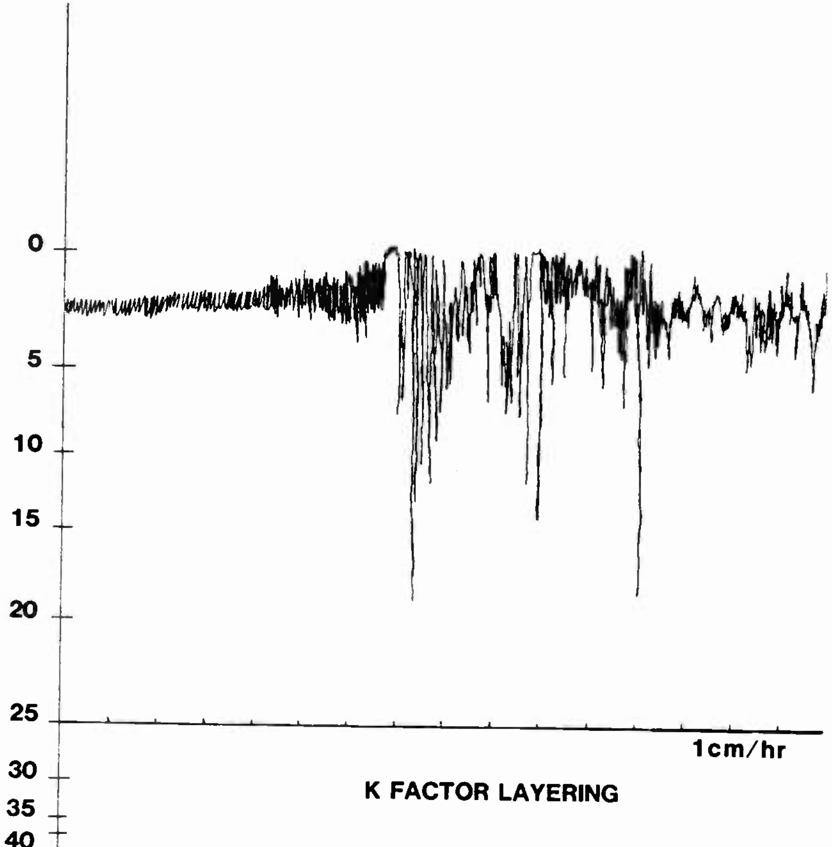
18 GHz radio systems should be examined when searching for new frequency bands in certain climatic regions. Their feasibility depends on individual applications, conditions, and the configuration of each system. In the Puget Sound Basin, the condition found to have the most adverse effect on system reliability is wet snowfall. This condition has more potential for causing a sustained outage than any other condition encountered in this study. Route diversity would certainly reduce the possibility of sustained outages caused by wet snow. System reliability consideration, of course, is the major factor in tradeoffs between system configuration and costs.

Other major considerations for engineering at 18 GHz should be rain cell size, snow moisture content, and the occurrence of conditions which cause ducting. As the light route 2 GHz spectrum fills, plans for 18 GHz light route radios will become more attractive. As more systems of this type are installed in different parts of the country, further knowledge will be gained through actual application and usage.

LITERATURE CITED

Freeman, R. L., Telecommunication Transmission Handbook, 2d ed., 1981.

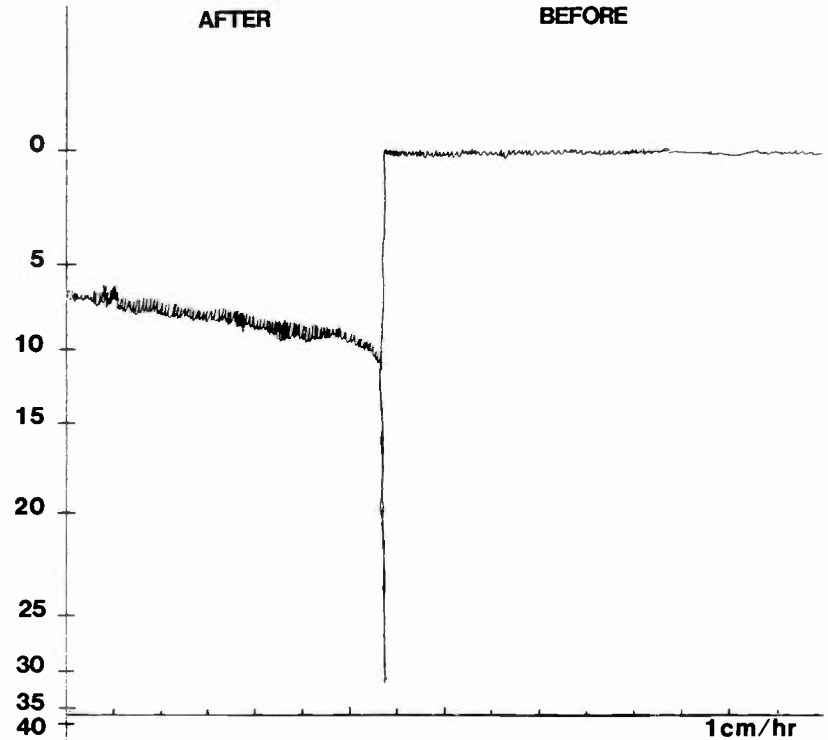
Laws, O. J., and Parsons, S. A., "The Relation of Raindrop Size to Intensity," Transactions of the American Geophysical Union, 1943.



K FACTOR LAYERING

EVERETT

fig.2



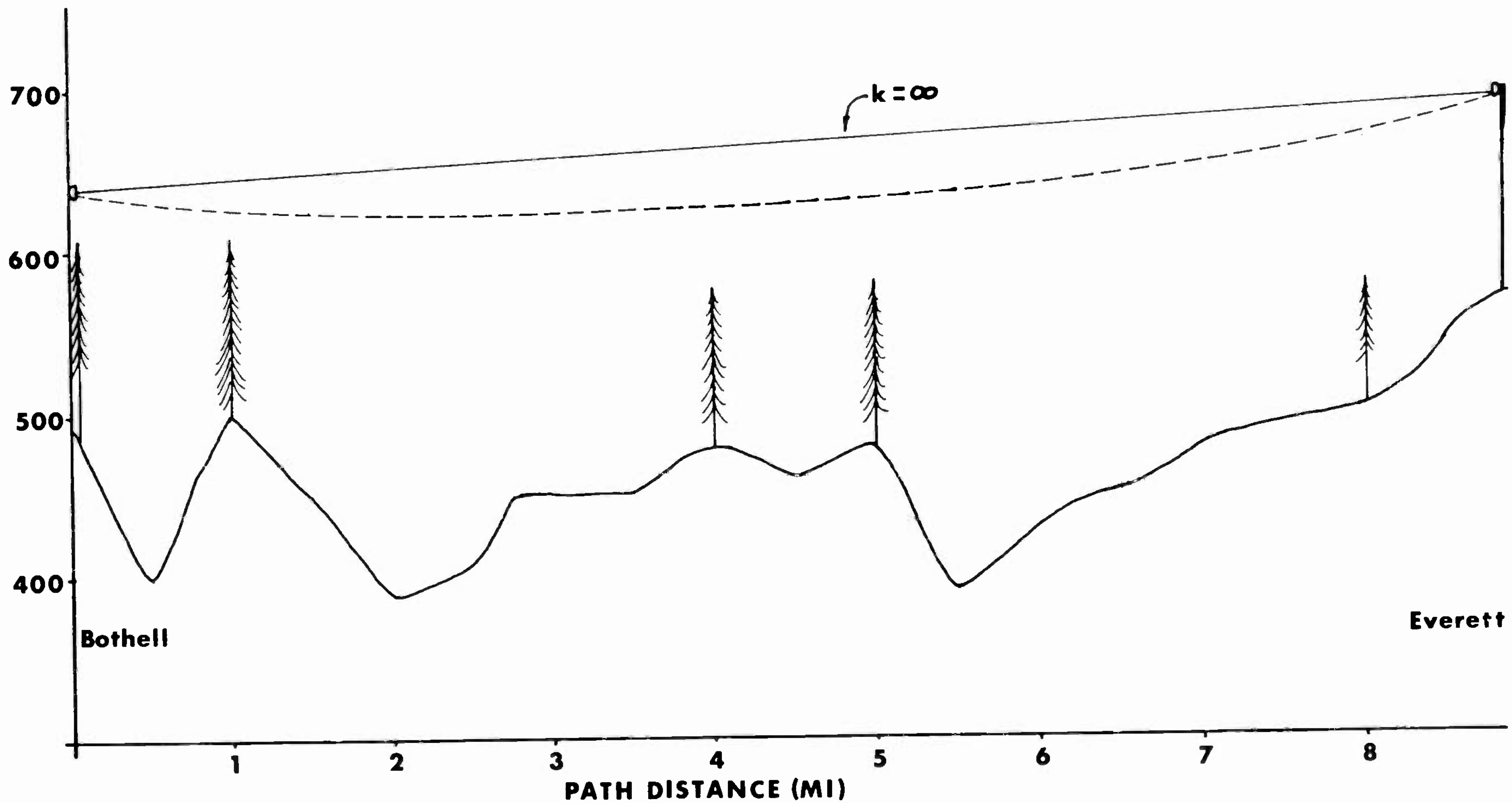
STRUBLE EFFECT

(FIRST RAIN)

EVERETT

fig.1

BOTHELL-EVERETT PATH PROFILE



--- Heavy Route Clearance (White)

fig.3

FIGURE 4

DATE 7-23-81

ENGINEER Steven R. Smith

MICROWAVE PATH DATA CALCULATION SHEET

CUSTOMER GTNW

PROJECT NO. MBO FREQUENCY 18 GHz

SYSTEM EPC - BOTH EQUIPMENT DM18

1	SITE		EPC	BOTH
2	LATITUDE		47-55-19N	47-47-35
3	LONGITUDE		122-13-34	122-14-16
4	ELEVATION (GRND ELEV.)	Ft.	570	484
5	TOWER HEIGHT (TOP OF BLDG)	Ft.	100	162
6	TOWER TYPE		SS	SS
7	AZIMUTH FROM TRUE NORTH.		183	3.5
8	PATH LENGTH	Mi.	8.9 MI	
9	PATH ATTENUATION	dB	141 DBM	
10	RIGID WAVEGUIDE	Ft.	-	-
11	FLEXIBLE WAVEGUIDE	Ft.	2	2
12	WAVEGUIDE LOSS	dB	.6	.6
13	ANT. ALIGN. LOSS	dB	2	2
14	CIRCULATOR OR HYBRID LOSS	dB	-	-
15	RADOME LOSS, TYPE*	dB	-	-
16	NEAR FIELD LOSS	dB		
17	CLOSE COUPLING LOSS (DOUBLE PASS.)	dB	-	-
18	TOTAL FIXED LOSSES	dB	4.6	4.6
19	TOTAL LOSSES	dB	150.2	
20	PARABOLA HEIGHT AGL	Ft.	104	160
21	PARABOLA DIAMETER	Ft.	4	4
22	MICROFLECTOR HEIGHT	Ft.	-	-
23	MICROFLECTOR SIZE, TYPE	Ft.	-	-
24	PARABOLA-MICROFLECTOR SEP.	Ft.	-	-
25	NEAR FIELD GAIN	dB		
26	ANTENNA SYSTEM GAIN	dB	44.7	44.7
27	TOTAL GAINS	dB	89.4	
28	NET PATH LOSS	dB	-60.8	
29	TRANSMITTER POWER	dBm	18.0	
30	MED. RECEIVED POWER (+ 2 dB)	dBm	-42.8	
31	RECEIVER NOISE THRESHOLD	dBm		
32	THEORECTICAL RF C/N RATIO	dB		
33	FM IMP. THRESHOLD (dBa)	dBm	-78	
34	FADE MARGIN (To FM Imp. Thresh.)	dB	35.2	
35	RELIABILITY SPACING	%		
36	POLARIZATION VERTICAL			

NOTES A = 99.99975

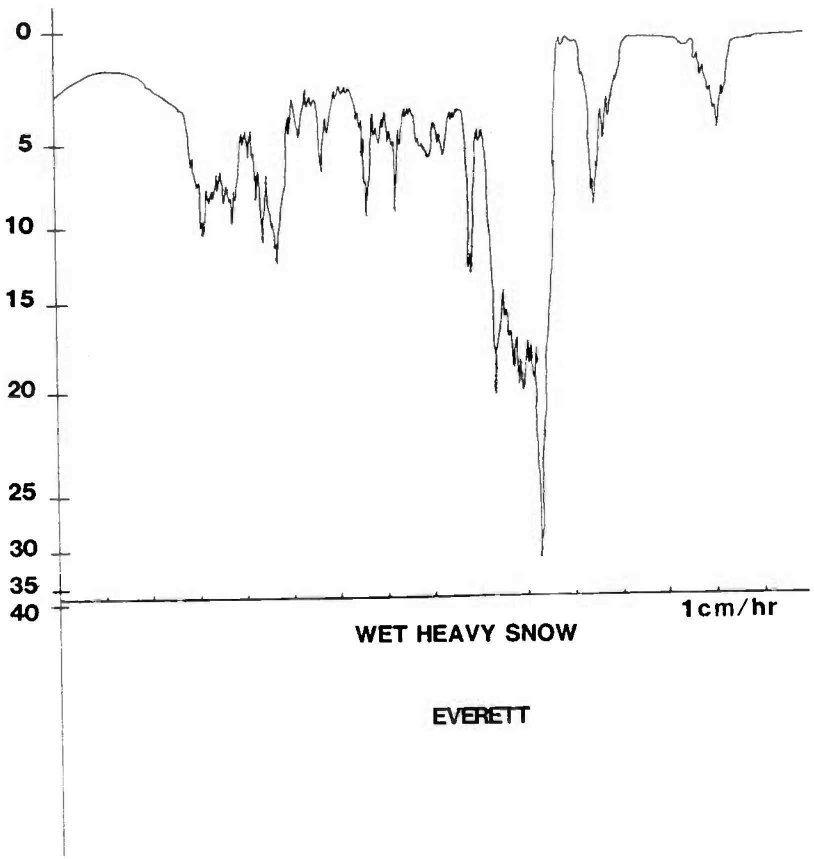

0
5
10
15
20
25
30
35
40
1cm/hr
WET HEAVY SNOW
EVERETT

fig.6

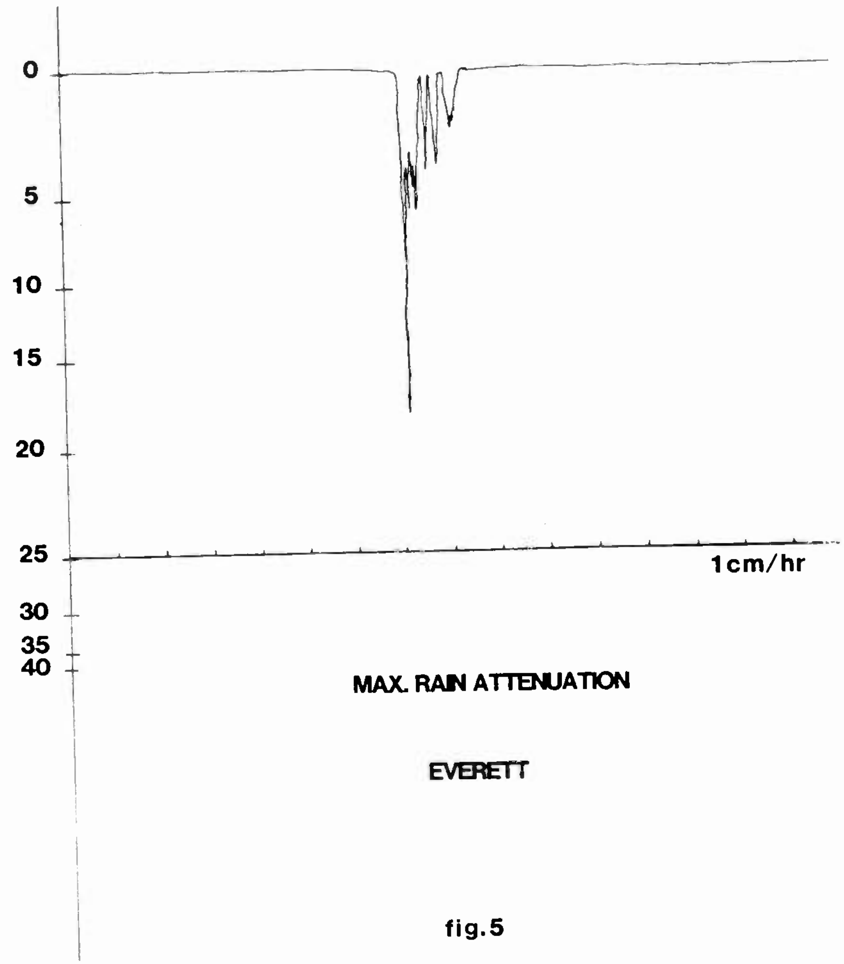

0
5
10
15
20
25
30
35
40
1cm/hr
MAX. RAIN ATTENUATION
EVERETT

fig.5

Report on Non-Translating Repeaters for ABSTL/ICR Microwave Systems

John Kean, Senior Engineer

National Public Radio

Washington, D.C.

The Commission has authorized the use of a special repeater, called a "booster", to retransmit microwave signals in cases where a line-of-sight path for microwave links is not possible. Since boosters receive and transmit on the same frequency, the Commission's action will provide more efficient use of the spectrum available to broadcasters for Auxiliary Broadcast Studio-Transmitter Links (STLs) and Intercity Relay (ICR) Radio Services. Previously, the Commission permitted only frequency-translating repeaters for these services.

This paper reviews the new rules issued in this proceeding (BC Docket 82-20; RM-2500) and discusses the application of boosters in several situations. A case study for a system designed by NPR is included.

Background

Two types of facilities are normally available to broadcasters to relay program material from the studio to a remotely-located transmitter site. These are either wireline circuits operated by common carrier services or station-owned STL or microwave facilities. Due to the demand for the highest-possible quality for program relay, and more recently the anticipation of increases in tariff rates for wireline services, the use of microwave STLs and ICRs has steadily increased. Because microwave frequencies are use in the STL service, an adequate line-of-sight must usually be maintained between antennas at the transmitter and receive sites.

A petition filed with the Commission several years ago by Marti Electronics, Inc., stated that some broadcasters have been unable to use this mode of interconnection because of physical obstructions that prevent establishing a clear line-of-sight transmission path between the two points (due to intervening hills, tall buildings or other obstacles). Furthermore, the ten 500 kHz channels now allocated for aural STL and ICR operations may not meet the need for service in large metropolitan areas. The use of microwave boosters would provide a measure of relief for broadcasters located in these areas, according to Marti, permitting more intensive use of the present spectrum allocation.

Broadcasters now use two types of radio relay devices to circumvent obstacles in their transmission paths: passive and active repeaters. Passive repeaters require no power input and simply "bend" the microwave signal to provide the necessary path clearance. Figure 1 shows three types of systems using passive elements.

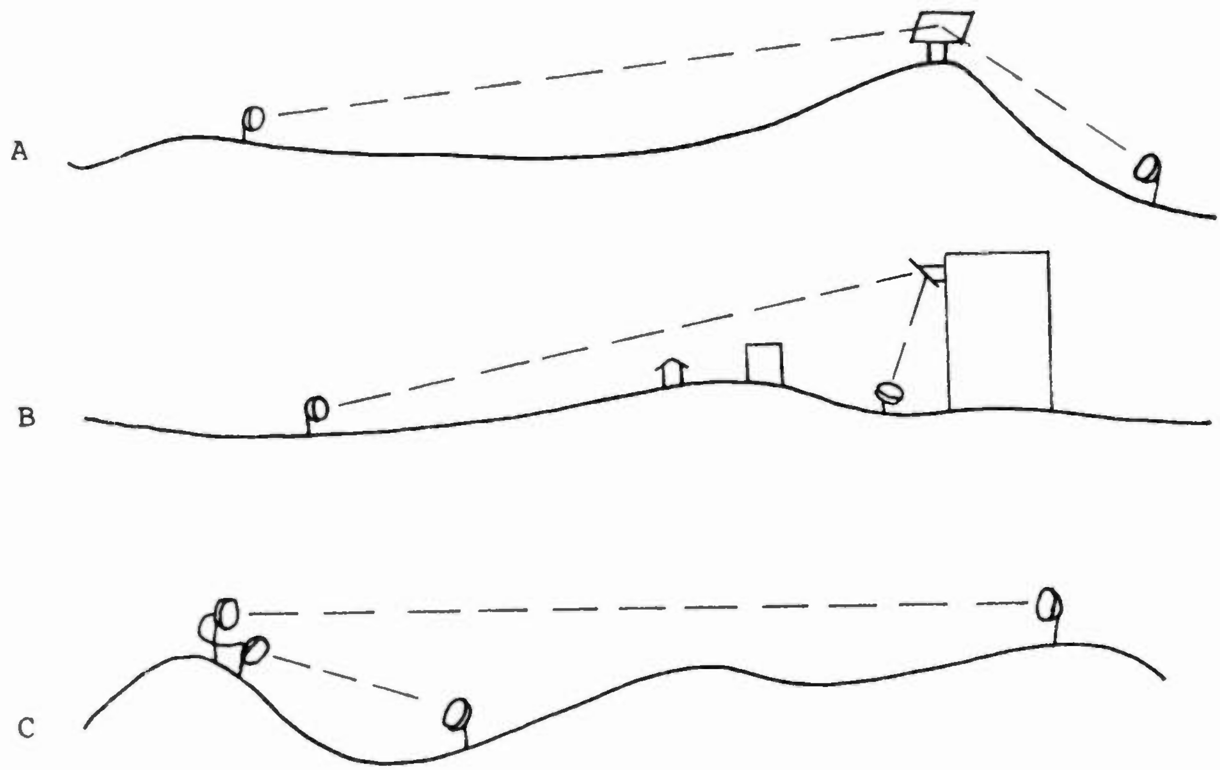


Fig. 1 - Types of passive two-hop antenna systems. A, system using a radic reflective "billboard" reflector; B, reflector elevated by building or other structure to serve as "periscope" antenna system; C, intermediate antennas connected back-to-back to re-direct signal path.

Passive elements are satisfactory for relatively short hops over and around obstructions, but as either first or second path becomes longer, the signal loss becomes critical. The Marti petition pointed out that because signal losses require large physical size of the antenna elements, passive systems can be impractical. This will be shown later in the case study.

Prior to this rule making, active repeaters have been required to convert the incoming carrier frequency before retransmitting a signal. Under the former rules, broadcasters typically used a "baseband" type repeater in which a standard microwave receiver is directly connected to second mircowave transmitter operating on another frequency. In some cases, a hetrodyne repeater has been used, in which the amplification and filtering is provided at an intermediate frequency and then the signal is translated to a new output frequency.

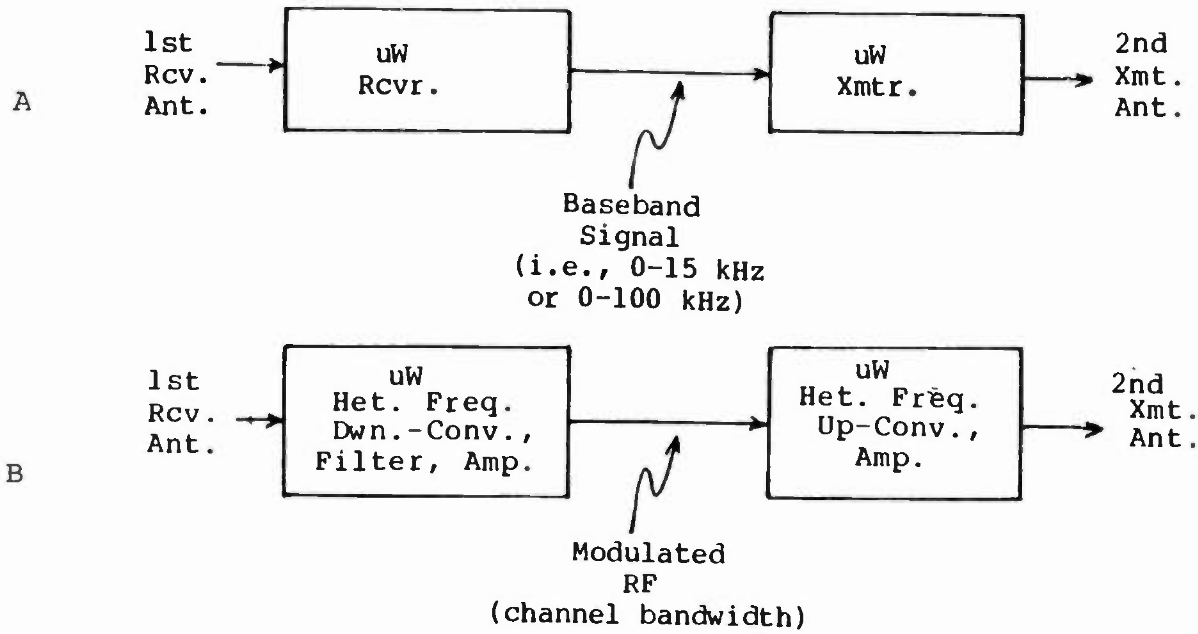


Fig 2 - Active, translating repeater functional blocks. A, baseband repeater; B, hetrodyne type repeater.

The advantages of boosters over baseband and hetrodyne repeaters may include:

- o one channel needed for operation, thereby doubling spectrum efficiency;
- o less costly, since unneeded circuity may be eliminated, e.g. frequency convertors and demodulation and remodulation stages;
- o with a large variety of "stock" microwave amplifiers, no more than necessary cost or transmit power need be used in the second hop;
- o less complex than a non-translating (booster) repeater.

The newly authorized microwave boosters share characteristics of both active and passive repeaters. But because they use the same frequency for both reception and retransmission, isolation between the input and output circuits must exceed the maximum overall gain of the repeater. They must depend on antenna directivity, physical spacing and shielding to avoid feedback or oscillation. Figure 3 shows the basic diagram for a booster and possible feedback mechanisms.

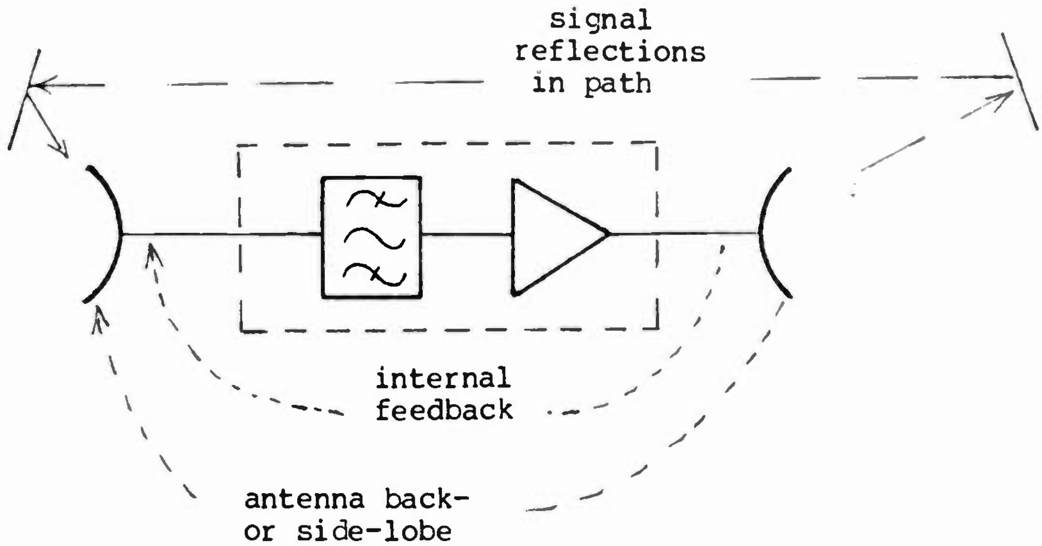
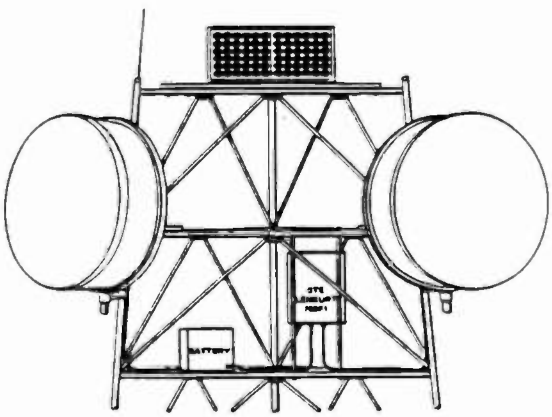
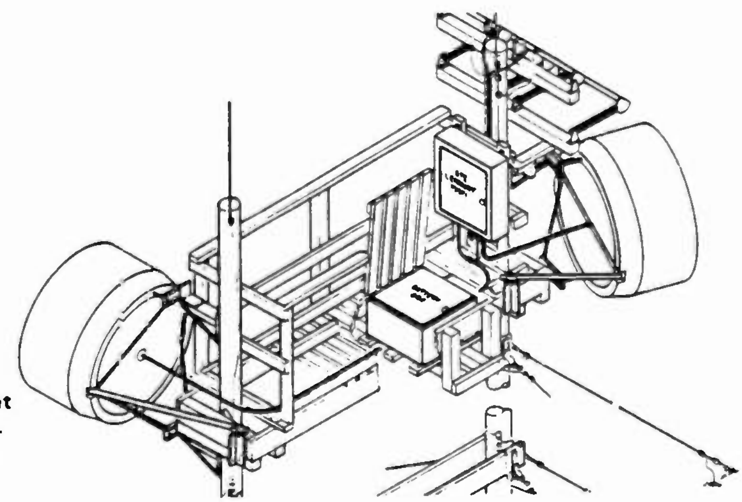


Fig 3 - Non-translating "booster" repeater with some potential feedback mechanisms.



RF repeater installations with antenna heights of 65 feet or more usually require a standard steel tower. For convenience, batteries and solar panels can be mounted at the tower base.



The 700F1 RF Repeater using a wooden structure is available as a self-contained RF repeater for use with customer-furnished antennas and power equipment, or as a complete package including repeater, antennas, solar generators, or battery charger and batteries.

Fig 4 - Views of the GTE Lunkurt 700F1 non-translating repeater system. Intended for microwave radio systems operating in the 1.7 to 2.3 GHz frequency band, this "booster" is capable of repeating up to 960 FDM or 120 PCM channels or a FM video signal. The system is designed for unattended, uncontrolled operation at remote locations along the radio path.

Summary of Amendment to Rules for Microwave Boosters

The FCC chose a relatively relaxed approach toward the operation of microwave boosters, discussed below:

- No minimum isolation value. The FCC indicated that the isolation between the input and output circuits (including the receive and transmit antennas) must be sufficiently greater than the gain of the booster so it may operate properly. Good design practice for is expected to insure reliable operation.

- Feedback detection and automatic shutoff not required. To quote the FCC: "appropriate observations must be made at the receiving end of the circuit to ensure proper operation". There is no requirement for remote control or fail-safe operation, since it is expected that the STL/ICR licensee will likely be the most seriously affected by a malfunction. The rules note that observations may be made by monitoring the broadcast station's transmitter signal, if applicable.

- Especially-narrow beam antennas or RF filters not required. Boosters can be made as resistant to cochannel or adjacent channel interference as translating repeaters when necessary, reasoned the Commission. They do not generally require the use of such specific spectrum-conserving measures.

- Carrier detection and booster shutoff not required. To help make boosters "a low-cost alternative to the problem of path obstruction", the FCC will not require either remote or automatic cutoff of a booster's operation when not in use, but "adequate safeguards [must be provided] to prevent improper operation".

- No maximum power output or gain limit. Although the Commission's NPRM had suggested one watt as a limit for boosters, the commentors disuaded the Commission. The power output must still be no more than "that required to render satisfactory service." However, authorizations will be given only for paths that cannot be covered by a single station.

- Type acceptance not required for booster equipment. Notification will be required only for ABSTL and ICR equipment operating in the 18 GHz frequency band. Despite the printed rules, the Commission may be unable to data process <u>any</u> applications until January, 1985, when revision to the computer programs will be complete. The Policy & Rules branch suggests that waivers of the type-acceptance rule will be submitted with applications until then.

- No ID requirement. Aural broadcast microwave booster stations will be assigned individual call signs, but station identification will be accomplished by retransmission of ID per 74.582(a).

Booster Case Study

A 950 MHz intercity relay system is to be installed by NPR to interconnect the studios of member station WNYC with one of the network's satellite uplinks. Due to the high-rise apartment buildings surrounding the uplink site in Brooklyn, a line-of-sight path from the studios in downtown New York is not possible. A path-bend near the uplink (receive site) is required to direct the signal over the buildings. Due to the angles and structures involved a "billboard" reflector is not feasible, although the path loss would be acceptable.

Figure 5(a) shows the overall path, which runs 7.6 miles on the first segment and is completed by a 0.2 mile "dog leg". The path-bend point (point B) is the equipment room atop a nearby apartment building as shown in Figure 5(b).

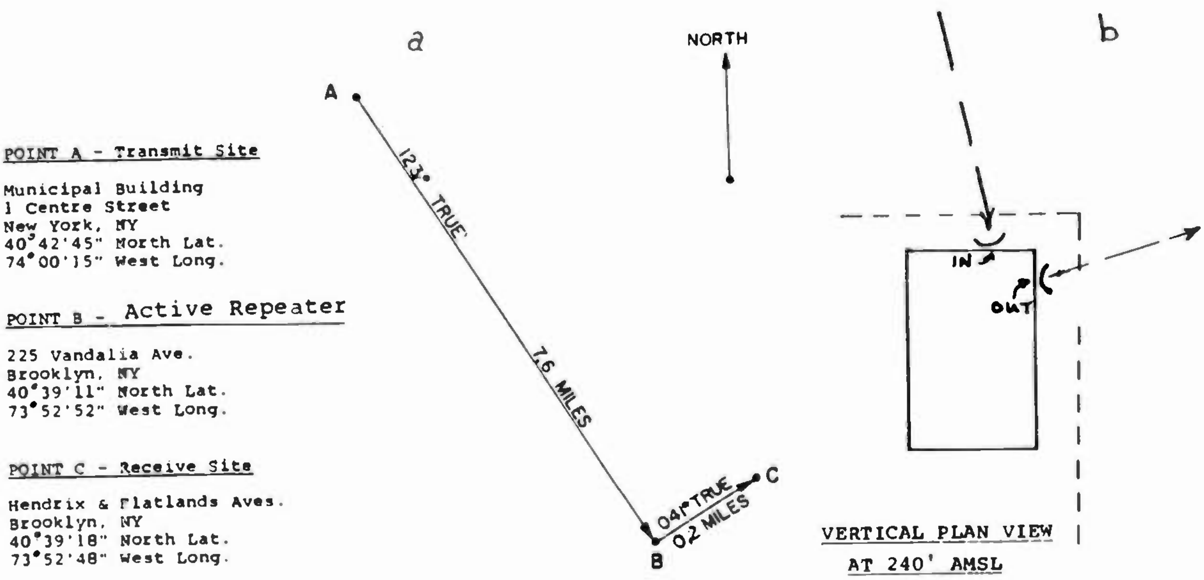


Fig 5(a) - Diagram for two-hop path used by NPR.
Fig 5(b) - Vertical plan view at 240' above ground for repeater.

Using a pair of six-foot parabolic antennas connected back-to-back (passively) at the path bend point, the following table shows received signal strength after various system gains and losses.

The nominal signal level resulting from a passive path bend is -72.9 dBm, or 50 microvolts in 50 ohms. While this would produce reasonably quiet performance with most STL receivers, there is too little margin for signal fading, component tolerances, immunity to interference, and so on.

The bottom line of the table shows that the received signal level would be -48.6 dBm, or 830 microvolts when a booster amplifier is used. This level provides more than adequate reliability for system operation.

ICR Path Level Study,
WNYC to NPR Satellite Uplink, New York City

950 MHz Xmtr.	TPO	+40.8	dBm		
WNYC ICR xmtr.	ant. gain	22.0	dBi		
Intermediate Rec.	" "	22.0			
" Xmt.	" "	22.0			
ICR Rec.	" "	19.0			
		+85.1	dBi		
Path Loss (to repeater)		114.4	dB		
" " (to receiver)		79.8			
Trans. Line loss, ICR xmtr.		0.5			
" " " , repeater in/out		0.6			
Connector " , (8 x 0.3 dB)		2.4			
		-198.8	dB		
Rcvd. signal, nominal rcvr. in		-72.9	dBm	50	uV
Booster gain, overall		+24.3	dB		
Rcvd. signal, nominal rcvr. in		-48.6	dBm	830	uV

Figure 6 is a simpified block diagram of the proposed booster. A small, commercialy-produced microwave preamplifier is used, having a gain of approximately 30 dB and a maximum output power of around +10 dBm (10 mW). It requires less than 100 milliamps at 15 volts, provided by a small power supply. A rechargeable battery pack can provide backup power for many hours if the AC supply is interrupted.

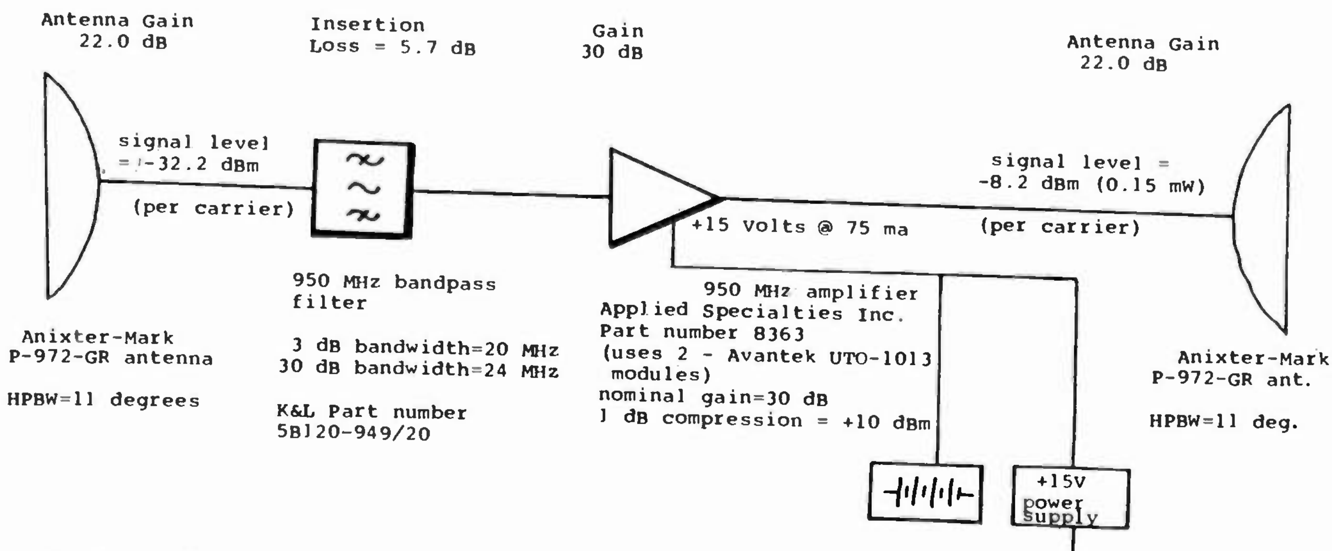


Fig 6 - Block diagram of proposed 950 MHz booster amplifier.

A tubular-type bandpass filter precedes by the amplifier to eliminate reradiation of strong UHF television carriers. The bandwidth is not sufficiently narrow to reject other STL frequencies, but this should not present a problem for several reasons. First, no STL/ICR carriers appear within the main beam of the antenna. Second, since the receiving antenna is quite directional, any carriers originating well off to the side will be highly attenuated. (It can be assumed that other 950 MHz transmitters also use directional antennas.) Third, the booster output is very low power and is radiated by a directional antenna that greatly narrows the possible interference pattern.

The amplifier, filter and input/output connectors are housed in a small, weatherproof box and will be mounted on the same wall as the antennas. The power supply and batteries will be placed inside the equipment room to protect against temperature extremes and corrosion. A few feet of low-voltage wire connects the power source to the booster.

The proposed booster system should be quite free from feedback due to the wide difference between the amplifier gain (24 dB, approximately) and the isolation of the antennas (at least 60 dB, as mounted). No control of amplification is provided, but due to the relatively low power of the booster and the directivity of both input and output antennas, interference is not expected to be a problem.

Conclusion

As a matter of interest, Figure 4 shows a non-translating booster which is commercially-made and approved for use by the FCC on microwave radio systems in the 1.7 to 2.3 GHz frequency band, manufactured by GTE Lenkurt, Inc. The model 700F1 RF Repeater is capable of repeating up to 960 FDM of 120 PCM channels or a FM video signal. The 700F1 equipment is comprised of broadband linear amplifiers, circulators, RF filters, isolators and hybrids, enclosed in a small weather-proof enclosure.

Nominal output level at the antenna port is +10 dBm or +20 dBm. Of special interest is its capability of amplifying in both directions on different frequencies (duplex operation). It used storage batteries which are solar-recharged. The circuitry provides 45 dB or 55 dB of gain; overall gain varies from 96 dB with 4-foot antennas to 129 dB with 15-foot antennas.

The GTE Lenkurt repeater does not require field tuning, field adjustments, according to advertising literature from the company. Recommended maintenance period is one year where the operation of the storage batteries and RF output power are checked. This system is evidence that well-planned microwave boosters can be used to complete difficult paths.

Boosters will not necessarily be the best choice in all microwave relay situations. Where signal levels are adequate, for example, a simple passive reflector of the "billboard" or "periscope" type mentioned earlier may be more cost-effective and is certainly simpler and more reliable than repeaters using amplifiers and directional antennas. In the completion of long paths, particularly in microwave systems which relay the composite stereo baseband, high gain and significant output power may be necessary, requiring careful input/output isolation to avoid feedback, and costly safeguards against malfunctions.

Where boosters are feasible, the lower cost, complexity, and spectrum requirements over that of translating repeaters should prove a favorable choice. The relaxed nature of the Commission's new rules will permit broadcasters a large degree of flexibility in planning and operating ABSTL/ICR systems employing boosters.

FIELD TEST RESULTS OF A NEW PHASE/AMPLITUDE CORRECTION SYSTEM FOR DIRECTIONAL AM ANTENNAS

Grant Bingeman, P.E. & Walter C. Jamison, P.E.
Continental Electronics Fisher Broadcasting Inc.
Dallas, Texas Seattle, Washington

Because an AM signal contains energy on both sides of the carrier frequency, it is important to consider the characteristics of the transmission medium at the sideband frequencies, not just at carrier. An ideal medium will not affect either the amplitude or phase of an AM signal. However, there are not many real antennas that look like ideal loads to real transmitters. Note that an average load may look average to an ideal transmitter, but bad to a real transmitter.

Rather than modify the transmission medium itself, it is sometimes more cost-effective to correct its phase and amplitude response externally. One can place a broadbanding network between the transmitter and the common-point of the directional antenna which reduces sideband VSWR to some nominal level while simultaneously providing a linear-phase (constant group-delay) frequency response (Figure 1).

The results of such correction are improved distortion, improved depth of modulation or frequency response, and improved separation; in other words, better quality audio. In some cases, such correction may also make the difference between compliance and non-compliance with FCC equipment performance measurement limits.

There are several considerations which make external, as opposed to internal, broadbanding attractive. Internal broadbanding, or modification of the phasing and coupling networks, may require extensive adjustment in order to maintain the licensed antenna monitor values, and it may also require field intensity measurements in order to verify the radiated pattern. Any power loss in discrete internal broadbanding networks cannot be made up by an increase in transmitter power. Hence the RMS or size of the pattern (i.e: coverage) may shrink. External broadbanding avoids all of these problems.

External broadbanding, if it is to be univerally applicable and thorough, consists of four steps (Figure 2). Step 1 requires a reduction in the system Zo. This is done in order to reduce losses and inductor size in the network. A typical transformation ratio is 50 to 15 ohms. Concurrent with this, the sideband impedances are rotated to series resonance. This is defined as approximately equal sideband pair resistances, and approximately equal magnitude, but opposite sign, sideband pair reactances. The lower sideband reactance must be negative (Figure 3). The sideband pairs of a 1000 kHz carrier are 998 and 1002 kHz, 995 and 1005 kHz, etc.

The second step requires transformation of the impedances obtained by Step 1 into parallel resonance. This is done by placing a parallel-resonant circuit in shunt. The result is fairly constant resistance across the desired passband, and equal, but opposite sign, sideband pair reactances. For parallel resonance, the lower sideband reactance must be positive (Figure 4). Since impedances plotted on a Smith chart are constrained by the laws of physics to travel in a clockwise direction with increasing frequency, it is impossible to obtain a perfect constant-resistance curve in Step 2. However, one can come close.

The third step consists of the insertion of a series-resonant circuit to cancel the sideband reactance obtained in Step 2. At this point very little reactance remains across the passband, and nearly constant resistance. We now have a broadband load.

In Step 4 the original system Zo is reconstructed with the correct phase shift to produce sideband impedance symmetry at the PA of the transmitter. As in Step 1, this consists simply of a tee network.

Four-stage broadbanding can typically improve a +/- 10 kHz sideband VSWR of 2.0 to 1.2 or less. Thus across the 20 kHz passband, the worst VSWR would be 1.2. The power loss in such a network would be near ten percent. In order to deliver 52.5 kilowatts to a common-point, a transmitter would have to produce about 58 kilowatts. To minimize power loss, specially constructed toroidal coils are necessary (Figure 5). A 25 microHenry toroidal coil with a Q of 1500 measures about 30 inches in diameter and about 18 inches in height. The losses in a broadbanding network designed to pull a 3.0 VSWR down to 1.2 are only about 12 percent, all other things being equal.

Another means of controlling losses is the use of a lumped-parameter version of a shorted quarter-wave stub in Step 2, as opposed to a simple parallel-resonant LC tank. The circulating current in a simple tank would be several hundred amperes in a 50 kilowatt, 2.0 VSWR system. At 1 mHz, the tank would require unrealistic component sizes -- about a half microHenry inductor and a .05 microFarad capacitor (that is a lot of parallelled vacuum caps). These problems are almost eliminated by use of a 90 degree network and a series-resonant circuit (Figure 6), where high current is traded for high voltage, but the component sizes are more easily realized. An additional advantage of the stub approach is its far greater adjustability. The shunt reactance-versus-frequency characteristic can be tailored to fit the real-world idiosyncrasies of a typical antenna, such as unequal sideband pair VSWRs.

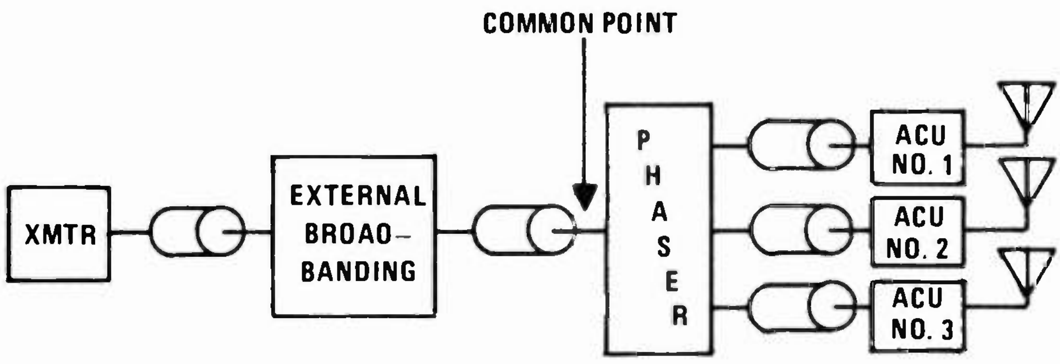


FIGURE 1

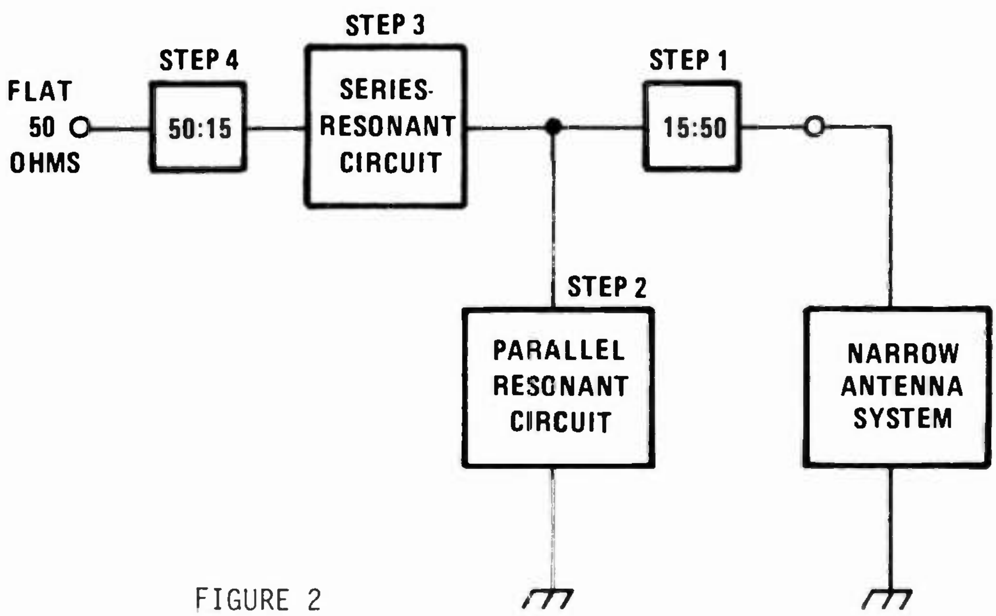
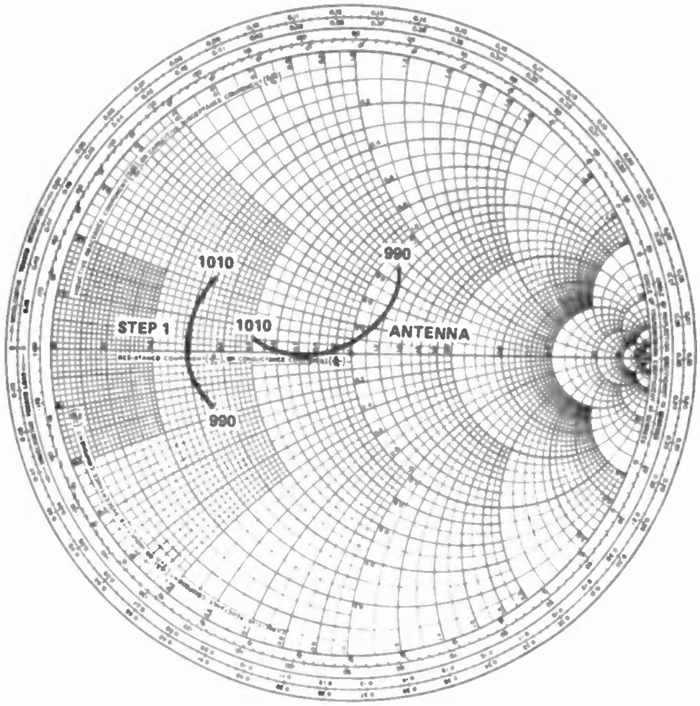


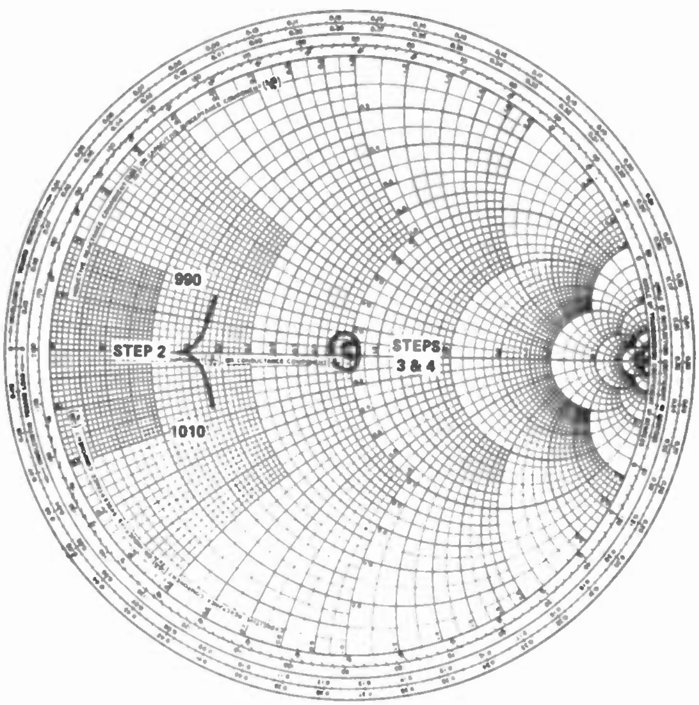
FIGURE 2

BROADBANDING NETWORK



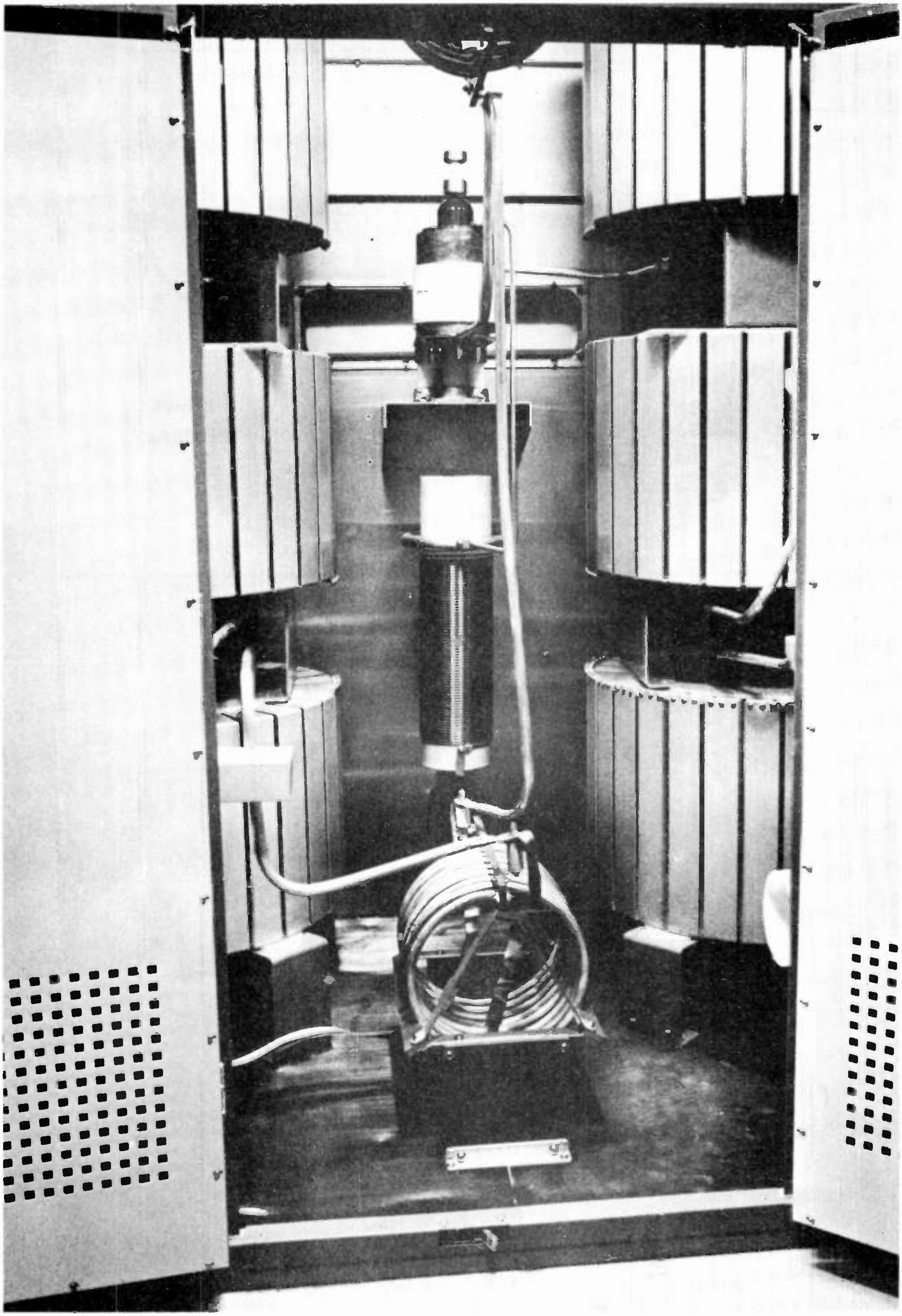
IMPEDANCE ROTATION
AND STEP DOWN

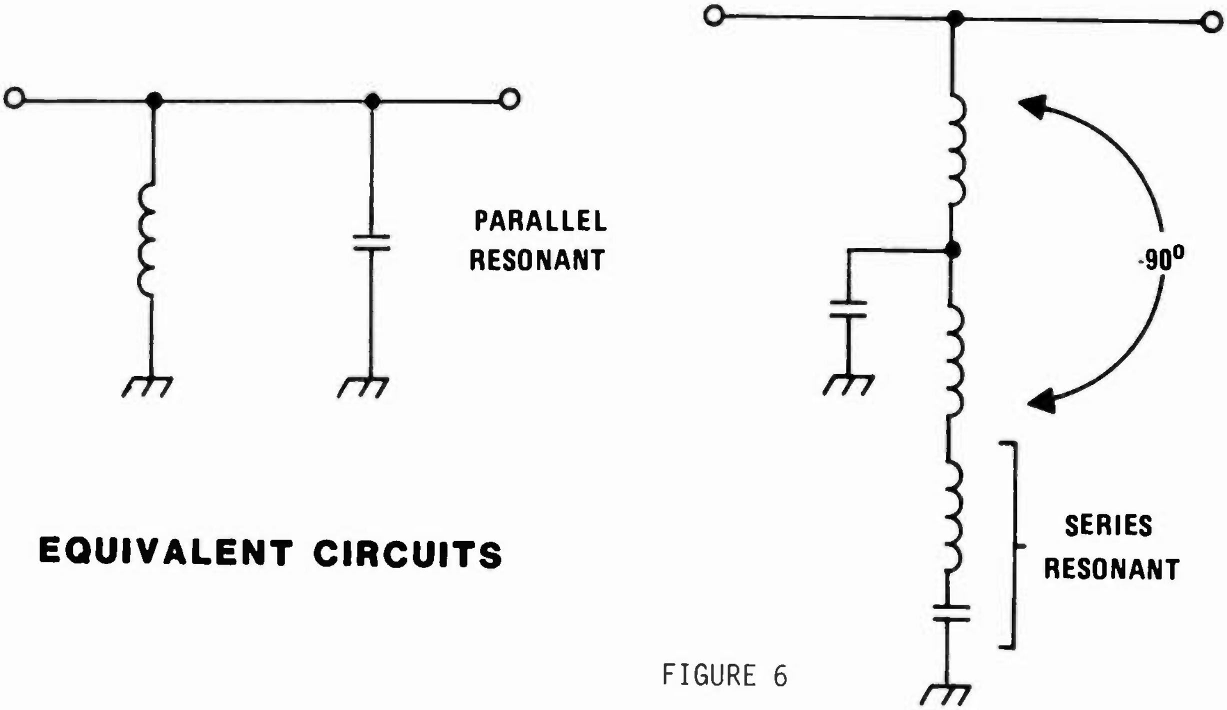
FIGURE 3



SHUNT TRANSFORMATION
AND SERIES CANCELLATION

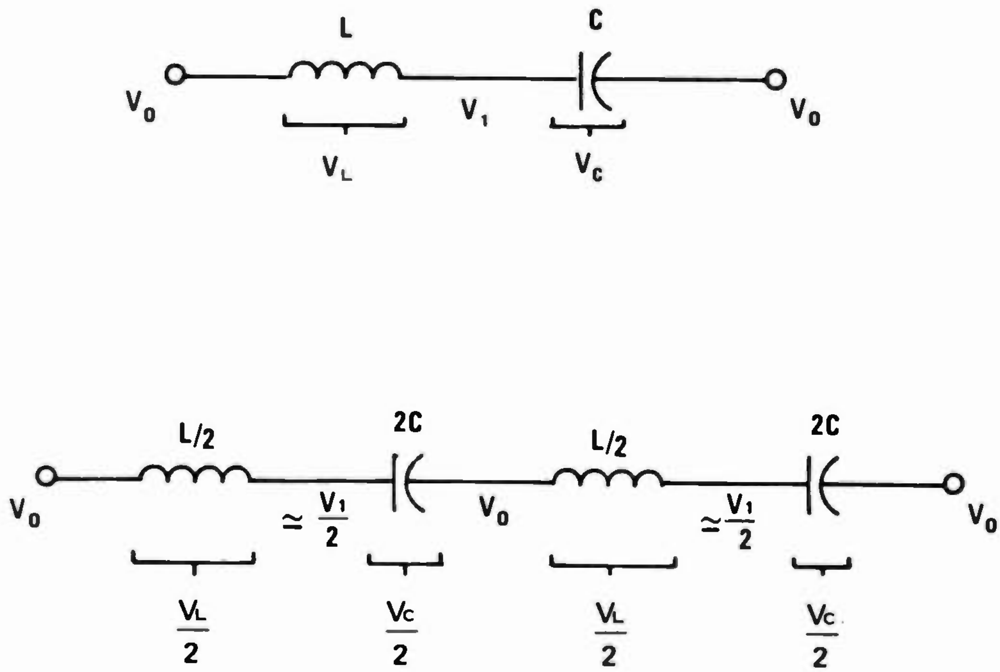
FIGURE 4





EQUIVALENT CIRCUITS

FIGURE 6



REDUCING VOLTAGE STRESS

FIGURE 7

Because the series-resonant circuits used in this type of network are required to have high loaded Qs, high voltages can develop at the LC junctions. It is sometimes desirable to reduce the junction voltage and the component voltages by dividing series resonance into two or more sections (Figure 7).

Shunt stray capacitance can cause problems for the series-resonant circuits by creating considerable impedance transformation. This effect can be controlled by the addition of an inductance in parallel with the stray capacitance, which partially tunes out the stray reactance (Figure 5). Careful consideration must be given to the voltage gradient and heating in these compensating coils.

Radio Station KOMO has been owned and operated by the Fisher family of Seattle since 1926. In 1948 KOMO moved into a new transmitter facility on Vashon Island located in Puget Sound, and increased power to 50 kilowatts (Figure 8). KOMO operates non-directional daytime and uses a three tower directional antenna at night.

The original directional common-point Q was considered satisfactory at the time, but it was somewhat high by today's standards. Traps installed at the base of each tower to prevent the re-radiation of another station just 90 kHz away have resulted in a VSWR increase to about 2.0 at the plus and minus 10 kHz points. Figure 9 was measured at the completion of an antenna system rehabilitation project several years ago.

Plans for installation of a new 50 kilowatt transmitter led to an investigation of how to improve the directional antenna load impedance. We wanted to:

1. optimize the performance of the new transmitter,
2. improve the modulation depth capability in directional mode for high modulating frequencies,
3. reduce the distortion at high modulating frequencies, and
4. optimize the AM stereo performance of the transmission system.

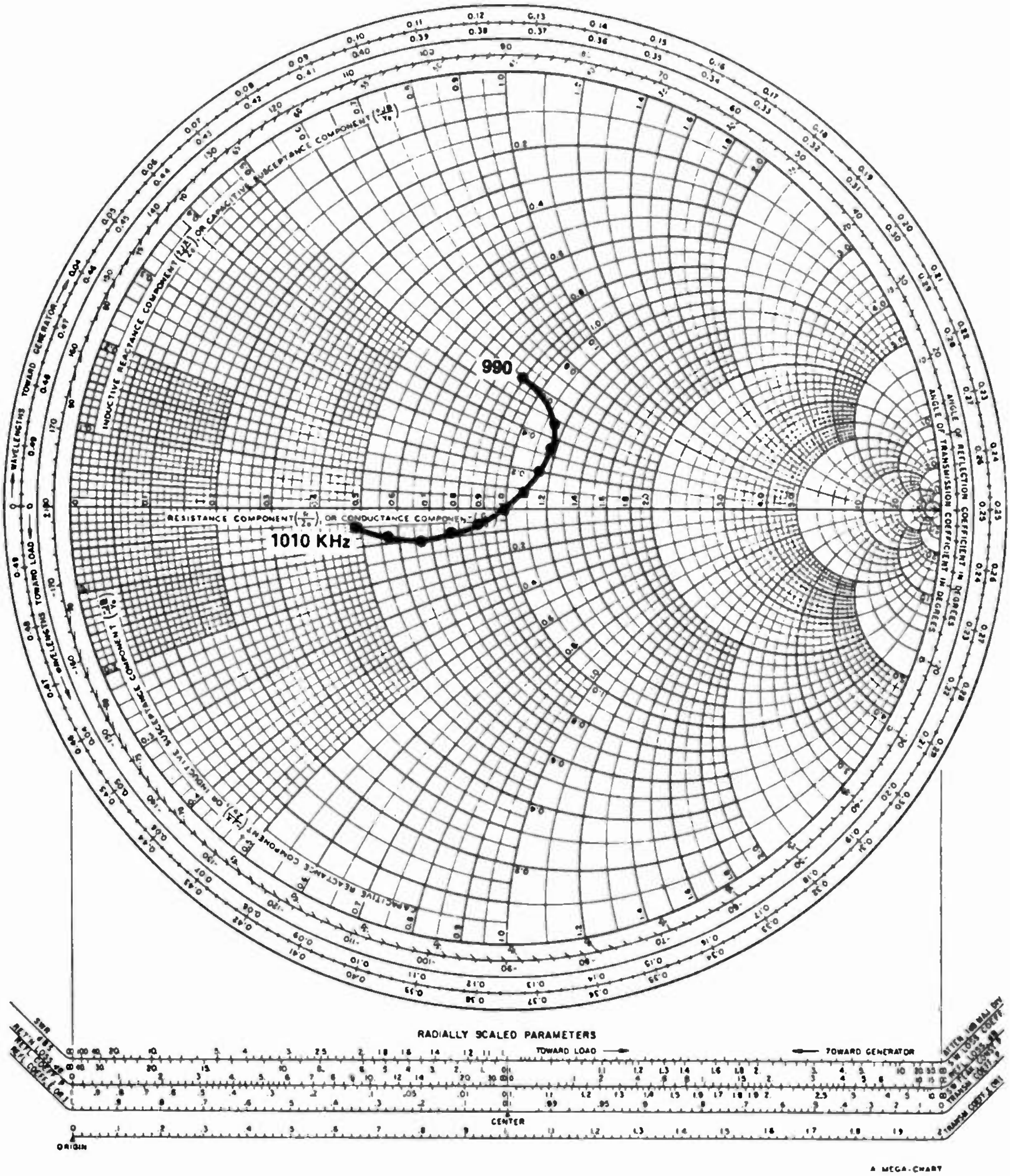
While discussing this with Continental Electronics, Grant Bingeman proposed a broadbanding network that would improve the transmitter load impedance bandwidth without changing the antenna system. KOMO then purchased a new Continental transmitter and broadbanding network.

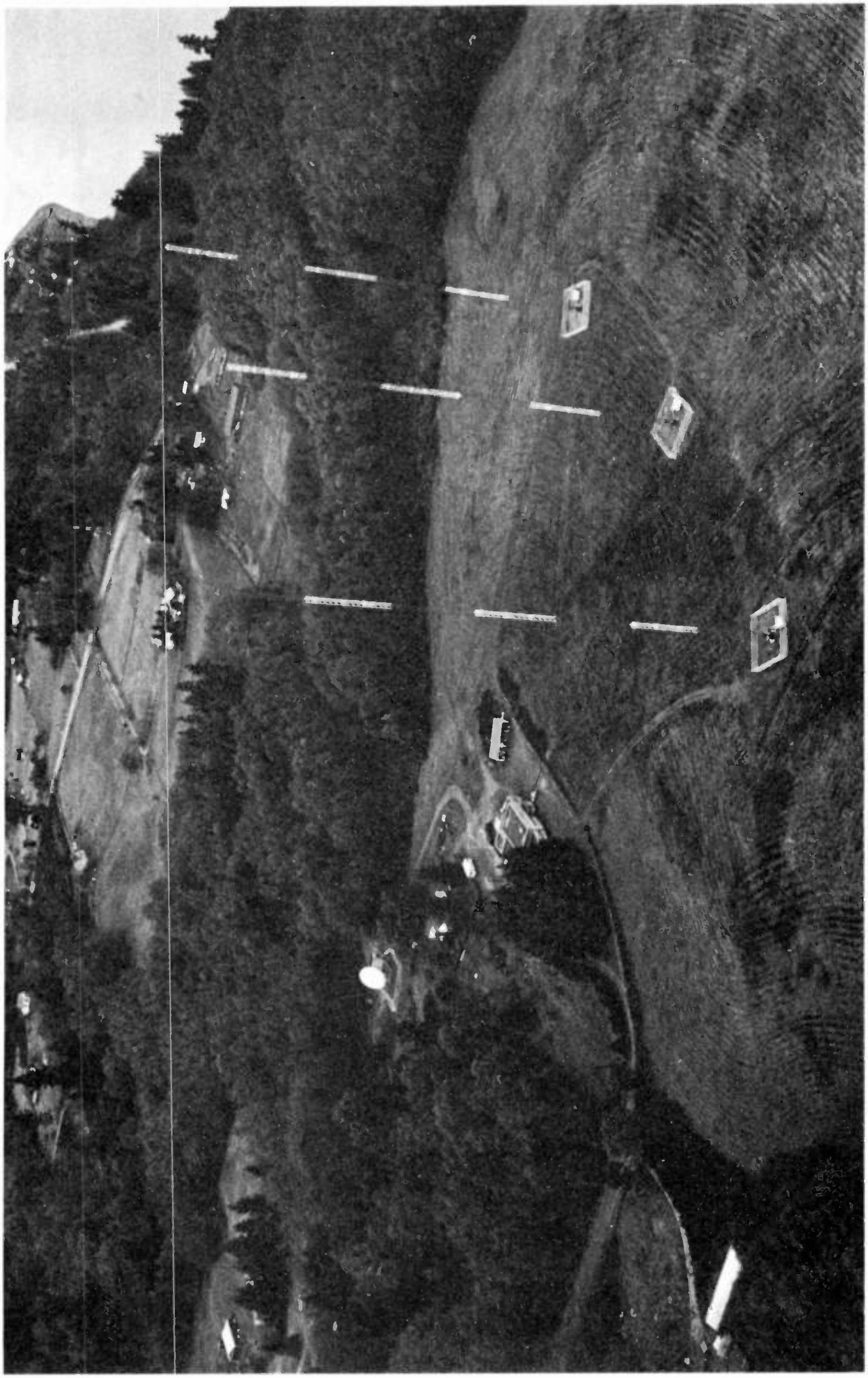
Installation was straightforward. Since we use it only in the directional mode, we installed the network in the basement under the phaser, so that it would be near the transmitter RF switching matrix (Figure 10). We provided a fan for air circulation and connected the door interlock switches into the antenna interlock circuit because of the high voltages within the cabinet.

Power losses in the broadbanding network require the transmitter to operate at approximately 58 kilowatts power output. This has not caused any problem with the transmitter.

NAME	TITLE KOMO COMMON POINT	DWG. NO. 9
SMITH CHART FORM 82-BSPR (9-66)	KAY ELECTRIC COMPANY, PINE BROOK, N.J. © 1966 PRINTED IN U.S.A.	DATE

IMPEDANCE OR ADMITTANCE COORDINATES






HOT

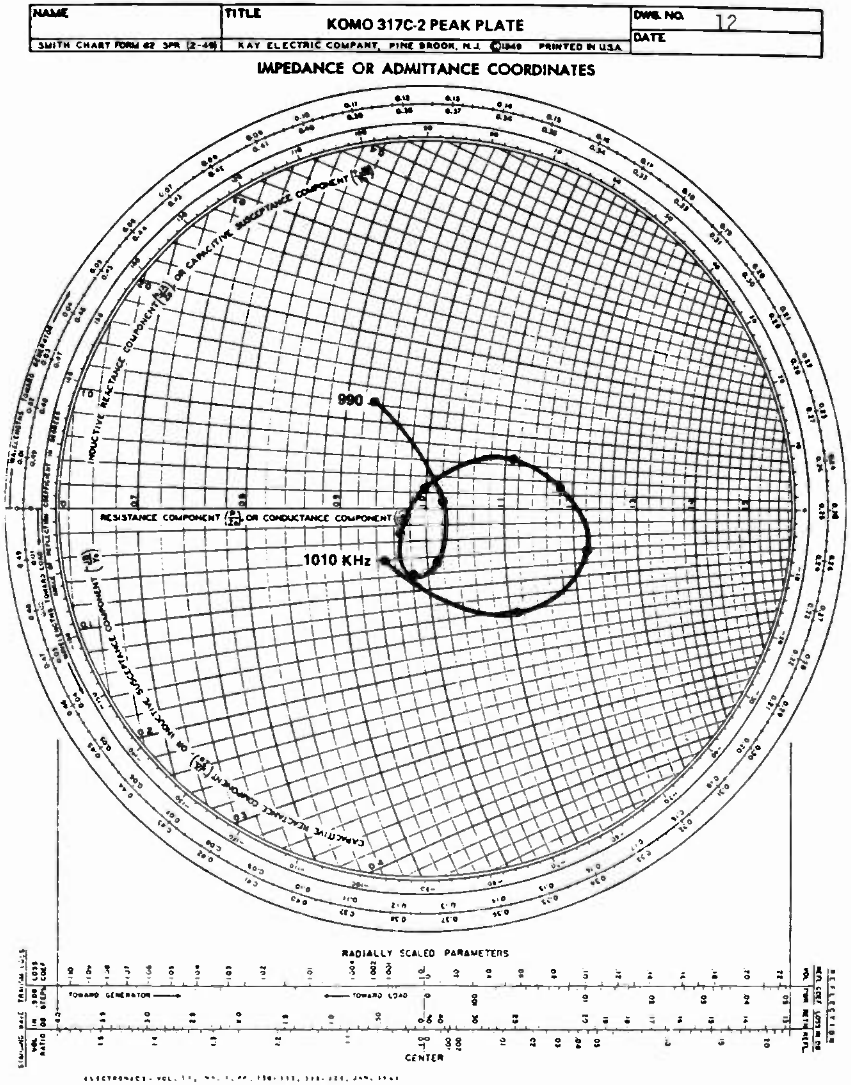
After adjustment, we operated the network at full power using tones and music modulation. One of the compensating coils ran too hot, so Continental supplied new larger coils for all three compensating coils. After installing the new coils and making minor adjustments, the network input impedance was measured, with the results plotted in Figure 11 (Note that this is an expanded-scale Smith chart). The peak tube load impedance is plotted in Figure 12.

The purpose of all this is to improve the audio frequency performance of the transmitter and antenna system. When audio measurements are made on a system with a high-Q load impedance, particular attention should be given to the location and method of obtaining the RF sample to measure. The sample should be representative of the signal in the directional antenna major lobes. Suppose the high-Q common-point impedances are oriented upward on a Smith chart, indicating a lower resistance for sidebands than for carrier. Supplying the sideband energy to fully modulate the carrier requires higher sideband currents at this point, than if resistance were constant. A current sampling transformer located at this point will indicate excessive modulation at higher frequencies. Addition of 90 degrees of delay would rotate the impedances 180 degrees on the Smith chart, the sideband resistance would be higher than at carrier, and a current sample will indicate too little modulation at high modulating frequencies. In a similar manner, if a voltage divider is used for sampling, the accuracy of the sample depends on both resistance and reactance remaining constant across the operating bandwidth.

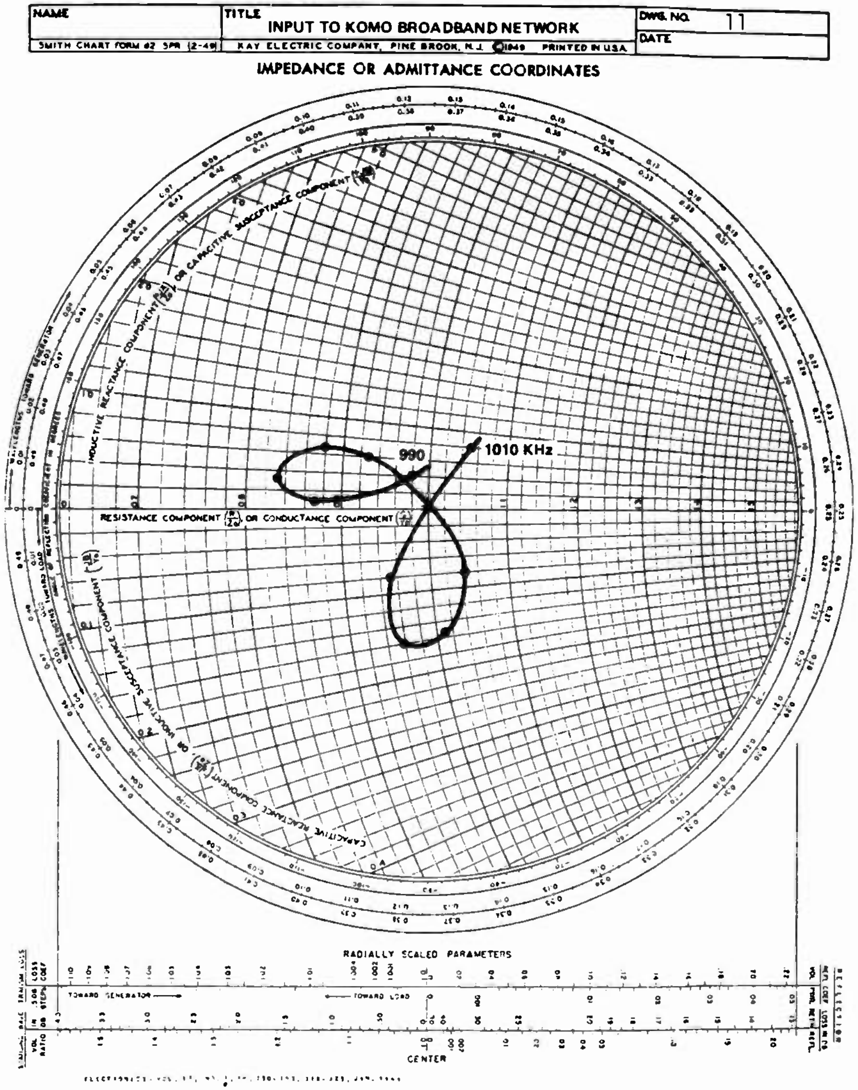
The KOMO common-point resistance measurements, Figure 9, show a large variation of resistance in the upper sideband and a large variation of reactance in the lower sideband, making that point unsuitable for either current sampling or voltage divider sampling (Figure 13). It has been our practice to use the signal from the tower 1 sample loop for audio response and distortion measurements. This tower has the highest power and current of the three towers and is believed to be most representative of the signal in the major lobes. With the broadbanding network in use, we would expect to see similar results at the tower loop and the transmitter sample loop except for out-of-band distortion components.

Figure 14 shows the distortion measurements at the transmitter and the tower. Also plotted on the graph are some of the distortion measurements made without the broadbanding network. They were made at 50 percent modulation due to the inability to fully modulate without the network in line. The transmitter VSWR protection circuit would trip the transmitter off when we attempted to modulate 50 percent at 10 kHz. The ski jump at 5 kHz, in the distortion measured at the tower loop, is probably caused by the filtering of the third and higher harmonics by the broadbanding network and the antenna.

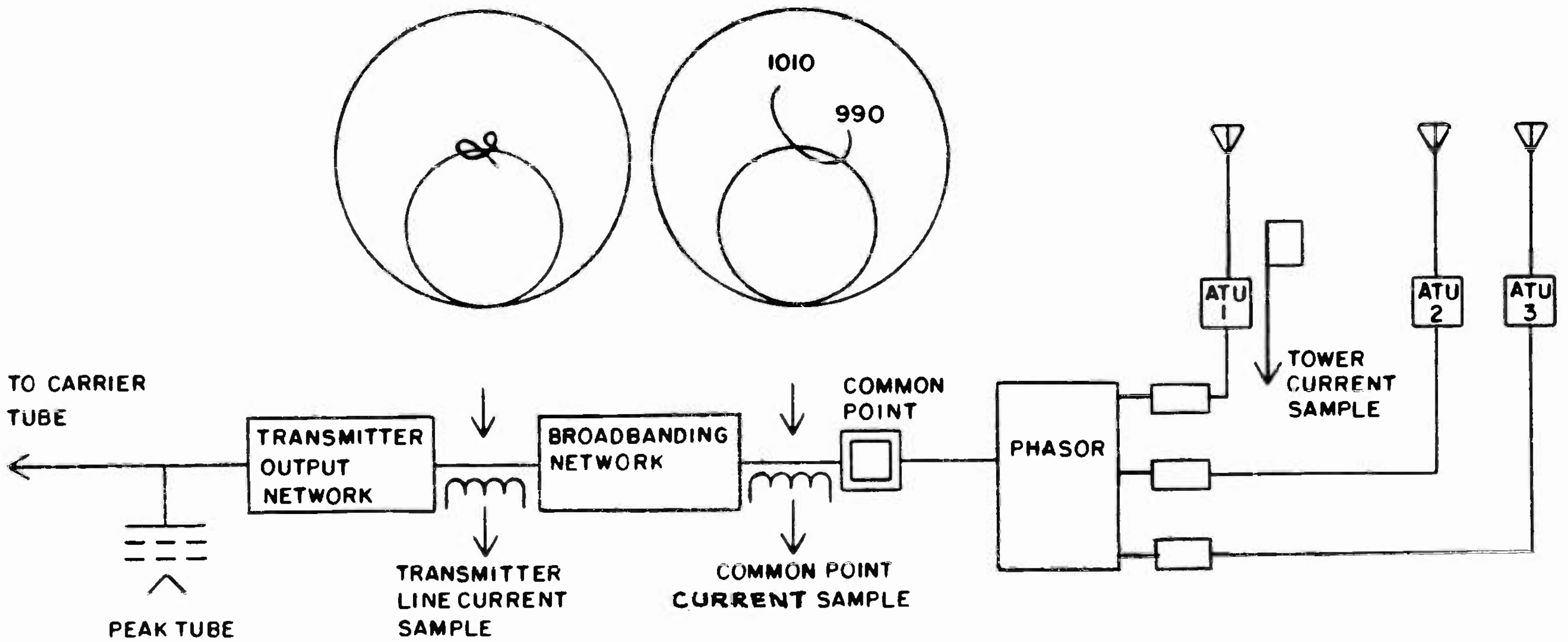
Figure 15 shows the frequency response at both points. The broadbanding network and the antenna show a roll-off of about 0.6 dB at 10 kHz. This may be the result of higher losses at sideband frequencies in the antenna and network, and the small transmitter load impedance variations remaining at the band edges. For comparison, the frequency response without the broadbanding network at 25 percent modulation is also shown.



NAME
TITLE
KOMO 317C-2 PEAK PLATE
DWG. NO. 12
DATE
SMITH CHART FORM 82 SPR (2-49)
KAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1949 PRINTED IN U.S.A.
IMPEDANCE OR ADMITTANCE COORDINATES
990
1010 KHz
RESISTANCE COMPONENT OR CONDUCTANCE COMPONENT
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER



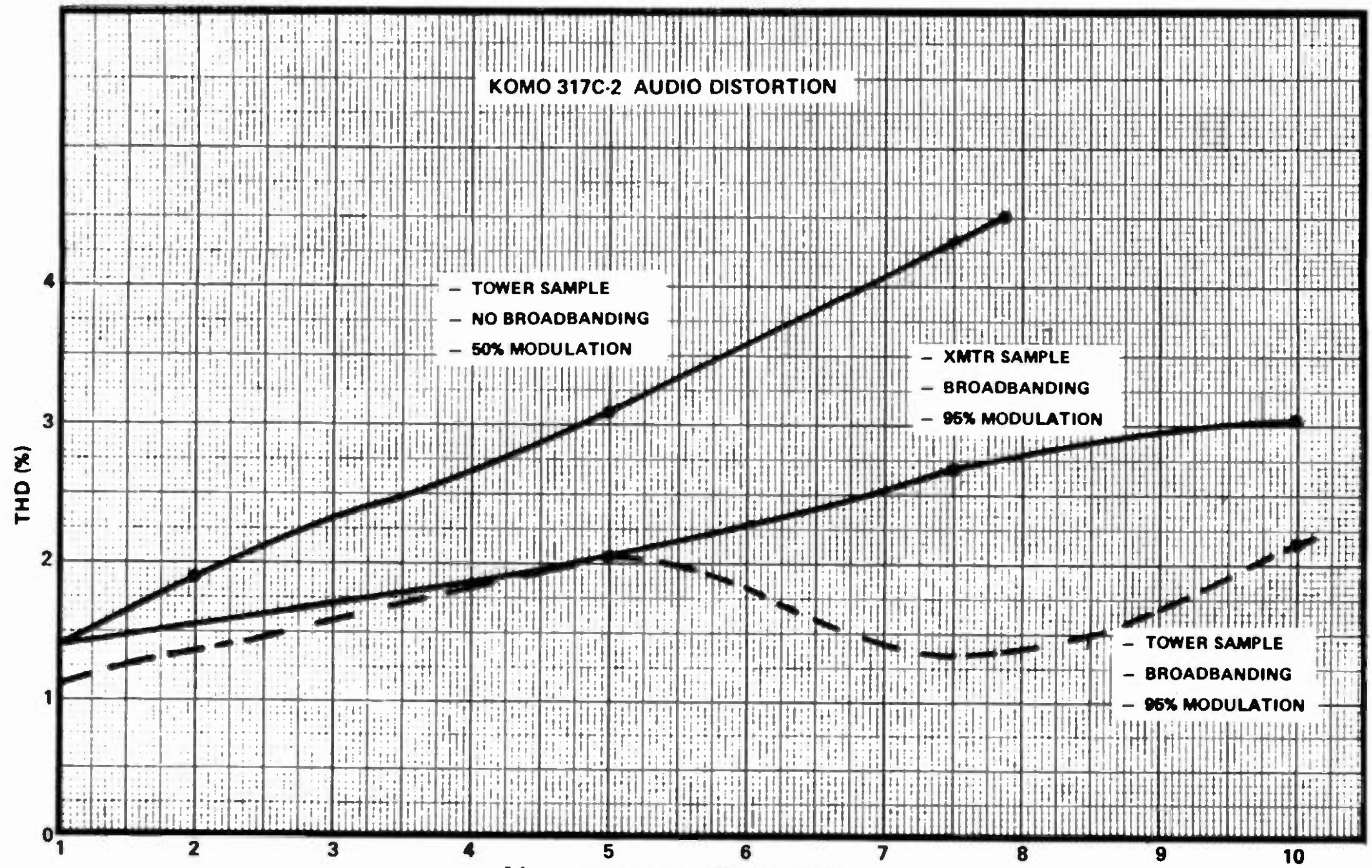
NAME
TITLE
INPUT TO KOMO BROADBAND NETWORK
DWG. NO. 11
DATE
SMITH CHART FORM 82 SPR (2-49)
KAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1949 PRINTED IN U.S.A.
IMPEDANCE OR ADMITTANCE COORDINATES
990
1010 KHz
RESISTANCE COMPONENT OR CONDUCTANCE COMPONENT
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER



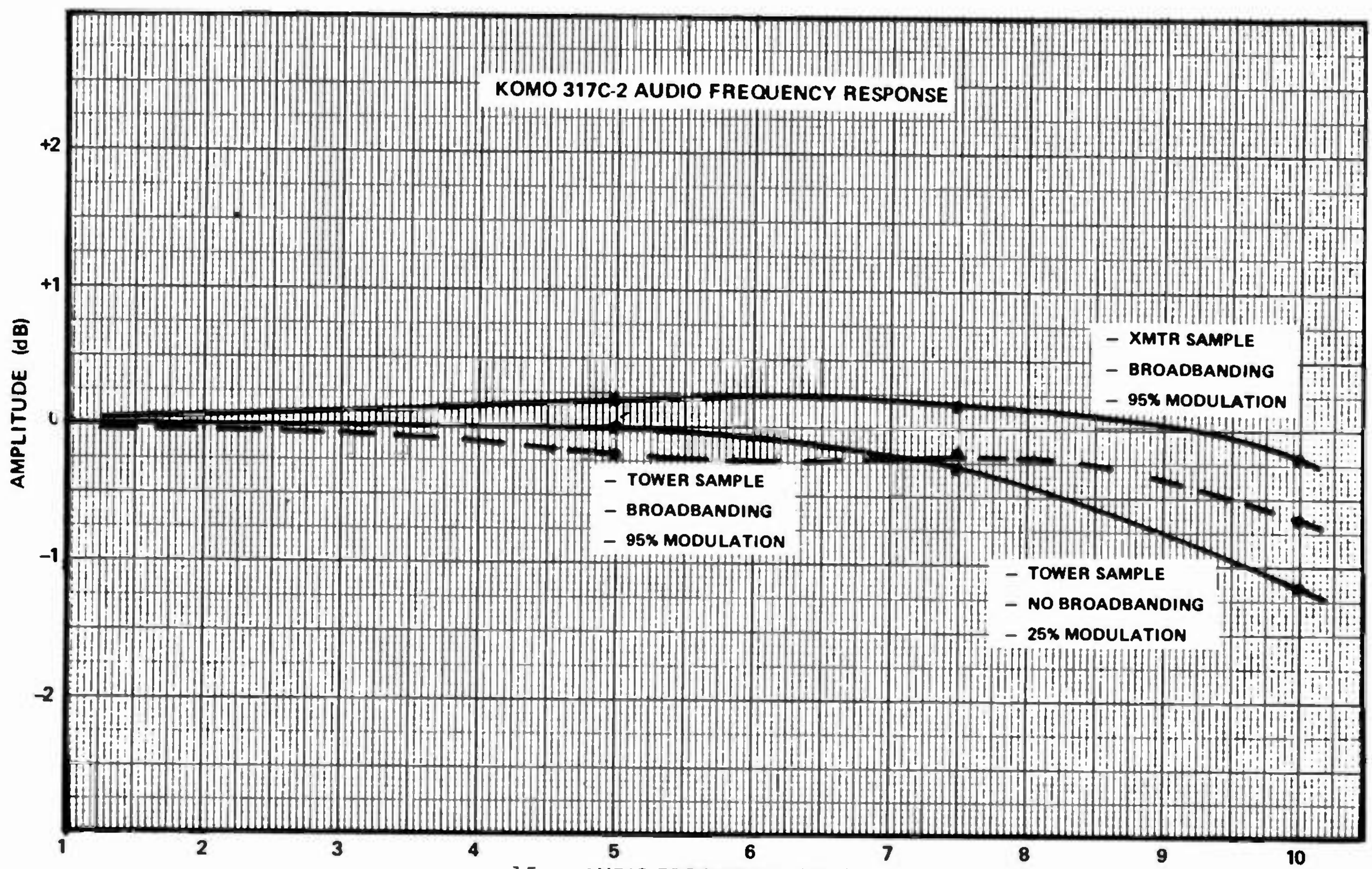
BLOCK DIAGRAM OF RF SYSTEM SHOWING MODULATION SAMPLE POINTS

FIGURE 13

DWN B. ARCHER
DATE 2-21-84



14. AUDIO FREQUENCY (KHz)



15. AUDIO FREQUENCY (KHz)

These distortion and frequency response measurements taken with the broadbanding network were all made at 95 percent modulation, while being compared with results measured at 25 and 50 percent modulation without the broadbanding network. This apples and peaches comparison points up a significant improvement in high frequency modulation depth, frequency response, and distortion.

In summary, four-stage external broadbanding will work with any antenna. Unlike single- or three-stage broadbanding, which operate on the sideband reactance only, four-stage broadbanding can pull down very high sideband VSWRs, since it operates on both resistance and reactance. Of course, you do not get something for nothing. A reduction in VSWR within a certain passband is done at the expense of VSWR outside that passband. But so long as one chooses the passband wide enough, this effect is unimportant.

In some cases it is possible to design a new array to have a broad passband. However, there are situations that force even a new array to have a narrow passband. For example, traps may be required at the tower bases to detune for a nearby existing station. Or the cost of land may force diplexing. Or the pattern may be high gain, or have a high RSS/RMS ratio. Sometimes the towers are required to be electrically short, or too tall, or too top-loaded. All of these factors can degrade bandwidth.

Regarding adjustability and long-term adaptability, minor changes in common-point impedance can be compensated, even when the toroidal coils themselves are not adjustable. This can be accomplished by adjusting the midpoint impedance (instead of 15 ohms, perhaps 13 or 17 ohms would be more suitable), or adjusting the impedance rotation, or adjusting the stub tee. The stub tee has three handles: you can adjust its phase shift, transformation ratio, or characteristic impedance. Thus many degrees of freedom exist to compensate for seasonal changes or normal deterioration in the antenna system.

The authors wish to thank Fisher Broadcasting Inc and Bob Plummer, Director of Engineering, for permission to publish the KOMO performance data, and Bob Holcomb, for collecting same.

Switched Diversity for Mobile FM Broadcast Reception

William R. Rambo
SRI International and CommTech International

ABSTRACT

A simple two-antenna, switched diversity system has been developed to reduce multipath interference in car stereo reception. Conventional car antennas are used and no circuit modifications are required in many current receivers. The use of frequency-selective fading for the primary control of switching greatly speeds the decision process and the audible transients usually associated with switched diversity are minimized. Performance is enhanced by incorporating sequential sampling to select the lesser disturbed antenna when both are simultaneously corrupted. The reception improvement generally accords with predictions based on typical fading statistics and is very substantial in many multipath conditions.

FM STEREO RECEIVER PERFORMANCE in automobiles is often degraded by multipath interference. The classic response to multipath in other services involves the incorporation of some form of multiple-antenna diversity reception. This paper summarizes the results of an exploration of the potentials of a simple two-antenna diversity system for improving car stereo reception. The first section deals with the basic multipath phenomena; the focus is on features important in the mobile reception of a VHF wideband FM transmission. The second section describes an experimental system designed with those features in mind. Design constraints of another type were imposed: The system must use conventional car antennas, must be compatible with current receivers, and must require no user intervention in operation. The final sections include a short discussion, largely qualitative, of performance tests of the experimental system, and conclusions drawn therefrom.

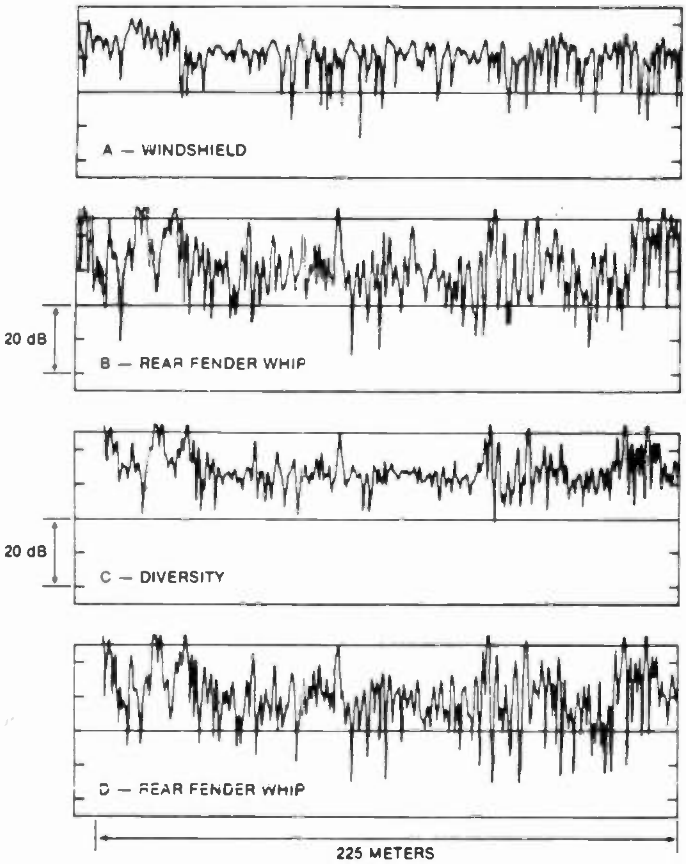
THE MULTIPATH/DIVERSITY CONNECTION

Experiments with diversity reception were reported as early as 1927. The technology has been incorporated in point-to-point communication systems for more than fifty years. More recently, interest has extended to the special problems encountered in mobile communication systems, particularly in systems operating in the 450-MHz band, and above. The intensity of this interest is evident in the literature; one treatise that provides a broad coverage of such systems lists several hundred references (1).* Investigations of diversity techniques for improving mobile reception of wideband VHF FM broadcast transmissions are naturally stimulated, and supported, by such a background (2).

FADE FEATURES -- The composite signal received by a car antenna moving through a multipath area is the sum of a number of components resulting from the reflection and defraction of the transmission, primarily by near-by objects and features -- buildings, trees, wires, hills, other cars, etc. As the car moves, the composite signal envelope displays a fluctuating pattern that can include momentary, deep fades accompanied by multipath noise and distortion in the output program material.

Traces A and B in Fig. 1 illustrate typical fade patterns. They were recorded simultaneously using two calibrated receivers, one fed by a windshield antenna with a substantial response to the horizontal polarization of the original transmission. The second receiver was fed by a vertical quarter-wave whip mounted on a rear fender. While the pattern fluctuations in the recording of

*Numbers in parentheses designate references at end of paper.



PAGE MILL ROAD — 25 km TO KARA, 105.7 MHz, 50 kw ERP

FIGURE 1 — SIGNAL AMPLITUDE PATTERN

the vertical antenna signal are (typically) more violent, the averaged signal strengths differed by less than a dB, suggesting the substantial loss of polarization purity in moderate multipath conditions.

Traces C and D were recorded in a second traverse of the route. Traces B and D correspond directly. Though there were variations in speed and precise route (a fraction of a wavelength displacement can be significant in terms of fade depth) the duplications in recording features are quite evident. This is to be expected since the multipath components arise from a fixed pattern of objects. Despite this pattern stability, the multipath interference impact may not be. The average disturbance level can vary from day-to-day (or hour-to-hour) in areas a sufficient distance from the transmitter to be subject to intervening propagation conditions.

The Fig. 1 recordings display two common pattern features. The dominant one is termed fast fading, this referring to the multipath associated, quasiperiodic fluctuations averaging less than a wavelength between minima. The second is so-called slow fading. It can occur over many tens of wavelengths and is ascribable to variations in the terrain profile and the general nature of the environment (3). The concern here is with the momentary, deep, fast fades.

If signal amplitude distribution data from sections of Fig. 1 A or B where the mean signal is sensibly constant were to be plotted on Fig. 2, the data points would fall along the curve marked M=1 (single antenna). This is indicative of Rayleigh distribution, and confirms the random nature of the multipath phenomenon essential to the diversity reception concept.

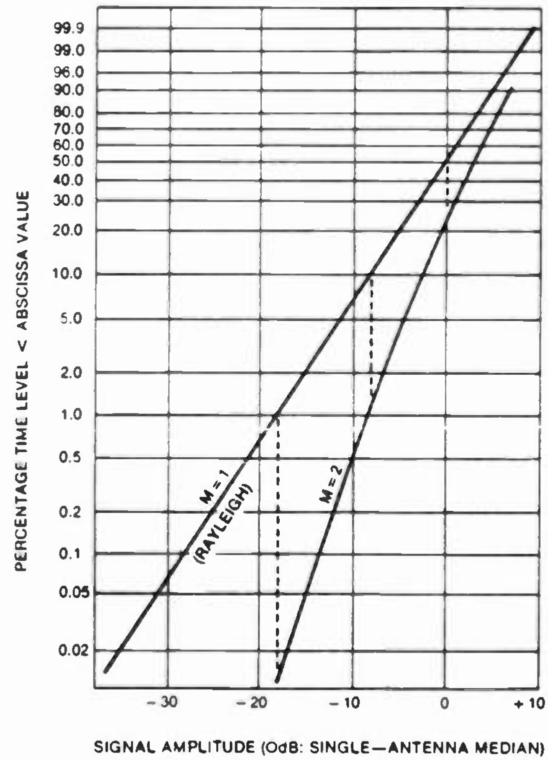
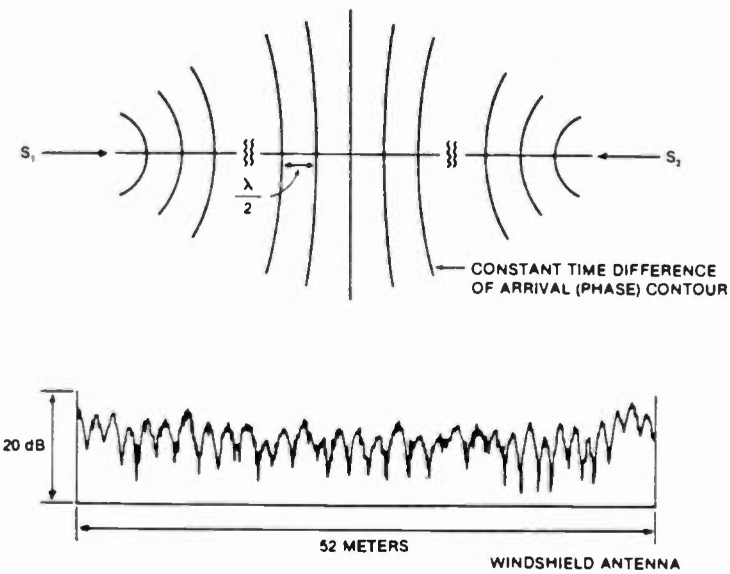


FIGURE 2 — PROBABILITY DISTRIBUTIONS FOR SINGLE ANTENNA (RAYLEIGH) AND TWO-ANTENNA SELECTION DIVERSITY

While the Raleigh distribution is typical it does not always obtain.Figure 3 illustrates a contrary example--one arising when only two rays dominate in the composite signal. The relative phase conditions for a fade are then satisfied along a hyperbolic family of lines of constant time difference of arrival. If the vehicle moves across the pattern, the periodic fluctuations noted in Fig. 3 result. But if by chance the vehicle travels along a fade line, a particularly disturbing long-duration fade can occur. In open country, it is sometimes possible to follow a fade line for much more than 100 meters. Rayleigh fading statistics obviously do not apply to the signal recorded in Fig. 3.



SRI PARKING LOT — 40 km TO KFOG, 104.5 MHz, 7.9 kw ERP

FIGURE 3 — TWO-RAY SIGNAL AMPLITUDE FLUCTUATIONS

THE DIVERSITY CONCEPT -- Diversity reception contemplates access to two (or more) signal samples displaying independent fading statistics. This independence arises in the two-antenna reception of FM broadcast transmissions through differences in antenna response (polarization/angle diversity), and/or by suitably separating the antennas (space diversity).* Space diversity is the usual choice. The wide angular disposition in the incoming multipath rays (and the inevitable path-length differences) leads to sufficiently decorrelated fading with quite practical spacings, down to less than a half-wavelength (< 1.5 meters).

The signal samples can be processed in a variety of ways. The simplest is to select the better one on some logical basis (selection diversity). The result of selecting the stronger signal is illustrated by Trace C, Fig. 1. The recorder was fed by a circuit selecting the stronger of the signals from the two antennas in the second traverse of the route. Horizontal lines have been drawn on Traces A, B and C at an arbitrary amplitude level. The fade penetrations of this level, 26 in the case of windshield antenna and 32 in the case of the vertical, correspond roughly in number to the audible multipath disturbances noted in the program material during the recording period. This level was penetrated only once in the recording of the diversity signal (Trace C). The diversity amplitude distribution data would plot in Fig. 2 along the line M=2, for two-antenna selection diversity.

*An ingenious alternative involves co-located orthogonal loops operated as active antennas(4)

Reference to Fig. 2 indicates an anticipation that a 10% likelihood of a disturbing fade event with a single antenna would be substantially improved to 1% with selection diversity. A 1% likelihood would be reduced to a near-perfect 0.01%. But if signal levels were such as to result in a 50% expectation with a single antenna, the diversity result would be a still-miserable 25%.

These comments pertain to a situation where two receivers provide continuous samples of the transmission. They can be utilized in selection diversity, or processed in more sophisticated, and effective, ways, e.g., equal-gain, or maximal-ratio combining. But the initial conditions imposed on this investigation limited the system to a single receiver. The signals from the two antennas then are only alternately available, thus forcing the use of a somewhat less effective form of diversity, so-called switched diversity. The antennas are selected alternately with the switching decision controlled by some form of logic based, typically, on an observed deterioration in some quality of the signal from the then-connected antenna.

FREQUENCY SELECTIVE FADING -- The speed of the recorder used in developing Fig. 1 was far too slow to display evidence of a multipath fading feature of singular importance in the reception of the wideband FM broadcast transmission. This is frequency selective fading. Because of the path length differences in the multipath rays, the many frequency components comprising the transmission are treated differently in terms of fading.

If the path length differences are large (microseconds), it is possible for some program frequencies to be in a deep fade while others are substantially unperturbed. It explains why in spot reception the unmodulated transmission may appear clear, but then with a disturbance that appears, and increases, with modulation (deviation). Frequency selective fading accompanies the composite signal deep fades pictured in Fig. 1. But an associated disturbance also may be present when the composite signal amplitude is little affected.

The negative features resulting from frequency selective fading are that the sharply defined fade points in Fig. 1 can become fade regions; the transit time through a fade is then extended; the likelihood of simultaneous fades at the two antennas increases. The positive features are the correlation of observed disturbances with frequency selective fading, and the high frequency content, extending to hundreds of kilohertz, in the disturbance. A rudimentary measure of the momentary magnitude of the disturbance itself becomes possible, in consequence the speed with which a switching decision is reached can be increased.

A TWO-ANTENNA SWITCHED-DIVERSITY SYSTEM

The practical design considerations initially imposed, and the multipath interference features herein described, controlled the system design pictured in block diagram form in Fig. 4.

The use of a single receiver dictated a switched-diversity format with switching at RF. The simplicity requirement (plus aesthetic considerations) held the design to two antennas. Fortunately, the nature of mobile multipath argued that two regular car antennas, in normal location, could provide usable space diversity.

The importance of frequency selective fading called for the retention of the full frequency spectrum of fade information in the control signal delivered to the switching decision circuitry. This information is available in the IF section of the receiver as amplitude perturbations (in contrast to the FM program format). Many current receivers use an integrated circuit IF system providing amplitude-sensitive signals for meter and mute control.* In suitable combination, these provide a measure of fade conditions over a very wide dynamic range. The information is usable despite some non-linearities in the control voltage vs signal amplitude relationship. Samples of these signals can be obtained by direct tap-off at the IC pins, thus meeting the imposed requirement of no internal circuit modifications in the receiver.

Two forms of non-multipath interference can corrupt the control information. One arises from certain combinations of the signal with ignition noise, the second from the switching transient that accompanies a sudden change, with switching, in signal amplitude and/or phase. In either case, the result is a pulse related to the impulse response of the receiver -- fortunately only a few microseconds in duration because of the wide broadcast channel bandwidth. While the interference reduction properties of the receiver discriminate against such interference in the output program material, these pulses in the control signal can appear as momentary fades, thereby potentially affecting the switching decision process quite apart from the true multipath conditions.

*Examples are the LM3189 and LA1140.

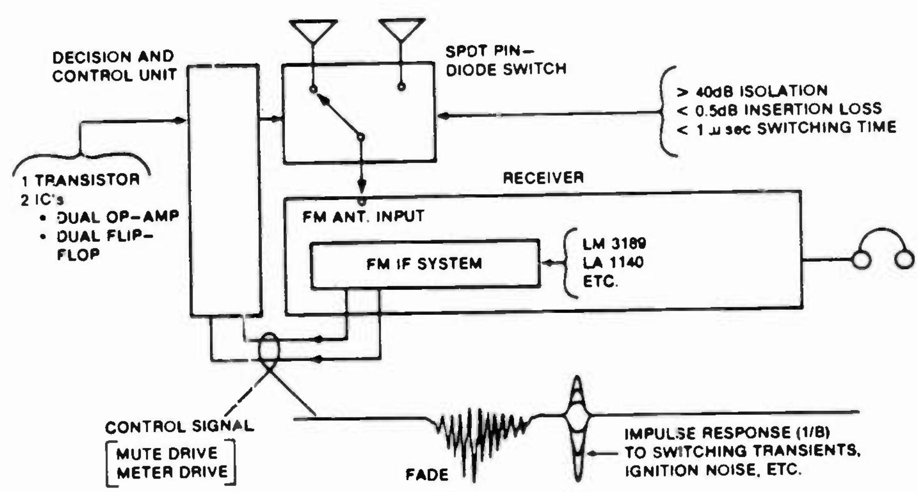


FIGURE 4 — SYSTEM BLOCK DIAGRAM

THE DECISION AND CONTROL UNIT -- Several design aspects require comment.

Initial switching occurs when the interference level evident in the control signal exceeds a pre-selected threshold. By retaining a full response to the interference information in the control signal, a decision to switch, and switching, can be accomplished in a few microseconds. This substantially eliminates the audible disturbance associated with switched diversity in narrow-band systems due to switching delay and switching transients.

For the reasons earlier discussed, the likelihood of simultaneous multipath interference in the signals from both antennas can be high. The system response is to measure the level of the disturbance initiating switching, compare it with the disturbance level in the newly connected antenna, and select the better of the two. The sampling is repeated periodically until one antenna clears. The sampling interval is short, perhaps 20 microseconds. There results a sort of hybrid selection diversity based on disturbance level. The improvement in performance is very noticeable in moderate multipath conditions.

In the absence of a multipath dictate, the system selection favors the stronger of the two signals. There is a minor danger. The weaker signal may remain clear while some level of multipath develops in the stronger. Until that level exceeds the difference in composite signal amplitudes, the selection process will continue to favor the stronger. Fortunately, this is not a usual occurrence, and is offset by the better utilization of the two-antenna availability in weak signal reception.

The final requirement, discrimination against impulse noise and switching transients in the control signal, arises from the first -- the inclusion in the switching decision process of the full fade frequency spectrum. The requirement is substantially met by a blanking gate generated at the onset of an isolated pulse (ignition noise), and by any antenna switching action. The gate duration, 10-12 microseconds, relates to the impulse response characteristics of the receiver.

CIRCUIT FEATURES -- The several system requirements are treated in circuit design as sketched in Fig. 5.

A series/shunt, pin-diode switch is driven by a Type D flip-flop (FF2 in Fig. 5). Each clocking produces a change of state that reverses current flow through the switch so as to connect the alternate antenna.

Switching is controlled by the receiver's meter and mute control signals. These are combined and amplified in the first op-amp (OA1). Gains are set to weigh properly the relative contributions of the two fade information sources. A fade appears in the output circuit as a positive-going signal superimposed on a nominal d-c level (V_N) that holds in quiescent intervals.

The second op-amp (OA2) acts in a switching cycle first as a comparator, then as a peak detector. The control signal is applied to the (+) input. A quiescent threshold level, V_T, which appears across C_1, is applied to the (-) input. The V_T-V_N difference sets the switching sensitivity, e.g., to trigger on a sharp 3-dB multipath fade. With V_T normally more positive than V_N, the op-amp output is normally LO.* A fade of sufficient amplitude to overcome the V_T-V_N difference causes the output of OA2, first operating as a comparator, to rise. There are two results. The collector of T_1 is brought down, arming the system for switching. The loop is then closed through D_1 and a charging current is delivered to C_1 by OA2 operating now as a peak detector. The C_1 (threshold) voltage is thus reset quickly (microseconds) to the peak value of the disturbance. Only then does the op-amp output voltage fall, the transistor collector rises, and switching occurs. (Both flip-flops are clocked by a waveform positive-going though V+/2.)

If the second antenna were concurrently experiencing a still larger disturbance, the resulting control signal amplitude would exceed the newly reset threshold. The cycle would repeat; the threshold would be reset to an even higher level. The lesser disturbed, original antenna would be reconnected, its disturbance level now being tolerated by the reset threshold. A reset threshold level decays toward V_T in accordance with the C_1-R_1 time constant, perhaps 0.1 seconds. If the antenna had not cleared, the disturbance level would be intercepted, the signals again sampled, and a new selection made.

The sampling rate is a function of the difference in disturbance levels. A sampling event averages perhaps 20 microseconds. It can be that long because of the action of the first flip-flop (FF1), clocked simultaneously with FF2. It produces a (typically) 10-microsecond blanking gate at $\overline{Q}$. This disables comparator action; the (+) input of OA2 is clamped LO by D_2 and the op-amp output voltage goes LO. Any switching transient dissipates during this gate interval. Only when the gate terminates are control data now only from the newly connected antenna, again presented to the comparator.

This flip-flop (FF1) can also be triggered at the SET (S) input by a positive-going waveform passing through V+/2. This triggering will happen at the onset of comparator action following a quiescent period, produced by the initial rise in the comparator output voltage. Circuit speeds are such that the comparator is disabled by the $\overline{Q}$ blanking gate before the

*The HI/LO labelings on Fig. 5 apply to the quiescent state.

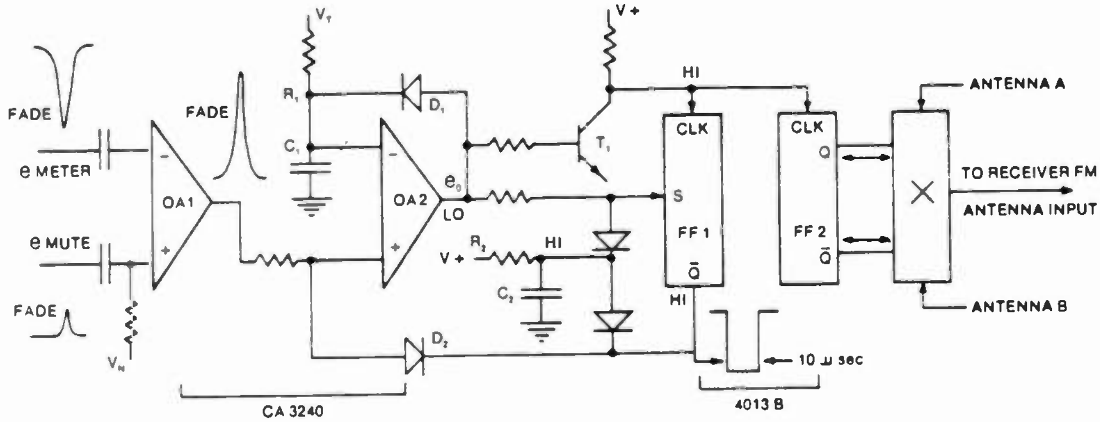


FIGURE 5 — CIRCUIT FEATURES

threshold reset, and antenna switching actions can be initiated. Thus an ignition noise impulse dissipates during the gate interval and does not cause switching.

The first ten microseconds of a true multipath fade are also lost by this gating action. But a feature of the disturbance reflected in frequency selective fading is that, if significant, the fade will extend beyond the gate interval, or will recur in a few microseconds. After the gate terminates, SET input triggering is prohibited for perhaps 200 microseconds, a condition brought about by the discharge of C_2 during the gate interval. The SET input is clamped, and SET triggering is prevented pending the recharge of C_2. The normal threshold reset, and antenna switching actions as earlier described, then can take place without delay.

While this operation is long in the telling, the circuit formulation accords with the original circuit simplicity goal. The principal active elements comprise a transistor, a dual op-amp, and a dual Type D flip-flop.

TEST RESULTS

Performance was evaluated only through qualitative listening tests. Because the multipath interference is both car position and program dependent, the practice was to record simultaneously the outputs of two identical receivers, one equipped with the diversity feature, and with one antenna shared.

In moderate multipath conditions, perhaps 100 audible events per 1000 meters of travel, typical in the local suburban area, the reduction in events was in the order of twenty to one -- as the fading statistics (Fig. 2) would suggest.

Also as the statistics suggest, performance deteriorated with increased multipath interference. The value of the selection of the lesser disturbed antenna was then evident. Replays through intervals of simultaneous interference typically showed a significant reduction in disturbance intensity, and in the "cleanliness" of the program reflected in both noise and distortion.

Some improvement was evident in heavy multipath conditions but the service still would be considered unsatisfactory by discriminating listeners. Diversity, of course, is of no practical assistance in fully shadowed areas. Five antenna pairs were used in the experiments. In all cases, spacing was at least 0.4 wavelengths. No pair was strongly superior. The windshield-rear fender whip combination was possibly the best, but any advantage over the windshield-rear window pair would probably fade in view of the aesthetic advantage of having no visible antennas.*

Some degree of polarization and angle diversity is inevitable in spaced car antennas as typically mounted. There is no advantage in seeking identical patterns, complementarity would be an advantage. LED indicators provided visible evidence of antenna usage and switching. It was commonplace to find one antenna clearly favored in driving in one direction along a street, and the other favored on the return trip. The advantage of having the two antennas was also evident in distant, weak signal reception.

*The test car was a Pontiac sedan with a standard GM windshield antenna.

CONCLUSIONS

Diversity reception can signficantly increase the practical service area of a broadcast transmission in multipath regions. It is a legitimate candidate for inclusion in quality systems.

In view of the limits imposed by Nature on the diversity improvement, the system cost must be commensurate with the gain in performance. The simple switched diversity approach accords with that requirement.

The LED indicators are worth incorporating in any installation. Diversity, like taking an aspirin, offers relief, not a cure. When things are going well, one wonders if the medicine is necessary. It is then satisfying to see the system so hard at work.

REFERENCES

(1) W. C. Jakes, Jr.
Microwave Mobile Communications
John Wiley & Sons, 1974 by Bell Telephone Labs., Inc.

(2) Robert A. Finger and Emil Torick
FM Receivers in Automobiles: A Case for Diversity Reception
IEEE Transactions on Consumer Electronics, Vol. CE-27
pp. 51-62, February 1981

(3) J.D. Parsons, Miguel Henze, P.A. Ratliff, and Michael J. Withers
Diversity Techniques for Mobile Radio Reception
IEEE Transactions on Vehicular Technology, Vol. VT-25, No. 3 pp. 75-85, August 1976

(4) S. Takeda, J. Shibata, M. Shimakata, and K. Aizawa
FM Multipath Distortion in Automobiles Has Been Significantly Reduced by a New Antenna System
IEEE Transactions on Consumer Electronics, Vol. CE-26
pp. 263-269, August 1980

Computer Optimized Directional Antenna Patterns Improve AM Coverage

Glen Clark
Glen Clark & Associates
Pittsburgh, PA

and

Edward Schober, P.E.
Radiotechniques
Haddon Heights, New Jersey

Optimization is the science of getting as much as you can, and paying as little for it as possible. There are a lot of things that we would all like to get as much of as we can, but one thing that broadcasters would universally like more of is coverage. To state the obvious, the larger and more strategically shaped the coverage area, the more possible listeners a station will have. More listeners beget higher ratings which in turn beget a higher revenue producing potential.

An excellent signal is not a substitute for proper programming; but, if all other factors are equal, the station with the better service area will achieve higher ratings.

We propose to show that advances in computer techniques make directional antenna designs now possible which contain a larger population within their service contours, sound better on the air and at the same time frequently cost less to build.

We anticipate that there are a number of non-technical people in the audience, and while we are describing a technical procedure, we will, at least initially, avoid technical terms when describing how and why the process will benefit the broadcaster.

As a brief example, we recently redesigned a proposed antenna system for a New England client. The new, optimized design increased the served daytime population by 43%, increased the served nighttime population by 23%, and required one less tower to build. The revised design made it possible to choose a location where real estate prices were significantly lower. The total acreage used was also less. Total savings to the client on the project: greater than $150,000.

If this sounds like a "something for nothing" situation and that there has to be a catch, there isn't one. It is simply a reflection of the value of information.

We are all familiar with trying to get "more". In a business negotiation, in order for you to get "more", someone else has to settle for "less". Business negotiations are usually what are known as "zero sum" situations because the sum of the gains of all parties involved is zero. You go home with an additional hundred dollars in your pocket only because someone else went home with a hundred dollars less.

Optimizations generally are not zero sum operations but fall into what are called "open ended" problems. In these situations, you get to keep more, not because someone else gave it up, but because you prevented it from being wasted. Frequently, open ended optimizations deal with information or specialized knowledge.

An early and simple example of optimizing an open ended situation is the farmers' practice of rotating their crops. A farmer has two large fields and raises a different type of crop in each. He finds that if in the following year he plants in field number one the type of crop that was in field two, and plants in field number two the type of crop that used to be in field number one, his harvest will be greater than if he had raised the crops in the same fields as before.

A simple understanding of how to let soil nutrients recover has allowed him to <u>get</u> more without <u>paying</u> more. He has "optimized" his planting. Notice also that it was not necessary for someone else to get less in order for him to get more.

It is also possible to optimize an open-ended situation to get the same reward as before, but to pay less for it. Most of us will recall the gas guzzlers that we drove prior to the Arab Oil Embargo. The reward then was that you moved from point to point at fifty-five miles per hour. The payment was one-twelfth of a gallon per mile. Today we still travel at fifty-five miles per hour but pay only one-thirtieth of a gallon per mile.

Several years of hundreds of frantic Detroit engineers toiling over computer models and in wind tunnels has netted us the knowhow to make combustion more efficient and reduce wind drag.

Optimization produces benefits for the user because it allows him to be more efficient, to make better use of the resources at hand.

For the station proposing a new directional antenna, an optimized pattern will usually:

1) serve a larger population
2) require fewer towers to build
3) require less real estate to construct, and occasionally
4) require less expensive real estate to construct.

A quarter-wave tower in the center of the AM broadcast band (1070 kHz) will be 230' tall. The ground system will include 3.8 acres. Assuming a modest price for the tower of $15,000 installed, and a real estate cost of only $2000 per acre, the savings by reducing the number of required towers by one, is greater than $22,600. (Economies in the phasor are also realized). Higher real estate values will produce even greater savings. It quickly becomes apparent that optimization is not simply an engineering nicety. It is a technique whose advantages can be immediately appreciated even by those who haven't the slightest interest in technical matters.

For those situations where it is possible to reduce the required number of towers, the cost of the optimization is completely outweighed by the reduced capital construction costs.

Even for the broadcaster with an existing directional antenna, and who has already erected more towers than he might need today, there is considerable possible benefit. He or she may be able to increase power with the present tower configuration, or with only minor changes in tower geometry. He or she may also be able to increase the population served while staying at the present power by letting out the nulls in the existing pattern. A computer optimization may simultaneously permit increased power of the station's Presunrise Authorization (PRSA), Post-Sunset Authorization (PSSA) or Critical Hours (CH) operation.

COMPARING THE PRESENT WITH THE PAST

A logical question for the operator of an existing directional antenna to ask at this point is "How do I know my present antenna isn't already optimized?" The answer is, if it has been on the air more than a very few years, it is very unlikely that it has been significantly optimized. For those who will be purchasing the NAB's bound edition of all of the convention's technical papers, there is a sixty second test in Appendix A of this paper, which will allow you to determine if your present pattern has been optimized.

In the 1930's, 40's and 50's, when many arrays were installed, computers were not even available. During the 60's and 70's some engineering offices availed themselves of the benefits of a computer. Most engineering programs were initially batch processed, and then later handled on an on-line or timesharing basis. Computer time was very expensive and charges were proportional to use. Unfortunately, optimizations consume large amounts of computer time. Even though the hardware was becoming available, one could run up a sizeable bill in very short order. It was still not economically advantageous to the client for the engineer to do more than a cursory optimization. Today, most major engineering offices have one or more computers in house, where costs can be more closely controlled.

The tendancy of most population centers to spread out over the years makes this greater availability of computational power a welcome addition. Where 20 years ago the population was neatly concentrated within or near the city limits, the proliferation of cars, Interstate Highways, beltways and "flight to the suburbs" have increased the number of directions in which a station must provide service.

More than a few general managers and owners have lamented that desirable suburbs are now located in areas that once were farmland or unpopulated. As it would have added considerable engineering cost to the antenna design, and there was no tangible benefit to be obtain from it at the time, no particular effort was made in the original design to serve these areas. Today, it would be very helpful if the coverage area could be increased to include them.

What was an acceptable directional pattern a number of years ago may leave the station at a serious disadvantage today.

HOW PATTERNS ARE DESIGNED

Understanding the differences between an optimized pattern and a non-optimum pattern requires a basic grasp of how patterns are designed. The following is not intended to be a rigorous description of the allocation process and some liberties have been taken in the interest of simplicity.

Directional antennas exist for two reasons, to put signal where you want it and to keep it from going where the FCC won't permit it. The Commission's restrictions usually serve the purpose of protectiong other station(s) on the same or nearby frequencies from interference. You will want to put signal in the most densly populated areas and are also required by law to provide a strong signal over your city of license. Your final pattern is a compromise of your goals in concert with the Commission's requirements, and physical and economic constraints.

The protections required by the Commission are called "limitations", "limits" or "maximums". The levels of signal that you place by choice or City of License requirement are called "desireds" or "minimums". Both of these are graphed on a common Engineering tool called the polar plot. The different directions from the center of the graph correspond to bearings on the compass. Straight up corresponds to due north; down on the graph corresponds to south. The further a point is from the center of the graph, the stronger the signal is in that direction.

Two typical sets of limits are shown here. Figure 1 corresponds to a daytime design problem. Figure 2 represents the nighttime limits encountered on a regional (Class III) channel.

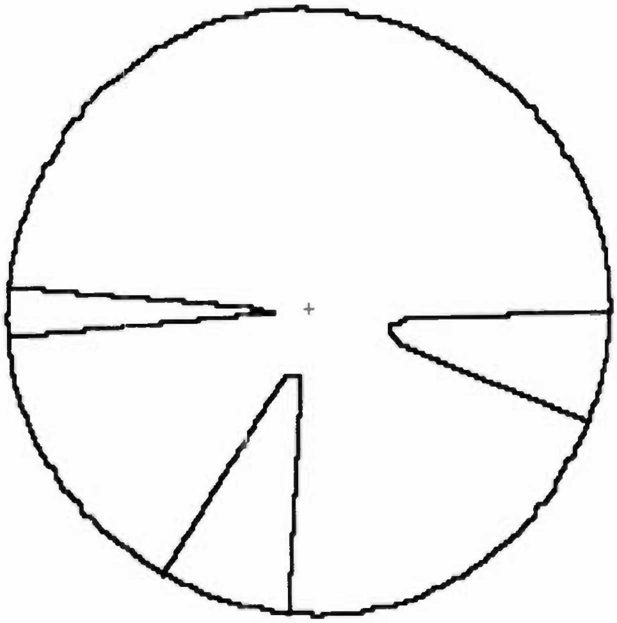


FIGURE 1 - A TYPICAL SET OF AM DAYTIME CONSTRAINTS

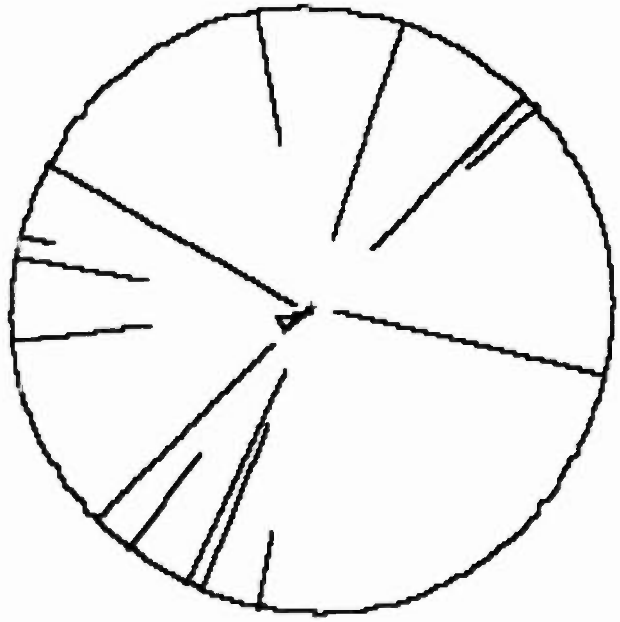


FIGURE 2 - A TYPICAL SET OF AM NIGHTTIME CONSTRAINTS

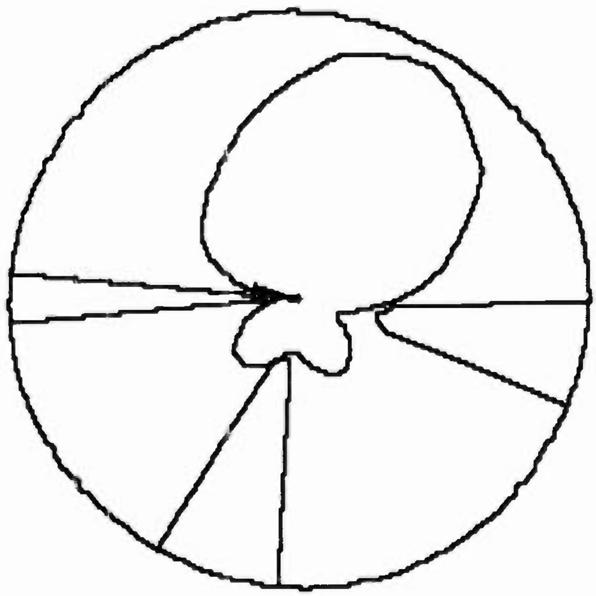


FIGURE 3 - A POSSIBLE SOLUTION TO THE DAYTIME CONSTRAINTS

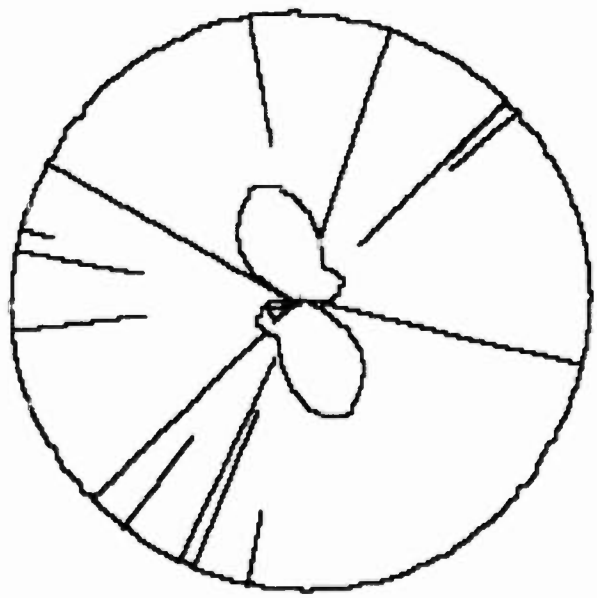


FIGURE 4 - A POSSIBLE SOLUTION TO THE NIGHTTIME CONSTRAINTS

Each limit has three characteristics: central bearing, maximum signal strength and angular width of the limit. Nighttime limitations also include a vertical angle above the horizon which is omitted here in the interest of simplicity. Notice that the depth of the limits vary from station to station depending on the distance between the stations, frequency difference between the stations, required protection levels and other factors.

A nighttime allocation situation on a Regional channel may have thirty or more limitations. Each limit will span at most a few degrees.

A daytime allocation situation will usually have only a few protections and each will span a broader arc.

The allocations engineer's job is to design a directional antenna pattern which meets these requirements. One possible solution to the daytime limits above is shown in Figure 3. A possible solution to the nighttime problem is shown in Figure 4.

Where the daytime pattern bulges to the top of the graph, it represents an increased signal to the north. This is known as a "major lobe". Where the daytime pattern is tucked in to the bottom of the graph, it represents reduced signal to the south. This is known as a "null".

The plot of a successful solution to any set of limitations and desireds is one which does not cross any of the radial lines (limitations and desireds). The solution will tuck inward to avoid a limitation and bulge outward to avoid a desired. Both these patterns meet that criteria.

In addition to the parameters on the polar plot, the allocations engineer will factor other information into his design, such as size and shape of available real estate, maximum economically feasible number of towers, FAA limitations of tower height, and location of the city of license with reference to the towers. Recently, it has also become commonplace to factor the geographic boundaries of the Arbitron rating areas into the design.

One consultant recently compared the perfect optimization to "blowing up a balloon in a Coke bottle". The balloon in all directions would expand to touch the inside of the bottle. It would be forced to assume the distinctive shape of the bottle. Similarly, the perfectly optimized pattern would in all directions touch the inside of all limitations, but in no direction would it exceed them. The pattern would assume the shape of the limitations.

Before the advent of the computer, engineers and broadcasters alike were grateful to find <u>any</u> solution to a given directional antenn pattern. So long as it <u>met</u> all limitations and provided good service to the city of license, the design process usually stopped there.

The result is that many limitations are overprotected. While conditions may require limiting the signal in a particular direction to 50 millivolts, the design may actually reduce the field to 10 millovolts. Other times, a spurious null occurs. While there is no limitation in that direction, a byproduct of meeting the rest of the requirements of the pattern is the unwanted null.

Before the easy accessability of computers and availability of optimal design methods, arrays were synthesized using a technique known as "pair multiplication". In the absence of substantial computing power, pair multiplication was the only way to get a handle on the staggering mathematics necessary to synthesize a large, multi-tower array.

While two-tower arrays could be designed almost mentally, how to meet a certain set of constraints with a three-tower array was less clear due to the much larger number of possible combinations. The addition of more towers further removed the optimal solution from view of even the most intuitive engineer. The pair-mulitplication procedure made larger arrays possible because it decreased the number of design choices for the engineer to a more manageable subset of all possible.

The disadvantage of pair-multiplication is that it leads to nulls which come in twos. A null on one side of an axis of the towers will have an identical twin on the other side of that axis. In-line arrays have one axis of symmetry. Pair-multiplication arrays in a parallelogram configuration have two axes of symmetry.

The pair multiplication array is efficient only where the limits are symmetrical. The difficulty is that the North American continent is not symmetrical. The United States are not symmetrical and neither the distribution of people or radio stations within the States is symmetrical. Consequently, the radiation limitations which a station will encounter will very seldom be symmetrical.

The result of trying to satisfy non-symmetrical limitations with antennas that can produce only symmetrical patterns is inevitably a loose fit: overprotected limitation(s) and spurious null(s).

Today there are two compelling reasons why the design process cannot stop while a pattern still contains overprotected limitations or spurious nulls.

As mentioned earlier, urban sprawl has placed valuable pockets of population in areas which formerly were lightly populated. A spurious null in that direction may have gone unnoticed years ago. Today it can be a significant handicap. Regardless of how innovative the promotion or how refined the music rotation, the simplest truth remains: if they can't hear you, they can't write your name in the diary.

Secondly, most FM stations are non-directional and have no nulls which restrict their service to any part of the listening area. If an AM station has deep nulls toward populated areas, it starts out with fewer possible listners than the FM station. It is in the AM broadcaster's interest to broadcast with as null-free a pattern as possible.

Perhaps an obvious question to ask is "Why don't they just let the nulls out which are too deep, and eliminate the spurious ones completely?" The not-so-obvious answer is that the shape of a directional pattern is controlled only indirectly. There is no adjustment knob which controls the signal to the north and another one which controls the signal to the east. The only controls available adjust the currents and phases in each of the towers. Increasing the current in a particular tower may let out one spurious null, but it will also make changes in the pattern in other directions, perhaps creating another null.

The goal then is to devise a comprehensive process which will monitor the pattern shape in all directions of interest and will increase the signal strength in those directions where it can be improved without causing it to violate an existing desired condition in any other direction.

Having thus adequately described the disease, and having given the treatment the name optimization, let us now look at two typical cases which have shown remarkable improvement after having been administered the treatment.

TWO BEFORE AND AFTER COMPARISONS

Two design problems and their original solutions are shown in Figures 5 and 7. The revised or optimized solutions are superimposed over the original limits in Figures 6 and 8.

Figures 5 and 6 represent the New England optimization described at the beginning of the paper. A four tower system was originally proposed which produced the pattern in Figure 5. The optimized three tower pattern is shown in Figure 6. Notice the main lobe of the optimized pattern is considerably broader and that it provides better service to the southwest.

Figures 7 and 8 represent a station operating nighttime hours and using a four tower parallelogram. It has been in operation since the late 1940's. This original pattern had significantly less field over the northwest quadrant than is allowable. Additionally, it had a sharp, deep null on a bearing of about 193° True. The limitation on this azimuth has a span of several degrees. The least expensive way to produce a broad null using pair multiplication is to make the null deeper than really required. This allows the null to spread out to the desired width by the time it reaches the maximum permissible signal strength.

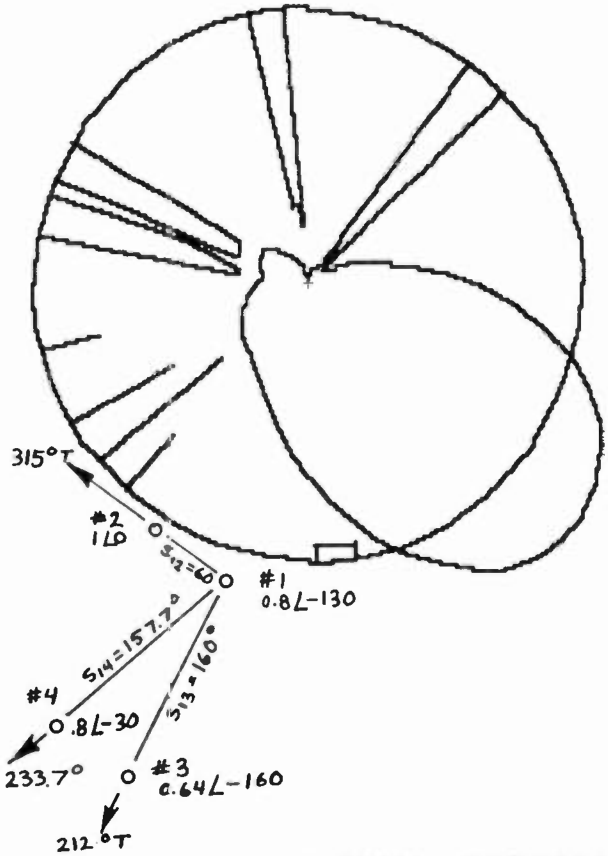


FIGURE 5 - THE ORIGINAL PROPOSED NEW ENGLAND NIGHTTIME

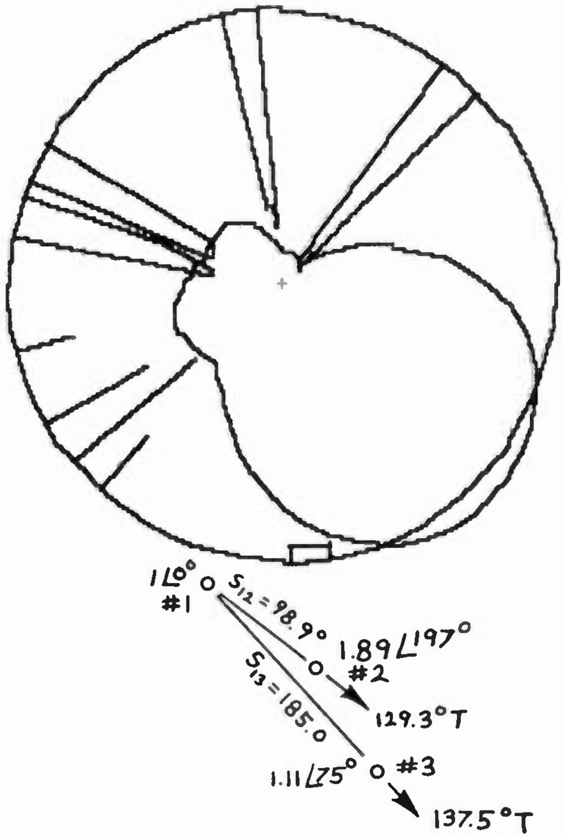


FIGURE 6 - THE OPTIMIZED NEW ENGLAND NIGHTTIME

SCALE: 1"=250 MV/M

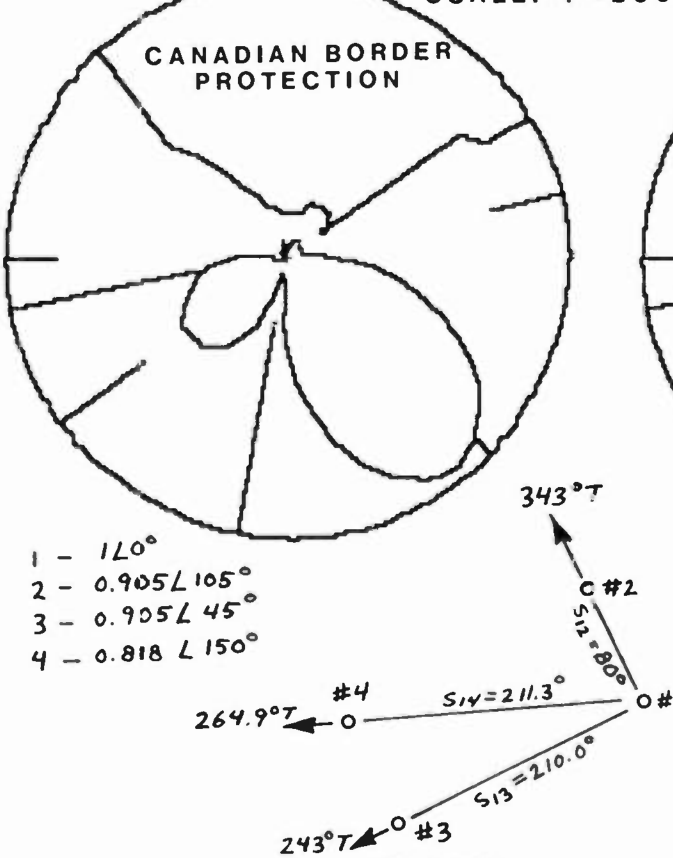


FIGURE 7 - THE ORIGINAL MIDWEST NIGHTTIME

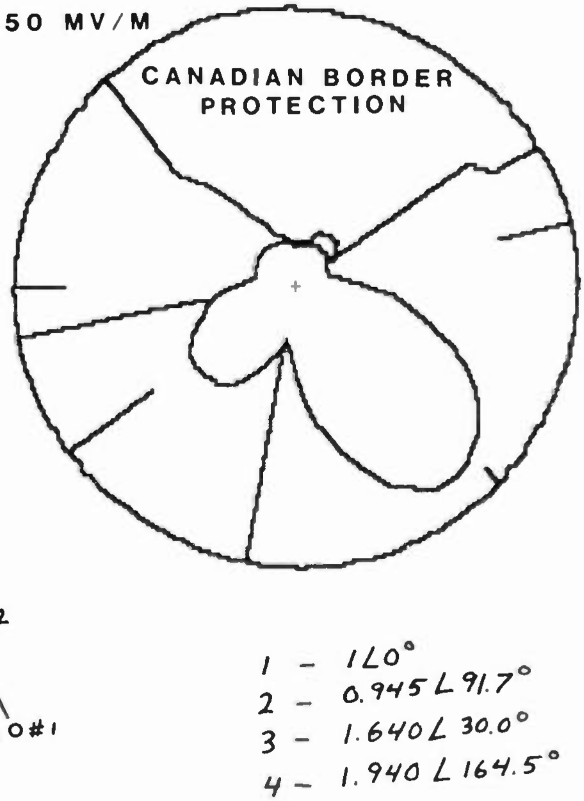


FIGURE 8 - THE OPTIMIZED MIDWEST NIGHTTIME

While the allowable field was 118 mV/m, the original design had a field in the center of the null of only 23 mV/m.

Given the client's restriction that no new construction was possible, a power increase was not feasible. However, substantial benefit was available at the same power level by merely changing some currents and phases. The redesigned pattern on the right achieves the same breadth of the 193° null at the critical field value, but the minimum field in the null has increased to 99 mV/m. Similarly the spurious null at 295°True in the original pattern has been let out to 79 mV/m. Where the highest field value in the original pattern over the arc from 285°True to 65°True was 38 mV/m, now the lowest field value encountered on the new pattern over that same arc is 67 mV/m.

Some of the benefits of filling in nulls where possible are not immediately evident. The higher the signal strength in a null, the less difficult it is to keep it in proper adjustment, reducing operational maintenance expenses. All other factors being equal, a sixty millivolt null is significantly easier to mainatin than a fifteen millivolt null.

Increasing the field in the nulls will also reduce the phasing effects which produce geographically localized areas of audio distortion. In addition to reducing or removing the distortion, minimizing the phasing effects will also improve the performance of AM stereo.

Under some circumstances, filling in the nulls while staying at the same power level can yield more real benefit to the station than a power increase. Power increases usually put more signal where you already have plenty. It is better to add a thousand real listeners in your own backyard than to add another thousand theoretical listeners three counties away.

You may question whether it is possible to get something for nothing. If the licensed power is not increased, won't the power added in the nulls, like a skin graft, have to come from some other part of the pattern? Some optimizations (usually of RMS-limited arrays) actually result in increased field in all directions. Even where the signal to fill in the nulls is moved from somewhere else in the pattern, where the signal has been reduced, the change is so small that it is undetectable. The law of marginal utility says that if you give a starving man a sandwich, you have given him something of considerable value. Give the same sandwich to someone who already has a lunch wagon and he will get little added benefit from it. Similarly, if you give forty millivolts to someone who has none, you have added a potential listner. Take forty millivolts away from a listner who already has six-hundred and it will be impossible for him to discern the loss.

TIMELY INCENTIVES TO OPTIMIZE

There are a number of reasons why a broadcaster might find it desirable and timely to optimize his directional pattern.

First, the delivery system for AM radio now allows it to be more competitive with FM than it has ever been before. AM stereo has only recently been approved. Better AM transmitter designs, particularly those not employing modulation transformers[1], deliver better quality audio than ever before. Receiver manufacturers have also finally begun taking the AM listener seriously, and are making receivers with some semblance of high frequency response.

Stations intending to broadband their directional antenna system, perhaps in conjunction with a move to AM stereo, should consider the related impact of optimization. If broadbanding and optimization are both to be done, the optimization will invariably change the broadbanding parameters. Broadbanding may also be easier after optimization since pattern solutions resulting in bad bandwidth designs are disacarded. If both improvements are planned, the optimization must be done first.

Optimization may make it possible to move the transmitter site to an area with substantially lower real estate prices. We know of one Canadian station which purchased a new site at a low price, built a new transmitter facility on it, and sold the old site for a cash profit in the low six figures.

Lastly, simple capital gains is a reason one would want to consider using an optimized pattern to obtain a power increase of an existing station. The most tangible way to increase the potential value of a station is to increase its service area.

RECENT RULE CHANGES MAKE IT EASIER

It frequently seems that life is perverse. Often, just about the time something becomes possible or desirable, new regulations make it more difficult. The exact opposite is true here. An impressive number of changes in Commission regulations makes it much easier to improve your coverage than it was previously. Keep in mind that the Commission's goal for all broadcasters is to "serve the public". So long as you protect other stations from interference, the FCC is happy to see you serve as many listeners as possible, consistent with efficient use of the spectrum.

One thing which you have been or will be hearing about is the signing of the new Canadian-American AM agreement last January. The protection which each country has been required to provide the other on Class I-A channels has been greatly reduced. This will considerably lessen the cost of arrays necessary for American stations to operate nighttime on Canadian Class I-A channels, and vice versa. Many new fulltime operations which were not possible under the old treaty now will be. While many of the changes approved are beyond the scope of this paper, it is safe to say that they will have a major effect upon available spectrum space.

Operational expenses for directional antenna systems have also been reduced. Years ago a broadcaster was required to have a First Class Licensee in charge of the transmitting apparatus whenever it was in a directional mode. Now, no special license is required for regular directional station operators. For those stations with approved sampling systems, remote control authorizations have been made simpler and the periodic reading requirements are reduced.

The institution of 2.5 kilowatts as an authorized power level has been since the installation of most arrays. Where there was previously sufficient allocation space to easily fit 1 kilowatt, but not quite enough to fit 5 kilowatts, the broadcaster had to content himself with 1 kilowatt because there was nothing in between. Many of these arrays could now be increased to 2.5 kilowatts.

Some AM stations may find the options have recently, or will soon, improve due to the actions of other nearby stations. Usually, enlargement of a nearby station will reduce <u>your</u> options to expand; however, the comparatively recent opening of the Clear Channels for additional allocations works in a reverse manner. Many stations have found that they could increase service or add nighttime if they were to relocate to one of the newly available Clear Channels. Those who are successful in their applications to move to the new frequencies will vacate their present channels. This will result in increased permissible radiation in some directions for the many stations which remain.

From all outward appearances, the chances to acquire the best of all possible directional patterns have never been more promising nor the results more desirable.

SEVERAL METHODS OF OPTIMIZATION

We have found five techniques useful in designing optimum arrays. Some are considerably more powerful than others. They are listed in order of increasing power:

1) random search program (also know as shotgun search).
2) local maximam seeking program (also known as steepest gradient method).
3) a combination of 1 and 2 above.
4) graphical methods adapted for computer manipulation.
5) Z-transform methods adapted for computer manipulation.

All five methods utilize the flowchart in Figure 9. A set of parameters is evaluated for the pattern which it will produce. That pattern is then compared against the problem definition to establish the quality of the solution. If the solution is not acceptable, the parameters are modified and the process repeats until it is interrupted or an acceptable solution is found. The five methods differ in how the parameters are modified between iterations.

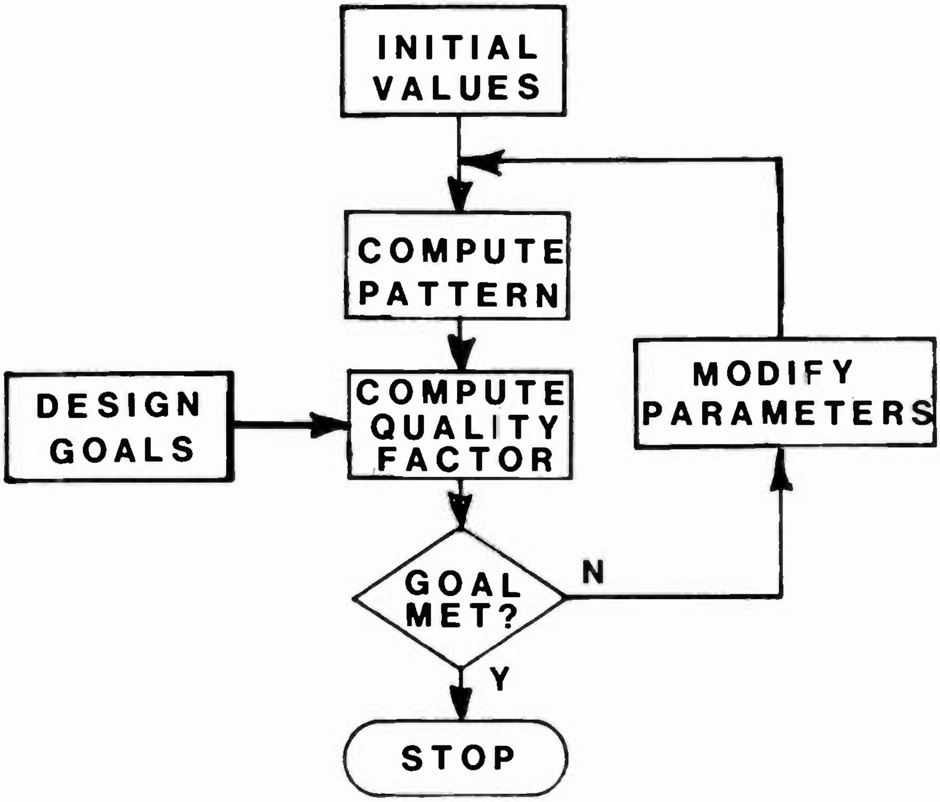


FIGURE 9 - THE BASIS FOR ALL FIVE COMPUTER METHODS

The simplest and most primative method is the random search. One simply places the calculations within several multiply nested DO loops (or FOR NEXT loops for those who speak BASIC) which exercise every possible combination of currents, phases and tower placements. This approach is much like mowing the lawn. In an organized pattern, you methodically go over every inch of the area.

The limitation is that the number of possible combinations is staggering. Increasing the step size will reduce the number of interations, but if the step size becomes too large, it becomes possible to skip over a solution. Assuming an agressive per-iteration time of one second and reasonable step sizes, a four-tower pattern with three unfixed tower locations will not complete a random search algorithm until the middle of the next century! The random search is useful only when the number of towers is confined to three or when the tower locations are fixed.

The steepest gradient method has been publicized widely in solutions to problems as diverse as designing cars or managing stock portfolios. Given an initial starting point, it will continue locating solutions with higher and higher quality factors. When it reaches a condition where there is no point which is adjacent to the most recent solution and which has a higher quality factor, it stops. It has reached a summit.

The problem is that the summit is only local maxima, or the best solution in the general area of the starting point. It is not a global maxima, or the best of all possible solutions. If there are additional more desirable summits or maxima within the permissible area, this method has no facility to move from one summit to the other. Accordingly, the quality of the resultant is dependant on the quality of the given starting point.

Our experience has shown that a thoroughly optimized pattern started from scratch will seldom resemble a pair multiplication solution in the placement of the towers. But as the starting point for this method is supplied manually by the engineer, and the engineer is likely to use pair multiplication to formulate his input, the results are confined to an area which may not include the one best solution.

A crude analogy might be to say that you can have any fruit in the orchard which you can pick from the road, while knowing that the best fruit in the orchard is usually far from the road.

The steepest gradient method is a useful tool for refining existing designs, but is nearly worthless when used alone on an unbounded problem with multiple degrees of freedom. It will seldom by itself reduce the required number of towers.

One solution to the problems presented so far is to use the above methods in conjunction with each other. One uses the random search with a very coarse step to find area(s) which have recurring high quality factors. After these areas are defined, the steepest gradient method is turned loose within them.

With some intuitive bounding of the degrees of freedom, this combined approach can produce useful results in a reasonable time frame. However, there is still no guarantee that the one best solution has been found. Assuming equal height towers and fixed placement of the reference tower, which has unity current at zero phase, a three tower array still has eight degrees of freedom. A four tower array has twelve degrees of freedom.

Additional computing power is always helpful, but it cannot completely offset what linear programmers call "the bane of dimensionality". Assuming an unreasonably coarse step size which will permit only six possible values for each parameter, with twelve degrees of freedom, the number of possible solutions is 2.17×10^9. (There are 3.15×10^7 seconds in a year). At one solution per second, the random search part of the procedure will be completed in 69 years. There is simply too much space to search to be able to do it thoroughly. A completely new approach is needed, something which makes what is going on more visible.

Laport[2] describes a method where the synthesis process can be represented graphically. While he describes a method for crossed-pair synthesis, the procedure can readily be adapted to the three tower case.

This graphical method makes it much more apparent how each parameter will affect the pattern. A computer program subroutine emulating the graphical method is inserted into the evaluation process in Figure 9, allowing the computer to "see" emerging trends. While this method's useful applications are limited to three and four tower systems, it has generated some three tower dog-leg designs which are sinfully efficient.

When more power is required than can be provided by any of the above methods, useful and efficient solutions have been found using a computerized adaptation of the method described by Jordan[3] and by Ramo, et al[4]. In it, the problem is mapped into the complex Z-plane.

As with most mapping and transform operations, procedures which would otherwise entail staggering amounts of computation are reduced to simple algebra. A filter problem which, by other methods, would be next to impossible to solve, becomes a matter of simply placing poles and zeroes on a sheet of the paper when the problem is mapped into the S-plane. Similarly, locating pattern nulls and determining their depth becomes a simple matter of placing lines across a paper when the problem is mapped into the Z-plane.

A computer subroutine emulating this procedure is also inserted into the evaluation procedure of Figure 9.

This procedure easily cracks nuts which are impervious to any of the first four methods described. An exciting side benefit of this method is its ability to simultaneously optimize the daytime and nighttime allocations situations to use the minimum number of towers.

Two parallel Z-planes are constructed, one representing the daytime problem, and the other, the nighttime problem. The zeroes representing reasonable solutions to each problem are placed on the appropriate plane. The program will attempt to solve both problems using the same zero(s) where possible. If a zero from the day pattern and a zero from the night pattern can be consolidated, the total number of zeroes (and the total number of towers) is reduced.

CONCLUSIONS

Optimization of directional AM antenna systems promises considerable benefit for the existing and the proposed facility alike. The procedure yields a number of positive benefits including higher performance and lower cost. There are no negative effects. Frequently, the cost of the optimization is more than offset by reductions in the cost of construction.

BIBLIOGRAPHY

(1) Glen T. Clark, "A Comparison of Current Broadcast Amplitude Modulation Techniques", IEEE Transactions on Broadcasting, Vol. BC-21, No. 2, June 1975.

(2) Edmund A. Laport, Radio Antenna Engineering. New York: McGraw-Hill, 1952, pp. 171-174.

(3) Edward C. Jordan, Electromagnetic Waves and Radiating Systems. Englewood Cliffs, NJ: Prentice-Hall, 1968, pp. 362-364.

(4) Ramo, Whinnery & Van Duzer, Fields and Waves in Communications Electronics. New York: Wiley, 1964, pp. 690-692.

APPENDIX A

IN-LINE ARRAYS: Unless the radiation limitations are symmetrical, any in-line array is a non-optimum design.

RANDOM OR DOG-LEG ARRAYS: are by definition an optimized array. More optimization may be possible with todays computer methods. The asymmetrical arrangement of the towers is a good starting point for further improvement.

PARALLELOGRAM ARRAYS: are completely un-optimized if all of the following conditions are met:

1) All opposite sides of the parallelogram are parallel (definition).

2) The sum of the theoretical phases of two diagonally opposite corner towers is equal to the sum of the phases of the other two corner towers.

3) The product of the theoretical field ratios of two diagonally opposite corner towers is equal to the product of the field ratios of the other two corner towers.

EXAMPLE: North tower operates with ratio 1.0 at 0°.
South tower operates with ratio 0.56 at 100°.
West tower operates with ratio 0.7 at 45°.
East tower operates with ratio 0.8 at 55°.

Do not use the operating or phase monitor parameters for this test. Only the theoretical parameters, such as from the posted license or Form 301 application, will produce valid results.

AM FIELD STRENGTH MEASUREMENTS BY AIR

Lyndon H. Willoughby

Diversified Broadcast Engineering, Inc.

San Antonio, Texas

Those elements which make a good AM transmitter site often result in an environment and terrain which does not lend itself easily to the gathering of field strength data as specified in Sections 73.151 and 73.186 of the Commission's Rules. The choice of transmitter location is often dictated strictly by the existing allocations window and required coverage area. When all things are considered one does not have much latitude in the choice of a new AM transmitter site.

In the case of existing AM sites, very often what orginally was an ideal, isolated location, has over the years seen open spaces within the first few miles become conjested with residential subdivisions, industrial parks and various other forms of land development.

If one considers in addition to the above the unique problems associated with certain areas of our country such as mountains, coast lines, military installations, national parks, forests and preserves, the difficulty in obtaining adequate field strength data can take on almost impossible proportions.

This paper addresses the procedure adapted in one such case. This is not the first or only time an aircraft was utilized in securing this type of data. Various individuals and companies offer aerial measurement procedures. If you find yourself in the position of considering aerial measurements, it should be carefully thought out and the decision made in concert with your consulting engineer and the FCC.

As you are probably aware, the cost of an aerial proof of performance is relatively expensive, with the preponderance of cost being the aircraft and pilot. The least expensive aircraft to rent is the single engine, fixed-wing type and the most expensive being the multi-passenger helicopter. Viewed strictly from the standpoint of obtaining field strength data, the helicopter offers many advantages over the fixed-wing aircraft. This is not say that data gathered using a fixed-wing aircraft is any less accurate, rather that the ability of the helicopter to execute certain manuevers tends to speed up data gathering and provides the means to calibrate, more properly the field strength meter.

The Commission's policy regarding aerial measurement of AM field strength requires that a measurement be made on the ground in the normal fashion wherever possible. With this policy in mind, aerial measurements offer little advantage to some stations and for the most part represent double work at a greatly increased cost.

The FCC has not specified one preferred procedure over any others. Each procedure is reviewed and stands on its own merit. However, there are certain criteria which must be satisfied and these criteria will be discussed later.

Should you decide to develop your own procedure, you should be aware of the fact that according to the Federal Aviation Administration, any alteration to an aircraft falls under their authority and review. There is a common misconception that if the aircraft owner determines an alteration is all right, that no other authority is necessary. This assumption is incorrect. The FAA is charged with the responsibility of protecting life and property where aircraft is involved and as such requires that any attachments to an aircraft have their approval. I am aware that some individuals have successfully circumvented the FAA, but it is still my responsibility to point out the law. If no accident occurs, it is quite possible that you can also successfully avoid involvement with the FAA, but for your own safety and the safety of the public at large, I strongly urge you to seek the FAA's cooperation in your aerial endeavors. Without their authority, insurance coverage does not exist and liability rests on the aircraft owner, the pilot, the air frame mechanic, your company and you individually.

Because of the nature of terrain surrounding the WNOE array, which is located in New Orleans, Louisiana, the ability to make adequate ground level measurements was impossible for virtually all compass bearings. Thus it was felt that use of a helicopter was the only feasible method to gather the required field intensity measurements. Conversations with FCC personnel led to the realization that one of the most important factors in performing measurements of this type is to remove the Field Intensity Meter from any adverse effects caused by it (the meter) being in the immediate environment of the helicopter.

Early developmental work on aerial measurements was performed with the meter located inside the aircraft. Various procedures which existed in 1979 were examined and subsequently rejected due to the belief that a better method could be developed to insure that the accuracy of the aerial measurements met or exceeded the accuracy obtained through normal ground measurement procedures. This method would have to satisfy four major criteria:

1.) There should be no error in the aerial F.I. measurements caused by the environmental effects of the helicopter.
2.) There should be a way to check and maintain F.I. Meter calibration while in flight.
3.) There should be a continuous hard copy readout of F.I. measurements obtained.
4.) There should be no navigational errors, and reasonable assurance that the location of the measurement was accurate.

FIELD EFFECTS OF METAL SURFACES ON FIELD INTENSITY METER

In order to satisfy criteria 1.) and 2.), it was decided to mount the loop antenna of the F.I. Meter below the helicopter, with the meter itself mounted in the helicopter cabin. The exact separation between the loop antenna and the helicopter fuselage was determined by constructing a metal surface somewhat larger than the surface area of the underside of a Bell 206 Jet Ranger helicopter. This surface was suspended from a wooden scaffold, 12 feet above the ground, and was raised and lowered using ropes and pulleys. A Potomac Instruments FIM-41 Field Intensity Meter was mounted on a tripod beneath the metal surface. A series of 126 measurements was then made, at frequencies varying from 600 kHz to 1580 kHz. The field intensities measured varied in strength from 4.25 volts to 130 microvolts. Measurements were first made without the surface and then with the surface, varying the height above the F.I. Meter by 12, 24, 36, 48, 60 and 72 inches. These measurements were then repeated at 36 and 72 inches and with the surface completely removed. The variation in readings from 48 inches to the readings with the surface completely removed varied as follows:

FIGURE 1

FREQUENCY kHz	F.I. MEASUREMENTS - mV/m 48"	60"	72"	Surface Removed
600	16.00	15.50	16.00	16.20
690	22.00	24.00	24.00	23.50
800	14.00	14.00	14.00	14.00
870	155.00	155.00	155.00	155.00
940	162.00	163.00	164.00	165.00
1060	4250.00	4000.00	4000.00	4000.00
1230	4.50	4.70	4.70	4.70
1280	47.00	47.00	47.25	47.50
1350	36.00	37.00	37.00	37.00
1400	.140	.142	.142	.142
1450	6.50	6.50	6.50	6.50
1510	1.50	1.60	1.60	1.55
1540	8.00	8.00	8.00	8.00
1580	.65	.65	.65	.65

As the above data shows, there is virtually no difference in field intensity beyond a distance of 48 inches.

After arriving at this separation of 48 inches, the Retractable Aircraft Mounting Bracket (U.S. Patent Pending) was designed. The bracket mounts to the forward cross tubes of a Bell 206 Jet Ranger helicopter. This bracket places the loop antenna of the Potomac FIM-41 Field Intensity Meter 49 inches below the skid pans and 73 inches below the fuselage of the helicopter. The loop antenna is electrically isolated from the bracket and the helicopter. The bracket is motorized and raises the loop into the stowed position for take off and landing, then lowers the loop into measurement position, where it is locked into place. In this configuration, the measured signal and the calibration signal are brought from the loop to the meter by special cables. These cables are fed to the meter via the normal contact points, therefore maintaining complete and absolute integrity. Since the normal antenna inputs are used rather than the "external RF" input, all calibration circuitry remains intact and operational. The external loop remains an integral part of the FIM-41. The meter is operated, calibrated and read exactly as it would be on the ground.

The loop antenna, complete with cables, connectors and the FIM-41 were sent to Potomac Instruments where Mr. Cliff Hall, the design engineer of the FIM-41, examined and calibrated the entire system and provided a calibration certificate.

FAA CERTIFICATION OF LOOP MOUNT

After all of the above had been accomplished, the necessary steps were taken to obtain Federal Aviation Administration certification for the RAM bracket, as required by FAR 43.7, 43.9, 43 Appendix B, and AC 43.9-1. This entailed submitting detailed drawings and specifications for approval by FAA engineers and General Aviation District Office personnel, as well as performing the following in-flight tests:

1.) Functional reliability testing of the bracket
2.) Wind loading properties
3.) Airfoil properties
4.) Vibration testing
5.) Functional in-flight testing
6.) Emergency landing procedure testing
7.) Auto-rotation performance testing
8.) Stress analysis

All of the above tests were performed while the aircraft was in the "Experimental" class. Upon acceptance by the FAA of Form 337, a "Restricted Airworthiness Certificate" was issued to Peterson Maritime Services, the owners of the helicopter.

This entire FAA certification procedure was an absolute requirement since the attachment of any external device to an aircraft must be approved by the FAA. Photographs of the RAM bracket equipment set-up in the helicopter are included as Figures 9 and 10 at the end of this text.

In order to determine what, if any, effect altitude might have on the aerial F.I. measurements, vertical F.I. measurements were made. For the vertical measurements, two points were chosen. The first was on the 15 degree radial to a distance of 8.86 miles from the array. The second was on the 73 degree radial at a distance 8.05 miles from the array.

VERTICAL TEST ON 15 DEGREE RADIAL

This point, 8.86 miles from the transmitter, is on the levee of the Mississippi River Gulf Outlet. A ground measurement was first made. Several attempts were then made to climb vertically, but due to the weight, the helicopter was unable to lift smoothly. The pilot burned up fuel and removed all non-essential equipment and the navigator. This sufficiently reduced the load, and the helicopter was able to climb vertically. The pilot used a 560' smoke stack in the distance as his guide to insure a perfectly vertical climb. The pilot called out the altitude in fifty foot increments and "tick

FIGURE 2
DETAILS OF RESTRICTED CATEGORY LIMITATIONS

DEPARTMENT OF TRANSPORTATION
FEDERAL AVIATION ADMINISTRATION

General Aviation District Office 8
FAA Bldg., Lakefront Airport
New Orleans, La. 70126

RESTRICTED CATEGORY OPERATING LIMITATIONS

Issue Date 10-2-81

Make Bell Model 206B

Registration No. N104PM Serial No. 2298

1. GENERAL

A. At all times when the aircraft is in the Restricted Category, it shall be operated in accordance with FAR 91.39.

B. Each conversion from Restricted Category to Standard and from Standard to Restricted Category shall be performed in accordance with copy of FAA Form 337, attached, and dated 10-2-81.

C. Each conversion shall be performed by a certificated airframe mechanic or appropriately rated repair station.

D. Each person performing a conversion shall record the work performed in the aircraft records in accordance with FAR 43.9.

2. The following placards are required for restricted operations in full view of the pilot and passengers.

A. Placard to be installed right side center console to read:

"3,000 lbs. maximum gross wt 70 mph maximum airspeed. Aircraft limited to pilot and two passengers".

B. Placard to be installed on antenna control box:

"Antenna operation from the stowed to extended position will be performed while the aircraft is at a minimum of 10 ft hover. Antenna will not be moved while aircraft is in forward flight. Radio frequency test will normally be performed at 60 mph. However, airspeed must not exceed 70 mph".

3. Other placards are also to be installed in accordance with FAR 45.23(b).

4. Only pilot and aerial survey equipment operators are authorized while aircraft is being operated in Restricted Category.

OFFICE: SW-GADO-8 LOCATION: New Orleans, LA.

FAA INSPECTOR: Earl A. McCarthy, Jr. DATE: 10-2-81
General Aviation Inspector

marks" were made on the chart.

We then hovered at approximately 25 feet and slowly rotated the helicopter for a maximum on bearing reading. This corresponded with the radio directional finding compass on board the helicopter, which was tuned to WNOE. The F.I. meter was then calibrated. When a altitude of 1200 feet was reached, the helicopter was rotated 90 degrees to an off bearing point. This complete segment was recorded on the chart recorder.

The chart of these measurements is shown as Figure 5. The rotations and meter calibration are clearly visible. A tabulation of the heights, reading and ratio to the ground measurement appear as Figure 3. The arithmetic average ratio for the 12 altitude measurements if 1.0017. The ratio on a point by point basis for the altitudes between 200 feet and 500 feet as compared to the ground was 1.0010.

VERTICAL TEST ON THE 73 DEGREE RADIAL

This point, 3.05 miles from the transmitter, is in a heavily wooded area in the F. Edward Hebert Agricultural Research Center. A ground measurement at the location was selected and a measurement made. The same procedure was followed in this series of measurements as it was for the 15 degree measurements.

A chart of the measurements is included in Figure 5. The rotations for maximum signal is clearly visible. A tabulation of the heights, readings and ratio to the ground measurement appears as Figure 4. The arithmetic average ratio for the elevel altitude measurements is 0.9958. The ratio on a point by point basis for the altitudes between 200 feet and 500 feet as compared to the ground was 1.000.

FIGURE 3
VERTICAL MEASUREMENT ON 15° RADIAL AT A DISTANCE
OF 8.86 MILES FROM THE ARRAY

Ground Measurement = 123 mV/m

Measurement No.	Altitude (Feet)	mV/m	Air Ratio - Ground
1	100	123	1.0000
2	200	123	1.0000
3	300	123	1.0000
4	400	123	1.0000
5	500	123.5	1.0041
6	600	123	1.0000
7	700	124	1.0081
8	800	124	1.0081
9	900	125	1.0163
10	1000	122	0.9919
11	1100	123	1.0000
12	1200	122	0.9919
	Average Ratio		1.0017
	Between 200 & 500 Ft.		1.0010 Avg

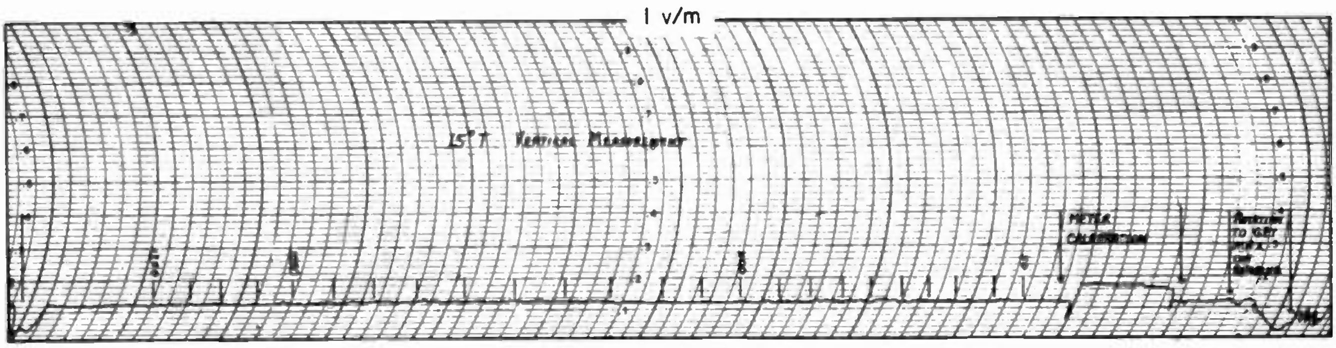
These slight variations are believed to be due to vibrations of the helicopter.

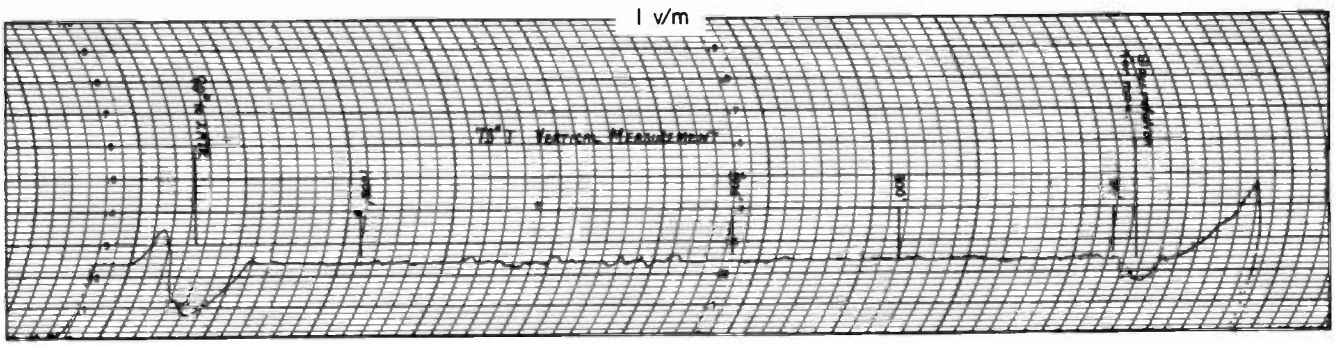
FIGURE 4
VERTICAL MEASUREMENTS ON 73° RADIAL AT A DISTANCE
OF 3.05 MILES FROM THE ARRAY

Ground Measurement = 240 mV/m

Measurement No.	Altitude (Feet)	mV/m	Ratio - Air Ground
1	100	240	1.00
2	200	240	1.00
3	300	240	1.00
4	400	240	1.00
5	500	240	1.00
6	600	240	1.00
7	700	240	1.00
8	800	230	.9583
9	900	240	1.00
10	1000	240	1.00
		Average	0.9958

FIGURE 5





The preceding data demonstrates that altitude (up to 1,000 feet) has virtually no effect on the measured field intensity when the radiators are 90 degrees tall. Therefore it was decided that flying at an altitude of 300 feet above any obstruction (which is the minimum altitude allowed for helicopters by the FAA) would cause no adverse effects to the aerial F.I. measurements.

PROCEDURE WHILE IN FLIGHT

At this stage, a calibration radial was measured in order to ascertain if any discrepancy existed between ground and air measurements. The 253 degree radial was chosen, since it is common to both modes of operation and its intervening terrain is characteristic of all types encountered along the other measured radials. Ground measurements were first obtained, in some instances by landing the helicopter and wading away a sufficient distance to obtain an accurate reading. Aerial measurement were then made using the following procedure:

1.) Landmarks (corresponding to each point of measurement) along the radial were marked on 7.5 minute topographic maps. It was the responsibility of the navigator (who had also made the ground measurements on that radial) to assist the pilot in recognizing these landmarks and to inform the back seat equipment operator when they were directly below the helicopter.
2.) F.I. Meter calibration was checked and maintained by the back seat equipment operator, who also marked the strip chart of the Esterline-Angus recorder (operating at a constant chart speed) when instructed to by the navigator.
3.) Proper aircraft heading was established prior to overflight of the transmitter building. Using the procedure outlined above, landmarks were called out and marked at distances between one and twenty miles from the transmitter. This method enabled each crew member to concentrate on his particular task, i.e., accurate navigation and accurate marking of the strip chart.
4.) Both the outbound and inbound legs were measured, however only inbound measurements were utilized in the analysis. By flying the radial outbound, the crew was able to familiarize themselves with the radial and assured a true course on the inbound run.
5.) F.I. measurements were extracted for each number marked on the strip chart. These point numbers and measurements were then matched with their appropriate mileage from the array.

Each aerial F.I. measurement was then ratioed to its corresponding ground measurement, summed, and an arithmetic mean ratio obtained. This entire measurement and evaluation procedure was performed on this radial while in the non-directional mode of operation. The arithmetic mean ratio for this mode is 0.997484. These measurements and ratios are included in Figure 6 and 7. Non-directional aerial measurements for all other radials were then made using this same procedure.

In most cases, a sufficient number of readily identifiable landmarks (in excess of 20) on each radial were not difficult to obtain, due to the heavily developed areas in the vicinity of the transmitter site and the prolific oil and gas wells (along with their supporting canal systems) in the outlying area. Landmarks were usually not only available at each measuring point, but also in between, enabling the crew to know their exact location at any given moment. These landmarks were partly comprised of well heads, bayous, canals, levees, power lines, channel markers, shorelines, pipelines, pumping stations, roads, buildings, etc.

FIGURE 6

WNOE RADIAL 253.0
NON-D AIR TO GROUND COMPARISON

POINT	DISTANCE (MILES)	GRND (MV/M)	AIR (MV/M)	RATIO
1	1.0	1050	1050	1.000
2	1.1	950	940	0.989
3	1.5	700	700	1.000
4	1.6	690	680	0.986
5	1.8	610	605	0.992
6	1.9	610	590	0.967
7	2.0	545	545	1.000
8	2.5	478	480	1.004
9	3.1	417	415	0.995
10	3.6	350	350	1.000
11	4.1	225	220	0.978
12	4.8	205	210	1.024
13	5.3	200	200	1.000
14	5.6	195	190	0.974
15	6.6	130	130	1.000
16	8.0	108	110	1.019
17	9.7	80	80	1.000
18	10.5	75	75	1.000
19	11.5	76	75	0.987
20	13.2	62	62	1.000
21	14.0	55	55	1.000
22	15.3	56	56	1.000
23	18.0	45	45	1.000
24	19.0	37	37	1.000
25	20.0	37	37	1.000

RADIAL AVERAGE: 0.997

FIGURE 7

WNOE RADIAL 253.0
DA-DAY AIR TO GROUND COMPARISON

POINT	DISTANCE (MILES)	GRND (MV/M)	AIR (MV/M)	RATIO
1	1.0	650	670	1.031
2	1.1	580	580	1.000
3	1.5	350	350	1.000
4	1.6	330	320	0.970
5	1.8	330	330	1.000
6	1.9	335	340	1.015
7	2.0	305	305	1.000
8	2.5	300	280	0.933
9	3.1	240	240	1.000
10	3.6	210	200	0.952
11	4.1	144	140	0.972
12	4.8	120	120	1.000
13	5.3	150	137	0.913
14	5.6	118	120	1.017
15	6.6	81	80	0.988
16	8.0	61	60	0.984
17	9.7	52	52	1.000
18	10.5	35	36	1.029
19	11.5	44	44	1.000
20	13.2	38	38	1.000
21	14.0	26	26	1.000
22	15.3	35	35	1.000
23	18.0	23	23	1.000
24	19.0	23	22	0.957
25	20.0	20	20	1.000

RADIAL AVERAGE: 0.990

INTERPOLATION BETWEEN LANDMARKS

In the few instances when interpolation between landmarks was necessary, the helicopter was flown so that a relatively constant ground speed of 60 mph was maintained between landmarks. This enables an accurate scale of chart length per land mile to be calculated for each interpolated section, since the distances involved were never more than five miles. For example, on the 253 degree radial non-directional strip chart (Figure 8), the distance between 18 miles and 20 miles is 11.75 centimeters. With a constant ground speed, 1 mile = 5.75 centimeters. To find 19 miles on the strip chart, proceed 5.75 centimeters to the right of the 18 mile point on the chart. As can be seen from the strip chart, the distance of the 19 mile point was also marked on the graph as it was flown over, it is believed that this method is extremely accurate. (Note: In order to facilitate navigation, a constant ground speed was not attempted for the entire 253 degree radial. Therefore, the aforementioned interpolation procedure does not apply for the entire length of the radial.) The same procedure was followed in gathering the DA-Day and DA-Night data.

Before final evaluation of the aerial F.I. measurements was completed, an air to ground comparison was made for each point that a ground measurement was available. This data was then ratioed in the same manner described previously for the 253 degree calibration radial. The average ratio for all points (NON-D, DA-Day and DA-Night) was 1.00103. As can be seen, excellent correlation exists between ground and air F.I. measurements. Since the variations between ground and air fields are well within the normal variation of any ground made remeasurement procedure, no factor was applied to the aerial measurements.

Several points should be made with regard to conclusions drawn from the procedure as it was applied to this particular station:

1.) WNOE utilizes 90 degree towers. In systems that utilize towers greater than 90 degrees in height, the altitude at which the measurement is made could become more critical. Preliminary tests should be made to determine the most desirable flight altitude.
2.) There are areas surrounding some tranmsitter sites where there are insufficient landmarks to assure accurate visual navigation. In these cases, additional navigational aids will be necessary, the use of navigational computers such as the Loran C provide excellent results. Most navigational computers calculate distance and bearing based on triangulation from reference point radio beacons and care must be exercised per the computer manufacturer's instructions.
3.) Close-in measurements (0 to 1 mile) were not incorporated in the WNOE proof of performance, however, in most cases these measurements will be required. If the terrain immediately around the site is impassable and aerial measurements are deemed necessary, a method utilizing a laser range meter and a theodolite (transit) can provide a very accurate determination of measurement locations.
4.) Ground conductivity around the New Orleans area consists of mostly 8 and 15 mmho. There was no documentation available which addresses areas of very low conductivity. Information volunteered during the research indicates that these low conductivity areas often form higher fields around 1000' above ground due to surface attenuation.

In conclusion, because of the Commission's position on aerial measurements versus ground measurements, the gathering of field strength data by air is financially prohibitive except in those cases where no alternative is available. This paper addresses the fact that accuracy and repeatability is attainable provided careful thought is applied to each step in the procedure. All of us who are involved with aerial measurements encourage further developments and improvement in the methods of gathering accurate field strength data by air.

ACKNOWLEDGEMENTS:

Daniel Miller and Ernest Kain, Jr. of WNOE Radio for their help and ideas, as well as being members of the measurement crew.
Kurt Voss of Diversified Broadcast Engineering, Inc., for assisting in the measurements and preparation of the final proof report.
Jim Bell, Chief Pilot; Mike Chapman, Pilot; Larry Metzler, Chief Mechanic; all of Peterson Maritime, Inc., who gave unselfishly of their time and without whom FAA clearance could not have been obtained.
Cliff Hall and David Harry of Potomac Instruments who took this project right in stride.
James Lytle and Earl McCarthy, Jr. of the FAA General Aviation District Office #8, New Orleans, who helped immeasurably in securing speedy FAA authority while insuring the safety of the crew as well as the general public.
John H. Mullaney, Sr. of John H. Mullaney Consulting Radio Engineering, Inc., for his comments and review of this procedure.
John Sadler of the Federal Communications Commission for his direction in aerial measurement criteria and acceptance of WNOE data for license grant.

FIGURE 8

253° NONDIRECTIONAL RADIAL
CHART RECORDER GRAPH

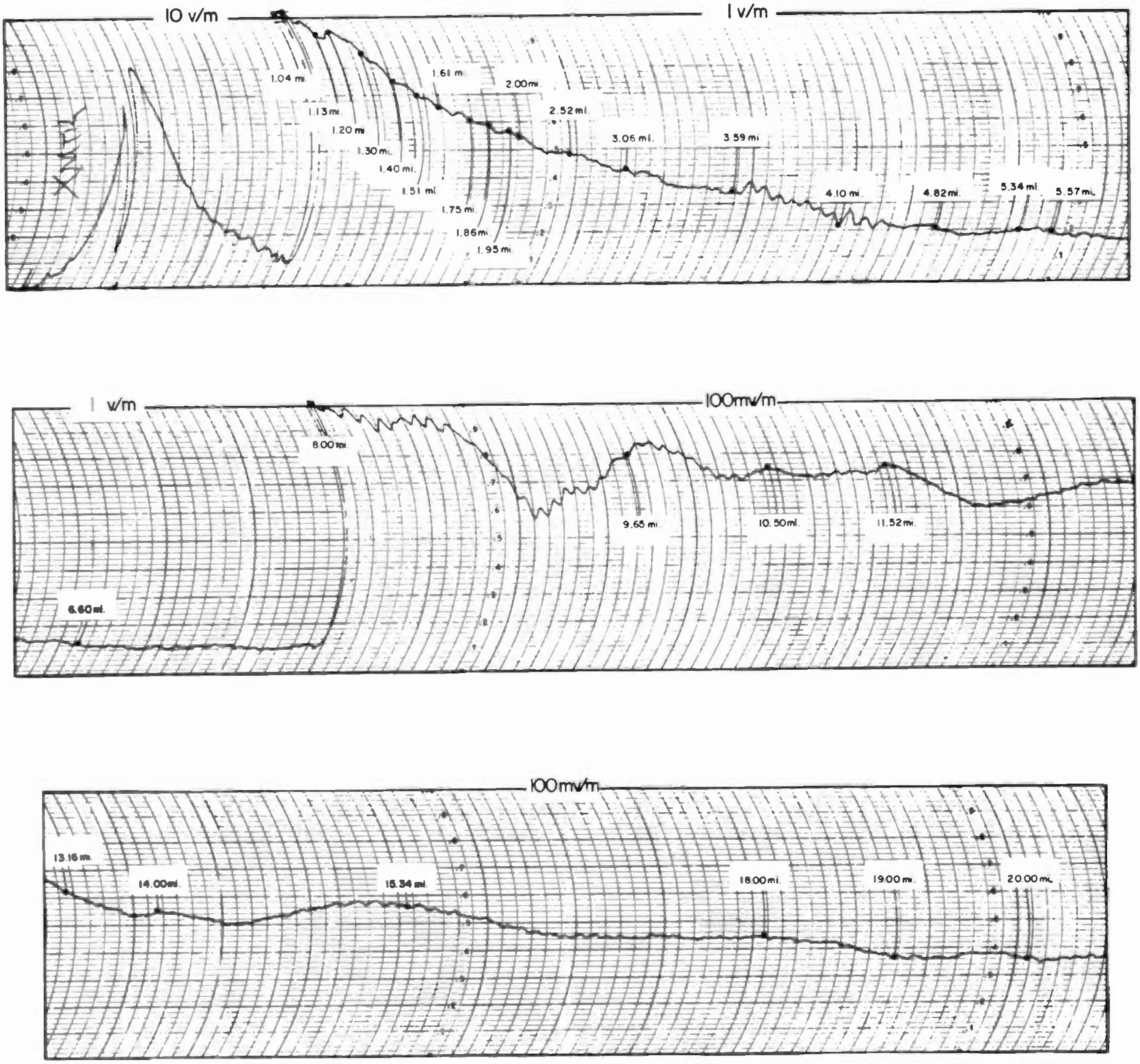


FIGURE 9

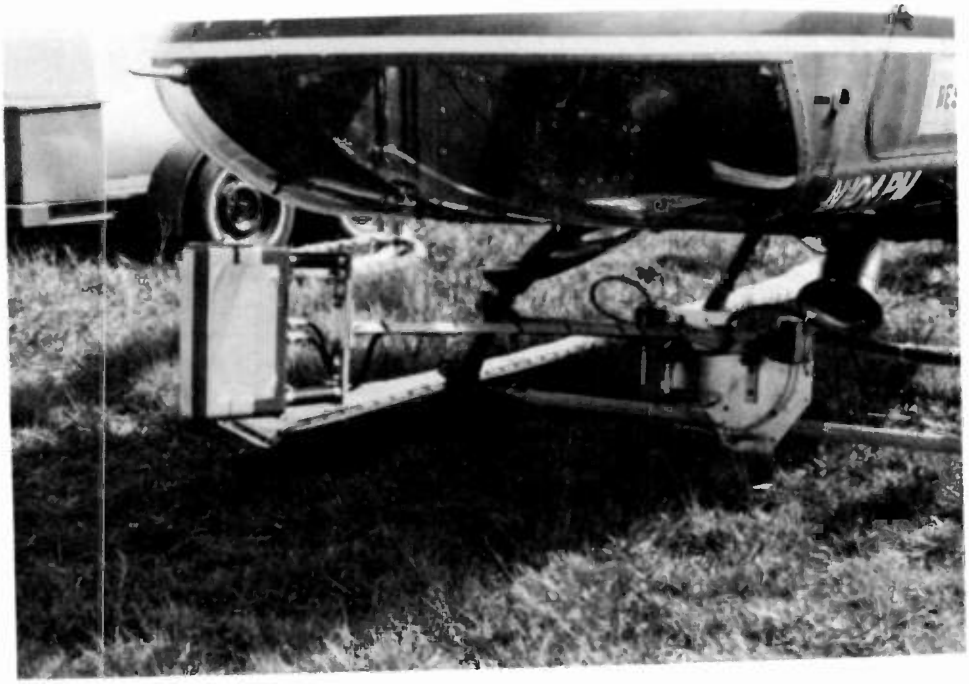
Loop In Measurement Position

Loop In Stowed Position

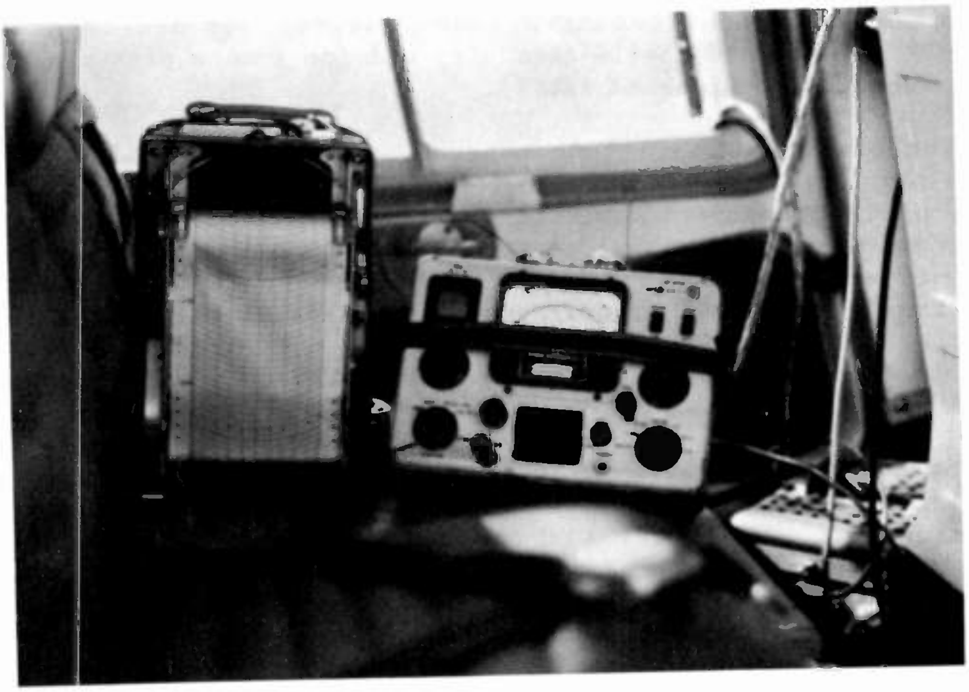
Loop In Measurement Position

Loop Being Lowered

FIGURE 10



Retractable Aircraft Mounting Bracket & Loop Assembly



Equipment Set Up

AMPLITUDE COMPANDORED SIDEBAND

WHAT'S IN IT FOR BROADCASTERS?

Ralph A. Haller

Federal Communications Commission

Washington, DC

BACKGROUND

If the demands for more uses of the radio spectrum are to be met, then either: 1) more spectrum must be made available, or 2) existing spectrum must be used more efficiently. Most of what can be considered useable spectrum by today's standards has already been allocated. This forces the second option, namely, spectrum efficient technologies. One type of new technology that offers promise in the land mobile community and for remote pickup uses is known as Amplitude Compandored Sideband (ACSB).

The ancestors of ACSB include conventional amplitude modulation (AM) and suppressed carrier single sideband (SSB). The first major step in the advancement of the radio art came when AM was developed. Amplitude modulation permitted transmission of actual sound information by "modulating" or changing the amplitude of the radio frequency (RF) carrier at a rate equal to the modulating frequency. This process produces sidebands that are displaced from the RF carrier by the modulating frequency (see figure 1a). AM signals are easy to produce, transmit, and receive. Unfortunately, AM signals lack high spectrum efficiency.

When analyzing AM, it becomes apparent that the carrier and one of the two sidebands do not convey any information that cannot be retrieved from the remaining sideband. This fact led to the development of suppressed carrier single sideband. In SSB, the transmitter basically produces an AM signal and then strips off the carrier and unwanted sideband. Only the remaining sideband gets transmitted (see figure 1b). At the receive end, a locally generated carrier must be combined with the received sideband to allow demodulation.

SSB has been used successfully by the military and by amateur radio operators for several years, primarily at high frequencies (HF). The major

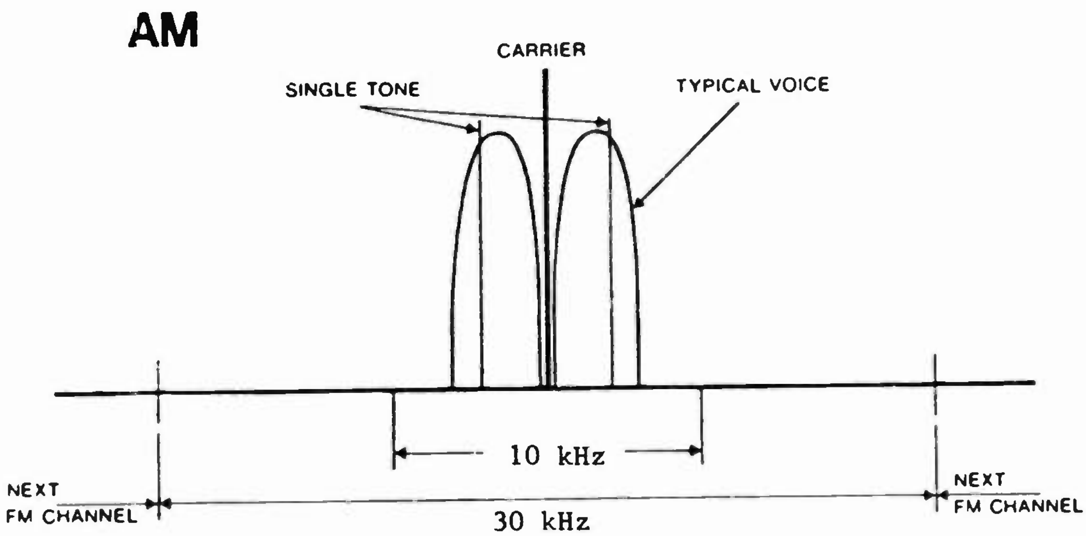


FIGURE: 1a

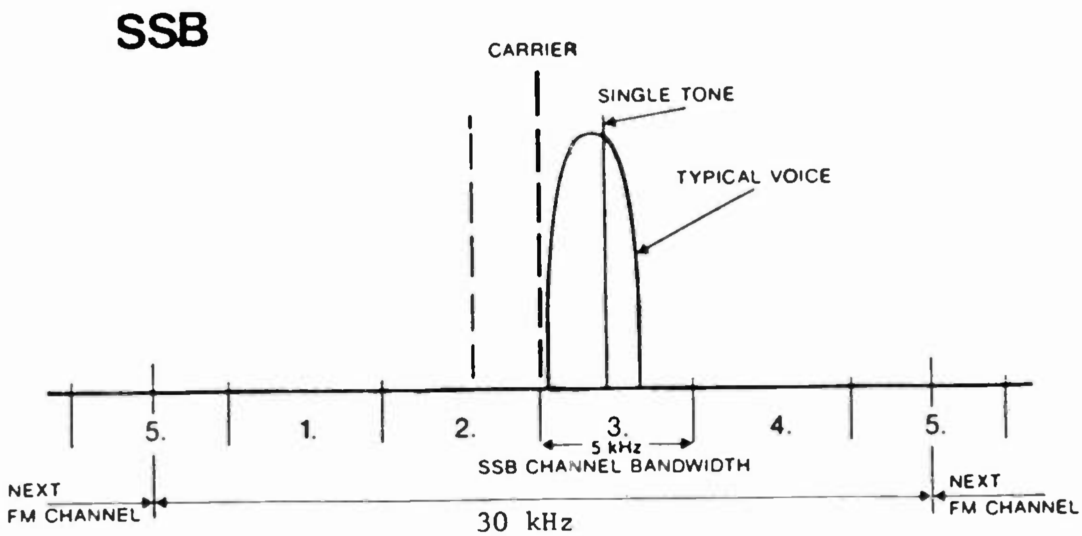


FIGURE: 1b

(Figures courtesy of Sideband Technology Incorporated) drawback to SSB has been the difficulty of tuning the receiver with no reference carrier present. Slight misadjustments result in the recovered voice sounding like "Donald Duck." The problem becomes even more difficult at very high frequencies (VHF) and above due to frequency instabilities in equipment.

Encouraged by the Federal Communications Commission (FCC) through a Notice of Inquiry and a funded study, Stanford University improved SSB technology significantly. Under the direction of Dr. Bruce B. Lusignan, an audio pilot tone and audio processing were added to traditional suppressed carrier SSB type signals. The result was a communications system that could be used at VHF, had an exceptionally low background noise level, and maintained the spectrum efficiency of conventional SSB.

CURRENT ACSB COMMERCIAL TECHNOLOGY

In the purest definition, ACSB would involve only the application of audio amplitude compression and expansion to existing SSB. The compression of audio (reduction of dynamic range) at the transmitter results in a high average level of modulation. In currently available ACSB equipment, the compression ratio is four-to-one. This means that a 40 decibel (dB) dynamic range in the input audio will be transmitted as only a 10 dB dynamic range. In a similar manner, the receivers perform the inverse of compression and expand the received signal, generally by a four-to-one factor, so that the recovered audio again has its 40 dB dynamic range.

The broadcasting industry has used transmission compression for years to improve the signal to noise ratio of the recovered audio at receivers. The major difference between ACSB and traditional compression as used in broadcast is that the ACSB uses very severe compression so receivers must perform at least some expansion of dynamic range to make the recovered audio acceptable to the end users. Broadcasters normally use far less compression, so receivers need not compensate for the compression. Figure 2 shows a graphic representation of the four-to-one compression and expansion process. Note that the expansion process reduces the relative power of the noise in the transmission path by four-to-one, resulting in a very quiet received signal.

The ACSB as developed by Stanford University uses upper sideband and includes an audio pilot tone. The pilot tone in use in today's commercial equipment is 3.1 kilohertz (kHz). Before transmission, the 3.1 kHz tone is mixed with the compressed audio (manufacturers have chosen slightly different methods of doing this, resulting in different operational tradeoffs). This tone provides two important functions in the receiver. First, because the tone is always 3.1 kHz above the suppressed RF carrier frequency, the receiver can incorporate an automatic frequency control (AFC) circuit to match exactly the receiver frequency to the transmitter. This solves the receiver tuning problem discussed earlier. Second, the pilot tone provides a known reference level for the receiver to use in an automatic gain control (AGC) circuit. Conventional SSB continually varies in amplitude as the modulation changes, making effective AGC at the receiver difficult. The pilot tone in ACSB provides the receiver with a known reference level for the AGC circuit.

The choice of 3.1 kHz does limit the audio bandwidth of the equipment to something less than 3 kHz. For on-air remote broadcasts, this bandwidth may not be fully acceptable, so future equipment developed specifically for the broadcasting industry may use a pilot tone of higher frequency to widen the

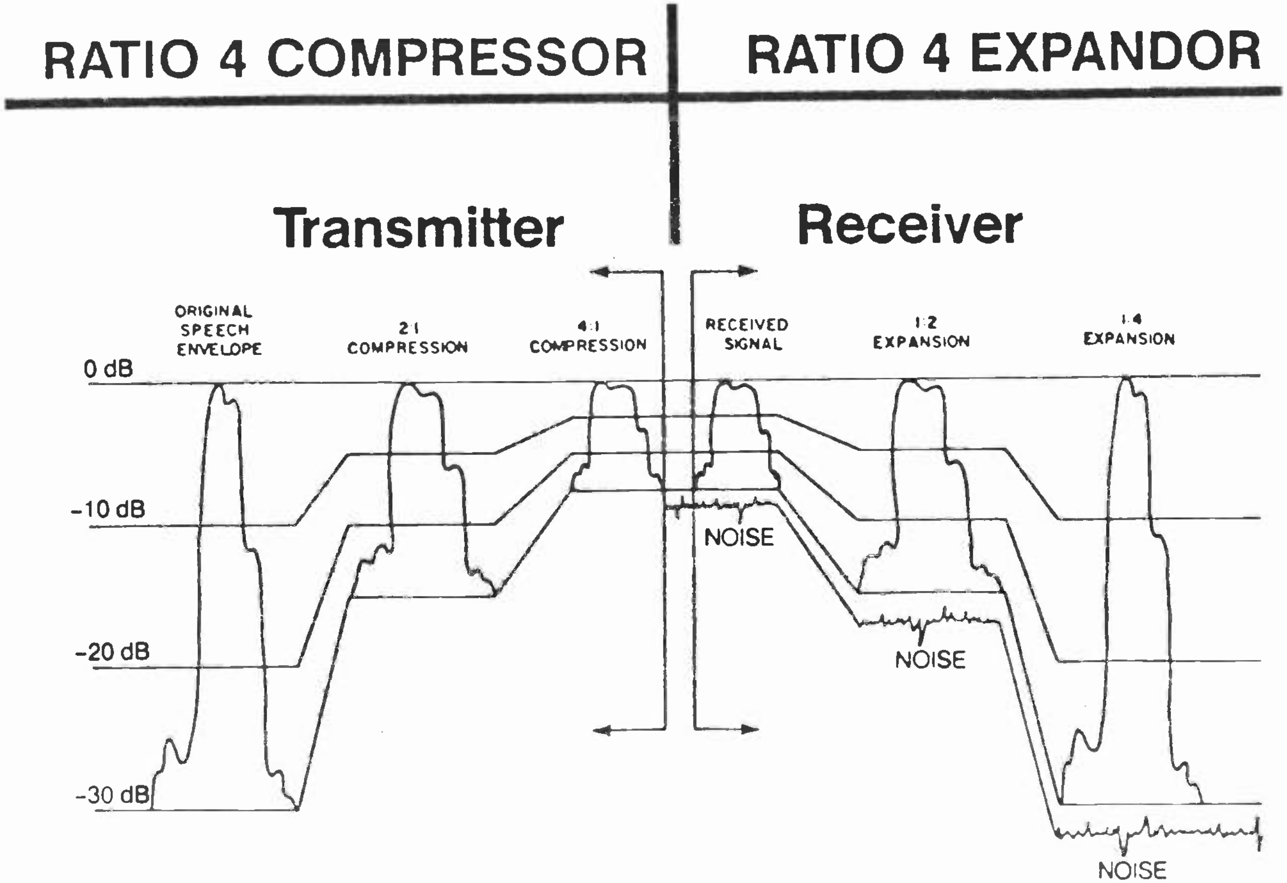


FIGURE 2: THE COMPANDORING PROCESS
(Courtesy of Sideband Technology Incorporated)

permissible audio bandwidth. ACSB can also be made to work without a pilot tone if highly stable frequency control elements are used in the transmitters and receivers. Nevertheless, the currently available equipment may be quite adequate for many remote pickup applications.

ACSB COMPARED TO FM

Most remote pickup equipment for the VHF and UHF (ultra high frequency) bands now employs frequency modulation (FM). FM can provide good audio fidelity along with high signal to noise ratios. FM receivers normally have limiter circuits to remove amplitude variations in the received signal, resulting in good discrimination against unwanted transmission noises. The primary disadvantage to FM rests in the bandwidth required to convey information. Significant sidebands may extend on each side of the carrier to three or four times the highest modulating frequency. Figure 3 shows a spectrum analyzer display of the RF spectrum from an FM transmitter with voice modulation.

The compandoring process gives ACSB most of the same characteristics that have been associated with FM; however, ACSB uses considerably less spectrum than FM to convey the same information. Figure 4 shows a spectrum analyzer display of the RF spectrum from an ACSB transmitter with voice modulation.

By comparing figure 3 with figure 4 (note the difference in horizontal scales), one can see that for the voice modulation used in the example, ACSB requires less bandwidth. Taking -40 dB as the reference level, the FM bandwidth is approximately 16 kHz versus only slightly over 5 kHz for the ACSB. Therefore, the spectrum savings alone make ACSB appear as an attractive alternative to FM in today's congested spectrum.

FCC LABORATORY TESTS

The FCC Laboratory conducted an extensive testing program to compare ACSB to FM and to determine how ACSB might be allowed in the land mobile services with a minimum disruption to current FM users. The results of the FCC work are contained in a report entitled, "Amplitude Compandored Sideband Compared to Conventional Frequency Modulation for VHF Mobile Radio: Laboratory and Field Testing Results," publication number FCC/OST TM83-7, published in October 1983. Although no attempt will be made here to describe fully the FCC's work, three principal results of the testing program merit some discussion.

First, the Commission found that co-channel interference (desired signal to undesired signal (D/U) level) was almost identical between ACSB and FM. FM systems have been credited with a "capture effect" with co-channel FM signals. For a just noticeable level of interference between two FM signals, the FCC found that the level of the undesired signal could be as high as 13 dB below the desired signal. Disruptive interference occurred at a D/U ratio of approximately 2.5 dB (at a desired signal level of -100 dB referenced to a milliwatt "dBm").1/ For ACSB the just noticeable level was at a D/U ratio of 9 dB and the disruptive level was 2.3 dB (again -100 dBm desired

1/ Ralph Haller and Hank Van Deursen, "Amplitude Compandored Sideband Compared to Conventional Frequency Modulation for VHF Mobile Radio: Laboratory and Field Testing Results," FCC, October, 1983, page 57.

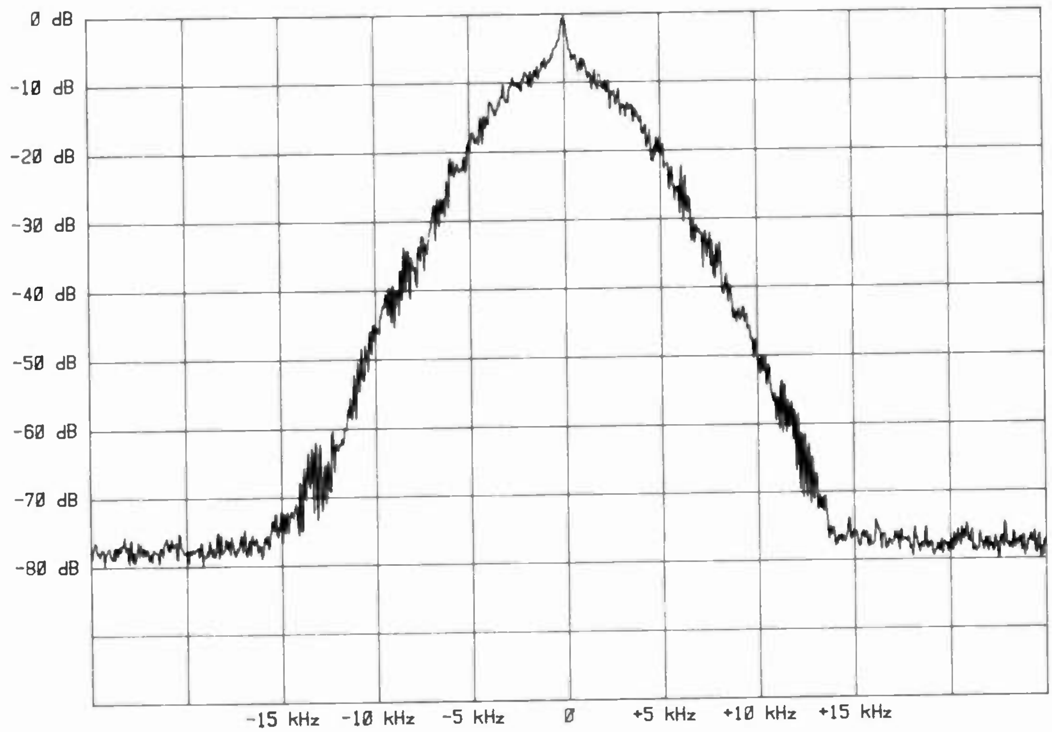


Figure 3
FM emission spectrum, VOICE

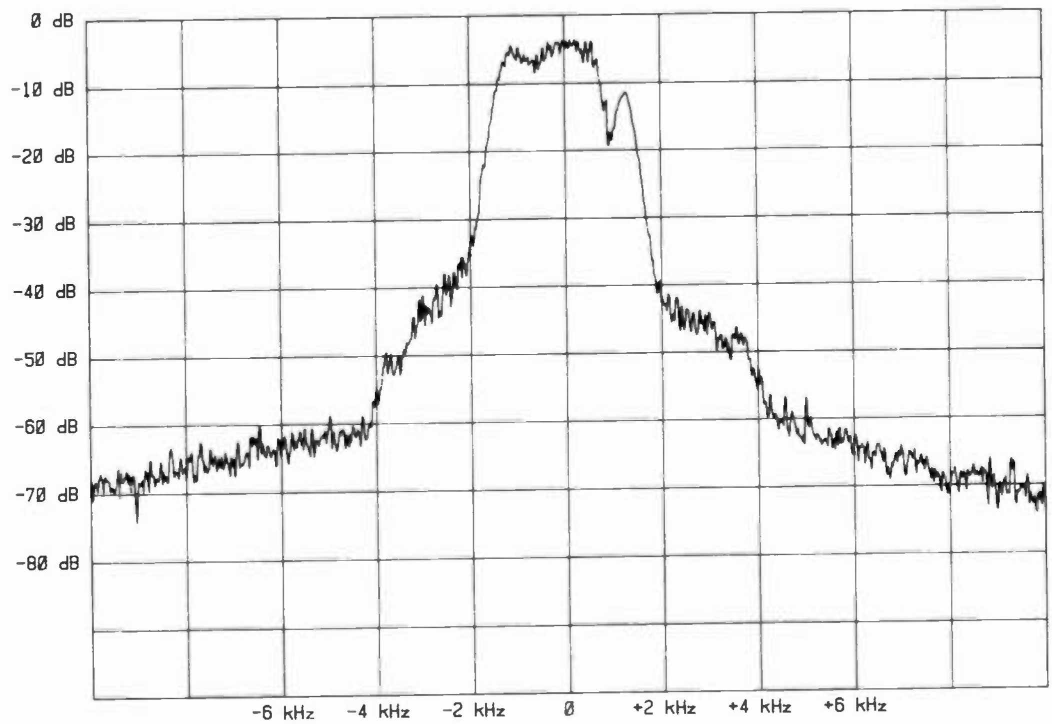


Figure 4
ACSB emission spectrum, VOICE

signal).2/ Notice that these results suggest that ACSB exhibits a capture effect similar to that of FM.

Second, existing FM channels in the remote pickup and land mobile services are usually 25 kHz or 30 kHz wide. The question of channel bandwidth for ACSB was addressed by the FCC. Based on an assumed required adjacent ACSB channel protection of 60 dB, the currently required bandwidth would be approximately 7 to 8 kHz.3/ This will likely improve to 5 kHz as technology advances. Currently, the limiting factors are: 1) the need for more selective filters in ACSB receivers, and 2) higher than desired adjacent channel emissions from ACSB transmitter power amplifiers. Very recent improvements in power amplifiers by one manufacturer have already helped the situation by about 9 dB.

Third, the FCC studied the potential impact of introducing ACSB into the existing FM two-way services. Specifically, several options exist for allowing ACSB on the air, but some will have more of an impact on existing users than others. The Commission found that, as one would expect, the more removed the center of the ACSB channel is from the center of the FM channel, the less interference will be mutually received by both users. The most advantageous placement of the ACSB channels would be immediately between existing FM channels (interlacing). For 25 kHz FM channels, this would mean that the ACSB channels would be centered 12.5 kHz from the FM channels. Likewise for 30 kHz FM channels, the ACSB channels would be removed by 15 kHz. Another alternative for 30 kHz FM spacing would be to place two ACSB channels between the FM channels, each at 12.5 kHz spacing from its closest FM channel. This last option is shown in figure 5.4/ One other option is simply to divide the existing FM channels into 5 kHz segments and allow for an orderly conversion to narrow-band technology. One important finding in the ACSB into FM interference case is that the interference tends to be more of a function of the FM receiver selectivity than a function of the exact characteristics of the ACSB transmitter.

Although interlacing provides the greatest isolation between the FM and ACSB services, it poses some problems. The difficulty occurs because the interlaced ACSB channels fall on the edges of two FM channels. Although the mutual interference between the FM and ACSB users will be minimal, if the existing FM channels are in different radio services, frequency coordination becomes more complex. Also, individual licensees cannot enjoy the maximum possible freedom in selecting transmission modes and frequencies because interlaced channels would require coordination (i.e., licensees could not, on their own, make the choice to use ACSB without coordinating with another radio service). The coordination exercise could become a very laborious task in areas such as Southern California and New York which have severe frequency congestion.

Therefore, the simple idea of dividing existing FM two-way channels into 5 kHz segments appears to be a highly viable, long-term solution, particularly in the VHF remote pickup service where adjacent FM channels are not always in the same service. This plan allows existing FM remote pickup licensees to make more choices of how to maximize spectrum efficiency. The plan also minimizes

2/ Ibid., p. 48.
3/ Ibid., p. 55.
4/ Ibid., p. 9.

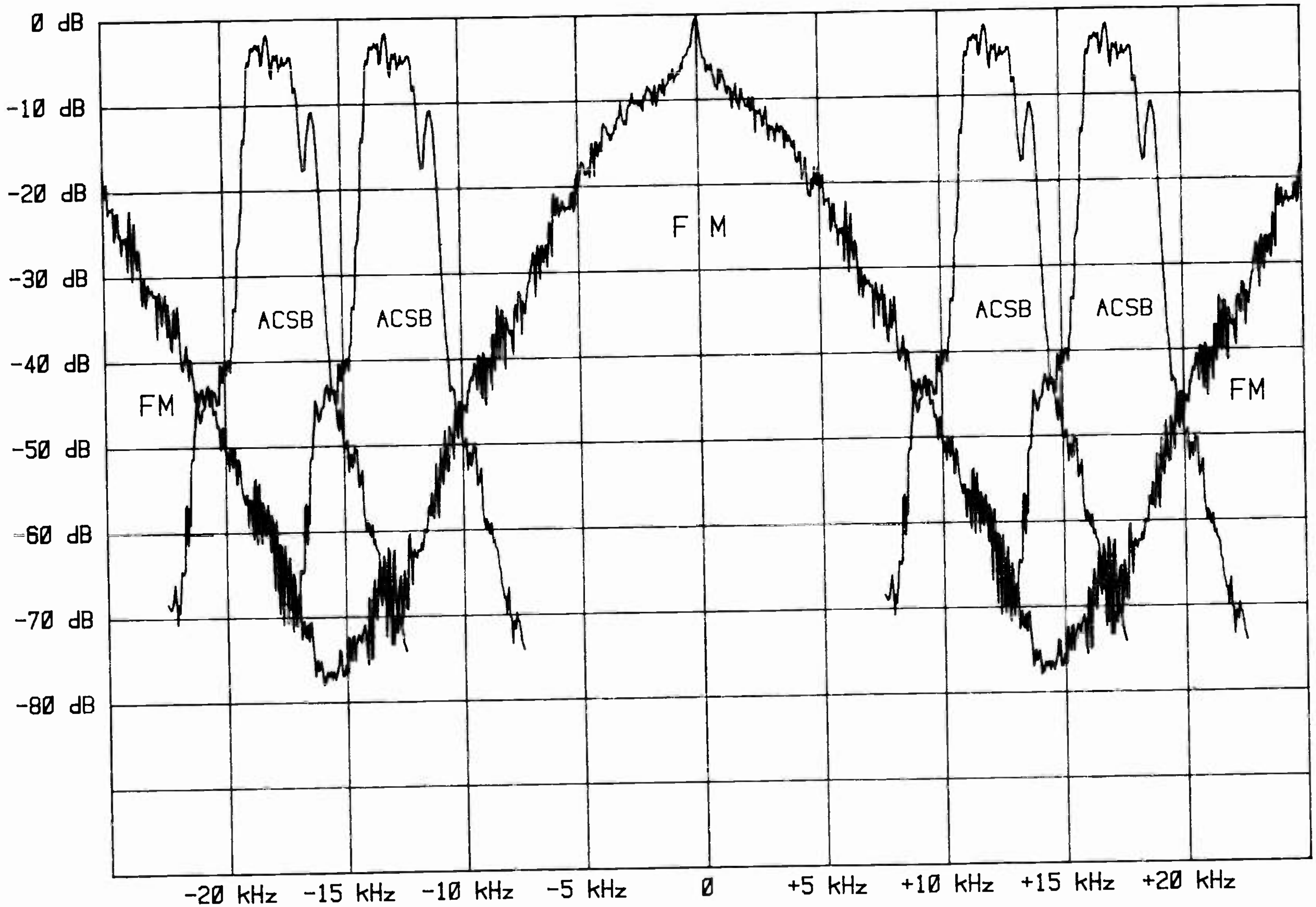


Figure 5
30 kHz spaced FM with 12.5 kHz offset ACSB drop-ins

the frequency coordination problems because coordination is necessary with only remote pickup licensees.

WHY SWITCH TO ACSB?

Licensees can convert existing FM two-way channels to four, five or six ACSB channels to relieve congestion in highly active areas. This can mean the equivalent of new spectrum and less congested channels in the remote pickup service. Certainly, as a minimum, waiting time for occupied channels to become free should be significantly reduced.

The tests at the FCC laboratory indicated that the co-channel interference effect of FM into ACSB was less than the effect of ACSB in FM, especially as the ACSB channels move away from the center of the FM channels. This should give the ACSB user a slight edge in getting the messages through on crowded channels.

ACSB transmitters need not support a continuous carrier, reducing their average power consumption. This can be highly desirable for extended remote broadcasts, particularly when powering the transmitters by batteries.

Also, repeaters can become a reality, even at 150 MHz, because of the "extra" ACSB channels. A repeater is usually located on a high terrain point or on a tower, giving it excellent line-of-sight coverage. The repeater receives weak mobiles or portables on one frequency and repeats the information on another channel. This type of operation extends the effective range of mobile and portable stations.

THE FUTURE

The continuing demand for spectrum must spawn advances in communications technology. In addition to ACSB, other transmission systems such as Linear Predictive Coding (LPC), narrow-band FM and spread spectrum are now being perfected by progressive manufacturers. Currently, ACSB equipment is being manufactured only for the 150 MHz band; however, work by Dr. McGeehan of Bath University, Bath, England, and perhaps others, should soon extend ACSB to at least 900 MHz. Licensees in the remote pickup radio service should be aware of these emerging communications technologies and experiment with them.

It is important to remember that licensees in other radio services also feel the pressures of spectrum crowding. They seek relief from the problem by improving their own transmission techniques and by looking for inefficiently used spectrum that can be shared. Broadcasters should be doing everything possible to lead the way in spectrum efficiency and not just follow the crowd. ACSB can provide a significant step forward in the remote pickup radio service.

(The opinions of the author are not necessarily those of the Federal Communications Commission.)

WIND LOADING AND ITS EFFECT ON DESIGN OF ANTENNAS

Seymour Donner

Edwards & Hjorth

New York, N. Y.

About twenty years ago I had the opportunity to check the design of a guyed tower in the thousand foot range, which was pretty tall in those days. As a result of our investigation we recommended that several elements in the structure be beefed up. The fabricator complained, saying that since the tower was designed by computer how could we possibly find anything wrong, especially since we were checking it with a manual system? The fact that we prevailed is not the point. Not very long after that there occurred a severe series of wind storms in the Pacific Northwest which toppled quite a few towers. Those two incidents together gave me some great insights into tower and antenna design. First, the computer results are only as good as what is inputted. which came to be known as GIGO--Garbage In/Garbage Out (I don't take credit for that name). Second, if the wind doesn't blow the tower won't go--down that is. (At least not from design failure.)

Which leads me to the subject of "Wind Loading and Its Effect on Design of Antennas". Wouldn't life be wonderful if the wind didn't blow; there would be zero effect and most of you could get back to more interesting pursuits.

The fact is that wind load is the <u>only</u> thing that seriously impacts the structural design of antennas. The weight is a factor but would be no more than about 15% of the overall load which includes P-Delta effects which we will talk about later.

For something as basic as wind loading it is somewhat surprising that more is not known about just how it impacts on an antenna. For instance, although we have some fairly good data concerning shape factors of individual elements such as a wind load on a cylindrical shape or perpendicular to or at 45° to a square shape we know a lot less about exactly what happens when one element is directly behind another. What do we use for a shape factor when small elements are attached to an otherwise smooth pipe?

In 1940 a group of three tower manufacturers; Blaw-Knox Company, Truscon Steel Company and Dresser-Ideco formed an organization called the Tower Manufacturers Association to have wind tunnel tests run on various configurations of a trussed tower section at the Daniel Guggenheim Airship Institute in Akron, Ohio. Partly as a result of that research the present Electronics Industry Association has standardized shape factors for square and triangular towers and for round and flat surfaces, but there are no industry standards for the more intricate configurations which make up the various types of antennas.

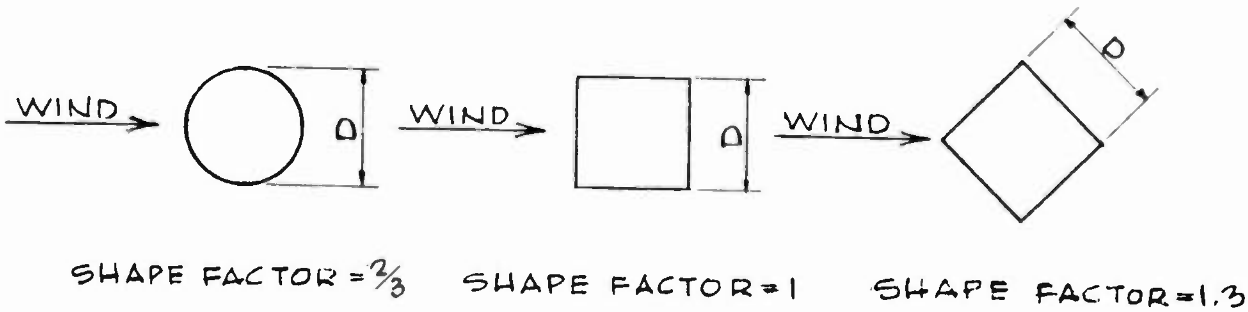
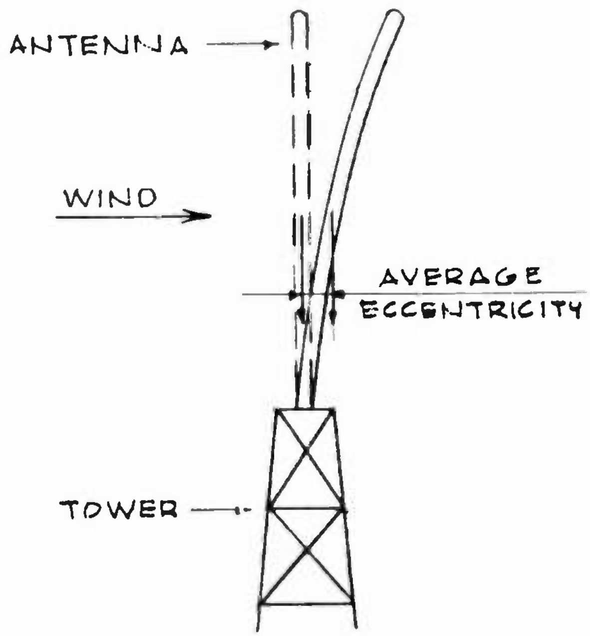


Figure 1

Now that we agree that we should do something about how to determine the wind load on these antennas we can get to the next step, which may be an even bigger problem. That is, what are the <u>effects</u> of the wind load on the antenna?

I already mentioned one which is the P-Delta effect. Picture if you will a typical TV broadcast antenna top-mounted on a tower. We know that when the wind blows the antenna will deflect, and we are very much concerned that the signal strength will decrease on the pattern fringes if it tilts too much. But also note that in this deflected position the weight of the antenna is no longer directly over the base of the antenna. This eccentricity causes a secondary bending stress which, if the antenna deflects sufficiently, could cause a structural failure which might be worse than a temporarily fuzzy picture.



P-DELTA EFFECT

Figure 2

I alluded to another problem, that of attaching small elements to an otherwise smooth pipe. On one hand the rougher surface will increase the coefficient of drag, thereby increasing the wind load on the pipe. On the other hand a smooth pipe can begin to vibrate at certain wind velocities, which is caused by vortex shedding--the Von Karman effect. Since this Von Karman effect can cause a rapid structural failure, care should be taken to prevent exposure of smooth surfaces and accept the minor penalty of increasing the drag on such a surface; on purpose if necessary. That is exactly what was done on the Sears Tower where a 100 foot high cylinder supporting the antennas above had wire rope spiraled around the circumference with the sole purpose of preventing vortex shedding.

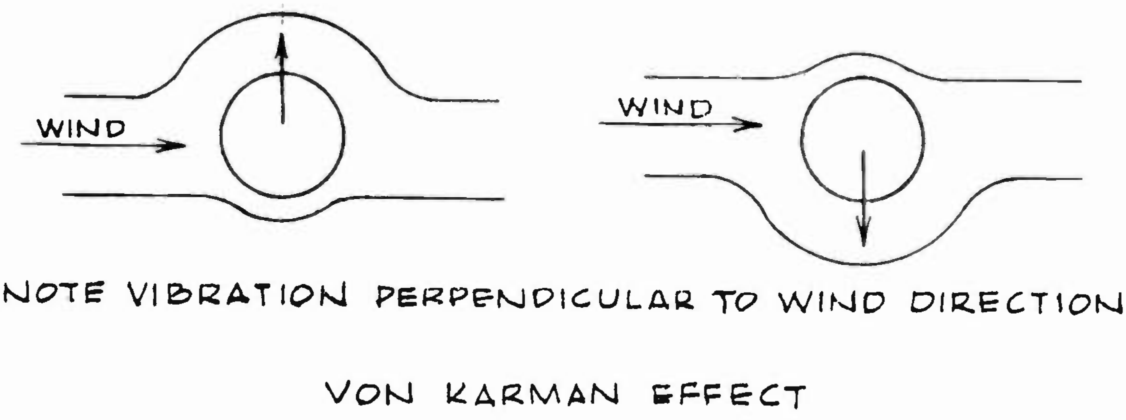


Figure 3

About now you are thinking to yourself that "This is all very interesting, but what does it have to do with me?" Well if you are a broadcaster, especially if you have anything to do with equipment purchase, you hopefully look forward to buying the best equipment available with price always a consideration. So when you are looking into replacing your existing broadcast antenna or building a new facility you make sure that you have your communications consultant review very carefully the electrical characteristics of the competing antennas. At some point your consultant will make field strength surveys, radiation hazard studies and anything else to insure proper broadcast function. If two antennas have equivalent electrical characteristics you will probably take the one that is lighter and has less wind load and therefore possibly cheaper. Such an antenna would have less negative impact on the support structure, which might make that structure less expensive also.

However, when the wind blows you are totally accepting the manufacturer's word that the antenna is structurally adequate. What if the smaller, lighter, less wind-loaded and cheaper antenna was designed with a less conservative approach, or worse, what if it would be overstressed at design loads? Shouldn't at least a structural comparison between the contending antennas be a part of your review procedure if not a full fledged structural analysis? The results of such studies could be good or bad in the short view, but if we consider the life cycle ramifications it can only be viewed as a positive step. First the least expensive antenna may indeed be satisfactory for all design parameters, and if it is also the lightest and has the least wind area that is an additional bonus. But if the reverse is true and the antenna is not properly designed then a potential for failure exists which should override all other considerations including cost.

A properly designed antenna may be more expensive initially and require the support structure to be stronger also which will affect the cost again. But such an antenna with its reduced risk of structural failure or poor reception during high wind periods would eventually be more economical if only because there would be less chance of premature replacement.

There are several possibilities available to reduce the chances of wide variations in structural properties in competing designs. The most basic is for the manufacturers to agree on certain criteria beyond that of simple shape factors for simple shapes. As a start I can offer the following possibilities:

1. Shadowing: When two surfaces of equal size are placed one behind the other in the direction of the wind there should be no reduction in wind area for the far surface if the least dimension is more than four times distant from the near surface. When they are the same distance apart as the least dimension then a factor of one-half for the rear surface is appropriate.

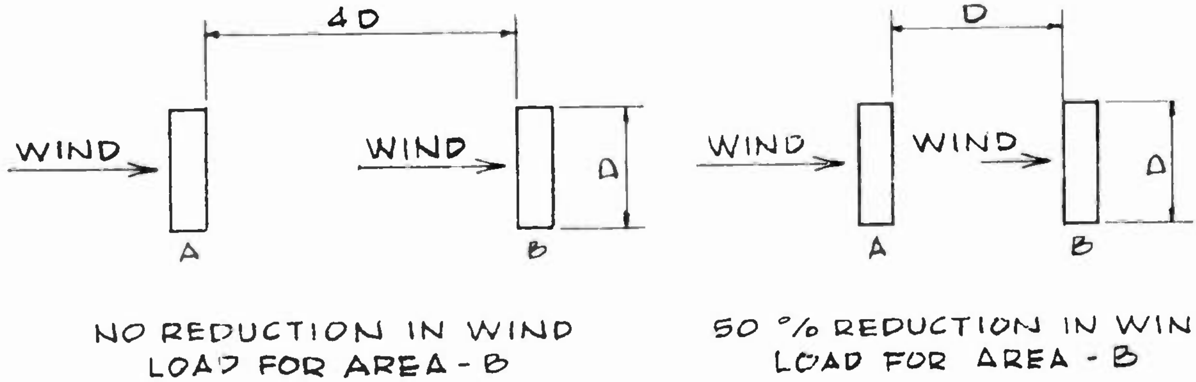
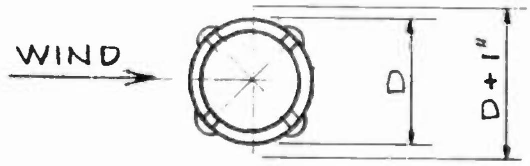


Figure 4

2. P-Delta effects: For top mounted or cantilevered broadcast antennas we suggest that actual stresses will be between 5% and 10% higher due to the effect of deflection.

3. Boundary layer conditions: Where miscellaneous items such as slot covers, pole steps, heater cables, etc. are extant on an otherwise smooth pipe we suggest that at least 1 inch should be added to the actual outside diameter of the pipe to compensate for those areas.



BOUNDARY LAYER CONDITION

Figure 5

Finally if sufficient interest is generated then some additional wind tunnel testing would be in order to confirm and standardize any or all of the factors mentioned.

I believe that by just highlighting some potential problems at this time you would be better served. Each of the factors mentioned can be discussed in much greater detail, but if I have sufficiently piqued your interest to consider the possibilities, that was my intention.

Acknowledgments:

I would like to thank Furman L. Anderson, Kline Iron & Steel Co.; Alan G. Davenport, University of Western Ontario; O. Ben Dov, Radio Corporation of America; and my partner at Edwards & Hjorth, Norman Karger, for their assistance in gathering certain factual information.

Use of Microprocessors for Monitoring, Control, and Diagnostics of Transmitters: What to Expect and What not to Expect

Whit Smith

Harris Corporation

Quincy, Illinois

Abstract

Due to the reliability and low cost of advancing computer technology, microprocessor based control systems are making their debut within broadcast products. With the rising cost of technical labor and the proliferation of unmanned radio stations and remote transmitter sites, computer monitoring, control, and diagnosis of transmitter equipment status has become an economic reality.

However, broadcasters are not sure what to expect and what not to expect from presently available and future systems. This paper discusses these subjects and presents capabilities and limitations of microprocessor based transmitter control systems.

Introduction

Traditional control systems in transmitters consist of relays and motorized potentiometers and variacs with more recently discrete digital IC logic replacing mechanical control. This transition from mechanical to electronic logic has been made with relative ease as the AND and OR gates and counters are easy analogs to series and parallelled relays, motorized potentiometers, and stepping relays. Now the design engineer can add extra features to the transmitter's control scheme without the previously necessary addition of a great deal of electromechanical equipment.

Indeed, the service benefits support this transition. Now all control logic can be placed on the same PC board or boards which already exist to house the power comparing, overload, and resistive divider circuitry. To troubleshoot a problem or replace a part, the technician usually needs only to access the PC board housing the related circuitry instead of having to crawl around inside the transmitter with a VOM checking the condition of power relays and contactors.

However, with the requirements of the next generations of transmitters discrete IC logic will not be able to handle the control requirements demanded by the marketplace. Although with enough discrete logic a design engineer could build almost any system, the cost and maintenance of such systems becomes harrowing. An example of discrete control logic getting out of hand is a dual TV transmitter configuration where multiple visual, aural, emergency multiplex RF paths exist in addition to multiple high voltage power supplies and control points. The number of PC boards, their location, and their internal communication become unwieldly. Once such systems are built, they are hard to modify to conform to customer desires thus driving up the cost of custom configurations and system upgrades.

As was seen in the products introduced at NAB 1983, microprocessor control logic is being introduced into the transmitter equipment of the broadcast industry. Now transmitters can incorporate more sophisticated control systems than previous technology allowed with relatively little additional overhead in the form of costs or maintenance.

Microprocessor Technology

Microprocessor based circuits can perform the same functions as complex discrete control circuits with fewer parts because the function description is stored as a program in a solid state memory instead of as the interconnection of a large number of fixed function ICs. The microprocessor IC itself is a generalized device which follows the specific instructions presented to it by an external or internal memory source. These instructions are followed very quickly. One hundred thousand to four million instructions can typically be executed per second. These instructions can be organized so that when interfaced to other circuitry, the resulting controller can count broomsticks or calculate linearized VSWR readings from forward and reflected RF power levels.

Now that the control functions of a transmitter can be stored in the form a computer program instead of etched into the traces of a PC board, the design engineer has the flexibility to maximize the performance of the transmitter controller given the same transducer inputs and controllable outputs as the previous discrete controller. The same information which allowed the discrete IC controller to display tube plate voltage, tube current, forward and reflected RF powers can now easily be used to automatically lower power into a safe operating area to protect the transmitter's transmission line from high VSWR or a tube from over-dissipation until the problem is corrected. Previously this information was used only to shut down the transmitter.

However, if the designer builds the transmitter to intelligently use the computer controller, the transmitter can do far more. The areas where microprocessor technology will be best seen include transmitter monitoring, control and soon diagnostics.

Monitoring

Most traditional broadcast transmitters are monitored via the use of mechanical D'Arsonival meters physically located on the transmitter and at the station. The local displays are usually several dedicated meters displaying major parameters with a separate multimeter displaying other values selected with a multiple position switch. Station personnel not at the transmitter site are able to view the parameters sent to them via their facilities controller. Typically fewer parameters are remoted than are available at the transmitter site. They are observed only one at a time due to limitations in the STL, telephone bandwidths, or other hardware.

By using microprocessor technology we can escape the limitations of mechanical meters and their switching mechanisms. The computer's transducers can be calibrated to internal voltage references relative to the mechanical meter which uses variable resistors and set screws. Also, the information recorded by the computer is stored as an absolute digital number and not as an analog value. Thus, two digital displays at different locations show the same reading, whereas two different mechanical meters usually show two different values unless they and their communications link are calibrated together frequently. The computer system has the quality of repeatability over mechanical meters in both time and with different hardware.

Typically, D'Arsonival meters display their measured parameters to approximately 2% of full scale accuracy due to mechanical limitations. Using different display technologies, the limitations now become the transducer and analog-to-digital converter resolution and accuracy. An inexpensive 10 bit A/D converter can view a 100% range to a resolution of 1024 parts or to approximately +/-0.1%. Better parts are generally used. (Usually only one such part is used in the hardware and it is shared among all the parameters measured.) These converters are referenced to one source, therefore, they all maintain the same accuracy. Thus the computer controller can provide more accuracy than previous technology allowed.

Through the use of intelligent digital displays such as multisegmented alphanumeric LEDs and LCDs (such as that on the HP-41 series calculators), bargraph displays, or even CRTs, the controller can now display parameters for best human perception. i.e., Filament voltage may best be presented as a 7 segment digital number whereas IPA reflected power would be better as a bargraph to facilitate the tuning process where relative readings and rate of change are important. An alphanumeric display capability gives the controller the ability to present the name of the displayed parameter along with its value. This frees the user from having to reference an isolated number back to its source and allows the designers a freedom to add new parameters without any hardware changes. When more than just a few parameters are measured, computer driven displays are cheaper than traditional displays. With CRTs and other 2 dimensional techniques, many parameters can be simultaneously displayed in any format desired such as bargraph, x-y graphs, or physical diagrams.

An added benefit of using the computer to handle the information includes the ability to more throughly analyze the information and to chronologically catalog the data. Given the DC input and RF output powers of a tube, the efficiency and dissipation can be calculated. These are examples of useful information which cannot be directly measured.

The measured information can also be stored over a period of days and weeks if desired. This would be useful for legal purposes as well as for predicting trends in component life. We have found it useful in Harris hardware to also keep a short term history of relevant transmitter parameters. This is useful in the cases of previously unexplained overloads and faults. Now we can look at a snapshot of transmitter conditions immediately preceeding the fault. In many cases, this has helped us to determine the causes of such events.

Control

Traditional control is in the form of Resistor/Capacitor timing circuits which flip contactors in hardware determined chronological sequences and op-amp type analog feedback loops for such systems as automatic power control. The only other control inputs came from external (typically human) sources. With the use of microprocessor technology, we can expand on this.

No longer must timing sequences depend on the accuracy of the RC time constants. Now the crystal controlled clock of the processor circuit can be employed to control the time between step/start contactor actuation, filament warm-up and rundown times, and any other time related functions. Each timing function no longer needs an independently tweaked and thermally sensitive resistive and capacitive element. Crystal accuracies are typically on the order of several parts per million. The high count frequencies allow for microsecond resolutions. This saves money for both the manufacturer in terms of components and initial calibration and the broadcaster by not providing another knob which needs to be adjusted.

The intelligence of the system can also be employed to go beyond simple sequencing. In the case of step/starting a power supply, the charging current or voltage across a capacitor can be used as a feedback mechanism to indicate when to close the main contactors and then remove the starter resistors. A software timer could be used as a safety backup to protect the starter hardware from a shorted power supply circuit. As was stated previously, output power can be raised and lowered to maintain the tube to within its safe operating area given the mentioned parameters as input data. Thus the transmitter can fail softly and not go completely off the air in case of any type of VSWR mismatch such as antenna icing.

The mechanisms of control include voltage controlled power supplies and regulators, motorized variacs, resistors, and reactive components, and relays. More solid state parts are becoming available to replace older mechanical methods. An example is the PIN diode RF switcher which has replaced RF relays for multiple exciter switching. Phase controlled triacs are also showing up in power supply circuits which previously used variacs and hand switched multiple-tap transformers. These electronically controlled devices are becoming cheaper and more reliable than their electromechanical counterparts, and they respond much faster.

The same sorts of displays which show the metered parameters can display the operation of the transmitter's control functions. For instance the user can see when a contactor has been activated. If the designer is clever, no auxiliary contacts for feedback from the contactor are necessary because the controller should be able to deduce a closure from the presence of voltage in another part of the circuit and indicate successful contact closure (or else indicate a fault).

Diagnostics

All that has been discussed so far is merely an extension of traditional transmitter control. The use of computer technology allows us to go one step further. We now have the capabilities to diagnose some of the problems which occur in transmitter hardware.

Many manufacturers of broadcast and other electronic equipment include fault tree diagrams with their product instruction books to aid technical personnel in identifying equipment problems. The computer can use the information it receives to passively proceed through a fault tree diagram residing in its software which can specify a fault or at least narrow down the possibilities. The transmitter can now inform station personnel of a possible problem and what it thinks the problem may be. The short history saved prior to shut down enables the diagnostic software to analyze not only the static readings but to look at the transmitter as it makes state transitions.

Active diagnostics are extensions of computer self-test methods. They require more monitoring transducers and the ability of the transmitter to exercise control functions as part of a test mode. This includes the "flexing" of contactors and motorized components to see if they are working properly. Well defined test signals such as TV's VITS or multiple tones for radio can be injected into the input of the system. Monitoring these signals throughout the RF chain with the appropriate demodulators and transducers allows the controller to diagnose the health of each stage. Such tests can be run extensively during station down times or continually as with the TV VITS which can be injected during each vertical retrace.

Although not yet implemented in any commercial broadcast equipment, this data could be used to automatically tune the RF stages of exciters and amplifiers and to dynamically bring the devices into specification. This could be to compensate for some failure or merely as part of the usual maintenance.

Historical diagnostics use the information gathered after long periods of time to indicate detrimental trends. The changing V-I characterisitics of a tube filament can be used to predict tube life. The power requirements of an ion pump for a klystron tube can tell the UHF operator about the quality of his vacuum seals.

The limitations of a transmitters' diagnostic capabilities lie in what it can measure, what it can control, and how well the designers have defined the possible faults. These are the manufacturers responsibilities. If the transmitter is not designed with diagnostics in mind then the controller will not be able to be very specific about the problems it may encounter. However, if a good job has been done, the technical personnel can drive to the transmitter site knowing what parts and test gear to carry. The extra capital cost incurred by the broadcaster for the diagnostic capability will pay for itself in terms of maintenance and technical personnel time.

Serial Remote

Another unheralded benefit of microprocessor control technology involves the use of serial communications relative to a dedicated wire or channel per parameter. The main benefit is that all information available to one end of a serial link may be sent to the other end over a single channel. This applies to hardware expansion at the transmitter site and to traditional remote control applications.

Serial communications involves the sending and receiving of digital codes in a time sequence recognizable by both the sending and receiving hardware over a single channel. The speed of transmission is a function of the channel bandwidth. A 300 bit/second channel can pass about 30 characters of text or update several metering channels per second. Typical methods use two different audio frequencies for voice grade bandwidth telephone channels (300 bits/sec), phase shift keying for higher speed channels requiring a carrier (9600 bits/sec over telephone leased lines), or voltage level changes on dedicated local lines (1M bits/sec). Algorithms exist which provide very good methods for error detection and correction (or retransmission requests). Examples of serial communication application include RTTY, computer terminals, and infared television remote controls.

At the transmitter site, this allows the controllers within multiple transmitter configurations to communicate with a minimum of external wiring relative to the bulky wiring harnesses which are presently used. Serial communications will soon be used within the transmitter itself to communicate commands and status between different parts of the transmitter. Again, this lowers wiring costs.

Serial communications between the studio and transmitter site allows studio personnel access to all information available to the transmitter's internal controller. Gone is the wire per channel limitation previously imposed. Instead of dedicating many channels of a STL to the status of discrete functions, one channel may carry all the information, and the controllers at each end will organize this information for display to interested parties. Present facilities controls use serial communications, and the newer transmitters can also use direct serial links.

This also offers the broadcaster an opportunity to contract out a new form of maintenance services. The parties providing this service contact the transmitter site at regular intervals and inquire as to the health of the hardware. If problems are diagnosed, they are reported to the station management or directly repaired as part of a service agreement. This is presently done with fixed aircraft navigation beacons on a worldwide scale.

The link possibilities are now also expanded. Commercially available modems exist for use over the public telephone network for data rates from 300 to 9600 bits/sec. Other media possibilities include a terrestrial microwave STL's channel, a transmitter SCA channel for one way status monitoring, or direct satellite networks. The use of the SCA channel can allow station personnel to monitor the status of their equipment from anywhere within the station's listening area with the use of a receiver, modem, and a personnel computer.

Maintainance Considerations

The resistance to using microprocessor based controls include the cost for any additional transducers and control devices, a different type of maintenance, and different failure modes which can be associated with the psychological barrier that has been presented by the industry to the use of microprocessor based controllers.

As was mentioned previously, traditional control and monitoring can be accomplished with the use of transducers and control circuits which cost about the same as similar discrete logic controllers. However, to achieve a reasonable degree of diagnostic capability, much more expensive hardware will need to be built into these transmitters. The industry has not yet taken that step.

If properly built, the microprocessor control circuits will have reliability figures many magnitudes higher than the rest of the transmitter. They do not draw much power, and they do not get hot. (In General Motor's repair manuals for late model cars, the authors tell the mechanics to try replacing everything before replacing the microprocessor based car control unit. They rarely fail even in an automotive environment.)

The people who understand RF hardware do not usually understand computer hardware. This is also true of the manufacturers. Traditionally, a technician fixes a transmitter with the use of a VOM and an oscilloscope. Few stations have much more equipment available. With relays or discrete digital logic, the technician has a chance of tracing a pushbutton depression through the control circuits to the final result of his action with his scope and meter. However, with computer controls, all signals are routed through the microprocessor's circuits which do not have fixed function paths. Conventional test equipment does not tell its user anything useful. Therefore, anxiety concerning owning microprocessor based equipment occurs. This does not need to be the case.

First, the controller will probably never break. Second, if it partially fails so that the self test software can still operate, the controller should try to tell you about its situation and guide service personnel through the repair procedure. However, if the core of the processor's circuits are gone, the technician will have to determine on his own which parts to replace to get the transmitter back on the air.

There are many ways to fix a broken microprocessor controller. Methods include board swapping and kernel isolation at the circuit level. The controller can also be treated as another black box and replaced like a car control unit.

Before buying microprocessor based transmitter hardware, the broadcaster should make sure that the equipment manufacturer fully supports his product. This support should include thorough and readable technical manuals, the availability of technical training for station maintenance personnel, and field service support. One should also check the hardware for ease of maintenance. The location of various controller components should be well defined and in an easy to reach place to facilitate board swapping.

Conclusion

Microprocessor based control circuits have arrived. They pose some new maintenance problems, however, with their reliability and benefits they will become universally used. They offer better control, monitoring, diagnostic, and communication capabilities than their predecessors, and their use facilitates the economics of automated stations and the rising costs of technical labor.

Thanks to:

D.L. Dickey
D.L. Grimes
G.M. Sanderson

of Harris Broadcast
Quincy Illinois

Digital Transmission of High-Quality Sound Programs - A Practical Approach

Chieu Nguyen, Kevin Flynn, Glenn Muth

M/A-COM DCC, INC.

Germantown, Maryland

INTRODUCTION TO DIGITAL TRANSMISSION

The distribution of network programming directly to affiliates via satellite is among the most attractive technical innovations available today. Satellite distribution can lower the networks' operating costs and simultaneously enable the networks to expand their affiliate base. This will, in turn, appeal to the advertisers who support the networks.

In addition, satellite digital program distribution offers flexibility through freedom from geographical constraints, a variety of program feeds, and very high audio quality. Program distribution needs no longer be constrained by the routing and directionality of the present system of landlines. Any affiliate is thus able to receive any network feed, regardless of location. This makes practical the idea of configuring subnetworks to carry sports, weather, news, and other programs of regional interest. Nationwide affiliate subnetworks targeted for special-interest audiences will also be feasible. In addition, it will be possible to target specific program segments to particular affiliates, individually. A nearly instantaneous reconfiguration capability can increase the flexibility of network operations dramatically.

Satellite distribution also makes possible a virtual explosion in the amount of network programming that is available. For the first time, the networks will be able to distribute multiple program channels -- including long-form programming - simultaneously.

Four system architectures are possible for the satellite distribution system:[1]

- Time Division Multiple Access (TDMA), single carrier at approximately 8 Mbps

- Time Division Multiplex (TDM), single carrier at 8.78 Mbps[2]

- Frequency Division Multiple Access (FDMA), multiple carriers, 1.544 Mbps each

- Single Carrier Per Channel (SCPC), multiple carriers at 64, 192, and 384 kbps.

This paper presents the FDMA multiple carrier system, the M/AESTRO (M/A-COM Earth STation Receive Only) system, which has been developed based upon 3-meter digital audio receive-only earth stations and T1 (1.544 Mbps) transmission format.

The important advantages of the M/AESTRO FDMA system are summarized below:[3]

- First, several uplink (transmitting) sites can access the same transponder. This allows, for example, a network to transmit directly from New York and/or Los Angeles without having to backhaul across the country. The affiliates can select from either carrier signal by simple thumbwheel or remote control, or carriers can alternately occupy the same frequency at different times.

- Second, since each T1 carrier requires only 20% of the satellite transponder power, network implementation and subsequent growth is handled incrementally in T1 steps. The advantages are very clear: a full transponder need not be utilized to handle a minimal number of initial program channels; only 20% would be required for four 15 kHz audio channels. Network growth is accommodated through the use of additional T1 signals, and the transponder usage (and its related cost) increases only as needed.

- Third, the interface between the satellite receiver and the audio demultiplexer is a digital T1 data stream. Thus, the earth station may be located separately from the studio facilities without any loss of audio performance, since the connection is done with standard T1 interfaces. The connection may be a simple twisted pair cable run over several thousand feet, or longer distance T1 facilities available from the local telephone company. Several studio locations or alternate network users can also share the same earth station and still benefit from the inherent quality of digital audio.

However, models of growth for the radio networks all but favor the incremental deployments of audio channels in T1 steps.[1] And every indication from the present situations of the major networks' TDM single carrier system has shown that the growths can be better fulfilled by the FDMA T1 approach.

RECEIVE-ONLY CONFIGURATION

The configuration of the receive-only earth station consists of the following subsystems:

- 3-meter Antenna.

- 120° K Low-Noise Amplifier (LNA).

- Downconverter to convert the received 4 GHz signal down to 70 MHz.

- QPSK Demodulator(s) and Forward Error Correction (FEC) Decoder(s) to demodulate the 70 MHz signal(s) to digital T1 signals. Rate 7/8 FEC is used.

- Optional Channel Select Oscillator (CSO) for agility in selecting the transmitted T1 channels.

- The resulting T1 signal(s) are sent to the Digital Program Terminal(s) to be demultiplexed and decoded back to audio programs.

TABLE 1

SUMMARY OF RF/IF SPECIFICATIONS

ITEM	VALUE
Channel Error Rate (a)	8×10^{-3}
Link Margin	2.4 dB
Information Bit Rate	1.544 Mbps
Information Error Rate (Minimum) (b)	1×10^{-5}
Information Error Rate (Typical)	Between 1×10^{-8} and 1×10^{-7}
Modulation	QPSK
Coding (Convolutional Code)	Rate 7/8, soft-decision

NOTE: For 5-meter antennas

(a) Reduces to 2.7×10^{-2}
(b) Improves to 1×10^{-8}

DIGITAL PROGRAM TERMINAL (DPT) (Figure 1)

Keypoints of Design

- Digital Interface to the RF/IF equipment. The interface conforms to T1 (1.544 Mbps) format as specified in AT&T Technical Advisory no. 34, no. 69, (Clear Channel Capability) and no. 70 (Extended Frame Format), as well as all compatibility requirements of Publication 43806 and Bulletin no. 143.

- Capacity of receiving up to four T1 inputs. Reference 3 describes channel capacity per one T1 input.

- Capability of selecting any audio channel from any of the four T1 inputs.
- Human Engineering features to ease the task of installing, operating, and maintaining the Terminals.
- Audio Performance is summarized below:

AUDIO PERFORMANCE

Full-load level	$\pm$21 dBm
THD	less than 0.3%
THD (+6 dB OVERLOAD with AGC)	less than 0.3%
Idle Noise	90 dB below full load
Gain match	within 0.3 dB
Phase match	within 4 degrees
Crosstalk	less than -75 dB
FREQUENCY RESPONSE	$\pm$0.3 dB (7.5 kHz/15 kHz Bandwidth)

General Descriptions (Figure 1)

Transmit Digital Program Terminal (TX DPT)

A block diagram of the TX DPT is shown in Figure 2(a). Audio inputs are converted into Pulse Code Modulation (PCM) digital format at the rate of 32 kilo samples/second for 15 kHz channels or 16 kilo samples/second for 7.5 kHz channels by the Program Audio Encoders. The digital PCM outputs of the Encoders are summed by the Multiplexer into one DS1 output.

Receive Digital Program Terminal

The Receive Digital Program Terminal can receive up to 4 DS1 inputs, up to 16 15 kHz channels or up to 32 7.5 kHz channels or combinations of the two types of channels. As shown in Figure 2(b), the synchronized 1.544 MHz timing and data buses from four Demultiplexers are common to all Program Audio Decoders, and selections of the proper audio channels are done at the individual Decoders.

DIGITAL MULTIPLEXING, DEMULTIPLEXING

T1 (1.544 Mbps) Bipolar Interface

The bipolar interface at T1 (1.544 Mbps) digital cross-connect is normally alternate mark inversion (AMI) signals. To provide clear channel capability, bipolar eight zero substitution (B8ZS) coding and decoding are as an option employed as shown in Figure 3(a). The B8ZS encoder is a state machine with the encoding sequence shown in Figure 3(b). Decoding must be transparent to the normal AMI signals, be able to detect the B8ZS automatically and must minimize multiplication effects of decoding under high T1 line bit errors.

Frame Structure and Synchronization Strategy

The T1 Extended Frame Format consists of 24 frames in a multiframe. A frame bit pattern of 001110 is inserted in frames 4, 8, 12, 16, 20, and 24.

To synchronize to the T1 frame and to correctly demultiplex the audio channels at the receive end, the synchronizer is designed as a state machine[4] with its sequences shown in Figure 4.

Synchronization Performance Analyses

In-Sync Performance

For m = 6 bits in the frame pattern, and n = 1 bit error tolerance in the synchronize's pattern matching as shown in its strategy, the misframe interval T_M is given as

$$T_M = \left\{ \frac{(m-n-1)!}{m!} \right\}^2 \cdot \left\{ P_e \right\}^{-2(n+1)} \cdot \left\{ \text{bit period} \right\}$$

At $P_e = 10^{-3}$ (RF/IF equipment dropout threshold), T_M is approximately once every two minutes.

At $P_e = 10^{-5}$ (minimum RF/IF equipment operating point), T_M is once every 55 thousand years.

At $P_e = 5 \times 10^{-7}$ (typical RF/IF equipment operating point), T_M is once every 800 million years.

Out-of-Sync Performance

The maximum average frame synchronization time is estimated as follows:

$$T_F(\max) = 4L + \left(2^{-m}\right) .2L + \left(2^{-m}\right) \cdot \left[(1 + p_o) 2L/p_o^{\,2} \right]$$

where $p_o = 1-2^{-m} \sum_{k=o}^{n} \binom{m}{n}$, the probability of m-bit code whose Hamming distance exceeds n

L = 3 millisecond, multiframe period

Note the effects of line errors are not included since these are negligibly small, and the sync verification time of two frames must be gone through every time before a temporary sync condition is declared. The first term of 4L represents the number of times multiframe synchronization must be declared before final sync is confirmed. T_F is approximately 400 millisec maximum.

ENCODING AND DECODING

The input audio is low-pass filtered then sampled and held for digitization by the analog-to-digital converter (A/D). At the decoder, the digital data is latched into a digital-to-analog (D/A) converter, and then sampled and held to remove glitches. Another low-pass filtering takes place, and then the audio is amplified for output. In the sections that follow, each important design criterion will be discussed, and the actual performance evaluated.

Sampling Frequency

Shannon's fundamental sampling theory[5] demonstrates that any time function that is band-limited to B Hz can be uniquely represented by its amplitude samples occurring every 1/2B seconds apart. The sampling rate for a digital audio system is therefore determined by the highest frequency that one desires to transmit. Many studies[6] have conclusively shown that, for music, the upper frequency limit beyond which no discernible improvement occurs is about 15 kHz. Theoretically then a 30 kHz sampling rate would work, however to compensate for the non-ideal response of real world filters, a sampling rate f_s of 32 kHz is used.

Anti-Aliasing Filter

Frequencies above $^{fs}/_2$ that are sampled produce distortion products in the pass-band that, once introduced, cannot be eliminated. This effect, called aliasing, occurs because the input signal is modulated by the sampling process, producing a spectrum that repeats at all multiples of f_s and overlapping occurs. The anti-aliasing filter used for M/AESTRO has a flat response up to 15 kHz, and provides more than 60 dB of attenuation for frequencies above 16 kHz. While theoretically this may seem barely adequate,[7] the fact that music and speech power levels at high frequencies are typically 10 to 20 dB lower than the levels in the base and midrange regions,[8] ensures that any aliasing distortion is inaudible.

Dynamic Range and Companding

The dynamic range of audio is considered as the difference of the maximum and minimum levels of the program material. Studies have shown that the natural dynamic range of music[9] and the human capability to detect tones at levels well below noise levels[10] require a dynamic range in excess of 100 dB in any reproduction-transmission system to faithfully represent the analog program.

The signal-to-quantization distortion ratio (S/D) for a full-load sinusoidal waveform is given as:[11]

$$\frac{S}{D} = 6.02\ n + 1.8\ \text{dB}$$

where n is the number of bits in the A/D.

Thus a resolution in excess of the present state-of-the-art 16-bit converters would be needed to fulfill the above dynamic range requirements. However, this would be prohibitively expensive. Converters with 16-bit resolution are used in M/AESTRO and, because of digital companding, only 15 bits are utilized, providing an idle noise level lower than 90 dB below full-load signals.

The quantizing error inherent in PCM systems gives rise to a flat wideband audio noise the amplitude of which changes with program levels. However, due to masking effects, typically noise and distortion cannot be heard if the simultaneous signal is 50 to 60 dB greater[12]. Therefore, full 16- or 15-bit resolution is not necessary at all times. Digital instantaneous companding provides a solution to reduce the transmission bit rate and still maintain the resolution of the linear A/D conversion.

The most popular companding law in widespread use is logarithmic compression. Other companders, i.e., hyberbolic companding, can provide improvement at certain input levels, but are too specialized for use with the broad volume range of program material experienced in program sound distribution systems. Logarithmic compression has the advantage that it produces much flatter signal-to-quantization distortion (S/D) over a wider dynamic range. In addition, logarithmic compression is suitable for digital transmission because it can be easily approximated by piecewise linear segments using digital circuitry.

As shown in Figure 5, a 13-segment A-Law companding is used in the M/AESTRO system to provide S/D of 53 to 56 dB over a 40 dB range. For levels below -40 dB the linear 15-bit resolution is retained. Although the distortion is increased (to approximately 0.2% in comparison to the 0.0012% level of full-load linear 16-bit conversion), it is not audible[13] and low levels maintain the 15-bit linear performance. Actual performance is shown for the 15-bit linear and 15 to 11 bit compressed sinusoidal waveforms in Figures 6(a) and 6(b) respectively.

Automatic Level Control

Input levels that exceed the 90-plus dB range occasionally occur and will be clipped by the A/D. Rapid increases in distortion levels are shown in Figure 7(a) for overload conditions (4- and 8-bit resolutions shown).[14] Moreover, since clipping occurs after the anti-aliasing filter stage, the high-order harmonics will be "folded" back in the passband upon decoding, causing much higher aliasing components.

To combat these conditions, a novel digital automatic gain control (AGC) is used. It is normally transparent to the audio path, and only reacts to peaks above full-load level, bringing them down instantaneously. The block diagram of the digital AGC is shown in Figure 7(b).

Audio levels up to 6 dB above full-load levels are digitized into 5 bits. A programmable read-only memory (PROM) examines these 5-bit overload levels and controls the 5-bit setting of the digitally-controlled attenuator. A decay clock frequency is also fed into the PROM to control the decay rate.

The algorithm is as follows:

PROM Input	PROM Output
$OV > \alpha p$	αn
$OV = \alpha p$	αp
$OV < \alpha p, \phi = 0$	αp
$OV < \alpha p, \phi = 1$	$\alpha p-1$

OV = overload level (5-bit word)
αp = present attenuation level (5-bit word)
αn = new attenuation level corresponding to OV
ϕ = decay clock

The actual performance of the AGC is illustrated in Figures 8(a) and 8(b). A 1 kHz sine wave burst is shown at the Decoder output with and without the AGC. Note that the correct gain is attained within 1 cycle, completely eliminating the clipping. The digital design has many advantages over existing analog AGC designs in areas of instantaneous attack without overshoot, programmable decay and characteristics, and does not suffer the inherent distortion and "breathing" effects of analog AGC counterparts.

Error Mitigation

The audible effects of digital bit errors on PCM transmission have long been recognized and investigated.[15] Error coding and correction can be employed in the RF transmission to reduce bit errors. However, errors cannot be totally eliminated (or corrected) when the operating bit error rate is high, and if occurring in the more significant bits, errors can produce very loud clicks that can be quite distracting. Efforts have been made to investigate several methods of reducing the audible effects of bit errors.[16],[17],[18] All of these involve inserting parity check bits on the digital sample at the transmit side to detect errors at the receive side. Major conclusions are summarized below:

a) For audio distribution, complete correction of transmission errors is not essential,[16] as penalties would have to be paid to implement a complex system to correct all errors. Also, additional bits are required to be transmitted per sample.

b) Error detection can be achieved by single-bit parity checking on seven of the most significant bits when the transmission bit error rate is 10^{-6} or better, since the probability of double errors in one sample is more than two orders of magnitude below the transmission error rate.[16]

c) Muting is one way to conceal digit errors; however, it can cause other audible effects if the attenuation time-constant is not properly designed.

d) Sample reuse (or zero-order extrapolation) involves replacing the erroneous digital sample by the previous sample. Experiments[17] have shown that sample reuse can raise the "perceptible" threshold of impairments due to bit errors by two orders of magnitude, i.e., from 10^{-7} to 10^{-5}.

Per the above discussions, a combination of parity check on seven most significant bits and sample reuse was selected as the most cost-effective way of concealing the effects of transmission errors. Note that at high frequencies, sample reuse may lose its effectiveness when compared to other schemes.[19] However, the simplicity and field-proven performance of sample reuse make it the most attractive and cost effective error mitigation technique.

Decoding

To eliminate turn-on and turn-off transients that can be devastating to the audio equipment connected to the M/AESTRO system, several circuits are included:

- Upon loss of synchronization, the audio decoder digitally decays to a very low PCM level, thus gradually muting the reconstructed audio output to provide a "soft" turnoff.

- Upon synchronization, the Demultiplexer unmutes the audio channels only when their respective PCM levels are very low, insuring a "soft" turnon.

- Upon power up, the decoder delays its audio output for about 5 seconds to avoid power transients.

Equalization for the frequency response distortion in nonimpulse sampling is achieved per the following equation.[11]

$$G(f) = \tau \frac{\sin \pi f \tau}{\pi f \tau}$$

The end-to-end frequency response is shown in Figure 9.

EVOLUTION FROM ANALOG TRANSMISSION

Several radio networks already use analog single-channel-per-carrier transmission that has occasionally performed poorly. The M/AESTRO equipment is compatible with most 3-meter receive dishes, and is offered as an upgrade to these systems to improve general performance and add capacity. In the same 22 dBw of satellite effective radiated power needed to support two analog FM SCPC signals for stereo, a digital T1 carrier will provide four 15 kHz circuits. Moreover, the audio performance of the four 15 kHz circuits is not degraded by transmission impairments such as multipath and cross-talk, and is generally superior to the equivalent analog transmission.

Another obvious advantage is that the T1 carrier can coexist in the same message traffic transponder with the analog FM signals. This means that space segment need not be repurchased from a new satellite vendor and the existing space can continue to be used.

VERSATILITY OF THE DIGITAL PROGRAM TERMINAL

Terrestrial and Fiber Optic Transmission

The link between the TX and RX DPTs does not need to be satellite transmission. Any means of transmission compatible to DS1 format can be used, such as microwave radio, metallic line transmission, etc. Fiber optic transmission typically carries large bandwidth data, the standard M1-2 and M1-3 digital multiplexers for the digital cross-connects can be used to pipeline the DPT transmissions into the fiber links.

Magnetic Recording of the DPT Digital Outputs

The 1.544 Mbps TTL level CLOCK and DATA outputs of the TX DPT can be used to interface to high-speed instrumentation recorders to provide a recording medium of the digital audio, hence making digital storage of any multichannel audio source possible.

Digital Audio Distribution within a Facility

Within one studio facility, long-cable transmissions of signals for distribution, recording, and monitoring purposes can be drastically deteriorated due to the analog nature of the transmission. The TX and RX DPTs can be utilized as an all digital audio distribution between various locations (sound room, muxing studio, recording room, radio tower, etc.) with good results since all cable connections carry the digital DS1 signals, as shown in Figure 10(a).

Remote Recordings or Live Transmissions of Live Events

Mobile mixing and recording equipment is expensive and may not be available for all live events (sports, concerts, etc.). The DPT system provides a means of recording or transmitting these live events using the existing sophisticated equipment at the studio. As shown in Figure 10(b), all audio sources are digitized by the TX DPTs (rough mixing and pre-equalization can be applied as needed), the DPT(s) then transmit via digital microwave or metallic lines to the studio. The RX DPTs then convert the transmitted data back to analog waveforms, for which final muxing and recordings (or live radio transmissions) can be applied.

ACKNOWLEDGMENTS

The authors wish to thank the Engineering Staff at AT&T Long Lines and members of the Technical Staff at Bell Laboratories for their constructive critique during the development of the M/AESTRO system.

REFERENCES

[1] A. Eklof, "Study of the operational and economic feasibility of satellite distribution for radio network programming," M/A-COM DCC Internal Study, January 1981.

[2] A. McBride, et. al., "Satellite distributed digital audio for network audio," NAB Conference, April 1982.

[3] S. Salamoff, "Digital audio distribution - A practical approach," BME, November 1982.

[4] R. Maruta, "A simple firmware realization of PCM framing systems," IEEE transaction on communications, August 1980, COM-28, No. 8, p. 1228.

[5] C. Shannon, B. Oliver, T. Pierce, "The philosophy of PCM," Proceedings of the IRE, Vol. 36, No. 11, November 1948.

[6] T. Muraoka, M. Iwahara, Y. Yamada, "Examination of audio-bandwidth requirements for optimum sound signal transmission," JAES, Vol. 29, No. 1/2, January/February 1981.

[7] A. Papoulis, "Error analysis in sampling theory," IEEE Proceedings, Vol. 54, No. 7, July 1966.

[8] D. Meares, K. Lansdowne, "Revised background noise criteria for broadcasting studios," BBC Research, August 1980.

[9] D. Meares, "Statistics of typical programme sound pressure levels in sound studios and their control rooms," BBC Research No. 1967/57.

[10] CCITT Recommendations 505 and H68.

[11] Bell Laboratories, "Transmission system for communications."

[12] Bos, DeBoer, "Masking and discrimination," JAES, Vol. 39, No. 4, 1966.

[13] Petri-Larmi, Otala, Leinonen, Lammasniemi, "Audibility of transient intermodulation distortion," IEEE Transaction ASSP, Vol. 28, No. 1, February 1980.

[14] Robert Gagliardi, "Introduction to communications engineering," Wiley & Sons, 1978.

[15] I. Dostis, "The effects of digital errors on PCM transmission of companded speech," BSTJ, December 1965, pp. 2227-2243.

[16] BBC Research Report 1972/18, "Pulse-code modulation for high-quality sound-sinal distribution: protection against digit errors."

[17] BBC Research Report 1971/37, "Pulse code modulation for high-quality sound-signal distribution: simulation of an advanced form of error protection."

[18] R. Steele, D. Goodman, "Detection and selective smoothing of transmission errors in linear PCM," BSTJ, Vol. 56, No. 3, March 1977.

[19] K. Tazaki, H. Kaneko, Y. Katagiu, "Realization and system design of a digital terminal equipment for sound program signal," presented at INTELCOM 1977, Atlanta, GA.

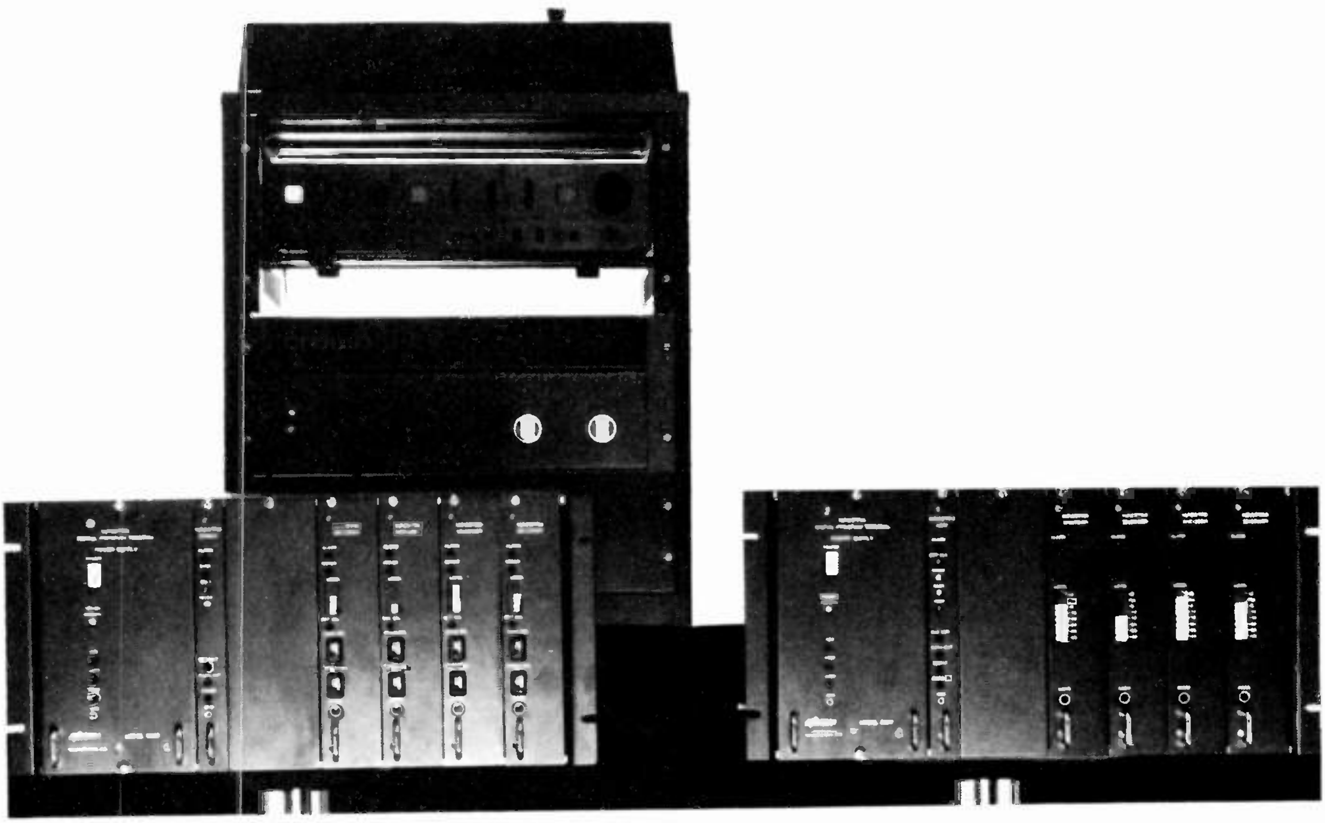
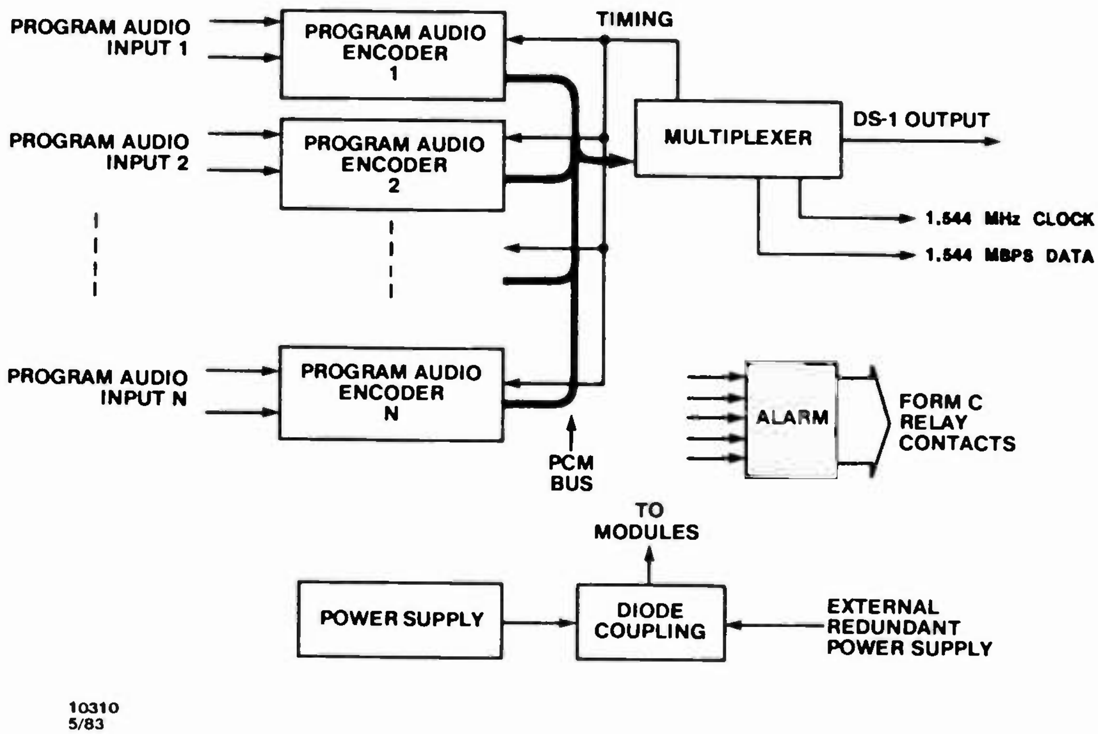


Figure 1. M/AESTRO Transmit & Receive Program Terminals



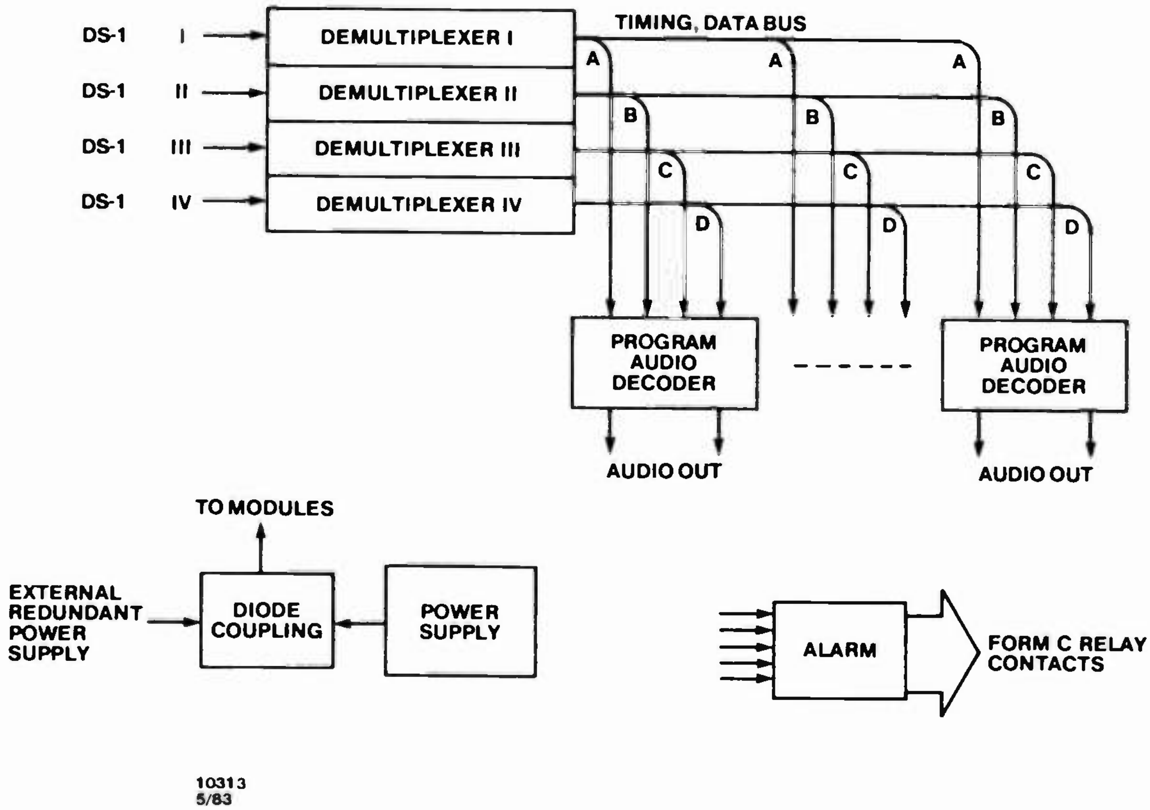
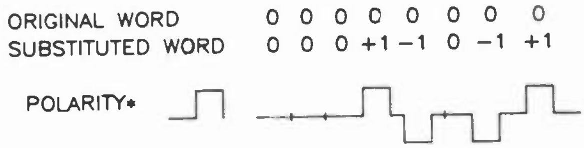
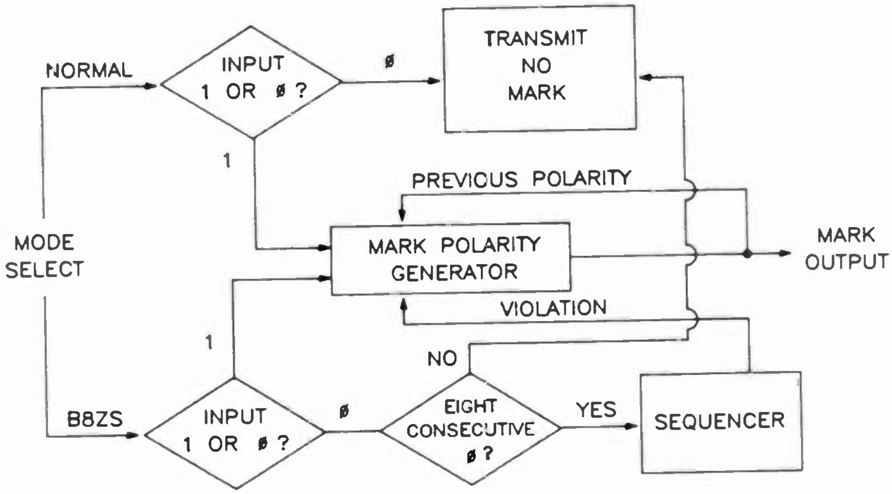


Figure 2. The M/AESTRO Digital Block Diagrams



*ASSUME LAST "ONE" TRANSMITTED WAS A POSITIVE "ONE". POLARITY IS REVERSED IF THE LAST "ONE" TRANSMITTED WAS A NEGATIVE "ONE". BIPOLAR VIOLATIONS ALWAYS OCCUR AT THE 4th AND 7th BIT POSITIONS OF THE SUBSTITUTED WORD.

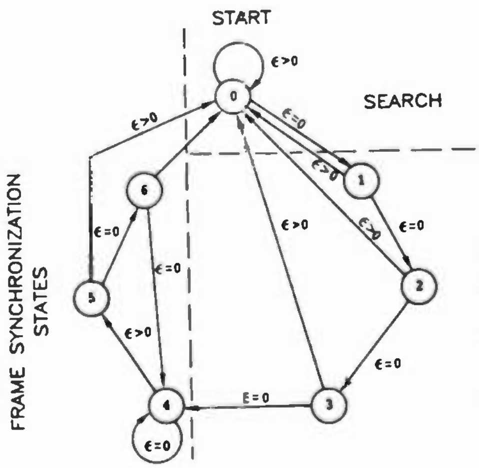
a) SAMPLE OF B8ZS CODE



b) ENCODING SEQUENCE

13949V
2/17/84

Figure 3. B8ZS Encoding



ε = NUMBER OF MISMATCH BETWEEN DETECTED PATTERN AND 001110

13952

ESF FRAME NUMBER	ESF BIT NUMBER	F-BITS ASSIGNMENTS FPS	FDL	CRC
1	0	–	m	–
2	193	–	–	CB_1
3	386	–	m	–
4	579	0	–	–
5	772	–	m	–
6	965	–	–	CB_2
7	1158	–	m	–
8	1351	0	–	–
9	1544	–	m	–
10	1737	–	–	CB_3
11	1930	–	m	–
12	2123	1	–	–
13	2316	–	m	–
14	2509	–	–	CB_4
15	2702	–	m	–
16	2895	1	–	–
17	3088	–	m	–
18	3281	–	–	CB_5
19	3474	–	m	–
20	3667	1	–	–
21	3860	–	m	–
22	4053	–	–	CB_6
23	4246	–	m	–
24	4439	0	–	–

A13953V
2/17/84

FPS–FRAMING PATTERN SEQUENCE (...001110...)
FDL–4kb/s FACILITY DATA LINK (message bits m)
CRC–CRC-6 BLOCK-CHECK FIELD (check bits CB_1–CB_6)

Figure 4. Frame Synchronization Strategy

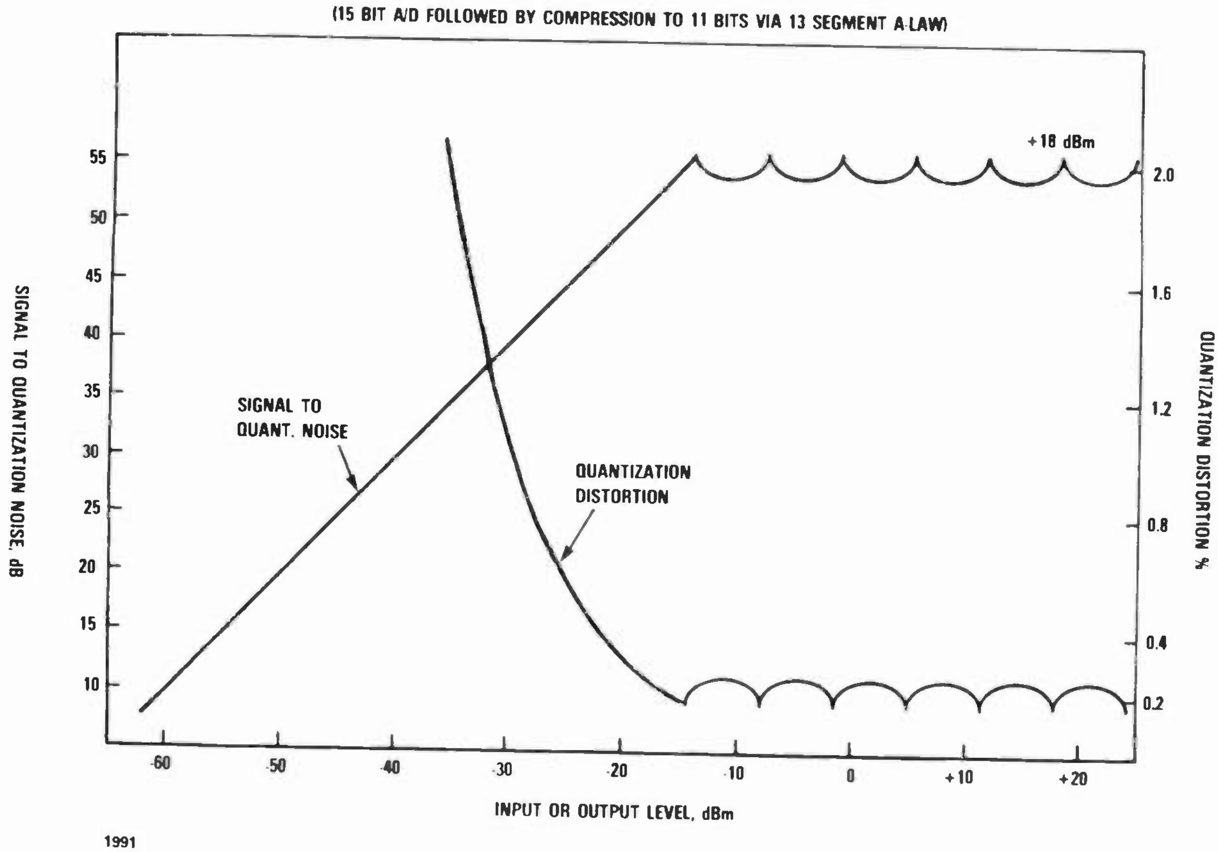


Figure 5. Quantization Noise Vs. Signal Level

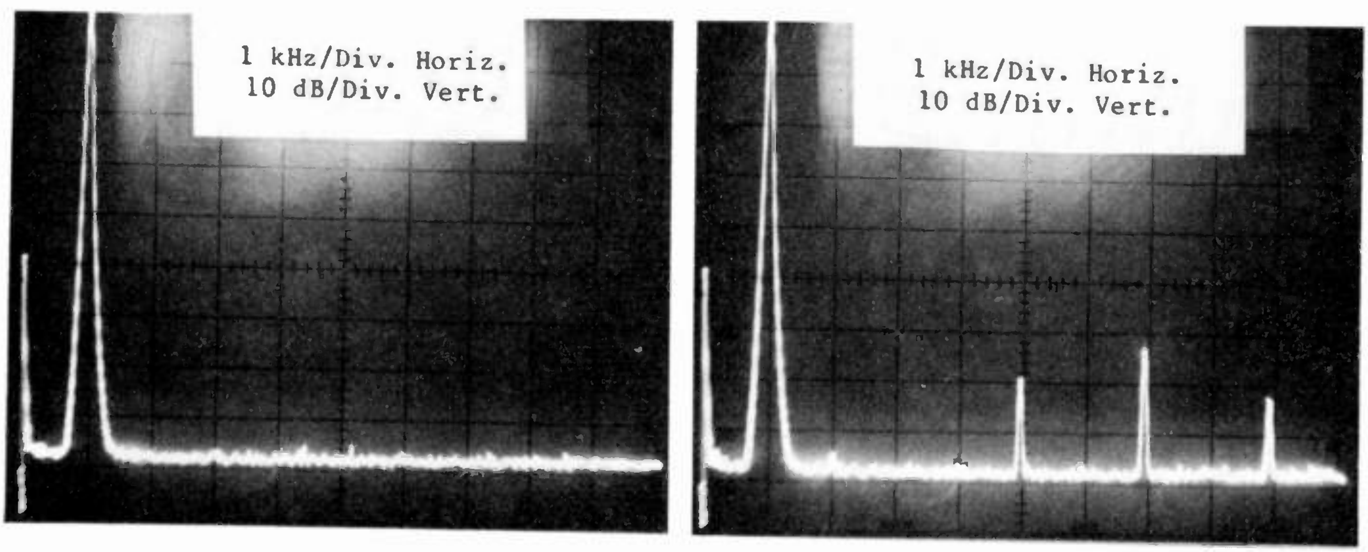


Figure 6. Distortion

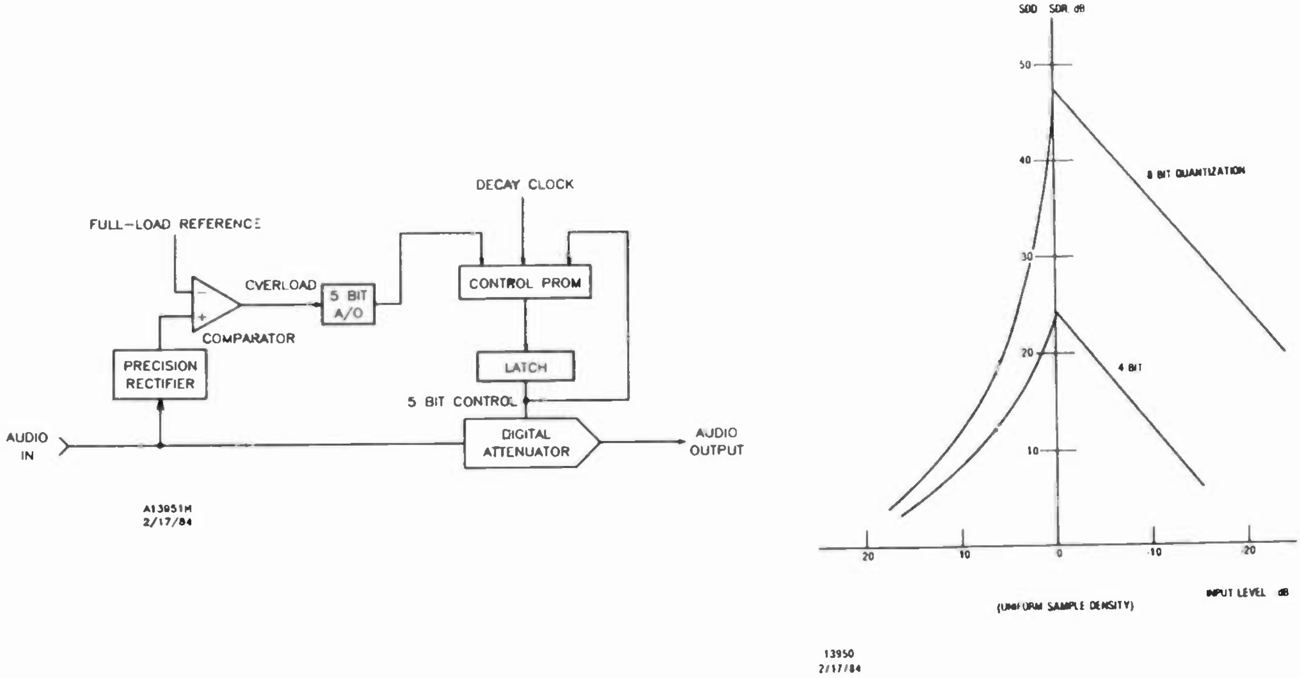


Figure 7. Effects of Overloads and AGC

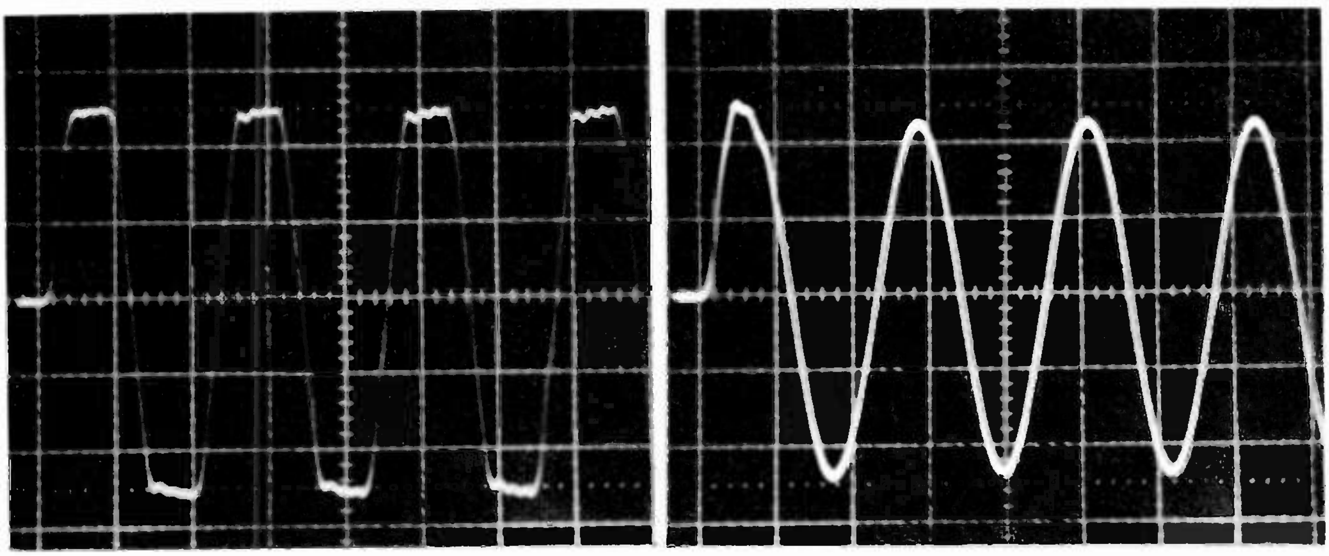


Figure 8. AGC Performance

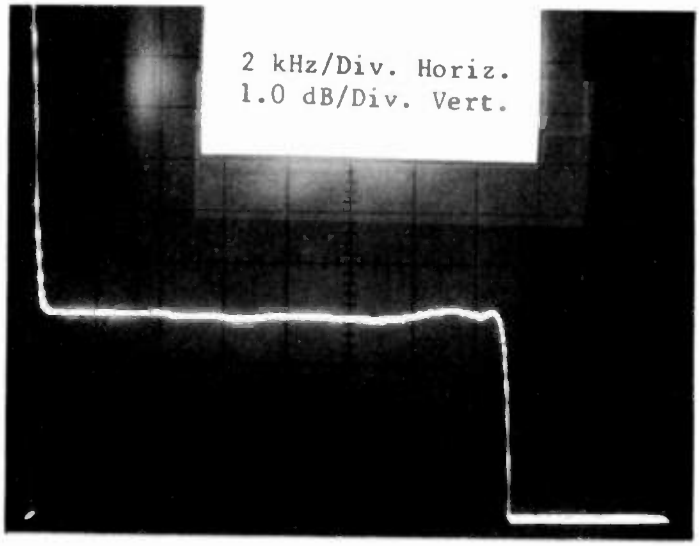


Figure 9. M/AESTRO End-to-End Frequency Response

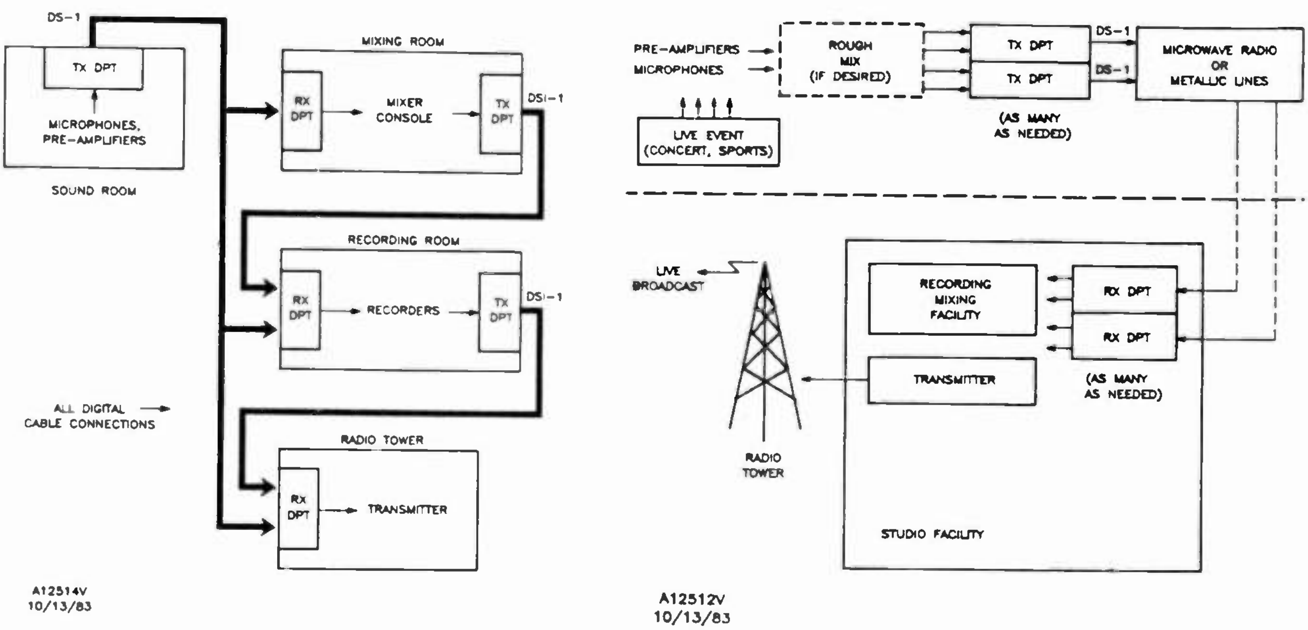


Figure 10. DPT Versatility

MINIMUM BANDWIDTH HIGH QUALITY DIGITAL AUDIO TRANSMISSION SYSTEM USING DC-PCM

Susumu Takahashi, Masanori Tanaka

Hiromi Kameda and Masaya Ishikawa

Sansui Electric Company Limited

Tokyo, Japan

ABSTRACT

The digital transmission system described in this paper was developed for transmitting high quality audio signals by a reduced-bit transmission channel. Basically, the system is a combination of near-instantaneous companding PCM (NIPCM) and differential PCM (DPCM). To realize such a combination, bits that are normally removed and truncated in NIPCM during compression are accumulated. Prototypes of encoder and decoder were subjected to measurement and listening tests. The signal-to-quantizing noise ratio (SNR) of this system is frequency dependent, varing from the level attainable by the number of quantizing bits to that of transmission bits. This response is well matched to the psychoacoustic characteristics of the human ear and efficient and high quality digital transmission system can be realized.

1. Introduction

Inspite of its high audio quality, a linear quantizing PCM system lacks in transmission efficiency. Such a system always requires transmission of full level data irrespective of the signal level. Various bit-rate reduction techniques, such as, instantaneous companding PCM (IPCM), near-instantaneous companding PCM (NIPCM), delta modulation PCM (DMPCM) and differential PCM (DPCM) have been developed to improve transmission efficiency.

In the USA, the Digital Audio Distribution System (DATS) developed by RCA American Communications, Inc., is currently operational (1). This system compresses 15-bit data to 11-bit by using floating point approximation companding techniques. Also, digitally controlled adaptive delta modulation (ADM) developed by Dolby Laboratories is known (2) for satellite application. In Europe and Japan, near-instantaneous companding audio multiplex (NICAM) developed by the British Broadcasting Corporation (BBC), originally for sound distribution links (3) - (B), and subsequently recommended by the European Broadcasting Union (EBU) for satellite broadcasting systems (9), is being considered. This is a NIPCM system which compresses 14-bit data to 10-bit transmission data. While it is reported that NICAM is satisfactory for high quality sound transmission (B), loss of quality is also reported (10), (11). Although not strictly applicable to high-quality audio signal transmission, DPCM is known to improve low frequency signal performance and, in addition, dramatically improve transmission entropy. Doi used music sources for his assessment (12).

With NICAM and DATS, their SNRs depend on the number of bits that are transmitted, and with ADM and DPCM, they depend on the sampling frequencies at higher frequencies. Therefore, it is generally believed that high fidelity transmission requires either increased number of bits or higher sampling frequency, and that improving transmission efficiency is contradictory to high fidelity transmission.

An efficient and high-quality digital transmission system is realized by combinging DPCM with NIPCM. Bits that are removed and truncated in NIPCM are accumulated to realize the combination. This hybrid system improves SNR of NIPCM at the rate of 6dB per octave at lower than 5.3kHz, reaching the maximum SNR of the quantizing bits at a low frequency. The new digital transmission system is called differential scale companding PCM, or DC-PCM for short.

In the following sections, configuration of the system using an 8-bit transmission channel is described, followed by a section on accumulation of removed bits and the performance of the DC-PCM systems.

2. System Configuration

It can be seen from the block diagram in Fig. 1, that the DC-PCM system embodies both DPCM and NIPCM. The explanation of this configuration, thereforem, concentrates only on the area that is new to this hybrid system.

It should be noted that DPCM as employed in this system does not use a predictor as does the more generally known DPCM system. In DIFFERENTIATOR (DPCM), the difference between data that is quantized to 14 bits by the analog-to-digital converter (ADC), and the immediately preceding data, is obtained. It is expressed in 15 bits in a 2's complement and the data is fed to NIPCM ENCODER and compressed to 8-bit data. ACCUMULATOR accumulates bits that are removed during this process. The function of the accumulator is described in greater detail in the following section. The 8-bit transmission data and the scale factor word are received by NIPCM DECODER, and expanded to 15 bits. INTEGRATOR restores the linear 14-bit data from the differential data, which is then fed to DAC for conversion to analog signals.

3. Accumulation of Removed Bits

Application of the DPCM technique to NIPCM has been hindered by two problems. First, transmission of differential data by NIPCM results in the accumulation of errors caused by truncated bits in the integrator of a receiver, and a shift or fluctuation in dc level. Second, when high and low frequency signals are transmitted at the same time, the low frequency signal becomes grossly distorted or is not transmitted at all.

With the system described in this paper these problems are resolved by accumulating the bits removed in NIPCM. The operation of the accumulator used in the DC-PCM system is depicted in Fig. 2. Instead of truncating bits that are removed by compression during the encoding process, they are retained and added to bits that are removed in the following compression process. When a carry is generated it is added to the compressed 8-bit data, which then becomes the transmission data.

When the compressed data before transmission is 01111111, addition of the carry from the accumulator makes the data 10000000, meaning the polarity of the data is reversed. This is because a positive 2's complement is used to express the data and to make system operation accurate and smooth. Therefore, polarity reversal of the most-significant bit (MSB) is not allowed, and, if such a state exists in a block, a measure like adding 1 to the scale prior to compression is necessary. Thus, accumulation of the removed bits can be summarized as follows:

The absolute error between the total sum of the original and decoded signals is equal to the value of the accumulator at the given moment. And the error is no larger than the value expressed by the accumulator bits at a given moment. These facts are achieved by adding a carry, which is generated in accordance with the value of the removed bits, to

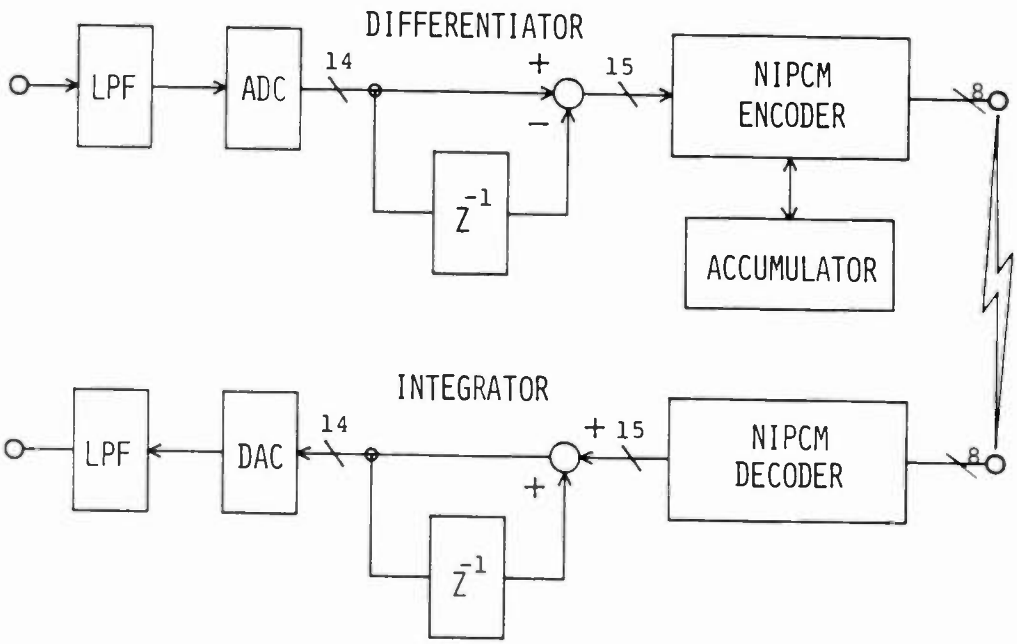


FIG. 1 BLOCK DIAGRAM OF THE 8-BIT DC-PCM SYSTEM

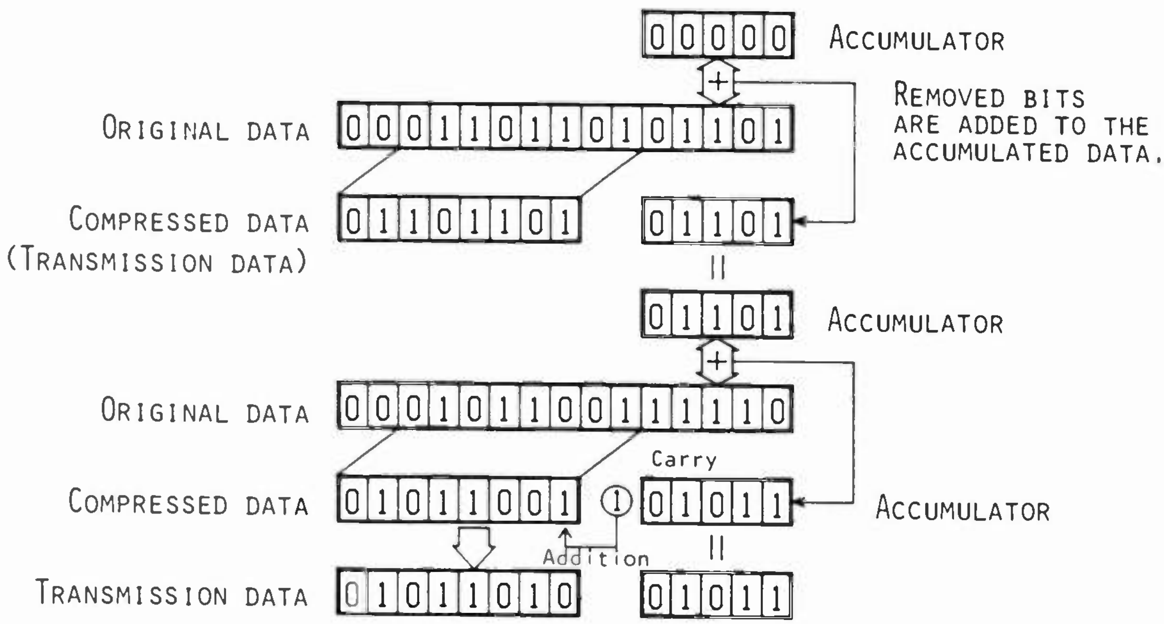


FIG. 2 OPERATION OF THE ACCUMULATOR

the transmission data.

Therefore, it can be said that function of the accumulator brings forth the following effects.

1) The decoded signals do not contain errors caused by the removed bits, and, thus, a dc level shift or fluctuation is suppressed.

2) Low frequency signals are properly reproduced even when their differential data is removed.

3) Signal-to-quantizing noise ratio

In the following sub-sections, measured results of the DC-PCM system using an 8-bit transmission channel and 48kHZ sampling frequency are discussed, followed by theoretical analysis of DC-PCM and NICAM using the same sampling frequency of 32kHz and transmission bits of 10. OdB in the measurement and analysis refers to the full scale of digital signals.

3.1) Measured results

Fig. 3 shows the measured SNR with respect to input levels, using frequency as a parameter. The curves of 5,000 Hz and 1,000 Hz display a sawtooth phenomenon when the input level is increased to a certain point. This is caused by the scale factor which grows progressively. The 100 Hz curve does not show such phenomenon, because compression does not remove bits at this frequency, making it identical to that of a 14-bit linear quantizing system. Fig. 4 shows SNR vs. frequency response. It is seen that DC-PCM displays 14-bit resolution at lower than 120 Hz even when using an 8-bit transmission channel. The small degradation is due to noise levels in the analog circuits.

3.2) Comparison of DC-PCM with NICAM

Fig. 5 shows theoretical SNR vs. frequency responses of near-instantaneous companding audio multiplex (NICAM) and differential scale companding PCM (DC-PCM). It is seen that the response of NICAM is frequency independent and its SNR is decided by the bits that are transmitted, i.e. 10 bits equal approximately 62dB. In contrast, DC-PCM exhibits frequency dependent response with its maximum SNR of about 86dB up to 637Hz, reaching its ceiling of 62dB at 5.3kHz. The improvement achievable by DC-PCM over NICAM is at the rate of 6dB per octave at lower than 5.3kHz. In other words, the minimum SNR at DC-PCM is decided by the number of bits used for transmission, and the maximum SNR by the

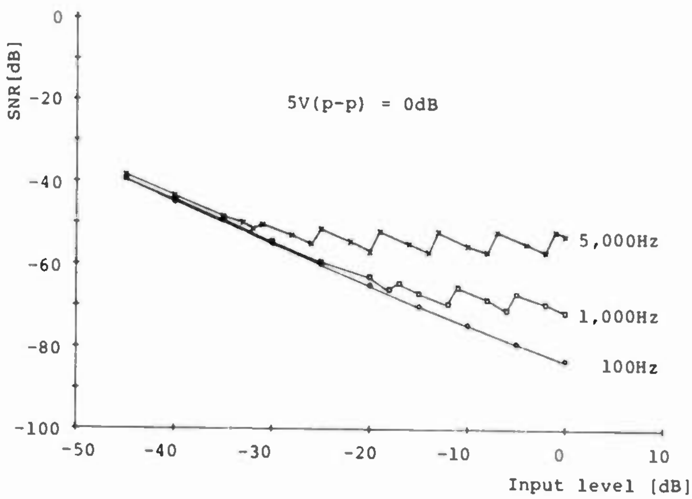


FIG. 3 SNR VS. INPUT LEVEL (MEASURED)

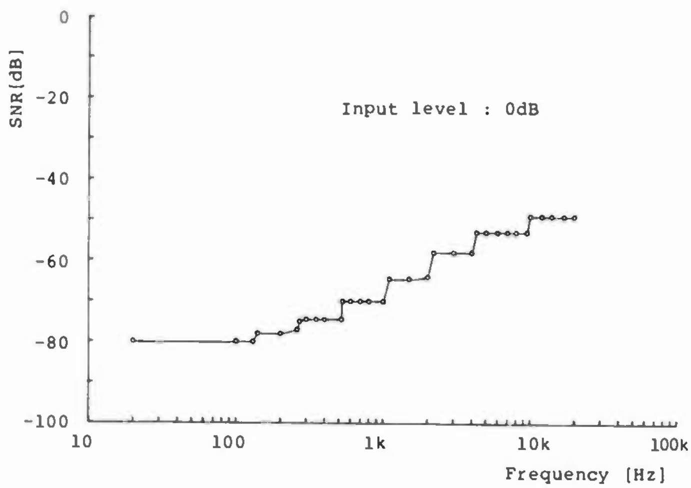


FIG. 4 SNR VS. FREQUENCY (MEASURED)

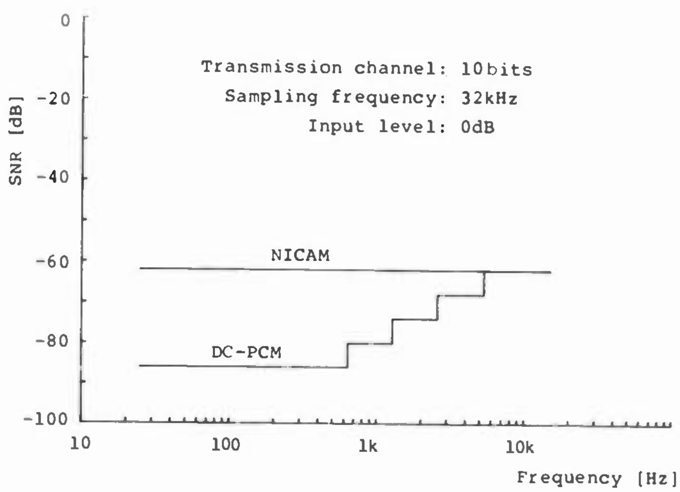


FIG. 5 SNR VS. FREQUENCY (THEORETICAL)

number of quantizing bits. Such characteristic is independent of the sampling frequency.

4) Listening Tests

A prototype of the DC-PCM system was compared with NIPCM and a linear 14-bit system, all of which sampling frequencies were 48kHz. We deliberately used an 8-bit transmission channel for NIPCM to judge the difference in quality under the same condition. With the 8-bit NIPCM, quantizing noise was obvious at low frequency range, particularly of high level. With the DC-PCM system, even under such condition, hardly any deterioration in noise was noted. In fact, even when the system was A/Bed with the linear 14-bit system, using music sources, hardly any deterioration was noted. Only a slight increase in distortion at higher frequencies was noted by a small percentage of people. The reason why the deterioration of SNR at higher frequencies is not objectionable may be because the SNR vs. frequency response is well matched to the characteristics of the auditory organ of the human ear (13). The human ear is less sensitive to higher frequencies than to lower ones (14). No attempts were made to compare DC-PCM with NICAM as developed by BBC, but it is assumed that a 10-bit DC-PCM will exhibit better performance (Fig. 5).

5) Conclusion

The near-instantaneous companding PCM system is now combined with differential PCM for more efficient and higher-quality digital transmission. We have confirmed that accumulation of the removed bits is an effective and realiable method for achieving such a combination. The new system greatly improves the signal-to-quantizing noise ratio of NIPCM or NICAM. On program materials no noticeable difference in subjective quality was noted even when an 8-bit transmission channel system was compared with a 14-bit linear system. It is clear, therefore, that the new differential scale companding PCM system (DC-PCM will offer significant advantages when applied to satellite transmission. Practical application, however, requires establishment of error correction and interpolation techniques and LSI integration of the circuitry.

6) ACKNOWLEDGEMENT

The authors would like to express their sincere gratitude to Kiroyiuki Kanzaki for his contribution to the development of this system.

7. References

[1] R. Langhans and M. Shumila,"Digital Audio Transmission System Using Satellite Distribution," Presented at the 74th AES convention, preprint no. 2018, October, 1983.

[2] K.J. Gundry, D.P. Robinson and C.C. Todd, "Recent Developments in Digital Audio Techniques," Presented at the 73rd AES convention, preprint no. 1956, March, 1983.

[3] J. R. Chew and J. G. Ingleton,"Pulse-code modulation for high-quality sound signal distribution: instantaneous companders using diodes as non-linear elements," BBC Research Department Report No. 1968/7, 1968.

[4] British Patent Application No. 10382/72, 6 March, 1972.

[5] D. W. Osborne,"Digital Sound Signals: Further investigation of Instantaneous and other Rapid Companding Systems," BBC Engineering, November, 1973.

[6] D. W. Osborne, M. G. Croll,"Digital Sound Signals: bit-rate reduction using an experimental digital compandor," BBC Research Department Report No. 1973/41, 1973.

[7] G. W. McNally,"The use of a programmed computer to perform real-time companding of high-quality sound signals," BBC Research Department Report No. 1978/14, 1978.

[8] N. H. C. Gilchrist,"Digital Sound Signals: tests to compare the performance of five companding systems for high-quality sound signals," BBC Research Department Report No. 1978/26, 1978.

[9] H. Mertens and D. Wood,"The C-MAC/packet system for direct satellite television," EBU Review No. 200, August, 1983.

[10] H. Hessenmüller and H. Jakubowski,"Digital sound transmission quality, Results of subjective tests; objective measurement methods," EBU Review No. 196, December, 1982.

[11] D. Wood and G. Billia, "EBU studies in digital audio," Presented at the 73rd AES convention, preprint No. 1975. March, 1983.

[12] T. Doi,"On bit reduction of digital audio systems," Presented at the 64th AES convention, preprint No. 1549. November, 1979.

[13] T. Muraoka, Y. Yamada and M. Yamazaki,"On auditory studies in the PCM recording-reproducing system utilizing a video tape recorder," Research Report, The Institute of Electronics and Communication Engineers of Japan, EA78-32, 1978.

[14] T. Doi and A. Iga, "Digital Audio," Published by Radio Technologies, Inc., p97, 1982.

IMPROVED NTSC-COMPATIBLE

Kerns H. Powers

Chairman, ATSC Technology Group on Improved-NTSC Compatible

RCA Corporation
David Sarnoff Research Center
Princeton, New Jersey 08540

The ATSC Technology Group on Improved NTSC-Compatible has four missions to accomplish:

(1) To assess the ultimate "benchmark" quality achievable in the home display with no incompatible changes in the current radiated-signal standards.

(2) To develop new transmission standards for NTSC that involve compatible changes to improve quality in new receivers but with acceptable impairments to old receivers.

(3) To foster a demonstration system that exhibits "Improved-NTSC benchmark quality" to be used as a standard for comparison with enhanced and high-definition system demonstrations.

(4) To provide input documents to the E-525 and HDTV Technology Groups containing recommendations to ensure compatibility with NTSC systems.

The first task undertaken by the Group was to prepare a complete block diagram of the Improved NTSC-Compatible system with identification of the interfaces where new standards or recommended practices are required. Figures 1 and 2 illustrate that the block diagram is separated into two components, that in the network chain and that at the home receiver. The network chain is illustrated because it contains all blocks from the production source camera to the transmitting antenna. Local station originations would, of course, involve fewer blocks. The interfaces requiring specification are indicated by arrows, and the stars distinguish those specifications that are either developed by or contributed by the ATSC Technology Group on Improved NTSC-Compatible.

The processes of improving the NTSC system are best discussed by starting at the display of Figure 2 and working backwards to the source. The ultimate display is a large one (0.5-1 m^2) with high resolution structural elements and scanning spot, driven by a high-line-rate scan converter that contains a field store and motion adaptive interpolation circuits to provide a lineless, flicker free, progressively scanned image without visible line crawl, interline flicker, or serrated edges on moving objects. The image sharpness is limited only by the horizontal and vertical resolution of 525-line transmission in a 4.2 MHz bandwidth rather than by the visible artifacts that derive from line interlace and the interleaved color subcarrier. The digital signal processing block provides three-dimensional filtering (H, V, T) on the incoming signal for sharpness enhancement and nonlinear processes to achieve ghost reduction, coring for S/N improvement, and crispening. The NTSC decoder block is a motion adaptive frame comb that virtually eliminates cross color and dot crawl in the high resolution display. The improved NTSC receiver contains multiple Hi-Fi loudspeakers to provide stereophonic or surround sound.

The network chain of Figure 1 is assumed to incorporate the "all-digital studio" concept implied by implementation of the world digital component standard of CCIR Recommendation 601. The video signals from the source camera are maintained in component form throughout the chain and are encoded into an NTSC composite signal only once, at the transmitter site. Tightened tolerances in existing NTSC radiated signal specifications might be applied here. The blocks labeled "predistortion box" and "ancillary signals" are intended to comprise not-yet-identified processes to improve the home display quality through complementary operations at transmitter and receiver for those receivers equipped to recognize the ancillary steering signal to perform the complementary process. The Technology Group is anxious to review new ideas and concepts in this area and would welcome suggestions from broadcasters, manufacturers, and others in our industry.

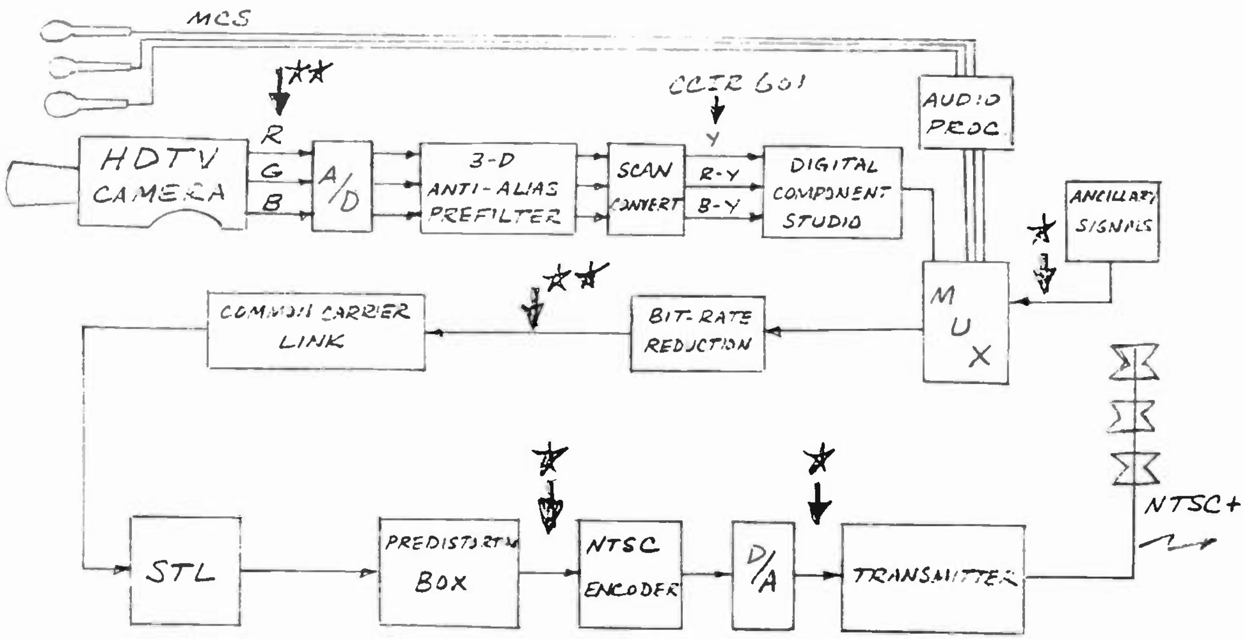
The Improved-NTSC signal is derived from a wideband high-line-rate camera through three-dimensional anti-alias prefiltering as described in the early work of Professor Broder Wendland of the University of Dortmund. Through vertical super sampling of the image combined with vertical peaking, brickwall filtering and subsampling, the so-called Kell factor can be increased to near unity which, when further combined with post filtering at the display, promises approximately a two-to-one increase in the effective (unimpaired) vertical resolution of the 525-line system.

The time frame for completion of the work of the Technology Group on Improved-NTSC is driven by that of the other Technology Groups. The improvements possible in NTSC are expected to be evolutionary in nature and will take place during a time period spanning a decade or more. These improvements can be implemented incrementally throughout the chain and are not constrained by the "chicken and egg" phenomenon. However, the demonstration of benchmark quality must be concomitant with and side by side with those of E-525 and HDTV, which are expected to commence by mid 1984.

A document has been prepared by the Technology Group in response to questions relevant to the use of NTSC encoding in new DBS service posed by the FCC Advisory Committee on DBS. A second document is being prepared to recommend to the HDTV Technology Group certain parameters for an HDTV production standard to ensure compatibility with CCIR Recommendation 601 (the world-wide digital component studio standard), thus compatibility with Improved-NTSC.

It is expected that such assistances to other committees will be provided whenever we can be of service.

IMPROVED NTSC



"AT THE NETWORK SOURCE"

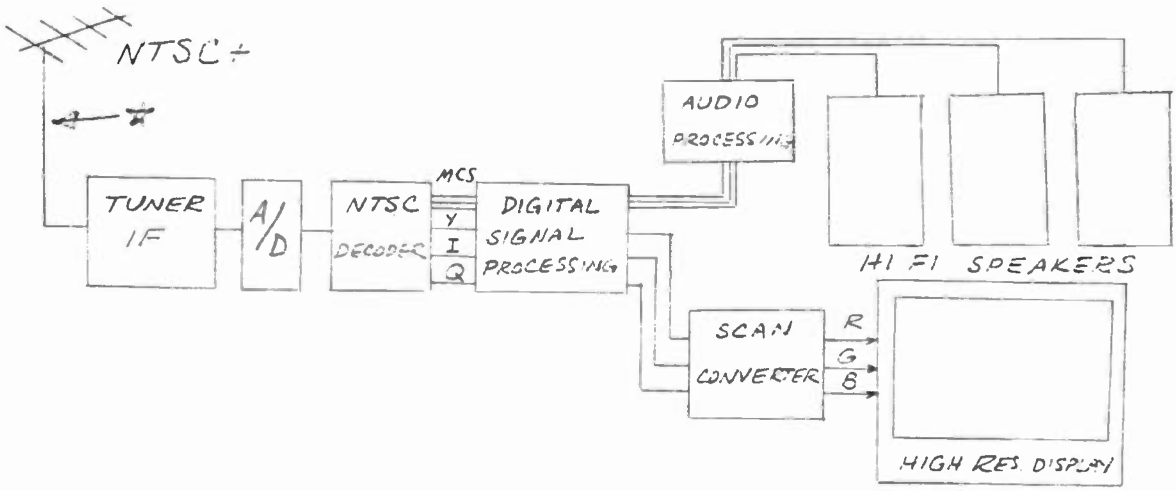
★ SPECIFICATION REQUIRED BY IMPROVED NTSC-COMPATIBLE TG

★★ SPECIFICATION BY OTHER GROUPS / INPUT FROM I-NTSC-C TG

December 15, 1983

Figure 1

IMPROVED NTSC



"AT THE RECEIVER"

★ SPECIFICATION REQUIRED BY IMPROVED NTSC-COMPATIBLE TG

December 15, 1983

Figure 2

HIGH DEFINITION TELEVISION: THE EUROPEAN BROADCASTER'S PERSPECTIVE

Charles Sandbank
Head of BBC Research Department
Chairman of EBU Specialist Group V1/HDTV

David Wood
Senior Engineer, EBU Technical Centre
Secretary of EBU Specialist Group V1/HDTV

Peter Rainger
Deputy Director of Engineering, BBC
Chairman of EBU Working Party V

1. EBU INVOLVEMENT IN HDTV

The great interest in high definition television in Japan for over ten years has resulted in exciting, well-engineered, high definition systems. The culmination of this work, as is well known, is a complete range of studio equipment working to an 1125/60/2:1 standard (8).

Until a few years ago however, in Europe, we were rather less convinced of the need to move into a new era of broadcasting. We had in any case, only in the past few years, completed the Europe-wide re-equipment for colour television. With some relief, the improvement in visual experience given by colour proved sufficient to induce the European consumer to pay five to six times more for his television set. Broadcasters could be forgiven for coming up for air for a while and being cautious about another leap forward. But the question of HDTV was not overlooked altogether. A notable study in the mid-1970s (1) developed some ideas for HDTV DBS systems. Whatever the potential of the system as a viable consumer product however, the bandwidth requirements seemed to indicate that further study would be unprofitable. There would be simply nowhere to broadcast the signal. And it must be said that available bandwidth still remains today a critical hurdle that must be overcome before HDTV broadcasting becomes a reality in Europe.

In 1979, an SMPTE study on HDTV (2) was published, and this and other factors prompted the members of EBU Working Party V (New Systems and Services) to consider the matter anew in 1980. One of the factors coming out of the SMPTE Report was the value of HDTV in the short term for the motion-picture industry. In essence, there was reason to believe that motion picture makers could expect significant savings in production costs if they used electronic picture origination. They could gain time by electronic editing, and there was a possibility of transmitting the electronic pictures by satellite to theatre outlets across the country. Alternatively, they could use tape-to-film transfer to obtain a film product. The expectation was that an HDTV system would allow definition which matched 35 mm film stock, and therefore, the audience's expectations of quality in a movie theatre could be maintained.

Here was a practical application for HDTV which was in no way constrained by the obstacles that seemed to prohibit its use in broadcasting. The available bandwidth was not an issue, and neither was the lack of appropriate low cost HDTV displays. If motion picture companies were to follow the opportunities offered, an HDTV standard would arise in the near future, and naturally one could expect it to be tailored specifically to the needs of the film production industry.

What would be the consequences of this for European broadcasters? First of all, we could cast our mind back to the evolution of film standards; where, broadly speaking, the standards chosen by Hollywood became worldwide standards, for reasons of market dominance, economics, etc. The same would probably happen for HDTV.

At some future point then, if and when it became possible to broadcast HDTV, there would be a de facto HDTV standard, which it would be necessary to adopt; or we would have to suffer the inconvenience of having at least two standards, one for film and one for (at least) European broadcasters.

The output of the motion picture companies represents a major source of broadcast material, and so transcoding of the electronic product could make a different standard even more unattractive.

European broadcasters have certainly seen the advantages of taking part in a standardization process at an early stage. We also know well the penalties of the lack of standardization. The digital 4:2:2 studio standard is now a world-wide standard, and digital video studio equipment interconnection will certainly be much simpler and more orderly than it might have been. On the other hand, current digital audio disc parameters are certainly not completely those European broadcasters would have preferred. In the future, whenever a digital audio disc player is used with a digital output in a digital broadcasting complex, we will need a sampling frequency converter to connect it to the rest of our digital audio studio equipment. Incidentally, the current market cost of a digital audio sampling rate converter is about five times the cost of a digital audio disc player itself.

Thus naturally anxieties, as well as a great intrinsic technical interest in the subject, prompted Working Party V to set up in 1980 an ad hoc Group to assemble evidence and information on current HDTV technology and prospects. Their report in 1981 was sufficient to convince the EBU Technical Committee that a Specialist Group was needed to embark on technical studies and propose a plan and timescale for the EBU's collective involvement in high definition television.

The NHK and CBS took part from the beginning in the EBU HDTV Specialist Group, and at a meeting in June 1982, they explained how they saw HDTV developing in North America and Japan.

As you know, there were (and are) hopes of broadcasting HDTV at least experimentally within the decade. For Europeans, this seemed quite impossible, because the only broadcast bands available in the near future (the 12 GHz band) had been already divided up with a channel width (27 MHz) that seemed to preclude

HDTV broadcasting. This aspect is examined further in Section 11, but whether or not Europeans envisaged moving into the age of HDTV, it was clear that the 525-world definitely did.

2. THE HDTV BROADCASTING CHAIN

An important element in the EBU's philosophy is that, with today's processing capability, there is no longer a need to maintain an exact one-to-one relationship between the technical format adopted for each of the three constituent parts of the broadcast chain. The chain is illustrated in Fig. 1 and comprises the studio complex, the transmission link, the signal emission and the display device. It is important to find the right balance between technical transparency and meeting the constraints of bandwidth and technology for each of the four elements of the chain. The four standards may be different, but each must be chosen with the characteristics of the others in mind. They are not necessarily tightly coupled. For example, the choice of display standard can be a receiver manufacturer's option and there can be a difference between expensive and cheaper receivers. The studio standard may be a source for transmissions using different emission standards with varying degrees of transparency to the original.

Broadly speaking, in the studio there are fewer constraints on system bandwidth, but the images should be capable of being processed without impairment. The processing includes such things as colour-matting, picture expansion and compression, etc. The general implication is that a guardband, or headroom, must be allowed for in picture definition over and above what is strictly necessary for technical quality. For the transmission link, there are more constraints on bandwidth, but this must be tempered with the possible need for some degree of picture processing downstream of the transmission link at a subsequent studio centre. For signal emission, minimising system bandwidth consistent with maintaining quality transparency becomes the primary requirement. In the receiver, bandwidth is again not a problem provided that semiconductor devices can operate at the appropriate frequencies or bit-rates.

The most pragmatic approach seemed to the EBU to be to first tackle the question of a studio standard. This does not mean that the other formats are forgotten, however, and some preliminary work on them is necessary. However, they will be simply one of the factors influencing the choice of studio standard rather than the determinant. The aim will be to avoid a studio standard that cannot be eventually converted to an emission standard, rather than to precisely define the emission standard at this stage.

3. A POSSIBLE TIMESCALE FOR STUDIES

In today's world of LSI, the necessary market size of ICs is certainly comfortably more than national. Also, we are more conscious of the value of worldwide standardization for cultural and political reasons and, in general, the operational requirements of broadcasters are always met more readily if world-wide standardization can be achieved.

To some extent, any worldwide standardization plan is constrained to operate in the timescale of the first users. With this in mind, the Specialist Group proposed the timescale outlined in Appendix 1 which envisaged defining the studio standard by about 1985 and the broadcast standard by about 1989. In such a case, any EBU proposals would have to be available somewhat earlier.

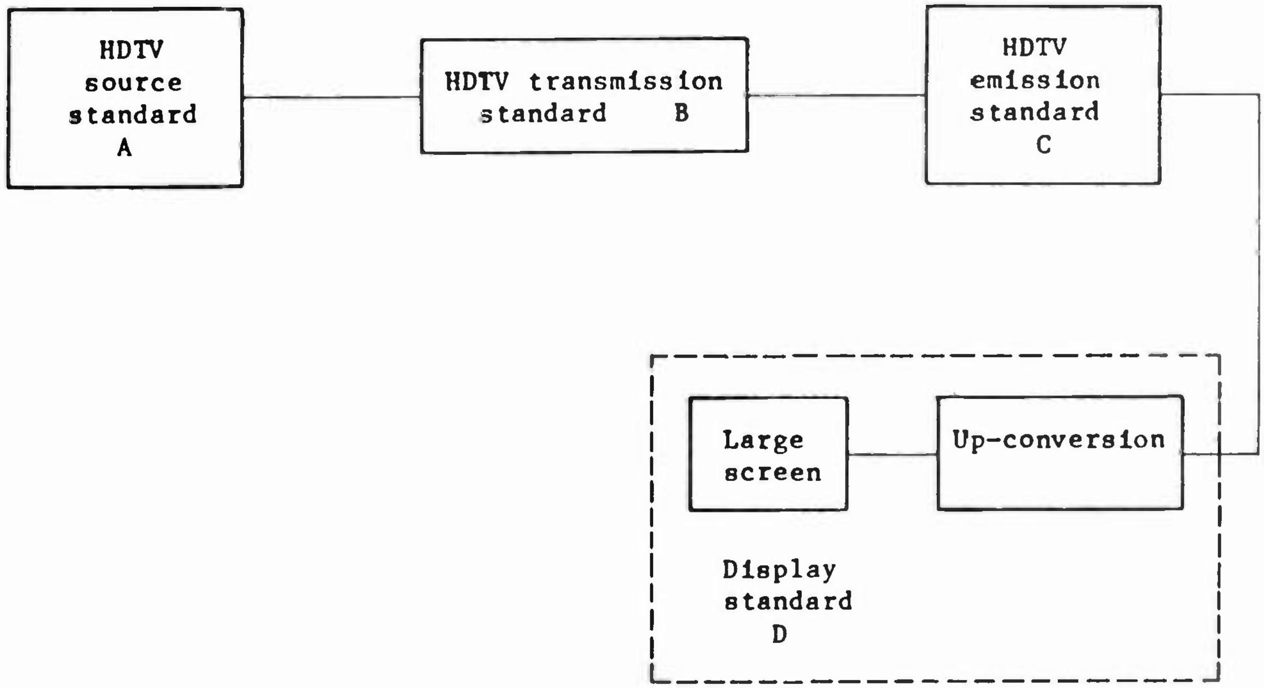


Fig. 1: The HDTV broadcasting chain

The hope was that these would be, as far as possible, worldwide standards, and indeed prompted by initiatives by the EBU, NANBA and the ABU, the Conference of all the world's nine Broadcasting Unions in March 1983, passed a resolution calling on all Broadcasting Unions to coordinate their studies with the aim of agreeing worldwide standards for HDTV. The Inter-Union Resolution is given as Appendix 2 to this article. As always in such cases, the aim is to submit proposals to the CCIR.

It would be impossible to predict the course of events of the years ahead, so there are no guarantees that these objectives will be met, but they certainly form the basis of our current planning.

4. FACTORS AFFECTING THE CHOICE OF AN HDTV STUDIO STANDARD

It is reasonable that the selection of parameters for the studio standard should take account of the following factors.

- The psyco-visual characteristics of the average viewer, based on the working assumption that a screen size of 1 metre or more is used at a viewing distance of about 3H. The aim would be a system close to the point where further increases in spatial and temporal resolution produce no worthwhile improvement, bearing in mind an appropriate degree of processing headroom. Coupled with this, is a consideration of the equivalence in quality to 35 mm film in movie theatre conditions. This latter factor is relevant to productions for threatre release and is therefore arguably, not a question for study by broadcasters. Nevertheless, it is imagined that it will certainly influence the final choice of parameters at the worldwide level.

- The practical availability of HDTV displays: these would not just be necessary for eventual use by the home consumer but also in order to conduct studio operations, and even before that, in the short term, to evaluate alternative HDTV systems.

- The constraints imposed by electronic cameras and studio operations, for example, limitations of camera integration time, noise performance, etc.

- The compatibility of the digital version of the system with the existing digital studio standard.

- The compatibility of the system with existing or envisaged conventional television systems.

- The possibilities for eventually broadcasting the HDTV signals.

- The relationship between HDTV and stereoscopic television.

These factors are considered in more detail in the following Sections 5-11.

5. BASIC QUALITY CONSIDERATIONS

Essentially, the objective of a high definition television system is to provide the viewer with a display which is indistinguishable from a window on the 'real world'. This means that the filtering action of the eye's response should be the limiting factor on the resolution, etc. that is perceived, rather than the transmission system and display. The spatio-temporal response of the eye is a function of intensity, viewing angle, wavelength and other factors, and therefore a complex multi-dimensional model is needed to specify it. This subject is still under study in many parts of the world but, according to some sources, as an example, a bright display would need a temporal bandwidth of the order of 70 Hz and a spatial bandwidth of 60 cycles/degree to be perfectly transparent. However, the eye's response is certainly not a step function and therefore it may be more practical to consider, for example, what volume under the eye's spatio-temporal response curves gives a reasonable compromise, or some other criterion. These concepts need to be clarified by theoretical and practical studies.

In broad terms, however, for a scanning system, the appropriate line rate seems likely to lie between about 1100 and 1500 lines/frame for a 2:1 interlace system, coupled with the corresponding horizontal resolution, although there are arguments for even higher values to allow for processing headroom.

An agreed method of subjective assessments for HDTV pictures must be developed. The EBU method (3) for subjective assessments of 625-line systems involves the regular use of a reference picture (usually RGB) to orientate observers. Defining and implementing a reference standard for HDTV presents problems. One possibility could be to use a directly projected 35 mm movie film or diapositive picture, but this would have colorimetry, contrast ratio, and response roll-off curve characteristics, which could be difficult to equate with electronic scanning systems.

A further question to be addressed is the minimum display rate for large area flicker to be imperceptible via peripheral vision, at high brightness levels. For a given brightness level there is a critical field frequency at which flicker is just perceptible. For example, at 200 candelas/m^2, this is about 57 Hz for a 2:1 interlace scanning system.

Other factors that must be considered include the minimum display field rate required for imperceptible interline twitter and the minimum field rate required for adequate motion portrayal, including considerations of processing headroom.

The question of picture aspect ratio is an important one, which requires psycho-visual analysis. The consequences on programme production and compatibility with existing aspect ratios also need to be considered. Clearly, the cinema industry has proved that aspect ratios of the order of 2:1 have great visual appeal, and possibly the figure which is close to the limit of creating a manifestly better sense of viewer involvement than current television aspect ratios is 5:3. The optimum aspect ratio may well be a compromise between conflicting interests.

The compromise between these various factors is made more difficult by the development of integrated circuit technology. We have reached a stage where the cost of circuits is not related to their complexity. One might almost say that the price is inversely proportional to the market size and independent of complexity. This is sufficiently near the truth to remove circuit complexity as one of the deciding issues as far as the domestic receiver is concerned. The limitations of what can be done are the more fundamental issues that have just been outlined.

The situation in terms of professional equipment is less clear. The smaller market size may mean that complexity is still a deciding issue.

6. DISPLAY SYSTEMS FOR HDTV

One of the fundamental problems in the study of HDTV is the current state of display technology. At the moment, systems other than those which are CRT-based seem to offer no great promise, and the point at which any breakthroughs in materials may occur is quite unknown. It could be in five years' time or in twenty-five years' time.

Research in image display technology requires a good deal of materials science know-how and capital investment; and, in general, EBU Members are not equipped to contribute significantly to this study. Our research is therefore largely confined to an analysis of the studies of the appropriate industrial laboratories.

For broadcast services, bearing the domestic viewing environment in mind, we will need a display of about 1 metre or so diagonal size, which will probably be viewed at about three times picture height.

Some relevant technologies are examined in the following paragraphs to evaluate the prospects for meeting the display requirements.

6.1 CRT-type displays

These may be either direct via phosphorescent tube or via a projection system. A number of HDTV projection systems have been built which would readily fulfil the screen size requirement, and there is some evidence of a modest market take-up of commercially available projection televisions for conventional services. However, there can certainly be no guarantees that domestic HDTV using projection display would be acceptable to the majority of consumers. In order to do so, the projector would have to be of sufficiently small size to fit comfortably into the average home viewing environment. This is not the case with current systems that we have seen. They would also have to have sufficient light output to be viewed in near daylight conditions, and have a wide viewing angle. As an example, a well known HDTV projection system gives a screen emission of about 25 ft-lamberts over an only modest viewing angle. This will probably need to be raised to 60-80 ft-lamberts over a wide viewing angle if it is to provide a service acceptable for the general public. It is true that the technology of projection displays is continuously improving, but one can be forgiven for wondering if there will be limitations of physics, which will prevent the ideal from being realised. We should also not forget that these systems must be within economic reach of the public, although perhaps cost considerations are mostly related to production volume, and therefore this problem may be solvable.

6.2 Flat screen electron beam systems

These have been the subject of considerable research and development and have yielded interesting results. There is, however, still much to be done before they could be considered for HDTV service, and not the least of the problems is evacuation.

6.3 LED displays

Many broadcast engineers have the inner feeling that a solid state flat screen display is the best and possibly the only long term solution to the problem. It could hang on the wall like a picture frame and cause no domestic inconvenience. It would be robust, suitable for high-volume production, etc. Unfortunately, at this time, there seems to be a fundamental problem of power conversion efficiency, which creates seemingly unsolvable heat disposal problems. There will, of course, always be a significant cost component associated with matrix addressing of elements distributed over a large area.

6.4 Scanning laser systems

These have substantial problems associated with the mechanical intricacy needed and the unwanted side effects such as speckle patterns due to the use of coherent light beams. Research into the solution of these problems is in progress in several countries.

6.5 Plasma-discharge devices, liquid crystal displays

There seem to be unacceptably low limits on the resolution possible with these types of display.

In summary then, current research gives little cause for optimism, apart from in the use of CRT-based displays, and even here the future route is not clearly marked. However, there may be an element of the chicken-and-egg situation. If high definition electronic signals are readily available, then more priority may be given to research aimed at the production of large, high resolution displays for the consumer market. On balance, however, if a working assumption is needed for the immediate future for the studio standard, it must be that the display will be based on electron beam scanning systems, and therefore all the defects of such systems must be taken into account.

7. CONSTRAINTS OF SIGNAL ORIGINATION

The problem of signal origination of HDTV pictures include physical limitations on camera pre-amplifier noise, camera integration time, and requirements for studio lighting levels. Currently developed HDTV cameras have more stringent requirements for incident light levels than conventional cameras even though they have much larger targets. This in turn places restrictions on production techniques, since cameras must be operated with a wider aperture and therefore a shallower depth of focus.

8. RELATIONSHIP TO THE 4:2:2 DIGITAL STANDARD

In 1981, a worldwide digital studio standard was adopted by the CCIR (5). This is termed the 4:2:2 standard, and the signal format is separate luminance and colour-difference (U and V) signals sampled at 13.5 MHz and 6.75 MHz respectively with 8 bits/sample resolution. This permits luminance bandwidths of the order of 6 MHz and 3 MHz respectively and a noise performance which is transparent to the source noise.

The CCIR members saw clearly that the natural evolution of studio equipment was towards digital video systems because of the advantages they offered in the maintenance of consistent performance and ease of processing techniques.

These same advantages would also apply to digital high definition pictures, and therefore we can conclude that the long term future at least must see a digital HDTV standard in operation.

The CCIR was careful to suggest that digital video standards should be chosen to be mutually compatible, by way of forming a family of digital standards. The idea was that conversion between them should be simple and not itself be a restriction on quality. Several ideas have been put forward to the CCIR for lower members of the family based on sub-multiples of the 4:2:2 sampling frequencies (i.e. 2:1:1, 3:1, 3:1/2:1/2). The sub-multiples concept could also possibly permit digital tape recorders to operate on both standards simply by switching tape speeds.

It seems entirely logical therefore that, if possible, a digital version of the HDTV standard should also conform to this notion of a family.

It is clear that the format for the HDTV signal will need to be different, in the constituent parts of the broadcast chain, in order to meet bandwidth and processing constraints. It seems inevitable that the changes of format will be achieved by means which include digital processing. It therefore seemed rational to the EBU group to consider and define the digital domain for the signal. This approach has a number of advantages, irrespective of whether in the early years the standard is implemented in digital form, because compatibility can be designed into the signal at the digital level, of course taking into account quality requirements, and specific needs of the analogue version, in areas such as blanking.

Whilst the CCIR studies proposed that digital television standards form a hierarchy of compatible standards, in reality there are two versions of the 4:2:2 standard, one 625/50 and the other 525/60. If, therefore, the hierarchy principle were applied absolutely, it would lead to two versions of the HDTV standard. This is not seen as a desirable objective, and the EBU group's proposal is for a single HDTV standard, which forms part of the hierarchy, in that it is designed to be compatible with the common elements of the 4:2:2 standard, in such a way as to allow transparent conversion to either one.

Working on this philosophy, the EBU group has attempted to delineate possible options for the HDTV signal in the digital domain. The initial views are given in Table 1. The choice of active picture area was made with compatibility and quality requirements in mind, but the required blanking intervals need to be studied further, and the overall line number, 1125, must be seen as heuristic. The Table is not intended to propose any specific standards but is an attempt to quantify the technological implications of possible parameters and to indicate the relationship of these with existing standards, particularly in the digital domain. It must be emphasised especially that the inclusion of field rates above 60 Hz is in the spirit of examining possibilities and to provide a framework for studies which need to be carried out.

These ideas are at an early stage, but nevertheless there would be everything to be gained from defining the main elements of the digital HDTV standard at the same time as the anlogue standard.

9. COMPATIBILITY WITH CONVENTIONAL TELEVISION SYSTEMS

This question may well prove the most critical in obtaining international

TABLE 1: DRAFT FRAMEWORK FOR A WORLDWIDE HDTV STANDARD

	EXISTING STANDARDS		SOME CANDIDATES FOR A WORLDWIDE HDTV STANDARD	
	625-line	525-line	Interlaced 1125-line (2)	Non-interlaced 1125-line (2)
Samples/ Active line Y(U/V)	720(360)	720(360)	1800(900) (3)	1800(900) (3)
Field rate (fields/sec)	50	60	Option a: 80 Option b: 60 Option c: 50	Option a: 80 Option b: 60 Option c: 50 Option d: 40
Sampling frequency MHz Y(U/V)	13.5(6.75)	13.5(6.75)	(1) (5) Option a: 74.5 + x (37.25 + x/2) Option b: 55.9 + x (27.9 + x/2) Option c: 46.6 + x (23.3 + x/2)	(1) Option a: 149 + 2x (74.5 + x) Option b: 111.7 + 2x (55.9 + x) Option c: 93.1 + 2x (46.6 + x) Option d: 74.5 + 2x (37.25 + x)
Number of active picture lines	575	483	1035 (4)	1035 (4)
Interlace	2:1	2:1	2:1	1:1
Aspect ratio	4:3	4:3	5:3	5:3
Bits/sample	8	8	8	8
f_H	15.625 kHz	15.734 kHz	(a) 45 kHz (b) 34 kHz (c) 28 kHz	(a) 90 kHz (b) 68 kHz (c) 56 kHz (d) 45 kHz

Note (1): The value of the variable x would relate to the horizontal blanking interval duration. The necessary amount of horizontal blanking remains to be studied.

Note (2): This allows for a blanking interval of 8% of picture height; but shorter intervals could be feasible.

Note (3): This allocation of samples/line allows a two-to-one increase in horizontal resolution in a 4:3 sub-picture.

Note (4): This choice is made to provide simple conversion to 575 and 483 active lines, since it is a multiple of a common factor (23).

Note (5): Similar sampling-frequencies in a non-interlaced system would permit about 517 active lines/picture.

agreement on an HDTV standard. It seems clear that programmes originated in the HDTV standard will be used as source material for many years by conventional television systems (or enhanced versions of them with the same line rates and field rates). This could be even more the case for the developing countries.

It therefore is vital to design a standard which will not give worse quality after standards conversion, than signals originated in the existing standards. The HDTV line rate chosen, provided it is sufficiently high, may not place fundamental restrictions on the quality of conversion. It will simply make the conversion algorithm more or less complex, and possibly mean the omission of an amount of picture area. However, this is not the case for field rate. If the HDTV field rate is relatively lose to the conventional field rate, the HDTV field rate chosen would have a marked effect on temporal aliasing in the converted picture, which could be impossible to remove.

If a signal originally 1125/60 is converted to a 625/60 signal, the temporal characteristics remain very similar. However, if a 1125/60 signal is converted to a 625/50 signal, then certain temporal aliasing components appear, which could give rise, for example, to movement judder on the picture. If a 1125/100 or 1125/50 signal is converted to 625/50, there would be no additional inband temporal aliasing components introduced (4).

Undoubtedly the most difficult problem in establishing a worldwide HDTV standard is to ensure that, on conversion, it suits both 60 Hz and 50 Hz standards adequately. One possibility might be to go above 60 Hz for a world-wide HDTV studio standard. If a high definition standard were to use 100 Hz or 120 Hz, high quality converted pictures could be obtained on both 50 Hz and 60 Hz standards, although the conversion in which the field rate change was 2:1 could be a little superior. However, such field rates would have to be paid for in terms of bandwidth and so possibly a compromise of 75 Hz or 80 Hz may be the answer to keeping temporal aliasing within the acceptable limits for both systems, and permitting a degree of operating headroom.

10. BROADCASTING HDTV SIGNALS

The problems of broadcasting HDTV signals are formidable. Typical sets of system parameters such as 1125/60/2:1 or 1250/50/2:1, would give bandwidths (without sophisticated compression techniques) of the same order: 20-30 MHz depending on the method of multiplexing the components. This certainly exceeds considerably the bandwidths currently allocated for vision in terrestrial channels. For Europeans, the next available frequency band will be the 12 GHz band for DBS and here, as previously mentioned, a plan has already been defined which has a 27 MHz channel spacing, and indeed a common European DBS system (C-MAC/packet) has been proposed by the EBU (7).

An approximately constant amplitude modulation method is needed for DBS, and the maximum available bandwidth (27 MHz channel spacing) for an analogue vision signal is likely to be in the range 9-13 MHz. The precise value would depend on an interference analysis, but a video bandwidth of 20 MHz or more would certainly not be feasible with a single European DBS channel in the 12 GHz band.

Other options open in theory would be to use higher frequency bands, with wider channel bandwidth allocations. The EBU is studying this possibility, with a view to initiating frequency planning if appropriate. Essentially, there are two possibilities: the use of the 23 GHz band and the use of the 42 GHz band.

One study (6) considered the use of digital coding plus bit-rate reduction for the HDTV signal. For the case of the 23 GHz band, the channel capacity would be limited to one or two HDTV channels per country and the transmitted power would need to be 750 W - 1.5 kW per channel, depending on the degree of bit-rate reduction possible.

For the case of the 42 GHz band, the corresponding figures would be 4-8 channels with transmit powers of 12 kW - 25 kW.

There is certainly no formal EBU position on the viability of using these bands at the moment, but it must be clear that a very high price will need to be paid in transmitter powers, and the possibility of launching such satellites must be in question in the foreseeable future.

This gives us very good reason for studying the possibilities for all three bands, including the 12 GHz band, at the same time.

Although the available video bandwidth in the 12 GHz band will be limited to something less than 13 MHz, there would remain ways of using the band to give viewers an improved quality service, which are entirely compatible with existing plans for DBS.

One option would be to use two channels for the total HDTV service. The signal could be partitioned in such a way that the first channel is receivable on a conventional receiver, whilst the information in a second channel complements it for HDTV reception. The partitioning could be done in the frequency domain or by vertical-temporal filters. (The latter would be the more suitable for the C-MAC/packet system.)

The maximum quality available in a single 27 MHz channel with the C-MAC/packet framework when using display up-conversion is unknown, but if this proves to be adequate for HDTV pictures displayed on domestic sized screens, it could provide one of the means of broadcasting HDTV source signals.

Recent studies such as those carried out by NHK (9), where the data channel is used to assist adaptive interpolation processes, now give rise to guarded optimism about the prospects for eventually achieving, within one non-compatible 27 MHz WARC channel, a transparency which might do justice to the EBU's quality objectives for HDTV as outlined in Section 5.

The 12 GHz broadcasting possibilities are therefore worth examining further, although there remains a good deal of work to be done on reconciling a transmission requirement for the optimum "high definition" service with the primary demands on the 12 GHz band.

11. STEREOSCOPIC TELEVISION

There have been arguments raised in the EBU for studying stereoscopic television simultaneously with HDTV. At the risk of simplification, the essential argument is that the 'presence effect' is really related to screen size, picture fidelity, and depth perception. Each of these elements has a bearing on the viewer's sense of participation in the scene. It is as if they might be considered as three orthogonal axes, and the 'presence effect' is a vector derived from the sum of three coordinates. Of course, different weightings may well apply to the three axes. Hypothetically, the same degree of presence effect might be achieved with much less resolution if the picture were stereoscopic than if it were monoscopic. These kinds of hypotheses have yet to be proven, but they certainly warrant study.

It would, however, be wrong to minimise either the display problems of stereoscopic pictures, or the problems of making stereoscopic programmes, but intuition tells us that unless we consider at least stereoscopic compatibility at this stage, our grandchildren may well call us shortsighted.

The only method of making an effective colour display for stereoscopic programmes identified by the EBU is the polarization plane method. Experiments have shown that this can give effective and high quality results, but of course still requires the use of polarized glasses. Work on the subject is continuing, including, for example, examination of eye fatigue with polarized glasses, but it is too early to see this as any kind of commitment on the part of the EBU to stereoscopic television.

12. CONCLUSIONS

The conclusions so far agreed by the EBU Specialists only concern the HDTV studio standard, and they simply represent a framework on which to build the complete specification in the coming years. The main elements are as follows.

A single worldwide HDTV studio standard should have the following characteristics:

- It should be based on the use of separate Y, R-Y, B-Y components.

- HDTV studio sources should provide signals with approximately twice the spatial resolution (e.g. twice the vertical resolution and twice the horizontal resolution) of studio sources using existing standards.

- The parameters should allow conversion to the existing 625/50 and 525/60 standards with a quality at least as good as that available from signals generated directly in the existing standards.

- The standard should be specified in both analogue and digital form from the outset even if the digital version is not immediately implemented.

Much work remains to be done; and, as mentioned earlier, if we are to meet the timescales of the first potential users, it must be done rather quickly.

The EBU believes that the international agreement on a 4:2:2 standard is extremely important and is anxious that the development of high definition standards should not put in question the agreement which has already been reached. As suggested, the new standards might be built upon colour component systems, the 4:2:2 family and its derivatives.

There must be steady pressure to make progress on high definition standards, even though its implementation is several years away. Even when we have achieved the objectives set out in the timetables produced by the working groups within the EBU, there will remain the process of design and manufacture of a range of related equipment and for most broadcasters further substantial investment in this new technology will be a long term process.

POSSIBLE TIMETABLE FOR STANDARDIZATION STUDIES

APPROPRIATE EVENTS IN STANDARDIZATION CYCLE		ACTION
	1982	
		Preparation of draft recommendation for "framework" of HDTV studio standard
	1983	
EBU Technical Committee		
CCIR Interim Meetings		
	1984	Preparation of draft recommendation for HDTV studio standard
	1985	
EBU Technical Committee		
CCIR* Final Meetings		
	1986	Preparation of draft recommendation for "framework" for HDTV broadcast (emission) standard
CCIR* Plenary Assembly		
	1987	
EBU Technical Committee		
CCIR* Interim Meetings		
	1988	Preparation of draft recommendation for HDTV broadcast (emission) standard
	1989	
EBU Technical Committee		
CCIR* Final Meetings		
	1990	
CCIR* Plenary Assembly		
	1991	

----

* Estimated.

APPENDIX 2

INTER-UNION RECOMMENDATION ON HDTV (ALGIERS, MARCH 1983)

DZ.T/41: HIGH DEFINITION TELEVISION

Considering:

- that there has been substantial progress made with high definition television technology of production equipment;

- that high definition television systems will require a resolution which is approximately equivalent to that of 35 mm film and corresponds to at least twice the horizontal and twice the vertical resolution of present television systems;

- that the advantages of a single HDTV worldwide standard include lower equipment costs for broadcasters and viewers, easier exchange of programmes and technical information, and encouragement to the ideal of international solutions to common technical problems;

- that multiple different standards will cause difficulties among broadcasters in the future;

- that there will be a need to distribute to existing television receivers, programmes originally produced in HDTV;

the Conference recommends:

1. that the Broadcasting Unions should encourage their Members to carry out studies on the preferred characteristics of a uniform world standard for a high definition television system;

2. that the Broadcasting Unions should coordinate their studies on high definition television systems;

3. that the Broadcasting Unions should concentrate these studies at first on a single HDTV production standard extending them later to transmission and broadcast;

4. that note should be taken in this development of the necessity for transcodeability to existing television standards throughout the world.

ACKNOWLEDGEMENTS

The opinions expressed in this article are those of the authors, but the EBU studies reported were made largely by members of EBU Specialist Group V1/HDTV whose work is gratefully acknowledged.

REFERENCES

1. Phillips, G.J. and Harvey. R.V. High-definition television for satellite broadcasting.
EBU Review - Technical, No. 170, (August 1978).

2. A study of high-definition television systems.
Report of the HDTV Study Group to the SMPTE Committee on New Technology, (July 1979).

3. Bernath, K., Kretz, F. and Wood, D. The EBU method for organising subjective tests of television picture quality.
EBU Review - Technical, No. 186, (April 1981).

4. Sandbank, C.P. and Moffat, M.E.B. High-definition television and compatibility with existing standards.
Journal of the SMPTE, (May 1983), pp. 552-561.

5. Recommendations and Reports of the CCIR, (1982).
Volume XI, Part 1, Recommendation 601.

6. CCIR Interim meetings of Study Groups 10 and 11, (1983).
Input document 10-11S/16.

7. CCIR Interim meetings of Study Groups 10 and 11, (1983).
Input document 10-11S/38.

8. Recommendations and Reports of the CCIR, (1982).
Volume XI, Part 1, Report 801-1.

9. A single channel HDTV broadcast system.
Non-published document prepared by NHK and submitted to the EBU (6 January 1984).

Proposed Solution to the

Channel 6/Educational FM Broadcast Interference Problem

Jules Cohen

Jules Cohen & Associates, P.C.

Washington, D. C.

Television channel 6 occupies the band of frequencies from 82 to 88 MHz. The FM broadcast band extends from 88 to 108 MHz, with the lowest 4 MHz of that band reserved for noncommercial educational FM broadcast stations. The potential for interference clearly exists and the experiences of channel 6 licensees document the fact that such interference is being received by television viewers.

By and large this interference phenomenon is in one direction only. The FM receiver designed for a signal only 0.2 MHz in width, is far less susceptible to interference than the television receiver with its 6 MHz bandwidth. Allegations are made, with some validity, that avoidance by the Federal Communications Commission of allocating adjacent television channels in the same community has permitted television receiver manufacturers to get by with producing sets having poor selectivity. That poor selectivity is cited by some as being the root cause of the FM interference problem and the solution lies in the production of better receivers. L. H. Hoke, of NAP Consumer Electronic Corp. in Knoxville, Tennessee, pointed out in his excellent paper presented at last year's NAB Engineering Conference that more than receiver selectivity is involved in the channel 6/FM interference problem. Furthermore, even if receivers with superior selectivity were made available tomorrow, approximately ten years would be required before the more than 83 million television households would be equipped with such receivers.

The FCC has been attempting to arrive at an equitable solution to the problem of channel 6/educational FM broadcasting interference. In response to a Petition for Rulemaking filed on May 12, 1972, by the Corporation for Public Broadcasting and a Petition by Operation Outreach, Inc., filed on December 2, 1975, the FCC on March 17, 1976, issued a Notice of Proposed Rulemaking./1 That rulemaking dealt also with matters other than interference, but proposed specific rules with respect to channel 6 interference from noncommercial FM broadcast stations. First and Second Reports and Order/2, /3 were adopted resolving issues other than channel 6 interference. The interference question was reserved for a later Report and Order. In a Further Notice of Proposed Rulemaking,/4 adopted June 7, 1978, the FCC stated that such a Report and Order was expected to be issued "soon." The Further Notice invited comments on the specification of classes for noncommercial educational FM assignments and on a Table of Assignments proposed by CPB. On May 13, 1982, the FCC adopted a Second Further Notice of Proposed Rulemaking/5 in Docket Number 20735. This Second Further Notice was devoted principally to the matter of TV channel 6 interference protection standards.

Review of the principles set forth in the Second Further Notice as the basis for channel 6 protection from FM interference created substantial concern in the television community. The criteria proposed to be adopted would, if implemented, result in substantial interference to channel 6 television stations from new noncommercial educational FM broadcast stations. The decision was made to convene an engineering committee which would study the interference problem and recommend suitable criteria which would protect television viewers while giving the noncommercial broadcaster as much leeway as possible in the establishment of new stations. Support for the activities of the committee came from the NAB, The Association of Maximum Service Telecasters, Taft Broadcasting Company, Storer Communications, Inc., McGraw-Hill Broadcasting Company, The Outlet Company, Chronicle Broadcasting Company, Capitol Cities Communications, Inc., and Cosmos Broadcasting Corporation.

National Public Radio and The Corporation for Public Broadcasting were invited to participate in the activities of the engineering committee and did, in fact, do so, although comments filed by those organizations in reply to the Second Further Notice of Proposed Rulemaking indicated less agreement with the conclusions of the committee than the majority believed to exist. In addition to representatives from NAB, NPR and CPB, participants in the committee included engineers from Taft Broadcasting Company, Capitol Cities Communications Inc., McGraw-Hill Broadcasting Company and the consulting engineering firms of A. D. Ring & Associates, Cohen & Dippell, P.C., and Jules Cohen & Associates, P.C.

Three principal techniques are advocated for minimizing channel 6 interference from FM stations: (1) collocation, (2) where collocation is not feasible, location of the FM station in an area of low population density, and (3) cross polarization.

Collocation

The purpose of collocation is to achieve the same propagation path for the television and FM stations, thus making possible the maintenance of a nearly constant desired-to-undesired signal ratio. If possible, the television and FM antennas should be mounted on the same tower. If use of the same tower is not feasible, separation of the two antenna systems by no more than 400 meters will preserve most of the benefits of collocation.

Since the objective of collocation is to maintain a near constant desired-to-undesired signal ratio, the maximum permissible effective radiated power of the FM station must be specified and the radiation patterns (both horizontal and vertical) must be correlated. That correlation of antenna patterns must take into account the relative heights of the two antennas. The determination of depression angles toward successive points along the path away from the TV and FM antennas is an essential element of that correlation.

Suggested maximum effective radiated power for FM stations conforming to the collocation criteria follows:

Channel	Power (dBk)	Power (kW)
203	5	3
204	7	5
205	9	8
206-215	10	10
216	12	16
217	15	32
218-220	20	100

It may be noted that channels 201 and 202 have been omitted from the Table. This is not an oversight. The carriers for FM stations operating on those channels are only .35 MHz and .55 MHz removed from the channel 6 aural carrier. Such frequency spacing would not be permitted for FM broadcast stations serving the same area, nor should they be permitted for an FM broadcast signal and the frequency-modulated aural signal of a TV station.

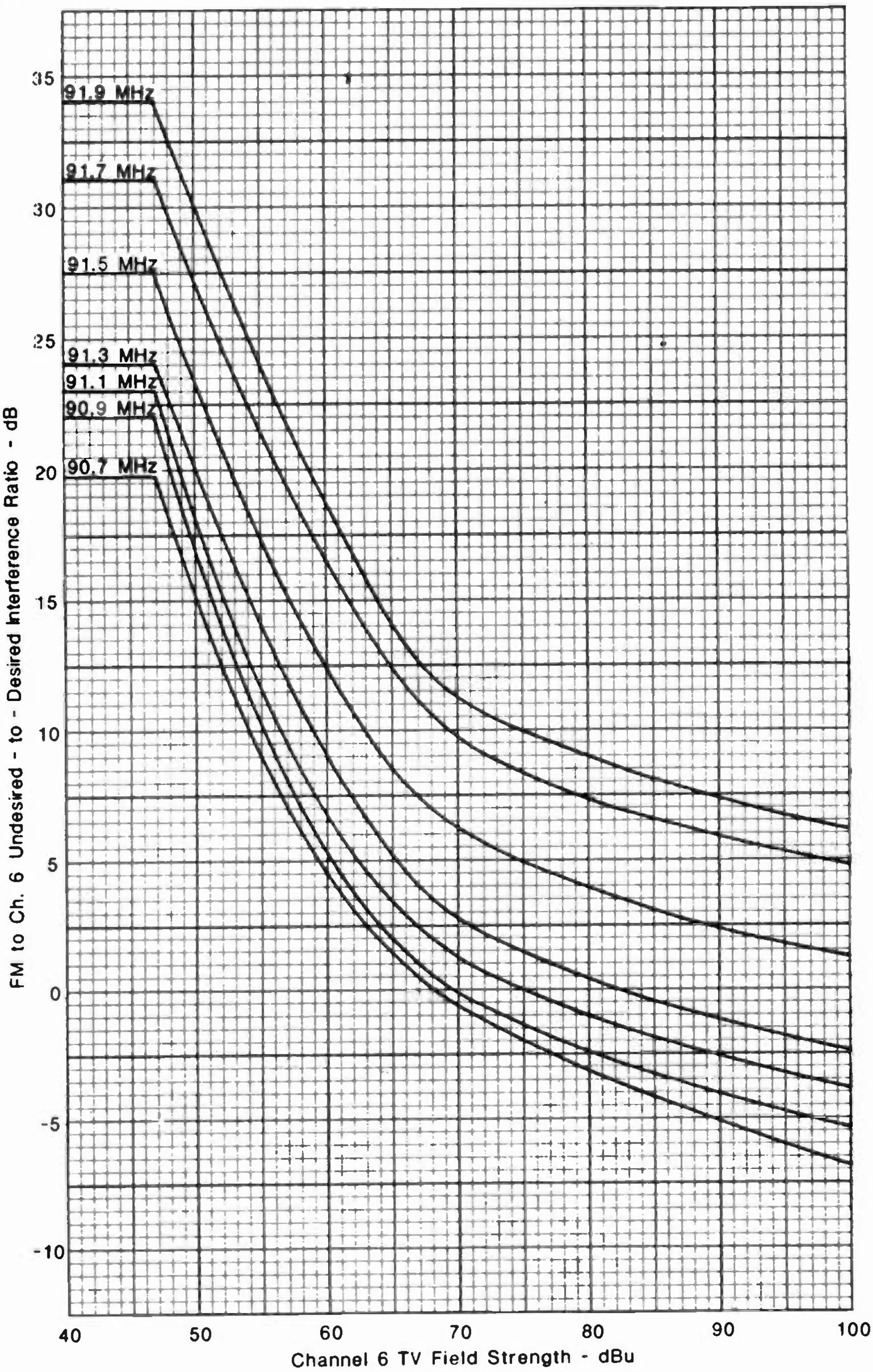
Alternate Locations

Where the FM broadcast station is not intended to serve the same principal community as the TV station, or other circumstances exist making collocation impractical, the FM broadcaster should be induced to locate in an area of relatively low population density by imposing a limit on the population which may be included within the area where a particular undesired-to-desired protection ratio is exceeded. Two undesired-to-desired protection ratios are suggested. These ratios are shown on four graphs accompanying this paper.

Figures 1 and 2 show the ratios for no perceptible interference for 70 percent of the 20 newer television receivers tested by the FCC laboratory./6 The second set of curves, identified as Figures 3 and 4, represent the point of perceptible interference for the median receiver.

The proposal has been made that the population permissible within the area bounded by the ratios shown in Figures 1 and 2 not be in excess of 3000. For the ratios shown in Figures 3 and 4, the population must not exceed 1000. The dual ratio is considered to be necessary in order that fewer people be allowed within that area where interference is most probable. After all, if only the first ratio applied, all 3000 persons could be crowded within the area where it is to be expected that 50 percent or more of television viewers would receive interference.

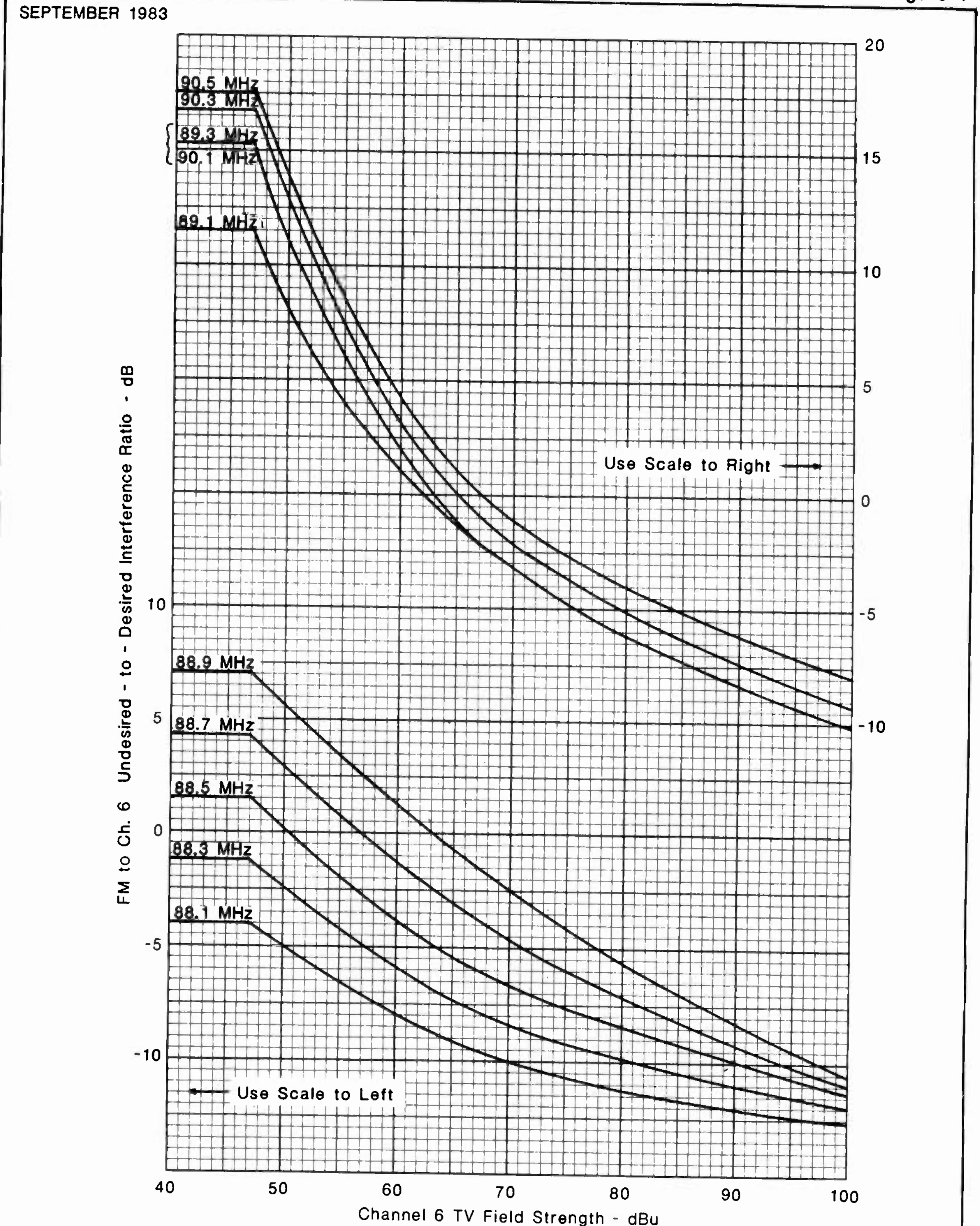
SEPTEMBER 1983



FM/TV 6 PROTECTION RATIOS
90.7 - 91.9 MHz

Jules Cohen & Associates, P.C. Consulting Electronics Engineers

Figure 1



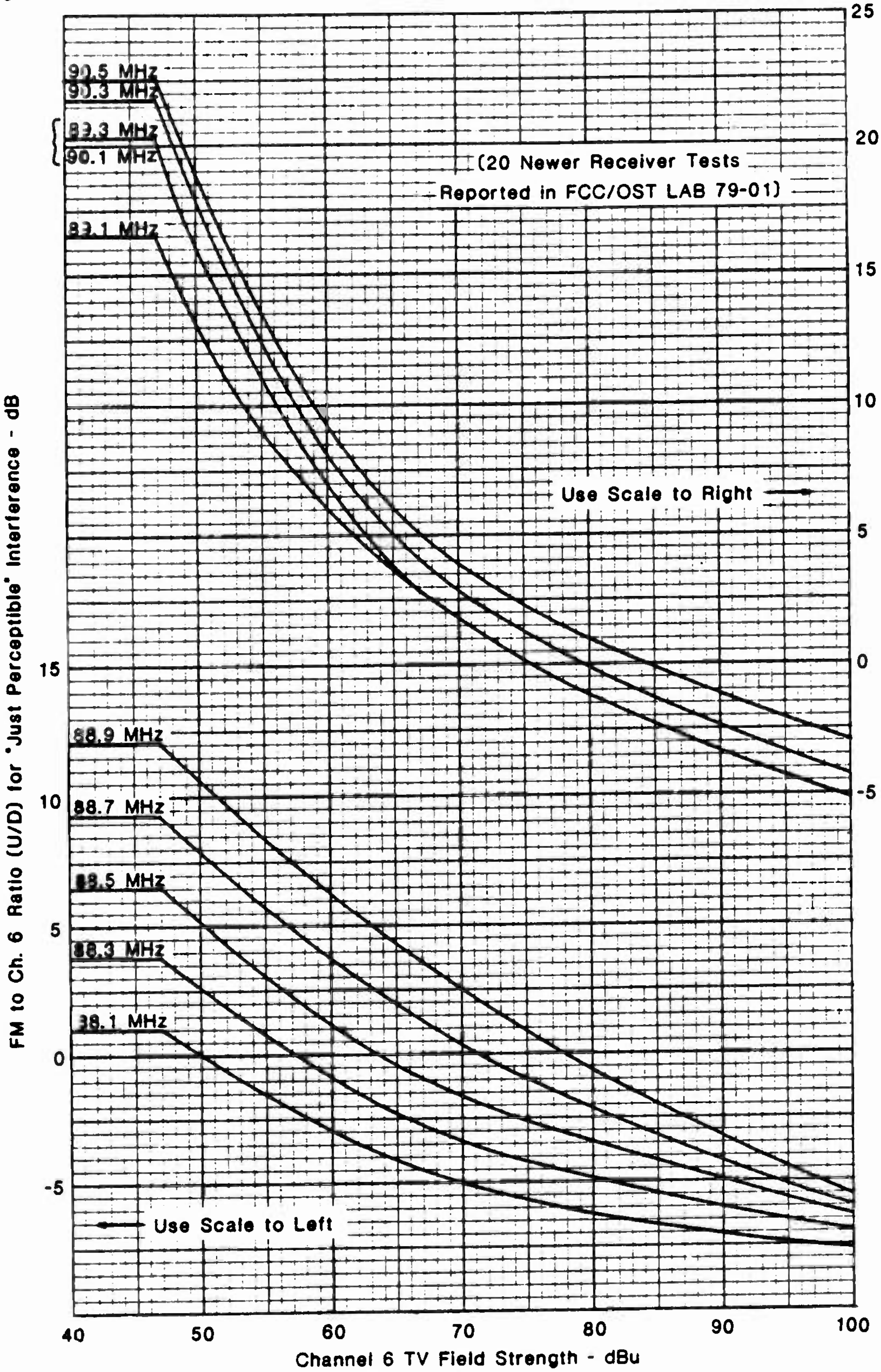
FM/TV 6 PROTECTION RATIOS

88.1 - 90.5 MHz

Jules Cohen & Associates, P.C. Consulting Electronics Engineers

Figure 3

AUGUST 1983

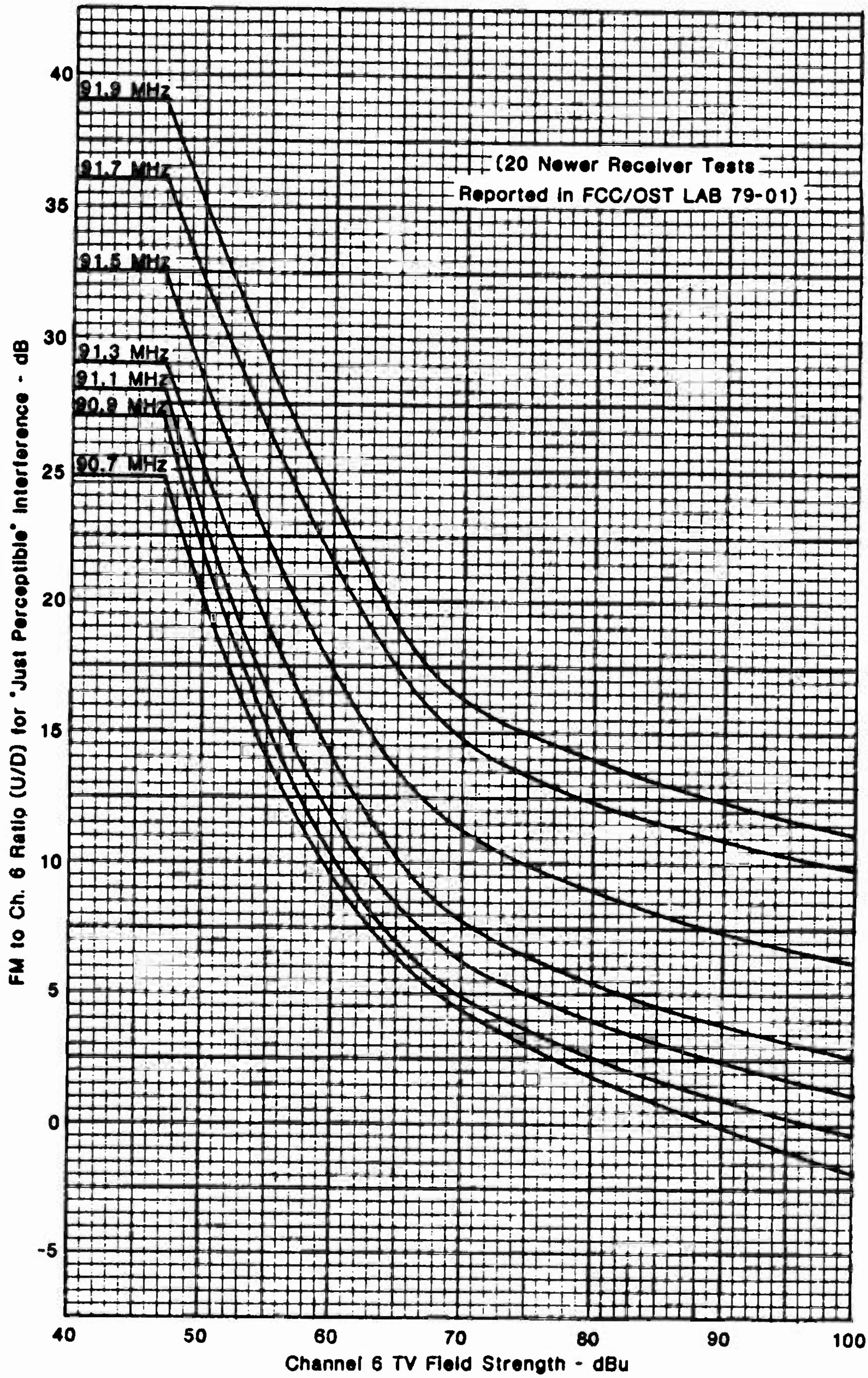


FM/TV 6 PROTECTION RATIOS
BASED ON MEDIAN RECEIVERS
88.1 - 90.5 MHz

Jules Cohen & Associates, P.C. Consulting Electronics Engineers

Figure 4

AUGUST 1983



FM/TV 6 PROTECTION RATIOS
BASED ON MEDIAN RECEIVERS
90.7 - 91.9 MHz

Jules Cohen & Associates, P.C. Consulting Electronics Engineers

An additional allowance is believed to be reasonable if the FM station is located in a rural area and the bearings to the television station and to the closest concentration of population are approximately 90 degrees. In general, allowance for the directivity of receiving antennas is not a wise idea because the reception pattern for an antenna in its usual rooftop location is quite different from what would be measured on a manufacturer's antenna range. The arrangement of the antenna range has been specifically selected to avoid reradiation effects from nearby objects and even from ground reflections. In the actual receiving location, reflections from the rooftop, trees, other buildings, power lines, water tanks and hillsides, all distort the reception pattern. However, in almost all instances, the pickup from directions approximately at right angles to the direction from which the maximum signal is received is suppressed. Where a showing can be made that the orthogonal relationship is achieved, it is believed that a 6 dB increase in the undesired-to-desired ratio would be appropriate.

Cross Polarization

An extensive test series was carried out by the firm of Cohen & Dippell, P. C. to determine the effect of cross polarizing the FM and TV signals. Tests were carried out in both urban and rural environments. Test results indicate that if a horizontally polarized receiving antenna is used, the television transmission is horizontally polarized and the FM station transmission is vertically polarized, discrimination of 16 dB is to be expected in rural areas and 10 dB in urban areas. In this case, the definition of "urban" has been taken to be a community with population of 50,000 or greater.

The question may be raised properly as to the effect on a circularly polarized television signal if the FM station operates with vertical polarization. The answer is believed to lie in the probability that, for the foreseeable future, outdoor antennas, particularly for VHF, are likely to remain horizontally polarized. A low VHF band circularly polarized antenna becomes quite bulky and difficult to install. As to indoor antennas which are generally benefited by a circularly polarized transmission, it is expected that the dissimilarity in polarization between the TV and FM signals is still likely to provide a setting which will discriminate more against the FM signal than would be the case with both signals with the same polarization, or even worse, with the TV signal horizontally polarized and the FM signal circularly polarized.

Omission of Rejection Filters

An allowance for the possible installation of rejection filters as proposed by the FCC is believed to be unwise. If the FM signal frequency is sufficiently removed from TV channel 6, the television receiver has a coaxial input and a good filter is installed by a competent technician, interference to the TV channel can be eliminated or reduced. However, experience shows that too few viewers sufficiently understand the problem and are motivated to have the necessary installation made. In addition, approximately 85 percent of the receivers currently in use employ balanced inputs, whereas the better rejection filters are all coaxial. The result is that even if the viewer understands the interference situation and seeks a remedy, he either installs a balanced stub tuner which is not likely to be satisfactory, or he is put to considerable expense to convert to an unbalanced coaxial input.

Conclusion

In this paper, specific criteria have been proposed for eliminating or reducing the problem of interference from FM stations to channel 6 through techniques of location, selection and cross polarization. Incorporation of an allowance for rejection filters is not recommended.

References

1. Federal Communications Commission; Notice of Proposed Rulemaking; Docket 20735; Vol. 41 Federal Register, page 16973, April 23, 1976.

2. Federal Communications Commission; First Report and Order; Docket 20735; Vol. 43 Federal Register, page 25821, June 15, 1978.

3. Federal Communications Commission; Second Report and Order; Docket 20735; Vol. 43 Federal Register, page 39704, September 6, 1978.

4. Federal Communications Commission; Further Notice of Proposed Rulemaking; Docket 20735; Vol. 43 Federal Register, page 27682, June 26, 1978.

5. Federal Communications Commission; Second Further Notice of Proposed Rulemaking; Vol. 47 Federal Register, page 24144, June 3, 1982.

6. Tests of TV Receivers for "Just Perceptible" Interference to TV CH. 6 from Educational FM Signals; Laboratory Division, Office of Science and Technology, FCC; FCC/OST Lab 79-01, September 1979.

Land Mobile Interference from TV Channels 14 and 69

Issues and Solutions

William C. King, Jr.

Jules Cohen & Associates, P. C.

Washington, D. C.

Most broadcast engineers have been confronted at one time or another with interference problems which required solutions. The first step in solving such problems is to determine the mechanism of the interference. Once the cause is known, an effective cure usually can be prescribed through physical isolation, through the application of simple filters, traps or shielding, or a combination of methods. In rare instances, where unfortunate frequency combinations result in direct interference to nearby assigned channels, a change of frequencies may be required. When interference problems involve comparable services, causes and cures usually are well understood and equitable solutions can be arranged. Moreover, objectionable interference to comparable services is relatively infrequent because allocation principles have been devised to avoid wherever possible circumstances that lead to serious problems.

Today, we live in a world of burgeoning demands for spectrum space. We have seen the loss to broadcasters of UHF channels 70 through 83 and the shared use with land mobile licensees of channels 14 through 20 in some large urban population centers. If land mobile channels are assigned to operate on adjacent channels, it is essential that appropriate engineering standards be devised where they do not exist, not only to reduce the potential for interference to other services, but also to enable licensees to understand the extent of their responsibility in the elimination of interference, should it occur.

If the public interest demands that full-service television stations and land mobile licensees operate in adjacent frequency bands to serve the same geographic area, then new engineering standards are needed for the protection of both services. Unless each service understands the needs and limitations of the other, and until appropriate allocation principles are devised and adhered to, interference conditions will arise which can be eliminated (if at all) only at great expense. Although interference to land mobile channels can occur at both ends of the UHF television spectrum, interference from channel 69 is aggravated and less easily eliminated by conventional means. Therefore, most of my remarks will deal with interference problems at the upper end of the spectrum.

Background

In reallocating the 806 to 890-megahertz band (formerly occupied by UHF television channels 70 through 83) to the Land Mobile Service, the Federal Communications Commission established two 15-megahertz bands for operation of conventional and trunked land mobile systems. Mobile and control stations are assigned to the band 806 to 821 megahertz (immediately adjacent to television channel 69). Base stations are assigned to the band 851 to 866 megahertz. Mobile stations are permitted by FCC rules to transmit with powers up to 100 watts, but typical mobile transmitters range in power from 10 to 30 watts. Base stations in urban centers are permitted to employ effective radiated powers up to 1000 watts and antenna heights above average terrain of up to 1000 feet (or more with appropriate adjustments in effective radiated power).

Because UHF propagation in the 806 to 890-megahertz band is most effective when line-of-sight conditions prevail, optimum coverage of an urban area usually results when the base station antenna is placed at the highest available elevation near the center of the market. Land mobile operations require that base stations receive (at the base station site) transmissions from mobile stations operating within the base station's service area. The sensitivity of base station receivers today is such that they are capable of receiving usable communications from mobile transmitters delivering signals of less than 0.5 microvolt to the receiver antenna input terminals. Base station antennas typically provide gains of from 7 to 9 dB (over a half-wave dipole). When one accounts for duplexer, transmission line and other losses and antenna gain, base station receiving installations are capable of providing satisfactory service when receiving mobile signal strengths of 5.4 microvolts per meter (14.6 dBu), or less.

In assigning frequencies to the Land Mobile Service, the FCC allocated the 15-megahertz band immediately adjacent to television channel 69 to mobile transmit (or base station receive) frequencies, commencing at 806.0125 megahertz (only 12.5 kilohertz above the upper edge of channel 69), and continuing upward at intervals of 25 kilohertz. Corresponding base station transmit frequencies are assigned to the 851 to 866-megahertz band commencing 45 megahertz (equivalent to more than seven television channels) above the lowest mobile frequency. Because base station receiving antennas usually are situated at a high elevation near the center of the service area, base station receivers are particularly susceptible to interference from any nearby channel 69 television station intended to serve the same market. Had the frequency bands for base and mobile operations been reversed, the potential for this type of interference would have been lessened.

Interference to Land Mobile Reception

Base station receivers situated in close proximity to a television channel 69 transmitting antenna are principally affected by three types of interference. The first type of interference results from out-of-band radiation from the television transmitter itself on or near the desired base station receive frequency. This kind of interference cannot be eliminated by filtering at the base station receiver because any attempt to do so would attenuate the desired mobile signals as well as undesired signals from the channel 69 transmitter. As will be demonstrated later, this kind of interference can occur even when the channel 69 transmitter is operating in accordance with FCC rules governing television operation. In fact, the problem is aggravated because no meaningful FCC standard exists to protect land mobile receivers operating in the band from 806 to 809 megahertz.

The second type of interference results from desensitization of the base station receiver by exceptionally high undesired signals not falling on the desired mobile frequency but originating within channel 69. This type of interference, which results from the reception of signals intended and expected to be transmitted during normal operation, cannot be filtered at the channel 69 transmitter. Any attempt to do so would impair channel 69 television transmission. Desensitization of the receiver occurs because the channel 69 signals, although separated from the desired land mobile frequency, are much stronger than the desired mobile signals and adversely affect receiver performance at the desired frequency. This kind of interference requires for its elimination reduction of the channel 69 signal strength at the land mobile receiver through filtering, if possible, through physical separation of the channel 69 transmitter and the land mobile receiver, or by increasing the frequency separation between desired and undesired signals.

The third type of interference is caused by the mixing of two or more strong undesired signals at the input to the land mobile receiver, resulting in the generation of intermodulation products which fall either on a desired frequency, or within the pass band of the receiver. Third order intermodulation products created by the mixing of the channel 69 visual, aural and color subcarrier frequencies can occur within the 806 to 815-megahertz band. Intermodulation interference originating in the receiver normally is eliminated by filtering one or more of the contributing frequencies at the land mobile receiver.

The use of cavity notch or band-pass filters to eliminate interference from nearby undesired frequencies cannot always be employed without sacrificing land mobile receiver performance. For example, if the undesired frequency is sufficiently near the desired frequency, a notch or band-pass filter may have a significant loss at the desired frequency and may not provide sufficient attenuation of the undesired frequency. If more than one filter is required for each receiver, as is likely in the case of a strong television signal, accumulated losses effectively reduce receiver sensitivity and limit the effective service range of the land mobile station.

UHF Transmitter Radiation Restrictions

FCC rules governing out-of-band radiation from television stations were developed primarily to provide adjacent channel protection to other television stations (normally spaced at least 55 miles apart at UHF) and were not modified to provide protection to the Land Mobile Service. The two sections of the rules applicable to out-of-band radiation from television stations are Section 73.687(a)(3) and Section 73.687(i)(1). Section 73.687(a)(3) applies to the visual transmitter only (not the combined visual and aural transmitters). Section 73.687(a)(3) requires that the field strength of the upper sideband not exceed a level of -20 dB for a modulating frequency of 4.75 MHz or greater, when measured in the manner prescribed in the rules. The -20 dB figure is not directly related to the visual carrier, but rather to the amplitude of the 200-kilohertz sideband. When the appropriate measurement procedure is employed, the -20 dB value with respect to modulating frequencies of 4.75 megahertz or greater is roughly equivalent to -38.0 dB when referred to peak visual carrier power.

Section 73.687(i)(1) requires that spurious emissions, including radio frequency harmonics, be maintained at as low a level as the state of the art permits, and that all emissions removed in frequency in excess of three megahertz above or below the respective channel edge be attenuated no less than 60 dB below the visual transmitted power. The Commission advises that the 60 dB value specified for television transmitters should be considered as a temporary requirement which may be increased at a later date. If interference is caused to any service, the FCC will require greater attenuation. The wording of subparagraphs (a)(3) and (i)(1) of Section 73.687 is the same wording which existed prior to reallocation of channels 70 through 83 to the Land Mobile Service. Moreover, the more stringent standard of Section 73.687(i)(1), which is the only standard applicable to combined visual and aural transmitters, applies only to frequencies in excess of three megahertz above the channel edge (809 MHz). Experience to date has shown that virtually all of the land mobile complaints attributable to the operation of channel 69 transmitters involve frequencies below 809 megahertz, in the region where there is no precise FCC requirement governing radiation from the combined visual and aural transmitters.

Illustrative Example

A hypothetical example can be used to illustrate the typical channel 69 interference problem. Assume that a channel 69 transmitter operates with peak visual effective radiated power of 5 megawatts (37.0 dBk) and with an aural effective radiated power of 500 kilowatts. Assume that a land mobile receiving antenna for the frequency 806.0125 megahertz is situated at a distance of three miles, in the major lobe of the channel 69 antenna, and so placed that free-space propagation conditions prevail (a not unreasonable assumption for land mobile receiving antennas). Assume further that the channel 69 visual transmitter surpasses by 6 dB the requirements of Section 73.687(a)(3) at 806.0125 megahertz. Then, under certain modulation conditions, the effective radiated power in the upper sideband due to the visual transmitter could be as high as -7.0 dBk (200 watts). Assume that the channel 69 transmissions are horizontally polarized and that the land mobile receiving antenna is designed for reception of vertically polarized signals. Typical cross polarization discrimination of land mobile receiving antennas is approximately 25 to 30 dB. Assume also that the land mobile receiver is capable of providing usable signals at desired field strengths as low as 14.6 dBu. Then, at the desired land mobile receiving antenna location, field strength from the channel 69 visual transmitter alone operating with 6 dB better suppression than required by FCC rules would be 56.3 dBu, or 41.7 dB greater than the desired land mobile signal strength. Obviously, reliable land mobile reception would be virtually impossible under such circumstances.

As noted earlier, this type of interference to the land mobile operation cannot be cured by filtering at the land mobile receiver. In order to reduce the received out-of-band transmissions from the channel 69 transmitter to a value even 6 dB below the desired land mobile signal strength, additional filtering at the channel 69 transmitter of more than 47 dB would be required. Since the Commission's rules require (and color fidelity impels) that the channel 69 transmitter visual response be essentially flat to at least 4.1 megahertz above the visual carrier, virtually all of the filter's attenuation would have to be provided in approximately 0.08 percent of the land mobile frequency, with negligible insertion loss at frequencies below 805.35 megahertz. Furthermore, comparable attenuation would have to be provided, not only at the single frequency of 806.0125 megahertz, but throughout the remainder of the land mobile frequency band. Television state of the art is such that it is not feasible to build a high-power filter capable of satisfying such stringent requirements. Even if that were possible, such a filter could not be used at the output of diplexed visual and aural transmitters for obvious reasons. At three megahertz above the channel edge, out-of-band radiation from the combined visual and aural transmitters necessarily would have to be attenuated to a level at least 92 dB below peak visual carrier in order not to exceed an effective signal strength at the land mobile receiver 6 dB below the minimum desired signal.

At the channel 69 aural carrier (only 0.033 percent below the desired land mobile frequency) the channel 69 aural carrier field strength would exceed the minimum desired land mobile field strength by more than 75 dB. If filtering of the aural carrier were necessary to achieve satisfactory land mobile reception, another problem arises. Cavity notch filters of the type often employed at lower frequencies are not applicable for frequencies in the 806-megahertz band because they cannot meet such stringent requirements. Instead, band-pass filters are customarily employed to reduce undesired out-of-band emissions affecting land mobile receivers. A typical dual-cavity band-pass filter for the 806-megahertz band would have an insertion loss of 4.0 dB at the desired land mobile frequency and produce a relative attenuation of only 6 dB at the channel 69 aural carrier, 0.26 megahertz below the desired land mobile frequency. Obviously, such a filter would be relatively ineffective. As this hypothetical model demonstrates, an incompatible situation exists which cannot be resolved without substantial compromise to either or both operations.

An FCC Admonition

On March 1, 1982, the FCC issued a Public Notice to channel 14 and channel 69 television permittees, reminding them of their obligation to protect existing land mobile facilities on adjacent frequencies from objectionable interference. In that Public Notice, the Commission cited Section 73.687(i)(1) of the Rules requiring permittees to avoid such interference to other services. However, as already noted, that Section is inapplicable to frequencies below 809 megahertz, where most interference has been experienced. In its March 1, 1982, Public Notice, the Commission stated that the grant of any construction permit for a new TV station or any license for modification of facilities on channels 14 or 69 will be conditioned to require that the permittee take adequate measures to provide protection against objectionable interference to existing land mobile facilities. Finally, in a warning to land mobile licensees, the Commission said: "Furthermore, land mobile applicants are advised that the frequencies immediately adjacent to TV Channels 14 and 69 may receive objectionable interference which cannot be completely eliminated [emphasis added]."

Engineering analysis shows that the channel 14 and channel 69 interference problems are not parallel situations. The frequency separation between the nearest television carrier and the nearest land mobile carrier expressed as a percentage of the land mobile frequency is nine times as great at channel 14 as at channel 69. A typical land mobile dual-cavity band-pass filter is 36 dB more effective (with 1 dB less insertion loss) at the nearest television carrier frequency at channel 14, than at channel 69. Attempts to sharpen filter response at frequencies above 806 megahertz result in increased insertion loss at the desired land mobile frequency, thus limiting performance of the land mobile base station receiver.

Where Do We Go From Here?

In summary, physical separations on the order of several miles may not be sufficient to avoid interference to land mobile channels nearest to the upper edge of channel 69. FCC rules governing out-of-band radiation for combined visual and aural television transmitters apply only to frequencies greater than three megahertz above the channel edge. It is entirely possible for television broadcast transmitters to comply fully with FCC regulations and still cause interference to nearby land mobile base station receivers operating on channels immediately adjacent to channel 69. Once physical plants have been constructed, the cost of achieving satisfactory physical or frequency separation can become substantial. Obviously, this kind of interference is more easily prevented than cured.

What is needed, of course, is a set of allocation standards governing the assignment of land mobile frequencies immediately adjacent to television channels 14 and 69 where those channels are allotted to the same geographical area. In an era of deregulation, additional standards are not likely to be promulgated by the Federal Communications Commission unless it is urged to do so. However, some sort of joint industry effort involving participation of engineers from both the broadcasting and the land mobile industries is needed in an effort to propose practical allocation standards capable of reducing the potential for interference, without inhibiting the full development and implementation of the frequency bands available to both services. Absent standards, industry engineers and licensees must realize that serious interference problems may occur which may not be easily eliminated by simple, relatively inexpensive, methods.

Broadcast Interference to Aeronautical Radionavigation and Communications

Ralph H. Justus

National Association of Broadcasters

Washington, D.C.

The television and FM radio broadcast services can pose interference problems to the proper operations of certain aeronautical radionavigation and communications facilities. This paper presents: an overview of the frequency management principles developed to date to protect the aeronautical services; activities both domestic and international, which seek to provide some degree of compatibility between these services; and how these developments may affect broadcasters.

I. Introduction

FM broadcasting (88-108 MHz) and TV broadcasting (54-72, 76-88, 174-216, and 470-806 MHz) facilities are potential sources of intermodulation and harmonic interference, as well as spurious and single frequency overload interference, to ground and airborne VHF (108-136 MHz) and UHF (239-399 MHz) air traffic control (ATC) communications and radionavigation receivers. This can cause communications problems ranging from nuisance background FM broadcasting audio to distorted and garbled reception. In airborne instrument landing system (ILS) localizers and VHF omnirange (VOR) receivers, the interference can produce the more serious problem of errors in course deviation, especially during the critical approach and landing phase, which would not be readily evident to the pilot.

Much of the literature on this subject has developed since the late 1970's, and has led to the adoption of domestic frequency management principles and practices intended to protect the aeronautical services. Internationally, the International Telecommunications Union (ITU) is in the process of determining the technical constraints to be used in planning for the 1979 WARC extension of the FM band up to 108 MHz in Regions 1 and certain parts of Region 3, and is looking towards the development of compatibility criteria to be applied on a world-wide basis.

Broadcasters need to become acquainted with these problems in order to be receptive to the needs of the aeronautical services. With the addition of hundreds of new FM assignments created by Docket 80-90, and with the upgrading of existing facilities, new elements of design constraint may need to be respected to insure at least minimal compatibility between these services.

II. FAA Guidelines and Procedures

The Federal Aviation Administration (FAA), in assigning frequencies to VHF/UHF air/ground communications facilities, has adopted guidelines to evaluate the effects of high powered TV and FM broadcasting stations (and other potential interference sources) on proposed frequencies.[1&2] Conversely, these same guidelines are used to evaluate the effects of proposed new or modified FM and TV stations on existing air/ground communications assignments. Some familiarity with these guidelines, and the engineering assumptions behind them, is then in order. (The analysis below is from the viewpoint of the FAA determining the availability of frequencies for its use. This analysis does not discuss the FAA's own intra-service frequency environment and analysis).

A. Co-site Interference to Ground Receivers

The effects of co-site interference usually appear in ground receiving equipment (except for the case of FM interference to airborne receivers, discussed later) and result from the effects of facilities only a short distance away. Because of their relatively high powers, all FM and TV broadcasting stations within 15 nmi are listed by the FAA as defining the "interference environment" for a subject frequency. The subject frequencies are then analyzed for potentially interfering harmonics, intermodulation (IM) products, and high powered effects from broadcast stations.

Protection of VHF/UHF communications ground receivers from FM & TV broadcast stations is a less formidable problem than protection of airborne receivers. The transmitter and receiver sites are known, and corrective measures are easily employed. These include: additional filtering of harmonics and spurious emissions at the transmitter output; filtering of spurious signals and intermodulation products at the receiver input; and selective filtering at the receiver input to avoid single frequency front-end overload.

1) Harmonic Study

Should the broadcast harmonics in Table 1 fall on the desired subject frequency, that frequency is then precluded from assignment.

Table 1

Subject Frequency	Broadcast Station Preclusion
VHF Assignments	VHF-TV: Aural and Visual 2nd Harmonic
UHF Assignments	VHF-TV and FM
	Ch. 7-11: 2nd Harmonic
	Ch. 5,6 & FM: 3rd Harmonic
	Ch. 2-6 & FM: 4th Harmonic
	Ch. 2-5: 5th Harmonic

2) Intermodulation (IM) Study

Third order IM products for every 2 and 3 signal combination of broadcast frequencies existing in the co-site environment in the general form:

$$fo = 2f1 - f2, \text{ or } \quad fo = f1 + f2 - f3$$

would potentially preclude assignment of a subject frequency. However, many of these IM frequency combinations may not have sufficient signal strength to cause interference. They are then analyzed for distance and power to determine if the power level of the interfering signal at the input to the victim ground receiver exceeds a -97 dBm threshold (for ATC communications receivers) or -87dBm threshold (for navigation receivers).

3) High Powered Effects

a. Spurious Interference:

Because of the greater frequency separation from the TV bands, spurious interference from TV stations to ground receivers is unlikely. However, since the FM band is adjacent to aeronautical frequencies, potential spurious interference from FM stations is possible, and depends on the combination of three major factors: receiver sensitivity, FM spurious emission limits, and FM station power. Spurious interference could result if:

$$EIRP - L_d - L_a - L_p - L_r - S_r = RS$$

where:

EIRP = Equivalent isotropic radiated power of the station in dBm. EIRP = ERP + 2.2 dB.
L_d = Free space transmission loss in dB at the ATC receiver frequency. L_d = 37.8 + 20 log (MHz) + 20 log (nmi.).
L_a = FM transmitting antenna vertical directivity (assumed 0 dB if unavailable).
L_p = Loss due to difference between broadcast and communications antenna polarizations (3 dB for horizontally polarized FM, or 0 dB for cross-polarized FM).
L_r = Receiver system on-frequency loss in dB (assumed 3 dB);
S_r = FM spurious radiation limit (assumed 80 dB for frequencies removed from the FM carrier by more than 600 kHz);
RS= Receiver sensitivity.

Example 1

As an example, assume an FM station proposes to operate on 97.5 MHz at 50 kw ERP from a transmitter site 2.0 nmi. from an ATC ground receiver on 118.3 Mhz at an airport. Then:

EIRP = 79.2 dBm (for 50 kw ERP)
L_d = 37.8 + 20log(118.30) + 20log(2) = 85.3 dB
L_a = 0 dB (information unavailable)

L_p = 0 dB (FM cross polarized)
S_r = 80 dB
RS = -97 dBm (typical)

Thus, 79.2 dBm - 85.3 dB - 3 dB - 80 dB = 89.1 dBm, approximately 8 dB greater than the typical ATC receiver sensitivity, and spurious interference is possible. The FM station could correct this by: moving its site to one 5.0 nmi away to provide 8 dB more free space loss; reducing power 8 dB,or; reducing its spurious emissions to 88 dB by filtering.

b. Single Frequency Front End Overload ("Brute Force" Interference)

The IF selectivity of the ground receiver will not provide any protection from single frequency front end overload because this effect occurs in the receiver RF section which will respond to most frequencies within the FM and TV broadcasting bands. A signal of from 0 to -10 dBm at the input to most receivers is sufficient to overload the RF section.

The FAA calculates the received power at its ground site from each FM and TV station within 5 nmi using the following equation:

$$P_r = EIRP - L_d - L_a - L_r - L_p$$

where:

P_r = The received power in dBm of the FM or TV signal at the FAA facility assuming a lossless isotropic receive antenna having no rejection of signals in the FM or TV bands;
EIRP = The equivalent isotropic radiated power of the FM or TV station in dBm. (EIRP = ERP + 2.2 dB);
L_d = Free space transmission loss in dB at the broadcast station frequency.
L_a = FM/TV transmitting antenna vertical directivity (assumed 0 dB if unavailable)
L_r = Receiver system on-frequency loss in dB. Assumed to be 3 dB.
L_p = Loss due to difference between broadcast and communication antenna polarizations. (3 dB for horizontally polarized FM/TV, or 0 dB for cross-polarized FM/TV).

The out-of-band rejection necessary to protect the ground receivers would be calculated as follows:

$$R = P_r\ (max) - (-10)$$

where:

R = The required out-of-band rejection in dB.
P_r (max) = The maximum received power at the FAA ground receiver site of all the broadcast stations within 5 nmi. in

dBm (Root-Sum-Squared for multiple stations).

The required amount of external out-of-band rejection can be obtained by the use of a more selective antenna, cavity filters, crystal filters, etc., as well as by distance separation of broadcast and ATC Communication sites.

B. <u>Protection of the Airborne Receiver from FM Interference</u>

Interference to airborne receivers is considerably more complex. Their locations continuously vary and their performance characteristics (selectivity, sensitivity, and rejection) differ enormously, ranging from the low performance general aviation equipment to the more sophisticated and immune equipment used by commercial air carriers. The FAA guidelines have, understandably, attempted to protect the "worst-case" performance level of general aviation equipment.

Interference to ATC communications is intermittent as an aircraft's location changes and passes through, perhaps, a maximum interference environment. A noisy or garbled transmission can simply be repeated. Interference to navigation receivers (ILS or VOR) poses a more serious problem as course deviation errors can result during critical approach and landing phases.

1) Intermodulation Interference

FM intermodulation usually occurs when the first RF stage of the airborne receiver is driven into nonlinear operation by the high power level of at least one FM station. When in nonlinear operation, the receiver RF is very susceptible to intermodulation from any combination of signals within the RF bandpass.

The FAA avoids this by precluding assignments of frequencies which are 2 and 3 signal third order IM products of existing FM stations, as well as 5th order combinations since they may become significant once the airborne receiver has been driven into the nonlinear region by a strong signal. The addition of fifth order products is intended to allow additional protection for the more susceptible airborne receivers.

Here, the FAA recommends extending its limit on defining the "interference environment" to consider all FM stations within 30 nmi. It then calculates all 2 and 3 signal third and fifth order intermodulation products of the form:

$$fo = 2f1 - f2$$
$$fo = f1 + f2 - f3$$
$$fo = 3f1 - 2f2$$
$$fo = 2f1 + f2 - 2f3$$

These frequencies are then precluded from assignment.

However, simply "precluding" assignment of a frequency is becoming a luxury the FAA cannot afford. Its available spectrum is becoming congested. If assignment of a frequency with an FM IM product is unavoidable, if a new FM station is planned, or if an existing FM station is being changed, the following Venn diagram approach is used to determine the potential for harmful interference to airborne receivers.

<u>Venn Diagram</u>

This assumes, based on test results of communication receivers, that undesired (FM) signal levels as low as -30 dBm can cause harmful interference to ATC communications receivers if another FM signal is also present at a level of -10 dBm.[3] In the case of navigation receivers, a -30 dBm FM signal can cause interference if another FM signal is present at a -20 dBm level. To determine where these levels exist, the -10 dBm, -20 dBm, and -30 dBm contours of the FM stations for the IM combination of interest are plotted using the following formula:

$$d = \frac{\log^{-1}\left((EIRP-P-C-L_r)/20\right)}{f}$$

where:

d = Contour radius in nmi.
EIRP = Effective isotropic radiated power of the FM station in dBm (EIRP = ERP + 2.2 dB)
P = Contour power level: -10, -20, or -30 dBm.
C = 37.8 for d in nmi.
L_r = Antenna rejection factor equal to:
<u>For communications receivers:</u>
10 dB for FM stations between 100 and 108 MHz and 10 dB + 2 dB/MHz for each MHz below 100 MHz for FM stations between 88 and 100 MHz.
<u>For navigation receivers:</u>
3 dB plus 1 dB/MHz below 108 MHz.
f = FM station frequency in MHz.

Peculiar to this analysis is the value set for L_r, the out-of-band rejection factor for the antenna system. The FAA defines the range of possible areas of interference by drawing contours both excluding and then including L_r. As the example below demonstrates, this leads to widely varying results.

<u>Example 2</u>

FM station #1 (107.9 MHz, 100 kw) and FM station #2 (103.9 MHz, 100 kw) are situated near an airport with an ILS localizer operating on 111.9 MHz, as shown in Figure 1. Where the FM stations' -20 dBm and -30 dBm contours overlap, intermodulation interference is likely.

Excluding L_r:

$$d_{\#1}(-20) = \frac{\log^{-1}((82.2 - (-20) - 37.8)/20)}{107.9} = 15.3 \text{ nmi.}$$

$$d_{\#2}(-30) = \frac{\log^{-1}((82.2 - (-30) - 37.8)/20)}{103.9} = 50.5 \text{ nmi.}$$

Including L_r:

$d_{\#1}(-20) = 10.9$ nmi, $L_r = 3$ dB

$d_{\#2}(-30) = 22.5$ nmi, $L_r = 7$ dB

These contours are then plotted as shown in Figure 1.

Where this predicted interference area covers an airport, an approach or departure path, or other areas of high air traffic density, the candidate frequency is not assigned by the FAA. If this study pertains to a new FM station, the FAA will file objections to the application for construction permit with the FCC.

2) Spurious or Front End Overload Interference

Protection of airborne receivers from spurious or front end overload interference from a single FM station is "achieved" by protecting airports and ILS approach paths from a -10 dBm FM contour. The FAA does this by objecting to FCC applications for new or modified (increased power or height) FM stations within a 5 nmi radius of an airport. This is derived from the Venn Diagram distance formula, above, which shows that the -10 dBm signal from an FM station on 107.9 MHz at 100 kw extends 5.0 nmi.

C. Antenna Rejection Factor

A common theme throughout the literature is the rejection capability of airborne receiver systems. Obviously, the frequency rejection performance of receiver systems impacts on the protection criteria. As shown in Example 2, above, the addition of a 3 and 7 dB antenna rejection figures dramatically affects the predicted interference area.

The antenna rejection factor values used in the Venn Diagram procedures are derived from studies of attenuation characteristics of a representative antenna with variable transmission line lengths.[4] See Figures 2 and 3. As the transmission line lengthens, the number of peak responses increases while simultaneously their amplitude decreases.

Also common throughout the literature is the recognition that a 10 dB increase of rejection capability of FM signals in airborne receivers would nearly eliminate FM intermodulation interference. For the worst-case condition (20 foot cable) the addition of a notch filter produced a significantly improved

FIGURE I
PREDICTED INTERFERENCE TO NAVIGATION RECEIVER

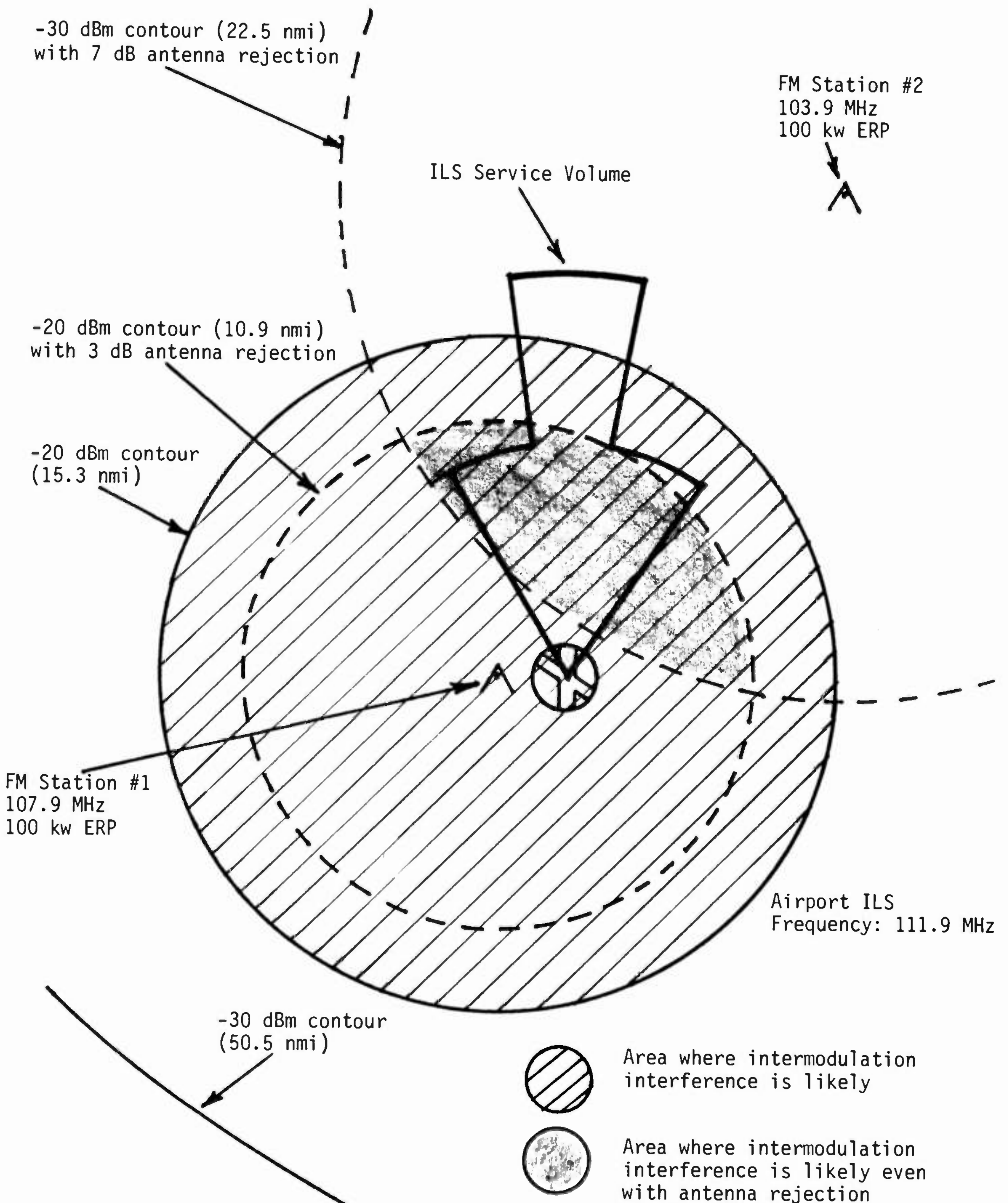


Figure 2 Antenna System Gain as a Function of Transmission Line Length: 20 Feet

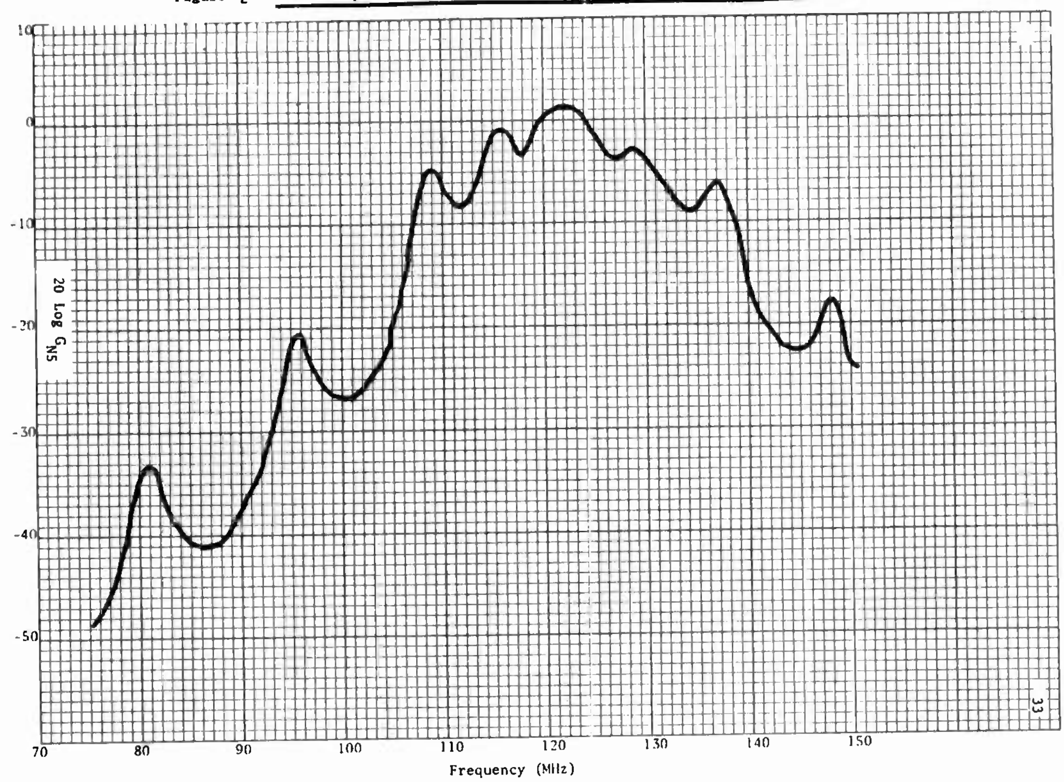


Figure 3 Antenna System Gain As a Function of Transmission Line Length: 118 Feet

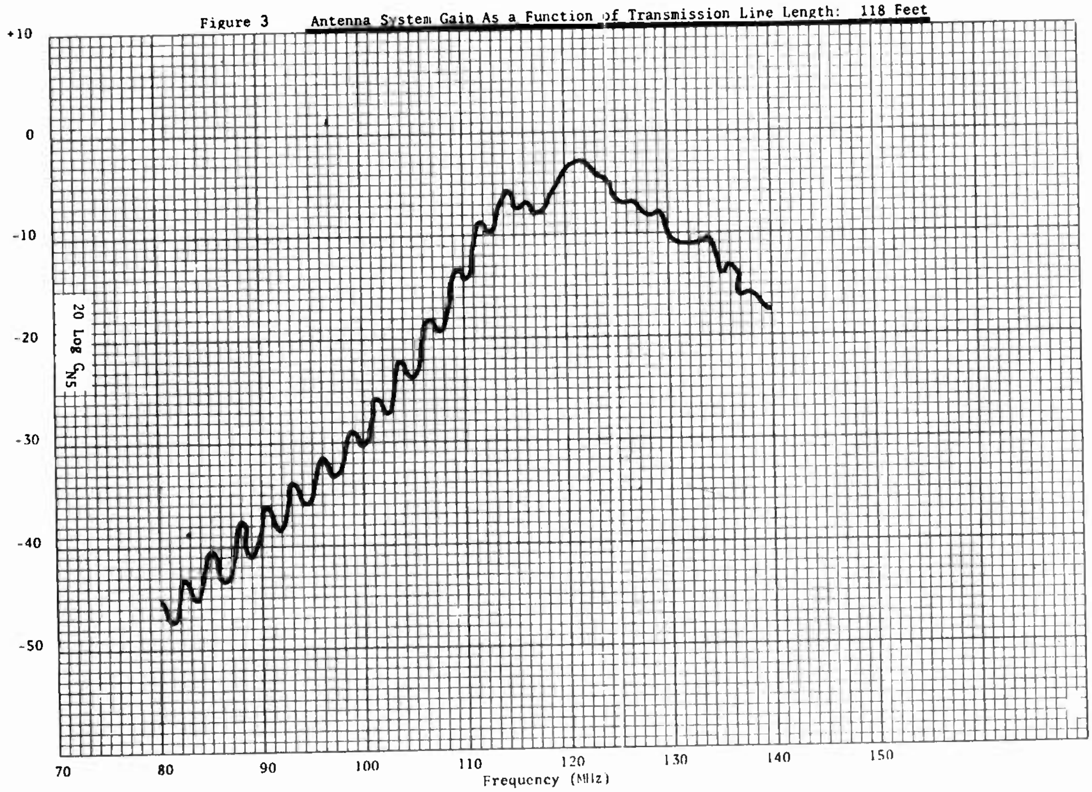
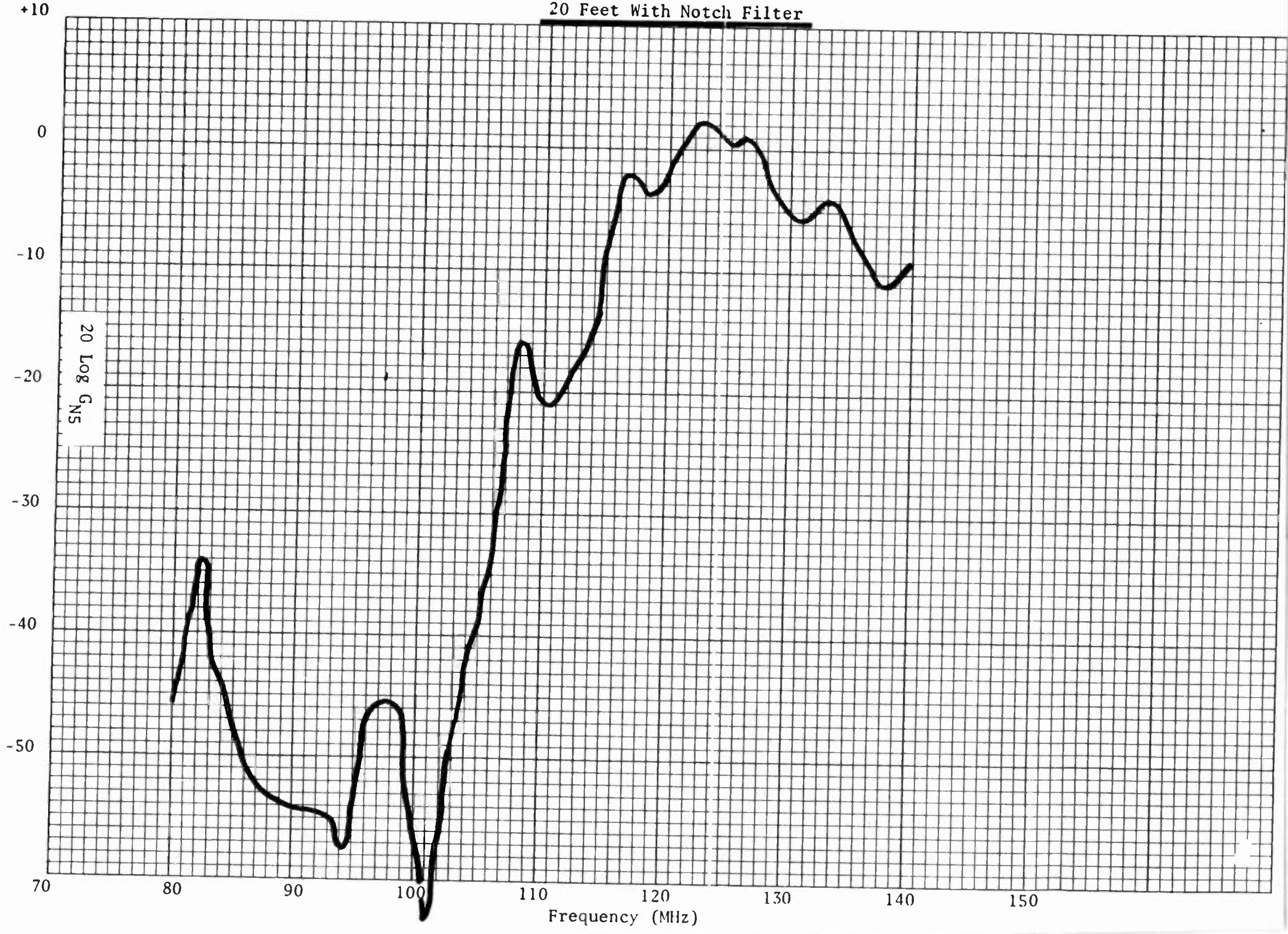


Figure 4 Antenna System Gain As a Function of Transmission Line Length: 20 Feet With Notch Filter



receiver immunity to third order intermodulation. See Figure 4.

The most prevalent form of interference (FM intermodulation) to the most critical avionics equipment (ILS and VOR receivers) can essentially be eliminated by improvement of this one factor - antenna system rejection. The effectiveness of a notch filter to cure receivers to the point of immunity has been demonstrated. Yet, the Radio Technical Commission on Aeronautics (RTCA) suggests that "the use of a notch filter must not be misconstrued as a general solution to interference problems."[4] It concludes that distance criteria, as well as frequency planning, is required to limit potentially interfering FM signals to a tolerable level in the vicinity of airports and approach paths. Accordingly, the FAA has adopted the guidelines presented above along with heightened coordination efforts with the FCC.

III. Future Domestic Activities

A. The FAA is preparing a Notice of Proposed Rulemaking, affecting Part 77.15 of its rules, which may expand its notification procedures to require prior notification if a proposed new broadcast construction or alteration of a tower (or facility) is:

1) above 30 MHz with an effective radiated power greater than 10 kw;
2) a change of frequency or ERP if the facility is within 3,000 feet of FAA ground facilities.

This would provide the FAA with a formal mechanism by which it would be notified of new and modified FM and TV applications. The FAA would then study these proposals for their potential interference impact on aeronautical communications and navigation facilities. It can then issue its determination of "Hazard" or "No Hazard" based on its study.

Previously this determination by the FAA has been limited to the physical impact on air space by a proposed tower structure. The FAA's intent behind these proposed rules is to alert broadcast applicants to potential problem areas.

B. The RTCA has recommended that FAA Advisory Circulars be issued warning the aviation community of potential limitations of some airborne equipment, and to encourage improvements in the performance of ILS and VOR receivers in the FM environment by both add-on devices, such as filters, and by an overall change in the minimum performance standards. However, little if any activity in these areas has taken place in the three years since the RTCA report was issued.

IV. International Developments

In 1979, the World Administrative Radio Conference (WARC) extended the VHF/FM broadcast band up to 108 MHz for Region 1 (Western Europe and Africa) and certain countries in Region 3.

The International Telecommunications Union (ITU)[5], in 1982, recognized that this might lead to interference problems for aeronautical radionavigation services in adjacent frequencies, and adopted certain technical criteria for broadcasting frequency assignment planning purposes which was intended to protect the aeronautical services. However, these criteria would impose severe restrictions on planning for FM broadcasting. The ITU recommended that the CCIR carry out further studies to determine what practical improvements in the immunity of airborne receivers can be made, and what degree of reduction of FM broadcast spurious emissions is feasible. These studies are being conducted now and the results will be reported to the second session of the Regional Administrative Conference for FM Sound Broadcasting to be held in Geneva in May, 1984.

The technical criteria adopted by the ITU in 1982 was essentially the Venn Diagram approach described above. However, it was found to be too restrictive for realistic FM broadcast planning in Region 1. For example, the only feasible transmitter sites for new FM stations in the United Kingdom were found to be either in the English Channel or in the Atlantic Ocean. Region 1 has a higher density of ILS and VOR installations, and consists of a large number of relatively smaller countries than in North America. The ITU believes that these differences might well justify different approaches to its frequency planning. The CCIR studies to date have recommended that the ITU adopt more stringent FM spurious emission attenuation requirements - at least 85 dB for transmitter powers above 8 kw.[6] Improvement of the immunity performance of airborne receivers was discussed at length with inconclusive results, except for the acknowledgment that add-on filters could improve a receiver's immunity from FM interference. However, this was not deemed at the time to be a practical solution.[7]

The CCIR has formed a new Joint Interim Working Party (JIWP 8-10/1) to determine if any improvement in compatibility between the FM and aeronautical services can be achieved, and to provide final guidance to the second session of the Conference. The JIWP will consider, further, that compatibility criteria may have to be applied on a world-wide basis.

V. Conclusions

The FAA guidelines were derived from the best available information, and provide a structured methodology to evaluate the potential for interference from nearby broadcast facilities. Guidelines, however, need not be strictly adhered to if it can be shown that the predicted interference can be eliminated by other means.

The FAA may determine that a proposed broadcast facility constitutes a "hazard" to air navigation, based on interference, and may formally object to the FCC grant of the broadcast construction permit application. However, if the broadcaster can show that it can eliminate the predicted interference by, for example, additional filtering of its out-of-band emissions or filtering of its signal at a ground receivers input, the FAA may be persuaded to remove its objections.

But there is little more a broadcaster can do to protect susceptible airborne receivers from interference, outside of choosing its transmitter site to avoid heavy air traffic areas. For this, we must look to the aeronautical industry to improve the immunity of its receivers. But this approach has, curiously, met with resistance. The recommendations of RCTA to employ effective add-on filters to existing receivers, and to re-examine the performance standards to improve the interference immunity of future receivers, have largely been ignored. However, these improvements would provide the means for a long range improvement in compatibility, and must be considered further.

It has been noted in international proceedings that there is a regrettable lack of contact between the work of the aeronautical and broadcast communities on improving compatibility between their services. One can read a certain degree of inter-service animosity between the lines of the CCIR reports. To the degree that an adversarial relationship exists between the broadcasting and aeronautical industries, the freedom to deploy their respective facilities with minimum restrictions is in jeopardy. This is unconscionable in an engineering arena.

It is unrealistic for the aeronautical industry to "blame" broadcasters for its interference problems, and insist on drastic broadcast power reductions, and distance/frequency separations, without considering the immunity improvements possible in its own receivers. It is equally unreasonable for the broadcast industry to claim "eminent domain" of allocated spectrum and rest entirely on improvements of the deficiencies in aeronautical receiver design as <u>the</u> solution to compatibility.

If compatibility between the broadcast and aeronautical services is to be achieved, a single set of protection criteria must be developed. But that will not happen without a reasonable degree of compromise and cooperation. Broadcasters may have to accept some constraints on choosing their transmitter sites to avoid airports and approach paths. They may also have to be receptive to reducing out-of-band spurious emissions well beyond what the FCC rules require.

On the other hand, the aeronautical industry must seriously consider improved receiver performance standards. Protection criteria are unnecessarily complicated by the wide variability of aircraft equipment characteristics. The lower performance levels of the general aviation equipment increases its susceptibility to interference, while keeping costs low. This may be a luxury which can no longer be afforded.

The JIWP 8-10/1 has an opportunity to make substantial and meaningful progress in the development of compatibility criteria between these services which may result in a world-wide standard. NAB looks forward to adoption of protection criteria in the near future which reasonably accommodate the needs of both the broadcasting and aeronautical services.

References:

1) FAA Order 6050.4B, Frequency Management Principles: Criteria and Procedures for Assigning VHF/UHF Air/Ground Communications Frequencies, Federal Aviation Administration, October 19, 1981.

2) FAA Order 6050.5B, Frequency Management Principles: Geographical Separation for NAVAID Frequency Assignments, Federal Aviation Administration, May 19, 1980.

3) Sawtelle, E.M. and Doug, J.G., Interference in Communications and Navigation Avionics from Commercial FM Stations, Final Report, FAA-RD-78-35, July, 1978.

4) RCTA/DO-176, FM Broadcast Interference Related to ILS, VOR and VHF Communications, Radio Technical Commission for Aeronautics, November, 1981.

5) Recommendations and Reports of the CCIR, 1982, Report No. 929, "Compatibility Between the Broadcasting Service in the Band of About 87-108 MHz and the Aeronautical Services in the Band 108-137 MHz", pg. 786, XVth Plenary Assembly, Geneva, 1982, Vol. VIII.

6) CCIR IWP 10/8 Report to Study Group 10, Doc. 10/55 (Rev. 1), September 23, 1983, "Improvement of the Level of Spurious Emissions Falling in the Frequency Bands Allocated to the Aeronautical Services from FM Broadcast Stations."

7) CCIR IWP 8/12 Report to Study Group 8, Doc. 8/13, August 2, 1983, "Improvement of the Immunity of Airborne Radionavigation Equipment to Interference from FM Broadcasting Stations."

A FRAMEWORK FOR A DECENTRALIZED RADIO SERVICE

Alex D. Felker
Kenneth Gordon[1]

Federal Communications Commission
Washington, D.C. 20554

In recent years it has become apparent that the Federal Communications Commission's basic spectrum management framework is increasingly strained and unwieldy in the face of a mushrooming demand for all types of communications. In the past, the FCC has been able to satisfy most requests for spectrum by either finding new (higher) frequencies or by helping new users reach a reasonable sharing arrangement with existing licensees. It is clear, however, that the Commission's ability to continue in this *ad hoc* fashion and still ensure that the underlying goals of the Communications Act are met as fully as possible has diminished sharply. Recently, some Commissioners have expressed interest in testing a *decentralized* spectrum management scheme, where users, not the FCC, would determine how spectrum should be used in an area. The staff has been instructed to prepare a concrete proposal to test the concept. This paper briefly summarizes the key policy and technical issues that will be addressed in the proposal.[2]

We begin by examining the problems of the present spectrum allocation scheme. This is followed by a summary of the decentralized allocation proposal. This section examines technical issues, outlines a plan for controlling interference at the license assignment stage, and identifies the major policy issues raised by the proposal. We conclude the paper with a summary and a list of important questions to be asked in a Notice of Proposed Rulemaking.

---

1 The views expressed in this paper are the authors' and do not necessarily reflect those of the Federal Communications Commission.

2 This paper is only a summary. For a more in depth analysis, the reader is directed to: Staff Report on *A Framework for a Decentralized Radio Service*, Alex Felker and Kenneth Gordon, Office of Plans and Policy, Federal Communications Commission (September 1983).

OVERVIEW

There are three specific areas where the current spectrum management system is ineffective: evaluating competing claims for spectrum; accounting for regional variation in spectrum use; and facilitating the introduction of new technologies. Each of these is discussed briefly below.

Evaluating Competing Claims for Spectrum

Pressures on available spectrum arise from several sources. First, the revolution in computer and communications technology has expanded enormously our ability to create and effectively use large quantities of information. The need to link the various users of this information -- locally, nationally, and even internationally -- is an important source of increased demand for spectrum. Mobile communications uses have also grown recently -- both as a result of the availability of better communications technology and because higher energy costs make it important to reduce inessential travel and direct personal contact. Finally, recent Commission actions that have increased the range of video delivery options have revealed the existence of a substantial latent demand for entertainment. The result of all these factors has been serious competition -- and at times acrimonious debate -- among the various claimants for spectrum.

The root of the problem is that the Commission has little independent and objective information by which to evaluate the competing claims for spectrum.[3] It must rely on its intuition and historical expertise, a process that worked tolerably well as long as unused spectrum was available. Those days are over. The FCC needs ways to choose among competing applicants efficiently, to encourage more intensive use of spectrum that has already been allocated, and to facilitate reallocation of spectrum as new uses arise and older uses become less important. Further refinement and tuning of the existing mechanism is not a long run solution. As long as spectrum is treated as a 'free good,' except for the administrative expenses of applying for it, no applicant has an incentive to limit his demand and seriously consider alternatives -- including making do with less spectrum. In some way, the

---

3 This problem is illustrated by the fact that there are three outstanding Notices of Proposed Rulemaking that propose to reallocate most of the land mobile reserve spectrum. (See, Notice of Proposed Rulemaking, Docket 83-26, 48 Fed. Reg. 12094 (March 23, 1983); Notice of Proposed Rulemaking, Docket 83-30, 48 Fed. Reg. 12253 (March 23, 1983); and Further Notice of Proposed Rulemaking, Docket 82-243, 48 Fed. Reg. 12267 (March 23, 1983).) Two of the reallocation proposals were initiated following petitions from leading communications firms that obviously believe the public will use these services. It is extremely difficult, however, to judge how the social worth of these services stacks up against that of other potential spectrum competitors -- either land mobile radio or some other service. Indeed, questions like these are frequently unanswerable without evidence of market demand.

Commission and spectrum users must be made aware of the fact that all use of scarce resources -- spectrum included -- is costly to society, and that these costs must be explicitly accounted for in the allocation process.

Regional Variations

A second problem of the present system comes from the increasing geographic variation in how spectrum should best be used. This fact brings into question the general policy of allocating spectrum for a single purpose on a national basis. The current allocation approach ignores the possibility of sharing in dimensions other than frequency, e.g., -- time and space. That is, the same frequency could be put to different uses at different times and/or places. It is also very likely that the demand conditions for spectrum vary according to locale. Moreover, since demand changes over time, periodic adjustments are likely to be necessary. We can be sure that the optimal spectrum allocation structure is more varied and varying than that reflected in a uniform national allocation that is set for long periods of time. For example, industry-specific allocations suited to one part of the country may be little used in others. And surely, as unanticipated uses of spectrum appear, it will sometimes be appropriate to displace older services. A mechanism that can smoothly accommodate these kinds of adjustments, preferably without politicizing them, is becoming increasingly important.

Introduction of New Technologies

Finally, while it is unlikely that technology can eliminate scarcity, it can no doubt ease spectrum pressure. The current system does not, unfortunately, facilitate the introduction of new communications techniques. Indeed, through its imposition of rigid technical standards the Commission may actually be preventing, or at least delaying, the implementation of new communications technologies. Simpler and more rapid adoption of some of these techniques could lead directly to reduced spectrum requirements. In the land mobile and broadcast auxiliary services, for example, amplitude compandored sideband (ACSB) requires less bandwidth than FM, but FM is used almost exclusively. Under the present system the process of adopting this new technique has been slow and tortuous. Other relatively new techniques also may make possible new, more valuable broadcast services: high definition television or digital audio, for example. But the development of such new services requires an entrepreneur to undergo a lengthy and arduous process of petitions and hearings that may have a highly uncertain outcome. The research and investment process is undoubtedly hampered by these circumstances. We know, for example, that it took fifteen years of FCC proceedings to bring cellular mobile radio to the stage it is today.

To summarize at this point, we see three substantial problems with the current spectrum management system. First, the Commission does not now have, and has no way of obtaining, sufficient information to independently and objectively evaluate competing claims for spectrum. Second, nationwide allocations do not recognize the regional differences in need for spectrum-consuming communications services. And third, the present system does not facilitate the introduction of new technologies that could either improve communications or reduce the need for spectrum. There exists, however, an allocation system that does not suffer from these inherent defects and that has proven reasonably effective for a wide range of resources used in our country: the market. Thus, an attractive alternative to establishing yet another radio service to meet the communications requirements of a specialized group of users, is to create a 'flexible' service wherein those most familiar with local needs -- local operators -- could adapt spectrum usage to their own circumstances.

A DECENTRALIZED RADIO SERVICE

The Commission's Office of Plans and Policy (OPP) has proposed that the Commission create a radio service that has no operational or technical rules beyond those needed to define interference limits. Entities licensed in this service would be allowed to decide on their own initiative, and in response to their own best business judgements, the types of communications offered on their authorized channels. Thus, a licensee would be free to offer a broadcast, paging, land mobile or any other service or combination of services. Licensees would also be free to adopt any available technology. The only limitation -- but a crucial one -- is that they not violate clearly specified interference rules. If the proposal is technically feasible, and OPP believes it is, it would provide a mechanism that both recognizes the variation of spectrum value across uses, time and locations and encourages licensees to act on those differences. Individual licensees would be able to tailor their use of spectrum in response to the demands of users in their communities --and adjust them as appropriate. Licensees would also have an incentive to use their assignments intensively, since they would be the beneficiaries.

Three possible issues arise in connection with the proposal. The first is the question of whether and how electrical interference can be managed. The second concerns issuing licenses and regulating users. The last is the question of (FCC) administrative cost. Each issue is discussed briefly below.

Interference Potential and Communications Usage

In order to establish a flexible service, the Commission must address several issues under the general category of interference. First, it must define licensees' interference rights and responsibilities in a manner that is applicable across a wide variety of system designs. Second, it must specify the responsibilities of mobile emitters. Third, it must establish a procedure that permits licensees to change from one system to another while still meeting their interference responsibilities. These three issues will be considered in turn.

Interference Limits: There are four primary factors that dictate the amount of interference one communications system generates to another:

1) the modulation and encoding/decoding techniques used in each system;
2) the response of the 'victim' receiver to interfering signals;
3) the power spectral density (PSD) of the desired and undesired signals;
4) the relative energies of the desired and undesired signals at the victim receiver's detector.[4]

---

4 Factors 3 and 4 are related. The energy of a signal is simply the integral across all frequencies of its PSD.

The fact that UHF-TV channels 14 through 20 are shared with land mobile services shows that there is no technical reason why different uses and technologies cannot utilize common frequencies. The significant difference between this sharing plan and sharing in a flexible service is that in the former case the factors of interference are specified by the FCC during the allocation process. In a flexible service, however, these factors would be specified by applicants. A decision would be made at the assignment stage as to whether a proposed system violated established interference criteria. The problem, therefore, is to create an administratively expedient method of making the assignment decision.

The method we advance is not really a new technique, but rather an adaptation of procedures already used by the FCC in the Low Power Television (LPTV) service. Here, the FCC specified at the allocation stage the interference factors listed above. Assignments are authorized only when an applicant demonstrates that a new station meets the specified predicted desired-to-undesired (D/U) signal strength ratios at the edge of existing stations' service areas (factor 4). One important aspect of this procedure is that interference rights are based on predicted, not actual, values. As a result, interference rights are based on a deterministic quantity. The chance of conflicts is therefore reduced significantly.

The protected contour overlap assignment methodology used in making LPTV assignments can be generalized with the following three additions. First, an assumed receiver response must be specified. Second, the desired and undesired powers and the interference criterion (D/U ratio) must be specified in a way that is independent of the relative frequency of RF carriers. Finally, non-colocated transmitters must be accounted for.

The specified receiver response will be necessarily dictated by where decentralization is introduced. If the new allocation is vacant spectrum, the choice of response characteristics can be arbitrary. On the other hand, if the new service is introduced into the UHF-TV band as we suggest later, then the specified receiver response should resemble that of a television receiver. Again, the exact choice is somewhat arbitrary and will depend on the FCC's perception of the optimum tradeoff between interference probability and new service. The important thing is that a response be defined. Its exact form is less critical to the success of the new service.

The receiver response is used to weight each spectral component in a signal's PSD so that those frequencies in the most sensitive portion of the receiver's passband are emphasized. By making several simplifying assumptions that have little effect on absolute accuracy, one can compute the <u>equivalent</u> field strength of a particular system at a given location as:

$$F = \left[\sum [S_i H(f_i)]^2\right]^{1/2} \tag{1}$$

where the summation is across all the system's transmitters and

F = equivalent field strength in microvolts/meter

S_i = field strength of i^{th} frequency transmitter in microvolts/meter[5]

$H(f_i)$ = receiver response at the i^{th} frequency

Note that the field strength of each transmitter is treated individually. Hence, the computed equivalent field strength does not depend upon the system's transmitters being colocated. The FCC uses a similar technique for calculating cochannel AM broadcast interference. For a particular location, the aggregate estimated field produced by several emitters is computed by taking the square root of the sum of the squares of the fields of each emitter (i.e., the root-sum-square, rss).

There are several ways a D/U criterion can be arrived at. It can, for example, be selected arbitrarily. In this case, licensees whose services require greater protection would have to convince their spectrum neighbors to operate at lower power levels. This approach is reasonable if the new allocation is vacant. In occupied bands, however, a better method would be to define protected contours and D/U ratios in a manner that protects the existing service. In the UHF-TV service, this approach would mean defining protected contours as either 64 dBμ (Grade B) or 74 dBμ (LPTV protected contour).[6] The D/U criterion would be between 45-50 dB, since these ratios are generally accepted as those needed to limit TV interference.[7]

In summary, the assignment procedure outlined in this section would base interference rights and responsibilities on the protected contour overlap concept introduced in the LPTV service. Contours would be found in an iterative fashion by systematically selecting a location, calculating the contour value using expression (1), and comparing this result with the specified value.

---

5 These values would be predicted using the FCC's propagation curves. See, 47 CFR Section 73.699 Figures 9 and 10.

6 Obviously, the FCC could select some lower value for a protected contour if it wished to provide a larger amount of interference protection to broadcast television stations.

7 For example, see, Television Allocations Study Organization (TASO), Engineering Aspects of Televsision Allocations, Report to the Federal Communications Commission, March 16, 1959; G.V. Waldo and W.A. Daniel, A Method of Estimating the Extent of Interference to TV Broadcast Service Caused by Low-power Transmitters Operating on VHF-TV Channels, FCC Report No. R-6306 (December 1963).

It is important to note that the interference rules in question are not meant to guarantee interference-free operation. Rather, licensees would be expected to design systems within the expected interference limits much the same way satellite systems are designed to operate with certain expected noise and interference limits.

Mobile Emitters: We believe that the transient nature, reduced range and relatively short transmitting times of mobile transmitters makes their interference potential small relative to fixed stations, and that it can be ignored for assignment purposes. But interference rights and responsibilities of a flexible service must account for mobile as well as fixed transmitters.

One way of addressing mobile interference is through the assignment of interference liability. Under this option, licensees would be liable for interference caused by mobiles transmitting outside the interference contour of associated fixed stations. While it provides users flexibility and encourages them to negotiate mutually beneficial arrangements, this option could be very time consuming and costly to enforce. In some microwave bands a slight variation of this option has been used successfully, wherein mobile emitters always operate on a secondary basis to fixed stations.[8] We recommend this approach for a flexible service. That is, mobiles interfering with fixed stations would always be responsible for resolving interference to fixed users.

Multi-Service Licensees: A flexible service would permit one licensee to offer a range of different communications services -- either simultaneously or at different times of the day. Because differing frequency/power combinations could be used, each service offering could have a different protected and interference contour. We recommend that only one system design be used for establishing protected and interference contours. Licensees could use different system architectures only if the resulting signal contours were within (or at most coincident with) those authorized.[9]

Legal/Regulatory

If flexibility were introduced into an existing service, there may be some procedural concerns about the rights of existing and/or earlier potential licensees. In particular, the Ashbacker decision might require the FCC to consider certain potential applicants who would have applied for the service had they known they could have used their authorization for many different

---

8 A good example of sharing between fixed and mobile users at microwave frequencies exists in the 2 and 7 GHz broadcast auxilary bands. Here, mobile news gathering units (ENG) share the band with studio-transmitter links (STL) and other fixed stations.

9 If all affected parties agreed, we see no reason why authorized contour values could not be exceeded.

communications purposes.[10] Of course no court has ever ruled on this specific proposal, since it has never been tried before. Hence, it is not clear whether the Ashbacker doctrine would apply to the facts here.[11] If it is ultimately determined that Ashbacker does apply, the Commission's procedural obligations could probably be met by restricting eligibility into a flexible service. That is, licensees authorized prior to a specified date would be permitted to offer only those services for which the channel had been originally allocated. Later licensees would have no operational restrictions.

We would expect that under our proposal a reasonably large number of channels will be assignable in all markets. If this is true, we do not believe any public interest goal will be served by imposing on these licensees economic, service or general technical regulation -- other than basic interference limits. Thus, we would recommend not imposing content or other service requirements on a licensee who chose to be a broadcaster, not limiting private radio services to eligible users, and forbearing from regulation if the licensee chose to operate as a common carrier.

Because the range of services that may be offered in a decentralized service is wide -- and moreover may change periodically -- we believe that the use of comparative hearings would be inappropriate. It would be difficult for example, to compare two applicants if one proposed a mobile radio service and the other a broadcast service. Therefore, to select a licensee from two or more mutually exclusive applicants, we recommend that a lottery be employed. We do recognize that the assignments in question could be used for broadcasting and that therefore minority preferences would probably be required by law. But preferences should be limited to these.

Administrative Cost

Because licensees in a decentralized service would be in a position to offer communications services that could be quite lucrative, we would expect a large number of applications. Just processing these applications could create a strain on the resources of the FCC. The agency is already experiencing severe pressure in processing LPTV and cellular applications. In both these cases the Commission and the Congress have recognized that expediting the licensing process will bring important services to the public, and that it is worth the expenditure of resources. A large number of applications for any service is an indication of the potential social value of that service. Thus, while handling a flood of applications is always a difficult management problem, it is an inevitable consequence when the Commission authorizes a new service that may significantly enhance social welfare. There are ways, of

---

10 Ashbacker Radio Corp. v. F.C.C., 326 U.S. 327 (1945). In the Ashbacker line of cases courts have ruled that when there are competing applications for new broadcast channels, the Commission must hold comparative hearings to determine the most worthy licensees.

11 In fact, the Commission has already expanded the range of services FM and TV broadcasters can offer in the SCA and teletext proceedings, without considering the Ashbacker doctrine.

course, to minimize the impact to the agency, some of which are already being employed.

Another cost component is enforcement. Costs should not be significantly higher in a decentralized service than with others. The interference cases that now require the most Commission resources are those where interference responsibilities are ill-defined. This would not be the case in the proposed service. Furthermore, with no operational and very few technical rules, the Field Bureau would not have to routinely monitor users in the service. FOB presence would be needed only to confirm that input parameters (e.g., power, antenna height and frequency) are such that the predicted outputs are proper.

SPECTRUM OPTIONS

Unlike a single-use service that could be dropped in between existing allocations, a flexible service would function best with an allocation that could be easily reconfigured. This fact suggest that the allocation chosen for testing the decentralization concept should have the following attributes. First it should be large enough to support several wideband assignments in every market. A relatively wideband assignment (e.g., 5 MHz) would support a wide range of services -- from tone-only paging to high speed data. A large per-assignment bandwidth, together with the desire to make at least three assignments per market, results in a minimum spectrum requirement of roughly 30 MHz (15 MHz for each of two adjacent markets). Second, in order to eliminate over-the-horizon propagation as an interference contributor, the allocation should be at UHF. Moreover, it should be no more than an octave higher in frequency than the existing UHF-TV band. This factor is important because we see the proposed service providing relief to the multitude of services currently vying for spectrum in and adjacent to the UHF-TV band. Finally, the ideal allocation would be unassigned. This would eliminate the procedural problems outlined in the previous section, reduce the problems associated with establishing interference limits and generally allow for a smoother implementation.

There are no frequency bands that meet all these criteria. There are, however, vacant frequencies from which a new allocation can be carved. Four possible sites for a flexible service are: the land mobile reserve bands; vacant UHF-TV channels, including those that are so-called 'taboo' channels; unoccupied spectrum currently allocated to aeronautical satellites; and new assignments in the FM broadcast band that will be available as a result of Commission action in Docket 80-90.

Any of these bands could adequately support a decentralized service. There are some differences among the options, however, that permit them to be ranked. For example, the FM assignments are less attractive than the other bands because they have less bandwidth and therefore less capacity. Land mobile operation above 1 GHz, while probably technically feasible, would be initially more expensive than on the lower frequencies. Finally, some parties estimate that land mobile users will require spectrum in addition to the land mobile reserve bands. Implementing the proposed service outside the reserve band could help provide capacity to meet future land mobile demand. At this juncture we favor testing flexibility in the UHF-TV band. It meets most of the ideal allocation criteria. Furthermore, vacant channels, even taboos, can be used for some communications purposes without significantly increasing the likelihood of additional interference. Ideally, we would include both occupied and vacant UHF-TV channels in the new service if the procedural problems outlined above can be resolved. Even if only vacant channels are considered, however, there are likely to be many UHF-TV channels in all markets that could be used by the new service. Thus, a variety of different system configurations could be possible.

CONCLUSIONS, RECOMMENDATIONS AND QUESTIONS FOR PUBLIC DISCUSSION

This paper has examined the feasibility of creating a decentralized spectrum allocations mechanism and concludes:

(1) A number of desirable results would follow the introduction of flexibility, including greater efficiency in spectrum use with the potential for more intensive spectrum use through the introduction of new communications technologies.

(2) A decentralized approach is technically possible. Today's sharing between land mobile and television in the UHF-TV band shows that some type of spectrum sharing by different services is feasible. A method of arranging interference protection has been suggested that uses the protected contour overlap concept developed in the LPTV proceeding. This method would ensure acceptably low interference levels while maintaining a great deal of operational flexibility. The arrangement presents an orderly approach to interference resolution, based on specific assignments of responsibility.

(3) Decentralization can be introduced in either an already-allocated frequency band or a new allocation. Unassigned UHF-TV channels presently appear to be the best spectrum selection for a flexible service. Its use does not present procedural problems, it represents a substantial amount of fallow spectrum, and we believe a flexible service could put much of it into operation quickly.

In sum, we believe that user flexibility is a spectrum management tool that should be tried. Many practical questions remain to be addressed and would benefit from public comment. Some of the most important are:

1. Should non-broadcast services be regulated in areas beyond interference? To the extent regulation is necessary, can the new service be regulated based on how assignments are used?
   Is the Commission free to exclude the service chosen from the comparative criteria used in mutually exclusive situations?

2. What are the correct interference criteria to use in a mixed communication usage environment? What are the relative merits of the approach outlined above? Are there other approaches that might be better? Should licensees be permitted to negotiate other interference criteria?

3. What are the appropriate mechanisms for private resolution of interference, and what role should the FCC play in providing a framework or forum for dispute resolution? For example, it is implicit in this proposal that licensees will notify the Commission of technical parameters and uses, to ensure they are consistent with our interference criteria and operational restrictions. Beyond providing this information to interested parties, should the Commission assume a larger role?

4. To what extent is equipment regulation (e.g., type acceptance) necessary, and if needed what form should it take?

5. In what part of the spectrum can the service be introduced? Does the proposal raise significant procedural inequities? To what extent can these inequities be resolved through a system of differential eligibility requirements?

national association
of broadcasters

national association
of broadcasters